DISTRIBUTED PEER-TO-PEER MULTIPLEXING USING AD HOC RELAY NETWORKS

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ABSTRACT

We consider an ad hoc network consisting of d source-destination pairs and R relaying nodes. Each source wishes to transmit its data to its corresponding destination through the relay network. Each relay in the network transmits a properly scaled version of its received signal thereby cooperating with other relays to deliver each source's data to the corresponding destination. Assuming a minimal cooperation among the relaying nodes, we design a distributed beamformer such that the total relay transmit power dissipated by all relays is minimized while, at the same time, the quality of services at all destinations are guaranteed to be above certain pre-defined thresholds. We show that using a semi-definite relaxation approach, the power minimization problem can be turned into a semi-definite programming (SDP) optimization, and therefore, it can be solved efficiently using interior point methods. Our results show that the distributed relay multiplexing is possible and may be beneficial depending on the channel conditions.

***Index Terms*—** Distributed beamforming, relay networks, semi-definite programming, convex feasibility problem, distributed signal processing.

1. INTRODUCTION

Signal fading is a major source of impairment in wireless networks which can significantly affect the reliability of communications. Implementing multiple antennas at the transmitter or receiver results in a remarkable performance gain as it can provide independently faded versions of the transmitted signal. In multiple-antenna communications, transmit and receive beamforming can be used to increase the capacity of the wireless channel through increasing the signalling range and mitigating the inter-user interference [1]. Despite all these potential benefits, implementing multiple transmit antennas in mobile terminals is impractical due to the size of the mobile unit, complexity and power limitations. To overcome these practical restrictions, user cooperative schemes are proposed in the literature [2]- [4]. In these schemes, users share their resources in order to transmit each others data. So each user acts as a relay for other users in specific time slots. In contrast to this paper, almost in all previous works related to cooperative communications, the mobile terminals cooperate to transmit data from a single source to a single destination [3]- [5]. In such cooperative schemes, single-antenna mobile terminals can be viewed as elements of an antenna array, thus acting collectively as a virtual multiple-antenna system. Therefore, one can exploit multiple-antenna communications techniques, such as beamforming and space-time coding, in the context of relay network. Several protocols have been proposed to achieve spatial diversity using cooperation [6], [7]. Amplify-and-forward (AF), decode-and-forward (DF), compress-and-forward and coded cooperation [8] are the most popular schemes. Space-time coding can also be used at relays to achieve full spatial diversity gain at the destination node [9].

Among these cooperative schemes, the AF approach is of particular interest as it can be easily implemented. In [5], a distributed beamforming technique has been developed for a relay network which consists of a transmitter, a receiver, and several relaying nodes. Assuming that the *instantaneous* channel state information is available at the relays and at the receiver, Jing and Jafarkhani design a distributed beamforming technique through maximization of the receiver signal-to-noise ratio (SNR) for a limited power per relay. In [10], a decentralized beamforming algorithm is developed based on the assumption that only the *second order statistics* of the channel coefficients are available at the receiver.

In this paper, we consider an ad hoc network consisting of d source-destination pairs and R relaying nodes. Each pair must communicate their data through the relay network. Our cooperative scheme consists of two phases. All sources transmit their symbols to the relays in the first phase. In the second phase, each relays amplifies and phase steers its received signal and forwards it to the destinations. Assuming that the correlation matrices of all communication channels are available, our goal is to optimize the complex gain of the relays such that the signal-to-interference-plus-noise ratios (SINRs) at all destinations are above specific targets. To this end, we minimize the sum of the powers dissipated by the relays. The major difference between our communication scheme and those considered in downlink beamforming literature (e.g. [11]) is that in our approach, all the sources transmit their signals at the same time. Therefore, each relay receives a mixture of all information symbols transmitted by all sources. However, in downlink beamforming schemes, the signal intended for each user is available separately at the transmitting antennas. We show that using a semi-definite relaxation approach, our power minimization problem can be turned into a semi-definite programming (SDP) optimization, and therefore, it can be solved efficiently using interior point methods. Interestingly enough, in all our numerical simulations, the solution to the SDP problem turned out to be rank one, and therefore, the solution to the power minimization problem can be easily obtained from that of the SDP problem.

2. SYSTEM MODEL

Consider a network which consists of d source-destination pairs and R relaying nodes. Each source in a pair wishes to transmit its information symbols to the corresponding destination. However, there is no direct connection between any source and its respective destination. All sources transmit their symbols to the relays in the first time slot. The relays receive mixtures of all source signals. The relay network is then responsible to deliver the data to the respective destination. In this paper, we consider an amplify-and-forward approach. That is, each relay retransmits an amplitude- and phase-adjusted version of its received signal.

Let f_{rp} denote the channel coefficient from the pth source to the rth relay and g_{rp} represent the channel coefficient from the rth relay to the pth destination. Then, the rth relay received signal x_r is given by

$$x_r = \sum_{p=1}^{d} f_{rp} s_p + \nu_r, \quad r = 1, \cdots, R \tag{1}$$

where ν_r is the noise at the rth relay node and s_p is the information symbol transmitted by the pth source. We use the following assumptions throughout the paper:

A1. The relay noise is spatially white, i.e., $E\{\nu_r \nu_{r'}^*\} = \sigma_\nu^2 \delta_{rr'}$, where $E\{\cdot\}$ is the statistical expectation, σ_ν^2 is the relay noise power, $(\cdot)^*$ denotes the complex conjugate, and $\delta_{rr'}$ is Kronecker's delta function.

A2. The power of the pth source is P_p, i.e., $E\{|s_p|^2\} = P_p$.

A3. The symbols transmitted by sources are uncorrelated, that is $E\{s_p s_q^*\} = P_p \delta_{pq}$.

A4. The rth relay noise ν_r and the information symbols $\{s_p\}_{p=1}^d$ are statistically independent.

Using vector notations, we can rewrite (1) as

$$\mathbf{x} = \sum_{p=1}^{d} \mathbf{f}_p s_p + \boldsymbol{\nu} \tag{2}$$

where $\mathbf{x} \triangleq [x_1 \quad x_2 \quad \ldots \quad x_R]^T$, $\boldsymbol{\nu} \triangleq [\nu_1 \quad \nu_2 \quad \ldots \quad \nu_R]^T$, $\mathbf{f}_p \triangleq [f_{1p} \quad f_{2p} \quad \ldots \quad f_{Rp}]^T$, and $(\cdot)^T$ denotes the transpose. The rth relay multiplies its received signal by a complex weight w_r^*. Hence, the vector of the signals transmitted by all relays can be expressed as

$$\mathbf{t} = \mathbf{W}^H \mathbf{x} \tag{3}$$

where rth entry of $\mathbf{t} \in \mathbb{C}^R$ is the signal transmitted by the rth relay, $\mathbf{W}$ is a diagonal matrix with its rth diagonal entry equal to w_r, and $(\cdot)^H$ is the Hermitian transpose.

Let us denote the vector of the channel coefficients from the relays to the ith destination as $\mathbf{g}_k = [g_{1k} \quad g_{2k} \quad \ldots \quad g_{Rk}]^T$. Then the signal received by the kth destination can be written as

$$\begin{aligned} y_k &= \mathbf{g}_k^T \mathbf{t} = \mathbf{g}_k^T \mathbf{W}^H \sum_{p=1}^{d} \mathbf{f}_p s_p + \mathbf{g}_k^T \mathbf{W}^H \boldsymbol{\nu} + n_k \\ &= \underbrace{\mathbf{g}_k^T \mathbf{W}^H \mathbf{f}_k s_k}_{\text{signal}} + \underbrace{\mathbf{g}_k^T \mathbf{W}^H \sum_{p=1, p\neq k}^{d} \mathbf{f}_p s_p}_{\text{interference}} + \underbrace{\mathbf{g}_k^T \mathbf{W}^H \boldsymbol{\nu} + n_k}_{\text{noise}} \end{aligned} \tag{4}$$

where n_k is the zero-mean additive noise at the kth destination with a variance of σ_n^2. We make one more assumption:

A5 The source signals $\{s_p\}_{p=1}^d$, the relay noise $\boldsymbol{\nu}$, kth destination noise n_k, the channel coefficients $\{\mathbf{g}_k\}_{k=1}^d$ and $\{\mathbf{f}_p\}_{p=1}^d$ are statistically independent.

3. POWER MINIMIZATION

In this section, we aim to find the relay weights $\{w_r\}_{r=1}^R$ such that the total transmit power dissipated by the relay network is minimized while maintaining the destinations' quality of services (QoSs) above certain pre-defined thresholds. We herein use the SINR as our measure of QoS. Therefore, our goal is to solve the following constrained optimization problem:

$$\begin{aligned} \min_{\mathbf{w}} \quad & P_T \\ \text{subject to} \quad & \text{SINR}_k \geq \gamma_k, \quad \text{for} \quad k = 1, 2, \ldots, d \end{aligned} \tag{5}$$

where P_T is the total relay transmit power and SINR_k is the SINR at the kth destination defined by

$$\text{SINR}_k = \frac{P_{\text{s}}^k}{P_{\text{i}}^k + P_{\text{n}}^k}. \tag{6}$$

Here, P_{s}^k, P_{i}^k, and P_{n}^k are the desired signal component power, the interference power, and the noise power at the kth destination, respectively. Using (3), the total transmit power is expressed as

$$P_T = E\left\{\mathbf{t}^H \mathbf{t}\right\} = E\left\{\mathbf{x}^H \mathbf{W} \mathbf{W}^H \mathbf{x}\right\} = \text{tr}\left\{\mathbf{W}^H E\left\{\mathbf{x}\mathbf{x}^H\right\} \mathbf{W}\right\} \tag{7}$$

where tr$\{\cdot\}$ denotes the trace operator. Let us define the correlation matrix of the relay transmitted signals as $\mathbf{R}_x \triangleq E\{\mathbf{x}\mathbf{x}^H\}$. Then, the total transmit power can be written as

$$P_T = \text{tr}\left\{\mathbf{W}^H \mathbf{R}_x \mathbf{W}\right\} = \sum_{r=1}^{R} |w_r|^2 \mathbf{R}_x(r,r) = \mathbf{w}^H \mathbf{D} \mathbf{w} \tag{8}$$

where $\mathbf{A}(r,s)$ represents the (r,s) entry of matrix $\mathbf{A}$, and $\mathbf{D} \triangleq \text{diag}\left([\mathbf{R}_x(1,1), \mathbf{R}_x(2,2), \cdots, \mathbf{R}_x(R,R)]\right)$. Throughout this paper, diag$(\mathbf{a})$ denotes a diagonal matrix with the elements of the vector $\mathbf{a}$ as its diagonal entries and diag$(\mathbf{A})$ returns the diagonal entries of the matrix $\mathbf{A}$ as a vector. Using (2) and assumptions A1-A4, the matrix $\mathbf{R}_x$ can be written as

$$\begin{aligned} \mathbf{R}_x &= \sum_{p,q=1}^{d} E\left\{\mathbf{f}_p \mathbf{f}_q^H\right\} E\{s_p s_q^*\} + \sigma_\nu^2 \mathbf{I} \\ &= \sum_{p=1}^{d} P_p E\left\{\mathbf{f}_p \mathbf{f}_p^H\right\} + \sigma_\nu^2 \mathbf{I} = \sum_{p=1}^{d} P_p \mathbf{R}_f^p + \sigma_\nu^2 \mathbf{I} \end{aligned} \tag{9}$$

where $\mathbf{R}_f^p \triangleq E\left\{\mathbf{f}_p \mathbf{f}_p^H\right\}$. Note that (8) implies that the transmitted power P_T depends on the variances of the channel coefficients of the source-relay paths as well as the relay noise powers. To derive an expression for the SINR at the kth destination, we express the desired signal component power P_{s}^k, the interference power P_{i}^k, and the noise power P_{n}^k in terms of $\{w_r\}_{r=1}^R$.

Using assumptions A1, A5 and (4), the noise power at the kth destination can be written as

$$\begin{aligned} P_{\text{n}}^k &= E\left\{\boldsymbol{\nu}^H \mathbf{W} \mathbf{g}_k^* \mathbf{g}_k^T \mathbf{W}^H \boldsymbol{\nu}\right\} + \sigma_n^2 \\ &= \text{tr}\left\{\mathbf{W}^H E\left\{\boldsymbol{\nu}\boldsymbol{\nu}^H\right\} \mathbf{W} E\left\{\mathbf{g}_k^* \mathbf{g}_k^T\right\}\right\} + \sigma_n^2 \\ &= \sigma_\nu^2 \text{tr}\left\{\mathbf{W}^H \mathbf{R}_g^k \mathbf{W}\right\} + \sigma_n^2 \end{aligned}$$

where $\mathbf{R}_g^k \triangleq E\left\{\mathbf{g}_k\mathbf{g}_k^H\right\}$. Thus, the noise power $P_{\rm n}^k$ is given by

$$P_{\rm n}^k = \sigma_\nu^2 \sum_{r=1}^{R} |w_r|^2 \mathbf{R}_g^k(r,r) + \sigma_n^2 = \mathbf{w}^H \mathbf{D}_k \mathbf{w} + \sigma_n^2 \qquad (10)$$

where $\mathbf{D}_k \triangleq \sigma_\nu^2 \ \mathrm{diag}\left(\left[\mathbf{R}_g^k(1,1), \mathbf{R}_g^k(2,2), \cdots, \mathbf{R}_g^k(R,R)\right]\right)$, and $\mathbf{w} \triangleq \mathrm{diag}(\mathbf{W})$.

The kth desired signal power can be written as

$$\begin{aligned} P_{\rm s}^k &= E\left\{\mathbf{g}_k^T \mathbf{W}^H \mathbf{f}_k \mathbf{f}_k^H \mathbf{W} \mathbf{g}_k^*\right\} E\{|s_k|^2\} \\ &= P_k E\left\{\mathbf{w}^H \mathrm{diag}(\mathbf{g}_k) \mathbf{f}_k \mathbf{f}_k^H \mathrm{diag}(\mathbf{g}_k^*) \mathbf{w}\right\} \\ &= P_k E\left\{\mathbf{w}^H (\mathbf{g}_k \odot \mathbf{f}_k)(\mathbf{f}_k^H \odot \mathbf{g}_k^H) \mathbf{w}\right\} \\ &= P_k \mathbf{w}^H E\{\mathbf{h}_k \mathbf{h}_k^H\} \mathbf{w} = \mathbf{w}^H \mathbf{R}_h^k \mathbf{w} \end{aligned} \qquad (11)$$

where $\odot$ stands for Schur-Hadamard (element-wise) multiplication, $\mathbf{h}_k \triangleq (\mathbf{g}_k \odot \mathbf{f}_k) = [f_{1k}\,g_{1k} \quad f_{2k}\,g_{2k} \quad \cdots \quad f_{Rk}\,g_{Rk}]^T$, and $\mathbf{R}_h^k \triangleq P_k E\{\mathbf{h}_k \mathbf{h}_k^H\}$. It is worth mentioning that the rth entry of the vector $\mathbf{h}_k$ represents the path gain between kth source to its corresponding destination via the rth relay (excluding the relay gain).

Using (4), the interference power at the kth destination can be also written as

$$\begin{aligned} P_{\rm i}^k &= E\left\{\mathbf{g}_k^T \mathbf{W}^H \left(\sum_{p,q \in \mathcal{D}-\{k\}} \mathbf{f}_p \mathbf{f}_q^H s_p s_q^*\right) \mathbf{W} \mathbf{g}_k^*\right\} \\ &= E\left\{\mathbf{w}^H \mathrm{diag}(\mathbf{g}_k) \left(\sum_{p \in \mathcal{D}-\{k\}} P_p \mathbf{f}_p \mathbf{f}_p^H\right) \mathrm{diag}(\mathbf{g}_k^*) \mathbf{w}\right\} \\ &= E\left\{\mathbf{w}^H \left(\sum_{p \in \mathcal{D}-\{k\}} P_p (\mathbf{g}_k \odot \mathbf{f}_p)(\mathbf{f}_p^H \odot \mathbf{g}_k^H)\right) \mathbf{w}\right\} \\ &= \mathbf{w}^H \mathbf{Q}_k \mathbf{w} \end{aligned} \qquad (12)$$

where $\mathcal{D} = \{1, 2, \cdots, d\}$, $\mathbf{Q}_k \triangleq E\left\{\sum_{p \in \mathcal{D}-\{k\}} P_p \mathbf{h}_k^p (\mathbf{h}_k^p)^H\right\}$ and $\mathbf{h}_k^p \triangleq \mathbf{g}_k \odot \mathbf{f}_p$. The vector $\mathbf{h}_k^p$ contains the coefficients of the multiple paths from the pth sources to the kth destination that are passing through the R relays.

Summarizing (8), (10) (11), and (12), the optimization problem in (5) can be rewritten as

$$\begin{aligned} &\min_{\mathbf{w} \in \mathbb{C}^R} \quad \mathbf{w}^H \mathbf{D} \mathbf{w} \\ &\text{subject to} \quad \frac{\mathbf{w}^H \mathbf{R}_h^k \mathbf{w}}{\mathbf{w}^H (\mathbf{Q}_k + \mathbf{D}_k) \mathbf{w} + \sigma_n^2} \geq \gamma_k, \ \text{for} \quad \forall k \in \mathcal{D} \end{aligned} \qquad (13)$$

or, equivalently, as

$$\begin{aligned} &\min_{\mathbf{w} \in \mathbb{C}^R} \quad \mathbf{w}^H \mathbf{D} \mathbf{w} \\ &\text{subject to} \quad \mathbf{w}^H (\mathbf{R}_h^k - \gamma_k (\mathbf{Q}_k + \mathbf{D}_k)) \mathbf{w} \geq \gamma_k \sigma_n^2, \ \forall k \in \mathcal{D}. \end{aligned} \qquad (14)$$

In general, the optimization problem in (14) is not convex and may not be amenable to a computationally affordable solution. We propose to use the semi-definite relaxation to approximately solve (14). To do so, let us define $\mathbf{X} \triangleq \mathbf{w}\mathbf{w}^H$. In this case, the problem in (14) becomes

$$\begin{aligned} &\min_{\mathbf{X} \in \mathbb{C}^{R \times R}} \quad \mathrm{tr}(\mathbf{D}\mathbf{X}) \\ &\text{subject to} \quad \mathrm{tr}(\mathbf{T}_k \mathbf{X}) \geq \gamma_k \sigma_n^2, \text{for} \quad \forall k \in \mathcal{D} \\ &\text{and} \quad \mathrm{rank}\,(\mathbf{X}) = 1, \quad \mathbf{X} \succeq \mathbf{0} \end{aligned} \qquad (15)$$

where $\mathrm{rank}(\cdot)$ denotes the rank of a matrix, $\mathbf{X} \succeq \mathbf{0}$ denotes that $\mathbf{X}$ is a positive semi-definite matrix, and $\mathbf{T}_k \triangleq \mathbf{R}_h^k - \gamma_k(\mathbf{Q}_k + \mathbf{D}_k)$. The rank constraint in (15) makes the above problem non-convex. We relax this constraint and solve the following convex problem

$$\begin{aligned} &\min_{\mathbf{X} \in \mathbb{C}^{R \times R}} \quad \mathrm{tr}(\mathbf{D}\mathbf{X}) \\ &\text{subject to} \quad \mathrm{tr}(\mathbf{T}_k \mathbf{X}) \geq \gamma_k \sigma_n^2 \text{ , for} \quad \forall k \in \mathcal{D} \\ &\text{and} \quad \mathbf{X} \succeq \mathbf{0} \end{aligned} \qquad (16)$$

The optimization problem in (16) is convex and can be efficiently solved using interior point based packages such as SeDuMi [15]. Due to the relaxation, the matrix $\mathbf{X}_{\rm opt}$ obtained by solving the optimization problem in (16) will not be of rank one in general. If $\mathbf{X}_{\rm opt}$ happens to be rank one, then its principal component will be the optimal solution to the original problem in (15). Otherwise, one has to resort to randomization techniques developed in [14] to obtain a rank-one solution from $\mathbf{X}_{\rm opt}$. Interestingly, in our extensive simulation results, we have never encountered a case where the solution to the SDP problem has a rank higher than one. We have yet to prove that the solution of the SDP problem (16) is rank one. This will be the focus of a future research.

4. SIMULATIONS

In our numerical examples, we consider a network with $R = 20$ relay nodes. The channel coefficients f_{rp} and $g_{r'p'}$ are assumed to be independent from each other for any p, p', r, and r'. We also assume that the channel coefficient f_{ip} can be written as $f_{rp} = \bar{f}_{rp} + \tilde{f}_{rp}$ where $\bar{f}_{rp}$ is the (known) mean of f_{rp} and $\tilde{f}_{rp}$ is a zero-mean random variable. It is assumed that $\tilde{f}_{rp}$ and $\tilde{f}_{r'p}$ are independent for $r \neq r'$. We choose $\bar{f}_{rp} = \frac{e^{j\theta_{rp}}}{\sqrt{1+\alpha_f}}$ and $\mathrm{var}(\tilde{f}_{rp}) = \frac{\alpha_f}{1+\alpha_f}$, where θ_{rp} is a uniform random variable chosen from the interval $[0 \ \ 2\pi]$ and α_f is a parameter which determines the level of uncertainty in the channel coefficient f_{rp}. Note that as $E\{|f_{rp}|^2\} = 1$, if α_f is increased, the variance of the random component $\tilde{f}_{rp}$ is increased while its mean is decreased. This, in turn, means that the level of the uncertainty in the channel coefficient f_{rp} is increased. Similarly, we model the channel coefficient g_{rp} as $g_{rp} = \bar{g}_{rp} + \tilde{g}_{rp}$ where $\bar{g}_{rp}$ is the mean of g_{rp} and $\tilde{g}_{rp}$ is a zero-mean random variable. We assume that $\tilde{g}_{rp}$ and $\tilde{g}_{r'p}$ are independent for $r \neq r'$. We choose $\bar{g}_{rp} = \frac{e^{j\phi_{rp}}}{\sqrt{1+\alpha_g}}$ and $\mathrm{var}(\tilde{g}_{rp}) = \frac{\alpha_g}{1+\alpha_g}$, where ϕ_{rp} is a uniform random variable chosen from the interval $[0 \ \ 2\pi]$ and α_g is a parameter which determines the level of uncertainty in the channel coefficient g_{rp}. Based on this channel modeling, we can write the (r, r') entry of the matrices $\mathbf{R}_f^p$, $\mathbf{R}_g^k$, $\mathbf{R}_h^k$, and $\mathbf{Q}_k$ respectively, as

$$\begin{aligned} \mathbf{R}_f^p(r,r') &= (\bar{f}_{rp}\bar{f}_{r'p}^* + \frac{\alpha_f}{1+\alpha_f}\delta_{rr'}) \\ \mathbf{R}_g^k(r,r') &= (\bar{g}_{rk}\bar{g}_{r'k}^* + \frac{\alpha_g}{1+\alpha_g}\delta_{rr'}) \\ \mathbf{R}_h^k(r,r') &= P_k \mathbf{R}_f^k(r,r') \odot \mathbf{R}_g^k(r,r') \\ \mathbf{Q}_k(r,r') &= \sum_{p=1, p\neq k}^{d} P_p \mathbf{R}_f^p(r,r') \odot \mathbf{R}_g^k(r,r')\,. \end{aligned}$$

Also, throughout our numerical examples, the transmit powers P_p are 10 dB above the noise power which is at 0 dB. Fig. 1 illustrates the minimum transmit power versus SINR threshold, for different

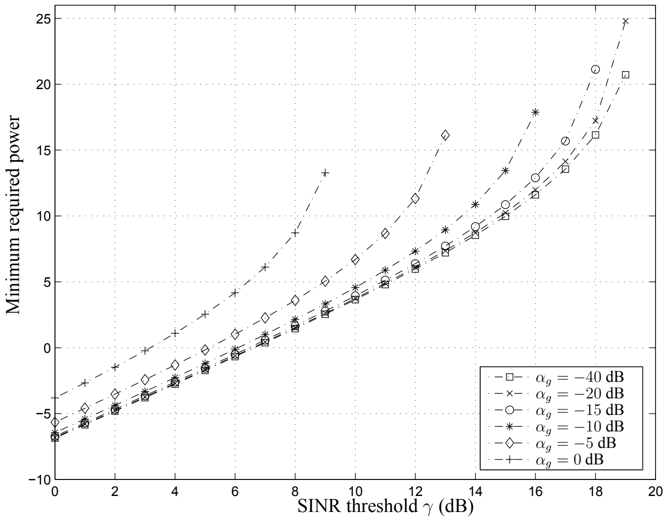

Fig. 1. Minimum transmit power versus SINR threshold, for different values of α_g and for $\alpha_f = -20$ dB.

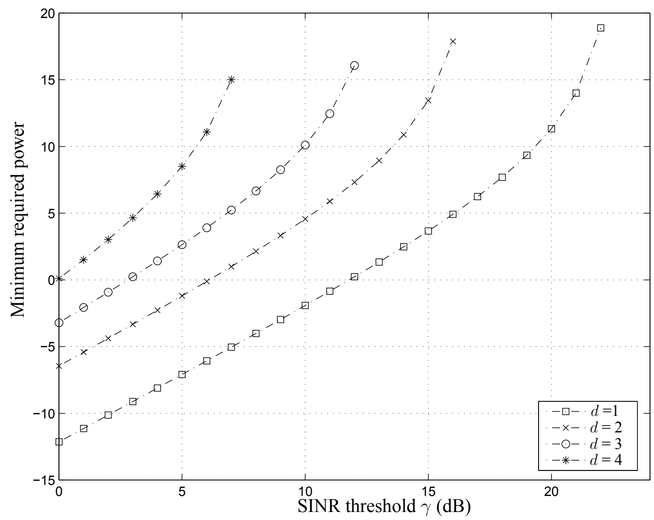

Fig. 2. Minimum transmit power versus SINR threshold, for different number of users, for $\alpha_f = -20$ dB and $\alpha_g = -10$ dB.

values of α_g, and for $\alpha_f = -20$ dB and $d = 2$. As one might expect, as the level of uncertainty in the channel coefficients is increased, the QoS constraints become more and more difficult to satisfy. It is worth mentioning that when $\alpha_g = -40$ dB, the forward channel coefficients are almost perfectly known.

In Fig. 2, the minimum transmit power versus SINR threshold is plotted for different number of source-destination pairs and for $\alpha_f = -20$ dB and $\alpha_g = -10$ dB. As can be seen from this figure, as the number of source-destination pairs is increased, it takes more power to satisfy the QoS constraints.

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