Rate Loss Analysis of Transmitter Cooperation with Distributed CSIT

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***Abstract*—We consider in this work the problem of determining the number of feedback bits which should be used to quantize the channel state information (CSI) in a broadcast channel (BC) with K transmit antennas (or equivalently K single-antenna transmitters (TXs)) and K single-antenna receivers (RXs). We focus on an extension of the conventional *centralized* CSI at the TX (CSIT) model, where instead of having a single channel estimate, or quantized version, perfectly shared by all the TX antennas, each TX receives its *own* estimate of the global multi-user channel. This CSIT configuration, denoted as *distributed* CSIT, is particularly suited to model the joint transmission from TXs which are not colocated. With centralized CSIT, a very important design guideline for the feedback link was provided by Jindal [Trans. Inf. Theory 2006] by providing a sufficient feedback rate to ensure that the rate loss stays below a maximum value. In the distributed CSIT setting, additional errors occur and the design guidelines for the centralized case are no longer valid. Consequently, we obtain a new relation between the rate loss and the number of feedback bits. Interestingly, the feedback rate derived in the distributed CSIT setting is roughly $K\log_2(K)$ bits larger than its counterpart in the centralized case. This highlights the critical impact of the CSIT distributedness over the performance.**

I. Introduction

In numerous scenarios of wireless communication, the nodes are now equipped with multiple antennas in order to increase the number of degree of freedom (DoF), or prelog factor, of the transmission [1]. While a DoF larger than one can be achieved without CSIT in point-to-point systems, the exploitation of the multiple antennas at the TX to achieve a DoF larger than one in multi-user settings heavily relies on the availability of sufficiently accurate CSIT [2]–[4].

Yet, obtaining the CSIT represents a challenge in fast fading channels. Indeed, in frequency division duplexing (FDD) systems, the channel estimate has to be fed back from the RXs which inevitably introduces some delays and degradations. Therefore, a large body of publications has focused on the problem of designing efficient feedback schemes and evaluating the impact of imperfect CSIT in the multiple-input single-output (MISO) BC (See [3], [5]–[7] and reference therein). Particularly relevant to this work are [3], [4] where the average loss induced by using limited feedback compared to the average rate achieved with perfect CSIT is upper bounded.

To further improve the performance and satisfy the always growing need for higher data rates, TX cooperation is currently being considered for next generation wireless networks [8]–[10]. With perfect sharing of the user's data message and the CSI, the different TXs can be seen as a unique virtual multiple-antenna array serving all RXs, in a MISO BC fashion. Yet, this type of MISO BC differs from the conventional ones in the way feedback is obtained at the TXs. Indeed, a channel estimate is obtained initially at a RX and has to be feedback to *every* cooperating TX in order to rely on conventional transmission schemes for the MISO BC. This can either be done by direct transmission from every RX to all the TXs in a broadcast fashion, e.g., using analog feedback [4], [11], or each RX can transmit its feedback solely to a single TX, which then forwards it to the other TXs, as currently envisioned for the future LTE systems [12].

In both scenarios, the conventional assumption of having a *single* imperfect channel estimate perfectly shared to all the TXs, which we call hereafter the *centralized* CSIT configuration, appears as rather optimistic in the case of non-colocated cooperating TXs. As a consequence, we consider here a novel CSIT configuration where every TX receives its *own* channel estimate of the multi-user channel. This knowledge is then used to compute the transmit coefficients locally without additional communication between the TXs. This setting, introduced in [13]–[15] as the *distributed CSIT* scenario, opens new problems and remains little studied despite its practical relevance. In [13], a precoding algorithm is provided for the two-user case. A DoF analysis is carried out in [15] while the CSIT dissemination problem is discussed in [16].

However, the performance have so far only been evaluated in terms of DoF which, although helpful to get insights, is not adapted for practical system design. In particular, it is not known whether the feedback schemes designed in [3] for the BC with centralized CSIT remain efficient when the CSIT is distributed. Answering this question is the main goal of this work.

Specifically, our main contributions are as follows. With centralized CSIT, we derive a sufficient feedback rate to ensure that the rate loss remains below a threshold value. Although this setting has already been studied in [3], another quantization scheme was considered such that a novel approach has to be developed. Turning to the distributed CSIT configuration, we provide also a sufficient feedback rate to ensure that the rate loss remain below a threshold value. The comparison of the two feedback rates allows to evaluate the impact of the CSIT distributedness.

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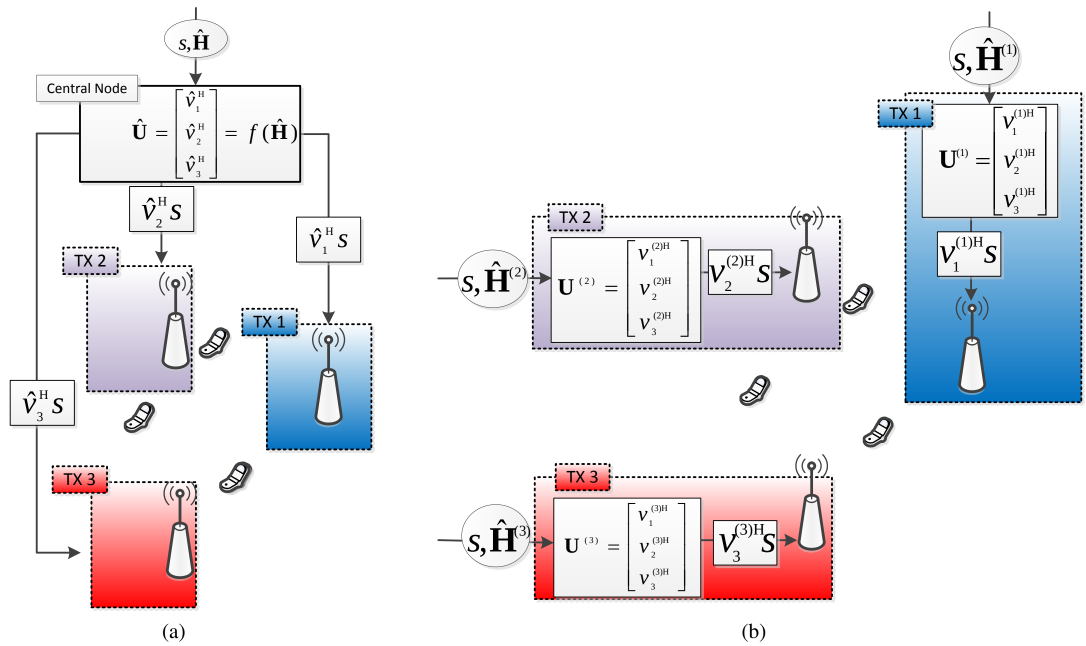

Fig. 1: The transmission with centralized CSIT/precoding is symbolically represented in Figure (a) while Figure (b) represents the distributed CSIT/precoding .

II. System Model

A. Transmission model

We consider two different CSIT models, the conventional single TX BC with *centralized CSIT* and the *distributed CSIT* BC corresponding to the joint transmission from non-colocated TXs. Both models will be detailed in Subsection II-B. Specifically, we study the transmission from K single-antenna TXs sharing the knowledge of the users' data symbols and using linear precoding to transmit to K single-antenna RXs which treat interference as noise. The channel from the K TXs to the K RXs is represented by the channel matrix $\mathbf{H}^{\mathrm{H}} \in \mathbb{C}^{K\times K}$ with its elements independent and identically distributed (i.i.d.) as $\mathcal{CN}(0,1)$ to model a uniform Rayleigh fading environment. We assume for simplicity that the pathloss is the same over all wireless links and the extension to channels with different pathloss is left for future works. The transmission is then described as

$$\begin{bmatrix} y_1 \\ \vdots \\ y_K \end{bmatrix} = \mathbf{H}^{\mathrm{H}}\boldsymbol{x} + \boldsymbol{\eta} = \begin{bmatrix} \boldsymbol{h}_1^{\mathrm{H}}\boldsymbol{x} \\ \vdots \\ \boldsymbol{h}_K^{\mathrm{H}}\boldsymbol{x} \end{bmatrix} + \begin{bmatrix} \eta_1 \\ \vdots \\ \eta_K \end{bmatrix} \tag{1}$$

where y_i is the signal received at the i-th RX, $\boldsymbol{h}_i^{\mathrm{H}} \in \mathbb{C}^{1\times K}$ the channel to the i-th RX, and $\boldsymbol{\eta} \triangleq [\eta_1, \ldots, \eta_K]^{\mathrm{T}} \in \mathbb{C}^{K\times 1}$ the normalized Gaussian noise ($\mathcal{CN}(0,1)$). We denote the average per-TX transmit power by P. We also call P the average SNR.

The transmitted signal $\boldsymbol{x} \in \mathbb{C}^{K\times 1}$ is obtained from the symbol vector $\boldsymbol{s} \triangleq [s_1, \ldots, s_K]^{\mathrm{T}} \in \mathbb{C}^{K\times 1}$ (i.i.d. $\mathcal{CN}(0,1)$) as

$$\boldsymbol{x} = \sqrt{P}\mathbf{U}\boldsymbol{s} = \sqrt{P}\begin{bmatrix} \boldsymbol{u}_1 & \ldots & \boldsymbol{u}_K \end{bmatrix} \begin{bmatrix} s_1 \\ \vdots \\ s_K \end{bmatrix} \tag{2}$$

where $\mathbf{U} \in \mathbb{C}^{K\times K}$ is the beamforming matrix and $\boldsymbol{u}_i \in \mathbb{C}^{K\times 1}$ is the beamforming vector used to transmit s_i to RX i. Our focus being on the high-SNR interference-limited regime, we are interested in canceling out the interference. Hence, we limit ourselves to studying zero-forcing (ZF) precoders. With perfect CSIT, the ZF precoder is denoted by $\mathbf{U}^\star \triangleq [\boldsymbol{u}_1^\star, \ldots, \boldsymbol{u}_K^\star]$ where

$$\boldsymbol{u}_i^\star \triangleq \left(\mathbf{I}_K - \mathbf{H}_i\left(\mathbf{H}_i^{\mathrm{H}}\mathbf{H}_i\right)^{-1}\mathbf{H}_i^{\mathrm{H}}\right)\boldsymbol{h}_i, \forall i \in \{1,\ldots,K\} \tag{3}$$

with

$$\mathbf{H}_i \triangleq \begin{bmatrix} \boldsymbol{h}_1 & \ldots & \boldsymbol{h}_{i-1} & \boldsymbol{h}_{i+1} & \ldots & \boldsymbol{h}_K \end{bmatrix}, \forall i \in \{1,\ldots,K\}. \tag{4}$$

Remark 1. The ZF beamformers $\boldsymbol{u}_i^\star$ are defined without power normalization. Yet, it can easily be seen that $\mathrm{E}[\|\boldsymbol{u}_i^\star\|^2] = 1, \forall i$ and $\mathrm{E}[\|\boldsymbol{e}_j^{\mathrm{H}}\mathbf{U}^\star\|^2] = 1, \forall j$. This means that a per-user as well as a per-TX power constraint is fulfilled on average. □

The metric of interest in this work is the sum rate averaged over the channel realizations and, in case of imperfect CSIT, the quantization error. With perfect CSI at the RX, the average rate of user i is written as

$$R_i \triangleq \mathrm{E}_{\mathcal{H},\mathcal{D}}\left[\log_2\left(1 + \frac{P|\boldsymbol{h}_i^{\mathrm{H}}\boldsymbol{u}_i|^2}{1 + \sum_{j\neq i} P|\boldsymbol{h}_i^{\mathrm{H}}\boldsymbol{u}_j|^2}\right)\right] \tag{5}$$

where the expectation $\mathrm{E}_{\mathcal{H},\mathcal{D}}[\bullet]$ denotes the expectation over the channel realizations and over the error resulting from the quantization scheme described in the following subsection.

B. Centralized versus distributed BC

Both CSIT scenarios are illustrated in Fig. 1.

1) BC with centralized CSIT: In the centralized CSIT configuration (see, e.g., [3], [4]), all the TX antennas are colocated and share *perfectly* a single imperfect channel estimate $\hat{\mathbf{H}} = [\hat{\boldsymbol{h}}_1, \ldots, \hat{\boldsymbol{h}}_K]$. We consider that the estimate $\hat{\boldsymbol{h}}_i$ of the channel of user i is obtained from the following quantization scheme which aims at minimizing the mean square error (MSE):

$$\hat{\boldsymbol{h}}_i = \operatorname*{argmin}_{\boldsymbol{w} \in \mathcal{W}_i} \|\boldsymbol{h}_i - \boldsymbol{w}\|^2, \quad \forall i \in \{1, \ldots, K\} \tag{6}$$

where $\mathcal{W}_i$ is a codebook containing 2^B elements. With random vector quantization (RVQ), the codebooks are randomly chosen and the variance of the estimation error can be shown to be equal to $\sigma^2 = C2^{-\frac{B}{K}}$ with $C > 0$ (when feeding back the channel vector without normalization) [3], [15]. For the sake of simplicity, we do not use the true distribution of the quantization error resulting from the RVQ but we instead model the quantization error as

$$\hat{\boldsymbol{h}}_i = \sqrt{1 - \sigma^2}\boldsymbol{h}_i + \sigma \boldsymbol{\delta}_i, \qquad \forall i \in \{1, \ldots, K\} \tag{7}$$

where $\boldsymbol{\delta}_i \in \mathbb{C}^{K \times 1}$ has i.i.d. $\mathcal{CN}(0, 1)$ elements and is independent of $\boldsymbol{h}_i$. We define then $\hat{\mathbf{H}}_i$ as

$$\hat{\mathbf{H}}_i \triangleq \begin{bmatrix} \hat{\boldsymbol{h}}_1 & \ldots & \hat{\boldsymbol{h}}_{i-1} & \hat{\boldsymbol{h}}_{i+1} & \ldots & \hat{\boldsymbol{h}}_K \end{bmatrix}, \forall i \in \{1, \ldots, K\}. \tag{8}$$

The ZF precoder $\mathbf{U}^{\mathrm{CCSI}} \triangleq [\boldsymbol{u}_1^{\mathrm{CCSI}}, \ldots, \boldsymbol{u}_K^{\mathrm{CCSI}}]$ is then obtained by applying (3) with the imperfect estimates $\hat{\mathbf{H}}_i$ and $\hat{\boldsymbol{h}}_i, \forall i$.

Remark 2. The commonly used Grassmannian quantization scheme chooses the vector $\boldsymbol{w}$ which maximizes $|\boldsymbol{h}_i^{\mathrm{H}} \boldsymbol{w}|$ [3], [4]. Yet, the figure of merit $|\boldsymbol{h}_i^{\mathrm{H}} \boldsymbol{w}|$ is invariant by multiplication of $\boldsymbol{w}$ by a complex unit norm number. This leads to inconsistencies between the channel estimates at the TXs which are disastrous for joint precoding in a distributed CSIT BC [15]. □

2) BC with distributed CSIT: In the BC with distributed CSIT, TX j has it own estimate of the global multi-user channel denoted by $\hat{\mathbf{H}}^{(j)} = [\hat{\boldsymbol{h}}_1^{(j)}, \ldots, \hat{\boldsymbol{h}}_K^{(j)}]$. Note that the channel estimate at every TX could potentially have a different accuracy but we study in this work the configuration where the estimates at all the TXs are of the *same* quality, i.e., they have the same statistical properties. The estimate $\boldsymbol{h}_i^{(j)}$ is obtained from the codebook $\mathcal{W}_i^{(j)}$ containing 2^B elements as

$$\hat{\boldsymbol{h}}_i^{(j)} = \operatorname*{argmin}_{\boldsymbol{w} \in \mathcal{W}_i^{(j)}} \|\boldsymbol{h}_i - \boldsymbol{w}\|^2, \qquad \forall i, j \in \{1, \ldots, K\}. \tag{9}$$

Similarly to the conventional BC described above, it holds that the variance of the estimation error is $\sigma^2 = 2^{-\frac{B}{K}}$ and we model the quantization error at TX j such that

$$\hat{\boldsymbol{h}}_i^{(j)} = \sqrt{1 - \sigma^2}\boldsymbol{h}_i + \sigma \boldsymbol{\delta}_i^{(j)}, \forall i, j \in \{1, \ldots, K\}. \tag{10}$$

where $\boldsymbol{\delta}_i^{(j)} \in \mathbb{C}^{K \times 1}$ has i.i.d. $\mathcal{CN}(0, 1)$ elements and is independent of $\boldsymbol{h}_i$. We assume furthermore that the quantization errors at the TXs are independent such that we have $\mathrm{E}[\boldsymbol{\delta}_i^{(j)} (\boldsymbol{\delta}_i^{(k)})^{\mathrm{H}}] = \delta_{ik} \mathbf{I}_K$.

Remark 3. A practical scenario where the CSIT is distributed arises when the CSI is broadcast from the RXs to the non-colocated TXs in an analog manner (analog feedback) [4], [11]. In fact, the digital quantization model used in this work is solely a model to represent the error in the CSIT due to the limited feeeback and the results can be directly extended to the case of analog feedback. In this case, the number of feedback bits can be related to the average transmit power (or bandwidth) on the feedback channel. □

The ZF precoder $\mathbf{U}^{(j)} \triangleq [\boldsymbol{u}_1^{(j)}, \ldots, \boldsymbol{u}_K^{(j)}]$ computed at TX j is then obtained from

$$\boldsymbol{u}_i^{(j)} \triangleq \left(\mathbf{I}_K - \hat{\mathbf{H}}_i^{(j)} \left((\hat{\mathbf{H}}_i^{(j)})^{\mathrm{H}} \hat{\mathbf{H}}_i^{(j)} \right)^{-1} (\hat{\mathbf{H}}_i^{(j)})^{\mathrm{H}} \right) \hat{\boldsymbol{h}}_i^{(j)} \tag{11}$$

where $\hat{\mathbf{H}}_i^{(j)}$ is defined as

$$\hat{\mathbf{H}}_i^{(j)} \triangleq \begin{bmatrix} \hat{\boldsymbol{h}}_1^{(j)} & \ldots & \hat{\boldsymbol{h}}_{i-1}^{(j)} & \hat{\boldsymbol{h}}_{i+1}^{(j)} & \ldots & \hat{\boldsymbol{h}}_K^{(j)} \end{bmatrix}. \tag{12}$$

Even though TX j computes the whole precoder $\mathbf{U}^{(j)}$, only its jth row is actually used in the transmission. Indeed, TX j controls only the jth antenna and emits $\boldsymbol{e}_j^{\mathrm{H}} \mathbf{U}^{(j)} \boldsymbol{s}$, where $\boldsymbol{e}_j$ denotes the j-th column vector of the identity matrix $\mathbf{I}_K$. The ZF precoder $\mathbf{U}^{\mathrm{DCSI}} \triangleq [\boldsymbol{u}_1^{\mathrm{DCSI}}, \ldots, \boldsymbol{u}_K^{\mathrm{DCSI}}]$ used for the transmission is then defined as

$$\boldsymbol{u}_i^{\mathrm{DCSI}} = \begin{bmatrix} \boldsymbol{e}_1^{\mathrm{T}} \boldsymbol{u}_i^{(1)} \\ \boldsymbol{e}_2^{\mathrm{T}} \boldsymbol{u}_i^{(2)} \\ \vdots \\ \boldsymbol{e}_K^{\mathrm{T}} \boldsymbol{u}_i^{(K)} \end{bmatrix}, \qquad \forall i \in \{1, \ldots, K\}. \tag{13}$$

III. Average Rate with Centralized Limited Feedback

We study in this section the feedback design in the conventional K-user MISO BC with centralized CSIT. The scaling of the number of feedback bits in terms of P is a well known result [3], [4], yet, it is obtained with a different ZF precoder and a different quantization scheme than considered here. Furthermore, we provide additional insights by studying the interference term $|\boldsymbol{h}_i^{\mathrm{H}} \boldsymbol{u}_j^{\mathrm{CCSI}}|^2$ for $i \neq j$, which represents the interference at RX i resulting from the transmission to RX j.

Proposition 1. *In the BC with centralized CSIT with* $\sigma^2 = 2^{-\frac{B}{K}}$*, there is a subspace* $\mathcal{A}_\sigma$ *of probability* 1 *such that*

$$E_{\mathcal{A}_\sigma}[|\boldsymbol{h}_i^{\mathrm{H}} \boldsymbol{u}_j^{\mathrm{CCSI}}|^2] \leq \sigma^2, \qquad \forall i \neq j \in \{1, \ldots, K\}. \tag{14}$$

Proof: See [17] for the proof. ■

The variance of the interference term $P|\boldsymbol{h}_i^{\mathrm{H}} \boldsymbol{u}_j^{\mathrm{CCSI}}|^2$ is a very important figure of merit since it represent the power of the leaked interference after ZF. Based on Proposition 1, the following sufficient CSIT allocation can be obtained.

Theorem 1. *At high SNR in the BC with centralized CSIT with* $\sigma^2 = 2^{-\frac{B}{K}}$*, the rate loss is upper bounded by* $\log_2(1 + b) + o(1)$ *bits if* $B = B^{\mathrm{CCSI}}$ *with*

$$B^{\mathrm{CCSI}} \triangleq K \log_2((K-1)P) - K \log_2(b). \tag{15}$$

Proof: See [17] for the proof. ■

It is intuitive that the distributedness of the CSIT leads to an increased amount of leaked interference since the precoding coefficients are less coordinated. However, this degradation has not yet been quantified and the feedback requirements with distributed CSIT are unknown.

IV. Average Rate with Distributed Limited CSI

The main goal of this section is to obtain a counterpart for the feedback design guideline provided in Theorem 1 in the case of distributed CSIT. Due to the distributed precoding, the distribution of the leaked interference $|\boldsymbol{h}_i^{\mathrm{H}}\boldsymbol{u}_j^{\mathrm{DCSI}}|^2$ is more complicated and cannot be easily obtained. To study this term, we start by evaluating the difference $\mathbf{U}^\star - \mathbf{U}^{(j)}$.

Proposition 2. *In the BC with distributed CSIT with $\sigma^2 = 2^{-\frac{B}{K}}$, it holds with probability one that*

$$\boldsymbol{u}_k^{(j)} = \boldsymbol{u}_k^\star + \boldsymbol{a}_k^{(j)} + o(\sigma), \qquad \forall k, j \in \{1, \dots, K\} \tag{16}$$

with

$$\mathrm{E}[\|\boldsymbol{a}_k^{(j)}\|^2] = (2K-1)\sigma^2. \tag{17}$$

Proof: See [17] for the proof. ■

As the precoding at each of the TX is the same as in the case of centralized CSIT, Proposition 2 is also valid for $\boldsymbol{u}_k^{\mathrm{CCSI}}$. We can now use Proposition 2 to state our main result.

Theorem 2. *At high SNR in the BC with distributed CSIT with $\sigma^2 = 2^{-\frac{B}{K}}$, the rate loss is upper bounded by $\log_2(1+b)+o(1)$ bits if $b > 2+2\log(K)$ and $B = B^{\mathrm{DCSI}}$ with*

$$B^{\mathrm{DCSI}} \triangleq K\log_2((2K-1)(K-1)P) - K\log_2\left(\frac{b-2-2\log(K)}{3+2\log(K)}\right). \tag{18}$$

Proof: See [17] for the proof. ■

Considering K large, it follows that the feedback rate should be roughly $K\log_2(K)$ larger when the CSIT is distributed compared to the centralized case so as to ensure that the rate loss is below the same threshold value. The second term in the right-hand side (RHS) of (18) is the reason behind the condition on b for using Theorem 2. Yet, it is believed to be an artifact from the proof and that both the restriction on the values taken by b and this additional term could be removed at the cost of the addition of a (small) constant in the expression of B^{DCSI}.

V. Simulation Results

We will verify by simulations the analytical results using 100 000 Monte-Carlo realizations of the channel and the quantization errors. First, we show in Fig. 2 the average value of the MSE $\|\boldsymbol{u}_i^{(j)} - \boldsymbol{u}_i^\star\|^2$ (resp. $\|\boldsymbol{u}_i^{\mathrm{CCSI}} - \boldsymbol{u}_i^\star\|^2$) as a function of the variance of the quantization error σ^2 when $K = 5$. Since we are interested in the high precision quantization, we consider small values of σ^2. The simulations can be seen to overlap exactly with the theoretical results from Proposition 2 as the slope of both curves is $(2K-1) = 9$.

We show then in Fig. 3 the average interference power after normalization by the average SNR P. As expected from Proposition 1, the average leaked interference are lower or equal than σ^2 in the centralized case. In contrast, it can be seen in the case of distributed CSIT that the expected leaked interference increases with σ^2 with a slope larger than one, in fact on the order of K. Comparing the feedback rates from Theorem 1 and Theorem 2, it is meaningful that the leaked interference have a different scaling in K.

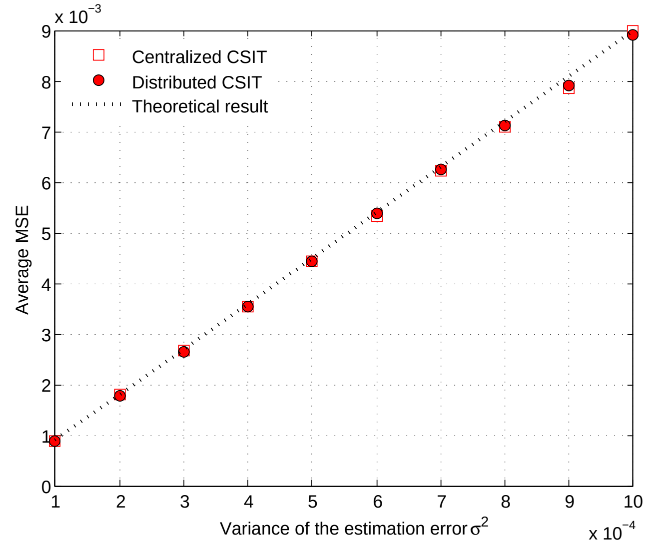

Fig. 2: Average MSE $\mathrm{E}[\|\boldsymbol{u}_i^{(j)} - \boldsymbol{u}_i^\star\|^2]$ (resp. $\mathrm{E}[\|\boldsymbol{u}_i^{\mathrm{CCSI}} - \boldsymbol{u}_i^\star\|^2]$) as a function of σ^2.

Remark 4. We see in the simulations that the average leaked interference seems to be exactly equal to σ^2 in the case of centralized CSIT. This is in fact a predictable result. Indeed, using a different modelization of the quantization error, where we have this time $\boldsymbol{h}_i = \hat{\boldsymbol{h}}_i + \sigma\boldsymbol{\delta}_i$ with the quantization error $\boldsymbol{\delta}_i$ independent of the estimate $\hat{\boldsymbol{h}}_i$, it is easily shown that the expectation of the inner product is exactly equal to σ^2. □

Finally, the average rate per user in the BC with distributed CSIT is shown in terms of the average SNR in Fig. 4 for $K = 15$. We show the average rate achieved when perfect CSIT is available at the TXs and we compare it to the rate achieved with limited feedback using different number of feedback bits. We consider a digital feedback with B^{DCSI} given in (18) for $b = 5+4\log(K)$ and with B^{CCSI} given in (15) with the same b. This choice of b has for consequence that the second term of the expression of B^{DCSI} given in (18) cancels out.

We can observe that the rate loss obtained using B^{DCSI} feedback bits leads to a rate loss smaller than 1 bit while the upper bound for the rate loss per user was $\log_2(1+b) = 4.0732$ bits. This is in agreement with our conjecture that the second term of the feedback rate in Theorem 2 could be replaced by a (small) constant. Indeed, according to this conjecture, using $b = 5+4\log(K)$ in the feedback rate of Theorem 2 corresponds approximately to using $b = 1$ in our conjectured feedback rate which gives a rate loss of 1 bit.

In contrast, the limited feedback scheme using B^{CCSI} leads to a rate loss of 8.2 bits, which is 4.1 bits higher than the

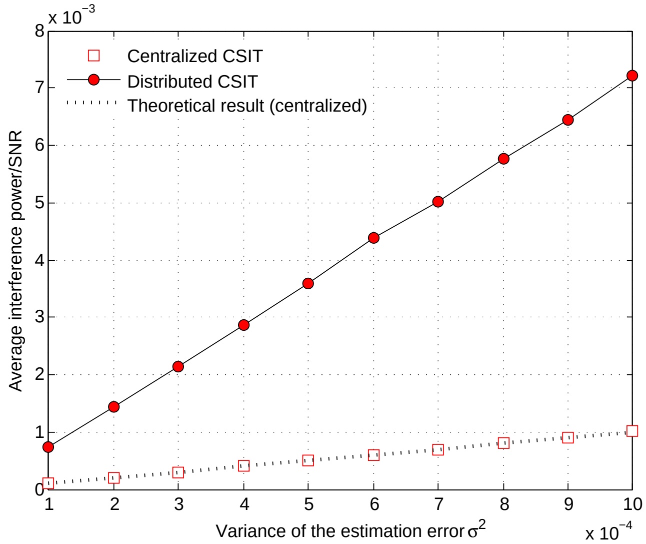

Fig. 3: Average value of the leaked interference after normalization by the average SNR $\mathrm{E}[|\boldsymbol{h}_i^{\mathrm{H}}\boldsymbol{u}_j^{\mathrm{DCSI}}|^2]$ (resp. $\mathrm{E}[|\boldsymbol{h}_i^{\mathrm{H}}\boldsymbol{u}_j^{\mathrm{CCSI}}|^2]$) as a function of σ^2.

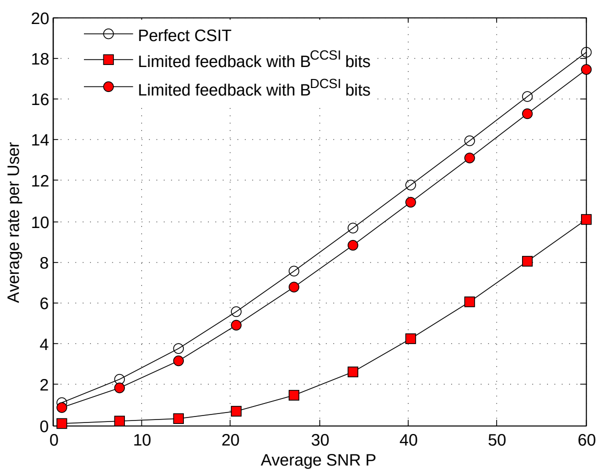

Fig. 4: Average rate per user versus the average SNR P in the distributed CSIT scenario with $B = B^{\mathrm{CCSI}}$ (Cf. (15)) and $B = B^{\mathrm{DCSI}}$ (Cf. (18)) for $b = 5 + 4\log(K)$, compared to the average rate per user obtained with perfect CSIT.

threshold value. This means that the number of feedback bits used is too small to satisfy the condition on the rate loss: Assuming that the CSIT is centralized, when it is in fact distributed, leads to an important performance degradation.

VI. Discussion of the Results

Considering a distributed CSIT configuration where every TX has its own channel estimate and cooperates with the other TXs without any additional exchange of information, we have derived a sufficient feedback rate which ensures that the rate loss compared to the transmission with perfect CSIT remains below a threshold value. Interestingly, the expression obtained in Theorem 2 can be seen to increase more quickly with the number of users K than its counterpart in the BC with "centralized" CSIT in Theorem 1. Hence, the CSI discrepancies resulting from the distributedness of the CSIT can lead to important performance degradations, if not taken into account in the feedback design. An upper-bound for the rate loss for a given amount of feedback has been derived and the derivation of a lower-bound is the focus of undergoing research. However, the statistical distribution of the interference with distributed CSIT makes this problem relatively intricate. Furthermore, the extension of the analysis to scenarios with different pathloss between every TX and every RX represents a challenging and interesting research problem.

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