

IMPERIAL

THE FIRST RATE-SPLITTING MULTIPLE ACCESS EXPERIMENTAL STUDY FOR UNICAST, MULTICAST AND JOINT COMMUNICATION AND SENSING

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A Thesis submitted in fulfillment of requirements for the degree of
Doctorate of Philosophy (PhD) Degree of Imperial College London and the
Diploma of Imperial College London (DIC) by research in

Wireless Communications and Signal Processing Lab
Communications and Signal Processing Group
Department of Electrical and Electronic Engineering
Imperial College London
2024

Abstract

In the recent Multiple Input Multiple Output (MIMO) system, numerous studies have demonstrated the theoretical and simulation-based advantages of Rate-Splitting Multiple Access (RSMA), showcasing its potential to enhance system performance across various scenarios. However, a gap remains in the form of experimental studies. These are essential to bridge the divide between theoretical models and practical implementations. This thesis bridges the research gap between the theoretical foundations and practical implementation of RSMA.

First, a prototype test-bed is developed to evaluate RSMA's performance, comprising a two-transmit-antenna base station and two single-antenna users. The experimental results demonstrate that RSMA outperforms Spatial Division Multiple Access (SDMA) and Non-Orthogonal Multiple Access (NOMA) in terms of both sum throughput and fairness performance. Notably, the advantages of RSMA become even more pronounced under practical conditions where the Channel State Information at Transmitter (CSIT) is quantized.

The second part of this thesis explores RSMA's performance in Non-Orthogonal Unicast and Multicast (NOUM) scenarios. When compared to Multi-User Linear Precoding (MULP), RSMA demonstrates superior flexibility in managing and utilizing both unicast and multicast streams. This adaptability enables RSMA better to balance the demands of unicast and multicast communication.

The third part of this thesis investigates RSMA's performance in overloaded scenarios. Specifically, a base station equipped with two transmit antennas serves two groups of users using multi-group multicasting. The results demonstrate that RSMA-based Multi-group Multicast (MGM) outperforms both NOMA-based MGM and SDMA-based MGM in terms of max-min fairness performance.

The final part of this thesis examines the application of RSMA in Integrated Sensing and Communication (ISAC) scenarios, specifically addressing the question of whether a dedicated sensing signal is necessary for ISAC. Both experimental and simulation results reveal that RSMA can achieve superior sensing performance without requiring a dedicated sensing signal, all while meeting the same throughput requirements.

Declaration of Originality

I hereby declare that the work presented in this thesis is my own unless otherwise stated. To the best of my knowledge the work is original and ideas developed in collaboration with others have been appropriately referenced.

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Acknowledgments

First and foremost, I would like to thank my supervisor, Prof. Bruno Clerckx, who always gives me beneficial advice and directions; his help and guidance make my PhD life smooth and happy, and I feel I can't thank you more for everything you did.

I would also like to extend my thanks to all my colleagues Ziang, Yumeng, Yuanwen, Hongyu, Sibio, Matteo, Kris, Yang, Jiawei, Yunnuo, Anup, Longfei, Rafael, Yijie and Onur in the Wireless Communication and Signal Processing Lab at Imperial College London, especially Ziang Liu, with whom I spent most of my PhD life. I hope this is only the start of our friendship and may continue long into the future. To Sundar Aditya and Junghoon Kim, who helped me a lot during my experiments and paper writing. Finally, I would like to thank Aidan O.T. Hogg, who inadvertently turned me into a beer fan over the past four years. I truly miss our pub chats—they were as enlightening as they were entertaining, and they've been an unforgettable part of this journey.

Finally, I would like to thank my family, including my loving parents, my sister, and my wife. Your unconditional love supported me in finishing this four-year journey.

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List of Acronyms

MIMO Multiple Input Multiple Output

MU Multiple Users

MISO Multiple Input Single Output

LoS Line-of-Sight

MCS Modulation Coding Scheme

ISAC Integrated Sensing and Communication

NOUM Non-Orthogonal Unicast and Multicast

SDMA Spatial Division Multiple Access

NOMA Non-Orthogonal Multiple Access

CSIT Channel State Information at Transmitter

CSI Channel State Information

OFDM Orthogonal Frequency-Division Multiplexing

TX Transmitter

RX Receiver

SDR Software Defined Radio

USRP Universal Software Radio Peripheral

WMMSE Weighted Minimized Min Square Error

NI National Instruments

PCIe Peripheral Component Interconnect Express

FPS Fine Phase Shift

ISM Industrial Scientific and Medical

QAM Quadrature Amplitude Modulation

MA Multiple Access

BS Base Station

FDMA Frequency Division Multiple Access

TDMA Time Division Multiple Access

CDMA Code Division Multiple Access

RSMA Rate-Splitting Multiple Access

LTE Long Term Evolution

mMIMO massive Multiple Input Multiple Output

4G Fourth Generation

5G Fifth Generation

6G Sixth Generation

OFDMA Orthogonal Frequency Multiple Access

OMA Orthogonal Multiple Access

MMSE Minimized Min Square Error

ISAC Integrated Sensing and Communication

MGM Multi-group Multicast

SNR Signal to Noise Ratio

SINR Signal to Interference Noise Ratio

MULP Multi-User Linear Precoding

SIC Successive Interference Cancellation

DoF Degree of Freedom

AWGN Additive White Gaussian Noise

NI National Instruments

UI User Interface

WLANS Wireless Local Area Networks

Wi-Fi Wireless Fidelity

VHT Very High Throughput

NR New Radio

3GPP 3rd Generation Partnership Project

SISO Single Input Single Output

OTFS Orthogonal Time Frequency Space

DFT Discrete Fourier Transform

AMC Adaptive Modulation and Coding

V2X Vehicle to Everything

BLER Block Error Rate

LTF Long Training Field

STF Short Training Field

QoS Quality of Service

LDPC Low-Density Parity Check

IoT Internet-of-Things

MRT Maximum Ratio Transmission

FMCW Frequency Modulated Continuous Wave

ZF Zero-Forcing

ADC Analog to Digital Converter

RIS Reconfigurable Intelligent Surfaces

UE User

List of Notations

$\mathbf{1}$	The all-one vector
$\mathbf{h}$	Column vector of channel
$\mathbf{p}$	Column vector of precoder
n	noise at user
W	Information Message
m	Bits per symbol
r	Coding rate
$\mathcal{I} = (m, r)$	Modulation and coding rate pair
$\mathcal{M}$	MCS level
s	Data symbols
$\mathbf{x}$	Transmit signals
$\cup$	Union operation
$\emptyset$	Empty set
$\mathbb{E}[\cdot]$	The expectation operator
$\mathbb{M}$	Set of MCS levels
$ \cdot $	The magnitude of scalars
$(\cdot)^H$	The Hermitian operator
$\ \cdot\ $	The Euclidean norm
$\text{Re}(\cdot)$	The real parts of a complex number
$\text{Im}(\cdot)$	The imaginary parts of a complex number
$\mathcal{CN}(0, \sigma^2)$	The circularly symmetric complex Gaussian distribution with zero mean and variance σ^2
$\mathbf{a}_\theta = [1 \ e^{j\pi \sin \theta} \ \dots \ e^{j\pi(N_T-1) \sin \theta}]^T$	A uniform linear array with N_T antenna elements and half-wavelength spacing, denotes the steering vector along direction θ .

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Introduction

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1.1 Overview

Multiple Access (MA) techniques serve an important role in the evolution of wireless communication systems, which enable the Base Station (BS) to allocate resources to users efficiently. These methods are critical in maximizing spectrum efficiency while ensuring reliable wireless connections across diverse applications [1]. The earliest MA techniques can be traced back to the 1970s [2] when Frequency Division Multiple Access Frequency Division Multiple Access (FDMA) and Time Division Multiple Access (TDMA) were introduced into wireless communication. FDMA and TDMA divide the spectrum into frequency bands or time slots for users' access. Since time and frequency are naturally orthogonal, the inter-user interference is managed in an orthogonal way, so both TDMA and FDMA are also called Orthogonal Multiple Access (OMA) [2]. Since OMA manages the inter-user interference in an orthogonal way, there is no inter-user interference. However, the disadvantage is also obvious: each orthogonal resource is allocated a dedicated user, which limits the maximum number of supportable users and results in low spectrum efficiency. Later,

to let multiple users access the cellular network simultaneously at the same frequency spectrum, Code Division Multiple Access (CDMA) is introduced, in which unique spread-spectrum codes are assigned to different users [3]. The exploring of code domain in CDMA increases the spectrum efficiency, but the maximum number of users is still limited by the number of user-specific codes.

With growing user density and data rate requirements, the limitations of the above-mentioned MA techniques met the capacity bottleneck and spurred the innovation in utilizing the spatial domain in Long Term Evolution (LTE) [4]. The emergence of MIMO technology transformed wireless communication by using multiple antennas at both ends of the link. By exploiting multipath propagation, MIMO increased the wireless system's capacity via spatial multiplexing, a new dimension—the Spatial domain [5]. In modern wireless communication systems (e.g., the Fourth Generation (4G), LTE and the Fifth Generation (5G)), MIMO has evolved to support multiple access schemes such as SDMA and Orthogonal Frequency Multiple Access (OFDMA) [6]. These combinations enable the BS to allocate frequency-time resources to multiple users, which significantly increases spectrum efficiency and wireless network capacity.

1.2 Multiple access in MIMO system

In the past decade, the prominent MA techniques that have been studied in literature are SDMA, NOMA and RSMA [7]. Each of them employs unique principles to enhance the capability, efficiency and flexibility of MIMO system. SDMA utilizes the spatial dimension in MIMO systems to distin-

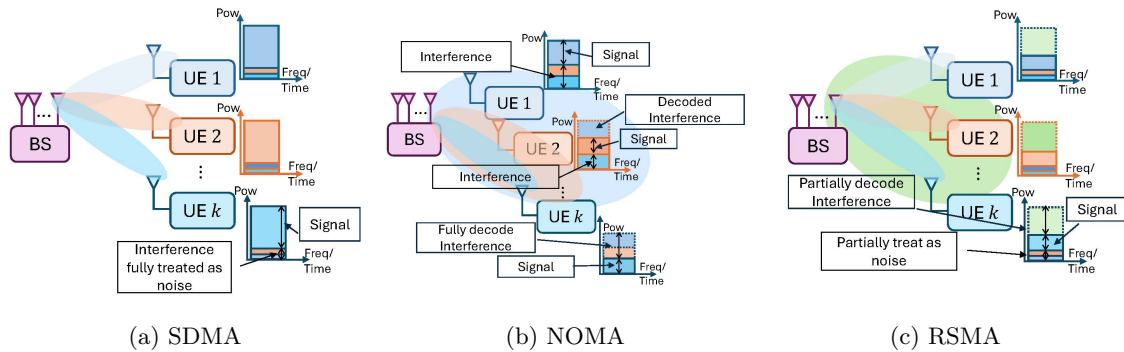


Figure 1.1: Multiple Access Techniques in MIMO

guish users by assigning them different spatial beams based on their channel spatial orthogonality or directions, as shown in Fig. 1.1a. The fundamental concept of SDMA lies in leveraging spatial separation provided by multiple Transmitter (TX) antennas and beamformers to serve multiple users simultaneously over the same frequency and time resources. SDMA can achieve high spectral

efficiency through the simultaneous transmission to multiple users and minimize the inter-user interference when users are spatially orthogonal. However, SDMA treats inter-user interference fully as noise. This feature requires BS to rely heavily on the accuracy of CSIT to null the inter-user interference effectively [8]. Also, SDMA cannot handle situations where users have highly correlated channels, and the number of users is larger than the number of transmitter antennas (overloaded situation).

Power domain NOMA, as shown in Fig. 1.1b, allows multiple users to share the same frequency and time resources by differential them in the power domain. Unlike SDMA, NOMA BS serves multiple users by superposition coding, which assigns different power levels based on their channel conditions [9]. The users in NOMA MIMO system are required to have Successive Interference Cancellation (SIC) receiver to decode inter-user interference. In a NOMA system, the power allocation is based on the user channel strengths (stronger channels are allocated with lower power), as shown in Fig. 1.1b, UE- k has the strongest channel and also required to decode all the inter-user interference fully. The advantage of NOMA compared to SDMA is the capability of supporting massive connectivity by overlapping users with the same time-frequency resources in different power levels [10]. The drawbacks of NOMA are also obvious. First of all, NOMA can not efficiently utilize the spatial domain, which causes a waste in multiplexing gain, as analyzed in [8], both sum multiplexing gain and max-min fairness gain of NOMA are incompatible with MUP. Second, NOMA requires the BS to acquire precise power control based on the CSI feedback, which is hard to maintain in a dynamic environment. Last but not least, users with stronger channels tend to receive lower power allocations, which can lead to unfair resource distribution, and this fairness is a persistent challenge in NOMA [11].

RSMA is a flexible and generalized MA framework that splits the users' messages into common and private parts; the common parts from all the users combined into a common message intended for all users, are decoded by all RXs [12]. The private messages are decoded by the specific users only. This approach allows the inter-user interference to be partially decoded by the user and partially treated as noise, which unifies two extremes (SDMA and NOMA) [13]. RSMA has proven to be more effective in managing inter-user interference compared to SDMA and NOMA, both in theoretical analyses [14] and experimental evaluations [15], [16]. The robustness of RSMA makes it outperform both SDMA and NOMA in various scenarios: First, in an overloaded MGM system, RSMA remains highest Degree of Freedom (DoF) among all three techniques [17], [18], and also highest measured throughput in an overloaded experimental setup [19]. Second, RSMA also shows its superiority in other applications, like NOUM [20], ISAC [21] and satellite communications [22].

Thirdly, RSMA can also maintain the performance gain over SDMA and NOMA with imperfect CSIT [23].

SDMA, NOMA, and RSMA represent distinct approaches to multiple access in MIMO systems, each tailored to specific challenges and scenarios. While SDMA excels in leveraging spatial dimensions, NOMA provides power-domain flexibility for massive connectivity. RSMA, as a unified framework, combines the strengths of SDMA and NOMA, offering enhanced flexibility and efficiency in managing interference and meeting diverse user demands. This makes RSMA a promising candidate for next-generation wireless networks such as the Sixth Generation (6G).

1.3 Research objectives and motivation

The evolution of wireless communication systems toward 6G necessitates innovative techniques to optimize spectral efficiency, energy utilization, and interference management. RSMA has emerged as a promising solution, offering enhanced flexibility and performance across diverse scenarios. Even studies in the past literature have shown RSMA's superiority in theory and simulations [7], there are no experimental studies to compare RSMA with the conventional MA (SDMA) and another potential candidate (NOMA). The research gap between theoretical simulations and hardware implementation is waiting to be filled by prototyping research. This section consolidates the prototype research objectives and motivations from four pivotal studies exploring RSMA's potential.

The first study focuses on building a 2×2 Multiple Users (MU)-MISO testbed for cross-comparing RSMA, SDMA and NOMA's performance. This study confirms RSMA's consistent performance gains across various conditions, including Line-of-Sight (LoS) environments and imperfect CSIT. By achieving higher throughput and fairness simultaneously, RSMA is established as a future-ready solution for handling multi-user interference in 6G.

Expanding RSMA's application, the second study investigates its utility in NOUM scenarios. This paradigm involves users simultaneously receiving shared (multicast) and unique (unicast) messages, which are common in applications like live streaming and vehicular communications. By leveraging RSMA's ability to combine multicast and unicast streams flexibly, the study demonstrates higher unicast sum rates compared to Mulp and orthogonal approaches. These findings underscore RSMA's capability to enhance data rates in environments with channel correlation.

The third study extends the scope to Overloaded MGM setups, where the number of users ex-

ceeds available antennas. Such configurations are critical for scenarios like satellite communications and federated learning. RSMA's hybrid strategy—partially decoding interference while treating residuals as noise—achieves superior fairness and throughput compared to SDMA and NOMA. The research highlights RSMA's adaptability in balancing competing demands for fairness and efficiency under complex network conditions.

The final study focuses on RSMA for ISAC. It addresses the pressing need to optimize precoder designs for systems that simultaneously serve communication users and sense surrounding targets. Unlike conventional SDMA, RSMA's capacity to handle interference more effectively is experimentally validated, revealing its superiority in scenarios with high inter-user interference levels. The study establishes RSMA as a robust technique for achieving an optimal balance between communication throughput and sensing accuracy, particularly without requiring dedicated sensing signals.

1.4 Thesis outline and contributions

The structure of this thesis is as follows:

- **Chapter 1:** A short introduction of MA techniques in MIMO system, and the research objectives and motivations.
- **Chapter 2:** A description of the system model of SDMA, NOMA and RSMA, and the optimization method used in the precoder design. The standards (IEEE 802.11g and IEEE 802.11ac-VHT) applied in the prototyping testbed to generate the OFDM signal. The hardware introduction about SDR equipment and instructions to build the testbed. In the last of this chapter, the background knowledge of the application (ISAC) in this work is introduced.
- **Chapter 3:** The details of the first prototype of the RSMA testbed and the first experimental comparison across SDMA, NOMA and RSMA. Through measurements using the prototype, the modulation and coding scheme-limited sum throughput maximization problem is solved empirically for RSMA, SDMA and NOMA for the two-user multiple-input single-output (MISO) scenario over (a) different pairs of line-of-sight channels that vary in terms of their relative pathloss and spatial correlation, and with (b) different channel state information quality. The observations show that RSMA achieves the highest sum throughput across all these cases, whereas SDMA and NOMA are effective only in some cases. Furthermore, RSMA

also achieves better fairness at a higher sum throughput than both SDMA and NOMA. These empirical results are consistent with theoretical predictions.

- **Chapter 4:** This chapter studies one of the applications of RSMA, NOUM, which is a variant of multi-antenna multi-user communications where the users desire a shared message (multicast) in addition to their respective unique messages (unicast). The multicast rate is capped in many emerging NOUM applications, such as live-event broadcasting, location-based services and vehicular communications. Given this constraint, the prototype experimentally shows that when the user channels are highly correlated, RSMA-based NOUM can meet the multicast rate while supporting a larger unicast sum rate than MUP-based NOUM and orthogonal unicast multicast.
- **Chapter 5:** This chapter presents the first-ever experimental evaluation of RSMA-based MGM, as well as the first-ever three-way comparison of RSMA-based, SDMA-based and NOMA-based MGM. Using a measurement setup involving a two-antenna transmitter and two groups of two single-antenna users per group, we consider the problem of realizing throughput (max-min) fairness across groups for each of three multiple access schemes, over nine experimental cases in a line-of-sight environment capturing varying levels of pathloss difference and channel correlation across the groups. Over these cases, the observations show that RSMA-based MGM achieves fairness at a higher throughput for each group than SDMA- and NOMA-based MGM. These findings validate RSMA-based MGM's promised gains from the theoretical literature.
- **Chapter 6:** This chapter studies a canonical use case of ISAC in MIMO systems involves a multi-antenna transmitter communicating with K users and sensing targets in its vicinity. For this setup, precoder design is of utmost importance, as the limited transmit power budget must be efficiently directed towards the desired directions (users and targets) to maximize both communications and sensing performance. This problem has been widely investigated analytically under various design choices, in particular (a) whether or not a dedicated sensing signal is needed, and (b) for different MIMO multiple access techniques, such as SDMA and RSMA. However, a conclusive answer on which design choice achieves the best ISAC performance, backed by experimental results, remains elusive. This chapter addresses this vacuum by experimentally evaluating and comparing the performance of four different design choices (i.e., with/without a dedicated sensing signal and using SDMA/RSMA) using SDRs to communicate with two users ($K = 2$) and sensing (ranging) one target. Over three scenarios that are representative of *vehicular* ISAC, covering different levels of inter-user

interference and separation/integration between sensing and communications, we show that RSMA without a dedicated sensing signal achieves better ISAC performance – i.e., higher sum throughput for the same radar Signal to Noise Ratio (SNR) – than SDMA with a dedicated sensing signal.

- **Conclusion:** The conclusion chapter summarizes the main achievements of this thesis. The limitations of works in this thesis are also discussed and will lead to future works.

1.5 Published material

The work in this thesis included two published journal papers, one conference paper and one ongoing journal paper.

Journal papers:

- X. Lyu, S. Aditya, J. Kim and B. Clerckx, “Rate-Splitting Multiple Access: The First Prototype and Experimental Validation of Its Superiority Over SDMA and NOMA,” in IEEE Transactions on Wireless Communications, vol. 23, no. 8, pp. 9986-10000, Aug. 2024, doi: 10.1109/TWC.2024.3367891.
- X. Lyu, S. Aditya and B. Clerckx, “Rate-Splitting Multiple Access for Overloaded Multi-Group Multicast: A First Experimental Study,” in IEEE Transactions on Broadcasting, early access, doi: 10.1109/TBC.2024.3475743.
- X. Lyu, S. Aditya and B. Clerckx, “Rate-Splitting Multiple Access for Integrated Sensing and Communications: A First Experimental Study,” ongoing journal paper, submitted to IEEE Transactions On Wireless Communications, (available: <https://arxiv.org/abs/2412.12037>).

Conference Paper:

- X. Lyu, S. Aditya and B. Clerckx, “Rate-Splitting Multiple Access for Non-Orthogonal Unicast Multicast: An Experimental Study,” 2024 IEEE 25th International Workshop on Signal Processing Advances in Wireless Communications (SPAWC), Lucca, Italy, 2024, pp. 591-595, doi: 10.1109/SPAWC60668.2024.10694383.

2

Fundamentals of Multiple Access, Applications and Test-bed

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2.1 Fundamentals about Multiple Access

This chapter provides fundamental knowledge on various MA techniques, downlink system models, experimental hardware setups, and the underlying principles of integrated sensing and communication.

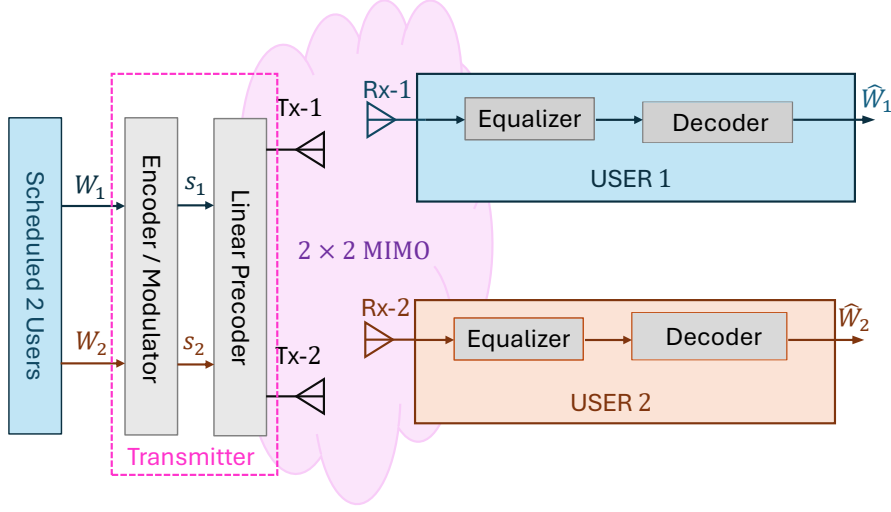


Figure 2.1: SDMA system model diagram

2.1.1 SDMA

The conventional MA, known as SDMA, is widely used in the modern wireless communication system. This section considers a system model of one BS with two single antenna users. As shown in Fig. 2.1, the transmitter (BS) schedules two users, W_1 and W_2 are the dedicated message for User (UE)-1 and UE-2, the message is encoded and modulated into streams s_1 and s_2 , respectively. The two streams symbols $\mathbf{s} = [s_1, s_2]$, and the expectation of symbols are defined as $\mathbb{E}[\mathbf{s}\mathbf{s}^H] = \mathbf{I}$. Both streams are linear precoded by precoders $\mathbf{P} = [\mathbf{p}_1, \mathbf{p}_2]$ into two streams and transmitted by two transmit antennas. The transmit power is constrained by the precoder power constraints $\|\mathbf{p}_1\|^2 + \|\mathbf{p}_2\|^2 \leq 1$. After precoding, the transmit signals are given by

$$\mathbf{x} = \mathbf{p}_1 s_1 + \mathbf{p}_2 s_2. \quad (2.1)$$

The channel vector of user- k is denoted as $\mathbf{h}_k$ so the received signal of user- k is:

$$y_k = \mathbf{h}_k^H \mathbf{x} + n_k, \quad \forall k \in \{1, 2\}, \quad (2.2)$$

where $n_k \sim \mathcal{CN}(0, \sigma_0^2)$ is the Additive White Gaussian Noise (AWGN) for UE- k . On the receiver side, each user decodes its own stream by treating other user's signals as interference. The resulting Signal to Interference Noise Ratio (SINR) of UE- k can be denoted as

$$\gamma_k = \frac{|\mathbf{h}_k^H \mathbf{p}_k|^2}{\sigma_0^2 + |\mathbf{h}_k^H \mathbf{p}_j|^2}, \quad \forall k \in \{1, 2\}, j \neq k, \quad (2.3)$$

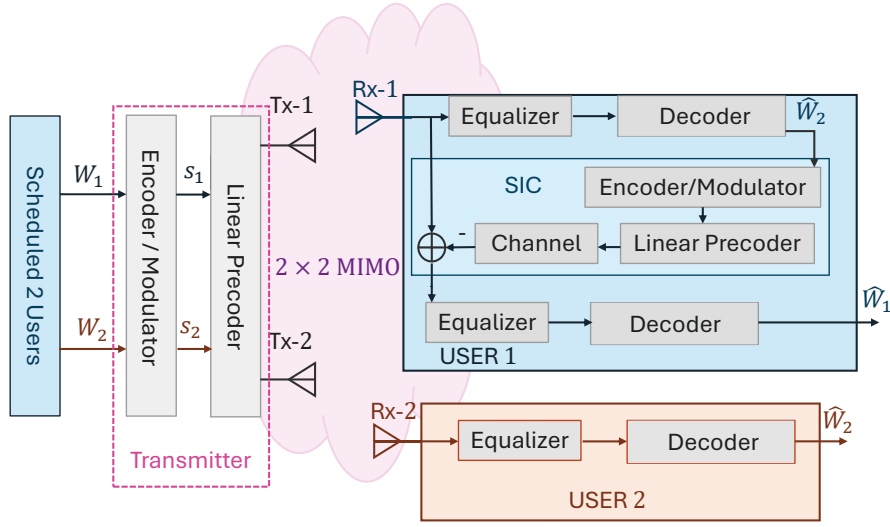


Figure 2.2: NOMA system model diagram

and by assuming Gaussian signalling, the capacity of UE- k is given by $R_k = \log_2(1 + \gamma_k)$, the sum capacity of SDMA in a two-users MISO system is $R_{sum} = \sum_{k=1,2} R_k$.

2.1.2 NOMA

NOMA normally refers to power domain NOMA, which using superposition coding at the TX side and the SIC receiver at the UEs' side. In the NOMA system, at least one UE needs to decode all other UEs' messages fully. Fig. 2.2 shows a system diagram of a two-user MISO system as same as Fig. 2.1, but the UE-1s' receiver is replaced with a SIC receiver¹. The received signal at UE- k can also be expressed by (2.2). In UE-1's decoding process, UE-1 decodes UE-2's message $\hat{W}_2$ first, then removes the interference from the received signal to decode its message. As for UE-2, the receiver decodes UE-2's message directly. The corresponding SINR for UE-1 assuming perfect SIC is expressed as:

$$\gamma_1 = \frac{|\mathbf{h}_1^H \mathbf{p}_1|^2}{\sigma_0^2}, \quad (2.4)$$

and the SINR expression of UE-2 is :

$$\gamma_2 = \frac{|\mathbf{h}_2^H \mathbf{p}_2|^2}{\sigma_0^2 + |\mathbf{h}_2^H \mathbf{p}_1|^2}. \quad (2.5)$$

By assuming Gaussian signalling, the capacity of UE- k is given by $R_k = \log_2(1 + \gamma_k), \forall k \in \{1, 2\}$, and the sum capacity of NOMA in a two-users MISO system is $R_{sum} = \sum_{k=1,2} R_k$.

¹In theory, both UEs are required to equip SIC receiver. In this system diagram, only UE-1 is required to perform SIC, for simplicity, the diagram only shows the SIC decoder structure of UE-1

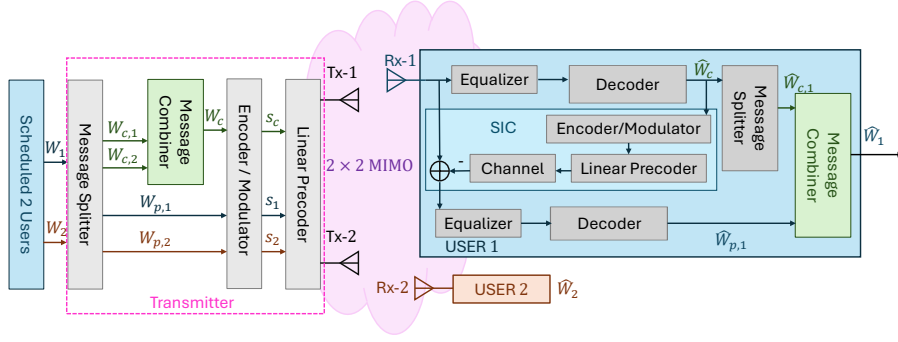


Figure 2.3: RSMA system model diagram

2.1.3 RSMA

Different from the previous two MA techniques' TX architecture, RSMA's TX contains a message splitter. It splits the message scheduled for UE- k (W_k) into common part ($W_{c,k}$) and private part ($W_{p,k}$). The message intends for UE- k can be expressed as $W_k = \{W_{c,k}, W_{p,k}\}$, which is the combination of UE- k 's common part and private part. The message combiner combines the common parts from two scheduled UEs into a common message W_c , which is encoded and precoded into the common stream s_c . Since s_c contains both UEs' messages, it must be coded by the same codebook shared among those two UEs and decoded by both UEs. The private parts $W_{p,1}$ and $W_{p,2}$ are encoded and precoded into private streams s_1 and s_2 individually. The streams $\mathbf{s} = [s_c, s_1, s_2]$ are precoded by precoders $\mathbf{P} = [\mathbf{p}_c, \mathbf{p}_1, \mathbf{p}_2]$ separately, into transmit signal :

$$\mathbf{x} = \mathbf{p}_c s_c + \mathbf{p}_1 s_1 + \mathbf{p}_2 s_2 \quad (2.6)$$

The expectation of streams symbols are defined as $\mathbb{E}[\mathbf{s}\mathbf{s}^H] = \mathbf{I}$, the transmit power constraints of the precoder vector is $\|\mathbf{p}_c\|^2 + \|\mathbf{p}_1\|^2 + \|\mathbf{p}_2\|^2 \leq 1$. By assuming the channel between the BS and UE- k is $\mathbf{h}_k$, the received signal at UE- k can be denoted by (2.2). As shown in Fig. 2.3, each UE- k decodes the common stream s_c first into $\hat{W}_c$ by treating the signal of the private streams as noise, then the decoded common message is sent to the message splitter and the SIC module simultaneously, the message splitter splits the decoded message $\hat{W}_c$ to obtain the part of the common message belongs to UE- k ($\hat{W}_{c,k}$). In the meantime, the SIC decoder subtracts the interference caused by the common stream and decodes the private stream s_k by treating other UEs' private stream as noise, the decoded private message intended to UE- k is $\hat{W}_{p,k}$. UE- k can reconstruct the message by combining the common part ($\hat{W}_{c,k}$) and the private part ($\hat{W}_{p,k}$). By assuming Gaussian signalling

and perfect SIC, the SINR of the common stream of UE- k is

$$\gamma_{c,k} = \frac{|\mathbf{h}_k^H \mathbf{p}_c|^2}{\sigma_0^2 + |\mathbf{h}_k^H \mathbf{p}_1|^2 + |\mathbf{h}_k^H \mathbf{p}_2|^2}, \forall k \in \{1, 2\}, \quad (2.7)$$

and the SINR of UE- k 's private stream is

$$\gamma_{p,k} = \frac{|\mathbf{h}_k^H \mathbf{p}_k|^2}{\sigma_0^2 + |\mathbf{h}_k^H \mathbf{p}_j|^2}, \forall k \in \{1, 2\}, j \neq k. \quad (2.8)$$

The common stream is required to be decoded by both UEs, so the capacity is limited by the minimum capacity among both UEs. By assuming Gaussian signalling, the capacity of the common stream is

$$R_c = \min(\log_2(1 + \gamma_{c,1}), \log_2(1 + \gamma_{c,2})). \quad (2.9)$$

The capacity of UE-1's private stream can be denoted as $R_{p,k} = \log_2(1 + \gamma_{p,k})$, then the sum capacity of RSMA in a two-user MISO system is written as:

$$R_{sum} = R_c + \sum_{k=1,2} R_{p,k}. \quad (2.10)$$

2.2 Fundamentals about Communication Applications

This section introduces the five main communication applications studied in this thesis: unicast, multicast, multi-group multicast, non-orthogonal unicast and multicast, and ISAC. This section also reviews the performance of RSMA in providing different communication applications compared to the conventional methods.

2.2.1 Basics about Unicast, Multicast

In wireless networks, when multiple users request content, the BS has two major methods to deliver it: unicast and multicast.

Unicast involves the BS transmitting a unique copy of the message to each individual user. This approach ensures a dedicated connection for each RX, providing personalized message delivery [24]. However, it can significantly increase the load on the BS and consume more network resources, such as bandwidth and transmission power, especially when the number of RXs requesting the same content grows.

Multicast, on the other hand, allows the BS to send a single copy of the message that can be received by all the users interested in that message. This method is highly efficient in scenarios where multiple users are requesting the same message simultaneously, such as live video streaming or public alerts [25]. By using multicast, the BS reduces redundancy, minimizes bandwidth usage, and optimizes resource allocation. Multicast has also faced several challenges, such as unreliable connections and limited throughput by the multicast group's weakest user.

Choosing between unicast and multicast depends on various factors, such as the number of users requesting the message, the nature of the data, and the network's capacity to handle simultaneous transmissions. While unicast offers individualized service at the cost of higher resource consumption, multicast is more scalable and efficient for delivering common data to multiple users.

2.2.2 Non-Orthogonal Unicast and Multicast

Non-orthogonal Unicast Multicast (NOUM) is a unique form of multi-antenna multi-user communications where users desire both multicast (shared) and unicast (unique) messages. These could arise in live-event broadcasting (where the unicast messages could be social media interactions), Vehicle to Everything (V2X) networks and Augmented Reality/Virtual Reality (AR/VR) applications (where the multicast message could be location-dependent information), etc. NOUM is a spectrally efficient way to jointly realize (physical layer) multicast and unicast in multi-antenna systems, with great potential for next-general wireless networks [26], provided adequate multicast and unicast data rates can be guaranteed.

The conventional way of realizing NOUM is by extending Mulp [27]–[30]; i.e., apart from one precoder per unicast message, there is also a multicast precoder. The precoder design amounts to superposition coding at the TX, with each user decoding two messages (the multicast message followed by its unicast message) via SIC.

The above scheme closely resembles RSMA, which has been shown to improve the performance of multi-antenna multi-user communications in several aspects, such as spectral efficiency, energy efficiency, mobility etc. [7], [31]. In particular, *unicast* RSMA involves the creation of $K + 1$ data streams and precoders for K users. The additional stream – known as the *common stream* – is composed of parts of each user's message, as illustrated in Table 2.1. Similar to Mulp-based NOUM, each user employs SIC to first decode the common stream (s_c), followed by its so-called *private stream* ($s_i, i \in \{1, 2\}$) to retrieve its desired message. This two-step retrieval process allows each user to partially decode and subtract the inter-user interference (through the common stream),

Stream	MULP-based NOUM	RSMA unicast	RSMA-based NOUM
s_c	W_0	$[W_{c,1}, W_{c,2}]$	$[W_0, W_{c,1}, W_{c,2}]$
s_1	W_1	$W_{p,1}$	$W_{p,1}$
s_2	W_2	$W_{p,2}$	$W_{p,2}$

Table 2.1: Comparison between MULP-based NOUM, RSMA unicast and RSMA-based NOUM for the two-user case. W_0 is the multicast message, and W_i is the unicast message for user i ($= 1, 2$). For RSMA, $W_i := [W_{c,i}, W_{p,i}]$. MULP-based NOUM is a special (and in general, suboptimal) case of RSMA-based NOUM.

and partially treat it as noise (i.e., the private streams of other users). This, in turn, increases the sum rate [15], [23].

In fact, MULP-based NOUM can be viewed as a special case of RSMA, where the common stream carries only the multicast message instead of a mixture of unicast messages. However, RSMA offers more flexibility for NOUM w.r.t the composition of the common stream, which translates to better rate [20] and latency [32] performance. To see how, two points need to be noted:

- (a) In many NOUM applications, the multicast rate is capped (e.g., video streaming, V2X safety messages, etc. [33]). Hence, a well-motivated objective is to maximize the unicast sum rate, subject to satisfying the multicast rate.
- (b) The common stream in RSMA unicast allows all users to partially decode and eliminate the interference from other users.

In light of (a), MULP-based NOUM only allocates enough resources (e.g., precoder power) to the common stream to meet the desired multicast rate. However, with RSMA-based NOUM, the common stream can be resourced to support a higher rate than the multicast rate, with the difference contributing to the unicast sum rate. This can be realized by designing the common stream to be a mixture of the multicast message and parts of each user's unicast message, as shown in Table 2.1. Just like (b), the unicast message components in the common stream help with interference suppression, which can increase the unicast sum rate beyond what is achievable with MULP-based NOUM.

2.2.3 Multi-Group-Multicast

Multi-group multicast (MGM) is a special case of multi-user wireless communications, wherein a multi-antenna transmitter (TX) jointly communicates with several *groups* of users simultaneously.

The g -th group (integer g) has N_g users, where $N_g \geq 1$ in general, and each user within this group desires the same message from the TX. The special case of a single group corresponds to multicast/broadcast communications, whereas $N_g = 1$ (for each g) corresponds to unicast communications.

MGM is a key physical (PHY) layer enabler of several *group* applications, such as video streaming, push-to-talk group radio communications, location-based services, safety-critical vehicular communications, etc [34]. The importance of these applications in future wireless networks is driving ongoing standardization efforts on how best to support multicast/broadcast service (MBS) in future wireless standards, most notably in the 3rd Generation Partnership Project (3GPP) [35].

Analogous to unicast, MGM can be realized through linear precoding (beamforming) at the TX, but with two key differences that need to be factored into the precoder design:

- MGM involves the additional constraint that each user within a group must be able to decode its desired message. In general, a wide range of *spatial geometries* may be possible in terms of group composition, resulting in a great variety of channels experienced by users in a group. However, a special case of considerable practical interest is where users within a group are closely situated and experience *similar* channels – spanning a spectrum from similar channel strengths at one end, all the way to high spatial correlation/alignment at the other. This could arise in applications like gaming, video conferencing, vehicular swarms, etc. Hence, we focus on this special case in this thesis.
- Additionally, in many MGM applications such as satellite communications [36] and Internet-of-Things (IoT) [37], the number of users typically exceeds the number of antennas at the TX, which is often referred to as an *overloaded* scenario. Interference management is particularly challenging in overloaded scenarios, since the interference power at all users cannot be suppressed to arbitrarily low levels through linear precoding² [17]. Hence, we further focus on the overloaded scenario in this thesis.

The above factors have given rise to the following precoder design approaches in the literature, motivated by well-known analogues for unicast:

- a) *SDMA-based MGM*: Most of the existing literature [38]–[44] on MGM focuses on this approach, which is an extension of Space Division Multiple Access (SDMA) for unicast. Thus,

²For instance, with N_T antennas at the TX and zero-forcing precoding, the interference can be nulled at no more than $N_T - 1$ users.

each group is treated as a super-user, resulting in one precoder per group, and the precoders are designed to suppress the power of the *inter-group* interference so that it can be treated as noise. While there are several well-motivated objectives for precoder design in MGM, arguably the most popular in the literature is (max-min) rate fairness across the different groups³. For SDMA-based MGM, this problem has been investigated analytically in many settings, namely [38], [39] (with a sum power constraint), [40] (with a per antenna power constraint), [41] (with low-complexity precoder design for large-scale transmit antennas) and [42] (for line-of-sight groups), [43] (for satellite communications), [48] (for cooperative integrated terrestrial-satellite communications). However, to the best of our knowledge, an experimental evaluation of SDMA-based MGM does not exist in the literature. The existing experimental studies have focused on implementing MGM at higher layers [49], [50], instead of the PHY layer.

- b) *NOMA-based MGM*: This approach is an extension of power domain Non-Orthogonal Multiple Access (NOMA), and is guided by the principle that decoding the interference rather than treating it as noise is more effective for overloaded scenarios, particularly for unicast [51]. Like above, this approach also yields one precoder per group, but for the two-group case, one of the groups employs successive interference cancellation (SIC) to decode and subtract the interference, just like the *stronger user* in unicast NOMA. Precoder optimization⁴ for NOMA-based MGM was investigated in [55], [56] (objective: minimizing transmit power subject to minimum rate constraints at each group). However, even for the overloaded scenario, NOMA-based MGM exhibited significant gains over SDMA-based MGM only when there is considerable disparity in channel strength across the groups [56]. Moreover, for groups of closely-spaced users with large angular separation between groups, it is plausible that SDMA-based MGM may still be effective in suppressing inter-group interference by exploiting the spatial domain, even in overloaded scenarios. The potential underperformance of NOMA-based MGM relative to SDMA-based MGM in such scenarios, despite a higher receiver complexity, is due to the former's inability to exploit the multi-antenna multiplexing gain [8], [17]. For a clearer picture of the circumstances under which NOMA-based MGM is superior to SDMA-based MGM (in terms of realizing max-min rate fairness across groups, say), an experimental study of the two schemes over a variety of group spatial geometries is needed, something that is missing in the literature.

³Other objectives for MGM precoder design include (i) minimizing transmit power subject to minimum rate constraints at each group [42]–[44], (ii) maximizing sum rate under a per antenna power constraint [45], [46], and (iii) extending zero-forcing and MMSE precoding for MGM [47].

⁴Note: NOMA-based MGM has also been investigated for single-antenna systems [52] and multi-antenna systems without precoder optimization [53], [54].

c) *RSMA-based MGM*: Rooted in the rate-splitting concept originally proposed for the two user single-antenna interference channel [57], Rate-Splitting Multiple Access (RSMA) has emerged as a powerful multiple access, interference management, and multi-user strategy for modern multi-antenna communication systems (see [7], [31] and the references therein). Inspired by its superiority over both SDMA and NOMA for unicast [13], [15], [58], RSMA-based MGM bridges a) and b) above to address their limitations. Specifically, each group *partially* decodes the inter-group interference and *partially* treats it as noise. All groups have the same signal processing complexity for retrieving their desired messages (one stage of SIC, regardless of the number of groups). Unlike a) and b), this approach yields one more precoder than the number of groups (details in Section 5.2). Precoder optimization for overloaded RSMA-based MGM⁵ was investigated in [17], [62], [63] (objective: max-min rate fairness across groups), where it was demonstrated analytically that RSMA-based MGM achieved a higher minimum rate than the other two approaches. This was validated using link-level simulations in [64], [65] (under perfect CSIT) and [63], [66], [67] (for satellite communications). However, for a clearer picture of the extent of such gains, an experimental comparison of the three MGM approaches over a variety of group spatial geometries is needed.

Indeed, it is surprising that an experimental evaluation of MGM for any of the above multiple access schemes is missing from the literature.

2.2.4 MIMO Integrated sensing and communications

ISAC is widely anticipated to be a prominent feature of future wireless networks, as evidenced by standardization activities on ISAC for 6G [68]. One strand of ISAC that has been extensively investigated by the research community is MIMO ISAC, as the MIMO paradigm is mature in both communications (comms) and radar [69]. In particular, a problem that has been widely investigated involves a multi-antenna TX simultaneously serving K comms users (UEs) and sensing targets in its vicinity, as shown in Fig. 2.4. For this problem, the comms and sensing performance is closely linked to the amount of power radiated towards the UEs and targets, respectively. As a result, the choice of precoders (beamforming) and multiple access techniques used at the TX has a direct bearing on ISAC performance. Indeed, in the context of ISAC, multiple access techniques make use of the resource dimensions (e.g., time, frequency, power, antenna, code, message, etc) to serve multiple physical users (communications) and virtual users (target to sense), ideally in the most

⁵Precoder optimization for overloaded RSMA *unicast* was investigated in [59]–[61].

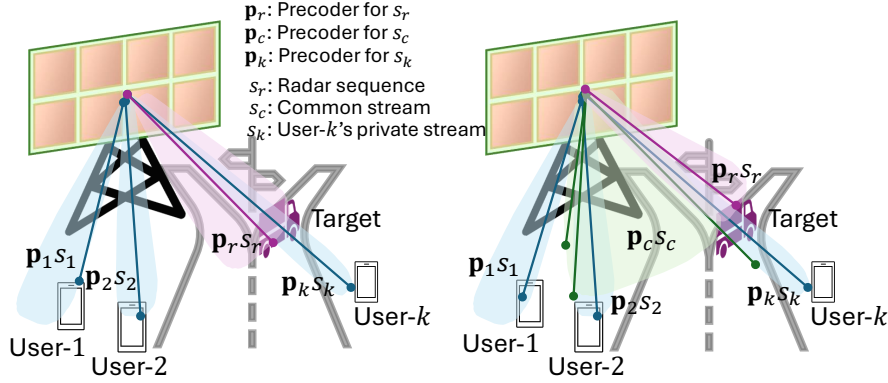


Figure 2.4: MIMO ISAC precoder design problem.

efficient way [70].

The ISAC precoder design problem in Fig. 2.4 to jointly optimize comms and sensing performance metrics has been studied extensively under several well-motivated settings – conventional MIMO with fully digital beamforming [71]–[75], massive Multiple Input Multiple Output (mMIMO) [76]–[79], vehicular systems [80], [81], millimeter-wave with hybrid beamforming [82]–[85], using low resolution analog-to-digital converters [86], etc., to name a few. There are two implicit assumptions across these works. Firstly, they can all be viewed as adapting SDMA – the prevailing multiple access technique used in 5G – for ISAC, and secondly, they either assume a dedicated sensing signal [76], [78], [79], [82], or reuse comms signals⁶ for sensing [72], [74], [75], [77], [80], [81], [83]–[86] as pre-fixed design choices without scrutinizing the merits of each choice. As a result, two fundamental questions can be posed w.r.t the ISAC precoder design problem, namely:

Q1. Are there alternatives to SDMA that could provide better ISAC performance⁷?

Q2. Are separate signals needed for comms and sensing for the best ISAC performance?

These are especially pertinent questions, as specifications related to the waveform and multiple access technique are among the first to be decided in every new generation of standards.

Regarding Q1, RSMA, which generalizes SDMA (and NOMA), has been shown both analytically and empirically to outperform SDMA (and NOMA) w.r.t comms [15], [19], [31]. Briefly, using RSMA to communicate with K UEs involves the design of $K + 1$ precoders to transmit $K + 1$ comms signals [31]. Likewise, using SDMA to communicate with K UEs and track a single target

⁶By this, we mean the data payload comprising channel-coded and modulated symbols.

⁷See Definition 1 in chapter 6 for our definition of ISAC performance.

using a dedicated sensing signal also involves the design of $K + 1$ precoders⁸. These similarities strongly motivate an RSMA v/s SDMA ISAC comparison.

As for Q2, it is important to clarify what sensing entails. For instance, if the sensing task involves detecting the presence/absence of a target or direction finding, then clearly these can be accomplished using comms signals with an energy detection receiver or subspace methods like MUSIC, respectively. Importantly, the sensing signal processing for these tasks does not depend on the *content* of the comms signal. For target ranging (i.e., distance estimation) on the other hand, the signal content impacts estimation performance as the range estimator is a matched filter. Since the answer to Q2 is non-trivial for ranging, we choose this as the sensing task in this thesis, although the terms sensing and ranging shall be used interchangeably. However, it has been shown that the ranging performance of comms signals matches that of dedicated sensing signals – e.g., Frequency Modulated Continuous Wave (FMCW) – asymptotically at large signal lengths [87].

Taken together, Q1 and Q2 motivate the following four-way ISAC comparison – RSMA/SDMA with/without dedicated sensing signal. This has been investigated in [21], [88]–[91] with link-level simulations indicating that RSMA without a dedicated sensing signal achieves the best ISAC performance – intuitively, the *extra* precoder (i.e., $K + 1$ precoders for K UEs) can be used for sensing⁹. In particular, [21] was the first to theoretically show that RSMA can be used for the dual purpose of boosting the communication sum rate of multiple users and simultaneously increasing the sensing performance, hence enlarging the communication-sensing performance trade-off. This is due to the common streams in RSMA that can be used to manage inter-user interference and sensing targets. However, a conclusive answer on which design choice achieves the best ISAC performance, backed by experimental results, is missing in the literature – experimental evaluations of ISAC for single-antenna systems is presented in [92]–[94] and for multi-antenna ISAC (SDMA) in [95].

2.3 Limitations of Theoretical Study and Challenges of Experimental Study

The evolutionary trajectory over successive generations of wireless networks has been towards achieving higher spectral efficiency, greater user fairness and higher energy efficiency. For upcoming

⁸Essentially, the target can be viewed as a virtual UE.

⁹This works only because the comms signal carried by the precoder can also be used for sensing. Thus, Q1 and Q2 in page 1 are closely related questions.

6G in particular, improvements in spectral and energy efficiency of at least 2x over 5G have been speculated [96]. To meet these targets, the underlying physical (PHY) layer multiple access technique must be capable of offering greater efficiency and flexibility, while coping with much higher levels of multi-user interference than SDMA – the state-of-the-art multiple access technique used in current wireless standards such as 3GPP 5G New Radio (NR) and IEEE 802.11. The two candidates widely believed to be capable of achieving this in future wireless networks are NOMA and RSMA. Despite the theoretical studies showing the superiority of RSMA over both NOMA and SDMA, there is no experimental study about the three-way cross-comparison.

2.3.1 NOMA and its implementation

NOMA is inspired by the fact that the capacity region of the K -user Single Input Single Output (SISO) broadcast channel is achieved by superposition coding at the TX and SIC at the RXs [97]. In SISO NOMA, users are ordered in decreasing order of their channel strength and the l -th strongest user decodes and subtracts the interference from the $K - l$ weaker users (i.e., SIC) before decoding its desired signal. The interference from the $l - 1$ stronger users is treated as noise¹⁰.

The optimality of NOMA for the SISO broadcast channel (in terms of achieving capacity) has been the driving force behind extensive investigations into NOMA for MIMO systems with the aim of improving spectral efficiency [99], [100] and fairness [101] in 5G networks and beyond [102]. However, for the multi-antenna broadcast channel, NOMA is not optimal, in general, in a capacity-achieving sense [8]. This is because with multiple antennas at the TX, the spatial domain can be used to suppress interference without SIC at the RXs through either MULP or non-linear precoding at the TX. MULP is the common approach adopted by SDMA [103]. Essentially, SDMA and NOMA represent two distinct interference management strategies; in SDMA, the interference is treated as noise, whereas in NOMA the interference from the weaker users is decoded by the stronger users [58]. Hence, NOMA provides substantial spectral efficiency gains over SDMA only when MULP cannot suppress the interference to near-noise levels. This occurs either in overloaded systems (i.e., more users than the number of antennas at the TX) or when the user channels in underloaded systems experience high spatial correlation. On the other hand, in non-overloaded systems where user channels do not experience high spatial correlation, NOMA can perform worse than SDMA [8]. While it can be argued that under high user density, groups of users with highly spatially correlated channels can be found with high probability, it does mean that MIMO-NOMA

¹⁰This is commonly referred to as power-domain NOMA to distinguish it from other variants, such as code-domain and frequency-domain NOMA [98]. In this chapter, NOMA refers only to power-domain NOMA.

requires a user grouping strategy based on their channels, which has given rise to two approaches – beamforming-based and cluster-based NOMA [104].

In terms of experimental evaluations, several papers have focussed on SISO-NOMA [105, Table 1]. MIMO-NOMA implementations are relatively fewer [106]–[108]. Link and system-level simulations were carried out in [106], along with measurements over emulated fading channels. Measurements involving over-the-air transmissions were carried out in [107], [108]. However, in these papers, the NOMA performance is compared with orthogonal multiple access instead of SDMA. To the best of our knowledge, no experimental comparison between NOMA and SDMA exists in the literature.

2.3.2 RSMA and its implementation

Unlike NOMA with its SISO roots, RSMA was designed for multi-antenna communications from the outset. For the K -user MISO case, RSMA produces $K + 1$ data streams (i.e., channel-coded and modulated symbol streams) at the TX from the K user messages [23] (details in Section 3.2.2). In contrast, both SDMA and NOMA generate K data streams. Similar to SDMA, each of the $K + 1$ streams is linearly precoded (i.e., Mulp) and transmitted over the air. At each RX, SIC is used (like NOMA) to decode *only two* streams to retrieve the desired message, regardless of a user's channel strength relative to the other users. Hence, unlike NOMA, no user grouping mechanism prior to data transmission is needed to realize spectral efficiency gains over SDMA.

RSMA's spectral efficiency gains over SDMA and NOMA stem from the fact that it is DoF optimal under perfect and imperfect CSI [8], [23]. Furthermore, RSMA has also been shown, in theory, to achieve spectral efficiency gains over SDMA and NOMA under varying interference levels, network loads and user deployments [58]. Apart from spectral efficiency, RSMA has also been shown to achieve – again, in theory – performance improvements over SDMA and NOMA in terms of other metrics such as energy efficiency, fairness, mixed-critical Quality of Service (QoS) and tight latency requirements, etc. (see [7], [31] and the references therein for a comprehensive survey). These performance improvements stem from RSMA's unique interference mitigation strategy, wherein each user *partially decodes* the interference and *partially treats it as noise*. In doing so, RSMA generalizes both SDMA and NOMA [13], [58]. Taken together, all of these features make RSMA an attractive candidate for PHY layer multiple access in 6G, provided the promised gains can be demonstrated in practice. In particular, an experimental comparison between RSMA, SDMA and NOMA is needed as a starting point to address the pressing question of which multiple

access technique is best equipped to deliver the spectral efficiency enhancements expected from 6G.

Despite its impressive benefits, the practical implementation of RSMA is still in its infancy, with the state-of-the-art focusing on the closely related issues of erroneous SIC, RX design, and Adaptive Modulation and Coding (AMC) for RSMA to overcome the combined effects of imperfect/outdated CSIT and finite blocklength [109]–[111]. Among these, RSMA link-level simulations were conducted in [109]–[111], feedback-based RSMA link adaptation was investigated in [112], while [12] considered a flexible variant of RSMA, where only a subset of the RXs employ SIC. Though significant, these practical aspects have also been tackled only through theory/simulations, and no experimental verification of RSMA's benefits has been reported till date.

2.4 Experimentation design and test-bed implementation

The primary objective of this thesis is to experimentally validate the theoretical gains observed in simulations. To achieve this, a testbed with MIMO testing capability is required. Specifically, the base station must feature at least two transmit antennas with synchronized RF chains and two single-antenna users. Each RF chain should allow for reconfigurable or customizable signal generation to implement the RSMA architecture on the hardware. Additionally, the hardware must meet the current wireless standards and criteria, such as bandwidth and carrier frequency requirements. Taking these considerations into account, we selected SDR devices from National Instruments (NI) and chose NI LabVIEW NXG 5.1 as the programming platform, running on a desktop computer. This chapter provides an overview of the software, hardware configurations, specifications, and the IEEE 802.11 family-based framework used for signal generation.

2.4.1 SDR test-bed implementation

The NI USRP is famous for its MIMO capability, one of the most famous prototypes is the mMIMO prototype in [113]. The mentioned prototype consists of 128 TX antennas and achieved a world-first spectral efficiency of 79.4 bits/s/Hz. NI provided some example software programs and the compatible hardware drives in the LabVIEW NXG, which can be used in the transceiver signal design to simplify the whole design procedure, Fig. 2.5 shows the UI of the software control panel and Fig. 2.6 shows the function blocks in the software. The input parameters in the control panel in Fig. 2.5 are sent and transferred through the building blocks in Fig. 2.6.

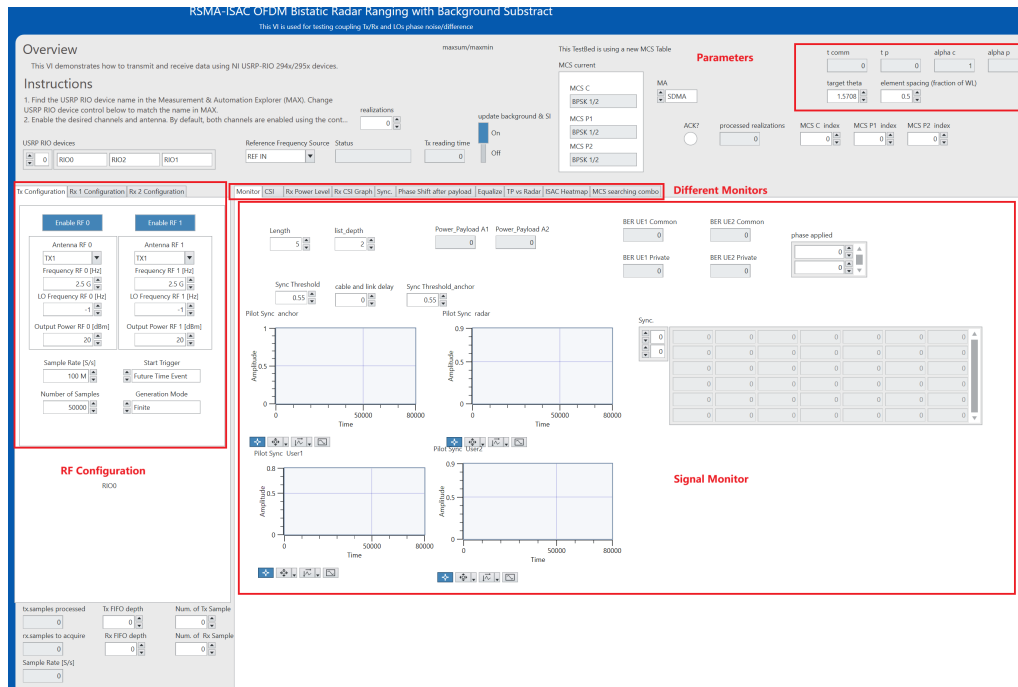


Figure 2.5: The User Interface of the software control panel

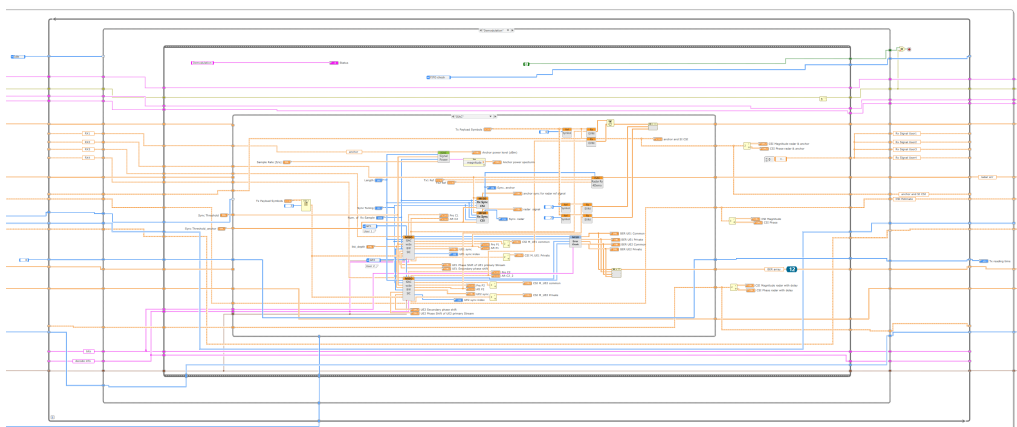


Figure 2.6: The function blocks diagram of the software running behind the UI

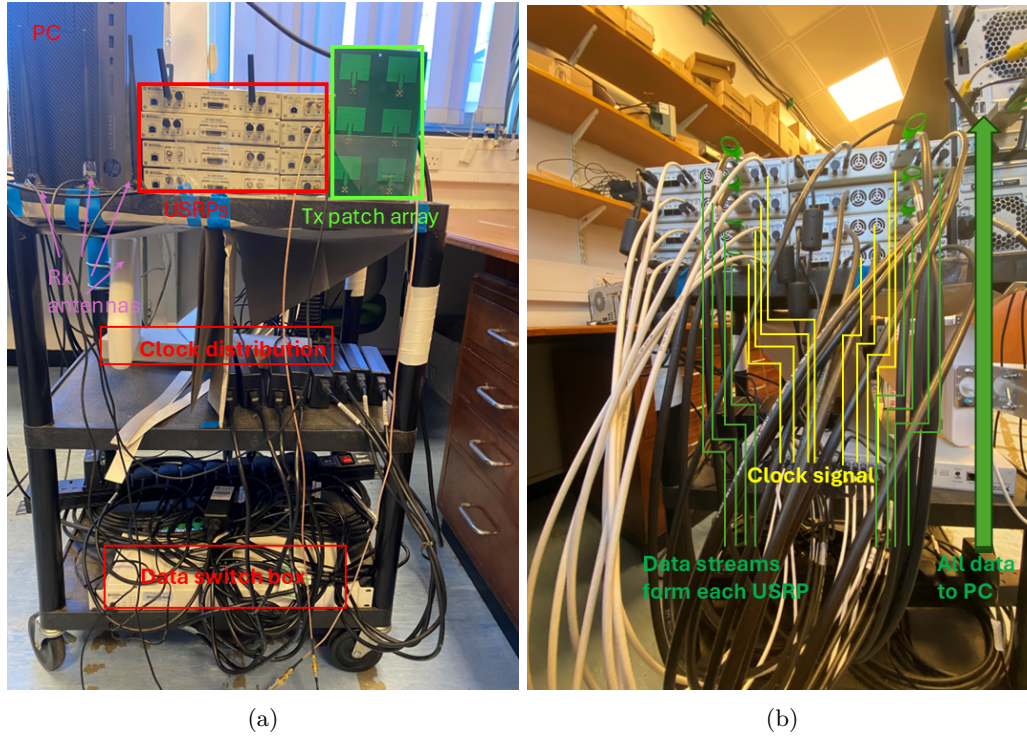


Figure 2.7: Pictures of the MIMO test-bed. The left shows the front side of the hardware used in this thesis. PC USRPs and antennas are on the top layer, a common clock reference distributor is on the second layer, and the bottom layer is the data switch box. The right shows the back side of the hardware and demonstrates the interconnection between the hardware. All USRPs reference signal input ports are connected to a common reference signal clock distributor for synchronization propose, and all the data transmitted/received from the USRPs are written/read through the PCIe cable to the switch box, the switch box schedule the data to be written/read by the PC.

In this thesis, we are using the same devices used in [113]. Fig. 2.7 shows the hardware used in this thesis, the front side (Fig. 2.7a) of the test bed shows the name of the main devices in the test bed and the back side (Fig. 2.7b) of the test bed shows the interconnection of main devices.

The main functions and specifications of the devices are listed below:

- A Host PC: Intel Core i7-8700 processor with a LabVIEW NXG 5.1 running on it. The role of the Host PC is running the function blocks for signal generation and controlling the configurable settings of the USRPs. After building the necessary blocks in LabVIEW NXG 5.1, the signals can be simply controlled in the UI of the software.
- USRP-2942R: One single USRP-2942R consists of one motherboard and two daughter boards (SBX-120), which provide 120 MHz bandwidth. The carrier frequency range of the RF front is from 400 MHz to 4.4 GHz. Each daughterboard provides one RF chain consisting of two ports; one of those ports can be used as TX/RX, and the other can be used as RX only. On the back side of the USRP, it has a reference clock input port, which is connected to a

common clock distributor (CDA-2990) for synchronization purposes. The back side of the USRP also has a PCIe port for reading/writing the data from the PC via a switch box (CSP-8910). By replacing the daughterboard of the USRP from SBX-120 to UBX-160, each port in USRP is able to work in a frequency range from 10 MHz to 6 GHz and 160 MHz bandwidth. This daughterboard swapping feature of USRP offers a high flexibility for USRP to be able to test different wireless communication applications.

- CDA-2990: An 8-channel clock distribution device helps to synchronize up to 8 USRPs in the system. It generates a 10 MHz pulse-per-second signal and distributes it through those eight ports.
- CPS-8910: A PCIe switch box designed for MIMO expansion configurations. It provides one PCIe×8 upstream port for data exchanging between it and PC, and 8 PCIe×4 ports for data exchanging between the USRPs with the switch box.
- PXIe-8381: An MXI-Express remote controller for CPS-8910 cabled over PCIe from a computer with a PCIe slot.
- Yagi-antenna: The directional antenna for receiving radar echo signals.
- Tx array: Each element on the transmitter antenna array is designed to work in a center frequency of 2.4 GHz, and the spacing between each element is half-wavelength.
- Omni-directional antenna: The Omni-directional antennas with extended cables are used as single antenna UEs.

2.4.2 IEEE 802.11 based frame structure

Wireless Local Area Networks (WLANs) play an important role in modern communication systems, driven by the demand for seamless connectivity, high data rates and robust performance. The IEEE 802.11 family of standards, commonly refer as Wireless Fidelity (Wi-Fi), defines the protocols and technologies underpinning WLANs operations. Among these, we find IEEE 802.11g (20 MHz) [114] and IEEE 802.11ac-VHT (160 MHz) [115] are the most suitable standards for our testbed, which have been deployed in millions of products.

IEEE 802.11g, operating in the legacy 20 MHz channel bandwidth, laid the groundwork for Wi-Fi technology by enabling reliable and efficient wireless communication. This configuration is widely utilized in scenarios prioritizing spectrum efficiency and compatibility with a broad range

of devices [116]. Its channel bandwidth of 20 MHz provides a balanced trade-off between spectral usage and data rate performance, making it suitable for dense deployment environments and ensuring broad interoperability.

In contrast, IEEE 802.11ac introduced significant advancements to address the growing demand for high-throughput applications, such as video streaming, cloud computing, and augmented reality [117]. A key enhancement of IEEE 802.11ac is the Very High Throughput (VHT) mode, which includes support for channel bandwidth up to 160 MHz. By utilizing contiguous or non-contiguous channel bonding, IEEE 802.11ac-VHT (160 MHz) dramatically increases the available bandwidth, enabling peak data rates exceeding 1 Gbps. Adopting advanced techniques, such as MU-MIMO and higher-order modulation (up to 256-Quadrature Amplitude Modulation (QAM)), further amplifies its capacity to deliver high-performance wireless connectivity in indoor and outdoor environments.

This evolution from 20 MHz channels to 160 MHz reflects the dynamic adaptation of IEEE 802.11 standards to the ever-expanding demands of wireless applications. Exploring the characteristics, performance trade-offs, and application scenarios of these configurations provides valuable insights into the optimization of WLANs systems for diverse use cases. For demonstrating RSMA's advantage in different applications, we need both of them. For the cross-comparison with SDMA and NOMA in unicasting, MGM and NOUM, we need the stability of IEEE 802.11g. For ISAC performance comparison, we need the bandwidth from IEEE 802.11 ac-VHT to get a finer resolution in target-ranging performance.

3

Rate-Splitting Multiple Access

Experimental Study of its

Superiority over SDMA and

NOMA

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3.1 Introduction

This chapter addresses the critical knowledge gap between the theoretical study and the experimental study by realizing the first-ever RSMA prototype using SDRs. Using our prototype, the study experimentally compares the throughput and fairness performance of RSMA, SDMA and NOMA. Our contributions are as follows:

- implementing RSMA on SDRs using a two-stage protocol. In Stage 1, orthogonal pilot signals are transmitted by the TX for CSI estimation at each RX. The estimated CSI is then fed back to the TX, either unquantized or quantized to emulate perfect and imperfect CSIT, respectively (Section 3.2.1). The CSIT is used to design the precoded RSMA signal that is transmitted in Stage 2 (Section 3.2.2). This study uses OFDM signals largely based on IEEE 802.11g PHY layer frames to implement the data payload in the RSMA signal (Section 3.3). The above two-stage protocol can also be used to implement SDMA and NOMA in our prototype with appropriate changes to Stage 2.
- For the two-user MISO scenario, this study formulates the practically relevant *MCS-limited sum throughput maximization problem* for RSMA, and establish its relationship to the canonical sum rate maximization problem that is widely analyzed in theory (Section 3.2.2). The MCS-limited sum throughput maximization problem can also be formulated for SDMA and NOMA (Table 3.1).
- Through measurements using the prototype, the study empirically solves the MCS-limited sum throughput maximization problem for RSMA, SDMA and NOMA for the two-user MISO scenario over line-of-sight channels in a lab environment, where the pair of channels differ in terms of their relative strength and spatial correlation (Section 3.4). The results show that
 - With unquantized CSI feedback, NOMA achieves significant throughput gains over SDMA when channels have high spatial correlation. On the other hand, when the channels have low spatial correlation, the SDMA throughput is higher than NOMA. RSMA, however, achieves the highest throughput not only under either extreme but also over channels that lie in between in terms of their spatial correlation. These outcomes are consistent with theoretical predictions in [13], [58].
 - With quantized CSI feedback, NOMA achieves significant throughput gains over SDMA even with channels having low spatial correlation, provided there is a sufficiently large

disparity between the channel strengths. The NOMA throughput performance is also at par with that of RSMA under these conditions. Essentially, the imperfect CSI disproportionately diminishes the weaker user's throughput for SDMA, whereas both RSMA and NOMA provide higher throughput to the weaker user via the common stream (see Section 3.2.2 for definition). However, when the channel strength disparity is not sufficiently large, RSMA achieves throughput gains over NOMA as well. These outcomes are consistent with theoretical predictions in [7], [8], [31].

- With both quantized and unquantized CSI feedback, RSMA achieves better fairness at a higher sum throughput than both SDMA and NOMA. This too is consistent with theoretical predictions in [7], [8], [31].

The rest of this chapter is organized as follows. In Section 3.2, this section introduces the RSMA system model for the two-user MISO scenario and formulate the practically relevant MCS-limited sum throughput maximization problem. In Section 3.3, the section provides a detailed description of our SDR-based RSMA implementation. In Section 3.4, the model empirically solves the MCS-limited sum throughput maximization problem for RSMA, SDMA and NOMA through measurements using our prototype and compare the sum throughput and fairness performance of the three schemes. Finally, Section 3.5 concludes this chapter.

3.2 System Model

Consider a two-antenna TX communicating with two single-antenna users/RXs, using OFDM¹ signals over N_c subcarriers. The communications takes place over two stages.

- **Stage 1:** First, the TX transmits orthogonal pilot signals for channel estimation at the RXs. The estimated CSI is then fed back to the TX through a separate link.
- **Stage 2:** Using the CSI acquired in Stage 1, the TX designs the precoder and the transmit signal to maximize the sum throughput.

This section describes each stage, in turn, in Sections 3.2.1 and 3.2.2.

¹In principle, RSMA can be used with any waveform. Our choice of OFDM is motivated by its prevalence in modern communications systems. However, in the context of 6G, alternatives to OFDM, such as Orthogonal Time Frequency Space (OTFS) have received considerable attention. RSMA for such waveforms – especially experimental evaluations – is a relevant topic yet to be investigated.

3.2.1 Stage 1: Channel Model and CSI feedback

Let $\mathbf{h}_i[k] := [h_{i1}[k], h_{i2}[k]]^T \in \mathbb{C}^{2 \times 1}$ denote the true frequency flat, slowly varying channel between the TX and RX i ($i = 1, 2$) over the k -th subcarrier ($k = 0, \dots, N_c - 1$). Similarly, let $\hat{\mathbf{h}}_i[k]$ denote the estimate of $\mathbf{h}_i[k]$, obtained at RX i through pilot signals transmitted by the TX.

To reduce feedback overhead, let $\hat{\mathbf{h}}_i$ denote the *wideband CSI* evaluated at RX i by averaging $\hat{\mathbf{h}}_i[k]$ over the subcarriers, i.e.,

$$\hat{\mathbf{h}}_i := \frac{1}{N_c} \sum_{k=0}^{N_c-1} \hat{\mathbf{h}}_i[k], \quad (i = 1, 2). \quad (3.1)$$

For CSI feedback of $\hat{\mathbf{h}}_i$ to the TX in the experiments, it considers two cases to emulate perfect and imperfect CSI at the TX.

- *Unquantized feedback:* Let $\hat{\mathbf{h}}_i^{\text{UQ}} = [\hat{h}_{i1}^{\text{UQ}} \hat{h}_{i2}^{\text{UQ}}]$ denote the unquantized version of $\hat{\mathbf{h}}_i$, where each complex $\hat{h}_{il}^{\text{UQ}}$ ($i, l \in \{1, 2\}$) is a 128-bit representation of $\hat{h}_{il}$ (i.e., a 64-bit double-precision floating-point number for the real and imaginary parts). This is intended to emulate perfect CSI at the TX.
- *Quantized feedback:* Let $\hat{\mathbf{h}}_i^{\text{Q}} = [\hat{h}_{i1}^{\text{Q}} \hat{h}_{i2}^{\text{Q}}]$ denote the quantized version of $\hat{\mathbf{h}}_i$, where each complex $\hat{h}_{il}^{\text{Q}}$ ($i, l \in \{1, 2\}$) is an 8-bit representation of $\hat{h}_{il}$ (i.e., 4 bits for the real and imaginary parts). The quantization mechanism is described in Appendix A, and is based on IEEE 802.11n [118]. It results in a more than 90% reduction in the CSI feedback overhead, and is intended to emulate imperfect CSI at the TX.

The feedback CSI is used by the TX to design the precoders, as described in the following subsection.

3.2.2 Stage 2: RSMA Signal Design

For the two-user MISO scenario, Fig. 3.1 depicts RSMA in operation. Let W_1 and W_2 denote the messages corresponding to RX 1 and RX 2, respectively. At the TX, each W_i ($i = 1, 2$) is split into common and private portions – denoted by $W_{c,i}$ and $W_{p,i}$, respectively – by the message splitter. The common portions of each RX's message (i.e., $W_{c,1}$ and $W_{c,2}$) are combined into a common message, which is then encoded and modulated to form a common stream, $s_c[k]$ ($k = 0, \dots, N_c - 1$) over the subcarriers. On the other hand, the private portion of RX i 's message is separately encoded

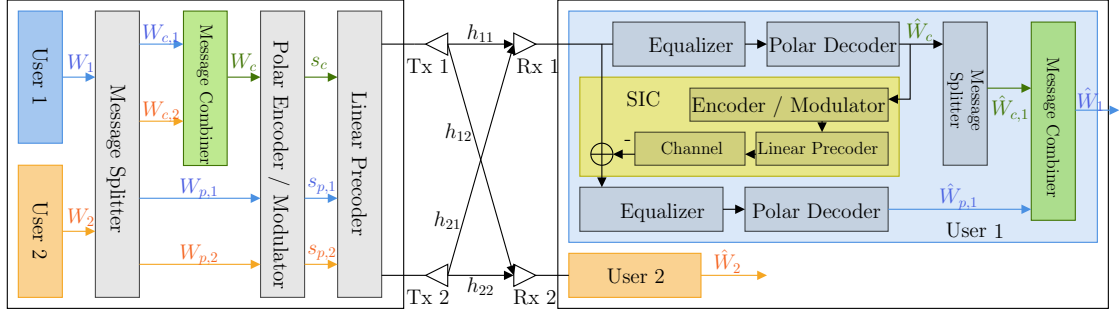


Figure 3.1: RSMA block diagram for the two-user MISO scenario.

and modulated to form a private stream, $s_i[k]$. All three streams (i.e., $s_c[k]$, $s_1[k]$ and $s_2[k]$) are linearly precoded to form the OFDM frequency domain transmit signal as follows:

$$\mathbf{x}[k] = \mathbf{p}_c s_c[k] + \mathbf{p}_1 s_1[k] + \mathbf{p}_2 s_2[k] \quad (k = 0, \dots, N_c - 1), \quad (3.2)$$

where $\mathbf{p}_c$ is referred to as the common stream precoder, and $\mathbf{p}_i$ the private stream precoder of RX i .

After appropriate Discrete Fourier Transform (DFT)-based processing, the received OFDM frequency domain signal, $y_i[k]$, at RX i is given by:

$$\begin{aligned} y_i[k] &= \mathbf{h}_i^H[k] \mathbf{x}[k] + n_i[k], \quad (k = 0, \dots, N_c - 1; \quad i = 1, 2) \\ &= \mathbf{h}_i^H[k] \mathbf{p}_c s_c[k] + \mathbf{h}_i^H[k] \mathbf{p}_1 s_1[k] + \mathbf{h}_i^H[k] \mathbf{p}_2 s_2[k] \\ &\quad + n_i[k], \end{aligned} \quad (3.3)$$

where $n_i[k] \sim \mathcal{CN}(0, \sigma^2)$ is the thermal noise. This study assumes that precoded pilot signals have been interspersed in some of the subcarriers of $\mathbf{x}[k]$ to help RX i estimate $\mathbf{h}_i^H[k] \mathbf{p}_c$ and $\mathbf{h}_i^H[k] \mathbf{p}_i$ upon receiving $y_i[k]$ ². Then, after suitable equalization (e.g., Minimized Min Square Error (MMSE)), RX i first decodes $s_c[k]$ to recover W_c , while treating the interference from the $s_1[k]$ and $s_2[k]$ as noise. The decoded estimate of W_c at RX i , denoted by $\hat{W}_c^{(i)}$, is simultaneously sent to:

- the message splitter to extract $\hat{W}_{c,i}$ – the decoded estimate of $W_{c,i}$, and
- the SIC module, which generates an estimate of $\mathbf{h}_i^H[k] \mathbf{p}_c s_c[k]$ and subtracts it from $y_i[k]$. The residue is once again subjected to equalization, this time to decode $s_i[k]$ by treating the interference from the other private stream as noise. Let $\hat{W}_{p,i}$ denote the decoded estimate of $W_{p,i}$.

²These pilots play a role similar to the demodulation reference signals (DM-RS) used in LTE and 5G NR.

Combining $\hat{W}_{c,i}$ and $\hat{W}_{p,i}$ (i.e., the inverse of the splitting operation at the TX), RX i obtains an estimate, $\hat{W}_i$, of its desired message, W_i .

3

Let $\mathbf{P} = [\mathbf{p}_c \ \mathbf{p}_1 \ \mathbf{p}_2]$. As a function of $\mathbf{P}$, from (3.3), the achievable rate (in bits/s/Hz) for error-free decoding of $s_c[k]$ at RX i , denoted by $R_{c,i}(\mathbf{P})$, is given by:

$$R_{c,i}(\mathbf{P}) = \min_k \log_2 \left(1 + \frac{|\mathbf{h}_i^H[k] \mathbf{p}_c|^2}{\sigma^2 + |\mathbf{h}_i^H[k] \mathbf{p}_1|^2 + |\mathbf{h}_i^H[k] \mathbf{p}_2|^2} \right), \quad (3.4)$$

where the minimum over k arises because the encoded and modulated common stream is embedded over subcarriers in the frequency domain. Additionally, since both RXs must first decode $s_c[k]$, it follows that the achievable rate for error-free decoding of $s_c[k]$ at *both* RXs, denoted by $R_c(\mathbf{P})$, is given by:

$$R_c(\mathbf{P}) = \min(R_{c,1}(\mathbf{P}), R_{c,2}(\mathbf{P})). \quad (3.5)$$

Assuming perfect SIC operation, the achievable rate for error-free decoding of $s_i[k]$ at RX i , similar to (3.4), is given by:

$$R_i(\mathbf{P}) = \min_k \log_2 \left(1 + \frac{|\mathbf{h}_i^H[k] \mathbf{p}_i|^2}{\sigma^2 + |\mathbf{h}_i^H[k] \mathbf{p}_j|^2} \right), \quad j \neq i. \quad (3.6)$$

Thus, the achievable rate at RX i for error-free recovery of W_i , denoted by $R_i^{\text{RSMA}}(\mathbf{P})$, is given by:

$$R_i^{\text{RSMA}}(\mathbf{P}) = (|W_{c,i}|/|W_c|)R_c(\mathbf{P}) + R_i(\mathbf{P}) \quad (3.7)$$

where the first term in the RHS captures the fraction of the common message, W_c , intended for RX i ($|W_{c,i}|$ and $|W_c|$ denote the size of $W_{c,i}$ and W_c , respectively). Consequently, the achievable RSMA sum rate, denoted by $R^{\text{RSMA}}(\mathbf{P})$, is:

$$\begin{aligned} R^{\text{RSMA}}(\mathbf{P}) &= R_1^{\text{RSMA}}(\mathbf{P}) + R_2^{\text{RSMA}}(\mathbf{P}) \\ &= R_c(\mathbf{P}) + R_1(\mathbf{P}) + R_2(\mathbf{P}) \end{aligned} \quad (3.8)$$

The canonical *sum rate maximization problem*, where the precoders at the TX are optimized to

maximize $R^{\text{RSMA}}(\mathbf{P})$ is formulated as follows:

$$OP_{\text{sr}}^{\text{RSMA}} : \max_{\mathbf{P}} R^{\text{RSMA}}(\mathbf{P}) \quad (3.9)$$

$$\text{s.t. } \text{tr}(\mathbf{P}\mathbf{P}^H) \leq P_t, \quad (3.10)$$

where (3.10) restricts the transmit power to P_t . Through the well-known Rate-Weighted Minimized Min Square Error (WMMSE) relationship [119], the non-convex problem in $OP_{\text{sr}}^{\text{RSMA}}$ can be transformed into a sequence of tractable sub-problems that can be iteratively solved till convergence to a local optimum. This is commonly known in the literature as the *WMMSE method* and despite only guaranteeing local optimality, it is widely recognized as a benchmark solution for the family of MIMO precoder design problems, of which $OP_{\text{sr}}^{\text{RSMA}}$ is a specific instance corresponding to RSMA. Let $\mathbf{P}_{\text{wmmse}}^{\text{RSMA}}$ denote the precoders obtained by applying the WMMSE method to solve $OP_{\text{sr}}^{\text{RSMA}}$.

Remark 1 (Imperfect CSI at the TX) $OP_{\text{sr}}^{\text{RSMA}}$ contains a slight abuse of notation for the sake of simplicity. Let $\hat{\mathbf{H}}^{\text{TX}} := [\hat{\mathbf{h}}_1^{\text{TX}} \ \hat{\mathbf{h}}_2^{\text{TX}}]$ denote the CSIT, where $\hat{\mathbf{h}}_i^{\text{TX}} \in \{\hat{\mathbf{h}}_i^{\text{UQ}}, \hat{\mathbf{h}}_i^{\text{Q}}\}$, as described in Section 3.2.1. Strictly speaking, to obtain the objective function of $OP_{\text{sr}}^{\text{RSMA}}$ at the TX, $\mathbf{h}_i[k]$ in (3.4) and (3.6) should be replaced by $\hat{\mathbf{h}}_i^{\text{TX}}$, so that the solution to $OP_{\text{sr}}^{\text{RSMA}}$ becomes a function of the CSIT.

In this chapter, the main performance metric in our measurements is the throughput (measured in bits/s). Over an effective bandwidth³, B , the achievable RSMA sum throughput, denoted by $T^{\text{RSMA}}(\mathbf{P})$, is related to $R^{\text{RSMA}}(\mathbf{P})$ in (3.8) via $T^{\text{RSMA}}(\mathbf{P}) = BR^{\text{RSMA}}(\mathbf{P})$. Hence, it follows that $T^{\text{RSMA}}(\mathbf{P}_{\text{wmmse}}^{\text{RSMA}})$ is a good benchmark in terms of the maximum achievable RSMA sum throughput. However, this benchmark is hard to realize in practice because:

1. In state-of-the-art communications systems, PHY-layer transmissions are restricted to a finite collection of modulation and coding scheme (MCS) levels (e.g., QPSK, rate 1/2), which caps the maximum achievable sum throughput, and
2. Error-free decoding of the transmitted messages at the RX cannot be guaranteed due to finite block length effects, which further diminishes the realized throughput.

This section addresses these two points by first characterizing the *MCS-limited* throughput. An MCS level can be defined by a pair (m, r) , where positive integer m denotes the bits per constel-

³The effective bandwidth is the portion of the total bandwidth available for data transfer, after accounting for signaling overheads (e.g., cyclic prefix in the case of OFDM).

lation symbol (e.g., 2 for QPSK) and $r \in (0, 1]$ denotes the code rate. Hence, the achievable rate (in bits/s/Hz) for MCS level (m, r) equals mr , and the corresponding throughput is Bmr . Let $\mathcal{M} = \{(m_c, r_c), (m_1, r_1), (m_2, r_2)\}$ denote the MCS levels chosen for $s_c[k]$, $s_1[k]$ and $s_2[k]$. Then, the *MCS-limited* RSMA sum throughput, denoted by $T_{\text{mcs}}^{\text{RSMA}}(\mathbf{P}, \mathcal{M})$, is given by:

$$\begin{aligned} T_{\text{mcs}}^{\text{RSMA}}(\mathbf{P}, \mathcal{M}) &= Bm_c r_c(\mathbf{p}_c) \times \mathbb{P}(\hat{W}_c^{(1)} = W_c; \hat{W}_c^{(2)} = W_c) \\ &\quad + Bm_1 r_1(\mathbf{p}_1) \times \mathbb{P}(\hat{W}_{p,1} = W_{p,1}) \\ &\quad + Bm_2 r_2(\mathbf{p}_2) \times \mathbb{P}(\hat{W}_{p,2} = W_{p,2}). \end{aligned} \quad (3.11)$$

The RHS of (3.11) indicates that the MCS levels of the streams are, in general, a function of the corresponding precoders. The probability terms capture the loss in throughput due to decoding errors – specifically, $\mathbb{P}(\hat{W}_c^{(1)} = W_c; \hat{W}_c^{(2)} = W_c)$ denotes the probability that W_c is correctly decoded at both RXs, while $\mathbb{P}(\hat{W}_{p,i} = W_{p,i})$ denotes the probability that $W_{p,i}$ is correctly decoded at RX i . Thus, similar to $OP_{\text{sr}}^{\text{RSMA}}$ in (3.9)-(3.10), the *MCS-limited RSMA sum throughput maximization problem*, which is of considerable practical relevance, can be defined as follows:

$$OP_{\text{mcs}}^{\text{RSMA}} : \max_{\mathbf{P}, \mathcal{M}} T_{\text{mcs}}^{\text{RSMA}}(\mathbf{P}_{\text{wmmse}}^{\text{RSMA}}, \mathcal{M}) \quad (3.12)$$

$$\mathcal{M} \in \mathbb{M}, \quad (3.13)$$

where $\mathbb{M}$ denotes the collection of permissible MCS levels for the three streams, which is typically pre-determined through standards (see Table 3.3 for the $\mathbb{M}$ used in our measurements). $OP_{\text{mcs}}^{\text{RSMA}}$ does not have a closed form solution, as the message decoding probabilities in (3.11) are difficult to characterize in terms of $\mathbf{P}_{\text{wmmse}}^{\text{RSMA}}$ and $\mathcal{M}$. However, these probabilities can be empirically evaluated through measurements. This motivates us to empirically solve $OP_{\text{mcs}}^{\text{RSMA}}$ through a brute force search⁴ over $\mathbb{M}$, which is the focus of our measurements in Section 3.4 using our SDR-based RSMA prototype. We conclude this subsection with the following remark:

Remark 2 (Wideband precoders) *In principle, narrowband precoders for each subcarrier (i.e., $\mathbf{p}_c[k]$, $\mathbf{p}_1[k]$, $\mathbf{p}_2[k]$: $k = 0, \dots, N_c - 1$) could increase the achievable rate in (3.4)-(3.8) by adapting to channel variations at subcarrier-level granularity. However, this would involve huge signalling overhead in both directions, as: (a) narrowband CSI (i.e., $\hat{\mathbf{h}}_i[k]$ for several, if not all, k) would need to be fed back to the TX for precoder design, and (b) more demodulation reference signals would have to be embedded in the transmit signal to estimate $\mathbf{h}_i^H[k]\mathbf{p}_c[k]$ and $\mathbf{h}_i^H[k]\mathbf{p}_i[k]$ at RX i*

⁴A more sophisticated empirical approach involves link adaptation, where the most suitable MCS level is determined by an ARQ-based mechanism [112]. This is left for future work.

for all k . Thus, to reduce the signalling overhead, we assume the same precoders, $\mathbf{P}$, across all the subcarriers.

3.2.3 RSMA v/s SDMA v/s NOMA

In general, the splitting of the user messages in Fig. 3.1 need not be equal (i.e., 50-50) between the common and private portions, nor does the ratio between the common and private portions need to be fixed across both messages. In fact, SDMA and NOMA are special cases of RSMA corresponding to specific choices of message splitting, as explained in the following remarks.

Remark 3 (SDMA as a special case of RSMA) *For the special case where the common stream is turned off – i.e., $W_i = W_{p,i}$ ($i = 1, 2$) in Fig. 3.1 – RSMA reduces to SDMA [13].*

Remark 4 (NOMA as a special case of RSMA) *Without loss of generality, suppose RX 2 has a weaker channel than RX 1. For the special case where the private stream of RX 2 is turned off, and no portion of W_1 is allocated to the common stream – i.e., $W_1 = W_{p,1}, W_2 = W_{c,2} = W_c$ in Fig. 3.1 – RSMA reduces to NOMA [13].*

This flexibility in message splitting provided by RSMA characterizes its interference management philosophy, as follows:

- **SDMA:** In the absence of a common stream, the interference experienced at RX i due to W_j ($j \neq i$) is treated as noise.
- **NOMA:** The common stream allows the stronger RX 1 to fully decode and subtract the interference it experiences (from W_2). The weaker RX 2 continues to treat the interference it experiences (from W_1) as noise, like SDMA.
- **RSMA:** The interference at RX i due to W_j ($j \neq i$) is *partially decoded* ($W_{c,j}$) and *partially treated as noise* ($W_{p,j}$). This feature enables RSMA to softly bridge between SDMA and NOMA depending on the channel conditions [13], [58]. Indeed, when the two channels are quasi-orthogonal, the solution to $OP_{\text{sr}}^{\text{RSMA}}$ in (3.9)-(3.10) is SDMA-like with very little power allocated to the common stream precoder. Similarly, when the channels are highly spatially correlated with one of them much stronger than the other, the solution to $OP_{\text{sr}}^{\text{RSMA}}$ is NOMA-like with considerable power allocated to the common stream precoder. Importantly, over the vast majority of channels that do not fall in either extreme, RSMA is expected to provide

significant gains over both SDMA and NOMA through judicious resource allocation between the common and private streams. We aim to demonstrate this experimentally in Section 3.4.

Based on Remarks 3 and 4, the corresponding expressions for SDMA and NOMA in (3.2)-(3.13) can be obtained by discarding $\mathbf{p}_c$ (SDMA) and $\mathbf{p}_2$ (NOMA). These expressions are provided in Table 3.1 for clarity. In Sections 3.4.2 and 3.4.3, we provide an experimental RSMA v/s SDMA v/s NOMA performance comparison by empirically solving OP_{mcs}^{ma} , $ma \in \{\text{RSMA, SDMA, NOMA}\}$ using our SDR-based RSMA prototype. In the following section, we present details of our prototype, including how the system model described above in Sections 3.2.1 and 3.2.2 is implemented.

3.3 RSMA Prototype using SDR

3.3.1 Hardware setup

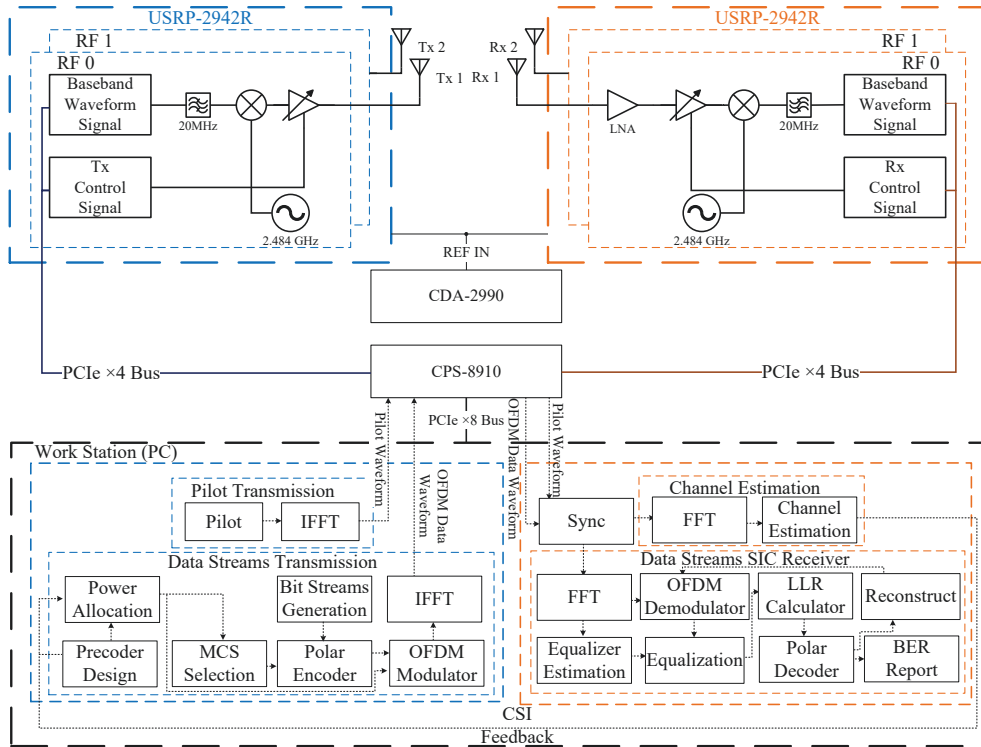


Figure 3.2: Block diagram of RSMA Prototype.

The RSMA prototype is built using National Instruments' (NI) Universal Software Radio Peripherals (USRPs). As shown in Fig. 3.2, two USRP-2942 units are used to realize the two-user MISO scenario. In particular, the RX antennas (TP-LINK TL-ANT2405C) are connected to its

Quantity	Expression
Message splitting	$W_i = W_{c,i} \cup W_{p,i} \ (i = 1, 2)$ $W_c = W_{c,1} \cup W_{c,2}$ RSMA: $W_{c,i}, W_{p,i} \neq \emptyset$, in general SDMA: $W_{c,i} = \emptyset, W_{p,i} = W_i$ NOMA: $W_{c,1} = \emptyset, W_{p,1} = W_1$ and $W_{c,2} = W_2, W_{p,2} = \emptyset$
Precoders	RSMA: $\mathbf{P} = [\mathbf{p}_c \ \mathbf{p}_1 \ \mathbf{p}_2]$ SDMA: $\mathbf{P} = [\mathbf{p}_1 \ \mathbf{p}_2]$ NOMA: $\mathbf{P} = [\mathbf{p}_c \ \mathbf{p}_1]$
Achievable common stream rate at RX i	RSMA: $R_{c,i}(\mathbf{P}) = \min_k \log_2 \left(1 + \frac{ \mathbf{h}_i^H[k]\mathbf{p}_c ^2}{\sigma^2 + \mathbf{h}_i^H[k]\mathbf{p}_1 ^2 + \mathbf{h}_i^H[k]\mathbf{p}_2 ^2} \right)$ $R_c(\mathbf{P}) = \min_i R_{c,i}(\mathbf{P})$ SDMA: $R_{c,i}(\mathbf{P}) = R_c(\mathbf{P}) = 0$ NOMA: $R_{c,i}(\mathbf{P}) = \min_k \log_2 \left(1 + \frac{ \mathbf{h}_i^H[k]\mathbf{p}_c ^2}{\sigma^2 + \mathbf{h}_i^H[k]\mathbf{p}_1 ^2} \right)$ $R_c(\mathbf{P}) = \min_i R_{c,i}(\mathbf{P})$
Achievable pvt. stream rate at RX i	RSMA: $R_i(\mathbf{P}) = \min_k \log_2 \left(1 + \frac{ \mathbf{h}_i^H[k]\mathbf{p}_i ^2}{\sigma^2 + \mathbf{h}_i^H[k]\mathbf{p}_j ^2} \right) \ (j \neq i)$ SDMA: $R_i(\mathbf{P}) = \min_k \log_2 \left(1 + \frac{ \mathbf{h}_i^H[k]\mathbf{p}_i ^2}{\sigma^2 + \mathbf{h}_i^H[k]\mathbf{p}_j ^2} \right) \ (j \neq i)$ NOMA: $R_1(\mathbf{P}) = \min_k \log \left(1 + \frac{ \mathbf{h}_1^H[k]\mathbf{p}_1 ^2}{\sigma^2} \right)$ $R_2(\mathbf{P}) = 0$
Achievable sum rate	RSMA: $R^{\text{RSMA}}(\mathbf{P}) = R_c(\mathbf{P}) + R_1(\mathbf{P}) + R_2(\mathbf{P})$ SDMA: $R^{\text{SDMA}}(\mathbf{P}) = R_1(\mathbf{P}) + R_2(\mathbf{P})$ NOMA: $R^{\text{NOMA}}(\mathbf{P}) = R_c(\mathbf{P}) + R_1(\mathbf{P})$
Sum rate max. problem	$OP_{\text{sr}}^{\text{ma}} : \max_{\mathbf{P}} R^{\text{ma}}(\mathbf{P}) \ (\text{ma} \in \{\text{RSMA}, \text{SDMA}, \text{NOMA}\})$ $\text{s.t. } \text{tr}(\mathbf{P}\mathbf{P}^H) \leq P_t$
Achievable sum throughput	$T^{\text{ma}}(\mathbf{P}) = BR^{\text{ma}}(\mathbf{P}),$
MCS levels	RSMA: $\mathcal{M} = \{(m_c, r_c), (m_1, r_1), (m_2, r_2)\}$ SDMA: $\mathcal{M} = \{(m_1, r_1), (m_2, r_2)\}$ NOMA: $\mathcal{M} = \{(m_c, r_c), (m_1, r_1)\}$
MCS-limited sum rate	RSMA: $R_{\text{mcs}}^{\text{RSMA}}(\mathbf{P}, \mathcal{M}) = m_c r_c(\mathbf{p}_c) + m_1 r_1(\mathbf{p}_1) + m_2 r_2(\mathbf{p}_2)$ SDMA: $R_{\text{mcs}}^{\text{SDMA}}(\mathbf{P}, \mathcal{M}) = m_1 r_1(\mathbf{p}_1) + m_2 r_2(\mathbf{p}_2)$ NOMA: $R_{\text{mcs}}^{\text{NOMA}}(\mathbf{P}, \mathcal{M}) = m_c r_c(\mathbf{p}_c) + m_1 r_1(\mathbf{p}_1)$
MCS-limited sum throughput	RSMA: $T_{\text{mcs}}^{\text{RSMA}}(\mathbf{P}, \mathcal{M}) = Bm_c r_c(\mathbf{p}_c) \times \mathbb{P}(\hat{W}_c^{(1)} = W_c; \hat{W}_c^{(2)} = W_c)$ $+ Bm_1 r_1(\mathbf{p}_1) \times \mathbb{P}(\hat{W}_{p,1} = W_{p,1})$ $+ Bm_2 r_2(\mathbf{p}_2) \times \mathbb{P}(\hat{W}_{p,2} = W_{p,2})$ SDMA: $T_{\text{mcs}}^{\text{SDMA}}(\mathbf{P}, \mathcal{M}) = Bm_1 r_1(\mathbf{p}_1) \times \mathbb{P}(\hat{W}_{p,1} = W_1)$ $+ Bm_2 r_2(\mathbf{p}_2) \times \mathbb{P}(\hat{W}_{p,2} = W_2)$ NOMA: $T_{\text{mcs}}^{\text{NOMA}}(\mathbf{P}, \mathcal{M}) = Bm_c r_c(\mathbf{p}_c) \times \mathbb{P}(\hat{W}_c^{(1)} = W_2; \hat{W}_c^{(2)} = W_2)$ $+ Bm_1 r_1(\mathbf{p}_1) \times \mathbb{P}(\hat{W}_{p,1} = W_1)$
MCS-limited sum throughput max. problem (Experiments in this chapter)	$OP_{\text{mcs}}^{\text{ma}} : \max_{\mathcal{M}} T_{\text{mcs}}^{\text{ma}}(\mathbf{P}_{\text{wmmse}}^{\text{ma}}, \mathcal{M})$ $\text{s.t. } \mathcal{M} \in \mathbb{M}$

Table 3.1: System model differences between RSMA, SDMA and NOMA w.r.t the terminology in Section 3.2.2.

No.	Name	Units Description
1.	Workstation	Running LabVIEW NXG
2.	NI USRP-2942	SDRs used to realize the TX and the RXs
3.	NI CPS-8910	Provides additional PCIe ports
4.	NI CDA-2990	8 Channel, 10 MHz Clock Distribution Device
5.	Linx Technologies ANT-W63WSx	TX antennas
6.	TP-LINK TL-ANT2405C	RX antennas

Table 3.2: List of hardware components.

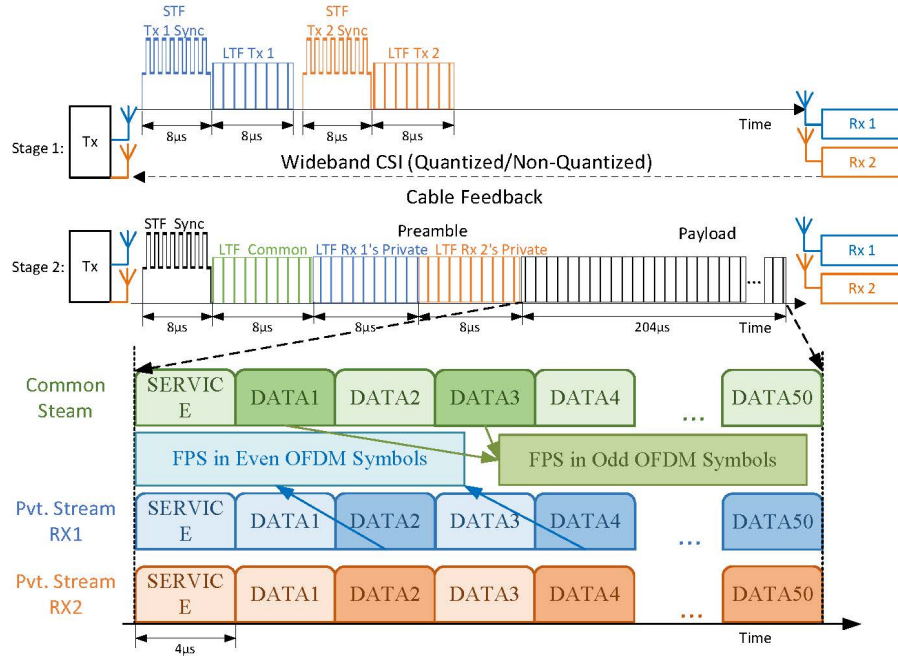


Figure 3.3: (Top and middle): The two-stage transmission protocol; (Bottom): OFDM symbol structure within the payload. In conventional 802.11g, every DATA symbol for an RX contains four pilot subcarriers for FPS. However, for RSMA, this approach would cause the pilot subcarriers in the common and private streams to interfere with each other. In particular, the inability to do FPS effectively for the common stream leads to large phase noise. To avoid this, the common (private) stream(s) carry non-zero pilots only in the odd (even) DATA symbols [120].

USRP using coaxial cables, which allows them to be moved around to create different channel configurations (Fig. 3.4). The USRPs share a common timing source (CDA-2990), and are connected to a workstation through which they are controlled using LabVIEW NXG. These connections are through PCIe cables, facilitated by a PCIe bus (CPS-8910). The total transmit power, P_t , is 23dBm. A list of hardware components is provided in Table 3.2.

3.3.2 RSMA Implementation Details

The USRPs are capable of operating in the 2.4GHz Industrial Scientific and Medical (ISM) band. Hence, we choose a center frequency of 2.484GHz for our transmissions⁵. For ease of practicalities like synchronization and channel estimation, it is convenient to implement signals based on well-known PHY layer standards (e.g., IEEE 802.11 frames) in a prototype. Hence, this prototype adopts several features of the IEEE 802.11g PHY layer frames in our signal design. Following contents describe in detail how Stages 1 and 2 from Sections 3.2.1 and 3.2.2, respectively, are implemented in the prototype:

Stage 1 Each TX antenna transmits a pilot signal orthogonally in time comprising a Short Training Field (STF, $8\mu s$ in duration) and a Long Training Field (LTF, $8\mu s$ in duration), as shown in the top portion of Fig. 3.3. The Short Training Field (STF) is used for synchronization and coarse frequency offset estimation, while the Long Training Field (LTF) is used to obtain a least squares estimate of the CSI at each RX over all the subcarriers (i.e., $\hat{\mathbf{h}}_i[k]$ in Section 3.2.1).

Stage 2 Here, the transmitted signal consists of a preamble followed by the data payload, as shown in the middle portion of Fig. 3.3.

- **Preamble:** The preamble consists of one STF and three LTFs. The function of the STF is the same as in the first stage, while the LTFs are precoded in order to estimate the *precoded CSI* for equalization at the RX; in particular, the first LTF is used to estimate $\mathbf{h}_i^H \mathbf{p}_c$ at RX i ($= 1, 2$) for decoding the common stream, the second to estimate $\mathbf{h}_1^H \mathbf{p}_1$ to decode the private stream meant for RX 1, and the third to estimate $\mathbf{h}_2^H \mathbf{p}_2$ to decode the private stream meant for RX 2.
- **Data Payload:** For the payload, we use OFDM signals over a total bandwidth of 20MHz with $N_c = 64$ subcarriers (labeled DATA in Fig. 3.3) and a cyclic prefix (CP) of 16 samples per OFDM symbol. Aligned with IEEE 802.11 frames, 52 subcarriers are used for communications while the rest serve as guard bands. Among these 52 subcarriers, 48 are used to carry data symbols, with the remaining used for fine phase shifting (FPS)⁶. This yields an

⁵This corresponds to channel no. 14 in the IEEE 802.11 family of standards for the 2.4GHz band. This prototype uses this channel to avoid ambient WiFi interference, as it is not commercially used in the UK.

⁶FPS is used to correct the common phase error across all subcarriers in one OFDM symbol [121].

MCS Index	Modulation (m)	Code Rate, r	Data Rate, Bmr (Mbps)
0	BPSK (1)	1/2	6
1	BPSK (1)	3/4	9
2	QPSK (2)	1/2	12
3	QPSK (2)	3/4	18
4	16QAM (4)	1/2	24
5	16QAM (4)	3/4	36
6	64QAM (6)	2/3	48
7	64QAM (6)	3/4	54
8	256QAM (8)	3/4	72
9	256QAM (8)	5/6	80

Table 3.3: MCS levels (largely based on IEEE 802.11g) implemented in our prototype. The data rate in the last column is equal to Bmr , where B is the effective bandwidth given by (3.14).

effective bandwidth of:

$$B = 20\text{MHz} \times \underbrace{(64/80)}_{\text{CP overhead}} \times \underbrace{(48/64)}_{\text{Guard band overhead}} = 12\text{MHz} \quad (3.14)$$

The payload consists of three superposed streams (one common, two private), each comprising 50 OFDM symbols (labeled DATA 1 through DATA 50, as shown in the bottom portion of Fig. 3.3).

- **MCS Implementation:** Table 3.3 lists the MCS levels, $\mathbb{M}$, implemented in our prototype. MCS indices 0 through 7 are identical to IEEE 802.11g, where 64-QAM is the largest supported constellation. To allow for the possibility of higher data rates due to RSMA's enhanced interference suppression, this prototype implements two more MCS levels supporting 256-QAM (MCS indices 8 and 9). After the preamble, the first OFDM symbol (labelled SERVICE in the bottom portion of Fig. 3.3) contains the MCS information of each stream. Unlike the IEEE 802.11 family of standards, which uses Low-Density Parity Check (LDPC) codes for error correction, this prototype implements Polar codes augmented with an 8-bit cyclic redundancy check [122], [123], with successive cancellation list decoding [124] using a list depth of 2.

An instance of Stage 1 and Stage 2, as described above and illustrated in Fig. 3.3, constitutes a single measurement run. To empirically solve $OP_{\text{mcs}}^{\text{ma}}$ ($\text{ma} \in \{\text{RSMA}, \text{SDMA}, \text{NOMA}\}$) from Table 3.1, we conduct 100 measurement runs. Let $D_{s,c}^{\text{ma}}$ denote the number of runs in which the

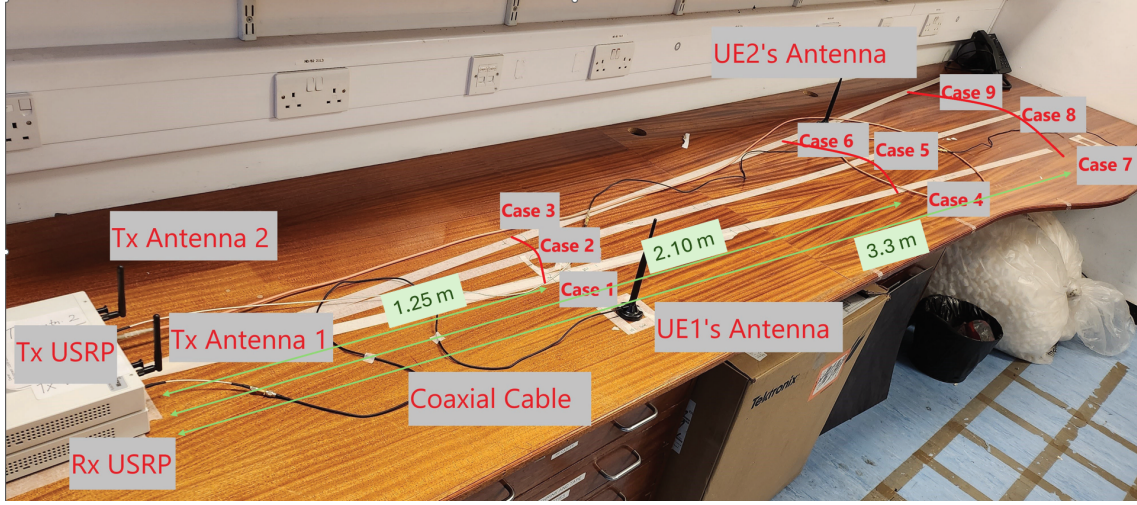


Figure 3.4: Layout of measurement cases. The RX antenna positions in the above snapshot correspond to Case 6.

common stream is successfully decoded by both RXs. Similarly, let $D_{s,i}^{\text{ma}}$ ($i = 1, 2$) denote the number of runs in which RX i successfully decodes its private stream. Consider the expressions for the MCS-limited sum throughput in Table 3.1 (penultimate row). Replacing the decoding probabilities therein with their empirical estimates, the *measured* sum throughput for each multiple access scheme is given by:

$$T_{\text{mcs}}^{\text{RSMA}}(\mathbf{P}_{\text{wmmse}}^{\text{RSMA}}, \mathcal{M}) = \frac{D_{s,c}^{\text{RSMA}}}{100} B m_c r_c + \sum_{i=1}^2 \frac{D_{s,i}^{\text{RSMA}}}{100} B m_i r_i \quad (3.15)$$

$$T_{\text{mcs}}^{\text{SDMA}}(\mathbf{P}_{\text{wmmse}}^{\text{SDMA}}, \mathcal{M}) = \frac{D_{s,1}^{\text{SDMA}}}{100} B m_1 r_1 + \frac{D_{s,2}^{\text{SDMA}}}{100} B m_2 r_2 \quad (3.16)$$

$$T_{\text{mcs}}^{\text{NOMA}}(\mathbf{P}_{\text{wmmse}}^{\text{NOMA}}, \mathcal{M}) = \frac{D_{s,c}^{\text{NOMA}}}{100} B m_c r_c + \frac{D_{s,1}^{\text{NOMA}}}{100} B m_1 r_1, \quad (3.17)$$

where B is given by (3.14) and the MCS levels are chosen from Table 3.3.

3.4 Measurement Results

This section first describes our measurement campaign in Section 3.4.1. Then, Section 3.4.2 presents measurement results pertaining to $OP_{\text{mcs}}^{\text{ma}}$ through which compares the measured sum throughput for RSMA, SDMA and NOMA. Then, in Section 3.4.3 presents a fairness comparison between RSMA, SDMA and NOMA.

Parameter	Notation (where applicable)	Value
Center frequency	f_c	2.484GHz
Transmit power	P_t	23dBm
TX antenna length		0.13m
Fraunhofer distance		0.28m
Total bandwidth		20MHz
Subcarriers	Total (N_c)	64
	Data	48
	Pilot (FPS)	4
	Guard band	12
CP length		16
Effective bandwidth	B	12MHz
OFDM symbols in payload		50
Experiment runs (per case)		100
Distances	RX 1 $\rightarrow$ TX	0.8m
	RX 2 $\rightarrow$ TX (Cases 1-3)	1.25m
	RX 2 $\rightarrow$ TX (Cases 4-6)	2.10m
	RX 2 $\rightarrow$ TX (cases 7-9)	3.30m
Channel coding		Polar codes
List depth		2
Environment		LoS

Table 3.4: List of parameters used in our experiments.

3.4.1 Measurement campaign

The measurements were carried out on a lab bench, as shown in Fig. 3.4. The TX position is fixed, along with the antenna of RX 1 which is placed 0.8m away from the TX⁷. The antenna position of RX 2 is varied as marked in Fig. 3.4 to create pairs of LoS channels with varying relative strengths and spatial correlation. For NOMA in particular, RX 1 is treated as the strong user and RX 2 the weak user. Every attempt was made to avoid any relative motion between the TXs and RXs during the measurements. For easy reference, a full list of all the parameters relevant to our measurements is provided in Table 3.4.

To capture the relative pathloss difference between the two channels, channel strength disparity

⁷As seen in Table 3.4, this is the closest TX-RX distance in our experiments. Since it exceeds the Fraunhofer distance, we are operating in the far-field.

	Relative pathloss		Spatial correlation	
	Cases	α [dB]	Cases	ρ
Low	1	-7.6	1	0.15
	2	-6.4	4	0.16
	3	-8.8	7	0.24
Medium	4	-14.5	2	0.48
	5	-13.0	5	0.46
	6	-13.0	8	0.35
High	7	-23.6	3	0.95
	8	-24.7	6	0.95
	9	-22.1	9	0.85

Table 3.5: The cases marked in Fig. 3.4 yield pairs of channels that vary in terms of their relative pathloss and spatial correlation, captured by α and ρ defined in (3.18) and (3.19), respectively. The above table lists the empirical average of α and ρ over all measurement runs for each case.

parameter, α , is introduced and defined in dB scale as follows:

$$\alpha \text{ [dB]} = 10 \log_{10} \frac{\|\hat{\mathbf{h}}_2\|}{\|\hat{\mathbf{h}}_1\|}. \quad (3.18)$$

where $\hat{\mathbf{h}}_i$ ($i = 1, 2$) is given by (3.1). Hence, RX 2 has a weaker channel when α is negative.

Similarly, to measure the spatial correlation between the two channels, we introduce the correlation parameter, $\rho \in [0, 1]$, defined as follows:

$$\rho = 1 - \frac{|\hat{\mathbf{h}}_1^H \hat{\mathbf{h}}_2|}{\|\hat{\mathbf{h}}_1\| \cdot \|\hat{\mathbf{h}}_2\|}. \quad (3.19)$$

Thus, $\rho = 0$ when the channels are completely aligned, whereas $\rho = 1$ when they are orthogonal.

Based on (3.18) and (3.19), we make the following remarks on the cases marked in Fig. 3.4.

Remark 5 For cases 1 through 3 in Fig. 3.4, we expect the relative pathloss difference between the two channels to be low by design, as the two RX antennas are fairly close together. In contrast, for cases 4 through 9, RX 2 is farther away from the TX than RX 1; hence, we expect α to be more negative by design. This is validated in Table 3.5, which lists the empirical average of α over all measurement runs for each case.

Remark 6 (Channels with high spatial correlation) In cases 1, 4 and 7, the channels are expected to be highly spatially correlated (i.e., small ρ) by design, as the two RXs are nearly in a

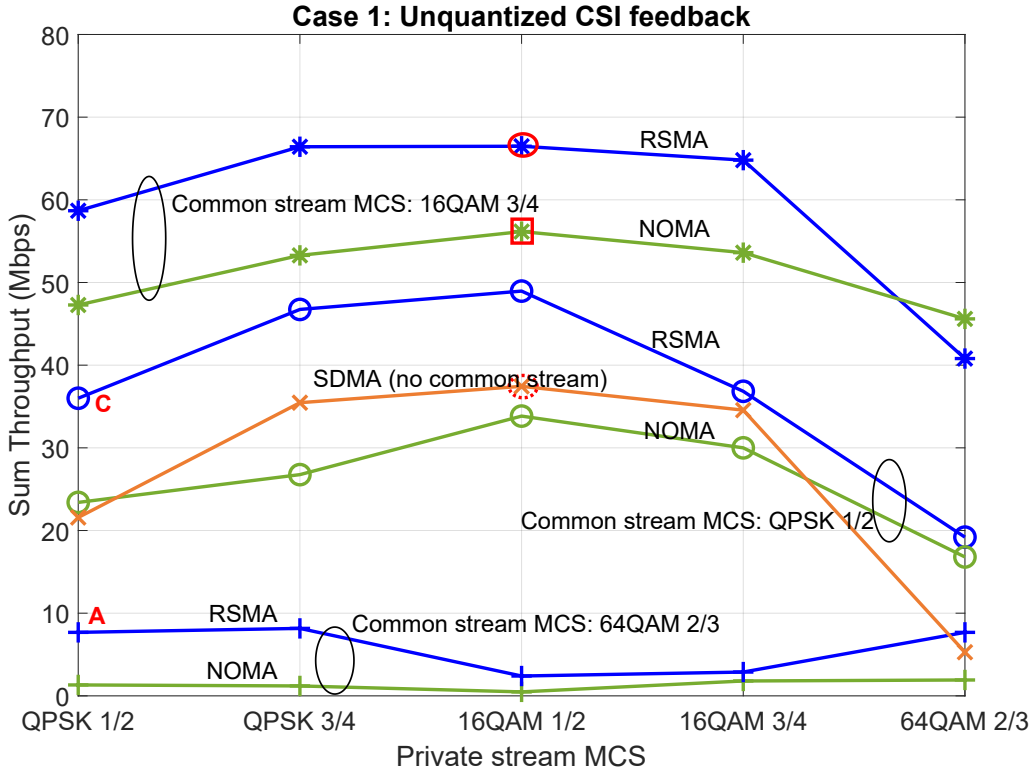


Figure 3.5: For Case 1, the large spatial correlation (i.e., small ρ , see Table 3.5) between the RX channels induces high mutual interference. Under these conditions, RSMA and NOMA can achieve significant throughput gains over SDMA with the help of the common stream, provided its MCS level is not too aggressive. The highest measured sum throughput above are: 66.48Mbps for RSMA, 56.16Mbps for NOMA, and 37.44Mbps for SDMA.

straight line to the TX. This is validated in Table 3.5, which lists the empirical average of ρ over all measurement runs for each case. RSMA and NOMA are expected to provide large throughput gains over SDMA in these cases, as the common stream allows for the decoding and elimination of interference.

Remark 7 (Channels with low spatial correlation) Similarly, in cases 3, 6 and 9, the spatial correlation between the two channels is expected to be relatively low (i.e., larger ρ) by design, due to the larger angular separation between the RXs as seen by the TX. This is validated in Table 3.5. The gain of RSMA over SDMA is expected to be modest, with the common stream being allocated lower power relative to the cases in Remark 6. NOMA is also unlikely to yield significant gains over SDMA; in fact, by forcibly imposing a common stream when the channels are not sufficiently aligned, the NOMA throughput performance can even be worse than SDMA.

3.4.2 RSMA v/s SDMA v/s NOMA: Sum Throughput Comparison

As an example measurement, the sum throughput performance for Case 1 as a function of the MCS levels is plotted in Fig. 3.5 with unquantized CSI feedback. For RSMA/SDMA, the MCS levels for the two private streams are chosen to be identical for ease of illustration⁸. For NOMA, the x -axis corresponds to the MCS of RX 1, as only RX 1 has a private stream. We make the following observations in Fig. 3.5:

- 1) For a given common stream MCS level for RSMA/NOMA, the sum throughput increases with the private stream MCS level up to a point as expected, before dropping when the MCS level is too aggressive. The same behavior is observed for SDMA as well, which does not have a common stream.
- 2) For a given private stream MCS level, a similar trend as above can also be seen w.r.t common stream MCS level for RSMA/NOMA. Specifically, the RSMA/NOMA sum throughput increases as the common stream MCS level is raised from QPSK rate 1/2 (MCS index 2 in Table 3.3) to 16QAM rate 3/4 (MCS index 5 in Table 3.3). However, the very next MCS level of 64QAM rate 3/4 for the common stream is too aggressive and due to the resulting error propagation, the RSMA/NOMA sum throughput is lower than SDMA. To demonstrate this, Fig. 3.6 shows the superposed received symbol constellation for RSMA at RX 1, corresponding to the MCS levels marked by points ‘C’ and ‘A’ in Fig. 3.5. The MCS levels at point C being conservative, a QPSK constellation (private stream) superposed on another QPSK constellation (common stream) is clearly visible in Fig. 3.6a. In a similar vein, if the MCS levels at point A were *not* aggressive, then we would expect to see 256 distinct symbol clusters in Fig. 3.6b, corresponding to a superposition of 64QAM (common stream) and QPSK (private stream) constellations. However, we only observe 81 distinct symbol clusters, which leads to high error rates for the common stream.
- 3) Points 1 and 2 above highlight the role of MCS selection in realizing sum throughput gains for RSMA and NOMA over SDMA. The highest measured sum throughput in Fig. 3.5 is the empirical solution of OP_{mcs}^{ma} in Table 3.1, with the additional constraint that the MCS levels for the two private streams in RSMA/SDMA are identical. For RSMA, this value is equal to 66.48Mbps (marked with a solid red circle); for NOMA, it is equal to 56.16Mbps (red square), and for SDMA it is only 37.44Mbps (dashed red circle).

⁸Since the two RXs are closely situated for this case, this is also an intuitively good choice. However, in general, the MCS levels of the private streams can be different depending on the channel realizations.

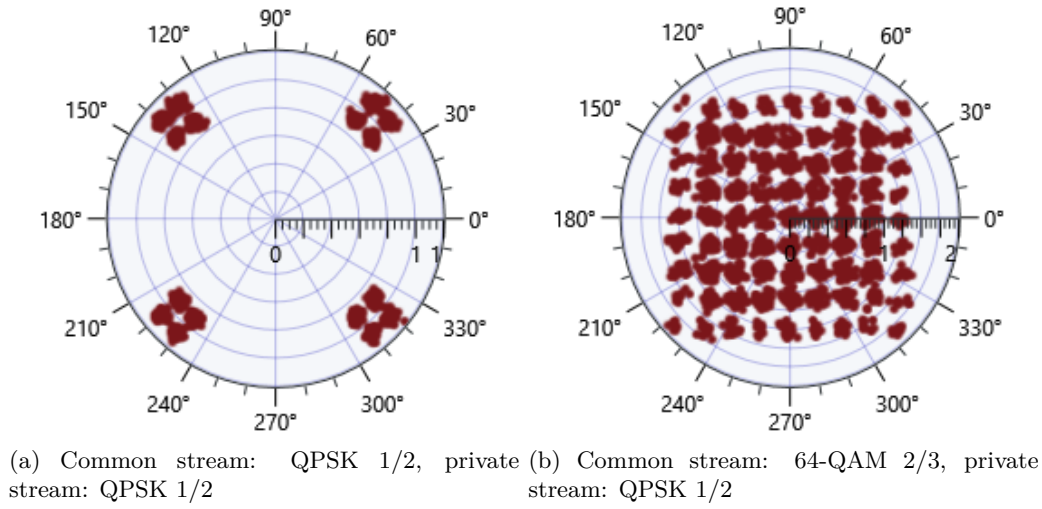


Figure 3.6: Superposed received symbol constellation for RSMA at RX 1 for the MCS levels in Fig. 3.5 corresponding to: (a) point ‘C’, which is an example of a conservative choice of MCS levels. Hence, the superposition of QPSK and QPSK constellations is clearly visible; (b) point ‘A’, which is an aggressive choice of MCS for the common stream. Instead of 256 distinct symbol clusters due to a superposition of 64-QAM and QPSK constellations, we only observe 81 distinct symbol clusters, which leads to high error rates for the common stream.

- 4) Finally, for all data points bar one, the RSMA sum throughput exceeds that of NOMA. Since RX 2’s channel is not much weaker than RX 1 for Case 1 (see α in Table 3.5), RSMA provides additional throughput gain through the private stream of RX 2.

Fig. 3.7 plots the empirical solution of OP_{mcs}^{ma} , broken down in terms of the contributions of the common and private streams. We make the following observations:

A. Unquantized CSI feedback:

- 1) As expected for RSMA, the contribution of the common stream (green bars) diminishes with increasing ρ (Remarks 6 and 7); in particular, for cases 1, 4 and 7, the common stream constitutes 51%, 62% and 62% of the sum throughput, respectively, whereas for cases 3, 6 and 9, this drops to 22%, 34% and 17%, respectively.
- 2) In cases 1, 4 and 7, where the channels have high spatial correlation (low ρ , see Table 3.5), SDMA’s sum throughput is significantly lower than those of RSMA and NOMA due to the high mutual interference. The gains of RSMA and NOMA in these cases stem from the decoding and elimination of interference through the common stream (Remark 6). In particular, RSMA achieves gains of 82%, 103% and 134%, respectively, over SDMA as shown by the up arrows in Fig. 3.7. In comparison, RSMA’s gain over NOMA for these cases is quite modest (23% for Case 1, none for Case 4 and 19% for Case 7).

- 3) In cases 3, 6 and 9, where the channels have low spatial correlation (high ρ , see Table 3.5), RSMA's gain over SDMA is modest due to the relatively lower mutual interference. On the other hand, the NOMA sum throughput is significantly worse than both RSMA and SDMA. In particular, RSMA achieves gains of 38%, 45%, and 19%, respectively, for these three cases over NOMA.

B. Quantized CSI feedback:

- 1) Unsurprisingly, all three multiple access schemes experience a sum throughput loss due to CSI quantization. However, SDMA and RSMA are more adversely affected than NOMA, especially in cases 4 through 9, where there is a significant pathloss difference between the two channels. For these cases, most of the throughput loss for RSMA and SDMA can be attributed to the private stream of RX 2 – the weaker user. In contrast, the robustness of NOMA to CSI quantization stems from the fact that RX 2 does not have a private stream. Furthermore, due to the very low private stream throughput for RX 2, the RSMA throughput performance is similar to NOMA for these cases.
- 2) However, in the absence of a large pathloss ($[-8.8, -6.4]$ dB) difference between the two channels – for instance, in cases 2 and 3 – the private stream throughput for RX 2 is still quite high, despite the loss from CSI quantization. Under these conditions, NOMA with its missing private stream for RX 2 continues to perform worse than both RSMA and SDMA, mirroring the performance seen with unquantized CSI feedback.
- 3) In all cases bar one, RSMA's gain over SDMA is much larger when compared to unquantized CSI feedback. This is because imperfect CSI has a more adverse effect on private stream throughputs due to the higher residual interference. Thus, via the common stream's throughput contribution, which stays pretty steady even with quantized CSI, RSMA's throughput loss is lower than that of SDMA.

In summary, RSMA achieves the best throughput performance across (a) different RX channels that vary in their relative strength and spatial correlation, and (b) different CSI quality. In some cases, SDMA is effective while in others NOMA is effective. RSMA is effective in all the cases, and thus effectively bridges between both SDMA and NOMA.

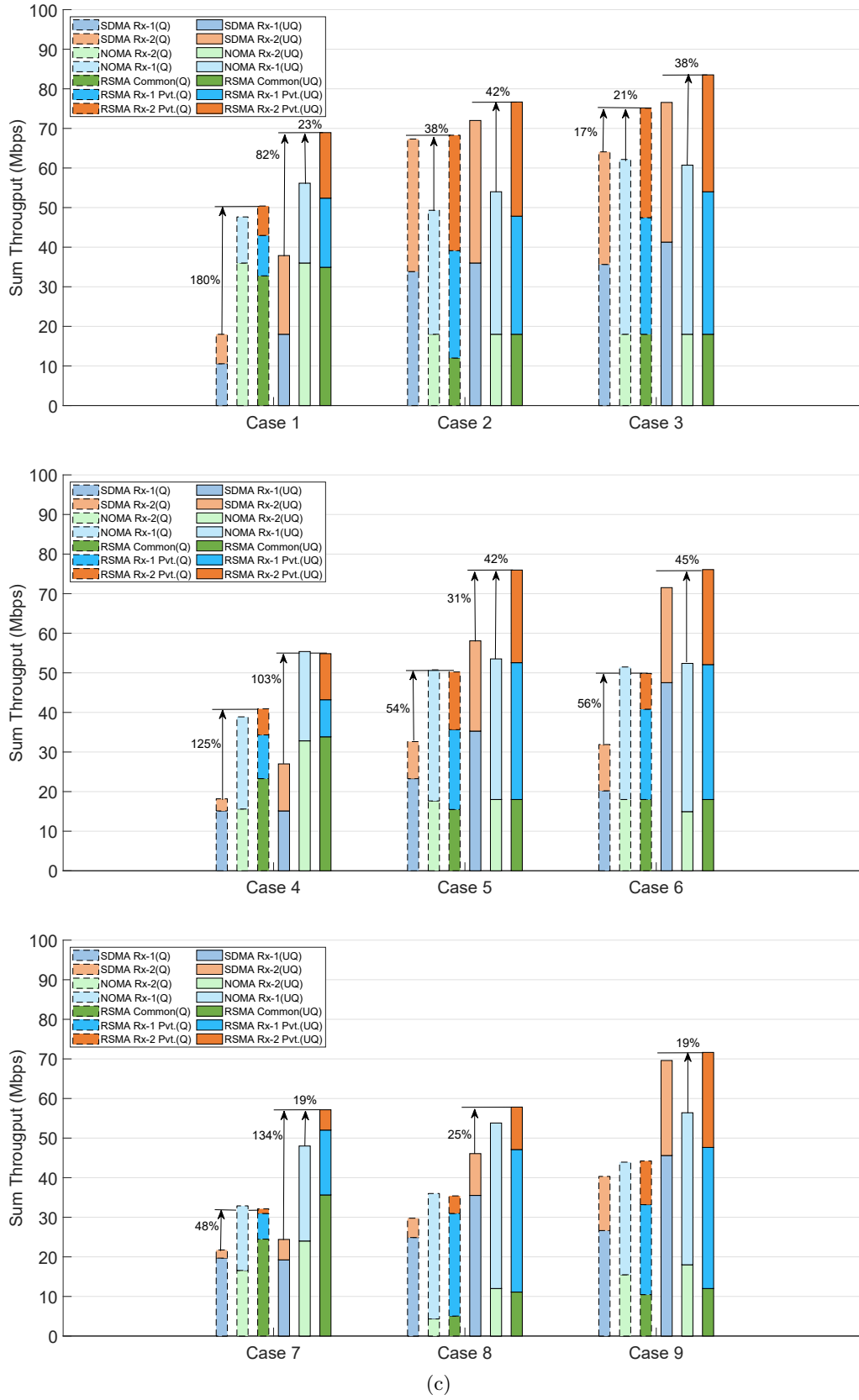


Figure 3.7: The maximum measured sum throughput under unquantized (UQ) and quantized (Q) CSI feedback for RSMA, SDMA and NOMA, which correspond to the empirical solutions of OP_{mcs}^{RSMA} , OP_{mcs}^{SDMA} and OP_{mcs}^{NOMA} , respectively. The arrows indicate RSMA's gain w.r.t. SDMA and NOMA.

3.4.3 RSMA v/s SDMA v/s NOMA: Fairness Comparison

In the previous subsection, we saw that for RSMA, the private stream throughput of RX 2 is severely diminished when there is a large pathloss difference between the two channels and imperfect CSI (bullet point B1). To increase RX 2's total throughput in such cases, a larger fraction (> 0.5) of the common stream can be allocated to it. This would not alter the RSMA sum throughput⁹, but would achieve better (max-min) fairness [7].

In this subsection, we experimentally verify this possibility of improved fairness afforded by RSMA. This is done in the following manner for each case in Fig. 3.7:

S1) if $R_1(\mathbf{P}_{\text{wmmse}}^{\text{RSMA}}) - R_2(\mathbf{P}_{\text{wmmse}}^{\text{RSMA}}) \leq R_c(\mathbf{P}_{\text{wmmse}}^{\text{RSMA}})$, then the common stream is partitioned such that $R_1(\mathbf{P}_{\text{wmmse}}^{\text{RSMA}}) + (|W_{c,1}|/|W_c|)R_c(\mathbf{P}_{\text{wmmse}}^{\text{RSMA}}) = R_2(\mathbf{P}_{\text{wmmse}}^{\text{RSMA}}) + (|W_{c,2}|/|W_c|)R_c(\mathbf{P}_{\text{wmmse}}^{\text{RSMA}})$ to ensure that both RXs have the same throughput;

S2) otherwise, when $R_1(\mathbf{P}_{\text{wmmse}}^{\text{RSMA}}) - R_2(\mathbf{P}_{\text{wmmse}}^{\text{RSMA}}) > R_c(\mathbf{P}_{\text{wmmse}}^{\text{RSMA}})$, the entire common stream is allocated to RX 2. This is similar to NOMA, albeit RX 2 has a private stream as well.

Remark 8 (Fairness while maximizing throughput) *User fairness in RSMA/SDMA/NOMA is often framed as an optimization problem, where the objective is to maximize the minimum user throughput through precoder design. Hence, the optimal precoders for fairness are, in general, different from the optimal precoders for sum throughput maximization. The fairness strategy in S1 and S2 differs from this viewpoint by trying to achieve fairness without changing the precoders. In other words, we are trying to achieve fairness while still maximizing the sum throughput. This ability to realize both objectives simultaneously is a unique feature of RSMA. S1 and S2 cannot be realized for NOMA and SDMA because the former does not have a private stream for RX 2 and the latter does not have a common stream.*

Fig. 3.8 compares the relationship between the sum throughput and the minimum throughput for RSMA (subject to S1 and S2), SDMA and NOMA. The case number is indicated beside each data point in the same color. The dashed $y = 2x$ line represents max-min fairness and the region above it is feasible for all three multiple access schemes. Thus, points that are closer (in terms of Euclidean distance) to the dashed line represent *fairer* outcomes. Based on this insight, we make the following remarks:

⁹For precisely this reason, the allocation of the common stream between the two RXs is not mentioned when discussing the sum throughput results in Section 3.4.2.

- When S1 is true, the corresponding RSMA point lies on the dashed line (i.e., achieves max-min fairness), as seen for all cases bar one in Figs. 3.8a and b. Otherwise, for S2, the RSMA point lies *as close as possible* above the dashed line, as seen for case 8 in Figs. 3.8a and b.
- In Fig. 3.8a, SDMA also achieves approximate fairness for some cases (namely 1, 2, 3 and 4), as does NOMA (case 7). However, the RSMA points for these cases lie to the *northeast* of the corresponding SDMA/NOMA points, which indicates fairness at a higher sum throughput.

In summary, RSMA is capable of achieving better fairness at a higher sum throughput than both SDMA and NOMA, regardless of CSI quality, through a judicious allocation of the common stream to the weaker RX.

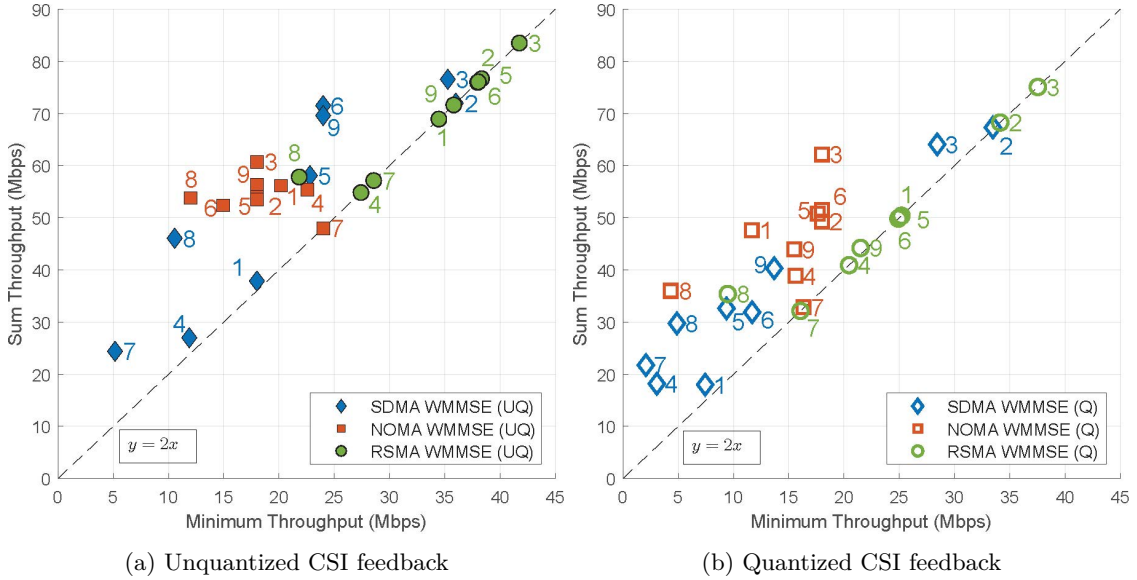


Figure 3.8: Fairness comparison between RSMA (subject to S1 and S2), SDMA and NOMA. The number beside each data point indicates the measurement case from Fig. 3.4. The dashed line ($y = 2x$) corresponds to max-min fairness and points that are closer (in terms of Euclidean distance) to this line represent fairer outcomes.

3.5 Conclusion

In this chapter, we realized the first-ever RSMA prototype by using SDRs and modifying the IEEE 802.11 standard. We implemented the key functions of the transceiver in the test bed, which includes CSI estimation, CSI quantization, precoder maximization and SIC receiver. We also solved the MCS-limited throughput maximization problem iteratively to validate RSMA's promised performance gains in the theoretical literature experimentally. To this effect, this study measured the throughput and fairness performance of RSMA, SDMA and NOMA using our prototype for

the two-user MISO scenario over (a) different pairs of channels varying in terms of their relative strength and spatial correlation, and (b) different CSI quality. Across these different cases, RSMA achieved the highest sum throughput, whereas SDMA and NOMA were each effective only in some cases. This chapter also showed that RSMA achieved better fairness at a higher sum throughput than both SDMA and NOMA. The imperfect SIC effects on the RSMA receiver are first observed experimentally in the experiments. The importance of selecting the most suitable MCS levels in MCS-limited throughput maximization problem is firstly considered in a RSMA prototype. As the first-ever experimental comparison between RSMA, SDMA and NOMA, the efforts in this chapter address the pressing question of which PHY layer multiple access technique is best equipped to deliver the performance enhancements expected from next-generation networks like 6G. Thus, the superior performance of RSMA in these experiments provides the impetus for further experimental efforts at a larger scale and in more challenging operating environments.

Rate-Splitting Multiple Access

Experimental Study for

Non-Orthogonal Unicast Multicast

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4.1 Introduction

NOUM is an important communication mode for providing both unicast (one-to-one) and multicast (one-to-many) services simultaneously. Despite the benefits of RSMA-based NOUM have been investigated analytically in [20], [32], an experimental validation of the same is absent in the literature. This motivates our efforts in this chapter, where we first characterize the conditions under which RSMA-based NOUM achieves a higher unicast sum rate than MUP-based NOUM while satisfying a given multicast rate. We then validate this experimentally and find that the

gains of RSMA-based NOUM are more pronounced with increasing channel correlation among the users.

The rest of this chapter is organized as follows. In Section 4.2, this section introduces the RSMA-based NOUM system model for a two-user MISO scenario and formulated into a MCS-limited throughput maximization problem with low complexity precoder design. In Section 4.3, the theoretical benefits of RSMA-based NOUM are analyzed. In Section 4.4, the description of the measurements and the results of each measurement are plotted on both rate region figures and bar chart figures. This section also compares the rate region achieved by RSMA-based NOUM and SDMA-based NOUM. In the last, Section 4.5 summarizes the benefits of RSMA-based NOUM and the key findings of this chapter.

4

4.2 System Model for RSMA-based NOUM

We follow the RSMA-based NOUM architecture of [20]. Consider the two-user MISO case. Let W_0 denote the multicast message, and W_i ($i = 1, 2$) the unicast message for user i . At the TX, each W_i is split into common and private portions, denoted by $W_{c,i}$ and $W_{p,i}$, respectively (in general, the splitting ratio need not be 50-50 nor does it have to be the same across both users). Let $W_c := [W_0, W_{c,1}, W_{c,2}]$ denote the common message formed by concatenating the multicast message along with the common portions of each unicast message. Through an appropriately chosen MCS level, W_c is transformed into a common stream, s_c . Similarly, $W_{p,1}$ and $W_{p,2}$ are separately encoded and modulated to form private streams, s_1 and s_2 , respectively. All three streams are linearly precoded to form the transmit signal:

$$\mathbf{x} = \mathbf{p}_c s_c + \mathbf{p}_1 s_1 + \mathbf{p}_2 s_2, \quad (4.1)$$

where $\mathbf{p}_c$ denotes the common stream precoder, and $\mathbf{p}_i$ the private stream precoder of user i . Let $\mathbf{h}_i$ denote the channel between the TX and the i -th user. The received signal, y_i , at the i -th user is given by:

$$\begin{aligned} y_i &= \mathbf{h}_i^H \mathbf{x} + n_i \quad (i = 1, 2) \\ &= \mathbf{h}_i^H \mathbf{p}_c s_c + \mathbf{h}_i^H \mathbf{p}_1 s_1 + \mathbf{h}_i^H \mathbf{p}_2 s_2 + n_i, \end{aligned} \quad (4.2)$$

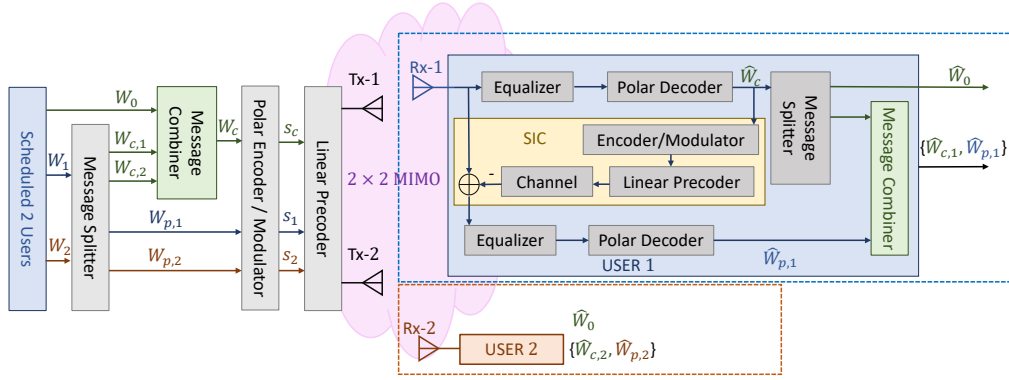


Figure 4.1: RSMA-NOUM block diagram for two-user MISO scenario

where n_i is the receiver thermal noise. Upon receiving y_i , the i -th user first decodes s_c , while treating the interference from the s_1 and s_2 as noise. This enables the i -th user to recover W_0 and a part of its unicast message ($W_{c,i}$). Then, through SIC, the i -th user decodes s_i to recover the rest of its unicast message ($W_{p,i}$) by treating the interference from the other private stream as noise. It is easy to see from (4.2) that the achievable multicast and unicast rates are a function of the precoders and MCS levels chosen for each of the three streams. We describe these next.

- **Precoder Design:** Let $\hat{\mathbf{h}}_i$ denote the estimate of $\mathbf{h}_i$ at the TX. Since s_c must be decoded by both users, $\mathbf{p}_c$ must have sufficient gain along each $\hat{\mathbf{h}}_i$. Thus, weighted Maximum Ratio Transmission (MRT) precoding is a reasonable good heuristic choice for $\mathbf{p}_c$, i.e.,

$$\mathbf{p}_c(t) := \sqrt{P_t(1-t)} \frac{(\hat{\mathbf{u}}_1 + \hat{\mathbf{u}}_2)}{\|\hat{\mathbf{u}}_1 + \hat{\mathbf{u}}_2\|}, \quad (4.3)$$

where $\hat{\mathbf{u}}_i$ denotes the unit vector along $\hat{\mathbf{h}}_i$. On the other hand, user i decodes s_i by treating the interference from s_j ($j \neq i$) as noise. Hence, to suppress this interference as much as possible, we consider Zero-Forcing (ZF) precoding for the private streams, i.e.,

$$\mathbf{p}_i(t) := \sqrt{P_t(t/2)} \hat{\mathbf{u}}_j^\perp \quad (j \neq i; i = 1, 2), \quad (4.4)$$

where $\mathbf{u}_i^\perp$ denotes the unit vector along the nullspace of $\hat{\mathbf{h}}_i$. In (4.3) and (4.4), the parameter $t \in [0, 1]$ determines the allocation of the transmit power, P_t , to the three streams.

- **MCS:** An MCS level can be defined by a pair $\mathcal{I} = (m, r)$, where positive integer m denotes the bits per modulation symbol (e.g., 2 for QPSK) and $r \in (0, 1]$ denotes the code rate. In general, the MCS level of each stream can be distinct. Hence, let $\mathcal{I}_c = (m_c, r_c)$, $\mathcal{I}_1 = (m_1, r_1)$ and $\mathcal{I}_2 = (m_2, r_2)$ denote the MCS levels chosen for s_c , s_1 and s_2 , respectively.

Let $\mathcal{M} = \{\mathcal{I}_c, \mathcal{I}_1, \mathcal{I}_2\}$. As a function of t and $\mathcal{M}$, let $R_c(t, \mathcal{M})$ and $R_i(t, \mathcal{M})$ denote the *MCS-limited rate* (in bits/s/Hz) for s_c and s_i , respectively. Their expressions are as follows:

$$R_c(t, \mathcal{M}) = m_c r_c \times \mathbb{P}(\hat{W}_c^{(1)} = W_c; \hat{W}_c^{(2)} = W_c) \quad (4.5)$$

$$R_i(t, \mathcal{M}) = m_1 r_1 \times \mathbb{P}(\hat{W}_{p,i} = W_{p,i}) \quad (i = 1, 2), \quad (4.6)$$

where the probability terms capture the rate loss due to decoding errors. In particular, $\mathbb{P}(\hat{W}_c^{(1)} = W_c; \hat{W}_c^{(2)} = W_c)$ in (4.5) denotes the probability that W_c is correctly decoded at both users, while $\mathbb{P}(\hat{W}_{p,i} = W_{p,i})$ in (4.6) denotes the probability that $W_{p,i}$ is correctly decoded at the i -th user. These probabilities are difficult to characterize in closed-form, but can be evaluated empirically (see Section 4.4).

Let $R_{\text{mult}}^{\text{RS}}(t, \mathcal{M})$ denote the multicast rate and $R_{\text{uni}}^{\text{RS}}(t, \mathcal{M})$ the unicast sum rate for RSMA-based NOUM. Their expressions are related to (4.5) and (4.6) as follows:

$$R_{\text{mult}}^{\text{RS}}(t, \mathcal{M}) = \frac{|W_0|}{|W_c|} R_c(t, \mathcal{M}) \quad (4.7)$$

$$R_{\text{uni}}^{\text{RS}}(t, \mathcal{M}) = \left(1 - \frac{|W_0|}{|W_c|}\right) R_c(t, \mathcal{M}) + \sum_{i=1}^2 R_i(t, \mathcal{M}) \quad (4.8)$$

where $|W|$ denotes the size of message W .

Remark 9 For the special case when $W_c = W_0$, RSMA-based NOUM reduces to Mulp-based NOUM (see Table 2.1). From (4.7) and (4.8), let $R_{\text{mult}}^{\text{MULP}}(t, \mathcal{M})$ and $R_{\text{uni}}^{\text{MULP}}(t, \mathcal{M})$ denote the corresponding multicast and unicast sum rate, respectively.

In the following section, we describe how RSMA-based NOUM can achieve higher unicast sum rate than Mulp-based NOUM, for fixed multicast rate.

4.3 Benefits of RSMA-based NOUM

From (4.3)-(4.6), the sum rate for RSMA-based NOUM, denoted by $R_{\text{sum}}(t, \mathcal{M})$, is given by:

$$R_{\text{sum}}(t, \mathcal{M}) := R_c(t, \mathcal{M}) + R_1(t, \mathcal{M}) + R_2(t, \mathcal{M}). \quad (4.9)$$

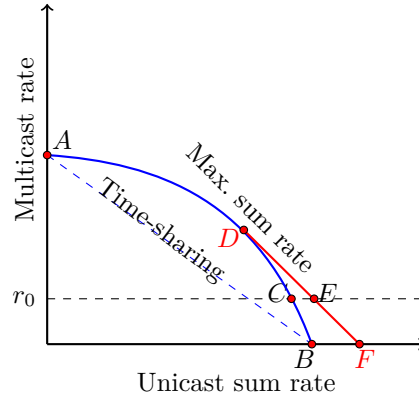


Figure 4.2: An illustration of the achievable rate region for MULP-based NOUM (blue curve) and RSMA-based NOUM (solid red line).

For fixed t , let

$$R_{\text{sum}}(t) := \max_{\mathcal{M} \in \mathbb{M}} R_{\text{sum}}(t, \mathcal{M}) = R_c(t) + R_1(t) + R_2(t), \quad (4.10)$$

where $\mathbb{M}$ denotes the collection of permissible MCS levels for the three streams, typically pre-determined through standards.

The achievable pairs of multicast and unicast sum rates for MULP-based and RSMA-based NOUM are shown in Fig. 4.2. The achievable region is convex due to time-sharing. Based on Fig. 4.2, we make the following remarks.

Remark 10 (Purely Multicast) *The y-axis points in the achievable region correspond to purely multicast, where only the common stream is transmitted. Point A represents the peak multicast rate, equal to $R_{\text{sum}}(0)$ in (4.10).*

Remark 11 (Unicast SDMA) *The x-axis from the origin till point B corresponds to unicast SDMA, with only the two private streams being transmitted. Point B represents the peak unicast SDMA sum rate, equal to $R_{\text{sum}}(1)$.*

Remark 12 (MULP-based NOUM) *For $t \in (0, 1)$, three streams are transmitted. Assuming perfect SIC and MCS levels of arbitrary granularity, the blue curve represents the rate boundary for MULP-based NOUM. In particular, the following multicast and unicast sum rates are achievable*

(point C):

$$R_{\text{mult}}^{\text{MULP}}(t) = R_c(t) \quad (4.11)$$

$$R_{\text{uni}}^{\text{MULP}}(t) = R_1(t) + R_2(t) \quad (4.12)$$

4

Remark 13 (Max. Sum Rate) Let R_{sum}^* denote the maximum sum rate obtained at t^* , corresponding to point D , i.e.,

$$R_{\text{sum}}^* := R_{\text{sum}}(t^*) = R_c(t^*) + R_1(t^*) + R_2(t^*). \quad (4.13)$$

We now characterize the conditions under which RSMA-based NOUM achieves a higher unicast sum rate than MULP-based NOUM for a fixed multicast rate.

Lemma 1 For $t^* \in [0, 1)$ and a multicast rate r_0 satisfying $r_0 < R_c(t^*)$, RSMA-based NOUM achieves a higher unicast sum rate than MULP-based NOUM.

From Remark 12, there exists t_0 that achieves the multicast rate, r_0 , through MULP-based NOUM, with the resulting unicast sum rate equal to $R_1(t_0) + R_2(t_0)$, i.e.:

$$R_{\text{mult}}^{\text{MULP}}(t_0) = R_c(t_0) = r_0 \quad (4.14)$$

$$R_{\text{uni}}^{\text{MULP}}(t_0) = R_1(t_0) + R_2(t_0) \quad (4.15)$$

Alternately, by choosing t^* and allocating – through (4.7) – only as much of the common stream as needed to satisfy the multicast rate, RSMA-based NOUM achieves the following rates:

$$\begin{aligned} R_{\text{sum}}(t^*) &= R_c(t^*) + R_1(t^*) + R_2(t^*) \\ &= R_{\text{mult}}^{\text{RS}}(t^*) + R_{\text{uni}}^{\text{RS}}(t^*) \end{aligned} \quad (4.16)$$

$$\text{where, } R_{\text{mult}}^{\text{RS}}(t^*) = r_0 \quad (4.17)$$

$$R_{\text{uni}}^{\text{RS}}(t^*) = \underbrace{R_c(t^*) - r_0}_{\geq 0} + R_1(t^*) + R_2(t^*) \quad (4.18)$$

The rest of the common stream is allocated to the unicast messages. This scheme corresponds to

point E in Fig. 4.2. Hence, we have

$$\begin{aligned}
 R_{\text{sum}}(t_0) &\leq R_{\text{sum}}(t^*) && \text{(Remark 13)} \\
 r_0 + R_{\text{uni}}^{\text{MULP}}(t_0) &\leq r_0 + R_{\text{uni}}^{\text{RS}}(t^*) && \text{(from (4.14)-(4.18))} \\
 \implies R_{\text{uni}}^{\text{MULP}}(t_0) &\leq R_{\text{uni}}^{\text{RS}}(t^*). && (4.19)
 \end{aligned}$$

Corollary 1 *For $r_0 = 0$ (i.e., purely unicast), we obtain the well-known result that the RSMA unicast sum rate is at least as large as the SDMA unicast sum rate, as captured by point F on the x -axis in Fig. 4.2.*

Corollary 2 *When $t^* = 1$, RSMA-based NOUM switches to unicast only (no common stream), and RSMA-based NOUM coincides with MULP-based NOUM. The RSMA-based NOUM region (red line) and MULP-based NOUM (blue line) coincide.*

In the following section, we experimentally validate Lemma 1.

4.4 Measurement Results

We realize a 2×2 MISO scenario for RSMA-based NOUM using a pair of NI USRP-2942 software-defined radios. We use OFDM signals for the three streams, with the MCS levels chosen from Table 3.3. For details of our implementation (e.g., channel estimation, OFDM frame structure, etc.), see [15]. We consider four cases as shown in Fig. 4.3, and described below:

- **Cases 1-3:** For the first three cases, the two users experience similar pathloss, but their channel correlation progressively increases from Case 1 (least) to Case 3 (highest), as the angular separation of the users w.r.t TX narrows.
- **Case 4:** To study the impact of different pathloss among users, we consider an additional case where user 2 is placed 2m further away from the TX than user 1. The channel correlation is similar to Case 3.

For each case and fixed $(t, \mathcal{M})$, we conduct 50 measurement runs, where each run involves transmitting a payload of 50 OFDM symbols per stream.

The measured throughputs for the common stream and the i -th user's private stream equals $BR_c(t, \mathcal{M})$ and $BR_i(t, \mathcal{M})$, where B denotes the bandwidth (equal to 12MHz). The probabilities

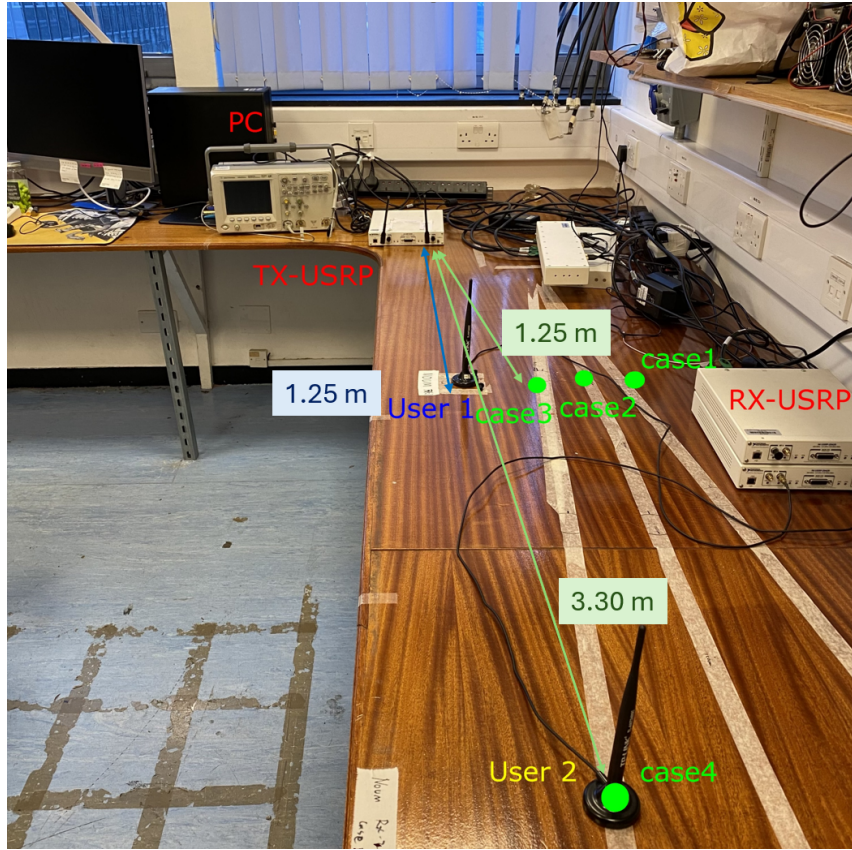


Figure 4.3: Layout of measurement cases. The location of the user antennas corresponds to Case 4.

in (4.5) and (4.6) are evaluated empirically (i.e., the fraction of measurement runs in which the common/private stream is successfully decoded by both/ i -th user). The maximization in (4.10) is done via brute-force search over the MCS levels in Table 3.3, and t varies from 0 to 1 in steps of 0.1. A list of parameters used for our measurements is provided in Table 4.1.

Fig. 4.4 plots the measured NOUM rate region (i.e., the empirical version of Fig. 4.2) for each case, and Fig. 4.5 plots the breakdown of the sum throughput in terms of the common and private stream contributions. We make the following observations:

1. Across Figs. 4.4a-d, the gap between the rate regions of RSMA- and Mulp-based NOUM increases with the channel correlation. As per Lemma 1, RSMA-based NOUM yields higher unicast throughput than Mulp-based NOUM when the multicast data rate is less than 6 Mbps (case 1), 18 Mbps (case 2) and 36 Mbps (cases 3, 4).
2. Between Fig. 4.4c and d, the main difference occurs at $t = 1$, which corresponds to unicast SDMA (Remark 11). The corresponding sum throughput in Case 4 (30Mbps) is lower than in Case 3 (36Mbps) due to the weaker channel of user 2 and the high interference from user

Parameter	Value
Center frequency	2.484GHz
Transmit power, P_t	23dBm
TX antenna length	0.13m
Fraunhofer distance	0.28m
Total/Data/Pilot/Guard band subcarriers	64/48/4/12
Cyclic prefix length	16
Bandwidth, B (excl. overheads)	12MHz
OFDM symbols in payload	50
Experiment runs (per case)	50
Distances: User 1 $\rightarrow$ TX	1.25m
User 2 $\rightarrow$ TX (Case 1-3, 4)	(1.25m, 3.30m)
Channel coding	Polar

Table 4.1: List of parameters used in our experiments.

1 due to the high channel correlation. This is evidenced by the lower MCS for user 2 ($\mathcal{I}_2 = 2$ in Fig. 4.5d v/s $\mathcal{I}_2 = 3$ in Fig. 4.5c). For $t \neq 1$, the common stream helps the weaker user 2 to partially decode and cancel the high interference it experiences from user 1, leading to improved sum rate performance. Thus, the rate-region for RSMA-based NOUM is identical in Figs. 4.4c and 4.4d. This suggests that the channel correlation is a stronger indicator of RSMA-based NOUM's rate performance than the relative pathloss difference between the user channels.

3. With increasing channel correlation, the highest sum rate is realized by allocating more power to the common stream. This is evidenced by t^* decreasing from 0.5 (case 1) to 0.3 (case 2) to 0.2 (cases 3, 4). Consequently, the contribution from the common stream to the max. sum throughput increases from 8% (case 1) to 33% (case 2) and then to 61% (cases 3, 4). The more the common stream contributes to the max. sum throughput, the larger the gap between the rate regions of RSMA- and Mulp-based NOUM, as there is greater flexibility in using the common stream for multicast and unicast.
4. For t approaching but not equal to 1, the measured throughput is 0 (e.g., $t = 0.9$ in all the cases). This is because the low power (only 10% of the power budget for $t = 0.9$) allocated to the common stream is insufficient to decode even the smallest MCS level ($\mathcal{I} = 0$ in Table 3.3). This causes imperfect SIC and error propagation while decoding the private streams.
5. For several values of t across Figs. 4.4a-d, the measured multicast and unicast throughputs

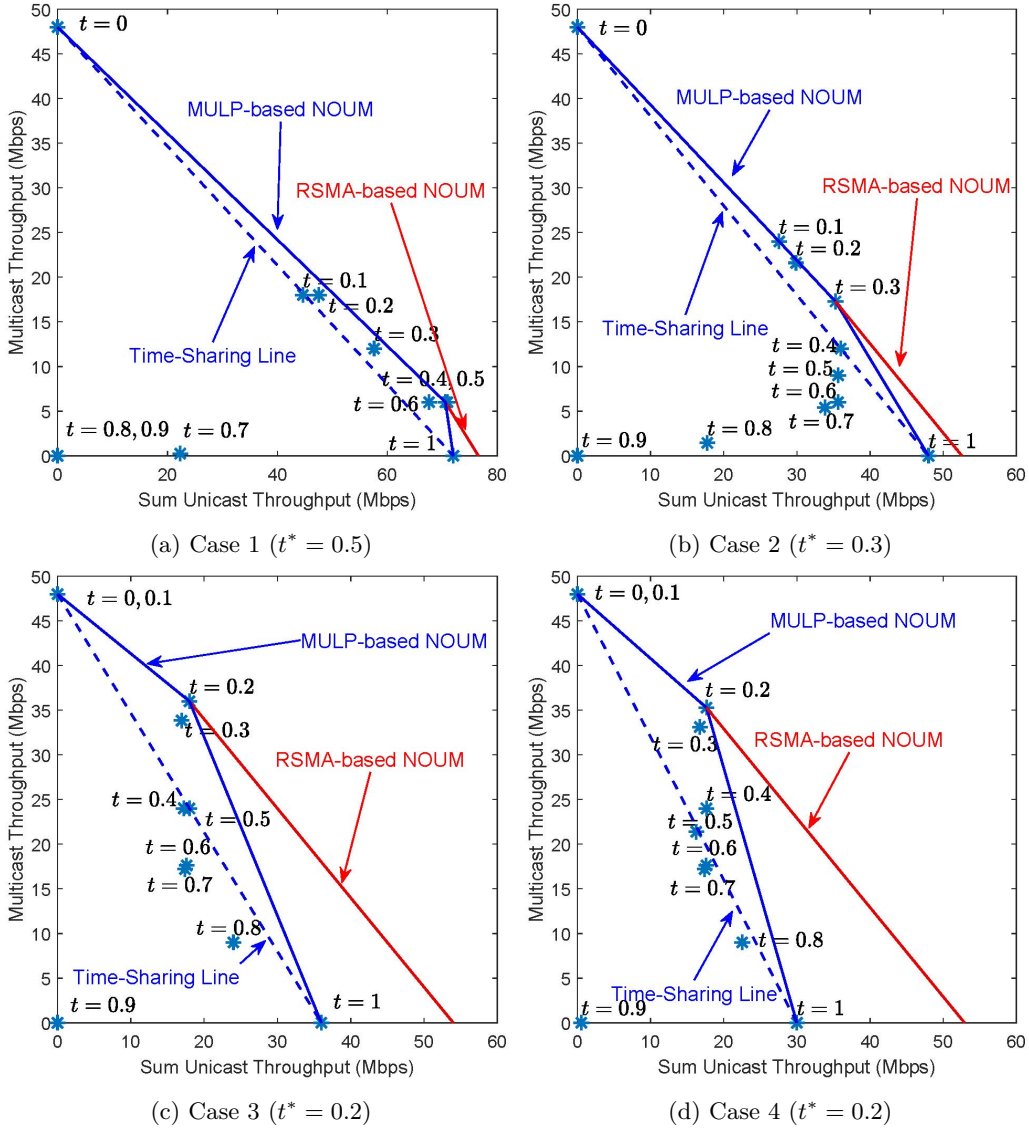


Figure 4.4: The measured NOUM rate region (i.e., the empirical version of Fig. 4.2). RSMA-based NOUM yields higher unicast throughput than MULP-based NOUM when the multicast data rate is less than 6 Mbps (Case 1), 18 Mbps (Case 2) and 36 Mbps (Case 3, 4).

are below the time-sharing line. This is due to the limited granularity of the MCS levels. As an example, consider the point corresponding to $t = 0.5$ for Case 2 in Fig. 4.4b. One way to push this point above the time-sharing line is by increasing the multicast throughput using a higher MCS level for the common stream. However, from Fig. 4.5b, the next highest MCS level for the common stream ($\mathcal{I}_C = 2$) is too aggressive, causing imperfect SIC and error propagation. More broadly, points 4 and 5 underscore the importance of the dual assumptions of perfect SIC and MCS levels of arbitrary granularity in Remark 12.

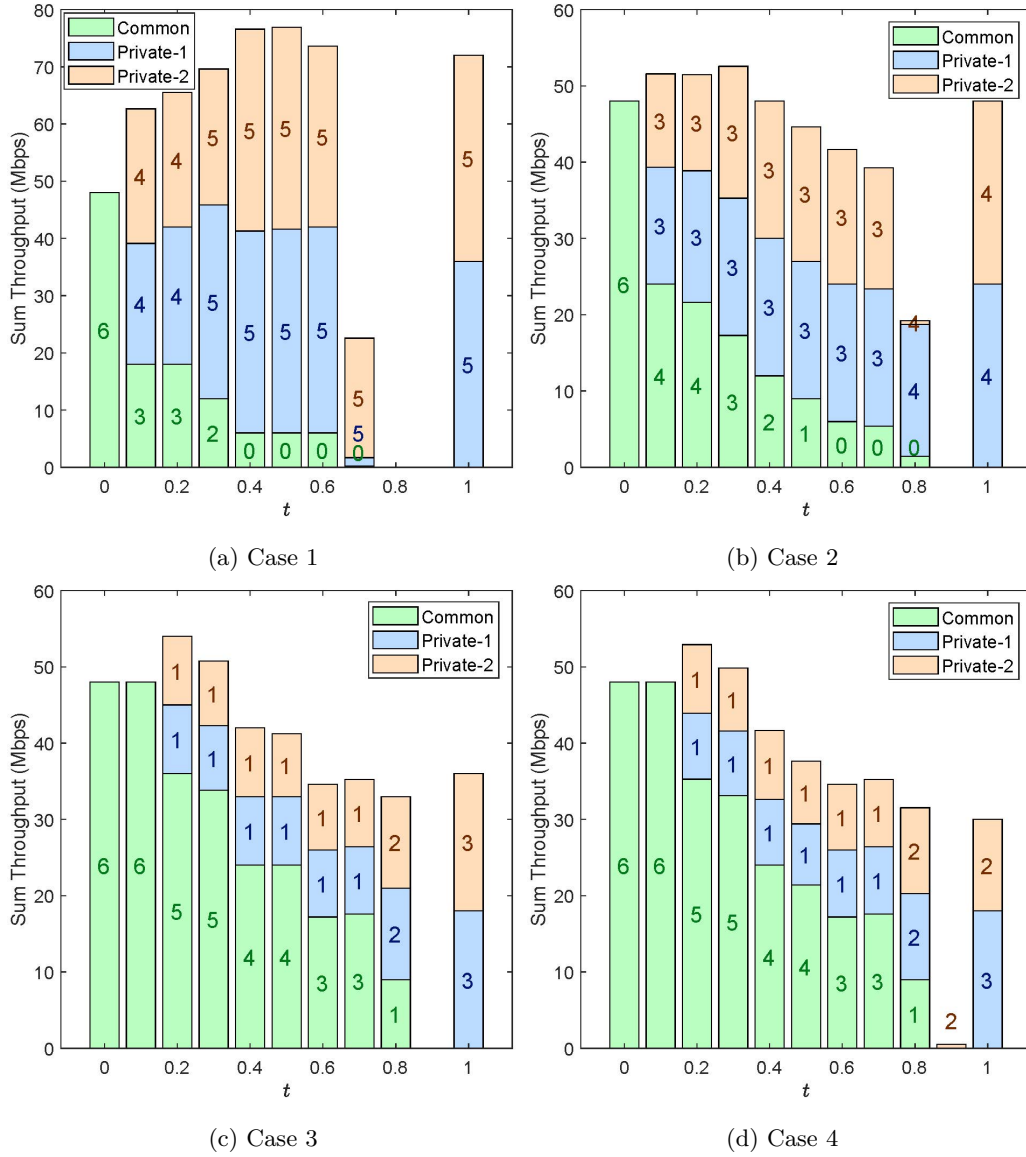


Figure 4.5: The measured sum throughput broken down in terms of common and private stream contributions. The number within each bar denotes the MCS index from Table 3.3 for the corresponding stream.

4.5 Summary

This chapter characterized and experimentally investigated RSMA-based NOUM for the two-user case. We saw that the benefits of RSMA-based NOUM are more pronounced with increasing channel correlation. This is because the common stream contributes an increasingly larger share to the sum throughput, which provides greater flexibility in using it for multicast and unicast. As a result, the gap between the rate regions of RSMA- and Mulp-based NOUM also increases with the channel correlation.

Rate-Splitting Multiple Access

Experimental Study for

Overloaded Multi-group

Multicasting

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5.1 Introduction

The three main approaches for MGM precoder design: SDMA-based MGM, NOMA-based MGM and RSMA-based MGM. SDMA-based MGM focuses on mitigating inter-group interference by treating other groups' interference as noise. The performance of SDMA-based MGM is investigated analytically in [38]–[44], [46]–[48]. NOMA-based MGM based on the power domain NOMA principle, which decodes the inter-group interference. NOMA-based MGM shows better performance than SDMA-based MGM in overloaded scenarios [55], [56]. RSMA-based MGM improves upon both by partially decoding and treating interference as noise, achieving superior fairness and minimum rate [17]. Even the theoretical studies have shown RSMA-based MGM outperforms SDMA-based MGM and NOMA-based MGM, experimental validation is still lacking. In this chapter, we address this gap by conducting the first-ever experimental comparison of the fairness performance of RSMA-, SDMA- and NOMA-based MGM. The contributions are as follows:

- For two groups with two users per group, we formulate the practically relevant *MCS-limited* max-min throughput fairness problem for RSMA-based MGM. In the process, we explain how RSMA-based MGM has an advantage over SDMA- and NOMA-based MGM in terms of realizing max-min fairness through a flexible allocation of the common stream (Section 5.2.2).
- Using RSMA prototype (Section 5.3), this chapter empirically solve the above-defined MCS-limited max-min fairness problem through measurements. In the experiments (Section 5.4), we realize nine cases, with the intention of capturing a wide variety in terms of pathloss difference and spatial correlation between the two groups (subject to closely situated users in each group). Over these nine cases, we observe that RSMA-based MGM achieves fairness at a higher minimum throughput than SDMA- and NOMA-based MGM. This is consistent with theoretical predictions [17], [62], [64].

5.2 System Model

Without loss of generality, we consider MGM involving two groups, with each group comprising two single-antenna users (i.e., $N_1 = N_2 = 2$). Each user belongs to exactly one group, and we assume that the composition of the groups is known beforehand to the TX. For brevity, let (u, g) denote user u in group g ($u, g \in \{1, 2\}$). All the users within a group desire the same message from

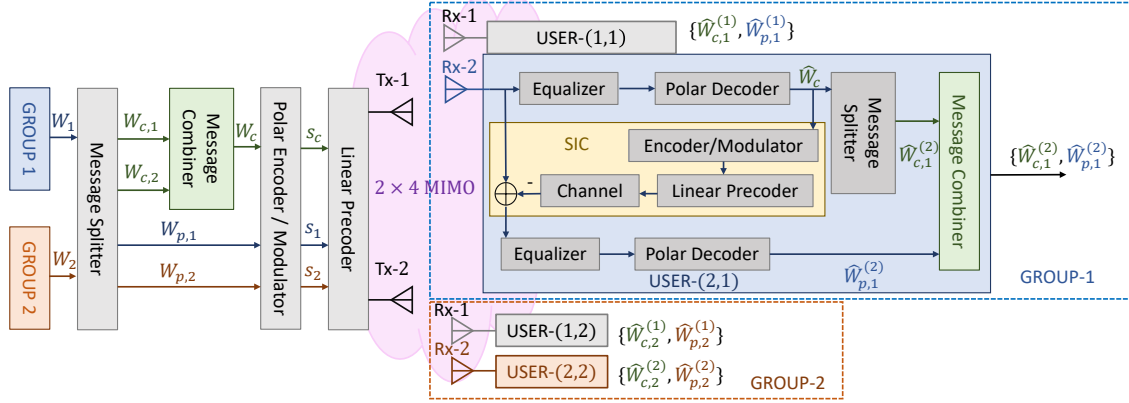


Figure 5.1: An illustration of RSMA-based MGM.

the TX, which is communicated over two stages – in the first stage, the TX acquires channel state information (CSIT), which is used in the second stage to design the RSMA-based MGM precoders and transmit signal. The two stages are described in more detail next.

5.2.1 Stage 1: CSIT Acquisition

We consider data transmission using OFDM signals over N_c subcarriers. Let N_t denote the number of antenna elements at the TX, and let $\mathbf{h}_{u,g}[k] \in \mathbb{C}^{N_t}$ denote the slowly varying, flat fading channel experienced by (u, g) over the k -th subcarrier ($k = 0, \dots, N_c - 1$). Through orthogonal pilots transmitted by the TX, (u, g) obtains an estimate (e.g., least squares) of $\mathbf{h}_{u,g}[k]$, denoted by $\hat{\mathbf{h}}_{u,g}[k]$. To reduce CSI feedback overhead, (u, g) evaluates the *wideband CSI* by averaging $\hat{\mathbf{h}}_{u,g}[k]$ over the subcarriers, i.e.,

$$\hat{\mathbf{h}}_{u,g} = \frac{1}{N_c} \sum_{k=0}^{N_c-1} \hat{\mathbf{h}}_{u,g}[k]. \quad (5.1)$$

5.2.2 Stage 2: RSMA-based MGM Signal Design

Fig. 5.1 depicts RSMA-based MGM in operation [17]. Let W_1 and W_2 denote the messages meant for groups 1 and 2, respectively. At the TX, each W_g ($g = 1, 2$) is split (by the message splitter block) into common and private portions denoted by $W_{c,g}$ and $W_{p,g}$, respectively (i.e., $W_g = W_{c,g} \cup W_{p,g}$). The two common portions – $W_{c,1}$ and $W_{c,2}$ – are then combined (by the message combiner block) into a common message, W_c . The three message components – W_c , $W_{p,1}$ and $W_{p,2}$ – are then individually encoded and modulated to form data streams $s_c[k]$, $s_1[k]$ and $s_2[k]$, respectively, over the subcarriers ($k = 0, \dots, N_c - 1$). $s_c[\cdot]$ is referred to as the *common*

stream, whereas $s_1[\cdot]$ ($s_2[\cdot]$) is referred to as the *private stream* of group 1 (group 2).

The streams are then linearly precoded, giving rise to the transmit signal $\mathbf{x}[k]$ over the k -th subcarrier, which can be expressed as follows:

$$\mathbf{x}[k] = \mathbf{p}_c s_c[k] + \mathbf{p}_1 s_1[k] + \mathbf{p}_2 s_2[k]. \quad (5.2)$$

where $\mathbf{p}_c$ is referred to as the *common stream precoder* and $\mathbf{p}_g$ the *private stream precoder* of group g . Let $\mathbf{P} = [\mathbf{p}_c, \mathbf{p}_1, \mathbf{p}_2]$ denote the collection of precoders.

Remark 14 $\mathbf{P}$ is designed as a function of the (imperfect) CSIT acquired in Stage 1 [see (5.8)-(5.9)]. However, to avoid messy notation, this functional dependence is not shown explicitly.

Remark 15 Assuming unit symbol power (i.e., $\mathbb{E}[|s_c[k]|^2] = \mathbb{E}[|s_1[k]|^2] = \mathbb{E}[|s_2[k]|^2] = 1$), $\|\mathbf{p}_c\|^2$ and $\|\mathbf{p}_g\|^2$ represent the power allocated to the common stream and group g 's private stream, respectively. In (5.2), the same precoders are applied across all subcarriers, which translates to uniform power allocation over the subcarriers. This suboptimal choice is driven by our use of imperfect CSIT in (5.1) to design $\mathbf{P}$. The optimal power allocation – dictated by waterfilling – requires knowledge of $\mathbf{h}_{u,g}[k]$ at the TX for all k .

The received signal at (u, g) , denoted by $y_{u,g}[k]$, is given by:

$$\begin{aligned} y_{u,g}[k] &= \mathbf{h}_{u,g}^H[k] \mathbf{x}[k] + n_{u,g}[k] \\ &= \mathbf{h}_{u,g}^H[k] \mathbf{p}_c s_c[k] + \mathbf{h}_{u,g}^H[k] \mathbf{p}_g s_g[k] + \mathbf{h}_{u,g}^H[k] \mathbf{p}_{g'} s_{g'}[k] \\ &\quad + n_{u,g}[k] \end{aligned} \quad (5.3)$$

From the perspective of (u, g) , the first and second terms in (5.3) both contain useful information – the former represents the common stream component, a part of which is meant solely for group g users, while the latter represents the private stream component meant solely for group g users. In contrast, the third term does not contain useful information, as it captures the private stream component meant for the other group $g' \neq g$. Finally, $n_{u,g}[k] \sim \mathcal{CN}(0, \sigma^2)$ denotes the receiver noise at (u, g) .

We assume that precoded pilot¹ signals have been interspersed across some subcarriers of $\mathbf{x}[k]$ to help (u, g) estimate $\mathbf{h}_{u,g}^H[k] \mathbf{p}_c$ and $\mathbf{h}_{u,g}^H[k] \mathbf{p}_g$ upon receiving $y_{u,g}[k]$, in order to decode the useful

¹These pilots play a role similar to the demodulation reference signals (DM-RS) used in LTE and 5G NR.

information in the first two terms of (5.3). Using its estimate of $\mathbf{h}_{u,g}^H[k]\mathbf{p}_c$, (u, g) first equalizes (5.3) to decode $s_c[k]$ and recover W_c , by treating the second and third (interference) terms in (5.3) as noise. The decoded estimate of W_c is simultaneously sent to:

- (a) the message splitter block to extract $\hat{W}_{c,g}^{(u)}$, which is the estimate of $W_{c,g}$ at (u, g) , and
- (b) the SIC block, which generates an estimate of $\mathbf{h}_{u,g}^H[k]\mathbf{p}_c s_c[k]$ and subtracts it from $y_{u,g}[k]$. Using its estimate of $\mathbf{h}_{u,g}^H[k]\mathbf{p}_g$, (u, g) then equalizes the resulting residue to decode $s_g[\cdot]$ and recover $W_{p,g}$, while treating the interference from the third term of (5.3) as noise. Let $\hat{W}_{p,g}^{(u)}$ denote the decoded estimate of $W_{p,g}$ at u . Hence, the estimate of W_g at (u, g) , denoted by $\hat{W}_g^{(u)}$, is given by $\hat{W}_g^{(u)} = \hat{W}_{c,g}^{(u)} \cup \hat{W}_{p,g}^{(u)}$.

For the design of $\mathbf{P}$, from the perspective of the TX, let $R_{c,g}(\mathbf{P})$ denote the highest (instantaneous) rate (in bits/s/Hz) supporting error-free decoding of $s_c[\cdot]$ among the users in group g , under the assumption that $\hat{\mathbf{h}}_{u,g}$ in (5.1) constitutes perfect CSIT. It has the following expression:

$$R_{c,g}(\mathbf{P}) = \min_u \log_2 \left(1 + \frac{|\hat{\mathbf{h}}_{u,g}^H \mathbf{p}_c|^2}{\sigma^2 + |\hat{\mathbf{h}}_{u,g}^H \mathbf{p}_1|^2 + |\hat{\mathbf{h}}_{u,g}^H \mathbf{p}_2|^2} \right) \quad (5.4)$$

$(g \in \{1, 2\}),$

where the minimum reflects the fact that every user in group g must decode $s_c[\cdot]$.

Remark 16 (Achievability) *Strictly speaking, $R_{c,g}(\mathbf{P})$ is not achievable as it may involve transmitting at a rate that is not supported by the true CSI $\{\mathbf{h}_{u,g}[k] : \forall k\}$. The ergodic rate, $\mathbb{E}_{\mathbf{h}, \hat{\mathbf{h}}}[R_{c,g}(\mathbf{P})]$, on the other hand is achievable, where the averaging is over the joint distribution of the true CSI and the CSIT [23] (subscripts omitted for simplicity). Since this joint distribution is unknown in practice, the ergodic rate is not a tractable metric for designing $\mathbf{P}$. Hence, despite not being technically achievable, $R_{c,g}(\mathbf{P})$ [and similarly, $R_{p,g}(\mathbf{P})$ in (5.6)] is still useful² for designing $\mathbf{P}$ [see OP_{mmf} in (5.8)-(5.9)]. For such $\mathbf{P}$, the empirically achievable rates in our experiments are determined by a suitable choice of modulation and coding scheme (MCS) levels.*

Since $s_c[\cdot]$ contains useful information for users in *both* groups, it follows that both groups must be able to decode $s_c[\cdot]$. Hence, subject to (5.4), the highest rate supporting error-free decoding of

²The ergodic rate, $\mathbb{E}_{\mathbf{h}, \hat{\mathbf{h}}}[R_{c,g}(\mathbf{P})]$, can be expressed as $\mathbb{E}_{\hat{\mathbf{h}}}[\mathbb{E}_{\mathbf{h}|\hat{\mathbf{h}}}[R_{c,g}(\mathbf{P})]]$, where the outer expectation is w.r.t the distribution of the CSIT, and the inner expectation is w.r.t the conditional distribution of the true CSI, given the CSIT. As an alternative to the *instantaneous* rate $R_{c,g}(\mathbf{P})$, the term $\mathbb{E}_{\mathbf{h}|\hat{\mathbf{h}}}[R_{c,g}(\mathbf{P})]$ – known as the *average rate* (and in general, not achievable) – can also be used to design $\mathbf{P}$ as a function of the CSIT [23]. However, like with the ergodic rate, the average rate is also not a tractable metric in practice, since the conditional distribution is unknown and obtaining an empirical estimate at the TX involves considerable overhead.

$s_c[\cdot]$ by both groups – denoted by $R_c(\mathbf{P})$ and referred to as the *common stream rate* – is given by:

$$R_c(\mathbf{P}) = \min_{g \in \{1,2\}} R_{c,g}(\mathbf{P}). \quad (5.5)$$

Assuming perfect SIC (in other words, assuming $R_c(\mathbf{P})$ is achievable for all users), let $R_{p,g}(\mathbf{P})$ denote the highest (instantaneous) rate supporting error-free decoding of $s_g[\cdot]$ among the users in group g , again under the assumption that (5.1) constitutes perfect CSIT. This is referred to as *group g 's private stream rate* and has an expression similar to (5.4), given by:

$$R_{p,g}(\mathbf{P}) = \min_u \log_2 \left(1 + \frac{|\hat{\mathbf{h}}_{u,g}^H \mathbf{P}_g|^2}{\sigma^2 + |\hat{\mathbf{h}}_{u,g}^H \mathbf{P}_{g'}|^2} \right) \quad (5.6)$$

$(g \in \{1,2\}, g' \neq g).$

Subject to (5.5) and (5.6), the highest rate supporting error-free recovery of W_g by all users in group g – denoted by $R_g(\mathbf{P})$ and referred to as the *net group g rate* – is given by :

$$R_g(\mathbf{P}) = \frac{|W_{c,g}|}{|W_c|} R_c(\mathbf{P}) + R_{p,g}(\mathbf{P}) \quad (g \in \{1,2\}). \quad (5.7)$$

where the first term is the fraction of the common message (W_c) intended for group g .

In general, the sizes of $W_{c,1}$ and $W_{c,2}$ (in bits) in Fig. 5.1 need not be equal, which can be used to enable RSMA-based MGM to achieve better fairness. To illustrate this, suppose group 2's private stream rate is smaller than that of group 1 [i.e., $R_{p,2}(\mathbf{P}) < R_{p,1}(\mathbf{P})$] because group 2 users have weaker channels than group 1 users. In such a scenario, allocating a larger fraction of the common stream to the weaker group 2 (by having $|W_{c,2}| > |W_{c,1}|$) pushes the net group rates – $R_1(\mathbf{P})$ and $R_2(\mathbf{P})$ – towards equality, as captured by the right-hand side of (5.7). This motivates the following scheme:

- S1) if $|R_{p,1}(\mathbf{P}) - R_{p,2}(\mathbf{P})| \leq R_c(\mathbf{P})$, then the common stream is allocated by adjusting the sizes of $W_{c,1}$ and $W_{c,2}$ such that:

$$R_{p,1}(\mathbf{P}) + \frac{|W_{c,1}|}{|W_c|} R_c(\mathbf{P}) = R_{p,2}(\mathbf{P}) + \frac{|W_{c,2}|}{|W_c|} R_c(\mathbf{P}),$$

to ensure that both groups have the same throughput (i.e., achieving max-min fairness);

- S2) otherwise, when $|R_{p,1}(\mathbf{P}) - R_{p,2}(\mathbf{P})| > R_c(\mathbf{P})$, the entire common stream is allocated to the group with the smaller private stream rate.

The design of $\mathbf{P}$ to realize max-min rate fairness across the two groups yields the following optimization problem:

$$OP_{\text{mmf}} : \max_{\mathbf{P}} \min\{R_1(\mathbf{P}), R_2(\mathbf{P})\} \quad (5.8)$$

$$\text{s.t. } \text{tr}(\mathbf{P}\mathbf{P}^H) \leq P_t, \quad (5.9)$$

where (5.9) is the TX power constraint. Due to the non-convexity of OP_{mmf} , a tractable algorithm that converges to its global optimal solution does not exist. However, a locally optimal solution that is widely recognized as the gold-standard can be obtained using the *WMMSE method* [17], which takes advantage of the well-known relationship between the achievable rate and the post-equalization symbol mean square error [119]. Let $\mathbf{P}^{\text{wm}} = [\mathbf{p}_c^{\text{wm}}, \mathbf{p}_1^{\text{wm}}, \mathbf{p}_2^{\text{wm}}]$ denote the solution to OP_{mmf} obtained by the WMMSE method. However, the resulting max-min fair rate is difficult to realize in practice because, in addition to Remark 16:

1. Data transmission is typically restricted to a finite collection of MCS levels, which caps the achievable rate/throughput, and
2. Error-free decoding of the streams at any user cannot be guaranteed due to a combination of finite block length effects and CSI estimation errors, which further reduces the measured throughput.

To account for imperfect CSIT, discrete MCS levels and decoding errors, we consider the *MCS-limited* throughput as the metric of interest throughout this chapter. An MCS level is defined by a pair (m, r) , where positive integer m denotes the bits per constellation symbol (e.g., 2 for QPSK) and $r \in (0, 1]$ denotes the code rate. Let $\mathcal{M} = \{(m_c, r_c), (m_1, r_1), (m_2, r_2)\}$ denote the set of MCS levels chosen for $s_c[\cdot]$, $s_1[\cdot]$ and $s_2[\cdot]$, respectively. Then, the corresponding *MCS-limited* throughputs (measured in bits/s) with WMMSE precoders are given by:

$$T_c^{\text{mcs}}(\mathbf{P}^{\text{wm}}, \mathcal{M}) = Bm_cr_c \times \mathbb{P}(\hat{W}_c = W_c), \quad (5.10)$$

$$T_{p,1}^{\text{mcs}}(\mathbf{P}^{\text{wm}}, \mathcal{M}) = Bm_1r_1 \times \mathbb{P}(\hat{W}_{p,1}^{(1)} = \hat{W}_{p,1}^{(2)} = W_{p,1}),$$

$$T_{p,2}^{\text{mcs}}(\mathbf{P}^{\text{wm}}, \mathcal{M}) = Bm_2r_2 \times \mathbb{P}(\hat{W}_{p,2}^{(1)} = \hat{W}_{p,2}^{(2)} = W_{p,2}).$$

In the above expressions, B denotes the effective bandwidth³, and the probability terms capture the loss in throughput due to decoding errors – specifically, the first expression denotes the probability

³The effective bandwidth is the portion of the total bandwidth available for data transferring, after accounting for signalling overhead (e.g., the cyclic prefix in OFDM, pilot subcarriers and guard band in IEEE 802.11 standard)

that W_c is correctly decoded by all groups, while the second and third expressions denote the probability that $W_{p,1}$ and $W_{p,2}$ are correctly decoded by groups 1 and 2, respectively.

Remark 17 *The message decoding probabilities in the right-hand side of (5.10) are a function of the chosen MCS level, as well as the SINR at each user, which in turn, depends on $\mathbf{P}^{\text{wm}}$. This functional dependence is not shown explicitly to avoid messy notation.*

Similar to (5.7), the *MCS-limited* throughput for group g with WMMSE precoders is given by:

$$T_g^{\text{mcs}}(\mathbf{P}^{\text{wm}}, \mathcal{M}) = \frac{|W_{c,g}|}{|W_c|} T_c^{\text{mcs}}(\mathbf{P}^{\text{wm}}, \mathcal{M}) + T_{p,g}^{\text{mcs}}(\mathbf{P}^{\text{wm}}, \mathcal{M}) \quad (5.11)$$

Thus, similar to OP_{mmf} in (5.8)-(5.9), the *MCS-limited* max-min throughput fairness problem for RSMA-based MGM, which is of considerable practical relevance, can be defined as follows:

$$OP_{\text{mmf}}^{\text{mcs}} : \max_{\mathcal{M}} \min\{T_1^{\text{mcs}}(\mathbf{P}^{\text{wm}}, \mathcal{M}), T_2^{\text{mcs}}(\mathbf{P}^{\text{wm}}, \mathcal{M})\} \quad (5.12)$$

$$\text{s.t. } \mathcal{M} \in \mathbb{M}, \quad (5.13)$$

where $\mathbb{M}$ denotes the collection of permissible MCS levels for the three streams, which is typically pre-determined through standards (see Table 3.3 for the $\mathbb{M}$ used in our measurements). The optimal $\mathcal{M}$ for $OP_{\text{mmf}}^{\text{mcs}}$ is a function of $\mathbf{P}^{\text{wm}}$ that is difficult to characterize because the message decoding probabilities in (5.10) do not have closed-form expressions in terms of $\mathbf{P}^{\text{wm}}$ and $\mathcal{M}$. However, these probabilities can be empirically evaluated through measurements. This motivates us to *empirically solve* $OP_{\text{mmf}}^{\text{mcs}}$ through a brute force search⁴ over $\mathbb{M}$, which is the focus of our measurements in Section 5.4.

We conclude this subsection by noting SDMA- and NOMA-based MGMs are special cases of RSMA-based MGM, corresponding to specific choices of message splitting, as explained in the following remarks.

Remark 18 *SDMA-based MGM is a special case of RSMA-based MGM, where the common stream is turned off – i.e., $W_g = W_{p,g}$, ($g = 1, 2$) in Fig. 5.1. The corresponding expressions in (5.2)-(5.13) for SDMA-based MGM can be obtained by setting $\mathbf{p}_c = \mathbf{0}$. Essentially, without a common stream, group g users treat the interference from $W_{g'}$ ($g' \neq g$) as noise.*

⁴A more sophisticated empirical approach involves link adaptation, where the most suitable MCS level is determined by an ARQ-based mechanism [112]. This is left for future work.

Remark 19 *NOMA-based MGM is a special case of RSMA-based MGM, where the private stream message of one group (group 2, say) is turned off (i.e., $W_{p,2} = \emptyset$), and no portion of the other message is allocated to the common stream (i.e., $W_1 = W_{p,1}, W_2 = W_{c,2} = W_c$ in Fig. 5.1). The corresponding expressions in (5.2)-(5.13) can be obtained by setting $\mathbf{p}_2 = \mathbf{0}$. For this example, group 1 users fully decode and subtract the interference from W_2 , whereas group 2 users treat the interference from W_1 as noise. In contrast, for RSMA-based MGM, group g users partially decode the interference from $W_{g' \neq g}$ ($W_{c,g'}$) and partially treat it as noise ($W_{p,g'}$).*

Remark 20 *The common stream allocation, as per condition S2 on page 4, is similar to NOMA-based MGM, albeit the weaker group also has a private stream.*

5.3 RSMA Prototype

5.3.1 Hardware setup

We implement RSMA-based MGM by using our Software Defined Radio (SDR) based RSMA prototype [15]. The TX and RXs are realized using National Instruments' (NI) USRP 2942 SDR units, which have two antennas/RF chains. Hence, we use three USRP 2942 units, one to realize a two-antenna TX ($N_t = 2$), and the other two to realize four single-antenna users that are assigned to two groups of two users each in our measurements in Section 5.4. In particular, the antennas of group 2 users are connected to their USRP through coaxial cables, which allow them to be moved around to realize different channel environments, as described in Section 5.4.1. The USRPs share a common timing source (CDA-2990), and are controlled by a workstation running LabVIEW NXG, through which the various blocks in Fig. 5.1 are realized. All connections (SDRs to workstation, SDRs to timing sources) are through PCIe cables, facilitated by a PCIe bus (CPS-8910). A list of hardware components is provided in Table 5.1.

5.3.2 RSMA-based MGM Implementation

We adopt several features of the IEEE 802.11g physical layer frames to implement the system model described in Section 5.2.

Stage 1 Each TX antenna transmits a pilot signal orthogonally in time comprising a Short Training Field (STF, $8\mu\text{s}$ in duration) and a Long Training Field (LTF, $8\mu\text{s}$ in duration), as shown

	Name	Description
1.	Workstation	Running LabVIEW NXG
2.	NI USRP-2942 (3 units)	SDRs used to realize TX and users
3.	NI CPS-8910	Provides additional PCIe ports
4.	NI CDA-2990	8 Channel, 10 MHz clock distribution device
5.	8dBi RP-SMA Male Wifi Antenna	TX antennas
6.	RP-SMA Male Wifi Antenna	Group 1 users' antennas
7.	Mini-Circuits ZAPD-2-272-S+	Power splitter
8.	TP-Link TL-ANT2405C	Group 2 users' antennas

Table 5.1: List of hardware components.

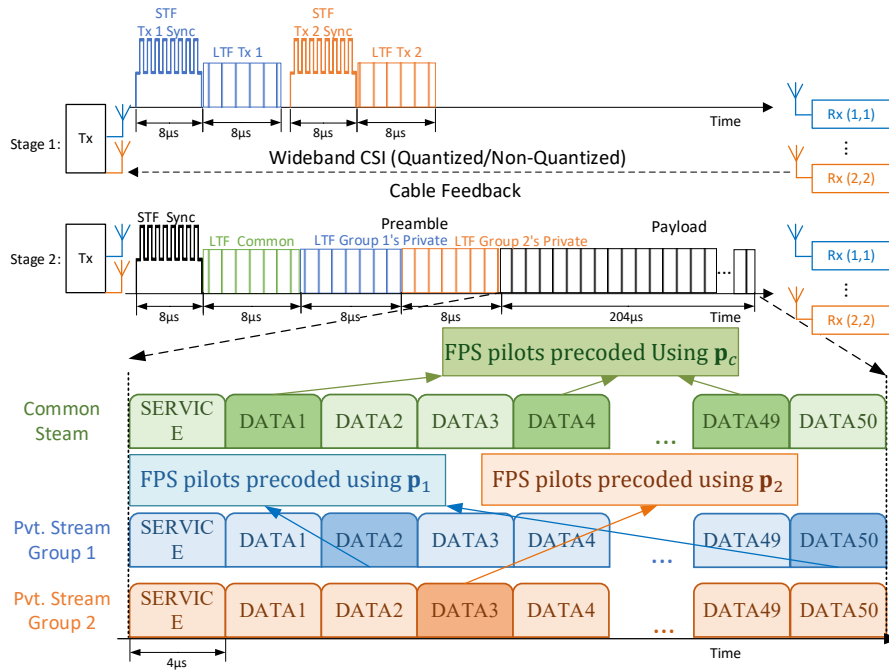


Figure 5.2: Signal structure within the two-stage transmission protocol used to implement RSMA-based MGM. In conventional (SDMA-based) 802.11g, every DATA (OFDM) symbol contains four *precoded* pilot subcarriers used to correct phase errors in the estimate of a user's precoded channel. This provides protection against decoding errors caused by rotated constellations post equalization, and is known as fine phase shift (FPS) [121]. However, in RSMA-based MGM, a user is associated with two precoders in general. Hence, we use $\mathbf{p}_c$ to precode the FPS pilots in DATA symbols 1, 4, 7 etc., $\mathbf{p}_1$ to precode the FPS pilots in DATA symbols 2, 5, 8 etc., and $\mathbf{p}_2$ to precode the FPS pilots in DATA symbols 3, 6, 9 etc. This is because the existence of the common/private stream precoders is channel-dependent, and the above scheme ensures that a user can correct phase errors at least once every 3 DATA symbols.

in the top portion of Fig. 5.2. The STF is used for synchronization and coarse frequency offset estimation, while the LTF is used for least squares based CSI estimation at each user (i.e., to obtain $\hat{\mathbf{h}}_{u,g}[k]$ in Section 5.2.1).

Stage 2 The transmitted signal consists of a preamble followed by the data payload, as shown in the bottom portion of Fig. 5.2.

- **Preamble:** The preamble consists of one STF and three LTFs. The function of the STF is the same as in Stage 1, while the LTFs are precoded in order to estimate the *precoded CSI* for equalization at the users. The first LTF is used by each (u, g) to estimate $\mathbf{h}_{u,g}^H \mathbf{p}_c$ for decoding the common stream. The second LTF is used by each user u in group 1 to estimate $\mathbf{h}_{u,1}^H \mathbf{p}_1$ to decode its private stream. Similarly, the third LTF is used by each user u' in group 2 to estimate $\mathbf{h}_{u',2}^H \mathbf{p}_2$ to decode its private stream.
- **Data Payload:** For the payload, we consider a total bandwidth of 20MHz with $N_c = 64$ subcarriers and a cyclic prefix (CP) of 16 samples per OFDM symbol. Aligned with IEEE 802.11 frames, 52 subcarriers are used for communications while the rest serve as guard bands. Among these 52 subcarriers, 48 are used to carry data symbols, with the remaining are pilot subcarriers for FPS, which are used to correct the common phase error across all subcarriers in one OFDM symbol [121]. This yields an effective bandwidth of 12 Mhz, based on (3.14). The payload consists of three superposed streams (one common, two private), each comprising 50 OFDM symbols.
- **MCS Implementation:** Table 3.3 lists the MCS levels, M , implemented in our prototype. For channel coding, we implement Polar codes augmented with an 8-bit cyclic redundancy check [122], [123], along with successive cancellation list decoding [124], with a list depth of 2. After the preamble, the first OFDM symbol (labelled SERVICE in the bottom portion of Fig. 5.2) contains the MCS information of each stream.

An instance of Stage 1 and Stage 2, as described above and illustrated in Fig. 5.2, constitutes a single measurement run. To empirically solve $OP_{\text{mmf}}^{\text{mcs}}$, we conduct 100 measurement runs. Let D_c denote the number of runs in which the common stream is successfully decoded by all users. Similarly, let D_g ($g \in \{1, 2\}$) denote the number of runs in which both users in group g successfully decode their private stream. Replacing the message decoding probabilities in (5.10) with their empirical estimates, the *measured* common and private stream throughputs for RSMA-based MGM

is given by:

$$\begin{aligned} T_c^{\text{mcs}}(\mathbf{P}^{\text{wm}}, \mathcal{M}) &= \frac{D_c}{100} B m_c r_c, \\ T_{p,1}^{\text{mcs}}(\mathbf{P}^{\text{wm}}, \mathcal{M}) &= \frac{D_1}{100} B m_1 r_1, \\ T_{p,2}^{\text{mcs}}(\mathbf{P}^{\text{wm}}, \mathcal{M}) &= \frac{D_2}{100} B m_2 r_2, \end{aligned} \quad (5.14)$$

where B is given by (3.14) and the MCS levels are chosen from Table 3.3.

Parameter	Notation	Value
Center frequency	f_c	2.484GHz ⁵
Transmit power	P_t	23dBm
No. of groups		2
No. of users per group		2
Total bandwidth		20MHz
Subcarriers	Total (N_c)	64
	Data	48
	Pilot (FPS)	4
	Guard band	12
CP length		16
Effective bandwidth	B	12MHz
OFDM symbols in payload		50
Experiment runs (per case)		100
TX antenna spacing		0.13m
Fraunhofer distance		0.28m
Distance between Group-1 RXs and TX		1.00m
Distance between Group-2 RXs and TX	Case 1-3	1.00-1.50m
	Case 4-6	2.00-2.30m
	Case 7-9	3.20-3.50m

Table 5.2: List of parameters used in our experiments.

Remark 21 (Message-splitting) *To demonstrate the gains of RSMA in overloaded MGM scenarios, it is sufficient to demonstrate that all users can decode and subtract the **entirety** of the common stream while treating the interference from both the private streams as noise. For this, it is sufficient and much simpler to implement the common and the two private streams as three separate data streams, rather than obtaining them from two data streams via message-splitting. Hence, in our implementation, we adopt this approach and retroactively assign a fraction of the measured*

common stream throughput to each groups, as per *S1* and *S2* on page 4 to mimic message-splitting at the TX. Of course, in a real-world implementation, message-splitting would need to take place.

5.4 Results

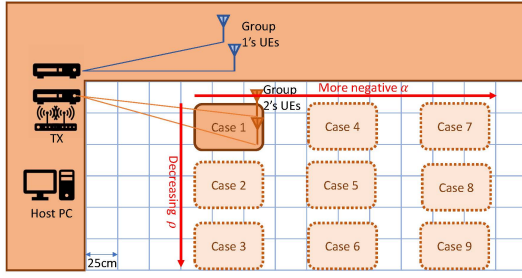
5.4.1 Measurement setup

In MGM, it is well known that a group's throughput is limited by its *weakest* user. Hence, in the existing literature, there is an assumption of admission control during the formation of the groups, so that users within a group have similar channel strength/link SNR. This motivates us to focus on the scenario where users within a group are closely situated, which could occur in applications like gaming, video conferencing, etc. In such cases, users within a group are likely to experience similar pathloss and large spatial correlations, whereas there could be considerable pathloss and spatial correlation differences across groups. Furthermore, as the relative performance of SDMA-, NOMA- and RSMA-based MGM is a function of the inter-group interference, it is important to realize a range of environments with varying levels of inter-group interference for a meaningful three-way comparison. This is intrinsically challenging for MGM, as notions of *weak/medium/strong* interference need to be defined over four pairs of links across the two groups instead of a single pair of links for two-user unicast. Motivated by our unicast measurements [15, Section IV], we rely on geometric notions of channel spatial correlation in line-of-sight environments as a predictor of inter-group interference levels. Thus, similar to [15], we consider nine measurement cases, as illustrated in Fig. 5.3a, where moving in the horizontal direction *should* increase the absolute value of inter-group pathloss difference, while moving in the vertical direction is *should* decrease the spatial correlation (and in turn, the interference levels) between the groups.

To quantify the inter-group pathloss difference between user u in group 1 and user u' in group 2, we define a parameter, $\alpha_{u,u'}$ (in dB scale) as follows:

$$\alpha_{u,u'} [\text{dB}] = 10 \log_{10} \frac{\|\hat{\mathbf{h}}_{u',2}\|}{\|\hat{\mathbf{h}}_{u,1}\|}, \quad (5.15)$$

where $\hat{\mathbf{h}}_{u,g}$ is given by (5.1). Since group 2 is farther from the TX than group 1, $\alpha_{u,u'}$ is non-positive and thus, a large negative value signifies greater pathloss difference among the two groups. Likewise, let $\rho_{u,u'}$ denote the spatial correlation between users u in group 1 and u' in group 2,



(a) The layout of the measurement environment, along with the TX and user positions for the nine cases. Intuitively, moving in the horizontal direction should increase the inter-group pathloss difference (i.e., α becoming more negative, as per (5.15)). Likewise, moving in the vertical direction should decrease the spatial correlation (i.e., ρ tending towards 0, as per (5.16)).



(b) The measurement environment (Case 4 is shown here).

Figure 5.3: Measurement setup

which is defined as follows:

$$\rho_{u,u'} = \frac{|\hat{\mathbf{h}}_{u,1}^H \hat{\mathbf{h}}_{u',2}|}{\|\hat{\mathbf{h}}_{u,1}\| \cdot \|\hat{\mathbf{h}}_{u',2}\|}. \quad (5.16)$$

Thus, $\rho_{u,u'} = 1$ signifies fully correlated (i.e., highly interfering) channels, while $\rho_{u,u'} = 0$ signifies orthogonal (i.e., zero interference) channels⁶. The expected trends for α and ρ are marked in Fig. 5.3a.

To realize Fig. 5.3a, group 1 users are fixed on a workbench 1m from the TX, while group 2 users are placed on a trolley (Fig. 5.3b) and moved around to realize the nine cases. The distance between the TX and group 2 is provided in Table 5.2. All users are in the far field. Table 5.3 captures the spread of the four α and ρ values for the nine cases. While the α spread is fairly narrow and follows the expected trend shown in Fig. 3a, the ρ spread is much larger for several cases because the intuition behind strong/medium/high levels of inter-group spatial correlation in Fig. 5.3a was based on only the (dominant) line-of-sight component, whereas the measured ρ is also sensitive to multipath, which is uncorrelated for each user. This highlights the difficulty of precisely controlling the extent of inter-group interference. Nevertheless, a shift towards 1 in the ρ spread consistent with increasing spatial correlation can be discerned when looking at the low and medium category of cases. No such insight can be drawn from the medium and high categories.

A full list of parameters used in our experiments is provided in Table 5.2.

⁶The intra-group spatial correlation can also be obtained from (5.16), when u and u' belong to the same group. However, we found this to be ≈ 1 throughout, consistent with closely-spaced (slightly larger than half wavelength) group member. Hence, it is not a useful parameter to explain performance differences across different cases.

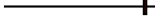
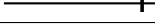
	Inter-group Pathloss Difference		Inter-group Spatial Correlation	
	Case	α spread $[-20,0]$	Case	ρ spread $[0,1]$
Low	1	 $[-0.39, -0.22]$	3	 $[0.18, 0.62]$
	2	 $[-1.18, -0.62]$	6	 $[0.17, 0.57]$
	3	 $[-1.63, -1.12]$	9	 $[0.15, 0.61]$
Medium	4	 $[-5.44, -4.12]$	2	 $[0.55, 0.76]$
	5	 $[-5.49, -4.09]$	5	 $[0.24, 0.76]$
	6	 $[-6.29, -5.02]$	8	 $[0.39, 0.88]$
High	7	 $[-11.07, -8.71]$	1	 $[0.44, 0.99]$
	8	 $[-18.96, -13.59]$	4	 $[0.43, 0.84]$
	9	 $[-18.99, -12.76]$	7	 $[0.42, 0.93]$

Table 5.3: The rectangles in the third and last columns capture the spread of the measured inter-group pathloss difference ($\alpha_{u,u'}$) and inter-group spatial correlation ($\rho_{u,u'}$). The left (right) end of each rectangle corresponds to the smallest (largest) value for each case. As expected, the α spread becomes more negative as the pathloss difference between the groups increases. But, the ρ spread is large for several cases due to sensitivity to multipath. Despite this, a shift towards 1 consistent with increasing spatial correlation can be discerned when looking at the low and medium categories of cases.

5.4.2 Fairness Comparison

To compare the fairness performance of RSMA-, NOMA- and SDMA-based MGM, Fig. 5.4 plots the sum throughput versus the minimum throughput for each case. The case number is indicated beside each data point. The black $y = 2x$ line represents max-min fairness, and the region above is feasible for all three multiple access schemes. Thus, points that are closer (in terms of Euclidean distance) to the black line represent *fairer* outcomes. Furthermore, a point that is either to the north, northeast or east of another point represents a *better* outcome. Based on these insights, we make the following observations:

- SDMA: From the spatial geometry of the user locations, cases 1, 4 and 7 have the highest amount of inter-group interference. For these cases, SDMA-based MGM achieves significantly lower minimum and sum throughput than NOMA- and RSMA-based MGM despite achieving near fairness. This is borne by the fact that the SDMA points for these cases lie to the southwest of the corresponding NOMA/RSMA points roughly along the $y = 2x$ line. This validates the well-known limitation of treating interference as noise in the high interference regime.

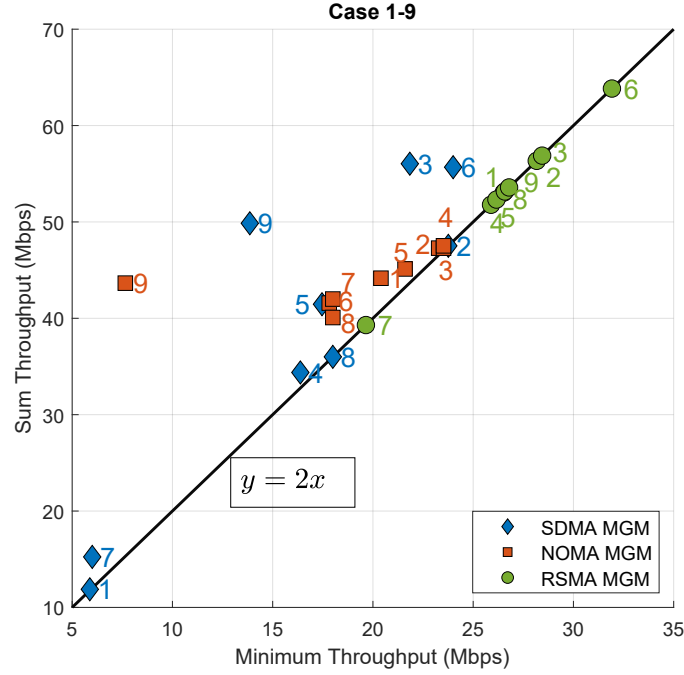


Figure 5.4: Fairness comparison between SDMA-based, NOMA-based and RSMA-based MGM. The number beside each data point indicates the measurement case. The black line ($y = 2x$) corresponds to max-min fairness, and points that are closer (in terms of Euclidean distance) to this line represent fairer outcomes.

- NOMA: With the exception of Case 9, NOMA-based MGM either achieves or comes close to achieving fairness for the other cases. It also achieves better outcomes than SDMA for cases 1, 4, 5, 7 and 8. Interestingly, it achieves strictly *worse* outcomes than SDMA for cases 6 and 9, despite SDMA's limitations on interference suppression in an overloaded scenario. It's worth noting that these cases are associated with the lowest inter-group interference (due to the relatively lower inter-group spatial correlation). This suggests that even for overloaded scenarios where interference cannot be fully suppressed to noise levels, a flexible mix of (partially) decoding the interference and (partially) treating it as noise, in response to the channel conditions is the ideal interference strategy, something RSMA achieves by design.
- RSMA: In Section 5.2, when condition S1 (see page 4) is true, the corresponding RSMA-based MGM point lies on the $y = 2x$ (i.e., achieves max-min fairness). Otherwise, for condition S2, the RSMA point lies *as close as possible* above the black line. For all cases, RSMA achieves max-min fairness, thereby implying that S1 is satisfied for each case (more on this while discussing Fig. 5.5). Furthermore, compared to SDMA, RSMA achieves strictly better performance for all cases. Likewise, when compared to NOMA, RSMA achieves strictly better performance for all cases bar one – the exception being case 7, where NOMA achieves a higher sum throughput than RSMA but not fairness (see Section 5.4.3-c for an explanation). To

explain RSMA's superior performance w.r.t SDMA and NOMA, it's important to gain some insight into the relative contributions of the common and private streams to each group's net throughput for each case. We will elaborate on this next.

Complementing Fig. 5.4, Fig. 5.5 plots the throughput of each group. For RSMA-based MGM, a group's throughput is broken down into common and private stream contributions. Table 5.4 lists the MCS levels and the block error rate (BLER) associated with each bar in Fig. 5.5. The arrows indicate the gain in the minimum throughput due to RSMA-based MGM. We make the following observations:

- For SDMA- and NOMA-based MGM, max-min fair precoding allocates more power to the weaker group's precoder in an attempt to equalize the post-precoding SINR at both groups. However, the post-precoding SINR can still be higher at the stronger group, depending on the extent of the inter-group pathloss difference. This SINR mismatch, depending on where it occurs w.r.t an MCS level's SINR boundaries, may permit a higher MCS level to be supported at the stronger group. This explains why fairness is not achieved in some cases for SDMA (3, 5, 7 and 9) and NOMA (6, 7, 8 and 9)⁷ – see Table 5.4.
- For RSMA-based MGM, in each subfigure, the contribution of the common stream to each group's throughput decreases from left to right across the three cases. This pattern was observed for unicast communications, as well [15]. Essentially, when the inter-group interference is high (as in cases 1, 4 and 7), most of the power is allocated to the common stream, which, in turn, contributes substantially to the net throughput. In other words, decoding most of the interference is favoured over suppressing it to noise levels. On the other hand, when the inter-group interference is lower (as in cases 3, 6 and 9), suppressing most of the interference to noise levels is preferred and hence, the power allocation to the private streams increases. As a result, the private streams contribute more to the net throughput.
- For RSMA-based MGM, there is a non-zero common stream contribution to each group's throughput in every case. Thus, S1 (see page 4) is satisfied for each case, and the allocation of the common stream to each group is done to ensure that both groups have the same net throughput. In other words, throughput contributions from both the common and the private streams for each group provides the required flexibility in terms of resource allocation to ensure that fairness is realized for each case.

⁷For the remaining SDMA and NOMA cases, the same MCS level is supported at both groups. The throughput variations for these cases (and any resulting lack of fairness) are due to block errors, which are difficult to control.

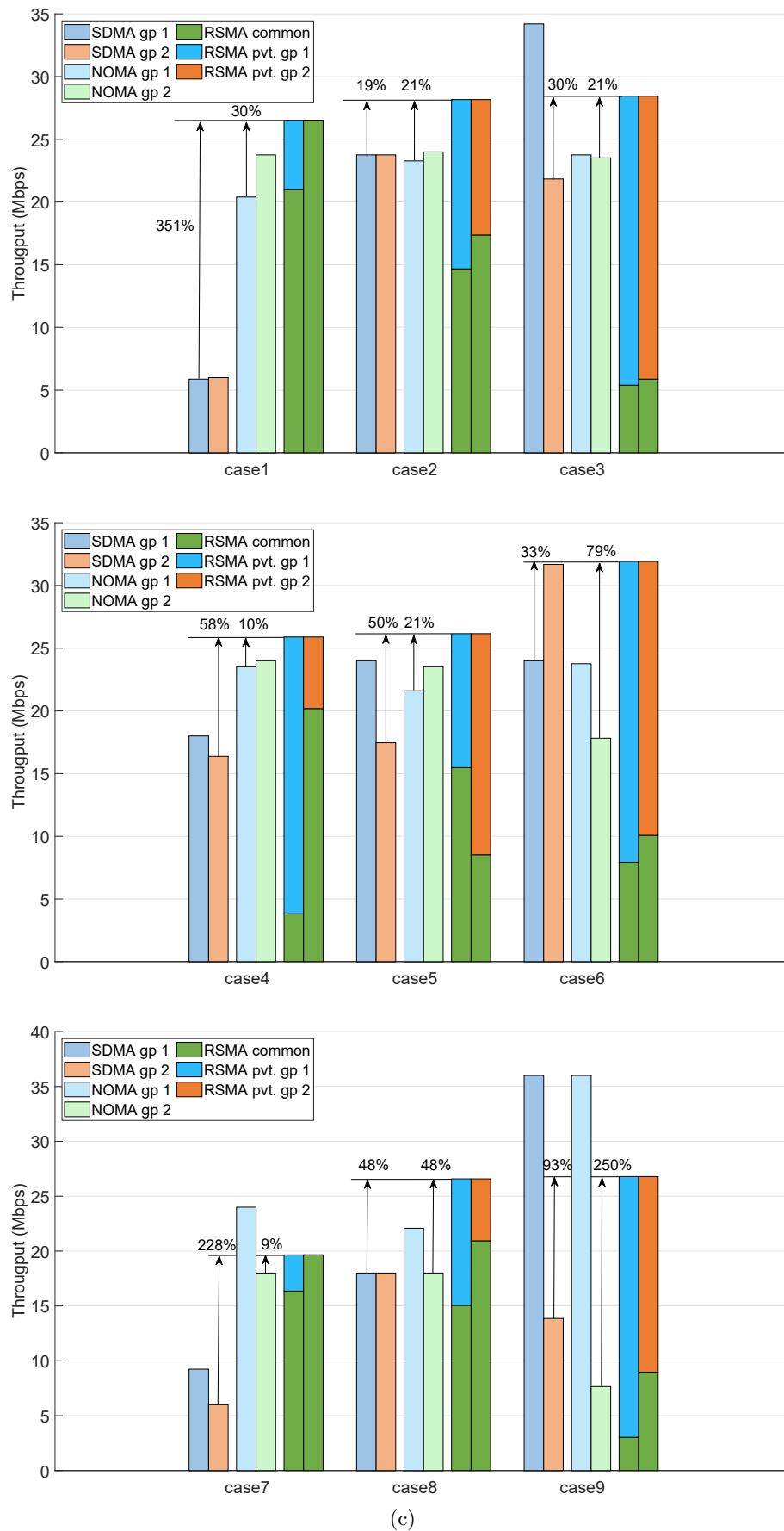


Figure 5.5: The throughput performance for both multicast groups of SDMA-, NOMA- and RSMA-based MGM.

Case	BLER (MCS index from Table 3.3)							
	SDMA		NOMA		RSMA			
	Group 1	Group 2	Group 1	Group 2	Common	Group 1 pvt.	Group 2 pvt.	
1	0.02 (0)	0 (0)	0.15 (4)	0.01 (4)	0.01 (6)	0.08 (0)	– –	
2	0.01 (4)	0.01 (4)	0.03 (4)	0 (4)	0.10 (5)	0.25 (3)	0.10 (2)	
3	0.05 (5)	0.09 (4)	0.01 (4)	0.02 (4)	0.06 (2)	0.04 (4)	0.06 (4)	
4	0 (3)	0.09 (3)	0.02 (4)	0 (4)	0 (4)	0.08 (4)	0.05 (0)	
5	0 (4)	0.03 (3)	0.11 (4)	0.02 (4)	0 (4)	0.11 (2)	0.02 (3)	
6	0 (4)	0.12 (5)	0.01 (4)	0.01 (3)	0 (3)	0 (4)	0.09 (4)	
7	0.23 (2)	0 (0)	0 (4)	0 (3)	0 (5)	0.45 (0)	– –	
8	0 (3)	0 (3)	0.08 (4)	0 (3)	0 (5)	0.04 (2)	0.06 (0)	
9	0 (5)	0.23 (3)	0 (5)	0.15 (1)	0 (2)	0.01 (4)	0.01 (3)	

Table 5.4: The MCS Index and the BLER corresponding to each throughput bar in Fig. 5.5.

- To further highlight the benefits of throughput contributions from both the common and the private streams for each group, consider cases 1 and 7 for RSMA-based MGM. Since the private stream contribution to (the weaker) group 2's throughput is zero, it appears at first glance as though RSMA-based MGM reduces to NOMA-based MGM. But there is a key difference. Unlike NOMA, the common stream can contribute to (the stronger) group 1's throughput as well, to the extent that most of group 1's throughput is due to the common stream. This is intuitive, since under high inter-group interference, most of the power is allocated to the common stream. In contrast, for NOMA-based MGM, group 1 needs to rely entirely on its private stream for its throughput. This forces a suboptimal resource allocation, where power has to be diverted from the common stream to group 1's private stream to achieve fairness. As a result, NOMA-based MGM achieves a lower minimum throughput than RSMA-based MGM for these cases. The flexibility offered by throughput contributions from both the common and the private streams for each group is also why RSMA-based MGM achieves better outcomes than both SDMA- and NOMA-based MGM.

In summary, RSMA-based MGM adapts well to variations in pathloss difference and interference between the groups and achieves superior fairness performance (i.e., fairness at a higher minimum throughput) when compared to SDMA- and NOMA-based MGM. We conclude this section by remarking on some seemingly counter-intuitive results in Fig. 5.5 and Table 5.4.

5.4.3 Counter-intuitive Results

SDMA, Case 6 Perhaps the most counter-intuitive observation in Fig. 5.5 is the supposedly weaker group 2 having a higher throughput (31.68Mbps) than the stronger group 1 (24Mbps) for SDMA-based MGM in Case 6. From Table 5.4, the MCS levels and the block error rate (BLER) corresponding to this particular case are:

- Group 1: 16QAM, rate 1/2 (level 4 in Table 3.3) with 0 BLER
- Group 2: 16QAM, rate 3/4 (level 5 in Table 3.3) with 0.12 BLER

It is clear from above that group 2's higher throughput stems from its ability to support a higher MCS level than group 1, albeit it is only one level higher and not error-free. The higher MCS level is due to the fact that more power is allocated to the group 2 precoder to achieve fairness. Ideally, under perfect CSIT, one would expect more power to be allocated to the group 2 precoder in such a way that both groups support the *same* MCS level, thereby achieving the same throughput. However, with imperfect CSIT, which is inherent in the experiments due to the use of wideband CSI in (5.1), we reckon it is not unusual for group 2 to support an MCS level one higher than group 1. Moreover, if we had imposed strict reliability requirements (i.e., zero BLER in the measurements), then both groups would have supported the same MCS level (16QAM, rate 1/2) and achieved the same throughput.

NOMA, Cases 1, 2, 4 and 5 Again, group 2 has a higher throughput than group 1 for these cases, although the difference is not as stark as above. This is because both groups support the same MCS level (see Table 5.4) which is expected, but the throughput variations are due to block errors that are difficult to control.

NOMA v/s RSMA, Case 7 For NOMA-based MGM, groups 1 and 2 supports MCS levels 4 and 3, respectively (both with zero BLER, see Table 5.4), and the resulting sum throughput is 42Mbps. Due to the high pathloss difference and the high inter-group interference associated with this case, RSMA-based MGM reduces to NOMA-based MGM, as evidenced by the lack of a private stream for group 2 (see Remark 19). But, unlike NOMA, the RSMA common stream can contribute to the throughput of *both* groups. Hence, nearly all of the transmit power is allocated to the common stream precoder. Consequently, the common stream alone should be able to support an MCS level whose throughput is at least 42Mbps (i.e., the NOMA sum throughput), which

would then be equally split between the two groups to achieve fairness. However, from Table 3.3, we see that there is no MCS level close to 42Mbps. Hence, in Table 5.4, we see that the common stream supports MCS level 5 (36Mbps) which is the nearest level lower than 42Mbps, but not MCS level 6 (48Mbps) which is the nearest level above 42Mbps. This is the reason why the NOMA sum throughput exceeds the RSMA sum throughput. However, even with conservative MCS level 5 for the common stream, RSMA-based MGM achieves a higher minimum throughput than NOMA-based MGM, which is the desired objective.

5.5 Summary

In this chapter, we presented the first-ever experimental evaluation of the fairness performance of RSMA-, SDMA- and NOMA-based MGM. We focused on the overloaded scenario with closely located group members, which is relevant for many applications. For two groups with two users per group, we realized nine cases that captured varying levels of interference and pathloss difference between the two groups. Over these nine cases, we observed that RSMA-based MGM achieved superior fairness performance (i.e., fairness at a higher minimum throughput) than SDMA- and NOMA-based MGM. This is consistent with theoretical predictions. The secret behind these gains stems from the fact that each group's throughput could, in general, have contributions from *both* the common and the private streams with the exact amounts of each dictated by the channel conditions.

6

Rate-Splitting Multiple Access for Integrated Sensing and Communications: A First Experimental Study

6

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6.1 Introduction

This chapter studies ISAC, a key feature in next-generation wireless systems, where a multi-antenna transmitter serves multiple users while simultaneously sensing nearby targets. The ISAC precoder design problem optimizes the power distribution between communication and sensing, which has been addressed under various settings, such as hybrid beamforming [82], [83], millimeter-wave ISAC [80]–[82], dedicated sensing signals [71]–[73], and reusing communication signals [74], [80], [81]. Two fundamental questions arise: (Q1) Can alternatives to SDMA, like RSMA, provide better ISAC performance? (Q2) Are separate signals needed for sensing and communication? Recent studies show RSMA outperforms SDMA under both perfect and imperfect CSIT, even when dedicated sensing signals are used [125]. For tasks like target ranging, communication signals can achieve similar performance to dedicated sensing signals asymptotically with large signal lengths (block length = 1024 symbols) [87]. These findings motivate a four-way ISAC comparison—RSMA/SDMA with/without dedicated sensing signals—where RSMA without a dedicated sensing signal consistently achieves the best performance [21], [91]. However, a comprehensive experimental validation of this comparison is still missing in the literature. In this chapter, we address this void, and the contributions are as follows:

- We formulate a signal model for RSMA ISAC with a dedicated sensing signal, which includes the other design choices (SDMA ISAC, RSMA ISAC without a dedicated sensing signal) as special cases. We then use the sum throughput (comms metric) and radar SNR (sensing metric) to define an *ISAC performance region* to compare the different design choices.
- Using software-defined radios (SDRs), we implement the above signal model using OFDM signals largely based on IEEE 802.11ac-VHT physical layer frames [115]. We consider a signal bandwidth of 100MHz, yielding a range resolution (bin size) of 1.5m.
- In our measurements, we consider three scenarios that are representative of simplified *vehicular ISAC*, covering different levels of inter-user interference and separation/integration between sensing and communications. Over these scenarios, we observe that:
 - a) The SDMA ISAC performance boundary lies in the interior of the RSMA ISAC performance region. Hence, RSMA ISAC outperforms SDMA ISAC.
 - b) Moreover, the RSMA ISAC performance boundary is achieved *without a dedicated sensing signal*. In contrast, a dedicated sensing signal is needed to achieve the SDMA performance boundary.

- c) The gap between the SDMA and RSMA ISAC performance regions is scenario dependent.

The rest of this chapter is organized as follows. In Section 6.2, we describe the signal model for RSMA ISAC with a dedicated sensing signal; the other design choices are special cases of this general case (see Section 6.2.3). Section 6.3 presents details of the SDR-based RSMA ISAC prototype, specifically the control signaling, OFDM payload structure, MCS levels etc. Section 6.4.1 motivates those measurement scenarios that are representative of vehicular ISAC. Capturing the peak performance of RSMA ISAC with a dedicated sensing signal involves a four-dimensional parameter search, which is cumbersome to realize in an experimental setting. This motivates us to pursue *simulation-aided parameter search*, described in Section 6.4.2, whereby we obtain a subset of well-chosen parameters that we use to measure and compare the ISAC performance of RSMA and SDMA, the results of which are presented in Section 6.4.3. Finally, Section 6.5 concludes this chapter.

6.2 System Model

Consider a TX with N_T antennas serving two-single antenna UEs and sensing (ranging) one target at its broadside as a monostatic radar in tracking mode. We consider OFDM signals over N_c subcarriers for both comms and sensing.

6.2.1 Transmit signal

RSMA comms For the two-UE case, RSMA comms involves the transmission of three linearly precoded signals. Specifically, let W_i denote the message meant for UE i ($= 1, 2$). At the TX, each W_i is split into common and private portions – denoted by $W_{c,i}$ and $W_{p,i}$, respectively. The common portions, $W_{c,1}$ and $W_{c,2}$, are combined into a common message, which is then encoded and modulated to form a *common stream*, $s_c[k]$ ($k = 0, \dots, N_c - 1$) over the subcarriers. Similarly, $W_{p,1}$ and $W_{p,2}$ are individually encoded and modulated to form *private streams*, $s_1[k]$ and $s_2[k]$, respectively. We assume zero-mean, unit-energy symbols for each stream. The three streams are then linearly precoded before transmission.

Remark 22 (SDMA as a special case of RSMA) *In the absence of message splitting (i.e., no s_c), RSMA reduces to SDMA.*

Sensing (Ranging) Ranging as a stand-alone functionality can be realized by transmitting a deterministic *sensing signal*, $s_r[k]$, aimed at the target using a precoder (beamformer) $\mathbf{p}_r$.

RSMA ISAC From a) and b) above, the RSMA ISAC transmit signal, $\mathbf{x}[k]$, can be modeled as follows:

$$\mathbf{x}[k] = \mathbf{p}_c[k]s_c[k] + \mathbf{p}_1[k]s_1[k] + \mathbf{p}_2[k]s_2[k] + \mathbf{p}_r[k]s_r[k], \quad (6.1)$$

where for subcarrier k , $\mathbf{p}_c[k]$ is referred to as the *common stream precoder*, $\mathbf{p}_i[k]$ the *private stream precoder* for UE i , and $\mathbf{p}_r[k]$ the *sensing precoder*. We assume a sum transmit power constraint, i.e.,

$$\sum_{k=0}^{N_c-1} (\|\mathbf{p}_c[k]\|^2 + \|\mathbf{p}_1[k]\|^2 + \|\mathbf{p}_2[k]\|^2 + \|\mathbf{p}_r[k]\|^2) = P_T \quad (6.2)$$

Remark 23 (Ranging using comms signals) To use the comms signals – $s_c[\cdot]$, $s_1[\cdot]$ and $s_2[\cdot]$ – for ranging [87], the respective precoders – $\mathbf{p}_c[\cdot]$, $\mathbf{p}_1[\cdot]$ and $\mathbf{p}_2[\cdot]$ – should strike a balance between radiating power towards the desired UE(s) and the target. This trade-off motivates the precoder design choices in the next subsection.

6.2.2 Precoder Design

Let $\mathbf{h}_i[k] \in \mathbb{C}^{N_T}$ denote the slowly varying, flat fading (comms) channel between the TX and UE i ($= 1, 2$). Through downlink pilot signals, we assume UE i obtains an estimate of $\mathbf{h}_i[k]$, denoted by $\hat{\mathbf{h}}_i[k]$, which is fed back to the TX. Let $\mathbf{u}_i[k] := \hat{\mathbf{h}}_i[k]/\|\hat{\mathbf{h}}_i[k]\|$ denote the unit vector along $\hat{\mathbf{h}}_i[k]$. Similarly, let $\mathbf{u}_0 := \mathbf{a}_0/\sqrt{N_T}$ denote the normalized steering vector along the TX's broadside, which is assumed to be .

Design of $\mathbf{p}_c[\cdot]$ In RSMA comms, each UE independently decodes $s_c[\cdot]$ first by treating the interference from $s_1[\cdot]$ and $s_2[\cdot]$ as noise. Hence, $\mathbf{p}_c[\cdot]$ must have sufficient gain at *both* UEs (for comms), as well as the target direction (for sensing, see Remark 23). This motivates a weighted

MRT precoder choice for $\mathbf{p}_c[\cdot]$, as follows:

$$\mathbf{p}_c[k] = \sqrt{P_T t_{\text{comms}}(1 - t_p)} \times \frac{(\sqrt{\alpha_c} \mathbf{u}_c[k] + \sqrt{1 - \alpha_c} \mathbf{u}_0)}{\sqrt{\sum_{k'} \|\sqrt{\alpha_c} \mathbf{u}_c[k'] + \sqrt{1 - \alpha_c} \mathbf{u}_0\|^2}} \quad (6.3)$$

$$\text{where } \mathbf{u}_c[k] = (\mathbf{u}_1[k] + \mathbf{u}_2[k]) / \|\mathbf{u}_1[k] + \mathbf{u}_2[k]\|. \quad (6.4)$$

In (6.4), through equi-weighted MRT along $\hat{\mathbf{h}}_1[k]$ and $\hat{\mathbf{h}}_2[k]$, the vector $\mathbf{u}_c[k]$ ensures some gain at both UEs, and the form of (6.3) further ensures some gain along the target direction ($\mathbf{u}_0$) as well. The parameter $\alpha_c \in [0, 1]$ captures the priority/weight assigned to comms. Finally, $t_{\text{comms}} \in [0, 1]$ captures the fraction of the transmit power allocated for comms, and $t_p \in [0, 1]$ represents the fraction of t_{comms} that is allocated to the private stream precoders.

Design of $\mathbf{p}_i[\cdot]$ ($i = 1, 2$) After decoding $s_c[\cdot]$, UE i subtracts its contribution (i.e., through successive interference cancellation) and then decodes $s_i[\cdot]$ by treating the interference from the other private stream as noise. Hence, $\mathbf{p}_i[\cdot]$ must have sufficient gain at UE i as well as the target direction. This motivates the following weighted MRT precoder choice for $\mathbf{p}_i[\cdot]$ similar to (6.3):

$$\mathbf{p}_i[k] = \sqrt{\frac{P_T t_{\text{comms}} t_p}{2}} \times \frac{(\sqrt{\alpha_p} \mathbf{u}_i[k] + \sqrt{1 - \alpha_p} \mathbf{u}_0)}{\sqrt{\sum_{k'} \|\sqrt{\alpha_p} \mathbf{u}_i[k'] + \sqrt{1 - \alpha_p} \mathbf{u}_0\|^2}}, \quad (6.5)$$

where like α_c in (6.3), $\alpha_p \in [0, 1]$ captures the priority/weight assigned to comms. In general, $\alpha_c \neq \alpha_p$. We further assume that the total private stream precoder power ($= t_{\text{comms}} t_p$) is equally divided between $\mathbf{p}_1[\cdot]$ and $\mathbf{p}_2[\cdot]$. This is a reasonably good choice when the UEs have similar channel strengths (i.e., $\|\hat{\mathbf{h}}_1\| \approx \|\hat{\mathbf{h}}_2\|$), which is the scenario we focus on in measurements.

Design of $\mathbf{p}_r[\cdot]$ Since $s_r[\cdot]$ is not an information-bearing signal, $\mathbf{p}_r[\cdot]$ needs to have high gain only along the target direction. From (6.3) and (6.5), $1 - t_{\text{comms}}$ is the fraction of the transmit power allocated solely for sensing. Hence,

$$\mathbf{p}_r[k] = \sqrt{\frac{P_T(1 - t_{\text{comms}})}{N_c}} \times \mathbf{u}_0. \quad (6.6)$$

Quantity	Notation	Description
Channels	$\mathbf{h}_i[k]$ ($i = 1, 2$)	Channel between the TX and UE i over subcarrier k
	$\hat{\mathbf{h}}_i[k]$	Estimate of $\mathbf{h}_i[k]$ (CSIT)
	$\mathbf{u}_i[k]$	Unit vector along $\hat{\mathbf{h}}_i[k]$
	$\mathbf{u}_0 := \mathbf{a}_0/\sqrt{N_T}$	Normalized steering vector along TX broadside
Precoders	$\mathbf{p}_i[k]$ ($i = 1, 2$)	Private stream precoder for UE i over subcarrier k
	$\mathbf{p}_c[k]$	Common stream precoder over subcarrier k
	$\mathbf{p}_r[k]$	Sensing precoder over subcarrier k
Signals	$s_i[k]$ ($i = 1, 2$)	Channel-coded and modulated private stream (comms) symbol for UE i over subcarrier k
	$s_c[k]$	Channel-coded and modulated common stream (comms) symbol over subcarrier k
	$s_r[k]$	Unit energy deterministic radar symbol over subcarrier k
Comms/sensing relative priorities	α_p ($1 - \alpha_p$)	Relative priority given to comms (sensing) for the private stream precoders ($\mathbf{p}_i$)
	α_c ($1 - \alpha_c$)	Relative priority given to comms (sensing) for the RSMA common stream precoder ($\mathbf{p}_c$). In general, $\alpha_c \neq \alpha_p$.
Power Allocation	$t_{\text{comms}} \in [0, 1]$	Fraction of power allocated to comms
	$1 - t_{\text{comms}}$	Fraction of power allocated to dedicated radar precoder ($\mathbf{p}_r$)
	$t_p \in [0, 1]$	Fraction of t_{comms} allocated to private stream precoders ($\mathbf{p}_i$)

Table 6.1: Summary of notation used in system model (Section 6.2).

Remark 24 Four parameters are used to capture the full extent of the comms-sensing trade-off in (6.3)-(6.6). In summary, they are:

- $t_{\text{comms}} \in [0, 1]$: Fraction of transmit power allocated for comms. $t_{\text{comms}} \rightarrow 0/1$ implies most of the power allocated for sensing/comms.
- $t_p \in [0, 1]$: Fraction of comms power (i.e., t_{comms}) allocated to the private streams allocated for comms. $t_p \rightarrow 0/1$ implies most of t_{comms} is allocated to the common/private stream(s).
- $\alpha_c \in [0, 1]$: As $\mathbf{p}_c[\cdot]$ can be used for both comms and sensing whenever it is allocated non-zero power, α_c is the priority given to comms. In particular, $\alpha_c \rightarrow 0/1$ implies very high priority given to sensing/comms.
- $\alpha_p \in [0, 1]$: Like α_c , α_p is the priority given to comms for the private stream precoders, $\mathbf{p}_1[\cdot]$ and $\mathbf{p}_2[\cdot]$. Again, $\alpha_p \rightarrow 0/1$ implies very high priority given to sensing/comms.

Remark 25 (Heuristic precoders) The precoders in (6.3)-(6.6) are sub-optimal, as they are not obtained by optimizing any performance metrics (e.g., sum throughput for comms, SNR of radar return for sensing). However, they are still useful because they are easy to implement, which in turn makes it relatively easier to obtain an **empirical ISAC performance region** (e.g., the achievable pairs of sum throughput and radar SNR – see Definition 1 in page 6 for a precise definition) to judge the comms/sensing trade-off associated with each design choice (i.e., RSMA/SDMA with/without dedicated signal). Nevertheless, the suboptimality of the precoders

means that the gap between the different ISAC performance region boundaries could potentially be enlarged through optimized precoders. This is left for future work.

Table 6.1 summarizes the notation used so far. Next, we focus on some special cases of (6.3)-(6.6) that are of interest.

6.2.3 Special Cases

RSMA ISAC without dedicated sensing signal – the general case

A key feature of (6.3)-(6.6) is the fact that comms precoders can be used for sensing, but not vice-versa, as $s_r[\cdot]$ offers no comms benefit. Hence, it is reasonable to question whether the power allocated to a dedicated sensing signal/precoder could be diverted to comms to significantly enhance comms performance, with minimal loss in sensing performance? To investigate this, t_{comms} can be set to 1 on (6.3)-(6.6) to yield an RSMA ISAC transmit signal involving three precoders, where each precoder is beneficial for both comms and sensing.

RSMA ISAC without dedicated sensing signal – a soft sensing/comms separation

While all precoders are beneficial for both comms and sensing in the previous special case, a balance can be struck wherein some of them prioritize sensing more than comms, while the others achieve the opposite effect. This can be realized, for instance, by setting $\alpha_c = 1 - \alpha_p$, where $\alpha_p \in [1/2, 1)$ (in addition to $t_{\text{comms}} = 1$). Thus, the private stream precoders prioritize comms, while the common stream precoder prioritizes sensing, similar to [90]. In this way, a soft separation of comms/sensing functionality for the precoders can be realized.

SDMA ISAC with dedicated sensing signal – the general case

With a dedicated sensing signal, $t_{\text{comms}} < 1$. Furthermore, from Remark 22, it follows that SDMA can be realized by setting $t_p = 1$.

SDMA ISAC with dedicated sensing signal – a hard sensing/comms separation

With a dedicated sensing signal/precoder, perhaps the comms precoder should *solely* prioritize comms to significantly enhance comms performance, with minimal loss in sensing performance?

This amounts to a *hard* comms/sensing separation among the precoders, in contrast to the soft separation from special case no. 2. This can be realized by setting $\alpha_p = 1$ (in addition to $t_{\text{comms}} < 1, t_p = 1$).

SDMA ISAC without dedicated sensing signal

Finally, this special case can be realized by setting $t_{\text{comms}} = 1$ and $t_p = 1$, giving rise to an SDMA ISAC transmit signal involving two precoders, each of which is beneficial for both comms and sensing.

Table 6.2 summarizes the special cases described above. Next, we introduce the performance metrics for comms and sensing performance and define the ISAC performance region to quantify ISAC performance.

6

6.2.4 Performance Metrics

Let $\mathbf{P} := \{\mathbf{p}_c[k], \mathbf{p}_1[k], \mathbf{p}_2[k], \mathbf{p}_r[k] : \forall k\}$ denote the collection of precoders.

Comms In the measurements, the comms performance metric is the *MCS-limited sum throughput*. An MCS level is characterized by a pair (m, r) , where positive integer m denotes the bits per constellation symbol (e.g., 1 for BPSK) and $r \in (0, 1]$ denotes the code rate.

Remark 26 (Necessary condition for decoding MCS level (m, r)) *An OFDM symbol stream over N_c subcarriers encoded using (m, r) can be successfully decoded at a UE **only if***

$$\frac{1}{N_c} \sum_{k=0}^{N_c-1} \log_2(1 + \text{SINR}[k]) \geq mr \quad (6.7)$$

where $\text{SINR}[k]$ denotes the signal-to-interference-plus-noise ratio of the symbol stream at subcarrier k at the UE.

The above inequality is a **necessary, but not sufficient condition** for successfully decoding (m, r) . This is because the left hand side represents the spectral efficiency (in bits/s/Hz) that can be achieved through Gaussian encoding, whereas the right hand side is the spectral efficiency achievable using (m, r) , which typically involves encoding over the binary field and modulation over discrete (non-Gaussian) constellations. Due to this mismatch, the minimum SNR needed to decode (m, r) is usually higher [126] than that given by (6.7).

Spl. case	Scheme	Parameters	Precoders	Comments
1.	RSMA ISAC w/o dedicated sensing signal – the general case	$t_{\text{comms}} = 1$ $\alpha_c \in [0, 1]$ $\alpha_p \in [0, 1]$ $t_p \in [0, 1]$	$\mathbf{p}_c[k] = \sqrt{P_T(1-t_p)} \times \frac{\sqrt{\alpha_c \mathbf{u}_c[k] + \sqrt{1-\alpha_c} \mathbf{u}_0}}{\sqrt{\sum_{k'} \ \sqrt{\alpha_c} \mathbf{u}_c[k'] + \sqrt{1-\alpha_c} \mathbf{u}_0\ ^2}}$ $\mathbf{p}_1[k] = \sqrt{P_T \frac{t_p}{2}} \times \frac{\sqrt{\alpha_p \mathbf{u}_1[k] + \sqrt{1-\alpha_p} \mathbf{u}_0}}{\sqrt{\sum_{k'} \ \sqrt{\alpha_p} \mathbf{u}_1[k'] + \sqrt{1-\alpha_p} \mathbf{u}_0\ ^2}}$ $\mathbf{p}_2[k] = \sqrt{P_T \frac{t_p}{2}} \times \frac{\sqrt{\alpha_p \mathbf{u}_2[k] + \sqrt{1-\alpha_p} \mathbf{u}_0}}{\sqrt{\sum_{k'} \ \sqrt{\alpha_p} \mathbf{u}_2[k'] + \sqrt{1-\alpha_p} \mathbf{u}_0\ ^2}}$	$\mathbf{u}_c[k] = \frac{\mathbf{u}_1[k] + \mathbf{u}_2[k]}{\ \mathbf{u}_1[k] + \mathbf{u}_2[k]\ }$. See Section 6.2.3. No dedicated radar signal/precoder. In general, common and pvt. streams have different priorities for comms/sensing (i.e., $\alpha_c \neq \alpha_p$).
2.	RSMA ISAC w/o dedicated sensing signal – a soft sensing/comms separation	$t_{\text{comms}} = 1$ $\alpha_c = 1 - \alpha_p$ $\alpha_p \in [1/2, 1]$ $t_p \in [0, 1]$	$\mathbf{p}_c[k] = \sqrt{P_T(1-t_p)} \times \frac{\sqrt{1-\alpha_p} \mathbf{u}_c[k] + \sqrt{\alpha_p} \mathbf{u}_0}{\sqrt{\sum_{k'} \ \sqrt{1-\alpha_p} \mathbf{u}_c[k'] + \sqrt{\alpha_p} \mathbf{u}_0\ ^2}}$ $\mathbf{p}_1[k] = \sqrt{P_T \frac{t_p}{2}} \times \frac{\sqrt{\alpha_p} \mathbf{u}_1[k] + \sqrt{1-\alpha_p} \mathbf{u}_0}{\sqrt{\sum_{k'} \ \sqrt{\alpha_p} \mathbf{u}_1[k'] + \sqrt{1-\alpha_p} \mathbf{u}_0\ ^2}}$ $\mathbf{p}_2[k] = \sqrt{P_T \frac{t_p}{2}} \times \frac{\sqrt{\alpha_p} \mathbf{u}_2[k] + \sqrt{1-\alpha_p} \mathbf{u}_0}{\sqrt{\sum_{k'} \ \sqrt{\alpha_p} \mathbf{u}_2[k'] + \sqrt{1-\alpha_p} \mathbf{u}_0\ ^2}}$	See Section 6.2.3. A special case of 1 above, where a soft sensing/comms separation is realized through the common (private) stream precoder(s) prioritizing sensing (comms).
3.	SDMA ISAC with dedicated sensing signal – the general case	$t_{\text{comms}} \in (0, 1)$ $t_p = 1$ $\alpha_p \in (0, 1)$	$\mathbf{p}_1[k] = \sqrt{P_T \frac{t_{\text{comms}}}{2}} \times \frac{\sqrt{\alpha_p} \mathbf{u}_1[k] + \sqrt{1-\alpha_p} \mathbf{u}_0}{\sqrt{\sum_{k'} \ \sqrt{\alpha_p} \mathbf{u}_1[k'] + \sqrt{1-\alpha_p} \mathbf{u}_0\ ^2}}$ $\mathbf{p}_2[k] = \sqrt{P_T \frac{t_{\text{comms}}}{2}} \times \frac{\sqrt{\alpha_p} \mathbf{u}_2[k] + \sqrt{1-\alpha_p} \mathbf{u}_0}{\sqrt{\sum_{k'} \ \sqrt{\alpha_p} \mathbf{u}_2[k'] + \sqrt{1-\alpha_p} \mathbf{u}_0\ ^2}}$ $\mathbf{p}_r[k] = \sqrt{P_T \frac{(1-t_{\text{comms}})}{N_c}} \times \mathbf{u}_0$	See Section 6.2.3. There exists a dedicated radar signal/precoder, but the comms precoders also have some priority for sensing.
4.	SDMA ISAC with dedicated sensing signal – a hard sensing/comms separation	$t_{\text{comms}} \in (0, 1)$ $t_p = 1$ $\alpha_p = 1$	$\mathbf{p}_1[k] = \sqrt{P_T \frac{t_{\text{comms}}}{2}} \times \mathbf{u}_1[k]$ $\mathbf{p}_2[k] = \sqrt{P_T \frac{t_{\text{comms}}}{2}} \times \mathbf{u}_2[k]$ $\mathbf{p}_r[k] = \sqrt{P_T \frac{(1-t_{\text{comms}})}{N_c}} \times \mathbf{u}_0$	See Section 6.2.3. A special case of 3 above, where the comms precoders do not prioritize sensing. Thus, a hard sensing/comms separation is realized.
5.	SDMA ISAC w/o dedicated sensing signal	$t_{\text{comms}} = 1$ $t_p = 1$ $\alpha_p \in (0, 1)$	$\mathbf{p}_1[k] = \sqrt{\frac{P_T}{2}} \times \frac{\sqrt{\alpha_p} \mathbf{u}_1[k] + \sqrt{1-\alpha_p} \mathbf{u}_0}{\sqrt{\sum_{k'} \ \sqrt{\alpha_p} \mathbf{u}_1[k'] + \sqrt{1-\alpha_p} \mathbf{u}_0\ ^2}}$ $\mathbf{p}_2[k] = \sqrt{\frac{P_T}{2}} \times \frac{\sqrt{\alpha_p} \mathbf{u}_2[k] + \sqrt{1-\alpha_p} \mathbf{u}_0}{\sqrt{\sum_{k'} \ \sqrt{\alpha_p} \mathbf{u}_2[k'] + \sqrt{1-\alpha_p} \mathbf{u}_0\ ^2}}$	See Section 6.2.3. No dedicated radar signal/precoder

Table 6.2: List of special cases described in Section 6.2.3.

Successful decoding of (m, r) yields a throughput (measured in bits/s) of Bmr , where B denotes the signal bandwidth (after accounting for signaling overheads, such as cyclic prefix).

The received signal, $y_i[k]$, at the i -th UE given by:

$$\begin{aligned} y_i[k] &= \mathbf{h}_i^H[k] \mathbf{x}[k] + n_i[k] \quad (i = 1, 2) \\ &= \mathbf{h}_i^H[k] \mathbf{p}_r[k] s_r[k] + \mathbf{h}_i^H[k] \mathbf{p}_c[k] s_c[k] \\ &\quad + \mathbf{h}_i^H[k] \mathbf{p}_1[k] s_1[k] + \mathbf{h}_i^H[k] \mathbf{p}_2[k] s_2[k] + n_i[k] \end{aligned} \quad (6.8)$$

We assume $s_r[\cdot]$ is known at the UEs and can be subtracted before decoding $s_c[\cdot]$, $s_1[\cdot]$ and $s_2[\cdot]$. Let $\mathcal{M} = \{(m_c, r_c), (m_1, r_1), (m_2, r_2)\}$ denote the MCS levels chosen for $(s_c[\cdot], s_1[\cdot], s_2[\cdot])$, respectively. After subtracting $s_r[\cdot]$, each UE independently decodes $s_c[\cdot]$, while treating the interference from $s_1[\cdot]$ and $s_2[\cdot]$ as noise. Thus, from Remark 26, the maximum MCS-limited common stream throughput, denoted by $T_c(\mathbf{P})$, is given by:

$$T_c(\mathbf{P}) = \max_{\mathcal{M} \in \mathbb{M}} Bm_c r_c \quad (6.9)$$

$$\text{s.t. } \min_i \frac{1}{N_c} \sum_{k=0}^{N_c-1} \log_2 (1 + \text{SINR}_{c,i}[k; \mathbf{P}]) > m_c r_c, \quad (6.10)$$

where (i) $\mathbb{M}$ denotes the collection of permissible MCS levels for the three streams, which is typically pre-determined through standards (see Table 6.4 for the $\mathbb{M}$ used in the measurements), (ii) $\text{SINR}_{c,i}[k; \mathbf{P}]$ denotes the common stream SINR at the k -th subcarrier at UE i , which is a function of $\mathbf{P}$ as follows:

$$\text{SINR}_{c,i}[k; \mathbf{P}] = \frac{|\mathbf{h}_i^H[k] \mathbf{p}_c[k]|^2}{\sum_{j=1}^2 |\mathbf{h}_i^H[k] \mathbf{p}_j[k]|^2 + \sigma^2}. \quad (6.11)$$

The minimum in (6.10) is due to both UEs needing to decode $s_c[\cdot]$. Furthermore, $T_c(\mathbf{P}) = 0$ when (6.10) is not satisfied even for the lowest MCS level.

After decoding $s_c[\cdot]$ and subtracting its contribution from $y_i[\cdot]$, UE i decodes $s_i[\cdot]$ while treating the interference from the other private stream ($s_j[\cdot]$, $j \neq i$) as noise. Similar to $T_c(\mathbf{P})$, the maximum MCS-limited private stream throughput at UE i , denoted by $T_i(\mathbf{P})$, is given by:

$$T_i(\mathbf{P}) := \max_{\mathcal{M} \in \mathbb{M}} Bm_i r_i \quad (6.12)$$

$$\text{s.t. } \frac{1}{N_c} \sum_{k=0}^{N_c-1} \log_2 (1 + \text{SINR}_{p,i}[k; \mathbf{P}]) > m_i r_i, \quad (6.13)$$

where $\text{SINR}_{p,i}[k; \mathbf{P}]$ denotes the private stream SINR at the k -th subcarrier for the i -th UE, which is a function of $\mathbf{P}$ as follows:

$$\text{SINR}_{p,i}[k; \mathbf{P}] = \frac{|\mathbf{h}_i[k]^H \mathbf{p}_i[k]|^2}{|\mathbf{h}_i[k]^H \mathbf{p}_j[k]|^2 + \sigma^2} \quad (j \neq i). \quad (6.14)$$

From (6.9)-(6.13), the MCS-limited sum throughput, $T_{\text{sum}}(\mathbf{P})$, has the following expression:

$$T_{\text{sum}}(\mathbf{P}) = T_c(\mathbf{P}) + \mathbb{1}(T_c(\mathbf{P}) > 0)(T_1(\mathbf{P}) + T_2(\mathbf{P})). \quad (6.15)$$

Remark 27 (Impact of Imperfect SIC and Finite MCS levels) *The indicator function in (6.15) captures the impact of imperfect SIC and finite MCS levels – i.e., if a UE cannot decode the common stream at the lowest MCS level, then it cannot decode its private stream at any MCS level, resulting in zero throughput. If at least one UE experiences such a **throughput collapse**, then RSMA is not viable (under the constraint of finite MCS levels), and we assume $T_{\text{sum}}(\mathbf{P}) = 0$. This phenomenon was experimentally observed in [127].*

It is worth noting that this is a marked departure from standard analytical models that assume perfect SIC and infinitely many MCS levels of arbitrary granularity (implicit when Gaussian encoding is assumed). Under these assumptions, the highest achievable RSMA sum throughput is never smaller than the highest achievable SDMA sum throughput [14]. However, with finite MCS levels, the SDMA sum throughput can exceed the RSMA sum throughput, especially when the common stream precoder is not allocated enough power to decode the lowest MCS level.

Remark 28 (SDMA Sum Throughput) *Since there is no common stream in SDMA (see Remark 22), the maximum MCS-limited SDMA sum throughput equals $T_1(\mathbf{P}) + T_2(\mathbf{P})$, where $\mathcal{M} = \{(m_1, r_1), (m_2, r_2)\}$.*

Remark 29 (Upper bound on measured throughputs) *Since (6.10) and (6.13) are necessary but not sufficient conditions for decoding non-Gaussian constellations (see Remark 26), $T_c(\mathbf{P})$, $T_i(\mathbf{P})$ ($i = 1, 2$) and $T_{\text{sum}}(\mathbf{P})$ are upper bounds for the measured common stream, private stream and sum throughputs, respectively.*

Ranging In measurements, performance metric for ranging is the *post-processing radar SNR*. We consider a single-antenna radar receiver (RX) colocated with the TX. For transmit signal $\mathbf{x}[k]$

given by (6.1), the signal at the radar RX, denoted by $y_r[k]$, is given by:

$$y_r[k] = \beta(\mathbf{a}_0^H \mathbf{x}[k]) \exp\left(j2\pi \frac{n_0}{N_c} k\right) + n_r[k], \quad (6.16)$$

where β denotes the signal attenuation, n_0 is the time domain sample delay, and $n_r[k] \sim \mathcal{CN}(0, \sigma_r^2)$ the RX noise. The maximum likelihood estimate of n_0 , denoted by $\hat{n}_0$, is given by:

$$\hat{n}_0 = \arg \max_n |Y_r[n]|, \quad (6.17)$$

$$\text{where } Y_r[n] = \sum_{k=0}^{N_c-1} y_r[k] (\mathbf{x}^H[k] \mathbf{a}_0) \exp\left(-j2\pi \frac{n}{N_c} k\right). \quad (6.18)$$

In (6.18), $Y_r[n]$ ($n = 0, \dots, N_c - 1$) is the N_c -point DFT of $\{y_r[k](\mathbf{x}^H[k] \mathbf{a}_0) : k = 0, \dots, N_c - 1\}$ delayed by n samples, and (6.17) amounts to identifying the location of the DFT peak.

The post-processing radar SNR, denoted by $\text{SNR}_{\text{rad}}(\mathbf{P})$, is defined as follows:

$$\text{SNR}_{\text{rad}}(\mathbf{P}) = \frac{|Y_r[\hat{n}_0]|^2}{\frac{1}{N_c-1} \sum_{n \neq \hat{n}_0} |Y_r[n]|^2} \quad (6.19)$$

The right-hand side of (6.19) is a function of $\mathbf{P}$, since $Y_r[\cdot]$ in (6.18) is a function of $\mathbf{x}[k]$, which in turn is a function of $\mathbf{P}$ (see (6.1)).

Definition 1 (RSMA ISAC performance region) *The ordered pair $(T_{\text{sum}}(\mathbf{P}), \text{SNR}_{\text{rad}}(\mathbf{P}))$ captures RSMA ISAC performance. Specifically, the four parameters $-(t_{\text{comms}}, t_p, \alpha_c, \alpha_p)$ – uniquely determine $\mathbf{P}$ (see (6.3)-(6.6) and Remark 24), which in turn determines $(T_{\text{sum}}(\mathbf{P}), \text{SNR}_{\text{rad}}(\mathbf{P}))$. Thus, the set of all feasible values for the four parameters corresponds to a collection of points $(T_{\text{sum}}(\mathbf{P}), \text{SNR}_{\text{rad}}(\mathbf{P}))$, which we define as the RSMA ISAC performance region, whose boundary (Pareto frontier) captures the limits of achievable ISAC performance for the heuristic precoders given by (6.3)-(6.6) – see also Remark 25.*

Remark 30 *Along the lines of Definition 1, the ISAC performance region can be defined for each of the special cases in Table 6.2. By definition, these performance regions are contained within the RSMA ISAC performance region defined above. However, the following questions are of particular interest:*

- *Can the boundary of the RSMA ISAC performance region in Definition 1 be achieved by one or more of the special cases in Table 6.2?*

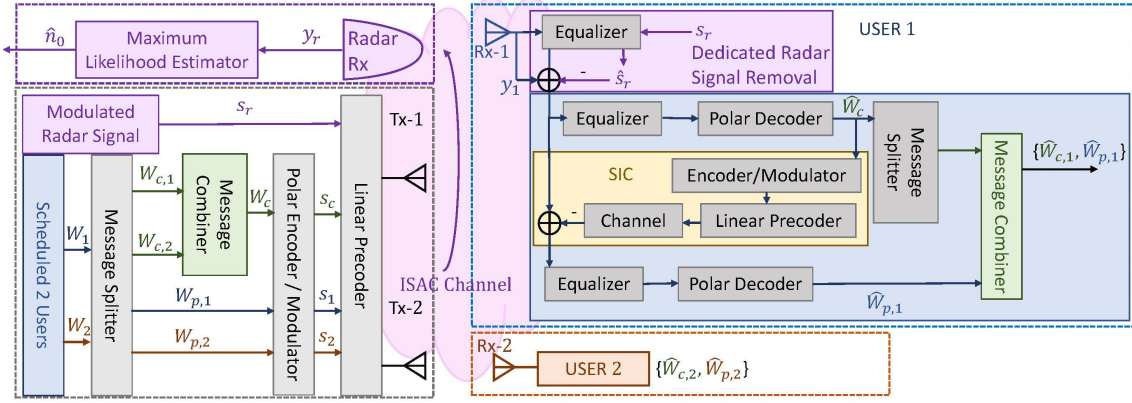


Figure 6.1: RSMA ISAC block diagram ($N_T = 2$) implemented using SDRs.

- How do the boundaries of the SDMA and RSMA ISAC performance regions w.r.t to one another?

6

Before addressing the questions in the previous remark in Section 6.4, we first describe our RSMA ISAC prototype in the following section.

6.3 RSMA ISAC Prototype

Fig. 6.1 depicts the RSMA ISAC block diagram, which we implement using software-defined radios (SDR). The TX, UEs and the radar RX are realized using National Instruments' (NI) USRP 2942 SDR units, which have two antennas/RF chains. Hence, we use three USRP 2942 units – one to realize the two-antenna TX, another to realize two single-antenna UEs, and a third to realize the (single-antenna) radar RX, where we use a Yagi-antenna pointing towards the target to obtain a strong radar return and suppress some of the self-interference (SI) from the TX $\rightarrow$ radar RX direct path. The TX and the radar RX are situated close together to mimic a monostatic configuration.

The USRPs share a common timing source (NI CDA-2990), and are controlled by a workstation running LabVIEW NXG, through which the various blocks in Fig. 6.1 are realized. All connections (SDRs to workstation, SDRs to timing sources) are through PCIe cables, facilitated by a PCIe bus (NI CPS-8910). A list of hardware components is provided in Table 6.3.

We realize the following two-stage protocol to implement the system model described in Section 6.2, adopting several features of the IEEE 802.11ac-VHT physical layer frames [115]:

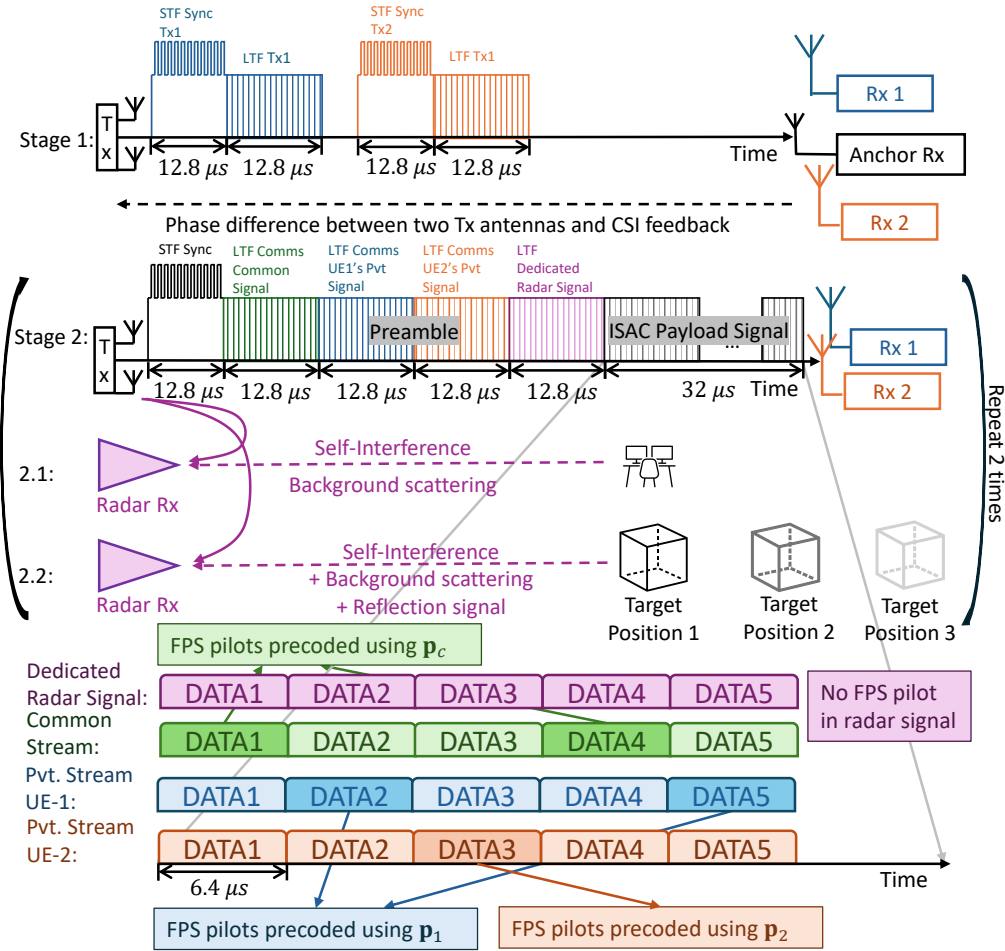


Figure 6.2: Signal structure within the two-stage transmission protocol used to implement RSMA ISAC in the prototype. In conventional, SDMA-based 802.11ac-VHT, where each UE has to decode only one data stream, every DATA (OFDM) symbol contains 16 *precoded* pilot subcarriers used to correct phase errors in the estimate of a UE's precoded channel. This protects against decoding errors caused by rotated constellations post equalization, and is known as fine phase shift (FPS) [121]. On the other hand, in RSMA, each UE needs to decode two data streams. To avoid FPS pilot contamination between overlapping streams, we use p_c to precode the FPS pilots in DATA symbols 1 and 4, similarly p_1 in DATA symbols 2 and 5, and p_2 in DATA symbol 3. The FPS pilots are unused in the stream carrying the dedicated radar signal.

	Name	Description
1.	Workstation	Running LabVIEW NXG
2.	NI USRP-2942 (3 units)	SDRs used to realize TX, UEs and radar RX
3.	NI CPS-8910	Provides additional PCIe ports
4.	NI CDA-2990	8 Channel, 10MHz clock distribution device
5.	Patch antennas (non-commercial, designed in house)	TX antennas
6.	TP-Link TL-ANT2405C (3 units)	Anchor and (two) UE antennas
7.	MM-ANT-NF-5G Yagi antenna	Radar RX antenna

Table 6.3: List of hardware components in RSMA prototype.

Stage 1 Each TX antenna transmits a pilot signal orthogonally in time comprising a Short Training Field (STF, $12.8\mu\text{s}$ in duration) and a Long Training Field (LTF, $12.8\mu\text{s}$ in duration), as shown in the top portion of Fig. 6.2. The STF is used for synchronization and coarse frequency offset estimation, while the LTF is used for channel estimation at each UE (i.e., $\hat{\mathbf{h}}_i[k]$ in Table 6.1).

Stage 2 Here, the transmitted signal consists of a preamble followed by the data payload, as shown in the bottom portion of Fig. 6.2.

- **Preamble:** The preamble consists of one STF and four LTFs. The function of the STF is the same as in Stage 1, while the LTFs are precoded in order to estimate the *precoded channels* for equalization at the UEs. The first LTF is used by UE i to estimate $\mathbf{h}_i^H \mathbf{p}_c$ ($i = 1, 2$) for decoding the common stream ($s_c[\cdot]$). The second LTF is used by UE 1 to estimate $\mathbf{h}_1^H \mathbf{p}_1$ to decode its private stream ($s_1[\cdot]$). Similarly, the third LTF is used by UE 2 to estimate $\mathbf{h}_2^H \mathbf{p}_2$ to decode its private stream ($s_2[\cdot]$). The fourth LTF is used by UE i to estimate $\mathbf{h}_i^H \mathbf{p}_r$ ($i = 1, 2$) in order to subtract the dedicated sensing signal ($s_r[\cdot]$) from (6.8).
- **ISAC Payload:** For the payload, we consider a total bandwidth of 100MHz with $N_c = 512$ subcarriers and a cyclic prefix (CP) of 128 samples per OFDM symbol. Aligned with IEEE 802.11ac-VHT frames, 468 subcarriers are used to carry data symbols, 16 subcarriers are used to correct the common phase error across all subcarriers in one OFDM symbol [121],

MCS Index $\mathcal{M}$	Modulation (m)	Code Rate r	Data Rate Bmr (Mbps)
0	BPSK (1)	1/2	36.5625
1	BPSK (1)	3/4	54.84375
2	QPSK (2)	1/2	73.125
3	QPSK (2)	3/4	109.6875
4	16QAM (4)	1/2	146.25
5	16QAM (4)	3/4	219.375
6	64QAM (6)	2/3	292.5
7	64QAM (6)	3/4	329.0625
8	256QAM (8)	3/4	438.75
9	256QAM (8)	5/6	487.5

Table 6.4: MCS levels (largely based on IEEE 802.11ac-VHT) implemented in the prototype. The data rate in the last column is equal to Bmr , where B is the effective bandwidth given by (6.20).

and the rest serve as guard bands. This yields an effective bandwidth of:

$$\begin{aligned}
 B &= 100\text{MHz} \times \underbrace{(512/640)}_{\text{CP overhead}} \times \underbrace{(468/512)}_{\text{Guard band overhead}} \\
 &= 73.125\text{MHz}
 \end{aligned} \tag{6.20}$$

The ISAC payload consists of at most four superposed streams (one dedicated sensing signal, one common, two private), each comprising 5 OFDM symbols.

- **MCS Implementation:** Table 6.4 lists the MCS levels, $\mathbb{M}$, implemented in the prototype. For channel coding, we implement Polar codes augmented with an 8-bit cyclic redundancy check [122], [123], along with successive cancellation list decoding [124], with a list depth of 2. After the preamble, the first OFDM symbol (labelled SERVICE in the bottom portion of Fig. 6.2) contains the MCS information of each stream.

- **Dedicated Sensing Signal:** For $s_r[\cdot]$, we consider a BPSK stream known to the UEs.

Parameter	Notation	Value
Center frequency	f_c	2.5GHz
Transmit power	P_t	23dBm
Wavelength	λ	0.12m
No. of UEs		2
No. of targets		1
Total bandwidth		100MHz
Subcarriers	Total (N_c)	512
	Data	468
	Pilot (FPS)	16
	Guard band	28
CP length		128
Effective bandwidth	B	73.125MHz
OFDM symbols in payload		5
Range resolution		1.5m
Radar $\rightarrow$ Target distances		2.25m ($n_0 = 1$)
		3.75m ($n_0 = 2$)
		5.25m ($n_0 = 3$)
TX $\rightarrow$ UE distances		≈ 1.5 m
TX antenna spacing	d	0.0625m
Fraunhofer distance (far-field criterion)		0.065m

Table 6.5: List of miscellaneous parameters in the experiments.

6.4 Experimental Results

6.4.1 Measurement Environment

the measurements were conducted in a fairly empty lecture room (Fig. 6.3), where the furniture was moved to the far end of the room, well away from the target, to reduce clutter. We used a metal box (1m high and 0.8m wide) placed at the broadside of the TX antennas to mimic a vehicular target. An anchor antenna was also placed at the TX broadside for phase calibration (see Appendix B).

We considered three scenarios as described below, capturing different levels of inter-UE interference and separation/integration between comms and sensing:

- S1. UEs and target mutually well separated** (in the angular domain), as shown in Fig. 6.3 (left). The larger the separation among UEs in the angular domain, the better SDMA can suppress the inter-UE interference. As a result, the RSMA gain – i.e., the extent to which its ISAC performance region is bigger than SDMA’s – could potentially be modest. Similarly, the extent of separation/integration between comms and sensing separation is dictated by the angular separation between the UEs and the target. A large angular separation may require a dedicated sensing signal to achieve the RSMA ISAC performance boundary.
- S2. UEs close together, but well separated from the target**, as shown in Fig. 6.3 (middle). By allocating more power to the common stream, RSMA can better suppress the high inter-UE interference when the UEs are not adequately separated in the angular domain. Hence, RSMA is expected to yield large throughput gains over SDMA in this scenario.
- S3. UEs and targets mutually close together**, as shown in Fig. 6.3 (right). The high level of integration between comms and sensing and the prominent role played by the common stream in suppressing the inter-UE interference makes this the most likely scenario where a dedicated sensing signal may not be needed for achieving the RSMA ISAC performance boundary.

It is worth pointing out that the geometry of these scenarios are representative of what could be encountered in peer-to-peer vehicular ISAC.

For each scenario, the goal of the measurements is to empirically characterize the RSMA ISAC performance region as per Definition 1. However, this can be quite cumbersome with four precoder parameters involved in determining each point $(T_{\text{sum}}(\mathbf{P}), \text{SNR}_{\text{rad}}(\mathbf{P}))$ in the RSMA ISAC performance region; for instance, with each parameter taking values over $[0, 1]$, a granularity of 0.1 per parameter yields 10^4 measurements. That said, it is more important to characterize the boundary of the RSMA ISAC performance region than its interior, as the boundary captures the limits of achievable ISAC performance. Thus, Remark 30 motivates the question: can we identify a smaller set of “well-chosen” precoder parameters corresponding to the boundaries of the RSMA and SDMA ISAC performance regions for a given scenario? We address this question next.

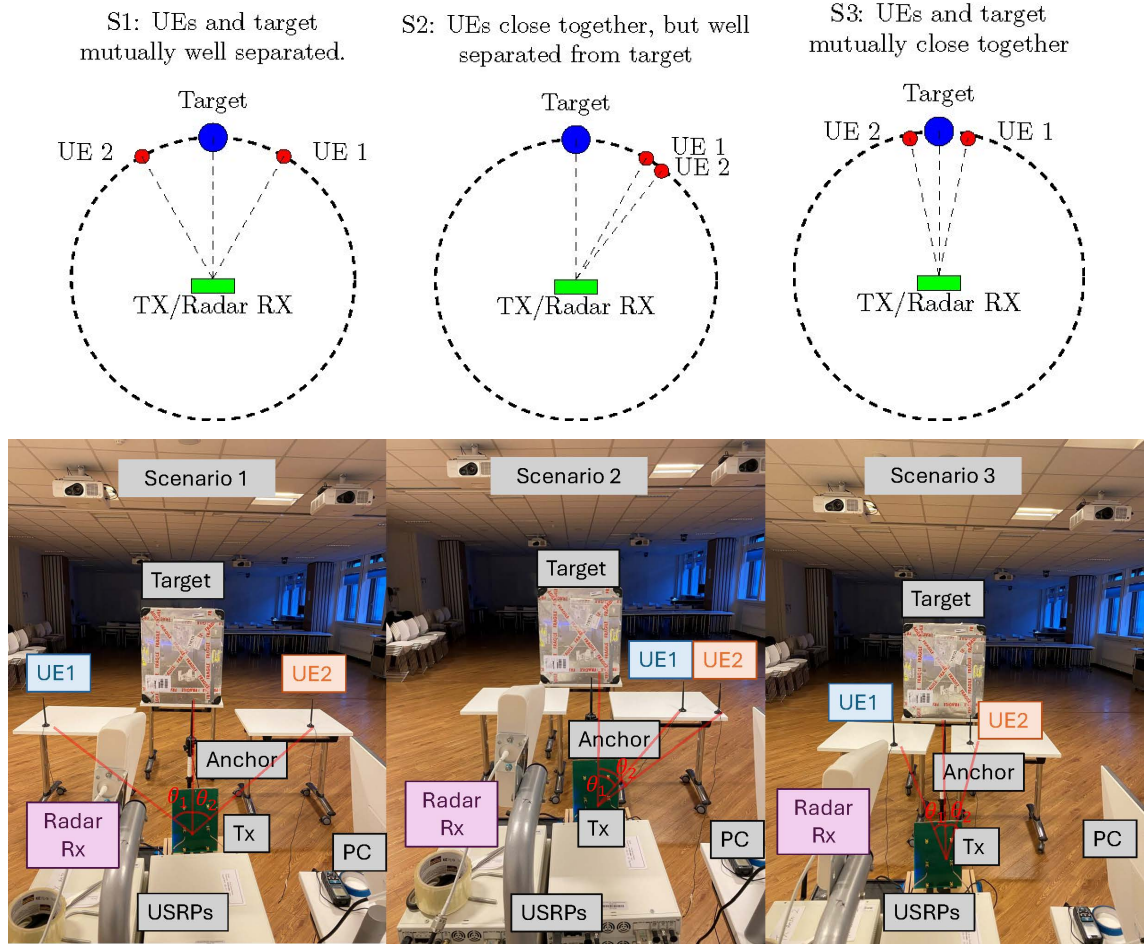


Figure 6.3: The three scenarios considered in the measurements.

6.4.2 Simulation-aided search for precoder parameters

Assuming the estimated UE channels ($\hat{\mathbf{h}}_i[k]$) as perfect CSIT¹, one can obtain a *simulated* RSMA ISAC performance region via a fine-grained sweep of the precoder parameters. We make this notion precise through the following definition.

Definition 2 (Simulated RSMA ISAC performance region) *The simulated RSMA ISAC performance region is the set of points $(T_{\text{sum}}(\mathbf{P}), G_0(\mathbf{P}))$, where $T_{\text{sum}}(\mathbf{P})$ is the **expected**, rather than measured, sum throughput (see Remark 29), and $G_0(\mathbf{P})$ is the power radiated towards the TX's broadside, given by*

$$G_0(\mathbf{P}) = \sum_{k=0}^{N_c-1} |\mathbf{a}_0^H \mathbf{x}[k]|^2, \quad (6.21)$$

¹The estimation of UE i 's CSIT is imperfect indeed. However, to simulate the ISAC performance region, the only CSIT we can obtain is the estimated CSIT.

In Definition 1, the sensing metric $-\text{SNR}_{\text{rad}}(\mathbf{P})$ – is straightforward to evaluate using measurement data (see 6.19). However, to evaluate $\text{SNR}_{\text{rad}}(\mathbf{P})$ in a simulation setting, we need to make several (potentially unrealistic) assumptions regarding the radar return (extent of signal attenuation, clutter etc.) which we wish to avoid. Hence, we use $G_0(\mathbf{P})$, which is proportional to $\text{SNR}_{\text{rad}}(\mathbf{P})$, as the sensing metric for the simulated RSMA ISAC performance region.

Ideally, the simulated RSMA ISAC performance region should be a good proxy for the *true* performance region – in particular, the precoder parameters corresponding to the RSMA/SDMA boundaries should have similar values over both regions; this then yields a smaller set of “well-chosen” parameter values that can be used for measuring RSMA and SDMA ISAC performance.

Fig. 6.4 plots the simulated RSMA ISAC region for each scenario by sweeping over the four precoder parameters in increments of 0.1, resulting in a reasonably comprehensive exploration of the parameter space. The points corresponding to the special cases in Table 6.2 are marked as well. The sum throughput values in these plots are based on *measured* UE channels – for each scenario, we placed the UEs approximately 1.5m away from the TX without any target present², and estimated the UE channels through only Stage 1 transmission in Fig. 6.2. We obtain the following insights from Fig. 6.4:

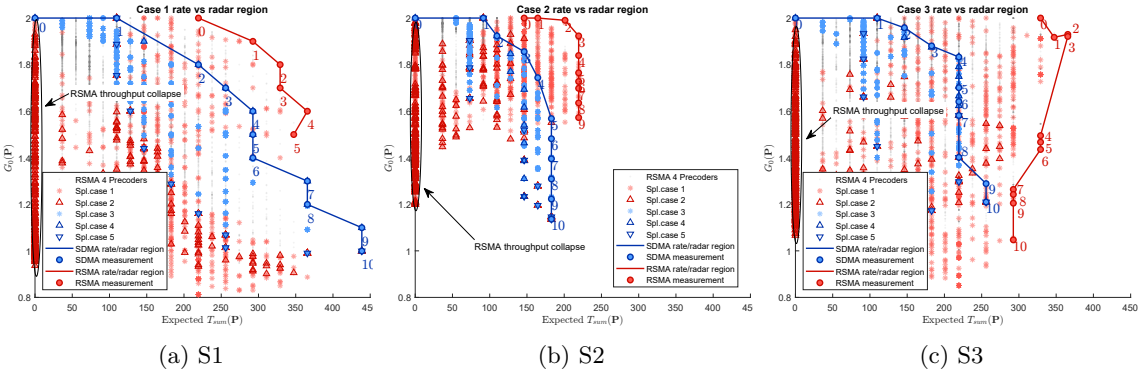


Figure 6.4: Simulated RSMA ISAC performance regions, as per Definition 2, for each of the scenarios in Fig. 6.3. The parameters associated with the points (marked 0, 1, etc.) on the SDMA and RSMA performance boundaries are listed in Table 6.6.

Impact of finite MCS levels Perhaps the most striking feature of Fig. 6.4 is the *non-convex* nature of the ISAC performance regions. In particular, the discretized nature of $T_{\text{sum}}(\mathbf{P})$ is a

²Indeed, the UE channels are affected by the target, which contributes a multipath component (TX $\rightarrow$ Target $\rightarrow$ UE) whose strength depends on the target range. However, for channels of similar strength, as in each of scenarios, $T_{\text{sum}}(\mathbf{P})$ is chiefly determined by the spatial correlation between the channels, which dictates the amount of inter-UE interference. The spatial correlation is not too sensitive to the target range; in fact, it remains qualitatively unchanged (i.e., low in S1, high in S2 and S3) even without a target. Thus, measuring the UE channels without a target present allows for the convenience of generating a single simulated RSMA ISAC performance region per scenario that is an effective proxy for the true performance region, regardless of the target range.

Meas. Index	S1					S2					S3				
	t_{comms}	α_c	α_p	$\mathcal{M}_1$	$\mathcal{M}_2$	t_{comms}	α_c	α_p	$\mathcal{M}_1$	$\mathcal{M}_2$	t_{comms}	α_c	α_p	$\mathcal{M}_1$	$\mathcal{M}_2$
0	0	-	-	-	-	0	-	-	-	-	0	-	-	-	-
1	0.1	-	1	0	0	0.1	-	0.8	0	0	0.4	-	0	0	0
2	0.2	-	1	1	1	0.2	-	0.8	0	0	0.1	-	1	0	0
3	0.3	-	1	1	2	0.3	-	0.8	0	0	0.2	-	0.9	0	0
4	0.4	-	1	3	3	0.4	-	0.8	1	1	0.4	-	1	1	1
5	0.5	-	1	3	3	0.5	-	0.8	1	1	0.5	-	0.9	1	1
6	0.6	-	1	3	4	0.6	-	0.8	1	1	0.6	-	0.9	1	1
7	0.7	-	1	3	4	0.7	-	0.8	1	1	0.9	-	0.7	1	1
8	0.8	-	1	3	4	0.8	-	0.8	1	1	1	-	0.7	1	1
9	0.9	-	1	3	4	0.9	-	0.8	1	1	1	-	0.9	2	2
10	1	-	1	4	4	1	-	1	2	2	1	-	1	2	2

(a) SDMA ($t_p = 1$)

Meas. Index	S1						S2						S3					
	t_p	α_c	α_p	$\mathcal{M}_c$	$\mathcal{M}_1$	$\mathcal{M}_2$	t_p	α_c	α_p	$\mathcal{M}_c$	$\mathcal{M}_1$	$\mathcal{M}_2$	t_p	α_c	α_p	$\mathcal{M}_c$	$\mathcal{M}_1$	$\mathcal{M}_2$
0	0	1	1	5	-	-	0	0	0	4	-	-	0	0	0	5	-	-
1	0.1	1	1	3	0	0	0.2	0	0.2	2	0	0	0.1	0	0.6	4	0	0
2	0.2	1	1	2	1	1	0.2	0	0.3	2	0	0	0.1	0	0.8	4	0	0
3	0.3	1	1	1	3	3	0.2	0	0.4	2	0	0	0.1	0	0.7	3	0	0
4	0.4	1	1	0	3	3	0.2	0	0.5	2	0	0	0	1	1	5	-	-
5	0.5	1	1	0	3	5	0.2	0	0.6	1	0	0	0.3	1	0.9	3	1	1
6	-	-	-	-	-	-	0.2	0	0.7	1	1	1	0.3	1	0.8	3	0	0
7	-	-	-	-	-	-	0.2	1	0.4	3	0	0	0.3	0.9	0.9	3	1	1
8	-	-	-	-	-	-	0.2	1	0.6	2	0	1	0.1	0.9	1	4	0	0
9	-	-	-	-	-	-	0.2	1	0.7	3	0	1	0	0.9	1	4	1	1
10	-	-	-	-	-	-	-	-	-	-	-	-	0.1	0.8	0.7	3	0	0

(b) RSMA without dedicated sensing signal ($t_{\text{comms}} = 1$)

Table 6.6: Precoder parameters associated with the boundary points of the simulated ISAC performance regions in Fig. 6.4 (the numbers marked 0, 1, and so on, in Fig. 6.4 correspond to the measurement index in the first column). The MCS levels shown are the ones that maximize the *expected* sum throughput, based on Remark 26. These parameters are used for the measurements in Section 6.4.3.

direct consequence of assuming finite MCS levels – instead of a one-to-one mapping between the throughput and SINR (i.e., $\log(1 + \text{SINR})$), a range of SINR values map to the same $T_{\text{sum}}(\mathbf{P})$.

Impact of imperfect SIC A significant proportion of RSMA points lie on the y -axis, where $T_{\text{sum}}(\mathbf{P}) = 0$. These capture the *throughput collapse* phenomenon, where for at least one of the UEs, the SINR is insufficient to decode the common stream at the lowest MCS level in Table 6.4 (see also Remark 27). Throughput collapse also explains why the RSMA ISAC performance boundary does not extend as far down as the SDMA boundary in the southeast region of Figs. 6.4a-b. The common stream SINR reduces as we move from the northwest to the southeast along the RSMA boundary; when it becomes low enough to trigger a throughput collapse, the boundary ends abruptly.

6

RSMA ISAC outperforms SDMA ISAC The SDMA ISAC performance boundary lies in the interior of the RSMA ISAC performance region, thereby addressing the second question posed in Remark 30. Hence, for well-chosen precoders (which boils down to well-chosen parameters), RSMA ISAC achieves points in the northeast region beyond the SDMA boundary. Furthermore, the extent to which RSMA ISAC outperforms SDMA ISAC is scenario-dependent. In scenario S1, where there is a relatively larger angular separation between the UEs, the gap between the RSMA and SDMA boundaries is narrower. The gap becomes much wider – especially around the northeast corner – as the UEs move closer together in S2 and S3.

RSMA ISAC performance boundary achieved without a dedicated sensing signal In each scenario, the points on the RSMA ISAC performance boundary are achieved when $t_{\text{comms}} = 1$, which corresponds to special case 1 in Table 6.2 where the entire transmit power is allocated for comms. This insight addresses the first question posed in Remark 30.

Dedicated sensing signal needed to achieve SDMA ISAC performance boundary Unlike in d), the points corresponding to SDMA ISAC without a dedicated sensing signal (i.e., special case 5 in Table 6.2) do not lie on the SDMA ISAC performance boundary.

Next, we seek to validate the above insights through measurements by selecting precoder parameters listed in Table 6.6 associated with the SDMA and RSMA ISAC boundary points marked 0, 1, etc., in Fig. 6.4.

6.4.3 Measurements

We conducted a total of 180 measurements with the following breakdown:

- RSMA – 81 measurements (6 boundary points for S1, 10 for S2 and 11 for S3, along with 3 different target ranges per boundary point to evaluate the effectiveness of ranging, especially for RSMA without a dedicated sensing signal; more details in paragraph a) below).
- SDMA – 99 measurements (11 boundary points each for S1, S2 and S3, along with 3 different target ranges per boundary point).

In these measurements, the UEs' locations were unchanged from the ones used to obtain Fig. 6.4. To suppress the radar self-interference (i.e., signal propagation along the direct path from the TX to the radar RX) and background clutter, we perform a simple background subtraction, where Stage 2 is repeated twice per measurement (illustrated as Stages 2.1 and 2.2 in Fig. 6.2). In the first instance, we obtain the radar return without any target present (Stage 2.1) to capture only the radar self-interference and any background clutter. This signal is then subtracted from the radar return obtained when the target is present (Stage 2.2). Table 6.6 lists the miscellaneous parameters used in the measurements.

Ranging performance of RSMA ISAC without a dedicated sensing signal The 100MHz signal bandwidth yields a range resolution of 1.5m. Hence, in the measurements, we considered target ranges of 2.25m ($n_0 = 1$), 3.75m ($n_0 = 2$) and 5.25m ($n_0 = 3$), corresponding to the middle of the second, third and fourth range bins, respectively³, as shown in Fig. 6.5.

Fig. 6.6 shows the ranging performance of RSMA without a dedicated sensing signal, where the heatmaps plot $\text{SNR}_{\text{rad}}(\mathbf{P})$ for each range bin. In each heatmap, the target is clearly identifiable at the correct range bin. Moreover, there is a graceful degradation in $\text{SNR}_{\text{rad}}(\mathbf{P})$ with increasing n_0 (left to right along each row), consistent with a progressively weaker radar return.

Measured ISAC performance region – RSMA v/s SDMA Fig. 6.7 plots the measured ISAC performance region for the precoder parameters listed in Table 6.6. To measure $T_{\text{sum}}(\mathbf{P})$, we identified the highest MCS levels that resulted in error-free decoding of the five OFDM symbols in the data payload (see Fig. 6.2) by using the MCS levels in Table 6.6 as a starting point and progressively decreasing/increasing the MCS levels if there were any/no decoding errors.

³ $n_0 = 0$ corresponds to the first range bin.

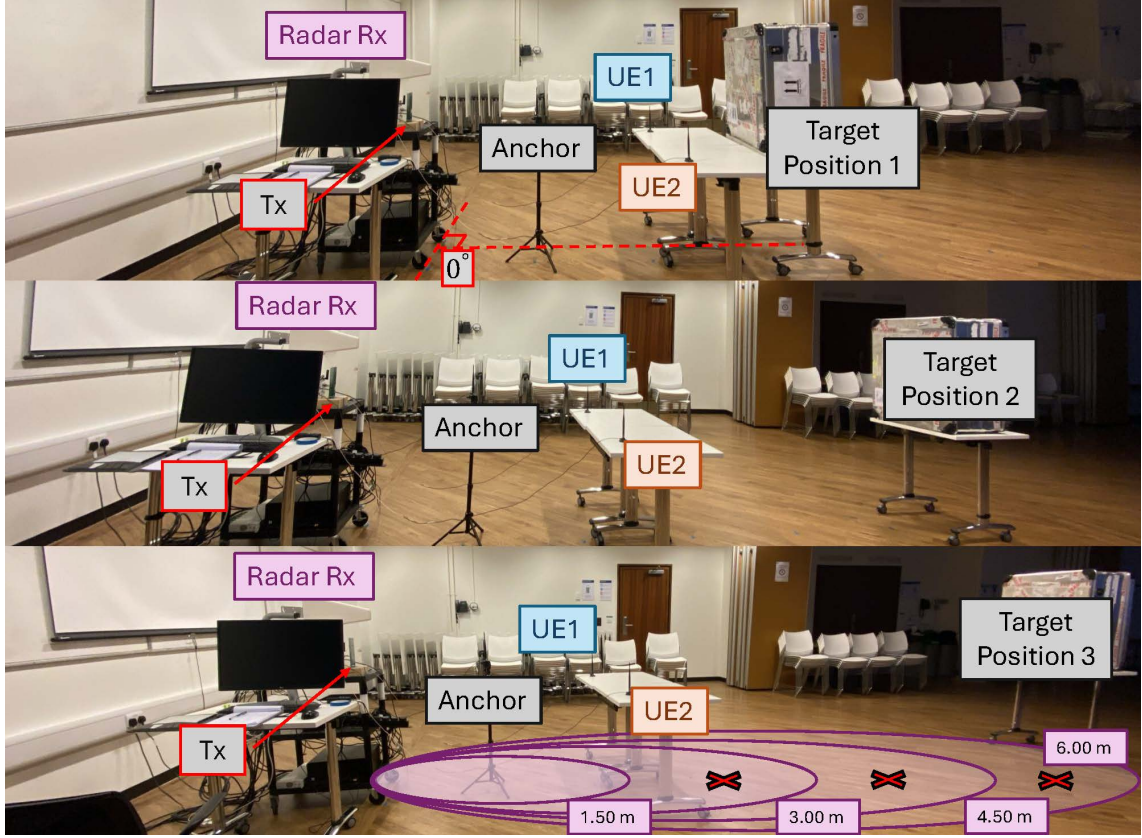


Figure 6.5: The three target distances considered for each scenario in Fig. 6.3 – 2.25m (top), 3.75m (middle) and 5.25m (bottom), which corresponds to the middle of the second ($n_0 = 1$), third ($n_0 = 2$) and fourth ($n_0 = 3$) range bins, respectively. The 100MHz signal bandwidth yields a range resolution (bin size) of 1.5m, which is marked in the bottom figure.

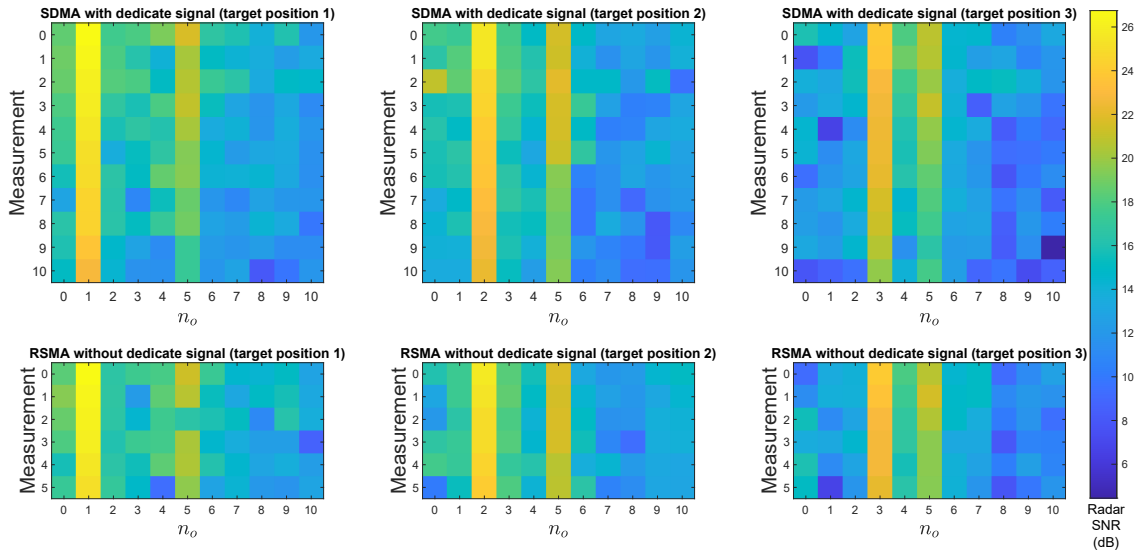


Figure 6.6: Heatmap plotting the measured $\text{SNR}_{\text{rad}}(\mathbf{P})$ for RSMA without a dedicated sensing signal across three different target distances for each scenario. The top row represents the results of SDMA with a dedicated sensing signal using 11 parameter settings; the bottom row represents the results of RSMA without a dedicated sensing signal using 6 parameter settings.

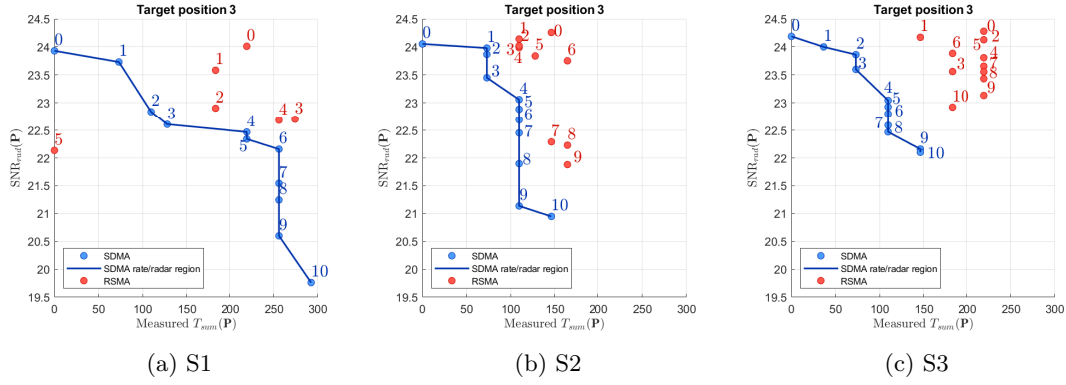


Figure 6.7: The measured ISAC performance region for RSMA and SDMA, for the precoder parameters and the MCS levels given in Table 6.6.

The following aspects of Fig. 6.4 are validated by Fig. 6.7:

- *RSMA outperforms SDMA in ISAC trade-off performance.* In Fig. 6.7, the measurement results demonstrate that RSMA consistently achieves superior trade-offs in ISAC performance compared to SDMA, with all the measured points (except point 5 in S1) located on the top right of the SDMA ISAC trade-off region as observed in the simulation.
- *Throughput collapse due to the impact of finite MCS levels and imperfect SIC.* The SDMA ISAC region seen in the Fig. 6.4 is mirrored in the measurements as shown in Fig. 6.7. The discretized achievable throughput (measured T_{sum}) aligns some of the results in the same line, even though they have different parameter settings and radar SNR performance. As a consequence of the imperfect SIC, point 5 (RSMA) in S1 is on the y-axis (measured $T_{sum} = 0$), when there is no lower MCS level for the common stream to use under the current SINR condition.

Finally, the following aspects of Fig. 6.7 diverge from Fig. 6.4.

- *The first RSMA measured results (red point 0) appear on the northwest of the second RSMA measured results among all three scenarios.* However, based on the simulation results in Fig. 6.4, the first measured point of RSMA (red point 0) should appear on the northeast of the second RSMA measured result (red point 1). To understand the reason for the throughput degradation, we need to look at the parameters Table 6.6 in measuring points 0 and 1. The main difference between those two measurements is in measuring point 0, there is no SIC involved since there is a common stream only ($t_p = 0$), the transmitter is broadcasting the common stream only to both UEs, and both UEs are only required to decode the common

stream, compared to measurements in point 1, private streams are activated ($t_p \neq 0$), both UEs are required to perform SIC to decode both common stream and the corresponding private stream. The reconstruction of the common stream symbols of UE i is based on the estimation of $\mathbf{h}_i^H \mathbf{p}_c$, the imperfect estimated precoded CSI causes imperfection⁴. The common stream can propagate residual interference to subsequent streams (private streams). This cascading effect may degrade the decoding performance due to the increase of SINR.

- *Red point 5 in S1 has shown a throughput collapse*, but in the simulation result shown in Fig. 6.4a, red point 5 is located on the northeast of the SDMA ISAC region. This phenomenon is caused by CSI estimation imperfection, but the parameter is decided by simulation, which assumes CSI is perfectly estimated. In the simulation, the calculated common stream SINR can support both UEs to decode the common stream under the parameter setting for point 5, but in practical measurement, the imperfection in CSIT estimation causes the mismatching between the achievable throughput and the corresponding parameter settings.

6

6.5 Summary

In this chapter, we present the first experimental study of RSMA ISAC using SDRs and cross-compared SDMA ISAC with RSMA ISAC in three different scenarios; key findings show that RSMA outperforms SDMA in all scenarios by achieving a larger ISAC trade-off region. Notably, RSMA without a dedicated sensing signal provides a broader ISAC region than SDMA, even if SDMA is assigned with a dedicated sensing signal. The results validate the superior functionality of the common stream in mitigating inter-user interference and enabling the same throughput for the same radar SNR. However, the imperfections introduced by the practical CSIT estimation and finite MCS levels limit both SDMA and RSMA performance in some instances, causing throughput collapse and the mismatching between the simulation and measurements, but RSMA outperforms SDMA in both simulation and experimental results. In conclusion, all findings show RSMA is a potential candidate for next-generation ISAC systems, especially in vehicular and multi-user communication scenarios. In future work, an optimized precoder design for a larger ISAC trade-off region is needed, and a better system model that considers imperfect SIC and CSIT estimation is also worth exploring in closing the gap between theoretical and experimental performance.

⁴This imperfection in SIC at UE i is caused by estimation error of precoded CSI ($\hat{\mathbf{h}}_i^H \mathbf{p}_c \approx \mathbf{h}_i^H \mathbf{p}_c$) which the main imperfection is introduced in the reconstruction procedure and channel estimation. However, the imperfect SIC mentioned in Section 6.4.2 is caused by the unsuccessful decoded common stream ($\hat{s}_c \neq \hat{s}$), which is introduced in the decoding common stream procedure.

Conclusions and Future Work

7.1 Conclusions

This work presents an experimental investigation into the performance of Multiple Access (MA) techniques, focusing on Rate-Splitting Multiple Access (RSMA) and its potential to address the challenges of next-generation wireless systems, particularly in Unicast, Non-Orthogonal Unicast and Multicast (NOUM), Multi-group Multicast (MGM) and Integrated Sensing and Communication (ISAC) applications. The study compares Rate-Splitting Multiple Access (RSMA) with Space Division Multiple Access (SDMA) and Non-Orthogonal Multiple Access (NOMA) across various metrics and scenarios, offering insights into their respective strengths and limitations. Through prototype implementations using Software Defined Radios (SDRs), this thesis highlights several key findings and is listed in Table 7.1.

In chapter 3, the superiority of RSMA is first shown experimentally in a hardware platform. RSMA outperforms SDMA and NOMA in a two-user Multi-user Multi-Input Single Output system in both sum throughput performance and user fairness performance. The throughput gain is even pronounced when the Channel State Information at Transmitter (CSIT) feedback is quantized. All those advantages of RSMA make it a promising candidate MA technique for next-generation PHY-layer.

Chapter 4 presents the characterization and experimental investigation of RSMA-based NOUM for the two-user case. The analysis reveals that the advantages of RSMA-based NOUM become more significant as channel correlation increases. This is attributed to the common stream con-

Chapter	System Model	Contributions
3	Unicast Two-user MU-MISO	Experimental validation of RSMA's superiority over SDMA and NOMA in sum throughput and fairness. Gains are more pronounced with quantized CSIT feedback.
4	NOUM Two-user MU-MISO	Demonstrates that RSMA-based NOUM benefits from increased channel correlation, expanding the rate region compared to MULP-based NOUM.
5	MGM Overloaded Two-groups	First experimental evaluation of max-min fairness in RSMA, showing superior fairness in scenarios with strong interference and pathloss variations.
6	ISAC RSMA-based MU-MISO	RSMA achieves a larger ISAC trade-off region than SDMA, even without a dedicated sensing signal. Practical challenges such as imperfect CSIT affect both techniques.

Table 7.1: Summary of main contributions of each chapters

7

tributing a larger portion to the sum throughput, enhancing flexibility in accommodating both multicast and unicast transmissions. Consequently, the rate region gap between RSMA-based NOUM and Multi-User Linear Precoder (MULP)-based NOUM widens with increasing channel correlation.

In chapter 5, the first experimental evaluation of the max-min fairness performance of RSMA-, SDMA-, and NOMA-based MGM, with a focus on overloaded scenarios involving closely located group members—situations relevant to various applications. The study considered two groups, each with two users, across nine cases that captured different levels of interference and pathloss variations between the groups. Results demonstrate that RSMA-based MGM achieves superior fairness performance, characterized by higher minimum throughput, compared to SDMA- and NOMA-based MGM. These findings align with theoretical predictions and are attributed to the ability of RSMA to allocate throughput contributions from both the common and private streams, with their proportions dynamically adjusted based on channel conditions.

In chapter 6, the first experimental study of RSMA-based ISAC using SDRs compares its performance with SDMA-based ISAC across three scenarios. The findings demonstrate that RSMA consistently outperforms SDMA by achieving a larger ISAC trade-off region. Notably, RSMA without a dedicated sensing signal achieves a broader ISAC region than SDMA, even when SDMA utilizes a dedicated sensing signal. The results confirm the effectiveness of RSMA's common stream in mitigating inter-user interference and maintaining throughput at the same radar SNR. However,

practical limitations, such as imperfect CSIT estimation and finite Modulation and Coding Scheme (MCS) levels, affect the performance of both SDMA and RSMA, occasionally causing throughput collapse and deviations from theoretical predictions. Despite these challenges, RSMA shows superior performance in both simulations and experimental evaluations.

In conclusion, RSMA demonstrates its potential as a leading candidate for next-generation systems in many applications, especially in vehicular networks ISAC, multi-user communications, and overloaded environments.

7.2 Future Work

7.2.1 RSMA test bed with larger scale transmitter antenna

Future work should prioritize scaling up the number of the Transmitter (TX) antennas in the current prototype. The current prototype only uses two TX antennas, which is way less than the number of 5G BSs' antennas. The future prototype test bed needs to incorporate a greater number of transmit antennas, enabling more comprehensive investigations into multi-antenna systems. Expanding the number of antennas at the transmitter would allow for more advanced precoding techniques, better spatial multiplexing capabilities, and enhanced handling of inter-user interference. Such configurations are particularly relevant for evaluating advanced Multiple Input Multiple Output (MIMO) scenarios, including massive MIMO. Additionally, incorporating massive MIMO configurations would provide insights into how emerging techniques such as RSMA, NOMA, and SDMA perform in large-scale antenna arrays. Future studies could evaluate the robustness of these techniques under pilot contamination, low-resolution Analog to Digital Converter (Analog to Digital Converter (ADC)), hardware impairments, and other real-world constraints, which are critical for practical deployment. Furthermore, with the use of larger antenna arrays, experiments could explore hybrid beamforming to balance the hardware complexity and performance, which is also a crucial beamforming technique in next-generation wireless communication.

7.2.2 RSMA test bed with link adaptation

Link adaptation plays a critical role in practical deployments by ensuring that communication systems operate efficiently under varying channel states, improving overall system performance. The future work of RSMA prototype should focus on incorporating link adaptation techniques

to dynamically adjust the transmission MCS levels based on the real-time channel conditions. Specifically, different from the link adaptation in conventional multiple access, link adaptation in RSMA needs to be applied to the common stream and the private streams to ensure communication reliability. Moreover, incorporating Adaptive Modulation and Coding (AMC) into the RSMA prototype would enable efficient use of the available resources, ensuring that each user is served with the appropriate MCS based on their channel quality. AMC combined with RSMA's flexibility in splitting and managing interference would allow the system to maintain robust communication even under imperfect CSI or rapidly changing environments. The studies about link adaptation of RSMA [112] and RSMA with AMC [109] focus on higher layer simulations rather than practical implementations. The future work should fill this research gap by realizing the link adaptation and AMC aided RSMA prototype.

7

7.2.3 RSMA test bed with mmWave transceiver

Millimeter-wave (mmWave) technology, operating in the 24–100 GHz range, is a critical enabler of 6G wireless networks due to its potential to offer massive bandwidth, ultra-high data rates, and low-latency communication. Integrating RSMA with mmWave transceivers would allow us to investigate its performance under the unique challenges and opportunities presented by mmWave propagation and hardware characteristics. Future studies can focus on the performance of RSMA under mmWave channel challenges. Specifically, the RSMA prototype with mmWave transceivers must account for significant signal attenuation due to the high operating frequency band. Moreover, the mmWave channel exhibits sparse multipath characteristics and rapid variations, especially in mobile scenarios. Extending the RSMA prototype to include real-time channel estimation and adaptation mechanism will be critical for adapting to the fast-changing mmWave channel. The mmWave transceiver can also be used in other 6G applications such as ISAC. Given the massive bandwidth available at mmWave frequencies, RSMA's integration with ISAC can be explored further. mmWave offers excellent range resolution for sensing tasks such as target detection, tracking, and environment mapping. The RSMA prototype can utilize mmWave signals to demonstrate its ability to simultaneously provide high data rates for communication and accurate sensing capabilities.

7.2.4 RSMA with RIS assisted communication

Future prototyping work on RSMA should also focus on its integration with Reflective Intelligent Surface (Reconfigurable Intelligent Surfaces (RIS)), a promising technology for enhancing wireless communication performance in 6G and beyond. RIS consists of large arrays of passive, low-cost, and energy-efficient reflecting elements that can manipulate the wireless propagation environment to improve signal quality, reduce interference, and enhance coverage. Combining RSMA with RIS creates a powerful synergy, enabling flexible interference management and efficient resource utilization in complex communication scenarios [128], [129]. However, several challenges must be addressed to develop a functional RIS-aided RSMA prototype and advance its real-world applicability. The first one is the Channel State Information (CSI) acquisition. Accurate CSI is critical for optimizing both RSMA precoders and the RIS phase shifts. However, obtaining full CSI in RIS-aided systems is highly challenging due to the passive nature of RIS elements, which do not perform signal processing. Secondly, hardware limitations and practical imperfections. Practical RIS hardware suffers from phase quantization, non-ideal reflection coefficients, and signal loss, which degrade the theoretical performance. RSMA-based prototypes must account for imperfections in hardware and compensate for non-ideal behaviours. Last but not least, joint optimization of precoders and RIS phase shift. The coupling effect between the RSMA precoders and RIS phase optimization makes the problem non-convex and computationally intensive. A new challenge is to develop real-time algorithms that jointly optimize the RSMA precoders and RIS phase shifts efficiently under hardware constraints. By tackling these challenges, RIS-aided RSMA prototype will enhance the interference management in 6G further.

7.2.5 RSMA prototype with antenna polarization

Future research on Rate-Splitting Multiple Access (RSMA) can be significantly enhanced by integrating antenna polarization techniques to optimize spectral efficiency, interference management, and user experience in next-generation wireless networks. Antenna polarization plays a crucial role in RSMA by influencing signal propagation, channel characteristics, and the separation of common and private message components. One promising direction is the development of polarization-aware precoding techniques, where RSMA precoders are designed to exploit dual-polarized or circularly polarized antennas to maximize the signal-to-interference-plus-noise ratio (SINR) at the receivers. By aligning the polarization states of transmitted and received signals, the system can minimize polarization mismatch losses, leading to higher achievable rates for both common and

private messages in RSMA-based networks. Additionally, polarization diversity can provide significant gains in user allocation strategies. Base Stations with orthogonally polarized antennas can further reduce inter-user interference, allowing more efficient power allocation and decoding. This is particularly advantageous in multi-user MIMO (MU-MIMO) scenarios, where polarization diversity helps improve channel orthogonality, enhancing spatial multiplexing and allowing multiple users to communicate simultaneously with reduced interference. Another crucial aspect is cross-polarization interference mitigation, where RSMA can benefit from the controlled use of polarization to suppress other users' interference. In dense and dynamic environments, users may experience signal degradation due to cross-polarization leakage, which RSMA can address through adaptive precoding and Link-adaptation. Furthermore, prototyping an over-the-air RSMA system with reconfigurable dual-polarized antennas will be essential to validate theoretical gains in practical deployments. Such a prototype would allow real-world testing of RSMA's robustness against polarization-dependent fading, enabling a deeper understanding of its performance in various environments, including indoor, urban, and vehicular communication scenarios.



CSIT Quantization Method in IEEE 802.11g

A

The quantized wideband CSI, $\hat{\mathbf{h}}_i^Q := [\hat{h}_{i1}^Q \ \hat{h}_{i2}^Q]$, is obtained from $\hat{\mathbf{h}}_i = [\hat{h}_{i1} \ \hat{h}_{i2}]$, as follows:

(a) The *linear scaler* is evaluated as follows:

$$M_h^{\text{lin}} := \frac{m_h}{10^{M_h/20}} \quad (\text{A.1})$$

$$\text{where } m_h := \max_{i,l} \{|\text{Re}(\hat{h}_{il})|, |\text{Im}(\hat{h}_{il})|\} \ (i, l \in \{1, 2\}), \quad (\text{A.2})$$

$$\text{and } M_h := \min \left\{ 7, \left\lfloor 20 \log_{10}(m_h) \right\rfloor \right\}. \quad (\text{A.3})$$

M_h in (A.3) is known as the scaling ratio and is quantized to three bits for feedback.

(b) Using (A.1), the real and imaginary parts $\hat{h}_{ij}$ are quantized using $N_b = 4$ bits in our prototype as follows.

$$\text{Re}(\hat{h}_{ij}^Q) := \text{round} \left(\frac{\text{Re}(\hat{h}_{ij})}{M_h^{\text{lin}}} (2^{N_b-1} - 1) \right) \quad (\text{A.4})$$

$$\text{Im}(\hat{h}_{ij}^Q) := \text{round} \left(\frac{\text{Im}(\hat{h}_{ij})}{M_h^{\text{lin}}} (2^{N_b-1} - 1) \right) \quad (\text{A.5})$$

The feedback overhead for the entire CSI equals $3 + 2N_b il$ bits, where the first term corresponds to the scaling ratio, and the factor of two in the second term accounts for the real and imaginary parts

of $\hat{h}_{ij}^Q$. For our prototype, this value equals 35 bits, as opposed to the $128il = 512$ bits without quantization.

At the TX, the CSI estimates, denoted by $\hat{h}_{il}^{\text{TX}}$, are recovered as follows:

$$\hat{h}_{il}^{\text{TX}} := \frac{\hat{h}_{il}^Q}{10^{M_h/20}} \quad (i, l \in \{1, 2\}). \quad (\text{A.6})$$

B

Phase compensation in ISAC Prototype

Each element in the transmitter antenna array experiences unique phase noise due to manufacturing variability. To ensure an accurate steering vector for the array, it is necessary to calibrate these elements so that there is no phase difference before the measurement. This calibration process in [130] requires placing an anchor antenna at the broadside of the transmitter array. Each array element is sequentially activated to transmit a reference signal sequence to the anchor receiver, allowing the measurement of the channel between each transmitter element and the anchor antenna individually. The measured channel between transmitter antenna- g to anchor on k -th subcarrier can be denoted as:

$$h_{a,g}[k] = \beta_a \underbrace{\exp(j2\pi \frac{n_a}{N_c} k)}_{\text{Propagation delay}} \cdot \underbrace{\exp(j\phi_{g,k})}_{\text{RF chain}} \cdot \underbrace{\exp(j2\pi \frac{dg}{\lambda} \sin(\theta_a))}_{\text{Extra propagation delay}}, \quad (\text{B.1})$$

where $g \in \{0, 1\}$, β_a denotes the signal attenuation between the transmitter antenna and the anchor, the first phase term comes from the propagation delay from the transmitter to anchor, the second term is the phase shift introduced by the RF chain hardware [131]. The third term comes from the extra propagation delay between the first antenna and the other antenna, which depends on the degree of arrival θ_a (0°). The calibration process will estimate the random phase difference caused by the RF components between TX-0 and TX-1 ($\Delta\hat{\phi} = \frac{1}{N_c} \sum_{k=0}^{N_c-1} (\hat{\phi}_{0,k} - \hat{\phi}_{1,k})$), then the estimation of phase difference $\Delta\hat{\phi}$ is applied on TX-1 to compensate the phase difference.

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