DECENTRALIZED BEAMFORMING FOR MULTI-CARRIER ASYNCHRONOUS BI-DIRECTIONAL RELAYING NETWORKS

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ABSTRACT

We consider an asynchronous two-way relay network, where multiple asynchronous relays cooperate to establish a connection between two transceivers. In such an asynchronous relay network, a certain signal path (originating from one transceiver and going through a certain relay) introduces a propagation and/or relaying delay to the corresponding relayed signal. We assume that such delays are different for different signal paths which correspond to different relays. Based on this model, the end-to-end communication link can be viewed as a multi-path channel, and thus, it can cause inter-symbol-interference (ISI) at the two transceivers when the data rate is sufficiently high. To tackle such an ISI, the two transceivers are herein assumed to employ orthogonal frequency division multiplexing (OFDM) technology. The relays however use amplify-and-forward relaying to materialize a distributed beamforming scheme. For such a communication scheme, we use a max-min fair design approach to optimally obtain the relay beamforming weights and the transceivers' subcarrier powers such that the smallest subcarrier signal-to-noise ratio (SNR) is maximized under a total power budget. Furthermore, we prove that this approach (which has been shown to equivalent to a SNR balancing scheme) leads to certain relay selection solution. We then present a semi-closed-form solution to obtain the relay beamforming weights and the associated maximum balanced SNR. Simulation results show that the performance of this solution is superior to an equal power allocation approach, where all relays and two transceivers consume the same level of power.

***Index Terms*—** Cooperative communication, asynchronous relay networks, SNR balancing, distributed beamforming, power allocation.

1. INTRODUCTION

In a two-way relay network where a reliable connection may not be achievable through direct link between two transceivers, the cooperation of multiple relays can enable a reliable communication between the two transceivers. Amplify-and-forward (AF), decode-and-forward (DF), estimate-and-forward (EF) and filter-and-forward (FF) relaying protocols are among some of the well-known relaying schemes. The AF relaying approach is of particular interest as it is considered to be the simplest among various relaying methods. Different features and applications of the AF method have been widely studied in [1] - [2]. Different bidirectional relaying schemes have been proposed and studied in [3–28]. For different relaying strategies, the problem of power allocation between the source and the relay node(s) has been well studied in the literature [29].

In this paper, assuming an AF protocol, we consider a two-way asynchronous relay network, where multiple relays are deployed to enable bidirectional communication between two transceivers. It is herein assumed that the propagation/relaying delay for each relay's signal path can be different from those for the other relays' paths. Such potentially different propagation/relaying delays cause inter-symbol-interference (ISI) at the two transceivers. As such, the network is assumed to employ orthogonal frequency division multiplexing (OFDM) technology at the two transceivers in order to combat the inter-symbol-interference (ISI) caused by lack of synchronization in different relays' signals. Note that the OFDM scheme could be used at the relays as well [7]. However, in this paper, to keep the relays as simple as possible, we propose to use the AF protocol at the relays and use the OFDM scheme only at the two transceiver. For such a network, under individual relay and transceiver power constraints, we presented a max-min fair approach in [30] to optimally design the beamforming weights and the transceivers' subcarrier powers. In [31], we used a max-min fair design approach under a total power constraint to obtain the relay beamforming weights as well the transceivers' subcarrier power loading at the two transceivers. A numerical approach was presented to obtain the solution to this max-min problem. No proof of global optimality was provided in [30].

In this paper, using the signal model developed in [31], we devise a semi-closed-from solution to the max-min fair design problem studied in [31]. We prove that this solution guarantees global optimality. We also prove that this approach leads to certain relay selection solution. More specifically, in the proposed solution. only relays, which contribute to a certain tap of the end-to-end channel impulse, participate in relaying and the remainder of the relays have to be turned off. Indeed, our solution has a semi-closed form solutions for the optimal beamformig weights of the active relays and for optimal power loading at the two transceivers.

Notations: Complex conjugate, transpose and Hermitian transpose are represented by $(\cdot)^*$, $(\cdot)^T$ and $(\cdot)^H$, respectively. Continuous-time and discrete-time convolution are denoted by $\star_{\mathrm{c}}$ and $\star_{\mathrm{d}}$ respectively. $E\{\cdot\}$ denotes the statistical expectation and $\mathrm{diag}\{\mathbf{a}\}$ stands for a diagonal matrix whose diagonal entries are given by the elements of vector $\mathbf{a}$.

2. SIGNAL MODEL

We consider a two-way AF relay network consisting of L single-antenna relay nodes to establish a bidirectional communication between two transceivers. We represent the signal transmitted by Transceiver p as

$$s_p(t) = \sum_{k=-\infty}^{\infty} s_p[k]\varphi(t - kT_{\mathrm{s}}) \quad p \in \{1, 2\} \tag{1}$$

where $\varphi(t)$ is the rectangular pulse shaping filter, $s_p[k]$ denotes the kth symbol transmitted by Transceiver p, and $T_{\rm s}$ is the symbol time period. The propagation delay for each flat fading channel path from Transceiver p, going through the lth relay, and ending at Transceiver q, for $p, q \in \{1, 2\}$ is denoted by τ_l. Let g_{lp} represent the flat fading channel coefficient between the lth relay and Transceiver p and let w_l stand for the complex relay weight of the lth relay. As such, $\alpha_l \triangleq w_l g_{lp} g_{lq}$ is the total path gain (or loss) of the lth relaying path. Hence, the impulse response of the end-to-end channel between Transceivers 1 and 2, denoted as $h(t)$, can be formulated as

$$h(t) = \sum_{l=0}^{L-1} \alpha_l \delta(t - \tau_l) \quad \text{for} \quad p, q \in \{1, 2\}$$

Sampling the signal received by Transceiver q at the rate of $1/T_{\rm s}$ gives us

$$r_q[nT_{\rm s}] = \left(\sum_{p=1}^{2} s_p(t) \star_{\rm c} h(t) \right) \Big|_{t=nT_{\rm s}} = \sum_{p=1}^{2} s_p[n] \star_{\rm d} h[n] \quad (2)$$

where $h[\cdot]$ is the equivalent discrete-time end-to-end channel impulse response and its nth tap is given by

$$h[n] \triangleq \sum_{l=0}^{L-1} \alpha_l \varphi(nT_{\rm s} - \tau_l). \quad (3)$$

At sufficiently high data rates, the frequency selectivity of the end-to-end channel causes ISI at the two transceivers. As depicted in Fig. 1, orthogonal frequency division multiplexing (OFDM) can be deployed at the transceivers to tackle ISI. In this figure, $\mathbf{T}_{\rm cp}$ and $\mathbf{R}_{\rm cp}$ denote the cyclic prefix insertion and deletion matrices, respectively, $\mathbf{F}$ represents the DFT matrix, while S/P and P/S represent serial-to-parallel and parallel-to-serial convertor. Let us introduce an $N \times L$ matrix $\mathbf{B}$ whose (n, l)th element is defined as

$$B(n, l) = \begin{cases} g_{lp} g_{lq}, & (n-1)T_{\rm s} \leq \tau_l \leq nT_{\rm s} \\ 0, & \text{otherwise.} \end{cases} \quad (4)$$

Assuming that the duration of $\varphi(t)$ is equal to $T_{\rm s}$, the lth relay contributes to the nth tap of $h[\cdot]$ only if $0 \leq nT_{\rm s} - \tau_{lpq} \leq T_{\rm s}$ or, equivalently, $(n-1)T_{\rm s} \leq \tau_l \leq nT_{\rm s}$. Using the latter inequality and approximating $\varphi(t)$ with a rectangular pulse, we can write the vector of discrete-time channel taps, as

$$\mathbf{h} \triangleq \Big[h[0], \ldots, h[N-1] \Big]^T = \mathbf{B}\mathbf{w} \quad (5)$$

where N denotes the length of discrete-time end-to-end channel impulse response $h[\cdot]$, which is assumed to be equal to the number of subcarriers, and $\mathbf{w} \triangleq [w_1, \ldots, w_L]^T$ is the complex vector of the relays' weights. Let $\gamma_l(t)$ denote the lth relay's noise which is assumed to be spatially and temporally white with variance 1. This noise is amplified by w_l and is received by Transceiver q with delay τ'_{lq}. Let us define the (m, l)th element of the $M \times L$ matrix $\mathbf{\Gamma}_q$ as

$$\Gamma_q(m, l) \triangleq \gamma_l(mT_{\rm s} - \tau'_{lq}), \; m = 1, \ldots, M, \; l = 1, \ldots, L \quad (6)$$

where $M = N + N_{\rm cp}$ is the length of the OFDM block and $N_{\rm cp}$ represents the length of the cyclic prefix. Defining the $M \times 1$ measurement noise vector of Transceiver q as $\mathbf{n}'_q$, the vector of the total noise at this transceiver can be represented as

$$\mathbf{n}_q \triangleq \mathbf{\Gamma}_q \mathbf{G}_q \mathbf{w} + \mathbf{n}'_q \;\; q \in \{1, 2\} \quad (7)$$

where $\mathbf{G}_q \triangleq \operatorname{diag}\{g_{1q}, \ldots, g_{Lq}\}$, for $q \in \{1, 2\}$. Let $\mathbf{s}_q \triangleq \Big[s_q[1], \ldots, s_q[N] \Big]^T$ be the vector of symbols transmitted by Transceiver q. The received signal at each transceiver can then be represented as

$$\mathbf{z}_q \triangleq \mathbf{A}_q \mathcal{C}_q \mathbf{s}_q + \mathbf{A}_p \mathcal{D} \mathbf{s}_p + \mathbf{F}\mathbf{R}_{\rm cp}\mathbf{n}_q \quad \text{for} \;\; p, q \in \{1, 2\} \quad (8)$$

where $\mathcal{D} \triangleq \operatorname{diag}\{\mathbf{F}\mathbf{h}\}$ is a diagonal matrix whose diagonal entries are the frequency responses of the end-to-end channel impulse response $h[\cdot]$ at the subcarrier frequencies, $\mathcal{C}_q$ is a diagonal matrix whose diagonal entries are the frequency responses of the "self-interference channel impulse response" at the subcarrier frequencies of Transceiver q [31], and $\mathbf{A}_q \triangleq \operatorname{diag}\{\sqrt{P_{iq}}\}_{i=1}^{i=N}$ is a diagonal matrix which determines the power loading of different subcarriers at Transceiver q. Note that P_{iq} is the allocated power to the ith subcarrier at Transceiver q. After self-interference cancelation [31], the residual signal

$$\tilde{\mathbf{z}}_q = \mathbf{A}_p \mathcal{D} \mathbf{s}_p + \mathbf{F}\mathbf{R}_{\rm cp}\mathbf{n}_q \quad \text{for} \;\; p, q \in \{1, 2\} \quad (9)$$

is used at Transceiver q to extract the desired symbol vector $\mathbf{s}_p$.

3. MAX-MIN FAIR DESIGN APPROACH

In this section, we aim to obtain jointly optimal relay beamforming weight vector as well as the transceivers' subcarrier powers by maximizing the worst SNR across all subcarriers subject to total power budget. We formulate this optimization problem as

$$\max_{\mathbf{p}_1, \mathbf{p}_2 \geq \mathbf{0}} \; \max_{\mathbf{w}} \; \min_{i \in \{1, \cdots, N\}} \; \min_{q \in \{1,2\}} \mathrm{SNR}_{iq}(\mathbf{w}) \quad (10)$$
$$\text{subject to } \frac{\mathbf{1}^T \mathbf{p}_1}{N} + \frac{\mathbf{1}^T \mathbf{p}_2}{N} + \sum_{l=1}^{L} \tilde{P}_l \leq P_{\max}$$

where $P_{\max}$ denotes the maximum power budget, $\tilde{P}_l$ represents the transmitted power of the lth relay, $\mathbf{p}_q \triangleq [P_{1q}, \ldots, P_{Nq}]^T$ is the vector of subcarrier powers at Transceiver q, and $\mathrm{SNR}_{iq}(\mathbf{w}) \triangleq = \frac{P_{ip} |\mathbf{f}_i^H \mathbf{B}_{pq} \mathbf{w}|^2}{\mathbf{w}^H \mathbf{D}_q \mathbf{w} + 1}$, $i = 1, \ldots, N$. and $p \neq q$. In light of the results of [31], the optimization problem (10) leads to all subcarrier SNRs being balanced, thereby leading to an SNR balancing scheme, and it can be written as

$$\min_{\mathbf{w}} \quad \sum_{i=1}^{N} \frac{1}{\phi_i(\mathbf{w})} \quad \text{subject to} \quad \|\mathbf{w}\|^2 \leq P_{\max} \quad (11)$$

where $\phi_i(\mathbf{w}) \triangleq \dfrac{\Big(P_{\max} - \mathbf{w}^H \mathbf{w} \Big) |\mathbf{w}^H \mathbf{a}_i|^2}{2 \Big(\mathbf{w}^H \mathbf{D}_1 \mathbf{w} + 1 \Big) \Big(\mathbf{w}^H \mathbf{D}_2 \mathbf{w} + 1 \Big)}$, according to the results of [3], represents the maximum achievable balanced SNR for a given beamforming relay weight vector $\mathbf{w}$, when the total power budget $P_{\max}$ is assigned to the ith subcarrier. Note that $\mathbf{a}_i \triangleq \mathbf{B}^H \mathbf{f}_i$, where $\mathbf{f}_i$ is the ith Vandermonde column of $\mathbf{F}^H$ and it is given by$\mathbf{f}_i = \frac{1}{\sqrt{N}} \begin{bmatrix} 1 & e^{\left(j \frac{2\pi(i-1)}{N} \right)} & \ldots & e^{\left(j \frac{2(N-1)(i-1)\pi}{N} \right)} \end{bmatrix}^T$ and $\mathbf{D}_q \triangleq \mathbf{G}_q^H \mathbf{\Gamma}_q^H \mathbf{R}_{\rm cp}^T \mathbf{f}_i \mathbf{f}_i^H \mathbf{R}_{\rm cp} \mathbf{\Gamma}_q \mathbf{G}_q$ is an $L \times L$ diagonal matrix, whose its lth diagonal element, as proven in [32], is given by $D_q(l, l) = |g_{lq}|^2$ for $l = 1, \ldots, L$.

Lemma: *The following inequality holds true for any set of positive numbers* $\{\phi_i\}_{i=1}^N$*:*

$$\sum_{i=1}^{N} \frac{1}{\phi_i} \geq \frac{N^2}{\sum_{i=1}^{N} \phi_i} \quad (12)$$

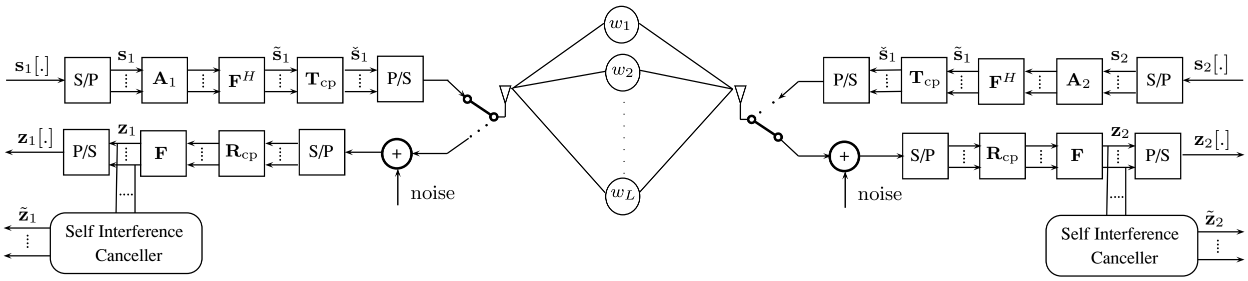

Fig. 1. System model

where equality holds when all $\{\phi_i\}_{i=1}^N$ are equal to each other.

Proof: The arithmetic mean of positive numbers $\{\phi_i\}_{i=1}^N$ is larger or equal to their harmonic mean:

$$\frac{1}{N}\sum_{i=1}^{N}\phi_i \geq \frac{1}{\frac{1}{N}\sum_{i=1}^{N}\frac{1}{\phi_i}}$$

and equality holds iff $\phi_i = \phi_j$, for $i \neq j$. ■

Note that if $\{\phi_i\}_{i=1}^N$ have a certain structure described as $\phi_i = \phi_i(\mathbf{w})$, the equality holds iff one can find such structured $\{\phi_i(\mathbf{w})\}_{i=1}^N$ that they are all equal, i.e., iff one can find a value for $\mathbf{w}$ such that all $\phi_i(\mathbf{w})$'s are all equal. Let $\mathcal{W}$ represent the set of such values of $\mathbf{w}$. Then, without any loss of optimality, we can rewrite the optimization in (11) as

$$\min_{\mathbf{w}} \sum_{i=1}^{N}\frac{1}{\phi_i(\mathbf{w})} \quad \text{subject to} \quad \|\mathbf{w}\|^2 \leq P_{\max} \quad \text{and } \mathbf{w}\in\mathcal{W} \tag{13}$$

Note that the feasible set of the optimization problem (13) is not empty as this feasible set includes $\mathbf{w} = \mathbf{0}$. Since for any $\mathbf{w} \in \mathcal{W}$, we have that $\sum_{i=1}^{N}\frac{1}{\phi_i(\mathbf{w})} = N^2(\sum_{i=1}^{N}\phi_i(\mathbf{w}))^{-1}$, we can rewrite (13) as

$$\min_{\mathbf{w}} \left(\sum_{i=1}^{N}\phi_i(\mathbf{w})\right)^{-1} \text{ s.t. } \|\mathbf{w}\|^2 \leq P_{\max} \quad \text{and } \mathbf{w}\in\mathcal{W} \tag{14}$$

or, equivalently, as

$$\begin{aligned}\max_{\mathbf{w}} \quad & \frac{\left(P_{\max} - \mathbf{w}^H\mathbf{w}\right)\sum_{i=1}^{N}|\mathbf{w}^H\mathbf{a}_i|^2}{2\left(\mathbf{w}^H\mathbf{D}_1\mathbf{w}+1\right)\left(\mathbf{w}^H\mathbf{D}_2\mathbf{w}+1\right)} \\ \text{subject to} \quad & \|\mathbf{w}\|^2 \leq P_{\max} \quad \text{and } \mathbf{w}\in\mathcal{W}\end{aligned} \tag{15}$$

Using Parseval's theorem, we can write

$$\sum_{i=1}^{N}|\mathbf{w}^H\mathbf{a}_i|^2 = \sum_{i=1}^{N}|\mathbf{w}^H\mathbf{B}^H\mathbf{f}_i|^2 = \|\mathbf{B}\mathbf{w}\|^2 = \mathbf{w}^H\mathbf{B}^H\mathbf{B}\mathbf{w} \tag{16}$$

Using (16), we can rewrite the optimization problem (15) as

$$\begin{aligned}\max_{\mathbf{w}} \quad & \frac{\left(P_{\max} - \mathbf{w}^H\mathbf{w}\right)\mathbf{w}^H\mathbf{B}^H\mathbf{B}\mathbf{w}}{2\left(\mathbf{w}^H\mathbf{D}_1\mathbf{w}+1\right)\left(\mathbf{w}^H\mathbf{D}_2\mathbf{w}+1\right)} \\ \text{subject to} \quad & \|\mathbf{w}\|^2 \leq P_{\max} \quad \text{and } \mathbf{w}\in\mathcal{W}\end{aligned} \tag{17}$$

The equality of $\phi_i(\mathbf{w})$'s characterizes the set $\mathcal{W}$ such that for any $\mathbf{w} \in \mathcal{W}$, $|\mathbf{w}^H\mathbf{a}_i|^2 = |\mathbf{w}^H\mathbf{a}_j|^2$ holds true for any $i \neq j$. In other words, the constraint $\mathbf{w} \in \mathcal{W}$ is equivalent to the following set of constraints on $\mathbf{w}$:

$$|\mathbf{f}_i^H\mathbf{B}\mathbf{w}|^2 = |\mathbf{f}_j^H\mathbf{B}\mathbf{w}|^2, \quad \text{for } i \neq j \tag{18}$$

The set of constraints in (18) requires the amplitude of the Fourier representation of the discrete-time FIR end-to-end channel impulse response $h[\cdot]$ to be constant [32]. Since allpass FIR filters can have only one non-zero tap, the constraint in (18) implies that $\mathbf{h}$ has only one non-zero tap. Therefore, when $\mathbf{w} \in \mathcal{W}$, only one of the elements of $\mathbf{h}$ is non-zero. This means that all relays corresponding to the zero taps of $h[\cdot]$ should be turned off (i.e., they should have zero weight). In other words, we should only turn on the relays which contribute to the only non-zero tap of $h[\cdot]$. Now, the crucial question is how to find the only non-zero tap of $h[\cdot]$ which results in the maximum value of the objective function in (17). To answer this question, we need to find all relay sets each of which contributes to one tap of the channel impulse response and compare the SNRs resulting from each set. For different taps, the set of relays which leads to maximum balanced SNR should be chosen and the rest of the relays have to be turned off. Such a relay selection scheme turns the end-to-end link into a frequency flat channel[1]. As such, the approach of [25], developed for two-way relay network with frequency flat channels, can be used to obtain the corresponding maximum balanced SNRs. In [3], a semi-closed-form solution is presented for this approach. According to (18) the potential non-zero elements of $\mathbf{h} = \mathbf{B}\mathbf{w}$ are determined by the rows of $\mathbf{B}$ which include at least one non-zero element. Defining $\mathbf{w}_n$ as the vector of those elements of $\mathbf{w}$ which contribute to the nth non-zero entry of $h[n]$, the optimal value of $\mathbf{w}_n$ has the following semi-closed-form solution:

$$\mathbf{w}_n(\mu_n) = \kappa(\mu_n)\sqrt{2\nu_n}\Big(2\mu_n\mathbf{D}_1 + 2\nu_n\mathbf{D}_2 + \mathbf{I}\Big)^{-1}\mathbf{b}_n \tag{19}$$

where $\nu_n \triangleq 0.5P_{\max} - \mu_n$, $\mathbf{b}_n^H$ is the vector of non-zero elements of the $(n+1)$th row of $\mathbf{B}$, $\kappa(\mu_n)$ is defined as $\kappa(\mu_n) \triangleq (\mathbf{b}_n^H\left(\mathbf{I}+2\mu_n\mathbf{D}_1\right)\left(2\mu_n\mathbf{D}_1+2\nu_n\mathbf{D}_2+\mathbf{I}\right)^{-2}\mathbf{b}_n)^{-1/2}$., and μ_n is the unique solution to the following equation which satisfies $0 \leq \mu_n \leq 0.5P_{\max}$:

$$\begin{aligned}&(P_{\max}-4\mu_n)\mathbf{b}_n^H(2\mu_n\mathbf{D}_1+(P_{\max}-2\mu_n)\mathbf{D}_2+\mathbf{I})^{-1}\mathbf{b}_n \\ &-\mu_n(P_{\max}-2\mu_n)\mathbf{b}_n^H(2\mu_n\mathbf{D}_1+(P_{\max}-2\mu_n)\mathbf{D}_2+\mathbf{I})^{-2} \\ &\times(2\mathbf{D}_1-2\mathbf{D}_2)\mathbf{b}_n = 0\end{aligned} \tag{20}$$

The bisection method developed in [3] can then be used to solve (20) and to compute μ_n. When those relays, which contribute to the

[1] Note that OFDM can still be used to provide multiplexity gain at the price of reducing the diversity gain of our communication system.

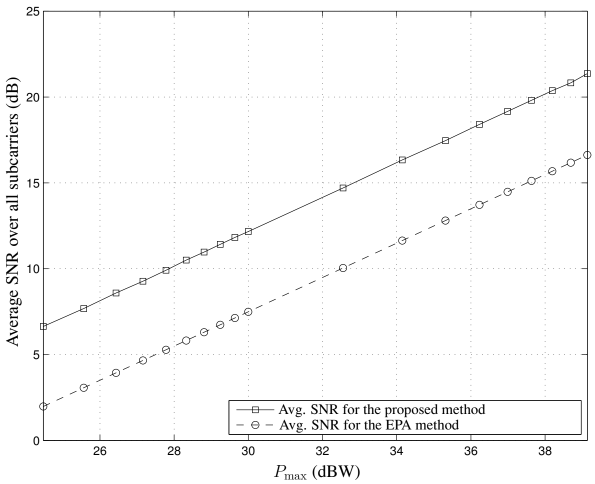

Fig. 2. The average subcarrier SNR versus the maximum available total transmit power.

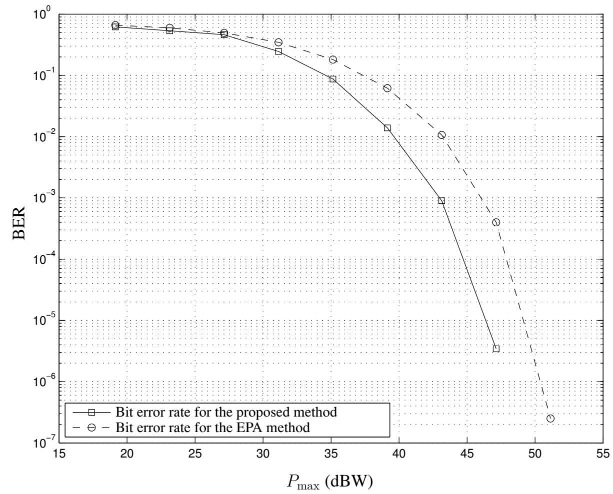

Fig. 3. Bit error rate versus versus the maximum available total transmit power.

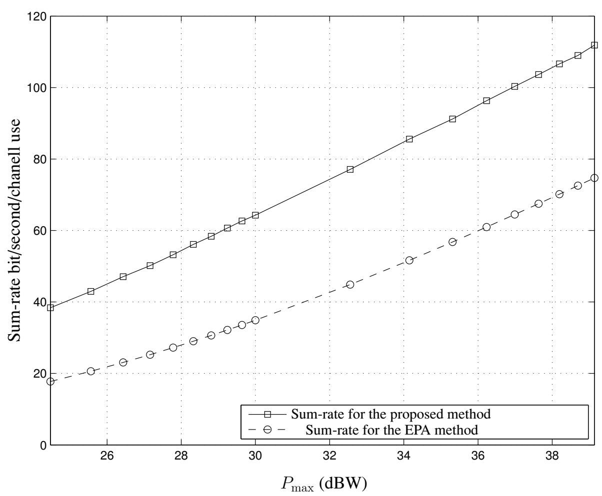

Fig. 4. Sum-rate versus maximum total available power

nth tap of the channel, are turned on and the rest of the relays are deactivated, the corresponding maximum balanced SNR is given by

$$\mathrm{SNR}_{\max}^{(n)} = \mu_n(P_{\max} - 2\mu_n)\mathbf{b}_n^H(2\mu_n\mathbf{D}_1 + 2\nu_n\mathbf{D}_2 + \mathbf{I})^{-1}\mathbf{b}_n \,.$$

The only non-zero tap of the impulse response channel is determined by the value of $n \in \{0, 1, \ldots, N-1\}$ which yields the largest value of $\mathrm{SNR}_{\max}^{(n)}$.

4. SIMULATION RESULTS

We consider an OFDM-based bidirectional communication scheme including 8 cooperating relays and 16 subcarriers. We also assume that the flat fading relay-transceiver channel coefficients follow i.i.d. complex Gaussian distribution with zero-mean and unit variance. All noises at the relays and transceivers are assumed to have a variance equal to one. As we are modeling an asynchronous network, the propagation delay for each relaying path is different from those for other relaying paths and it is drawn, at each simulation run, from a uniform distribution in the interval $[0, 8T_s]$. We compare the performance of our proposed SNR balancing scheme with an equal power allocation (EPA) approach, where the total available power is equally allocated to all the relays and to the two transceivers. As depicted in Fig. 2, the average maximum balanced SNR achieved by our proposed method is about 4.5 dB higher than that achieved by the EPA method. The performances of both methods in terms of bit error rate (BER) have been shown in Fig. 3. As can be seen from this figure, our proposed approach outperforms the EPA technique. Fig. 4 compares the sum-rate of our proposed algorithm with that of the EPA technique. As depicted in this figure, compared to the EPA method, our SNR balancing approach offers a higher sum-rate for any given power budget.

5. CONCLUSION

In this work, we considered an asynchronous bidirectional relay network with propagation/realying that can be different from one relay to another relay. Under the assumption that the transceivers are equipped with OFDM to combat ISI, an SNR balancing algorithm is introduced to jointly design the optimal relay beamformer as well as the subcarrier power loading at the two transceivers via maximization of the smallest SNR across all subcarriers. We proved that such an SNR balancing approach results in a single-tap end-to-end channel impulse response and leads us to select a subset of the relays which all contribute to the only non-zero tap of the impulse response of the end-to-end channel. To obtain the relay beamforming weights and the associated maximum balanced SNR, a semi-closed-form solution was then presented. Our simulation results showed that our proposed method outperforms the equal power allocation approach (where all nodes consume the same level of power) in terms of average SNR, bit error rate, and sum-rate.

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