Non-Regenerative Multi-Antenna Two-Way and Multi-Way Relaying

Vom Fachbereich Elektrotechnik und Informationstechnik
der Technischen Universität Darmstadt
zur Erlangung des akademischen Grades eines
Doktor-Ingenieurs (Dr.-Ing.)
genehmigte Dissertation

von
Dipl.-Ing. Holger Degenhardt
Geboren am 07.01.1985 in Lindenfels

Referent:	Prof. Dr.-Ing. Anja Klein
Korreferent:	Prof. Dr.-Ing. Armin Dekorsy
Tag der Einreichung:	22.05.2014
Tag der mündlichen Prüfung:	29.10.2014

D 17

Darmstadt 2014

Danksagung

Zum Gelingen dieser Arbeit haben eine Vielzahl von Menschen beigetragen, denen ich mit den folgenden Zeilen danken möchte.

Mein besonderer Dank richtet sich an meine Betreuerin Prof. Dr.-Ing. Anja Klein, welche mir die Möglichkeit zur Promotion an ihrem Lehrstuhl gegeben hat. Zum einen gab Sie mir die Möglichkeit ein interessantes Forschungsthema im Rahmen eines Schwerpunktprogramms der Deutschen Forschungsgemeinschaft wissenschaftlich zu bearbeiten. Hierbei wurde ich von Ihr vorbildlich unterstützt und hatte somit auch die Möglichkeit an vielen internationalen Konferenzen aktiv teilzunehmen. Zum anderen gab Sie mir die Möglichkeit aktiv in der Lehre mitzuwirken. Durch die Betreuung von Lehrveranstaltungen und von acht studentischen Abschlussarbeiten sowie von mehreren Seminargruppen konnte ich interessante Erfahrungen sammeln und mich auch außerhalb meiner wissenschaftlichen Fähigkeiten stetig weiterentwickeln. Des Weiteren möchte ich mich bei Ihr und bei Dr. Yue Rong für meinen Forschungsaufenthalt an der Curtin University of Technology in Australien bedanken. Außerdem geht mein Dank an Prof. Dr.-Ing. Armin Dekorsy für die Übernahme des Korreferats.

Ein großes Dankeschön geht auch an meine ehemaligen Kollegen des Fachgebietes für Kommunikationstechnik. Durch den guten Zusammenhalt und die hervorragende Arbeitsatmosphäre hatte ich eine sehr schöne Zeit am Fachgebiet. Zudem werden mir die lustigen und zum Teil auch abenteuerlichen KT-Ausflüge in sehr guter Erinnerung bleiben. Meinen ehemaligen Kollegen möchte ich zudem für die vielen Diskussionen und Gespräche danken, durch die so manches Problem ganz einfach gelöst werden konnte. Ein Dank geht in diesem Zusammenhang auch an die Studenten, deren Abschlussarbeiten ich betreut habe oder die mich als studentische Hilfskraft bei der Bearbeitung meines Forschungsthemas unterstützt haben.

Besonders hervorheben möchte ich die Unterstützung meiner Familie und meiner Freunde. Sie haben mich sowohl während meines Studiums als auch während meiner Promotion bedingungslos unterstützt und durch zahlreiche Unternehmungen für die nötige Ablenkung gesorgt. Sie haben diese Arbeit erst möglich gemacht und dafür möchte ich mich bei Ihnen von ganzem Herzen bedanken.

Darmstadt im Oktober 2014
Holger Degenhardt

Kurzfassung

Relaisverfahren sind höchst vorteilhaft, um in drahtlosen Kommunikationssystemen Abschattungseffekte zu überwinden, Reichweiten zu erhöhen, die Energieeffizienz zu verbessern und den erzielbaren Datendurchsatz zu steigern. Um den erzielbaren Datendurchsatz weiter zu steigern, können Mehrantennentechniken genutzt werden. In dieser Arbeit werden Sendestrategien sowie Filterentwürfe für drei verschiedene nicht-regenerative Mehrantennen-Relais-Szenarien vorgeschlagen. Um Relaisverfahren in zukünftigen zellularen Szenarien zu untersuchen, wird ein zellulares Mehrnutzer-Relaisszenario betrachtet, in welchem eine Mehrantennen-Basisstation mit mehreren Mehrantennen-Mobilstationen bidirektional kommuniziert. Um Relaisverfahren in zukünftigen Ad-Hoc-Netzwerken und Sensornetzwerken zu untersuchen, werden ein Mehrpaar-Relaisszenario und ein Mehrgruppen-Mehrwege-Relaisszenario betrachtet. In dem Mehrpaar-Relaisszenario kommunizieren mehrere Mehrantennen-Knoten paarweise bidirektional miteinander. In dem Mehrgruppen-Mehrwege-Relaisszenario besteht jede Gruppe aus mehreren Mehrantennen-Knoten und jeder dieser Knoten teilt seine Daten mit allen anderen Knoten in seiner Gruppe. In allen Szenarien senden die Knoten während einer Vielfachzugriffsphase zeitgleich zur Relaisstation. Anschließend sendet die Relaisstation während mehrerer Broadcast (BC) Phasen linear verarbeitete Versionen dieser empfangenen Signale zu den Knoten. In dem zellularen Mehrnutzer-Relaisszenario und dem Mehrpaar-Relaisszenario wird nur eine BC Phase benötigt, da bidirektionale Kommunikationen betrachtet werden. In dem Mehrgruppen-Mehrwege-Relaisszenario werden mehrere BC Phasen benötigt, da jeder Knoten die Nachrichten von allen anderen Knoten in seiner Gruppe empfangen muss.

Jeder Knoten benötigt in der Regel eine unterschiedliche Datenrate zum Senden und Empfangen. So ist zum Beispiel die benötigte Datenrate in der Abwärtsstrecke von der Basisstation zu den Mobilstationen normalerweise höher als die benötigte Datenrate in der Aufwärtsstrecke von den Mobilstationen zur Basisstation. Um dies zu berücksichtigen, werden asymmetrische Datenraten-Forderungen (ADRF) eingeführt. Jedoch ist das Problem, die Summenrate für die betrachteten Szenarien mit und ohne Berücksichtigung der eingeführten ADRF zu maximieren, nicht konvex und die Suche nach einer optimalen Lösung hat eine sehr hohe Berechnungskomplexität. Aus diesem Grund wird in dieser Arbeit für jedes betrachtete Szenario eine Zerlegung des Problems der Summenratenmaximierung vorgeschlagen. Basierend auf diesen Problemzerlegungen werden die folgenden Verfahren eingeführt.

Im zellularen Mehrnutzer-Relaisszenario können die Sende- und Empfangssignale der Basisstation gemeinsam über alle Antennen verarbeitet werden. Für dieses Szenario

wird ein Filterentwurf für das Sendeempfangsfilter der Relaisstation vorgeschlagen, welcher ausnutzt, dass die Sendesignale der Mobilstationen gemeinsam an der Basisstation verarbeitet werden können. Für den vorgeschlagenen Filterentwurf werden die Fähigkeiten der Selbst- und der Schrittweisen-Interferenz-Auslöschung an den Knoten ausgenutzt. Zudem wird eine analytische Lösung basierend auf der Minimierung des gewichteten mittleren quadratischen Fehlers hergeleitet. Zusätzlich wird ein Entwurf des Sendefilters an der Basisstation vorgeschlagen, der die Fähigkeit der Schrittweisen-Interferenz-Auslöschung an den Mobilstationen ausnutzt. Außerdem wird ein Verfahren eingeführt, welches den gemeinsamen Entwurf der Filter an den Knoten und des Filters an der Relaisstation ermöglicht. Weiterhin werden zwei Sendestrategien vorgeschlagen, welche die Sendeleistungen an den Knoten und die Sendeleistungsverteilungen an der Basisstation und an der Relaisstation anpassen, um die betrachteten ADRF zu erfüllen. Zudem wird bei einer der vorgeschlagenen Sendestrategien eine Subträger-Zuweisung durchgeführt, um die Anzahl der gleichzeitig gesendeten Datenströme unter Berücksichtigung der ADRF anzupassen. Durch numerische Ergebnisse wird gezeigt, dass die Performanz der vorgeschlagenen Sendestrategien kombiniert mit den vorgeschlagenen Filterentwürfen an den Knoten und an der Relaisstation signifikant besser ist als die Performanz konventioneller Verfahren. So benötigen die vorgeschlagenen Verfahren zum Beispiel bis zu drei Antennen an der Relaisstation weniger als konventionelle Verfahren, um dieselbe Summenrate zu erzielen.

Im Mehrpaar-Relaisszenario können weder die Sende- noch die Empfangssignale, die zu unterschiedlichen Paaren gehören, gemeinsam an einem Knoten verarbeitet werden. Für dieses Szenario wird ein Filterentwurf für das Sendeempfangsfilter der Relaisstation vorgeschlagen, welcher die Interferenz zwischen den Knoten verschiedener Paare unterdrückt und somit eine gleichzeitige Kommunikation aller Paare ermöglicht. Zudem nutzt der vorgeschlagene Filterentwurf die Fähigkeiten der Selbst- und der Schrittweisen-Interferenz-Auslöschung an den Knoten aus. Basierend auf der Minimierung des gewichteten mittleren quadratischen Fehlers wird für den vorgeschlagenen Filterentwurf eine analytische Lösung hergeleitet. Zusätzlich werden zwei Verfahren vorgeschlagen, um die Sende- und Empfangsfilter an den Knoten zu entwerfen. Weiterhin werden zwei Sendestrategien vorgeschlagen, welche die Sendeleistungen an den Knoten und die Sendeleistungsverteilung an der Relaisstation anpassen, um die betrachteten ADRF zu erfüllen. Zudem wird bei einer der vorgeschlagenen Sendestrategien eine vollständige Suche durchgeführt, um die Anzahl der gleichzeitig gesendeten Datenströme unter Berücksichtigung der ADRF zu optimieren. Durch numerische Ergebnisse wird gezeigt, dass die Performanz der vorgeschlagenen Sendestrategie, welche die Anzahl der gleichzeitig gesendeten Datenströme optimiert, kombiniert mit den vorgeschlagenen Filterentwürfen an den Knoten und an der Relaisstation signifikant besser ist als die Performanz konventioneller Verfahren. So benötigt das vorgeschlagene Verfahren zum Beispiel bis zu drei Antennen an der Relaisstation weniger als konventionelle Verfahren, um dieselbe Summenrate zu erzielen.

Im Mehrgruppen-Mehrwege-Relaisszenario kann die Auswahl der Signale, die in den einzelnen BC Phasen ausgesendet werden, optimiert werden. Dies stellt eine zusätzliche Herausforderung im Vergleich zu den anderen zwei Relaisszenarien dar. Zudem kann an jedem Knoten eine gemeinsame zeitliche Verarbeitung der Empfangssignale aus den verschiedenen BC Phasen durchgeführt werden. Für dieses Szenario werden zwei Sendestrategien vorgeschlagen, welche die analoge Netzwerkcodierung verwenden, um die Fähigkeit der räumlichen Signalverarbeitung an den Knoten und an der Relaistation auszunutzen sowie um die Fähigkeit der zeitlichen Signalverarbeitung an den Knoten auszunutzen. Außerdem nutzen die vorgeschlagenen Sendestrategien die Fähigkeiten der Selbst- und der Schrittweisen-Interferenz-Auslöschung an den Knoten aus. Um eine effiziente Anwendung der vorgeschlagenen Sendestrategien zu ermöglichen, wird ein Entwurf des Sendeempfangsfilters der Relaisstation vorgeschlagen, welcher die Umsetzung von analoger Netzwerkcodierung ermöglicht. Der vorgeschlagene Filterentwurf basiert auf der Minimierung des gewichteten mittleren quadratischen Fehlers und es wird eine analytische Lösung für das Sendeempfangsfilter hergeleitet, welche Mithilfe von Gewichtsparametern angepasst werden kann. Zusätzlich wird ein Verfahren eingeführt, welches den gemeinsamen Entwurf der Empfangsfilter an den Knoten und des Sendeempfangsfilters an der Relaisstation ermöglicht. Durch numerische Ergebnisse wird gezeigt, dass die Performanz der vorgeschlagenen Sendestrategien kombiniert mit dem vorgeschlagenen gemeinsamen Entwurf der Empfangsfilter an den Knoten und des Sendeempfangsfilters an der Relaisstation signifikant besser ist als die Performanz konventioneller Verfahren. So benötigen die vorgeschlagene Verfahren zum Beispiel bei Betrachtung einer einzelnen Gruppe mit zehn Knoten bis zu sechs Antennen an der Relaisstation weniger als konventionelle Verfahren, um dieselbe Summenrate zu erzielen.

Abstract

Relaying techniques are highly beneficial in wireless communication systems to overcome shadowing effects, to increase the communication range, to improve the energy efficiency and to increase the achievable throughput. To further increase the achievable throughput, multi-antenna techniques can be exploited. In this thesis, transmit strategies and filter designs for three different non-regenerative multi-antenna relaying scenarios are proposed. To investigate relaying in future cellular networks, a cellular multi-user relaying scenario is considered where a multi-antenna base station wants to bidirectionally communicate with several multi-antenna mobile stations. To investigate relaying in future ad-hoc and sensor networks, a multi-pair relaying scenario and a multi-group multi-way relaying scenario are considered. In the multi-pair relaying scenario, several pairs of multi-antenna nodes want to perform bidirectional pairwise communications. In the multi-group multi-way relaying scenario, each group consists of several multi-antenna nodes and each node wants to share its data with all other nodes within its group. In all scenarios, the nodes simultaneously transmit to the relay station during one multiple access phase. Afterwards, the relay station retransmits linearly processed versions of the received signals during several broadcast (BC) phases to the nodes. In the cellular multi-user and in the multi-pair relaying scenario, one BC phase is required due to considering bidirectional communications. In the multi-group multi-way relaying scenario, several BC phases are required because each node has to receive the messages of all other nodes within its group.

To consider that each node typically requires different data rates for transmission and reception, e.g., the required data rates in downlink are typically higher than the required data rates in uplink, asymmetric data rate (ADR) requirements are introduced. However, the problem of maximizing the sum rate with and without considering the introduced ADR requirements is non-convex for the considered scenarios and searching for an optimal solution has a very high computational complexity. Thus, in this thesis, a decomposition of the sum rate maximization problem is proposed for each considered scenario. Based on the proposed decompositions, the following low-complexity approaches are introduced.

In the cellular multi-user relaying scenario, joint spatial processing over all antennas at the base station can be performed for transmission and reception. For this scenario, a relay transceive filter design is proposed which exploits that the signals transmitted by the mobile stations can be jointly processed at the base station. For the proposed filter design, the self-interference and successive interference cancellation capabilities of the nodes are exploited and an analytical solution based on minimizing the weighted

mean square error is derived. Furthermore, a successive interference cancellation aware transmit filter design at the base station is proposed. Additionally, an approach to enable a joint design of the filters at the nodes and at the relay station is introduced. Moreover, two low-complexity transmit strategies are proposed which adjust the transmit powers of the mobile stations and the transmit power distributions at the base station and at relay station to tackle the considered ADR requirements. Additionally, one of the proposed transmit strategies performs a low-complexity subcarrier allocation to adjust the numbers of simultaneously transmitted data streams with respect to the considered ADR requirements. By numerical results, it is shown that the proposed transmit strategies combined with the proposed filter designs at the nodes and at the relay station significantly outperform conventional approaches. For instance, for the considered configurations, the proposed approaches require up to three antennas less at the relay station than conventional approaches to achieve the same sum rate.

In the multi-pair relaying scenario, neither the transmit signals nor the receive signals of nodes which belong to different pairs can be jointly processed at one node. For this scenario, a relay transceive filter design is proposed which suppresses the interferences between nodes of different pairs and thus, enables the simultaneous communication of all pairs. Furthermore, the proposed filter design exploits the capability of the nodes to perform self-interference and successive interference cancellation. The proposed relay transceive filter design is based on minimizing the weighted mean square error and an analytical solution is derived. Furthermore, two approaches for designing the transmit and receive filters at the multi-antenna nodes are introduced. Moreover, two low-complexity transmit strategies are proposed which adjust the transmit powers of the nodes and the transmit power distribution at the relay station to tackle the considered ADR requirements. Additionally, one of the proposed transmit strategies performs an exhaustive search to optimize the numbers of simultaneously transmitted data streams with respect to the considered ADR requirements. By numerical results, it is shown that the proposed transmit strategy which additionally optimizes the numbers of simultaneously transmitted data streams combined with the proposed filter designs at the nodes and at the relay station significantly outperforms conventional approaches. For instance, for the considered configurations, the proposed approach requires up to three antennas less at the relay station than conventional approaches to achieve the same sum rate.

In the multi-group multi-way relaying scenario, the selection of the signals which are retransmitted in each BC phase can be optimized which is an additional challenge compared to the other two relaying scenarios. Furthermore, the nodes can additionally perform joint temporal receive processing over the received signals of the different BC phases. For this scenario, two low-complexity transmit strategies are proposed which

utilize analog network coding to exploit the spatial processing capabilities of the nodes and of the relay station as well as the capability of the nodes to perform temporal receive processing over the received signals of the different BC phases. Additionally, the proposed transmit strategies exploit the capability of the nodes to perform self-interference and successive interference cancellation. To enable an efficient application of the proposed transmit strategies, an analog network coding aware relay transceive filter design is proposed. The relay transceive filter design is based on minimizing the weighted mean square error and an analytical solution is derived which can be adjusted via the considered weighting parameters. Additionally, a joint approach for designing the receive filters at the nodes together with the proposed analog network coding aware relay transceive filter is introduced. By numerical results, it is shown that the proposed transmit strategies combined with the proposed joint filter design at the nodes and at the relay station significantly outperform conventional approaches. For instance, if a single group with ten nodes is considered, the proposed approaches require up to six antennas less at the relay station than conventional approaches to achieve the same sum rate.

Contents

Chapter 1
Introduction

1.1 Multi-Antenna Two-Hop Relaying

The demand for high data rate wireless access rapidly increases. In urban as well as in rural areas, people desire high data rate wireless access to share personal or public data such as pictures and videos. Furthermore, machine to machine (M2M) communications will play an important role in future wireless communication systems [Fet12]. In M2M communication systems, the number of nodes which want to share information can be much higher than in a cellular system and this can significantly increase the required throughput.

In conventional wireless communication systems, transmissions are performed from source to destination nodes via the direct link channel between these nodes. To recover the transmitted signal at the destination, the required ratio between the received signal power and the received interference and noise power, termed signal-to-interference-plus-noise ratio (SINR), has to be sufficiently high [Kam11]. Thus, the transmit power at the source has to be increased to overcome shadowing effects or to increase the communication range, because the received signal power decreases at least with the increasing distance between source and destination squared [TV05]. Furthermore, to increase the achievable data rates, the transmit power has to be increased as well such that a higher SINR is achieved at the receiver. However, arbitrarily increasing the transmit power is not possible due to practical issues. For example, an increased transmit power would reduce the battery lifetime of mobile devices, increase the energy costs per bit, cause higher interference to neighboring nodes which reuse the same resources and impact the electromagnetic compatibility.

To overcome the aforementioned problems, relaying is considered as an energy- and cost-efficient solution to increase the communication range and the achievable throughput in future wireless communication systems [HBG+13, WT12, SDR12]. In relaying, one or several intermediate nodes, termed relay stations, support the communications between the source and the destination nodes. An example for a simple relaying scenario is shown in Figure 1.1. In this scenario, the nodes S_1 and S_2 want to bidirectionally exchange information. The direct link between the nodes is assumed to be weak because it is affected by strong shadowing effects and thus, the communications are performed via an intermediate relay station, termed RS.

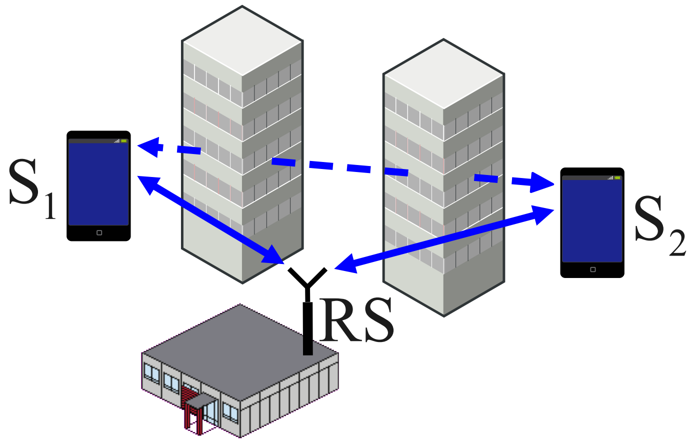

Figure 1.1. Single-pair bidirectional relaying.

At the intermediate relay stations, different signal processing approaches can be applied. Non-regenerative relaying, also known as amplify-and-forward, and regenerative relaying, also known as decode-and-forward, are two of the most prominent signal processing approaches for relaying [TH07, OJWB09, WT12, SDR12]. In non-regenerative relaying, the relay retransmits a linearly processed version of the received signals [TH07]. In regenerative relaying, the relay decodes the received signals and re-encodes these signals before the retransmission [OJWB09].

The main disadvantages of non-regenerative compared to regenerative relaying are that the received noise is retransmitted by the relay stations and different desired signals are spatially superimposed because a combination on bit-level is not possible. Thus, the available transmit power at the relay stations is used less efficiently. One main advantage of non-regenerative relaying compared to regenerative relaying is that no decoding and re-encoding is required at the relay stations which reduces the latency, reduces the required processing power and simplifies security problems [TH07]. Furthermore, due to not requiring a decoding of the signals at the relay stations, a spatial separation of the superimposed received signals is also not required which can reduce the required number of antennas at the relay stations and can increase the achievable data rates.

To further increase the achievable throughput, multiple-input multiple-output (MIMO) techniques can be considered [GJJV03, PNG03]. MIMO techniques can be used to increase the achievable throughput by spatially multiplexing different data streams. Moreover, MIMO techniques can be used at the intermediate relay stations to spatially separate the communications of different nodes. In case of multi-antenna nodes, successive interference cancellation (SIC) can be considered at the receiving nodes to improve the estimation of multiple simultaneously received data streams [TV05, ZCW05].

Due to the aforementioned advantages of non-regenerative relaying and MIMO techniques, this thesis focuses on non-regenerative MIMO relaying. Furthermore, this thesis focuses on scenarios where the communications between several nodes are supported by a single multi-antenna relay station. The transmissions which are performed via the relay station can be regarded as two-hop transmissions where the first hop for each transmitted signal refers to the transmission from its source to the relay station and the second hop refers to its transmission from the relay station to its destination. Additional hops between source and destination are not considered because of the noise propagation in non-regenerative relaying which degrades the performance.

Due to the high dynamic range between the transmitted and the received signal powers, it is typically assumed that the nodes and the relay station cannot transmit and receive simultaneously which is referred to as half-duplex constraint [RW07]. To resolve this problem, the most prominent relaying protocols, which are briefly described in the next paragraph, use orthogonal time resources for transmission and reception. The durations of the reception and the transmission times at the relay station are equal because due to considering non-regenerative relaying, the relay station retransmit a linearly processed version of the received signals.

The two most prominent relaying protocols for the scenario of Figure 1.1, where the nodes S_1 and S_2 want to bidirectionally exchange information via a relay station, are termed one-way and two-way relaying [Ung09]. For this single-pair scenario, the one-way relaying protocol requires four time slots to exchange the information between the nodes [Ung09,SDR12]. In the first time slot, S_1 transmits its signal to the relay station. In the second time slot, the relay station retransmits a processed version of the received signal to S_2. In the third time slot, S_2 transmits its signal to the relay station and in the fourth time slot, the relay station retransmits a processed version of the received signal to S_1. For the same scenario, the two-way relaying protocol only requires two time slots to exchange the information between the nodes [RW05]. In the first time slot, both nodes S_1 and S_2 simultaneously transmit their signals to the relay station. In the second time slot, the relay station retransmits a processed version of the superimposed received signals back to the nodes. The nodes estimate the desired signals after subtracting their own transmit signal from the received superposition [RW07]. This is often referred to as self-interference cancellation [Ung09]. Due to requiring only half the number of time slots, the spectral efficiency of two-way relaying can be almost twice the spectral efficiency of one-way relaying.

In this thesis, three different relaying scenarios are investigated to cover cellular networks on the one hand and to cover ad-hoc and sensor networks on the other hand.

To investigate relaying in cellular networks, a cellular multi-user relaying scenario is considered. In this scenario, a multi-antenna base station wants to bidirectionally communicate with several multi-antenna mobile stations via an intermediate multi-antenna relay station. To investigate relaying in ad-hoc and sensor networks, a multi-pair relaying scenario and a multi-group multi-way relaying scenario are considered. In multi-pair relaying, several pairs of multi-antenna nodes want to perform bidirectional pairwise communications via an intermediate multi-antenna relay station. In multi-group multi-way relaying, several groups are considered which want to communicate via an intermediate multi-antenna relay station. Each group consists of several multi-antenna nodes and each node wants to share its data with all other nodes within its group. Typical multi-pair and multi-group multi-way relaying applications are video conferences, file sharing or multiplayer gaming as well as M2M, emergency or sensor applications. To enable high data rate transmissions, transmit strategies which are based on extensions and modifications of the two-way relaying protocol are introduced for the different scenarios. Furthermore, different approaches for efficiently designing the relay transceive filter and the transmit (Tx) and receive (Rx) filters at the nodes are proposed.

Considering bidirectional communications between the base station and the mobile stations in the cellular relaying scenario or communications between the nodes in the multi-pair relaying scenarios, each node typically requires different data rates for transmission and reception. Considering a cellular multi-user relaying scenario for instance, the required data rates in downlink are typically higher than the required data rates in uplink. Considering a file sharing application in the multi-pair relaying scenario for instance, the file sizes are typically different and thus, each node requires a different data rate rate for transmission if all files should be exchanged simultaneously. To handle such requirements, asymmetric data rate (ADR) requirements are introduced for the cellular multi-user and for the multi-pair two-way relaying scenario. ADR requirements can also be considered in the multi-group multi-way relaying scenario. However, for multi-group multi-way relaying, the focus is on the development of Tx strategies which efficiently combine the temporal processing capabilities of the nodes with the spatial processing capabilities of the relay station because this has not been investigated, so far. Nevertheless, the extension of these Tx strategies to consider ADR requirements is briefly described.

1.2 State of the Art

This section presents a review of the state of the art with regard to the different non-regenerative multi-antenna two-hop relaying scenarios considered in this thesis.

The basic building block for the considered two-hop relaying scenarios is the single-antenna single-pair relaying scenario which has been introduced in [Meu71]. In [Meu71], unidirectional communications between a single-antenna source node and a single-antenna destination node are supported by a single-antenna relay station. To perform the communications, the direct link between the source and the destination node as well as the two links between these nodes and the relay station are considered. Although this relaying scenario has been introduced many years ago, the capacity is still unknown [CG79]. To achieve diversity gains by utilizing all possible links, cooperative transmit strategies can be applied [SEA03, LW03, LTW04]. These strategies are based on jointly optimizing the transmissions of the source node and the relay station.

To increase the achievable throughput, MIMO techniques can be considered [GJJV03, PNG03]. Multi-antenna relaying scenarios where unidirectional communications are performed between a multi-antenna source node and a multi-antenna destination node with the help of an intermediate multi-antenna relay station have been investigated in [WZHM05, FT07, MnMVA07, Ron10, HW07, TH07, RTH09] and references therein. In [WZHM05], a full-duplex regenerative multi-antenna relay station is considered and capacity bounds for cooperative transmit strategies are computed. In [TH07, MnMVA07], transceiver designs for considering a non-regenerative half-duplex multi-antenna relay station are investigated considering the direct link. In [Ron10], the same scenario is investigated and the transmit filter of the source and the relay transceive filter are jointly optimized. In [FT07], the direct link is neglected and different regenerative and non-regenerative relaying schemes are investigated assuming that several half-duplex multi-antenna relay stations are located in between the source and the destination node. In [HW07, RTH09], the authors focus on non-regenerative one-way relaying schemes considering a single half-duplex multi-antenna relay station and neglecting the direct link.

For several applications such as video conferences or file sharing, bidirectional communications are required. For bidirectional communications, the two-way relaying protocol, which enables bidirectional communications between two half-duplex single-antenna or multi-antenna nodes via an intermediate half-duplex multi-antenna relay station, was proposed in [RW05, RW07] to overcome the duplexing loss of conventional one-way relaying schemes. To achieve this, the two-way relaying protocol applies analog network coding (ANC) at the relay station [ACLY00, KGK07]. The received signals are linearly combined before the retransmission taking into account that self-interference cancellation can be performed at the nodes before estimating the desired signals. The filter design for non-regenerative two-way relaying has been investigated in [Ung09, ZLCC09, RH09, LLSL09, XH11, Ron12, WT12] and references therein. In these papers, the direct link between the nodes is neglected because half-duplex nodes are considered which is favorable for practical implementations. In [Ung09], different relay transceive filter approaches for multi-antenna nodes are investigated considering one-way and two-way relaying. The focus is on the design of zero-forcing (ZF) and minimum mean square error (MMSE) relay transceive filters. The design of these filters is based on the derivations for conventional MIMO Rx and Tx filters which are presented in [Joh04]. Additionally, for MIMO two-way relaying, a relay transceive filter is introduced in [Ung09] which exploits that the nodes can perform self-interference cancellation. Exploiting self-interference cancellation for the relay transceive filter design is referred to as self-interference aware relay transceive filter design in the following. In [ZLCC09], the capacity region for non-regenerative two-way relaying is analyzed and optimal beamforming is investigated assuming single antenna nodes and a multi-antenna relay station. In [RH09], a transmit strategy is introduced which maximizes the weighted sum of the Frobenius norms of the effective channels considering two multi-antenna nodes which bidirectionally communicate via an intermediate multi-antenna relay station. In [LLSL09], the same scenario is considered and a gradient based relay transceive filter approach for sum rate maximization is presented. In [XH11, Ron12, WT12], the joint design of the spatial Tx filters at the nodes and of the relay transceive filter is investigated for MIMO two-way relaying considering different objective functions such as MMSE or weighted sum rate. Examples of regenerative relaying schemes for bidirectional communications can be found in [RW06, RW07, PY07, HKE+07, OJWB09, WWD13, WLW+14] and references therein.

In cellular scenarios, where a multi-antenna base station wants to bidirectionally exchange information with several multi-antenna mobile stations, a multi-antenna relay station can be integrated to increase the coverage and the throughput. Such scenarios are referred to as cellular multi-user relaying in the following. Non-regenerative cellular multi-user two-way relaying with single antenna mobile stations and a multi-antenna base station has been considered in [TS09, DKTL11, ZRH11, SYLV11, SYLV12, WTH12]. In [TS09, DKTL11], approaches which combine the idea of signal alignment for the Tx filter design at the base station with the idea of ZF for the relay transceive filter design are presented assuming that the base station and the relay station are equipped with the same number of antennas. In [ZRH11], three sub-optimal algorithms which are based on channel inversion, block diagonalization as well as on ZF dirty paper coding are presented for the filter design at the base station and at the relay station. In [SYLV11, SYLV12], an alternating optimization between the filters at the base station and the transceive filter at the relay station is proposed to maximize the sum rate under the constraints that the interferences between the Tx signals of different mobile stations should be zero. In this approach, the optimization of the relay transceive filter and the optimization of the Tx filter at the base station results in non-convex problems and thus, analytical solutions cannot be obtained. To resolve this problem, a suboptimal low-complexity approach is additionally presented in [SYLV12]. In [WTH12], quality-of-service requirements should be ensured for each mobile station in downlink while in the uplink, either the mean squared error (MSE) should be minimized or the sum rate should be maximized. For this approach, analytical solutions can neither be obtained for the Tx filter at the base station nor for the relay transceive filter. Regenerative two-way relaying in such a single cell two-way relaying scenario has for example been investigated in [WM07, EW08]. In non-regenerative two-way relaying, an MMSE based relay transceive filter design takes into account the noise powers at the nodes and at the relay station as well as the interference powers at the nodes. Furthermore, if an MMSE based relay transceive filter design is considered, low-complexity solutions can be typically derived. However, for non-regenerative cellular multi-user two-way relaying, an MMSE based relay transceive filter design exploiting the capabilities of the multi-antenna nodes to perform self-interference cancellation and SIC has not been presented in previous works so far. Furthermore, ADR requirements have not been considered. Moreover, the aforementioned publications only consider single antenna mobile stations.

In ad-hoc networks, where multiple nodes want to simultaneously perform pairwise bidirectional communications, an intermediate multi-antenna relay station can be used to coordinate the transmissions and to spatially separate the communications of the different pairs. Such scenarios are referred to as multi-pair relaying in the following. Non-regenerative multi-pair two-way relaying considering single antenna nodes has been investigated in [YZGK10, AK10b, LDLG11, ZDP⁺11, TW12]. In [YZGK10, AK10b, LDLG11, ZDP⁺11], different relay transceive filters based on the idea of ZF block-diagonalization have been proposed to exploit the self-interference cancellation capability of the nodes. In [TW12], relay transceive filters based on semidefinite relaxation and on ZF have been proposed to maximize the minimum achievable data rate among all nodes exploiting self-interference cancellation. The approach based on semidefinite relaxation achieves a good performance, but results in a high computational complexity. The approach based on ZF has a lower computational complexity, but it performs significantly worse than the semidefinite relaxation approach. For non-regenerative multi-pair two-way relaying considering multi-antenna nodes, relay transceive filters based on ZF and MMSE have been proposed in [JS10]. These filters suppress self-interferences and thus, all received signals have to be spatially separated at the relay station. An MMSE based relay transceive filter design for multi-pair two-way relaying exploiting the capabilities of the multi-antenna nodes to perform self-interference cancellation and SIC has not been presented in previous works so far. Furthermore, ADR requirements have not been considered. Moreover, the optimization of the numbers of simultaneously transmitted data streams has not been investigated.

Considering multiplayer gaming, video conferences or emergency applications usually the data exchange between multiple nodes which belong to a specific group is required. In these scenarios, each node of a group wants to share its data with all other nodes within its group and an intermediate multi-antenna relay station can be used to coordinate and support the communications of the nodes and to spatially separate the communications of different groups. Such scenarios are referred to as multi-group multi-way relaying in the following. Different schemes and approaches for regenerative multi-group multi-way relaying have been considered in [OJK10, OJK11, OKJ12, HIR12, AK11b]. In [GYGP09, GYGP13, AK10a, CZ12, AK11a], single antenna nodes are considered and different non-regenerative multi-group multi-way relaying scenarios are investigated. In [GYGP09, GYGP13], the full-duplex multi-group multi-way relay channel is investigated and time division multiple access (TDMA) is applied to separate the communications of different groups. Non-regenerative multi-way relaying via a half-duplex multi-antenna relay station for a single group scenario is considered in [AK10a, CZ12]. In these scenarios, the communications are performed in one multiple access (MAC) and several broadcast (BC) phases. In [AK10a], a transmit strategy based on analog network coding (ANC) is proposed to reduce the required number of BC phases. For the proposed transmit strategy, all received signals have to be spatially separated at the relay station. In [CZ12], joint receive processing over all BC phases with SIC is considered at each node to estimate the desired signals. The proposed scheme assumes random linear processing at the relay station to enable the communications within a single group. Considering random linear processing at the relay station, the communications of different groups cannot be spatially separated. In [AK11a], multi-group multi-way relaying is considered and different relay transceive filters are proposed to spatially separate the groups and to enable the multi-way communications within each group. The proposed filters separate all received signals at the relay station. An approach to efficiently combine the spatial processing at the relay station and the temporal processing at the nodes has not been presented in previous works so far. Furthermore, ANC has not fully been exploited by the proposed Tx strategies in [AK10a, AK11a] because part of the signals which can be canceled at the nodes are intentionally suppressed at the relay station. Moreover, an MMSE based relay transceive filter design which can exploit the capabilities of the nodes to perform self-interference cancellation, temporal receive processing and SIC has not been presented. Besides, a joint spatial filter design between the filters at the nodes and at the relay station has not been investigated. Additionally, the aforementioned publications only consider single antenna nodes.

1.3 Open Issues

In this section, the open issues arising from the review of the state of the art are summarized for the cellular multi-user two-way relaying scenario, the multi-pair two-way relaying scenario and the multi-group multi-way relaying scenario.

In the cellular multi-user two-way relaying scenario, joint spatial processing over all antennas at the base station can be performed for transmission and reception. Thus, the Tx signals in the downlink from the base station to the mobile stations can be jointly processed and the Rx signals in the uplink from the mobile stations to the base station can be jointly processed. This is the main difference compared to the considered multi-pair two-way relaying scenario, where neither the Tx signals nor the Rx signals of nodes which belong to different pairs can be jointly processed. Nevertheless, the cellular multi-user two-way relaying scenario and the multi-pair two-way relaying scenario have several open issues in common. However, these open issues have to be tackled individually for each scenario due to the aforementioned difference caused by the spatial processing capabilities at the base station.

To maximize the sum rate in the cellular multi-user and the multi-pair two-way relaying scenario, the Tx and Rx filters of the nodes, the Tx powers of the nodes, the relay transceive filter and the numbers of simultaneously transmitted data streams of the nodes have to be jointly optimized over all subcarriers. Due to the high computational complexity of finding an optimal solution for this problem, suboptimal low-complexity approaches are required which achieve high sum rates. For these low-complexity approaches, the spatial processing capabilities at the nodes and at the relay station as well the capability of the nodes to perform self-interference cancellation and SIC shall be exploited. Thus, the open issues arising from the review of the state of the art are as follows:

1: How to efficiently decompose the aforementioned problem of maximizing the sum rate into different low-complexity subproblems?

2: How to define a system model which considers multi-antenna nodes which can perform linear Tx and Rx processing, self-interference cancellation and SIC?

3: How to perform a low-complexity design of the relay transceive filter and of the Tx and Rx filters of the multi-antenna nodes which exploits the self-interference cancellation and the SIC capabilities of the nodes?

Furthermore, as mentioned in Section 1.1, ADR requirements shall be considered in both scenarios. With respect to the consideration of ADR requirements, the additional open issues arising from the review of the state of the art are as follows:

4: How to extend the relay transceive filter design and the Tx filter design at the nodes to enable an adjustment of these filters with respect to the consideration of ADR requirements?

5: How to adjust the relay transceive filter and the Tx filters of the nodes to tackle specific ADR requirements?

6: How to optimize the numbers of simultaneously transmitted data streams?

In the multi-group multi-way relaying scenario, several BC phases are required to enable the communications between the nodes. Thus, the selection of the signals which are retransmitted in each BC phase can be optimized which is an additional challenge compared to the considered two-way relaying scenarios. Furthermore, in the multi-group multi-way relaying scenario, the nodes can additionally perform joint temporal Rx processing over the received signals of the different BC phases.

To maximize the sum rate in the multi-group multi-way relaying scenario, the selection of the signals which are retransmitted in each BC phase has to be optimized. Furthermore, the spatial Tx and Rx filters of the nodes, the temporal Rx filters of the nodes and the relay transceive filters have to be optimized with respect to the selected signals for each BC phase. Due to the high computational complexity of finding an optimal solution for this problem, suboptimal low-complexity approaches are required which achieve high sum rates. For these low-complexity approaches, the spatial processing capabilities at the nodes and at the relay station as well the capability of the nodes to perform joint temporal Rx processing over the received signals of the different BC phases shall be exploited. Additionally, the capability of the nodes to perform self-interference cancellation and SIC shall be exploited. Thus, the open issues arising from the review of the state of the art are as follows:

7: How to efficiently decompose the aforementioned problem of maximizing the sum rate into different low-complexity subproblems?

8: How to define a system model which considers multi-antenna nodes which can perform spatial Tx and Rx processing, temporal Rx processing, self-interference cancellation and SIC?

9: How to perform a low-complexity design of the relay transceive filter and of the Tx and Rx filters of the multi-antenna nodes which enables the utilization of the self-interference cancellation, SIC and temporal Rx processing capabilities of the nodes?

10: How to efficiently combine the temporal processing capabilities of the nodes and the spatial processing capabilities of the relay station? How to select the signals which are retransmitted in each BC phase such that the self-interference cancellation and the SIC capabilities as well as the temporal processing capabilities of the nodes are exploited?

1.4 Contributions and Thesis Overview

In this section, an overview of the thesis is presented by summarizing the main contributions which solve the open problems introduced in Section 1.3. In the following, the contents along with the main contributions of each chapter are briefly described.

In Chapter 2, the considered scenarios are briefly described and the assumptions which are valid throughout this thesis are introduced.

In Chapter 3, the cellular multi-user two-way relaying scenario is investigated. The main contributions which are presented in this chapter are as follows:

1a: A decomposition of the problem of maximizing the sum rate into different low-complexity subproblems is proposed.

2a: A system model for cellular multi-user two-way relaying considering multi-antenna nodes which can perform self-interference cancellation, linear Tx and Rx processing and SIC is introduced.

3a: A novel low-complexity relay transceive filter design is proposed which exploits that the signals transmitted by the mobile stations can be jointly processed at the base station. For the proposed filter design, the self-interference cancellation and the SIC capabilities of the nodes are exploited and an analytical solution based on minimizing the weighted MSE is derived. Furthermore, a SIC aware Tx filter design at the base station is proposed which exploits the capability of the mobile stations to perform SIC. Additionally, an approach to enable a joint design of the filters at the nodes and at the relay station is presented.

4a: To enable an adjustment of the achievable data rates in downlink from the base station to the mobile stations, the Tx filter design at the base station is extended by considering weighting parameters which enable an adjustment of the Tx power distribution at the base station. Furthermore, to enable an adjustment of the achievable data rates in uplink from the mobile stations to the base station, the Tx filter design at the mobile stations is extended by considering weighting parameters which enable an adjustment of the Tx powers of the mobile stations. Additionally, to enable an adjustment of the relay transceive filter with respect to the consideration of ADR requirements, the relay transceive filter design is extended by considering additional weighting parameters which enable an adjustment of the Tx power distribution at the relay station.

5a: A new low-complexity Tx strategy is proposed which adjusts the aforementioned weighting parameters with respect to the considered ADR requirements. By adjusting the weighting parameters at the base station and at the relay station, the achievable data rates in downlink are adjusted. Furthermore, by adjusting the weighting parameters at the mobile stations and at the relay station, the achievable data rates in uplink are adjusted.

6a: A new low-complexity Tx strategy is proposed which adjusts the aforementioned weighting parameters and additionally adjusts the numbers of simultaneously transmitted data streams. To adjust the numbers of simultaneously transmitted data streams, a low-complexity subcarrier allocation approach is proposed.

In Chapter 4, the multi-pair two-way relaying scenario is investigated. In principle, the presented contributions for multi-pair two-way relaying are similar to the contributions for cellular multi-user two-way relaying. However, due to considering spatially separated communication pairs in the multi-pair two-way relaying scenario, the filter designs at the nodes and at the relay station as well as the proposed Tx strategies have to be modified. For the filter designs at the nodes and at the relay station, it has to be considered that each node can only jointly process the Tx signals which are intended for its partner as well as the Rx signals which are transmitted by its partner and thus, the relay transceive filter has to suppress the interferences between different pairs. Furthermore, due to spatially separating different pairs, the adjustment of the weighting parameters and the optimization of the numbers of simultaneously transmitted data streams have to be modified to tackle the considered ADR requirements. Thus, the main contributions which are presented in this chapter are as follows:

1b: A decomposition of the problem of maximizing the sum rate into different low-complexity subproblems is proposed.

2b: A system model for multi-pair two-way relaying considering multi-antenna nodes which can perform self-interference cancellation, linear Tx and Rx processing and SIC is introduced.

3b: A novel low-complexity relay transceive filter design is proposed which considers that the interferences between different pairs have to be suppressed. For the proposed filter design, the self-interference cancellation and the SIC capabilities of the nodes are exploited and an analytical solution based on minimizing the weighted MSE is derived. Furthermore, low-complexity approaches for designing the Tx and Rx filters at the multi-antenna nodes are presented.

4b: To enable an adjustment of the relay transceive filter with respect to the consideration of ADR requirements, the relay transceive filter design is extended by considering weighting parameters which enable an adjustment of the Tx power distribution at the relay station. Furthermore, to enable an adjustment of the Tx filters at the nodes with respect to the consideration of ADR requirements, the Tx filter design at the nodes is extended by considering weighting parameters which enable an adjustment of the Tx powers of the nodes.

5b: A new low-complexity Tx strategy is proposed which adjusts the aforementioned weighting parameters with respect to the considered ADR requirements. For this approach, the Tx powers of the nodes are only adjusted if the considered ADR requirements cannot be fulfilled by adjusting the Tx power distribution at the relay station.

6b: A new Tx strategy is proposed which adjusts the aforementioned weighting parameters and additionally optimizes the numbers of simultaneously transmitted data streams. To optimize the numbers of simultaneously transmitted data streams, an approach based on performing an exhaustive search over all possible combinations is proposed.

In Chapter 5, the multi-group multi-way scenario is investigated. In this scenario, novel Tx strategies compared to the cellular multi-user and the multi-pair two-way relaying scenario are required because multiple BC phases are considered. For each of these BC phases, the signals which are retransmitted have to be selected. Furthermore, temporal receive processing can be performed at the nodes. To enable an efficient combination of the temporal processing capabilities of the nodes with the spatial processing capabilities of the relay station, a novel relay transceive filter design is required. The main contributions which are presented in this chapter are as follows:

7: A decomposition of the problem of maximizing the sum rate into different low-complexity subproblems is proposed.

8: A system model for multi-group multi-way relaying considering multi-antenna nodes which can perform spatial Tx and Rx processing, temporal Rx processing, self-interference cancellation and SIC is introduced.

9: A novel low-complexity relay transceive filter design is proposed for each BC phase. The proposed filter design exploits the self-interference cancellation and the SIC capabilities of the nodes and enables an efficient combination of the temporal processing capabilities of the nodes with the spatial processing capabilities of the relay station. For the proposed relay transceive filter design, an analytical solution based on minimizing the weighted MSE is derived. Furthermore, a new approach for jointly optimizing the spatial Rx filters of the nodes and the relay transceive filter is proposed.

10: Two novel Tx strategies are proposed which efficiently combine the temporal processing capabilities of the nodes and the spatial processing capabilities of the relay station. For both Tx strategies, the signals which are retransmitted in each BC phase are selected such that the self-interference cancellation and the SIC capabilities as well as the temporal processing capabilities of the nodes are exploited.

In Chapters 3-5, the sum rate performances of the proposed approaches are compared to the performances of conventional approaches through numerical simulations.

Finally, the main conclusions of this thesis and a brief outlook are presented in Chapter 6.

Chapter 2

Considered Scenarios and Assumptions

2.1 Considered Scenarios

In this section, the three different two-hop relaying scenarios which are considered in this thesis are described. Due to the high dynamic range between the transmitted and the received signal powers, it is assumed that the nodes cannot transmit and receive simultaneously as explained in Section 1.1. To separate the transmit and the receive phases at each node and at the relay station, time division duplex (TDD) is considered, i.e., orthogonal time resources are used for transmission and reception.

First, the cellular multi-user two-way relaying scenario as shown in Figure 2.1 is described. The scenario consists of $K \geq 2$ half-duplex multi-antenna nodes, termed S_1, S_2,..., S_K and of an intermediate half-duplex multi-antenna relay station, termed RS. Node S_1 is assumed to be a base station and nodes S_2, S_3,..., S_K are assumed to be mobile stations. In this scenario, S_1 performs bidirectional communications with each mobile station via RS and one communication cycle consists of one MAC and one BC phase. In the MAC phase, all nodes simultaneously transmit to RS and in the BC phase, RS retransmits a linearly processed version of the superimposed received signals back to the nodes. Afterwards, each node estimates the desired signals after performing self-interference cancellation and linear receive processing with SIC. SIC is considered at the nodes because if they are equipped with multiple antennas, multiple data streams are simultaneously received during the BC phase and thus, the consideration of SIC improves the performance [TV05, ZCW05].

Secondly, the multi-pair two-way relaying scenario as shown in Figure 2.2 is described. The scenario consists of $K \geq 2$ half-duplex multi-antenna nodes, termed S_1, S_2,..., S_K, where K is assumed to be an even number, and of an intermediate half-duplex multi-antenna relay station, termed RS. In this scenario, the nodes S_{1+2i} and S_{2+2i}, $i = 0, 1, ..., K/2 - 1$, perform pairwise bidirectional communications via RS and one communication cycle consists of one MAC and one BC phase. In the MAC phase, all nodes simultaneously transmit to RS and in the BC phase, RS retransmits a linearly processed version of the superimposed received signals back to the nodes. Afterwards, each node estimates the desired signals after performing self-interference cancellation and linear receive processing with SIC.

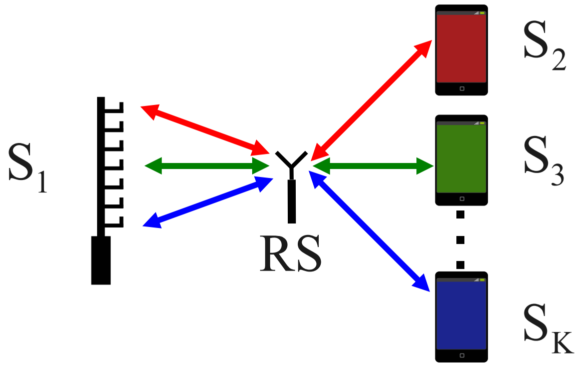

Figure 2.1. Cellular multi-user two-way relaying scenario.

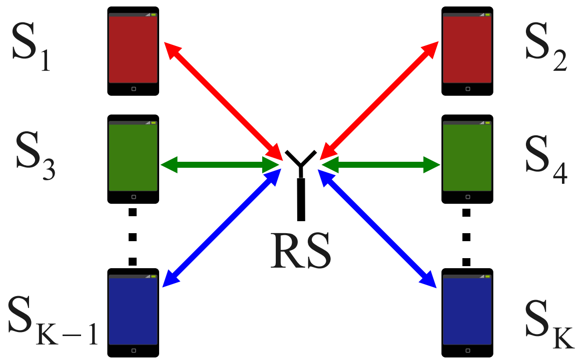

Figure 2.2. Multi-pair two-way relaying scenario.

Thirdly, the multi-group multi-way relaying scenario as shown in Figure 2.3 is described. The scenario consists of $G \geq 1$ groups with $N \geq 2$ nodes per group and of an intermediate half-duplex multi-antenna relay station, termed RS. In this scenario, each node has to transmit a message to all other nodes within its group via RS and thus, each node has to receive $N-1$ independent messages. To receive $N-1$ independent messages at each node, at least $N-1$ BC phases are required. Thus, one communication cycle consists of one MAC and several BC phases. In the MAC phase, all nodes simultaneously transmit to RS and in the BC phases, RS retransmits linearly processed versions of the superimposed received signals back to the nodes. Afterwards, each node estimates the desired signals after performing self-interference cancellation and linear receive processing with SIC. In case of $N = 2$, the multi-pair two-way relaying scenario and the multi-group multi-way relaying scenario are the same.

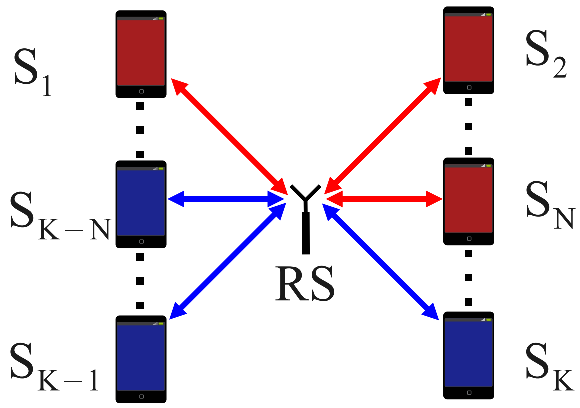

Figure 2.3. Multi-group multi-way relaying scenario.

2.2 Assumptions

In this section, the system assumptions are described which are valid throughout this thesis unless otherwise stated. This thesis is a continuation and an extension of [Ung09, Ama11] and thus, the following assumptions are based on [Ung09, Ama11]. Compared to [Ung09, Ama11], different scenarios are investigated, multiple subcarriers are considered and ADR requirements are introduced.

Throughout this thesis, the equivalent low-pass domain is considered [Pro00]. Signals and radio channels are represented by their complex valued samples in the frequency domain. Each sample is valid for one specific time-frequency unit. Further on, the operators $\mathrm{tr}(\cdot)$, $\otimes$ denote the sum of the main diagonal elements of a matrix and the Kronecker product of matrices, respectively. The operators $\mathrm{diag}[\cdot]$ and $\mathrm{diag}[\cdot]^{-1}$ denote the construction of a block diagonal matrix where the diagonal elements are given by the square matrices within the brackets and the construction of a vector consisting of the diagonal elements of the matrix within the brackets, respectively. The operator $\Re[\cdot]$ denotes the real part of a scalar or a matrix and $\mathrm{E}[\cdot]$ denotes the expectation over the random variables within the brackets. The operators $|\cdot|$, $||\cdot||_2$, $||\cdot||_\mathrm{F}$ denote the norm of a complex number, the Euclidean norm of a complex vector and the Frobenius norm of a complex matrix, respectively. The operators $(\cdot)^\mathrm{T}$, $(\cdot)^*$ and $(\cdot)^\mathrm{H}$ denote the conjugate, the transpose and the conjugate transpose, respectively, of a scalar, vector or matrix. The vectorization operator $\mathrm{vec}(\mathbf{Z})$ stacks the columns of matrix $\mathbf{Z}$ into a vector. The operator $\mathrm{vec}^{-1}_{M,N}(\cdot)$ is the revision of the operator $\mathrm{vec}(\cdot)$, i.e., a vector of length MN is sequentially divided into N smaller vectors of length M which are combined to a matrix with M rows and N columns. Furthermore, $\mathbf{I}_M$ denotes an identity matrix of size M. Moreover, $\mathbf{I}_{1:N,M}$ denotes the first N rows of $\mathbf{I}_M$ and $\mathbf{I}_{M,1:N}$ denotes the first

N columns of $\mathbf{I}_M$. Vectors and matrices are denoted by lower and upper case boldface letters, respectively.

The assumptions which are valid throughout this thesis unless otherwise stated are as follows:

- RS is assumed to be non-regenerative, i.e., linear signal processing is performed at RS. Furthermore, it is assumed that RS is equipped with $L > 1$ antennas.
- The nodes and RS are assumed to be half-duplex and thus, they cannot transmit and receive simultaneously. To separate the transmit and the receive phases at each node and at RS, TDD is considered.
- It is assumed that the received signals at RS are synchronized.
- For the wireless channels between the nodes and RS, the following assumptions are valid. These assumptions have been widely used in two-way and multi-way relaying [Ung09, Ama11, ZLCC09, RH09, XH11, Ron12, WT12].
  - An orthogonal frequency division multiplexing (OFDM) system consisting of $C \geq 1$ perfectly orthogonal subcarriers is assumed [RMBG99, NP00], i.e., inter-carrier interference does not exist in the system. The bandwidth of each subcarrier is assumed to be much smaller than the minimum coherence bandwidth of the different channels between the nodes and between the nodes and RS. Based on this, frequency flat fading subcarrier channels are assumed and thus, the channel transfer function for each transmit and receive antenna pair of each subcarrier can by modeled by a complex fading coefficient in frequency domain.
  - A quasi-static channel model is assumed, i.e., it is assumed that the aforementioned fading coefficients are constant in time during one communication cycle with consists of one MAC and several BC phases dependent on the considered relaying scenario. Between two communication cycles, the fading coefficients can change completely.
  - Channel reciprocity is assumed. This is inherently obtained by the previous assumption of constant fading coefficients during one communication cycle.
  - Additive white Gaussian noises (AWGN) with zero mean and variances σ_{n}^2 and $\sigma_{\mathrm{n,RS}}^2$ are assumed at the receive antennas of the nodes and at the receive antennas of RS, respectively. For simplicity of the notation but without loss of generality (w.l.o.g.), it is assumed that the noise variances are equal at all nodes.

- Perfect global channel state information (CSI) is assumed at all nodes and at RS and this CSI corresponds to the instantaneous fading coefficients for each subcarrier.

- It is assumed that RS has a Tx power constraint per subcarrier. The maximum Tx power of RS on each subcarrier is given by P_{RS}.

- For the cellular multi-user two-way relaying scenario, it is additionally assumed that:

  - Each mobile station S_k is equipped with $M_k = M \geq 1$ antennas, $k = 2, 3, ..., K$.
  - The base station S_1 is equipped with $M_1 = (K-1)M$ antennas.
  - Each mobile station has a Tx power constraint per subcarrier. The maximum Tx power of each mobile station on each subcarrier is given by P_{node}.
  - The base station has a Tx power constraint per subcarrier. The maximum Tx power of the base station on each subcarrier is given by $P_{\mathrm{BS}} = (K-1)P_{\mathrm{node}}$.
  - Each mobile station S_k simultaneously transmits between $m_{k,c} = 0$ and $m_{k,c} = M$ data streams on subcarrier c, $c = 1, 2, ..., C$.
  - The base station simultaneously transmits $m_{1,c} = (K-1)M$ data streams on subcarrier c.
  - The Tx signal vector $\mathbf{s}_{k,c} \in \mathbb{C}^{m_{k,c}\times 1}$ of mobile station S_k on subcarrier c satisfies $\mathrm{E}[\mathbf{s}_{k,c}\mathbf{s}_{k,c}^{\mathrm{H}}] = \mathbf{I}_{m_{k,c}}$.
  - The Tx signal vector $\mathbf{s}_{1,c} \in \mathbb{C}^{m_{1,c}\times 1}$ of S_1 on subcarrier c satisfies $\mathrm{E}[\mathbf{s}_{1,c}\mathbf{s}_{1,c}^{\mathrm{H}}] = \mathbf{I}_{m_{1,c}}$.
  - The transmitted data streams of the mobile stations and the base station are statistically independent.

- For the multi-pair two-way relaying scenario and the multi-group multi-way relaying scenario, it is additionally assumed that:

  - Each node is equipped with $M \geq 1$ antennas.
  - Each node has a Tx power constraint per subcarrier. The maximum Tx power of each node on each subcarrier is given by P_{node}.
  - Each node S_k simultaneously transmits between $m_{k,c} = 0$ and $m_{k,c} = M$ data streams on subcarrier c, $k = 1, 2, ..., K$, $c = 1, 2, ..., C$.

  - The transmitted data streams of the nodes are statistically independent and the Tx signal vector $\mathbf{s}_{k,c} \in \mathbb{C}^{m_{k,c}\times 1}$ of node S_k on subcarrier c satisfies $\mathrm{E}[\mathbf{s}_{k,c}\mathbf{s}_{k,c}^{\mathrm{H}}] = \mathbf{I}_{m_{k,c}}$.

- If the terms maximum achievable data rate or achievable sum rate are used throughout this thesis, this does not mean the information-theoretic capacity. In this thesis, maximum achievable data rate or achievable sum rate means the maximum data rate or sum rate, respectively, that can be achieved considering Gaussian codebooks, the proposed filter designs and a given decoding order.

Chapter 3

Cellular Multi-User Two-Way Relaying

3.1 Problem Overview and Decomposition

In this chapter, the cellular multi-user two-way relaying scenario as shown in Figure 2.1 is investigated. To investigate this scenario, a system model for cellular multi-user two-way relaying considering multi-antenna nodes which can perform self-interference cancellation, linear receive processing and SIC is introduced. In such a multi-user single-cell scenario, the required data rates in downlink are typically higher than the required data rates in uplink which is considered by introducing ADR requirements.

To maximize the sum rate under the aforementioned ADR requirements, the Tx and Rx filters of the nodes, the Tx powers of the nodes, the relay transceive filter and the numbers of simultaneously transmitted data streams of the nodes have to be jointly optimized over all subcarriers. Due to the high computational complexity of finding an optimal solution for this problem, suboptimal approaches based on a problem decomposition are proposed in this chapter. To obtain such suboptimal approaches which fulfill the aforementioned ADR requirements whilst achieving high sum rates, the following steps are proposed:

1. It is proposed to decouple the overall problem into three different subproblems as shown in Figure 3.1. The considered subproblems are the design of a Tx strategy, the design of the relay transceive filter and the design of the Tx and Rx filters at the nodes.

2. To tackle the ADR requirements, it is proposed to couple the filter design at the nodes and at RS with the design of the Tx strategy by introducing the following weighting parameters:

    - $v_{\mathrm{BS},k}$: To adjust the fraction of the Tx power used at S_1 and at RS to perform transmissions from S_1 to S_k, $0 \leq v_{\mathrm{BS},k} \leq 1$, $k = 2, 3, ..., K$,
    - $v_{\mathrm{MS},k}$: To adjust the fraction of the Tx power used at RS to perform transmissions from S_k to S_1, $0 \leq v_{\mathrm{MS},k} \leq 1$,
    - p_k: To adjust the Tx power of S_k, $0 \leq p_k \leq 1$.

These weighting parameters are considered for the Tx filter design at the nodes and for the relay transceive filter design. By these weighting parameters, the Tx powers of the mobile stations and the Tx power distributions at S_1 and at RS are adjusted via the Tx strategy.

3. It is proposed to focus on low-complexity solutions for the different subproblems.

Based on these steps, suboptimal low-complexity approaches for the different subproblems are proposed as shown in Figure 3.1.

For the design of a Tx strategy which fulfills the ADR requirements whilst achieving high sum rates, two different approaches are proposed. The power adapted Tx strategy considers that each mobile station transmits M data streams and the base station S_1 transmits M_1 data streams on each subcarrier. Based on this, the Tx powers of the nodes and the Tx power distributions at S_1 and at RS are adjusted via the aforementioned weighting parameters. Thus, an optimization of the numbers of simultaneously transmitted data streams of the nodes is not considered for this Tx strategy. The subcarrier allocation Tx strategy is an extension of the power adapted Tx strategy. A suboptimal low-complexity optimization of the numbers of simultaneously transmitted data streams of the nodes is performed by considering a subcarrier allocation approach. The proposed subcarrier allocation approach aims for increasing the sum rates under the considered ADR requirements by reducing the number of subcarriers on which each mobile station transmits. By reducing the number of subcarriers on which each mobile station transmits, less signals are simultaneously received at RS during the MAC phase and thus, the spatial separation of the different signals is simplified. Additionally, the Tx powers of the nodes and the Tx power distributions at S_1 and at RS are adjusted similar to the power adapted Tx strategy via the aforementioned weighting parameters.

For the relay transceive filter design, three different approaches are investigated. For comparison, a weighted ZF and a weighted MMSE approach are considered. The weighted ZF and the weighted MMSE approach are straightforward extensions of the state of the art to tackle the ADR requirements by considering the aforementioned weighting parameters in the relay transceive filter design. In addition to these approaches, a weighted self-interference cancellation and SIC aware relay transceive filter is proposed. To obtain an analytical solution for the relay transceive filter design which ensures high sum rates whilst considering the ADR requirements, a weighted MMSE based approach is proposed. The proposed relay transceive filter design exploits the capability of the nodes to perform self-interference cancellation and SIC. However, the proposed relay transceive filter depends on the Tx and Rx filters at the nodes and vice versa. Thus, to overcome this problem, a joint design approach is proposed based on performing an alternating optimization between the Tx and Rx filters at the nodes and the relay transceive filter.

For the Tx and Rx filter design at the nodes, three different approaches are investigated. In case Diag, diagonal Tx and Rx filters are considered at the nodes. This case is investigated because it is a straightforward extension of the state of the art to consider ADR requirements and multi-antenna mobile stations and thus, it can be used for comparison. In case Rx, an extension of case Diag is investigated by considering MMSE Rx filters instead of diagonal Rx filters at the nodes. In case Rx&Tx, a weighted SIC aware Tx filter design at S_1 is proposed combined with diagonal Tx filters at the mobile stations and MMSE Rx filters at all nodes. The proposed SIC aware Tx filter at S_1 is designed such that the capability of the mobile stations to perform SIC is exploited.

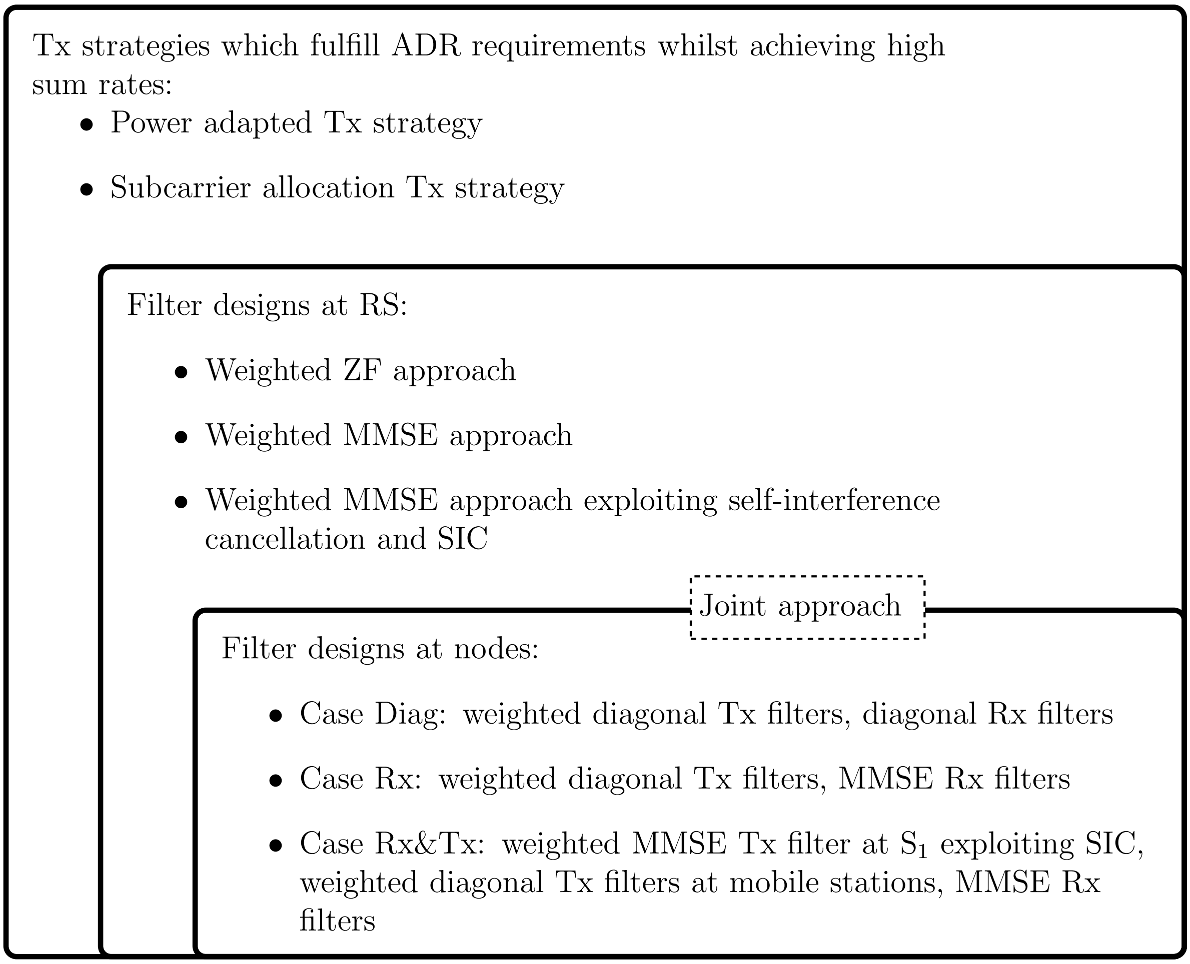

Figure 3.1. Overview of the proposed and investigated approaches for cellular multi-user two-way relaying.

The proposed relay transceive filter design depends on the Tx and Rx filters at the nodes and vice versa. Furthermore, the proposed Tx strategies are based on the relay transceive filter design and on the filter design at the nodes. Moreover, the relay transceive filter and the filters at the nodes depend on the weighting parameters and the subcarrier allocation which are computed based on the Tx strategies. Thus, the computation of the different filters, the weighting parameters and the subcarrier allocation is performed as shown in Figure 3.2. First, all weighting parameters are assumed to be one, i.e., no weighting is considered. Furthermore, all mobile stations are assumed to transmit M data streams per subcarrier and S_1 is assumed to transmit M_1 data streams per subcarrier. Based on these assumptions, the Tx and Rx filters at the nodes are computed according to case Diag because only in this case, the Tx and Rx filter design at the nodes is independent of the relay transceive filter design. Secondly, the relay transceive filter is computed considering the Tx and Rx filters at the nodes of the previous step. Thirdly, if case Rx or case Rx&Tx is considered, the Tx and Rx filters at the nodes are updated considering the relay transceive filter of the previous step. Additionally, an alternating optimization between the relay transceive filter and the Tx and Rx filters at the nodes can be performed. Fourthly, the weighting parameters which are considered for the Tx filter design at the nodes and for the relay transceive filter design are adjusted. Furthermore, if the subcarrier allocation Tx strategy is considered, a subcarrier allocation is performed. To adjust the weighting parameters and to perform a subcarrier allocation, the relay transceive filter and the Tx and Rx filters at the nodes have to be updated after each step. Finally, weighting parameters and a subcarrier allocation which fulfill the ADR requirements whilst achieving high sum rates are selected.

The rest of the chapter is organized as follows. In Section 3.2, the system model for the considered cellular multi-user two-way relaying scenario is presented. In Section 3.3, the different cases for the Tx and Rx filter design at the nodes are presented and the different relay transceive filters are proposed. In Section 3.4, the Tx strategies are proposed and in Section 3.5, the performance of the proposed approaches is investigated by numerical results. Several parts of this chapter have been originally published by the author in [DUK11, DK12b]. Compared to [DUK11, DK12b], the system model and the filter designs are extended to consider and to exploit SIC at the nodes, respectively. Furthermore, the Tx strategies are presented in more detail.

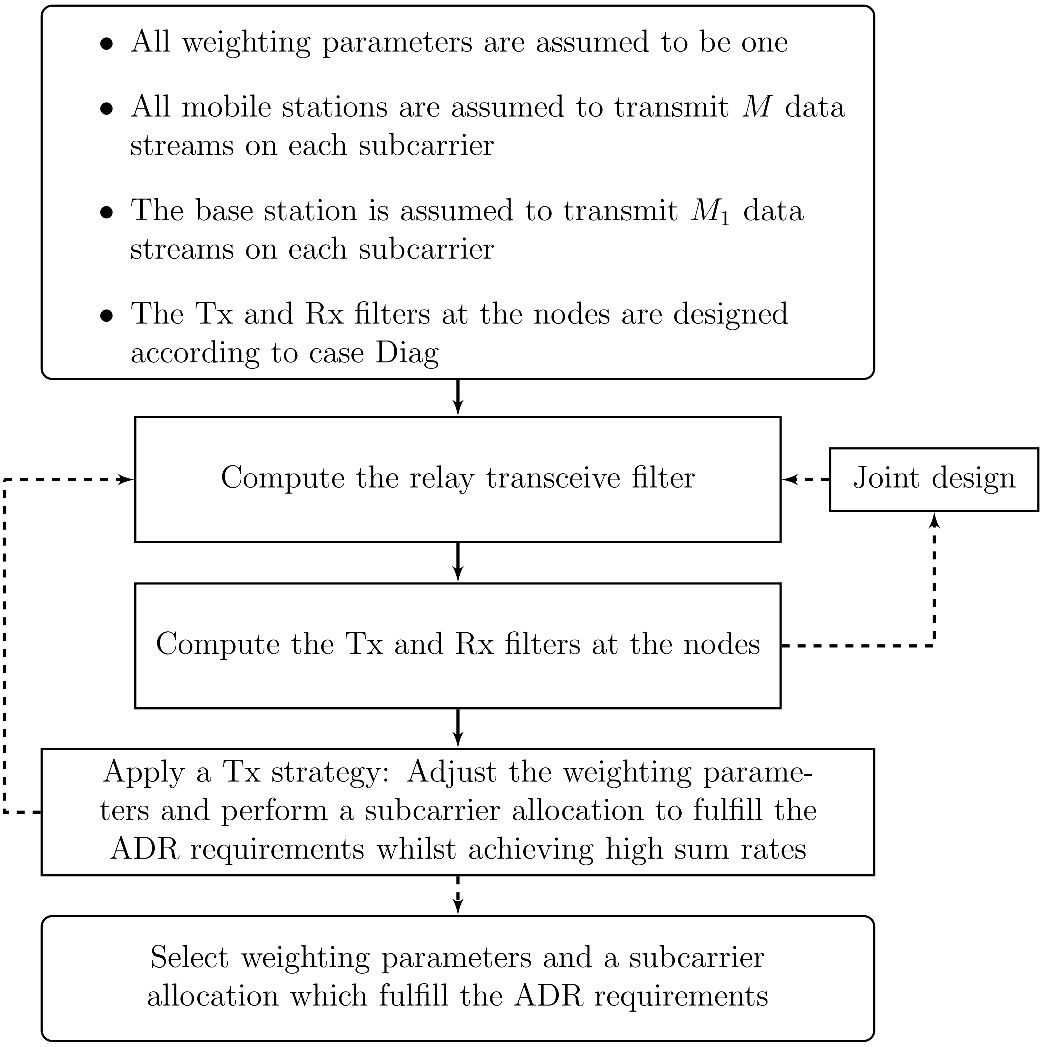

Figure 3.2. Flow chart for the computation of the filters at the nodes and at RS considering the proposed Tx strategies.

3.2 System Model

In this section, the system model for the considered cellular multi-user two-way relaying scenario as shown in Figure 3.3 is presented. As described in Section 2.1, the scenario consists of a half-duplex multi-antenna base station S_1 and $K-1$ half-duplex multi-antenna mobile stations S_k, $k = 2, 3, ..., K$. The bidirectional communications between S_1 and the mobile stations are performed via RS.

In the MAC phase, all nodes simultaneously transmit to RS and the superposition of these transmit signals is received at RS. Before the transmission, the Tx signal vector $\mathbf{s}_{k,c} \in \mathbb{C}^{m_{k,c}\times 1}$ of node S_k on subcarrier c is filtered by the Tx filter $\mathbf{Q}_{k,c} \in \mathbb{C}^{M_k\times m_{k,c}}$, with $||\mathbf{Q}_{k,c}||_{\mathrm{F}}^2 \leq P_{\mathrm{Node}}$ for $k \neq 1$ and $||\mathbf{Q}_{1,c}||_{\mathrm{F}}^2 \leq (K-1)P_{\mathrm{Node}}$, $k = 1, 2, ..., K$, $c = 1, 2, ..., C$. The transmitted symbols of S_1 on subcarrier c which are intended for S_k are described by the vector $\mathbf{s}_{1,k,c} \in \mathbb{C}^{M\times 1}$ and the corresponding columns of the Tx filter at S_1 are given by $\mathbf{Q}_{1,k,c} \in \mathbb{C}^{M_1\times M}$, i.e., $\mathbf{s}_{1,c} = \left[\mathbf{s}_{1,2,c}^{\mathrm{T}}, \mathbf{s}_{1,3,c}^{\mathrm{T}}, ..., \mathbf{s}_{1,K,c}^{\mathrm{T}}\right]^{\mathrm{T}}$ and $\mathbf{Q}_{1,c} = [\mathbf{Q}_{1,2,c}, \mathbf{Q}_{1,3,c}, ..., \mathbf{Q}_{1,K,c}] \in \mathbb{C}^{M_1\times M_1}$. Furthermore, let $\mathbf{n}_{\mathrm{RS},c} \in \mathbb{C}^{L\times 1}$ represent the complex white Gaussian noise vector at RS on subcarrier c and let $\mathbf{H}_{k,c} \in \mathbb{C}^{L\times M_k}$ denote the channel from node S_k to RS on subcarrier c. Now, the received signal at RS on subcarrier c can be written as

$$\mathbf{y}_{\mathrm{RS},c} = \sum_{k=1}^{K} \mathbf{H}_{k,c}\mathbf{Q}_{k,c}\mathbf{s}_{k,c} + \mathbf{n}_{\mathrm{RS},c}. \tag{3.1}$$

In the BC phase, RS retransmits a linearly processed version of the superimposed received signals back to the nodes. The received signal $\mathbf{y}_{\mathrm{RS},c}$ is linearly processed at RS using the transceive filter matrix $\mathbf{G}_c \in \mathbb{C}^{L\times L}$. Using the receive filters $\mathbf{D}_{k,c} \in \mathbb{C}^{M_k\times M_k}$ for $k \neq 1$ and $\mathbf{D}_{1,c} \in \mathbb{C}^{\sum_{i=2}^{K}(m_{i,c})\times M_1}$, the received signal at node S_k on subcarrier c is given by

$$\mathbf{y}_{k,c} = \mathbf{D}_{k,c}(\mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{y}_{\mathrm{RS},c} + \mathbf{n}_{k,c}), \tag{3.2}$$

where $\mathbf{n}_{k,c} \in \mathbb{C}^{M_k\times 1}$ represents the complex white Gaussian noise vector at S_k on subcarrier c. The Rx filter at S_1 to receive the transmitted symbols of S_k on subcarrier c is given by $\mathbf{D}_{1,k,c}$, i.e., $\mathbf{D}_{1,c} = \left[\mathbf{D}_{1,2,c}^{\mathrm{T}}, \mathbf{D}_{1,3,c}^{\mathrm{T}}, ..., \mathbf{D}_{1,K,c}^{\mathrm{T}}\right]^{\mathrm{T}}$.

The received useful signals and interferences at the nodes in the BC phase are illustrated in Figure 3.4. S_1 receives the useful signals from the mobile stations, back-propagated self-interference and noise. Each mobile station receives its intended useful signals, interference from the signals intended for the other mobile station, termed BS-MS-interference, interference from the signals transmitted by the other mobile stations

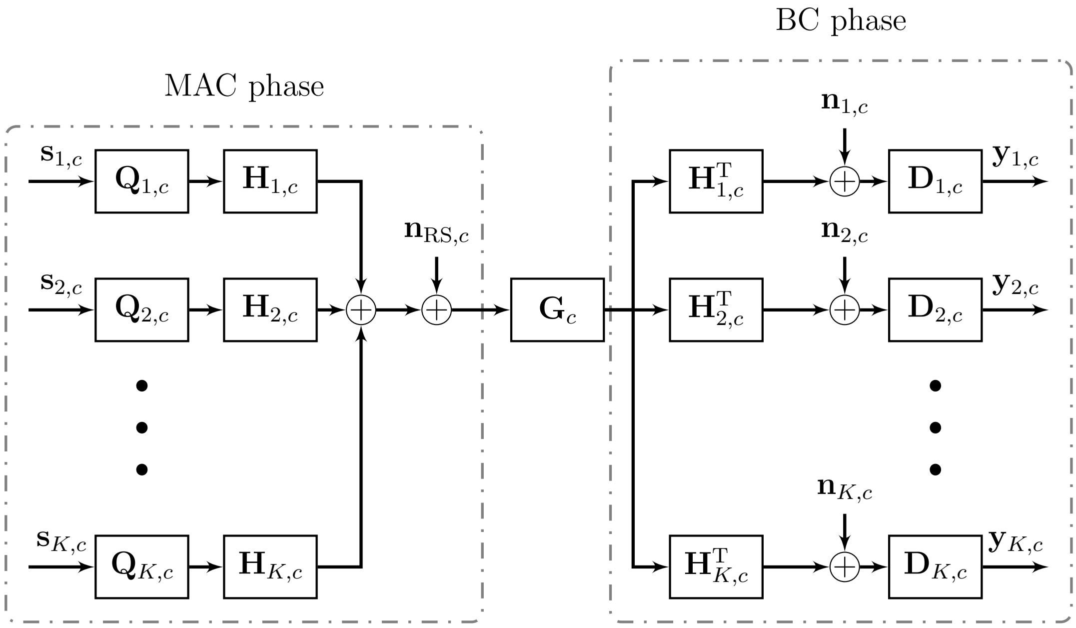

Figure 3.3. System model for cellular multi-user two-way relaying.

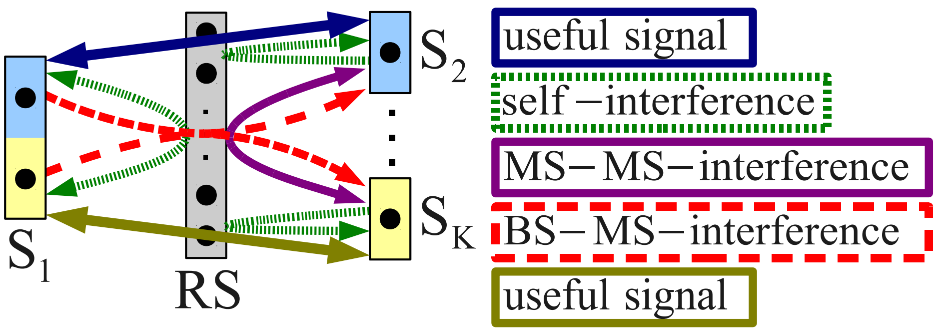

Figure 3.4. Compositions of the received signals out of useful signals and interferences in a cellular multi-user two-way relaying scenario.

which are retransmitted by RS, termed MS-MS-interference, and back-propagated self-interference as well as noise.

To simplify the notation, the overall channel coefficients for the transmission of the m_1^{th} data stream from S_1 to S_k, $m_1 = 1, 2, ..., M$, and for the transmission of the m_k^{th} data stream from S_k to S_1,$m_k = 1, 2, ..., m_{k,c}$, on subcarrier c can be written as

$$h_{\mathrm{ov},1,k,m_1,c} = \mathbf{d}_{k,m_1,c}\mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{H}_{1,c}\mathbf{q}_{1,k,m_1,c}, \tag{3.3a}$$

$$h_{\mathrm{ov},k,1,m_k,c} = \mathbf{d}_{1,k,m_k,c}\mathbf{H}_{1,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{H}_{k,c}\mathbf{q}_{k,m_k,c}, \tag{3.3b}$$

respectively, where $\mathbf{q}_{1,k,m_1,c}$ and $\mathbf{q}_{k,m_k,c}$ are the m_1^{th} and the m_k^{th} column vectors of $\mathbf{Q}_{1,k,c}$ and $\mathbf{Q}_{k,c}$, respectively, and where $\mathbf{d}_{1,k,m_k,c}$ and $\mathbf{d}_{k,m_1,c}$ are the m_k^{th} and the the m_1^{th} row vectors of $\mathbf{D}_{1,k,c}$ and $\mathbf{D}_{k,c}$, respectively.

As mentioned before, perfect self-interference cancellation and perfect SIC are assumed at all nodes. To exploit the SIC capabilities of the nodes for the Tx filter design at S_1 and for the relay transceive filter design, a fixed decoding order is required. This decoding order has to be independent of the considered filters at the nodes and at RS. Thus, it is proposed that the signals of the mobile stations are decoded at S_1 in increasing order of the corresponding indices of the mobile stations, i.e., the signal $\mathbf{s}_{2,c}$ is decoded first and the signal $\mathbf{s}_{K,c}$ is decoded last on subcarrier c. When decoding the signal $\mathbf{s}_{j,c}$, $j = 2, 3, ..., K$, at S_1 on subcarrier c, the transmit signals $\mathbf{s}_{i,c}$, $i = 2, ..., j-1$, are assumed to be perfectly canceled in advance. Furthermore, it is assumed that the data streams transmitted by S_l are decoded at S_k in increasing order of the corresponding indices, i.e., the data stream $m = 1$ is decoded first and the data stream $m = m_{l,c}$ for $l \neq 1$ and $m = M$ for $l = 1$ is decoded last.

Considering the aforementioned decoding order and writing the estimate of $\mathbf{s}_{1,c}$ at the mobile stations as

$$\hat{\mathbf{s}}_{1,c} = [\hat{\mathbf{s}}_{1,2,c}^{\mathrm{T}}, \hat{\mathbf{s}}_{1,3,c}^{\mathrm{T}}, ..., \hat{\mathbf{s}}_{1,K,c}^{\mathrm{T}}]^{\mathrm{T}}, \tag{3.4}$$

the signal at mobile station S_k for estimating the m^{th} data stream from S_1 on subcarrier c with $m = 1, 2, ..., M$ after self-interference cancellation and SIC can be written as

$$\begin{aligned}\hat{s}_{1,k,m,c} = {} & \mathbf{d}_{k,m,c}\mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{H}_{1,c}\left(\sum_{i=m}^{M}\mathbf{q}_{1,k,i,c}s_{1,k,i,c} + \sum_{j=2,j\neq k}^{K}\mathbf{Q}_{1,j,c}\mathbf{s}_{1,j,c}\right) \\ & + \mathbf{d}_{k,m,c}\mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c\left(\sum_{j=2,j\neq k}^{K}\mathbf{H}_{j,c}\mathbf{Q}_{j,c}\mathbf{s}_{j,c} + \mathbf{n}_{\mathrm{RS},c}\right) + \mathbf{d}_{k,m,c}\mathbf{n}_{k,c},\end{aligned} \tag{3.5}$$

where $s_{1,k,m,c}$ is the m^{th} element of $\mathbf{s}_{1,k,c}$ and $\hat{s}_{1,k,m,c}$ is the m^{th} element of $\hat{\mathbf{s}}_{1,k,c}$. Due to considering perfect self-interference cancellation and SIC, the terms

$\mathbf{d}_{k,m,c}\mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{H}_{k,c}\mathbf{Q}_{k,c}\mathbf{s}_{k,c}$ and $\sum_{i=1}^{m-1}\mathbf{d}_{k,m,c}\mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{H}_{1,c}\mathbf{q}_{1,k,i,c}s_{1,k,i,c}$ are neglected in (3.5), respectively.

Furthermore, writing the estimate of $\mathbf{s}_{k,c}$ as $\hat{\mathbf{s}}_{k,c}$, the signal at S_1 for estimating the m^{th} data stream from mobile station S_k on subcarrier c with $m = 1, 2, ..., m_{k,c}$ after self-interference cancellation and SIC can be written as

$$\hat{s}_{k,m,c} = \mathbf{d}_{1,k,m,c}\mathbf{H}_{1,c}^{\mathrm{T}}\mathbf{G}_c\left(\mathbf{H}_{k,c}\sum_{i=m}^{m_{k,c}}\mathbf{q}_{k,i,c}s_{k,i,c} + \sum_{j=k+1}^{K}\mathbf{H}_{j,c}\mathbf{Q}_{j,c}\mathbf{s}_{j,c} + \mathbf{n}_{\mathrm{RS},c}\right) + \mathbf{d}_{1,k,m,c}\mathbf{n}_{k,c}, \tag{3.6}$$

where $s_{k,m,c}$ is the m^{th} element of $\mathbf{s}_{k,c}$ and $\hat{s}_{k,m,c}$ is the m^{th} element of $\hat{\mathbf{s}}_{k,c}$.

Considering perfect self-interference cancellation and SIC, the expected signal, interference and noise powers when estimating the m^{th} data stream of S_1 at S_k on subcarrier c can be written as

$$P_{\mathrm{S},1,k,m,c} = |h_{\mathrm{ov},1,k,m,c}|^2, \tag{3.7a}$$
$$P_{\mathrm{I},1,k,m,c} = \sum_{j=1,j\neq k}^{K}||\mathbf{d}_{k,m,c}\mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{H}_{j,c}\mathbf{Q}_{j,c}||_2^2 - \sum_{i=1}^{m}|\mathbf{d}_{k,m,c}\mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{H}_{1,c}\mathbf{q}_{1,k,i,c}|^2, \tag{3.7b}$$
$$P_{\mathrm{N},1,k,m,c} = \sigma_{\mathrm{n,RS}}^2||\mathbf{d}_{k,m,c}\mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c||_2^2 + \sigma_{\mathrm{n}}^2||\mathbf{d}_{k,m,c}||_2^2, \tag{3.7c}$$

respectively, where $k = 2, 3, ..., K$. Moreover, the expected signal, interference and noise powers when estimating the m^{th} data stream of S_k at S_1 on subcarrier c can be written as

$$P_{\mathrm{S},k,1,m,c} = |h_{\mathrm{ov},k,1,m,c}|^2, \tag{3.8a}$$
$$P_{\mathrm{I},k,1,m,c} = \sum_{j=k}^{K}||\mathbf{d}_{1,k,m,c}\mathbf{H}_{1,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{H}_{j,c}\mathbf{Q}_{j,c}||_2^2 - \sum_{i=1}^{m}|\mathbf{d}_{1,k,m,c}\mathbf{H}_{1,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{H}_{k,c}\mathbf{q}_{k,i,c}|^2, \tag{3.8b}$$
$$P_{\mathrm{N},k,1,m,c} = \sigma_{\mathrm{n,RS}}^2||\mathbf{d}_{1,k,m,c}\mathbf{H}_{1,c}^{\mathrm{T}}\mathbf{G}_c||_2^2 + \sigma_{\mathrm{n}}^2||\mathbf{d}_{1,k,m,c}||_2^2, \tag{3.8c}$$

respectively.

For performance comparison in Section 3.5, the achievable sum rate, cf. [Ung09, Ama11, DUK11, DK12b], is considered and the corresponding equations are presented in the following. Assuming that Gaussian codebooks are used for each data stream, the maximum achievable data rate after linear receive processing, self-interference cancellation and SIC for the m^{th} data stream from S_l to S_j on subcarrier c is given by

$$C_{l,j,m,c} = \frac{1}{2}\log_2(1 + P_{\mathrm{S},l,j,m,c}(P_{\mathrm{I},l,j,m,c} + P_{\mathrm{N},l,j,m,c})^{-1}), \tag{3.9}$$

where the factor $\frac{1}{2}$ is due to the fact that two time slots are required to perform all transmissions and where $l = 1$ if $j \neq 1$ and $l \neq 1$ if $j = 1$, $j, l = 1, 2, ..., K$. Using this notation, the maximum achievable data rates for the transmissions from S_1 to S_k and from S_k to S_1 are given by

$$C_{1,k} = \sum_{c=1}^{C} \sum_{m=1}^{M} C_{1,k,m,c}, \tag{3.10a}$$

$$C_{k,1} = \sum_{c=1}^{C} \sum_{m=1}^{m_{k,c}} C_{k,1,m,c}, \tag{3.10b}$$

respectively. Thus, the achievable sum rate of the system is given by

$$C_{\mathrm{sum}} = \sum_{k=2}^{K} \left(C_{1,k} + C_{k,1}\right). \tag{3.11}$$

As mentioned in Section 3.1, ADR requirements shall be considered. For simplicity of the notation, it is assumed that the instantaneous data rates in downlink from S_1 to each mobile station are required to be equal. Furthermore, it is assumed that the instantaneous data rates in downlink are required to be r times higher than the instantaneous data rates in uplink from each mobile station to S_1, $r \geq 1$. Thus, the constraints

$$r = C_{1,k}/C_{k,1}, k = 2, 3, ..., K, \tag{3.12a}$$

$$C_{1,k} = C_{1,l}, l, k = 2, 3, ..., K, \tag{3.12b}$$

have to be fulfilled for the sum rate under ADR requirements. The extension to consider more general ADR requirements is straightforward for the presented approaches. To fulfill the constraints (3.12a) and (3.12b), either the lowest maximum achievable data rate from S_1 to any mobile station or the lowest maximum achievable data from any mobile station to S_1 is the limiting data rate. Considering the constraint (3.12a), the maximum achievable data rate between S_1 and S_k, $k \neq 1$, can be written as

$$C_{\mathrm{ADR},1,k} = \min\left(C_{1,k}, rC_{k,1}\right), \tag{3.13}$$

and the maximum achievable data rate between S_k and S_1 can be written as

$$C_{\mathrm{ADR},k,1} = \frac{1}{r} C_{\mathrm{ADR},1,k}. \tag{3.14}$$

Considering both constraints (3.12a) and (3.12b), the achievable sum rate under the considered ADR requirements is given by

$$C_{\mathrm{ADR,sum}} = (K-1)(1+1/r) \cdot \min_k C_{\mathrm{ADR},1,k}. \tag{3.15}$$

3.3 Filter Design

3.3.1 Introduction

In this section, three different low-complexity approaches for designing the Tx and Rx filters at the nodes are proposed as described in Section 3.1. Furthermore, three low-complexity approaches for designing the relay transceive filter are proposed as described in Section 3.1. As mentioned in Section 3.1, the filter design at the nodes depends on the design of the relay transceive filter and vice versa. Thus, to overcome this problem, an alternating optimization is proposed as illustrated in Figure 3.2. Furthermore, as mentioned in Section 3.1, weighting parameters are considered for the Tx filter design at the nodes and for the relay transceive filter design to enable an adjustment of these filters via the proposed Tx strategies which are introduced in Section 3.4. Utilizing these weighting parameters, the proposed Tx strategies perform an adjustment of the Tx powers of the mobile stations and of the Tx power distributions at S_1 and at RS to fulfill the ADR requirements.

As described in Section 3.1, the following weighting parameters are considered for the Tx filter design at the nodes which is presented in Section 3.3.2 and for the relay transceive filter design which is presented in Section 3.3.3:

- To adjust the fraction of the Tx power used at S_1 and at RS to perform transmissions from S_1 to S_k, a weighting parameter $\upsilon_{\mathrm{BS},k}$ is considered, $0 \leq \upsilon_{\mathrm{BS},k} \leq 1$, $k = 2, 3, ..., K$. Furthermore, a weighting matrix $\mathbf{V}_{\mathrm{BS}}$ and a weighting vector $\mathbf{v}_{\mathrm{BS}}$ are defined which contain these weighting parameters:

$$\mathbf{V}_{\mathrm{BS}} = \mathrm{diag}\left([\upsilon_{\mathrm{BS},2}, \upsilon_{\mathrm{BS},3}, ..., \upsilon_{\mathrm{BS},K}]\right) \otimes \mathbf{I}_M, \tag{3.16a}$$

$$\mathbf{v}_{\mathrm{BS}} = \mathrm{diag}^{-1}\left(\mathbf{V}_{\mathrm{BS}}\right). \tag{3.16b}$$

- To adjust the fraction of the Tx power used at RS to perform transmissions from S_k to S_1, a weighting parameter $\upsilon_{\mathrm{MS},k}$ is considered, $0 \leq \upsilon_{\mathrm{MS},k} \leq 1$, $k = 2, 3, ..., K$. Furthermore, a weighting vector $\mathbf{v}_{\mathrm{MS},c}$ is defined for each subcarrier c which contains these weighting parameters:

$$\mathbf{v}_{\mathrm{MS},c} = \left[\mathrm{diag}^{-1}\left(\upsilon_{\mathrm{MS},2} \otimes \mathbf{I}_{m_{2,c}}\right), ..., \mathrm{diag}^{-1}\left(\upsilon_{\mathrm{MS},K} \otimes \mathbf{I}_{m_{K,c}}\right)\right]^{\mathrm{T}}, \tag{3.17}$$

- To adjust the Tx power of S_k, a weighting parameter p_k is considered, $0 \leq p_k \leq 1$, $k = 2, 3, ..., K$.

For the filter design in Section 3.3.2 and Section 3.3.3, the parameters $v_{\mathrm{BS},k}$, $v_{\mathrm{MS},k}$ and p_k are assumed to be given. The optimization of the different parameters is described in Section 3.4.

3.3.2 Transmit and Receive Filter Design at Nodes

3.3.2.1 Case Diag

In case Diag, the Tx and Rx filters at the nodes are diagonal matrices. To consider the weighting parameters introduced in Section 3.3.1 and to fulfill the Tx power constraints introduced in Section 2.2, the Tx filters are designed as

$$\mathbf{Q}_{1,c} = \sqrt{\frac{(K-1)P_{\mathrm{node}}}{||\mathbf{V}_{\mathrm{BS}}||_{\mathrm{F}}^2}}\mathbf{V}_{\mathrm{BS}}, \tag{3.18a}$$

$$\mathbf{Q}_{k,c} = p_{k,c}\sqrt{\frac{P_{\mathrm{node}}}{m_{k,c}}}\mathbf{I}_{M,1:m_{k,c}} \text{ for } k \neq 1, \tag{3.18b}$$

where $k = 2, 3, ...K$. The Rx filters are designed as unweighted diagonal matrices considering the number of desired data streams. Thus, the Rx filters are given by

$$\mathbf{D}_{1,k,c} = \mathbf{I}_{1:m_{k,c},M_1}, \tag{3.19a}$$

$$\mathbf{D}_{k,c} = \mathbf{I}_M. \tag{3.19b}$$

3.3.2.2 Case Rx

In case Rx, the Tx filters at the nodes are given by (3.18) and the Rx filter at node S_k, $k = 2, 3, ..., K$, is designed to minimize the MSE for the transmission from S_1 to S_k and the Rx filter at the node S_1 is designed to minimize the MSE for the transmissions from all mobile stations to S_1. For the Rx filter design, perfect self-interference cancellation and perfect SIC are assumed. Thus, the Rx filters at the nodes are designed as MMSE-SIC receivers which are described for conventional multi-user uplink systems in [TV05]. Due to the MMSE based design, the Rx filters depend on the overall channels and therewith, they depend on the relay transceive filters. In this section, the Rx filters at the nodes are designed assuming that the relay transceive filters are known in advance. An alternating optimization between the Rx filters at the nodes and the relay transceive filter on each subcarrier is presented in Section 3.3.4.

Using (3.4) and (3.5), the MSE for the transmission from S_1 to S_k on subcarrier c after self-interference cancellation and SIC can be expressed as

$$\mathrm{MSE}_{1,k,c} = \mathrm{E}\left\{\|\mathbf{s}_{1,k,c} - \hat{\mathbf{s}}_{1,k,c}\|_2^2\right\}. \tag{3.20}$$

Using (3.6), the MSE for the transmission from all mobile stations to S_1 on subcarrier c after self-interference cancellation and SIC can be expressed as

$$\mathrm{MSE}_{k,c} = \mathrm{E}\left\{\sum_{k=2}^{K} \|\mathbf{s}_{k,c} - \hat{\mathbf{s}}_{k,c}\|_2^2\right\}. \tag{3.21}$$

However, to compute the Rx filters at the mobile stations, it is proposed to neglect MS-MS-interferences and BS-MS-interferences because these interferences shall only be tackled by the relay transceive filter design and by the Tx filter design at S_1. Thus, to compute the Rx filters based on an MMSE design considering SIC but neglecting MS-MS-interferences and BS-MS-interferences, two channel matrices are introduced as follows:

$$\mathbf{H}_{\mathrm{MS},k,m,c} = \mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{H}_{1,c}\left[\mathbf{q}_{1,k,m,c}, \mathbf{q}_{1,k,m+1,c}, ..., \mathbf{q}_{1,k,M,c}\right], \tag{3.22a}$$

$$\mathbf{H}_{\mathrm{BS},k,m,c} = \mathbf{H}_{1,c}^{\mathrm{T}}\mathbf{G}_c\left[\mathbf{H}_{k,c}\left[\mathbf{q}_{k,m,c}, \mathbf{q}_{k,m+1,c}, ..., \mathbf{q}_{k,m_{k,c},c}\right], \mathbf{H}_{k+1,c}\mathbf{Q}_{k+1,c}, ..., \mathbf{H}_{K,c}\mathbf{Q}_{K,c}\right]. \tag{3.22b}$$

Using (3.22a) and considering (3.20), the rows of the SIC aware MMSE Rx filter at mobile station S_k neglecting MS-MS-interferences and BS-MS-interferences on subcarrier c based on the derivations for conventional MIMO Rx filters [Joh04] are given by

$$\mathbf{d}_{k,m,c} = \frac{\left(\mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{H}_{1,c}\mathbf{q}_{1,k,m,c}\right)^{\mathrm{H}}\left(\mathbf{H}_{\mathrm{MS},k,m,c}\mathbf{H}_{\mathrm{MS},k,m,c}^{\mathrm{H}} + \mathbf{N}_{k,c}\right)^{-1}}{\|\left(\mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{H}_{1,c}\mathbf{q}_{1,k,m,c}\right)^{\mathrm{H}}\left(\mathbf{H}_{\mathrm{MS},k,m,c}\mathbf{H}_{\mathrm{MS},k,m,c}^{\mathrm{H}} + \mathbf{N}_{k,c}\right)^{-1}\|_2^2}, \tag{3.23}$$

with $\mathbf{N}_{k,c} = \sigma_{\mathrm{n}}^2\mathbf{I}_M + \sigma_{\mathrm{n,RS}}^2\mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{G}_c^{\mathrm{H}}\mathbf{H}_{k,c}^{*}$ and where interferences of all signals which are decoded in previous decoding steps are assumed to be perfectly canceled by applying SIC.

Furthermore, using (3.22b) and considering (3.21), the rows of the SIC aware MMSE Rx filter at S_1 on subcarrier c based on the derivations for conventional MIMO Rx filters [Joh04] are given by

$$\mathbf{d}_{1,k,m,c} = \frac{\left(\mathbf{H}_{1,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{H}_{k,c}\mathbf{q}_{k,m,c}\right)^{\mathrm{H}}\left(\mathbf{H}_{\mathrm{BS},k,m,c}\mathbf{H}_{\mathrm{BS},k,m,c}^{\mathrm{H}} + \mathbf{N}_{1,c}\right)^{-1}}{\|\left(\mathbf{H}_{1,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{H}_{k,c}\mathbf{q}_{k,m,c}\right)^{\mathrm{H}}\left(\mathbf{H}_{\mathrm{BS},k,m,c}\mathbf{H}_{\mathrm{BS},k,m,c}^{\mathrm{H}} + \mathbf{N}_{1,c}\right)^{-1}\|_2^2}, \tag{3.24}$$

with $\mathbf{N}_{1,c} = \sigma_{\mathrm{n}}^2\mathbf{I}_{M_1} + \sigma_{\mathrm{n,RS}}^2\mathbf{H}_{1,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{G}_c^{\mathrm{H}}\mathbf{H}_{1,c}^{*}$.

3.3.2.3 Case Rx&Tx

In case Rx&Tx, the Tx and Rx filters at the mobile stations are given by (3.18b) and (3.23), respectively, the Rx filters at S_1 are given by (3.24) and the Tx filters at S_1 are designed to minimize the weighted MMSE on each subcarrier for the transmission from S_1 to all mobile stations. Due to the MMSE based design, the Rx filters at the nodes and the Tx filters at S_1 depend on the overall channels and therewith, they depend on the relay transceive filters. In this section, the Rx filters at the nodes and the Tx filters at S_1 are designed assuming that the relay transceive filters are known in advance. An alternating optimization between the filters at the nodes and the relay transceive filter on each subcarrier is presented in Section 3.3.4.

To consider the noise powers at the nodes with respect to the power constraint at S_1, it is proposed to consider an additional receive coefficient β_c at all mobile stations and to solve the joint optimization problem of β_c and $\mathbf{Q}_{1,c}$ on subcarrier c as it is considered for the conventional MIMO Tx filter design in [JUN05]. Thus, the general equation for the joint optimization problem of the receive coefficient β_c and the weighted SIC aware MMSE Tx filter $\mathbf{Q}_{1,c}$ on subcarrier c can be written as

$$\{\beta_c, \mathbf{Q}_{1,c}\} = \underset{\beta_c, \mathbf{Q}_{1,c}}{\arg\ \min}\ \mathrm{E}\left\{\left\|\sqrt{\mathbf{V}_{\mathrm{BS}}}\left(\mathbf{s}_{1,c} - \beta_c \hat{\mathbf{s}}_{1,c}\right)\right\|_2^2\right\}, \tag{3.25a}$$

s.t.

$$\mathrm{tr}\left(\mathbf{Q}_{1,c}\mathbf{Q}_{1,c}^{\mathrm{H}}\right) \leq (K-1)P_{\mathrm{node}}, \tag{3.25b}$$

where $\hat{\mathbf{s}}_{1,c}$ is the estimate of $\mathbf{s}_{1,c}$ after linear receive filtering and SIC at the different mobile stations as defined in (3.4) and (3.5).

To obtain a simplified expression for the MSE which can be used to derive an analytical solution for the Tx filter at S_1, let the matrices $\boldsymbol{\Upsilon}_{k,c}$, $\boldsymbol{\Theta}_{k,c}$ and $\boldsymbol{\Theta}_{k,m,c}$ be given by

$$\boldsymbol{\Upsilon}_{k,c} = \mathbf{H}_{k,c}\mathbf{Q}_{k,c}\mathbf{Q}_{k,c}^{\mathrm{H}}\mathbf{H}_{k,c}^{\mathrm{H}}, \tag{3.26a}$$

$$\boldsymbol{\Theta}_{c} = \mathrm{diag}\left[\mathbf{D}_{2,c}, \mathbf{D}_{3,c}, ..., \mathbf{D}_{K,c}\right]\left[\mathbf{H}_{2,c}, \mathbf{H}_{3,c}, ..., \mathbf{H}_{K,c}\right]^{\mathrm{T}}\mathbf{G}_c\mathbf{H}_{1,c}, \tag{3.26b}$$

$$\boldsymbol{\Theta}_{k,m,c} = \mathbf{R}_{(k-2)M+m}\boldsymbol{\Theta}_{c}, \tag{3.26c}$$

where $\mathbf{R}_i \in \mathcal{C}^{M_1 \times M_1}$ is a diagonal matrix with all elements equal to zero except for the ith diagonal element which is equal to one. Using this notation, the weighted MSE for the transmission from S_1 to the mobile stations with respect to the Tx filter at S_1 on subcarrier c can be written as

$$\begin{aligned}
&\mathrm{E}\left\{\left\|\sqrt{\mathbf{V}_{\mathrm{BS}}}\left(\mathbf{s}_{1,c}-\beta_c\hat{\mathbf{s}}_{1,c}\right)\right\|_2^2\right\}\\
&= M_1 - 2\Re\left[\mathrm{tr}\left(\beta_c\boldsymbol{\Theta}_c\mathbf{Q}_{1,c}\mathbf{V}_{\mathrm{BS}}\right)\right] + |\beta_c|^2\mathrm{tr}\left(\boldsymbol{\Theta}_c\mathbf{Q}_{1,c}\mathbf{V}_{\mathrm{BS}}\mathbf{Q}_{1,c}^{\mathrm{H}}\boldsymbol{\Theta}_c^{\mathrm{H}}\right)\\
&- |\beta_c|^2\mathrm{tr}\left(\sum_{k=2}^{K}\sum_{n=2}^{M}\sum_{m=1}^{n-1} v_{\mathrm{BS},k}\boldsymbol{\Theta}_{k,n,c}\mathbf{Q}_{1,c}\mathbf{R}_{(k-2)M+m}\mathbf{Q}_{1,c}^{\mathrm{H}}\boldsymbol{\Theta}_{k,n,c}^{\mathrm{H}}\right)\\
&+ |\beta_c|^2\mathrm{tr}\left(\sum_{k=2}^{K}\sum_{l=2,l\neq k}^{K} v_{\mathrm{BS},k}\mathbf{D}_{k,c}\mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c\boldsymbol{\Upsilon}_{l,c}\mathbf{G}_c^{\mathrm{H}}\mathbf{H}_{k,c}^{*}\mathbf{D}_{k,c}^{\mathrm{H}}\right)\\
&+ |\beta_c|^2\mathrm{tr}\left(\sum_{k=2}^{K} v_{\mathrm{BS},k}\left(\sigma_{\mathrm{n,RS}}^2\mathbf{D}_{k,c}\mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{G}_c^{\mathrm{H}}\mathbf{H}_{k,c}^{*}\mathbf{D}_{k,c}^{\mathrm{H}} + \sigma_{\mathrm{n}}^2\mathbf{D}_{k,c}\mathbf{D}_{k,c}^{\mathrm{H}}\right)\right). \qquad (3.27)
\end{aligned}$$

Using (3.27), the objective function (3.25a) is non-convex since $\mathbf{Q}_{1,c}$ and β_c appear jointly in third order degree or higher. However, β_c can be assumed to be positive real-valued and the MSE of (3.27) as well as the constraint (3.25b) are convex with respect to $\mathbf{Q}_{1,c}$. Based on the assumption that β_c is positive real valued, a unique solution for problem (3.25) can be obtained by using Lagrangian optimization [BV04, Joh04, Ung09]. With $F(\mathbf{Q}_{1,c},\beta_c,c) = \mathrm{E}\left\{\left\|\sqrt{\mathbf{V}_{\mathrm{BS}}}\left(\mathbf{s}_{1,c}-\beta_c\hat{\mathbf{s}}_{1,c}\right)\right\|_2^2\right\}$, using (3.27), the Lagrangian function with the Lagrangian multiplier η_c results in

$$L\left(\mathbf{Q}_{1,c},\beta_c,\eta_c\right) = F(\mathbf{Q}_{1,c},\beta_c,c) - \eta_c\left(\mathrm{tr}\left(\mathbf{Q}_{1,c}\mathbf{Q}_{1,c}^{\mathrm{H}}\right) - (K-1)P_{\mathrm{node}}\right). \qquad (3.28)$$

From the Lagrangian function, the Karush-Kuhn-Tucker (KKT) conditions, which are necessary conditions for a global optimum, can be derived and η_c can be computed, which is presented in detail in Appendix A.1. The KKT conditions can be written as

$$\frac{\partial L}{\partial \mathbf{Q}_{1,c}} = \frac{\partial F(\mathbf{Q}_{1,c},\beta_c,c)}{\partial \mathbf{Q}_{1,c}} - \eta_c\mathbf{Q}_{1,c}^{*} = \mathbf{0}, \qquad (3.29a)$$

$$\frac{\partial L}{\partial \beta_c} = \frac{\partial F(\mathbf{Q}_{1,c},\beta_c,c)}{\partial \beta_c} = 0, \qquad (3.29b)$$

$$\eta_c\left(\mathrm{tr}\left(\mathbf{Q}_{1,c}\mathbf{Q}_{1,c}^{\mathrm{H}}\right) - (K-1)P_{\mathrm{node}}\right) = 0, \qquad (3.29c)$$

and the Lagrangian multiplier η_c results in

$$\eta_c = -\frac{|\beta_c|^2\sum_{k=2}^{K}\mathrm{tr}\left(\mathbf{A}_{k,c}\right)}{(K-1)P_{\mathrm{node}}}, \qquad (3.30)$$

where

$$\begin{aligned}
\mathbf{A}_{k,c} &= v_{\mathrm{BS},k}\cdot\mathrm{tr}\left(\sum_{l=2,l\neq k}^{K}\mathbf{D}_{k,c}\mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c\boldsymbol{\Upsilon}_{l,c}\mathbf{G}_c^{\mathrm{H}}\mathbf{H}_{k,c}^{*}\mathbf{D}_{k,c}^{\mathrm{H}}\right)\\
&+ v_{\mathrm{BS},k}\cdot\mathrm{tr}\left(\sigma_{\mathrm{n,RS}}^2\mathbf{D}_{k,c}\mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{G}_c^{\mathrm{H}}\mathbf{H}_{k,c}^{*}\mathbf{D}_{k,c}^{\mathrm{H}} + \sigma_{\mathrm{n}}^2\mathbf{D}_{k,c}\mathbf{D}_{k,c}^{\mathrm{H}}\right). \qquad (3.31)
\end{aligned}$$

Now, considering the derivations in Appendix A.1, a unique solution can be obtained for $|\beta_c|^2$. Thus, restricting β_c to be positive real-valued, a unique solution can be obtained for problem (3.25).

Considering the first KKT condition (3.29a) and using (3.31), a matrix $\mathbf{M}_c$ is defined as

$$\mathbf{M}_c = \mathbf{V}_{\mathrm{BS}}\mathbf{I}_{M_1} \otimes \mathbf{\Theta}_c^{\mathrm{H}}\mathbf{\Theta}_c - \sum_{k=2}^{K} v_{\mathrm{BS},k} \sum_{n=2}^{M}\sum_{m=1}^{n-1} \mathbf{R}_{(k-2)M+m}^{\mathrm{T}} \otimes \mathbf{\Theta}_{k,n,c}^{\mathrm{H}}\mathbf{\Theta}_{k,n,c} \\ + \mathbf{I}_{M_1} \otimes \frac{\sum_{k=2}^{K} \mathrm{tr}\left(\mathbf{A}_{k,c}\right)}{(K-1)P_{\mathrm{node}}}\mathbf{I}_{M_1}. \tag{3.32}$$

Now, an analytical solution can be obtained for the weighted SIC aware MMSE Tx filter at S_1. Using the auxiliary matrix $\widetilde{\mathbf{Q}}_{1,c}$ given by

$$\widetilde{\mathbf{Q}}_{1,c} = \mathrm{vec}_{L,L}^{-1}\left(\mathbf{M}_c^{-1}\ \mathrm{vec}\left(\mathbf{\Theta}_{k,c}^{\mathrm{H}}\mathbf{V}_{\mathrm{BS}}^{\mathrm{H}}\right)\right), \tag{3.33}$$

and using

$$\beta_c = \sqrt{\frac{\mathrm{tr}\left(\widetilde{\mathbf{Q}}_{1,c}\widetilde{\mathbf{Q}}_{1,c}^{\mathrm{H}}\right)}{(K-1)P_{\mathrm{node}}}}, \tag{3.34}$$

the weighted SIC aware MMSE Tx filter at S_1 which solves problem (3.25) can be written as

$$\mathbf{Q}_{1,c} = \frac{1}{\beta_c}\mathrm{vec}_{L,L}^{-1}\left(\mathbf{M}_c^{-1}\ \mathrm{vec}\left(\mathbf{\Theta}_c^{\mathrm{H}}\mathbf{V}_{\mathrm{BS}}^{\mathrm{H}}\right)\right). \tag{3.35}$$

3.3.3 Transceive Filter Design at Relay Station

3.3.3.1 Weighted Zero-Forcing (WZF) Approach

In this section, a weighted ZF (WZF) approach is introduced which does neither exploit the self-interference cancellation capabilities nor the SIC capabilities of the nodes due to spatially separating all signals. This approach is an extension of the ZF approach for single-pair two-way relaying with limited capabilities at the nodes of [Ung09] and it is only introduced for comparison due to the lack of state of the art approaches which consider the introduced ADR requirements. Compared to the ZF approach of [Ung09], the WZF approach contains weighting parameters to adjust the Tx power distribution at RS with respect to the considered ADR requirements. Furthermore, it is extended to be applicable to the cellular multi-user two-way relaying scenario. As in [Ung09], the Tx and Rx filters at the nodes are assumed to be diagonal matrices. Thus, they are designed according to case Diag which is presented in Section 3.3.2.1.

Defining overall channels for the MAC and for the BC phase on subcarrier c as

$$\mathbf{H}_{\mathrm{MAC},c} = \left[\mathbf{H}_{1,c}\mathbf{Q}_{1,c}, \mathbf{H}_{2,c}\mathbf{Q}_{2,c}, ..., \mathbf{H}_{K,c}\mathbf{Q}_{K,c}\right], \tag{3.36a}$$

$$\mathbf{H}_{\mathrm{BC},c} = \left[\mathbf{H}_{2,c}\mathbf{D}_{2,c}^{\mathrm{T}}, \mathbf{H}_{3,c}\mathbf{D}_{3,c}^{\mathrm{T}}, ..., \mathbf{H}_{K,c}\mathbf{D}_{K,c}^{\mathrm{T}}, \mathbf{H}_{1,c}\mathbf{D}_{1,c}^{\mathrm{T}}\right]^{\mathrm{T}}, \tag{3.36b}$$

respectively, and defining a diagonal weighting matrix $\mathbf{V}_c$ using (3.16b) and (3.17) as

$$\mathbf{V}_c = \mathrm{diag}\left[\mathbf{v}_{\mathrm{BS}}, \mathbf{v}_{\mathrm{MS},c}\right], \tag{3.37}$$

the WZF transceive filter at RS on subcarrier c using the derivation of [Ung09] can be written as

$$\mathbf{G}_c = \frac{1}{\alpha_{\mathrm{ZF},c}}\mathbf{H}_{\mathrm{BC},c}^{\mathrm{H}}\left(\mathbf{H}_{\mathrm{BC},c}\mathbf{H}_{\mathrm{BC},c}^{\mathrm{H}}\right)^{-1}\mathbf{V}_c\left(\mathbf{H}_{\mathrm{MAC},c}^{\mathrm{H}}\mathbf{H}_{\mathrm{MAC},c}\right)^{-1}\mathbf{H}_{\mathrm{MAC},c}^{\mathrm{H}}, \tag{3.38}$$

where $\alpha_{\mathrm{ZF},c}$ is a parameter to fulfill the power constraint at RS and it is given by

$$\alpha_{\mathrm{ZF},c} = \sqrt{\frac{\mathrm{tr}\left(\widetilde{\mathbf{G}}_c\left(\sum_{k=1}^{K}\left(\mathbf{H}_{k,c}\mathbf{Q}_{k,c}\mathbf{Q}_{k,c}^{\mathrm{H}}\mathbf{H}_{k,c}^{\mathrm{H}}\right) + \sigma_{\mathrm{n,RS}}^2\mathbf{I}_L\right)\widetilde{\mathbf{G}}_c^{\mathrm{H}}\right)}{P_{\mathrm{RS}}}}, \tag{3.39}$$

with the auxiliary matrix $\widetilde{\mathbf{G}}_c$ given by

$$\widetilde{\mathbf{G}}_c = \mathbf{H}_{\mathrm{BC},c}^{\mathrm{H}}\left(\mathbf{H}_{\mathrm{BC},c}\mathbf{H}_{\mathrm{BC},c}^{\mathrm{H}}\right)^{-1}\mathbf{V}_c\left(\mathbf{H}_{\mathrm{MAC},c}^{\mathrm{H}}\mathbf{H}_{\mathrm{MAC},c}\right)^{-1}\mathbf{H}_{\mathrm{MAC},c}^{\mathrm{H}}. \tag{3.40}$$

3.3.3.2 Weighted MMSE (WMMSE) Approach

In this section, a weighted MMSE (WMMSE) approach is introduced which does neither exploit the self-interference cancellation capabilities nor the SIC capabilities of the nodes due to suppressing all interferences with respect to minimizing the MSE. This WMMSE approach is an extension of the MMSE approach for single-pair two-way relaying with limited capabilities at the nodes of [Ung09] and it is only introduced for comparison due to the lack of state of the art approaches which consider the introduced ADR requirements. Compared to the MMSE approach of [Ung09], the WMMSE approach contains weighting parameters to adjust the Tx power distribution at RS with respect to the considered ADR requirements. Furthermore, it is extended to be applicable to the cellular multi-user two-way relaying scenario. Compared to the WZF approach of Section 3.3.3.1, the noise powers at the nodes and at RS are taken into account for the WMMSE relay transceive filter design. As in [Ung09], the Tx and Rx filters at the nodes are assumed to be diagonal matrices. Thus, they are designed according to case Diag which is presented in Section 3.3.2.1.

Using the overall channels for the MAC and for the BC phase on subcarrier c as defined in (3.36) and using (3.37), the WMMSE transceive filter at RS on subcarrier c using the derivation of [Ung09] can be written as

$$\mathbf{G}_c = \frac{1}{\alpha_{\mathrm{MMSE},c}} \widetilde{\mathbf{G}}_c, \tag{3.41}$$

with the auxiliary matrix $\widetilde{\mathbf{G}}_c$ given by

$$\widetilde{\mathbf{G}}_c = \left(\mathbf{H}_{\mathrm{BC},c}^{\mathrm{H}} \mathbf{V}_c \mathbf{H}_{\mathrm{BC},c} + \frac{\mathrm{tr}\left(\mathbf{V}_c\right) \sigma_{\mathrm{n}}^2}{P_{\mathrm{RS}}} \mathbf{I}_L \right)^{-1} \mathbf{H}_{\mathrm{BC},c}^{\mathrm{H}} \mathbf{V}_c \mathbf{H}_{\mathrm{MAC},c}^{\mathrm{H}} \left(\mathbf{H}_{\mathrm{MAC},c} \mathbf{H}_{\mathrm{MAC},c}^{\mathrm{H}} + \sigma_{\mathrm{n,RS}}^2 \mathbf{I}_L \right)^{-1}, \tag{3.42}$$

and with $\alpha_{\mathrm{MMSE},c}$ given by

$$\alpha_{\mathrm{MMSE},c} = \sqrt{\frac{\mathrm{tr}\left(\widetilde{\mathbf{G}}_c \left(\sum_{k=1}^{K} \left(\mathbf{H}_{k,c} \mathbf{Q}_{k,c} \mathbf{Q}_{k,c}^{\mathrm{H}} \mathbf{H}_{k,c}^{\mathrm{H}} \right) + \sigma_{\mathrm{n,RS}}^2 \mathbf{I}_L \right) \widetilde{\mathbf{G}}_c^{\mathrm{H}} \right)}{P_{\mathrm{RS}}}}. \tag{3.43}$$

3.3.3.3 SIC-Aware Weighted MMSE (WMMSE-SIC) Approach

In this section, a weighted MMSE approach for the transceive filter design at RS is presented which exploits the self-interference cancellation capabilities and the SIC capabilities of the nodes. The approach, termed WMMSE-SIC, is an extension of the MMSE approach for single-pair two-way relaying with local CSI at the nodes of [Ung09]. Compared to the MMSE approach of [Ung09], the WMMSE-SIC approach contains weighting parameters to adjust the Tx power distribution at RS with respect to the considered ADR requirements. Furthermore, it exploits the SIC capabilities of the nodes to increase the achievable sum rates. Moreover, it is extended to be applicable to the cellular multi-user two-way relaying scenario by considering BS-MS-interferences and MS-MS-interferences. For this WMMSE-SIC approach, all three cases of Tx and Rx filter design at the nodes presented in Section 3.3.2 are considered. Thus, the design of the relay transceive filter depends on the Tx and Rx filters at the nodes and vice versa. In the following, the WMMSE-SIC transceive filter at RS is derived assuming given Tx and Rx filters at the nodes to obtain an analytical solution. However, this analytical solution depends on the Tx and Rx filters at the nodes. To jointly design the WMMSE-SIC relay transceive filter and the Tx and Rx filters at the nodes, an alternating optimization between the filters at the nodes and the relay transceive filter on each subcarrier is presented in Section 3.3.4.

To include the weighting parameters, the MSE for each direction of transmission is separated for the derivation of the WMMSE-SIC relay transceive filter. After separating the different MSEs, the weighting parameters $v_{\mathrm{BS},k}$ and $v_{\mathrm{MS},k}$ are multiplied with the corresponding MSE to enable an adjustment of the fraction of the Tx power at RS which is used to retransmit the data streams from S_1 to S_k and from S_k to S_1, respectively, $k = 2, 3, ..., K$. Furthermore, to consider the noise powers at the nodes with respect to the power constraint at RS, it is proposed to consider an additional receive coefficient α_c at all nodes and to solve the joint optimization problem of α_c and $\mathbf{G}_c$ on subcarrier c as it is considered for the conventional MIMO Tx filter design in [JUN05]. Thus, the general equation for the joint optimization problem of the receive coefficient α_c and the WMMSE-SIC relay transceive filter $\mathbf{G}_c$ on subcarrier c can be written as

$$\{\alpha_c, \mathbf{G}_c\} = \underset{\alpha_c, \mathbf{G}_c}{\arg\ \min}\ \mathrm{E}\left\{\sum_{k=2}^{K} v_{\mathrm{BS},k} \left\|\mathbf{s}_{1,k,c} - \alpha_c \hat{\mathbf{s}}_{1,k,c}\right\|_2^2 + \sum_{k=2}^{K} v_{\mathrm{MS},k} \left\|\mathbf{s}_{k,c} - \alpha_c \hat{\mathbf{s}}_{k,c}\right\|_2^2\right\}, \tag{3.44a}$$

s.t.

$$\mathrm{tr}\left(\mathbf{G}_c \left(\sum_{k=1}^{K} \left(\mathbf{H}_{k,c}\mathbf{Q}_{k,c}\mathbf{Q}_{k,c}^{\mathrm{H}}\mathbf{H}_{k,c}^{\mathrm{H}}\right) + \sigma_{\mathrm{n,RS}}^2 \mathbf{I}_L\right) \mathbf{G}_c^{\mathrm{H}}\right) \leq P_{\mathrm{RS}}, \tag{3.44b}$$

where $\hat{\mathbf{s}}_{1,k,c}$ and $\hat{\mathbf{s}}_{k,c}$ are the estimates of $\mathbf{s}_{1,k,c}$ and $\mathbf{s}_{k,c}$ after linear receive filtering, self-interference cancellation and SIC at S_k and S_1 considering (3.5) and (3.6), respectively.

To obtain a simplified expression of the MSE, let matrices $\boldsymbol{\Upsilon}_{k,c}$, $\boldsymbol{\Upsilon}_{k,m,c}^{(\mathrm{BS})}$, $\boldsymbol{\Upsilon}_{k,m,c}^{(\mathrm{MS})}$ and $\boldsymbol{\Upsilon}_c$ be given by

$$\boldsymbol{\Upsilon}_{k,c} = \mathbf{H}_{k,c}\mathbf{Q}_{k,c}\mathbf{Q}_{k,c}^{\mathrm{H}}\mathbf{H}_{k,c}^{\mathrm{H}}, \tag{3.45a}$$

$$\boldsymbol{\Upsilon}_{k,m,c}^{(\mathrm{BS})} = \mathbf{H}_{1,c}\mathbf{q}_{1,k,m,c}\mathbf{q}_{1,k,m,c}^{\mathrm{H}}\mathbf{H}_{1,c}^{\mathrm{H}}, \tag{3.45b}$$

$$\boldsymbol{\Upsilon}_{k,m,c}^{(\mathrm{MS})} = \mathbf{H}_{k,c}\mathbf{q}_{k,m,c}\mathbf{q}_{k,m,c}^{\mathrm{H}}\mathbf{H}_{k,c}^{\mathrm{H}}, \tag{3.45c}$$

$$\boldsymbol{\Upsilon}_{c} = \sum_{k=1}^{K} \left(\mathbf{H}_{k,c}\mathbf{Q}_{k,c}\mathbf{Q}_{k,c}^{\mathrm{H}}\mathbf{H}_{k,c}^{\mathrm{H}}\right) + \sigma_{\mathrm{n,RS}}^2 \mathbf{I}_L. \tag{3.45d}$$

Using these notations, the MSE for the transmission from S_1 to S_k, $k = 2, 3, ...K$ on subcarrier c can be written as

$$\begin{aligned}\mathrm{E}\left\{\|\mathbf{s}_{1,k,c}-\alpha_c\hat{\mathbf{s}}_{1,k,c}\|_2^2\right\} &= M - 2\Re\left[\mathrm{tr}\left(\alpha_c\mathbf{D}_{k,c}\mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{H}_{1,c}\mathbf{Q}_{1,k,c}\right)\right] \\ &+ |\alpha_c|^2\mathrm{tr}\left(\sum_{l=1,l\neq k}^{K}\mathbf{D}_{k,c}\mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c\boldsymbol{\Upsilon}_{l,c}\mathbf{G}_c^{\mathrm{H}}\mathbf{H}_{k,c}^{*}\mathbf{D}_{k,c}^{\mathrm{H}}\right) \\ &- |\alpha_c|^2\mathrm{tr}\left(\sum_{n=2}^{M}\sum_{m=1}^{n-1}\mathbf{d}_{k,n,c}\mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c\boldsymbol{\Upsilon}_{k,m,c}^{(\mathrm{BS})}\mathbf{G}_c^{\mathrm{H}}\mathbf{H}_{k,c}^{*}\mathbf{d}_{k,n,c}^{\mathrm{H}}\right) \\ &+ |\alpha_c|^2\mathrm{tr}\left(\sigma_{\mathrm{n,RS}}^2\mathbf{D}_{k,c}\mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{G}_c^{\mathrm{H}}\mathbf{H}_{k,c}^{*}\mathbf{D}_{k,c}^{\mathrm{H}} + \sigma_{\mathrm{n}}^2\mathbf{D}_{k,c}\mathbf{D}_{k,c}^{\mathrm{H}}\right). \end{aligned} \quad (3.46)$$

Furthermore, the MSE for the transmission from S_k to S_1, $k = 2, 3, ...K$ on subcarrier c can be written as

$$\begin{aligned}\mathrm{E}\left\{\|\mathbf{s}_{k,c}-\alpha_c\hat{\mathbf{s}}_{k,c}\|_2^2\right\} &= m_{k,c} - 2\Re\left[\mathrm{tr}\left(\alpha_c\mathbf{D}_{1,k,c}\mathbf{H}_{1,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{H}_{k,c}\mathbf{Q}_{k,c}\right)\right] \\ &+ |\alpha_c|^2\mathrm{tr}\left(\sum_{l=k}^{K}\mathbf{D}_{1,k,c}\mathbf{H}_{1,c}^{\mathrm{T}}\mathbf{G}_c\boldsymbol{\Upsilon}_{l,c}\mathbf{G}_c^{\mathrm{H}}\mathbf{H}_{k,c}^{*}\mathbf{D}_{1,k,c}^{\mathrm{H}}\right) \\ &- |\alpha_c|^2\mathrm{tr}\left(\sum_{n=2}^{m_{k,c}}\sum_{m=1}^{n-1}\mathbf{d}_{1,k,n,c}\mathbf{H}_{1,c}^{\mathrm{T}}\mathbf{G}_c\boldsymbol{\Upsilon}_{k,m,c}^{(\mathrm{MS})}\mathbf{G}_c^{\mathrm{H}}\mathbf{H}_{1,c}^{*}\mathbf{d}_{1,k,n,c}^{\mathrm{H}}\right) \\ &+ |\alpha_c|^2\mathrm{tr}\left(\sigma_{\mathrm{n,RS}}^2\mathbf{D}_{1,k,c}\mathbf{H}_{1,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{G}_c^{\mathrm{H}}\mathbf{H}_{1,c}^{*}\mathbf{D}_{1,k,c}^{\mathrm{H}} + \sigma_{\mathrm{n}}^2\mathbf{D}_{1,k,c}\mathbf{D}_{1,k,c}^{\mathrm{H}}\right). \end{aligned} \quad (3.47)$$

Using (3.46) and (3.47), the objective function (3.44a) is non-convex since $\mathbf{G}_c$ and α_c appear jointly in third order degree or higher. However, α_c can be assumed to be positive real-valued and the MSEs of (3.46) and (3.47) as well as the constraint (3.44b) are convex with respect to $\mathbf{G}_c$. Based on the assumption that α_c is positive real valued, a unique solution for problem (3.44) can be obtained by using Lagrangian optimization [BV04, Joh04, Ung09]. With $F_{\mathrm{BS}}(\mathbf{G}_c, \alpha_c, k, c) = v_{\mathrm{BS},k}\mathrm{E}\left\{\|\mathbf{s}_{1,k,c}-\alpha_c\hat{\mathbf{s}}_{1,k,c}\|_2^2\right\}$, using (3.46), and $F_{\mathrm{MS}}(\mathbf{G}_c, \alpha_c, k, c) = v_{\mathrm{MS},k}\mathrm{E}\left\{\|\mathbf{s}_{k,c}-\alpha_c\hat{\mathbf{s}}_{k,c}\|_2^2\right\}$, using (3.47), the Lagrangian function with the Lagrangian multiplier η_c results in

$$L\left(\mathbf{G}_c, \alpha_c, \eta_c\right) = \sum_{k=2}^{K}\left(F_{\mathrm{BS}}(\mathbf{G}_c, \alpha_c, k, c) + F_{\mathrm{MS}}(\mathbf{G}_c, \alpha_c, k, c)\right) - \eta_c\left(\mathrm{tr}\left(\mathbf{G}_c\boldsymbol{\Upsilon}_c\mathbf{G}_c^{\mathrm{H}}\right) - P_{\mathrm{RS}}\right). \quad (3.48)$$

From the Lagrangian function, the KKT conditions, which are necessary conditions for a global optimum, can be derived and η_c can be computed, which is presented in detail in Appendix A.2. The KKT conditions can be written as

$$\frac{\partial L}{\partial \mathbf{G}_c} = \sum_{k=2}^{K} \frac{\partial \left(F_{\mathrm{BS}}(\mathbf{G}_c, \alpha_c, k, c) + F_{\mathrm{MS}}(\mathbf{G}_c, \alpha_c, k, c)\right)}{\partial \mathbf{G}_c} - \eta_c \, \mathbf{G}_c^* \boldsymbol{\Upsilon}_c^{\mathrm{T}} = \mathbf{0}, \tag{3.49a}$$

$$\frac{\partial L}{\partial \alpha_c} = \sum_{k=2}^{K} \frac{\partial \left(F_{\mathrm{BS}}(\mathbf{G}_c, \alpha_c, k, c) + F_{\mathrm{MS}}(\mathbf{G}_c, \alpha_c, k, c)\right)}{\partial \alpha_c} = 0, \tag{3.49b}$$

$$\eta_c \left(\mathrm{tr}\left(\mathbf{G}_c \boldsymbol{\Upsilon}_c \mathbf{G}_c^{\mathrm{H}}\right) - P_{\mathrm{RS}}\right) = 0, \tag{3.49c}$$

and the Lagrangian multiplier η_c results in

$$\eta_c = -\frac{|\alpha_c|^2 \sigma_{\mathrm{n}}^2 \left(\sum_{k=2}^{K} \upsilon_{\mathrm{BS},k} \mathrm{tr}\left(\mathbf{D}_{k,c} \mathbf{D}_{k,c}^{\mathrm{H}}\right) + \sum_{k=2}^{K} \upsilon_{\mathrm{MS},k} \mathrm{tr}\left(\mathbf{D}_{1,k,c} \mathbf{D}_{1,k,c}^{\mathrm{H}}\right)\right)}{P_{\mathrm{RS}}}. \tag{3.50}$$

Now, considering the derivations in Appendix A.2, a unique solution can be obtained for $|\alpha_c|^2$. Thus, restricting α_c to be positive real-valued, a unique solution can be obtained for problem (3.44).

Considering the first KKT condition (3.49a), a matrix $\mathbf{K}_c$ is defined as

$$\begin{aligned}
\mathbf{K}_c = &\sum_{k=2}^{K} \upsilon_{\mathrm{BS},k} \sum_{l=1, l \neq k}^{K} \boldsymbol{\Upsilon}_{l,c}^{\mathrm{T}} \otimes \mathbf{H}_{k,c}^* \mathbf{D}_{k,c}^{\mathrm{H}} \mathbf{D}_{k,c} \mathbf{H}_{k,c}^{\mathrm{T}} \\
&- \sum_{k=2}^{K} \upsilon_{\mathrm{BS},k} \sum_{n=2}^{M} \sum_{m=1}^{n-1} \boldsymbol{\Upsilon}_{k,m,c}^{(\mathrm{BS})\mathrm{T}} \otimes \mathbf{H}_{k,c}^* \mathbf{d}_{k,n,c}^{\mathrm{H}} \mathbf{d}_{k,n,c} \mathbf{H}_{k,c}^{\mathrm{T}} \\
&+ \sum_{k=2}^{K} \upsilon_{\mathrm{BS},k} \sigma_{\mathrm{n,RS}}^2 \mathbf{I}_L \otimes \mathbf{H}_{k,c}^* \mathbf{D}_{k,c}^{\mathrm{H}} \mathbf{D}_{k,c} \mathbf{H}_{k,c}^{\mathrm{T}} \\
&+ \sum_{k=2}^{K} \upsilon_{\mathrm{MS},k} \sum_{l=k}^{K} \boldsymbol{\Upsilon}_{l,c}^{\mathrm{T}} \otimes \mathbf{H}_{1,c}^* \mathbf{D}_{1,k,c}^{\mathrm{H}} \mathbf{D}_{1,k,c} \mathbf{H}_{1,c}^{\mathrm{T}} \\
&- \sum_{k=2}^{K} \upsilon_{\mathrm{MS},k} \sum_{n=2}^{m_{k,c}} \sum_{m=1}^{n-1} \boldsymbol{\Upsilon}_{k,m,c}^{(\mathrm{MS})\mathrm{T}} \otimes \mathbf{H}_{1,c}^* \mathbf{d}_{1,k,n,c}^{\mathrm{H}} \mathbf{d}_{1,k,n,c} \mathbf{H}_{1,c}^{\mathrm{T}} \\
&+ \sum_{k=2}^{K} \upsilon_{\mathrm{MS},k} \sigma_{\mathrm{n,RS}}^2 \mathbf{I}_L \otimes \mathbf{H}_{1,c}^* \mathbf{D}_{1,k,c}^{\mathrm{H}} \mathbf{D}_{1,k,c} \mathbf{H}_{1,c}^{\mathrm{T}}, \\
&+ \boldsymbol{\Upsilon}_c^{\mathrm{T}} \otimes \frac{\sigma_{\mathrm{n}}^2 \sum_{k=2}^{K} \left(\upsilon_{\mathrm{BS},k} \mathrm{tr}\left(\mathbf{D}_{k,c} \mathbf{D}_{k,c}^{\mathrm{H}}\right) + \upsilon_{\mathrm{MS},k} \mathrm{tr}\left(\mathbf{D}_{1,k,c} \mathbf{D}_{1,k,c}^{\mathrm{H}}\right)\right)}{P_{\mathrm{RS}}} \mathbf{I}_L.
\end{aligned} \tag{3.51}$$

Now, an analytical solution can be obtained for the WMMSE-SIC relay transceive filter which solves problem (3.44) using (3.49), (3.50) and (3.51). With the auxiliary matrix $\widetilde{\mathbf{G}}_c$ given by

$$\widetilde{\mathbf{G}}_c = \mathrm{vec}_{L,L}^{-1} \left(\mathbf{K}_c^{-1} \, \mathrm{vec} \left(\sum_{k=2}^{K} \left(\upsilon_{\mathrm{BS},k} \mathbf{H}_{k,c}^* \mathbf{D}_{k,c}^{\mathrm{H}} \mathbf{Q}_{1,k,c}^{\mathrm{H}} \mathbf{H}_{1,c}^{\mathrm{H}} + \upsilon_{\mathrm{MS},k} \mathbf{H}_{1,c}^* \mathbf{D}_{1,k,c}^{\mathrm{H}} \mathbf{Q}_{k,c}^{\mathrm{H}} \mathbf{H}_{k,c}^{\mathrm{H}} \right) \right) \right), \tag{3.52}$$

and using

$$\alpha_c = \sqrt{\frac{\mathrm{tr}\left(\widetilde{\mathbf{G}}_c \boldsymbol{\Upsilon}_c \widetilde{\mathbf{G}}_c^{\mathrm{H}}\right)}{P_{\mathrm{RS}}}}, \tag{3.53}$$

the WMMSE-SIC filter at RS which solves problem (3.44) is given by

$$\mathbf{G}_c = \frac{1}{\alpha_c}\mathrm{vec}_{L,L}^{-1}\left(\mathbf{K}_c^{-1}\ \mathrm{vec}\left(\sum_{k=2}^{K}\left(v_{\mathrm{BS},k}\mathbf{H}_{k,c}^{*}\mathbf{D}_{k,c}^{\mathrm{H}}\mathbf{Q}_{1,k,c}^{\mathrm{H}}\mathbf{H}_{1,c}^{\mathrm{H}} + v_{\mathrm{MS},k}\mathbf{H}_{1,c}^{*}\mathbf{D}_{1,k,c}^{\mathrm{H}}\mathbf{Q}_{k,c}^{\mathrm{H}}\mathbf{H}_{k,c}^{\mathrm{H}}\right)\right)\right). \tag{3.54}$$

The derived WMMSE-SIC relay transceive filter minimizes the weighted MSE for given Tx and Rx filters at the nodes considering that the nodes can perform self-interference cancellation and SIC.

3.3.4 Joint Filter Design at Nodes and at Relay Station

Considering the three cases for the Tx and Rx filter design at the nodes presented in Section 3.3.2, the Tx and Rx filters at the nodes are independent of the relay transceive filter design in case Diag and are dependent on the relay transceive filter design in case Rx and case Rx&Tx. Furthermore, the proposed WMMSE-SIC relay transceive filter depends on the Tx and Rx filters at the nodes.

For case Diag, a joint optimization is not required because the Tx and Rx filters at the nodes do not depend on the relay transceive filter design. In this case, the WMMSE-SIC relay transceive filter $\mathbf{G}_c$ can be computed on each subcarrier c according to (3.54) considering the Tx and Rx filters as defined in (3.18) and (3.19), respectively.

For case Rx and case Rx&Tx, the filters at the nodes depend on the relay transceive filter and vice versa. Thus, an alternating optimization between the WMMSE-SIC relay transceive filter and the Tx and Rx filters at the nodes is proposed as follows:

1) Compute the transceive filter $\mathbf{G}_c$ using (3.54) and considering (3.18) and (3.19) for the Tx and Rx filters at the nodes, respectively,

2) Compute the Rx filters according to (3.23) and (3.24),

3) For case Rx&Tx, compute the Tx filter at S_1 according to (3.35) considering the Rx filters computed in step 2),

4) Compute the transceive filter $\mathbf{G}_c$ using (3.54) and considering the Rx and Tx filters computed in step 2) and step 3), respectively,

5) Continue from step 2) using the relay transceive filter computed in step 4) until convergence.

3.4 Transmit Strategies for the Consideration of ADR Requirements

3.4.1 Introduction

In this section, two Tx strategies are proposed which are specifically designed to fulfill the considered ADR requirements whilst achieving high sum rates. First, a Tx strategy, termed power adapted (PA) Tx strategy, is proposed which adapts the Tx powers of the mobile stations and the Tx power distributions at RS and at S_1 for each direction of transmission. Secondly, a Tx strategy which additionally optimizes the allocation of the different subcarriers to the mobile stations, termed subcarrier allocation (SA) Tx strategy, is proposed. For all Tx strategies, it is assumed that the required data rates in downlink are r times higher than the required data rates in uplink as introduced in Section 3.2.

3.4.2 Power Adapted (PA) Transmit Strategy

The PA Tx strategy is based on adjusting the Tx powers of the mobile stations and the Tx power distributions at RS and at S_1 for each direction of transmission. The Tx power distributions at S_1 and at RS are adjusted via the weighting parameters $v_{\mathrm{BS},k}$ and $v_{\mathrm{MS},k}$ which have been considered for the Tx filter design at S_1 in Section 3.3.2 and for the relay transceive filter design in Section 3.3.3. The Tx powers of the mobile stations are adjusted via the weighting parameters $p_{k,c}$ which have been considered for the Tx filter design at the mobile stations (3.19b). To achieve high sum rates under the ADR requirements (3.15), the weighting parameters which achieve the highest sum rate according to (3.15) have to be determined. However, this would require a joint optimization of all weighting parameters which has a high computational complexity. Thus, to reduce the computational complexity, a suboptimal low-complexity approach is proposed. In this approach, the joint adjustment of the weighting parameters $v_{\mathrm{BS},k}$, $v_{\mathrm{MS},k}$ and $p_{k,c}$ is separated and performed iteratively as follows:

First, the relay transceive filter and the filters at the nodes are initialized and the achievable data rates are computed as follows:

1) Compute the Tx and Rx filters at the nodes and the relay transceive filter as proposed in Section 3.3 on each subcarrier c assuming $p_{k,c} = v_{\mathrm{BS},k} = v_{\mathrm{MS},k} = 1$, $k = 2, 3, ..., K$,

2) Compute the achievable data rates $C_{1,k}$ and $C_{k,1}$ (3.10) using the filters of step 1).

Secondly, the weighting parameters $v_{\mathrm{BS},k}$ are adjusted such that the achievable data rates in downlink from S_1 to each mobile station are equal. To achieve this, the weighting parameter $v_{\mathrm{BS},k_{\min}}$ with $k_{\min} = \arg\min_k C_{1,k}$ which corresponds to the minimum of the achievable data rates in downlink is set to $v_{\mathrm{BS},k_{\min}} = 1$. Furthermore, the weighting parameter $v_{\mathrm{BS},k_{\max}}$ with $k_{\max} = \arg\max_k C_{1,k}$ which corresponds to the maximum of the achievable data rates in downlink is reduced. By this approach, the differences between the achievable data rates in downlink are decreased in each iteration until all data rates in downlink are equal. The precise steps are as follows:

3) Compute $k_{\min} = \arg\min_k C_{1,k}$ and set $v_{\mathrm{BS},k_{\min}} = 1$.

4) Compute the filters at the nodes and the relay transceive filter for the weighting parameters of step 3).

5) Compute the achievable data rates $C_{1,k}$ and $C_{k,1}$ (3.10) using the filters of step 4).

6) Reduce the maximum downlink data rate from S_1 to any mobile station. To achieve this, reduce the weighting parameter $v_{\mathrm{BS},k_{\max}}$ with $k_{\max} = \arg\max_k C_{1,k}$ to fulfill the condition

$$\frac{1}{K-1}\sum_{l=2}^{K} C_{1,l} - \frac{\epsilon}{2} < C_{1,k_{\max}} < \frac{1}{K-1}\sum_{l=2}^{K} C_{1,l} + \frac{\epsilon}{2}, \tag{3.55}$$

where ϵ can be selected according to the required accuracy. To adjust the weighting parameter $v_{\mathrm{BS},k_{\max}}$, the bisection method is applied. The Tx and Rx filters at the nodes and the relay transceive filter are recalculated after each update of $v_{\mathrm{BS},k_{\max}}$. Furthermore, the achievable data rates $C_{1,k}$ and $C_{k,1}$ (3.10) are recalculated after each update of the filters.

Thirdly, the weighting parameters $v_{\mathrm{MS},k}$ are adjusted such that the achievable data rates in uplink from each mobile station to S_1 are equal. To achieve this, the weighting parameter $v_{\mathrm{MS},k_{\max}}$ with $k_{\max} = \arg\max_k C_{k,1}$ which corresponds to the maximum of the achievable data rates in uplink is reduced. Furthermore, the weighting parameter $v_{\mathrm{MS},k_{\min}}$ with $k_{\min} = \arg\min_k C_{k,1}$ which corresponds to the minimum of the achievable data rates in uplink is increased. By this approach, the differences between the achievable data rates in uplink are decreased in each iteration until all data rates in uplink are equal. The precise steps are as follows:

7) Reduce the maximum uplink data rate from any mobile station to S_1 if this data rate weighted with r is higher than the average downlink data rate, i.e., if $\max_k rC_{k_{\max},1} > \frac{1}{K-1}\sum_{l=2}^{K} C_{1,l}$. To achieve this, reduce $v_{\mathrm{MS},k_{\max}}$ with $k_{\max} = \arg\max_k C_{k,1}$ to fulfill the condition

$$\frac{1}{K-1}\sum_{l=2}^{K} C_{1,l} - \frac{\epsilon}{2} < rC_{k_{\max},1} < \frac{1}{K-1}\sum_{l=2}^{K} C_{1,l} + \frac{\epsilon}{2}, \tag{3.56}$$

or until the conditions

$$rC_{k_{\max},1} > \frac{1}{K-1}\sum_{l=2}^{K} C_{1,l} - \frac{\epsilon}{2}, \tag{3.57a}$$

$$v_{\mathrm{MS},k_{\max}} < \delta, \tag{3.57b}$$

are fulfilled, where $0 < \delta < 1$ ensures that the MSE for this direction of transmission has a sufficient impact on the considered MMSE based relay transceive filter designs. This is required due to decoupling the adjustment of $v_{\mathrm{MS},k}$ from the adjustment of $p_{k,c}$. For an efficient adjustment of $p_{k,c}$ in steps 11) and 12), it is required that $v_{\mathrm{MS},k} > 0\ \forall k$. In this thesis, $\delta = 0.1$ is selected based on numerical results. To adjust $v_{\mathrm{MS},k_{\max}}$, the bisection method is applied. Similar to step 6), the Tx and Rx filters at the nodes and the relay transceive filter are recalculated after each update of $v_{\mathrm{MS},k_{\max}}$. Furthermore, the achievable data rates $C_{1,k}$ and $C_{k,1}$ (3.10) are recalculated after each update of the filters.

8) Increase the minimum uplink data rate from any mobile station to S_1 if this data rate weighted with r is lower than the average downlink data rate, i.e., if $\min_k rC_{k,1} < \frac{1}{K-1}\sum_{l=2}^{K} C_{1,l}$. To achieve this, increase $v_{\mathrm{MS},k_{\min}}$ with $k_{\min} = \arg\min_k C_{k,1}$ to fulfill the condition

$$\frac{1}{K-1}\sum_{l=2}^{K} C_{1,l} - \frac{\epsilon}{2} < rC_{k_{\min},1} < \frac{1}{K-1}\sum_{l=2}^{K} C_{1,l} + \frac{\epsilon}{2}, \tag{3.58}$$

or until the conditions

$$rC_{k_{\min},1} < \frac{1}{K-1}\sum_{l=2}^{K} C_{1,l} + \frac{\epsilon}{2}, \tag{3.59a}$$

$$\upsilon_{\mathrm{MS},k_{\min}} = 1, \tag{3.59b}$$

are fulfilled. To adjust $\upsilon_{\mathrm{MS},k_{\min}}$, the bisection method is applied. Similar to the previous steps, the Tx and Rx filters at the nodes and the relay transceive filter are recalculated after each update of $\upsilon_{\mathrm{MS},k_{\min}}$. Furthermore, the achievable data rates $C_{1,k}$ and $C_{k,1}$ (3.10) are recalculated after each update of the filters.

Fourthly, the previous steps are repeated until the ADR requirements are fulfilled, i.e., until the achievable data rates in downlink are equal, the achievable data rates in uplink are equal and the ratio between the achievable data rates in downlink and in uplink is r taking into account an inaccuracy of ϵ. However, if the achievable data rates in uplink are not equal or the ratio between the achievable data rates in downlink and in uplink is not fulfilled even if $\upsilon_{\mathrm{MS},k_{\max}} \leq \delta$ with $k_{\max} = \arg\max_k C_{k,1}$, an adjustment of the Tx powers of the nodes is required by adjusting the weighting parameters $p_{k,c}$. In this case, the previous steps are not repeated. The precise conditions are as follows:

9) Continue from step 3) considering the updated weighting parameters $\upsilon_{\mathrm{BS},k}$ and $\upsilon_{\mathrm{MS},k}$ until the conditions

$$\max_k C_{1,k} < \min_k C_{1,k} + \epsilon, \tag{3.60a}$$

$$\max_k C_{k,1} < \min_k C_{k,1} + \epsilon, \tag{3.60b}$$

$$\max_k C_{1,k} - \epsilon < \max_k rC_{k,1} < \max_k C_{1,k} + \epsilon, \tag{3.60c}$$

are fulfilled or until the conditions

$$\max_k C_{1,k} < \min_k C_{1,k} + \epsilon, \tag{3.61a}$$

$$\upsilon_{\mathrm{MS},k_{\max}} \leq \delta, \tag{3.61b}$$

are fulfilled, where $k_{\max} = \arg\max_k C_{k,1}$.

Fifthly, the Tx powers of the nodes are adjusted if the conditions (3.61) are fulfilled and the conditions (3.60) are not fulfilled. The Tx powers of the nodes are adjusted via the weighting parameters $p_{k,c}$. To achieve this, the weighting parameter $p_{k_{\max}}$ with $k_{\max} = \arg\max_k C_{k,1}$ which corresponds to the maximum of the achievable data rates in uplink is reduced. Furthermore, the weighting parameter $p_{k_{\min}}$ with $k_{\min} = \arg\min_{k} C_{k,1}$ which corresponds to the minimum of the achievable data rates in uplink is increased. By this approach, the differences between the achievable data rates in uplink are decreased in each iteration until all data rates are equal. The precise steps are as follows:

11) Reduce the maximum uplink data rate from any mobile station to S_1 if this data rate weighted with r is higher than the average downlink data rate, i.e., if $\max_{k} rC_{k,1} > \frac{1}{K-1}\sum_{l=2}^{K} C_{1,l}$. To achieve this, reduce $p_{k_{\max}}$ with $k_{\max} = \arg\max_{k} C_{k,1}$ to fulfill the condition

$$\sum_{l=2}^{K} C_{1,l} - \frac{\epsilon}{2} < r\,(K-1)\,C_{k_{\max},1} < \sum_{l=2}^{K} C_{1,l} + \frac{\epsilon}{2}. \tag{3.62}$$

Similar to the previous steps, recalculate the filters and the achievable data rates after each adjustment.

12) Increase the minimum uplink data rate from any mobile station to S_1 if this data rate weighted with r is lower than the average downlink data rate, i.e., if $\min_{k} rC_{k,1} < \frac{1}{K-1}\sum_{l=2}^{K} C_{1,l}$. To achieve this, increase $p_{k_{\min}}$ with $k_{\min} = \arg\min_{k} C_{k,1}$ to fulfill the condition

$$\sum_{l=2}^{K} C_{1,l} - \frac{\epsilon}{2} < r\,(K-1)\,C_{k_{\min},1} < \sum_{l=2}^{K} C_{1,l} + \frac{\epsilon}{2}, \tag{3.63}$$

or until the conditions

$$r\,(K-1)\,C_{k_{\min},1} < \sum_{l=2}^{K} C_{1,l} + \frac{\epsilon}{2}, \tag{3.64a}$$

$$p_{k_{\min}} = 1, \tag{3.64b}$$

are fulfilled. Similar to the previous steps, recalculate the filters and update the achievable data rates after each adjustment.

Finally, the previous steps are repeated until the ADR requirements are fulfilled taking into account an inaccuracy of ϵ. The precise step is as follows:

13) Continue from step 3) considering the updated weighting parameters until the conditions (3.60) are fulfilled.

After performing the aforementioned iterative adjustment of the weighting parameters $v_{\mathrm{BS},k}$, $v_{\mathrm{MS},k}$ and $p_{k,c}$, the conditions (3.60) are fulfilled. Thus, taking into account an inaccuracy of ϵ, the ADR requirements are fulfilled by the proposed PA Tx strategy.

3.4.3 Subcarrier Allocation (SA) Transmit Strategy

In this section, the SA Tx strategy is introduced. So far, it is assumed that each mobile station simultaneously transmits M data streams on each subcarrier. However, to improve the spatial separation of the different signals at RS and thus, to reduce MS-MS-interferences, the numbers of simultaneously transmitted data streams can be optimized for each mobile station on each subcarrier. By reducing MS-MS-interferences, the achievable data rates in downlink which are the limiting data rates in case of ADR requirements with $r > 1$ can be increased. In this section, a low-complexity approach is proposed which assumes that each mobile station S_k can either transmit $m_{k,c} = M$ or $m_{k,c} = 0$ data streams simultaneously on subcarrier c. This approach is referred to as SA Tx strategy because each mobile station only transmits on the subcarriers where $m_{k,c} = M$ is selected. After performing a subcarrier allocation, the weighting parameters are adjusted considering the PA Tx strategy of Section 3.4.2.

The SA Tx strategy is illustrated in Figure 3.5. First, the Tx and Rx filters at the nodes and the transceive filter at RS are computed on each subcarrier c as described in Section 3.3 assuming $m_{k,c} = M$ and $p_{k,c} = v_{\mathrm{BS},k} = v_{\mathrm{MS},k} = 1$, $k = 2, 3, ..., K$, $c = 1, 2, ..., C$.

Secondly, the Tx power distributions at S_1 and at RS are adjusted by changing the weighting parameters $v_{\mathrm{BS},k}$ to achieve $\max\limits_k C_{1,k} < \min\limits_k C_{1,k} + \epsilon$ assuming $p_{k,c} = v_{\mathrm{MS},k} = 1$. To achieve this, step 3) to step 6) of the PA Tx strategy of Section 3.4.2 are repeated until the condition is fulfilled.

Thirdly, the subcarrier allocation is performed for each mobile station. The steps to perform the subcarrier allocation for each mobile station S_k are as follows:

1) Define a subset $\mathcal{SA}_k$ which contains the subcarriers c on which S_k transmits $m_{k,c} = M$ data streams. Furthermore, compute the achievable data rate in uplink from S_k on each subcarrier c as

$$C_{k,1,c} = \sum_{m=1}^{m_{k,c}} C_{k,m,c}. \tag{3.65}$$

2) Determine the subcarrier $c_{\min,k} = \arg \min\limits_{c \epsilon \mathcal{SA}_k} C_{k,1,c}$ on which S_k achieves the lowest data rate.

3) Compute if S_k can fulfill the ADR requirements without transmitting on subcarrier $c_{\min,k}$ by testing the condition

$$C_{1,k} < r(C_{k,1} - C_{k,1,c_{\min,k}}). \tag{3.66}$$

If condition (3.66) is fulfilled, set the number of simultaneously transmitted data streams of S_k on this subcarrier to $m_{k,c_{\min,k}} = 0$ and save the subcarrier index, i.e., $c_{\mathrm{save},k} = c_{\min,k}$. Thus, subcarrier $c_{\min,k}$ is no longer allocated to S_k which reduces the MS-MS-interferences on this subcarrier.

4) Return to step 1) considering the next mobile station until all mobile stations have been considered.

5) If the subcarrier allocation has changed for any mobile station, recalculate the relay transceive filter and the Tx and Rx filters at the nodes. Furthermore, adjust the Tx power distributions at S_1 and at RS by changing the weighting parameters $v_{\mathrm{BS},k}$ to achieve $\max_k C_{1,k} < \min_k C_{1,k} + \epsilon$ assuming $p_{k,c} = v_{\mathrm{MS},k} = 1$. To achieve this, repeat step 3) to step 6) of the PA Tx strategy of Section 3.4.2 until $\max_k C_{1,k} < \min_k C_{1,k} + \epsilon$ is fulfilled.

6) Ensure that the achievable uplink data rate of each mobile station still fulfills the condition $rC_{k,1} \geq C_{1,k}$. If $rC_{k,1} < C_{1,k}$ and not all subcarriers are allocated to S_k, set $m_{k,c_{\mathrm{save},k}} = M$.

7) Determine if the subcarrier allocation has changed for any mobile station. If the subcarrier allocation has changed, return to step 1). If the subcarrier allocation has not changed, the subcarrier allocation is finished.

Finally, if the subcarrier allocation is finished, the PA Tx strategy of Section 3.4.2 is applied to adjust the weighting parameters $v_{\mathrm{BS},k}$, $v_{\mathrm{BS},k}$ and $p_{k,c}$ considering that only part of the subcarriers are allocated to each mobile station.

3.5 Performance Analysis

In this section, the performances of the Tx strategies presented in Section 3.4 are investigated through numerical simulations considering the different filter designs presented in Section 3.3. All channels are assumed to be i.i.d. Rayleigh fading channels with zero-mean and unit variance and the noise variances at the nodes and at RS are assumed to be equal, i.e., $\sigma^2_{\mathrm{n,RS}} = \sigma^2_{\mathrm{n}}$. Furthermore, $C = 8$ orthogonal subcarriers are

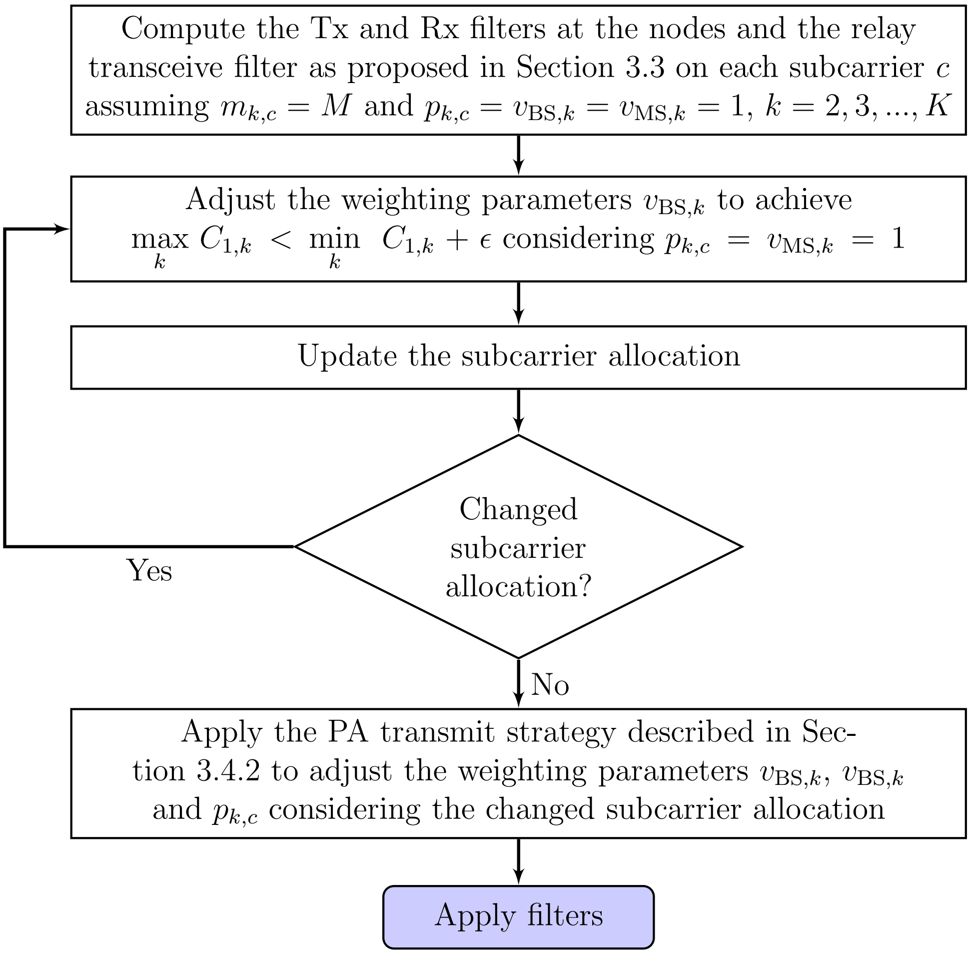

Figure 3.5. Flowchart of the SA transmit strategy

considered because this is sufficient to compare the performances of both Tx strategies due to only considering $K \leq 5$ nodes and ADR requirements r with $1 \leq r \leq 5$ for the numerical simulations. All simulation results are averaged over 1000 independent channel realizations and the maximum Tx power at RS is set to be equal to the maximum Tx power at node S_1, i.e., $P_{\mathrm{RS}} = P_{\mathrm{BS}} = (K-1)P_{\mathrm{node}}$. The ratio between the maximum Tx power P_{node} at the mobile stations and the noise level σ_{n}^2 is termed average SNR.

For the numerical simulations, two different configurations of the cellular multi-user two-way relaying scenario are investigated. In configuration A, $K = 3$ nodes are considered and each mobile station is equipped with $M = 2$ antennas. In configuration B, $K = 5$ nodes are considered and each mobile station is equipped with $M = 1$ antenna. These two configurations are investigated because the number of nodes K and the number of antennas M per node are selected such that a comparison between both scenarios is possible. In both configurations, S_1 is equipped with $M_1 = 4$ antennas and simultaneously transmits $m_{1,c} = 4$ data streams on each subcarrier to RS during the MAC phase.

For performance comparison, the approaches listed in Table 3.1 are considered. Due to the lack of state of the art approaches which consider the introduced ADR requirements, the WZF and the WMMSE relay transceive filter designs have been introduced as a straightforward extension from the state of the art in Section 3.3. These approaches are used to show the performance gain of the proposed WMMSE-SIC relay transceive filter design compared to conventional ZF or MMSE based relay transceive filter designs. Thus, the filters at the nodes are not optimized if a WZF or WMMSE filter at RS is considered, i.e., only case Diag is considered for the filter design at the nodes. To compare the performances of the different approaches versus the average SNR and versus the number L of antennas at RS, an ADR requirement of $r = 1$ is considered. For $r = 1$, the achievable data rates in up- and downlink have to be equal which enables a straightforward comparison of the different filter design approaches at the nodes and at RS. Afterwards, to compare the performances of the different Tx strategies, the performances of the approaches are compared versus different ADR requirements r assuming a fixed number L of antennas at RS and a fixed average SNR.

Name	Transmit Strategy	Filter Design at Nodes	Filter Design at RS
WZF:PA	PA of Section 3.4.2	case Diag of Section 3.3.2.1	WZF of Section 3.3.3.1
WZF:SA	SA of Section 3.4.3	case Diag of Section 3.3.2.1	WZF of Section 3.3.3.1
WMMSE:PA	PA of Section 3.4.2	case Diag of Section 3.3.2.1	WMMSE of Section 3.3.3.2
WMMSE:SA	SA of Section 3.4.3	case Diag of Section 3.3.2.1	WMMSE of Section 3.3.3.2
WMMSE-SIC:PA	PA of Section 3.4.2	case Diag of Section 3.3.2.1	WMMSE-SIC of Section 3.3.3.3
WMMSE-SIC:SA	SA of Section 3.4.3	case Diag of Section 3.3.2.1	WMMSE-SIC of Section 3.3.3.3
Rx+WMMSE-SIC:PA	PA of Section 3.4.2	case Rx of Section 3.3.2.2	WMMSE-SIC of Section 3.3.3.3
Rx+WMMSE-SIC:SA	SA of Section 3.4.3	case Rx of Section 3.3.2.2	WMMSE-SIC of Section 3.3.3.3
RxTx+WMMSE-SIC:PA	PA of Section 3.4.2	case Rx&Tx of Section 3.3.2.3	WMMSE-SIC of Section 3.3.3.3
RxTx+WMMSE-SIC:SA	SA of Section 3.4.3	case Rx&Tx of Section 3.3.2.3	WMMSE-SIC of Section 3.3.3.3

Table 3.1. Overview of the considered approaches for performance comparison.

The performances of Rx+WMMSE-SIC:PA and Rx+WMMSE-SIC:SA are omitted in some of the figures to improve the readability because the approaches perform in between WMMSE-SIC:PA and RxTx+WMMSE-SIC:PA and WMMSE-SIC:SA and RxTx+WMMSE-SIC:SA, respectively.

Figures 3.6 and 3.7 show the average achievable sum rates versus the average SNR for configuration A and configuration B, respectively, considering an ADR requirement of $r = 1$, i.e., the achievable data rates in up- and downlink have to be equal. For these simulations, $L = 8$ antennas at RS are assumed. For all approaches, the achievable sum rate increases for increasing the average SNR. Furthermore, the achievable data rates are higher in configuration B because due to considering single antenna mobile stations which have the same Tx power constraint than the multi-antenna mobile stations in configuration A, the Tx power per data stream in uplink is higher in configuration B. Moreover, the performances of the PA Tx strategy and the SA Tx strategy are similar for all considered filter designs due to an ADR requirement of $r = 1$ and due to considering $L = 8$ antennas at RS which enables that all simultaneously received signals can be spatially separated at RS.

The WZF approaches, i.e., WZF:PA and WZF:SA, perform worst due to spatially separating all simultaneously received signals at RS without considering the impact of noise. Especially for a low average SNR, the WMMSE approaches, i.e., WMMSE:PA and WMMSE:SA, perform better than the WZF approaches due to considering the impact of noise by minimizing the MSE instead of ZF all interferences. However, all approaches which consider the proposed WMMSE-SIC transceive filter at RS significantly outperform the conventional WZF and WMMSE approaches.

The third best performance is achieved by the approaches considering case Diag for the filter design at the nodes and a WMMSE-SIC relay transceive filter design, i.e., by the approaches WMMSE-SIC:PA and WMMSE-SIC:SA. Especially, for medium to high average SNR, these approaches significantly outperform the WMMSE approach due to exploiting the self-interference cancellation and SIC capabilities of the nodes for the relay transceive filter design. For low average SNR, the WMMSE-SIC and the WMMSE approaches are noise limited and thus, the performance gain of the WMMSE-SIC approaches is smaller. For configuration A, the gain of the WMMSE-SIC:PA approach compared to the WMMSE:PA approach is approximately 15% for an average SNR of 5dB and 34% for an average SNR of 15dB. For configuration B, the gain of the WMMSE-SIC:PA approach compared to the WMMSE:PA approach is approximately 17% for an average SNR of 5dB and 31% for an average SNR of 15dB.

The gain of the WMMSE-SIC approaches can be further increased by optimizing the Tx and Rx filters at the nodes. Accordingly, the second best performance is achieved by the Rx+WMMSE-SIC:PA and Rx+WMMSE-SIC:SA approaches. The best performance is achieved by the RxTx+WMMSE-SIC:PA and RxTx+WMMSE-SIC:SA approaches due to considering a joint optimization between the Tx and Rx filters at the nodes and the transceive filter at RS. The gain of additionally optimizing the Tx

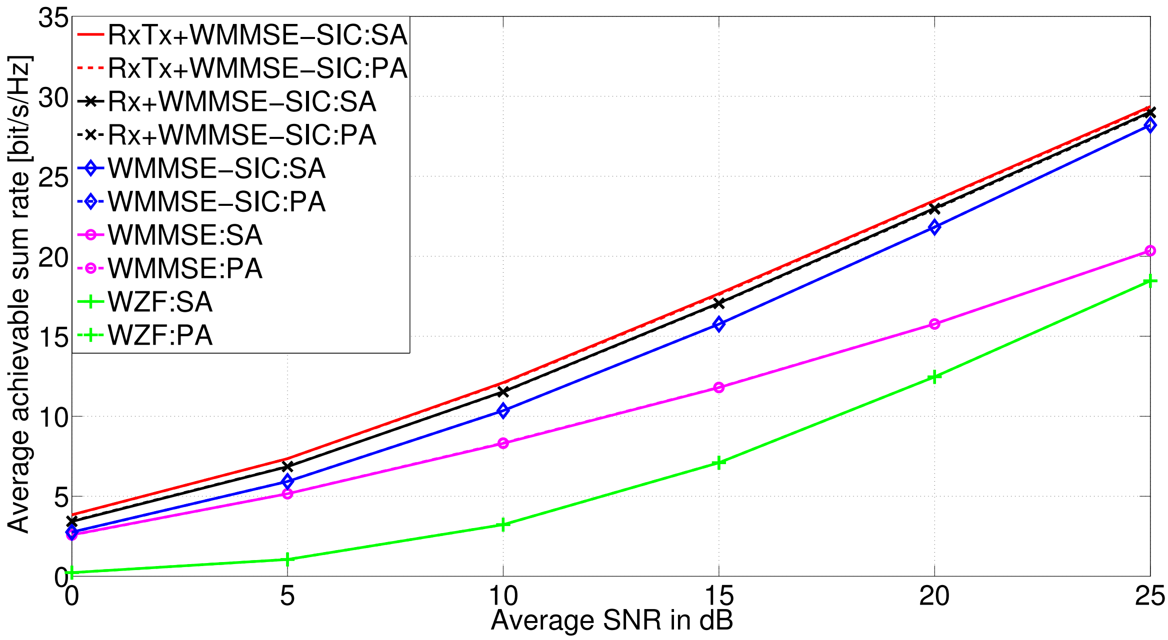

Figure 3.6. Average achievable sum rates versus average SNR, configuration A, $r = 1$, $L = 8$.

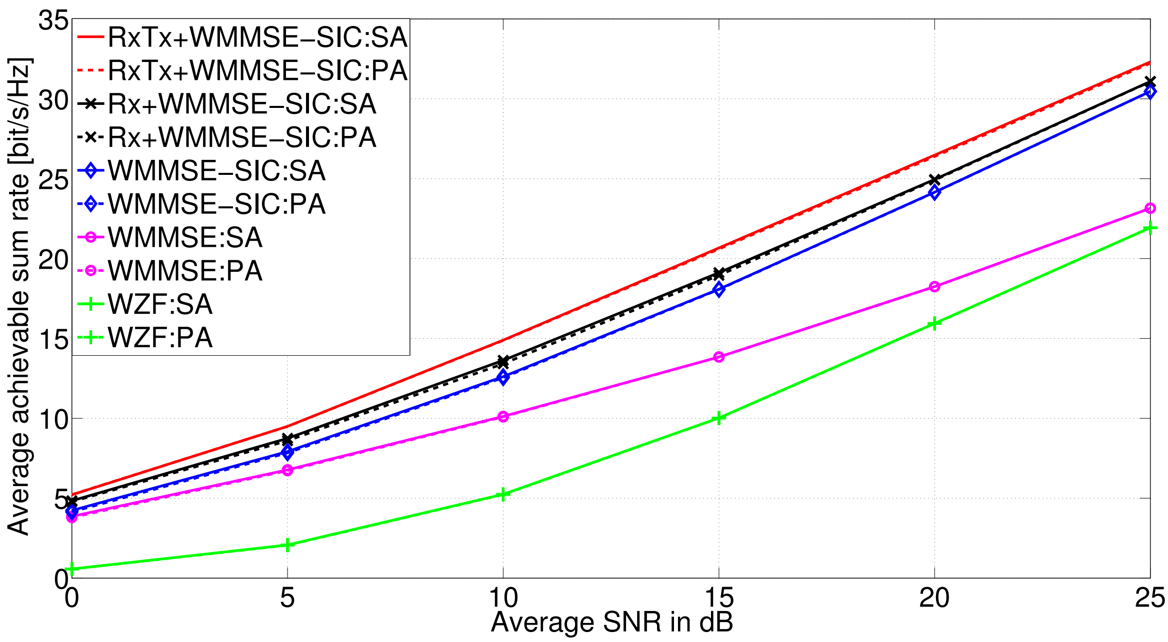

Figure 3.7. Average achievable sum rates versus average SNR, configuration B, $r = 1$, $L = 8$.

filter at S_1 is higher in configuration B than in configuration A because in configuration A spatial receive processing and SIC can be performed at the mobile stations which cannot be performed in configuration B due to considering single antenna mobile stations. For configuration A, the gain of the RxTx+WMMSE-SIC:PA approach compared to the WMMSE-SIC:PA approach is approximately 12% for an average SNR of 15dB. For configuration B, the gain of the RxTx+WMMSE-SIC:PA approach compared to the WMMSE-SIC:PA approach is approximately 14% for an average SNR of 15dB. Furthermore, for an average SNR of 15dB, the gain of the RxTx+WMMSE-SIC:PA approach compared to the WMMSE:PA approach is approximately 50% and 49% considering configuration A and B, respectively.

Figures 3.8 and 3.9 show the average achievable sum rates versus the number L of antennas at RS for configuration A and configuration B, respectively, considering an ADR requirement of $r = 1$ to compare the performances of the different filter designs at RS and at the nodes. For these simulations, the average SNR is 15dB. For all approaches, the achievable sum rate increases for increasing the number L of antennas at RS. Furthermore, for $L \geq 8$ antennas at RS the performances of the PA Tx strategy and the SA Tx strategy are similar for all considered filter designs because $L \geq 8$ antennas are required to spatially separate all simultaneously received signals at RS.

The WZF approaches, i.e., WZF:PA and WZF:SA, achieve the worst performance. For applying a WZF filter at RS, $L \geq 8$ antennas are required because all received signals have to be spatially separated. Especially for low number L of antennas at RS, the WMMSE approaches, i.e., WMMSE:PA and WMMSE:SA, perform better than the WZF approaches due to considering the impact of noise by minimizing the MSE instead of ZF all interferences. For large number $L > 10$ of antennas at RS, the WZF approaches perform similar than the WMMSE approaches because the loss in signal power compared to the noise power, termed noise enhancement, due to spatially separating all signals at RS in case of a WZF relay transceive filter decreases for increasing the number L of antennas at RS. The performance gap between the WMMSE approaches and the WZF approaches is slightly higher in configuration A than in configuration B because the Tx power per data stream in uplink is higher in configuration B and thus, the impact of the noise enhancement is smaller. In configuration A, the WMMSE:SA approach performs better than the WMMSE:PA approach for $L < 7$ antennas at RS and in configuration B, the WMMSE:SA approach performs better than the WMMSE:PA approach for $L < 8$ antennas at RS. In these cases, the number L of antennas at RS is too small to spatially separate all received signals and thus, a gain is achieved by performing the proposed subcarrier allocation which reduces the number of simultaneously transmitted data streams per subcarrier. The gain of the SA Tx strategy compared to the PA Tx strategy is higher in configuration B than in configuration A because due to considering more mobile stations in configuration B, the MS-MS-interferences are higher than in configuration A if all mobile stations are simultaneously transmitting. Nevertheless, all approaches which consider the proposed WMMSE-SIC transceive filter at RS significantly outperform the conventional WZF and WMMSE approaches due to exploiting the self-interference cancellation and SIC capabilities of the nodes for the relay transceive filter design.

Similar to the previous performance results, the third best performance is achieved by the approaches considering case Diag for the filter design at the nodes and a WMMSE-SIC relay transceive filter design, i.e., by the approaches WMMSE-SIC:PA and WMMSE-SIC:SA. Considering configuration A, the gain of WMMSE-SIC:PA compared to WMMSE:PA is approximately 34% for $L = 8$ antennas at RS and the gain of WMMSE-SIC:PA compared to WZF:PA is approximately 123% for $L = 8$ antennas at RS. Considering configuration B, the gain of WMMSE-SIC:PA compared to WMMSE:PA is approximately 31% for $L = 8$ antennas at RS and the gain of WMMSE-SIC:PA compared to WZF:PA is approximately 83% for $L = 8$ antennas at RS.

These gains can be further increased by optimizing the Tx and Rx filters at the nodes for considering the proposed WMMSE-SIC relay transceive filter. Accordingly, the second best performance is achieved by the Rx+WMMSE-SIC:PA and Rx+WMMSE-SIC:SA approaches and the best performance is achieved by the RxTx+WMMSE-SIC:PA and RxTx+WMMSE-SIC:SA approaches. The gain of optimizing the filter design at the nodes increases for decreasing the number of antennas at RS because a joint filter design between the nodes and RS enables RS to efficiently process a higher number of simultaneously received signals. Due to the same reason, the gain of the SA Tx strategy compared to the PA Tx strategy decreases from considering case Diag over case Rx to case Rx&Tx for the filter design at the nodes and a WMMSE-SIC transceive filter at RS. Furthermore, the gain of the SA Tx strategy compared to the PA Tx strategy considering a WMMSE-SIC transceive filter at RS is higher in configuration B than in configuration A because due to considering more mobile stations in configuration B, the impact of MS-MS-interference is higher than in configuration A. The gain of RxTx+WMMSE-SIC:PA compared to WMMSE-SIC:PA decreases for increasing the number L of antennas at RS, e.g., considering configuration A, the gain is approximately 35% for $L = 6$ and 12% for $L = 8$ and considering configuration B, the gain is approximately 58% for $L = 6$ and 14% for $L = 8$. Compared to the WMMSE:PA approach, the proposed RxTx+WMMSE-SIC:PA approach achieves approximately the same sum rate with three antennas less at RS due to exploiting the self-interference cancellation and the SIC capabilities of the nodes for the relay transceive filter design.

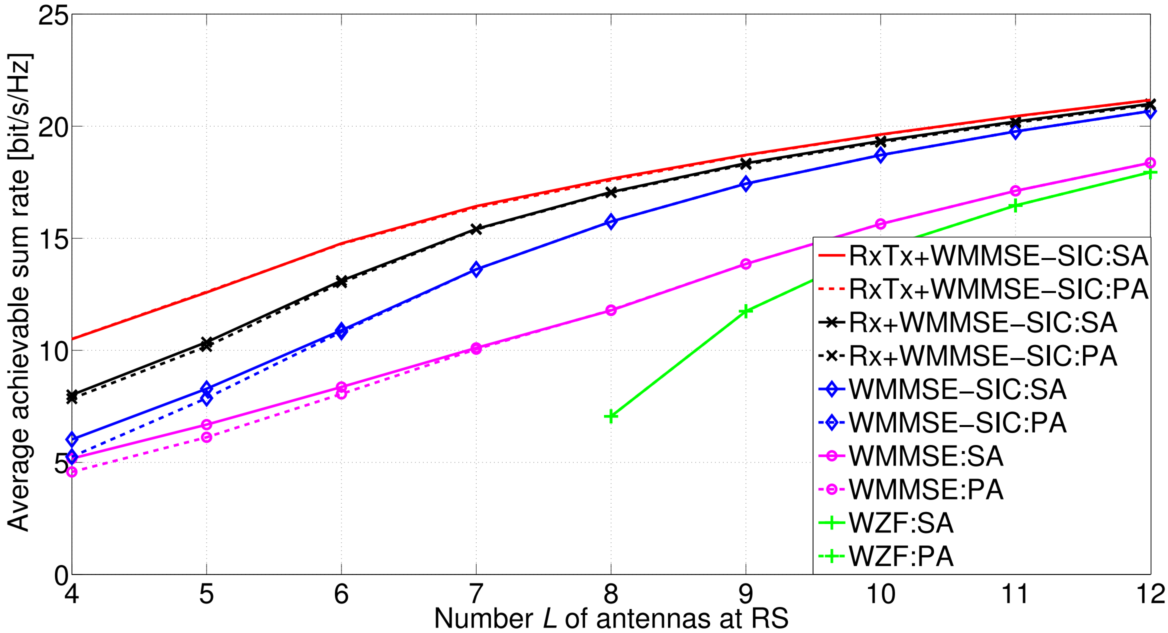

Figure 3.8. Average achievable sum rates versus number L of antennas at RS for an average SNR = 15dB, configuration A, $r = 1$.

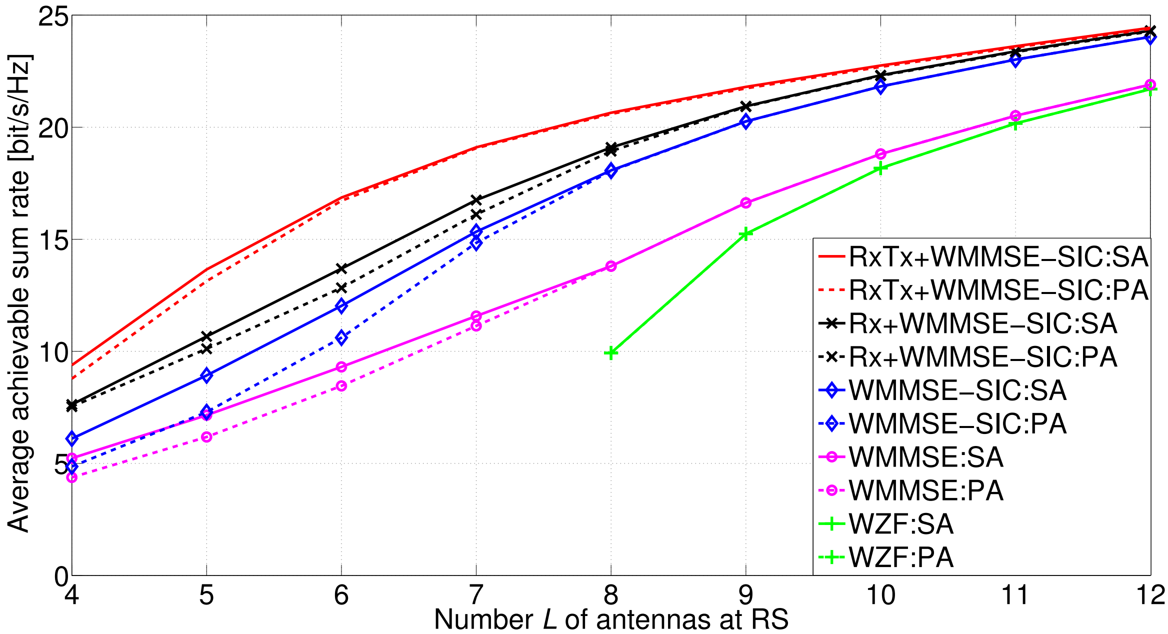

Figure 3.9. Average achievable sum rates versus number L of antennas at RS for an average SNR = 15dB, configuration B, $r = 1$.

Figures 3.10 and 3.11 show the average achievable sum rates versus the ADR requirement r for configuration A and configuration B, respectively, considering $L = 8$ antennas at RS and an average SNR of 15dB. Thus, the achievable data rates in downlink have to be r times higher than the achievable data rates in uplink. For all considered filter designs, the SA Tx strategy outperforms the PA Tx strategy and the gain increases for increasing the ADR ratio r.

The WZF:PA approach achieves the worst performance. Considering $r = 1$, the performances of WZF:PA and WZF:SA are the same. However, for WZF:SA the achievable sum rate increases for increasing r until $r \approx 3$ whereas the achievable sum rate for WZF:PA decreases. The sum rate increases for WZF:SA until $r \approx 3$ because the number of signals which are simultaneously received on each subcarrier is decreased for increasing r due to allocating less subcarriers to each mobile station. Thus, less signals have to be spatially separated at RS which improves the performance. For $r > 3$ considering WZF:SA and for WZF:PA, the sum rate decreases because the achievable data rate in downlink is limiting the sum rate. Considering configuration A, the gain of WZF:SA compared to WZF:PA is approximately 96% for $r = 3$. Similar to the previous performance results, the WMMSE:PA approach achieves higher sum rates than the WZF:PA approach due to considering the noise at RS when separating the received signals instead of ZF all interferences.

For WMMSE:SA, the achievable sum rate increases for increasing r until $r \approx 2$ due to the same reason as for WZF:SA. For $r > 3$, the performance of WZF:SA and WMMSE:SA is similar because the number of simultaneously received signals at RS on each subcarrier is small compared to the number L of antennas at RS and thus, the impact of the noise enhancement of the ZF approach is small. The approaches considering the proposed WMMSE-SIC filter design at RS outperform the other approaches. However, the gain of considering a WMMSE-SIC filter at RS decreases for increasing r compared to WMMSE:SA because the data rate in downlink is the limiting data rate for the WMMSE-SIC approaches over the entire range of $r > 1$ and thus, the achievable sum rate decreases for increasing r. Furthermore, the gain of the SA Tx strategy compared to the PA Tx strategy is smaller for considering a WMMSE-SIC filter at RS than for considering a WMMSE or a WZF filter at RS because due to exploiting the self-interference cancellation capabilities and SIC at the nodes, the suppression of MS-MS-interferences is improved in case of a WMMSE-SIC filter design at RS.

Nevertheless, RxTx+WMMSE-SIC:SA achieves the best performance. Considering configuration A, the gain of RxTx+WMMSE-SIC:SA compared to RxTx+WMMSE-SIC:PA is in between $3\% - 4\%$ for $2 \leq r \leq 5$ and considering configuration B, the gain of RxTx+WMMSE-SIC:SA compared to RxTx+WMMSE-SIC:PA is in between

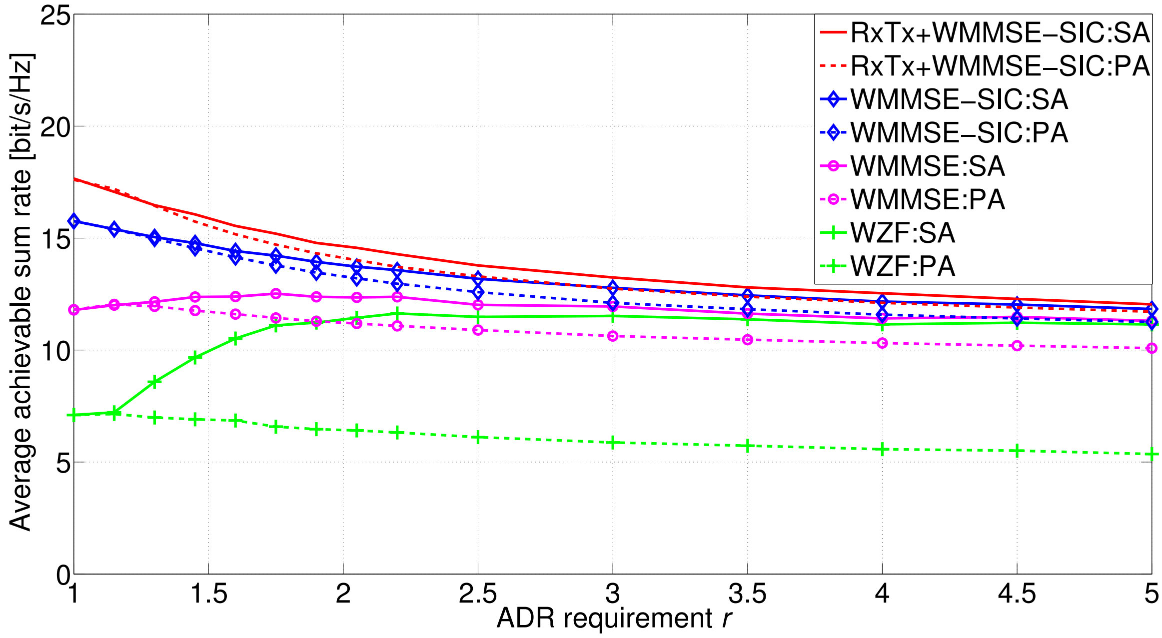

Figure 3.10. Average achievable sum rates versus ADR requirement r for an average SNR = 15dB, configuration A, $L = 8$.

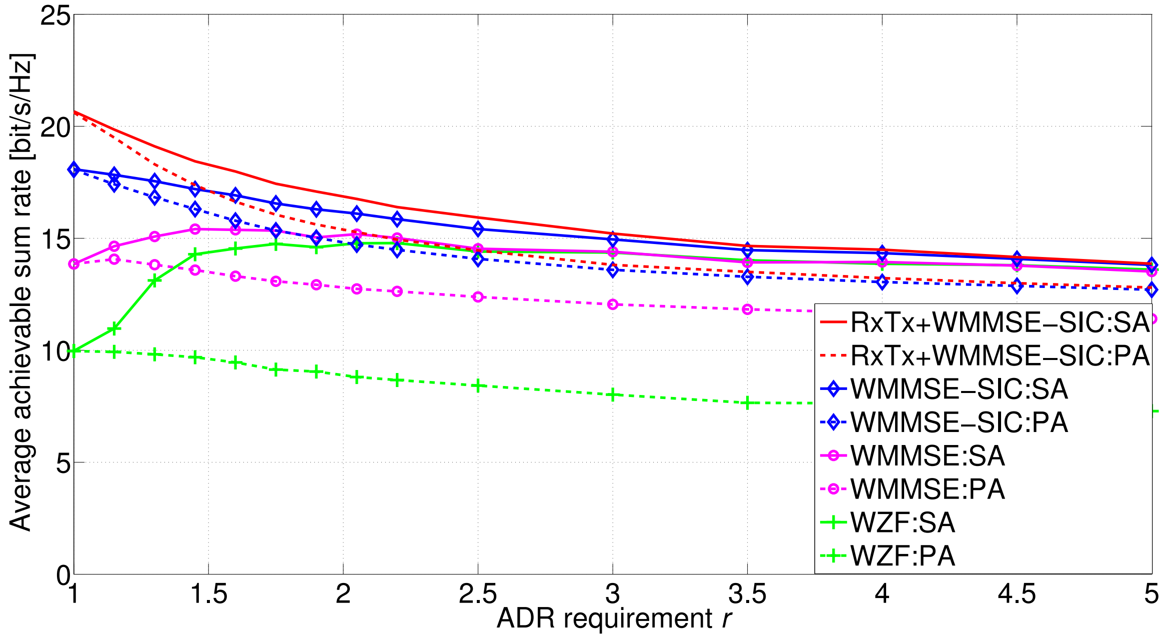

Figure 3.11. Average achievable sum rates versus ADR requirement r for an average SNR = 15dB, configuration B, $L = 8$.

8%−10% for $2 \leq r \leq 5$. The gain is larger in configuration B because due to considering more mobile stations, the impact of MS-MS-interferences on the achievable sum rate is increased. Considering configuration B, the gain of RxTx+WMMSE-SIC:SA compared to WZF:SA is approximately 10% for $r = 2.5$ and the gain of RxTx+WMMSE-SIC:SA compared to WZF:PA is approximately 89% for $r = 2.5$.

To summarize, the proposed RxTx+WMMSE-SIC:SA approach significantly outperforms all other approaches. An overview of selected performance gains of the RxTx+WMMSE-SIC:SA approach compared to approaches using a conventional WMMSE relay transceive filter design is presented in Table 3.2 considering configuration A and B. For the proposed RxTx+WMMSE-SIC:SA approach, the transceive filter at RS exploits the self-interference cancellation and the SIC capabilities of the nodes. Furthermore, the Tx and Rx filters at the nodes are designed jointly with the transceive filter at RS using an alternating optimization approach. Moreover, the SA Tx strategy performs a combined subcarrier allocation and weighting parameter adjustment to tackle the ADR requirements. This significantly increases the achievable sum rate compared to conventional MMSE or ZF based approaches. Additionally, the proposed WMMSE-SIC based relay transceive filter design requires less antennas at RS to achieve the same performance than considering a conventional MMSE or ZF based relay transceive filter design.

Table 3.2. Selected performance gains of the proposed RxTx+WMMSE-SIC:SA approach.

Config.	SNR	L	r	Conv. Approach	Proposed Approach	Perf. Gain
A	5dB	8	1	WMMSE:SA	RxTx+WMMSE-SIC:SA	43%
A	15dB	8	1	WMMSE:SA	RxTx+WMMSE-SIC:SA	50%
A	15dB	6	1	WMMSE:SA	RxTx+WMMSE-SIC:SA	77%
A	15dB	8	2.5	WMMSE:SA	RxTx+WMMSE-SIC:SA	15%
A	15dB	8	2.5	WMMSE:PA	RxTx+WMMSE-SIC:SA	26%
B	5dB	8	1	WMMSE:SA	RxTx+WMMSE-SIC:SA	40%
B	15dB	8	1	WMMSE:SA	RxTx+WMMSE-SIC:SA	49%
B	15dB	6	1	WMMSE:SA	RxTx+WMMSE-SIC:SA	81%
B	15dB	8	2.5	WMMSE:SA	RxTx+WMMSE-SIC:SA	10%
B	15dB	8	2.5	WMMSE:PA	RxTx+WMMSE-SIC:SA	29%

Chapter 4

Multi-Pair Two-Way Relaying

4.1 Problem Overview and Decomposition

In this chapter, the multi-pair two-way relaying scenario as shown in Figure 2.2 is investigated. To investigate this scenario, a system model for multi-pair two-way relaying considering multi-antenna nodes which can perform self-interference cancellation, linear receive processing and SIC is introduced. In such a multi-pair scenario, the required data rate for the transmission from S_{2i+1} to S_{2i+2} is typically different compared to the required data rate for the transmission from S_{2i+2} to S_{2i+1}, $i = 0, 1, ..., K/2 - 1$. Furthermore, the required data rates of the different pairs are typically different as well. Thus, in this thesis, ADR requirements are considered within each pair and among different pairs.

To maximize the sum rate under the aforementioned ADR requirements, the Tx and Rx filters of the nodes, the Tx powers of the nodes, the relay transceive filter and the numbers of simultaneously transmitted data streams of the nodes have to be jointly optimized. Due to the high computational complexity of finding an optimal solution for this problem, suboptimal approaches based on a problem decomposition are proposed in this chapter. To obtain such suboptimal approaches which fulfill the aforementioned ADR requirements whilst achieving high sum rates, the following steps are proposed:

1. It is proposed to decouple the overall problem into three different subproblems as shown in Figure 4.1. The considered subproblems are the design of a Tx strategy, the design of the relay transceive filter and the design of the Tx and Rx filters at the nodes.

2. To tackle the ADR requirements, it is proposed to couple the filter design at the nodes and at RS with the design of the Tx strategy by introducing the following weighting parameters:

   - w_k: To adjust the fraction of the Tx power used at RS to perform transmission from S_k to S_l, $0 \leq w_k \leq 1$, $k = 1, 2, ..., K$, $l = k - 1 + 2 \cdot \mathrm{mod}_2 k$,
   - p_k: To adjust the Tx power of S_k, $0 \leq p_k \leq 1$.

These weighting parameters are considered for the Tx filter design at the nodes and for the relay transceive filter design. By these weighting parameters, the Tx powers of the nodes and the Tx power distribution at RS are adjusted via the Tx strategy.

3. It is proposed to focus on low-complexity solutions for the different subproblems.

Based on these steps, suboptimal low-complexity approaches for the different subproblems are proposed as shown in Figure 4.1.

For the design of a Tx strategy which fulfills the ADR requirements whilst achieving high sum rates, two different approaches are proposed. The power adapted Tx strategy considers that each node either transmits M data streams or one data stream. Based on this, the Tx powers of the nodes and the Tx power distribution at RS are adjusted via the aforementioned weighting parameters. Thus, an optimization of the numbers of simultaneously transmitted data streams of the nodes is not considered for this Tx strategy. The optimized streams Tx strategy is an extension of the power adapted Tx strategy because it optimizes the numbers of simultaneously transmitted data streams of the nodes by performing an exhaustive search. Additionally, the Tx powers of the nodes and the Tx power distribution at RS are adjusted similar to the power adapted Tx strategy via the aforementioned weighting parameters.

For the relay transceive filter design, three different approaches are investigated. For comparison, a weighted ZF and a weighted MMSE approach are considered. The weighted ZF and the weighted MMSE approach are straightforward extensions of the state of the art to tackle the ADR requirements by considering the aforementioned weighting parameters in the relay transceive filter design. In addition to these approaches, a weighted self-interference cancellation and SIC aware relay transceive filter is proposed. To obtain an analytical solution for the relay transceive filter design which ensures high sum rates whilst considering the ADR requirements, a weighted MMSE based approach is proposed. The proposed relay transceive filter design exploits the capability of the nodes to perform self-interference cancellation and SIC. The proposed relay transceive filter depends on the Tx and Rx filters at the nodes.

For the Tx and Rx filter design at the nodes, two different approaches are proposed. For the local Tx and Rx filter design, only the channel between the node and RS is considered and thus, the impact of the filter design on the performances of other nodes is neglected. For the global Tx and Rx filter design, the impact of the filter design on the performances of other nodes is considered by taking all channels between

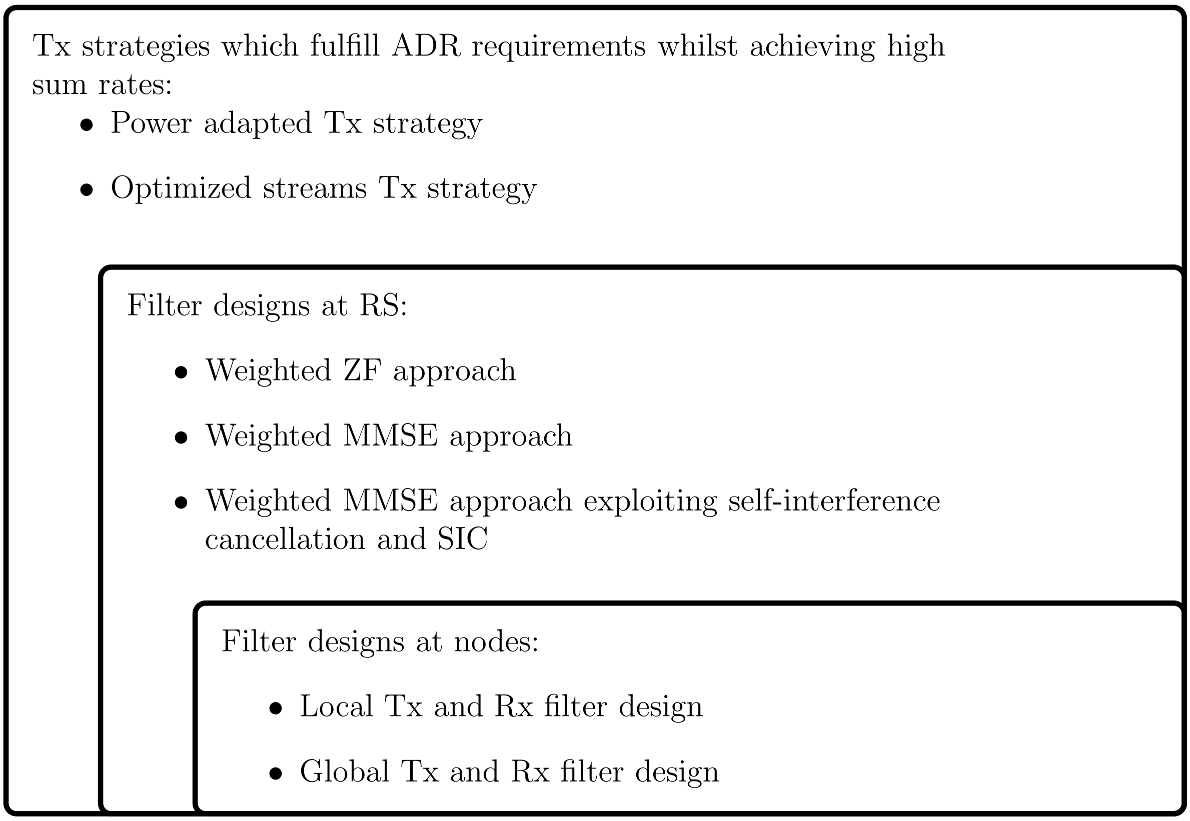

Figure 4.1. Overview of the proposed and investigated approaches for multi-pair two-way relaying.

the nodes and RS into account for designing the Tx and Rx filters at each node. For both approaches, the Tx and Rx filters at the nodes are designed independent of the relay transceive filter to reduce the computational complexity. Furthermore, the aforementioned weighting parameters p_k are considered for the Tx filter design at the nodes to enable an adjustment of the Tx powers of the nodes with respect to the considered ADR requirements.

The proposed relay transceive filter design depends on the Tx and Rx filters at the nodes. Furthermore, the proposed Tx strategies are based on the relay transceive filter design and on the filter design at the nodes. Moreover, the relay transceive filter and the filters at the nodes depend on the weighting parameters and on the numbers of simultaneously transmitted data streams which are computed based on the Tx strategies. Thus, the computation of the different filters, the weighting parameters and the numbers of simultaneously transmitted data streams is performed as shown in Figure 3.2. First, all weighting parameters are assumed to be one, i.e., no weighting is considered. Furthermore, all nodes are assumed to either transmit one data streams or M data streams. Based on these assumptions, the Tx and Rx filters at the nodes are computed according to the proposed local or global the Tx and Rx filter design. Secondly, the relay transceive filter is computed considering the Tx and Rx filters at the nodes of the previous step. Thirdly, the weighting parameters which are considered for the Tx filter design at the nodes and for the relay transceive filter design are adjusted. Furthermore, if the optimized streams Tx strategy is considered, an optimization of the numbers of simultaneously transmitted data streams is performed. To adjust the weighting parameters and to perform an optimization of the numbers of simultaneously transmitted data streams, the relay transceive filter and the Tx and Rx filters at the nodes have to be updated after each step. Finally, weighting parameters and numbers of simultaneously transmitted data streams which fulfill the ADR requirements whilst achieving high sum rates are selected.

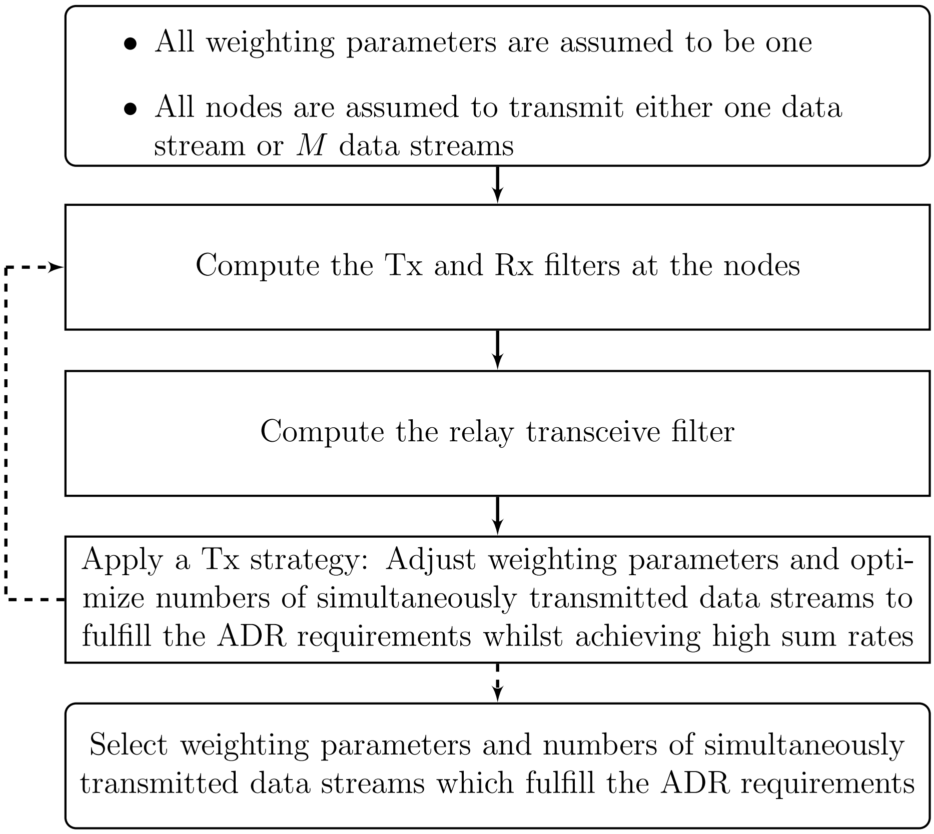

Figure 4.2. Flow chart for the computation of the filters at the nodes and at RS considering the proposed Tx strategies.

The rest of the chapter is organized as follows. In Section 4.2, the system model for the considered multi-pair two-way relaying scenario is presented. In Section 4.3, different filter design approaches for the design of the Tx and Rx filters at the nodes and for the design of the relay transceive filter are proposed. In Section 4.4, different low-complexity transmit strategies are proposed to tackle the ADR requirements and in Section 4.5, the performance of the proposed approaches is investigated by numerical results. Several parts of this chapter have been originally published by the author in [DK12a,DK12c]. Compared to [DK12a,DK12c], the system model and the filter designs are extended to consider and to exploit SIC at the nodes, respectively. Furthermore, the Tx strategies are presented in more detail. In addition, a pilot transmission scheme and a robust self-interference cancellation aware relay transceive filter design for multi-pair two-way relaying with single antenna nodes has been published by the author in [DHK12].

4.2 System Model

In this section, the system model for the considered multi-pair two-way relaying scenario as shown in Figure 4.3 is presented. As described in Section 2.1, the scenario consists of K half-duplex multi-antenna nodes S_k which perform pairwise bidirectional communications via RS, $k = 1, 2, ..., K$. As shown in Figure 2.2, S_1 and S_2, S_3 and S_4,...,S_k and S_l form bidirectional communication pairs, $l = k - 1 + 2 \cdot \mathrm{mod}_2 k$. For this scenario, only a single subcarrier is considered, i.e., $C = 1$, because the consideration of multiple subcarriers has no impact on the performance of the proposed Tx strategies. In the following, the subcarrier index c is omitted.

In the MAC phase, all nodes simultaneously transmit to RS and the superposition of these transmit signals is received at RS. Before the transmission, the Tx signal vector $\mathbf{s}_k \in \mathbb{C}^{m_k \times 1}$ of node S_k is filtered by the Tx filter $\mathbf{Q}_k \in \mathbb{C}^{M \times m_k}$, with $||\mathbf{Q}_k||_\mathrm{F}^2 \leq P_\mathrm{Node}$. Let $\mathbf{n}_\mathrm{RS} \in \mathbb{C}^{L \times 1}$ represent the complex white Gaussian noise vector at RS and let $\mathbf{H}_k \in \mathbb{C}^{L \times M}$ denote the channel from S_k to RS. Now, the received signal at RS can be written as

$$\mathbf{y}_\mathrm{RS} = \sum_{k=1}^{K} \mathbf{H}_k \mathbf{Q}_k s_k + \mathbf{n}_\mathrm{RS}. \tag{4.1}$$

In the BC phase, RS retransmits a linearly processed version of the superimposed received signals back to the nodes. The received signal $\mathbf{y}_\mathrm{RS}$ is linearly processed at RS using the transceive filter matrix $\mathbf{G} \in \mathbb{C}^{L \times L}$. Using the receive filter $\mathbf{D}_k \in \mathbb{C}^{m_l \times M}$, $l = k - 1 + 2 \cdot \mathrm{mod}_2 k$, the received signal at S_k is given by

$$\mathbf{y}_k = \mathbf{D}_k(\mathbf{H}_k^\mathrm{T} \mathbf{G} \mathbf{y}_\mathrm{RS} + \mathbf{n}_k), \tag{4.2}$$

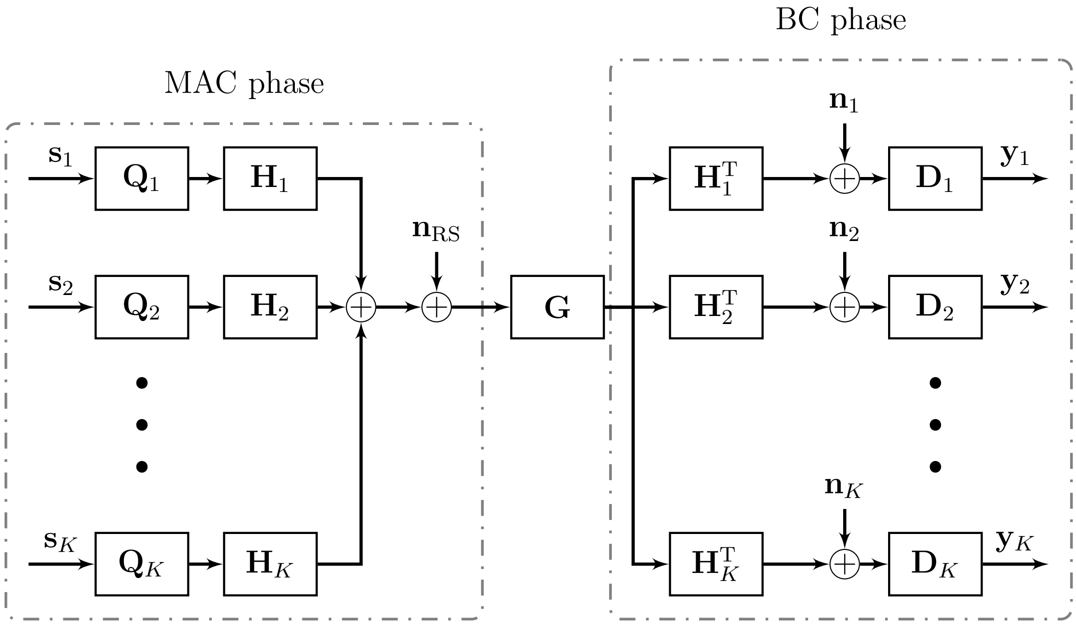

Figure 4.3. System model for multi-pair two-way relaying.

where $\mathbf{n}_k \in \mathbb{C}^{M\times 1}$ represents the complex white Gaussian noise vector at S_k.

The received useful signals and interferences at the nodes in the BC phase are illustrated in Figure 4.4. Each node S_k receives a useful signal from S_l, interferences from the signals intended for nodes of other pairs, termed inter-pair-interference, and back-propagated self-interference as well as noise.

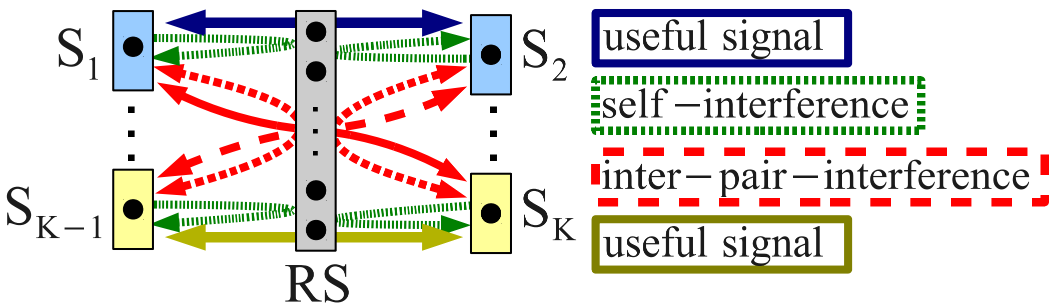

Figure 4.4. Compositions of the received signals out of useful signals and interferences in a bidirectional multi-pair two-way relaying scenario.

To simplify the notation, the overall channel coefficient for the transmission of the m^{th} data stream from S_l to S_k, $0 < m \leq m_l$, can be written as

$$h_{\mathrm{ov},l,k,m} = \mathbf{d}_{k,m}\mathbf{H}_k^{\mathrm{T}}\mathbf{G}\mathbf{H}_l\mathbf{q}_{l,m}, \tag{4.3}$$

where $\mathbf{d}_{k,m}$ is the m^{th} row vector of $\mathbf{D}_k$ and $\mathbf{q}_{l,m}$ is the m^{th} column vector of $\mathbf{Q}_l$.

As mentioned before, perfect self-interference cancellation and perfect SIC are assumed at all nodes. To exploit the SIC capabilities of the nodes for the relay transceive filter design, a fixed decoding order is required. This decoding order has to be independent of the considered filters at the nodes and at RS. Thus, it is proposed that the data streams transmitted by S_l are decoded at S_k in increasing order of the corresponding indices, i.e., the data stream $m = 1$ is decoded first and the data stream $m = m_l$ is decoded last. Writing the estimate of $\mathbf{s}_l$ as $\hat{\mathbf{s}}_l$, the signal at S_k for estimating the m^{th} data stream from S_l with $m = 1, 2, ..., m_l$ after self-interference cancellation and SIC can be written as

$$\hat{s}_{l,m} = \mathbf{d}_{k,m}\mathbf{H}_k^{\mathrm{T}}\mathbf{G}\left(\mathbf{H}_l\sum_{i=m}^{m_l}\mathbf{q}_{l,i}s_{l,i} + \sum_{j=1,j\neq k,l}^{K}\mathbf{H}_j\mathbf{Q}_j\mathbf{s}_j + \mathbf{n}_{\mathrm{RS}}\right) + \mathbf{d}_{k,m}\mathbf{n}_k, \tag{4.4}$$

where $s_{l,m}$ is the m^{th} element of $\mathbf{s}_l$ and $\hat{s}_{l,m}$ is the m^{th} element of $\hat{\mathbf{s}}_l$. Due to considering perfect self-interference cancellation and SIC, the terms $\mathbf{d}_{k,m}\mathbf{H}_k^{\mathrm{T}}\mathbf{G}\mathbf{H}_k\mathbf{Q}_k\mathbf{s}_k$ and $\sum_{i=1}^{m-1}\mathbf{d}_{k,m}\mathbf{H}_k^{\mathrm{T}}\mathbf{G}\mathbf{H}_l\mathbf{q}_{l,i}s_{l,i}$ are neglected in (4.4), respectively.

Considering perfect self-interference cancellation and SIC, the expected signal, interference and noise powers when estimating the m^{th} data stream of S_l at S_k can be written as

$$P_{\mathrm{S},l,k,m} = |h_{\mathrm{ov},l,k,m}|^2, \tag{4.5a}$$

$$P_{\mathrm{I},l,k,m} = \sum_{j=1,j\neq k}^{K}||\mathbf{d}_{k,m}\mathbf{H}_k^{\mathrm{T}}\mathbf{G}\mathbf{H}_j\mathbf{Q}_j||_2^2 - \sum_{i=1}^{m}|\mathbf{d}_{k,m}\mathbf{H}_k^{\mathrm{T}}\mathbf{G}\mathbf{H}_l\mathbf{q}_{l,i}|^2, \tag{4.5b}$$

$$P_{\mathrm{N},l,k,m} = \sigma_{\mathrm{n,RS}}^2||\mathbf{d}_{k,l,m}\mathbf{H}_k^{\mathrm{T}}\mathbf{G}||_2^2 + \sigma_{\mathrm{n}}^2||\mathbf{d}_{k,l,m}||_2^2, \tag{4.5c}$$

respectively.

For performance comparison in Section 4.5, the achievable sum rate, cf. [Ung09, Ama11, DUK11, DK12b], is considered and the corresponding equations are presented in the following. Assuming that Gaussian codebooks are used for each data stream, the maximum achievable data rate after linear receive processing and SIC for the m^{th} data stream from S_l to S_k is given by

$$C_{l,k,m} = \frac{1}{2}\log_2(1 + P_{\mathrm{S},l,k,m}(P_{\mathrm{I},l,k,m} + P_{\mathrm{N},l,k,m})^{-1}), \tag{4.6}$$

where the factor $\frac{1}{2}$ is used due to the fact that two time slots are required to perform all transmissions. The maximum achievable data rate for the transmission from S_l to

S_k is given by

$$C_l = \sum_{m=1}^{m_l} C_{l,k,m}. \tag{4.7}$$

Thus, the achievable sum rate of the system is given by

$$C_{\text{sum}} = \sum_{l=1}^{K} C_l. \tag{4.8}$$

As mentioned in Section 4.1, ADR requirements shall be considered. For simplicity of the notation, it is assumed that the instantaneous data rate required for the transmission from S_k has to be r_k times higher than the instantaneous data rate required for the transmission from S_1 which is used as a reference data rate, $r_k \geq 0$, $k = 1, 2, ..., K$. Thus, the constraint

$$r_k = C_k / C_1, k = 1, 2, ..., K, \tag{4.9}$$

with $r_1 = 1$ has to be fulfilled for the sum rate under ADR requirements. To fulfill this constraint, the lowest maximum achievable data rate from any node S_k divided by r_k is the limiting data rate. Thus, considering the constraint (4.9), the achievable sum rate under the ADR requirements is given by

$$C_{\text{ADR,sum}} = \min_k \frac{C_k}{r_k} \sum_{i=1}^{K} r_i. \tag{4.10}$$

4.3 Filter Design

4.3.1 Introduction

In this section, two low-complexity approaches for designing the Tx and Rx filters at the nodes are proposed as described in Section 4.1. Furthermore, three low-complexity approaches for designing the relay transceive filter are proposed as described in Section 4.1 assuming predefined Tx and Rx filters at the nodes. As mentioned in Section 4.1, weighting parameters are considered for the Tx filter design at the nodes and for the relay transceive filter design to enable an adjustment of these filters via the proposed Tx strategies which are introduced in Section 4.4. Utilizing these weighting parameters, the proposed Tx strategies perform an adjustment of the Tx powers of the nodes and of the Tx power distribution at RS to fulfill the ADR requirements.

As mentioned in Section 4.1, the weighting parameter p_k, $0 \leq p_k \leq 1$, is used to adapt the Tx power of S_k, $k = 1, 2, ..., K$. Furthermore, to adapt the fraction of the Tx power at RS which is assigned to the transmission of $\mathbf{s}_k$ from S_k to S_l, the weighting parameter w_k, $0 \leq w_k \leq 1$, is used, $l = k - 1 + 2 \cdot \mathrm{mod}_2 k$. Additionally, a diagonal weighting matrix $\mathbf{W}$ which contains the weighting parameters w_k is defined as

$$\mathbf{W} = \mathrm{diag}\left[\left[\mathrm{diag}^{-1}\left(w_1 \otimes \mathbf{I}_{m_1}\right), \mathrm{diag}^{-1}\left(w_2 \otimes \mathbf{I}_{m_2}\right), ..., \mathrm{diag}^{-1}\left(w_K \otimes \mathbf{I}_{m_K}\right)\right]^{\mathrm{T}}\right]. \quad (4.11)$$

Additionally, the numbers m_k of simultaneously transmitted data streams are considered as optimization parameters for the filter design.

For the filter design in Section 4.3.2 and Section 4.3.3, the parameters w_k and p_k as well as the numbers m_k of simultaneously transmitted data streams are assumed to be given. The optimization of the different parameters is described in Section 4.4.

4.3.2 Transmit and Receive Filter Design at Nodes

4.3.2.1 Local Transmit and Receive Filter Design

The local Tx and Rx filter design at each node is based on the channel between the node and RS. Thus, the knowledge of local CSI is sufficient for the filter design. This case has also been investigated in other publications, e.g., [JS10]. Each node selects the strongest singular vectors of its channel for transmission and reception. Considering the SVD of the channel $\mathbf{H}_k = \mathbf{U}_k \mathbf{\Sigma}_k \mathbf{V}_k^{\mathrm{H}}$, the Tx and Rx filters are given by

$$\mathbf{Q}_k = p_k \sqrt{\frac{P_{\mathrm{node}}}{m_k}} \mathbf{V}_{k,1:m_k}, \quad (4.12a)$$

$$\mathbf{D}_k = \mathbf{V}_{k,1:m_l}^{\mathrm{T}}, \quad (4.12b)$$

respectively, where $\mathbf{V}_{k,1:m_k}$ contains the m_k singular vectors of $\mathbf{V}_k$ which correspond to the m_k strongest singular values of $\mathbf{\Sigma}_k$.

4.3.2.2 Global Transmit and Receive Filter Design

The global Tx and Rx filter design at each node is based on taking all channels between the nodes and RS into account. To achieve this, the filter design at each node is performed based on spatial Tx and Rx subchannels allocated to each pair by RS. The subchannels are obtained by multiplying the channels between the nodes and RS with spatial filters which are designed to reduce inter-pair interferences by taking all channels between the nodes and RS into account. To obtain the Tx and Rx subchannels for each pair, an iterative optimization is performed. This iterative optimization can be separated into four steps.

First, the Tx and Rx filters at the nodes are computed as described in Section 4.3.2.1. Thus, the Tx and Rx filters are given by (4.12a) and (4.12b), respectively.

Secondly, Tx and Rx subchannels are determined for each pair. To compute the Tx and Rx subchannels for each pair, an MMSE extension of the ZF block-diagonalization (ZFBD) idea is proposed. ZFBD approaches have been introduced for downlink spatial multiplexing in [SSH04] and have been considered for multi-user two-way relaying in [YZGK10, AK10b, LDLG11]. Using ZFBD, the interferences caused by the other pairs are forced to zero in each subchannel. To reduce the noise enhancement compared to ZFBD approaches, it is proposed to allow some inter-pair interferences in each spatial subchannel according to the MMSE principle instead of completely suppressing inter-pair interferences. To compute the Tx subchannels based on the MMSE principle, let $\tilde{\mathbf{H}}_{\mathrm{Tx,j}}$ denote the transmit channel matrix of all nodes not belonging to the j^{th} pair with $j = 1, 2, ..., K/2$. Thus, $\tilde{\mathbf{H}}_{\mathrm{Tx,j}}$ can be written as

$$\tilde{\mathbf{H}}_{\mathrm{Tx},\mathbf{j}} = [\mathbf{H}_1\mathbf{Q}_1, \mathbf{H}_2\mathbf{Q}_2, ..., \mathbf{H}_{2j-2}\mathbf{Q}_{2j-2}, \mathbf{H}_{2j+1}\mathbf{Q}_{2j+1}, ..., \mathbf{H}_K\mathbf{Q}_K], \tag{4.13}$$

where the Tx filters of (4.12a) are used at the nodes. The spatial Tx subchannel of S_k is based on the SVD of $\tilde{\mathbf{H}}_{\mathrm{Tx},j}$ with $j = \lceil \frac{k}{2} \rceil$. Let the SVD of $\tilde{\mathbf{H}}_{\mathrm{Tx},j}$ be given by $\tilde{\mathbf{H}}_{\mathrm{Tx},j} = \tilde{\mathbf{U}}_{\mathrm{Tx},\mathbf{j}} \tilde{\mathbf{\Sigma}}_{\mathrm{Tx},\mathbf{j}} \tilde{\mathbf{V}}^{\mathrm{H}}_{\mathrm{Tx},\mathbf{j}}$. Now, the MMSE based Tx subchannel of S_k can be written as

$$\mathbf{H}_{\mathrm{Tx},k} = \left(\tilde{\mathbf{U}}_{\mathrm{Tx},j}\mathbf{D}_{\mathrm{Tx},j}\right)^H \mathbf{H}_k, \tag{4.14a}$$

$$\text{with } \mathbf{D}_{\mathrm{Tx},j} = \left(\tilde{\mathbf{\Sigma}}_{\mathrm{Tx},j}\tilde{\mathbf{\Sigma}}^{\mathrm{T}}_{\mathrm{Tx},j} + \frac{\sigma^2_{\mathrm{n,RS}}}{P_{\mathrm{node}}}\mathbf{I}_L\right)^{-\frac{1}{2}}. \tag{4.14b}$$

To compute the Rx subchannels based on the MMSE principle, let $\tilde{\mathbf{H}}_{\mathrm{Rx},\mathbf{j}}$ denote the Rx channel matrix of all nodes not belonging to the jth pair which is given by

$$\tilde{\mathbf{H}}_{\mathrm{Rx},\mathbf{j}} = [\mathbf{D}_1\mathbf{H}_1^{\mathrm{T}}, \mathbf{D}_2\mathbf{H}_2^{\mathrm{T}}, ..., \mathbf{D}_{2j-2}\mathbf{H}_{2j-2}^{\mathrm{T}}, \mathbf{D}_{2j+1}\mathbf{H}_{2j+1}^{\mathrm{T}}, ..., \mathbf{D}_{2K}\mathbf{H}_{2K}^{\mathrm{T}}], \tag{4.15}$$

where the Rx filters of (4.12b) are used at the nodes. The spatial Rx subchannel of S_k is based on the SVD of $\tilde{\mathbf{H}}_{\mathrm{Rx},j}$ with $j = \lceil \frac{k}{2} \rceil$. Let the SVD of $\tilde{\mathbf{H}}_{\mathrm{Rx},j}$ be given by $\tilde{\mathbf{H}}_{\mathrm{Rx},j} = \tilde{\mathbf{U}}_{\mathrm{Rx},\mathbf{j}} \tilde{\mathbf{\Sigma}}_{\mathrm{Rx},\mathbf{j}} \tilde{\mathbf{V}}^{\mathrm{H}}_{\mathrm{Rx},\mathbf{j}}$. Now, the MMSE based Rx subchannel of S_k can be written as

$$\mathbf{H}_{\mathrm{Rx},k} = \mathbf{H}_k^T \tilde{\mathbf{V}}_{\mathrm{Rx},j}\mathbf{D}_{\mathrm{Rx},j}, \tag{4.16a}$$

$$\text{with } \mathbf{D}_{\mathrm{Rx},j} = \left(\tilde{\mathbf{\Sigma}}^{\mathrm{T}}_{\mathrm{Rx},j}\tilde{\mathbf{\Sigma}}_{\mathrm{Rx},j} + \frac{\sigma^2_{\mathrm{n}}}{P_{\mathrm{RS}}}\mathbf{I}_L\right)^{-\frac{1}{2}}. \tag{4.16b}$$

Thirdly, the Tx and Rx filters of each node are recomputed based on the corresponding Tx and Rx subchannels, respectively. Let the SVD of $\mathbf{H}_{\mathrm{Tx},k}$ be given by $\mathbf{H}_{\mathrm{Tx},k} = \mathbf{U}_{\mathrm{Tx},k}\mathbf{\Sigma}_{\mathrm{Tx},k}\mathbf{V}^{\mathrm{H}}_{\mathrm{Tx},k}$ and let the SVD of $\mathbf{H}_{\mathrm{Rx},k}$ be given by $\mathbf{H}_{\mathrm{Rx},k} = \mathbf{U}_{\mathrm{Rx},k}\mathbf{\Sigma}_{\mathrm{Rx},k}\mathbf{V}^{\mathrm{H}}_{\mathrm{Rx},k}$. Furthermore, let each node select the m_k strongest singular vectors of its Tx subchannel for transmission and let each node select the m_l strongest singular vectors of its Rx subchannel for reception. Thus, using the aforementioned notation, the Tx and Rx filters at S_k are recomputed according to

$$\mathbf{Q}_k = p_k\sqrt{\frac{P_{\mathrm{node}}}{m_k}}\mathbf{V}_{\mathrm{Tx},k,1:m_k}, \tag{4.17a}$$

$$\mathbf{D}_k = \mathbf{U}^{\mathrm{H}}_{\mathrm{Rx},k,1:m_l}, \tag{4.17b}$$

respectively, where $\mathbf{V}_{\mathrm{Tx},k,1:m_k}$ contains the m_k singular vectors of $\mathbf{V}_{\mathrm{Tx},k}$ which correspond to the m_k strongest singular values of $\mathbf{\Sigma}_{\mathrm{Tx},k}$ and $\mathbf{U}_{\mathrm{Rx},k,1:m_l}$ contains the m_l singular vectors of $\mathbf{U}_{\mathrm{Rx},k}$ which correspond to the m_l strongest singular values of $\mathbf{\Sigma}_{\mathrm{Rx},k}$.

Fourthly, the iterative optimization continues for a finite number of times from the second step considering the recomputed Tx and Rx filters of (4.17).

The described iterative optimization can be performed at RS and investigations showed that, in general, five repetitions are sufficient. Afterwards, pilot assisted channel estimation can be used to estimate the Tx and Rx subchannels at each node.

4.3.3 Transceive Filter Design at Relay Station

4.3.3.1 Weighted Zero-Forcing (WZF) Approach

In this section, a weighted ZF (WZF) approach is introduced which does neither exploit the self-interference cancellation capabilities nor the SIC capabilities at the nodes due to spatially separating all signals. This approach is an extension of the ZF approach for single-pair two-way relaying with limited capabilities at the nodes of [Ung09] and it is only introduced for comparison due to the lack of state of the art approaches which consider the introduced ADR requirements. Compared to the ZF approach of [Ung09], the WZF approach contains weighting parameters to adjust the Tx power distribution at RS with respect to the considered ADR requirements. Furthermore, it is extended to be applicable to the multi-pair two-way relaying scenario. The design of the WZF relay transceive filter depends on the Tx and Rx filters at the nodes.

For the WZF approach, the overall channels for the MAC and for the BC phase are defined as follows

$$\mathbf{H}_{\mathrm{MAC}} = \left[\mathbf{H}_1\mathbf{Q}_1, \mathbf{H}_2\mathbf{Q}_2, ..., \mathbf{H}_{K-1}\mathbf{Q}_{K-1}, \mathbf{H}_K\mathbf{Q}_K\right], \tag{4.18a}$$

$$\mathbf{H}_{\mathrm{BC}} = \left[\mathbf{H}_2\mathbf{D}_2^{\mathrm{T}}, \mathbf{H}_1\mathbf{D}_1^{\mathrm{T}}, \mathbf{H}_4\mathbf{D}_4^{\mathrm{T}}, \mathbf{H}_3\mathbf{D}_3^{\mathrm{T}}..., \mathbf{H}_K\mathbf{D}_K^{\mathrm{T}}, \mathbf{H}_{K-1}\mathbf{D}_{K-1}^{\mathrm{T}}\right]^{\mathrm{T}}, \tag{4.18b}$$

respectively, using the Tx and Rx filters at the nodes of Section 4.3.2.

Using these definitions and considering the weighting matrix $\mathbf{W}$ (4.11), the WZF transceive filter at RS using the derivation of [Ung09] can be written as

$$\mathbf{G} = \frac{1}{\alpha_{\mathrm{ZF}}}\mathbf{H}_{\mathrm{BC}}^{\mathrm{H}}\left(\mathbf{H}_{\mathrm{BC}}\mathbf{H}_{\mathrm{BC}}^{\mathrm{H}}\right)^{-1}\mathbf{W}\left(\mathbf{H}_{\mathrm{MAC}}^{\mathrm{H}}\mathbf{H}_{\mathrm{MAC}}\right)^{-1}\mathbf{H}_{\mathrm{MAC}}^{\mathrm{H}}, \tag{4.19}$$

where α_{ZF} is a parameter to fulfill the power constraint at RS and it is given by

$$\alpha_{\mathrm{ZF}} = \sqrt{\frac{\mathrm{tr}\left(\widetilde{\mathbf{G}}\left(\sum_{k=1}^{K}\left(\mathbf{H}_k\mathbf{Q}_k\mathbf{Q}_k^{\mathrm{H}}\mathbf{H}_k^{\mathrm{H}}\right) + \sigma_{\mathrm{n,RS}}^2\mathbf{I}_L\right)\widetilde{\mathbf{G}}^{\mathrm{H}}\right)}{P_{\mathrm{RS}}}}, \tag{4.20}$$

with the auxiliary matrix $\widetilde{\mathbf{G}}$ given by

$$\widetilde{\mathbf{G}} = \mathbf{H}_{\mathrm{BC}}^{\mathrm{H}}\left(\mathbf{H}_{\mathrm{BC}}\mathbf{H}_{\mathrm{BC}}^{\mathrm{H}}\right)^{-1}\mathbf{W}\left(\mathbf{H}_{\mathrm{MAC}}^{\mathrm{H}}\mathbf{H}_{\mathrm{MAC}}\right)^{-1}\mathbf{H}_{\mathrm{MAC}}^{\mathrm{H}}. \tag{4.21}$$

4.3.3.2 Weighted MMSE (WMMSE) Approach

In this section, a weighted MMSE (WMMSE) approach is introduced which does neither exploit the self-interference cancellation capabilities nor the SIC capabilities at the nodes due to suppressing all interferences with respect to minimizing the MSE. This WMMSE approach is an extension of the MMSE approach for single-pair two-way relaying with limited capabilities at the nodes of [Ung09] and it is only introduced for comparison due to the lack of state of the art approaches which consider the introduced ADR requirements. Compared to the MMSE approach of [Ung09], the WMMSE approach contains weighting parameters to adjust the Tx power distribution at RS with respect to the considered ADR requirements. Furthermore, it is extended to be applicable to the multi-pair two-way relaying scenario. Compared to the WZF approach of Section 4.3.3.1, the noise powers at the nodes and at RS are taken into account for the WMMSE relay transceive filter design. The design of the relay transceive filter depends on the Tx and Rx filters at the nodes.

Using the overall channels for the MAC and for the BC phase as defined in (4.18) and considering the weighting matrix $\mathbf{W}$ (4.11), the WMMSE transceive filter at RS using the derivation of [Ung09] can be written as

$$\mathbf{G} = \frac{1}{\alpha_{\mathrm{MMSE}}}\widetilde{\mathbf{G}}, \tag{4.22}$$

with the auxiliary matrix $\widetilde{\mathbf{G}}$ given by

$$\widetilde{\mathbf{G}} = \left(\mathbf{H}_{\mathrm{BC}}^{\mathrm{H}}\mathbf{W}\mathbf{H}_{\mathrm{BC}} + \frac{\mathrm{tr}\left(\mathbf{W}\right)\sigma_{\mathrm{n}}^{2}}{P_{\mathrm{RS}}}\mathbf{I}_{L}\right)^{-1}\mathbf{H}_{\mathrm{BC}}^{\mathrm{H}}\mathbf{W}\mathbf{H}_{\mathrm{MAC}}^{\mathrm{H}}\left(\mathbf{H}_{\mathrm{MAC}}\mathbf{H}_{\mathrm{MAC}}^{\mathrm{H}} + \sigma_{\mathrm{n,RS}}^{2}\mathbf{I}_{L}\right)^{-1}, \tag{4.23}$$

and with α_{MMSE} given by

$$\alpha_{\mathrm{MMSE}} = \sqrt{\frac{\mathrm{tr}\left(\widetilde{\mathbf{G}}\left(\sum_{k=1}^{K}\left(\mathbf{H}_{k}\mathbf{Q}_{k}\mathbf{Q}_{k}^{\mathrm{H}}\mathbf{H}_{k}^{\mathrm{H}}\right) + \sigma_{\mathrm{n,RS}}^{2}\mathbf{I}_{L}\right)\widetilde{\mathbf{G}}^{\mathrm{H}}\right)}{P_{\mathrm{RS}}}}. \tag{4.24}$$

4.3.3.3 SIC-Aware Weighted MMSE (WMMSE-SIC) Approach

In this section, a weighted MMSE approach for the transceive filter design at RS is presented which exploits the self-interference cancellation capabilities and the SIC capabilities of the nodes. The approach, termed WMMSE-SIC, is an extension of the MMSE approach for single-pair two-way relaying with local CSI at the nodes of [Ung09]. Compared to the MMSE approach of [Ung09], the WMMSE-SIC approach contains weighting parameters to adjust the Tx power distribution at RS with respect to the considered ADR requirements. Furthermore, it exploits the SIC capabilities of the nodes to increase the achievable sum rates. Moreover, it is extended to be applicable to the multi-pair two-way relaying scenario by considering inter-pair interferences. The design of the relay transceive filter depends on the Tx and Rx filters at the nodes. In the following, the WMMSE-SIC transceive filter at RS is derived assuming given Tx and Rx filters at the nodes.

To include the weighting parameters, the MSE for each direction of transmission is separated for the derivation of the WMMSE-SIC relay transceive filter. After separating the different MSEs, the weighting parameters w_k which have been introduced in Section 4.3.1 are multiplied with the corresponding MSE to enable an adjustment of the fraction of the Tx power at RS which is used to retransmit the signal $\mathbf{s}_k$ at RS. Furthermore, to consider the noise powers at the nodes with respect to the power constraint at RS, it is proposed to consider an additional receive coefficient α at all nodes and to solve the joint optimization problem of α and $\mathbf{G}$ as it is considered for the conventional MIMO Tx filter design in [JUN05]. Thus, the general equation for the joint optimization problem of the receive coefficient α and the WMMSE-SIC relay transceive filter $\mathbf{G}$ can be written as

$$\{\alpha, \mathbf{G}\} = \underset{\alpha,\mathbf{G}}{\arg\min}\ \mathrm{E}\left\{\sum_{k=1}^{K} w_k \left\|\mathbf{s}_k - \alpha\hat{\mathbf{s}}_k\right\|_2^2\right\}, \tag{4.25a}$$

s.t.

$$\mathrm{tr}\left(\mathbf{G}\left(\sum_{k=1}^{K}\left(\mathbf{H}_k\mathbf{Q}_k\mathbf{Q}_k^{\mathrm{H}}\mathbf{H}_k^{\mathrm{H}}\right) + \sigma_{\mathrm{n,RS}}^2\mathbf{I}_L\right)\mathbf{G}^{\mathrm{H}}\right) \leq P_{\mathrm{RS}}, \tag{4.25b}$$

where $\hat{\mathbf{s}}_k$ (4.4) is the estimate of $\mathbf{s}_k$ at S_l after linear receive filtering, self-interference cancellation and SIC, $l = k - 1 + 2 \cdot \mathrm{mod}_2 k$.

To obtain a simplified expression of the MSE, let matrices $\boldsymbol{\Upsilon}_k$, $\boldsymbol{\Upsilon}_{k,n}$ and $\boldsymbol{\Upsilon}$ be given by

$$\boldsymbol{\Upsilon}_k = \mathbf{H}_k\mathbf{Q}_k\mathbf{Q}_k^{\mathrm{H}}\mathbf{H}_k^{\mathrm{H}} + \frac{1}{K-1}\sigma_{\mathrm{n,RS}}^2\mathbf{I}_L, \tag{4.26a}$$

$$\boldsymbol{\Upsilon}_{k,n} = \mathbf{H}_k\mathbf{q}_{k,n}\mathbf{q}_{k,n}^{\mathrm{H}}\mathbf{H}_k^{\mathrm{H}}, \tag{4.26b}$$

$$\boldsymbol{\Upsilon} = \sum_{k=1}^{K}\mathbf{H}_k\mathbf{Q}_k\mathbf{Q}_k^{\mathrm{H}}\mathbf{H}_k^{\mathrm{H}} + \sigma_{\mathrm{n,RS}}^2\mathbf{I}_L. \tag{4.26c}$$

Using these notations, the MSE for the transmission from S_k to S_l can be written as

$$\begin{aligned}\mathrm{E}\left\{\left\|\mathbf{s}_k - \alpha\hat{\mathbf{s}}_k\right\|_2^2\right\} = {} & m_k - 2\Re\left[\mathrm{tr}\left(\alpha\mathbf{D}_l\mathbf{H}_l^{\mathrm{T}}\mathbf{G}\mathbf{H}_k\mathbf{Q}_k\right)\right] \\ & + |\alpha|^2\mathrm{tr}\left(\sum_{i=1,i\neq l}^{K}\mathbf{D}_l\mathbf{H}_l^{\mathrm{T}}\mathbf{G}\boldsymbol{\Upsilon}_i\mathbf{G}^{\mathrm{H}}\mathbf{H}_l^{*}\mathbf{D}_l^{\mathrm{H}}\right) \\ & - |\alpha|^2\mathrm{tr}\left(\sum_{n=2}^{m_k}\sum_{m=1}^{n-1}\mathbf{d}_{l,n}\mathbf{H}_l^{\mathrm{T}}\mathbf{G}\boldsymbol{\Upsilon}_{k,m}\mathbf{G}^{\mathrm{H}}\mathbf{H}_l^{*}\mathbf{d}_{l,n}^{\mathrm{H}}\right) \\ & + |\alpha|^2\mathrm{tr}\left(\sigma_{\mathrm{n}}^2\mathbf{D}_l\mathbf{D}_l^{\mathrm{H}}\right).\end{aligned} \tag{4.27}$$

Using 4.27, the objective function (4.25a) is non-convex since $\mathbf{G}$ and α appear jointly in third order degree or higher. However, α can be assumed to be positive real-valued and the MSE (4.27) as well as the constraint (4.25b) are convex with respect to $\mathbf{G}$. Based on the assumption that α is positive real valued, a unique solution for problem (4.25) can be obtained by using Lagrangian optimization [BV04, Joh04, Ung09]. With $F(\mathbf{G}, \alpha, k) = w_k\mathrm{E}\left\{\left\|\mathbf{s}_k - \alpha\hat{\mathbf{s}}_k\right\|_2^2\right\}$, using (4.27), the Lagrangian function with the Lagrangian multiplier η results in

$$L(\mathbf{G}, \alpha, \eta) = \sum_{k=1}^{K}\left(F(\mathbf{G}, \alpha, k)\right) - \eta\left(\mathrm{tr}\left(\mathbf{G}\boldsymbol{\Upsilon}\mathbf{G}^{\mathrm{H}}\right) - P_{\mathrm{RS}}\right). \tag{4.28}$$

From the Lagrangian function, the KKT conditions, which are necessary conditions for a global optimum, can be derived and η can be computed, which is presented in detail in Appendix A.3. The KKT conditions can be written as

$$\frac{\partial L}{\partial \mathbf{G}} = \sum_{k=1}^{K} \frac{\partial F(\mathbf{G}, \alpha, k)}{\partial \mathbf{G}} - \eta \, \mathbf{G}^* \boldsymbol{\Upsilon}^{\mathrm{T}} = \mathbf{0}, \tag{4.29a}$$

$$\frac{\partial L}{\partial \alpha} = \sum_{k=1}^{K} \frac{\partial F(\mathbf{G}, \alpha, k)}{\partial \alpha} = 0, \tag{4.29b}$$

$$\eta \left(\mathrm{tr}\left(\mathbf{G} \boldsymbol{\Upsilon} \mathbf{G}^{\mathrm{H}} \right) - P_{\mathrm{RS}} \right) = 0, \tag{4.29c}$$

and the Lagrangian multiplier η results in

$$\eta = -\frac{|\alpha|^2 \sigma_{\mathrm{n}}^2 \sum_{k=1}^{K} w_k \mathrm{tr}\left(\mathbf{D}_k \mathbf{D}_k^{\mathrm{H}} \right)}{P_{\mathrm{RS}}}. \tag{4.30}$$

Now, considering the derivations in Appendix A.3, a unique solution can be obtained for $|\alpha|^2$. Thus, restricting α to be positive real-valued, a unique solution can be obtained for problem (4.25).

Considering the first KKT condition (4.29a), a matrix $\mathbf{K}$ is defined as

$$\begin{aligned} \mathbf{K} = & \sum_{k=1}^{K} w_k \sum_{i=1, i \neq l}^{K} \boldsymbol{\Upsilon}_i^{\mathrm{T}} \otimes \mathbf{H}_l^* \mathbf{D}_l^{\mathrm{H}} \mathbf{D}_l \mathbf{H}_l^{\mathrm{T}} + \boldsymbol{\Upsilon}^{\mathrm{T}} \otimes \frac{\sigma_{\mathrm{n}}^2 \sum_{k=1}^{K} w_k \mathrm{tr}\left(\mathbf{D}_k \mathbf{D}_k^{\mathrm{H}} \right)}{P_{\mathrm{RS}}} \mathbf{I}_L \\ & - \sum_{k=1}^{K} w_k \sum_{n=2}^{m_k} \sum_{m=1}^{n-1} \boldsymbol{\Upsilon}_{k,m}^{\mathrm{T}} \otimes \mathbf{H}_l^* \mathbf{d}_{l,n}^{\mathrm{H}} \mathbf{d}_{l,n} \mathbf{H}_l^{\mathrm{T}}. \end{aligned} \tag{4.31}$$

Now, an analytical solution can be obtained for the WMMSE-SIC relay transceive filter which solves problem (4.25) using (4.29), (4.30) and (4.31). With the auxiliary matrix $\widetilde{\mathbf{G}}$ given by

$$\widetilde{\mathbf{G}} = \mathrm{vec}_{L,L}^{-1} \left(\mathbf{K}^{-1} \ \mathrm{vec} \left(\sum_{k=1}^{K} w_k \mathbf{H}_l^* \mathbf{D}_l^{\mathrm{H}} \mathbf{Q}_k^{\mathrm{H}} \mathbf{H}_k^{\mathrm{H}} \right) \right), \tag{4.32}$$

and using

$$\alpha = \sqrt{\frac{\mathrm{tr}\left(\widetilde{\mathbf{G}} \boldsymbol{\Upsilon} \widetilde{\mathbf{G}}^{\mathrm{H}} \right)}{P_{\mathrm{RS}}}}, \tag{4.33}$$

the WMMSE-SIC filter at RS which solves problem (4.25) is given by

$$\mathbf{G} = \frac{1}{\alpha} \mathrm{vec}_{L,L}^{-1} \left(\mathbf{K}^{-1} \ \mathrm{vec} \left(\sum_{k=1}^{K} w_k \mathbf{H}_l^* \mathbf{D}_l^{\mathrm{H}} \mathbf{Q}_k^{\mathrm{H}} \mathbf{H}_k^{\mathrm{H}} \right) \right). \tag{4.34}$$

The derived WMMSE-SIC relay transceive filter minimizes the weighted MSE for given Tx and Rx filters at the nodes considering that the nodes can perform self-interference cancellation and SIC.

4.4 Transmit Strategies for the Consideration of ADR Requirements

4.4.1 Introduction

In this section, two Tx strategies are proposed which are specifically designed to fulfill the considered ADR requirements whilst achieving high sum rates. First, a Tx strategy, termed power adapted (PA) Tx strategy, is proposed which adapts the Tx powers of the nodes and the Tx power distribution at RS for each direction of transmission. Secondly, a Tx strategy which additionally optimizes the numbers of simultaneously transmitted data streams, termed optimized streams (OS) Tx strategy, is proposed. For all Tx strategies, the ADR requirements introduced in Section 4.2 are considered.

4.4.2 Power Adapted (PA) Transmit Strategy

The PA Tx strategy is based on adjusting the Tx powers of the nodes and the Tx power distribution at RS for each direction of transmission. For this approach, it is assumed that each node S_k either transmits $m_k = M$ data streams or $m_k = 1$ data stream simultaneously because an optimization of the numbers of simultaneously transmitted data streams is not considered, $k = 1, 2, ..., K$. The Tx power distribution at RS is adjusted via the weighting parameters w_k which have been considered for the relay transceive filter design in Section 4.3.3. The Tx powers of the nodes are adjusted via the weighting parameters p_k which have been considered for the Tx filter design at the nodes in Section 4.3.2. To achieve high sum rates under the ADR requirements (4.10), the weighting parameters which achieve the highest sum rate according to (4.10) have to be determined. However, this would require a joint optimization of all weighting parameters which has a high computational complexity. Thus, to reduce the computational complexity, a suboptimal low-complexity approach is proposed. In this approach, the joint adjustment of the weighting parameters w_k and p_k is separated and performed iteratively as follows:

First, the relay transceive filter and the filters at the nodes are initialized and the achievable data rates are computed as follows:

1) Compute the Tx and Rx filters at the nodes and the relay transceive filter as proposed in Section 4.3 assuming all weights $p_k = w_k = 1, k = 1, 2, ..., K$,

2) Compute the achievable data rates C_k (4.7) using the filters of step 1).

Secondly, the weighting parameters w_k are adjusted to increase the achievable sum rate (4.10). To determine the data rate which limits the achievable sum rate, the data rates computed in step 2) are weighted by one over the corresponding ADR requirements. Using these weighted data rates, the weighting parameter $w_{k_{\min}}$ with $k_{\min} = \arg\min_k C_k/r_k$ which corresponds to the data rate which limits the sum rate is set to $w_{k_{\min}} = 1$. Furthermore, the weighting parameter $w_{k_{\max}}$ with $k_{\max} = \arg\max_k C_k/r_k$ which corresponds to the maximum of the weighted data rates is reduced. By this approach, the differences between the weighted data rates are decreased in each iteration until all weighted data rates are equal. The precise steps are as follows:

3) Compute $k_{\min} = \arg\min_k \; C_k/r_k$ and set $w_{k_{\min}} = 1$.

4) Compute the relay transceive filter and the achievable data rates for the weighting parameters of step 3).

5) Compute $k_{\max} = \arg\max_k \; C_k/r_k$ and reduce $w_{k_{\max}}$ to fulfill the condition

$$\frac{1}{K}\sum_{i=1}^{K}\frac{C_i}{r_i} - \frac{\epsilon}{2} \leq \frac{C_{k_{\max}}}{r_{k_{\max}}} \leq \frac{1}{K}\sum_{i=1}^{K}\frac{C_i}{r_i} + \frac{\epsilon}{2}, \tag{4.35}$$

or until the condition

$$w_{k_{\max}} < \delta, \tag{4.36}$$

is fulfilled, where ϵ can be selected according to the required accuracy and where $0 < \delta < 1$ ensures that the MSE for this direction of transmission has a sufficient impact on the considered MMSE based relay transceive filter designs. Considering (4.36) is required due to decoupling the adjustment of $w_{k_{\max}}$ from the adjustment of $p_{k_{\max}}$. For an efficient adjustment of $p_{k_{\max}}$ in step 9), it is required that $w_k > 0 \; \forall k$. In this thesis, $\delta = 0.05$ is selected based on numerical results. To adjust $w_{k_{\max}}$, the bisection method is applied. The relay transceive filter is recalculated after each update of $w_{k_{\max}}$. Furthermore, the achievable data rates C_k (4.7) are recalculated after each update of the relay transceive filter.

6) Compute $k_{\max} = \arg\max_k \; C_k/r_k$ and continue from step 3) considering the updated weighting parameters if the conditions

$$\max_k \frac{C_k}{r_k} > \min_k \frac{C_k}{r_k} + \epsilon, \tag{4.37a}$$

$$w_{k_{\max}} > \delta, \tag{4.37b}$$

are fulfilled.

Thirdly, the Tx powers at the nodes are adjusted if the conditions (4.37) are not fulfilled to further increase the achievable sum rate (4.10). The Tx powers at the nodes are adjusted via the weighting parameters p_k. To achieve this, the weighting parameter $p_{k_{\max}}$ with $k_{\max} = \arg\max_k C_k/r_k$ which corresponds to the maximum of the weighted data rates is reduced. Furthermore, the weighting parameter $p_{k_{\min}}$ with $k_{\min} = \arg\min_k C_k/r_k$ which corresponds to the minimum of the weighted data rates is increased. By this approach, the differences between the weighted data rates are decreased in each iteration until all weighted data rates are equal. The precise steps are as follows:

7) Compute $k_{\min} = \arg\min_k \; C_k/r_k$ and set $p_{k_{\min}} = 1$.

8) Compute the relay transceive filter and the achievable data rates for the weighting parameters of step 3.1).

9) Compute $k_{\max} = \arg\max_k \; C_k/r_k$ and reduce $p_{k_{\max}}$ to fulfill the condition

$$\frac{1}{K}\sum_{i=1}^{K}\frac{C_i}{r_i} - \frac{\epsilon}{2} \leq \frac{C_{k_{\max}}}{r_{k_{\max}}} \leq \frac{1}{K}\sum_{i=1}^{K}\frac{C_i}{r_i} + \frac{\epsilon}{2}. \tag{4.38}$$

To adjust $p_{k_{\max}}$, the bisection method is applied. The relay transceive filter is recalculated after each update of $p_{k_{\max}}$. Furthermore, the achievable data rates C_k (4.7) are recalculated after each update of the relay transceive filter.

Finally, the previous steps are repeated until the ADR requirements are fulfilled taking into account an inaccuracy of ϵ. The precise step is as follows:

10) Continue from step 3) considering the updated weighting parameters until

$$\max_k \frac{C_k}{r_k} < \min_k \frac{C_k}{r_k} + \epsilon. \tag{4.39}$$

After performing the aforementioned iterative adjustment of the weighting parameters w_k and p_k, the condition (4.39) is fulfilled. Thus, taking into account an inaccuracy of ϵ, the ADR requirements are fulfilled by the proposed PA Tx strategy.

4.4.3 Optimized Streams (OS) Transmit Strategy

The OS Tx strategy combines an optimization of the numbers m_k of simultaneously transmitted data streams with the PA transmit strategy of Section 4.4.2. The intention of the OS transmit strategy is to reduce inter-pair interference and therewith to increase the achievable data rates. Initially, each node transmits $m_k = M$ data streams and the superposition of MK data streams is received at RS. If less signals are simultaneously received at RS, the spatial separation of the different pairs is simplified and therewith inter-pair interference can be reduced.

In the considered multi-pair two-way relaying scenario, M^K combinations for the numbers m_k of simultaneously transmitted data streams exist. To obtain the numbers m_k of simultaneously transmitted data streams which achieve the highest sum rate (4.10), an exhaustive search over all possible combinations is performed as illustrated in Figure 4.5. First, for each possible input vector $\mathbf{m} = [m_1, m_2, ..., m_K]$ containing the numbers m_k of simultaneously transmitted data streams, the Tx and Rx filters at the nodes and the relay transceive filter are computed as described in Section 4.3 assuming $p_k = v_k = 1$, $k = 1, 2, ..., K$. Secondly, for each $\mathbf{m}$, the weighting parameters w_k and p_k are adjusted using the PA transmit strategy of Section 4.4.2. Thirdly, for each $\mathbf{m}$, the achievable sum rate (4.10) is computed. Finally, the vector $\mathbf{m}$ and the corresponding parameters w_k and p_k which achieve the highest sum rate (4.10) are selected. Considering these parameters, the Tx and Rx filters at the nodes and the relay transceive filter are recomputed as described in Section 4.3.

4.5 Performance Analysis

In this section, the performances of the Tx strategies presented in Section 4.4 are investigated through numerical simulations considering the different filter designs presented in Section 4.3. All channels are assumed to be i.i.d. Rayleigh fading channels with zero-mean and unit variance and the noise variances at the nodes and at RS are assumed to be equal, i.e., $\sigma^2_{\mathrm{n,RS}} = \sigma^2_{\mathrm{n}}$. All simulation results are averaged over 1000 independent channel realizations. The maximum Tx power at RS is set to be K times the maximum Tx power at node S_1 because the signals of all nodes have to be retransmitted at RS during the BC phase, i.e., $P_{\mathrm{RS}} = KP_{\mathrm{node}}$. The ratio between the maximum Tx power P_{node} at the nodes and the noise level σ^2_{n} is termed average SNR. The ADR requirements are given by the vector $\mathbf{r} = (r_1, r_2, ..., r_k)$.

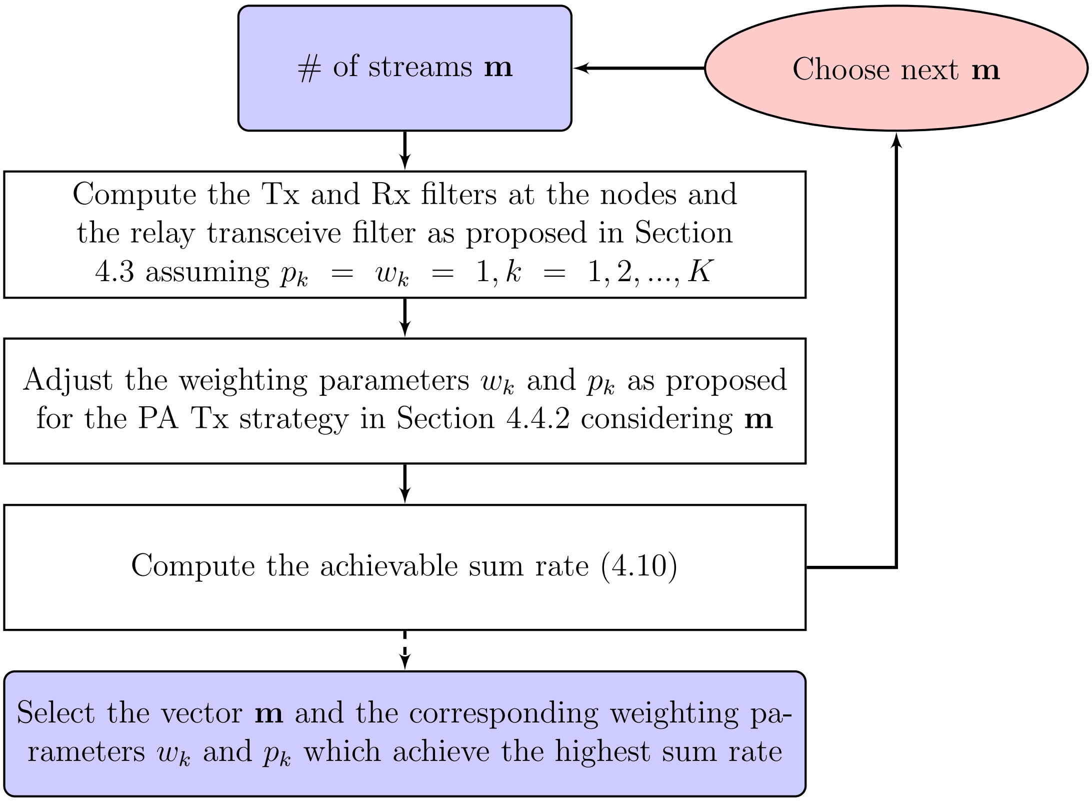

Figure 4.5. Flowchart of the OS transmit strategy

For the numerical simulations, one configuration of the multi-pair two-way relaying scenario is investigated. In this configuration, $K = 4$ nodes are considered and each node is equipped with $M = 2$ antennas. This configuration is investigated because it enables a comprehensive comparison of the different filter designs and the different Tx strategies presented in Section 4.3 and Section 4.4, respectively. Considering $M > 2$ antennas at each node would increase the gain of the proposed approaches compared to the straightforward extensions from the state of the art.

For performance comparison, the approaches listed in Table 4.1 are considered. Due to the lack of state of the art approaches which consider the introduced ADR requirements, the WZF and the WMMSE relay transceive filter designs have been introduced as a straightforward extension from the state of the art in Section 4.3. These approaches are used to show the performance gain of the proposed WMMSE-SIC relay transceive filter design compared to conventional ZF or MMSE based relay transceive filter designs. The performance of the OS transmit strategy compared to the PA transmit strategy is compared for considering the WMMSE-SIC transceive filter at RS because the other transceive filters at RS perform worse and a comparison of the different transceive filters at RS is already provided by considering the PA transmit strategy. Furthermore, the local filter design approach of Section 4.3.2.1 at the nodes performs slightly worse than the global filter design approach of Section 4.3.2.2. Thus, the performance of both approaches is only compared using the WMMSE-SIC transceive filter at RS and considering the OS transmit strategy.

Name	**Transmit Strategy**	**Filter Design at Nodes**	**Filter Design at RS**
GL:WZF:PA (max.streams)	PA of Section 4.4.2, $m_k = M$	Global of Section 4.3.2.2	WZF of Section 4.3.3.1
GL:WMMSE:PA (max.streams)	PA of Section 4.4.2, $m_k = M$	Global of Section 4.3.2.2	WMMSE of Section 4.3.3.2
GL:WMMSE-SIC:PA (max.streams)	PA of Section 4.4.2, $m_k = M$	Global of Section 4.3.2.2	WMMSE-SIC of Section 4.3.3.3
GL:WMMSE-SIC:PA (1 stream)	PA of Section 4.4.2, $m_k = 1$	Global of Section 4.3.2.2	WMMSE-SIC of Section 4.3.3.3
LO:WMMSE-SIC:OS	OS of Section 4.4.3	Local of Section 4.3.2.1	WMMSE-SIC of Section 4.3.3.3
GL:WMMSE-SIC:OS	OS of Section 4.4.3	Global of Section 4.3.2.2	WMMSE-SIC of Section 4.3.3.3

Table 4.1. Overview of the considered approaches for performance comparison.

Figure 4.6 shows the average achievable sum rates versus the number L of antennas at RS considering $\mathbf{r} = (1, 1, 1, 1)$. For these simulations, the average SNR is 15dB. For all approaches, the achievable sum rate increases for increasing the number L of antennas at RS. Neglecting GL:WMMSE-SIC:PA (1 streams), the WZF approach, i.e., GL:WZF:PA (max.streams), performs worst due to spatially separating all simultaneously received signals at RS without considering the impact of noise. For this approach, $L \geq 8$ antennas are required because all received signals have to be spatially separated. The WMMSE approach, i.e., GL:WMMSE:PA (max.streams), performs better than the WZF approach due to considering the impact of noise by minimizing the MSE instead of ZF all interferences. For large number $L > 10$ of antennas at RS, the WZF approach performs similar as the WMMSE approach because the loss in signal power compared to the noise power, termed noise enhancement, due to spatially separating all signals at RS in case of a WZF relay transceive filter decreases for increasing the number L of antennas at RS.

The approaches GL:WMMSE-SIC:PA (max.streams), LO:WMMSE-SIC:OS and GL:WMMSE-SIC:OS, which consider the proposed WMMSE-SIC relay transceive filter design, outperform the conventional WZF and WMMSE relay transceive filter design approaches. The gain of GL:WMMSE-SIC:PA (max.streams) compared to GL:WMMSE:PA (max.streams) is approximately 18% for $L = 8$ antennas at RS and compared to GL:WZF:PA (max.streams), the gain is approximately 63% for $L = 8$ antennas at RS.

Based on the proposed WMMSE-SIC relay transceive filter design, the different Tx strategies are compared. In general, the achievable sum rate for considering the OS transmit strategy is higher than for considering the PA transmit strategy because the OS transmit strategy performs a tradeoff between increasing the multiplexing gain and reducing the inter-pair interferences with respect to the ADR requirements. However, for $L < 7$ antennas at RS, the WMMSE-SIC approach GL:WMMSE-SIC:PA (1 stream) considering the PA transmit strategy achieves the same performance as the WMMSE-SIC approach considering the OS transmit strategy, i.e., GL:WMMSE-SIC:OS, due to considering an ADR requirement of $\mathbf{r} = (1, 1, 1, 1)$. Furthermore, for $L > 8$ antennas at RS, the WMMSE-SIC approach GL:WMMSE-SIC:PA (max.stream) achieves the same performance as GL:WMMSE-SIC:OS due to the considered ADR requirement. For $L < 8$ antennas at RS, a better performance is achieved if each node transmits one data stream instead of transmitting $M = 2$ data streams and for $L \geq 8$ antennas at RS, a better performance is achieved if each node transmits two data streams instead of transmitting one data stream because in both cases, the number L of antennas at RS is at least as high as the number of simultaneously received data streams.

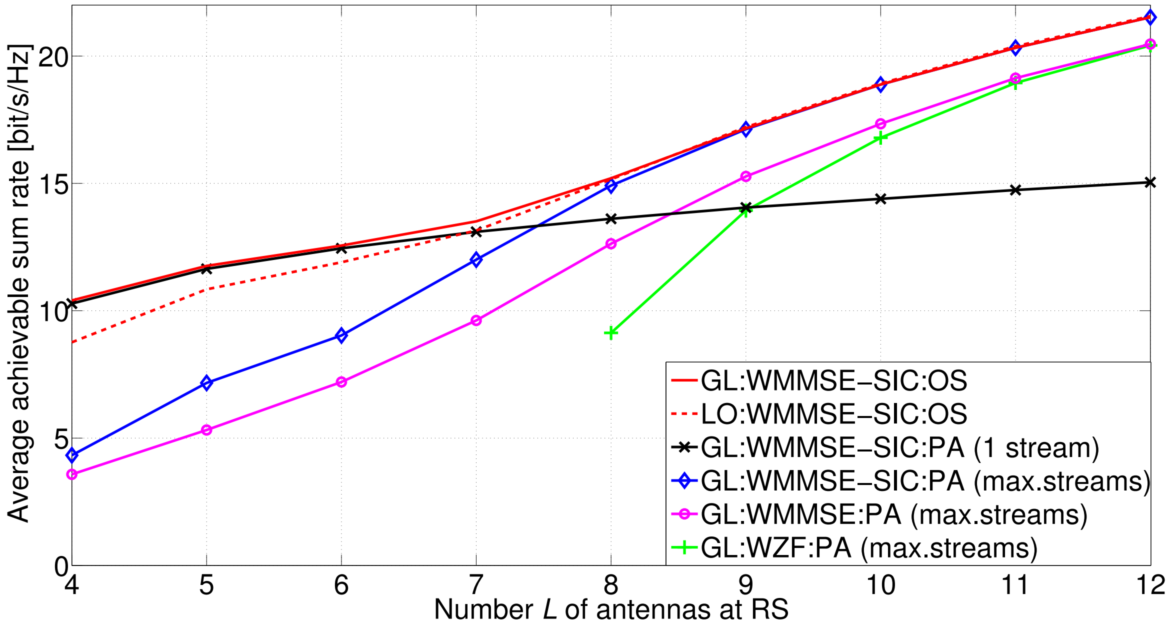

Figure 4.6. Average achievable sum rates versus number L of antennas at RS for an average SNR = 15dB, $K = 4$, $M = 2$, $\mathbf{r} = (1, 1, 1, 1)$.

The performance gain of the global Tx and Rx filter design at the nodes, i.e., GL:WMMSE-SIC:OS, compared the local filter design, i.e., LO:WMMSE-SIC:OS, decreases for increasing the number L of antennas at RS. For $L = 4$ antennas, the gain of GL:WMMSE-SIC:OS compared to LO:WMMSE-SIC:OS is approximately 19% and for $L > 8$ antennas at RS the gain tends towards zero.

Figure 4.7 shows the average achievable sum rates versus the number L of antennas at RS considering $\mathbf{r} = (1, 1/2, 1, 1/2)$. For these simulations, the average SNR is 15dB. For all approaches, the achievable sum rate increases for increasing the number L of antennas at RS. Neglecting GL:WMMSE-SIC:PA (1 streams), the WZF approach, i.e., GL:WZF:PA (max.streams), performs worst due to spatially separating all simultaneously received signals at RS without considering the impact of noise. For this approach, $L \geq 8$ antennas are required because all received signals have to be spatially separated. The WMMSE approach, i.e., GL:WMMSE:PA (max.streams), performs better than the WZF approach due to considering the impact of noise by minimizing the MSE instead of ZF all interferences. For large number $L > 10$ of antennas at RS, the WZF approach performs similar as the WMMSE approach because the noise enhancement due to spatially separating all signals at RS in case of a WZF relay transceive filter decreases for increasing the number L of antennas at RS.

The approaches GL:WMMSE-SIC:PA (max.streams), LO:WMMSE-SIC:OS and

GL:WMMSE-SIC:OS, which consider the proposed WMMSE-SIC relay transceive filter design, outperform the conventional WZF and WMMSE relay transceive filter design approaches. The gain of GL:WMMSE-SIC:PA (max.streams) compared to GL:WMMSE:PA (max.streams) is approximately 16% for $L = 8$ antennas at RS and compared to GL:WZF:PA (max.streams), the gain is approximately 68% for $L = 8$ antennas at RS.

Based on the proposed WMMSE-SIC relay transceive filter design, the different Tx strategies are compared. In general, the achievable sum rate for considering the OS transmit strategy is higher than for considering the PA transmit strategy because the OS transmit strategy performs a tradeoff between increasing the multiplexing gain and reducing the inter-pair interferences with respect to the ADR requirements. Thus, each node transmits a different number of data streams to tackle the ADR requirements. For $L \leq 5$ antennas at RS, the WMMSE-SIC approach GL:WMMSE-SIC:PA (1 stream) considering the PA transmit strategy achieves the same performance as the WMMSE-SIC approach considering the OS transmit strategy, i.e., GL:WMMSE-SIC:OS, because for $L \leq 5$ antennas at RS it is not beneficial if any node transmits more than one data stream under the considered ADR requirements. The performance gain of the approach GL:WMMSE-SIC:OS compared to the approach GL:WMMSE-SIC:PA (max.streams) decreases for increasing the number L of antennas at RS. For $L = 7$ and $L = 8$ antennas at RS, the performance gains of the approach GL:WMMSE-SIC:OS compared to the approach GL:WMMSE-SIC:PA (max.streams) are approximately 22% and 13%, respectively. For $L < 7$ antennas at RS, a better performance is achieved if each node transmits one data stream instead of transmitting $M = 2$ data streams and for $L \geq 7$ antennas at RS, a better performance is achieved if each node transmits two data streams instead of transmitting one data stream.

The performance gain of the global Tx and Rx filter design at the nodes, i.e., GL:WMMSE-SIC:OS, compared the local filter design, i.e., LO:WMMSE-SIC:OS, decreases for increasing the number L of antennas at RS. For $L = 4$ antennas, the gain of GL:WMMSE-SIC:OS compared to LO:WMMSE-SIC:OS is approximately 13% and for $L > 8$ antennas at RS the gain tends towards zero. In general, the proposed GL:WMMSE-SIC:OS approach significantly outperforms the conventional WZF and WMMSE approaches for the considered ADR requirements. The gain of GL:WMMSE-SIC:OS compared to GL:WMMSE:PA (max.streams) is approximately 31% for $L = 8$ antennas at RS and compared to GL:WZF:PA (max.streams), the gain is approximately 89% for $L = 8$ antennas at RS.

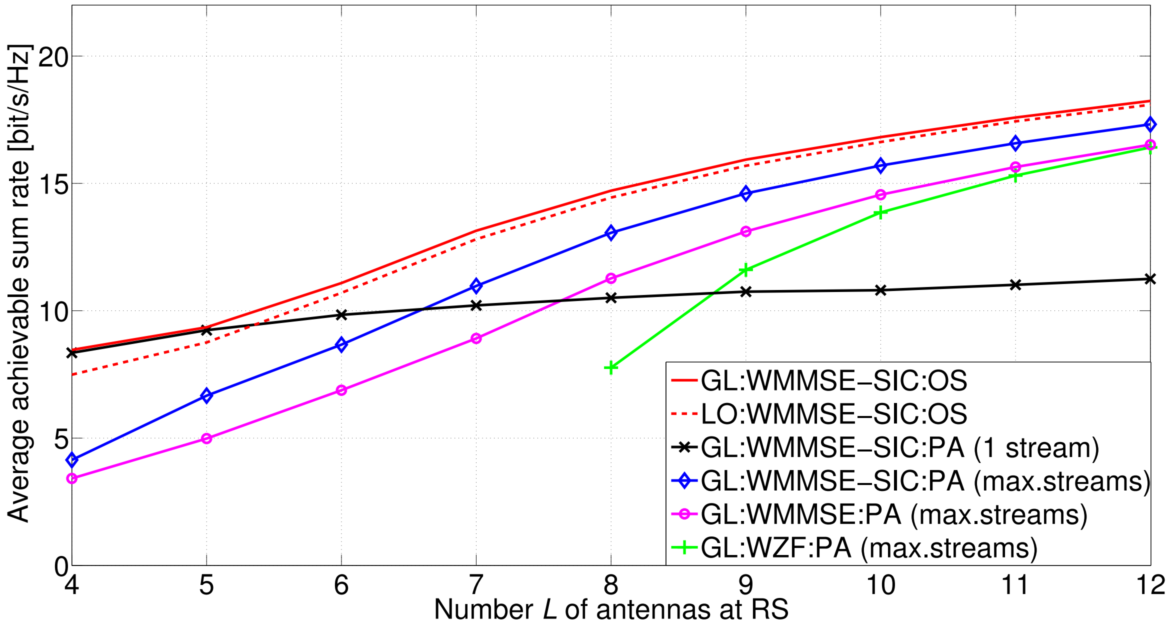

Figure 4.7. Average achievable sum rates versus number L of antennas at RS for an average SNR = 15dB, $K = 4$, $M = 2$, $\mathbf{r} = (1, 1/2, 1, 1/2)$.

Figure 4.8 shows the average achievable sum rates versus the number L of antennas at RS considering $\mathbf{r} = (1, 1/2, 1/2, 1/4)$. For these simulations, the average SNR is 15dB. For all approaches, the achievable sum rate increases for increasing the number L of antennas at RS. Neglecting GL:WMMSE-SIC:PA (1 streams), the WZF approach, i.e., GL:WZF:PA (max.streams), performs worst due to spatially separating all simultaneously received signals at RS without considering the impact of noise. For this approach, $L \geq 8$ antennas are required because all received signals have to be spatially separated. The WMMSE approach, i.e., GL:WMMSE:PA (max.streams), performs better than the WZF approach due to considering the impact of noise by minimizing the MSE instead of ZF all interferences. For large number $L > 10$ antennas at RS, the WZF approach performs similar as the WMMSE approach because the noise enhancement due to spatially separating all signals at RS in case of a WZF relay transceive filter decreases for increasing the number L of antennas at RS.

The approaches GL:WMMSE-SIC:PA (max.streams), LO:WMMSE-SIC:OS and GL:WMMSE-SIC:OS, which consider the proposed WMMSE-SIC relay transceive filter design, outperform the conventional WZF and WMMSE relay transceive filter design approaches. The gain of GL:WMMSE-SIC:PA (max.streams) compared to GL:WMMSE:PA (max.streams) is approximately 5% for $L = 8$ antennas at RS and compared to GL:WZF:PA (max.streams), the gain is approximately 58% for $L = 8$ antennas at RS.

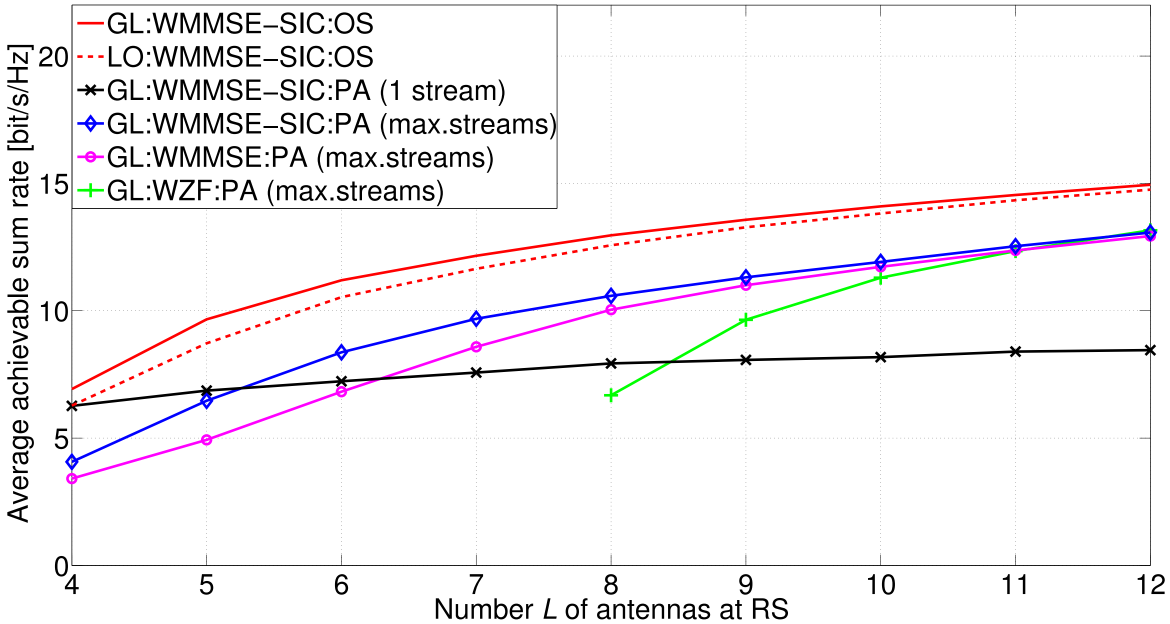

Figure 4.8. Average achievable sum rates versus number L of antennas at RS for an average SNR = 15dB, $K = 4$, $M = 2$, $\mathbf{r} = (1, 1/2, 1/2, 1/4)$.

Based on the proposed WMMSE-SIC relay transceive filter design, the different Tx strategies are compared. In general, the achievable sum rate for considering the OS transmit strategy is higher than for considering the PA transmit strategy because using the OS Tx strategy, each node can transmit a different number of data streams to tackle the ADR requirements. The performance gain of the approach GL:WMMSE-SIC:OS compared to the approach GL:WMMSE-SIC:PA (max.streams) decreases for increasing the number L of antennas at RS. For $L = 6$ and $L = 10$ antennas at RS, the performance gains of the approach GL:WMMSE-SIC:OS compared to the approach GL:WMMSE-SIC:PA (max.streams) are approximately 34% and 18%, respectively. For $L < 6$ antennas at RS, a better performance is achieved if each node transmits one data stream instead of transmitting $M = 2$ data streams and for $L \geq 6$ antennas at RS, a better performance is achieved if each node transmits two data streams instead of transmitting one data stream.

The performance gain of the global Tx and Rx filter design at the nodes, i.e., GL:WMMSE-SIC:OS, compared the local filter design, i.e., LO:WMMSE-SIC:OS, decreases for increasing the number L of antennas at RS. For $L = 5$ antennas, the gain of GL:WMMSE-SIC:OS compared to LO:WMMSE-SIC:OS is approximately 11% and for $L > 10$ antennas at RS the gain tends towards zero. In general, the proposed GL:WMMSE-SIC:OS approach significantly outperforms the conventional WZF and WMMSE approaches for the considered ADR requirements. The gain of GL:WMMSE-SIC:OS compared to GL:WMMSE:PA (max.streams) is approximately 29% for $L = 8$ antennas at RS and compared to GL:WZF:PA (max.streams), the gain is approximately 94% for $L = 8$ antennas at RS. The proposed approach GL:WMMSE-SIC:OS requires approximately three antennas less at RS than the approach GL:WMMSE:PA (max.streams) to achieve the same sum rate.

Comparing Figures 4.6-4.8, the gain of the OS transmit strategy compared to the PA transmit strategy increases for increasing the asymmetry in the ADR requirements. Furthermore, the intersection between GL:WMMSE-SIC:PA (max.streams) and GL:WMMSE-SIC:PA (1 stream) shifts to smaller numbers L of antennas at RS if the asymmetry in the ADR requirements is increased because in this case it is beneficial if the node which requires the highest transmit data rate transmits two data streams instead of one data stream.

Figure 4.9 shows the average achievable sum rates versus the average SNR considering $\mathbf{r} = (1, 1/2, 1/2, 1/4)$ and $L = 7$. For applying a WZF filter at RS, $L \geq 8$ antennas would be required and thus, the GL:WZF:PA (max.streams) approach is not considered. For all other approaches, the achievable sum rate increases for increasing the average SNR. Neglecting GL:WMMSE-SIC:PA (1 streams), the WMMSE approach, i.e., GL:WMMSE:PA (max.streams), performs worst due to not exploiting the self-interference cancellation and the SIC capabilities at the nodes.

The approaches GL:WMMSE-SIC:PA (max.streams), LO:WMMSE-SIC:OS and GL:WMMSE-SIC:OS, which consider the proposed WMMSE-SIC relay transceive filter design, outperform the conventional WMMSE relay transceive filter design approach. The gain of GL:WMMSE-SIC:PA (max.streams) compared to GL:WMMSE:PA (max.streams) is approximately 12% for an average SNR = 15dB.

Based on the proposed WMMSE-SIC relay transceive filter design, the different Tx strategies are compared. In general, the achievable sum rate for considering the OS transmit strategy is higher than for considering the PA transmit strategy due to optimizing the number of simultaneously transmitted data streams of each node with respect to the ADR requirements. The performance gain of the approach GL:WMMSE-SIC:OS compared to the approach GL:WMMSE-SIC:PA (max.streams) increases for increasing the average SNR. For an average SNR of 5dB and 15dB, the performance gains of the approach GL:WMMSE-SIC:OS compared to the approach GL:WMMSE-SIC:PA (max.streams) are approximately 22% and 25%, respectively. For an average SNR < 5dB, a better performance is achieved if each node transmits one data stream instead of transmitting $M = 2$ data streams and for an average

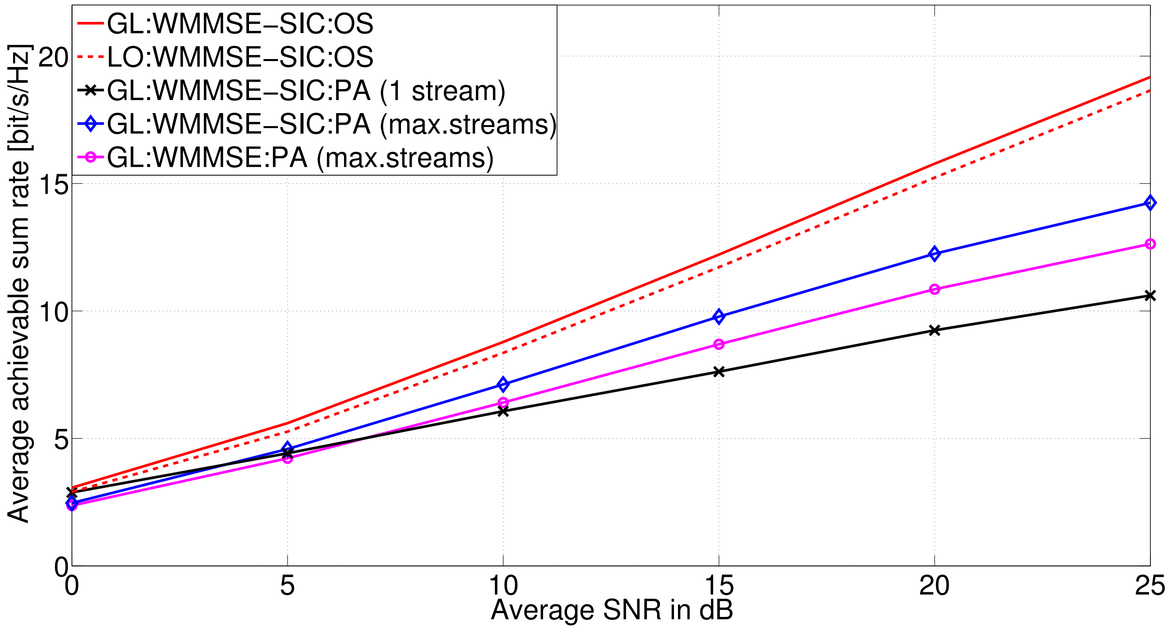

Figure 4.9. Average achievable sum rates versus average SNR, $K = 4$, $M = 2$, $L = 7$, $\mathbf{r} = (1, 1/2, 1/2, 1/4)$.

SNR > 5dB, a better performance is achieved if each node transmits two data streams instead of transmitting one data stream.

The performance gain of the global Tx and Rx filter design at the nodes, i.e., GL:WMMSE-SIC:OS, compared the local filter design, i.e., LO:WMMSE-SIC:OS, is approximately 5% for an average SNR of 10dB. In general, the proposed GL:WMMSE-SIC:OS approach significantly outperforms the conventional WMMSE approach for the considered ADR requirements. The gain of GL:WMMSE-SIC:OS compared to GL:WMMSE:PA (max.streams) is approximately 37% for an average SNR of 10dB.

To summarize, the approach considering the proposed WMMSE-SIC transceive filter at RS of Section 4.3.3.3 combined with the OS transmit strategy of Section 4.4.3 and using the global Tx and Rx filter design of Section 4.3.2.2 significantly outperforms all other approaches. For this approach, the transceive filter at RS exploits the self-interference cancellation and SIC capabilities at the nodes. Furthermore, the Tx and Rx filters at the nodes are designed to minimize inter-pair interference. Moreover, the OS transmit strategy performs an optimization of the numbers of simultaneously transmitted data streams and adjusts the Tx powers of the nodes as well as the Tx power distribution at RS to tackle the ADR requirements. This significantly increases the achievable sum rate compared to conventional MMSE or ZF based approaches. An overview of selected performance gains of the proposed GL:WMMSE-SIC:OS approach compared to the GL:WMMSE:PA (max. streams) approach is presented in Table 4.2 considering $K = 4$ and $M = 2$.

Table 4.2. Selected performance gains of the proposed GL:WMMSE-SIC:OS approach.

SNR	L	$\mathbf{r}$	Conv. Approach	Proposed Approach	Perf. Gain
15dB	6	(1,1,1,1)	GL:WMMSE:PA	GL:WMMSE-SIC:OS	74%
15dB	8	(1,1,1,1)	GL:WMMSE:PA	GL:WMMSE-SIC:OS	20%
15dB	6	(1,1/2,1,1/2)	GL:WMMSE:PA	GL:WMMSE-SIC:OS	61%
15dB	8	(1,1/2,1,1/2)	GL:WMMSE:PA	GL:WMMSE-SIC:OS	31%
15dB	6	(1,1/2,1/2,1/4)	GL:WMMSE:PA	GL:WMMSE-SIC:OS	64%
5dB	7	(1,1/2,1/2,1/4)	GL:WMMSE:PA	GL:WMMSE-SIC:OS	33%
15dB	7	(1,1/2,1/2,1/4)	GL:WMMSE:PA	GL:WMMSE-SIC:OS	40%

Chapter 5

Multi-Group Multi-Way Relaying

5.1 Problem Overview and Decomposition

In this chapter, the multi-group multi-way relaying scenario as shown in Figure 2.3 is investigated. As explained in Section 2.1, one communication cycle consists of one MAC and several BC phases. To enable that each node can decode the messages of all other nodes within its group, at least $N-1$ BC phases are required because each node has to receive $N-1$ independent messages. Considering more than $N-1$ BC phases typically decreases the achievable data rates and thus, $N-1$ BC phases are considered in this thesis. Due to multiple BC phases, temporal Rx processing can be performed at the nodes. Thus, a system model for multi-group multi-way relaying considering multi-antenna nodes which can perform self-interference cancellation and SIC as well as linear spatial and temporal Rx processing is introduced to investigate this scenario.

To maximize the sum rate, the selection of the signals which are retransmitted in each BC phase has to be optimized. Furthermore, the spatial Tx and Rx filters of the nodes, the temporal Rx filters of the nodes and the relay transceive filters have to be optimized with respect to the selected signals for each BC phase. Due to the high computational complexity of finding an optimal solution for this problem, suboptimal approaches based on a problem decomposition are proposed in this chapter. To obtain such suboptimal approaches which achieve high sum rates, the following steps are proposed:

1. It is proposed to decouple the overall problem into three different subproblems as shown in Figure 5.1. The considered subproblems are the design of a Tx strategy which selects the signals which are retransmitted in each BC phase, the design of the relay transceive filters for each BC phase and the design of the spatial and temporal filters of the nodes.

2. It is proposed to focus on low-complexity solutions for the different subproblems.

Based on these steps, suboptimal low-complexity approaches for the different subproblems are proposed as shown in Figure 5.1.

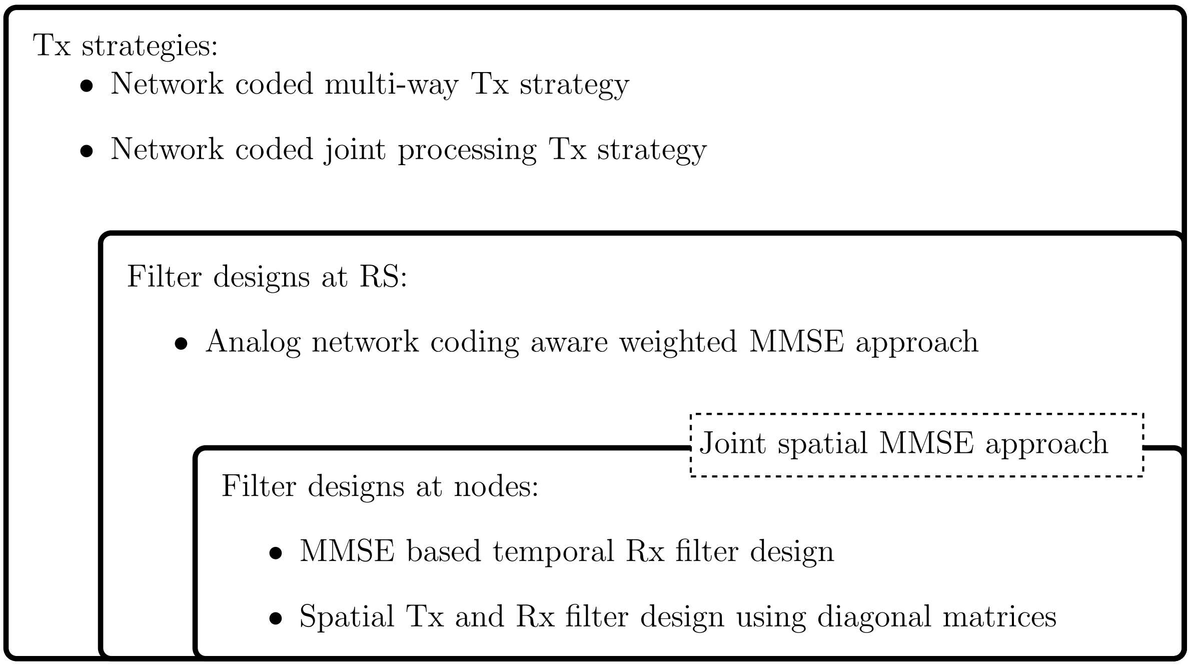

Figure 5.1. Overview of the proposed and investigated approaches for multi-group multi-way relaying.

For the design of a Tx strategy which achieves high sum rates, two different approaches are proposed. The network coding multi-way (NCMW) Tx strategy selects the signals which are retransmitted in each BC such that the self-interference cancellation and SIC capabilities of the nodes can be exploited. For this Tx strategy, joint temporal Rx processing over all BC phases is not considered at the nodes. However, to exploit the temporal processing capabilities of the nodes, it is considered that SIC can be performed over the received signals of all BC phases, i.e., if a signal is successfully decoded at one node in one BC phase, this signal can be canceled at this node before decoding the remaining signals of other BC phases. In the following, this is referred to as known-interference cancellation. The network coding joint processing (NCJP) Tx strategy selects the signals which are retransmitted in each BC such that the capability of the nodes to jointly process the received signals of all BC phases can be exploited in addition to the self-interference cancellation and SIC capabilities of the nodes. Both Tx strategies significantly differ from the Tx strategies which have been proposed in Chapter 3 and Chapter 4 because due to considering multiple BC phases, a selection of the retransmitted signals is required for each BC phase.

For the relay transceive filter design, an ANC aware weighted MMSE approach is proposed which enables an efficient application of the proposed Tx strategies. The proposed relay transceive filter design exploits the capability of the nodes to perform self-interference cancellation and SIC. Additionally, it enables the consideration of ANC based approaches which exploit the temporal processing capabilities of the nodes. The proposed relay transceive filter depends on the spatial Tx and Rx filters at the nodes. To improve the performance, a joint design approach is proposed based on performing an alternating optimization between the spatial filters at the nodes and the relay transceive filter of each BC phase.

For the filter design at the nodes, two approaches are considered. For the temporal Rx filter design at the nodes, a SIC aware MMSE based approach is introduced to jointly process the received signals of all BC phases at each node. This approach depends on the relay transceive filters of all BC phases. For the spatial filter design at the nodes, an optimization is only performed in case of the aforementioned joint design approach. Otherwise, diagonal matrices are considered which are independent of the relay transceive filters.

The proposed relay transceive filter design depends on the spatial Tx and Rx filters at the nodes. Furthermore, the temporal Rx filters of nodes depend on the relay transceive filters. Moreover, the relay transceive filters depend on the selected signals for each BC phase which are determined based on the Tx strategies. Thus, the computation of the different filters and the selection of the signals for each BC phase are performed as shown in Figure 5.2. First, the signals for each BC phase are selected based on the considered Tx strategy. Secondly, the spatial Tx and Rx filters at the nodes are computed as diagonal matrices. Thirdly, the relay transceive filter for each BC phase is computed considering the Tx and Rx filters at the nodes of the previous step. Fourthly, if a joint spatial filter design is considered, the Rx filters at the nodes are updated for each BC phase considering the corresponding relay transceive filter of the previous step. Additionally, an alternating optimization between the relay transceive filter and the Rx filters at the nodes is performed for each BC phase. Finally, the temporal Rx filters at the nodes are computed.

In the following, the relay transceive filter design is introduced before the Tx strategies because the description of the proposed Tx strategies is based on the signal categories which are introduced for the relay transceive filter design. Thus, the rest of the chapter is organized as follows. In Section 5.2, the system model for the considered multi-group multi-way scenario is presented. In Section 5.3, the approaches for the design of the spatial and temporal filters at the nodes and for the design of the relay transceive filter are proposed. In Section 5.4, the Tx strategies are proposed and in Section 5.5, the performance of the proposed approaches is investigated by numerical results. Several parts of this chapter have been originally published by the author in [DK13b, DRK13, DK13a].

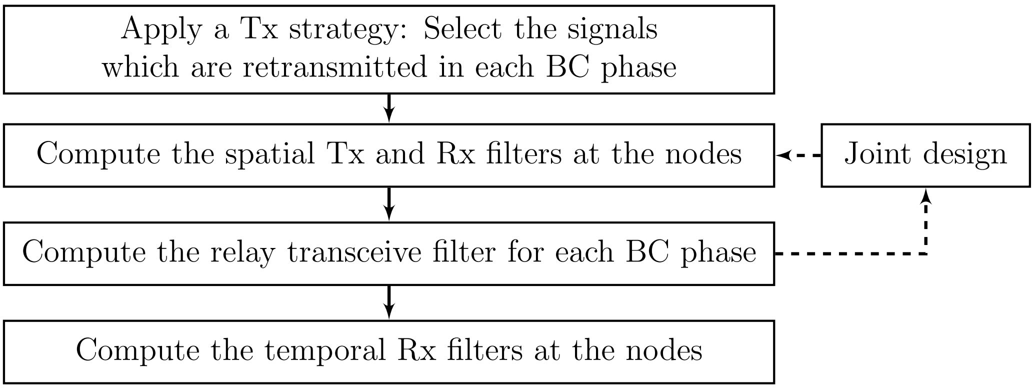

Figure 5.2. Flow chart for the computation of the filters at the nodes and at RS considering the proposed Tx strategies.

5.2 System Model

In this section, the system model for the considered multi-group multi-way relaying scenario as shown in Figure 5.3 is presented. As described in Section 2.1, the scenario consists of G groups with N half-duplex multi-antenna nodes per group. The nodes are termed S_k, $k = 1, 2, ..., K$, where $K = NG$ is the total number of nodes. It is assumed that the nodes S_j, S_{j+1}, ..., S_{j+N-1} with $j = 1+(g-1)N$, are assigned to the g^{th} group, $g = 1, 2, ..., G$. The communications between the nodes of one group are performed via RS. For this scenario, only a single subcarrier is considered, i.e., $C = 1$, because the consideration of multiple subcarriers has no impact on the performance of the proposed Tx strategies. In the following, the subcarrier index c is omitted. Furthermore, it is assumed that each node simultaneously transmits $m_k = M$ data streams.

In the MAC phase, all nodes simultaneously transmit to RS and the superposition of these transmit signals is received at RS. Before the transmission, the Tx signal vector $\mathbf{s}_k \in \mathbb{C}^{M\times 1}$ of node S_k is filtered by the Tx filter $\mathbf{Q}_k \in \mathbb{C}^{M\times M}$, with $||\mathbf{Q}_k||_{\mathrm{F}}^2 \leq P_{\mathrm{Node}}$. Let $\mathbf{n}_{\mathrm{RS}} \in \mathbb{C}^{L\times 1}$ represent the complex white Gaussian noise vector at RS and let $\mathbf{H}_k \in \mathbb{C}^{L\times M}$ denote the channel from S_k to RS. Now, the received signal at RS can be written as

$$\mathbf{y}_{\mathrm{RS}} = \sum_{k=1}^{K} \mathbf{H}_k \mathbf{Q}_k \mathbf{s}_k + \mathbf{n}_{\mathrm{RS}}. \tag{5.1}$$

In the $N-1$ subsequent BC phases, RS retransmits different linearly processed versions of the superimposed received signals back to the nodes. In time slots $t = 2, 3, ..., N$,

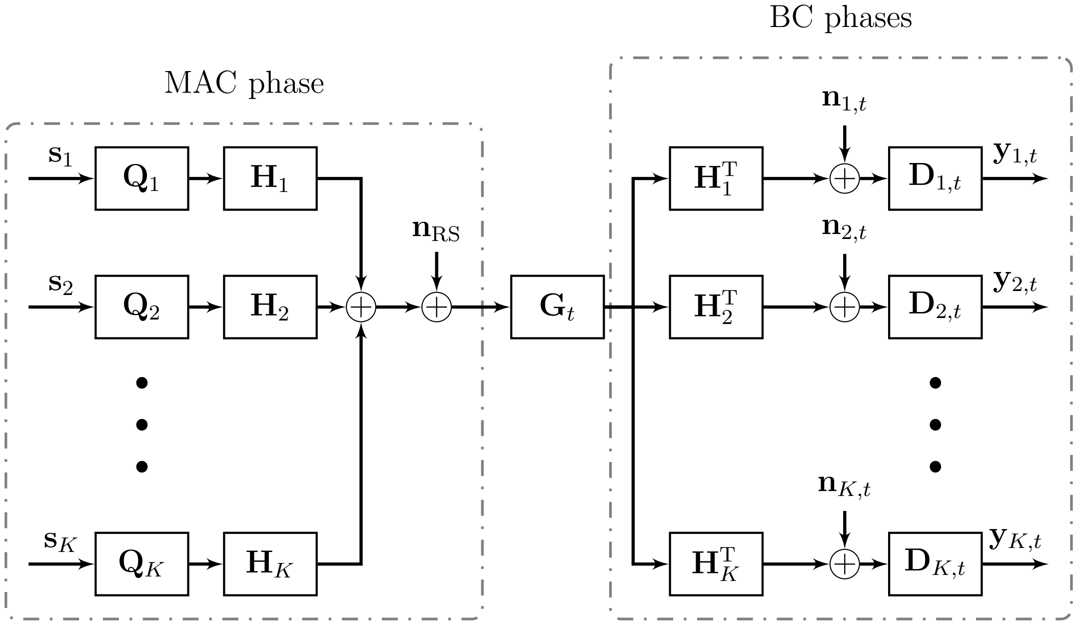

Figure 5.3. System model for MGMW relaying.

the received signal $\mathbf{y}_{\mathrm{RS}}$ is linearly processed at RS using the transceive filter matrix $\mathbf{G}_t \in \mathbb{C}^{L\times L}$. Using the receive filter $\mathbf{D}_{k,t} \in \mathbb{C}^{M\times M}$, the received signal in time slot t at node S_k is given by

$$\mathbf{y}_{k,t} = \mathbf{D}_{k,t}(\mathbf{H}_k^{\mathrm{T}}\mathbf{G}_t\mathbf{y}_{\mathrm{RS}} + \mathbf{n}_{k,t}), \tag{5.2}$$

where $\mathbf{n}_{k,t} \in \mathbb{C}^{M\times 1}$ represents the complex white Gaussian noise vector at S_k in time slot t.

Due to multiple BC phases, joint temporal Rx processing at the nodes over the different BC phases can be applied. To consider joint temporal Rx processing, different matrices and vectors have to be defined to simplify the descriptions. First, channel matrices $\mathbf{H}_{\mathrm{ov},l,k,t}$ are defined to describe the overall channel coefficients for the transmission from S_l to S_k, $k, l = 1, 2, ..., K$, in time slot t as

$$\mathbf{H}_{\mathrm{ov},l,k,t} = \mathbf{D}_{k,t}\mathbf{H}_k^{\mathrm{T}}\mathbf{G}_t\mathbf{H}_l\mathbf{Q}_l. \tag{5.3}$$

Secondly, matrices $\mathbf{A}_{l,k} \in \mathbb{C}^{(N-1)M\times M}$ are defined using (5.3) to describe the overall channel coefficients for the transmission from S_l to S_k during all $N-1$ BC phases as

$$\mathbf{A}_{l,k} = (\mathbf{H}^{\mathrm{T}}_{\mathrm{ov},l,k,2}, \mathbf{H}^{\mathrm{T}}_{\mathrm{ov},l,k,3}, ..., \mathbf{H}^{\mathrm{T}}_{\mathrm{ov},l,k,N})^{\mathrm{T}}. \tag{5.4}$$

Thirdly, using $\mathbf{n}_{\mathrm{ov},k,t} = \mathbf{H}_k^{\mathrm{T}}\mathbf{G}_t\mathbf{n}_{\mathrm{RS}} + \mathbf{n}_{k,t}$ to describe the received noise at S_k in time slot t, a vector $\mathbf{n}_{\mathrm{ov},k} \in \mathbb{C}^{(N-1)M\times 1}$ is defined which contains the received noise at S_k during all BC phases as

$$\mathbf{n}_{\mathrm{ov},k} = \left[(\mathbf{D}_{k,2}\mathbf{n}_{\mathrm{ov},k,2})^{\mathrm{T}}, (\mathbf{D}_{k,3}\mathbf{n}_{\mathrm{ov},k,3})^{\mathrm{T}}, ..., (\mathbf{D}_{k,N}\mathbf{n}_{\mathrm{ov},k,N})^{\mathrm{T}}\right]^{\mathrm{T}}. \tag{5.5}$$

Now, using a matrix $\mathbf{W}_k \in \mathbb{C}^{(N-1)M \times (N-1)M}$ to perform joint linear Rx processing at S_k, the received signals at S_k after joint linear processing over the $N-1$ BC phases can be written as

$$\mathbf{y}_k = \mathbf{W}_k(\mathbf{A}_{1,k}, \mathbf{A}_{2,k}, ..., \mathbf{A}_{K,k}) \cdot (\mathbf{s}_1, \mathbf{s}_2, ..., \mathbf{s}_K)^{\mathrm{T}} + \mathbf{W}_k\mathbf{n}_{\mathrm{ov},k}. \tag{5.6}$$

Remark: To decouple the temporal and the spatial processing for the relay transceive filter design, the receive filters $\mathbf{D}_{k,t}$ and the Rx processing matrix $\mathbf{W}_k$ are designed separately. This approach enables a joint spatial filter design between the relay transceive filter and the Rx filters at the nodes as proposed in Section 5.3.5.

Let $\mathcal{S}_k$ be a subset which contains the $N-1$ indices of the nodes which are in the same group as S_k. To estimate the m^{th} data stream of S_l at S_k, $l \in \mathcal{S}_k$ and $m = 1, 2, ..., M$, the row vector $\mathbf{w}_{l,k,m}$ of $\mathbf{W}_k$ is used which corresponds to the joint Rx processing vector for estimating this data stream. To consider SIC, let $\mathcal{SIC}_{l,k}$ be a subset which contains the indices of the nodes whose transmit signals are already decoded at S_k before estimating the transmit signal of S_l and let this subset include the index k to consider perfect self-interference cancellation. Using $\mathbf{a}_{l,k,m}$ which is the m^{th} column vector of $\mathbf{A}_{l,k}$ (5.4), the expected signal, interference and noise powers when estimating the m^{th} data stream of S_l at S_k can be written as

$$P_{\mathrm{S},l,k,m} = |\mathbf{w}_{l,k,m}\mathbf{a}_{l,k,m}|^2, \tag{5.7}$$

$$P_{\mathrm{I},l,k,m} = \sum_{j=1, j\notin\mathcal{SIC}_{l,k}}^{K} ||\mathbf{w}_{l,k,m}\mathbf{A}_{j,k}||_2^2 - \sum_{i=1}^{m} |\mathbf{w}_{l,k,m}\mathbf{a}_{l,k,i}|^2, \tag{5.8}$$

$$P_{\mathrm{N},l,k,m} = \mathrm{E}[\mathbf{w}_{l,k,m}\mathbf{n}_{\mathrm{ov},k}\mathbf{n}_{\mathrm{ov},k}^{\mathrm{H}}\mathbf{w}_{l,k,m}^{\mathrm{H}}], \tag{5.9}$$

respectively.

For performance comparison in Section 5.5, the maximum achievable sum rate of multi-way relaying, cf. [AK10a, AK11a], is considered and the corresponding equations are presented in the following. Assuming that Gaussian codebooks are used for each data stream, the maximum achievable data rate for the m^{th} data stream from S_l to S_k, $l \in \mathcal{S}_k$, is given by

$$C_{l,k,m} = \frac{1}{N}\log_2(1 + P_{\mathrm{S},l,k,m}(P_{\mathrm{I},l,k,m} + P_{\mathrm{N},l,k,m})^{-1}), \tag{5.10}$$

where the factor $\frac{1}{N}$ is used due to the fact that N time slots are required to perform all transmissions. However, the maximum achievable multi-way rate for the m^{th} data stream of S_l is determined by the minimum over all maximum achievable data rates from S_l to any other node within the same group. Thus, it is given by

$$C_{l,m} = (N-1) \cdot \min_{\forall k \in \mathcal{S}_l} C_{l,k,m}, \tag{5.11}$$

where the factor $N-1$ is used because the Tx signal of S_l is transmitted to $N-1$ nodes. Thus, the achievable sum rate of the multi-way relay system is given by

$$C_{\text{sum}} = \sum_{k=1}^{K} \sum_{m=1}^{M} C_{k,m}. \tag{5.12}$$

5.3 Filter Design

5.3.1 Introduction

In this section, low-complexity approaches for designing the temporal Rx filters at the nodes and for designing the spatial Tx and Rx filters at the nodes are presented as described in Section 5.1. Furthermore, a low-complexity approach for designing the relay transceive filter is proposed as described in Section 5.1 assuming predefined spatial Tx and Rx filters at the nodes. Moreover, an iterative approach for jointly designing the spatial filters at the nodes and the relay transceive filter is proposed.

5.3.2 Temporal Receive Filter Design at Nodes

In this section, the design of the temporal Rx processing matrix $\mathbf{W}_k$ at the nodes is presented. Joint temporal Rx processing over the different BC phases is performed in conjunction with SIC at each node to improve the performance. To determine the temporal Rx processing matrix $\mathbf{W}_k$ at node S_k, an MMSE based filter design as considered in [CZ12] is applied. For this filter design, perfect self-interference cancellation and perfect SIC are assumed.

Thus, the row vector $\mathbf{w}_{l,k,m}$ to perform an MMSE based filtering of the m^{th} data stream transmitted by S_l at S_k, $l \in \mathcal{S}_k$, over the $N-1$ BC phases based on the derivations for conventional MMSE Rx filters [Joh04] is given by

$$\mathbf{w}_{l,k,m} = \mathbf{a}_{l,k,m}^{\mathrm{H}} \left(\sum_{j=1, j \notin \mathcal{SIC}_{l,k}}^{K} \mathbf{A}_{j,k} \mathbf{A}_{j,k}^{\mathrm{H}} - \sum_{i=1}^{m} \mathbf{a}_{l,k,i} \mathbf{a}_{l,k,i}^{\mathrm{H}} + \mathbf{N}_k \right)^{-1}, \tag{5.13}$$

where

$$\mathbf{N}_k = \mathrm{diag}\left(\mathbf{D}_{k,2}\mathbf{H}_k^{\mathrm{T}}\mathbf{G}_2\mathbf{G}_2^{\mathrm{H}}\mathbf{H}_k^{*}\mathbf{D}_{k,2}^{\mathrm{H}}, ..., \mathbf{D}_{k,N}\mathbf{H}_k^{\mathrm{T}}\mathbf{G}_N\mathbf{G}_N^{\mathrm{H}}\mathbf{H}_k^{*}\mathbf{D}_{k,N}^{\mathrm{H}}\right)\sigma_{n,\mathrm{RS}}^2 + \mathrm{diag}\left(\mathbf{D}_{k,2}\mathbf{D}_{k,2}^{\mathrm{H}}, ..., \mathbf{D}_{k,N}\mathbf{D}_{k,N}^{\mathrm{H}}\right)\sigma_n^2. \tag{5.14}$$

The computation of the temporal Rx processing matrix $\mathbf{W}_k$ using (5.13) and (5.14), depends on the relay transceive filters $\mathbf{G}_t$, $t = 2, ..., N$. Furthermore, it depends on the spatial Tx and Rx filters of the nodes. To overcome this problem, the relay transceive filters and the Tx and Rx filters of the nodes are computed before the computation of the temporal Rx processing matrix as described in Section 5.1 and as illustrated in Figure 5.2.

5.3.3 Spatial Transmit and Receive Filter Design at Nodes

In this section, the spatial Tx and Rx filter design at the nodes is presented independent of the considered relay transceive filter. To design the Tx and Rx filters independently of the relay transceive filter, weighted identity matrices are considered.

Thus, the Tx filters are designed as diagonal matrices as follows

$$\mathbf{Q}_k = \sqrt{\frac{P_{\mathrm{node}}}{M}}\mathbf{I}_M. \tag{5.15}$$

Furthermore, the Rx filters are designed as identity matrices as follows

$$\mathbf{D}_{k,t} = \mathbf{I}_M. \tag{5.16}$$

5.3.4 ANC-Aware Weighted MMSE (WMMSE-ANC) Relay Transceive Filter Design

In this section, a weighted ANC aware MMSE based transceive filter design at RS is proposed. The approach, termed WMMSE-ANC, is an extension of the relay transceive filter approach presented in Section 4.3.3.3. Compared to the approach presented in Section 4.3.3.3, the proposed WMMSE-ANC transceive filter at RS is applicable to scenarios with $N \geq 2$ nodes per group whereas the approach of Section 4.3.3.3 is only applicable to scenarios with $N = 2$ nodes per group. Furthermore, it enables the consideration of ANC based approaches which exploit the temporal processing capabilities of the nodes. To achieve this, four different signal categories are considered for the proposed WMMSE-ANC transceive filter design which are introduced in the following.

First, desired signals at the nodes are considered in each BC phase. The desired signal at node S_k in time slot t is given by $\mathbf{s}_{l_{k,t}}$, where the index $l_{k,t}$ is based on the Tx strategy as described in detail in Section 5.4. For the retransmission of the desired signals, it is proposed to exploit ANC. By ANC, it is meant that instead of spatially separating the different desired signals of each group in time slot t given by

$$\mathbf{s}_{l_{k,t}}, \mathbf{s}_{l_{k+1,t}}, ..., \mathbf{s}_{l_{k+N-1,t}} \tag{5.17}$$

with $k = 1 + (g-1)N$ and g the group index, it is proposed to spatially superimpose the desired signals of each group in each BC phase and to recover the individual signals by utilizing the temporal processing capabilities of the nodes.

Secondly, suppressed signals are considered in each BC phase. The retransmission of these signals has to be suppressed by the WMMSE-ANC transceive filter at RS and thus, these signals have to be spatially separated from the desired signals. To achieve this, an MMSE based separation is proposed which considers the noise at RS. The indices of the suppressed signals at node S_k in time slot t are contained in the vector $\mathbf{o}_{k,t} = \left[o_{1,k,t}, o_{2,,k,t}, ..., o_{N_{\mathrm{o},k,t},t}\right]$, where $N_{\mathrm{o},k,t}$ is the number of suppressed signals in time slot t. The signals which are considered as suppressed signals are presented in Section 5.4 for each Tx strategy.

Thirdly, self- and known-interference signals (SKISs) are considered. SKISs are assumed to be known at the nodes and thus, are assumed to be perfectly canceled at the nodes before performing temporal Rx processing and SIC. Thus, no power should be wasted at RS neither for retransmitting SKISs nor for spatially separating SKISs from the desired signals. Based on this, the proposed WMMSE-ANC transceive filter at RS does not intentionally suppress SKISs but considers SKISs with respect to the power constraint at RS. Thus, spatially processed linear combinations of the desired signals and the SKISs are retransmitted by RS in each BC phase. The SKISs at node S_k in time slot t are collected in the subset $\mathcal{SKIS}_{k,t}$. The signals which are considered as SKISs are presented in Section 5.4 for each Tx strategy.

Fourthly, the remaining signals (RMSs) at the nodes are considered. A signal $\mathbf{s}_l$ is considered as RMS at node S_k in time slot t if $\mathbf{s}_l$ is neither considered as desired signal nor as suppressed signal nor as SKIS. Due to considering joint temporal Rx processing over all BC phases at the nodes, the RMSs can be used to improve the overall performance because each RMS in one BC phase is a desired signal in another BC phase.

Based on this, we propose that RMSs are treated in the same way as SKISs for the spatial processing at RS. However, RMSs are not known at the nodes and thus, cannot be canceled before performing temporal Rx processing. Therefore, the interferences caused by RMSs have to be reduced or canceled at the nodes when estimating a desired signal by performing temporal Rx processing and SIC. The proposed approach exploits ANC and instead of spatially separating the RMSs from the desired signals, spatially processed linear combinations of these signals are retransmitted by RS in each BC phase.

Considering the introduced signal categories and weighting parameters $v_{k,t}$, $0 \leq v_{k,t} \leq 2$, it is proposed that the spatial processing at RS should minimize the weighted MSE for the transmission of the desired signals in time slot t given by

$$\mathbf{G}_t = \underset{\mathbf{G}_t}{\arg\min}\ \mathrm{E}\left[\sum_{k=1}^{K} v_{k,t}||\mathbf{s}_{l_{k,t}} - \hat{\mathbf{s}}_{l_{k,t}}||_2^2\right], \tag{5.18a}$$

$$\text{s.t.} \ \sum_{k=1}^{K} ||\mathbf{G}_t\mathbf{H}_k\mathbf{Q}_k||_\mathrm{F}^2 + ||\mathbf{G}_t||_\mathrm{F}^2\sigma_{\mathrm{n,RS}}^2 \leq P_\mathrm{RS}, \tag{5.18b}$$

where $\hat{\mathbf{s}}_{l_{k,t},k}$ is the estimate of $\mathbf{s}_{l_{k,t}}$ at node S_k assuming that the SKISs and the RMSs can be perfectly canceled at the nodes due to exploiting the temporal Rx processing capabilities and SIC. Thus, $\hat{s}_{l_{k,t},k}$ only contains the desired signal, the suppressed signals and noise and it is given by

$$\hat{\mathbf{s}}_{l_{k,t},k} = \mathbf{D}_{k,t}\mathbf{H}_k^\mathrm{T}\mathbf{G}_t\left(\mathbf{H}_{l_{k,t}}\mathbf{s}_{l_{k,t}} + \sum_{i=1}^{N_{\mathrm{o},k,t}} \mathbf{H}_{o_{i,k,t}}\mathbf{s}_{o_{i,k,t}} + \mathbf{n}_\mathrm{RS}\right) + \mathbf{D}_{k,t}\mathbf{n}_{l,t}. \tag{5.19}$$

An intuitive explanation of the proposed spatial processing at RS with respect to the introduced signal categories can be given as follows. Considering (5.18a), the weighted MSE for the transmission of the desired signals is minimized. In the estimate of each desired signal (5.19), the impact of the suppressed signals $\mathbf{s}_{o_{i,k,t}}$ is considered and thus, the retransmission of these signals is suppressed at RS to minimize (5.18a), $i = 1, 2, ..., N_{\mathrm{o},k,t}$. The SKISs and the RMSs are not considered in the estimate of each desired signal (5.19) because it is assumed that these signals can be suppressed or canceled by utilizing the temporal processing capabilities of the nodes. However, the SKISs and the RMSs are considered in the power constraint at RS (5.18b). Thus, no power is wasted at RS with respect to minimizing the MSE (5.18a) neither for retransmitting SKISs and RMSs nor for spatially separating SKISs and RMSs from the desired signals.

If the relay transceive filter $\mathbf{G}_t$ is designed to minimize (5.18), the solution for $\mathbf{G}_t$ does not consider the noise powers at the nodes. However, the noise powers at the nodes should be considered with respect to the power constraint at RS to increase the achievable data rates. To achieve this, it is proposed to consider an additional receive coefficient α_t at all nodes and to solve the joint optimization problem of α_t and $\mathbf{G}_t$ as it is considered for MIMO Tx filter design in [Joh04, JUN05].

Thus, the joint optimization problem of the weighted ANC aware relay transceive filter $\mathbf{G}_t$ and the receive coefficient α_t in time slot t with respect to the Tx power constraint at RS is given by

$$\{\alpha_t, \mathbf{G}_t\} = \underset{\alpha_t, \mathbf{G}_t}{\arg\ \min}\ \mathrm{E}\left[\sum_{k=1}^{K} v_{k,t} ||\mathbf{s}_{l_{k,t}} - \alpha_t \hat{\mathbf{s}}_{l_{k,t},k}||_2^2\right], \tag{5.20a}$$

$$\text{s.t.} \sum_{k=1}^{K} ||\mathbf{G}_t \mathbf{H}_k \mathbf{Q}_k||_\mathrm{F}^2 + ||\mathbf{G}_t||_\mathrm{F}^2 \sigma_{\mathrm{n,RS}}^2 \leq P_{\mathrm{RS}}, \tag{5.20b}$$

where the weighting parameters $v_{k,t}$ are assumed to be known. The computation of these weighting parameters is described in Section 5.4.

Using the matrices $\mathbf{\Upsilon}^{(k)}$ and $\mathbf{\Upsilon}$ given by

$$\mathbf{\Upsilon}^{(k)} = \mathbf{H}_k \mathbf{Q}_k \mathbf{Q}_k^\mathrm{H} \mathbf{H}_k^\mathrm{H}, \tag{5.21a}$$

$$\mathbf{\Upsilon} = \sum_{k=1}^{K} \mathbf{H}_k \mathbf{Q}_k \mathbf{Q}_k^\mathrm{H} \mathbf{H}_k^\mathrm{H} + \sigma_{n,\mathrm{RS}}^2 \mathbf{I}_L, \tag{5.21b}$$

respectively, the MSE for the transmission of $\mathbf{s}_{l_{k,t}}$ to S_k in time slot t can be written as

$$\begin{aligned}
\mathrm{E}\left[||\mathbf{s}_{l_{k,t}} - \alpha_t \hat{\mathbf{s}}_{l_{k,t},k}||_2^2\right] = {} & M - 2\Re\left[\mathrm{tr}\left(\alpha_t \mathbf{D}_{k,t} \mathbf{H}_k^\mathrm{T} \mathbf{G}_t \mathbf{H}_{l_{k,t}} \mathbf{Q}_{l_{k,t}}\right)\right] \\
& + \mathrm{tr}\left(|\alpha_t|^2 \mathbf{D}_{k,t} \mathbf{H}_k^\mathrm{T} \mathbf{G}_t \mathbf{\Upsilon}^{(l_{k,t})} \mathbf{G}_t^\mathrm{H} \mathbf{H}_k^* \mathbf{D}_{k,t}^\mathrm{H}\right) \\
& + \mathrm{tr}\left(|\alpha_t|^2 \sum_{i=1}^{N_{\mathrm{o},k,t}} \mathbf{D}_{k,t} \mathbf{H}_k^\mathrm{T} \mathbf{G}_t \mathbf{\Upsilon}^{(o_{i,k,t})} \mathbf{G}_t^\mathrm{H} \mathbf{H}_k^* \mathbf{D}_{k,t}^\mathrm{H}\right) \\
& + \mathrm{tr}\left(|\alpha_t|^2 \sigma_{n,\mathrm{RS}}^2 \mathbf{D}_{k,t} \mathbf{H}_k^\mathrm{T} \mathbf{G}_t \mathbf{G}_t^\mathrm{H} \mathbf{H}_k^* \mathbf{D}_{k,t}^\mathrm{H}\right) \\
& + \mathrm{tr}\left(\sigma_n^2 |\alpha_t|^2 \mathbf{D}_{k,t} \mathbf{D}_{k,t}^\mathrm{H}\right).
\end{aligned} \tag{5.22}$$

Using (5.22), the objective function (5.20a) is non-convex since $\mathbf{G}_t$ and α_t appear jointly in third-order degree or higher. However, α_t can be assumed to be positive real-valued and the MSE of (5.22) as well as the constraint (5.20b) are convex with respect to $\mathbf{G}_t$. Based on the assumption that α_t is positive real valued, a unique solution for problem (5.20) can be obtained by using Lagrangian optimization [BV04, Joh04, Ung09]. With $F(\mathbf{G}_t, \alpha_t, k, t) = \mathrm{E}\left[||\mathbf{s}_{l_{k,t}} - \alpha_t \hat{\mathbf{s}}_{l_{k,t},k}||_2^2\right]$, using (5.22), the Lagrangian function for the MMSE problem (5.20a) considering the power constraint (5.20b) with the Lagrangian multiplier η_t results in

$$L\left(\mathbf{G}_t, \alpha_t, \eta_t\right) = \sum_{k=1}^{K} \left(v_{k,t} F(\mathbf{G}_t, \alpha_t, k, t)\right) - \eta_t \left(\mathrm{tr}\left(\mathbf{G}_t \boldsymbol{\Upsilon} \mathbf{G}_t^{\mathrm{H}}\right) - P_{\mathrm{RS}}\right). \tag{5.23}$$

From the Lagrangian function, the KKT conditions can be derived and η_t can be computed, which is presented in detail in Appendix A.4. The KKT conditions can be written as

$$\frac{\partial L}{\partial \mathbf{G}_t} = \sum_{k=1}^{K} v_{k,t} \frac{\partial F(\mathbf{G}_t, \alpha_t, k, t)}{\partial \mathbf{G}_t} - \eta_t \ \mathbf{G}_t^* \boldsymbol{\Upsilon}^{\mathrm{T}} = \mathbf{0}, \tag{5.24a}$$

$$\frac{\partial L}{\partial \alpha_t} = \sum_{k=1}^{K} v_{k,t} \frac{\partial F(\mathbf{G}_t, \alpha_t, k, t)}{\partial \alpha_t} = 0, \tag{5.24b}$$

$$\eta_t \left(\mathrm{tr}\left(\mathbf{G}_t \boldsymbol{\Upsilon} \mathbf{G}_t^{\mathrm{H}}\right) - P_{\mathrm{RS}}\right) = 0, \tag{5.24c}$$

and the Lagrangian multiplier η_t results in

$$\eta_t = -\frac{|\alpha_t|^2 \sigma_n^2 \sum_{k=1}^{K} v_{k,t} \mathrm{tr}\left(\mathbf{D}_{k,t} \mathbf{D}_{k,t}^{\mathrm{H}}\right)}{P_{\mathrm{RS}}}. \tag{5.25}$$

Now, considering the derivations in Appendix A.4, a unique solution can be obtained for $|\alpha_t|^2$. Thus, restricting α_t to be positive real-valued, a unique solution can be obtained for problem (5.20).

Considering the first KKT condition (5.24a), a matrix $\mathbf{K}_t$ is defined as

$$\begin{aligned}\mathbf{K}_t = &\sum_{k=1}^{K} v_{k,t} \boldsymbol{\Upsilon}^{(l_{k,t})^{\mathrm{T}}} \otimes \left(\mathbf{H}_k^* \mathbf{D}_{k,t}^{\mathrm{H}} \mathbf{D}_{k,t} \mathbf{H}_k^{\mathrm{T}}\right) \\ &+ \sum_{k=1}^{K} v_{k,t} \sum_{i=1}^{N_{\mathrm{o},k,t}} \boldsymbol{\Upsilon}^{(o_{i,k,t})^{\mathrm{T}}} \otimes \left(\mathbf{H}_k^* \mathbf{D}_{k,t}^{\mathrm{H}} \mathbf{D}_{k,t} \mathbf{H}_k^{\mathrm{T}}\right) \\ &+ \sum_{k=1}^{K} v_{k,t} \left[\sigma_{n,\mathrm{RS}}^2 \mathbf{I}_L \otimes \left(\mathbf{H}_k^* \mathbf{D}_{k,t}^{\mathrm{H}} \mathbf{D}_{k,t} \mathbf{H}_k^{\mathrm{T}}\right)\right] + \boldsymbol{\Upsilon}^{\mathrm{T}} \otimes \frac{\sigma_n^2 \sum_{k=1}^{K} v_{k,t} \mathrm{tr}\left(\mathbf{D}_{k,t} \mathbf{D}_{k,t}^{\mathrm{H}}\right)}{P_{\mathrm{RS}}} \mathbf{I}_L.\end{aligned} \tag{5.26}$$

Now, an analytical solution can be obtained for the WMMSE-ANC relay transceive filter which solves problem (5.20) using (5.24), (5.25) and (5.26). With the auxiliary matrix $\widetilde{\mathbf{G}}_t$ given by

$$\widetilde{\mathbf{G}}_t = \mathrm{vec}_{L,L}^{-1} \left(\mathbf{K}_t^{-1} \ \mathrm{vec}\left(\sum_{k=1}^{K} v_{k,t} \mathbf{H}_k^* \mathbf{D}_{k,t}^{\mathrm{H}} \mathbf{Q}_{l_{k,t}}^{\mathrm{H}} \mathbf{H}_{l_{k,t}}^{\mathrm{H}}\right)\right), \tag{5.27}$$

and using

$$\alpha_t = \sqrt{\frac{\mathrm{tr}\left(\widetilde{\mathbf{G}}_t \boldsymbol{\Upsilon} \widetilde{\mathbf{G}}_t^{\mathrm{H}}\right)}{P_{\mathrm{RS}}}}, \tag{5.28}$$

the WMMSE-ANC filter at RS which solves problem (5.20) is given by

$$\mathbf{G}_t = \frac{1}{\alpha_t} \cdot \mathrm{vec}_{L,L}^{-1} \left(\mathbf{K}_t^{-1} \ \mathrm{vec} \left(\sum_{k=1}^{K} \upsilon_{k,t} \mathbf{H}_k^* \mathbf{D}_{k,t}^{\mathrm{H}} \mathbf{Q}_{l_{k,t}}^{\mathrm{H}} \mathbf{H}_{l_{k,t}}^{\mathrm{H}} \right) \right). \tag{5.29}$$

The derived WMMSE-ANC relay transceive filter minimizes the weighted MSE for given Tx and Rx filters at the nodes considering that the nodes can perform self-interference cancellation, temporal Rx processing and SIC.

5.3.5 Joint Spatial Filter Design at Nodes and at RS

The WMMSE-ANC relay transceive filter depends on the Tx and Rx filters at the nodes. To jointly design the relay transceive filter and the Rx filters at the nodes, an alternating MMSE based optimization is proposed as follows:

1) Compute the relay transceive filters $\mathbf{G}_t$ according to (5.29) assuming $\mathbf{Q}_k = \sqrt{\frac{P_{\mathrm{node}}}{M}} \mathbf{I}_M$ and $\mathbf{D}_{k,t} = \mathbf{I}_M$, $k = 1, 2, ..., K$ and $t = 2, 3, ..., N$.

2) Compute MMSE Rx filters at the nodes considering the relay transceive filters of step 1). To compute the Rx filters, the overall MIMO channel for the transmission from $\mathrm{S}_{l_{k,t}}$ to S_k and the overall noise at S_k in time slot t can be written as

$$\mathbf{H}_{\mathrm{ov},k,t} = \mathbf{H}_k^{\mathrm{T}} \mathbf{G}_t \mathbf{H}_{l_{k,t}} \mathbf{Q}_{l_{k,t}}, \tag{5.30}$$

$$\mathbf{N}_{k,t} = \sigma_{\mathrm{n,RS}}^2 \mathbf{H}_k^{\mathrm{T}} \mathbf{G}_t \mathbf{G}_t^{\mathrm{H}} \mathbf{H}_k^* + \mathbf{I}_M \sigma_{\mathrm{n}}^2, \tag{5.31}$$

respectively. Thus, the MMSE Rx filters based on the derivations for conventional MIMO Rx filters [Joh04] can be written as

$$\mathbf{D}_{k,t} = \frac{\mathbf{H}_{\mathrm{ov},k,t}^{\mathrm{H}} \left(\mathbf{H}_{\mathrm{ov},k,t} \mathbf{H}_{\mathrm{ov},k,t}^{\mathrm{H}} + \mathbf{N}_{k,t} \right)^{-1}}{||\mathbf{H}_{\mathrm{ov},k,t}^{\mathrm{H}} \left(\mathbf{H}_{\mathrm{ov},k,t} \mathbf{H}_{\mathrm{ov},k,t}^{\mathrm{H}} + \mathbf{N}_{k,t} \right)^{-1} ||_{\mathrm{F}}}. \tag{5.32}$$

3) Continue from step 1) using the Rx filters of step 2) until convergence.

5.4 Transmit Strategies

5.4.1 Introduction

In this section, two Tx strategies are proposed. Conventional Tx strategies for multi-way relaying typically exploit either the spatial processing capabilities of RS [AK10a, AK11a] or the temporal processing capabilities of the nodes [CZ12]. However, an efficient combination of the spatial processing capabilities of RS with the temporal processing capabilities of the nodes by utilizing ANC can significantly increase the performance. In the following, two Tx strategies are introduced which utilize ANC to efficiently combine the spatial processing capabilities of RS with the temporal processing capabilities of the nodes. First, a Tx strategy, termed NCMW Tx strategy, is proposed which selects the signals which are retransmitted in each BC such that the self-interference cancellation and the SIC capabilities of the nodes are exploited. As explained in Section 5.1, this Tx strategy exploits the temporal processing capabilities of the nodes by considering known-interference cancellation. Secondly, a Tx strategy, termed NCJP Tx strategy, is proposed which selects the signals which are retransmitted in each BC such that the capability of the nodes to jointly process the received signals of all BC phases is exploited. Additionally, this Tx strategy also exploits the self-interference cancellation and SIC capabilities of the nodes.

To provide a general overview, one cycle of the proposed Tx strategies is illustrated in Fig. 5.4. In the MAC phase, all nodes simultaneously transmit to RS. Afterwards, the processing at RS is performed in three steps. First, the desired signals, the SKISs, the RMSs and the suppressed signals are determined at RS for the different BC phases based on the considered Tx strategy. Secondly, the ANC aware relay transceive filters for the different BC phases and the spatial Rx filters at the nodes are computed based on the preselected signals. Thirdly, RS retransmits the received signals to the nodes in $N-1$ different BC phases after linearly processing these signals with the corresponding relay transceive filter for each BC phase. Using $\mathbf{y}_{\mathrm{RS},c}$ of (5.1), the retransmitted linear combination in time slot t is given by $\mathbf{G}_t\mathbf{y}_{\mathrm{RS}}$. Finally, the temporal processing capabilities of the nodes are exploited based on the considered Tx strategy to estimate all desired signals after performing self-interference cancellation and SIC.

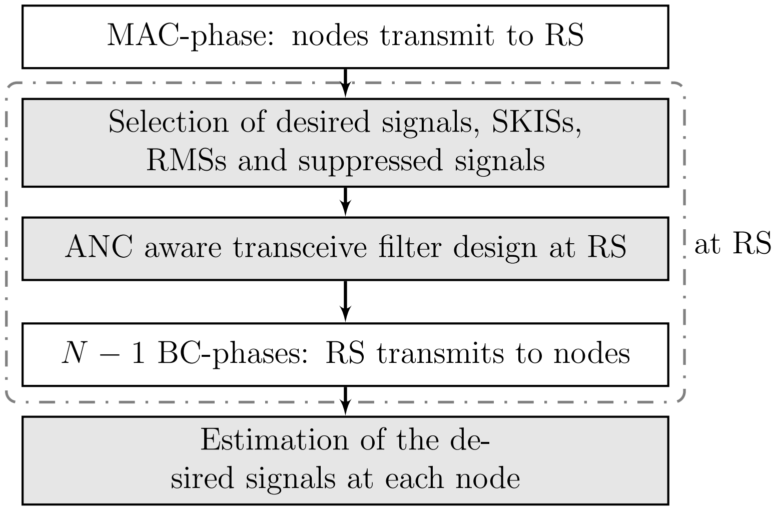

Figure 5.4. Overview of one cycle of the Tx strategies.

5.4.2 Network Coded Multi-Way (NCMW) Transmit Strategy

In this section, the NCMW Tx strategy is introduced. This strategy is based on the hybrid uni-/multicasting strategy of [AK11a]. In each BC phase, one signal is unicasted to one node and another signal is multicasted to the remaining nodes of the group. However, in contrast to the hybrid uni-/multicasting strategy of [AK11a], the unicasted signal is not considered as interference at the nodes which receive the multicasted signal and vice versa. Thus, the transmission of the unicasted signal to the nodes which intentionally receive the multicasted signal and vice versa has not to be suppressed by the relay transceive filter. For the proposed NCMW Tx strategy, the capability of the nodes to jointly process the received signals of all BC phases is not exploited. Thus, the temporal processing matrix $\mathbf{W}_k$ is assumed to be $\mathbf{W}_k = \mathbf{I}_{M(N-1)}$, $k = 1, 2, ...K$. Nevertheless, the temporal processing capabilities of the nodes are exploited due to considering known-interference cancellation. In the following, the desired signals, the SKISs, the suppressed signals and the RMSs at the nodes are introduced.

First, the desired signals at the nodes in each BC phase are considered. For the NCMW Tx strategy, it is assumed that the signal $\mathbf{s}_{k+t-1}$, termed MC signal, is desired at the nodes S_k, S_{k+1}, ...,S_{k+t-2},S_{k+t},..., S_{k+N-1} of group g in time slot t, where $k = 1 + (g-1)N$. Additionally, the signal $\mathbf{s}_k$, termed UC signal, is desired at node S_{k+t-1} in time slot t. Thus, the index of the desired signal at node S_j in time slot t is given by

$$l_{j,t} = \begin{cases} k & \text{if} \quad j = k+t-1, \\ k+t-1 & \text{if} \quad j \neq k+t-1, \end{cases} \tag{5.33}$$

where $j = k, k+1, ..., k+N-1$ are the indices of the nodes of the g^{th} group. Due to changing the MC signal in each BC phase, every transmit signal $\mathbf{s}_j$ is desired at each node within group g in one of the BC phases. Using this approach, the relay Tx power is focused on as few signals as possible because only one MC and one UC signal are desired within each group in each BC phase. For the retransmission of the desired signals, ANC is exploited. By ANC, it is meant that instead of spatially separating both desired signals as considered in [AK10a, AK11a], the relay transceive filter spatially superimposes the desired UC and MC signals in each BC phase such that the individual signals can be recovered by utilizing the temporal processing capabilities of the nodes. Optimizing the selection of the UC signal can further improve the performance.

Secondly, the SKISs at the nodes in each BC phase are considered. For the NCMW Tx strategy, the UC signal shall not be considered as interference at the nodes which receive the MC signal and vice versa. To achieve this, the UC signal will be decoded first at each node and known-interference cancellation will be applied before estimating the MC signals. Thus, the UC signal is considered as SKIS at the nodes which receive the MC signal. Furthermore, to exploit the self-interference cancellation capabilities of the nodes, the signal $\mathbf{s}_j$ is considered as SKIS at node S_j. This ensures that the MC signal is not considered as interference at the node which receives the UC signal. Moreover, to further exploit the known-interference cancellation capabilities of the nodes, MC signals of previous time slots are considered as SKISs at the nodes which receive the MC signal in time slot t. To summarize, the indices of the SKISs at node S_j in time slot t are given by

$$\mathcal{SKIS}_{j,t} = \begin{cases} \{k+1, k+2, ..., k+t-2, j\} & \text{if} \quad j = k+t-1, \\ \{k, k+1, ..., k+t-2, j\} & \text{if} \quad j \neq k+t-1, \end{cases} \tag{5.34}$$

where $j = k, k+1, ..., k+N-1$.

Thirdly, the suppressed signals are considered. For the NCMW Tx strategy, all signals which are neither considered as a desired signal nor as self- or known-interference signal at node S_j in time slot t are considered as a suppressed signal. Thus, the indices of the suppressed signals at node S_j in time slot t are given by

$$\mathbf{o}_{j,t} = \begin{cases} [1, 2, ..., k-1, k+t, ..., j-1, j+1, ..., K] & \text{if} \quad j \geq k+t, \\ [1, 2, ..., k-1, k+1, ..., j-1, j+1, ..., K] & \text{if} \quad j = k+t-1, \\ [1, 2, ..., k-1, k+t, ..., K] & \text{if} \quad j < k+t-1, \end{cases} \tag{5.35}$$

where $j = k, k+1, ..., k+N-1$.

RMSs are not considered for the proposed NCMW Tx strategy because each signal is either considered as a desired signal, as a SKIS or as a suppressed signal.

Based on the aforementioned consideration of SKISs, the decoding order for the NCMW Tx strategy at node S_j is defined as

$$\mathbf{q}_j = \begin{cases} (k+1, k+2, ..., k+N-1) & \text{if} \quad j = k, \\ (k, k+1, ..., j-1, j+1, ..., k+N-1) & \text{if} \quad j \neq k, \end{cases} \tag{5.36}$$

where $j = k, k+1, ..., k+N-1$. Thus, the subset $\mathcal{SIC}_{l,j}$, which is considered in (5.8) to compute the expected interference power, is given by

$$\mathcal{SIC}_{l,j} = \begin{cases} (j, q_{j,1}, q_{j,2}, ..., q_{j,l-k-1}) & \text{if} \quad j = k, \\ (j, q_{j,1}, q_{j,2}, ..., q_{j,l-k}) & \text{if} \quad j \neq k, \end{cases} \tag{5.37}$$

where $q_{j,i}$ is the i^{th} element of $\mathbf{q}_j$ (5.36) and where $l = k, k+1, ..., k+N-1$ is the index of the transmit signal which shall be estimated at S_j.

To illustrate the proposed NCMW Tx strategy, a scenario consisting of $G = 1$ group with $N = 4$ nodes per group is exemplarily considered. The group consists of the nodes S_1, S_2, S_3 and S_4. Similar to the hybrid uni-/multicasting strategy of [AK11a], in time slot t, the UC signal $\mathbf{s}_1$ is desired at S_t and the MC signal $\mathbf{s}_t$ is desired at the remaining nodes of the group. However, as already mentioned before, the UC signal is not considered as interference at the nodes which receive the MC signal and vice versa due to exploiting ANC. The considered desired signals, SKISs and suppressed signals are summarized in Table 5.1. In each BC phase $t = 2, 3, ..., N$, it is assumed that the nodes can subtract the back propagated self-interferences. Furthermore, it is assumed that the nodes can subtract known-interferences, i.e., interferences which are known at the nodes due to successful decoding of the corresponding signals in a previous time slot, e.g., each node knows the multicasted signals of the previous time slots because successful decoding is assumed. Additionally, the unicasted signal is always assumed to be known at the nodes which receive the multicasted signal. This assumption is valid because the unicasted signal is received and estimated at S_k in time slot $t = k$ and afterwards known-interference cancellation can be applied to the signals received in previous or subsequent time slots.

For the computation of the relay transceive filter in Section 5.3.4, weighting parameters $v_{k,t}$, $0 \leq v_{k,t} \leq 2$, have been considered, $k = 1, 2, ..., K$, $t = 2, 3, ..., N$. To achieve high sum rates (5.12), the weighting parameters $v_{k,t}$ which achieve the highest sum rate according to (5.12) have to be determined. However, this would require a joint

Table 5.1. NCMW Tx strategy for a group of $N = 4$ nodes.

	receiving node	S_1	S_2	S_3	S_4
$t = 2$	desired signal	$\mathbf{s}_2$	$\mathbf{s}_1$	$\mathbf{s}_2$	$\mathbf{s}_2$
	self-interference	$\mathbf{s}_1$	$\mathbf{s}_2$	$\mathbf{s}_3$	$\mathbf{s}_4$
	known-interference	-	-	$\mathbf{s}_1$	$\mathbf{s}_1$
	suppressed signals	$\mathbf{s}_3$, $\mathbf{s}_4$	$\mathbf{s}_3$, $\mathbf{s}_4$	$\mathbf{s}_4$	$\mathbf{s}_3$
$t = 3$	desired signal	$\mathbf{s}_3$	$\mathbf{s}_3$	$\mathbf{s}_1$	$\mathbf{s}_3$
	self-interference	$\mathbf{s}_1$	$\mathbf{s}_2$	$\mathbf{s}_3$	$\mathbf{s}_4$
	known-interference	$\mathbf{s}_2$	$\mathbf{s}_1$	-	$\mathbf{s}_1$, $\mathbf{s}_2$
	suppressed signals	$\mathbf{s}_4$	$\mathbf{s}_4$	$\mathbf{s}_2$, $\mathbf{s}_4$	-
$t = 4$	desired signal	$\mathbf{s}_4$	$\mathbf{s}_4$	$\mathbf{s}_4$	$\mathbf{s}_1$
	self-interference	$\mathbf{s}_1$	$\mathbf{s}_2$	$\mathbf{s}_3$	$\mathbf{s}_4$
	known-interference	$\mathbf{s}_2$, $\mathbf{s}_3$	$\mathbf{s}_1$, $\mathbf{s}_3$	$\mathbf{s}_1$, $\mathbf{s}_2$	-
	suppressed signals	-	-	-	$\mathbf{s}_2$, $\mathbf{s}_3$

optimization of all weighting parameters which has a high computational complexity. Thus, to reduce the computational complexity, a suboptimal low-complexity approach is proposed. In this approach, the joint adjustment of the weighting parameters is separated and performed iteratively as follows:

First, the relay transceive filter is initialized assuming all weighting parameters $v_{k,t} = 1$, $k = 1, 2, ..., K$, $t = 2, 3, ..., N$. Furthermore, the achievable data rates $C_{l,k}$ given by

$$C_{l,k} = \sum_{m=1}^{M} C_{l,k,m}, \tag{5.38}$$

with $C_{l,k,m}$ of (5.10) are computed.

Secondly, the weighting parameters are adjusted to increase the achievable sum rate (5.12). Considering the achievable data rates (5.38), the index $k_{\min,l} = \arg \min_{\forall k \in \mathcal{S}_l} C_{l,k}$ which corresponds to the data rate $C_{l,k_{\min,l}}$ which limits the sum rate is determined for each Tx signal $\mathbf{s}_l$, $l = 1, 2, ..., K$. Furthermore, the index $k_{\max,l} = \arg \max_{\forall k \in \mathcal{S}_l} C_{l,k}$ which corresponds to the maximum of the achievable data rates $C_{l,k_{\max,l}}$ is determined for each Tx signal $\mathbf{s}_l$. Using the computed indices, the precise steps to adjust the weighting parameters are as follows:

- For the MC signals of group g with the indices $l = 2 + (g-1)N, 3 + (g-1)N, ..., gN$, $v_{k_{\min,l},l-j}$ with $j = 1 + (g-1)N$ is increased by the same value than $v_{k_{\max,l},l-j}$ is decreased until

$$v_{k_{\min,l},l-j} > 2 - \delta \text{ or} \tag{5.39a}$$

$$v_{k_{\max,l},l-j} < \delta, \tag{5.39b}$$

or until

$$C_{l,k_{\max,l}} - \epsilon < C_{l,k_{\min,l}} < C_{l,k_{\max,l}} + \epsilon, \tag{5.40}$$

where ϵ can be selected according to the required accuracy and where $0 < \delta < 1$ ensures that the MSE for this direction of transmission has a sufficient impact on the considered MMSE based relay transceive filter design. In this thesis, $\delta = 0.05$ is selected based on numerical results. If $v_{k_{\min,l},l-j} > 2 - \delta$, $v_{k_{\max,l},l-j}$ is decreased until either (5.39b) or (5.40) is fulfilled. If $v_{k_{\max,l},l-j} < \delta$, $v_{k_{\min,l},l-j}$ is increased until either (5.39a) or (5.40) is fulfilled.

- For the UC signal of group g with the index $l = 1 + (g-1)N$, $v_{k_{\min,l},k_{\min,l}}$ is increased by the same value than $v_{k_{\max,l},k_{\max,l}}$ is decreased until

$$v_{k_{\min,l}k_{\min,l}} > 2 - \delta \text{ or} \tag{5.41a}$$

$$v_{k_{\max,l},k_{\max,l}} < \delta, \tag{5.41b}$$

or until (5.40) is fulfilled. If $v_{k_{\min,l},k_{\min,l}} > 2 - \delta$, $v_{k_{\max,l},k_{\max,l}}$ is decreased until either (5.41b) or (5.40) is fulfilled. If $v_{k_{\max,l},k_{\max,l}} < \delta$, $v_{k_{\min,l},k_{\min,l}}$ is increased until either (5.41a) or (5.40) is fulfilled.

To adjust the weighting parameters, the bisection method can be applied. The relay transceive filter is recalculated after each update of the weighting parameters. Furthermore, the achievable data rates $C_{l,k}$ (5.38) are recalculated after each update of the relay transceive filter.

Thirdly, the second step is repeated for a finite number of times considering the updated weighting parameters. The number of repetitions depends on the required accuracy and in this thesis, three repetitions are considered.

5.4.3 Network Coded Joint Processing (NCJP) Transmit Strategy

In this section, the NCJP Tx strategy is introduced. The proposed NCJP Tx strategy exploits ANC and efficiently combines the spatial processing capabilities of RS and the temporal processing capabilities of the nodes. To exchange all messages in $N-1$ BC phases, different linear combinations of the transmitted signals have to be received at each node in each BC phase. The proposed NCJP Tx strategy is based on retransmitting spatially processed linear combinations of all received signals such that the spatial processing capabilities of RS are utilized efficiently and the temporal processing capabilities of the nodes can be exploited. In the following, the desired signals, the suppressed signals, the SKISs and the RMSs at the nodes are introduced.

First, the desired signals at the nodes in each BC phase are considered. Similar to the Tx strategies presented in [AK10a, AK11a], it is proposed that the signal $\mathbf{s}_{m_{g,t}}$, termed MC signal, is desired at all nodes S_k, S_{k+1},...,$\mathrm{S}_{m_{g,t}-1}$,$\mathrm{S}_{m_{g,t}+1}$,...,S_{k+N-1} of group g in time slot t, where $m_{g,t}$ is the index of the MC signal for group g in time slot t, $k = 1 + (g-1)N$. This MC signal changes in each BC phase and the indices of the different MC signals are contained in the vector $\mathbf{m}_g = (m_{g,2}, ..., m_{g,N})$. Additionally, it is proposed that the transmit signal $\mathbf{s}_{u_g}$, termed UC signal, is desired at node $\mathrm{S}_{m_{g,t}}$, where u_g is the index of the UC signal of group g, $u_g \neq m_{g,t} \forall t$. The UC signal is the same in all BC phases. To summarize, the index of the desired signal at node S_j in time slot t is given by

$$l_{j,t} = \begin{cases} u_g & \text{if} \quad j = m_{g,t}, \\ m_{g,t} & \text{if} \quad j \neq m_{g,t}, \end{cases} \tag{5.42}$$

where $j = k, k+1, ..., k+N-1$. The selection of the MC and the UC signals for each group is described at the end of this section. Due to changing the MC signal in each BC phase, every transmit signal $\mathbf{s}_{j,c}$ is desired at each node of group g in one of the BC phases neglecting node S_j which transmits this signal. Using this approach, the Tx power at RS is focused on as few signals as possible because only one MC and one UC signal are desired at all nodes of group g in each BC phase. For the retransmission of the desired signals, ANC is exploited because instead of spatially separating both desired signals as considered in [AK10a, AK11a], the desired UC and MC signals are spatially superimposed in each BC phase and the individual signals are recovered by utilizing the temporal processing capabilities of the nodes.

Secondly, the suppressed signals are considered. For the NCJP Tx strategy, one signal within each group is considered as a suppressed signal in each BC phase. The consideration of one suppressed signal within each group increases the temporal processing gain at the nodes in case of $N > 2$ because it reduces the linear dependencies between the retransmitted signals of the different BC phases. Suppressing more than one signal or all remaining signals as considered in [AK10a, AK11a] is not beneficial because it would reduce the temporal processing gain and would require more antennas at RS for spatially separating desired and suppressed signals. Thus, it is proposed that only one signal $\mathbf{s}_{o_{g,t}}$ is considered as suppressed signal at each node within group g in time slot t, where $o_{g,t}$ is the index of the suppressed signal. Additionally, the transmit signals of the nodes which belong to other groups have to be suppressed. Thus, the indices of the suppressed signals at node S_j in time slot t are given by

$$\mathbf{o}_{j,t} = [1, 2, ..., k-1, o_{g,t}, k+N, ..., K], \tag{5.43}$$

where $j = k, k+1, ..., k+N-1$ and $k = 1+(g-1)N$. The suppressed signal within each group $\mathbf{s}_{o_{g,t}}$ changes in each time slot and the indices of the suppressed signals within each group are contained in the vector $\mathbf{o}_g = (o_{g,2}, ..., o_{g,N}), u_g \neq o_{g,t} \neq m_{g,t} \forall t = 2, ..., N$. The selection of the indices $o_{g,t}$ for each group g and each time slot t is described at the end of this section.

Thirdly, the SKISs are considered. Considering the nodes of group g, $\mathbf{s}_{j,c}$ is considered as self-interference at node S_j in time slot t if $j \neq o_{g,t}$, where $j = k, k+1, ..., k+N-1$ with $k = 1 + (g-1)N$. Known-interference signals are not considered for the NCJP Tx strategy. Thus, the indices of the SKISs at node S_j are given by

$$\mathcal{SKIS}_{j,t} = \begin{cases} \emptyset & \text{if} \quad j = o_{g,t}, \\ \{j\} & \text{if} \quad j \neq o_{g,t}. \end{cases} \tag{5.44}$$

Fourthly, the RMSs at the nodes are considered. The signal $\mathbf{s}_r$ is considered as RMS at node S_j in time slot t if

$$\{r\} \cap \{l_{j,t}, \mathbf{o}_{j,t}, j\} = \emptyset \quad \text{for } j \neq o_{g,t}, \tag{5.45a}$$

$$\{r\} \cap \{l_{j,t}, \mathbf{o}_{j,t}\} = \emptyset \quad \text{for } j = o_{g,t}, \tag{5.45b}$$

where $j = k, k+1, ..., k+N-1$ and $k = 1+(g-1)N$. The proposed approach exploits ANC and instead of spatially separating the RMSs from the desired signals as considered in [AK10a, AK11a], spatially processed linear combinations of these signals are retransmitted by RS in each BC phase.

An exemplary overview of the proposed NCJP Tx strategy is given in Table 5.2 for a scenario consisting of $G = 1$ group with $N = 4$ nodes. At node S_k, the signal $\mathbf{s}_k$ is self-interference. Thus, it is not shown in Table 5.2 because it can be perfectly canceled before performing temporal Rx processing. In this example, the UC signal $\mathbf{s}_1$ is desired at node S_t in time slot t which is marked by u in the table. Furthermore, the MC signal $\mathbf{s}_{m_{1,t}}$ is desired at the remaining nodes in each time slot which is marked by m in the table. Additionally, the signal $\mathbf{s}_{o_{1,t}}$ is considered as suppressed signal at each node in time slot t. The suppressed signal is marked by o in the table. The RMSs which are only considered with respect to the power constraint at RS in each BC phase are marked by $*$. The individual signals are recovered at the nodes by performing joint temporal Rx processing over the received signals of all BC phases as described in Section 5.3.2.

Table 5.2. Proposed NCJP Tx strategy for a multi-way group of $N = 4$ nodes, $u_1 = 1$, $\mathbf{m}_1 = (2, 3, 4)$, $\mathbf{o}_1 = (3, 4, 2)$.

	signals at S_1			signals at S_2			signals at S_3			signals at S_4		
t	$\mathbf{s}_2$	$\mathbf{s}_3$	$\mathbf{s}_4$	$\mathbf{s}_1$	$\mathbf{s}_3$	$\mathbf{s}_4$	$\mathbf{s}_2$	$\mathbf{s}_1$	$\mathbf{s}_4$	$\mathbf{s}_2$	$\mathbf{s}_3$	$\mathbf{s}_1$
2	m	o	$*$	u	o	$*$	m	$*$	$*$	m	o	$*$
3	$*$	m	o	$*$	m	o	$*$	u	o	$*$	m	$*$
4	o	$*$	m	$*$	$*$	m	o	$*$	m	o	$*$	u

For the NCJP Tx strategy, the weighting parameters $v_{k,t}$ which have been considered for the computation of the relay transceive filter in Section 5.3.4 are computed as described for the NCMW Tx strategy in Section 5.4.2. The only difference which has to be considered is that the indices of the UC and of the MC signals are different for the NCJP Tx strategy compared to the NCMW Tx strategy. Thus, the indices of the corresponding weighting parameters are also different which has to be considered.

SIC Decoding Order

To exploit the SIC capabilities of the nodes for the relay transceive filter design, a fixed decoding order is required. For the proposed NCJP Tx strategy, the following decoding order is proposed.

First, the UC signal is decoded at the nodes which receive the UC signal as a desired signal in one of the BC phases because due to not suppressing the UC signal in any of the BC phases, the average receive power of the UC signal is higher than that of any MC signal over all BC phases. Furthermore, the UC signal interferes with every MC signal because it is either considered as desired signal, as SKIS or as RMS within each group for the spatial processing at RS.

Secondly, the different MC signals are decoded at all nodes. The different MC signals are received equally strong on average. However, the signal to interference ratio increases on average in each decoding step which increases the achievable data rates. Thus, it is proposed that the decoding order of the MC signals should be equal at all nodes because the minimum over all maximum achievable data rates from one node to all other nodes within the same group limits the maximum achievable multi-way rate (5.11). Thus, it is proposed that the MC signals are decoded in decreasing order of the respective indices of the transmitting nodes.

In summary, the decoding order for nodes S_j, $j \neq u_g$, is

$$\mathbf{q}_{j,c} = (u_g, a, b, ..., c), \tag{5.46}$$

where $gN \geq a > b > c \geq 1 + (g-1)N$, $\{a, b, ..., c\} \cap \{j, u_g\} = \emptyset$ describe the indices of the nodes in decreasing order excluding j and u_g. For node S_{u_g}, the decoding order is

$$\mathbf{q}_{u_g} = (a, b, ..., c), \tag{5.47}$$

where $gN \geq a > b > c \geq 1 + (g-1)N$, $\{a, b, ..., c\} \cap \{u_g\} = \emptyset$.

Thus, the subset $\mathcal{SIC}_{l,j}$, which is considered in (5.8) to compute the expected interference power, is given by

$$\mathcal{SIC}_{l,j} = \begin{cases} (j, q_{j,1}, q_{j,2}, ..., q_{j,N-1-l+k}) & \text{if} \quad j = u_g \text{ and } l > j, \\ (j, q_{j,1}, q_{j,2}, ..., q_{j,N-2-l+k}) & \text{if} \quad j = u_g \text{ and } l < j, \\ (j) & \text{if} \quad j \neq u_g \text{ and } l = u_g, \\ (j, u_g, q_{j,2}, q_{j,3}, ..., q_{j,N-2-l+k}) & \text{if} \quad j \neq u_g \text{ and } l < u_g \text{ and } l < j, \\ (j, u_g, q_{j,2}, q_{j,3}, ..., q_{j,N-1-l+k}) & \text{if} \quad j \neq u_g \text{ and } l > u_g \text{ and } l < j, \\ (j, u_g, q_{j,2}, q_{j,3}, ..., q_{j,N-1-l+k}) & \text{if} \quad j \neq u_g \text{ and } l < u_g \text{ and } l > j, \\ (j, u_g, q_{j,2}, q_{j,3}, ..., q_{j,N-l+k}) & \text{if} \quad j \neq u_g \text{ and } l > u_g \text{ and } l > j, \end{cases} \tag{5.48}$$

where $q_{j,i}$ is the i^{th} element of $\mathbf{q}_j$ using (5.46) and (5.47) and where $l = k, k+1, ..., k+N-1$ is the index of the transmit signal which shall be estimated at S_j, $l \neq j$.

Selection of UC signal

The selection of the UC signals $\mathbf{s}_{u_g}$ has an impact on the achievable data rates. To determine the UC signal $\mathbf{s}_{u_g}$ which has to be selected to achieve the highest sum rate, an exhaustive search over the signals transmitted by all nodes within each group has to be performed. However, this has a high computational complexity. Thus, a suboptimal low-complexity approach is proposed which is based on the cross-correlations between the different channels $\mathbf{H}_j$ over which the signals $\mathbf{s}_j$ are transmitted from the nodes of group g to RS, $j = k, k+1, ..., k+N-1$ with $k = 1 + (g-1)N$. The intention of the suboptimal approach is to select a UC signal which is transmitted over a channel which is highly correlated with all other channels of group g because the UC signal is either considered as desired signal, as SKIS or as RMS within each group for the relay transceive filter design. The sum of the cross-correlations between the channel $\mathbf{H}_j$ and all other channels within group g is given by

$$c_j = \sum_{l=k}^{k+N-1} \frac{||\mathbf{H}_j^{\mathrm{H}} \mathbf{H}_l||_{\mathrm{F}}^2}{||\mathbf{H}_j^{\mathrm{H}} \mathbf{H}_j||_{\mathrm{F}} ||\mathbf{H}_l^{\mathrm{H}} \mathbf{H}_l||_{\mathrm{F}}}. \tag{5.49}$$

Using these cross-correlations, the index u_g of the UC signal of group g is determined according to $u_g = \arg\max_j c_j$.

Selection of MC and suppressed signals

The selection of the MC signals $\mathbf{s}_{m_{g,t}}$ with respect to the selection of the suppressed signals $\mathbf{s}_{o_{g,t}}$ influences the achievable MSE of (5.18a) and thus, influences the achievable data rates. Thus, an approach to obtain a suitable sorting of $\mathbf{m}_g$ and $\mathbf{o}_g$ which contain the indices of the signals which should be multicasted and suppressed within group g, respectively, is introduced. To achieve a low MSE for the MC signal, the correlation between the channel $\mathbf{H}_{m_{g,t}}$ over which the MC signal is transmitted from $\mathrm{S}_{m_{g,t}}$ to RS and the channel $\mathbf{H}_{o_{g,t}}$ over which the suppressed signal $\mathbf{s}_{o_{g,t}}$ is transmitted from $\mathrm{S}_{o_{g,t}}$ to RS should be as low as possible in each BC phase. Without loss of generality, it is proposed to keep the sorting of $\mathbf{o}_g$ fixed and to change the sorting of $\mathbf{m}_g$ to achieve low correlations. Thus, $\mathbf{m}_g$ can be sorted according to

$$\mathbf{m}_g = \arg\min_{\mathbf{m}_g} \sum_{t=2}^{N} \frac{||\mathbf{H}_{m_{g,t}}^{\mathrm{H}} \mathbf{H}_{o_{g,t}}||_{\mathrm{F}}^2}{||\mathbf{H}_{m_{g,t}}^{\mathrm{H}} \mathbf{H}_{m_{g,t}}||_{\mathrm{F}} ||\mathbf{H}_{o_{g,t}}^{\mathrm{H}} \mathbf{H}_{o_{g,t}}||_{\mathrm{F}}}. \tag{5.50}$$

To obtain the sorting of $\mathbf{m}_g$ which minimizes (5.50) is a combinatorial problem. Thus, a stepwise low-complexity algorithm is proposed to obtain a suitable sorting of $\mathbf{m}_g$ for the proposed NCJP Tx strategy as follows:

1) Define a set $\mathcal{N}_{\mathrm{MC}}$ which contains all indices of $\mathbf{m}_g$.

2) For $t = 2$ to $t = N$:
Set $m_{g,t} = \arg\min_{m_i} \frac{||\mathbf{H}_{m_i}^{\mathrm{H}} \mathbf{H}_{o_{g,t}}||_{\mathrm{F}}^2}{||\mathbf{H}_{m_i}^{\mathrm{H}} \mathbf{H}_{m_i}||_{\mathrm{F}} ||\mathbf{H}_{o_{g,t}}^{\mathrm{H}} \mathbf{H}_{o_{g,t}}||_{\mathrm{F}}}$, $m_i \in \mathcal{N}_{\mathrm{MC}}$, $i = 2, 3, ..., N$.

3) Remove $m_{g,t}$ from the set $\mathcal{N}_{\mathrm{MC}}$.

4) If $m_{g,N} = o_{g,N}$, perform a reallocation for $m_{g,N}$:
Set $a = \arg\min_{i} \frac{||\mathbf{H}_{m_{g,i}}^{\mathrm{H}} \mathbf{H}_{o_{g,N}}||_{\mathrm{F}}^2}{||\mathbf{H}_{m_{g,i}}^{\mathrm{H}} \mathbf{H}_{m_{g,i}}||_{\mathrm{F}} ||\mathbf{H}_{o_{g,N}}^{\mathrm{H}} \mathbf{H}_{o_{g,N}}||_{\mathrm{F}}}$, $i = 2, 3, ..., N-1$,
Set $b = m_{g,a}$, $m_{g,a} = m_{g,N}$ and $m_{g,N} = b$.

5.5 Performance Analysis

In this section, the performances of the Tx strategies presented in Section 5.4 are investigated through numerical simulations considering the filter designs presented in Section 5.3. All channels are assumed to be i.i.d. Rayleigh fading channels with zero-mean and unit variance and the noise variances at the nodes and at RS are assumed to be equal, i.e., $\sigma_{\mathrm{n,RS}}^2 = \sigma_{\mathrm{n}}^2$. All simulation results are averaged over 1000 independent channel realizations. The maximum Tx power at RS is set to be equal to the maximum Tx power at node S_1, i.e., $P_{\mathrm{RS}} = P_{\mathrm{node}}$. The ratio between the maximum Tx power P_{node} at the nodes and the noise level σ_{n}^2 is termed average SNR.

For the numerical simulations, four different configurations of the multi-group multi-way relaying scenario are investigated. In configurations A, B and C, a single-group multi-way relaying scenario is considered, i.e., $G = 1$. In configurations A and B, $N = 4$ nodes are considered. In configuration A, each node is equipped with $M = 1$ antenna and in configuration B each node is equipped with $M = 2$ antennas. In configuration C, $N = 10$ nodes are considered and each node is equipped with $M = 1$ antenna. In configuration D, a multi-group multi-way relaying scenario is considered. In this configuration, $G = 2$ groups with $N = 4$ single-antenna nodes per group are considered, i.e., $M = 1$. These four configurations are investigated because they enable a comprehensive comparison of the proposed filter designs and the proposed Tx strategies with other state of the art approaches.

For performance comparison, the following approaches are considered.

- MMSE-SIC: joint temporal Rx processing approach of [CZ12] considering random beamforming at RS,
- U/MC:ZF: hybrid uni-/multicasting Tx strategy of [AK11a] considering a ZF filter at RS,
- U/MC:MMSE: hybrid uni-/multicasting Tx strategy of [AK11a] considering an MMSE filter at RS,
- NCMW: proposed NCMW Tx strategy of Section 5.4.2 considering the proposed WMMSE-ANC relay transceive filter of Section 5.3.4 and the spatial filters at the nodes of Section 5.3.3,
- NCMW-Joint: proposed NCMW Tx strategy of Section 5.4.2 considering the joint spatial filter design approach at nodes and at RS of Section 5.3.5,
- NCJP: proposed NCJP Tx strategy of Section 5.4.3 considering the proposed WMMSE-ANC relay transceive filter of Section 5.3.4, the spatial filters at the nodes of Section 5.3.3 and the temporal Rx filters at the nodes of Section 5.3.2,
- NCJP-Joint: proposed NCJP Tx strategy of Section 5.4.3 considering the joint spatial filter design approach at nodes and at RS of Section 5.3.5 and the temporal Rx filters at the nodes of Section 5.3.2.

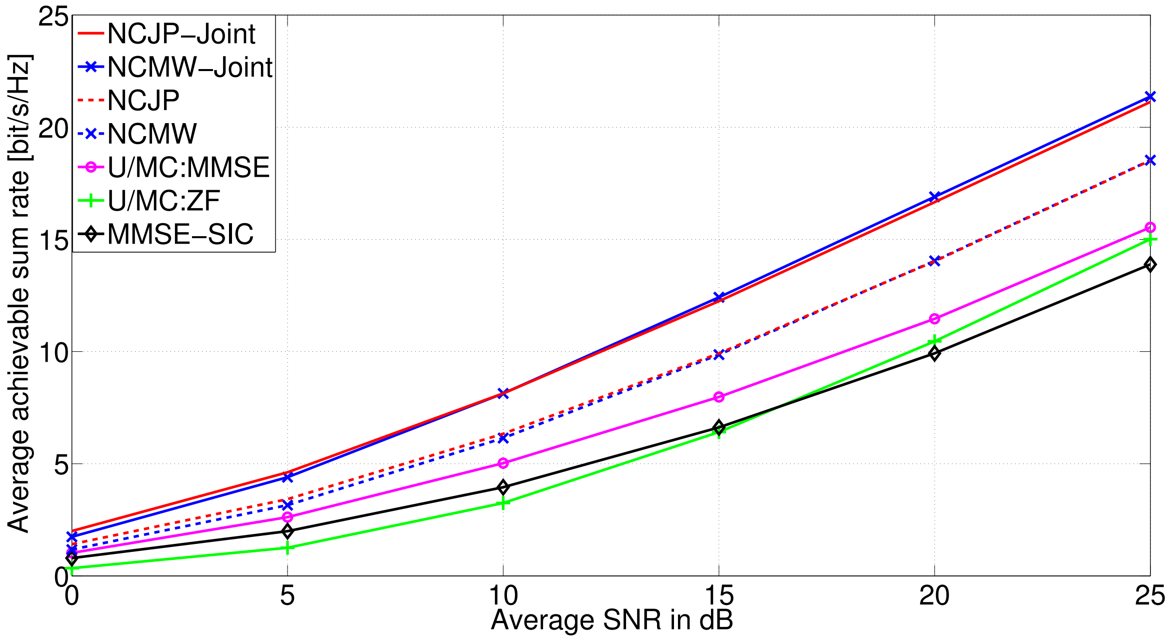

Figure 5.5. Average achievable sum rates versus average SNR for configuration A considering $L = 4$ antennas at RS.

Fig. 5.5 shows the average achievable sum rates versus the average SNR for configuration A. For these simulations, $L = 4$ antennas at RS are assumed. For all approaches, the achievable sum rate increases for increasing the average SNR. The approaches which are based on the proposed NCMW and NCJP Tx strategies, i.e., NCMW, NCMW-Joint, NCJP and NCJP-Joint, outperform all other approaches over the entire SNR range. MMSE-SIC benefits less from an increase of the average receive SNR at RS than the other approaches because random beamforming is considered at RS. The performance of U/MC:ZF improves compared to the U/MC:MMSE and MMSE-SIC for increasing the average SNR because the impact of the noise enhancement due to the spatial separation of all signals at RS decreases. The approaches which are based on the NCJP Tx strategy perform slightly better than the approaches which are based on the NCMW Tx strategy for low average SNRs because more signals are considered as suppressed signals for the NCMW Tx strategy and thus, more signals have to be spatially separated at RS. For an average SNR of 10dB, the gains of the proposed NCJP-Joint approach compared to U/MC:MMSE and MMSE-SIC are approximately 62% and 106%, respectively.

Figure 5.6 shows the average achievable sum rates versus the number L of antennas at RS for configuration A. For these simulations, the average SNR is 15dB. For all approaches, the achievable sum rate increases for increasing the number L of antennas at RS. To spatially separate the received signals at RS, $L \geq N$ antennas are

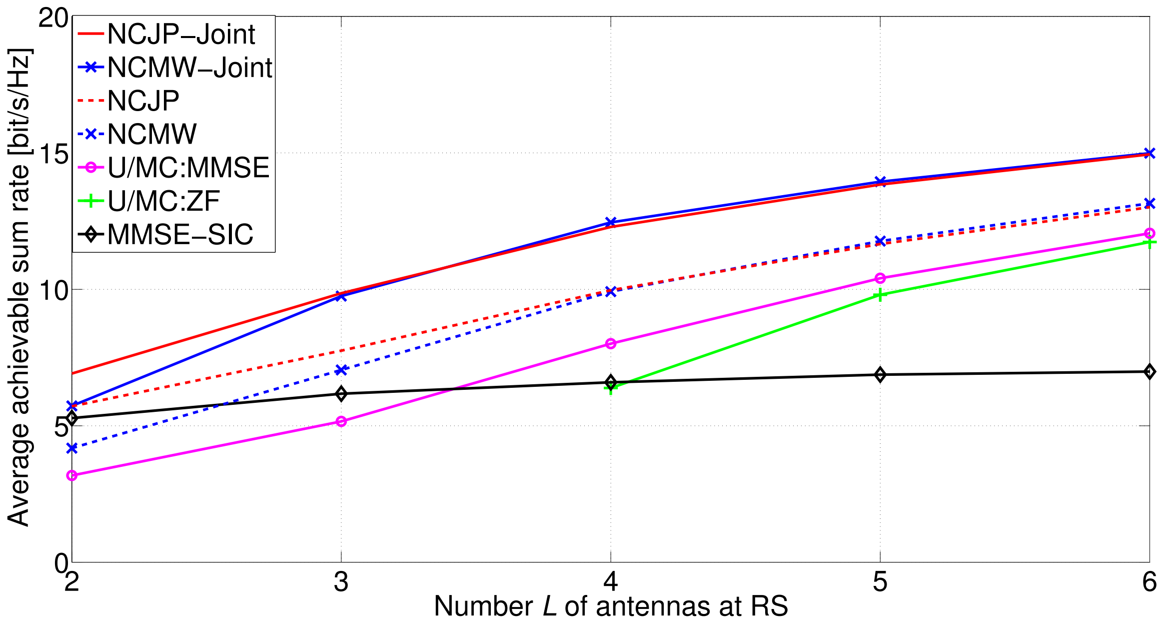

Figure 5.6. Average achievable sum rates versus number L of antennas at RS for configuration A considering an average SNR of 15dB.

required. Thus, the U/MC:ZF approach of [AK11a] starts from $L = 4$ antennas. For U/MC:MMSE of [AK11a], a solution can be obtained for all L due to an MMSE based separation of the signals. However, for $L < 4$ antennas at RS, the performance of the U/MC:MMSE approach is worse than the performance of the MMSE-SIC approach of [CZ12] because the U/MC:MMSE approach does not exploit the joint temporal Rx processing capabilities of the nodes. For $L > 4$ antennas at RS, the U/MC:MMSE approach performs better than the MMSE-SIC approach of [CZ12] because the U/MC:MMSE approach exploits the spatial processing capabilities of RS.

The approaches NCMW-Joint and NCJP-Joint, outperform all other approaches because the NCMW Tx strategy efficiently exploits the self-interference cancellation and the SIC capabilities of the nodes and the NCJP Tx strategy additionally exploits the joint temporal Rx processing capabilities of the nodes. The performance gains of NCMW, NCMW-Joint, NCJP and NCJP-Joint increase compared to U/MC:MMSE for decreasing the number L of antennas at RS, e.g., the gain of NCJP-Joint compared to U/MC:MMSE is approximately 33% for $L = 5$ and 91% for $L = 3$ antennas at RS. The performance gains of NCMW, NCMW-Joint, NCJP and NCJP-Joint increase compared to MMSE-SIC of [CZ12] for an increasing number L of antennas at RS because the proposed Tx strategies efficiently utilize the spatial processing capabilities of RS, e.g., the gain of NCJP-Joint compared to MMSE-SIC is approximately 60% for

$L = 3$ and 101% for $L = 5$ antennas at RS. For $L < 4$ antennas at RS, the NCJP approach performs better than the NCMW approach and for $L < 3$ antennas at RS, the NCJP-Joint approach performs better than the NCMW-Joint approach because the gain of additionally exploiting the joint temporal Rx processing capabilities of the nodes is higher for a low number L of antennas at RS. The gain of considering a joint spatial filter design for the proposed Tx strategies decreases for increasing the number L of antennas at RS, e.g., the gain of NCJP-Joint compared to NCJP is approximately 27% for $L = 3$ and 19% for $L = 5$ antennas at RS.

Due to efficiently combining the spatial processing capabilities of RS with the temporal processing capabilities of the nodes, the approach NCJP-Joint requires two antennas less at RS than the U/MC:MMSE approach to achieve approximately the same sum rate.

Figure 5.7 shows the average achievable sum rates versus the number L of antennas at RS for configuration B. For these simulations, the average SNR is 15dB. For all approaches, the achievable sum rate increases for increasing the number L of antennas at RS. To spatially separate the received signals at RS, $L \geq NM$ antennas are required. Thus, the U/MC:ZF approach of [AK11a] starts from $L = 8$ antennas. For U/MC:MMSE of [AK11a], a solution can be obtained for all L due to an MMSE based separation of the signals. However, for $L < 8$ antennas at RS, the performance of the U/MC:MMSE approach is worse than the performance of the MMSE-SIC approach of [CZ12] because the U/MC:MMSE approach does not exploit the joint temporal Rx processing capabilities of the nodes. For $L > 8$ antennas at RS, the U/MC:MMSE approach performs better than the MMSE-SIC approach of [CZ12] because the U/MC:MMSE approach exploits the spatial processing capabilities of RS.

The approaches NCMW-Joint and NCJP-Joint outperform all other approaches because the NCMW Tx strategy efficiently exploits the self-interference cancellation and the SIC capabilities of the nodes and the NCJP Tx strategy additionally exploits the joint temporal Rx processing capabilities of the nodes. The performance gains of the approaches which are based on the proposed NCMW and NCJP Tx strategies increase compared to U/MC:MMSE for decreasing the number L of antennas at RS, e.g., the gain of NCJP-Joint compared to U/MC:MMSE is approximately 40% for $L = 10$ and 111% for $L = 6$ antennas at RS. For $L \leq 6$, the performance of NCMW is worse than the performance of MMSE-SIC due to not exploiting the joint temporal Rx processing capabilities of the nodes. The performance gains of NCMW, NCMW-Joint, NCJP and NCJP-Joint increase compared to MMSE-SIC for an increasing number L of antennas at RS because the proposed Tx strategies efficiently utilize the spatial processing capabilities of RS, e.g., the gain of NCJP-Joint compared to MMSE-SIC is approximately

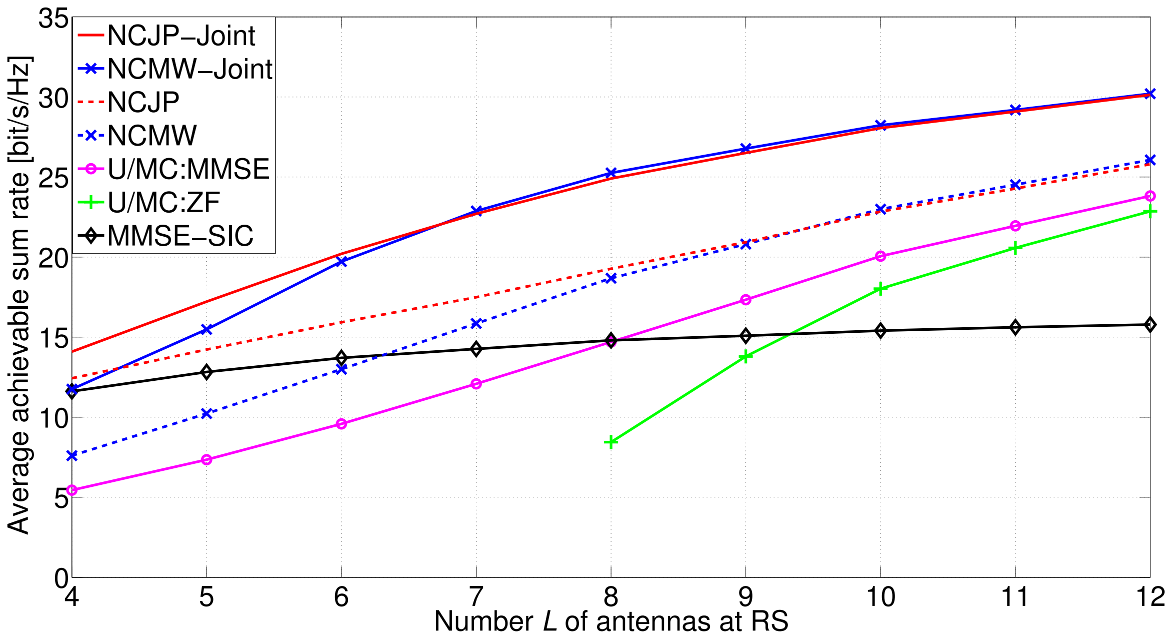

Figure 5.7. Average achievable sum rates versus number L of antennas at RS for configuration B considering an average SNR of 15dB.

47% for $L = 6$ and 82% for $L = 10$ antennas at RS. For $L < 9$ antennas at RS, the NCJP approach performs better than the NCMW approach and for $L < 7$ antennas at RS, the NCJP-Joint approach performs better than the NCMW-Joint approach because the gain of additionally exploiting the joint temporal Rx processing capabilities of the nodes is higher for a low number L of antennas at RS. The gain of considering a joint spatial filter design for the proposed Tx strategies decreases for increasing the number L of antennas at RS, e.g., the gain of NCJP-Joint compared to NCJP is approximately 27% for $L = 6$ and 23% for $L = 10$ antennas at RS.

Due to efficiently combining the spatial processing capabilities of RS with the temporal processing capabilities of the nodes, the approach NCJP-Joint requires four antennas less at RS than the U/MC:MMSE approach to achieve approximately the same sum rate. This number has doubled compared to the performance results for configuration A because in configuration B, the nodes are equipped with $M = 2$ antennas whereas the nodes are equipped with $M = 1$ antenna in configuration A. If the nodes are equipped with more antennas, the gain of performing a joint spatial filter design and the gain of performing joint temporal Rx processing increases.

Figure 5.8 shows the average achievable sum rates versus the number L of antennas at RS for configuration C. For these simulations, the average SNR is 15dB. For all approaches, the achievable sum rate increases for increasing the number L of anten-

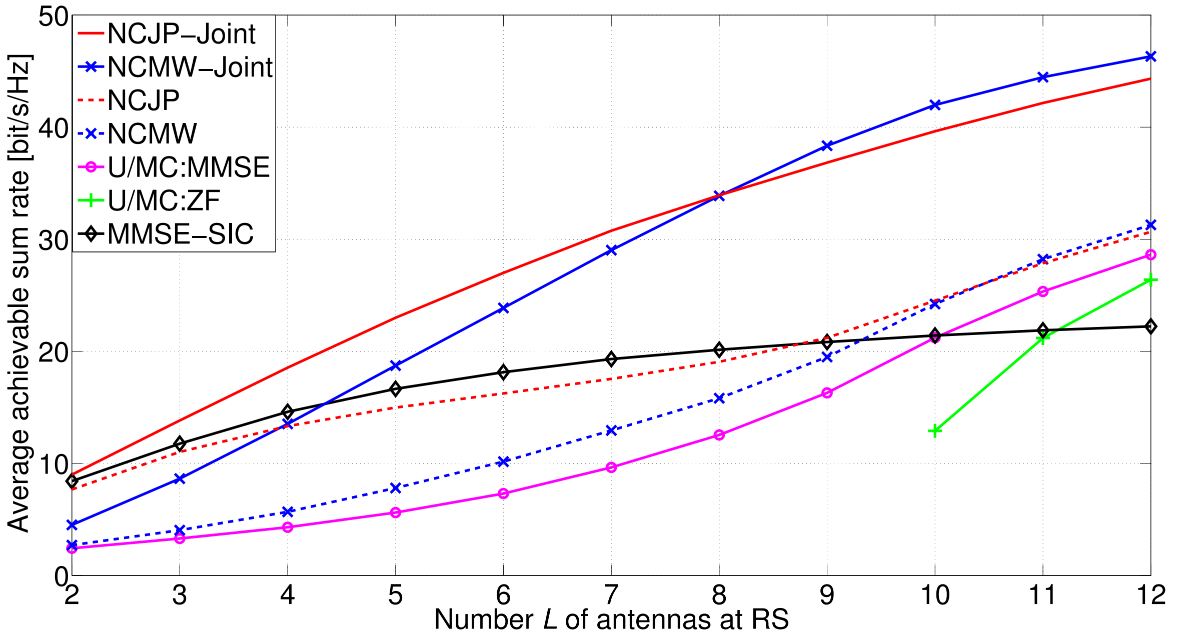

Figure 5.8. Average achievable sum rates versus number L of antennas at RS for configuration C considering an average SNR of 15dB.

nas at RS. To spatially separate the received signals at RS, $L \geq NM$ antennas are required. Thus, the U/MC:ZF approach of [AK11a] starts from $L = 10$ antennas. For U/MC:MMSE of [AK11a], a solution can be obtained for all L due to an MMSE based separation of the signals. However, for $L < 10$ antennas at RS, the performance of the U/MC:MMSE approach is worse than the performance of the MMSE-SIC approach of [CZ12] because the U/MC:MMSE approach does not exploit the joint temporal Rx processing capabilities of the nodes. For $L > 10$ antennas at RS, the U/MC:MMSE approach performs better than the MMSE-SIC approach of [CZ12] because the U/MC:MMSE approach exploits the spatial processing capabilities of RS.

For $L \geq 9$ antennas at RS, the NCJP and the NCMW approach perform worse than the MMSE-SIC approach of [CZ12] because the considered WMMSE-ANC relay transceive filter has to cope with $N - 1 = 9$ different channel rotations to retransmit the desired MC signal in each BC phase. This degrades the performance of the proposed WMMSE-ANC relay transceive filter design. The NCJP-Joint and the NCMW-Joint approaches overcome this problem because a joint spatial filter design is considered and thus, the spatial Rx filters at the nodes compensate the different channel rotations. If more nodes N per group are considered, the gain of performing a joint spatial filter design increases. Due to efficiently combining the spatial processing capabilities of RS with the temporal processing capabilities of the nodes, the NCJP-Joint approach outperforms all other approaches for $L < 8$ antennas at RS. For $L > 8$ antennas at RS, the NCMW-Joint approach outperforms the NCJP-Joint approach because for high number $L > 8$ of antennas at RS, the NCMW-Joint approach exploits the spatial processing capabilities of RS better than the NCJP-Joint approach.

The approach NCJP-Joint requires five to six antennas less at RS than the U/MC:MMSE approach to achieve approximately the same sum rate. This number has approximately tripled compared to the performance results for configuration A because in configuration C, $N - 1 = 9$ BC phases are considered to perform the communications whereas $N - 1 = 3$ BC phases are considered in configuration A. If more BC phases are considered, the gain of performing joint temporal Rx processing increases.

The gain of NCJP-Joint compared to MMSE-SIC increases for increasing the number L of antennas at RS, e.g., the gain is approximately 27% for $L = 4$ and 68% for $L = 8$ antennas at RS. Furthermore, the gain of NCJP-Joint compared to U/MC:MMSE decreases for increasing the number L of antennas at RS, e.g., the gain is approximately 331% for $L = 4$ and 85% for $L = 10$ antennas at RS. Moreover, the gain of NCMW-Joint compared to MMSE-SIC increases for increasing the number L of antennas at RS, e.g., the gain is approximately 12% for $L = 5$ and 96% for $L = 10$ antennas at RS.

Figure 5.9 shows the average achievable sum rates versus the number L of antennas at RS for configuration D. For these simulations, the average SNR is 15dB. The approach MMSE-SIC achieves the worst performance because this approach does not enable a spatial separation of multiple groups and thus, the interferences between the different groups limit the achievable data rates. For all other approaches, the achievable sum rate increases for increasing the number L of antennas at RS. To spatially separate the received signals at RS, $L \geq K$ antennas are required. Thus, the U/MC:ZF approach of [AK11a] starts from $L = 8$ antennas. For U/MC:MMSE of [AK11a], a solution can be obtained for all L due to an MMSE based separation of the signals.

The approaches which are based on the proposed NCMW and NCJP Tx strategies, i.e., NCMW, NCMW-Joint, NCJP and NCJP-Joint, outperform all other approaches because the NCMW Tx strategy efficiently exploits the self-interference cancellation and the SIC capabilities of the nodes and the NCJP Tx strategy additionally exploits the joint temporal Rx processing capabilities of the nodes. The performance gains of NCMW, NCMW-Joint, NCJP and NCJP-Joint, increase compared to U/MC:MMSE for decreasing the number L of antennas at RS, e.g., the gain of NCJP-Joint compared to U/MC:MMSE is approximately 31% for $L = 10$ and 108% for $L = 6$ antennas at RS. For $L < 8$ antennas at RS, the NCJP approach performs better than the

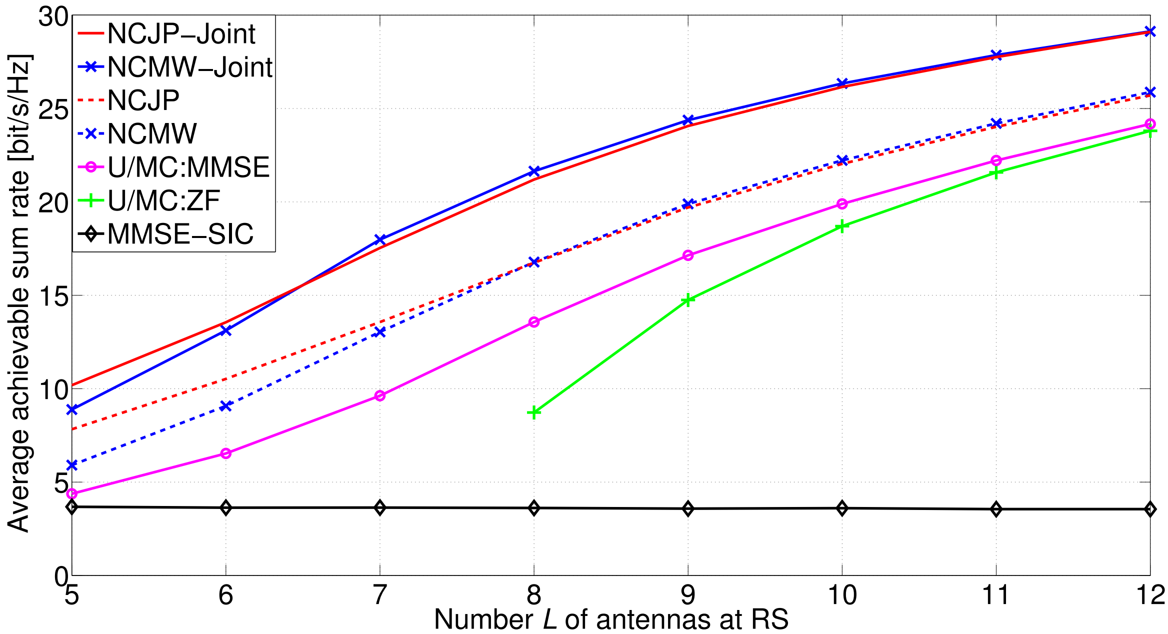

Figure 5.9. Average achievable sum rates versus number L of antennas at RS for configuration D considering an average SNR of 15dB.

NCMW approach and for $L < 7$ antennas at RS, the NCJP-Joint approach performs better than the NCMW-Joint approach because the gain of additionally exploiting the joint temporal Rx processing capabilities of the nodes is higher for a low number L of antennas at RS. The gain of considering a joint spatial filter design for the proposed Tx strategies decreases for increasing the numbers L of antennas at RS, e.g., the gain of NCJP-Joint compared to NCJP is approximately 29% for $L = 6$ and 19% for $L = 10$ antennas at RS.

Due to efficiently combining the spatial processing capabilities of RS with the temporal processing capabilities of the nodes, the approach NCJP-Joint requires two to three antennas less at RS than the U/MC:MMSE approach to achieve approximately the same sum rate. This is similar as in configuration A because $N = 4$ single antenna nodes per group are considered in both configurations.

To summarize, the proposed NCJP Tx strategy combined with the proposed joint spatial filter design between the relay transceive filter and the Rx filters of nodes significantly outperforms the state of the art approaches. The proposed NCJP Tx strategy efficiently combines the spatial processing capabilities of RS with the temporal processing capabilities of the nodes by utilizing ANC. Furthermore, the transceive filter at RS exploits the self-interference cancellation, the SIC and the temporal processing capabilities of the nodes. An overview of selected performance gains of the proposed

NCJP-Joint approach compared to the approaches U/MC:MMSE and MMSE-SIC is presented in Table 5.3.

Table 5.3. Selected performance gains of the proposed NCJP-Joint approach.

Config.	SNR	L	Conv. Approach	Proposed Approach	Perf. Gain
A	5dB	4	U/MC:MMSE	NCJP-Joint	76%
A	5dB	4	MMSE-SIC	NCJP-Joint	132%
A	15dB	4	U/MC:MMSE	NCJP-Joint	53%
A	15dB	4	MMSE-SIC	NCJP-Joint	86%
B	15dB	8	U/MC:MMSE	NCJP-Joint	68%
B	15dB	8	MMSE-SIC	NCJP-Joint	68%
C	15dB	5	U/MC:MMSE	NCJP-Joint	310%
C	15dB	5	MMSE-SIC	NCJP-Joint	38%
C	15dB	10	U/MC:MMSE	NCJP-Joint	85%
C	15dB	10	MMSE-SIC	NCJP-Joint	85%
D	15dB	6	U/MC:MMSE	NCJP-Joint	108%
D	15dB	8	U/MC:MMSE	NCJP-Joint	56%

Chapter 6

Summary and Outlook

6.1 Summary

In this thesis, different filter design approaches and different Tx strategies are proposed for non-regenerative cellular multi-user two-way relaying, multi-pair two-way relaying and multi-group multi-way relaying.

In Chapter 1, the concept of multi-antenna two-hop relaying is introduced and an overview of the state of the art is presented. Based on that, the open issues are identified and formulated. Afterwards, the main contributions of this thesis are summarized and an overview of this thesis is provided.

In Chapter 2, the considered scenarios are briefly described and the assumptions which are valid throughout this thesis are introduced.

In Chapter 3, a non-regenerative cellular multi-user two-way relaying scenario is investigated considering half-duplex multi-antenna nodes and a half-duplex multi-antenna relay station. To investigate this scenario, a system model is introduced which considers that the nodes can perform self-interference cancellation and SIC. Furthermore, ADR requirements are introduced to consider that the required data rates in downlink are typically different than the required data rates in uplink. Due to the high computational complexity of maximizing the sum rate under the considered ADR requirements, a problem decomposition is proposed and based on this decomposition, suboptimal low-complexity approaches are introduced to design the Tx and Rx filters of the nodes, to design the relay transceive filter, to adjust the Tx powers of the nodes and to adjust the numbers of simultaneously transmitted data streams of the nodes. For the Tx and Rx filter design at the nodes, MMSE based Rx filters are introduced and an analytical solution for a SIC aware weighted MMSE based Tx filter of the base station is derived. For the relay transceive filter design, an analytical solution for a weighted self-interference cancellation and SIC aware relay transceive filter is derived which exploits that the signals transmitted by the mobile stations can be jointly processed at the base station. The derived filters can be adjusted via the considered weighting parameters. Furthermore, a joint approach for designing the Tx and Rx filters at the nodes together with the proposed self-interference cancellation and SIC aware relay transceive filter is introduced based on performing an alternating optimization between the different filters. Moreover, two Tx strategies are proposed to tackle the ADR requirements. The PA Tx strategy adjusts the Tx powers of the nodes and the Tx power distributions at the base station and at RS via the considered weighting parameters to fulfill the ADR requirements. The SA Tx strategy performs a subcarrier allocation to adjust the numbers of simultaneously transmitted data streams. Additionally, it adjusts the Tx powers of the nodes and the Tx power distributions at the base station and at RS to fulfill the ADR requirements. By numerical results, it is shown that the proposed Tx strategies combined with the proposed joint filter design approach significantly outperform conventional approaches. For instance, for the considered configurations, the proposed approaches require up to three antennas less at RS than conventional approaches to achieve the same sum rate. Considering the same number of antennas at RS, the proposed approaches achieve significantly higher sum rates than the conventional approaches. Moreover, for low numbers of antennas at RS and if the required data rates in downlink are higher than the required data rates in uplink, the proposed SA Tx strategy outperforms the proposed PA Tx strategy due to performing a subcarrier allocation.

In Chapter 4, a non-regenerative multi-pair two-way relaying scenario is investigated considering half-duplex multi-antenna nodes and a half-duplex multi-antenna relay station. To investigate this scenario, a system model is introduced which considers that the nodes can perform self-interference cancellation and SIC. Furthermore, ADR requirements are introduced to consider that the required data rates are typically different for each direction of transmission. Due to the high computational complexity of maximizing the sum rate under the considered ADR requirements, a problem decomposition is proposed and based on this decomposition, suboptimal low-complexity approaches are introduced to design the Tx and Rx filters of the nodes, to design the relay transceive filter, to adjust the Tx powers of the nodes and to adjust the numbers of simultaneously transmitted data streams of the nodes. For the Tx and Rx filter design at the nodes, two different approaches are proposed. The local Tx and Rx filter design at each node is based on the channel between the node and RS. The global Tx and Rx filter design at each node is based on taking all channels between the nodes and RS into account. For the relay transceive filter design, an analytical solution for a weighted self-interference cancellation and SIC aware relay transceive filter is derived which spatially separates the communications of different pairs. Furthermore, two Tx strategies are proposed to tackle the ADR requirements. The PA Tx strategy adjusts the Tx powers of the nodes and the Tx power distribution at RS to fulfill the ADR requirements. The OS Tx strategy performs an optimization of the numbers of simultaneously transmitted data streams. Additionally, it adjusts the Tx powers of the nodes and the Tx power distribution at RS to fulfill the ADR requirements. By numerical results, it is shown that the proposed OS Tx strategy combined with the proposed global Tx and Rx filter design at the nodes and the proposed WMMSE-SIC transceive filter at RS significantly outperforms conventional approaches. For instance, for the considered configurations, the proposed approach requires up to three antennas less at RS than conventional approaches to achieve the same sum rate. Considering the same number of antennas at RS, the proposed approach achieves significantly higher sum rates than the conventional approaches.

In Chapter 5, a non-regenerative multi-group multi-way relaying scenario is investigated considering half-duplex multi-antenna nodes and a half-duplex multi-antenna relay station. To investigate this scenario, a system model is introduced which considers that the nodes can perform self-interference cancellation, SIC and joint temporal Rx processing over multiple BC phases. Due to the high computational complexity of maximizing the sum rate, a problem decomposition is proposed and based on this decomposition, suboptimal low-complexity approaches are introduced to select the signals which are retransmitted in each BC phase, to design the relay transceive filter, to design the spatial Rx filters of the nodes and to design the temporal Rx filters of the nodes. To select the signals which are retransmitted in each BC phase, two different Tx strategies are proposed which utilize ANC to exploit the spatial processing capabilities of the nodes and of RS as well the capability of the nodes to perform temporal Rx processing over the received signals of the different BC phases. Additionally, the proposed Tx strategies exploit the capability of the nodes to perform self-interference cancellation and SIC. To design a relay transceive filter which enables an efficient application of the proposed Tx strategies, an analytical solution for an ANC aware weighted relay transceive filter is derived. Furthermore, a joint approach for designing the Rx filters at the nodes together with the proposed ANC aware weighted relay transceive filter is introduced based on performing an alternating optimization between the different filters. For the temporal Rx filter design, an MMSE based approach utilizing SIC is presented. By numerical results, it is shown that the proposed Tx strategies combined with the proposed joint spatial filter design significantly outperform conventional approaches. For instance, if a single group configuration with ten nodes is considered, the proposed approaches require up to six antennas less at RS than conventional approaches to achieve the same sum rate. Considering the same number of antennas at RS, the proposed approaches achieve significantly higher sum rates than the conventional approaches. Moreover, for low numbers of antennas at RS, the proposed NCJP Tx strategy outperforms the proposed NCMW Tx strategy due to exploiting the capability of each node to jointly process the received signals of the different BC phases.

6.2 Outlook

In this thesis, only a limited selection of topics is investigated. In the following, some additional topics which are closely related to the findings of this thesis are briefly discussed. For some of these topics, initial investigations have already been carried out by the author.

In this thesis, perfect CSI is assumed for the computation of the Tx and Rx filters at the nodes and for the computation of the relay transceive filter. Furthermore, the considered self-interference cancellation and SIC capabilities of the nodes are based on considering perfect CSI. However, in realistic scenarios, the channels have to be estimated [HH03]. Thus, the available CSI is not perfect due to estimation errors caused by noisy measurements, quantization and / or outdated CSI. In [DHK12], non-regenerative multi-pair two-way relaying with imperfect CSI is investigated by the author considering single antenna nodes. To obtain CSI at RS and at the nodes, a pilot transmission scheme for multi-pair two-way relaying is proposed. Furthermore, it is assumed that the nodes can subtract the back-propagated self-interference and the cases of perfect and imperfect self-interference cancellation are considered. Additionally, a robust self-interference aware relay transceive filter is introduced which minimizes the mean square error between the estimated and the transmitted signals if the proposed pilot transmission scheme is applied. This is a first approach for investigating channel estimation and robust filter design in multi-pair two-way relaying. Future work could extend the proposed pilot transmission scheme and the proposed robust relay transceive filter design to cellular multi-user two-way relaying, multi-pair two-way relaying, and multi-group multi-way relaying considering multi-antenna nodes. Furthermore, different pilot transmission schemes could be investigated and the proposed Tx strategies could be extended to consider imperfect CSI. Moreover, the overhead caused by pilot assisted channel estimation could be considered for the development of novel Tx strategies.

In this thesis, all communications between the nodes are performed via RS. However, if the direct links between the nodes are strong, transmissions could also be performed via the direct links. However, if two-way relaying is applied, transmissions via the direct links can no longer be considered due to the half-duplex constraint of the nodes. Thus, a 3-phase two-way protocol is considered in [LWZ12] which enables direct link transmissions. However, the achievable sum rate of the 3-phase two-way relaying protocol is worse than the achievable sum rate of conventional two-way relaying for weak direct links and is worse than the achievable sum rate of pure direct link transmissions for strong direct links. To overcome this problem, a hybrid approach for combining non-regenerative MIMO two-way relaying and direct link transmissions is proposed by the author in [DHK13]. In [DHK13], a single pair two-way relaying scenario is considered and it is assumed that all nodes are equipped with the same number of antennas. A TWR/DT time-sharing approach is proposed which combines the transmissions via two-way relaying and via the direct link using orthogonal resources in time to fulfill the ADR requirements. Furthermore, a TWR/DT select approach is proposed which performs the bidirectional transmissions either via two-way relaying or via the direct link depending on which scheme achieves a higher sum rate for the instantaneous channel conditions under the considered ADR requirements. By numerical results, it is shown that the proposed TWR/DT time-sharing and the proposed TWR/DT select approach achieve higher sum rates than conventional schemes which are based on either performing pure two-way relaying or pure direct link transmissions. Future work could investigate the utilization of the direct links for cellular multi-user two-way relaying, multi-pair two-way relaying, and multi-group multi-way relaying.

For the multi-group multi-way relaying scenario investigated in this thesis, it is assumed that all nodes simultaneously transmit in the MAC phase. However, the consideration of several MAC phases can improve the performance in case of a limited number of antennas at RS. In future work, only some of the nodes could be selected to simultaneously transmit in the MAC phase, which would simplify the spatial processing at RS and could improve the performance. Thus, Tx strategies could be developed which optimize the number of MAC phases and which optimize the selection of the nodes which simultaneously transmit in each MAC phase. For these Tx strategies, the WMMSE-ANC relay transceive filter which is proposed in this thesis could be utilized. Furthermore, if only some of the nodes are transmitting in each MAC phase, transmissions via the direct links between the nodes could be considered. If the temporal Rx processing capabilities of the nodes are exploited, the consideration of direct link transmissions would reduce the required number of BC phases by one. Thus, Tx strategies which efficiently utilize direct link transmissions for multi-group multi-way relaying could be developed.

In this thesis, ADR requirements and multiple subcarriers have not been considered for the multi-group multi-way relaying scenario. In future work, multiple subcarriers and ADR requirements could be considered. Thus, a Tx strategy could be proposed which performs a subcarrier allocation to tackle the ADR requirements. On each subcarrier, only some of the nodes could be selected to simultaneously transmit in the MAC phase. This selection could be based on the channel conditions of the nodes and on the amount of data which each node has to transmit. In the subsequent BC phases, different linearly processed versions of the received signals could be retransmitted to all nodes. For such a Tx strategy, the same number of nodes should be selected for transmission on each subcarrier such that the same number of BC phases is required.

Appendix

A.1 Derivation of the Lagrangian Multiplier η_c for the Tx Filter Design at S_1 for Cellular Multi-User Two-Way Relaying

In the following, the derivation of the Lagrangian multiplier η_c, which is considered for the Tx Filter design at S_1 presented in Section 3.3.2.3, is sketched. The derivation of η_c is based on [Joh04, Ung09]. For the derivation of the Lagrangian multiplier η_c, the KKT conditions (3.29) are considered, where

$$\begin{aligned}\frac{\partial F(\mathbf{Q}_{1,c},\beta_c,c)}{\partial \mathbf{Q}_{1,c}} =& -\beta_c \boldsymbol{\Theta}_c^{\mathrm{T}} \mathbf{V}_{\mathrm{BS}}^{\mathrm{T}} \\ &+ |\beta_c|^2 \boldsymbol{\Theta}_c^{\mathrm{T}} \boldsymbol{\Theta}_c^{*} \mathbf{Q}_{1,c}^{*} \mathbf{V}_{\mathrm{BS}}^{\mathrm{T}} \\ &- |\beta_c|^2 \sum_{k=2}^{K}\sum_{n=2}^{M}\sum_{m=1}^{n-1} v_{\mathrm{BS},k} \boldsymbol{\Theta}_{k,n,c}^{\mathrm{T}} \boldsymbol{\Theta}_{k,n,c}^{*} \mathbf{Q}_{1,c}^{*} \mathbf{R}_{(k-2)M+m}^{\mathrm{T}}, \end{aligned} \tag{A.1a}$$

and

$$\begin{aligned}\frac{\partial F(\mathbf{G}_c,\beta_c,k,c)}{\partial \beta_c} =& -\operatorname{tr}\left(\boldsymbol{\Theta}_c \mathbf{Q}_{1,c} \mathbf{V}_{\mathrm{BS}}\right) \\ &+ \beta_c^{*} \operatorname{tr}\left(\boldsymbol{\Theta}_c \mathbf{Q}_{1,c} \mathbf{V}_{\mathrm{BS}} \mathbf{Q}_{1,c}^{\mathrm{H}} \boldsymbol{\Theta}_c^{\mathrm{H}}\right) \\ &- \beta_c^{*} \operatorname{tr}\left(\sum_{k=2}^{K}\sum_{n=2}^{M}\sum_{m=1}^{n-1} v_{\mathrm{BS},k} \boldsymbol{\Theta}_{k,n,c} \mathbf{Q}_{1,c} \mathbf{R}_{(k-2)M+m}^{\mathrm{T}} \mathbf{Q}_{1,c}^{\mathrm{H}} \boldsymbol{\Theta}_{k,n,c}^{\mathrm{H}}\right) \\ &+ \beta_c^{*} \operatorname{tr}\left(\sum_{k=2}^{K}\sum_{l=2,l\neq k}^{K} v_{\mathrm{BS},k} \mathbf{D}_{k,c} \mathbf{H}_{k,c}^{\mathrm{T}} \mathbf{G}_c \boldsymbol{\Upsilon}_{l,c} \mathbf{G}_c^{\mathrm{H}} \mathbf{H}_{k,c}^{*} \mathbf{D}_{k,c}^{\mathrm{H}}\right) \\ &+ \beta_c^{*} \operatorname{tr}\left(\sum_{k=2}^{K} v_{\mathrm{BS},k} \left(\sigma_{\mathrm{n,RS}}^2 \mathbf{D}_{k,c} \mathbf{H}_{k,c}^{\mathrm{T}} \mathbf{G}_c \mathbf{G}_c^{\mathrm{H}} \mathbf{H}_{k,c}^{*} \mathbf{D}_{k,c}^{\mathrm{H}} + \sigma_{\mathrm{n}}^2 \mathbf{D}_{k,c} \mathbf{D}_{k,c}^{\mathrm{H}}\right)\right). \end{aligned} \tag{A.2a}$$

Using the second KKT condition (3.29b), β_c^{*} can be written as

$$\beta_c^{*} = \frac{\operatorname{tr}\left(\boldsymbol{\Theta}_c \mathbf{Q}_{1,c} \mathbf{V}_{\mathrm{BS}}\right)}{\mathbf{B}_c}, \tag{A.3}$$

with

$$\begin{aligned}\mathbf{B}_c = &\operatorname{tr}\left(\mathbf{\Theta}_c\mathbf{Q}_{1,c}\mathbf{V}_{\mathrm{BS}}\mathbf{Q}_{1,c}^{\mathrm{H}}\mathbf{\Theta}_c^{\mathrm{H}}\right)\\ &-\operatorname{tr}\left(\sum_{k=2}^{K}\sum_{n=2}^{M}\sum_{m=1}^{n-1}\upsilon_{\mathrm{BS},k}\mathbf{\Theta}_{k,n,c}\mathbf{Q}_{1,c}\mathbf{R}_{(k-2)M+m}^{\mathrm{T}}\mathbf{Q}_{1,c}^{\mathrm{H}}\mathbf{\Theta}_{k,n,c}^{\mathrm{H}}\right)\\ &+\operatorname{tr}\left(\sum_{k=2}^{K}\sum_{l=2,l\neq k}^{K}\upsilon_{\mathrm{BS},k}\mathbf{D}_{k,c}\mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{\Upsilon}_{l,c}\mathbf{G}_c^{\mathrm{H}}\mathbf{H}_{k,c}^{*}\mathbf{D}_{k,c}^{\mathrm{H}}\right)\\ &+\operatorname{tr}\left(\sum_{k=2}^{K}\upsilon_{\mathrm{BS},k}\left(\sigma_{\mathrm{n,RS}}^2\mathbf{D}_{k,c}\mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{G}_c^{\mathrm{H}}\mathbf{H}_{k,c}^{*}\mathbf{D}_{k,c}^{\mathrm{H}}+\sigma_{\mathrm{n}}^2\mathbf{D}_{k,c}\mathbf{D}_{k,c}^{\mathrm{H}}\right)\right).\end{aligned}\tag{A.4}$$

Now, β_c can be inserted in the first KKT condition (3.29a). Afterwards, the condition can be multiplied by $\mathbf{Q}_{1,c}^{\mathrm{T}}$ and the trace operator can be applied. Furthermore, the transpose operation and some algebraic manipulations can be performed, yielding

$$\eta_c\left(\operatorname{tr}\left(\mathbf{Q}_{1,c}\mathbf{Q}_{1,c}^{\mathrm{H}}\right)\right) = -|\beta_c|^2\sum_{k=2}^{K}\mathbf{A}_{k,c},\tag{A.5}$$

where

$$\begin{aligned}\mathbf{A}_{k,c} = &\,\upsilon_{\mathrm{BS},k}\cdot\operatorname{tr}\left(\sum_{l=2,l\neq k}^{K}\mathbf{D}_{k,c}\mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{\Upsilon}_{l,c}\mathbf{G}_c^{\mathrm{H}}\mathbf{H}_{k,c}^{*}\mathbf{D}_{k,c}^{\mathrm{H}}\right)\\ &+\upsilon_{\mathrm{BS},k}\cdot\operatorname{tr}\left(\sigma_{\mathrm{n,RS}}^2\mathbf{D}_{k,c}\mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{G}_c^{\mathrm{H}}\mathbf{H}_{k,c}^{*}\mathbf{D}_{k,c}^{\mathrm{H}}+\sigma_{\mathrm{n}}^2\mathbf{D}_{k,c}\mathbf{D}_{k,c}^{\mathrm{H}}\right).\end{aligned}\tag{A.6}$$

Using the third KKT condition (3.29c) the Lagrangian multiplier η_c results in

$$\eta_c = -\frac{|\beta_c|^2\sum_{k=2}^{K}\operatorname{tr}\left(\mathbf{A}_{k,c}\right)}{(K-1)P_{\mathrm{node}}}.\tag{A.7}$$

A.2 Derivation of the Lagrangian Multiplier η_c for the WMMSE-SIC Transceive Filter Design at RS for Cellular Multi-User Two-Way Relaying

In the following, the derivation of the Lagrangian multiplier η_c, which is considered for the WMMSE-SIC relay transceive filter design presented in Section 3.3.3.3, is sketched. The derivation of η_c is based on [Joh04, Ung09]. For the derivation of the Lagrangian multiplier η_c, the KKT conditions (3.49) are considered, where

$$\begin{aligned}
\frac{1}{v_{\mathrm{BS},k}} \cdot \frac{\partial F_{\mathrm{BS}}(\mathbf{G}_c, \alpha_c, k, c)}{\partial \mathbf{G}_c} =& -\alpha_c \mathbf{H}_{k,c}\mathbf{D}_{k,c}^{\mathrm{T}}\mathbf{Q}_{1,k,c}^{\mathrm{T}}\mathbf{H}_{1,c}^{\mathrm{T}} \\
&+ |\alpha_c|^2 \sum_{l=1,l\neq k}^{K} \mathbf{H}_{k,c}\mathbf{D}_{k,c}^{\mathrm{T}}\mathbf{D}_{k,c}^{*}\mathbf{H}_{k,c}^{\mathrm{H}}\mathbf{G}_c^{*}\mathbf{\Upsilon}_{l,c}^{\mathrm{T}} \\
&- |\alpha_c|^2 \sum_{n=2}^{M}\sum_{m=1}^{n-1} \mathbf{H}_{k,c}\mathbf{d}_{k,n,c}^{\mathrm{T}}\mathbf{d}_{k,n,c}^{*}\mathbf{H}_{k,c}^{\mathrm{H}}\mathbf{G}_c^{*}\mathbf{\Upsilon}_{k,m,c}^{(\mathrm{BS})\mathrm{T}} \\
&+ |\alpha_c|^2 \sigma_{\mathrm{n,RS}}^2 \mathbf{H}_{k,c}\mathbf{D}_{k,c}^{\mathrm{T}}\mathbf{D}_{k,c}^{*}\mathbf{H}_{k,c}^{\mathrm{H}}\mathbf{G}_c^{*}, \qquad \text{(A.8a)} \\
\frac{1}{v_{\mathrm{MS},k}} \cdot \frac{\partial F_{\mathrm{MS}}(\mathbf{G}_c, \alpha_c, k, c)}{\partial \mathbf{G}_c} =& -\alpha_c \mathbf{H}_{1,c}\mathbf{D}_{1,k,c}^{\mathrm{T}}\mathbf{Q}_{k,c}^{\mathrm{T}}\mathbf{H}_{k,c}^{\mathrm{T}} \\
&+ |\alpha_c|^2 \sum_{l=k}^{K} \mathbf{H}_{1,c}\mathbf{D}_{1,k,c}^{\mathrm{T}}\mathbf{D}_{1,k,c}^{*}\mathbf{H}_{1,c}^{\mathrm{H}}\mathbf{G}_c^{*}\mathbf{\Upsilon}_{l,c}^{\mathrm{T}} \\
&- |\alpha_c|^2 \sum_{n=2}^{m_{k,c}}\sum_{m=1}^{n-1} \mathbf{H}_{1,c}\mathbf{d}_{1,k,n,c}^{\mathrm{T}}\mathbf{d}_{1,k,n,c}^{*}\mathbf{H}_{1,c}^{\mathrm{H}}\mathbf{G}_c^{*}\mathbf{\Upsilon}_{k,m,c}^{(\mathrm{MS})\mathrm{T}} \\
&+ |\alpha_c|^2 \sigma_{\mathrm{n,RS}}^2 \mathbf{H}_{1,c}\mathbf{D}_{1,k,c}^{\mathrm{T}}\mathbf{D}_{1,k,c}^{*}\mathbf{H}_{1,c}^{\mathrm{H}}\mathbf{G}_c^{*}, \qquad \text{(A.8b)}
\end{aligned}$$

and

$$\begin{aligned}
\frac{1}{v_{\mathrm{BS},k}} \cdot \frac{\partial F_{\mathrm{BS}}(\mathbf{G}_c, \alpha_c, k, c)}{\partial \alpha_c} =& -\mathrm{tr}\left(\mathbf{D}_{k,c}\mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{H}_{1,c}\mathbf{Q}_{1,k,c}\right) \\
&+ \alpha_c^{*}\mathrm{tr}\left(\sum_{l=1,l\neq k}^{K} \mathbf{D}_{k,c}\mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{\Upsilon}_{l,c}\mathbf{G}_c^{\mathrm{H}}\mathbf{H}_{k,c}^{*}\mathbf{D}_{k,c}^{\mathrm{H}}\right) \\
&- \alpha_c^{*}\mathrm{tr}\left(\sum_{n=2}^{M}\sum_{m=1}^{n-1} \mathbf{d}_{k,n,c}\mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{\Upsilon}_{k,m,c}^{(\mathrm{BS})}\mathbf{G}_c^{\mathrm{H}}\mathbf{H}_{k,c}^{*}\mathbf{d}_{k,n,c}^{\mathrm{H}}\right) \\
&+ \alpha_c^{*}\mathrm{tr}\left(\sigma_{\mathrm{n,RS}}^2\mathbf{D}_{k,c}\mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{G}_c^{\mathrm{H}}\mathbf{H}_{k,c}^{*}\mathbf{D}_{k,c}^{\mathrm{H}} + \sigma_{\mathrm{n}}^2\mathbf{D}_{k,c}\mathbf{D}_{k,c}^{\mathrm{H}}\right), \\
&\qquad \text{(A.9a)} \\
\frac{1}{v_{\mathrm{MS},k}} \cdot \frac{\partial F_{\mathrm{MS}}(\mathbf{G}_c, \alpha_c, k, c)}{\partial \alpha_c} =& -\mathrm{tr}\left(\mathbf{D}_{1,k,c}\mathbf{H}_{1,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{H}_{k,c}\mathbf{Q}_{k,c}\right) \\
&+ \alpha_c^{*}\mathrm{tr}\left(\sum_{l=k}^{K} \mathbf{D}_{1,k,c}\mathbf{H}_{1,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{\Upsilon}_{l,c}\mathbf{G}_c^{\mathrm{H}}\mathbf{H}_{k,c}^{*}\mathbf{D}_{1,k,c}^{\mathrm{H}}\right) \\
&- \alpha_c^{*}\mathrm{tr}\left(\sum_{n=2}^{m_{k,c}}\sum_{m=1}^{n-1} \mathbf{d}_{1,k,n,c}\mathbf{H}_{1,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{\Upsilon}_{k,m,c}^{(\mathrm{MS})}\mathbf{G}_c^{\mathrm{H}}\mathbf{H}_{1,c}^{*}\mathbf{d}_{1,k,n,c}^{\mathrm{H}}\right) \\
&+ \alpha_c^{*}\mathrm{tr}\left(\sigma_{\mathrm{n,RS}}^2\mathbf{D}_{1,k,c}\mathbf{H}_{1,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{G}_c^{\mathrm{H}}\mathbf{H}_{1,c}^{*}\mathbf{D}_{1,k,c}^{\mathrm{H}} + \sigma_{\mathrm{n}}^2\mathbf{D}_{1,k,c}\mathbf{D}_{1,k,c}^{\mathrm{H}}\right). \\
&\qquad \text{(A.9b)}
\end{aligned}$$

Using the second KKT condition (3.49b), α_c^* can be written as

$$\alpha_c^* = \frac{\sum_{k=2}^{K} \left(\upsilon_{\mathrm{BS},k}\mathrm{tr}\left(\mathbf{D}_{k,c}\mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{H}_{1,c}\mathbf{Q}_{1,k,c}\right) + \upsilon_{\mathrm{MS},k}\mathrm{tr}\left(\mathbf{D}_{1,k,c}\mathbf{H}_{1,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{H}_{k,c}\mathbf{Q}_{k,c}\right)\right)}{\sum_{k=2}^{K}\mathbf{B}_{k,c}}, \tag{A.10}$$

with

$$\begin{aligned}
\mathbf{B}_{k,c} = {} & \upsilon_{\mathrm{BS},k}\mathrm{tr}\left(\sum_{l=1,l\neq k}^{K}\mathbf{D}_{k,c}\mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c\boldsymbol{\Upsilon}_{l,c}\mathbf{G}_c^{\mathrm{H}}\mathbf{H}_{k,c}^{*}\mathbf{D}_{k,c}^{\mathrm{H}}\right)\\
& - \upsilon_{\mathrm{BS},k}\mathrm{tr}\left(\sum_{n=2}^{M}\sum_{m=1}^{n-1}\mathbf{d}_{k,n,c}\mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c\boldsymbol{\Upsilon}_{k,m,c}^{(\mathrm{BS})}\mathbf{G}_c^{\mathrm{H}}\mathbf{H}_{k,c}^{*}\mathbf{d}_{k,n,c}^{\mathrm{H}}\right)\\
& + \upsilon_{\mathrm{BS},k}\mathrm{tr}\left(\sigma_{\mathrm{n,RS}}^2\mathbf{D}_{k,c}\mathbf{H}_{k,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{G}_c^{\mathrm{H}}\mathbf{H}_{k,c}^{*}\mathbf{D}_{k,c}^{\mathrm{H}} + \sigma_{\mathrm{n}}^2\mathbf{D}_{k,c}\mathbf{D}_{k,c}^{\mathrm{H}}\right)\\
& + \upsilon_{\mathrm{MS},k}\mathrm{tr}\left(\sum_{l=k}^{K}\mathbf{D}_{1,k,c}\mathbf{H}_{1,c}^{\mathrm{T}}\mathbf{G}_c\boldsymbol{\Upsilon}_{l,c}\mathbf{G}_c^{\mathrm{H}}\mathbf{H}_{k,c}^{*}\mathbf{D}_{1,k,c}^{\mathrm{H}}\right)\\
& - \upsilon_{\mathrm{MS},k}\mathrm{tr}\left(\sum_{n=2}^{m_{k,c}}\sum_{m=1}^{n-1}\mathbf{d}_{1,k,n,c}\mathbf{H}_{1,c}^{\mathrm{T}}\mathbf{G}_c\boldsymbol{\Upsilon}_{k,m,c}^{(\mathrm{MS})}\mathbf{G}_c^{\mathrm{H}}\mathbf{H}_{1,c}^{*}\mathbf{d}_{1,k,n,c}^{\mathrm{H}}\right)\\
& + \upsilon_{\mathrm{MS},k}\mathrm{tr}\left(\sigma_{\mathrm{n,RS}}^2\mathbf{D}_{1,k,c}\mathbf{H}_{1,c}^{\mathrm{T}}\mathbf{G}_c\mathbf{G}_c^{\mathrm{H}}\mathbf{H}_{1,c}^{*}\mathbf{D}_{1,k,c}^{\mathrm{H}} + \sigma_{\mathrm{n}}^2\mathbf{D}_{1,k,c}\mathbf{D}_{1,k,c}^{\mathrm{H}}\right).
\end{aligned} \tag{A.11}$$

Now, α_c can be inserted in the first KKT condition (3.49a). Afterwards, the condition can be multiplied by $\mathbf{G}_c^{\mathrm{T}}$ and the trace operator can be applied. Furthermore, the transpose operation and some algebraic manipulations can be performed, yielding

$$\eta_c\left(\mathrm{tr}\left(\mathbf{G}_c\Upsilon_c\mathbf{G}_c^{\mathrm{H}}\right)\right) = -|\alpha_c|^2\sigma_{\mathrm{n}}^2\left(\sum_{k=2}^{K}\upsilon_{\mathrm{BS},k}\mathrm{tr}\left(\mathbf{D}_{k,c}\mathbf{D}_{k,c}^{\mathrm{H}}\right) + \sum_{k=2}^{K}\upsilon_{\mathrm{MS},k}\mathrm{tr}\left(\mathbf{D}_{1,k,c}\mathbf{D}_{1,k,c}^{\mathrm{H}}\right)\right). \tag{A.12}$$

Now, using the third KKT condition (3.49c) the Lagrangian multiplier η_c results in

$$\eta_c = -\frac{|\alpha_c|^2\sigma_{\mathrm{n}}^2\left(\sum_{k=2}^{K}\upsilon_{\mathrm{BS},k}\mathrm{tr}\left(\mathbf{D}_{k,c}\mathbf{D}_{k,c}^{\mathrm{H}}\right) + \sum_{k=2}^{K}\upsilon_{\mathrm{MS},k}\mathrm{tr}\left(\mathbf{D}_{1,k,c}\mathbf{D}_{1,k,c}^{\mathrm{H}}\right)\right)}{P_{\mathrm{RS}}}. \tag{A.13}$$

A.3 Derivation of the Lagrangian Multiplier η for the WMMSE-SIC Transceive Filter Design at RS for Multi-Pair Two-Way Relaying

In the following, the derivation of the Lagrangian multiplier η, which is considered for the WMMSE-SIC relay transceive filter design presented in Section 4.3.3.3, is sketched.

The derivation of η is based on [Joh04, Ung09]. For the derivation of the Lagrangian multiplier η, the KKT conditions (4.29) are considered, where

$$\begin{aligned}\frac{1}{w_k}\cdot\frac{\partial F(\mathbf{G},\alpha,k)}{\partial\mathbf{G}} =& -\alpha\mathbf{H}_l\mathbf{D}_l^{\mathrm{T}}\mathbf{Q}_k^{\mathrm{T}}\mathbf{H}_k^{\mathrm{T}} \\ &+|\alpha|^2\sum_{i=1,i\neq l}^{K}\mathbf{H}_l\mathbf{D}_l^{\mathrm{T}}\mathbf{D}_l^{*}\mathbf{H}_l^{\mathrm{H}}\mathbf{G}^{*}\boldsymbol{\Upsilon}_i^{\mathrm{T}} \\ &-|\alpha|^2\sum_{n=2}^{m_k}\sum_{m=1}^{n-1}\mathbf{H}_l\mathbf{d}_{l,n}^{\mathrm{T}}\mathbf{d}_{l,n}^{*}\mathbf{H}_l^{\mathrm{H}}\mathbf{G}^{*}\boldsymbol{\Upsilon}_{k,m}^{\mathrm{T}}\end{aligned} \tag{A.14}$$

and

$$\begin{aligned}\frac{1}{w_k}\cdot\frac{\partial F(\mathbf{G},\alpha,k)}{\partial\alpha} =& -\mathrm{tr}\left(\mathbf{D}_l\mathbf{H}_l^{\mathrm{T}}\mathbf{G}\mathbf{H}_k\mathbf{Q}_k\right) \\ &+\alpha^{*}\mathrm{tr}\left(\sum_{i=1,i\neq l}^{K}\mathbf{D}_l\mathbf{H}_l^{\mathrm{T}}\mathbf{G}\boldsymbol{\Upsilon}_i\mathbf{G}^{\mathrm{H}}\mathbf{H}_l^{*}\mathbf{D}_l^{\mathrm{H}}\right) \\ &-\alpha^{*}\mathrm{tr}\left(\sum_{n=2}^{m_k}\sum_{m=1}^{n-1}\mathbf{d}_{l,n}\mathbf{H}_l^{\mathrm{T}}\mathbf{G}\boldsymbol{\Upsilon}_{k,m}\mathbf{G}^{\mathrm{H}}\mathbf{H}_l^{*}\mathbf{d}_{l,n}^{\mathrm{H}}\right) \\ &+\alpha^{*}\mathrm{tr}\left(\sigma_{\mathrm{n}}^2\mathbf{D}_l\mathbf{D}_l^{\mathrm{H}}\right).\end{aligned} \tag{A.15}$$

Using the second KKT condition (4.29b), α^* can be written as

$$\alpha^{*}=\frac{\sum_{k=1}^{K}w_k\mathrm{tr}\left(\mathbf{D}_l\mathbf{H}_l^{\mathrm{T}}\mathbf{G}\mathbf{H}_k\mathbf{Q}_k\right)}{\sum_{k=1}^{K}w_k\mathbf{B}_k}, \tag{A.16}$$

with

$$\begin{aligned}\mathbf{B}_k =& \mathrm{tr}\left(\sum_{i=1,i\neq l}^{K}\mathbf{D}_l\mathbf{H}_l^{\mathrm{T}}\mathbf{G}\boldsymbol{\Upsilon}_i\mathbf{G}^{\mathrm{H}}\mathbf{H}_l^{*}\mathbf{D}_l^{\mathrm{H}}\right)-\mathrm{tr}\left(\sum_{n=2}^{m_k}\sum_{m=1}^{n-1}\mathbf{d}_{l,n}\mathbf{H}_l^{\mathrm{T}}\mathbf{G}\boldsymbol{\Upsilon}_{k,m}\mathbf{G}^{\mathrm{H}}\mathbf{H}_l^{*}\mathbf{d}_{l,n}^{\mathrm{H}}\right) \\ &+\mathrm{tr}\left(\sigma_{\mathrm{n}}^2\mathbf{D}_l\mathbf{D}_l^{\mathrm{H}}\right).\end{aligned} \tag{A.17}$$

Now, α can be inserted in the first KKT condition (4.29a). Afterwards, the condition can be multiplied by $\mathbf{G}^{\mathrm{T}}$ and the trace operator can be applied. Furthermore, the transpose operation and some algebraic manipulations can be performed, yielding

$$\eta\left(\mathrm{tr}\left(\mathbf{G}\boldsymbol{\Upsilon}\mathbf{G}^{\mathrm{H}}\right)\right)=-|\alpha|^2\sigma_{\mathrm{n}}^2\sum_{k=1}^{K}w_k\mathrm{tr}\left(\mathbf{D}_k\mathbf{D}_k^{\mathrm{H}}\right). \tag{A.18}$$

Now, using the third KKT condition (4.29c) the Lagrangian multiplier η results in

$$\eta=-\frac{|\alpha|^2\sigma_{\mathrm{n}}^2\sum_{k=1}^{K}w_k\mathrm{tr}\left(\mathbf{D}_k\mathbf{D}_k^{\mathrm{H}}\right)}{P_{\mathrm{RS}}}. \tag{A.19}$$

A.4 Derivation of the Lagrangian Multiplier η_t for the WMMSE-ANC Transceive Filter Design at RS for Multi-Group Multi-Way Relaying

In the following, the derivation of the Lagrangian multiplier η_t, which is considered for the WMMSE-SIC relay transceive filter design presented in Section 5.3.4, is sketched. The derivation of η_t is based on [Joh04, Ung09]. For the derivation of the Lagrangian multiplier η_t, the KKT conditions (5.24) are considered, where

$$\begin{aligned}\frac{\partial F(\mathbf{G}_t, \alpha_t, k, t)}{\partial \mathbf{G}_t} =& -\alpha_t \mathbf{H}_{k,c} \mathbf{D}_{k,t}^{\mathrm{T}} \mathbf{Q}_{l_{k,t},c}^{\mathrm{T}} \mathbf{H}_{l_{k,t},c}^{\mathrm{T}} \\ &+ |\alpha_t|^2 \mathbf{H}_{k,c} \mathbf{D}_{k,t}^{\mathrm{T}} \mathbf{D}_{k,t}^{*} \mathbf{H}_{k,c}^{\mathrm{H}} \mathbf{G}_t^{*} \mathbf{\Upsilon}^{(k)^{\mathrm{T}}} \\ &+ |\alpha_t|^2 \mathbf{H}_{k,c} \mathbf{D}_{k,t}^{\mathrm{T}} \mathbf{D}_{k,t}^{*} \mathbf{H}_{k,c}^{\mathrm{H}} \mathbf{G}_t^{*} \mathbf{\Upsilon}^{(o_t)^{\mathrm{T}}} \\ &+ |\alpha_t|^2 \mathbf{H}_{k,c} \mathbf{D}_{k,t}^{\mathrm{T}} \mathbf{D}_{k,t}^{*} \mathbf{H}_{k,c}^{\mathrm{H}} \mathbf{G}_t^{*} \sigma_{n,\mathrm{RS}}^2, \end{aligned} \tag{A.20}$$

and

$$\begin{aligned}\frac{\partial F(\mathbf{G}_t, \alpha_t, k, t)}{\partial \alpha_t} =& -\operatorname{tr}\left(\mathbf{D}_{k,t} \mathbf{H}_{k,c}^{\mathrm{T}} \mathbf{G}_t \mathbf{H}_{l_{k,t},c} \mathbf{Q}_{l_{k,t},c}\right) \\ &+ \alpha_t^* \operatorname{tr}\left(\mathbf{D}_{k,t} \mathbf{H}_{k,c}^{\mathrm{T}} \mathbf{G}_t \mathbf{\Upsilon}^{(k)} \mathbf{G}_t^{\mathrm{H}} \mathbf{H}_{k,c}^{*} \mathbf{D}_{k,t}^{\mathrm{H}}\right) \\ &+ \alpha_t^* \operatorname{tr}\left(\mathbf{D}_{k,t} \mathbf{H}_{k,c}^{\mathrm{T}} \mathbf{G}_t \mathbf{\Upsilon}^{(o_t)} \mathbf{G}_t^{\mathrm{H}} \mathbf{H}_{k,c}^{*} \mathbf{D}_{k,t}^{\mathrm{H}}\right) \\ &+ \alpha_t^* \operatorname{tr}\left(\mathbf{D}_{k,t}\left(\mathbf{H}_{k,c}^{\mathrm{T}} \mathbf{G}_t \mathbf{G}_t^{\mathrm{H}} \mathbf{H}_{k,c}^{*} \sigma_{n,\mathrm{RS}}^2 + \sigma_n^2\right) \mathbf{D}_{k,t}^{\mathrm{H}}\right). \end{aligned} \tag{A.21}$$

Using the second KKT condition (5.24b), we can write α_t^* as

$$\alpha_t^* = \frac{\sum_{k=1}^{K} v_{k,t} \operatorname{tr}\left(\mathbf{D}_{k,t} \mathbf{H}_{k,c}^{\mathrm{T}} \mathbf{G}_t \mathbf{H}_{l_{k,t},c} \mathbf{Q}_{l_{k,t},c}\right)}{\sum_{k=1}^{K} v_{k,t} b_{k,t}}, \tag{A.22}$$

where

$$\begin{aligned} b_{k,t} =& \operatorname{tr}\left(\mathbf{D}_{k,t} \mathbf{H}_{k,c}^{\mathrm{T}} \mathbf{G}_t \mathbf{\Upsilon}^{(k)} \mathbf{G}_t^{\mathrm{H}} \mathbf{H}_{k,c}^{*} \mathbf{D}_{k,t}^{\mathrm{H}}\right) + \operatorname{tr}\left(\mathbf{D}_{k,t} \mathbf{H}_{k,c}^{\mathrm{T}} \mathbf{G}_t \mathbf{\Upsilon}^{(o_t)} \mathbf{G}_t^{\mathrm{H}} \mathbf{H}_{k,c}^{*} \mathbf{D}_{k,t}^{\mathrm{H}}\right) \\ &+ \operatorname{tr}\left(\mathbf{D}_{k,t}\left(\mathbf{H}_{k,c}^{\mathrm{T}} \mathbf{G}_t \mathbf{G}_t^{\mathrm{H}} \mathbf{H}_{k,c}^{*} \sigma_{n,\mathrm{RS}}^2 + \sigma_n^2\right) \mathbf{D}_{k,t}^{\mathrm{H}}\right). \end{aligned} \tag{A.23}$$

Now, α_t can be inserted in the first KKT condition (5.24a). Afterwards, the condition can be multiplied by $\mathbf{G}^{\mathrm{T}}$ and the trace operator can be applied. Furthermore, the transpose operation and some algebraic manipulations can be performed, yielding

$$\eta_t(\operatorname{tr}(\mathbf{G}_t \mathbf{\Upsilon} \mathbf{G}_t^{\mathrm{H}}) = -|\alpha_t|^2 \sigma_n^2 \sum_{k=1}^{K} v_{k,t} \operatorname{tr}\left(\mathbf{D}_{k,t} \mathbf{D}_{k,t}^{\mathrm{H}}\right). \tag{A.24}$$

Now using the third KKT condition (5.24c), the Lagrangian multiplier η_t results in

$$\eta_t = -\frac{|\alpha_t|^2 \sigma_n^2 \sum_{k=1}^{K} v_{k,t} \operatorname{tr}\left(\mathbf{D}_{k,t} \mathbf{D}_{k,t}^{\mathrm{H}}\right)}{P_{\mathrm{RS}}}. \tag{A.25}$$

List of Acronyms

ADR	Asymmetric Data Rate
ANC	Analog Network Coding
AWGN	Additive White Gaussian Noise
BC	Broadcast
CSI	Channel State Information
KKT	Karush-Kuhn-Tucker
M2M	Machine to Machine
MAC	Multiple Access
MIMO	Multiple Input Multiple Output
MMSE	Minimum Mean Square Error
MSE	Mean Square Error
NCJP	Network Coding Joint Processing
NCMW	Network Coding Multi-Way
OFDM	Orthogonal Frequency Division Multiplexing
OS	Optimized Streams
PA	Power Adapted
RMSs	Remaining Signals
Rx	Receive
SA	Subcarrier Allocation
SIC	Successive Interference Cancellation
SINR	Signal-to-Interference plus Noise Ratio
SKISs	Self- and Known-Interference Signals
SNR	Signal-to-Noise Ratio
SVD	Singular Value Decomposition

TDD Time Division Duplex

TDMA Time Division Multiple Access

Tx Transmit

w.l.o.g. Without Loss of Generality

WMMSE Weighted Minimum Mean Square Error

WMMSE-ANC Analog Network Coding Aware Weighted Minimum Mean Square Error

WMMSE-SIC Self-Interference and Successive Interference Cancellation Aware Weighted Minimum Mean Square Error

WZF Weighted Zero-Forcing

ZF Zero-Forcing

ZFBD Zero-Forcing Block-Diagonalization

List of Symbols

$\arg\max_x y$	Returns the value of x that maximizes y
$\arg\min_x y$	Returns the value of x that minimizes y
$\alpha_{\mathrm{MMSE},c}$	Factor to fulfill the power constraint at the relay station on subcarrier c considering the WMMSE relay transceive filter design
$\alpha_{\mathrm{ZF},c}$	Factor to fulfill the power constraint at the relay station on subcarrier c considering the WZF filter design
α_c	Additional receive coefficient at all nodes considered for the WMMSE-SIC relay transceive filter design on subcarrier c
α_t	Additional receive coefficient at all nodes considered for the WMMSE-SIC relay transceive filter design in time slot t
$\mathbf{A}_{l,k}$	Matrix containing the overall channel coefficients for the transmission from S_l to S_k during all broadcast phases
β_c	Additional receive coefficient at all mobile stations considered for the transmit filter design at S_1
C	Number of subcarriers
$\mathbb{C}$	Set of complex numbers
$C_{l,k,m,c}$	Maximum achievable data rate for the m^{th} data stream from S_l to S_k on subcarrier c
$C_{l,m}$	Maximum achievable multi-way rate for the transmission of the m^{th} data stream from S_l
$C_{k,l}$	Maximum achievable data rate for the transmission from S_k to S_l
C_k	Maximum achievable data rate for the transmission from S_k
C_{sum}	Achievable sum rate
$C_{\mathrm{ADR},k,l}$	Maximum achievable data rate for the transmission from S_k to S_l considering the asymmetric data rate requirements
$C_{\mathrm{ADR,sum}}$	Achievable sum rate under the asymmetric data rate requirements
$\mathbf{D}_{k,c}$	Receive filter of node S_k on subcarrier c
$\mathbf{D}_{k,t}$	Receive filter of node S_k in time slot t
$\mathbf{d}_{k,m,c}$	m^{th} row vector of $\mathbf{D}_{k,c}$
$\mathbf{D}_{1,k,c}$	Rows of the receive filter of S_1 used to filter the transmit signal of S_k on subcarrier c
$\mathbf{d}_{1,k,m,c}$	m^{th} row vector of $\mathbf{D}_{1,k,c}$
$\mathrm{diag}[\cdot]$	Returns a block diagonal matrix where the diagonal elements are given by the square matrices within the brackets

$\mathrm{diag}[\cdot]^{-1}$	Returns a vector which consists of the elements of the main diagonal of the matrix within the brackets
$\mathrm{E}[\cdot]$	Expectation operator
G	Number of groups in the multi-group multi-way relaying scenario
$\mathbf{G}_c$	Relay transceive filter on subcarrier c
$\mathbf{G}_t$	Relay transceive filter in time slot t
$\mathbf{H}_{k,c}$	Matrix containing the channel coefficients for transmissions from S_k to the relay station on subcarrier c
$\mathbf{H}_{\mathrm{BC},c}$	Matrix containing the overall channel coefficients for the BC phase on subcarrier c
$\mathbf{H}_{\mathrm{BS},k,m,c}$	Matrix containing the overall channel coefficients for the transmissions from all mobile stations to S_1 on subcarrier c considering successive interference cancellation
$\mathbf{H}_{\mathrm{MAC},c}$	Matrix containing the overall channel coefficients for the MAC phase on subcarrier c
$\mathbf{H}_{\mathrm{MS},k,m,c}$	Matrix containing the overall channel coefficients for the transmission from S_1 to S_k on subcarrier c considering successive interference cancellation
$\mathbf{H}_{\mathrm{ov},l,k,t}$	Matrix containing the overall channel coefficients for the transmission from S_l to S_k in time slot t
$h_{\mathrm{ov},l,k,m,c}$	Overall channel coefficient for the transmission of the m^{th} data stream from S_l to S_k on subcarrier c
$\tilde{\mathbf{H}}_{\mathrm{Rx,j}}$	Receive channel matrix of all nodes not belonging to the j^{th} pair
$\mathbf{H}_{\mathrm{Tx},k}$	Receive subchannel of S_k
$\tilde{\mathbf{H}}_{\mathrm{Tx,j}}$	Transmit channel matrix of all nodes not belonging to the j^{th} pair
$\mathbf{H}_{\mathrm{Tx},k}$	Transmit subchannel of S_k
$\mathbf{I}_X$	Identity matrix of size X
$\mathbf{I}_{1:N,M}$	First N row vectors of $\mathbf{I}_M$
$\mathbf{I}_{M,1:N}$	First N column vectors of $\mathbf{I}_M$
K	Number of nodes
L	Number of antennas at the relay station
M_k	Number of antennas at S_k
M	Number of antennas at the nodes excluding the base station
$m_{k,c}$	Number of simultaneously transmitted data streams of S_k on subcarrier c
$m_{g,t}$	Index of the multicast signal of group g in time slot t
$\mathbf{m}_g$	Vector containing the indices of the multicast signals of group g
$\mathrm{MSE}_{1,k,c}$	Mean square error for the transmission from S_2 to S_k on subcarrier c

$\mathrm{MSE}_{k,c}$	Mean square error for the transmission from S_k to S_1 on subcarrier c
N	Number of nodes per group in the multi-group multi-way relaying scenario
$\mathbf{n}_{\mathrm{RS},c}$	Complex white Gaussian noise vector at the relay station on subcarrier c
$\mathbf{n}_{k,c}$	Complex white Gaussian noise vector at S_k on subcarrier c
$\mathbf{n}_{k,t}$	Complex white Gaussian noise vector at S_k in time slot t
$\mathbf{n}_{\mathrm{ov},k,t}$	Vector containing the received noise at S_k in time slot t
$\mathbf{n}_{\mathrm{ov},k}$	Vector containing the received noise at S_k during all broadcast phases
$\mathbf{N}_{k,c}$	Matrix containing the received noise at S_k on subcarrier c
$\mathbf{N}_{k,t}$	Matrix containing the received noise at S_k in time slot t
η_c	Lagrangian multiplier
$o_{g,t}$	Index of the suppressed signal of group g in time slot t
$\mathbf{o}_g$	Vector containing the indices of the suppressed signals of group g
P_{BS}	Maximum transmit power of the base station on each subcarrier
P_{node}	Maximum transmit power of each node on each subcarrier excluding the base station
P_{RS}	Maximum transmit power of the relay station on each subcarrier
$P_{\mathrm{S},l,k,m,c}$	Expected signal power when estimating the m^{th} data stream of S_l at S_k on subcarrier c
$P_{\mathrm{I},l,k,m,c}$	Expected interference power when estimating the m^{th} data stream of S_l at S_k on subcarrier c
$P_{\mathrm{N},l,k,m,c}$	Expected noise power when estimating the m^{th} data stream of S_l at S_k on subcarrier c
p_k	Weighting parameter to adjust the transmit power of S_k
$\mathbf{Q}_{k,c}$	Transmit filter of node S_k on subcarrier c
$\mathbf{q}_{k,m,c}$	m^{th} column vector of $\mathbf{Q}_{k,c}$
$\mathbf{Q}_{1,k,c}$	Columns of the transmit filter of S_1 used to filter the transmit signal $\mathbf{s}_{1,k,c}$ intended for S_k on subcarrier c
$\mathbf{q}_{1,k,m,c}$	m^{th} column vector of $\mathbf{Q}_{1,k,c}$
r	Factor describing the asymmetric data rate requirement between the downlink and uplink
r_k	Factor describing the asymmetric data rate requirement between the required data rates at S_1 and at S_k
$\mathbf{r}$	Vector describing the asymmetric data rate requirements
$\mathbf{s}_{k,c}$	Transmit signal vector of node S_k on subcarrier c
$s_{k,m,c}$	m^{th} element of $\mathbf{s}_{k,c}$

$\hat{\mathbf{s}}_{k,c}$	Estimate of $\mathbf{s}_{k,c}$
$\hat{\mathrm{s}}_{k,m,c}$	m^{th} element of $\hat{\mathbf{s}}_{k,c}$
$\mathbf{s}_{1,k,c}$	Transmit signal vector containing the symbols of S_1 which are intended for S_k on subcarrier c
$\mathrm{s}_{1,k,m,c}$	m^{th} element of $\mathbf{s}_{1,k,c}$
$\hat{\mathbf{s}}_{1,k,c}$	Estimate of $\mathbf{s}_{1,k,c}$
$\hat{\mathrm{s}}_{1,k,m,c}$	m^{th} element of $\hat{\mathbf{s}}_{1,k,c}$
u_g	Index of the unicast signal of group g
$v_{\mathrm{BS},k}$	Weighting parameter to adjust the fraction of the transmit power used to perform transmissions from S_1 to S_k
$\mathbf{v}_{\mathrm{BS}}$	Vector containing the weighting parameters $v_{\mathrm{BS},k}$
$\mathbf{V}_{\mathrm{BS}}$	Matrix containing the weighting parameters $v_{\mathrm{BS},k}$
$\mathbf{V}_c$	Diagonal weighting matrix
$v_{\mathrm{MS},k}$	Weighting parameter to adjust the fraction of the transmit power used to perform transmissions from S_k to S_1
$\mathbf{v}_{\mathrm{MS},c}$	Vector containing the weighting parameters $v_{\mathrm{MS},k}$ for subcarrier c
$\mathbf{V}_k$	Matrix containing the right-singular vectors of $\mathbf{H}_k$
$\mathbf{V}_{k,1:m}$	Matrix containing the m strongest singular vectors of $\mathbf{V}_k$
w_k	Weighting parameter to adjust the fraction of the transmit power used at the relay station to retransmit $\mathbf{s}_k$
$\mathrm{vec}(\cdot)$	Stacks the columns of a matrix into a vector
$\mathrm{vec}_{M,N}^{-1}(\cdot)$	A vector of length MN is sequentially divided into N smaller vectors of length M which are combined to a matrix with M rows and N columns
$\mathbf{W}$	Diagonal weighting matrix
$\mathbf{W}_k$	Temporal receive processing matrix at S_k
$\mathbf{y}_{k,c}$	Received signal at S_k on subcarrier c
$\mathbf{y}_{k,t}$	Received signal at S_k in time slot t
$\mathbf{y}_{\mathrm{RS},c}$	Received signal at the relay station on subcarrier c
$\Re[\cdot]$	Real part of a scalar or a matrix
$(\cdot)^{\mathrm{T}}$	Transpose of a vector or matrix
$(\cdot)^{\mathrm{H}}$	Conjugate transpose of a vector or matrix
$(\cdot)^{*}$	Conjugate of a scalar, vector, or matrix
$\lceil\cdot\rceil$	Rounds a scalar up to the next integer
$(\cdot)^{-1}$	Inverse of a square matrix
$\lvert\cdot\rvert$	Absolute value of a scalar
$\lvert\lvert\cdot\rvert\rvert_2$	Euclidean norm or 2-norm of a vector

$\|\| \cdot \|\|_{\mathrm{F}}$	Frobenius norm of a matrix

Bibliography

[ACLY00] R. Ahlswede, N. Cai, S.-Y. Li, and R. Yeung, "Network information flow," *IEEE Transactions on Information Theory*, vol. 46, no. 4, pp. 1204–1216, 2000.

[AK10a] A. Amah and A. Klein, "Beamforming-based physical layer network coding for non-regenerative multi-way relaying," *EURASIP Journal on Wireless Communications and Networking, Special Issue on Physical Layer Network Coding for Wireless Cooperative Networks*, 2010.

[AK10b] ——, "Pair-aware transceive beamforming for non-regenerative multi-user two-way relaying," in *Proc. IEEE International Conference on Acoustics Speech and Signal Processing*, 2010, pp. 2506–2509.

[AK11a] ——, "Non-regenerative multi-antenna multi-group multi-way relaying," *EURASIP Journal on Wireless Communications and Networking*, 2011.

[AK11b] ——, "Regenerative multi-group multi-way relaying," *IEEE Transactions on Vehicular Technology*, vol. 60, no. 7, pp. 3017–3029, 2011.

[Ama11] A. Amah, "Multi-antenna multi-group multi-way relaying," Ph.D. dissertation, TU Darmstadt, February 2011.

[BKW+09] S. Berger, M. Kuhn, A. Wittneben, T. Unger, and A. Klein, "Recent advances in amplify-and-forward two-hop relaying," *IEEE Communications Magazine*, vol. 47, no. 7, pp. 50–56, 2009.

[BV04] S. Boyd and L. Vandenberghe, *Convex Optimization.* Cambridge, U.K.: Cambridge Univ. Press, 2004.

[CACC08] S. Christensen, R. Agarwal, E. Carvalho, and J. Cioffi, "Weighted sum-rate maximization using weighted MMSE for MIMO-BC beamforming design," *IEEE Transactions on Wireless Communications*, vol. 7, no. 12, pp. 4792–4799, 2008.

[CG79] T. Cover and A. Gamal, "Capacity theorems for the relay channel," *IEEE Transactions on Information Theory*, vol. 25, no. 5, pp. 572–584, 1979.

[CZ12] J. Cao and Z. Zhong, "Non-regenerative multi-way relaying: Ordered MMSE-SIC receivers exploiting temporal diversity," in *Proc. IEEE 75th Vehicular Technology Conference*, 2012, pp. 1–5.

[DHK12] H. Degenhardt, F. Hohmann, and A. Klein, "Pilot transmission scheme and robust filter design for non-regenerative multi-pair two-way relaying," in *Proc. IEEE International Symposium on Personal, Indoor and Mobile Radio Communication*, 2012.

[DHK13] ——, "A hybrid approach combining non-regenerative MIMO two-way relaying and direct link transmission," in *Proc. International ITG Workshop on Smart Antennas*, 2013.

[DK11] H. Degenhardt and A. Klein, "Non-regenerative multi-antenna two-hop relaying under an asymmetric rate constraint," in *Proc. IEEE Wireless Communications and Networking Conference*, 2011, pp. 1454–1459.

[DK12a] ——, "Filter design with optimized numbers of data streams for multi-pair two-way relaying under asymmetric rate requirements," in *Proc. International ITG Workshop on Smart Antennas*, 2012, pp. 294–300.

[DK12b] ——, "Hybrid one-/two-way transmission scheme for bidirectional multi-user relaying under asymmetric rate requirements," in *Proc. International OFDM Workshop*, 2012, pp. 1–7.

[DK12c] ——, "Self-interference aware MIMO filter design for non-regenerative multi-pair two-way relaying," in *Proc. IEEE Wireless Communications and Networking Conference*, 2012.

[DK13a] ——, "Known-interference aware iterative MMSE filter design for non-regenerative multi-way relaying," in *Proc. IEEE International Workshop on Computational Advances in Multi-Sensor Adaptive Processing*, 2013, pp. 488–491.

[DK13b] ——, "A network coding approach to non-regenerative multi-antenna multi-group multi-way relaying," in *Proc. IEEE Vehicular Technology Conference*, 2013.

[DKTL11] Z. Ding, I. Krikidis, J. Thompson, and K. Leung, "Physical layer network coding and precoding for the two-way relay channel in cellular systems," *IEEE Transactions on Signal Processing*, vol. 59, no. 2, pp. 696–712, 2011.

[DRK13] H. Degenhardt, Y. Rong, and A. Klein, "Non-regenerative multi-way relaying: Combining the gains of network coding and joint processing," *IEEE Transactions on Wireless Communications*, vol. 12, no. 11, pp. 5692–5703, 2013.

[DUK11] H. Degenhardt, T. Unger, and A. Klein, "Self-interference aware MMSE filter design for a cellular multi-antenna two-way relaying scenario," in *Proc. International Symposium on Wireless Communication Systems*, 2011, pp. 261–265.

[EW08] C. Esli and A. Wittneben, "Multiuser MIMO two-way relaying for cellular communications," in *Proc. IEEE International Symposium on Personal, Indoor and Mobile Radio Communications*, Sep. 2008.

[Fet12] G. P. Fettweis, "A 5G wireless communications vision," *Microwave Journal*, December 2012.

[FT07] Y. Fan and J. Thompson, "MIMO configurations for relay channels: Theory and practice," *IEEE Transactions on Wireless Communications*, vol. 6, no. 5, pp. 1774–1786, 2007.

[GJJV03] A. Goldsmith, S. Jafar, N. Jindal, and S. Vishwanath, "Capacity limits of MIMO channels," *IEEE Journal on Selected Areas in Communications*, vol. 21, no. 5, pp. 684–702, 2003.

[GYGP09] D. Gunduz, A. Yener, A. Goldsmith, and H. Poor, "The multi-way relay channel," in *Proc. IEEE International Symposium on Information Theory*, 2009, pp. 339–343.

[GYGP13] ——, "The multi-way relay channel," *IEEE Transactions on Information Theory*, vol. 59, no. 1, 2013.

[HBG+13] Y. Hua, D. W. Bliss, S. Gazor, Y. Rong, and Y. Sung, "Guest editorial: Theories and methods for advanced wireless relays - Issue II," *IEEE Journal on Selected Areas in Communications*, vol. 31, no. 8, pp. 1361–1367, 2013.

[HH03] B. Hassibi and B. Hochwald, "How much training is needed in multiple-antenna wireless links?" *IEEE Transactions on Information Theory*, vol. 49, no. 4, pp. 951–963, 2003.

[HIR12] C. Hausl, O. Iscan, and F. Rossetto, "Resource allocation for asymmetric multi-way relay communication over orthogonal channels," *EURASIP Journal on Wireless Communications and Networking*, vol. 2012, no. 1, p. 20, 2012.

[HKE+07] I. Hammerström, M. Kuhn, C. Esli, J. Zhao, A. Wittneben, and G. Bauch, "MIMO two-way relaying with transmit CSI at the relay," in *Proc. IEEE Signal Processing Advances in Wireless Communications*, Jun. 2007, p. 5.

[HW07] I. Hammerström and A. Wittneben, "Power allocation schemes for amplify-and-forward MIMO-OFDM relay links," *IEEE Transactions on Wireless Communications*, vol. 6, no. 8, pp. 2798–2802, 2007.

[Joh04] M. Joham, "Optimization of linear and nonlinear transmit signal processing," Ph.D. dissertation, TU München, June 2004.

[JS10] J. Joung and A. Sayed, "Multiuser two-way amplify-and-forward relay processing and power control methods for beamforming systems," *IEEE Transactions on Signal Processing*, vol. 58, no. 3, pp. 1833–1846, 2010.

[JUN05] M. Joham, W. Utschick, and J. Nossek, "Linear transmit processing in MIMO communications systems," *IEEE Transactions on Signal Processing*, vol. 53, no. 8, pp. 2700–2712, 2005.

[Kam11] K.-D. Kammeyer, *Nachrichtenübertragung*, 5th ed. Vieweg+Teubner, Reihe Informations-/Kommunikationstechnik, Wiesbaden, Germany, 2011.

[KGK07] S. Katti, S. Gollakota, and D. Katabi, "Embracing wireless interference: Analog network coding," in *Proc. ACM SIGCOMM*, 2007, pp. 397–408.

[LDLG11] C. Y. Leow, Z. Ding, K. Leung, and D. Goeckel, “On the study of analogue network coding for multi-pair, bidirectional relay channels,” *IEEE Transactions on Wireless Communications*, vol. 10, no. 2, pp. 670–681, 2011.

[LLSL09] K.-J. Lee, K. W. Lee, H. Sung, and I. Lee, “Sum-rate maximization for two-way MIMO amplify-and-forward relaying systems,” in *Proc. IEEE Vehicular Technology Conference*, 2009, pp. 1–5.

[LTW04] J. Laneman, D. Tse, and G. W. Wornell, “Cooperative diversity in wireless networks: Efficient protocols and outage behavior,” *IEEE Transactions on Information Theory*, vol. 50, no. 12, pp. 3062–3080, 2004.

[LW03] J. Laneman and G. W. Wornell, “Distributed space-time-coded protocols for exploiting cooperative diversity in wireless networks,” *IEEE Transactions on Information Theory*, vol. 49, no. 10, pp. 2415–2425, 2003.

[LWD13] F. Lenkeit, D. Wübben, and A. Dekorsy, “Reliability-aware iterative detection scheme (RAID) for distributed IDM space-time codes in relay systems,” *EURASIP Journal on Advances in Signal Processing: Special Issue on Advanced Distributed Wireless Communication Techniques - Theory and Practise*, April 2013.

[LWZ12] G. Li, Y. Wang, and P. Zhang, “Linear MMSE processing design for 3-phase two-way cooperative MIMO relay systems,” *IEEE Signal Processing Letters*, vol. 19, no. 7, pp. 443–446, 2012.

[Meu71] E. C. V. D. Meulen, “Three-terminal communication channels,” *Advances in Applied Probability*, vol. 3, pp. 120–154, 1971.

[MnMVA07] O. Muñoz Medina, J. Vidal, and A. Agustín, “Linear transceiver design in nonregenerative relays with channel state information,” *IEEE Transactions on Signal Processing*, vol. 55, no. 6, pp. 2593–2604, 2007.

[NP00] R. v. Nee and R. Prasad, *OFDM for Wireless Multimedia Communications*, 1st ed. Norwood, MA, USA: Artech House, Inc., 2000.

[OJK10] L. Ong, S. Johnson, and C. Kellett, “An optimal coding strategy for the binary multi-way relay channel,” *IEEE Communications Letters*, vol. 14, no. 4, pp. 330–332, 2010.

[OJK11] ——, “The capacity region of multiway relay channels over finite fields with full data exchange,” *IEEE Transactions on Information Theory*, vol. 57, no. 5, pp. 3016–3031, 2011.

[OJWB09] T. J. Oechtering, E. A. Jorswieck, R. F. Wyrembelski, and H. Boche, “On the optimal transmit strategy for the MIMO bidirectional broadcast channel,” *IEEE Transactions on Communications*, vol. 57, no. 12, pp. 3817–3826, Dec. 2009.

[OKJ12] L. Ong, C. M. Kellett, and S. J. Johnson, "On the equal-rate capacity of the AWGN multiway relay channel," *IEEE Transactions on Information Theory*, vol. 58, no. 9, pp. 5761–5769, 2012.

[PNG03] A. Paulraj, R. Nabar, and D. Gore, *Introduction to Space-Time Wireless Communications.* Cambridge, U.K.: Cambridge Univ. Press, 2003.

[PP12] K. B. Petersen and M. S. Pedersen, *The Matrix Cookbook.* Technical University of Denmark, 2012.

[Pro00] J. G. Proakis, *Digital Communications*, 4th ed. McGraw-Hill Science/Engineering/Math, 2000.

[PY07] P. Popovski and H. Yomo, "Physical network coding in two-way wireless relay channels," in *Proc. IEEE International Conference on Communications*, 2007, pp. 707–712.

[RH09] F. Roemer and M. Haardt, "Algebraic norm-maximizing (ANOMAX) transmit strategy for two-way relaying with MIMO amplify and forward relays," *IEEE Signal Processing Letters*, vol. 16, no. 10, pp. 909–912, 2009.

[RMBG99] H. Rohling, T. May, K. Bruninghaus, and R. Grunheid, "Broad-band OFDM radio transmission for multimedia applications," *Proc. of the IEEE*, vol. 87, no. 10, pp. 1778–1789, 1999.

[Ron10] Y. Rong, "Optimal joint source and relay beamforming for MIMO relays with direct link," *IEEE Communications Letters*, vol. 14, no. 5, pp. 390–392, 2010.

[Ron12] ——, "Joint source and relay optimization for two-way linear non-regenerative MIMO relay communications," *IEEE Transactions on Signal Processing*, vol. 60, no. 12, pp. 6533–6546, 2012.

[RTH09] Y. Rong, X. Tang, and Y. Hua, "A unified framework for optimizing linear nonregenerative multicarrier MIMO relay communication systems," *IEEE Transactions on Signal Processing*, vol. 57, no. 12, pp. 4837–4851, 2009.

[RW05] B. Rankov and A. Wittneben, "Spectral efficient signaling for half-duplex relay channels," in *Conference Record of the Thirty-Ninth Asilomar Conference on Signals, Systems and Computers*, Oct. 2005, pp. 1066–1071.

[RW06] ——, "Achievable rate regions for the two-way relay channel," in *Proc. IEEE International Symposium on Information Theory*, 2006, pp. 1668–1672.

[RW07] ——, "Spectral efficient protocols for half-duplex fading relay channels," *IEEE Journal on Selected Areas in Communications*, vol. 25, no. 2, pp. 379–389, 2007.

[SDR12] L. Sanguinetti, A. D'Amico, and Y. Rong, "A tutorial on the optimization of amplify-and-forward MIMO relay systems," *IEEE Journal on Selected Areas in Communications*, vol. 30, no. 8, pp. 1331–1346, 2012.

[SEA03] A. Sendonaris, E. Erkip, and B. Aazhang, "User cooperation diversity. Part I. System description," *IEEE Transactions on Communications*, vol. 51, no. 11, pp. 1927–1938, 2003.

[SSH04] Q. Spencer, A. Swindlehurst, and M. Haardt, "Zero-forcing methods for downlink spatial multiplexing in multiuser MIMO channels," *IEEE Transactions on Signal Processing*, vol. 52, no. 2, pp. 461–471, 2004.

[SYLV11] C. Sun, C. Yang, Y. Li, and B. Vucetic, "Transceiver optimization for multi-user multi-antenna two-way relay channels," in *Proc. IEEE International Conference on Acoustics, Speech and Signal Processing*, 2011, pp. 3064–3067.

[SYLV12] ——, "Transceiver design for multi-user multi-antenna two-way relay cellular systems," *IEEE Transactions on Communications*, vol. 60, no. 10, pp. 2893–2903, 2012.

[TH07] X. Tang and Y. Hua, "Optimal design of non-regenerative MIMO wireless relays," *IEEE Transactions on Wireless Communications*, vol. 6, no. 4, pp. 1398–1407, 2007.

[TS09] S. Toh and D. Slock, "A linear beamforming scheme for multi-user MIMO AF two-phase two-way relaying," in *Proc. IEEE International Symposium on Personal, Indoor and Mobile Radio Communications*, 2009, pp. 1003–1007.

[TV05] D. Tse and P. Viswanath, *Fundamentals of Wireless Communication*. Cambridge, U.K.: Cambridge Univ. Press, 2005.

[TW12] M. Tao and R. Wang, "Linear precoding for multi-pair two-way MIMO relay systems with max-min fairness," *IEEE Transactions on Signal Processing*, vol. 60, no. 10, pp. 5361–5370, 2012.

[Ung09] T. Unger, "Multi-antenna two-hop relaying for bi-directional transmission in wireless communication systems," Ph.D. dissertation, TU Darmstadt, June 2009.

[WLW+14] M. Wu, F. Ludwig, M. Woltering, D. Wuebben, A. Dekorsy, and S. Paul, "Analysis and implementation for physical-layer network coding with carrier frequency offset," in *Proc. International ITG Workshop on Smart Antennas*, 2014, pp. 1–8.

[WM07] L. Weng and R. Murch, "Multi-user MIMO relay system with self-interference cancellation," in *Proc. IEEE Wireless Communications and Networking Conference*, 2007, pp. 958–962.

[WT12] R. Wang and M. Tao, "Joint source and relay precoding designs for MIMO two-way relaying based on MSE criterion," *IEEE Transactions on Signal Processing*, vol. 60, no. 3, pp. 1352–1365, 2012.

[WTH12] R. Wang, M. Tao, and Y. Huang, "Linear precoding designs for amplify-and-forward multiuser two-way relay systems," *IEEE Transactions on Wireless Communications*, vol. 11, no. 12, pp. 4457–4469, 2012.

[WWD13] M. Wu, D. Wubben, and A. Dekorsy, "Physical-layer network coding in coded OFDM systems with multiple-antenna relay," in *Proc. IEEE Vehicular Technology Conference*, 2013, pp. 1–5.

[WZHM05] B. Wang, J. Zhang, and A. Host-Madsen, "On the capacity of MIMO relay channels," *IEEE Transactions on Information Theory*, vol. 51, no. 1, pp. 29–43, 2005.

[XH11] S. Xu and Y. Hua, "Optimal design of spatial source-and-relay matrices for a non-regenerative two-way MIMO relay system," *IEEE Transactions on Wireless Communications*, vol. 10, no. 5, pp. 1645–1655, 2011.

[YZGK10] E. Yilmaz, R. Zakhour, D. Gesbert, and R. Knopp, "Multi-pair two-way relay channel with multiple antenna relay station," in *Proc. IEEE International Conference on Communications*, 2010, pp. 1–5.

[ZCW05] A. Zanella, M. Chiani, and M. Win, "MMSE reception and successive interference cancellation for MIMO systems with high spectral efficiency," *IEEE Transactions on Wireless Communications*, vol. 4, no. 3, pp. 1244–1253, 2005.

[ZDP+11] Z. Zhao, Z. Ding, M. Peng, W. Wang, and K. Leung, "A special case of multi-way relay channel: When beamforming is not applicable," *IEEE Transactions on Wireless Communications*, vol. 10, no. 7, pp. 2046–2051, 2011.

[ZLCC09] R. Zhang, Y.-C. Liang, C. C. Chai, and S. Cui, "Optimal beamforming for two-way multi-antenna relay channel with analogue network coding," *IEEE Journal on Selected Areas in Communications*, vol. 27, no. 5, pp. 699–712, 2009.

[ZRH11] J. Zhang, F. Roemer, and M. Haardt, "Beamforming design for multi-user two-way relaying with MIMO amplify and forward relays," in *Proc. IEEE International Conference on Acoustics, Speech and Signal Processing*, 2011, pp. 2824–2827.

Author's Publications

In this section, the international journal and conference publications of the author on non-regenerative two-way and multi-way relaying are listed. These publications have been produced during the Dr.-Ing. candidacy of the author.

[DRK13] H. Degenhardt, Y. Rong, and A. Klein, "Non-regenerative multi-way relaying: Combining the gains of network coding and joint processing," *IEEE Transactions on Wireless Communications*, vol. 12, no. 11, pp. 5692-5703, 2013.

[DK13b] H. Degenhardt and A. Klein, "A network coding approach to non-regenerative multi-antenna multi-group multi-way relaying," in *Proc. IEEE Vehicular Technology Conference*, 2013.

[DK13a] H. Degenhardt and A. Klein, "Known-interference aware iterative MMSE filter design for nonregenerative multi-way relaying," in *IEEE International Workshop on Computational Advances in Multi-Sensor Adaptive Processing*, 2013.

[DHK13] H. Degenhardt, F. Hohmann, and A. Klein, "A hybrid approach combining non-regenerative MIMO two-way relaying and direct link transmission," in *Proc. International ITG Workshop on Smart Antennas*, 2013.

[DK12c] H. Degenhardt and A. Klein, "Self-interference aware MIMO filter design for non-regenerative multi-pair two-way relaying," in *Proc. IEEE Wireless Communications and Networking Conference*, 2012.

[DK12b] H. Degenhardt and A. Klein, "Hybrid one-/two-way transmission scheme for bidirectional multi-user relaying under asymmetric rate requirements," in *Proc. International OFDM Workshop*, 2012.

[DK12a] H. Degenhardt and A. Klein, "Filter design with optimized numbers of data streams for multi-pair two-way relaying under asymmetric rate requirements," in *Proc. International ITG Workshop on Smart Antennas*, 2012.

[DHK12] H. Degenhardt, F. Hohmann, and A. Klein, "Pilot transmission scheme and robust filter design for non-regenerative multi-pair two-way relaying," in *Proc. IEEE International Symposium on Personal, Indoor and Mobile Radio Communication*, 2012.

[DUK11] H. Degenhardt, T.Unger, and A. Klein, "Self-interference aware MMSE filter design for a cellular multi-antenna two-way relaying scenario," in *Proc. International Symposium on Wireless Communication Systems*, 2011.

[DK11] H. Degenhardt and A. Klein, "Non-regenerative multi-antenna two-hop relaying under an asymmetric rate constraint," in *Proc. IEEE Wireless Communications and Networking Conference*, 2011.

Lebenslauf

Name:	Holger Degenhardt
Geburtsdatum:	07.01.1985
Geburtsort:	Lindenfels

Schulausbildung

1995 - 2004	Altes Kurfürstliches Gymnasium Bensheim, Schulabschluss: Allgemeine Hochschulreife

Studium

10/2004 - 01/2010	Studium der Elektro- und Informationstechnik an der Technischen Universität Darmstadt, Studienabschluss: Diplom

Berufstätigkeit

02/2010 - 06/2014	Wissenschaftlicher Mitarbeiter am Fachgebiet für Kommunikationstechnik, Institut für Nachrichtentechnik, Technische Universität Darmstadt

Darmstadt, 22. Mai 2014

Holger Degenhardt