Joint Beamforming and Power Optimization for D2D Underlaying Cellular Networks

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***Abstract*—This paper studies the optimal joint beamforming and power control strategy for device-to-device (D2D) communication underlaying multiuser multiple-input multiple-output cellular networks. We consider multiple antennas at the base station (BS) and a single antenna at each cellular user (CU), D2D transmitter (DT) and D2D receiver (DR). We aim to minimize the total transmission power of the system by jointly designing the transmit beamforming at the BS and the transmit powers for both BS and DTs, while satisfying the signal-to-interference-plus-noise ratio based quality-of-service constraints for both CUs and DRs. Due to the non-convex nature of the problem, we apply the semidefinite relaxation technique to find the optimal solution, which always satisfies the rank-one constraint. We also investigate three sub-optimal fixed beamforming schemes: zero-forcing (ZF), regularized ZF and hybrid maximum ratio transmission-ZF, where the focus is to minimize the total transmission power while reducing complexity. When perfect channel information is not available, we propose a robust transmit power minimization strategy with ZF beamforming which only requires limited feedback based channel direction information at the BS. Finally, computer simulation results are presented to demonstrate the effectiveness of the proposed schemes.**

***Index Terms*—Beamforming, multiple-input multiple-output (MIMO), D2D, semidefinite relaxation (SDR).**

I. Introduction

In a device-to-device (D2D) wireless communication, two neighbouring or nearby users are allowed to communicate directly with limited or no participation of the cellular base station (BS) [1]. D2D has the potential to support high-data rate communication and low latency applications effectively [2]. D2D has been shown to improve resource utilization, spectral efficiency, energy efficiency [3] and cellular coverage [4] of wireless networks. Due to these attractive features, it has been envisioned as a key technology for the fifth generation (5G) communication systems. Some applications of D2D include online gaming, video streaming and multimedia downloading [5].

In D2D enabled cellular systems, depending on the spectrum sharing, D2D can be divided into two main types: in-band and out-of-band. D2D uses the same licensed cellular band in in-band, whereas in out-of-band D2D operates in a different band than the cellular band. In-band D2D can be further classified into two categories: overlay and underlay. Overlay refers to the approach where cellular users (CUs) and D2D receivers (DRs) use orthogonal spectrum resources, whereas underlay refers to the framework where both CUs and DRs utilize the same time/frequency resources. This study focuses on the in-band D2D underlaying cellular systems, and therefore, below we present some relevant studies that have been carried.

In D2D underlaying cellular networks, the uplink resource sharing mode is considered more favorable than the downlink resource sharing mode due to some attractive features [6]. For example, in the downlink resource sharing, the user requires an additional Tx chain, which will increase the hardware cost and complexity. However, using either downlink resource sharing or uplink resource sharing ends up increasing the interference in the network, if interference management schemes are not employed [6]. In scenarios where D2D links are close to the BS, the downlink resource sharing mode outperforms the uplink resource sharing mode. In addition to this, most of the existing studies using uplink resource sharing for the D2D communication [6], [7] assumes a lighter traffic in the uplink and a heavier downlink traffic. However, some uplink applications with high data rate requirements are becoming popular in the future networks such as video conferencing and VoIP. Hence, according to [8], the traffic between uplink and downlink is becoming less asymmetric. Under these circumstances, the performance of the downlink resource sharing mode becomes more superior [8]. Therefore, the resource allocation problem for D2D underlaying the downlink cellular network is also important and should be investigated.

Interference management is a challenging task in the D2D underlaying cellular system due to the interference between D2D pairs and the cross interference between D2D and CUs [4]. Therefore, power control schemes have an immense importance in D2D underlaying cellular systems. By employing power control schemes, it is possible to minimize interference between CUs and DRs. Both centralized and distributed power control algorithms are proposed in [9] considering single-input single-output (SISO) links. It has been observed that the bottleneck of D2D underlaying cellular networks is the cross-layer interference and not the interference between D2D pairs. For the centralized power control in [9], the authors

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This article has been accepted for publication in a future issue of this journal, but has not been fully edited. Content may change prior to final publication. Citation information: DOI 10.1109/TVT.2018.2845748, IEEE Transactions on Vehicular Technology

determine the transmit power of users by maximizing the signal-to-interference-plus-noise ratio (SINR) of CUs while satisfying individual SINR constraints of DRs. Similarly, a power control scheme is investigated in [10] that regulates D2D transmit powers to protect the SINR degradation of CUs below a certain threshold. A sum-rate maximization scheme for designing DT powers is introduced in [11] using a deterministic network model. To minimize the amount of interference from the single D2D pair, a dynamic power control scheme is developed in [12]. All the studies discussed above consider single antenna SISO links, whereas current cellular networks are deploying multiple antennas also known as multiple-input multiple-output (MIMO) technology.

There are several studies that investigate D2D underlaying multiuser (MU) MIMO cellular systems. The work in [5] provides a comprehensive and systematic framework to address the issue of joint beamforming and power control in D2D underlaying cellular networks. This study only considers a single D2D pair and the statistical channel state information (CSI) with slowly varying channels. An MU MIMO system with multiple D2D pairs is studied in [13], where an optimization problem is designed which maximizes the overall rate of the system by employing maximum ratio transmission (MRT) and zero-forcing (ZF) beamforming schemes.

D2D can also be integrated with other key 5G technologies in order to meet the high data rate demand. Recently, a combination of MIMO beamforming, non-orthogonal multiple access and D2D technologies is investigated in [14]. It has been shown that together these three technologies can provide significant improvements in the overall system throughput.

Most of the studies discussed so far assume perfect CSI at the BS. However, in practice only imperfect or partial CSI is available at the BS. In frequency division duplex systems, limited feedback schemes are used to equip the BS with the quantized or imperfect CSI [15]. For D2D underlaying MU MIMO cellular systems, both perfect CSI and quantized CSI are investigated with conventional MRT and interference cancellation (IC) based beamforming schemes in [16].

Despite existing efforts such as [5] [16], there have been no prior studies that deal with the optimal joint beamforming and power control strategy with multiple D2D pairs guaranteeing QoS requirements of both CUs and DRs. Therefore, in this paper we propose an optimal joint beamforming and power control scheme for D2D communication underlaying MU MIMO downlink cellular networks under the centralized resource control framework [9] with perfect CSI and limited feedback. In particular, we aim to minimize the total transmit power in the network by jointly optimizing the beamforming vectors for CUs and the transmit power optimization for BS and DTs, while satisfying SINR based QoS requirements for both CUs and DRs. According to [17], the fast growth in the number of wireless devices in future will increase CO_2-equivalent emissions significantly. Also, 80% of the total power is consumed by the base station which needs to be redesigned in order to reduce the consumed power [17]. The goal is to manage the cost and revenue of running the BS. Therefore, to achieve this, we have selected the total transmission power in the network as an objective function. Our contributions are summarized as follows:

- With perfect CSI, we propose to use the semidefinite relaxation (SDR) technique [18] to find the optimal beamforming and power allocation solution. For the formulated problem, we prove that the SDR solution satisfies the rank-one constraint. Therefore, the rank relaxation does not lead to loss of optimality.
- In addition, we investigate three low-complexity suboptimal fixed beamforming schemes: ZF, regularized ZF (RZF) and hybrid MRT-ZF schemes. A closed-form optimal power allocation solution is derived for ZF beamforming, whereas second-order cone programming is used to find the solution of the hybrid MRT-ZF beamforming scheme.
- Next, when perfect CSI is not available at the BS, we propose a limited feedback based robust transmit power optimization strategy with ZF beamforming. The advantage is that it only requires quantized channel direction information (CDI) feedback but not the channel quality indicator (CQI). To our best knowledge, this is the first study that investigates robust power control for the D2D underlaying MU MIMO with limited feedback.

In order to evaluate the performance of the proposed schemes, we adopt the Poisson point process (PPP) to distribute CUs and DRs in the network as the PPP model has shown to be accurate for SINR distribution in urban cellular networks [19].

The remainder of this paper is organized as follows. Section II presents the D2D underlaying cellular system model and the formulation of the joint beamforming and power optimization problem. Section III provides the optimal solution using SDR and also proves that SDR is tight for our problem. Section IV presents suboptimal solutions with fixed beamforming schemes. Section V introduces the limited feedback based robust power optimization strategy with ZF beamforming. Numerical results for the proposed schemes are presented in Section VI. Finally, Section VII concludes the paper.

Notations: We use $(\cdot)^H$, $(\cdot)^T$, $(\cdot)^{-1}$ and $(\cdot)^\dagger$ to denote the conjugate transpose, the transpose, the inverse and the right pseudoinverse, respectively. $\|\cdot\|$ and $|\cdot|$ stand for vector and scalar norms, respectively. The mean, variance and covariance operations are denoted by $\mathbb{E}[\cdot]$, var$(\cdot)$ and cov$(\cdot,\cdot)$, respectively.

II. System Model and Problem Formulation

We consider a hybrid network consisting of CUs and D2D pairs in a given cell $\mathcal{A}$. The total number of CUs in a given cell is denoted by C, whereas the total number of D2D pairs in the cell is represented by D. We assume that each BS is equipped with M transmit antennas, whereas CUs, DTs and the associated DRs have a single antenna. We focus on the downlink transmission with no out-of-cell interference. In the given cell, the centralized radio resource allocation approach is considered where perfect CSI is first assumed available at the BS. The BS transmits the signal s_n with normalized power to the n^{th} CU after performing downlink precoding with the weight vector denoted by $\mathbf{w}_n$. The n^{th} CU experiences

This article has been accepted for publication in a future issue of this journal, but has not been fully edited. Content may change prior to final publication. Citation information: DOI 10.1109/TVT.2018.2845748, IEEE Transactions on Vehicular Technology

interference due to DTs and other CUs in the cell. Therefore, the received signal at the n^{th} CU can be written as:

$$\begin{aligned} y_n =& \sqrt{P_n L_n} \mathbf{h}_n^H \mathbf{w}_n s_n + \sum_{i \neq n}^{C} \sqrt{P_i L_n} \mathbf{h}_n^H \mathbf{w}_i s_i \\ &+ \sum_{k=1}^{D} \sqrt{p_k L_{k,n}} g_{k,n} s_k^d + n_n, \end{aligned} \tag{1}$$

where P_i denote the transmit power from the BS to the i^{th} CU. The transmit power of the k^{th} DT is given by p_k. The channel from the BS to the n^{th} CU is denoted by $\mathbf{h}_n \in \mathbb{C}^M$. The interfering channel from the k^{th} DT to the n^{th} CU is given by $g_{k,n}$. L_n and $L_{k,n}$ represent the path losses from the serving BS to the n^{th} CU and from the k^{th} DT to the n^{th} CU, respectively. s_k^d is the transmitted signal from the k^{th} DT to the k^{th} DR with unit power. The noise at the n^{th} CU is denoted by n_n which is assumed to have a Gaussian distribution with zero mean and variance σ_n^2. Thus, the SINR of the n^{th} CU is

$$\text{SINR}_n = \frac{P_n L_n |\mathbf{h}_n^H \mathbf{w}_n|^2}{L_n \sum_{i \neq n} P_i |\mathbf{h}_n^H \mathbf{w}_i|^2 + \sum_{k=1} p_k L_{k,n} |g_{k,n}|^2 + \sigma_n^2}. \tag{2}$$

The received signal at the m^{th} DR from the serving DT can be written as

$$\begin{aligned} y_m^d =& \sqrt{p_m L_m^d} g_m^d s_m^d + \sum_{i=1}^{C} \sqrt{P_i \bar{L}_m} \bar{\mathbf{h}}_m^H \mathbf{w}_i s_i \\ &+ \sum_{j \neq m}^{D} \sqrt{p_j L_{j,m}^d} g_{j,m}^d s_j^d + n_m^d, \end{aligned} \tag{3}$$

where the channel between the m^{th} D2D pair is denoted by g_m^d and the interfering channel from the j^{th} DT to the m^{th} DR is denoted by $g_{j,m}^d$. The interfering channel from the BS to the m^{th} DR is given by $\bar{\mathbf{h}}_m$. L_m^d and $L_{j,m}^d$ denote the path losses from the serving DT to the m^{th} DR and from the j^{th} DT to the m^{th} DR, respectively. Similarly, $\bar{L}_m$ represents the path loss from the BS to the m^{th} DR. The noise term at the DR is represented by n_m^d which is also assumed to be a Gaussian distributed random variable with zero mean and variance $\sigma_m^{d^2}$. The SINR of the m^{th} DR can be expressed as

$$\text{SINR}_m^d = \frac{p_m L_m^d |g_m^d|^2}{\sum_{i=1} P_i \bar{L}_m |\bar{\mathbf{h}}_m^H \mathbf{w}_i|^2 + \sum_{j \neq m} p_j L_{j,m}^d |g_{j,m}^d|^2 + \sigma_m^{d^2}}. \tag{4}$$

In D2D underlaying cellular systems, the major challenge is to mitigate the interference due to the added D2D links. It is well known that by employing an efficient power control scheme, interference in the network can be well controlled. Therefore, in this study, we specifically focus on a power control approach that not only reduces interference but also improves the energy efficiency of the network. The main idea is to allocate minimum transmit powers to all CUs and DRs in the network, while maintaining the acceptable QoS requirements for both CUs and DRs. We provide both optimal and sub-optimal strategies to achieve the minimum total transmit power in the cell. In the optimal method, the aim is to jointly design beamforming vectors at the BS and transmit powers for both BS and DTs. Therefore, we can formulate the optimization problem as

$$\begin{aligned} \min_{\{\mathbf{v}_n, p_m\}} \quad & \sum_{n=1}^{C} \|\mathbf{v}_n\|^2 + \sum_{m=1}^{D} p_m \\ \text{s.t.} \quad & \frac{L_n |\mathbf{h}_n^H \mathbf{v}_n|^2}{L_n \sum_{i \neq n} |\mathbf{h}_n^H \mathbf{v}_i|^2 + \sum_{k=1} p_k L_{k,n} |g_{k,n}|^2 + \sigma_n^2} \geq \gamma_n, \forall n, \\ & \frac{p_m L_m^d |g_m^d|^2}{\sum_{i=1} \bar{L}_m |\bar{\mathbf{h}}_m^H \mathbf{v}_i|^2 + \sum_{j \neq m} p_j L_{j,m}^d |g_{j,m}^d|^2 + \sigma_m^{d^2}} \geq \gamma_m^d, \forall m, \end{aligned} \tag{5}$$

where $\mathbf{v}_n = \sqrt{P_n} \mathbf{w}_n$. γ_n and γ_m^d denote the SINR targets for the n^{th} CU and for the m^{th} DR, respectively. Note that the problem (5) is non-convex due to quadratic terms involving $\{\mathbf{v}_n\}$. In the following section, we first present the optimal solution for (5) using the SDR technique. In this study, similar to [20], [21], we have not limited the transmitting nodes by a maximum transmit power constraint, however, the formulation of the problem can be easily extended to include the maximum transmit power constraint by adding linear constraints. According to [21], the extended model does not affect the convexity and complexity of the problem.

III. The Optimal Solution using SDR

In this section, we provide the optimal solution for the problem (5) using the SDR technique [18]. In SDR, the constraint $\text{Rank}(\mathbf{X}_n) = 1$, where $\mathbf{X}_n = \mathbf{v}_n \mathbf{v}_n^H$, is usually dropped in order to obtain the relaxed version of the problem. Using $P_{\text{T}} = \sum_{n=1}^{C} \text{Tr}(\mathbf{X}_n) + \sum_{m=1}^{D} p_m$, the SDR of problem (5) can be written as

$$\begin{aligned} \min_{\{\mathbf{X}_n \succeq 0, p_m \geq 0\}} \quad & P_{\text{T}} \\ \text{s.t.} \quad & \frac{L_n \mathbf{h}_n^H \mathbf{X}_n \mathbf{h}_n}{\gamma_n} - \sum_{i \neq n}^{C} L_n \mathbf{h}_n^H \mathbf{X}_i \mathbf{h}_n \geq \sum_{k=1}^{D} p_k L_{k,n} |g_{k,n}|^2 + \sigma_n^2, \forall n, \\ & \frac{p_m L_m^d |g_m^d|^2}{\gamma_m^d} - \sum_{i=1}^{C} \bar{L}_m \bar{\mathbf{h}}_m^H \mathbf{X}_i \bar{\mathbf{h}}_m \geq \sum_{j \neq m}^{D} p_j L_{j,m}^d |g_{j,m}^d|^2 + \sigma_m^{d^2}, \forall m. \end{aligned} \tag{6}$$

The problem (6) is a convex problem and can be solved by interior-point algorithm [22] using CVX, a package for specifying and solving convex programs [23]. Let $\{\mathbf{X}_n^*\}$ and $\{p_m^*\}$ denote the optimal solution to the problem (6), then the optimal solution $\mathbf{v}_n^*$ to the problem (5) can be obtained from the eigenvalue decomposition of $\mathbf{X}_n^*$, $\forall n$, if $\text{Rank}(\mathbf{X}_n^*) = 1$. Here, the vector $\mathbf{v}_n^*$ corresponds to the eigenvector corresponding to the maximum eigenvalue of $\mathbf{X}_n^*$. On the other hand, if $\text{Rank}(\mathbf{X}_n^*) > 1$, then the optimal solutions $\{\mathbf{X}_n^*\}$ and $\{p_m^*\}$ to problem (6) may not necessarily be optimal for problem (5). Following the derivation approach presented in [24], here we prove that the solution to the problem (6) always satisfies $\text{Rank}(\mathbf{X}_n^*) = 1, \ \forall n$. As the problem (6) is convex (and satisfies the Slater's condition), therefore, a strong duality holds i.e., the duality gap is zero. Let $\{\lambda_n\}$ and $\{\mu_n\}$ denote the dual variables corresponding to the CU and DR SINR

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constraints in the problem (6), respectively. We can write the partial Lagrangian of the problem (6) as

$$L(\{\mathbf{X}_n, p_m, \lambda_n, \mu_m\}) \triangleq \sum_{n=1}^{C} \mathrm{Tr}\,(\mathbf{X}_n) + \sum_{m=1}^{D} p_m - \sum_{n=1}^{C} \lambda_n \times \quad (7)$$
$$\left(\frac{L_n \mathbf{h}_n^H \mathbf{X}_n \mathbf{h}_n}{\gamma_n} - L_n \sum_{i \neq n}^{C} \mathbf{h}_n^H \mathbf{X}_i \mathbf{h}_n - \sigma_n^2 - \sum_{k=1}^{D} p_k L_{k,n} |g_{k,n}|^2\right) -$$
$$\sum_{m=1}^{D} \mu_m \left(\frac{p_m L_m^d |g_m^d|^2}{\gamma_m^d} - \sum_{i=1}^{C} \bar{L}_m \bar{\mathbf{h}}_m^H \mathbf{X}_i \bar{\mathbf{h}}_m - \sum_{j \neq m}^{D} p_j L_{j,m}^d |g_{j,m}^d|^2 - \sigma_m^{d^2}\right).$$

Therefore, with the partial Lagrangian (7), the Lagrange dual function of problem (6) is given by [22, Section 5.1.2]

$$\min_{\mathbf{X}_n \succeq 0, p_m > 0} L(\{\mathbf{X}_n, p_m, \lambda_n, \mu_m\}), \quad (8)$$

which can be explicitly expressed as

$$\min_{\mathbf{X}_n \succeq 0, p_m > 0} \Bigg\{ \sum_{n=1}^{C} \mathrm{Tr}\,(\mathbf{A}_n \mathbf{X}_n) + \sum_{m=1}^{D} p_m + \sum_{n=1}^{C} \lambda_n \times$$
$$\left(\sum_{k=1}^{D} p_k L_{k,n} |g_{k,n}|^2 - \sigma_n^2\right) - (1 + \gamma_m^d) \sum_{m=1}^{D} \mu_m p_m L_m^d |g_m^d|^2$$
$$+ \sum_{m=1}^{D} \gamma_m^d \mu_m \left(\sum_{j=1}^{D} p_j L_{j,m}^d |g_{j,m}^d|^2 + \sigma_m^{d^2}\right)\Bigg\}, \quad (9)$$

where by letting $\{\lambda_n^*\}$ and $\{\mu_m^*\}$ denote the optimal dual solution to the problem (6), the matrix $\mathbf{A}_n^*$ can be written as

$$\mathbf{A}_n^* = \mathbf{I}_M + \sum_{j=1}^{C} \lambda_j^* L_j \mathbf{h}_j \mathbf{h}_j^H + \sum_{q=1}^{D} \mu_q^* \gamma_q^d \bar{L}_q \bar{\mathbf{h}}_q \bar{\mathbf{h}}_q^H$$
$$- \lambda_n^* L_n \left(\frac{1}{\gamma_n} + 1\right) \mathbf{h}_n \mathbf{h}_n^H. \quad (10)$$

It is evident from (9) that $\mathbf{X}_n^*$ must be a solution to the problem below

$$\min_{\mathbf{X}_n \succeq 0} \mathrm{Tr}(\mathbf{A}_n^* \mathbf{X}_n) = 0. \quad (11)$$

To obtain a lower bounded dual optimal value, we should have $\mathbf{A}_n^* \succeq 0,\ \forall n$ [24]. This results in the optimal value of the problem (11) equals to zero, i.e., $\mathrm{Tr}(\mathbf{A}_n^* \mathbf{X}_n^*) = 0$, and with both $\mathbf{A}_n^* \succeq 0$ and $\mathbf{X}_n^* \succeq 0$, we have

$$\mathbf{A}_n^* \mathbf{X}_n^* = \mathbf{0},\ \forall n, \quad (12)$$

From (10), it is evident that $\mathbf{A}_n^*$ has at most one zero eigenvalue, and therefore, $\mathrm{Rank}(\mathbf{A}_n^*) \geq M - 1, \forall n$. Then according to (12), $\mathrm{Rank}(\mathbf{X}_n^*) = 1$ when the problem is feasible. This completes the proof.

By incorporating maximum power constraints at the BS and DTs, it can be shown that the rank-one solution also holds for the modified SDR problem. The maximum power constraints for BS and DTs are given by

$$\sum_{u=1}^{C} \mathrm{Tr}\,(\mathbf{X}_u) \leq P_b^{\max} \quad (13)$$

and

$$p_m \leq p_d^{\max}, \quad (14)$$

respectively, where $P_b^{\max}$ and $p_d^{\max}$ denote the maximum total power available at the BS and DTs, respectively. The dual variables associated with the maximum transmit power constraints at the BS and DTs are represented by β_b and ζ_m, respectively, where $\beta_b \geq 0$ and $\zeta_m \geq 0$. The Lagrange dual function of the modified problem is given by

$$\min_{\mathbf{X}_n \succeq 0, p_m > 0} L(\{\mathbf{X}_n, p_m, \lambda_n, \mu_m, \beta_b, \zeta_m\}) = \quad (15)$$
$$\min_{\mathbf{X}_n \succeq 0, p_m > 0} \Bigg\{ \sum_{n=1}^{C} \mathrm{Tr}\left(\hat{\mathbf{A}}_n \mathbf{X}_n\right) + \sum_{m=1}^{D} p_m + \sum_{n=1}^{C} \lambda_n \times$$
$$\left(\sum_{k=1}^{D} p_k L_{k,n} |g_{k,n}|^2 - \sigma_n^2\right) - (1 + \gamma_m^d) \sum_{m=1}^{D} \mu_m p_m L_m^d |g_m^d|^2 +$$
$$\sum_{m=1}^{D} \gamma_m^d \mu_m \left(\sum_{j=1}^{D} p_j L_{j,m}^d |g_{j,m}^d|^2 + \sigma_m^{d^2}\right) - \beta_b P_b^{\max} + \zeta_m (p_m - P_d^{\max})\Bigg\}.$$

By denoting β_b^* as the optimal dual solution corresponding to the maximum BS power constraint in the modified problem, we can write the matrix, $\hat{\mathbf{A}}_n^*$ as

$$\hat{\mathbf{A}}_n^* = (1 + \beta_b^*)\,\mathbf{I}_M + \sum_{j=1}^{C} \lambda_j^* L_j \mathbf{h}_j \mathbf{h}_j^H + \sum_{q=1}^{D} \mu_q^* \gamma_q^d \bar{L}_q \bar{\mathbf{h}}_q \bar{\mathbf{h}}_q^H$$
$$- \lambda_n^* L_n \left(\frac{1}{\gamma_n} + 1\right) \mathbf{h}_n \mathbf{h}_n^H. \quad (16)$$

Note that the matrix $\hat{\mathbf{A}}_n^*$ has a similar structure to that of $\mathbf{A}_n^*$ in (10), therefore using the similar approach as before, we can observe that for $\hat{\mathbf{A}}_n^* \succeq 0$ and $\mathbf{X}_n^* \succeq 0$, we have $\hat{\mathbf{A}}_n^* \mathbf{X}_n^* = 0$. This implies that the matrix $\hat{\mathbf{A}}_n^*$ has at most one zero eigenvalue, and therefore, from $\hat{\mathbf{A}}_n^* \mathbf{X}_n^* = 0, \forall n$, we can conclude that $\mathrm{Rank}(\mathbf{X}_n) = 1, \forall n$.

IV. Sub-optimal Solutions with Fixed Beamforming

In this section, we present the sub-optimal solution for the problem (5) using fixed beamforming techniques: ZF, RZF and the hybrid MRT-ZF beamforming scheme. Here, the aim is to minimize the total transmission power at BS and DTs while guaranteeing QoS requirements for both CUs and DRs with reduced complexity.

A. ZF Beamforming

In this section, we present the sub-optimal solution for the problem (5) using the ZF beamforming scheme. The constraint $M \geq (C + D)$ must be satisfied in order to implement ZF beamforming at the BS. In this study, the BS not only eliminates interference at each CU caused due to the other CUs but also at each DR caused due to the CUs in the network. Thus, with ZF beamforming we have $\mathbf{h}_n^H \mathbf{w}_i = 0$ and $\bar{\mathbf{h}}_m^H \mathbf{w}_i = 0$. The ZF beamforming vector for the n^{th} CU can be computed by solving the following optimization problem

$$\begin{aligned} \max_{\{\mathbf{w}_n\}} \quad & \left|\mathbf{h}_n^H \mathbf{w}_n\right|^2 \\ \text{s.t.} \quad & \mathbf{H}_n^H \mathbf{w}_n = \mathbf{0}_{(M-1)\times 1} \quad \forall n, \\ & \|\mathbf{w}_n\|^2 = 1, \quad \forall n, \end{aligned} \quad (17)$$

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where $\mathbf{H}_n = [\mathbf{h}_1, \ldots, \mathbf{h}_{n-1}, \mathbf{h}_{n+1}, \ldots, \mathbf{h}_C, \bar{\mathbf{h}}_1, \ldots, \bar{\mathbf{h}}_D]$. The solution to the problem (17) is given by [25]

$$\mathbf{w}_n^{(\text{ZF})} = \frac{\left(\mathbf{I}_M - \mathbf{H}_n^{\dagger}\mathbf{H}_n\right)\mathbf{h}_n}{\left\|\left(\mathbf{I}_M - \mathbf{H}_n^{\dagger}\mathbf{H}_n\right)\mathbf{h}_n\right\|}. \quad (18)$$

Denoting $P_{\text{T}} = \sum_{n=1}^{C} P_n + \sum_{m=1}^{D} p_m$, $G_{i,j} = L_i \left|\mathbf{h}_i^H \mathbf{w}_j\right|^2$, $\bar{g}_{i,j} = L_{i,j}|g_{i,j}|^2$, $\bar{g}_{i,j}^d = L_{i,j}^d|g_{i,j}^d|^2$ and $\bar{g}_i^d = L_i^d|g_i^d|^2$, we can transform the problem (5) using (18) into the following formulation

$$\begin{aligned} &\min_{\{P_n, p_m\}} \quad P_{\text{T}} \\ &\text{s.t.} \quad P_n G_{n,n} \geq \gamma_n \sum_{k=1} p_k \bar{g}_{k,n} + \gamma_n \sigma_n^2, \quad \forall n, \\ &\qquad p_m \bar{g}_m^d \geq \gamma_m^d \sum_{j \neq m} p_j \bar{g}_{j,m}^d + \gamma_m^d \sigma_m^{d^2}, \quad \forall m. \end{aligned} \quad (19)$$

Note that the problem (19) is convex because it comprises of a linear objective function and linear terms with respect to the transmit powers, in the two constraints. By solving the problem (19) using the CVX software, we can obtain the optimal transmit powers at the BS, denoted by $\{P_n^*\}$, and also the optimal transmit powers for DTs, represented by $\{p_m^*\}$.

To derive the closed-form solution of the problem (19), we observe that $\{p_m\}$ can be decoupled from $\{P_n\}$, so we can optimize them separately. We can first obtain the solution for the minimum $\{p_m^*\}$ which can be shown to have a unique solution satisfying

$$\frac{p_m \bar{g}_m^d}{\sum_{j \neq m} p_j \bar{g}_{j,m}^d + \sigma_m^{d^2}} \geq \gamma_m^d, \quad \forall m. \quad (20)$$

Expressing (20) in the matrix form, we get

$$\left[\mathbf{I}_D - \mathbf{F}\right]\mathbf{p}^{\text{d}} \geq \mathbf{u}, \quad (21)$$

where $\mathbf{p}^{\text{d}} = [p_1, p_2, \ldots, p_D]^T$ and the m^{th} entry of $\mathbf{u}$ is $u_m = \gamma_m^d \sigma_m^{d^2} / \bar{g}_m^d$. The entries of matrix $\mathbf{F}$ are given by

$$\mathbf{F}_{m,j} = \begin{cases} 0, & \text{if } j = m, \\ \frac{\gamma_m^d \bar{g}_{j,m}^d}{\bar{g}_m^d}, & \text{if } j \neq m. \end{cases} \quad (22)$$

Therefore, we can write the problem (19) as

$$\begin{aligned} &\min \sum_{m=1}^{D} p_m \\ &\text{s.t.} \quad \left[\mathbf{I}_D - \mathbf{F}\right]\mathbf{p}^{\text{d}} \succeq \mathbf{u}. \end{aligned} \quad (23)$$

When the spectral radius of $\mathbf{F}$ is less than one, the matrix $\mathbf{I}_D - \mathbf{F}$ is invertible [20], thus yielding the optimal solution as

$$\mathbf{p}^{\text{d}^*} = \left[\mathbf{I}_D - \mathbf{F}\right]^{-1}\mathbf{u}. \quad (24)$$

Note that in [26], an alternative distributed solution to solve the problem (23) is presented which can also achieve the optimal solution by iteratively updating the power of each DR without performing matrix inversion.

Once we get the optimal solution for $\{p_m^*\}$ from (24), the transmit power for the n^{th} CU, P_n, can be found as

$$P_n = \frac{\gamma_n(\sum_{k=1} p_k \bar{g}_{k,n} + \sigma_n^2)}{G_{n,n}}. \quad (25)$$

B. RZF Beamforming

RZF beamforming was introduced in [27] as an alternative to ZF beamforming because the latter is not robust for the ill-conditioned channel matrix. RZF beamforming also performs well in the low signal-to-noise ratio (SNR) regime, compared to ZF beamforming [27]. The RZF beamforming vector for the n^{th} CU is given by

$$\mathbf{w}_n^{(\text{RZF})} = \frac{\left(\mathbf{Y}_n\mathbf{Y}_n^{\dagger} + \eta_n\mathbf{I}_M\right)^{-1}\mathbf{h}_n}{\left\|\left(\mathbf{Y}_n\mathbf{Y}_n^{\dagger} + \eta_n\mathbf{I}_M\right)^{-1}\mathbf{h}_n\right\|}, \quad (26)$$

where $\mathbf{Y}_n = [\mathbf{h}_1, \ldots, \mathbf{h}_C, \mathbf{h}_{n,1}, \ldots, \mathbf{h}_{n,D}]$ and η_n is a regularization parameter. In this study, we use $\eta_n = (C+D)/\sigma_n^2$ which is similar to the regularization parameter used in [27]. Ideally η_n should be optimized, however, this is a difficult task which requires the consideration of factors like propagation environment, antenna correlation, etc. Therefore, to keep the complexity low, we use the regularization parameter derived in [27] which considers equal power allocation among the users (i.e. $\eta_n = \eta$, $\forall n$). Unlike ZF beamforming, in RZF beamforming we have $G_{i,j} \neq 0$ when $i \neq j$. Therefore, the problem (5) with RZF beamforming becomes

$$\begin{aligned} &\min_{\{P_n, p_m\}} \quad P_{\text{T}} \\ &\text{s.t. } P_n G_{n,n} - \gamma_n \sum_{i \neq n} P_i G_{n,i} \geq \gamma_n \sum_{k=1} p_k \bar{g}_{k,n} + \gamma_n \sigma_n^2, \ \forall n, \\ &p_m \bar{g}_m^d - \gamma_m^d \sum_{j \neq m} p_j \bar{g}_{j,m}^d \geq \gamma_m^d \sum_{i=1} P_i G'_{m,i} + \gamma_m^d \sigma_m^{d^2}, \ \forall m, \end{aligned} \quad (27)$$

where $G'_{m,i} = \bar{L}_m|\bar{\mathbf{h}}_m^H \mathbf{w}_i|^2$. Similar to (19), the problem (27) also comprises of a linear objective function and linear constraints, hence, making it a convex optimization problem. The method in IV.A applies to this case except that $\{P_n\}$ and $\{p_m\}$ are coupled together, so we need to optimize the whole power vector $\{P_1, \cdots, P_n, \cdots, P_C, p_1, \cdots, p_m, \cdots, p_D\}$. The same technique can be used to solve for any other fixed beamforming such as MRT where the beamforming vector for the n^{th} CU is given by $\mathbf{w}_n^{\text{MRT}} = \mathbf{h}_n/\|\mathbf{h}_n\|$.

C. Hybrid MRT-ZF

In this subsection, we linearly combine MRT and ZF beamforming schemes to obtain a hybrid beamforming scheme which achieves the optimal trade-off between the two schemes [21], [28]. Denoting $\hat{\mathbf{w}}_n^{\text{MRT}}$ and $\hat{\mathbf{w}}_n^{\text{ZF}}$ as unnormalized MRT and ZF beamforming vectors, respectively. Then according to [21], the hybrid MRT-ZF beamforming vector for the n^{th} CU is given by

$$\mathbf{w}_n^{(\text{hyb})} = \frac{\sqrt{x_n}\hat{\mathbf{w}}_n^{(\text{MRT})} + \sqrt{y_n}\hat{\mathbf{w}}_n^{(\text{ZF})}}{\|\sqrt{x_n}\hat{\mathbf{w}}_n^{(\text{MRT})} + \sqrt{y_n}\hat{\mathbf{w}}_n^{(\text{ZF})}\|}. \quad (28)$$

Let us denote the unnormalized hybrid beamforming vector for the n^{th} CU as $\mathbf{v}_n^{\text{hyb}} = \sqrt{P_n}\mathbf{w}_n^{\text{hyb}}$. We can write the quantity $\bar{G}_{i,j} = L_i|\mathbf{h}_i^H\mathbf{v}_j^{\text{hyb}}|^2$ as

$$\bar{G}_{i,j} = \begin{cases} x_j L_i|\mathbf{h}_i^H\mathbf{h}_j|^2, & \text{if } i \neq j, \\ L_i|\sqrt{x_i}\mathbf{h}_i^H\mathbf{h}_i + \sqrt{y_i}\mathbf{h}_i^H\left(\mathbf{I}_M - \mathbf{H}_i^{\dagger}\mathbf{H}_i\right)\mathbf{h}_i|^2, & \text{if } i = j. \end{cases} \quad (29)$$

Substituting $Z_{i,j} = |\mathbf{h}_i^H\mathbf{h}_j|$, $z_i = \mathbf{h}_i^H\left(\mathbf{I}_M - \mathbf{H}_i^{\dagger}\mathbf{H}_i\right)\mathbf{h}_i$ and $r_i = \sqrt{x_i y_i}$, we can rewrite (29) as

$$\bar{G}_{i,j} = \begin{cases} x_j L_i Z_{i,j}^2, & \text{if } i \neq j, \\ L_i(x_i Z_{i,i}^2 + y_i z_i^2 + 2r_i Z_{i,i} z_i), & \text{if } i = j. \end{cases} \quad (30)$$

Note that for the $i = j$ case, $Z_{i,i} \geq 0$ and $z_i \geq 0$. Denoting $\hat{\mathbf{w}}_n^{\text{hyb}} = \sqrt{x_n}\hat{\mathbf{w}}_n^{(\text{MRT})} + \sqrt{y_n}\hat{\mathbf{w}}_n^{(\text{ZF})}$, the transmit power to the n^{th} CU with hybrid beamforming can be written as [21]

$$P_n = \|\hat{\mathbf{w}}_n^{\text{hyb}}\|^2 = x_n Z_{n,n} + y_n z_n + 2r_n z_n. \quad (31)$$

In order to convert the original problem (5) with hybrid beamforming to a convex optimization problem, we relax the constraint $r_n = \sqrt{x_n y_n}$ by using $x_n y_n \geq r_n^2$. Then the optimization problem (5) with hybrid MRT-ZF beamforming can be expressed as

$$\begin{aligned} \min_{\{x_n, y_n, r_n, p_m\}} \quad & \sum_{n=1}^{C} x_n Z_{n,n} + y_n z_n + 2r_n z_n + \sum_{m=1}^{D} p_m \qquad (32) \\ \text{s.t. } & L_n(x_n Z_{n,n}^2 + y_n z_n^2 + 2r_n Z_{n,n} z_n) - \gamma_n L_n \sum_{i \neq n} x_i Z_{n,i}^2 \\ & \geq \gamma_n \sum_{k=1} p_k \bar{g}_{k,n} + \gamma_n \sigma_n^2, \forall n, \\ & p_m \bar{g}_m^d - \gamma_m^d \sum_{j \neq m} p_j \bar{g}_{j,m}^d \geq \gamma_m^d \bar{L}_m \sum_{i=1} x_i Z_{m,i}^{'2} + \gamma_m^d \sigma_m^{d^2}, \forall m, \\ & x_n y_n \geq r_n^2, \forall n, \\ & x_n \geq 0, \; y_n \geq 0, \forall n, \end{aligned}$$

where $Z_{m,i}^{'} = |\bar{\mathbf{h}}_m^H \mathbf{h}_i|$. We reformulate the constraint $x_n y_n \geq r_n^2$ into a second-order cone constraint, so that (32) can be written as a convex optimization problem as

$$\begin{aligned} \min_{\{x_n, y_n, r_n, p_m\}} \quad & \sum_{n=1}^{C} x_n Z_{n,n} + y_n z_n + 2r_n z_n + \sum_{m=1}^{D} p_m \qquad (33) \\ \text{s.t. } & L_n(x_n Z_{n,n}^2 + y_n z_n^2 + 2r_n Z_{n,n} z_n) - \gamma_n L_n \sum_{i \neq n} x_i Z_{n,i}^2 \\ & \geq \gamma_n \sum_{k=1} p_k \bar{g}_{k,n} + \gamma_n \sigma_n^2, \forall n, \\ & p_m \bar{g}_m^d - \gamma_m^d \sum_{j \neq m} p_j \bar{g}_{j,m}^d \geq \gamma_m^d \bar{L}_m \sum_{i=1} x_i Z_{m,i}^{'2} + \gamma_m^d \sigma_m^{d^2}, \forall m, \\ & \left\| \begin{pmatrix} 2r_n \\ x_n - y_n \end{pmatrix} \right\| \leq x_n + y_n, \forall n, \\ & x_n \geq 0, \; y_n \geq 0, \forall n. \end{aligned}$$

Note that as the problem (33) belongs to the second-order programming and thus convex, it can also be solved efficiently by using off-the-shelf algorithms [29] in CVX. Note that due to the relaxation $x_n y_n \geq r_n^2$, the problem (33) may provide a solution x_n, y_n, r_n which may not be a valid solution. For such instances, one can use the solution $\{x_n, y_n\}$ of problem (33) to construct the direction of the beamforming vectors given by (28), and then use the approach given in problem (27) to obtain the transmit powers. However, if the latter provides an infeasible solution, ZF beamforming can be utilized.

V. Limited Feedback Based Robust Power Control with ZF Beamforming

So far we have assumed that perfect CSI is available at the BS. Since this assumption is not feasible in practice, we propose to use limited feedback as an effective technique to provide the BS with the quantized CSI [15]. In limited feedback schemes, a rate constrained feedback link is used between the UE and the BS. Each user estimates the downlink channel $\mathbf{h}$ and quantizes the CDI, given by $\hat{\mathbf{h}} = \mathbf{h}/\|\mathbf{h}\|$. The quantization is performed using a codebook containing 2^B codewords or quantized channel vectors, given by $\mathcal{C} = [\mathbf{c}_1, \mathbf{c}_2, \ldots, \mathbf{c}_{2^B}]$, where B is the number of feedback bits. Both users and the BS maintain the same codebook. The codeword is selected on the basis of a minimum chordal distance [30], i.e, the codeword yielding the minimum chordal distance with the channel vector is selected. The minimum chordal distance based codeword selection for the n^{th} is given by

$$\tilde{\mathbf{h}}_n = \arg \min_{1 \leq i \leq 2^B} \sqrt{1 - \left|\hat{\mathbf{h}}_n^H \mathbf{c}_i\right|^2}. \quad (34)$$

The user feeds back the index of the selected codeword to the BS where the BS selects the quantized channel entry from the codebook corresponding to the received index. In this section, we only present the ZF beamforming based limited feedback scheme because it can be implemented efficiently with only having the CDI at the BS, whereas all the other fixed beamforming schemes discussed in this paper require both CDI and CQI, CQI $= \|\mathbf{h}\|$, at the BS. In this study, we rely on the well-known random vector quantization (RVQ) codebook [31] for CDI quantization due to their simplicity and adaptability. In the RVQ codebook, each codeword is randomly and independently generated from M dimensional unit norm complex Gaussian vectors.

We compute the optimal robust transmit powers for ZF beamforming by formulating an optimization problem that minimizes the total transmission power while satisfying the expected SINR constraints. Here, the reason to use the expected SINR constraints is due to the fact that the knowledge of the instantaneous channels is not available at the BS, hence, it is not able to compute instantaneous SINRs for both CUs and DRs. We begin by deriving the expected SINRs at both CUs and DRs with limited feedback and ZF beamforming. Here, we assume that $M = C + D$. The ZF beamforming vector for the n^{th} CU with limited feedback CSI is given by

$$\tilde{\mathbf{w}}_n^{(\text{ZF})} = \frac{\left(\mathbf{I}_M - \tilde{\mathbf{H}}_n^{\dagger}\tilde{\mathbf{H}}_n\right)\tilde{\mathbf{h}}_n}{\left\|\left(\mathbf{I}_M - \tilde{\mathbf{H}}_n^{\dagger}\tilde{\mathbf{H}}_n\right)\tilde{\mathbf{h}}_n\right\|}, \quad (35)$$

where $\tilde{\mathbf{H}}_n = [\tilde{\mathbf{h}}_1, \ldots, \tilde{\mathbf{h}}_{n-1}, \tilde{\mathbf{h}}_{n+1} \ldots, \tilde{\mathbf{h}}_C, \tilde{\bar{\mathbf{h}}}_1, \ldots, \tilde{\bar{\mathbf{h}}}_D]$ with $\tilde{\mathbf{h}}_n$ denoting the quantized CDI from the BS to the n-th CU and $\tilde{\bar{\mathbf{h}}}_m$ denoting the quantized interfering CDI from the

BS to the m-th DR. Note that with limited feedback CSI, ZF beamforming is not able to eliminate the interference completely and the expected SINR of the n^{th} CU can be approximated[1] by (36) given at the top of the next page, where $\hat{\mathbf{h}}_n$ denotes CDI of the n^{th} CU channel. We assume that entries of n^{th} CU channel, $\mathbf{h}_n$, follow an independent and identically distributed (i.i.d.) Gaussian distribution with zero mean and unit variance, and therefore, in (36), the quantity $\|\mathbf{h}_n^H\|^2$ is Gamma distributed with parameters M and 1, i.e., $\|\mathbf{h}_n^H\|^2 \sim \text{Gamma}(M,1)$, so we have $\mathbb{E}\left[\|\mathbf{h}_n^H\|^2\right] = M$. Both $\hat{\mathbf{h}}_n$ and $\tilde{\mathbf{w}}_n^{(\text{ZF})}$ are independent and isotropically distributed in $\mathbb{C}^M$, and therefore the quantity $|\hat{\mathbf{h}}_n^H \tilde{\mathbf{w}}_n^{(\text{ZF})}|^2$ has a Beta distribution with parameters 1 and $M-1$, i.e., $|\hat{\mathbf{h}}_n^H \tilde{\mathbf{w}}_n^{(\text{ZF})}|^2 \sim \text{Beta}(1, M-1)$, making $\mathbb{E}[|\hat{\mathbf{h}}_n^H \tilde{\mathbf{w}}_n^{(\text{ZF})}|^2] = 1/M$ [37], [38]. $|g_{k,n}|^2$ follows an exponential distribution with the rate parameter equals to 1, and therefore, we have $\mathbb{E}[|g_{k,n}|^2] = 1$. Although we have assumed a centralized control framework in this paper, for the limited feedback approach considered in this section, the knowledge of D2D channels (both desired and interfering) is not required at the BS to design a long-term robust power control scheme, which reduces the feedback overhead of the system. We can decompose the vector $\hat{\mathbf{h}}_n$ as

$$\hat{\mathbf{h}}_n = (\cos\theta)\tilde{\mathbf{h}}_n + (\sin\theta)\mathbf{f}_n, \tag{37}$$

where $\mathbf{f}_n$ stands for the error vector due to the quantization process and θ denotes the angle between $\hat{\mathbf{h}}_n$ and $\tilde{\mathbf{h}}_n$. Using (37), we can express $\mathbb{E}\left[|\hat{\mathbf{h}}_n^H \tilde{\mathbf{w}}_i^{(\text{ZF})}|^2\right] = \mathbb{E}[\sin^2\theta]\mathbb{E}\left[|\mathbf{f}_n^H \tilde{\mathbf{w}}_i^{(\text{ZF})}|^2\right]$, where from [37] $\mathbb{E}[\sin^2\theta] = ((M-1)/M)2^{-B/(M-1)}$ and $|\mathbf{f}_n^H \tilde{\mathbf{w}}_i^{(\text{ZF})}|^2 \sim \text{Beta}(1, M-2)$ with $\mathbb{E}[|\mathbf{f}_n^H \tilde{\mathbf{w}}_i^{(\text{ZF})}|^2] = 1/(M-1)$. Therefore, (36) can be written as

$$\mathbb{E}\left[\widetilde{\text{SINR}}_n\right] \approx \frac{P_n L_n}{L_n \sum_{i\neq n} P_i 2^{\frac{-B}{M-1}} + \sum_{k=1} p_k L_{k,n} + \sigma_n^2}. \tag{38}$$

For the D2D SINR, unlike the CU case, numerator and denominator are independent. Denoting $\tilde{S}_m = p_m L_m^d |g_m^d|^2$, $\tilde{I}_{\text{CU}} = \sum_{i=1} P_i \bar{L}_m |\bar{\mathbf{h}}_m^H \tilde{\mathbf{w}}_i|^2$ and $\tilde{I}_{\text{D2D}} = \sum_{j\neq m} p_j L_{j,m}^d |g_{j,m}^d|^2$. Therefore, we can compute the expected SINR of the m^{th} DR as

$$\begin{aligned}
\mathbb{E}\left[\widetilde{\text{SINR}}_m^d\right] &= \mathbb{E}\left[\frac{\tilde{S}_m}{\tilde{I}_{\text{CU}} + \tilde{I}_{\text{D2D}} + \sigma_m^{d^2}}\right] && (39)\\
&= \mathbb{E}\left[\tilde{S}_m\right]\mathbb{E}\left[\frac{1}{\tilde{I}_{\text{CU}} + \tilde{I}_{\text{D2D}} + \sigma_m^{d^2}}\right] && (40)\\
&\overset{(a)}{\geq} \mathbb{E}\left[\tilde{S}_m\right]\mathbb{E}\left[\tilde{I}_{\text{CU}} + \tilde{I}_{\text{D2D}} + \sigma_m^{d^2}\right]^{-1}, && (41)
\end{aligned}$$

where (a) follows from the Jensen's inequality as the random variable, $X = \tilde{I}_{\text{CU}} + \tilde{I}_{\text{D2D}} + \sigma_m^{d^2}$, is strictly positive and $1/\tilde{S}_m$ is convex. Equation (41) gives the lower bound on the expected SINR of the m^{th} D2D receiver. We assume that the channel between each D2D pair follows the Rician fading channel and the interfering channel from any DT to CUs and non-serving DRs follows the i.i.d. Rayleigh fading distribution with zero mean and unit variance. Therefore, according to [39], $\mathbb{E}\left[|g_m^d|^2\right] = {}_1F_1(-1,1;-K)/(K+1)$, where ${}_1F_1(\cdot)$ is the Kummer confluent hypergeometric function and K denotes the Rician K factor. Therefore, the lower bounded expected SINR of the m^{th} DR can be simplified to

$$\mathbb{E}\left[\widetilde{\text{SINR}}_m^d\right] \geq \frac{p_m L_m^d {}_1F_1(-1,1;-K)/(K+1)}{\bar{L}_m \sum_{i=1} P_i 2^{\frac{-B}{M-1}} + \sum_{j\neq m} p_j L_{j,m}^d + \sigma_m^{d^2}}. \tag{42}$$

Using (38) and (42), we formulate the optimization problem to obtain the optimal robust transmit powers with limited feedback and ZF beamforming as

$$\begin{aligned}
&\min_{\{P_n, p_m\}} \quad P_{\text{T}} && (43)\\
&\text{s.t.} \quad \frac{P_n L_n}{L_n \sum_{i\neq n} P_i 2^{\frac{-B}{M-1}} + \sum_{k=1} p_k L_{k,n} + \sigma_n^2} \geq \gamma_n, \ \forall n,\\
&\quad \frac{p_m L_m^d {}_1F_1(-1,1;-K)/(K+1)}{\bar{L}_m \sum_{i=1} P_i 2^{\frac{-B}{M-1}} + \sum_{j\neq m} p_j L_{j,m}^d + \sigma_m^{d^2}} \geq \gamma_m^d, \quad \forall m.
\end{aligned}$$

Note that problem (43) is a convex and linear optimization problem which can be solved efficiently using the power control scheme presented in Section IV. The limited feedback-based D2D underlaying cellular communication approach presented in this section can be easily extended to other codebooks with known mean quantization errors.

VI. Numerical Results

In this section, we present numerical results to assess the proposed schemes for the D2D underlaying MU MIMO cellular communication system. We compare the total transmission power requirements for the SDR scheme given by (6) and the fixed beamforming schemes: ZF, RZF and hybrid MRT-ZF given by (19), (27) and (33), respectively. The noise power is assumed to be equal for all the users (both CUs and DUs) such that $\sigma_n^2 = \sigma_m^{d^2} = \sigma^2 = -70$ dBm. The reason for selecting a higher noise power is due to the fact that 5G communication systems will have higher bandwidths compared to the conventional systems, which in turn increases the thermal noise [40]. In addition, other losses such as receiver noise figure (10 dBm) and implementation losses (5 dBm) also accounts for the higher noise power at the receiver. The desired and interfering channel vectors from the BS are assumed to follow the i.i.d. Rayleigh fading distribution with zero mean and unit variance. The SISO channel between the D2D pair follows the Rician fading distribution where the Rician K factor is 9 dB. The interfering channels from the given DT to the other non-serving DRs are assumed to be distributed according to the i.i.d. Rayleigh fading distribution with zero mean and unit variance. For simplicity, the SINR targets for all the CUs are assumed to be the same, i.e., $\gamma_n, \forall n$. Similarly, the SINR targets for all the DRs in the network are also considered to be equal, i.e., $\gamma_m^d, \forall m$. The path loss between the BS and

[1]To derive the expected SINR, similar to [32]–[36], we also rely on the expected SINR approximation given by (36). According to [34], the expected SINR approximation matches well with the true expected SINR for both large and moderate values of M. For the accuracy of the SINR approximation, we refer the readers to [34, Section III-B]. The accuracy of the approximation when used in the rate analysis is discussed in [35, Lemma I].

This article has been accepted for publication in a future issue of this journal, but has not been fully edited. Content may change prior to final publication. Citation information: DOI 10.1109/TVT.2018.2845748, IEEE Transactions on Vehicular Technology

$$\mathbb{E}\left[\widetilde{\mathrm{SINR}}_n\right] \approx \frac{P_n L_n \mathbb{E}\left[\|\mathbf{h}_n^H\|^2\right] \mathbb{E}\left[|\hat{\mathbf{h}}_n^H \tilde{\mathbf{w}}_n|^2\right]}{L_n \mathbb{E}\left[\|\mathbf{h}_n^H\|^2\right] \sum_{i \neq n} P_i \mathbb{E}\left[|\hat{\mathbf{h}}_n^H \tilde{\mathbf{w}}_i|^2\right] + \sum_{k=1} p_k L_{k,n} \mathbb{E}\left[|g_{k,n}|^2\right] + \sigma_n^2}. \tag{36}$$

CU/DR is given by $22\log_{10}(d) + 42 + 20\log_{10}(f_c/5)$ [41] (same path loss used between the DT and the CU and other non serving DRs), where d is the distance in meters (m) and f_c is the carrier frequency in GHz, here $f_c = 1.9$ GHz. Similarly, the path loss between the D2D pair is given by $16.9\log_{10}(d) + 46.8 + 20\log_{10}(f_c/5)$ [41].

A. Network Model

A hybrid network model consisting of both D2D pairs and CUs is considered in this study. The average number of BSs per unit area is denoted by λ_b and each BS is placed in the center of a hexagonal cell with an area given by $1/\lambda_b$. There are two types of UEs: cellular CUs and DTs which are uniformly distributed in a hexagonal grid of an area $\mathcal{A}$, where we denote the spatial location of the UEs by $\{x_i\}$. Therefore, the PPP $\{x_i\}$ is represented by $\tilde{\Phi} \in \mathbb{R}^2$ having an intensity λ_u. The type of UE, denoted by $\{t_i\}$, is assumed to be i.i.d. Bernoulli random variables with $\mathbb{P}(t_i = 1) = q$. In this study, the i^{th} UE is considered to be a DT if $t_i = 1$, otherwise a CU. Each DR is randomly positioned on the circle centered at the associated DT, where the distance between the D2D pair is uniformly distributed from 0 to R.

B. Simulation Setup

In the simulations, we follow the network model described above. The BSs are placed in the center of the hexagonal cell (with radius of 650m) in the hexagonal grid of area $\mathcal{A} = 4$km $\times$ 4km. The area of each hexagon is $1/\lambda_b$, where $\lambda_b = 0.911$ per km^2. The simulation steps are explained as follows:

- Generate a Poisson random variable, denoted by N, with parameter $\lambda_u \mathcal{A}$, where λ_u is the average number of UEs per unit area. Here, we set $\lambda_u = 10\lambda_b$. Next, generate N points that are uniformly distributed in $\mathcal{A}$.
- From these N points, DTs are selected using Bernoulli random variables with probability 0.2.
- Each DR is randomly placed on the circle centered at the associated DT with uniformly distributed radius between 20m to 100m.

A snapshot of the hybrid network for a given instance is shown in Fig. 1 comprising of BSs, CUs and DTs with the network parameters defined above. For clarity, we omit the positions of DRs which are isotropically placed in the circle centered at the serving DTs. The results presented in this section are the average of 1000 randomly generated instances.

C. Impact of SINR targets on the performance

In Fig. 2, the total transmit power, $P_T = \sum_n P_n + \sum_m p_m$, is plotted against different values of the CU target SINR, γ_n, with fixed value of DR SINR target $\gamma_m^d = 0$ dB. We

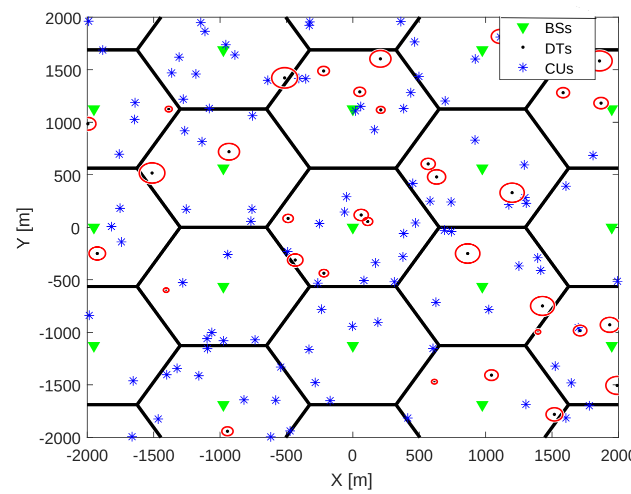

Fig. 1. A snapshot of the hybrid network model. For clarity, we omit plotting the location of DRs, where each DR is randomly located in the red circle centered around the serving DT.

observe that lower γ_n values yield lower transmit powers because the SINR constraints are more easily satisfied. The SDR scheme requires the least total transmit power compared to the fixed beamforming schemes. The ZF scheme always results in higher transmit powers compared to the other schemes presented. This is because the ZF scheme eliminates interference completely and cannot exploit it to satisfy the SINR constraints. Meanwhile, the RZF problem results in a large number of infeasibility instances, especially at higher γ_n values. For example, we note that in the RZF scheme, for the γ_n values of 0 dB, 2 dB and 4 dB, the infeasible instances are 10%, 70% and 90%, respectively. This indicates that structure of beamforming vectors are vital to decide the solution of the power minimization problem. For RZF beamforming, to satisfy large SINR thresholds is difficult and as this requires almost full cancellation of the interference [21], which is not possible with RZF beamforming vectors, thus resulting in a large number of infeasible instances. Results of fixed beamforming schemes demonstrate that the hybrid MRT-ZF scheme is the best compromise.

To examine the effects of the SINR threshold, γ_m^d, on the total transmit power, in Fig. 3, we plot P_T values against the range of γ_m values by setting the value of $\gamma_n = 0$ dB. We observe similar trends as in Fig. 2, where the SDR problem outperforms the fixed beamforming schemes by requiring the minimum total transmit power. The hybrid MRT-ZF scheme performs slightly better than the RZF scheme in the low to medium SINR range. However, it is notable that for higher values of γ_m^d (i.e. $\gamma_m^d > 10$ dB), the RZF scheme surpasses the performance of the hybrid MRT-ZF scheme. The reason for

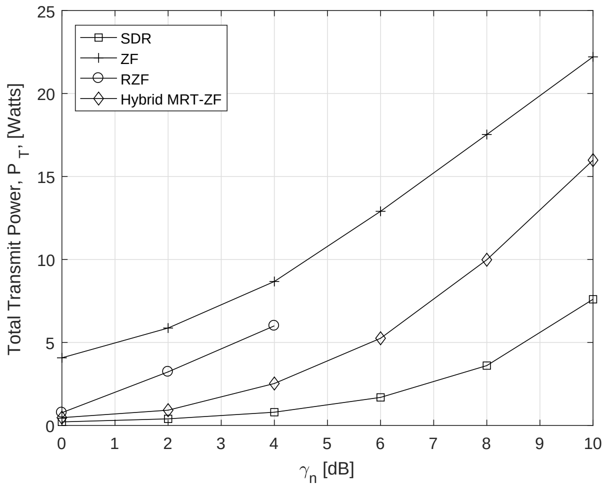

Fig. 2. The total transmit power versus CU SINR targets; $\gamma_m^d = 0$ dB, $C = 6$, $D = 2$, and $M = 8$.

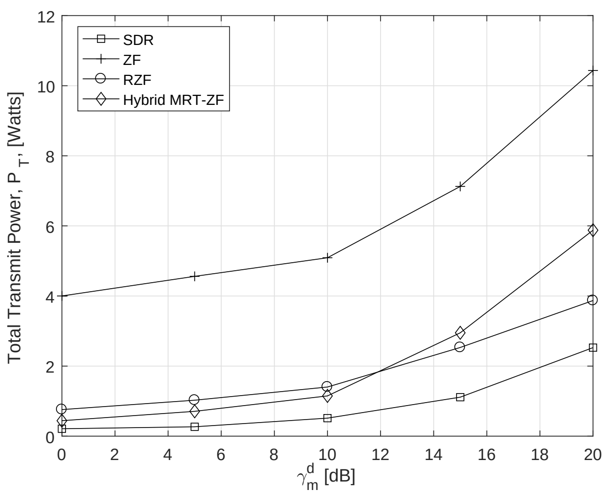

Fig. 3. The total transmit power versus DR SINR targets; $\gamma_n = 0$ dB, $C = 6$, $D = 2$, and $M = 8$.

this trend is that with large SINR threshold, γ_m^d, the solution of the hybrid MRT-ZF problem (29) results in values of x_n and y_n which in turn make $\mathbf{w}_n^{(\text{hyb})}$ in (24) to be inclined more towards ZF beamforming, which may yield negligible or no interference. Therefore, at higher γ_m^d values, unlike RZF beamforming, the hybrid MRT-ZF beamforming scheme is not able to exploit the interference to minimize the total transmission power. One interesting point is that unlike Fig. 2, the RZF scheme yields feasible solutions for the range of γ_m^d values considered. This implies that the infeasibility of the RZF scheme mainly depends on the γ_n values and not the γ_m^d values. It is also observed that increasing γ_m^d values does not increase the total transmit power requirements significantly unlike the case seen in Fig. 2 with increasing γ_n.

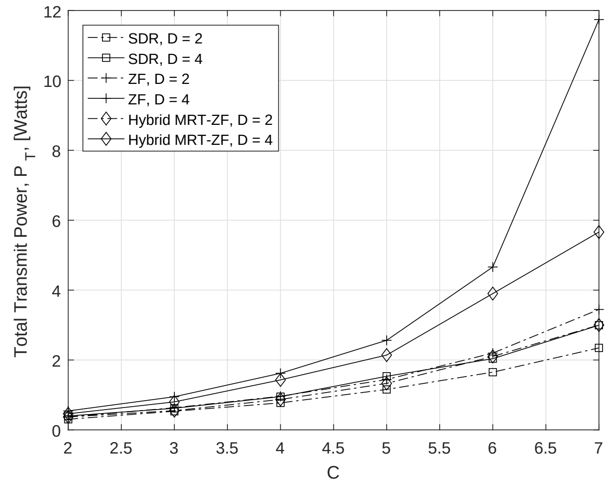

Fig. 4. The total transmit power versus the number of CUs; $\gamma_n = 10$ dB, $\gamma_m^d = 0$ dB, and $M = 12$.

D. Impact of the number of CUs and DRs on the performance

The total transmit power is plotted against varying number of CUs in Fig. 4 for SDR, ZF and hybrid MRT-ZF beamforming schemes with $D = 2$ and $D = 4$. For all the schemes, the transmit power values with the $D = 4$ case are higher than the $D = 2$ case. However, an interesting observation in the total transmit power is noticed for the ZF scheme with $D = 4$, where the total transmit power requirement increases significantly compared to the $D = 2$ case. For example, in the ZF scheme with $C = 8$ and $D = 2$, the total transmit power is 5.9 watts which is 70% less than the total transmit power with $D = 4$, i.e., 20 watts. This implies that the increase in the number of D2D pairs in the network has a detrimental effect on the performance of the ZF scheme. On the other hand, the total transmit power requirement with the SDR scheme increases slightly as the number of D2D pairs increases. This suggests that the performance of the SDR scheme is much less sensitive to the increasing number of D2D pairs in the network than the ZF and hybrid MRT-ZF schemes.

E. Impact of the number of antennas on the performance

Recently, the massive MIMO technology has drawn a considerable interest in the field of wireless communications due to its ability to improve spectral efficiency and energy efficiency [42]. Therefore, it is of interest to investigate the impact of the number of transmit antennas on the performance of the schemes studied in this paper. For this purpose, in Fig. 5, we evaluate the total transmit power for SDR, ZF and hybrid MRT-ZF schemes by varying the total number of transmit antennas at the BS. We note that the total transmit power requirement of the network decreases dramatically as the number of transmit antennas, M, increases. This is due to the fact that with increasing number of transmit antennas, the channel gain also increases, and thus, SINR constraints are more easily satisfied with lower transmit powers.

This article has been accepted for publication in a future issue of this journal, but has not been fully edited. Content may change prior to final publication. Citation information: DOI 10.1109/TVT.2018.2845748, IEEE Transactions on Vehicular Technology

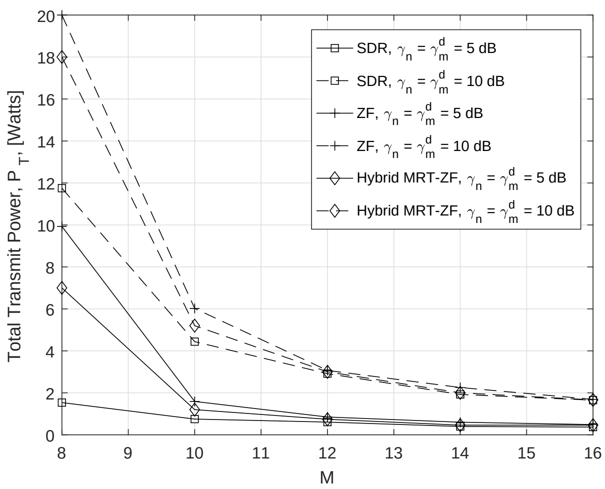

Fig. 5. The total transmit power versus the number of transmit antennas; $C = 6$ and $D = 2$.

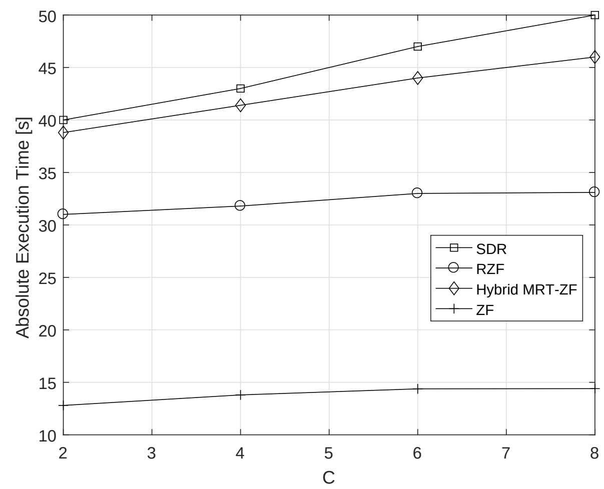

Fig. 7. Time required to compute the powers for SDR and fixed beamforming schemes for various values of C with $D = 2$, $\gamma_n = 0$ dB, $\gamma_m^d = 0$ dB, and $M = 12$.

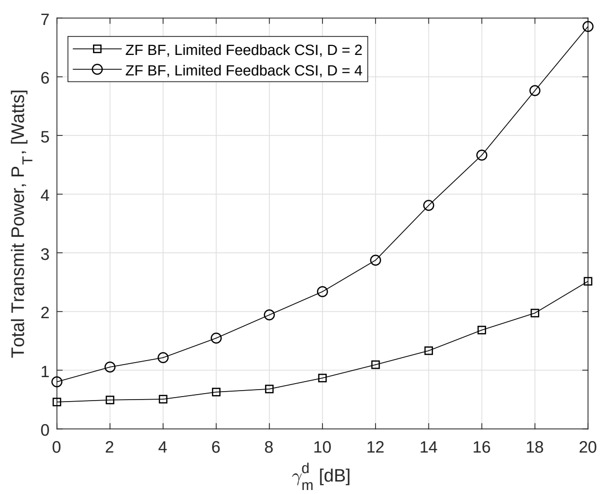

Fig. 6. The total transmit power versus DR SINR targets; $\gamma_n = 0$ dB, $C = 2$ and $M = C + D$.

F. Impact of Limited Feedback

In Fig. 6, we compare the performance of the ZF beamforming scheme with limited feedback based CSI at the BS. The limited feedback based transmit power optimization problem is given by (43). It is evident from Fig. 6 that the limited feedback ZF beamforming scheme with $D = 2$ has lower total transmit power requirement compared to the limited feedback ZF beamforming scheme with $D = 4$. We have also observed that the limited feedback approach is highly infeasible when $\gamma_n > 14$ dB and for large number of CUs.

G. Computational Complexity

Finally, we evaluate the complexity of computing the solutions for all the schemes by observing the total execution time. In Fig. 7, the absolute execution time (in seconds) is shown for various values of C with $D = 2$. Here, the absolute execution time is presented for 20 problem instances. The execution time for computing beamforming vectors and transmit powers for the SDR scheme is significantly higher than the other schemes. One reason for this high computational complexity is that solving a SDR problem in the optimal scheme is much more complicated than solving a second-order cone programming problem or a linear power control problem. As expected, the ZF problem has the lowest computational complexity compared to all other schemes. The hybrid MRT-ZF beamforming scheme has a higher complexity compared to other fixed beamforming schemes, as it has to rely on RZF when the second-order cone constraints in (33) are not tight.

From the numerical results, we observed that the SDR problem has the lowest transmit power requirements, but at the cost of the higher computational complexity. On the other hand, the ZF problem requires higher transmit powers to satisfy the SINR constraints but it needs significantly less execution time compared to the SDR scheme. For large M values (i.e. large number of transmit antennas), ZF, hybrid MRT-ZF and SDR schemes have the similar transmit power performance. Thus making the less computationally complex scheme, such as ZF, a suitable contender for D2D underlaying communication systems with massive MIMO.

VII. Conclusion

This paper investigated the D2D underlaying MU MIMO cellular system and proposed the joint beamforming and transmit power strategies. We studied both the optimal solution using SDR and fixed beamforming schemes with the aim to reduce the total transmit power consumption. It is observed that the SDR scheme has the best performance as it reduces the total transmit power significantly. One interesting point noted is that as the number of transmit antennas at the BS increases, the total transmit power of the ZF scheme approaches that of the SDR scheme. The proposed robust limited feedback based power optimization with the ZF beamforming scheme also shows promising performance compared to the performance with perfect CSI, given the fact that it can satisfy the average

This article has been accepted for publication in a future issue of this journal, but has not been fully edited. Content may change prior to final publication. Citation information: DOI 10.1109/TVT.2018.2845748, IEEE Transactions on Vehicular Technology

SINR requirements and does not require full CSI for the optimization of transmit powers at the BS. For future work, it will be interesting to integrate different technologies such as massive MIMO, millimeter wave with D2D which may further conserve the total transmit power requirements.

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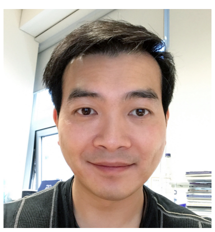

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