

SUM-RATE OPTIMAL NETWORK BEAMFORMING AND
POWER ALLOCATION FOR SINGLE-CARRIER
ASYNCHRONOUS BIDIRECTIONAL RELAY NETWORK

by

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Abstract

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We study the problem of sum-rate maximization, under a total transmit power budget, for an *asynchronous single-carrier bidirectional (two-way) network*. The network consists of two single-antenna transceivers which wish to exchange information with the help of multiple single-antenna amplify-and-forward (AF) relays. We assume that the network is asynchronous meaning that different transceiver-relay links cause significantly different propagation delays in the signal they convey. As a result, the end-to-end channel is not amenable to a frequency flat model, rather a multi-path channel model with multiple taps appears to be more appropriate. Such a multi-path model for the end-to-end channel raises the issue of inter-symbol-interference (ISI) at the two transceivers. In a block transmission/reception scheme, ISI leads to inter-block interference (IBI), which could result in loss in the sum-rate of the network, if it is not considered in the design of the system. Considering a block transmission/reception scheme and assuming a total transmit power budget, we maximize the sum-rate of this ISI end-to-end channel over the relay complex weights and transceivers' transmit powers. We rigorously prove that such a sum-rate maximization problem leads to a relay selection scheme, where only those relays which contribute to one tap of the end-to-end channel impulse response are turned on and the rest of the relays are switched off. Indeed, we prove that at the optimum, the end-to-end channel impulse response (CIR) has only one non-zero tap, rendering the end-to-end channel frequency flat. We present the optimal value of the vector of the weights of the active relays and the optimal values of the transceivers' transmit powers in a semi-closed form.

Dedication

To my family

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List of Acronyms

AF Amplify-and-Forward

BER Bit Error Rate

CDMA Code-Division Multiple Access

CF Compress-and-Forward

CIR Channel Impulse Response

CMIMO Cooperative Multiple-Input Multiple-Output

CP Cyclic Prefix

CSI Channel State Information

DF Decode-and-Forward

EF Estimate-and-Forward

EPA Equal Power Allocation

EWM Element-Wise Minimum

FF Filter-and-Forward

FIR Finite Impulse Response

IBI Inter-Block Interference

ICT Information and Communication Technology

ISI Inter-Symbol Interference

LTI Linear Time-Invariant

MABC Multiple Access Broadcast

MCFO Multiple-Carrier Frequency Offsets

MIMO Multiple-Input Multiple-Output

MMSE Minimum Mean-Squared Error

MSE Mean Square Error

MTO Multiple-Timing Offsets

OFDM Orthogonal Frequency Division Multiplexing

PNC Physical-layer Network Coding

QoS Quality of Service

SDMA Space-Division Multiple Access

SIC Self-Interference Cancellation

SISO Single-Input Single-Output

SNR Signal to Noise Ratio

SOCP Second-Order Concave cone Programming

TDBC Time Division Broadcast

ZF Zero Forcing

Chapter 1

Introduction

Future generations of the wireless communication systems continuously strive to meet the demands for higher transmission rates. On the other hand, the energy consumed in the information and communication technology (ICT) is significantly growing. More specifically, as studied in [1], the energy consumption of mobile networks is growing at higher rates compared to the average energy rate of energy consumption in the ICT sector as a whole, and if no effective actions are taken, wireless communication systems will increasingly consume more energy in the future.

Numerous theoretical research efforts and practical implementations of multi-antenna wireless communication systems, called multi-input multi-output (MIMO) schemes, proved that this technique is energy-efficient while adding *diversity* to the system. This multi-antenna technology, along with wideband signaling format, which reduces inter-symbol-interference (ISI), such as orthogonal frequency-division multiplexing (OFDM) and code-division multiple access (CDMA), improve the performance of the system. Since the intro-

duction of MIMO systems, diversity is of primary importance as it enables the transmission of several versions of the signal. In this case, the likelihood of receiving the signal intact is increased. To achieve diversity, in ad-hoc networks, cooperative communication was introduced, whereby the users cooperate with each other in a cell, not only with the base station [2]. This newly emerged technology is commonly accepted as a future key technique, to cope with harsh channel conditions, yet still emulates transmit diversity.

1.1 Cooperative Communication

MIMO techniques increase the throughput with a low energy consumption in wireless networks using multiple antennas. Alternatively, MIMO systems enable *spatial multiplexing* (multi-antenna nodes simultaneously send/receive several data streams) and/or *spatial diversity* (multi-antenna nodes simultaneously send/receive one data stream over several antennas), thereby achieving a higher capacity and/or a higher transmission reliability, respectively. Although transmit diversity is advantageous in a cellular network, where the antennas are separated from each other by at least half of the operating wavelength, it may not be practical for small devices such as mobile stations and sensors. To meet the demand for diversity in the presence of hardware limitations, a new class of techniques, known as *cooperative communications*, were introduced by the standard task group in IEEE 802.16j. The basic idea behind this technique rests in the observation that each single-antenna node in the wireless environment share their antennas, thereby materializing a virtual MIMO scheme. In cooperative communication systems, a wireless user that lies within a certain proximity of other users is assumed to send/receive signals as well as being a cooperative

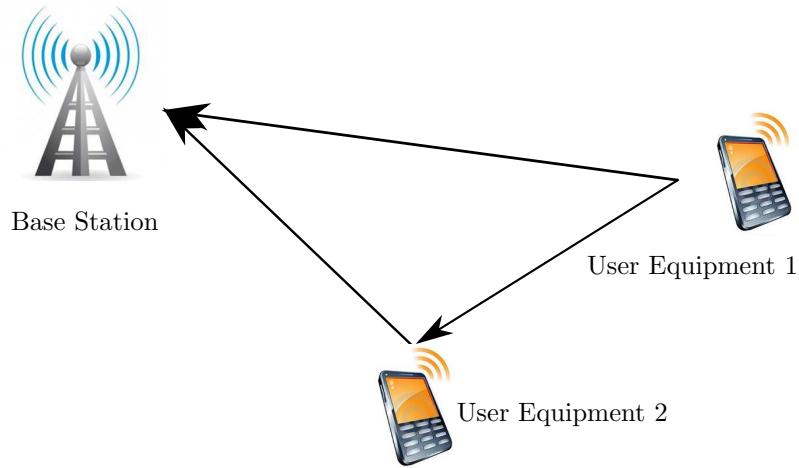


Figure 1.1: Cooperative Communication

node for other users, often referred to as a *relay* or a partner. A simplified demonstration of a cooperative communication scheme appears in Fig. 1.1. The power consumption in a cooperative communication scheme can be viewed from two perspectives. From the base station point of view, no additional cost is associated in the cooperative communication, as its antennas are omnidirectional. From the users' point of view, more power is needed at each user's node as it has to transmit its own information as well as the other users' information. However, considering the diversity gains, achieved at the users, less transmit power is required [2]. With the same reasoning, although the rate for each user might be reduced (sending its own bits and the partner's), the spectral efficiency of each user improves, due to the cooperative diversity which leads to higher channel rates [3,4].

In the course of development of cooperative communication scheme, cooperative multiple-input multiple-output (CMIMO) was introduced. In CMIMO schemes, each relay node is equipped with one or more antennas to create a multi-antenna node. This new method reaps the benefits of both MIMO and cooperative communication systems [4,5].

The cooperative communication has the potential to offer improved coverage and capacity over single-hop communications. Typically, it is envisaged that this technology would have a significant impact on the commercial world.

1.2 Motivation

1.2.1 Technical issues in Relay-Based Communication

Aforementioned arguments make the cooperative communication an interesting candidate in situations where multi-path and fast fading is present. However, there are many challenges in the design, the efficiency and the reliability of cooperative schemes.

Based on how the group of users in a cooperative communication are activated and supported, we consider two wireless network models, centralized and decentralized¹ networks. Although this perspective is insufficient to distinctly categorize all communication networks, it does clarify the trade off between centralized and decentralized schemes. In a centralized approach, operations within a cell are controlled by a centralized fusion center e.g., base station. Alternatively, in the decentralized manner, packets can be forwarded by intermediate nodes [6, 7] and each node is not controlled by the base station. The focus of the centralized network is on improving total system performance, while in the decentralized manner, each node improves its own performance rather than total system performance. To achieve high capacity in a distributed fashion, a distributed algorithm can be used to support individual node's performance. Although inconsiderate distributed algorithm might be beneficial to each node, it can dramatically decrease the system per-

¹It is also called *distributed* network.

formance and eventually halt the system. In this thesis, we intend to design a distributed communication system to enable a two-way of cooperative relay-assisted communication scheme. Meanwhile, we apply a total transmit power constraint to the problem in order to optimize a given performance criterion such as sum-rate (summation of the users' rates). Therefore, the challenge here is to develop a scheme that treats all users fairly while the total system performance is also considered.

Earlier communication protocols were mostly based on one-way relaying where the signals are transmitted in one direction from a source to a destination. In one-way relaying, the communication is established through two steps. First, the transmitter sends the blocks of information symbols to the relay nodes. In the second step, the relays send the processed version of their received signals to the receiver. To efficiently exploit the spectrum, two-way relaying can be used where two transceivers simultaneously exchange information.

Any relay-assisted communication with more than one relay node can be viewed as a multi-path channel where the signal is retransmitted by the relay nodes and is received at the other side of the channel. Due to the different relay-transmitter distance, the retransmitted signal arrives at the other transceiver with different delays. Simply assuming an identical delay in each path, is only applicable for low data rates, and when data rates are sufficiently high, such an assumption results in interference between the symbols at the destination. Designing a system to maximize the rate without considering this *asynchronous* behavior lowers the overall quality of service (QoS) of the relay network.

In this paper, we model an asynchronous two-way relaying scheme consisting of two single-antenna transceivers and several single-antenna relay nodes. We assume that there is no direct link between the two transceivers and that the propagation delay of each relaying path is different from those of the other relaying paths. Therefore, variants delayed versions

of the signal are received at the transceivers, and thus, the end-to-end channel can indeed be viewed as a multi-path link. The multi-path nature of the end-to-end channel can cause ISI between adjacent transmitted symbols. In a block transmission and reception scheme, ISI results in inter-block-interference (IBI) and intra-block-interference. In such a system, power allocation and rate maximization is a of significant importance. Hence, our motivation is to study sum-rate maximization through optimal power allocation and network beamforming, for asynchronous two-way relay network.

1.3 Objective

The above mentioned issues were the key driving force for us to investigate cooperative communication systems and provide a solution in development of these systems. Many different factors can be used as a criterion to evaluate the performance of wireless communication networks. Examples are bit-error-rate (BER), signal-to-noise ratio (SNR), sum-rate and etc. Considering sum-rate as one of the main interests of the system designers in the relay-based communication, we aim to jointly find the optimal relay beamforming weights and transceivers' power which maximize the sum-rate under the total power constraint in an asynchronous relay network. Our objective is to improve the data rate of the transmission in an asynchronous two-way relay scenario by maximizing the sum-rate, while the power of two transceivers and the beamforming weights of the relay nodes are considered as the optimization variables of the problem.

1.4 Methodology

We consider a two-way relay communication system model consisting of two transceivers and several relay nodes. The relays aim to establish a bidirectional connection between the two transceivers. To model an asynchronous communication system, we assign different delays to each relaying path. In this context, knowing the relay beamforming weights of the relay nodes is enough to find the impulse response of the end-to-end channel. We formulate the sum-rate maximization problem to jointly obtain the relay beamforming weights and the power of the two transceivers under a total power constraint. By scrutinizing the problem and using the appropriate mathematical tool, we provide an optimal solution for the system parameters. We prove that the sum-rate maximization problem has a semi-closed-form solution which expresses the optimal relay beamforming weight vector in terms of optimal transmit powers. Moreover, we show that this optimization problem leads to a relay selection scheme where only those relays that contribute to one tap of the end-to-end channel are turned on and the rest of relays are turned off. Then according to the proposed solution, an algorithm is introduced to efficiently assign relay beamforming weights and achieve the maximum sum-rate without the requirement of the synchronization at the relay nodes.

1.5 Contribution

A full version of this thesis has been submitted to *IEEE Transactions in Signal Processing* as [8]

- M. Askari and S. ShahbazPanahi, “Sum-rate optimal network beamforming and

power allocation for single-carrier asynchronous bidirectional relay network,” *IEEE Trans. Signal Process.*, vol. submitted, pp. xxxx-xxxx, July 2015.

Also the contents of this work will be presented at Forthty-Ninth Asilomar conference as [9]

- M. Askari and S. ShahbazPanahi, “Sum-rate maximization for asynchronous two-way relay network,” in *Proc. 49th Annual Asilomar Conf. on Signal, Systems, and computer*, November 2015, pp. xxx-xxx.

1.6 Outline of Thesis

This section describes the thesis structure. Generally, the thesis investigates the optimal performance of the asynchronous bidirectional cooperative relaying scheme to improve the system rate while the energy is limited. The remainder of this thesis is organized as follows.

The literature review is presented in Chapter 2, where we focus on the most relevant works which have been the main inspiration of this thesis. Chapter 3 demonstrates the channel model which is used in the problem definition. In this chapter, we present the sum-rate maximization problem followed by its mathematical derivations of the solution in details. The solution leads to a jointly optimal relay selection and power allocation algorithm. Subsection 3.3 presents the simulation results and the comparison of the proposed algorithm with a non-optimal (but simple) scheme, namely equal power allocation (EPA) technique. The conclusion and future research directions relevant to the work in this thesis are presented in Chapter 4.

1.7 Notations

Throughout this thesis we use the following notations. We use uppercase and lowercase boldface letters to show vector and matrices. The statistical expectation is expressed by $E\{\cdot\}$ and $\det\{\cdot\}$ represents the determinant of matrix. We use $\text{diag}(\mathbf{v})$ to show a diagonal matrix whose diagonal entries are the elements of the vector \mathbf{v} . We describe the logarithm function with base 2, as $\log(\cdot)$. Complex conjugate, transpose and Hermitian operations transpose are denoted as $(\cdot)^*$, $(\cdot)^T$, and $(\cdot)^H$ respectively. Also, $\|\cdot\|$ denotes the Euclidean norm of a vector, $|\cdot|$ stands for the amplitude of a complex number, $[\cdot]_{k,l}$ denotes the $(k,l)^{th}$ entry of a matrix, $[\cdot]_k$ denotes the k^{th} entry of a vector, \mathbf{I}_M represents the identity matrix of size M , and $\mathbf{0}_{M \times M}$ denotes the all-zero matrix of size $M \times M$.

Chapter 2

Literature Review

Numerous relay-assisted communication schemes have been developed since a three terminal communication scheme was first examined by Van Der Meulen in [10]. Subsequent advances in design were later made to achieve higher reliability and rate in these schemes. These advances are too extensive to be included here; we thus only refer to those most relevant to two-way relay networks.

Several relaying schemes have been proposed and studied in the literature. These schemes include amplify-and-forward (AF), estimate-and-forward (EF), decode-and-forward (DF), compress-and-forward (CF) and filter-and-forward (FF). In this study, we focus on AF relaying as this strategy has the benefits of no decoding complexity at the relay, is easily deployable in practical applications and offers low relaying delay. Amplifying gain, the key parameter used in AF relaying, can be fixed or can depend on the channel state information (CSI). We review several relay-assisted communication designs with AF protocol method which are studied in the context of distributed beamforming.

One challenge in the design of wireless network nodes arises from their inability of sending and receiving signals simultaneously at the same time (or in the frequency band), which is referred to as full-duplex transmission. Full-duplex transmission can cause *loop interference*, i.e., coupling from a node's transmission to its own reception [11, 12]. This coupling effect is not trivial as the received signal is several times weaker than the amplified transmitted signal and can be easily corrupted. Due to this problem, a node can only have one of the two modes of reception or transmission, i.e., where the signals are transmitted and received in two time slots (or frequency bands). This transmission method is called half-duplex, and it is regarded as more practical [12]. Although full-duplex only needs one channel for two-hop transmission, it is not trivial for mobile hand-held relay nodes. Using such half-duplex transmission, we study two different relaying protocols namely, one-way and two-way relaying protocols.

2.1 One-Way Relaying

In conventional one-way relaying schemes, the signals are relayed in one direction from a source to a destination. Sending information symbols in one-way relay-assisted transmission is possible through two phases [13]. In the first phase, the source node conveys the signal while the destination node is idle. In the second phase, the relay node forwards its received signal to the destination while the source node is idle. Earlier studies in relay-assisted communication only included uni-directional communications (e.g., [14–28]).

2.2 Two-Way Relaying

In order to establish a bidirectional communication with half-duplex relays between two transceivers, one can use two successive one-way relaying. This straightforward approach requires four time slots to exchange information symbols between two transceivers, and hence, suffers from spectral loss due to the repetitive transmissions from the relays. Research activities on two-way relaying schemes have developed different strategies to decrease the spectral loss and enhance the system performance (e.g., [29–32]). More sophisticated two-way relaying protocols using half-duplex relays can be classified into two main categories. One is the three-phase two-way relaying protocol namely, time division broadcast (TDBC), and the other one is two-phase relaying protocol called, multiple access broadcast (MABC). In the TDBC approach, during the first and second time slots, the relay node receives the transmitters' transmitted signals in sequence one after another. The signals received from transceivers' terminal are combined or coded at the relay node before the broadcast phase. In the MABC approach, each relay node receives both transceivers' signal in the first time slot and transmits the relayed signal in the second time slot.

In both TDBC and MABC protocols, the relay nodes should be equipped with more complicated hardware to combine the data. Due to the half-duplex constraint, the MABC protocol cannot use the direct link between the end nodes as these nodes cannot transmit and receive simultaneously. On the contrary, in the TDBC scheme, the signals are exchanged in three transmission phases and it enables the two transceivers at the end-to-end channel to benefit from a direct communication link to improve system performance. Some studies on the TDBC scheme show that due to the utilization of the direct link, the outage performance of the TDBC is superior to that of the MABC method [29,33]. However, this

comparison ignores some drawbacks in the TDBC technique, such as low spectral efficiency and higher signal interference at the relay nodes. In this work, we focus our attention on the MABC communication scheme.

In a synchronous bi-directional relay network with frequency flat relay-transceiver channels, the channel from one transceiver to the other one can be modeled as a frequency flat link. In such a model, assuming the relays use the AF relaying scheme, the gain of the end-to-end channel is determined by the relays' complex beamforming weights and the relay-transceiver channel coefficients.

In a more realistic model of relay networks, the propagation/relaying delays of different relaying paths can be significantly different from each other. As a result, the signals going through different relaying paths will arrive at the receiving transceivers at significantly different times, thereby rendering the end-to-end channel *asynchronous*. In such scenarios, a frequency flat model (i.e. a single channel gain) may not be the right model to characterize the end-to-end channel, rather a multi-path multi-tap channel model appears to be more appropriate. Then, at sufficiently high data rates, the challenge then becomes how to equalize such a multi-path end-to-end channel, thereby eliminating or suppressing the ISI at the two transceivers.

In earlier studies, it has been assumed that relays are coordinating perfectly; therefore, symbols are accurately synchronized in time at the destination. This assumption is quite costly in ad hoc wireless networks due to their lack of centralized control [34]. In a synchronous scheme, the relays require knowledge of their corresponding path delays as well as maximum delay in the network and they have to be equipped with variable-length memory blocks to compensate their corresponding delay according to the maximum delay. This can be quite costly in terms of complexity and extra destination-to-relay feedback [35].

It should be noted that there is another kind of asynchrony in relay-based communication which occurs for each relay node independent of the other relay nodes. In the MABC communication scheme, the signals transmitted simultaneously from two transceivers may not be synced when arriving at a relay node [36]. Physical-layer network coding (PNC) is known as a perfect attempt to turn the situation. It is proved in [37] and [38] that PNC may actually benefit from this kind of asynchrony in a relay node. This misalignment behavior is not the purpose of this thesis and for the rest of this thesis we assume that the signals, arrived at each relay node from two transceivers, are perfectly synchronized and the term asynchronous only refers to the first kind of misalignment.

The rest of this literature review focuses on synchronous and asynchronous two-way relay networks and optimization techniques applied in these schemes. It also provides a brief discussion of the known results and the researches conducted previously.

2.2.1 Synchronous Two-Way Relay Networks

Various novel synchronization techniques have been developed to mitigate the ISI problem caused by time-asynchronous nature of the nodes in high data-rate relay networks. Some studies have developed techniques to synchronize the relay nodes by using time or carrier frequency offset estimation. The authors of [39] propose an algorithm for estimation of channel and synchronization impairments in a half-duplex space-division multiple-access (SDMA) SISO cooperative network. In order to achieve synchronization in DF and AF relaying network, the multiple-carrier frequency offsets (MCFOs), multiple-timing offsets (MTOs) and channel gains are estimated. This kind of synchronization can add to the complexity at the relays as it requires decoding at the relays.

The study in [40] is an example of a synchronous MABC relaying scheme for a network with multiple relays where all relay nodes are deployed in each transmission. In [40], in order to optimally find the beamforming weight coefficients and transceivers' transmit powers, two different methods are introduced: 1) total transmit power minimization (both at transceivers and relays), 2) SNR balancing method. In the total power minimization, first the total power consumed in the network is minimized under the constraints on two transceivers' SNR while using an iterative steepest descent algorithm to solve the problem. It is also proved that the cost function of the total transmit power minimization has a unique solution. The authors in [40] also solve the problem of minimizing total relay powers with the same constraints on the two transceivers' SNR. For this problem, they assume that the transceivers' powers are given. It is shown that this problem is a second-order convex cone programming (SOCP) type, and has a unique solution. In the SNR balancing method, the smaller SNR of the two transceivers is maximized subject to a total transmit power constraint. This approach can lead to a power allocation scheme where half of the minimum total transmit power will be allocated to the two transceivers and the remaining half will be shared among the relays. However, for the total power minimization, this happens only when a symmetric relaying scheme with equal SNR thresholds at the two transceivers is assumed.

Applying individual per node power constraint to the SNR balancing problem in [40], the authors of [41] maximize the lower SNR of the two communication links in the same MABC scheme. They assume that each relay and each of the transceivers has its own power constraint. Using max-min fair design for transceiver's received SNR, the authors in [41] solve the optimization problem and determine the transmit powers of the two users as well as relay distributed beamforming weights. Due to the computational complexity of

the optimal solution to their problem, the authors have proposed two suboptimal solutions, element-wise-minimum (EWM) solution and one-dimensional (1-D) solution, both of which provide low complexity along with close-to-optimal performance. The simulation results confirm that the proposed power control and relay beamforming method can improve the power efficiency of the network while achieving near optimal performance.

Some subsequent work aimed at maximizing sum-rate in synchronous two-way relaying scheme identical to the one assumed in [40] and [41]. The authors in [42] obtain the achievable beamforming rate region under the total transmit power constraint. Furthermore, to obtain the jointly optimal relay beamforming weights as well as transceivers' powers, two approaches are established and compared. One is sum-rate maximization and the other one is max-min fair design approach. One may expect that sum-rate maximization would result in a water-filling type of solution, which outperforms that gained by a max-min fair design in terms of sum-rate. However, the results in [42] show that for the considered two-way relaying scheme under a total power constraint, the two approaches turn out to be equivalent. Finally, to find the solution for the sum-rate optimal relay beamforming weights, a semi-closed form solution is derived.

There are also other published results on performance comparison between the two synchronous beamforming schemes in the TDBC and MABC methods, [43]. As mentioned earlier, the TDBC method can benefit from the direct link between the two transceivers, but has lower bandwidth efficiency compared to the MABC approach. In this comparison, the power consumed in the whole TDBC network is maximized under QoS constraints for two conditions, with or without a direct link. It is shown that TDBC approach can outperform the MABC method especially in the case when a direct link exists between the two transceivers, but the TDBC has a lower bandwidth efficiency compared to the MABC

scheme.

2.2.2 Asynchronous Two-Way Relay Networks

The studies reported in [34, 44–55] have made important contributions to the research on asynchronous two-way relay networks. However, they do not study the very same two-way relay network that this thesis considers. In what follows, we briefly highlight the differences between our work and each of these studies.

In [44], the authors assume that each relay obtains its to-be-transmitted signal by FIR-filtering its received signal, whereas in the network we consider, the relays use simple AF relaying protocol. Moreover, the authors of [44] assume linear post-channel equalization schemes (such as zero-forcing (ZF) and minimum mean squared error (MMSE) receivers) and investigate the diversity of these schemes, while in our work, we consider both pre-channel and post-channel equalizations and focus on obtaining the sum-rate-optimal values of the pre-channel and post-channel equalizers and the relays' AF coefficients.

The study in [47] considers an asynchronous two-way relay network of two transceivers and a multi-antenna relay with the aim to develop a timing offset and channel estimation technique and use that for relay re-synchronization. Our approach however assumes no relay re-synchronization and relies on pre- and post-channel equalization to combat the ISI produced at the two transceivers due to lack of relay synchronization. Considering a two-way relay network with multiple single-antenna relays, the communication scheme of [48] relies on cyclic prefix (CP) insertion at the transceivers. Then, each relay discards the CP from its received signal, cyclicly shifts the residual signal, re-appends the CP, and retransmits the so-obtained signal. At their receiver front-ends, the transceivers employ

linear post-channel equalization schemes (such as ZF and MMSE receivers). For such a communication scheme, the authors investigate the respective BER performances and develop a BER-optimal power allocation scheme. Different from [48], we do not assume any equalization, consider AF relaying, without any cyclic shift, for the sake of relay simplicity, and use sum-rate as our optimality criterion.

The authors of [50] study a bidirectional DF relay network and devise post-channel equalization schemes (such as MMSE and MMSE decision feedback equalizers) for a scenario where multiple frequency offsets are present at the relays. Unlike [50], we herein consider AF relaying without any equalization.

It should be noted that there has been remarkable research in asynchronous one-way relay networks; however, these approaches cannot be directly applied to two-way relay networks.. The studies in [34, 51, 56, 57] present a communication scheme for an asynchronous one-way relay network, which consists of one source, one destination, and multiple relays. The authors of [51] has proposed a simple orthogonal space-time transmission scheme to achieve full spatial diversity without the requirement of symbol level synchronization at one-way relay nodes. Moreover, the authors of [34] have designed a minimum MMSE receiver for combining received signals in the multiple-relay channel. The results show that the performance can be retained while relaxing the symbol synchronization requirements. The authors of [56] and [57] prove that their asynchronous DF-based one-way relay scheme achieves full diversity under arbitrary delays and provide optimized coding gain despite of the delays. Our study is different from [34, 51, 56, 57] in the sense that since we assume a bidirectional AF relaying scheme, we need to take into account the delays corresponding to the transceiver-relay links. The scheme of [34, 51, 56, 57] is however a one-way DF-based scheme, and thus, it takes into consideration only the delays between the relays and the

destination.

Different from the existing approaches cited above, some other works focus on techniques used to combat the time asynchronism issue without coding at the relay nodes.

A multi-carrier asynchronous two-way relay network is studied in [58] and [59]. In [58], two algorithms for joint subcarrier power allocation and network beamforming in an OFDM based system are provided where each subcarrier is used to enable bidirectional transmission among one pair of transceivers. The first algorithm is deployed in two steps; Step 1: finding distributed beamforming weights at the relays under the per-relay power constraint, while maximizing the power-normalized SNR at one of the transceivers on one of the subcarriers and Step 2: determining the transceivers' subcarrier powers through maximizing the smallest subcarrier SNR. In the second algorithm, while maximizing the SNR, the authors maximize the minimum SNR subject to a total power constraint. According to the simulated results and mathematical proof, the authors conclude that the second approach is equivalent to an SNR balancing technique, which leads to a relay selection strategy where only the relays corresponding to the optimum tap of the channel are chosen. The authors in [59] have characterized the achievable SNR region and the corresponding rate region, in the same system setup as demonstrated in [58]. Moreover, it is proved in [59] that assuming equal rates at different subcarriers at each transceiver leads to a semi-closed-form solution for beamforming weights, for transceivers' subcarrier powers, and for the boundaries of the SNR region.

The problem of maximizing sum-rate for the same multi-carrier asynchronous bidirectional relay network as in [58], has been studied in [60]. The problem is solved under a total power constraint and it is proved that it leads to the same solution as the max-min SNR fair approach in [58]. The authors arrived at an unexpected result where max-min

fair design leads to a sum-rate optimal solution in the context of asynchronous two-way relay network. The validity of the optimal solution is mathematically proved and shown in simulation results.

2.3 Summery

In the literature review, we surveyed the development and major issues associated with asynchronous cooperative communications. In general, this thesis looks at the performance of an asynchronous two-way relay network in order to improve the rate and the energy efficiency. In the next chapter, we will discuss the sum-rate maximization problem when synchronization is not present in the system. We assume that the relaying delays of different relaying paths are known. The problem of estimating these delays does not fit in the scope of this paper. Assuming that these delays are known, our goal is to mitigate the effects of these delays. Also, we do not take synchronization into account, thereby avoiding complexity at the relays. To the best of our knowledge, this problem has not been addressed previously.

Chapter 3

Sum-Rate Maximization

3.1 Preliminaries

We consider a *single-carrier* bidirectional multi-relay network, where two single-antenna transceivers wish to exchange information with the help of L single-antenna AF relay nodes. We assume that there is no direct link between the two transceivers and that the propagation delay of each relaying path is different from those of the other relaying paths. Therefore, as different delayed and attenuated versions of the signal are received at the transceivers via different relaying paths, the end-to-end channel can indeed be viewed as a multi-path link. The multi-path nature (time dispersiveness) of the end-to-end channel can cause ISI at the two transceivers. In a block transmission and reception scheme, ISI results in IBI and intra-block-interference. To mitigate such IBI, one can insert CP or a guard interval between consecutive blocks of information. Intra-block-interference, on the

other hand, has to be tackled by precoding and/or equalization at the transmit and receive front-end of the two transceivers, respectively. Fig. 3.1 shows a two-way AF relay network with an ISI inducing end-to-end channel, where CP insertion and removal operations are used to eliminate IBI. In such a communication setting, our goal is to solve for the optimum relay beamforming weights and the transceivers' transmit powers, such that the sum-rate is maximized subject to a constraint on the total power consumed in the entire network. In doing so, the intra-block-interference has to be explicitly taken into account.

To formulate the sum-rate maximization problem, we begin with a description of the channel model and then model the signals transmitted by the transceivers. Next, we provide the noise model at the relay nodes and at the transceivers and express the total noise received at each transceiver in terms of system parameters. Then, describing the processing performed at the receivers, the signal received at the each receiver is modeled in detail. In the final subsection, an expression for the total power consumed in the network system is presented. The channel and noise model that will be presented in the following subsections were originally presented in [58] but we need to review them for the sake of clarity. Note, however, the communication scheme and the optimality criterion that we herein study is different from those considered in [58]. Indeed, the communication scheme in this thesis is a single-carrier and the design criterion is sum-rate while [58] assumes a multi-carrier scheme and uses a max-min SNR fair design approach to network beamforming and sub-carrier power loading.

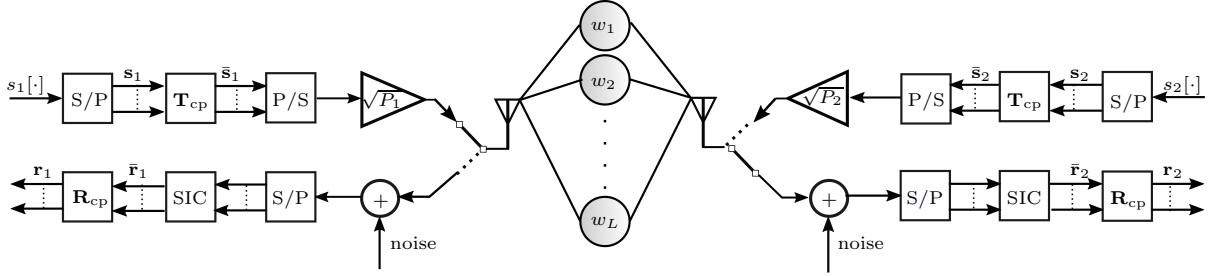


Figure 3.1: System Block Diagram

3.1.1 Channel Modeling

The purpose of this section is to represent the continuous-time channel with an equivalent discrete-time CIR. The link between each transceiver and each relay is considered to be frequency flat and reciprocal. The complex coefficient g_{lq} represents the frequency flat channel between Transceiver q and the l -th relay, for $q \in \{1, 2\}$ and $l \in \{1, 2, \dots, L\}$. Assigning complex beamforming weight w_l to the l -th relay, the total amplification/attenuation factor b_l of the signal path going through the l -th relay can be presented as

$$c_l \triangleq w_l g_{l1} g_{l2}. \quad (3.1)$$

The time delay corresponding to the l -th relaying path is denoted as τ_l . This time delay incorporates the propagation delay of the entire path and is the sum of the three delays: one delay corresponds to the propagation from one transceiver to the l -th relay node; the second delay corresponds to the propagation from that relay node to the other transceiver; and the third delay corresponds to processing and relaying performed at the this relay. Under the discrete-time modeling assumed above, we define $\check{\tau}_l$ as the discrete-time propagation delay of the l -th relaying path, originating from Transceiver 1 and ending

at Transceiver 2, or vice versa. Note that \check{n}_l satisfies $(\check{n}_l - 1)T_s < \tau_l \leq \check{n}_l T_s$, where T_s is the symbol period. With above definitions, the n -th tap of the impulse response of the linear time-invariant (LTI) end-to-end channel between Transceivers 1 and 2, denoted as $h[n]$, can be written as

$$h[n] = \sum_{l=1}^L c_l \delta[n - \check{n}_l], \quad \text{for } 0 \leq n \leq N - 1 \quad (3.2)$$

where N is the maximum number of taps of $h[\cdot]$. Note that according to the CIR in (3.1), $h[\cdot]$ depends on the relay weight vector $\mathbf{w} \triangleq [w_1 \ w_2 \ \dots \ w_L]^T$. For the sake of notation simplicity, we do not explicitly show the dependency of $h[\cdot]$ on the relay complex weight vector \mathbf{w} . Nevertheless, we define the vector of the channel coefficients as

$$\mathbf{h}(\mathbf{w}) \triangleq \left[h[0] \ h[1] \ \dots \ h[N - 1] \right]^T, \quad (3.3)$$

thereby explicitly emphasizing the dependency of these coefficients on \mathbf{w} . Using (3.1) and (3.2), the contribution of the l -th relaying path to the n -th tap of $h[\cdot]$ can be represented by the $(n+1, l)$ -th element of the $N \times L$ matrix \mathbf{B} , which is defined for $n = 0, 1, \dots, N - 1$ and $l = 0, 1, \dots, L$ as

$$B(n+1, l) \triangleq \begin{cases} g_{l1}g_{l2}, & \frac{n}{T_s} \leq n < \frac{n}{T_s} + 1 \\ 0, & \text{otherwise.} \end{cases} \quad (3.4)$$

Using (3.3) and (3.4), the vector of the taps of the end-to-end CIR can be written as

$$\mathbf{h}(\mathbf{w}) = \mathbf{B}\mathbf{w}. \quad (3.5)$$

For the rest of this section, we use $\mathbf{h}(\mathbf{w})$ to represent the vector of the taps of the impulse response of the channel between the two transceivers.

3.1.2 Transmitted Signal Modeling

As shown in Fig. 3.1, the input serial symbols are converted into parallel, using the “S/P” block. We represent the i -th block of the resulting $N_s \times 1$ symbol vector transmitted by Transceiver q as

$$\mathbf{s}_q(i) = \left[s_q[iN_s] \ s_q[iN_s + 1] \ \cdots \ s_q[iN_s + N_s - 1] \right]^T \quad (3.6)$$

where $s_q[k]$ is the k -th symbol transmitted by Transceiver q . CP is now added to $\mathbf{s}_q(i)$ by pre-multiplying $\mathbf{s}_q(i)$ with the matrix $\mathbf{T}_{\text{cp}} \triangleq [\mathbf{I}_{\text{cp}}^T \ \mathbf{I}_{N_s}^T]^T$, where \mathbf{I}_{cp} is the matrix of the last N rows of the $N_s \times N_s$ identity matrix \mathbf{I}_{N_s} , and N is the length of the vector of the taps of the equivalent discrete-time end-to-end CIR. Indeed, the length of the cyclic prefix is equal to the length of the end-to-end CIR. The output of the cyclic prefix insertion block

is denoted as $\bar{\mathbf{s}}_q(i)$ and can be written as

$$\begin{aligned}
\bar{\mathbf{s}}_q(i) &\triangleq \left[\bar{s}_q[iN_t] \bar{s}_q[iN_t + 1] \cdots \bar{s}_q[iN_t + N_t - 1] \right]^T \\
&\triangleq \mathbf{T}_{\text{cp}} \mathbf{s}_q(i) \\
&= \left[s_q[(i+1)N_s - N] \cdots s_q[(i+1)N_s - 1] s_p[iN_s] \right. \\
&\quad \left. \cdots s_q[(i+1)N_s - 1] \right]^T \tag{3.7}
\end{aligned}$$

where $N_t \triangleq N + N_s$ is the length of the transmitted blocks and $\bar{s}_q[iN_t + k]$ is the k -th entry of $\bar{\mathbf{s}}_q(i)$, for $k = 0, 1, \dots, N_s - 1$. The data block $\bar{\mathbf{s}}_q(i)$ is passed through the parallel-to-serial conversion block, denoted as ‘‘P/S’’, and is thus converted into serial symbols. The serial signal is multiplied by $\sqrt{P_q}$, where P_q is the transmit power of Transceiver q , and then transmitted over the wireless channel.

3.1.3 Received Noise Modeling

At the l -th relay node, we assume that the noise is spatially and temporally white with zero mean and variance σ^2 and this noise is represent as $v_l[n]$. Using AF relaying protocol, this noise is multiplied by w_l at the relay, passes through the channel g_{lq} , and arrives at Transceiver q , for $q = 1, 2$, with a delay of n'_{lq} samples, where n'_{lq} satisfies $\frac{\tau'_{lq}}{T_s} \leq n'_{lq} < \frac{\tau'_{lq}}{T_s} + 1$ and τ'_{lq} is the delay of propagation between Transceiver q and the l -th relay. Therefore, the noise at Transceiver q is a superposition (denoted as $\xi_q[n]$) of amplified, phase-adjusted, and delayed versions of the relay noises plus the transceiver measurement noise, represented

as $\gamma'_q[n]$. That is, $\gamma_q[n]$ can be modeled as

$$\begin{aligned}\gamma_q[n] &= \xi_q[n] + \gamma'_q[n] \\ &= \sum_{l=1}^L w_l g_{lq} v_l[n - n'_{lq}] + \gamma'_q[n] = \mathbf{v}_{n,q}^T \mathbf{G}_q \mathbf{w} + \gamma'_q[n]\end{aligned}\quad (3.8)$$

where $\mathbf{G}_q \triangleq \text{diag}\{g_{1q}, g_{2q}, \dots, g_{Lq}\}$ is a diagonal matrix of channel coefficients corresponding to the links between Transceivers q and the relays and the $L \times 1$ vector $\mathbf{v}_{n,q}$ of the delayed relay noises is defined as

$$\mathbf{v}_{n,q} \triangleq \left[v_1[n - n'_{1q}] \ v_2[n - n'_{2q}] \ \cdots \ v_L[n - n'_{Lq}] \right]^T. \quad (3.9)$$

Using the following definitions:

$$\boldsymbol{\gamma}_q(i) \triangleq \left[\gamma_q[iN_t] \ \gamma_q[iN_t + 1] \ \cdots \ \gamma_q[iN_t + N_t - 1] \right]^T \quad (3.10)$$

$$\boldsymbol{\xi}_q(i) \triangleq \left[\xi_q[iN_t] \ \xi_q[iN_t + 1] \ \cdots \ \xi_q[iN_t + N_t - 1] \right]^T \quad (3.11)$$

$$\boldsymbol{\gamma}'_q(i) \triangleq \left[\gamma'_q[iN_t] \ \gamma'_q[iN_t + 1] \ \cdots \ \gamma'_q[iN_t + N_t - 1] \right]^T, \quad (3.12)$$

we can write (3.8) in vector form as

$$\boldsymbol{\gamma}_q(i) = \boldsymbol{\xi}_q(i) + \boldsymbol{\gamma}'_q(i) = \mathbf{\Upsilon}_q(i) \mathbf{G}_q \mathbf{w} + \boldsymbol{\gamma}'_q(i) \quad (3.13)$$

where $\Upsilon_q(i) \triangleq \begin{bmatrix} \mathbf{v}_{iN_t,q} & \mathbf{v}_{iN_t+1,q} & \cdots & \mathbf{v}_{(iN_t+N_t-1),q} \end{bmatrix}^T$ is an $N_t \times L$ matrix whose l -th column is the l -th relay noise, after going through the delay between this relay and Transceiver q , corresponding to the i -th received block.

3.1.4 Received Signal Modeling

At the other side of the channel, the received signal is passed through the ‘‘S/P’’ block and is turned into blocks (vectors) of received signals. The self-interference cancellation is then performed. This type of interference is known to the transmitters and could be canceled. Self-interference cancellation, denoted as ‘‘SIC’’ is a technique which allows each transmitter to cancel its own transmitted signal which is relayed back to that transceiver. To removed the cyclic prefix, a cyclic prefix removal block, denoted by the matrix $\mathbf{R}_{cp} \triangleq [\mathbf{0}_{N_s \times N} \quad \mathbf{I}_{N_s}]$ is used. Assuming $E\{|s_p[k]|^2\} = 1$ and $E\{s_p[k]\} = 0$, for $p \in \{1, 2\}$ and for any k , the i -th signal block received at the output of the self-interference cancellation block at Transceiver q can be expressed as [61]

$$\bar{\mathbf{r}}_q(i) = \sqrt{P_{\bar{q}}}\mathbf{H}_0(\mathbf{w})\bar{\mathbf{s}}_{\bar{q}}(i) + \sqrt{P_{\bar{q}}}\mathbf{H}_1(\mathbf{w})\bar{\mathbf{s}}_{\bar{q}}(i-1) + \gamma_q(i). \quad (3.14)$$

Here, $\bar{q} = 2$, for $q = 1$, and $\bar{q} = 1$, when $q = 2$, $P_{\bar{q}}$ is the transmit power of Transceiver \bar{q} and we have used the following definitions:

$$\mathbf{H}_0(\mathbf{w}) \triangleq \begin{bmatrix} h[0] & 0 & 0 & \cdots & 0 \\ \vdots & h[0] & 0 & \cdots & 0 \\ h[N-1] & \cdots & \ddots & \cdots & \vdots \\ \vdots & \ddots & \cdots & \ddots & 0 \\ 0 & \cdots & h[N-1] & \cdots & h[0] \end{bmatrix} \quad (3.15)$$

$$\mathbf{H}_1(\mathbf{w}) \triangleq \begin{bmatrix} 0 & \cdots & h[N-1] & \cdots & h[1] \\ \vdots & \ddots & 0 & \ddots & \vdots \\ 0 & \cdots & \ddots & \cdots & h[N-1] \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ 0 & \cdots & 0 & \cdots & 0 \end{bmatrix}. \quad (3.16)$$

The received signal vector $\bar{\mathbf{r}}_q(i)$ is multiplied by the cyclic prefix removal matrix \mathbf{R}_{cp} , and thus its first N entries are discarded. One can easily verify that $\mathbf{R}_{\text{cp}}\mathbf{H}_1(\mathbf{w}) = \mathbf{0}$, hence the IBI-inducing matrix $\mathbf{H}_1(\mathbf{w})$ is eliminated by a cyclic prefix removal operation. Using (3.14), we can write the vector $\mathbf{r}_q(i)$, at the output of the cyclic prefix removal block of

Transceiver q as

$$\begin{aligned} \mathbf{r}_q(i) &\triangleq \mathbf{R}_{\text{cp}} \bar{\mathbf{r}}_q(i) = \sqrt{P_{\bar{q}}} \mathbf{R}_{\text{cp}} \mathbf{H}_0(\mathbf{w}) \mathbf{T}_{\text{cp}} \mathbf{s}_{\bar{q}}(i) + \mathbf{R}_{\text{cp}} \boldsymbol{\gamma}_q(i) \\ &= \sqrt{P_{\bar{q}}} \tilde{\mathbf{H}}(\mathbf{w}) \mathbf{s}_{\bar{q}}(i) + \tilde{\boldsymbol{\gamma}}_q(i). \end{aligned} \quad (3.17)$$

Here $\tilde{\boldsymbol{\gamma}}_q(i) \triangleq \mathbf{R}_{\text{cp}} \boldsymbol{\gamma}_q(i)$ is an $N_s \times 1$ vector of noise and $\tilde{\mathbf{H}}(\mathbf{w}) \triangleq \mathbf{R}_{\text{cp}} \mathbf{H}_0(\mathbf{w}) \mathbf{T}_{\text{cp}}$ is an $N_s \times N_s$ circulant matrix whose (k, l) -th entry is given by $\tilde{h}[(k - l) \bmod N_s]$, where $\tilde{h}[n] = h[n]$ if $0 \leq n \leq N - 1$ and $\tilde{h}[n] = 0$ if $N \leq n \leq N_s - 1$. Note that, in general, the matrix $\tilde{\mathbf{H}}(\mathbf{w})$ may not be diagonal, hence it produces intra-block-interference in the received block.

3.1.5 Total Transmit Power Derivations

Assuming that P_1 and P_2 denote the transceivers' transmit powers, we now derive the relay power in terms of P_1 and P_2 and the relay weight vector \mathbf{w} . As illustrated in Fig. 3.1, the $N_t \times 1$ vector $\bar{\mathbf{x}}_l(i)$ of the i -th signal block relayed by the l -th relay can be written as

$$\begin{aligned} \bar{\mathbf{x}}_l(i) &\triangleq \left[\bar{x}_l[iN_t] \ \bar{x}_l[iN_t + 1] \ \cdots \ \bar{x}_l[iN_t + N_t - 1] \right]^T \\ &= w_l \left(\sqrt{P_1} g_{l1} \bar{\mathbf{s}}_1(i) + \sqrt{P_2} g_{l2} \bar{\mathbf{s}}_2(i) + \mathbf{v}_l(i) \right) \end{aligned} \quad (3.18)$$

where the vector $\mathbf{v}_l(i) \triangleq [v_l[iN_t] \ v_l[iN_t + 1] \ \cdots \ v_l[iN_t + N_t - 1]]^T$ is the i -th block of measurement noise at the l -th relay and $\bar{x}_l[k]$ is the signal transmitted by the l -th relay at time k . We assume that $\mathbf{v}_l(\cdot)$ is a stationary zero-mean random vector process whose entries are uncorrelated and have variances equal to σ^2 . Using (3.18), the average transmit

power of the l -th relay is then written as

$$\begin{aligned}
\tilde{P}_l &\triangleq \frac{1}{N_t} \mathbb{E} \{ \bar{\mathbf{x}}_l^H(i) \bar{\mathbf{x}}_l(i) \} \\
&= \frac{|w_l|^2}{N_t} \mathbb{E} \left\{ \left[\sqrt{P_1} g_{l1}^* \bar{\mathbf{s}}_1^H(i) + \sqrt{P_2} g_{l2}^* \bar{\mathbf{s}}_2^H(i) + \mathbf{v}_l^H(i) \right] \times \left[\sqrt{P_1} g_{l1} \bar{\mathbf{s}}_1(i) + \sqrt{P_2} g_{l2} \bar{\mathbf{s}}_2(i) + \mathbf{v}_l(i) \right] \right\} \\
&= \frac{P_1 |g_{l1}|^2 |w_l|^2}{N_t} \mathbb{E} \{ \bar{\mathbf{s}}_1^H(i) \bar{\mathbf{s}}_1(i) \} + \frac{P_2 |g_{l2}|^2 |w_l|^2}{N_t} \mathbb{E} \{ \bar{\mathbf{s}}_2^H(i) \bar{\mathbf{s}}_2(i) \} + \frac{|w_l|^2}{N_t} \mathbb{E} \{ \mathbf{v}_l^H(i) \mathbf{v}_l(i) \} \\
&= |w_l|^2 (|g_{l1}|^2 P_1 + |g_{l2}|^2 P_2 + \sigma^2)
\end{aligned} \tag{3.19}$$

where we have assumed that $\bar{\mathbf{s}}_1(\cdot)$, $\bar{\mathbf{s}}_2(\cdot)$, and $\mathbf{v}_l(\cdot)$ are zero-mean mutually independent stationary random vectors. Note that, to arrive at (3.19), we assume that the communication time frame is much longer than the maximum of the time differences between arrivals of transceiver signals at the relays [58]. Using (3.19), the total transmit power consumed in the entire network can be obtained as

$$\begin{aligned}
P_{total} &\triangleq P_1 + P_2 + \sum_{l=1}^L \tilde{P}_l \\
&= P_1 + P_2 + \sum_{l=1}^L |w_l|^2 (|g_{l1}|^2 P_1 + |g_{l2}|^2 P_2 + \sigma^2) \\
&= P_1 (1 + \|\mathbf{G}_1 \mathbf{w}\|^2) + P_2 (1 + \|\mathbf{G}_2 \mathbf{w}\|^2) + \sigma^2 \mathbf{w}^H \mathbf{w}.
\end{aligned} \tag{3.20}$$

In our design, we constrain P_{total} to be less than, or equal to the maximum available power P_{max} in the system.

3.2 Sum-rate Maximization

In this section, our goal is to optimally obtain the relay beamforming weight vector \mathbf{w} , and the transceivers' transmit powers P_1 and P_2 such that the sum-rate is maximized under a total transmit power budget. Note that the model in (3.17) can be viewed as a MIMO channel. The data-rate corresponding to the MIMO channel in (3.17) can be written as [62]

$$\frac{1}{2} \log \det \left\{ \mathbf{I} + P_{\bar{q}} \mathbf{R}_q^{-1/2}(\mathbf{w}) \tilde{\mathbf{H}}(\mathbf{w}) \tilde{\mathbf{H}}^H(\mathbf{w}) \mathbf{R}_q^{-1/2}(\mathbf{w}) \right\} \quad (3.21)$$

where the factor $\frac{1}{2}$ signifies the fact that relaying scheme implemented in two time slots, while $\mathbf{R}_q(\mathbf{w})$ is the correlation matrix of the noise in (3.17) and can be written as [63]

$$\begin{aligned} \mathbf{R}_q(\mathbf{w}) &\triangleq \mathbb{E}\{\tilde{\gamma}_q(i)\tilde{\gamma}_q^H(i)\} \\ &= \mathbf{R}_{\text{cp}} \mathbb{E}\{\gamma_q(i)\gamma_q^H(i)\} \mathbf{R}_{\text{cp}}^H \\ &= \sigma^2 (\mathbf{w}^H \mathbf{G}_q^H \mathbf{G}_q \mathbf{w} + 1) \mathbf{I}_{N_s} \\ &= \sigma^2 (\|\mathbf{G}_q \mathbf{w}\|^2 + 1) \mathbf{I}_{N_s}. \end{aligned} \quad (3.22)$$

To obtain the optimal transmit powers at the two transceivers as well as the relay beamforming weight vector, ignoring the $\frac{1}{2}$ factor in (3.21), the problem of maximizing the

sum-rate under a total power constraint can be formulated as

$$\begin{aligned} & \max_{P_1, P_2, \mathbf{w}} \sum_{q=1}^2 \log \det \left\{ \mathbf{I} + P_{\bar{q}} \mathbf{R}_q^{-1/2}(\mathbf{w}) \tilde{\mathbf{H}}(\mathbf{w}) \tilde{\mathbf{H}}^H(\mathbf{w}) \mathbf{R}_q^{-1/2}(\mathbf{w}) \right\} \\ & \text{s.t.} \quad P_{total} \leq P_{max} \\ & \quad \quad P_1 \geq 0, P_2 \geq 0. \end{aligned} \quad (3.23)$$

To solve (3.23), note that the $N_s \times N_s$ circulant matrix $\tilde{\mathbf{H}}(\mathbf{w})$ can be decomposed as

$$\tilde{\mathbf{H}}(\mathbf{w}) = \mathbf{F}^H \mathbf{D}(\mathbf{w}) \mathbf{F}. \quad (3.24)$$

Here

$$\mathbf{D}(\mathbf{w}) \triangleq \text{diag} \left\{ H(e^{j0}), H(e^{j\frac{2\pi}{N_s}}), \dots, H(e^{j\frac{2\pi(N_s-1)}{N_s}}) \right\} \quad (3.25)$$

is an $N_s \times N_s$ diagonal matrix of the values of frequency response of the end-to-end channel at integer multiples of $\frac{1}{N_s}$, whereas $H(e^{j2\pi f}) \triangleq \sum_{n=0}^{N-1} h[n] e^{-j2\pi f n}$ is the frequency response of the end-to-end channel at the normalized frequency f , and \mathbf{F} is the $N_s \times N_s$ DFT matrix whose (k, k') -th element is defined as $F_{k,k'} = N_s^{-\frac{1}{2}} e^{-j2\pi(k-1)(k'-1)/N_s}$, for $k, k' \in \{1, \dots, N_s\}$.

To write matrix $\mathbf{D}(\mathbf{w})$ explicitly in terms of the relay beamforming weights, we define $\mathbf{f}_k \triangleq \frac{1}{\sqrt{N_s}} \left[1 \quad e^{j\frac{2\pi(k-1)}{N_s}} \quad \dots \quad e^{j\frac{2\pi(k-1)(N-1)}{N_s}} \right]^T$ which is the k -th row of \mathbf{F}^H , for $k = 1, 2, \dots, N_s$.

Hence, we can write

$$\begin{aligned} \mathbf{D}(\mathbf{w}) &\triangleq \text{diag} \left\{ H(e^{j0}), H(e^{j\frac{2\pi}{N_s}}), \dots, H(e^{j\frac{2\pi(N_s-1)}{N_s}}) \right\} \\ &= \sqrt{N_s} \text{diag} \left\{ \mathbf{f}_1^H \tilde{\mathbf{h}}, \mathbf{f}_2^H \tilde{\mathbf{h}}, \dots, \mathbf{f}_{N_s}^H \tilde{\mathbf{h}} \right\} \end{aligned} \quad (3.26)$$

where

$$\begin{aligned} \tilde{\mathbf{h}}(\mathbf{w}) &\triangleq [\tilde{h}[0] \tilde{h}[1], \dots, \tilde{h}[N_s - 1]]^T \\ &= [\mathbf{h}^T(\mathbf{w}) \mathbf{0}_{1 \times N_s - N}]^T = [(\mathbf{B}\mathbf{w})^T \mathbf{0}_{1 \times N_s - N}]^T \end{aligned} \quad (3.27)$$

is the zero-padded version of the channel vector $\mathbf{h}(\mathbf{w})$. Let us define $\beta_q(\mathbf{w}) = 1 + \|\mathbf{G}_q \mathbf{w}\|^2$, for $q = 1, 2$, and denote the i -th diagonal entry of $\mathbf{D}(\mathbf{w})$ as $D_{ii}(\mathbf{w})$. Using (3.22) and (3.24), we can now simplify the objective function in (3.23) as

$$\begin{aligned}
& \sum_{q=1}^2 \log \det \left\{ \mathbf{I} + P_{\bar{q}} \mathbf{R}_q^{-1/2}(\mathbf{w}) \tilde{\mathbf{H}}(\mathbf{w}) \tilde{\mathbf{H}}^H(\mathbf{w}) \mathbf{R}_q^{-1/2}(\mathbf{w}) \right\} = \\
& \sum_{q=1}^2 \log \det \left\{ \mathbf{I} + \frac{P_{\bar{q}}}{\sigma^2 \beta_q(\mathbf{w})} \mathbf{F}^H \mathbf{D}(\mathbf{w}) \underbrace{\mathbf{F} \mathbf{F}^H}_{\mathbf{I}} \mathbf{D}^H(\mathbf{w}) \mathbf{F} \right\} = \\
& \sum_{q=1}^2 \log \det \left\{ \mathbf{I} + \frac{P_{\bar{q}}}{\sigma^2 \beta_q(\mathbf{w})} \mathbf{F}^H \mathbf{D}(\mathbf{w}) \mathbf{D}^H(\mathbf{w}) \mathbf{F} \right\} = \\
& \sum_{q=1}^2 \log \det \left\{ \mathbf{F}^H \left(\mathbf{I} + \frac{P_{\bar{q}}}{\sigma^2 \beta_q(\mathbf{w})} \mathbf{D}(\mathbf{w}) \mathbf{D}^H(\mathbf{w}) \right) \mathbf{F} \right\} = \\
& \sum_{q=1}^2 \log \det \left\{ \mathbf{I} + \frac{P_{\bar{q}}}{\sigma^2 \beta_q(\mathbf{w})} \mathbf{D}(\mathbf{w}) \mathbf{D}^H(\mathbf{w}) \right\} = \\
& \sum_{q=1}^2 \log \prod_{i=1}^{N_s} \left(1 + \frac{P_{\bar{q}}}{\sigma^2 \beta_q(\mathbf{w})} |D_{ii}(\mathbf{w})|^2 \right) = \\
& \sum_{q=1}^2 \log \prod_{i=1}^{N_s} \left(1 + \frac{P_{\bar{q}}}{\sigma^2 \beta_q(\mathbf{w})} N_s |\mathbf{f}_i^H \tilde{\mathbf{h}}(\mathbf{w})|^2 \right). \tag{3.28}
\end{aligned}$$

Using (3.28), the optimization problem (3.23) can be written as

$$\begin{aligned}
& \max_{P_1, P_2, \mathbf{w}} \quad \sum_{q=1}^2 \sum_{i=1}^{N_s} \log \left(1 + \frac{P_{\bar{q}}}{\sigma^2 \beta_q(\mathbf{w})} N_s |\mathbf{f}_i^H \tilde{\mathbf{h}}(\mathbf{w})|^2 \right) \\
& \text{subject to} \quad P_1 \beta_1(\mathbf{w}) + P_2 \beta_2(\mathbf{w}) + \sigma^2 \mathbf{w}^H \mathbf{w} \leq P_{max} \\
& \quad \quad \quad P_1 \geq 0, P_2 \geq 0. \tag{3.29}
\end{aligned}$$

To solve (3.29), we first solve the maximization over P_1 and P_2 while assuming that \mathbf{w} is fixed, and later obtain the optimal value of \mathbf{w} . To do so, we rewrite (3.29) as

$$\begin{aligned} & \max_{\mathbf{w}} \max_{P_1, P_2} \sum_{i=1}^{N_s} \log \left(1 + \frac{P_1}{\sigma^2 \beta_2(\mathbf{w})} \alpha_i(\mathbf{w}) \right) \left(1 + \frac{P_2}{\sigma^2 \beta_1(\mathbf{w})} \alpha_i(\mathbf{w}) \right) \\ & \text{subject to} \quad P_1 \beta_1(\mathbf{w}) + P_2 \beta_2(\mathbf{w}) \leq \gamma(\mathbf{w}) \\ & \quad \quad \quad P_1 \geq 0, P_2 \geq 0 \end{aligned} \quad (3.30)$$

where we used the following definitions: $\alpha_i(\mathbf{w}) \triangleq N_s |\mathbf{f}_i^H \tilde{\mathbf{h}}(\mathbf{w})|^2$, and $\gamma(\mathbf{w}) \triangleq P_{max} - \sigma^2 \mathbf{w}^H \mathbf{w}$. Note that at the optimum, we assume that $\gamma(\mathbf{w}) \geq 0$, otherwise P_1 and/or P_2 have to be negative, thereby contradicting the non-negative conditions on P_1 and/or P_2 . Note that for any fixed \mathbf{w} , under the constraints in (3.30), we can write

$$\begin{aligned} & \max_{P_1, P_2} \sum_{i=1}^{N_s} \log \left(1 + \frac{P_1}{\sigma^2 \beta_2(\mathbf{w})} \alpha_i(\mathbf{w}) \right) \left(1 + \frac{P_2}{\sigma^2 \beta_1(\mathbf{w})} \alpha_i(\mathbf{w}) \right) \leq \\ & \sum_{i=1}^{N_s} \max_{P_1, P_2} \log \left(1 + \frac{P_1}{\sigma^2 \beta_2(\mathbf{w})} \alpha_i(\mathbf{w}) \right) \left(1 + \frac{P_2}{\sigma^2 \beta_1(\mathbf{w})} \alpha_i(\mathbf{w}) \right) \end{aligned} \quad (3.31)$$

The inequality in (3.31) is satisfied with equality if and only if the solution in terms of P_1 and P_2 to the following maximization:

$$\begin{aligned} & \max_{P_1, P_2} \quad \log \left(1 + \frac{P_1}{\sigma^2 \beta_2(\mathbf{w})} \alpha_i(\mathbf{w}) \right) \left(1 + \frac{P_2}{\sigma^2 \beta_1(\mathbf{w})} \alpha_i(\mathbf{w}) \right) \\ & \text{s.t.} \quad P_1 \beta_1(\mathbf{w}) + P_2 \beta_2(\mathbf{w}) = \gamma(\mathbf{w}) \\ & \quad \quad \quad P_1 \geq 0, P_2 \geq 0. \end{aligned} \quad (3.32)$$

is independent of i . To show that such P_1 and P_2 exist, we now solve (3.32). The solution to the optimization problem (3.32) can be found using the Lagrangian multiplier method. For fixed \mathbf{w} , we define the Lagrangian as

$$\begin{aligned} \mathcal{L}(P_1, P_2; \mathbf{w}) = & \log \left(1 + \frac{P_1}{\sigma^2 \beta_2(\mathbf{w})} \alpha_i(\mathbf{w}) \right) \left(1 + \frac{P_2}{\sigma^2 \beta_1(\mathbf{w})} \alpha_i(\mathbf{w}) \right) \\ & + \lambda (P_1 \beta_1(\mathbf{w}) + P_2 \beta_2(\mathbf{w}) - \gamma(\mathbf{w})) \end{aligned} \quad (3.33)$$

where λ is the Lagrangian multiplier coefficient. The optimal values of P_1 and P_2 in terms of λ can be obtained by differentiating (3.33) with respect to P_1 and P_2 and equating the derivatives to zero. That is, we can write

$$\frac{\partial}{\partial P_1} \mathcal{L}(P_1, P_2; \mathbf{w}) = \frac{\frac{\alpha_i(\mathbf{w})}{\sigma^2 \beta_2(\mathbf{w})}}{\left(1 + \frac{P_1 \alpha_i(\mathbf{w})}{\sigma^2 \beta_2(\mathbf{w})} \right) \ln 2} + \lambda \beta_1 = 0. \quad (3.34)$$

Solving for P_1 , we obtain

$$P_1 = \frac{-1}{\beta_1(\mathbf{w}) \lambda \ln 2} - \frac{\sigma^2 \beta_2(\mathbf{w})}{\alpha_i(\mathbf{w})}. \quad (3.35)$$

Similarly, we can obtain the optimal value of P_2 in terms of λ as

$$P_2 = \frac{-1}{\beta_2(\mathbf{w}) \lambda \ln 2} - \frac{\sigma^2 \beta_1(\mathbf{w})}{\alpha_i(\mathbf{w})}. \quad (3.36)$$

Substituting P_1 and P_2 from (3.35) and (3.36) into the first constraint in (3.32), we can find the optimal value of λ as

$$\lambda = \frac{-2\alpha_i(\mathbf{w})}{(\gamma(\mathbf{w})\alpha_i(\mathbf{w}) + 2\sigma^2\beta_1(\mathbf{w})\beta_2(\mathbf{w})) \ln 2}. \quad (3.37)$$

Using (3.37) along with (3.35) and (3.36), we obtain the optimal values of P_1 and P_2 , for any given feasible value of \mathbf{w} , as

$$P_1 = \frac{\gamma(\mathbf{w})}{2\beta_1(\mathbf{w})} \geq 0 \quad (3.38)$$

$$P_2 = \frac{\gamma(\mathbf{w})}{2\beta_2(\mathbf{w})} \geq 0. \quad (3.39)$$

As can be seen from (3.38) and (3.39), the optimal values of P_1 and P_2 , are independent of the value of i . Hence, the inequality in (3.31) holds with equality. Note that at the optimum, since $\gamma(\mathbf{w}) \geq 0$, P_1 and P_2 in (3.38) and (3.39) will be nonnegative, hence they belong to the feasible set of the optimization problem (3.30). As a result, the optimization in (3.30) can be written as

$$\begin{aligned} & \max_{\mathbf{w}} \sum_{i=1}^{N_s} \log \left(1 + \frac{\gamma(\mathbf{w})\alpha_i(\mathbf{w})}{2\sigma^2\beta_1(\mathbf{w})\beta_2(\mathbf{w})} \right)^2 \\ & \text{subject to } \sigma^2 \mathbf{w}^H \mathbf{w} \leq P_{max}. \end{aligned} \quad (3.40)$$

or

$$\begin{aligned} \max_{\mathbf{w}} \quad & \log \prod_{i=1}^{N_s} \left(1 + \frac{\gamma(\mathbf{w})\alpha_i(\mathbf{w})}{2\sigma^2\beta_1(\mathbf{w})\beta_2(\mathbf{w})} \right)^2 \\ \text{subject to} \quad & \sigma^2 \mathbf{w}^H \mathbf{w} \leq P_{max}. \end{aligned} \quad (3.41)$$

To solve the optimization problem (3.41), let us define

$$\phi_i(\mathbf{w}) \triangleq 1 + \frac{\gamma(\mathbf{w})\alpha_i(\mathbf{w})}{2\sigma^2\beta_1(\mathbf{w})\beta_2(\mathbf{w})}. \quad (3.42)$$

Using (3.42), the optimization problem (3.41) can be equivalently written as

$$\begin{aligned} \max_{\mathbf{w}} \quad & \prod_{i=1}^{N_s} \phi_i(\mathbf{w}) \\ \text{subject to} \quad & \mathbf{w}^H \mathbf{w} \leq \frac{P_{max}}{\sigma^2} \end{aligned} \quad (3.43)$$

or, equivalently, as

$$\begin{aligned} \max_{\mathbf{w}} \max_{\alpha} \quad & \prod_{i=1}^{N_s} \phi_i(\mathbf{w}) \\ \text{subject to} \quad & \sum_{i=1}^{N_s} \phi_i(\mathbf{w}) = \alpha \\ \text{and} \quad & \mathbf{w}^H \mathbf{w} \leq \frac{P_{max}}{\sigma^2} \end{aligned} \quad (3.44)$$

where we have added an auxiliary variable α . In order to further simplify (3.44), we use the following lemma.

Lemma 1: Suppose that $\phi_i \geq 0$ for all $i \in \{1, 2, \dots, N_s\}$. The solution to the following maximization problem:

$$\begin{aligned} & \max_{\{\phi_i\}_{i=1}^{N_s}} && \prod_{i=1}^{N_s} \phi_i \\ & \text{subject to} && \sum_{i=1}^{N_s} \phi_i = \alpha \end{aligned} \quad (3.45)$$

is given by $\phi_i = \frac{\alpha}{N_s}$ for all i , i.e. all ϕ_i s are equal at the optimum.

Using lemma 1 along with the fact that all $\phi_i(\mathbf{w})$'s in (3.44) are positive, we conclude that for any \mathbf{w} , the inner maximization is upper-bounded by the situation where all $\phi_i(\mathbf{w})$'s are equal. If we can find a value for \mathbf{w} such that all $\phi_i(\mathbf{w})$'s are equal, then this upper bound is achievable. We now show that such a value for \mathbf{w} exists. That is we show the set \mathcal{W} , which represents all values of \mathbf{w} such that $\{\phi_i(\mathbf{w})\}_{i=1}^{N_s}$ are equal, is not empty. According to its definition, the set \mathcal{W} can be written as

$$\begin{aligned} \mathcal{W} &\triangleq \{\mathbf{w} \mid \phi_k(\mathbf{w}) = \phi_{k'}(\mathbf{w}), \forall k \neq k'\} \\ &= \left\{ \mathbf{w} \mid |\mathbf{f}_k^H \tilde{\mathbf{h}}(\mathbf{w})| = |\mathbf{f}_{k'}^H \tilde{\mathbf{h}}(\mathbf{w})|, \forall k \neq k' \right\}. \end{aligned} \quad (3.46)$$

Noting that each relay in the system contributes only to one tap of the end-to-end CIR, we define \mathcal{R}_n as the set of weight vectors of those relays which contribute to the n -th tap of the channel impulse response. It has been shown in [63] that $\mathcal{W} = \bigcup_{n=0}^{N-1} \mathcal{R}_n$. In other

words, if $\mathbf{w} \in \mathcal{R}_n$, then only those relays which contribute to the n -th tap of the end-to-end CIR are turned on and the remainder of the relays are switched off. Note that the sets $\{\mathcal{R}_n\}_{n=0}^{N-1}$ are mutually exclusive, i.e., $\mathcal{R}_n \cap \mathcal{R}_{n'} = \emptyset$, for $n \neq n'$, hence the optimal value of \mathbf{w} belongs to one of these sets. Restricting \mathbf{w} such that it belongs to the set \mathcal{W} , we can write the optimization problem (3.44) as

$$\begin{aligned}
& \max_{\mathbf{w}} \max_{\alpha} \left(\frac{\alpha}{N_s} \right)^{N_s} \\
& \text{subject to} \quad \mathbf{w}^H \mathbf{w} \leq \frac{P_{max}}{\sigma^2}, \\
& \quad \sum_{i=1}^{N_s} \phi_i(\mathbf{w}) = \alpha \\
& \text{and} \quad \mathbf{w} \in \bigcup_{n=0}^{N-1} \mathcal{R}_n
\end{aligned} \tag{3.47}$$

or, equivalently, as

$$\begin{aligned}
& \max_{\mathbf{w}} \left(\frac{\sum_{i=1}^{N_s} \phi_i(\mathbf{w})}{N_s} \right)^{N_s} \\
& \text{subject to} \quad \mathbf{w}^H \mathbf{w} \leq \frac{P_{max}}{\sigma^2} \\
& \text{and} \quad \mathbf{w} \in \bigcup_{n=0}^{N-1} \mathcal{R}_n.
\end{aligned} \tag{3.48}$$

The optimization problem (3.48) can be now simplified as

$$\begin{aligned}
& \max_{\mathbf{w}} && \sum_{i=1}^{N_s} \phi_i(\mathbf{w}) \\
& \text{subject to} && \mathbf{w}^H \mathbf{w} \leq \frac{P_{max}}{\sigma^2} \\
& \text{and} && \mathbf{w} \in \bigcup_{n=0}^{N-1} \mathcal{R}_n.
\end{aligned} \tag{3.49}$$

Using (3.42), we write the objective function in (3.49) as

$$\begin{aligned}
\sum_{i=1}^{N_s} \phi_i(\mathbf{w}) &= N_s + \frac{\gamma(\mathbf{w}) \sum_{i=1}^{N_s} |\mathbf{f}_i^H \tilde{\mathbf{h}}(\mathbf{w})|^2}{2\sigma^2 \beta_1(\mathbf{w}) \beta_2(\mathbf{w})} \\
&= N_s + \frac{\gamma(\mathbf{w}) \|\tilde{\mathbf{h}}(\mathbf{w})\|^2}{2\sigma^2 \beta_1(\mathbf{w}) \beta_2(\mathbf{w})}
\end{aligned} \tag{3.50}$$

where in the second equality, we have used the Parseval's theorem:

$$\sum_{i=1}^{N_s} |\mathbf{f}_i^H \tilde{\mathbf{h}}(\mathbf{w})|^2 = \|\tilde{\mathbf{h}}(\mathbf{w})\|^2. \tag{3.51}$$

Using (3.50), we can write the optimization problem (3.49) as

$$\begin{aligned}
& \max_{\mathbf{w}} && N_s + \frac{\gamma(\mathbf{w})\|\tilde{\mathbf{h}}(\mathbf{w})\|^2}{2\sigma^2\beta_1(\mathbf{w})\beta_2(\mathbf{w})} \\
& \text{subject to} && \mathbf{w}^H\mathbf{w} \leq \frac{P_{max}}{\sigma^2} \\
& \text{and} && \mathbf{w} \in \bigcup_{n=0}^{N-1} \mathcal{R}_n.
\end{aligned} \tag{3.52}$$

As the sets $\{\mathcal{R}_n\}_{n=0}^{N-1}$ are mutually exclusive, the optimal value of \mathbf{w} can be found by decomposing the optimization problem (3.52) into a set of maximum N subproblems, each of which assumes that \mathbf{w} belongs to one of the sets $\{\mathcal{R}_n\}_{n=0}^{N-1}$. The optimal value of \mathbf{w} is eventually found by finding the solutions to these problems and determining that which results in the maximum possible value for the cost function. In other words, we can write the optimization problem (3.52) as

$$\begin{aligned}
& \max_{0 \leq n \leq N-1} \max_{\mathbf{w}} && N_s + \frac{\gamma(\mathbf{w})\|\tilde{\mathbf{h}}(\mathbf{w})\|^2}{2\sigma^2\beta_1(\mathbf{w})\beta_2(\mathbf{w})} \\
& \text{subject to} && \mathbf{w}^H\mathbf{w} \leq \frac{P_{max}}{\sigma^2} \\
& \text{and} && \mathbf{w} \in \mathcal{R}_n.
\end{aligned} \tag{3.53}$$

We use L_n to denote the number of the relays which contribute to the n -th tap of the end-to-end CIR and let \mathbf{w}_n represent the $L_n \times 1$ vector of the weights of those relays which contribute to the n -th tap of the end-to-end CIR. If $\mathbf{w} \in \mathcal{R}_n$, then, using (3.27), we can

write

$$\|\tilde{\mathbf{h}}(\mathbf{w})\|^2 = \|\mathbf{h}(\mathbf{w})\|^2 = \mathbf{w}^H \mathbf{B}^H \mathbf{B} \mathbf{w} = \mathbf{w}_n^H \mathbf{b}_n \mathbf{b}_n^H \mathbf{w}_n \quad (3.54)$$

where \mathbf{b}_n^H is an $1 \times L_n$ vector which captures the non-zero entries of the $(n+1)$ -th row of matrix \mathbf{B} , for $n = 0, 1, \dots, N-1$. Using (3.54) along with the definition of $\gamma(\mathbf{w})$, $\beta_1(\mathbf{w})$ and $\beta_2(\mathbf{w})$ and ignoring the constant N_s , we can write the optimization problem (3.53) as

$$\begin{aligned} & \max_{0 \leq n \leq N-1} \max_{\mathbf{w}_n} \frac{(P_{max} - \sigma^2 \mathbf{w}_n^H \mathbf{w}_n) \mathbf{w}_n^H \mathbf{b}_n \mathbf{b}_n^H \mathbf{w}_n}{2\sigma^2(1 + \|\mathbf{G}_1^{(n)} \mathbf{w}_n\|^2)(1 + \|\mathbf{G}_2^{(n)} \mathbf{w}_n\|^2)} \\ & \text{subject to} \quad \mathbf{w}_n^H \mathbf{w}_n \leq \frac{P_{max}}{\sigma^2} \end{aligned} \quad (3.55)$$

where $\mathbf{G}_q^{(n)}$, for $q \in \{1, 2\}$, is an $L_n \times L_n$ diagonal matrix whose diagonal entries are a subset of the diagonal entries of \mathbf{G}_q which correspond to those relays that contribute to the n -th tap of the end-to-end CIR. We can rewrite the optimization problem (3.55) in the following form:

$$\begin{aligned} & \max_{0 \leq n \leq N-1} \max_{\mathbf{w}_n} \frac{(P_{max} - \sigma^2 \mathbf{w}_n^H \mathbf{w}_n) \mathbf{w}_n^H \mathbf{b}_n \mathbf{b}_n^H \mathbf{w}_n}{2\sigma^2(1 + \mathbf{w}_n^H \mathbf{Q}_1^{(n)} \mathbf{w}_n)(1 + \mathbf{w}_n^H \mathbf{Q}_2^{(n)} \mathbf{w}_n)} \\ & \text{subject to} \quad \mathbf{w}_n^H \mathbf{w}_n \leq \frac{P_{max}}{\sigma^2} \end{aligned} \quad (3.56)$$

where $\mathbf{Q}_q^{(n)} \triangleq (\mathbf{G}_q^{(n)})^H \mathbf{G}_q^{(n)}$, for $q = 1, 2$. It is now well-known that the inner maximization problem (3.56) is amenable to a semi-closed-form solution for the optimal value of \mathbf{w}_n ,

denoted as \mathbf{w}_n° and is given by

$$\mathbf{w}_n^\circ = k_n \sqrt{2\nu_n} \left(2\mu_n \mathbf{Q}_1^{(n)} + 2\nu_n \mathbf{Q}_2^{(n)} + \mathbf{I} \right)^{-1} \mathbf{b}_n. \quad (3.57)$$

Here, the following definitions are used:

$$\nu_n \triangleq 0.5P_{max}/\sigma^2 - \mu_n \quad (3.58)$$

$$k_n \triangleq \left(\mathbf{b}_n^H \left(\mathbf{I} + 2\mu_n \mathbf{Q}_1^{(n)} \right) \left(\mathbf{I} + 2\mu_n \mathbf{Q}_1^{(n)} + 2\nu_n \mathbf{Q}_2^{(n)} \right)^{-2} \mathbf{b}_n \right)^{-\frac{1}{2}} \quad (3.59)$$

where the parameter μ_n is the *unique* solution to (3.60). equation:

$$\begin{aligned} & (P_{max}/\sigma^2 - 4\mu_n) \mathbf{b}_n^H (2\mu_n \mathbf{Q}_1^{(n)} + (P_{max}/\sigma^2 - 2\mu_n) \mathbf{Q}_2^{(n)} + \mathbf{I})^{-1} \mathbf{b}_n - \\ & \mu_n (P_{max}/\sigma^2 - 2\mu_n) \mathbf{b}_n^H (2\mu_n \mathbf{Q}_1^{(n)} + (P_{max}/\sigma^2 - 2\mu_n) \mathbf{Q}_2^{(n)} + \mathbf{I})^{-2} (2\mathbf{Q}_1^{(n)} - 2\mathbf{Q}_2^{(n)}) \mathbf{b}_n = 0 \end{aligned} \quad (3.60)$$

which satisfies $0 \leq \mu_n \leq 0.5P_{max}/\sigma^2$. A simple bisection method can be used to obtain the value of μ_n . With the so obtained μ_n , the values of ν_n and κ_n can be calculated as in (3.58) and (3.59), respectively, and for any given n , the optimal value of the beamforming weight vector can be eventually obtained as in (3.57). After finding $\{\mathbf{w}_n^\circ\}_{n=0}^{N-1}$, the optimal n is determined by finding the value of \mathbf{w}_n , which leads to the largest value of the objective

function in (3.56). Therefore, the optimal value of n can be obtained as

$$n^\circ = \arg \max_{0 \leq n \leq N-1} \frac{(P_{max} - \sigma^2 \|\mathbf{w}_n^\circ\|) |\mathbf{b}_n^H \mathbf{w}_n^\circ|^2}{2\sigma^2(1 + \mathbf{w}_n^{\circ,H} \mathbf{Q}_1^{(n)} \mathbf{w}_n^\circ)(1 + \mathbf{w}_n^{\circ,H} \mathbf{Q}_2^{(n)} \mathbf{w}_n^\circ)}. \quad (3.61)$$

Expressed differently, this approach is a process of searching for the set of relays which contribute only to one tap of the end-to-end CIR while resulting in the highest value of sum-rate should their beamforming weights be chosen optimally. In the search for the optimum n among all taps, once the corresponding relay beamforming vector is found and the corresponding sum-rate is calculated, the value of n with the maximum sum-rate is introduced as the optimal n . Finally, using (3.38) and (3.39), we obtain the optimal value of the transceivers' transmit powers as

$$P_1^\circ = \frac{\gamma(\mathbf{w}_{\text{opt}})}{2\beta_1(\mathbf{w}_{\text{opt}})} = \frac{P_{max} - \sigma^2 \|\mathbf{w}_{\text{opt}}\|^2}{2(1 + \|\mathbf{G}_1 \mathbf{w}_{\text{opt}}\|^2)} \quad (3.62)$$

$$P_2^\circ = \frac{\gamma(\mathbf{w}_{\text{opt}})}{2\beta_2(\mathbf{w}_{\text{opt}})} = \frac{P_{max} - \sigma^2 \|\mathbf{w}_{\text{opt}}\|^2}{2(1 + \|\mathbf{G}_2 \mathbf{w}_{\text{opt}}\|^2)} \quad (3.63)$$

where \mathbf{w}_{opt} denotes the optimal relay weight vector. If the l -th relay contributes to the tap n° of the end-to-end CIR, then the l -th entry of \mathbf{w}_{opt} is equal to the element of $\mathbf{w}_{n^\circ}^\circ$ which corresponds to the l -th relay. If the l -th relay does not contribute to the tap n° of the end-to-end CIR, then the l -th entry of \mathbf{w}_{opt} is zero.

Our proposed algorithm is described as Algorithm I. It should be emphasized that, since the algorithm selects a different relay subset for each value of n , the beamforming weights are obtained for those relay nodes, not all the relays. Moreover, because of the structure of matrix \mathbf{B} , each relay only contributes to one tap of the CIR. Hence, although

n is chosen between the N taps of the CIR, the maximum number of nonzero taps is not more than L .

3.3 Simulation Results

The general setup of the simulation results section contains two parts. In the first part, two single-antenna transceivers are communicate with each other with the help of $L = 60$ single-antenna relay nodes in the absence of symbol synchronization at the relays. The size of the signal blocks is assumed to be $N_s = 64$ symbols with a cyclic prefix length of $N = 8$. In the simulations, we have first generated different sets of delays which can be interpreted as different relay locations, and for each delay set (fixed relay location) different channels are randomly generated. The coefficients of the frequency flat channel are generated from independent complex Gaussian distribution with zero-mean and variance inversely proportional to the path loss. The pass loss corresponding to the propagation from/to any transceiver to/from any relay is set to be proportional to the corresponding delay to the power of 3. In other words, the noise variances of each channel coefficients are inversely proportional to the corresponding delay to the power of 3. The delay of propagation of each transceiver-relay link is generated as a random variable which is uniformly distributed from T_s to $4T_s$. Clearly, the delay of each relaying path has a triangular distribution in the interval $[2T_s, 8T_s]$.

Fig. 3.2 demonstrates the maximum sum-rate per sub-channel (divided by N_s) achieved in the two-way relaying scheme described in this thesis, versus the total transmit power P_{max} . In this figure, the performance of the proposed scheme is compared with that of the

Algorithm 1 : Sum-rate Optimal Network Beamforming and Power Allocation Algorithm

- 1: Set \mathbf{B} , \mathbf{G}_q for $q \in \{1, 2\}$, σ^2 and P_{max} as the input parameters of the algorithm.
- 2: Set $n = 0$.
- 3: If no relay contributes to the n -th tap of $h[\cdot]$ (i.e., if the $(n + 1)$ -th row of matrix \mathbf{B} is zero), go to Step 10.
- 4: Let the $1 \times L_n$ vector \mathbf{b}_n^H capture the non-zero entries of the $(n + 1)$ -th row of matrix \mathbf{B} , where L_n is the number of non-zero entries of the $(n + 1)$ -th row of matrix \mathbf{B} . For $q = 1, 2$, define $\mathbf{Q}_q^{(n)} \triangleq (\mathbf{G}_q^{(n)})^H \mathbf{G}_q^{(n)}$, where $\mathbf{G}_q^{(n)}$, for $q \in \{1, 2\}$, is an $L_n \times L_n$ diagonal matrix whose diagonal entries are a subset of the diagonal entries of \mathbf{G}_q which correspond to those relays which contribute to the n -th tap of the end-to-end CIR.
- 5: Use a bisection algorithm to obtain μ_n in the interval $[0 \ 0.5P_{max}/\sigma^2]$ such that

$$(P_{max}/\sigma^2 - 4\mu_n)\mathbf{b}_n^H \left(2\mu_n \mathbf{Q}_1^{(n)} + (P_{max}/\sigma^2 - 2\mu_n)\mathbf{Q}_2^{(n)} + \mathbf{I} \right)^{-1} \mathbf{b}_n - \\ \mu_n(P_{max}/\sigma^2 - 2\mu_n)\mathbf{b}_n^H \left(2\mu_n \mathbf{Q}_1^{(n)} + (P_{max}/\sigma^2 - 2\mu_n)\mathbf{Q}_2^{(n)} + \mathbf{I} \right)^{-2} \left(2\mathbf{Q}_1^{(n)} - 2\mathbf{Q}_2^{(n)} \right) \mathbf{b}_n = 0.$$

- 6: Calculate $\nu_n = 0.5P_{max}/\sigma^2 - \mu_n$.
- 7: Calculate κ_n using

$$\kappa_n = \left(\mathbf{b}_n^H \left(\mathbf{I} + 2\mu_n \mathbf{Q}_1^{(n)} \right) \left(\mathbf{I} + 2\mu_n \mathbf{Q}_1^{(n)} + 2\nu_n \mathbf{Q}_2^{(n)} \right)^{-2} \mathbf{b}_n \right)^{-\frac{1}{2}}.$$

- 8: Having obtained the values for κ_n , μ_n and ν_n , calculate \mathbf{w}_n° as

$$\mathbf{w}_n^\circ = \kappa_n \sqrt{2\nu_n} \left(2\mu_n \mathbf{Q}_1^{(n)} + 2\nu_n \mathbf{Q}_2^{(n)} + \mathbf{I} \right)^{-1} \mathbf{b}_n.$$

- 9: Calculate the cost function $f_n(\mathbf{w}_n^\circ)$ as

$$f_n(\mathbf{w}_n^\circ) = \frac{(P_{max} - \sigma^2 \|\mathbf{w}_n^\circ\|^2) |\mathbf{b}_n^H \mathbf{w}_n^\circ|^2}{2\sigma^2 \left(\mathbf{w}_n^{\circ,H} \mathbf{Q}_1^{(n)} \mathbf{w}_n^\circ + 1 \right) \left(\mathbf{w}_n^{\circ,H} \mathbf{Q}_2^{(n)} \mathbf{w}_n^\circ + 1 \right)}.$$

- 10: Set $n = n + 1$, if $n \geq N$ go to the next step, otherwise go to Step 3.
- 11: Find the value of n which yields the maximum $f_n(\mathbf{w}_n^\circ)$, that is

$$n^\circ = \arg \max_{0 \leq n \leq N-1} f_n(\mathbf{w}_n^\circ)$$

- 12: Let \mathbf{w}_{opt} denote the optimal relay weight vector. If the l -th relay contributes to the tap n° of the end-to-end CIR, then the l -th entry of \mathbf{w}_{opt} is equal to the element of \mathbf{w}_n° which corresponds to the l -th relay. If the l -th relay does not contribute to the tap n° of the end-to-end CIR, then the l -th entry of \mathbf{w}_{opt} is zero.

13: Calculate the transceivers' transmit powers as

$$P_1^\circ = \frac{P_{max} - \sigma^2 \|\mathbf{w}_{opt}\|^2}{2(1 + \|\mathbf{G}_1 \mathbf{w}_{opt}\|^2)},$$

$$P_2^\circ = \frac{P_{max} - \sigma^2 \|\mathbf{w}_{opt}\|^2}{2(1 + \|\mathbf{G}_2 \mathbf{w}_{opt}\|^2)}.$$

so-called equal power allocation (EPA) approach in which the total power of the system is uniformly allocated across all nodes in the network. In other words, in the EPA approach, each node in the system consumes $(\frac{1}{L+2})$ fraction of the total available power budget. As can be seen from this figure, the sum-rate of Algorithm 1 significantly outperforms the EPA method, owing to its optimality.

Fig. 3.3 displays the average value of the total relay power of the proposed method versus the total available transmit power. For comparison, the performance of the EPA method is also given in Fig. 3.3. We observe that, in the EPA method, the power consumed in the relay nodes is higher than that in our proposed method by approximately 3 dB. This is equivalent to saying that the total relay power of the EPA method is nearly twice the power consumed by the relays in our proposed method. Therefore, in the EPA technique, the total relay power is unnecessarily high as there is no optimality in the system. On the other hand, in the proposed method, only half of the total available power is used by the relays.

To evaluate the reliability of the proposed algorithm, we plot the end-to-end bit-error-rate (BER) curves in Fig. 3.4 for both the proposed method and for the EPA method. This comparison allows the determination of how much performance can be retained while optimizing the power allocation among the nodes. It can be seen from Fig. 3.4 that the

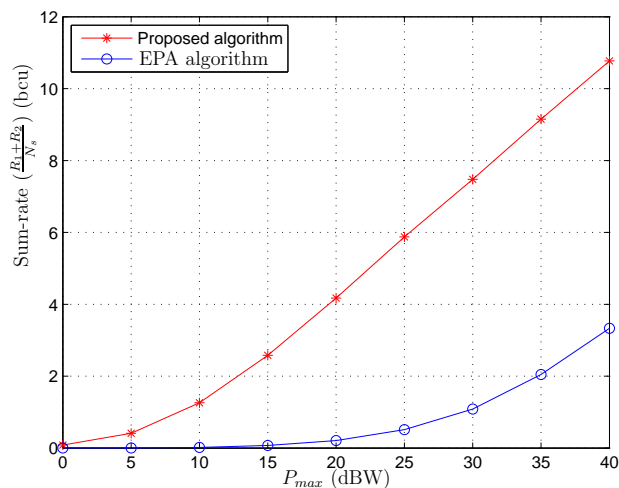


Figure 3.2: The sum-rate curves versus the total available transmit power, P_{max} ; for the proposed single-carrier scheme and for EPA method.

proposed method can achieve higher reliability compared to the EPA method.

In the second part of the simulation results, we study the performance of our proposed algorithm under different number of relays in Fig. 3.5. We assume a fixed value for the total available power, $P_{max} = 20$ dBw and $N_s = 64$, and the sum-rate is shown per sub-channel use. In Fig. 3.5, the more relays are used in the system, a higher sum-rate can be achieved.

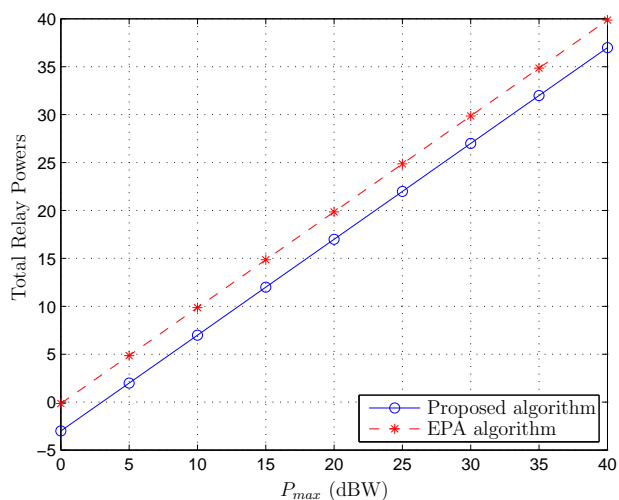


Figure 3.3: Total consumed relay power versus the total available transmit power, P_{max} ; for different schemes.

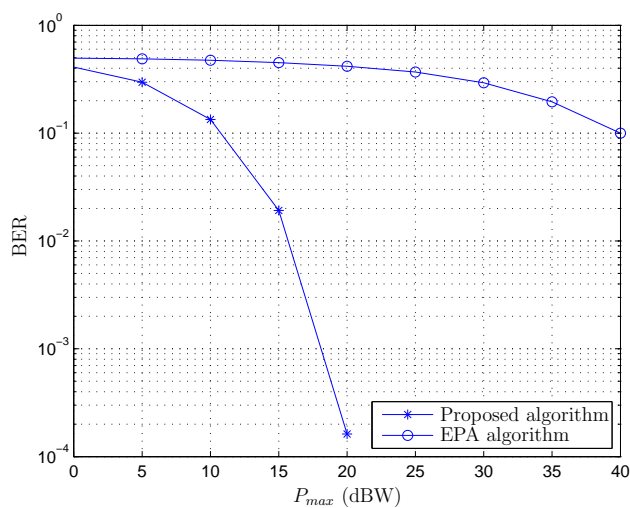


Figure 3.4: Bit error rate versus total available transmit power, P_{max} , for different methods.

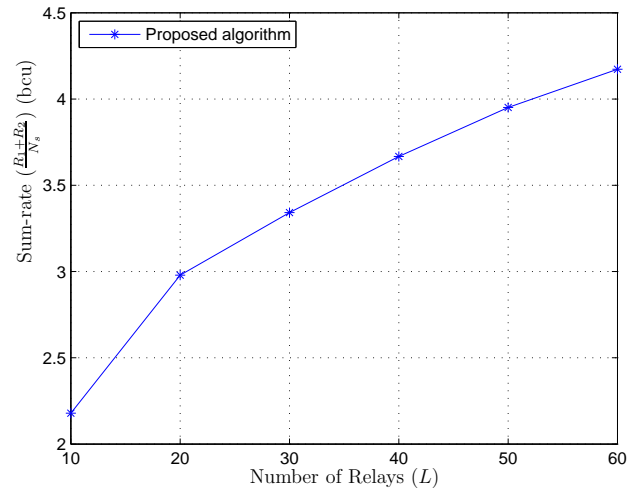


Figure 3.5: The sum-rate curves per sub-channel versus the number of relays for the proposed single-carrier scheme.

Chapter 4

Conclusion and Future Work

4.1 Conclusion

We considered an asynchronous single-carrier bidirectional (two-way) network which consists of two single-antenna transceivers and multiple single-antenna relay nodes. The two transceivers wish to exchange information with the help of the relays nodes. We assume that the network is asynchronous meaning that different transceiver-relay channels cause significantly different propagation delays in the signal they convey. As a result, the end-to-end channel is not amenable to a frequency flat model, rather a multi-path channel model with multiple taps appears to be more appropriate. Such a multi-path model for the end-to-end channel raises the issue of inter-symbol-interference (ISI) at the two transceivers. In a block transmission/reception scheme, ISI leads to inter-symbol-interference (IBI), which could result in loss in the sum-rate of the network, if different propagation delays is not

considered in the design of the system. Considering a block transmission/reception scheme and assuming a total transmit power budget, we maximized the sum-rate of this ISI end-to-end channel over the relay complex weights and transceivers' transmit powers. We rigorously proved that such a sum-rate maximization problem leads to a relay selection scheme, where only those relays which contribute to one tap of the end-to-end channel impulse response are turned on and the rest of the relays are switched off. Indeed, we prove that at the optimum, the end-to-end channel impulse response (CIR) has only one non-zero tap, rendering the end-to-end channel frequency flat. We presented the optimal value of the vector of the weights of the active relays and the optimal values of the transceivers' transmit powers.

4.2 Future work

Future research activities driving from the work presented in this thesis could include following studies.

- A dual problem where the total power consumed in the system is minimized under individual rate constraints for each transceiver seems to be an interesting topic in this system model.
- Another power minimization problem for our developed single carrier communication can be studied under two mean square error (MSE) constraints for each transceiver.
- In this thesis, we assumed a SISO model communication scheme where the transceivers and the relays are equipped with single antennas. Another research in this area could

involve the extension of the proposed scheme to a MIMO system model where the relays and the two transceivers are equipped with multiple antennas. Designing OFDM-based communication scheme for such MIMO system is an open area of research.

- We anticipate future research focus will move toward the systems with multi-users. In asynchronous bidirectional communication scheme, the asynchronous relay channel is shared among all users. This makes the system more complicated. However, the idea is likely to become a fertile ground for future work.

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