Net DoF analysis for the K-user MISO IC with outdated and imperfect channel feedback

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***Abstract*—The net degrees of freedom (DoF) describe the efficiency of a protocol in terms of network throughput at high signal-to-noise ratio (SNR). As the conventional DoF, it captures how the achievable user data rate grows as SNR increases, but also takes into account the limited channel state information at the transmitters (CSIT), and required protocol overheads. This work studies the impact of short channel coherence time for the *K*-user Multiple-Input Single-Output (MISO) Interference Channel (IC) by means of the net DoF. We propose different protocols, combining techniques that exploit different levels of CSIT (delayed, current or no CSIT), and elucidating which performs better as a function of user mobility.**

I. INTRODUCTION

Bandwidth scarcity pushes towards more efficient networks for next generation wireless systems, where simmultaneous transmissions are allowed, but following some kind of interference management procedure. For example, the interference channel (IC) depicted in Fig. 1 models the downlink of a cellular system, where K transmitter-receiver pairs are scheduled on the same time-frequency resource block. An intuitive way of evaluating the impact of different conditions, e.g. channel feedback delay or antenna setting, is by means of the degrees of freedom (DoF) [1] of the channel, representing the slope of channel capacity for high signal-to-noise ratio (SNR).

Promising advances have been recently made in this direction by exploiting zero-forcing (ZF) and interference alignment concepts [1]. However, it is assumed perfect accuracy on the channel state information available at the transmitters side (CSIT), being not realistic in practical scenarios in general. For example, in frequency-division duplexing systems (FDD), receivers estimate the channels after a training period, and then report it with some delay. The worst case arises when this time lag is longer than the channel coherence time, since the channel has completely changed, and channel feedback results completely outdated. Recently, Maddah Ali et al. [2] have introduced a new framework for the broadcast channel (BC) where this knowledge is still useful (delayed CSIT). Their scheme works independently of the delay by aligning the interference signals along the space-time domain. Such ideas have been extended to the IC for some special cases, e.g. the MISO case in [3], where transmitters have $M \geq K$ antennas and communicate with single-antenna receivers. Notice that either for current or delayed CSIT, having more than K antennas at the transmitters does not increase the DoF.

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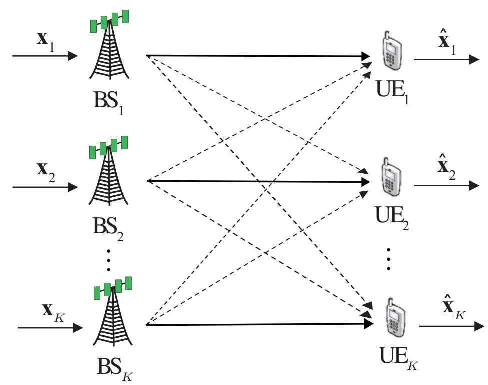

Fig. 1. The K-user MISO IC. Each base station BS$_i$, equipped with K antennas, has a message for each single-antenna user equipment UE$_i$.

Quality of feedback and incurred overheads can severely degrade the gains promised by MIMO techniques. First, the quality of the CSI depends on resources dedicated for feedback, either for digital or analog feedback, see [4] and [5], respectively. Both works derive the required accuracy in the CSIT for getting the theoretical DoF. Second, CSI-based strategies assume each user equipment (UE) knows some particular channels observed at other UEs. We denote by *dedicated training* the procedure for acquiring the required CSI at the receivers (CSIR). Note that both types of training and channel feedback involve overheads affecting the net throughput.

The *net DoF* metric [6] accounts for delays in feedback, its associated quality, along with the required overheads, which are neglected on conventional DoF. Although having more CSI at the nodes one would expect that higher data rates are feasible in the system, CSI provision may be too costly depending on the coherence time value, i.e. user mobility. The net DoF for the K-user MISO BC as a function of the channel coherence time were studied in [6] and [7]. The former analyzes the net DoF assuming finite digital feedback, while [7] assumes perfect analog feedback, thus no feedback errors are considered. Moreover, not all dedicated training overheads are accounted in these two works. The main conclusion of these works is that delayed CSIT-based strategies are useful for medium-high user mobility (short channel coherence time), providing a better trade-off by exploiting the CSI at a reduced cost. The present work studies if the conclusion for the BC, where there are K channels to be estimated, applies also for the IC, where the number of links becomes K^2, and there is no transmitter cooperation. The main contributions are:

- We propose a a common framework for dealing with feedback quality regardless of the feedback procedure.

- The net DoF of 4 protocols are derived when applied to the K-user MISO IC as a function of channel coherence time and the quality of channel feedback. They are constructed on the basis of three low complexity strategies designed for the cases of current, delayed, and no CSIT. Results show that the net DoF decrease piecewise linearly with user velocity.
- The CSIR distribution method proposed in [7] for the BC is extended to the IC.
- Beyond theoretical net DoF analysis, we provide net sum-rate results for the system working at finite SNR and finite feedback quality.

Notation: Row vectors are underlined ($\underline{\mathbf{x}}$), $\mathbb{1}$ is the indicator function, $(a)^+ = \max(a, 0)$, and $a \wedge b = \min(a, b)$.

II. System Model

In a K-user MISO IC, K transmitter/receiver pairs are simmultaneously scheduled, see Fig. 1. Each base station (BS) is equipped with K antennas, a maximum transmission power P_{DL}, and its associated to a single-antenna moving UE with transmission power P_{UL}.

A. Downlink Transmission

Each BS_i delivers b symbols to UE_i along T_{TX} time slots. All symbols are transmitted during all time slots. The signal received at UE_j during *symbol slot* s is described by:

$$y_j^{(s)} = \sum_{i=1}^{K} \underline{\mathbf{h}}_{j,i}^{(s)} \mathbf{V}_i^{(s)} \mathbf{x}_i + n_j^{(s)}, \quad (1)$$

where $\mathbf{x}_i \in \mathbb{C}^{b\times 1}$ contains the intended symbols to UE_i, $n_j^{(s)}$ is the $\mathcal{CN}(0, 1)$ additive white Gaussian noise term, and $\mathbf{V}_i^{(s)} \in \mathbb{C}^{K\times b}$ is the precoding matrix used by BS_i during time slot s, ensuring the transmit power constraint

$$\text{tr}\big(\big(\mathbf{V}_i^{(s)}\big)\big(\mathbf{V}_i^{(s)}\big)^H\big) \leq P_{\text{DL}}, \quad (2)$$

where P_{DL} also represents the downlink SNR since unit-power noise and channels are assumed. Finally, $h_{j,i}^{(s)}(a) \sim \mathcal{CN}(0, 1)$, the ath element of $\underline{\mathbf{h}}_{j,i}^{(s)} \in \mathbb{C}^{1\times K}$, denotes the gain from antenna a of BS_i to UE_j during time slot s. It is assumed that UEs are moving approximattely at constant velocity $v = \frac{\lambda}{T_c}$, where λ is a constant. Then, channels are i.i.d. across *time slots* and constant within each *time slot* of duration T_c *symbol slots*.

The communication lasts for T symbol slots, after which all b transmitted symbols can reliably be decoded. Let R denote the sum-rate (SR) of the network. Then, the *data DoF*, i.e. number of delivered signaling dimensions, write as

$$d = \lim_{P_{\text{DL}}\to\infty} \frac{R}{\log_2 P_{\text{DL}}}. \quad (3)$$

B. Channel Acquisition

We assume perfect quality CSIR, obtained either by means of training or dedicated training. For simplicity those phases are developed in a TDMA fashion, thus there is no interference. On the other hand, two finite quality procedures are envisioned in the literature for channel feedback (FB) from UE to BSs. Note that since MISO channels are vectors, they are described by its magnitude and direction. However, only direction but not magnitude is required for the strategies considered in this work, thus only $K-1$ coefficients per channel will be sent.

1) Digital feedback (DFB): In this case, the CSI is quantized into B bits and then transmitted through an error-free control channel. The quantization is performed using a quantization codebook available at all nodes, containing 2^B unitary Gaussian vectors. Note that since the phase of a Gaussian vector is uniformly distributed, for a sufficiently high B the vectors are uniformly distributed along the unit sphere.

Each channel vector is quantized to the closest codeword in terms of the chordal distance. The total number of bits to be reported per UE is equal to $B \times (K-1)$, where the latter corresponds to the number of channels to be reported, whichever the precoding scheme.

Finite number of bits B yields imperfect channel knowledge, thus not all interference can be removed and residual interference penalizes the system performance. [4] analyzed the DoF for imperfect CSIT, concluding that exact CSIT DoF cannot be attained unless B scales with P_{DL} in dB.

2) Analog feedback (AFB): In this alternative way of feed back, UEs transmit predefined pilot sequences precoded by channel estimations during

$$T_{\text{FB}} = \gamma \cdot K\,(K-1)^2 \quad (4)$$

symbol slots, see [5] for details, where γ is the number of repetitions of the channel feedback transmission, which allows to improve the feedback quality by coherently combining repetitions of the same uplink transmission. The accuracy of channel feedback, and thus achieved DoF, also depends on the maximum UE transmission power P_{UL} and the noise power. Specifically, the authors of [5] showed that exact CSIT DoF cannot be reached unless P_{UL} scales with P_{DL}.

3) A common framework: Feedback rate and DoF describe the resources consumed, and they are defined as follows:

$$F = \begin{cases} \frac{B}{T}, & \text{DFB} \\ \frac{T_{\text{FB}}}{T}\log_2 P_{\text{UL}}, & \text{AFB} \end{cases}, \; d_{\text{FB}} = \lim_{P_{\text{DL}}\to\infty} \frac{F}{\log_2 P_{\text{DL}}}. \quad (5)$$

Let $\hat{\underline{\mathbf{h}}}_{j,i}^{(s)}$ denote the estimation available at the BSs for the exact channel $\underline{\mathbf{h}}_{j,i}^{(s)}$. Then, the feedback error is given by

$$\tilde{\underline{\mathbf{h}}}_{j,i}^{(s)} = \underline{\mathbf{h}}_{j,i}^{(s)} - \hat{\underline{\mathbf{h}}}_{j,i}^{(s)} \sim \mathcal{CN}\big(0, P_{\text{DL}}^{-\epsilon}\big) \quad (6)$$

independently of which feedback method is used, where the exponent $\epsilon \in [0, 1]$ describes feedback quality. Hence, the quality is determined by the decaying rate of the error variance w.r.t. P_{DL}: $\epsilon = 1$ implies exact CSIT, whereas $\epsilon = 0$ entails completely inaccurate CSIT, equivalent to having no CSIT. Based on [4][5], we found that ϵ writes for each method as:

$$\epsilon = \begin{cases} \frac{B}{\big(K-1\big)\log_2 P_{\text{DL}}}, & \text{DFB} \\ \alpha + \beta, & \text{AFB} \end{cases} \quad (7)$$

with $\alpha = \frac{\log P_{\text{UL}}}{\log P_{\text{DL}}}$, $\beta = \frac{\log \gamma}{\log P_{\text{DL}}}$. Notice that finite number of repetitions γ does not enhance the DoF of the system, since $\beta \to 0$ when $P_{\text{DL}} \to \infty$. But, it might be useful for improving the sum-rate at low and medium SNR values.

III. Net DoF Analysis

The four protocols depicted in Fig. 2 are studied in terms of net DoF [6] and net sum-rate:

$$d_{\mathrm{N}} = d - d_{\mathrm{FB}} = \lim_{P_{\mathrm{DL}}\to\infty} \frac{R_{\mathrm{N}}}{\log_2 P_{\mathrm{DL}}}, \quad , \quad R_{\mathrm{N}} = R - F, \quad (8)$$

where the resources devoted to carry out the channel acquisition at both sides are accounted for each case.

A. TDMA (no CSIT)

The protocol frame is shown in Fig. 2-a. The training period lasts for one symbol slot since only one antenna is used at the BS, and no channel feedback is required. Hence

$$d_{\mathrm{N}}^{(\mathrm{TDMA})} = \frac{T_{\mathrm{c}} - 1}{T_{\mathrm{c}}} = 1 - \frac{1}{T_{\mathrm{c}}}. \quad (9)$$

B. ZF-TDMA (current and no CSIT)

In this case the downlink channel usage is divided in 4 parts (see Fig. 2-b). First, a pilot-based training is carried out during K^2 symbol slots since all antennas of all BSs are to be used. Next, feedback is reported and BSs must wait for T_{w} symbol slots. One possibility is to exploit this *dead time* to transmit without current CSIT, e.g. TDMA. As far as the authors know, the idea of using the dead times of ZF was first introduced in [8] for the BC. In this case, this adds $T_{\mathrm{w}} \wedge \left(T_{\mathrm{c}} - K^2\right)^+$ DoF per block to the net DoF count.

After the feedback waiting time, each BS computes the ZF precoder. However, UEs have only local knowledge and cannot compute it. Therefore, one symbol slot of dedicated training is scheduled, providing each UE_i the product $\underline{\mathbf{h}}_{i,i}^{(s)} \cdot \mathbf{V}_i^{(s)}$.

The data DoF achieved by ZF-TDMA are given by

$$d^{(\text{ZF-TDMA})} = d_1 + d_2, \quad (10)$$

$$d_1 = \epsilon \cdot \left(1 - \frac{K^2 + 1 + T_{\mathrm{w}}}{T_{\mathrm{c}}}\right)^+, \; d_2 = \frac{T_{\mathrm{w}} \wedge \left(T_{\mathrm{c}} - K^2\right)^+}{T_{\mathrm{c}}},$$

where d_1 are the DoF achieved by the ZF part, taking into account that the DoF with imperfect current CSIT collapse to $K \cdot \epsilon$ [4], whereas d_2 are the DoF achieved by the TDMA part. On the other hand, $K(K-1)$ channels are reported, thus the feedback DoF are given by

$$d_{\mathrm{FB}}^{(\text{ZF-TDMA})} = \gamma \cdot \alpha \cdot \frac{K(K-1)^2}{T_{\mathrm{c}}} \mathbb{1}(T_{\mathrm{c}} > K^2 + 1 + T_{\mathrm{w}}). \quad (11)$$

The net DoF expression follows from combining (8), (10)-(11).

C. TDMA-pairs scheme (delayed CSIT)

1) Review: For the K-user MISO IC with delayed CSIT, a low complexity two-phase precoding scheme was proposed by Torrellas et al. in [3], delivering K symbols per user after $T_{\mathrm{TX}}^{(\mathrm{TP})} = K + \binom{K}{2}$ slots, thus achieving $\frac{2}{K+1}$ DoF per user. We denote it as the time division multiple access (TDMA)-pairs (TP) scheme, since the first phase is carried out in a TDMA fashion, whereas only a pair of transmitters are active per slot during the second phase. According to this, the signal observed by UE_j during the ith first phase slot writes as

$$y_j^{(i)} = \underline{\mathbf{h}}_{j,i}^{(i)} \mathbf{V}_i^{(i)} \mathbf{x}_i + n_j^{(i)}. \quad (12)$$

Hence, after the first phase UE_i obtains one linear combination of desired symbols, and a linear combination of *overheard interference* (OHI) containing the symbols of each $\mathrm{UE}_j, j \neq i$, which would be useful if delivered to UE_i.

During the second phase only pairs of users are active (i.e. two pairs BS-UE), while the rest remain inactive. Each BS retransmits only using one antenna the OHI generated during the first phase at the other active UE. Then, assuming slot s is dedicated to users i and j, UE_j obtains

$$y_j^{(s)} = h_{j,j}^{(s)}(1)\, \underline{\mathbf{h}}_{i,j}^{(j)} \mathbf{V}_j^{(j)} \mathbf{x}_j + h_{j,i}^{(s)}(1)\, \underline{\mathbf{h}}_{j,i}^{(i)} \mathbf{V}_i^{(i)} \mathbf{x}_i + n_j^{(s)},$$

where it can be seen that the interference can be removed using the first phase OHI, see (12), since they are *aligned*. Repeating this procedure for each pair, K interference-free and independent linear combinations of desired symbols are provided to each UE, thus all symbols can be linearly decoded.

2) Net DoF derivation: In addition to the time slots for data transmission, and some extra blocks are used for dedicated training (see Fig. 2-d). Each of these time slots may occur separately in time. Data transmission blocks are in turn divided in two parts: data transmission and downlink training. The training during the first phase lasts for K symbol slots, since only one BS is active at a time and has K antennas. In contrast, during the second phase there are two simmultaneously active BSs but using only one antenna each. Hence, training consumes only 2 symbol slots, and the $K-2$ symbol slots exceed and can be used to transmit dedicated training coefficients, as will be explained later.

For the proper decoding of desired signals, every UE needs all the CSI $\hat{\underline{\mathbf{h}}}_{i,j}^{(j)}, \forall i \neq j$ used by its associated BS for precoding, and the direct channel $\underline{\mathbf{h}}_{j,j}^{(j)}$, although the latter is already known thanks to the downlink training phase. The aim of the dedicated training period is to deliver to every UE the remaining required CSI. In case of transmitting one channel coefficient at a time, it would take $K^2(K-1)$ symbol slots to deliver the $K(K-1)$ channels. Next, we show that when using digital feedback the number of symbol slots required for dedicated training can be halved. Inspired by [8] and the TP scheme, we exploit the same concept of overheard interference (namely overheard CSI) to transmit the required CSIR. Our approach will be described by means of an example, and later we specify how it is used with the proposed protocol frame.

Consider the channel $\hat{\underline{\mathbf{h}}}_{1,2}^{(2)}$ $\left(\hat{\underline{\mathbf{h}}}_{2,1}^{(1)}\right)$, available at UE_1 (UE_2) thanks to the downlink training phase, and desired at UE_2 (UE_1). Assume that $\hat{\underline{\mathbf{h}}}_{1,2}^{(2)}$, $\hat{\underline{\mathbf{h}}}_{2,1}^{(1)}$ are transmitted at a time from BS_1 and BS_2, respectively. Therefore, each UE obtains a linear combination of $\left\{\hat{\underline{\mathbf{h}}}_{1,2}^{(2)}, \hat{\underline{\mathbf{h}}}_{2,1}^{(1)}\right\}$ and the required non-local CSI may be obtained by using the previous available local CSI. Note that this approach is only reliable if receivers have access to the same CSI as transmitters, which is the case of digital, but not analog feedback. Now, since $\binom{K}{2} \cdot (K-2)$ symbol slots are saved along the whole second phase, some extra symbol slots are required for dedicated training, written in general as

$$T_{\mathrm{DT}}^{(\mathrm{TP})} = \binom{K}{2} \cdot \left(\mu K - K + 2\right), \quad (13)$$

with $\mu = 1, 2$ for digital or analog feedback, respectively. The duration of this process can be longer than the coherence time.

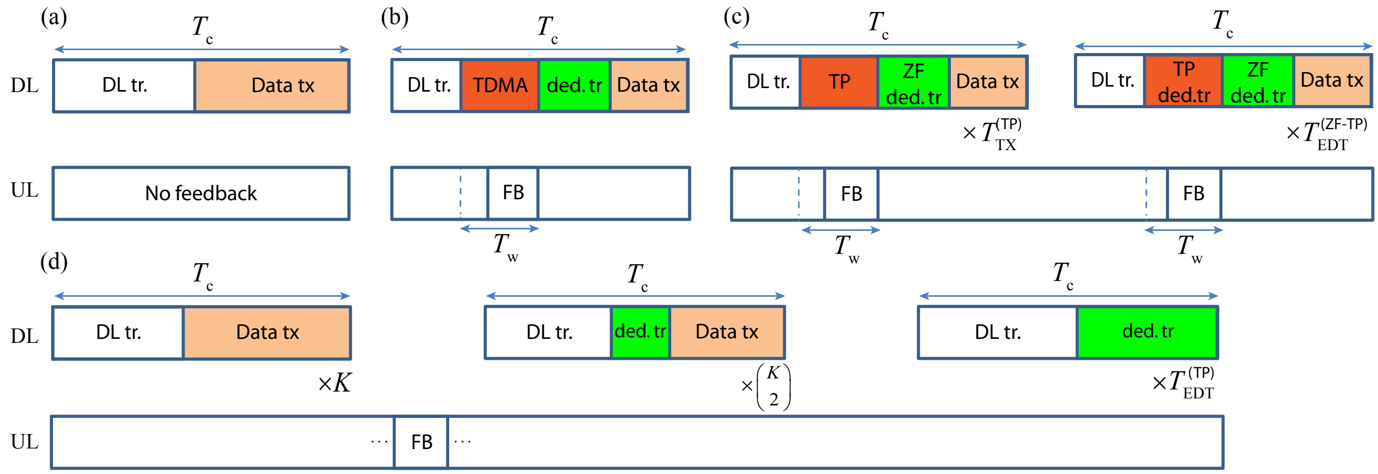

Fig. 2. Protocol frames for the downlink and uplink of (a) TDMA, (b) ZF-TDMA, (c) ZF-TP, and (d) pure TP.

Then, since each block requires downlink training, the number of blocks for completing the dedicated training process is

$$T_{\rm EDT}^{\rm (TP)} = \left\lceil \frac{T_{\rm DT}^{\rm (TP)}}{T_{\rm c} - K} \right\rceil, \tag{14}$$

assuming K symbol slots for downlink training per block. Then, data and feedback DoF are written as

$$d^{\rm (TP)} = K \frac{(T_{\rm c} - K)\,(1+\epsilon\,(K-1))}{T_{\rm c} \cdot T_{\rm TX}^{\rm (TP)} + K \cdot T_{\rm EDT}^{\rm (TP)} + T_{\rm DT}^{\rm (TP)}}, \tag{15}$$

$$d_{\rm FB}^{\rm (TP)} = \gamma \cdot \alpha \frac{K\,(K-1)^2}{T_{\rm c} \cdot T_{\rm TX}^{\rm (TP)} + K \cdot T_{\rm EDT}^{\rm (TP)} + T_{\rm DT}^{\rm (TP)}}, \tag{16}$$

where we use that $K\frac{1+\epsilon(K-1)}{K+\binom{K}{2}}$ DoF are achieved with imperfect delayed CSIT [3]. Note that the same feedback as ZF-TDMA is required but at a reduced rate, thus reducing also the feedback DoF.

D. ZF-TP (current and delayed CSIT)

Similarly to ZF-TDMA (Section III-B), and ZF-MAT in [8], the TP scheme may be carried out during the dead times of ZF, since it does not require current CSIT, see its protocol frame in Fig. 2-c. The number of blocks for dedicated training are

$$T_{\rm EDT}^{\rm (ZF\text{-}TP)} = \left\lceil \binom{K}{2} \frac{K}{T_{\rm w} \wedge \left(T_{\rm c} - K^2\right)} \right\rceil, \tag{17}$$

Since each block offers $T_{\rm w}$ symbol slots for ZF, we have

$$d^{\rm (ZF\text{-}TP)} = d_1 + K \frac{T_{\rm w} \wedge \left(T_{\rm c} - K^2\right)^+}{T_{\rm c}} \frac{1+\epsilon\,(K-1)}{T_{\rm TX}^{\rm (TP)} + T_{\rm EDT}^{\rm (ZF\text{-}TP)}}.$$

with $d_{\rm FB}^{\rm (ZF\text{-}TP)} = d_{\rm FB}^{\rm (ZF\text{-}TDMA)}$ and d_1 given in (11).

IV. Simulation Results

The protocols have been evaluated in terms of net DoF and net sum-rate assuming LTE-based system parameters: carrier frequency $f_{\rm c} = 3.5$ GHz, $T_{\rm s} = \frac{1}{168}$ msec (see Section IV-E in [6] for details), and feedback delay equal to an LTE frame, i.e. 10 msec, equivalent to $T_{\rm w} = 1680$.

A. Net DoF

Fig. 3 depicts net DoF as a function of UE velocity $v = \frac{c}{4 f_{\rm c}\, T_{\rm c}\, T_{\rm s}}$. Those results assume $K = 4$ users, $\epsilon = 0.7$, and $\gamma = 1$, since repeating the feedback transmission cannot improve the net DoF. This translates with, for instance $P_{\rm DL} = 20$ dB, to $B \approx 14$ bits or $P_{\rm UL} = 14$ dB. Note that digital presents gains w.r.t. analog feedback when using the TP scheme, as explained in Section III-C.

Two regions are clearly observed, separated at approximately $v_{\rm T} = \frac{\lambda}{T_{\rm w}} \approx 7.5$ km/h. This threshold corresponds to the velocity where coherence time is comparable to the delay, since in such a case ZF is no more reliable and the net DoF are severely reduced. ZF-TP performs the best (closely followed by ZF-TDMA) below $v_{\rm T}$. Otherwise, pure TP exhibits the best performance. Notice that ZF-TP is outperformed by ZF-TDMA for $v > 60$ km/h. This is essentially because the former transmits an excessive amount of feedback whereas the latter reduces to simply TDMA since ZF cannot be done.

Finally, notice that the net DoF for any protocol are well approximated as piecewise linear functions, whose cut at the y-axis corresponds to the conventional DoF with imperfect CSIT. Due to space limitation, the derivation of those linear functions is avoided, although it can be handled straightforwardly by simple approximations on the net DoF expressions.

B. Net Sum Rate

Simulation results depicted in Fig. 4 allow to elucidate when previous conclusions apply to finite SNR and finite feedback quality. Results are averaged over 200 channel realizations. Transmissions are repeated such that all protocols employ the same number of channels. The same system parameters as in Section IV-A are used, but now we fix $B = 14$ bits, $P_{\rm UL} = 14$ dB for all SNR values, $v = 5$ km/h, and evaluate two possible number of feedback repetitions $\gamma = \{1, 10\}$.

In this case, ZF is reliable, since $v < v_{\rm T}$. Therefore, ZF-TP (closely followed by ZF-TDMA) provides the best net sum-rate for low-moderate SNR values ($P_{\rm DL} < 40$ dB). However, both ZF-based protocols decay severely as $P_{\rm DL}$ increases, since the feedback error cannot be bounded using finite B or $P_{\rm UL}$. For

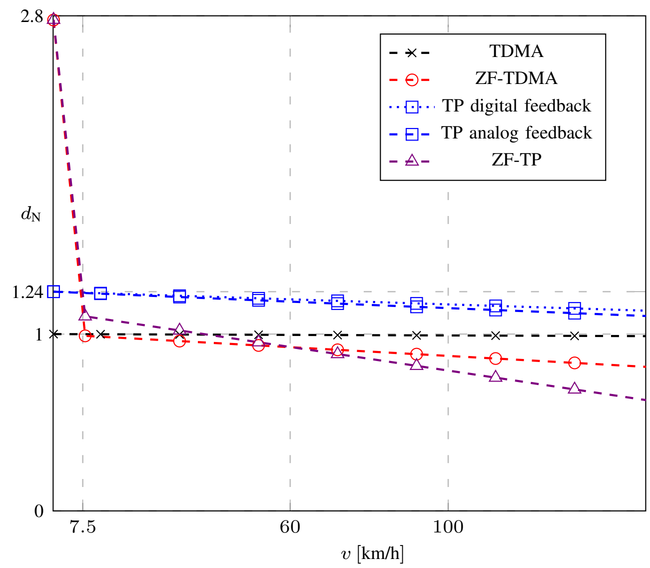

Fig. 3. Net DoF vs. UE velocity for $\epsilon = 0.7$, $K = 4$ users and $T_w = 1680$. Net DoF are independent of the feedback method except for the TP protocol. $\gamma = 1$ is set since additional repetitions do not provide net DoF gains.

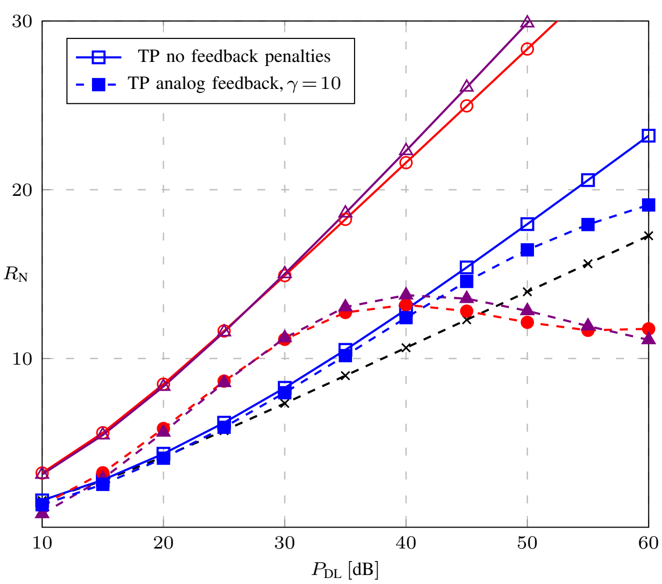

Fig. 4. Net sum-rate [bps/Hz] vs. SNR for $v = 5$ km/h. Colors and markers identify protocols as in Fig. 3. Line styles denote different feedback settings, with $B = 14$ bits used for digital, and $P_{UL} = 14$ dB for analog feedback.

analog feedback, this might be partially solved by increasing the number of repetitions γ, i.e. improving feedback quality at the cost of feedback load. This is beneficial for high SNR, but not for low SNR, where such amount of feedback penalizes system performance. Consequently, we conclude that there is a trade-off between γ and P_{DL}. In contrast, the TP protocol provides lower performance for the low-medium SNR regime, but it is more resilient to feedback quality, and becomes the best protocol in terms of net sum-rate for $P_{DL} > 40$ dB.

V. Conclusions

The net DoF for the K-user MISO IC have been studied as a function of channel coherence time and the quality of feedback. This allows to elucidate which type of transmission schemes (based on current/delayed/no CSIT) performs better depending on user mobility. We have proposed different protocols and derived its net DoF, as well as a method for enhanced feedback transfer, and a common framework for net DoF analysis regardless of the feedback method. Protocols using the TP scheme, which only requires delayed CSIT, perform the best for any level of mobility. For low mobility, combining ZF and TP (current and delayed CSIT) provides the best net DoF value. Otherwise, the TP protocol (pure delayed CSIT) achieves the best performance. Moreover, we found that the net DoF are piece-wise linear w.r.t. user velocity.

Simulations show that the net sum-rate of the system is severely penalized with finite SNR and finite feedback quality. This may be partially bypassed in case of analog feedback by repeating the uplink transmission, which provides net sum-rate gains at moderate SNR. Our results also show that the TP protocol is more resilient to imperfect CSIT, and thus more suitable for scenarios with high SNR conditions.

Possible lines of future work include deriving the net DoF for other schemes, such as [9][10]. These schemes are designed for intermediate scenarios where channel feedback is acquired before channel changes, i.e. $T_{FB} < T_c$, and combine both current and delayed CSIT.

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