FILTER-AND-FORWARD DISTRIBUTED BEAMFORMING FOR RELAY NETWORKS IN FREQUENCY SELECTIVE FADING CHANNELS

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ABSTRACT

A half-duplex distributed beamforming technique for relay networks with frequency selective fading channels is developed. The network relays use the filter-and-forward (FF) strategy to compensate for the transmitter-to-relay and relay-to-destination channels using finite impulse response (FIR) filters. With the channel state information (CSI) being available at the receiver, the transmit relay power is minimized subject to the destination quality-of-service (QoS) constraint. This distributed beamforming problem is shown to have a closed-form solution. Simulation results demonstrate substantial improvements in terms of the relay transmitted power and feasibility of the destination QoS constraint as compared to amplify-and-forward (AF) distributed beamforming techniques.

***Index Terms*—** Cooperative communications, distributed beamforming, relay networks, filter-and-forward

1. INTRODUCTION

Recently, cooperative wireless communication techniques have attracted much attention in the literature because they can exploit cooperative diversity without any need of employing multiple antennas at each user [1]-[4]. In such user cooperative schemes, different users share their communication resources to assist each other in transmitting the information data through the network.

Different relaying strategies have been proposed to achieve spatial diversity via user cooperation. Two most popular relaying approaches are the amplify-and-forward (AF) and decode-and-forward (DF) strategies [1], [2]. In the AF scheme, relays forward scaled versions of their received signals, while in the DF scheme, relays decode their received messages first, and then retransmit their re-encoded versions. Due to its simplicity, the AF relaying strategy is of considerable practical interest.

Transmit power minimization problems are important for wireless relay networks as, typically, the user battery life is rather limited. Moreover, the relay transmit powers have to be limited to reduce the network interference. Many of recent works have developed algorithms for minimizing or limiting the relay transmit powers in relay networks [3]-[6]. For example, several distributed beamforming techniques using the AF strategy have been recently proposed [5], [6] that minimize the total relay transmit power subject to the receive signal-to-noise ratio (SNR) constraint or, alternatively, maximize the receive SNR subject to relay transmit power constraints. All of these distributed beamforming techniques assume the transmitter-to-relay and relay-to-destination channels to be frequency flat. If these channels become frequency selective, then there is a significant amount of inter-symbol interference (ISI) which makes it difficult to directly extend the techniques of [5] and [6] to the frequency selective fading channel case.

This work was supported by the European Research Council under Advanced Investigator Grant Scheme.

In this paper, we consider a relay network of one transmitter, one destination, and multiple relays with frequency selective finite impulse response (FIR) transmitter-to-relay and relay-to-destination channels. To compensate for the effect of these channels, a filter-and-forward (FF) strategy is employed at each relay by means of FIR adaptive filtering. The distributed beamforming technique proposed below minimizes the transmit power of the relays subject to the quality-of-service (QoS) constraint at the destination. In contrast to the flat fading case (where the QoS has been characterized in terms of SNR), we use the signal-to-interference-plus-noise ratio (SINR) as a measure of QoS. It is shown that this distributed beamforming problem has a simple closed-form solution. Simulation results demonstrate substantial performance improvements of the proposed distributed beamforming approach relative to the AF beamforming technique of [6].

2. SIGNAL MODEL

Let us consider a half-duplex relay network with one single-antenna transmitting source, one single-antenna destination and R single-antenna FF relays, as shown in Fig. 1. We assume that there is no direct link between the transmitter and destination, and each transmission consists of two stages. In the first stage, the transmitter broadcasts its data to the relays. The signals received at the relays are passed through the relay FIR filters to compensate for the transmitter-to-relay and relay-to-destination frequency selective channels. In the second stage, the outputs of each relay filter are sent to the destination. The destination is assumed to have the full channel state information (CSI) and, based on this knowledge, it determines the filter coefficients of each relay according to a certain beamforming criterion. We also assume that there is a low-rate feedback link from the destination to each relay that is used to inform the relays about their optimal weight coefficients.

The transmitter-to-relay and relay-to-destination channels can be modeled as linear FIR filters

$$\mathbf{f}(\omega) = \sum_{l=0}^{L_f-1} \mathbf{f}_l e^{-j\omega l}, \qquad \mathbf{g}(\omega) = \sum_{l=0}^{L_g-1} \mathbf{g}_l e^{-j\omega l} \qquad (1)$$

respectively, where $\mathbf{f}_l$ and $\mathbf{g}_l$ are the $R \times 1$ channel impulse response

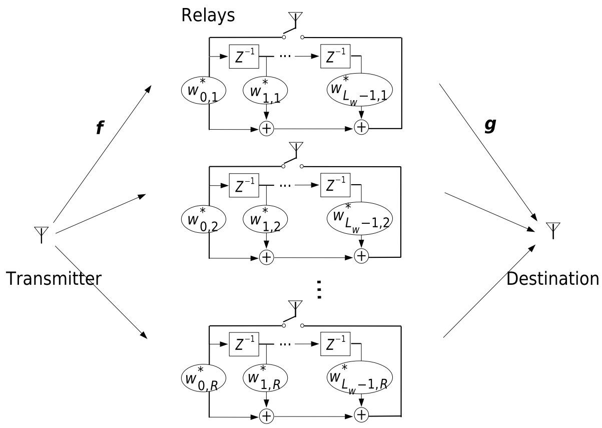

Fig. 1. System model of the relay network.

vectors corresponding to the lth effective tap of the transmitter-to-relay and relay-to-destination channels, $\mathbf{f}(\omega)$ and $\mathbf{g}(\omega)$ are the $R \times 1$ vectors of channel frequency responses, and L_f and L_g are the channel lengths. The $R \times 1$ vector $\mathbf{r}(n) = [r_1(n), \cdots, r_R(n)]^T$ of signals received by the relays at time n can be written as

$$\mathbf{r}(n) = \sum_{l=0}^{L_f-1} \mathbf{f}_l s(n-l) + \boldsymbol{\eta}(n) \tag{2}$$

where $s(n)$ denotes the signal transmitted by the source, $\boldsymbol{\eta}(n) = [\eta_1(n), \cdots, \eta_R(n)]^T$ is the $R \times 1$ vector of relay noise, and $(\cdot)^T$ denotes the transpose. Introducing

$$\begin{aligned} \mathbf{F} &\triangleq [\mathbf{f}_0, \cdots, \mathbf{f}_{L_f-1}] \\ \mathbf{s}(n) &\triangleq [s(n), s(n-1), \cdots, s(n-L_f+1)]^T \end{aligned}$$

we can rewrite (2) as

$$\mathbf{r}(n) = \mathbf{F}\mathbf{s}(n) + \boldsymbol{\eta}(n). \tag{3}$$

The signal $\mathbf{t}(n) = [t_1(n), \cdots, t_R(n)]^T$ transmitted by the relays to the destination can be expressed as

$$\mathbf{t}(n) = \sum_{l=0}^{L_w-1} \mathbf{W}_l^H \mathbf{r}(n-l) \tag{4}$$

where $\mathbf{W}_l = \mathrm{diag}\{w_{l,1}, \cdots, w_{l,R}\}$ is the diagonal matrix of relay filter impulse responses corresponding to the lth effective filter taps of the relays, L_w is the length of the relay FIR filters, and $(\cdot)^H$ denotes the Hermitian transpose. Hereafter, for any vector $\boldsymbol{x}$, $\mathrm{diag}\{\boldsymbol{x}\}$ denotes the diagonal matrix containing the entries of $\boldsymbol{x}$ on its main diagonal. For any square matrix $\boldsymbol{X}$, $\mathrm{diag}\{\boldsymbol{X}\}$ denotes the vector formed from the diagonal entries of $\boldsymbol{X}$. Let

$$\begin{aligned} \mathbf{W} &\triangleq [\mathbf{W}_0, \cdots, \mathbf{W}_{L_w-1}]^T \\ \boldsymbol{\mathcal{F}} &\triangleq \begin{bmatrix} \mathbf{f}_0 & \mathbf{f}_1 & \cdots & \mathbf{f}_{L_f-1} & \overbrace{\mathbf{0}_{R\times 1} \quad \cdots \quad \mathbf{0}_{R\times 1}}^{L_w-1 \text{ columns}} \\ \mathbf{0}_{R\times 1} & \mathbf{f}_0 & \mathbf{f}_1 & \cdots & \mathbf{f}_{L_f-1} \quad \cdots \quad \mathbf{0}_{R\times 1} \\ & & \ddots & & \\ \mathbf{0}_{R\times 1} & \mathbf{0}_{R\times 1} & \cdots & \mathbf{f}_0 & \mathbf{f}_1 \quad \cdots \quad \mathbf{f}_{L_f-1} \end{bmatrix} \\ \tilde{\mathbf{s}}(n) &\triangleq [s(n), s(n-1), \cdots, s(n-L_f-L_w+2)]^T \\ \tilde{\boldsymbol{\eta}}(n) &\triangleq [\boldsymbol{\eta}^T(n), \boldsymbol{\eta}^T(n-1), \cdots, \boldsymbol{\eta}^T(n-L_w+1)]^T \end{aligned}$$

where $\mathbf{0}_{N\times M}$ is the $N \times M$ matrix of zeros. Using these notations, we can rewrite (4) as

$$\mathbf{t}(n) = \mathbf{W}^H \boldsymbol{\mathcal{F}} \tilde{\mathbf{s}}(n) + \mathbf{W}^H \tilde{\boldsymbol{\eta}}(n). \tag{5}$$

The received signal at the destination can be written as

$$y(n) = \sum_{l=0}^{L_g-1} \mathbf{g}_l^T \mathbf{t}(n-l) + v(n) \tag{6}$$

where $v(n)$ is the receiver noise. Using (5) together with the fact that the matrices $\mathbf{W}_l$ are diagonal, (6) can be rewritten as

$$\begin{aligned} y(n) &= \sum_{l=0}^{L_g-1} \mathbf{w}^H (\mathbf{I}_{L_w} \otimes \mathbf{G}_l) \boldsymbol{\mathcal{F}} \tilde{\mathbf{s}}(n-l) \\ &\quad + \sum_{l=0}^{L_g-1} \mathbf{w}^H (\mathbf{I}_{L_w} \otimes \mathbf{G}_l) \tilde{\boldsymbol{\eta}}(n-l) + v(n) \end{aligned} \tag{7}$$

where $\mathbf{I}_N$ is the $N \times N$ identity matrix, $\mathbf{w} \triangleq [\mathbf{w}_0^T, \cdots, \mathbf{w}_{L_w-1}^T]^T$, $\mathbf{w}_l \triangleq \mathrm{diag}\{\mathbf{W}_l\}$, $\mathbf{G}_l \triangleq \mathrm{diag}\{\mathbf{g}_l\}$ and $\otimes$ denotes the Kronecker product. To write (7) in a more compact form, we introduce

$$\begin{aligned} \boldsymbol{\mathcal{G}} &\triangleq \left[\mathbf{I}_{L_w} \otimes \mathbf{G}_0, \cdots, \mathbf{I}_{L_w} \otimes \mathbf{G}_{L_g-1}\right] \\ \breve{\mathbf{F}} &\triangleq [\boldsymbol{\mathcal{F}}_0^T, \cdots, \boldsymbol{\mathcal{F}}_{L_g-1}^T]^T \\ \boldsymbol{\mathcal{F}}_l &\triangleq [\overbrace{\mathbf{0}_{RL_w\times 1}, \cdots, \mathbf{0}_{RL_w\times 1}}^{l \text{ columns}}, \boldsymbol{\mathcal{F}}, \overbrace{\mathbf{0}_{RL_w\times 1}, \cdots, \mathbf{0}_{RL_w\times 1}}^{(L_g-1-l) \text{ columns}}], \\ & \quad l = 0, \cdots, L_g-1 \\ \tilde{\mathbf{I}} &\triangleq [\breve{\mathbf{I}}_0^T, \cdots, \breve{\mathbf{I}}_{L_g-1}^T]^T \\ \breve{\mathbf{I}}_l &\triangleq [\overbrace{\mathbf{0}_{RL_w\times R}, \cdots, \mathbf{0}_{RL_w\times R}}^{l \text{ blocks}}, \mathbf{I}_{RL_w}, \overbrace{\mathbf{0}_{RL_w\times R}, \cdots, \mathbf{0}_{RL_w\times R}}^{(L_g-1-l) \text{ blocks}}], \\ & \quad l = 0, \cdots, L_g-1 \\ \breve{\mathbf{s}}(n) &\triangleq [s(n), s(n-1), \cdots, s(n-L_f-L_w-L_g+3)]^T \\ \breve{\boldsymbol{\eta}}(n) &\triangleq [\boldsymbol{\eta}^T(n), \boldsymbol{\eta}^T(n-1), \cdots, \boldsymbol{\eta}^T(n-L_w-L_g+2)]^T. \end{aligned}$$

Using these notations, (7) can be rewritten as

$$y(n) = \mathbf{w}^H \boldsymbol{\mathcal{G}} \breve{\mathbf{F}} \breve{\mathbf{s}}(n) + \mathbf{w}^H \boldsymbol{\mathcal{G}} \tilde{\mathbf{I}} \breve{\boldsymbol{\eta}}(n) + v(n). \tag{8}$$

Let $\bar{\mathbf{f}}$ and $\bar{\mathbf{F}}$ denote the first column and the residue of $\breve{\mathbf{F}}$, respectively, so that $\breve{\mathbf{F}} = [\bar{\mathbf{f}}, \bar{\mathbf{F}}]$. Then, (8) can be rewritten as

$$\begin{aligned} y(n) &= \mathbf{w}^H \boldsymbol{\mathcal{G}} [\bar{\mathbf{f}}, \bar{\mathbf{F}}] \begin{bmatrix} s(n) \\ \bar{\mathbf{s}}(n) \end{bmatrix} + \mathbf{w}^H \boldsymbol{\mathcal{G}} \tilde{\mathbf{I}} \breve{\boldsymbol{\eta}}(n) + v(n) \\ &= \underbrace{\mathbf{w}^H \boldsymbol{\mathcal{G}} \bar{\mathbf{f}} s(n)}_{\text{signal}} + \underbrace{\mathbf{w}^H \boldsymbol{\mathcal{G}} \bar{\mathbf{F}} \bar{\mathbf{s}}(n)}_{\text{ISI}} + \underbrace{\mathbf{w}^H \boldsymbol{\mathcal{G}} \tilde{\mathbf{I}} \breve{\boldsymbol{\eta}}(n) + v(n)}_{\text{noise}} \end{aligned} \tag{9}$$

where

$$\bar{\mathbf{s}}(n) \triangleq [s(n-1), \cdots, s(n-L_f-L_w-L_g+3)]^T.$$

In Equation (9),

$$\begin{aligned} y_s(n) &= \mathbf{w}^H \boldsymbol{\mathcal{G}} \bar{\mathbf{f}} s(n) & (10) \\ y_i(n) &= \mathbf{w}^H \boldsymbol{\mathcal{G}} \bar{\mathbf{F}} \bar{\mathbf{s}}(n) & (11) \\ y_n(n) &= \mathbf{w}^H \boldsymbol{\mathcal{G}} \tilde{\mathbf{I}} \breve{\boldsymbol{\eta}}(n) + v(n) & (12) \end{aligned}$$

are, respectively, the signal, the ISI, and the noise components at the destination. Note that for the sake of simplicity of our algorithm developed in the next section, we do not consider here *block processing* to coherently combine all the delayed by multipath copies of the signal.

The signal component in (10) can be rewritten as

$$\begin{aligned} y_s(n) &= \mathbf{w}_0^H \mathbf{G}_0 \mathbf{f}_0 s(n) \\ &= \mathbf{w}_0^H (\mathbf{g}_0 \odot \mathbf{f}_0) s(n) \\ &= \mathbf{w}_0^H \mathbf{h}_0 s(n) \end{aligned} \tag{13}$$

where $\odot$ denotes the Schur-Hadamard matrix product and $\mathbf{h}_0 \triangleq \mathbf{g}_0 \odot \mathbf{f}_0$.

3. FILTER-AND-FORWARD RELAY BEAMFORMING

The distributed beamforming problem we consider in this paper corresponds to minimizing the total relay transmitted power P_t subject to the QoS constraint at the destination. It can be written as

$$\min_{\mathbf{w}} \ P_t \qquad \text{s.t.} \quad \text{SINR} \geq \gamma \tag{14}$$

where γ is the minimal required SINR at the destination. Using (5), the transmitted power of the ith relay can be written as

$$\begin{aligned} p_i &= \mathrm{E}\{|t_i(n)|^2\} \\ &= \mathrm{E}\{\mathbf{e}_i^T \mathbf{W}^H \boldsymbol{\mathcal{F}} \tilde{\mathbf{s}}(n) \tilde{\mathbf{s}}^H(n) \boldsymbol{\mathcal{F}}^H \mathbf{W} \mathbf{e}_i\} \\ &\quad + \mathrm{E}\{\mathbf{e}_i^T \mathbf{W}^H \tilde{\boldsymbol{\eta}}(n) \tilde{\boldsymbol{\eta}}^H(n) \mathbf{W} \mathbf{e}_i\} \\ &= P_s \mathbf{e}_i^T \mathbf{W}^H \boldsymbol{\mathcal{F}} \boldsymbol{\mathcal{F}}^H \mathbf{W} \mathbf{e}_i + \sigma_\eta^2 \mathbf{e}_i^T \mathbf{W}^H \mathbf{W} \mathbf{e}_i \end{aligned} \tag{15}$$

where $\mathbf{e}_i$ is the ith column of the identity matrix. In (15), we assume that

$$\begin{aligned} \mathrm{E}\{\tilde{\mathbf{s}}(n) \tilde{\mathbf{s}}^H(n)\} &= P_s \mathbf{I}_{L_f + L_w - 1} \\ \mathrm{E}\{\tilde{\boldsymbol{\eta}}(n) \tilde{\boldsymbol{\eta}}^H(n)\} &= \sigma_\eta^2 \mathbf{I}_{RL_w} \end{aligned}$$

where P_s is the source transmitted power and σ_η^2 is the relay noise variance.

Using $\mathbf{E}_i \triangleq \mathrm{diag}\{\mathbf{e}_i\}$, Equation (15) can be rewritten as

$$\begin{aligned} p_i &= P_s \mathbf{w}^H \left(\mathbf{I}_{L_w} \otimes \mathbf{E}_i\right) \boldsymbol{\mathcal{F}} \boldsymbol{\mathcal{F}}^H \left(\mathbf{I}_{L_w} \otimes \mathbf{E}_i\right)^H \mathbf{w} \\ &\quad + \sigma_\eta^2 \mathbf{w}^H \left(\mathbf{I}_{L_w} \otimes \mathbf{E}_i\right) \left(\mathbf{I}_{L_w} \otimes \mathbf{E}_i\right)^H \mathbf{w} \end{aligned} \tag{16}$$

The total relay transmit power can be expressed as

$$P_t = \sum_{i=1}^{R} p_i = \mathbf{w}^H \mathbf{D} \mathbf{w} \tag{17}$$

where

$$\mathbf{D} = P_s \sum_{i=1}^{R} \left(\mathbf{I}_{L_w} \otimes \mathbf{E}_i\right) \boldsymbol{\mathcal{F}} \boldsymbol{\mathcal{F}}^H \left(\mathbf{I}_{L_w} \otimes \mathbf{E}_i\right)^H + \sigma_\eta^2 \mathbf{I}_{RL_w}.$$

The SINR at the destination can be written as

$$\text{SINR} = \frac{\mathrm{E}\{|y_s(n)|^2\}}{\mathrm{E}\{|y_i(n)|^2\} + \mathrm{E}\{|y_n(n)|^2\}}. \tag{18}$$

Using (13), we have

$$\begin{aligned} \mathrm{E}\{|y_s(n)|^2\} &= \mathrm{E}\{|\mathbf{w}_0^H \mathbf{h}_0 s(n)|^2\} \\ &= P_s \mathbf{w}_0^H \mathbf{h}_0 \mathbf{h}_0^H \mathbf{w}_0 \\ &= P_s \mathbf{w}^H \mathbf{A}^H \mathbf{h}_0 \mathbf{h}_0^H \mathbf{A} \mathbf{w} \\ &= \mathbf{w}^H \mathbf{Q}_s \mathbf{w} \end{aligned} \tag{19}$$

where

$$\begin{aligned} \mathbf{A} &\triangleq [\mathbf{I}_R, \mathbf{0}_{R \times (L_w - 1)R}] \\ \mathbf{Q}_s &\triangleq P_s \mathbf{A}^H \mathbf{h}_0 \mathbf{h}_0^H \mathbf{A}. \end{aligned}$$

Using (11), we obtain

$$\begin{aligned} \mathrm{E}\{|y_i(n)|^2\} &= \mathrm{E}\{\mathbf{w}^H \boldsymbol{\mathcal{G}} \bar{\mathbf{F}} \bar{\mathbf{s}}(n) \bar{\mathbf{s}}^H(n) \bar{\mathbf{F}}^H \boldsymbol{\mathcal{G}}^H \mathbf{w}\} \\ &= P_s \mathbf{w}^H \boldsymbol{\mathcal{G}} \bar{\mathbf{F}} \bar{\mathbf{F}}^H \boldsymbol{\mathcal{G}}^H \mathbf{w} \\ &= \mathbf{w}^H \mathbf{Q}_i \mathbf{w} \end{aligned} \tag{20}$$

where

$$\mathbf{Q}_i \triangleq P_s \boldsymbol{\mathcal{G}} \bar{\mathbf{F}} \bar{\mathbf{F}}^H \boldsymbol{\mathcal{G}}^H.$$

Making use of (12), we also have

$$\begin{aligned} \mathrm{E}\{|y_n(n)|^2\} &= \mathrm{E}\{\mathbf{w}^H \boldsymbol{\mathcal{G}} \tilde{\mathbf{I}} \tilde{\boldsymbol{\eta}}(n) \tilde{\boldsymbol{\eta}}^H(n) \tilde{\mathbf{I}}^H \boldsymbol{\mathcal{G}}^H \mathbf{w}\} + \sigma_\upsilon^2 \\ &= \sigma_\eta^2 \mathbf{w}^H \boldsymbol{\mathcal{G}} \tilde{\mathbf{I}} \tilde{\mathbf{I}}^H \boldsymbol{\mathcal{G}}^H \mathbf{w} + \sigma_\upsilon^2 \\ &= \mathbf{w}^H \mathbf{Q}_n \mathbf{w} + \sigma_\upsilon^2 \end{aligned} \tag{21}$$

where

$$\mathbf{Q}_n \triangleq \sigma_\eta^2 \boldsymbol{\mathcal{G}} \tilde{\mathbf{I}} \tilde{\mathbf{I}}^H \boldsymbol{\mathcal{G}}^H.$$

Using (17) and (19)-(21), the problem in (14) can be expressed as

$$\min_{\mathbf{w}} \ \mathbf{w}^H \mathbf{D} \mathbf{w} \quad \text{s.t.} \quad \frac{\mathbf{w}^H \mathbf{Q}_s \mathbf{w}}{\mathbf{w}^H \mathbf{Q}_i \mathbf{w} + \mathbf{w}^H \mathbf{Q}_n \mathbf{w} + \sigma_\upsilon^2} \geq \gamma. \tag{22}$$

Introducing

$$\tilde{\mathbf{w}} \triangleq \mathbf{D}^{1/2} \mathbf{w}, \quad \mathbf{Q} \triangleq \mathbf{D}^{-1/2} (\mathbf{Q}_s - \gamma \mathbf{Q}_i - \gamma \mathbf{Q}_n) \mathbf{D}^{-1/2}$$

we can rewrite (22) as

$$\min_{\tilde{\mathbf{w}}} \ \|\tilde{\mathbf{w}}\|^2 \qquad \text{s.t.} \quad \tilde{\mathbf{w}}^H \mathbf{Q} \tilde{\mathbf{w}} \geq \gamma \sigma_\upsilon^2. \tag{23}$$

The constraint function in (22) can be used for checking the feasibility of the problem for any given value of γ. In particular, for all the values of γ that lead to *negative semidefinite* $\mathbf{Q}$, the problem in (23) is infeasible. It can be also easily proved that the constraint in (23) can be replaced by the equality constraint $\tilde{\mathbf{w}}^H \mathbf{Q} \tilde{\mathbf{w}} = \gamma \sigma_\upsilon^2$.

If the problem in (23) is feasible, then its solution can be written in a closed form as

$$\tilde{\mathbf{w}} = \beta \, \mathcal{P}\{\mathbf{Q}\} \tag{24}$$

where $\mathcal{P}\{\cdot\}$ denotes the principal eigenvector of a matrix and

$$\beta = \left(\frac{\gamma \sigma_\upsilon^2}{\mathcal{P}\{\mathbf{Q}\}^H \mathbf{Q} \, \mathcal{P}\{\mathbf{Q}\}} \right)^{1/2}.$$

The optimum beamforming weight vector and the minimum total relay transmit power can, therefore, be written as

$$\begin{aligned} \mathbf{w} &= \beta \, \mathbf{D}^{-1/2} \mathcal{P}\{\mathbf{Q}\} && (25) \\ P_{t,\min} &= \gamma \sigma_\upsilon^2 / \lambda_{\max}\{\mathbf{Q}\} && (26) \end{aligned}$$

where $\lambda_{\max}\{\cdot\}$ denotes the principal eigenvalue of a matrix.

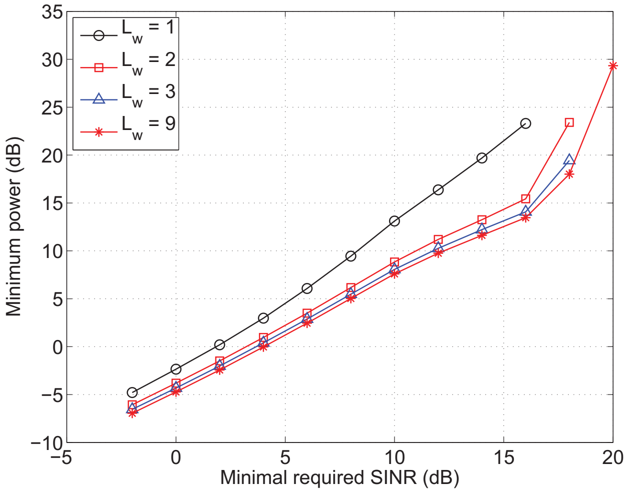

Fig. 2. Total relay transmit power versus minimal required SINR.

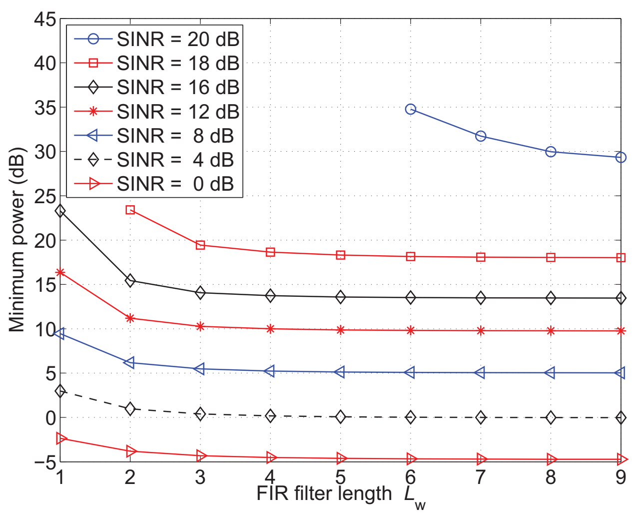

Fig. 3. Total relay transmit power versus relay filter length L_w.

4. SIMULATION RESULTS

In our simulations, we consider a relay network with $R = 10$ relays and quasi-static frequency selective transmitter-to-relay and relay-to-destination channels with the lengths $L_f = L_g = 3$. The channel impulse response coefficients are modeled as i.i.d. zero-mean unit-variance complex Gaussian random values. The relay and destination noises are assumed to have the same powers, and the source transmitted power is 10 dB with respect to the noise. The transmitted signals are assumed to be binary phase shift keying (BPSK) modulated.

Fig. 2 displays the total relay transmit power versus the minimal required SINR at the destination for different lengths of the relay filters. It can be seen that at high values of SINR, some points are missing. These missing points correspond to the case when the problem in (23) (or, equivalently, in (14)) is infeasible, that is, when the QoS requirement cannot be satisfied.

In the trivial case of $L_w = 1$, the FF strategy reduces to the conventional AF approach and, correspondingly, the proposed method boils down to that of [6]. It can be seen from Fig. 2 that using the FF approach at the relays one can substantially reduce the total relay transmit power as compared to the AF case and improve the QoS feasibility range of the original distributed beamforming problem. These improvements are monotonic with the increase of L_w: for example, the QoS constraint with the $\mathrm{SINR} = 20$ dB is not achievable for the relay filter lengths $L_w = 1, 2, 3$, but it can be achieved for $L_w = 9$.

Fig. 3 shows the total relay transmit power versus the relay filter length L_w for different values of the minimal required SINR at the destination. Similar to the previous figure, Fig. 3 demonstrates that the performance (in terms of the relay transmit power) and feasibility of the QoS constraint can be substantially improved by using the FF approach, and these improvements become more pronounced when increasing the relay filter length.

5. CONCLUSION

The problem of distributed network beamforming has been considered in the case of frequency selective transmitter-to-relay and relay-to-destination fading channels. A novel approach to relay beamforming has been proposed using the filter-and-forward strategy. According to this strategy, a finite impulse response filter is used at each relay to compensate for the effects of the transmitter-to-relay and relay-to-destination channels. In our approach, the total transmit power of the relays is minimized subject to the destination quality-of-service constraint. It has been shown that this problem has a closed-form solution. Our simulations validate substantial performance improvements as compared to the existing amplify-and-forward relay beamforming technique.

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