



University of
Nottingham

UK | CHINA | MALAYSIA

Near-field MIMO for Wireless Communication in Complex Propagation Environments

Thesis submitted **March 2021**, to the University of Nottingham
for the award of the degree **Doctor of Philosophy**.

Mir Lodro

Supervised by
Dr. Steve Greedy and Prof. Ana Vukovic

George Green Institute for Electromagnetics Research-GGIEMR
University of Nottingham, UK

I hereby declare that I have all necessary rights and consents to publicly
distribute this dissertation via the University of Nottingham's e-dissertation
archive.

*Dedicated to my father and my mother for the their unconditional love and their
kind support throughout my life.
My kids Abeera and Seerat and my wife Nahdia for the beautiful memories.*

Abstract

Communication between chips on integrated packages takes place using conducting wires, vias, microwave transmission lines, inductive and capacitive coupling. Inductive and capacitive coupling for data transfer is impractical for high data rate chip-to-chip communication because of limited scalability and poor performance caused by chip misalignments. Therefore, the solution to overcome this bottleneck is the usage of wireless transmission of information for chip-to-chip or inter-chip communication. Chips are packaged within metal encasing or metal enclosures with electronic circuitry acting as a source of scatterers with a natural tendency to form LOS and NLOS scenarios. Wireless communication between chips under this scenario forms a reverberant cavity where attenuated and phase-shifted multipath fading signals can arrive from directions at different time delays. Such an environment has a high potential to exhibit rich multipath fading characteristics. Therefore, extensive channel measurements are necessary for the statistical characterization of the wireless chip-to-chip channel. An SDR based 2x2 MIMO communication link is created to perform measurements in the mode-stirred metal enclosure by keeping in mind the fact that wireless chip-to-chip communication is a reverberant environment. Wireless chip-to-chip communication takes place in a confined environment with distance and form-factor limitations. These limitations further aggravate MIMO wireless chip-to-chip communication pushing MIMO communication to use closely spaced transmit and receive antenna elements. Because of form-factor and design limitations, transmit and receive antenna arrays are placed in the near-field of each other. Therefore, the

best effort has been made to develop a 2x2 MIMO communication link using SDR based testbed and recreate this environment in a mode-stirred metal enclosure. MIMO channel measurements are conducted in LOS and NLOS multipath fading scenarios in the presence of transmitting and receiving antenna correlations. QPSK digital receiver has been used to understand the near-field link based on BER performance and EVM measurements in a mode-stirred environment. SDR based 2x2 MIMO communication link is tested in metal enclosure in LOS and in NLOS environment when it's empty and when it was loaded with RF absorbers. Two channel BER and EVM results were extracted in LOS and NLOS scenarios both in presence of RF absorbers and in absence of RF absorbers. The experiment for 2x2 MIMO communication link was also repeated in presence of MIMO interference. From the measurements it was shown that the two USRP channel BER and EVM decreases with the increase in Tx interference gain.

Acknowledgement

First and the foremost, I am thankful to my PhD supervisors Dr. Steve Greedy and Prof. Ana Vukovic who guided me through rewarding journey of PhD life. I am also thankful to research team at GGIEMR specially Dr. Chris Smartt, Prof. Dave Thomas, Prof. Ian Harrison, Dr. Gabriele Gradoni, Prof. Gregor Tanner, Dr. Sandy Phang, Dr. Ismail Maricar, Dr. Armando Mendez. Specially, I am thankful to Dr. Gabriele Gradoni for his outstanding support and guidance on metal enclosure related measurements. I am deeply thankful to the USRP community and the GNU Radio community for their support which proved to be helpful for the design of USRP based near-field MIMO testbed.

List of Abbreviations

- ADC: Analog-to-Digital Converter
- AFD: Average Fade Duration
- AGC: Automatic Gain Control
- AIC: Akaike Information Criterion
- BER: Bit Error Ratio
- C2C: Chip-to-Chip
- CIR: Channel Impulse Response
- CDF: Cumulative Distribution Function
- CCDF: Complementary Cumulative Distribution Function
- DAC: Digital-to-Analog Converter
- DDC: Digital Down-conversion
- DUC: Digital Up-conversion
- EGC: Equal Gain Combining
- EVM: Error Vector Magnitude
- IF: Intermediate Frequency
- IID: Independent and Identically Distributed

- KPI: Key Performance Indicators
- LCR: Level Crossing Rate
- LFRX: Low Frequency Receiver
- LFSR: Linear Feedback Shift Register
- LFTX: Low Frequency Transmitter
- LoRa: Long Range
- LOS: Line-of-Sight
- MER: Modulation Error Rate
- MIMO: Multiple Input Multiple Output
- MISO: Multiple Input Single Output
- MPC: Multipath Component
- MRC: Maximum Ratio Combining
- NFC: Near-field Communication
- NFMI: Near-field Magnetic Induction
- NR: New Radio
- NLOS: Non-line-of-Sight
- OFDM: Orthogonal Frequency Division Multiplexing
- PDF: Probability Density Function
- PSD: Power Spectral Density
- QPSK: Quadrature Phase-Shift Keying
- RFIC: Radio Frequency IC

- RFID: Radio Frequency Identification
- SIMO: Single Input Multiple Output
- SISO: Single Input Single Output
- SDR: Software Defined Radio
- SNR: Signal-to-Noise Ratio
- SVD: Singular Value Decomposition
- UAV: Unmanned Aerial Vehicle
- USRP: Universal Software Radio Peripheral
- V2V: Vehicle to Vehicle Communication
- WLAN: Wireless Local Area Network
- XPC: Cross Polarization Coupling
- XPD: Cross Polarization Discrimination
- XPR: Cross Polarization Ratio

List of Symbols

- $\mathbf{H}$: MIMO channel matrix
- $\mathbf{y}$: Received signal vector
- $\mathbf{x}$: Transmitted signal vector
- γ : SNR
- ρ : MIMO SNR
- ρ_{th} : Diversity threshold
- K : Rician factor
- $k(\mathbf{H})$: Condition number
- $\mathbf{R}_H$: MIMO system correlation matrix
- $\mathbf{R}_R$: Receive correlation matrix
- $\mathbf{R}_T$: Transmit correlation matrix
- $f_X(x)$: Probability density function of r.v. X
- $f_X(\bullet|\bullet)$: Conditional distribution
- $F_X(x)$: Cumulative distribution function of r.v. X
- $|h(i)|^2$: Power gain of i th branch
- Q : Q-factor

- $Q(\bullet)$: Q-function
- B : Bandwidth
- C : Channel Capacity
- N_T : Number of transmitter antennas
- N_R : Number of receiver antennas
- $H(f)$: MIMO frequency transfer function
- $H(f, t)$: MIMO frequency transfer function with t snapshots
- $h(\tau)$: Impulse response of channel
- $h(\tau, t)$: Impulse response of channel with t snapshots
- σ_{max} : Maximum Singular Value
- σ_{min} : Minimum Singular Value
- $I(\mathbf{x}; \mathbf{y})$: Mutual information between r.v. $\mathbf{x}$ and output r.v. $\mathbf{y}$
- λ_i : i th eigenvalue of MIMO channel

List of Publications

Journal

1. "Near-Field Image Transmission and EVM Measurements in Rich Scattering Environment in Metal Enclosure" Progress In Electromagnetics Research M, Vol. 101, 139–147, 2021
2. "Near-field BER and EVM Measurement at 5.8 GHz in Mode-stirred Metal Enclosure" The Applied Computational Electromagnetics Society (ACES) Journal, vol. 35,no.9,2020.
3. "Statistical Characterization of Wireless MIMO Channels in Mode-Stirred Enclosures" submitted IEEE ACCESS
4. "2x2 Near-field MIMO Prototype for BER and EVM Measurements in Metal Enclosure" submitted in PIER M

Conference

1. "Channel Characterization of Reconfigurable Intelligent Radio Environment" 2021 Joint IEEE International Symposium on Electromagnetic Compatibility, Signal and Power Integrity, EMC Europe, pp.1-5, 2021
2. "Time-Domain Wireless Channel Sounding Using Software Defined Radio" IEEE International Conference on Mathematics and Emerging Technologies (iCoMET 2020), pp.1-5,2020

3. "SDR based Implementation of MSK and GMSK Receiver in Rich Multipath Environment" IEEE International Conference on Mathematics and Emerging Technologies (iCoMET 2020),pp.1-5,2020
4. "BER measurement and OTA performance of QPSK receiver in an anechoic chamber" Loughborough Antennas and Propagation Conferences (LAPC2018), United Kingdom
5. "Statistical Channel Modelling of 5G mmWave MIMO Wireless Communication" IEEE International Conference on Mathematics and Emerging Technologies (iCoMET 2018),pp.1-5,2018.
6. "Image Transmission using OSTBC-encoded 16-QAM over Time-selective Multipath Fading Channels" Loughborough Antennas and Propagation Conferences 2017 (LAPC 2017), United Kingdom

Contents

Abstract	ii
Acknowledgment	iv
List of Abbreviations	v
List of Symbols	viii
List of Publications	x
1 Introduction	1
1.1 Wireless Chip-to-Chip communication	1
1.2 Near-field and Far-field Communication	3
1.2.1 Far-field or Fraunhofer Region	4
1.2.2 Reactive near-field	4
1.2.3 Near-field Magnetic Communication	5
1.2.4 BER Measurement of Intra-chip Communication	6
1.3 Aims and Objectives	8
1.4 Contribution	9
1.5 Thesis Structure Outline	10
1.6 Conclusion	12
Bibliography	14

2	SDR Based Near-field MIMO Testbed	18
2.1	Software Defined Radio	18
2.2	SDR based Near-field Testbed	21
2.3	MIMO Channel Sounding	22
2.4	USRP based Testbed Limitations	25
2.5	USRP X310 Architecture	30
2.6	Digital Upconversion and Downconversion	31
2.6.1	Digital Up-conversion (DUC)	31
2.6.2	Digital Down-conversion (DDC)	33
2.7	Baseband Model for Frequency Sweep	34
2.8	USRP Classification	38
2.9	RF Daughterboards	40
2.10	Baseband Waveform Development Platforms	41
2.10.1	GNU Radio	42
2.11	Conclusion	43
	Bibliography	44
3	Channel Measurement, Modeling and Diver-	
	sity	47
3.1	Introduction	47
3.2	Classical Fading Distributions	47
3.3	Gaussian or Normal Distribution	48
3.4	Rayleigh Distribution	49
3.5	Rician Distribution	49
3.6	Nakagami-m Distribution	50
3.7	Log-Normal Distribution	50
3.8	Weibull Distribution	51
3.9	Multipath Fading	52

3.9.1	Small-scale fading	53
3.9.2	Large-scale fading	53
3.9.3	Path Loss	55
3.9.4	PDP/RMS Delay Spread	56
3.9.5	Coherence bandwidth	56
3.9.6	Coherence Time	57
3.9.7	Coherence Distance	58
3.9.8	Doppler Spread	58
3.9.9	Frequency flat fading	59
3.9.10	Frequency selective fading	59
3.10	Second Order Statistics	59
3.10.1	Level Crossing Rate (LCR)	60
3.10.2	Average Fade Duration (AFD)	60
3.11	Diversity	61
3.12	Selection Combining (SC)	62
3.13	Equal Gain Combining (EGC)	64
3.14	Maximum Ratio Combining (MRC)	65
3.14.1	BER with MRC Diversity	65
3.14.2	Diversity capacity analysis	66
3.15	Conclusion	66
Bibliography		68

4 Near-field MIMO BER and EVM Measurements in Metal Enclosure 71

4.1	Reverberation Chamber	71
4.2	Q-factor and Rician Factor Measurement	72
4.3	Near-field Measurements using QPSK Receiver	73
4.3.1	Automatic Gain Control	73

4.3.2	Transmit and Receive RC Filter	75
4.3.3	Coarse Frequency Compensation	76
4.3.4	Symbol Synchronization	78
4.3.5	Fine Frequency Compensation	80
4.3.6	Preamble Detector	82
4.3.7	Frame Synchronization	83
4.3.8	Phase Ambiguity Resolution	85
4.4	BER and EVM Measurements	86
4.5	MIMO BER and EVM Measurements	87
4.6	MIMO BER and EVM Measurements in Presence of MIMO Inter- ference	97
4.7	BER Measurement in Empty and Loaded Metal Enclosure in Pres- ence of MIMO Interference	102
4.8	2x2 MIMO BER and EVM Measurements in NLOS	109
4.9	Conclusion	115
Bibliography		116
 5 Near-field MIMO Wireless Communication		118
5.1	MIMO System Model	118
5.2	MIMO Channel Capacity with Unknown CSI at Transmitter . . .	119
5.2.1	Ergodic Capacity	120
5.2.2	Wishart Matrix	120
5.3	MIMO System Correlation Matrix	123
5.4	Kronecker Model	125
5.5	Semi-correlated MIMO Channel Capacity	127
5.5.1	MIMO Channel Capacity in Presence of Tx Correlation . .	127
5.5.2	MIMO Channel Capacity in Presence of Rx Correlation . .	128
5.6	Diagonally-Correlated MIMO Channels	129
5.7	Keyhole or Degenerate MIMO Channels	132

5.7.1	Single Keyhole Channel	132
5.7.2	Multi-keyhole MIMO Channel	133
5.8	Antenna Correlation	135
5.8.1	Uniform Correlation	136
5.8.2	Exponential Correlation	136
5.8.3	Bessel Correlation	136
5.9	MIMO Performance Metrics	136
5.9.1	Diversity Gain	137
5.9.2	Diversity Measure	137
5.9.3	K-factor	138
5.10	Experimental Setup I	139
5.11	Measurement Results	140
5.12	Conclusion	148
Bibliography		150
 6 Near-field MIMO in Metal Enclosure		 155
6.1	MIMO Channel Sounding	155
6.2	Channel Impulse Response Measurement	156
6.3	2x2 Near-field MIMO Channel Matrix $\mathbf{H}$	157
6.4	MIMO System Model	163
6.5	Rank and Condition Number	163
6.6	Near-field Capacity with Polarization Diversity	164
6.6.1	Cross-polarized and co-polarized channels	165
6.6.2	Spatial MIMO	165
6.6.3	Cross-polar Discrimination (XPD) factor	165
6.6.4	LOS/NLOS MIMO capacity	166
6.7	MIMO Measurements in Metal Enclosure	167
6.8	Conclusion	172

Bibliography	176
7 Near-field MIMO Capacity Measurements	180
7.1 Introduction	180
7.2 Wideband Channel capacity	181
7.3 Conclusion	186
Bibliography	188
8 Conclusion and Future Work	190
8.1 Conclusion	190
8.2 Future Work	192

Chapter 1

Introduction

1.1 Wireless Chip-to-Chip communication

Wireless chip-to-chip communication has gained attention from researchers worldwide due to increased transistor integration and availability of large number of cheap devices and thrust for high data rates. The need for short-range communication such as board-to-board communication, chip-to-chip communication and intra-chip communication has increased. There are many technological advancements yet to progress in the areas of antennas, materials and channel modeling for the realization of wireless chip-to-chip communication. Wireless chip-to-chip communication is a complex propagation environment with all possible propagation scenarios similar to real-world mobile communication. However, there are many technical challenges to realize wireless chip-to-chip communications. Extensive channel modeling and wireless communication system measurements tailored to wireless chip-to-chip communication are necessary. Recently, channel modeling measurement campaigns have investigated wireless chip-to-chip channels in desktop size metal enclosure of dimension $30.5\text{ cm} \times 30.5\text{ cm} \times 9.6\text{ cm}$ [1] [2] where the authors have measured power delay profiles and frequency correlation functions. The channel modeling has been performed at 300 GHz in Terahertz spectrum. THz part of the electromagnetic spectrum can offer high bandwidth to support high data-rate wireless links. however, it is susceptible to molecular

absorption and high path loss because of blockages. The channel modeling is performed using single Tx and Rx antennas in metal enclosure. However, there is need to investigate THz communication systems in metal enclosure for targeted end-to-end system performance. This is a very challenging task not only from hardware perspective but also baseband processing of real-time data. High data-rates and spectral-efficiency and energy-efficient modulation and PHY layer techniques along with usage of multi-antenna techniques can be used for WC2C links. The wireless links on chip or among multiple chips can take place in highly reflective environment or in complex electromagnetic environment in presence of materials with different electromagnetic properties. In addition spectral-efficient modulation techniques the increased spectral-efficiency can be achieved by using multiple antennas at the transmit and receive side. Authors in [3] have designed RF data transmission for inter-chip communication in the near-field. the authors have designed monopole antenna and done experimental demonstration using VNA. The authors have measured BER values of 10^{-4} and 10^{-10} at distances of 0.3175 cm and 0.635 cm respectively. In another study [4] authors have demonstrated integrated OOK transceiver for wireless chip-to-chip communication and they have demonstrated a BER of 10^{-12} for transmit power of 6.8 dBm. The measurements were conducted at 210 GHz in the far-field at a distance of 1 cm using integrated single transmit and single receive antenna. The introduction of new knowledge domain of intelligent reflecting surfaces can foster the implementation aspects of wireless chip-to-chip communication [5]. Intelligent reflecting surfaces consists of low-cost passive reflecting elements which can change the amplitude and/or phase of the impinging EM waves [6][7]. The progress in the miniaturization of the passive elements and increased phase-shift resolution can increase the performance of wireless chip-to-chip channels. Transistor integration continue every year. This is predicated by Moor's law as shown in Fig.1.1. Moore's law says that the number of transistors in integrated circuits (ICs) doubles every two year. This means more chips shall be available to perform specific task in many of

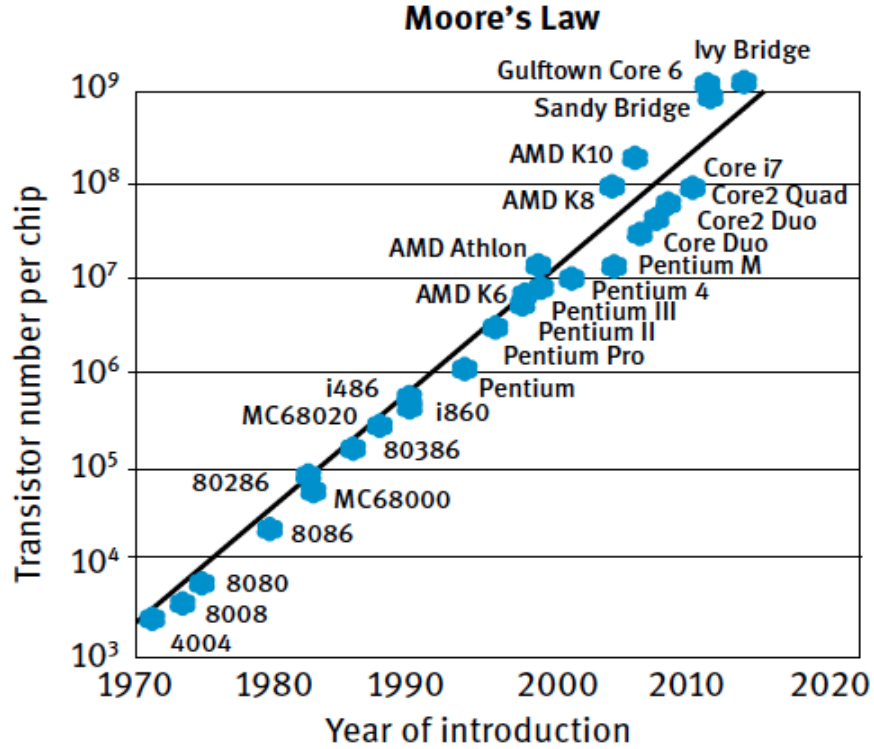


Figure 1.1: Moore's Law [8]. Moore's law introduced by Gordon Moore in 1965 and it's valid until today. Lately, Moore's law has remained in news highlights that it's going to be dead. However, chip maker companies have kept the Moore's alive.

the devices that we normally use which points out to internet of things revolution or device-to-device communication. In this condition, information transfer can take place near-field directly with little intervention from access point.

1.2 Near-field and Far-field Communication

Near-field of the antenna is also known as reactive near-field and is determined with the help of regions surrounding the antennas. In terms of distance near-field exists less than the distance $2D^2/\lambda$ [9]. Where D is the largest dimension of radiating antenna. This is true for electrically-large antennas whose linear dimension is greater than the distance $\lambda/2$. Whereas for electrically small antennas whose dimension is less than $D < \lambda/2$ the near-field boundary at distance R is defined as $R \gg \lambda$. The field surrounding antenna can be divided into three main regions:

- Reactive near-field

- Radiating near-field or Fresnel region
- Far-field or Fraunhofer region

1.2.1 Far-field or Fraunhofer Region

Traditionally, communication for mobile cellular communication, wireless local area networks, Bluetooth, and Radar takes place in far-field. However, the near-field communication using inductors, capacitors, and RF antennas has gained attention for wireless chip-to-chip communication [10][11][12][13][14][15]. Previously inductive and capacitive coupling has been used for low data-rate wireless communication at very short distances. However, near-field communication at chip-scale which requires high data rate transmission in compact and complex environments i.e. metal enclosures. In far-field the radiation dies off as the inverse of the square of a distance from antenna i.e. $1/R^2$. In this region electric $\mathbf{E}$ and magnetic field $\mathbf{H}$ are orthogonal to each other and to the direction of propagation. The conditions to be in far-field are as follows:

- $R < \frac{2D^2}{\lambda}$ which means the distance should be less than twice the square of the linear dimension of antenna divided by operating wavelength
- $R \gg D$: this condition means the distance should be much greater than the linear-dimension of radiating antenna
- $R \gg \lambda$: this condition implies the distance should be much greater than the operating wavelength.

1.2.2 Reactive near-field

It exists in the vicinity of antenna. In this region the fields are dominantly reactive and the $\mathbf{E}$ field and $\mathbf{H}$ field are out of phase by 90° to each other. Boundary for near-field is [9]:

$$R < 0.62 \sqrt{\frac{D^3}{\lambda}} \quad (1.1)$$

There is also another field known as the radiative near-field region. It's also known as the Fresnel region. Commonly it's a region between reactive near-field and far-field and its boundary can be given as [16]:

$$0.62\sqrt{\frac{D^3}{\lambda}} < R < \frac{2D^2}{\lambda} \quad (1.2)$$

There are many applications that exploit near-field of antenna such as near-field communication for body-implantable devices, near-field communication for wireless information and power transfer [17], near-field links for high data-rate transfer [18].

1.2.3 Near-field Magnetic Communication

Near-field magnetic communication works in the near-field region for data transmission. It works on the principle of magnetic induction communication where primary coil is excited which generates magnetic field and data transmission takes place via magnetic induction when a secondary magnetic coil is placed in the near-field of primary coil. Magnetic inductive communication is used for low-rate communication for various applications such as wireless information and power transfer for biomedical implantable devices [19][20], magnetic-induction based communication system for underground applications [21][22][23], magnetic-induction communication in underwater environment [24][25][26]. Magnetic induction communication is useful for low-rate wireless communication, hence it can be useful for control information transfer for wireless chip-to-chip communication. Authors in [21] have created USRP based measurement setup for the evaluation of magnetic induction communication for low-rate underground wireless sensor network. Authors have measured received power, path loss, and packet error rate of magnetic induction communication link through materials with high loss and harsh EM environment. Another application for magnetic induction communication is demonstrated for underwater communication. Magnetic induction com-

munication in propagation medium has advantage of negligible delay, predictable channel response. However, it has disadvantage of high path loss, alignment errors and low data rate which can only be used for low-rate control and feedback signals in context of wireless chip-to-chip communication. A passive RFID tag when inductively coupled can obtain power and can transmit data to inductively coupled RFID reader. NFC is short-range communication protocol for low-rate data exchange at operating frequency of 13.56 MHz. There are various NFC modes under which NFC communication takes place such as reader/write mode, card emulator mode and peer-to-peer mode. NFC can work with existing RFID infrastructure using card-emulator mode [27][28]. All these technologies based on magnetic induction communication have been used for low-rate links in the near-field and the far-field. Another very relevant study on near-field magnetic induction MIMO communication for high-data rate has been studied by authors in [29]. Authors have confirmed NFMI MIMO communication is similar to that achieved with conventional far-field RF MIMO systems. Further, authors are of the opinion that their technique does not require complicated MIMO signal processing such as SVD decomposition and channel estimation. The main contribution of authors is reduction of cross-talk, decoupling parallel streams and the usage of less complex signal processing schemes. The authors have created near-field magnetic- induction 2x2 MIMO communication and investigated near-field MIMO capacity at operating frequency of 13.56 MHz.

1.2.4 BER Measurement of Intra-chip Communication

BER measurement is one of the KPIs that is used for the data communication between chips. Y.P.Zhang has performed BER measurements for wireless intra-chip communication in in SISO mode [30]. The author has shown that BER performance decreases with the distance and the transmission data rate. Authors have shown BER less than 10^{-3} at a distance of 20 mm for data rate of 2 Gbps. For a fixed data rate of 2 Gbps, the data rate decreases with the increase in

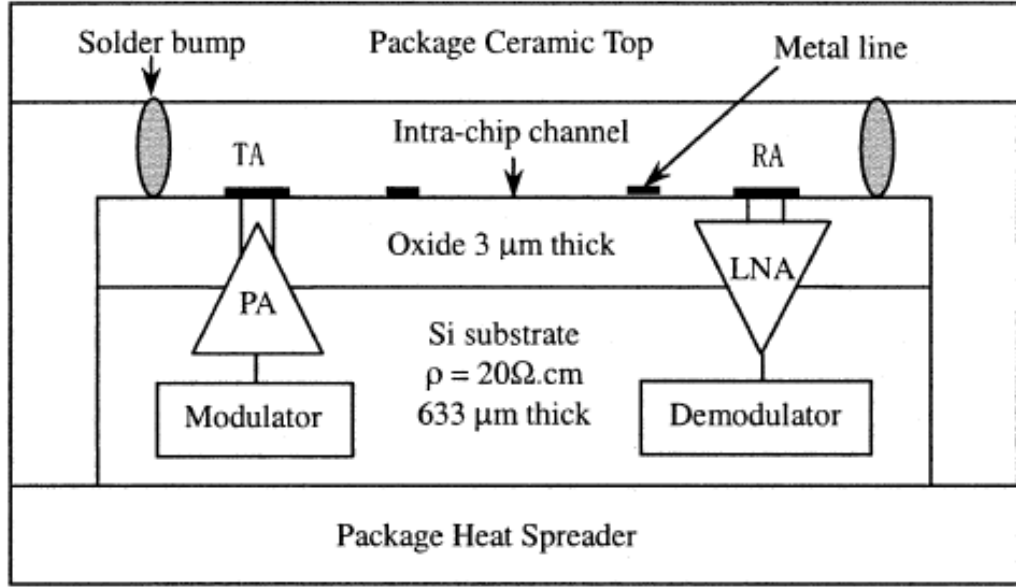


Figure 1.2: BPSK based intra-chip wireless communication architecture [30].

distance between Tx and Rx. Additionally, they have shown BER is degraded as the data rate is increased. Authors have shown BER measurement for SISO mode of intrachip communication which suffers from NLOS link resulting in poor signal quality and BER measurements. In contrast, MIMO communication in similar setting shall exploit NLOS and rich-scattering which increases signal quality by virtue of diversity and total system throughput using spatial multiplexing.

Fig.1.2 shows a typical chip structure that shows the package ceramic top, package heat spreader, metal lines, and solder bumps. Additionally, it also shows some wireless transmitter and receiver components such as power amplifier, low noise amplifier, modulators, and demodulators. It can be seen that this is typical case of a complex intrachip wireless propagation environment. This type of environment has the potential to exhibit rich multipath scattering because of the presence of solders balls, metal lines, heat spreaders, and top ceramic package. Additionally, in this environment propagation scenarios can arise that can be responsible for single MIMO keyhole or multi-keyhole channels as well as diagonally correlated channels. Floyd et al. has used wireless interconnect for clock distribution and has performed SNR estimation for on-chip wireless clock distribution

[31]. Moreover, authors have concluded that the switching noise and thermal noise has been reported to be dominant in such scenarios.

1.3 Aims and Objectives

Wireless chip-to-chip communication has gained recent attraction from researchers within computer architecture, chip manufacturers, computing and wireless communication domains. It has scope for small form-factor communication such as multi-chip modules, board-to-board communication and processor to memory links. Similar to more popular cellular and WLAN communication, this form of communication relies on fundamental understanding of radio in single and multi-antenna settings. The KPI like MIMO channel capacity, EVM and BER measurement are important to understand such form of wireless communication in rich propagation environment. Small-scale fading is used to differentiate MIMO correlated channels, therefore, it is important to understand MIMO channel capacity. Ideally, two MIMO channels should have equal power but independent small-scale fading.

The aims and objectives of this thesis are as follows:

- Aims
 - To develop 2x2 MIMO software-defined-radio based testbed for the measurement campaigns in the reverberation chamber and metal enclosures in the near-field. Short-distance propagation measurements under different channel conditions in LOS and NLOS are performed in a mode-stirred metal enclosure to address problems in mode-stirred wireless chip-to-chip communication channels.
- Objectives
 - Perform extensive channel measurement and 2x2 MIMO BER and

EVM measurements in the reverberation chamber using a software-defined-radio testbed.

- Perform near-field MIMO measurements in the reverberation chamber under different multipath fading scenarios and radio propagation conditions.
- Perform MIMO channel measurements under different stirring and loading scenarios. Present statistical characterization of the wireless chip-to-chip channel.
- Assess diversity performance of wireless chip-to-chip channel in a mode-stirred metal enclosure. Present statistical characterization of second-order statistics of the wireless chip-to-chip channel in different propagation environments.
- Measure channel parameters and present statistical model for example Rician K-factor, path-loss, path-loss exponent, RMS delay spread, coherence bandwidth.
- Measure wideband MIMO channel capacity in a mode-stirred metal enclosure in rich isotropic multipath fading scenario.
- Perform MIMO channel measurement with polarization diversity.
- Demonstrate near-field MIMO wireless communication system and perform near-field MIMO BER and EVM measurements in empty metal enclosure and when the metal enclosure is loaded with absorbers.
- Perform 2x2 MIMO BER and EVM measurements in metal enclosure in presence of MIMO interference.

1.4 Contribution

The main contribution is the development of SDR based MIMO testbed for near-field channel measurement and characterization of the wireless chip-to-chip channel in metal enclosure. 2x2 MIMO BER and EVM measurements are performed

under different loading conditions. Demonstration of digital wireless communication receiver and BER measurement in the near-field reverberant environment. Empirical measurements and characterization of near-field MIMO wireless chip-to-chip channel in a mode-stirred metal enclosure. Using 2x2 MIMO measurement setup two USRP channel BER and EVM measurements are performed in the presence of MIMO interference. It is shown that the EVM and BER degrades as MIMO interference increases. Additionally, other channel parameters such as RMS delay spread, coherence bandwidth are measured for different transmit and receive inter-element distances. Finally, near-field MIMO channel capacity measurement and the effect of transmit correlation and receive correlation on MIMO capacity.

1.5 Thesis Structure Outline

The thesis is divided into seven chapters starting from introduction to conclusion and future work. More specifically, this thesis consists of the following chapters with brief introduction of each chapters.

Chapter 1 is introduction of the thesis research issues in the area of chip-to-chip communication. It presents a review of the fundamentals of wireless chip-to-chip or wireless inter-core communication. It reviews existing forms of chip-to-chip communication and interconnects. It also reviews recent literature in the area of wireless chip-to-chip communication in a mode-stirred environment. It also gives a brief review of previous measurements conducted in a confined environment including ultra-wideband measurements in PC casings and compact devices. The near-field environment shall play a role for densely deployed high data-rate links which shall be part of the 6G THz wireless communication revolution. This chapter presents the aims and objectives of the thesis. It also presents the contribution of the thesis and outlines the thesis structure.

Chapter 2 is about the software-defined-radio based testbed and usage of Ettus USRP radios and signal processing libraries such as GNU Radio and MATLAB

for the development of baseband communication system models. This chapter gives a detailed overview and description of the measurement setup, propagation environments, and other instrumentation used during measurements. Two baseband development waveforms are used that are Simulink and GNU Radio. Both baseband development waveforms were used depending on the quick waveform accessibility, ease of use, latency and portability.

Chapter 3 highlights measurement, modeling, and diversity channels for chip-to-chip channels in mode-stirred metal enclosures. Here different diversity techniques such as selection combining, equal gain combining, and maximum ratio combining are discussed. It's important to know fundamental channel parameters that are useful for the design of wireless chip-to-chip communication system. The parameters such as RMS delay spread, coherence bandwidth, Doppler spread, Rician K-factor, and path loss are discussed. These parameters are useful for channel modeling and characterization and the development of a digital wireless communication system.

Chapter 4 is about 2x2 MIMO QPSK receiver for BER and EVM measurements in metal enclosure in different fading conditions and in the presence of different levels of interference. The baseband model of the QPSK digital receiver is explained. The importance of frequency synchronization and timing synchronization in communication systems is stressed. Therefore digital synchronization in the baseband model and the synchronization signals from Octoclock-G are compared. The design stages of the QPSK digital receiver is explained with the help of measurements. BER measurements and EVM measurements are conducted using SDR-based QPSK digital receiver in different propagation scenarios. 2x2 MIMO BER and EVM measurements of transmit diversity and receive diversity techniques are also presented.

Chapter 5 is about near-field MIMO wireless communication. This details MIMO system parameters and discusses the information-theoretic perspective of MIMO communication. MIMO capacity of doubly-correlation channels is dis-

cussed i.e correlation present at the transmitter as well as at the receiver. Eigenvalue decomposition of the measured Gramian matrix $\mathbf{H}\mathbf{H}^H$ is presented. For proper MIMO capacity gain MIMO channel should not be rank-deficient. Therefore, relevant MIMO channel parameters such as condition number and rank of the channel are discussed. Additionally, different forms of MIMO correlation are discussed to address the effect of correlation on MIMO channel capacity.

Chapter 6 is about MIMO measurements using polarization diversity and the results on MIMO capacity with Tx and Rx correlation. Polarization diversity is helpful from practical perspectives if the transmitter and receiver are to be integrated in limited form-factor devices. Polarization diversity helps reduce transmit and receive correlation, however it results in power gain loss of subchannels in MIMO channel matrix because of orthogonal polarization. MIMO channels in polarization diversity exhibit non-identical fading subchannels of MIMO channel. Cross-polarization discrimination (XPD), cross-polarization coupling (XPC) and cross-polarization ratio (XPR) are discussed. The effect of XPD and XPC on MIMO capacity is important to understand as the XPD may affect MIMO channel capacity differently in different propagation environment and at different SNRs. Measurement results are presented to support the usage of polarization diversity in near-field MIMO communication. BER and EVM measurements are performed using polarization diversity in LOS and NLOS environment.

Chapter 7 is about wideband MIMO capacity measurements in a mode-stirred metal enclosure in different propagation scenarios and channel conditions. The measurements are performed under LOS and NLOS channel conditions, in the presence of Tx correlation and Rx correlation.

1.6 Conclusion

Wireless chip-to-chip communication has been investigated with state-of-the-art literature on wireless chip-to-chip communication. Wireless chip-to-chip communication in the light of Moore's law has been explained. possible spectrum candi-

dates and previous prototypes such as near-field magnetic induction communication and measurements for wireless network on chip are discussed. Cutting-edge issues of wireless chip-to-chip communication are explored. Possibilities of MIMO keyhole channels in wireless chip-to-chip or board-to-board communication use cases have been discussed and relevant issues are highlighted. In addition to issues of wireless chip-to-chip communication, the possibility of near-field communication is assessed and the research foundation has been established to develop SDR based near-field MIMO prototype which is discussed in chapter 2.

Bibliography

- [1] Jinbang Fu, Prateek Juyal, and Alenka Zajić. Modeling of 300 ghz chip-to-chip wireless channels in metal enclosures. *IEEE Transactions on Wireless Communications*, 19(5):3214–3227, 2020.
- [2] Jinbang Fu, Prateek Juyal, and Alenka Zajić. Thz channel characterization of chip-to-chip communication in desktop size metal enclosure. *IEEE Transactions on Antennas and Propagation*, 67(12):7550–7560, 2019.
- [3] Bahaa Radi, Ajaypal Singh Dhillon, and Odile Liboiron-Ladouceur. Demonstration of inter-chip rf data transmission using on-chip antennas in silicon photonics. *IEEE Photonics Technology Letters*, 32(11):659–662, 2020.
- [4] Siavash Moghadami, Fleura Hajilou, Priyanka Agrawal, and Shahab Ardalan. A 210 ghz fully-integrated ook transceiver for short-range wireless chip-to-chip communication in 40 nm cmos technology. *IEEE Transactions on Terahertz Science and Technology*, 5(5):737–741, 2015.
- [5] Mohammadreza F Imani, Sergi Abadal, and Philipp del Hougne. Smart on-chip electromagnetic environment. *arXiv preprint arXiv:2109.03284*, 2021.
- [6] Marco Di Renzo, Alessio Zappone, Merouane Debbah, Mohamed-Slim Alouini, Chau Yuen, Julien De Rosny, and Sergei Tretyakov. Smart radio environments empowered by reconfigurable intelligent surfaces: How it works, state of research, and the road ahead. *IEEE Journal on Selected Areas in Communications*, 38(11):2450–2525, 2020.
- [7] Linglong Dai, Bichai Wang, Min Wang, Xue Yang, Jingbo Tan, Shuangkaisheng Bi, Shenheng Xu, Fan Yang, Zhi Chen, Marco Di Renzo, et al. Reconfigurable intelligent surface-based wireless communications: Antenna design, prototyping, and experimental results. *IEEE Access*, 8:45913–45923, 2020.

- [8] Karsten König and Andreas Ostendorf. *Optically induced nanostructures: biomedical and technical applications*. Walter de Gruyter GmbH & Co KG, 2015.
- [9] John L Volakis. *Antenna engineering handbook*. McGraw-Hill Education, 2007.
- [10] Yen-Sheng Chen, Shih-Yuan Chen, and Hsueh-Jyh Li. Analysis of antenna coupling in near-field communication systems. *IEEE Transactions on antennas and propagation*, 58(10):3327–3335, 2010.
- [11] Hristomir H Yordanov and Peter Russer. Integrated on-chip antennas for chip-to-chip communication. In *2008 IEEE Antennas and Propagation Society International Symposium*, pages 1–4. IEEE, 2008.
- [12] Seahee Hwangbo, Yong-Kyu Yoon, and Aric B Shorey. Millimeter-wave wireless chip-to-chip (c2c) communications in 3d system-in-packaging (sip) using compact through glass via (tgv)-integrated antennas. In *2018 IEEE 68th Electronic Components and Technology Conference (ECTC)*, pages 2074–2079. IEEE, 2018.
- [13] Xavier Timoneda, Albert Cabellos-Aparicio, Dionysios Manassis, Eduard Alarcón, and Sergi Abadal. Channel characterization for chip-scale wireless communications within computing packages. In *2018 Twelfth IEEE/ACM International Symposium on Networks-on-Chip (NOCS)*, pages 1–8. IEEE, 2018.
- [14] Yazan Al-Alem, Ahmed A Kishk, and Raed Shubair. Wireless chip to chip communication link budget enhancement using hard/soft surfaces. In *2018 IEEE Global Conference on Signal and Information Processing (GlobalSIP)*, pages 1013–1014. IEEE, 2018.
- [15] Ihsan El Masri, Thierry Le Gougec, Pierre-Marie Martin, Rozenn Allanic, and Cedric Quendo. Electromagnetic characterization of the intrachip prop-

- agation channel in *ka*-and *v*-bands. *IEEE Transactions on Components, Packaging and Manufacturing Technology*, 9(10):1931–1941, 2019.
- [16] Constantine A Balanis. *Antenna theory: analysis and design*. John wiley & sons, 2015.
 - [17] Christoph Degen. Inductive coupling for wireless power transfer and near-field communication. *EURASIP Journal on Wireless Communications and Networking*, 2021(1):1–20, 2021.
 - [18] Sebastian Stoecklin, Elias Rosch, Adnan Yousaf, and Leonhard Reindl. Very high bit rate near-field communication with low-interference coils and digital single-bit sampling transceivers for biomedical sensor systems. *Sensors*, 20(21):6025, 2020.
 - [19] Han-Joon Kim, Hiroshi Hirayama, Sanghoek Kim, Ki Jin Han, Rui Zhang, and Ji-Woong Choi. Review of near-field wireless power and communication for biomedical applications. *IEEE Access*, 5:21264–21285, 2017.
 - [20] Werner Meskens, Erika J Van Baelen, and Tony M Nygard. Magnetic induction communication system for an implantable medical device, October 20 2015. US Patent 9,166,655.
 - [21] Xin Tan, Zhi Sun, and Ian F Akyildiz. Wireless underground sensor networks: Mi-based communication systems for underground applications. *IEEE Antennas and Propagation Magazine*, 57(4):74–87, 2015.
 - [22] Shih-Chun Lin, Ian F Akyildiz, Pu Wang, and Zhi Sun. Distributed cross-layer protocol design for magnetic induction communication in wireless underground sensor networks. *IEEE Transactions on Wireless Communications*, 14(7):4006–4019, 2015.
 - [23] Zhi Sun and Ian F Akyildiz. Magnetic induction communications for wireless underground sensor networks. *IEEE transactions on antennas and propagation*, 58(7):2426–2435, 2010.

- [24] Hongzhi Guo, Zhi Sun, and Pu Wang. Multiple frequency band channel modeling and analysis for magnetic induction communication in practical underwater environments. *IEEE Transactions on Vehicular Technology*, 66(8):6619–6632, 2017.
- [25] Ian F Akyildiz, Pu Wang, and Zhi Sun. Realizing underwater communication through magnetic induction. *IEEE Communications Magazine*, 53(11):42–48, 2015.
- [26] Yuzhou Li, Shengnan Wang, Cheng Jin, Yu Zhang, and Tao Jiang. A survey of underwater magnetic induction communications: Fundamental issues, recent advances, and challenges. *IEEE Communications Surveys & Tutorials*, 21(3):2466–2487, 2019.
- [27] Vedat Coskun, Busra Ozdenizci, and Kerem Ok. A survey on near field communication (nfc) technology. *Wireless personal communications*, 71(3):2259–2294, 2013.
- [28] Vedat Coskun, Busra Ozdenizci, and Kerem Ok. The survey on near field communication. *Sensors*, 15(6):13348–13405, 2015.
- [29] Han-Joon Kim, Jinho Park, Kyoung-Sub Oh, Jihwan P Choi, Jae Eun Jang, and Ji-Woong Choi. Near-field magnetic induction mimo communication using heterogeneous multipole loop antenna array for higher data rate transmission. *IEEE Transactions on Antennas and Propagation*, 64(5):1952–1962, 2016.
- [30] Yue Ping Zhang. Bit-error-rate performance of intra-chip wireless interconnect systems. *IEEE Communications letters*, 8(1):39–41, 2004.
- [31] Brian A Floyd, Chih-Ming Hung, et al. Intra-chip wireless interconnect for clock distribution implemented with integrated antennas, receivers, and transmitters. *IEEE Journal of Solid-State Circuits*, 37(5):543–552, 2002.

Chapter 2

SDR Based Near-field MIMO Testbed

2.1 Software Defined Radio

Software-defined-radio is a re-programmable and reconfigurable platform that can be used for the development and demonstration of state-of-the-art wireless communication systems, for example 5G new radio (NR), wireless local area network (WLAN) [1], Zigbee, long range wide area network (LoRaWAN) [2], Radio Frequency identification (RFID) tags and backscatter communication [3][4][5], spectrum sensing and detection in TV white space etc. This SDR can be programmed to perform channel measurement and characterization. This work demonstrates the potential of SDR for communication algorithm development and RF testing of wireless communication systems as well as for channel measurements and characterization. The main advantage of SDRs over conventional measurement equipment is their portable, re-programmable, and scalable characteristics. Due to these characteristics, the SDR based testbed can be quickly prepared and deployed in difficult propagation environments. Few examples are vehicle-to-vehicle (V2V) communication, Air-to-Ground, and Air-to-Air channel modeling in unmanned aerial vehicle (UAV) communication [6][7][8], Tunnels, and Gold mine communication [9]. In SDR baseband signal processing is carried on host PC using baseband development platforms such as GNU Radio or MATLAB. Additionally, digital upconversion and downconversion are performed at a high-performance

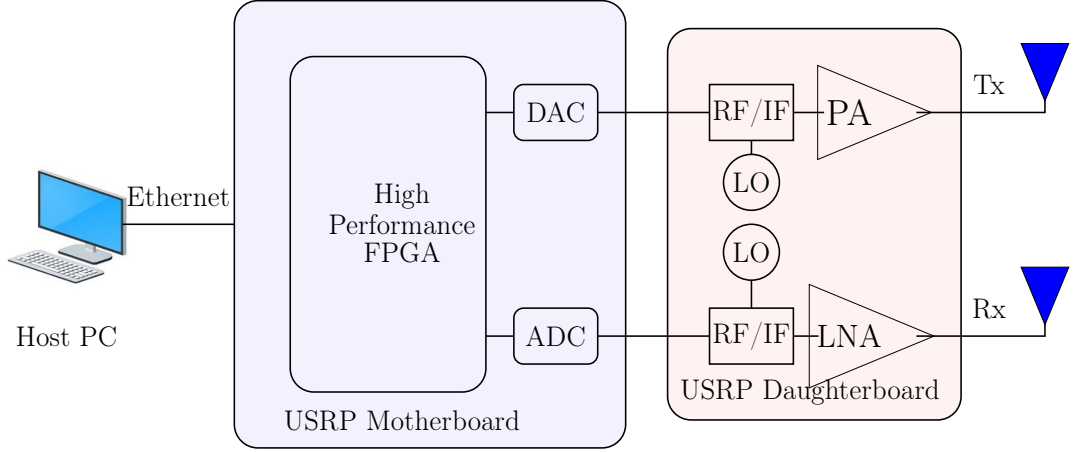


Figure 2.1: Software Defined Radio Schematic.

FPGA side as shown in Fig. 2.1. This architecture of SDR gives increased degree of freedom for the design of a reconfigurable wireless communication system. SDR has two main components one is motherboard and the other components is RF daughterboard. RF frequency range is defined by RF daughterboards. There are various RF daughterboards that can be interfaced with Ettus USRPs. For example UBX-160, UBX-40, TwinRx-80, WBX-120, SBX-120, CBX-120, CBX-40, LFTX-LFRX as shown in Fig.2.7c. These RF daughterboards have different bandwidth and RF frequency range. For example UBX-160 has maximum bandwidth of 160 MHz and its frequency ranges from 70 MHz to 6 GHz. LFRX-LFTX can receive signals from DC to 30 MHz. These RF cards can be useful for applications that use low frequency spectrum such as power-line communication and EMC emission testing and interference analysis.

Fig. 2.3 shows front view of USRP based 2x2 MIMO testbed. It shows two X310 USRPs front panel, two UBX-160 RF daughterboards per USRPs. Octoclock-G is used to supply 10 MHz frequency and 1 PPS time synchronization signals to both the USRPs. Fig. 2.4 shows top view of the testbed showing UBX-160 daughterboards, two SFP+ 10GigE ports, 1Gig Ethernet interface adaptor and RF daughterboard cables leading to the front panel of each USRP.

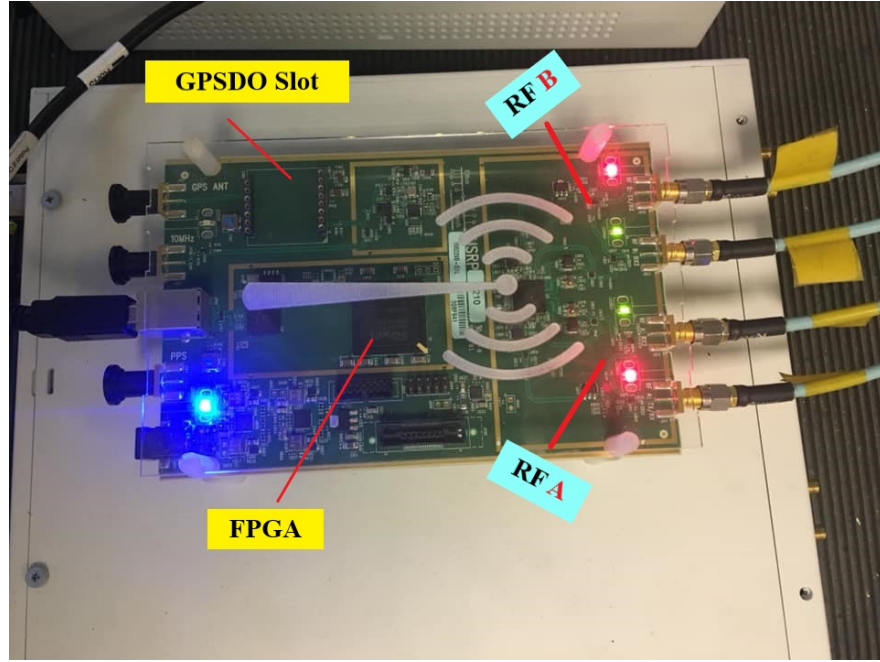


Figure 2.2: Top view of B210 USRP. This USRP allow 2x2 Full-duplex MIMO communication as it's visible from the two red TX LEDs and green RX LEDs which shows simultaneous operation of two transmitters and two receivers respectively.

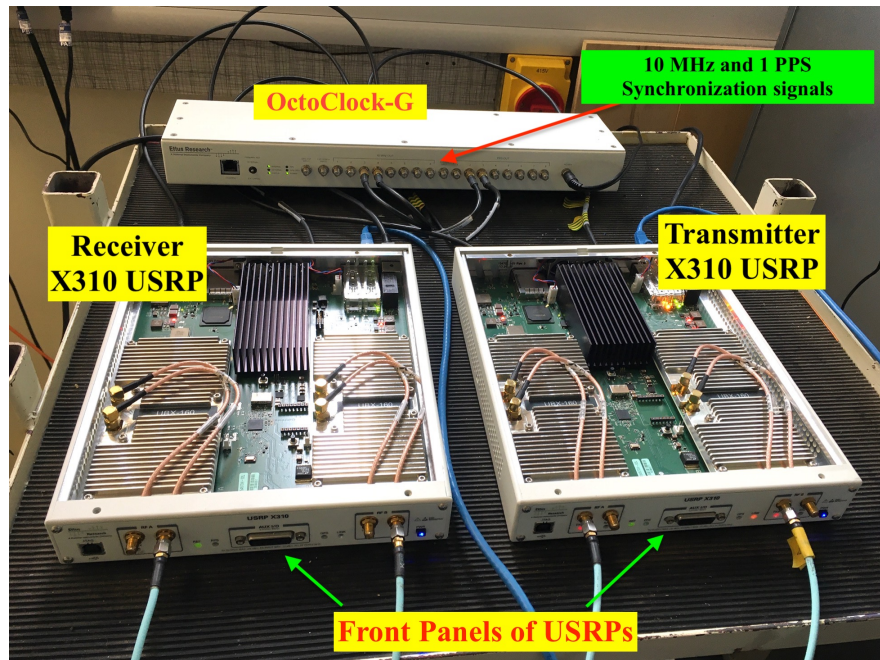


Figure 2.3: Front view of 2x2 MIMO measurement setup based on two X310 USRPs. The measurement setup uses two X310 USRPs with UBX-160. Each of the X310 USRPs is equipped with two UBX-160 RF cards and one OctoClock-G which supplies synchronization signals to each USRP.

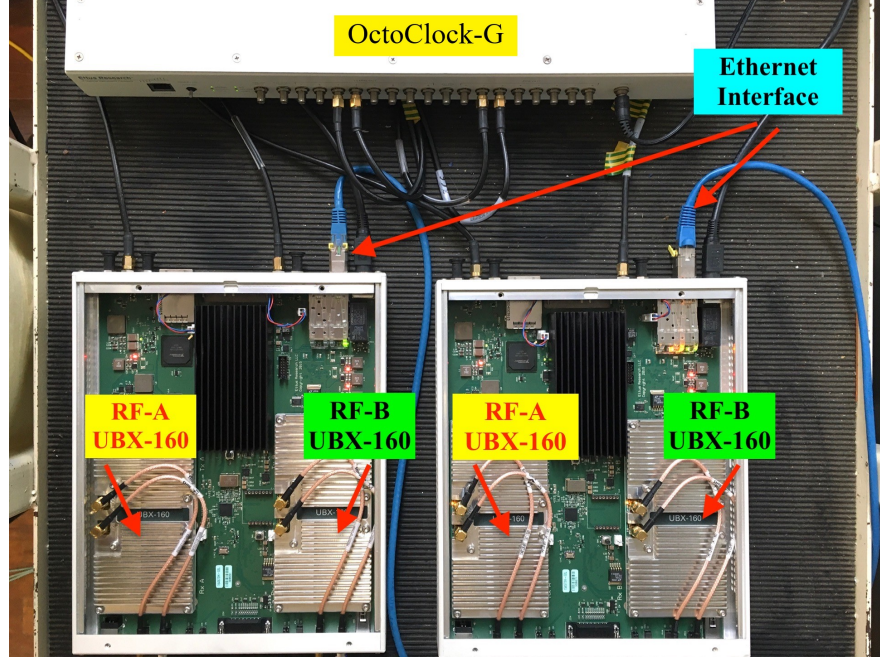


Figure 2.4: Top view of 2x2 MIMO measurement setup based on two X310 USRPs. The measurement setup uses two X310 USRPs with UBX-160. Each of the X310 USRPs is equipped with two UBX-160 RF cards and one OctoClock-G which supplies synchronization signals to each USRP.

2.2 SDR based Near-field Testbed

The near-field testbed is developed using two X310 USRPs and one B210 USRP for 2x2 near-field MIMO measurements. Measurements are performed both in the time-domain as well as in the frequency-domain using the frequency-swept approach. Measurements are performed using a time-domain approach i.e. transmitting orthogonal sequences like PN code sequence, Gold code sequence, and Zadoff-Chu sequence. These sequences are widely used for performing channel sounding and characterization of the wireless channel. In the frequency-swept way of channel sounding which is the same as to measure channel transfer function in frequency domain i.e. measure S21 parameters and forward transmission channel. USRPs are configured to be used for the measurement of S21 parameters by sweeping over a band of interests. USRPs are configured to measure forward transmission i.e. S21 parameters but not the S11 parameters. For S11 parameter measurement, an additional RF component known as circulator can be deployed to measure the reflection coefficient, however, this can reduce the efficiency of USRP based

testbed significantly. Previously, SDR-based near-field communication testbed has been used to conduct BER measurements using magnetically coupled coils [10]. However, the operating frequency was set to 10 MHz to excite magnetically coupled loops for BER measurements. Additionally, the propagation environment doesn't consider the rich-scattering environment or metal enclosure.

2.3 MIMO Channel Sounding

Complete sketch of LOS and NLOS MIMO channel sounder for channel measurements, BER and EVM measurements is shown in the Fig. 2.5 and Fig. 2.6 respectively. Two USRPs are connected to independent host PCs for signal generation and transmission over fading channel in the metal enclosure. A 2x2 MIMO system using independent USRPs as transmitter and receiver can also be developed by connecting both USRPs to single high-performance PC. This will require host PC to have dual NIC installed or one 10G Ethernet card with two interface ports. In this way data acquisition can be performed on the same PC and packet mismatched can be avoided because of different baseband model runtimes and hence increasing EVM and BER measurement accuracy for wireless communication system measurements. Channel transfer function is measured by sweeping LO frequency of Tx and Rx USRP simultaneously over a specified bandwidth. The Tx USRP and Rx USRP LO frequency can be tuned by inputting linearly increasing frequency points from Simulink or GNU Radio. However, this may result in oscillator stability issues. Alternative and more reliable way is to programmatically control Simulink and measure power at each frequency point. However, this procedure can take longer measurement time but it's more tractable way to measured received power at both the channels.

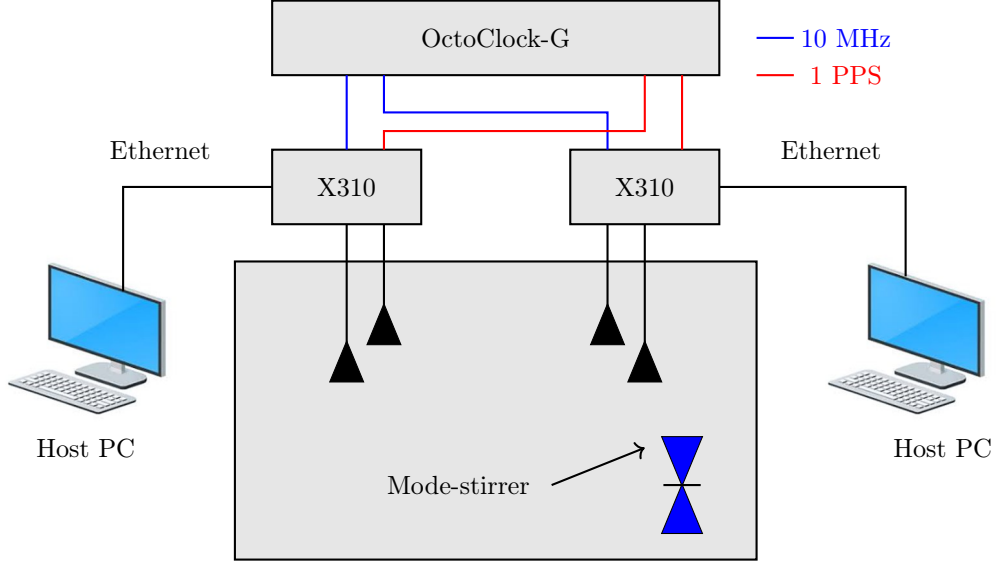


Figure 2.5: Complete sketch of LOS MIMO channel measurement setup in metal enclosure. The metal enclosure with dimensions of $h \times \ell \times w$ of $45\text{ cm} \times 37\text{ cm} \times 55\text{ cm}$ houses a mode-stirrer and two monopole antennas for the transmitter and two monopoles for the receiver.

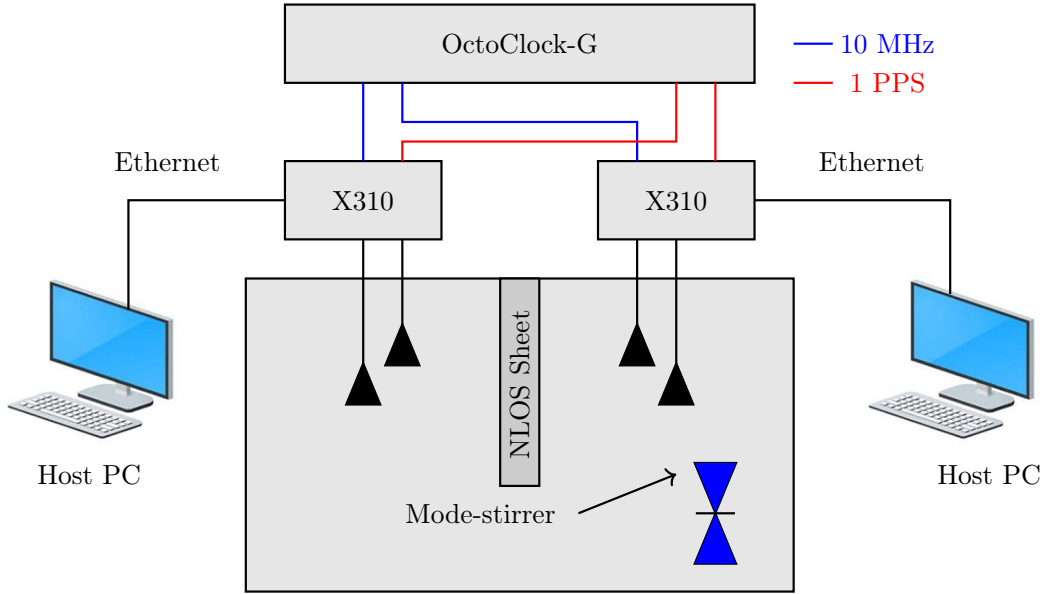


Figure 2.6: Complete sketch of NLOS 2x2 MIMO USRP-based channel measurement setup in metal enclosure. The metal enclosure with dimensions of $h \times \ell \times w$ of $45\text{ cm} \times 37\text{ cm} \times 55\text{ cm}$ houses a mode-stirrer, NLOS metal sheet and two monopole antennas for the transmitter and two monopoles for the receiver.

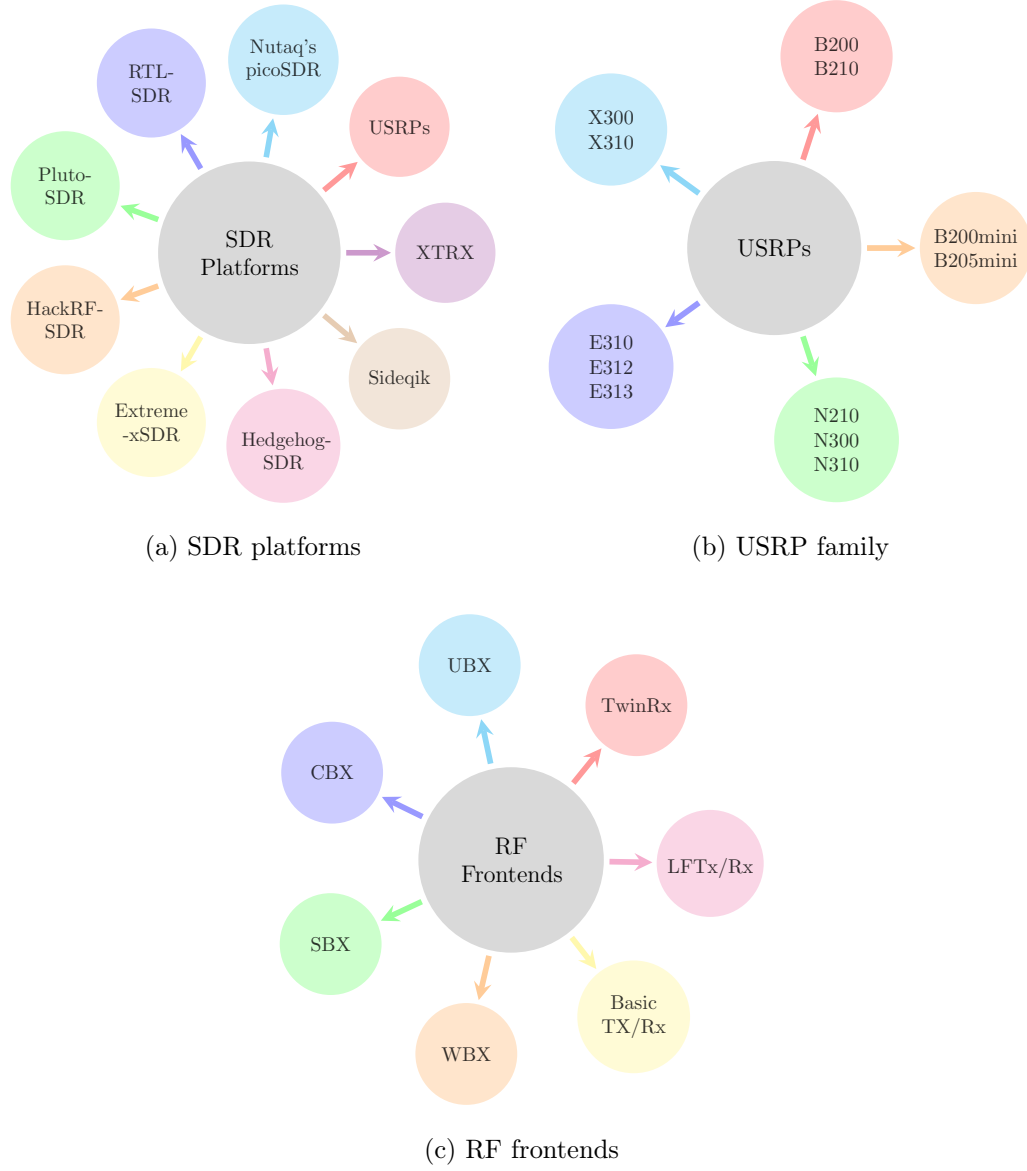


Figure 2.7: Classification of SDR platforms, USRPs and SDR RF front-ends. At the time of writing there are other SDRs with signal intelligence and high-performance GPUs such as artificial intelligence radio transceiver (AIR-T) from DeepWave Digital. For high bandwidth Cyan SDR from Per Vices offers RF bandwidth upto 3 GHz.

2.4 USRP based Testbed Limitations

USRP is a cost-effective SDR by Ettus Research for testing and developing wireless communication prototypes, however, it has some drawbacks that must be taken into consideration before initiating USRP based measurement campaign. Selection of SDR such as USRP depends on various factors such as dynamic range, RF frequency range, number of channels, onboard processing capability, high resource FPGA, bandwidth, interfacing with baseband development software, size, and the cost. All of the following issues which are mostly USRP related issues were faced during the testbed development and measurement campaigns.

Crosstalk

USRPs have an issue of cross-talk because of leakage from USRP antenna ports and the circuitry. Some USRP RF cards such as UBX-160 come with shielding to improve the performance. However, even with UBX-160 daughter-boards cross-talk issue is difficult to avoid especially when the transmission and reception take place on the same RF daughter-board. This type of cross-talk is known as intra-board cross-talk. Additionally, cross-talk from one USRP port to another USRP port was also observed. In order to avoid such cross-talk it's better to terminate port with an attenuator. In order to increase isolation two separate USRPs can be used and connected with the same PC. However, it shall increase the total cost of SDR based testbed. Single B210 USRP can be configured as 2x2 full-duplex MIMO communication, however, there will be problem of crosstalk among TX/RX0 and TX/RX1 port to RX1 and RX2 port. In terms of RF isolation, X310 USRP has an advantage over B210 USRP because of X310 has increased daughterboard shielding. It should be of special concern when the Tx and Rx antennas are mounted directly on USRP then higher interference should be expected in full-duplex mode. This is known as full-duplex cross-talk. Full duplex has problem of self-interference and various strategies are available in the literature for mitigating self-interference in prototyping full-duplex SDR prototypes

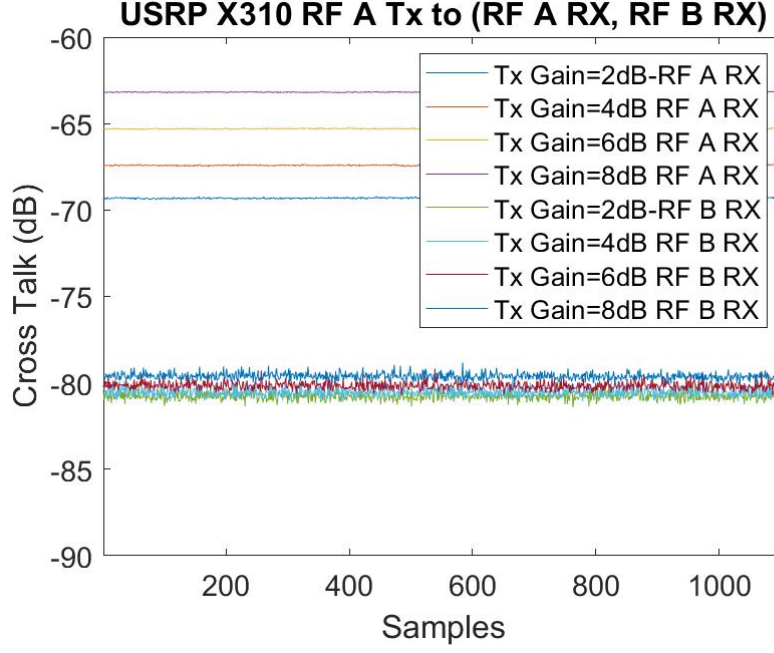


Figure 2.8: USRP X310 crosstalk from RF A TX to RF A RX and RF B RX.

[11]. Fig. 2.8 shows that there is more cross-talk when the receiver is configured on same UBX-160 RF daughterboard whereas the cross-talk is comparatively much lower at RX of RF B UBX-160 daughterboard. The low cross-talk values are observed for all values of Tx gains.

Figure 2.9 shows power measured at channel 1 and channel 2 of one USRPs using frequency sweep approach. Power is measured using loop back cable with 30 dB attenuator in order to avoid any possible damage to receiving RF daughterboard. The maximum input power of USRP X310 is -15 dBm and the maximum transmit power of X310 is +20 dBm. It can be seen the UBX-160 RF boards have same power over two frequency sweeps and it's also consistent with the UBX-160 data characterization sheet by Ettus Research. Conducted power measurements are useful to find out start and stop time of the sweeps since two USRPs don't start at the same time, however because of the external synchronization signals two USRPs lock with each other and produce correct streaming data but sweep 0 has not sufficient number of samples i.e. frequency points. Therefore, sweep 1 and sweep 2 has been considered to find out channel transfer function. There are two full-frequency sweeps each of 251 frequency points except the sweep 0 which has

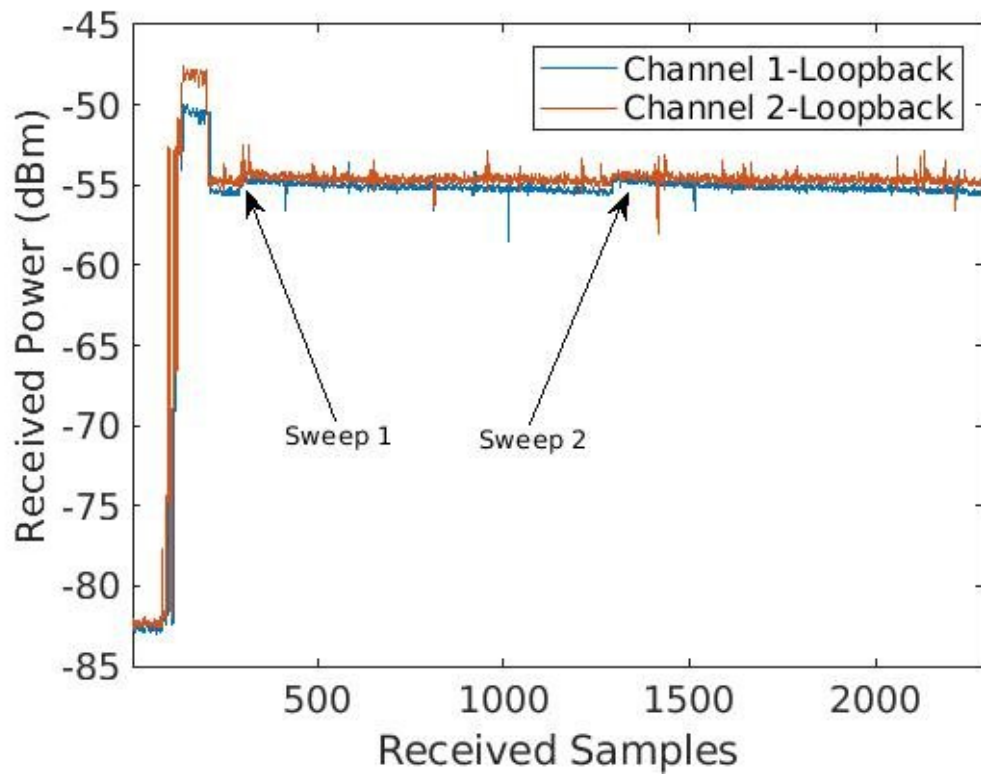


Figure 2.9: Loop-back power measurement using frequency sweep approach. The two USRPs were connected to same PC and were operated in frequency-swept approach. The Tx gains of each channel of USRP was set to 10 dB respectively and Rx gain of each channel of USRP was set to 0 dB. Tx channel of one USRP was connected directly to Rx channel of another USRP. Tx and Rx channels of the USRPs were connected with loop-back cable with 30 dB attenuator to avoid damage to the RF daughterboard.

insufficient samples and was wasted during USRP and Simulink initialization and start-up time. It is important to note that there is random startup delay when using X310 USRP and B210 USRP together for MIMO measurements. There are four updates per frequency points. LO stabilization is 0.1 second and the data update period is 0.025.

Bandwidth

Another limitation of the USRPs is their limited bandwidth which restricts USRPs to the applications that require less bandwidth. Therefore, UWB channel measurements and modeling that require high bandwidth can be challenging to perform with this device. Total available bandwidth is defined as a minimum of host bandwidth, RF daughtercard bandwidth, and the FPGA bandwidth. RF bandwidth is defined by RF daughterboard on USRP. For example, UBX-160 board can provide a bandwidth of 160 MHz. Therefore, UBX-160 can cover many wireless standards such as WLAN, Bluetooth, Zigbee, LoRA, LTE-A and 5G-NR communication. In order to fully utilize the RF daughterboard bandwidth it is necessary to use high-performance PC, high-speed interface of the PC, proper CPU governor configuration, power management, and interpolation/decimation factors as suggested by Ettus Research manual. It's also dependent on the complexity of waveform design which can give rise to underflow and overflow which results in loss of the measurement data thereby affecting reliability of the measurement results. In case of limited PC resources and recurrent underflow and overflow, the best solution could be to reduce the sampling rate and simulation measurement time.

Return Loss

USRPs can only measure forward transmission i.e. S_{21} parameters, these devices can't measure return loss i.e. S_{11} because they can't measure reflected wave on the same port. In order to measure reflected power more USRP ports should

be available for the measurement which directly affects the number of required RF daughterboards which increases testbed cost and reduces the efficiency of the measurement setup. For S11 measurements Circulator has also been suggested to be used with USRPs [12].

Calibration

USRPs are not calibrated devices as compared to VNA, therefore these devices require power calibration on a regular basis. Power calibration for 2x2 MIMO can be performed by connecting TX and RX port of USRP with 30 dB attenuator. For example for multi-channel phased array applications such as source localization, beamforming, and interference cancellation, multi-channel USRPs can be calibrated in time and frequency by supplying 1s PPS time and 10 MHz frequency reference signal from Octoclock-G. Additionally, at every retune of USRP initial random phase offset may be expected which can be corrected by supplying reference signal from transmitting USRP. USRPs can be configured for automated calibration of USRPs at additional cost of hardware. Automated calibration of SDR-based dual-polarized radar is given in [13].

DC component

It appears because of imperfect filtering in the radio transceiver. It's dominant at low SNR value. It is removed by using a carefully designed band-reject or notch filter in the baseband model. It is common problem in radio receivers and is prominent specially in direct-conversion SDR receivers.

Host PC resource

It's extremely important to choose host PC wisely for USRP based measurements especially if the application requirement is high data rate. USRP-based testbeds are known to have the problem of overflow and underflow which can cause data loss in the measurements that are sensitive to overflow and underflow such as

BER and EVM measurements. There are useful ways to find out when underflow or overflow occurs. One way is to activate underflow and overflow options in SDRuTransmitter and SDRuReceiver blocks in Simulink and another way is to check the terminal where letter U and O may be printed on the screen if there is any underflow or overflow. Underflow and overflow problem can be avoided by addressing issues like CPU power management, less complex baseband waveform, increasing buffer size, reducing visualizers in the baseband model, reducing data logging in the baseband model or data logging with decimation and time intervals.

2.5 USRP X310 Architecture

X310 USRP by Ettus Research is a high-performance multi-channel and scalable SDR which operates in full-duplex and supports 2x2 MIMO communication. It has two daughterboard slots, high-speed interfaces, GPSDO slot, GPIO, 1 GB DDR3 memory and a high performance Kintex-7 FPGA. At the time of writing NI Ettus Research has introduced new X410 USRP with four transmit and receive channel, on-board quad-core ARM-Cortex-A53 processing subsystem. It has also support for 100 GbE high-speed interface for data transfer. FPGA on X310 USRP is used to process digital baseband IQ samples. ADC/DAC on X310 USRP are used to rate transition and the data transfer takes place between host PC and X310 USRP over 1 Gig Ethernet cable or 10 Gig SFP+ interface. The samples are converted to continuous signals using a DAC, then mixed with complex RF carrier. For Bus series USRP (using AD9361 SDR) the base sample rate is 61.44 MHz. To obtain a 240 kps IO rate, it can be configured to sample at 5.28 MHz and perform 22 decimation/interpolation. The sample time would then be set to $(1/5.28\text{e}6/22)$. X310 USRPs and B210 USRP can be configured for different master clock rates and based on the interpolation and decimation factor desired sampling rate can be obtained. Sampling rate of 400 kps can be achieved for master clock rate of 200 MS/s and decimation/interpolation of 500. Sample rate

can be obtained as:

$$\text{Sample Rate} = \frac{\text{Clock Rate}}{\text{Interpolation/Decimation}} \quad (2.1)$$

Fig. 2.10 shows internal view of X310 USRP motherboard with no RF daughterboards. It has two slots for RF daughterboards where we can insert two RF daughterboards based on the RF frequency range requirement such as UBX-160 for wider frequency range (70 MHz to 6 GHz) and LFTX/LFRX for DC-30 MHz coverage. It has also slot for GPSDO for more accurate frequency measurements.

2.6 Digital Upconversion and Downconversion

Software-defined radio transceivers can have a variety of architectures, with one of the most important factors being separation between digital and analog processing. Digital upconversion and downconversion is interpolation and decimation operation that takes place at FPGA side. Appropriate interpolation and decimation factor is selected at the transmitter and receiver side to match the baseband sampling rate and the DAC/ADC sampling rate. Interpolation and decimation operation is accompanied with anti-imaging and anti-aliasing filters respectively [14]. Based on interpolation and decimation factors anti-imaging and anti-aliasing filters are configured. Digital upconversion and digital down conversion is also performed at baseband if the sampling rate is insufficient or the communication blocks require low-sample rate for downstream processing. Digital upconversion and digital down-conversion requires filtering to avoid imaging and aliasing. Therefore, it is not a computational friendly operation.

2.6.1 Digital Up-conversion (DUC)

In transmit chain of the USRP sampling rate is increased using filtering operations. The sampling rate increases from baseband sampling rate to a rate that matches DAC conversion rate. Multirate filters are required to increase and de-

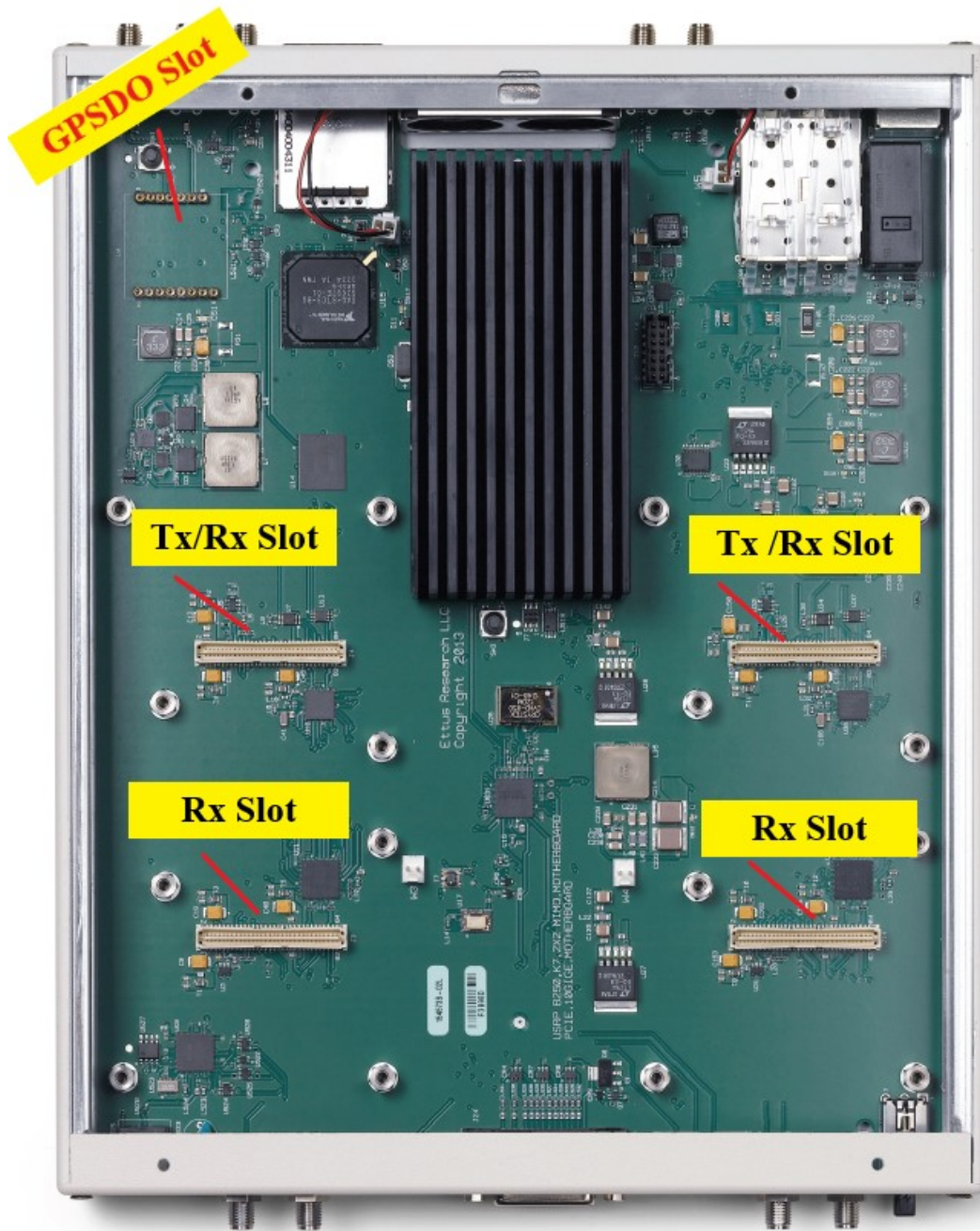


Figure 2.10: X310 USRP internal view without RF daughterboards

crease the sampling rate. The process of upsampling takes place when sampling rate is increased. The effect of upsampling is to generate images at integer multiples of the original sampling rate as shown in Fig 2.11 and it must be subsequently removed using a lowpass filter. The combined operation of upsampling and low-pass filtering is known as *interpolation* [14][15]. The images can be removed using low-pass filter as shown in Fig.2.11d. The spectral regrowth or the images are low-pass filtered with attenuation of approximately 75 dB.

2.6.2 Digital Down-conversion (DDC)

In the receive chain of the USRP sampling rate is reduced from ADC sampling rate to a rate that matches baseband sampling rate. Rate conversion to baseband sampling rate is useful to process baseband samples for reliable KPI extraction. In the second case at the receiver sampling rate is reduced using downsampling operation. This must be preceded by lowpass filter, in order to remove the frequency that would otherwise be aliased into the baseband region. The combined operation of lowpass filtering and downsampling is known as *decimation* [14][15].

Lowpass multi-rate filters are required for DUC and DDC. The design of filter depends on the interpolation or decimation ratio, with higher ratios demanding more exacting filter responses. If the ratio between system rate and symbol rate is large then a lowpass filter with sharp cut-off is required which implies the filter is computationally expensive. Rather than performing filtering with a single filter, a more efficient alternative is to partition the interpolation or decimation task into a cascade of filters. For an overall decimation ratio of 200, it would be efficient to use three different filters in cascade decimating the incoming signal by 20, 2, and 5 respectively. There could be other possible combinations. Proper design of DUC and DDC is required for computational efficient hardware implementation on FPGAs or ASIC signal processing systems. Multi-rate elements are used to simulate digital modulation and demodulation between baseband and IF. FIR and IIR filters can be used within filter cascade, with FIR typically preferred for

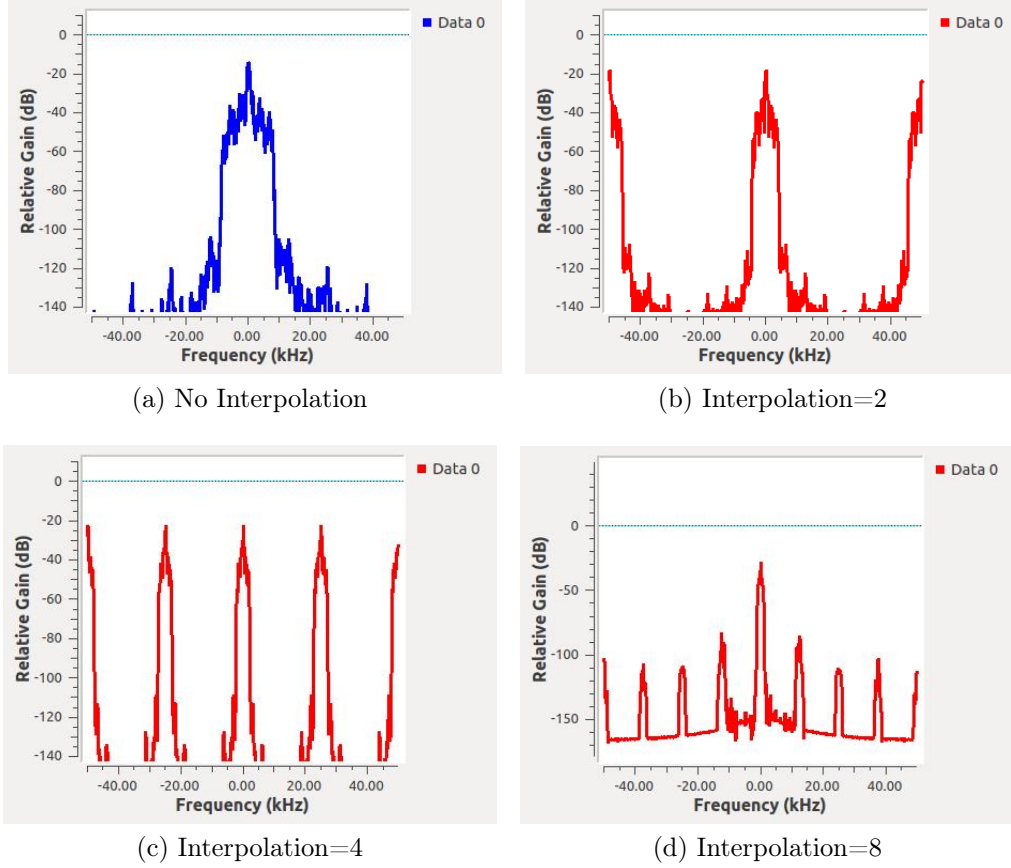


Figure 2.11: Effect of interpolation (a) without interpolation (b) with interpolation factor of 2 (c) with interpolation factor of 4 (d) with interpolation factor of 8. There is spectral regrowth or images after interpolation of the original signal. The number of images are related to interpolation factor.

its linear phase response characteristics. Polyphase methods are normally applied to optimize the structure.

2.7 Baseband Model for Frequency Sweep

Frequency transfer function of MIMO channels can be measured by sweeping the LOs of two USRP synchronously and by transmitting sinusoids using two USRP channels. The received power of each MIMO channel i.e. direct channels and crosstalk channels can be measured at the baseband. Fig. 2.12 shows baseband model to perform frequency sweep of two LOs and measured received signal magnitude. This model can be used to perform MIMO measurement using virtual MIMO technique where the channel is measured in two time instances. In one

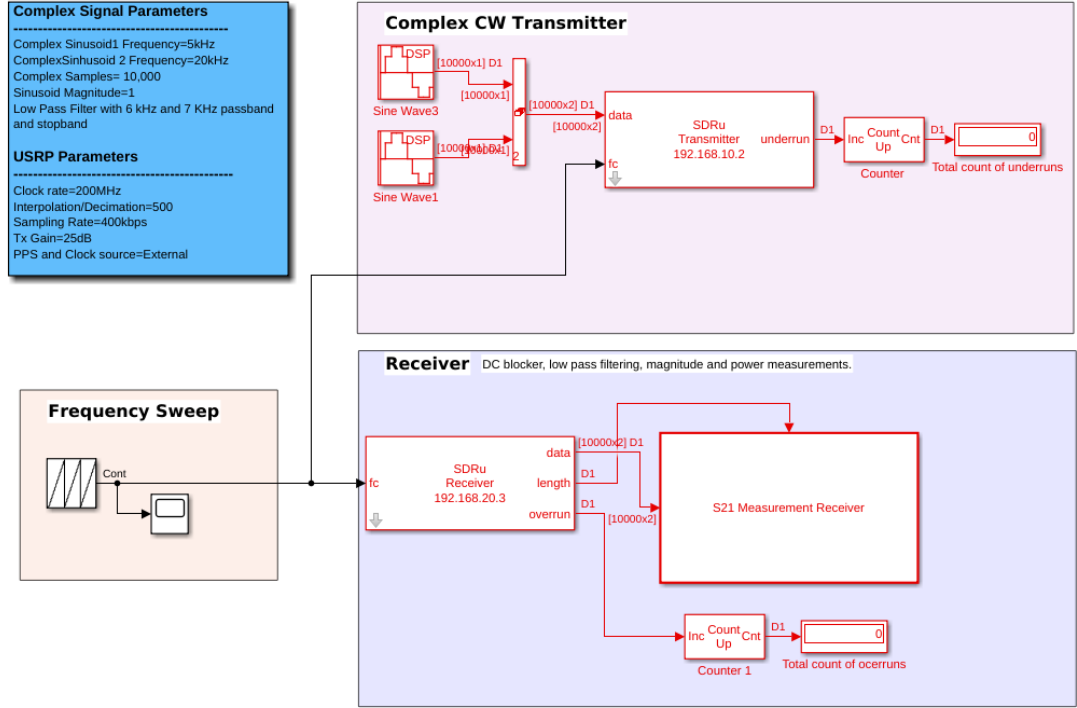


Figure 2.12: Simulink model for frequency transfer function measurement. The Tx and Rx baseband models were created in the same Simulink canvas. SDRu-Transmitter and SDRuReceiver blocks accept the sweep sequence at the same time.

time instant channel between transmit antenna 1 and receive antenna 1 and receive antenna 2 is measured and in another instant channel between transmit antenna 2 and receive antenna 1 and receive antenna 2 is measured. The same model can be configured to extract 2x2 MIMO channel matrix with one time slot. Full MIMO channel measurement can be obtained with the same baseband model without switching Tx channels. This can be achieved by transmitting two baseband complex Sinusoids at different frequencies and sweeping the LOs synchronously. In the baseband model lowpass filter and bandpass filter can be employed to measure the magnitude of two received Sinusoids. This can be done by transmitting two orthogonal Sinusoids and using bandpass filters at the baseband. Since two USRP channels are active at the same time, therefore, this way to extract MIMO channel can include the effect of correlation and mutual coupling of Tx and Rx antennas. Fig. 2.13 shows four frequency sweeps each frequency sweep of 200 MHz bandwidth. Channel measurement using first sweep

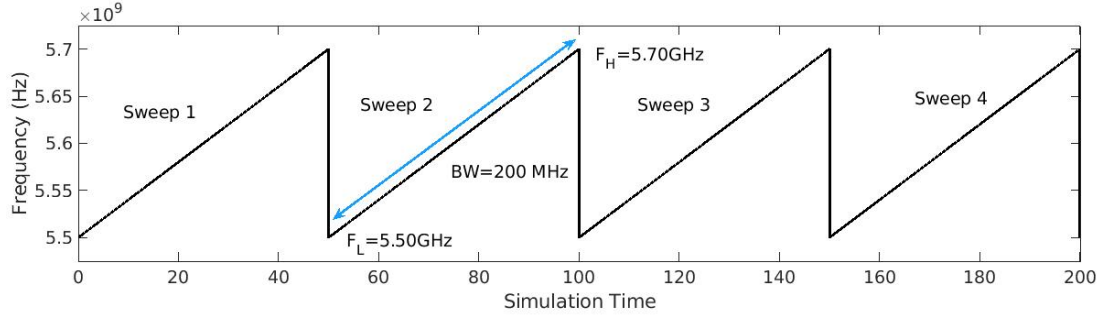


Figure 2.13: Number of frequency sweeps.

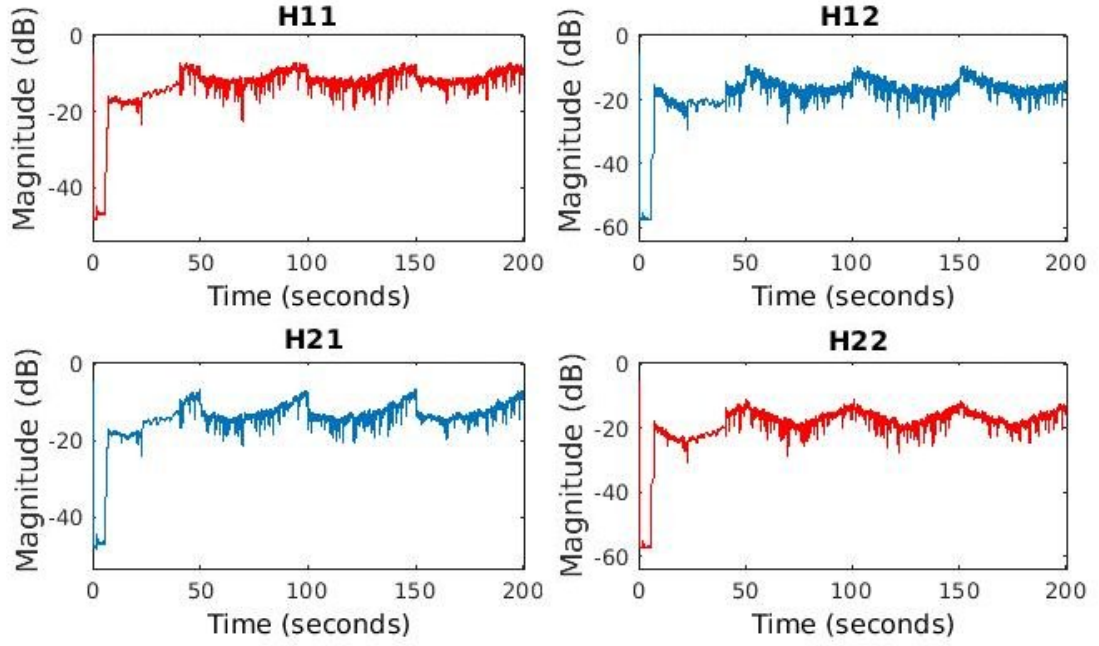


Figure 2.14: MIMO channel measurement sweeps as a function of time. H11 is CTF between Tx1-Rx1, H12 is CTF between Tx1-Rx2, H21 is CTF between Tx2-Rx1, H22 is CTF between Tx2-Rx2.

are not considered in the data analysis as it contains transport delays, Simulink initialization time and time-duration in which two USRPs attain synchronization, therefore insufficient channel measurement samples are obtained. Fig.2.14 shows MIMO channel measurements as a function of time. The channel power of four elements of MIMO channel $\mathbf{H}$ are shown with three sweeps in each plot.

Fig. 2.15 shows frequency transfer function of metal enclosure at eight different stirrer positions. The frequency bandwidth is 100 MHz centered at 5.6 GHz. There are four sweeps of 100 MHz bandwidth. The frequency transfer function is measured at eight different stirrer positions when the mode-stirrer was stationary.

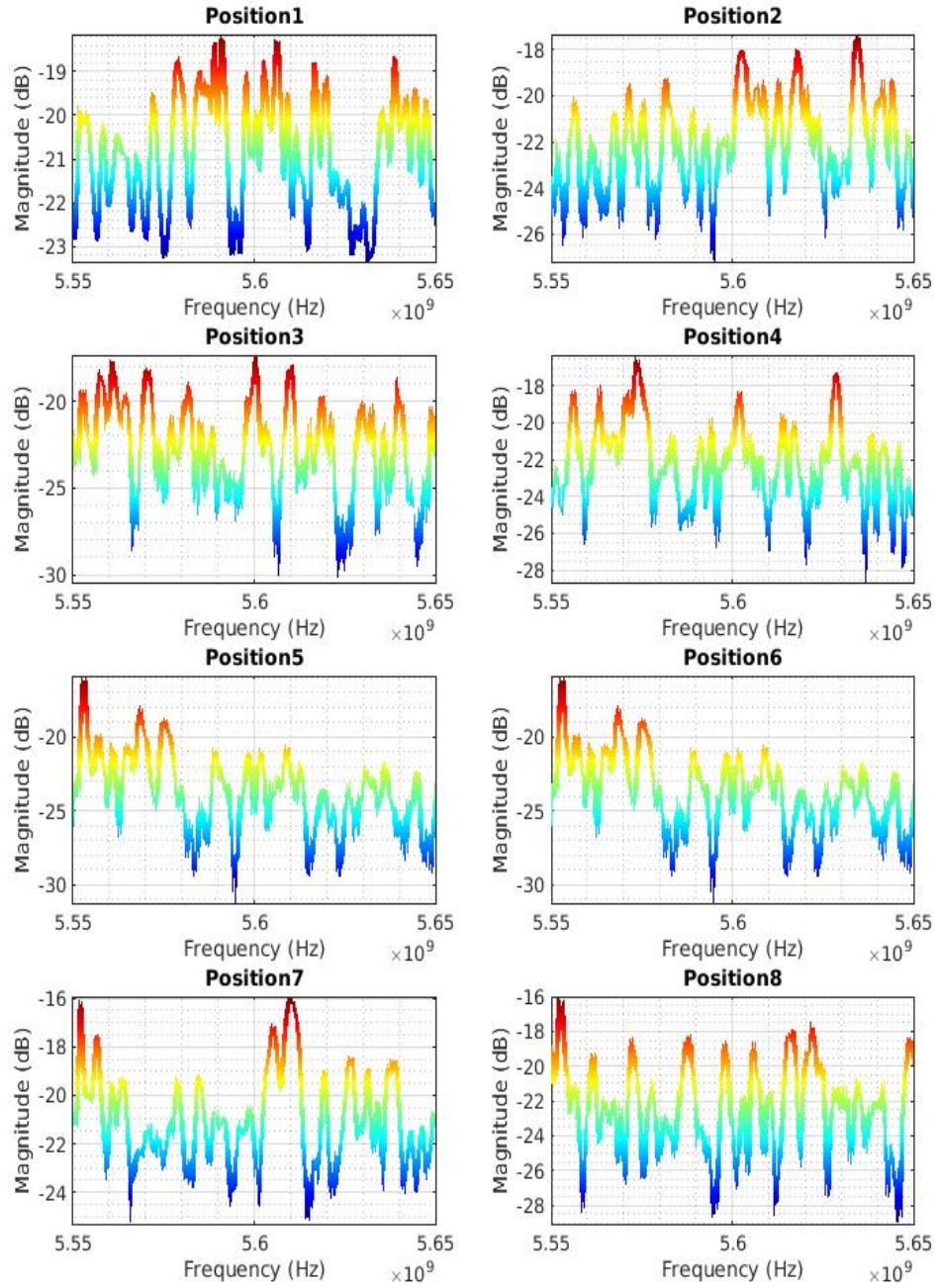


Figure 2.15: Measured frequency response using X310 USRP in metal enclosure at eight mode-stirrer positions. Each subplot of frequency response has three sweeps. The frequency response measurements were conducted in stationary environment at each mode-stirrer positions.

Hence, the prototype measurements results show a changing EM field distribution inside the enclosure. Multiple sweeps of channel transfer function are repeatable in stationary propagation environment for fixed stirrer position. This confirms qualitative accuracy of the MIMO prototype and the baseband model created for MIMO transfer function measurement.

Fig.2.16 shows frequency transfer function in metal enclosure over bandwidth of 100 MHz centered around 5.6 GHz. It can be seen that there are different levels of reflection at stirrer positions. The effect of changing stirrer position was also observed on the BER of two SDR channels when 2x2 MIMO communication receiver was tested. There were instance of low-gain channels where the BER and EVM degraded. Initially, the BER performance was considered as an effect of underflow and overflow and after a careful investigation the problem of low gain values at few stirrer positions was observed. Different magnitude levels at each stirrer position resulted in BER that would vary for each stirrer positions.

2.8 USRP Classification

USRP is classified into subfamilies depending on the operating frequency range, FPGA board, duplexing capability etc. Some of the popular USRPs are X-series, N-series USRPs, B-series, and E-series USRP. Figure .2.7b shows the classification of USRPs. There is a variety of off-the-shelf SDR devices like HackRF, BladeRF, PlutoSDR, Nutaq's PicoSDR, Hedgehog, XTRX, extreme SDR, LimeSDR, Air-T, kiWiSDR, RedPitaya and RTL-SDR etc. However, Ettus has produced a family of SDRs known as universal software radio peripheral (USRP). These USRPs have varying parameters like bandwidth, number of channels, processing capability, ADC/DAC resolution etc. USRP can be targeted for variety of applications some are mentioned in chapter 1. The selection of USRPs is usually based on the number of channels, bandwidth, dynamic range, transmit power, availability of on-board processor, high performance FPGA and operating frequency range. The important parameters of different USRP devices from Ettus Research are

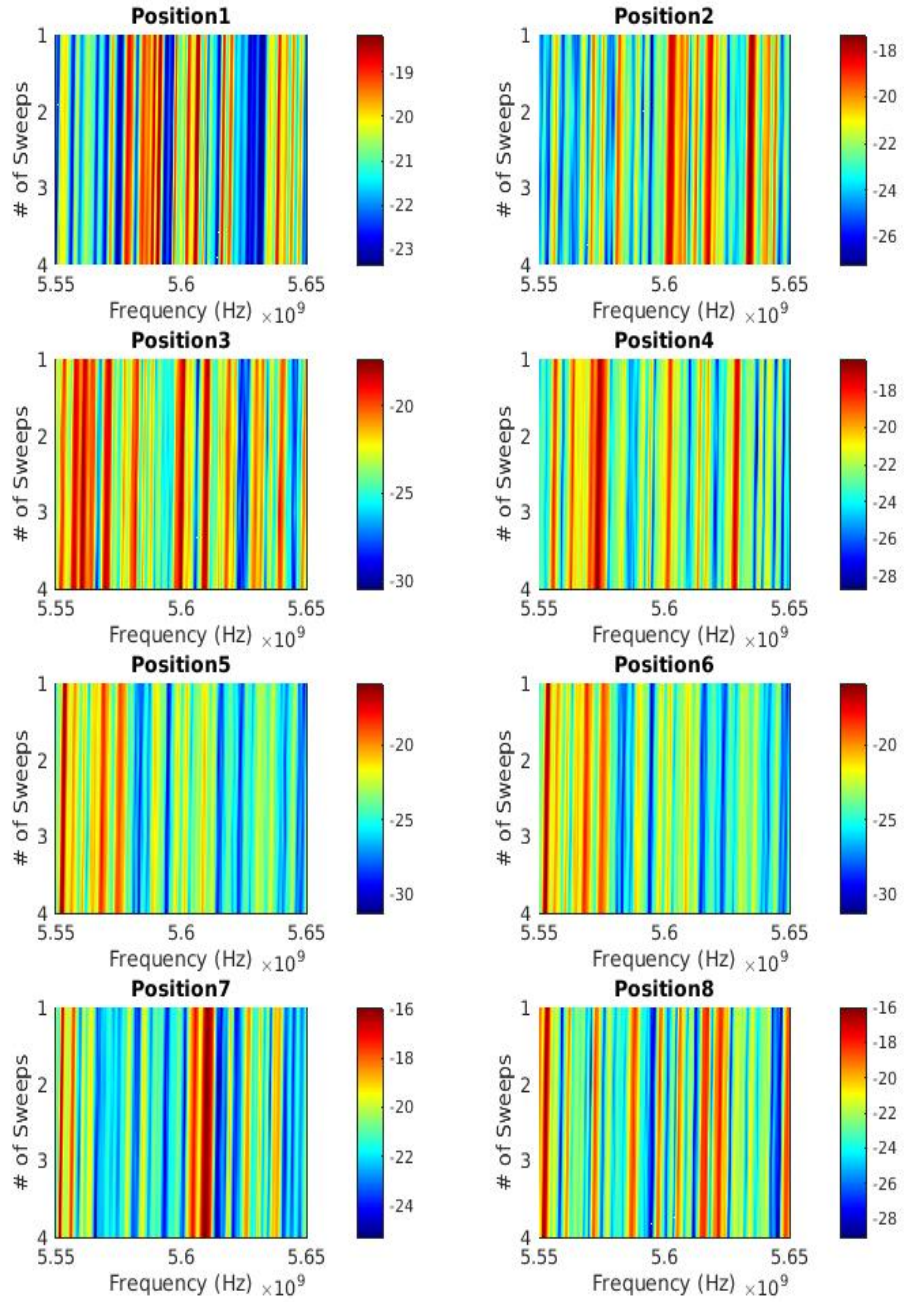


Figure 2.16: Measured frequency response using X310 USRP in metal enclosure at eight mode-stirrer positions. Each subplot of frequency response has three sweeps. The frequency response measurements were conducted in stationary environment at each mode-stirrer positions.

Table 2.1: USRP family of SDR devices.

Parameters	Low Cost B2xx	Embedded E3xx	General Purpose Nxx	High Performance X3xx
Frequency(Hz)	70 MHz- 6 GHz	70 MHz- 6 GHz	10 MHz-6 GHz	10 MHz-6 GHz
Bandwidth	32 MHz in 2x2	32 MHz in 2x2	40 MHz	160 MHz
Channels	2 Tx, 2 Rx	2 Tx, 2 Rx	1 Tx, 1 Rx	2Tx, 2 Rx
DAC/ADC Rate	DAC:12 bit, 61.44 MS/s ADC:12 bit, 61.44 MS/s	DAC:12 bit, 61.44 MS/s ADC:12 bit, 61.44 MS/s	DAC: 16bit, 400 MS/s ADC: 14 bit, 100 MS/s	DAC: 16 bit, 800 MS/s ADC: 14bit, 200 MS/s
Noise Figure	<8 dB	< 8 dB	5-7 dB,	<8 dB
Processing	Host PC	ARM Processor	Host PC	Host PC
MIMO	Phase Coherent	Phase Coherent	Phase Coherent	Phase Coherent
Latency	<i>ms</i> to Host	μs to FPGA and μs to Host	<i>ms</i> to Host	μs to FPGA and μs to Host
Bus	USB3, USB2	1 GbE (ARM to Host)	1 GigE	1/10 GbE
Calibration	None	None	None	None

summarised in Table 2.1.

2.9 RF Daughterboards

RF daughterboards are an important component of any SDR. RF daughterboards are usually limited by the bandwidth and the operating frequency. There are various types of RF frontends which are summarized in Fig.2.7c. Most of RF daughterboards by Ettus Research has an operating frequency range from 70 MHz to 6 GHz which depends on integrated RFIC in the USRP. Few popular RFICs are AD9361, AD9364. AD9371 etc. In our SDR based MIMO testbed two UBX-160 RF daughterboards per X310 USRP are used. The architecture of the wideband full-duplex UBX-160 RF daughterboard is shown in the Fig. 2.17. UBX-160 uses low pass filters in its transmit and receive RF chain and it has direct conversion path from 500 MHz to 6 GHz. Recently, SDR are being devel-

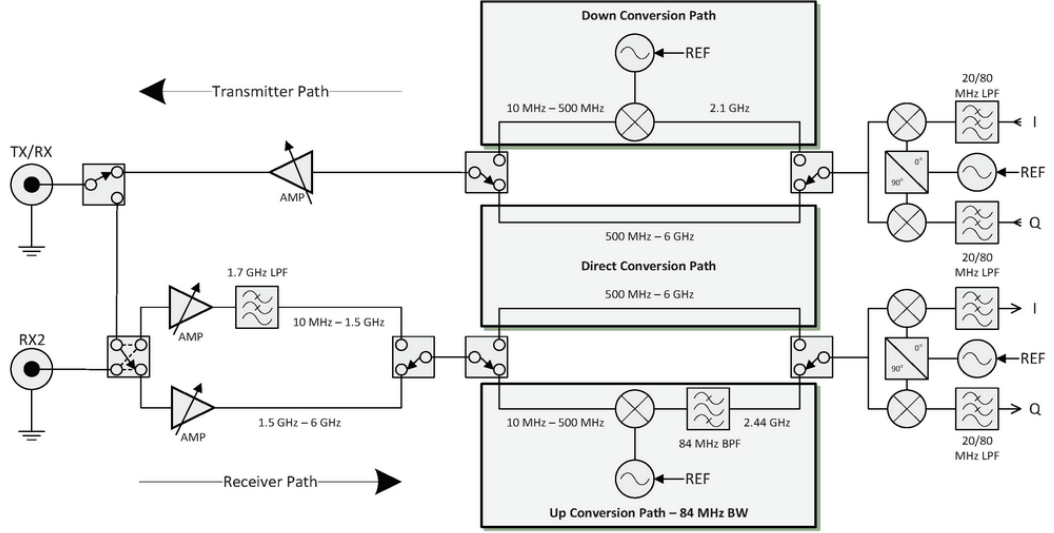


Figure 2.17: UBX-160 RF board architecture.

oped for mmWave measurements and prototyping such as mmWave measurement setup from national instruments and multi-channel hedgehog SDR by BAE systems. However, in order to use X310 USRPs at higher frequencies upconverters can be used. There are few examples where SDR based measurements have been performed at higher frequencies such as 60 GHz and 28 GHz. Authors in [16] have performed an excellent review of mmWave upconverter usage and addressed the prevalent problem of phase noise. Authors in [17] have used SiverIMA up-converter and performed BER measurement at 60 GHz using X310 USRP for mmWave vehicle-to-vehicle communication. Authors in [18] have designed and tested RFNoC testbed for 60 GHz measurements using GNU Radio framework. USRP based platform for 26/28 GHz has been used in work [19] for beamforming applications.

2.10 Baseband Waveform Development Platforms

Baseband development is one of the main parts for SDR based testbed. The choice of baseband waveform development platform plays a major role in data portability and post-processing of the data. There are three well-known baseband waveform development platforms that are used for SDR based signal processing

and communication algorithm implementation. GNU Radio, MATLAB/Simulink and LabView are used for baseband waveform development. This work has extensively used GNU Radio and MATLAB/Simulink for MIMO BER and EVM measurements in metal enclosure. Selection of baseband waveform development platform depends on various factors such as host architecture OS, low code complexity, ease of the deployment and type of USRP. For example for UAV communication Raspberry Pi 4 and GNU Radio combination shall be suitable and for high data-rate reverberation chamber based measurement NUC and/or other high performance host PC with MATLAB/Simulink, GNU Radio or LabView combination shall meet the requirements.

2.10.1 GNU Radio

GNU Radio is a signal processing library used for the development of baseband communication models. Similar to Simulink GNU radio offer GRC canvas. GRC is a graphical interface that sits on top of the normal GNU Radio programming environment that is in Python. GRC translates the flowgraph we create in the GUI canvas here into a Python script, so when we execute a flowgraph, we are really running a Python program. The IQ data generated by GNU Radio baseband model is sent to SDR platform over Ethernet cable, USB 3.0, SFP+ or PCIe for further hardware processing and RF transmission. GNU Radio is an open-source platform for communications and signal processing development that provides numerous native and OOT projects such as IEEE 802.11a/g/p WLAN, IEEE 802.15.4 Zigbee, LPWAN, LoRA, Radar, radio astronomy. Additionally, the custom blocks can be produced using Python programming and there is an excellent GNU Radio community support available through GNU Radio forum. Signal processing blocks are written in C++ and the binding of these blocks is performed using Python. Native embedded python block can be used for development of own signal processing blocks in GRC. Signal flow graph python code can be generated and later can be parameterized and used directly from the terminal.

2.11 Conclusion

In this chapter the details of SDR-based near-field MIMO testbed is presented. Simulink based baseband model has been explained to measure 2x2 MIMO channel matrix using frequency-swept approach. Two approaches to extract MIMO channel are explained one is using virtual MIMO when one Tx antenna is active and two Rx antenna are active and another when both the channels of USRP are active. The later has benefit of capturing mutual coupling and correlation. Measured channel frequency response in metal enclosure at eight stirrer position is presented as an example. The importance of reprogrammable, portable nature of SDRs, and baseband development has been highlighted. Important digital receiver concepts that are frequently used in SDR based development are explained and related to the practical implementation of the wireless communication receiver. The process of digital upsampling and downsampling has been produced with the help of GNU Radio measurement. Near-field MIMO testbed mainly comprises of Ettus X310 USRPs, hence the architecture of X310 USRP is also explained. Different RF daughtercards that can be useful for the RF propagation are also enlisted. Finally, near-field MIMO testbed limitations such as crosstalk, bandwidth, underflow and overflow that can affect measurements are explained.

Bibliography

- [1] Ramanathan Subramanian, Benjamin Drozdenko, Eric Doyle, Rameez Ahmed, Miriam Leiser, and Kaushik Roy Chowdhury. High-level system design of ieee 802.11 b standard-compliant link layer for matlab-based sdr. *IEEE Access*, 4:1494–1509, 2016.
- [2] Hussein Kwasme and Sabit Ekin. Rssi-based localization using lorawan technology. *IEEE Access*, 7:99856–99866, 2019.
- [3] Luca Catarinucci, Danilo De Donno, Riccardo Colella, Fabio Ricciato, and Luciano Tarricone. A cost-effective sdr platform for performance characterization of rfid tags. *IEEE Transactions on Instrumentation and Measurement*, 61(4):903–911, 2011.
- [4] Han Ding, Chen Qian, Jinsong Han, Jian Xiao, Xingjun Zhang, Ge Wang, Wei Xi, and Jizhong Zhao. Close-proximity detection for hand approaching using backscatter communication. *IEEE Transactions on Mobile Computing*, 18(10):2285–2297, 2018.
- [5] Florian Galler, Stefan Grebien, Thomas Faseth, Klaus Witrisal, Gottfried Magerl, and Holger Arthaber. Extension of an sdr uhf rfid testbed for mimo and monostatic time of flight based ranging. *IEEE Journal of Radio Frequency Identification*, 1(1):32–38, 2017.
- [6] Wahab Khawaja, Ozgur Ozdemir, and Ismail Guvenc. Uav air-to-ground channel characterization for mmwave systems. In *2017 IEEE 86th Vehicular Technology Conference (VTC-Fall)*, pages 1–5. IEEE, 2017.
- [7] Guojin Zhang, Xuesong Cai, Wei Fan, and Gert Frølund Pedersen. A usrp-based channel sounder for uav communications. *Eucap 2020*, 2019.
- [8] Xuesong Cai, Nanxin Wang, José Rodríguez-Piñeiro, Xuefeng Yin, Antonio Pérez Yuste, Wei Fan, Guojin Zhang, Gert Frølund Pedersen, and

- Li Tian. Low altitude air-to-ground channel characterization in lte network. In *2019 13th European Conference on Antennas and Propagation (EuCAP)*, pages 1–5. IEEE, 2019.
- [9] Rickard Nilsson and Jaap van de Beek. Channel measurements in an open-pit mine using usrps: 5g—expect the unexpected. In *2016 IEEE Wireless Communications and Networking Conference*, pages 1–6. IEEE, 2016.
- [10] Sukhyun Hwang, Han-Joon Kim, Kyung Tae Kim, and Ji-Woong Choi. Testbed implementation of near-field magnetic mimo communication system using sdr. In *2016 International Symposium on Antennas and Propagation (ISAP)*, pages 610–611. IEEE, 2016.
- [11] MinKeun Chung, Min Soo Sim, Jaeweon Kim, Dong Ku Kim, and Chan-Byoung Chae. Prototyping real-time full duplex radios. *IEEE Communications Magazine*, 53(9):56–63, 2015.
- [12] Ahmed Hussain, Andrés Alayón Glazunov, Bjarni Þór Einarsson, and Per-Simon Kildal. Antenna measurements in reverberation chamber using usrp. *IEEE Transactions on Antennas and Propagation*, 64(3):1152–1157, 2016.
- [13] Jeremy Hershberger, Thomas Pratt, and Robert Kossler. Automated calibration of a dual-polarized sdr radar. In *2016 IEEE Conference on Antenna Measurements & Applications (CAMA)*, pages 1–4. IEEE, 2016.
- [14] Alexander M Wyglinski, Robin Getz, Travis Collins, and Di Pu. *Software-Defined Radio for Engineers*. Artech House, 2018.
- [15] Robert W Stewart, Kenneth W Barlee, Dale SW Atkinson, and Louise H Crockett. *Software defined radio using MATLAB & Simulink and the RTL-SDR*. Number 1. Strathclyde Academic Media, 2015.
- [16] Adnan Quadri, Huacheng Zeng, and Y Thomas Hou. A real-time mmwave communication testbed with phase noise cancellation. In *IEEE INFOCOM*

2019-IEEE Conference on Computer Communications Workshops (INFOCOM WKSHPS), pages 455–460. IEEE, 2019.

- [17] Adrian Loch, Arash Asadi, Gek Hong Sim, Joerg Widmer, and Matthias Hollick. mm-wave on wheels: Practical 60 ghz vehicular communication without beam training. In *2017 9th International Conference on Communication Systems and Networks (COMSNETS)*, pages 1–8. IEEE, 2017.
- [18] Adriana Moreno, Jesus Omar Lacruz, and Joerg Widmer. Open source rfnoc-based testbed for millimeter-wave experimentation using usrp software defined radios. In *2020 IEEE International Symposium on Circuits and Systems (ISCAS)*, pages 1–5. IEEE, 2020.
- [19] Martin Danneberg, Roberto Bomfin, Ahmad Nimr, Zhongju Li, and Gerhard Fettweis. Usrc-based platform for 26/28 ghz mmwave experimentation. In *2020 IEEE Wireless Communications and Networking Conference Workshops (WCNCW)*, pages 1–6. IEEE, 2020.

Chapter 3

Channel Measurement, Modeling and Diversity

3.1 Introduction

Wireless chip-to-chip channel has potential to exhibit all forms of propagation configurations and fading scenarios. Wireless C2C channel can appear in different forms such as LOS, NLOS or partially blocked, the fading can be isotropic or non-isotropic. There are many factors that can shape wireless C2C channel to name a few are ;if not all, operating frequency, antenna placement, type of antennas, number of antennas, the number of scatterers around Tx and Rx antennas, type of polarization, near-field scattering. In order to develop empirical model for wireless C2C channel it's important to apply fitting to measured data using different statistical models. Akaike information criteria (AIC), generalized likelihood ratio test (GLRT) and Kolmogorov-Smirnov (KS) tests are used to determine type of distribution that closely fits the measurement data. Following sections explain classical fading distributions with first and second order parameters.

3.2 Classical Fading Distributions

Classical fading distribution Rayleigh and Rician distributions are use to describe field distribution of wireless channel. Following distributions are useful which are used for various fading scenarios [1][2][3]. Classical fading distribution are charac-

terized by PDF, CDF, moment generating function, characteristic function, mean and variances respectively [4].

- CDF: CDF of random variable X is defined as:

$$f_X(x) = P[X \leq x] = \int_{-\infty}^x p_U(u) du \quad (3.1)$$

- MGF: MGF of random variable X is defined as:

$$M_X(s) = \int_{-\infty}^{\infty} \exp(sx) p_X(x) dx \quad (3.2)$$

- CF: Characteristic function of random variable X is defined as $C_X(\omega) = E[\exp(j\omega X)]$ which can be expressed as:

$$C_X(\omega) = \int_{-\infty}^{\infty} \exp(j\omega x) f_X(x) dx \quad (3.3)$$

CF can be obtained from MGF by putting $s = j\omega$.

3.3 Gaussian or Normal Distribution

This distribution is very popular and widely used distribution in wireless communication. This is evident from central limit theorem which says sum of large number of continuous random variables is Gaussian distributed. A random variable X is said to be Gaussian distributed if the r.v. has following PDF:

$$f_X(x) = \frac{1}{\sqrt{2\pi}\sigma} \exp\left(-\frac{(x-\mu)^2}{2\sigma^2}\right) \quad (3.4)$$

and its CDF is:

$$f_X(x) = 1 - Q\left(\frac{x-\mu}{\sigma}\right) \quad (3.5)$$

In short Gaussian distributed random variable X with mean μ and variance σ^2 is written as $X \sim \mathcal{N}(\mu, \sigma^2)$.

3.4 Rayleigh Distribution

Rayleigh distribution is used to fit the measured data when the received signal is a sum of all the reflected multipath components and no line-of-sight component is present [5]. Indoor and outdoor NLOS fading channels are characterized using Rayleigh fading. The PDF of Rayleigh distribution is

$$p(\alpha) = \frac{\alpha}{\sigma^2} e^{-\frac{\alpha^2}{2\sigma^2}}; \alpha \geq 0 \quad (3.6)$$

and the CDF of Rayleigh distribution is given by:

$$P(\alpha) = \begin{cases} 1 - e^{-\frac{\alpha^2}{2\sigma^2}}; & \alpha > 0 \\ 0 & \end{cases} \quad (3.7)$$

Envelope of Rayleigh distribution can be obtained using two i.i.d Gaussian random variables with mean $\mu = 0$ and variance $\sigma^2 = 1$.

$$\alpha = \sqrt{X_1^2 + X_2^2}; \quad X_1 \sim \mathcal{N}(0, 1); X_2 \sim \mathcal{N}(0, 1) \quad (3.8)$$

Mean and variance of Rayleigh fading envelope is:

$$E(\alpha) = \sigma \sqrt{\frac{\pi}{2}}; Var(\alpha) = \left(2 - \frac{\pi}{2}\right) \sigma^2 \quad (3.9)$$

3.5 Rician Distribution

Rician distribution is used to fit the measured data when line-of-sight or stronger reflections are present in the received signal in addition to multipath fading components. The PDF of Rician distribution is:

$$p(\alpha) = \frac{2(1+K)e^{-K}\alpha}{\Omega} e^{-\frac{(1+K)\alpha^2}{\Omega}} I_0 \left(2\sqrt{K}\alpha \sqrt{\frac{1+K}{\Omega}} \right); \alpha \geq 0 \quad (3.10)$$

Envelope of Rician distribution is obtained using two IID Gaussian random variables i.e.

$$h = \sqrt{X_1^2 + X_2^2}; \quad X_1 \sim \mathcal{N}(m_1, \sigma^2), X_2 \sim \mathcal{N}(m_2, \sigma^2) \quad (3.11)$$

Where $K = \frac{m_1^2 + m_2^2}{2\sigma^2}$ and $\Omega = m_1^2 + m_2^2 + 2\sigma^2$.

3.6 Nakagami-m Distribution

The Nakagami fading is a general model obtained from experimental data fitting. Main benefit of Nakagami-m is that it can model range of fading statistics only by changing its parameter m [6]. The PDF of Nakagami- m fading distribution is:

$$f_\alpha(\alpha) = \frac{2m^m \alpha^{2m-1}}{\Omega^m \Gamma(m)} \exp\left(-\frac{m\alpha^2}{\Omega}\right); \alpha \geq 0; m \geq 1/2 \quad (3.12)$$

Where $\Omega = E(\alpha^2)$. Channel gains can be modeled by Nakagami-m. For $m = 1/2$ the distribution is Gaussian, for $m = 1$ the distribution is Rayleigh fading and large values of m , the distribution is like log-normal fading. Rayleigh and Rician distribution can be obtained by changing the parameter m . In a nutshell, Nakagami-m distribution can model Rayleigh, Rician, and Suzuki distribution with an appropriate selection of parameters[7][8].

3.7 Log-Normal Distribution

Log-normal distribution is typically used to characterize shadow fading. A random variable is said to be log-normally distributed if the logarithm of it follows normal distribution. PDF of log-normal distribution is given as:

$$f_X(x) = \frac{1}{x} \frac{1}{\sigma \sqrt{2\pi}} \exp\left(-\frac{(\ln x - \mu)^2}{2\sigma^2}\right) \quad (3.13)$$

Similarly, its CDF:

$$F_X(x) = \Phi\left(\frac{\ln x - \mu}{\sigma}\right) \quad (3.14)$$

Equivalently, CDF can be given in terms of complementary error function

$$F_X(x) = \frac{1}{2} \operatorname{erfc}\left(-\frac{\ln x - \mu}{\sigma\sqrt{2}}\right) \quad (3.15)$$

3.8 Weibull Distribution

Weibull distribution is a special type of distribution. Its application to wireless communication was lately found [9]. The PDF of Weibull distribution is related to Rayleigh fading by $y = x^{2/\beta}$. The PDF of Weibull fading distribution can be written as

$$p_r(r) = \frac{\beta}{\Omega} r^{\beta-1} \exp\left(-\frac{r^\beta}{\Omega}\right) \quad (3.16)$$

The Weibull CDF of received signal r

$$F_r(r) = 1 - \exp\left(-\frac{r^\beta}{\Omega}\right) \quad (3.17)$$

For $\beta = 2$ is reduced to Rayleigh fading. The derivative of r

$$\dot{r} = \frac{2}{\beta} r^{1-(\beta/2)} \dot{x} \quad (3.18)$$

General definition of average LCR and average fade duration (AFD).

$$N(r) = \int_0^\infty \dot{r} p_{\dot{r},r}(\dot{r}, r) d\dot{r} \quad (3.19)$$

$$\tau(r) = \frac{F_r(r)}{N(r)} \quad (3.20)$$

Where $F_r(r)$ is CDF of r and $p_{\dot{r},r}(\dot{r}, r)$ is joint PDF of sampled received signal r and its derivative $\dot{r}$. Closed-form expression of average LCR and average AFD of

the Weibull is obtained as

$$N(\rho) = \sqrt{2\pi} f_d \left(\frac{\rho}{\sqrt{a}} \right) \exp \left[- \left(\frac{\rho}{\sqrt{a}} \right)^\beta \right] \quad (3.21)$$

$$\tau(\rho) = \frac{1 - \exp \left[- \left(\frac{\rho}{\sqrt{a}} \right)^\beta \right]}{\sqrt{2\pi} f_d \left(\frac{\rho}{\sqrt{a}} \right)^{\beta/2} \exp \left[- \left(\frac{\rho}{\sqrt{a}} \right)^\beta \right]} \quad (3.22)$$

Where $\rho = r/r_{rms}$ and $r_{rms} = \sqrt{E < r^2 >} = \Omega^1/\beta/\sqrt{a}$ and $a = 1/\Gamma(1 + 2\beta)$. Channel capacity and second order statistics of Weibull fading distribution is investigated by authors in [10].

3.9 Multipath Fading

Received signal is sum of LOS/NLOS multipath components that results from scattering by objects present around transmitter and/or receiver. Wireless channel is modeled with impulse response which records attenuated, delayed and phase-shifted copies of the original signal. Channel impulse response is a random quantity can be written as [6]:

$$h(t) = a_0\delta(t - \tau_0) + a_1\delta(t - \tau_1) + \dots + a_{N-1}\delta(t - \tau_{N-1}) \quad (3.23)$$

Equivalently:

$$h(t) = \sum_{i=0}^{N-1} a_i\delta(t - \tau_i) \quad (3.24)$$

Received signal $y(t)$ is convolution of transmitted signal $x(t)$ and channel impulse response $h(t)$. Mathematically, it can be written as:

$$y(t) = h(t) * x(t) + n(t) \quad (3.25)$$

$$= \sum_{i=0}^{N-1} a_i x(t - \tau_i) e^{j2\pi f_c(t - \tau_i)} \quad (3.26)$$

If two received signals have same attenuation but different time delay then two signals are out of phase. In multipath fading channel impulse response is phasor and is sum of real and imaginary components [11].

$$h(\tau) = \sum_{i=0}^{N-1} a_i e^{-j2\pi f_c \tau_i} = x + jy = ae^{j\phi} \quad (3.27)$$

$$= \sum_{i=0}^{N-1} a_i \cos(2\pi f_c \tau_i) - ja_i \sin(2\pi f_c \tau_i) \quad (3.28)$$

Hence,

$$\Re\{h\} = \sum_{i=0}^{N-1} a_i \cos(2\pi f_c \tau_i) \quad \text{and} \quad \Im\{h\} = - \sum_{i=0}^{N-1} b_i \cos(2\pi f_c \tau_i) \quad (3.29)$$

3.9.1 Small-scale fading

Small-scale fading is instantaneous random variation of multipath fading and changes in a distance of $\sim \lambda/2$ [12] and it is often modeled as Rayleigh or Rician fading. Small-scale fading parameter is one of the useful statistics for differentiating orthogonality in MIMO channels. Fig.3.1 shows received channel power gain of NLOS channel in metal enclosure in stationary channel condition. Envelope of received NLOS signal is modeled with Rayleigh fading.

3.9.2 Large-scale fading

Large-scale fading or shadow fading is a large-scale variation of fading and often modeled with a log-normal distribution. It is a result of large-scale path loss between transmitter and receiver. Fig.3.2 shows received channel power gain when the stirrer moves which corresponds to changing path-loss and propagation environment.

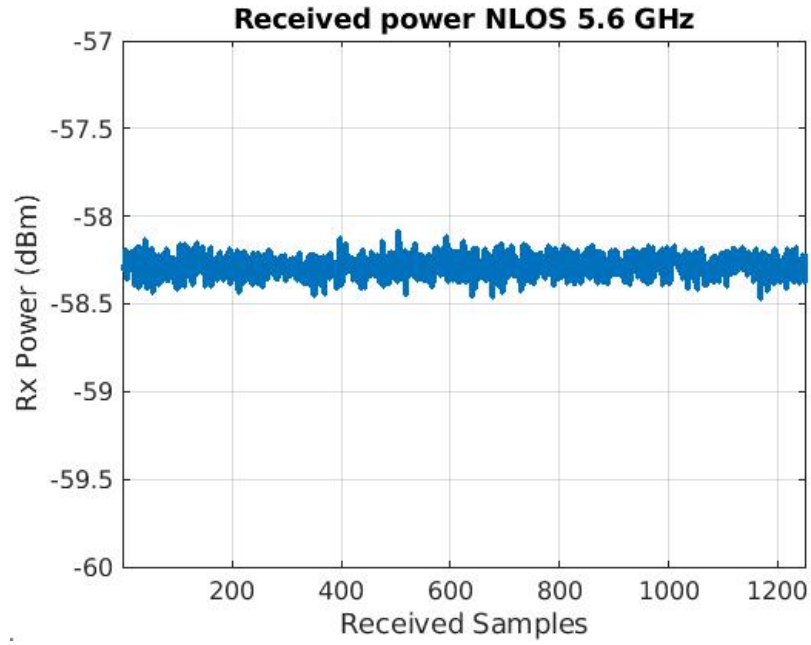


Figure 3.1: NLOS received power at 5.6 GHz in metal enclosure. The Tx and Rx gain of the USRP were 10 dB and 0 dB respectively.

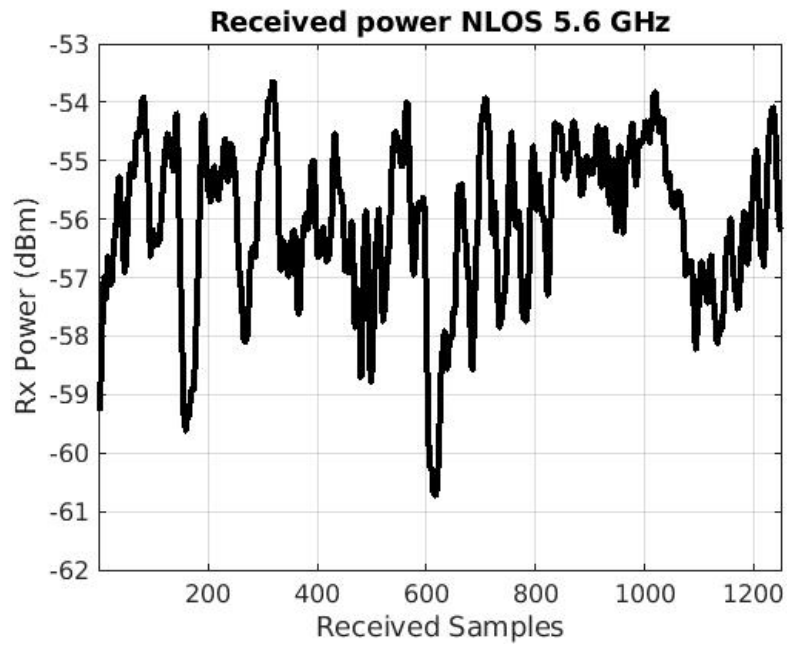


Figure 3.2: NLOS received power at 5.6 GHz in metal enclosure when the stirrer was turned on. The Tx and Rx gain of the USRP were 10 dB and 0 dB respectively.

3.9.3 Path Loss

Free space path loss is function of distance d and wavelength λ . Free-space power received by receiver at distance d [13].

$$P_r(d) = \frac{P_t G_t G_r \lambda^2}{(4\pi)^2 d^2 L} \quad (3.30)$$

Where G_t , G_r , and L are transmit antenna gain, receive antenna gain and system loss respectively. Path loss quantifies attenuation suffered by signal. Logarithmically, it's difference of power received to power transmitted.

$$PL(\text{dB}) = 10 \log \frac{P_t}{P_r} = -10 \log \left[\frac{G_t G_r \lambda^2}{(4\pi)^2 d^2} \right] \quad (3.31)$$

For transmit and receive antenna gains $G_t = G_r = 1$, path loss can be written as:

$$PL_0 = \left(\frac{4\pi d}{\lambda} \right)^2 \quad (3.32)$$

From equation (3.32) it's clear that path loss is directly proportional to the square of distance (d) and inversely proportional to square of the signal wavelength (λ). Therefore, path loss increases as the distance between transmitter and receiver increases. It also increases with increase in signal frequency which implies that path loss is higher for small wavelength signal. Measured path loss from frequency transfer function is obtained by averaging over time and frequency:

$$\overline{PL(d)} = \frac{1}{MN} \sum_{i=1}^N \sum_{j=1}^M |H(f_i, t_j, d)|^2 \quad (3.33)$$

M is number of frequency response snapshots over time and N is number of observed frequencies.

$$\overline{PL} = PL(d_0) + 10n \log_{10} \left(\frac{d}{d_0} \right) + \chi_\sigma \quad (3.34)$$

n is pathloss exponent and χ_σ is shadow fading modeled with lognormal distribution with mean 0 and variance σ .

3.9.4 PDP/RMS Delay Spread

PDP is defined as power of number of received multipaths as a function of time delay. RMS delay spread can be given by following expression [11]:

$$\tau_{RMS} = \sqrt{\frac{\sum_k P(\tau_k) \tau_k^2}{\sum_k P(\tau_k)} - \frac{\sum_k P(\tau_k) \tau_k}{\sum_k P(\tau_k)}} \quad (3.35)$$

$$\tau_{RMS} = \sqrt{\overline{\tau^2} - \bar{\tau}^2} \quad (3.36)$$

Where $P(\tau_k)$ is power of k th MPC arriving at τ_k , $\bar{\tau}$ is mean excess delay and $\overline{\tau^2}$ is second central moment of power delay profile.

$$\bar{\tau} = \frac{\sum_k P(\tau_k) \tau_k}{\sum_k P(\tau_k)} \quad (3.37)$$

RMS delay spread is inversely related to coherence bandwidth. Therefore, RMS delay spread has direct implications on the total bit rate for communication system independent of frequency-selectivity of the wireless channel.

3.9.5 Coherence bandwidth

It's statistical measure of range of frequencies over which the channel can be considered flat[14]. Coherence bandwidth is measured as auto-correlation of frequency transfer function $H(f)$ [15][16]. Mathematically

$$R(\delta f) = \int_{f_{min}}^{f_{max} - \delta f} H(f) H^*(f + \delta f) df \quad (3.38)$$

Where f_{max} and f_{min} are maximum and minimum frequencies respectively. It's bandwidth over which frequency spectrum is considered flat or frequency transfer function is correlated. Coherence bandwidth is determined by measuring

the width of frequency auto-correlation function at particular threshold usually above 0.5. It's helpful for the selection of design parameters of wireless communications systems in any radio propagation environment. This parameter can affect complexity of wireless chip-to-chip communication system as the inclusion and selection of equalizer as a design component for combating the inter-symbol interference (ISI) depends on this parameter. It's inversely related to RMS delay spread and coherence bandwidth in terms of RMS delay spread at threshold of 0.5 is given by [11]:

$$B_c \approx \frac{1}{5\tau_{rms}} \quad (3.39)$$

And at threshold of 0.9

$$B_c \approx \frac{1}{50\tau_{rms}} \quad (3.40)$$

There is extensive literature available for the measurement of coherence bandwidth in different radio propagation environments such as [17] presents RMS delay spread and coherence bandwidth measurements in indoor propagation environment. Coherence bandwidth has also been considered equivalent to average mode-bandwidth in mode-stirred reverberant environment [18]. It's also known that average-mode bandwidth can be measured faster in repeatable measurement setup such as metal enclosures or reverberation chamber.

3.9.6 Coherence Time

Coherence time is inversely proportional to Doppler frequency. It's measured as time duration over which channel impulse response is considered to be time invariant [19]. Doppler spectrum and coherence time are inversely related quantities [5]. Constant of proportionality that relates two quantities can be calculated by directly measuring coherence time and Doppler spread. Doppler spectrum is measured using following relationship [20]:

$$S_D(f_D) = \int_{\tau} S_D(f_D; \tau) d\tau \quad (3.41)$$

Table 3.1: Frequency, time and space correspondence of channel transfer function.

	Frequency	Time	Space
Dependency	f	t	r
Coherence	Coherence bandwidth (B_c)	Coherence time (T_c)	Coherence distance (D_c)
Spectral Domain	Delay (τ)	Doppler Frequency (f)	Wave vector (k)
Spectral Width	Delay spread (T_{spread})	Doppler spread (D_{spread})	Wavevector spread (k)

Where $S_D(f_D; \tau)$ is also known as scattering function in Radar systems.

3.9.7 Coherence Distance

Coherence distance D_c is defined as distance over which channel remains unchanged. This is also known as spatial coherence. For a wireless channel transfer function $H(f, t, r)$ which is function of frequency f , time t , and space r , the following table 3.1 summarizes dependency relationships.

$$\underbrace{H(f, t, \underline{r})}_{\text{Transfer - function}} \implies \underbrace{\Phi_H(\Delta f, \Delta t, \Delta \underline{r})}_{\text{Autocorrelation-function}} \xrightarrow{FT} \underbrace{S_H(\tau, f, \underline{k})}_{\text{Scattering - function}} \quad (3.42)$$

3.9.8 Doppler Spread

Doppler is introduced in the channel when transmitter and/or receiver are in motion. It's can also be introduced with motion of scatterers that comprise a wireless channel. Doppler shift of received signal is [6]:

$$f_d = \frac{vf_c}{c} \cos \theta \quad (3.43)$$

Where v is velocity of moving object, f_c is center frequency and θ is the angle between moving object and the mobile station. RMS bandwidth $\psi(D_o)(f)$ is called Doppler spread and is given by:

$$f_{RMS} = \sqrt{\frac{\int (f - \bar{f})^2 \psi_{D_o}(f) df}{\int \psi_{D_o}(f) df}} \quad (3.44)$$

Where $\bar{f}$ is average frequency of Doppler spectrum.

$$\bar{f} = \frac{\int f \psi_{Do}(f) df}{\int \psi_{Do}(f) df} \quad (3.45)$$

Hence, the coherence time T_c of channel is

$$T_c = \frac{1}{f_{RMS}} \quad (3.46)$$

It is the best measure of how channel changes in time which implies higher the coherence time, the slower the channel fluctuations.

3.9.9 Frequency flat fading

Frequency flat fading is defined as bandwidth over which spectrum remains constant or is correlated. This type of fading has fixed channel attenuation or path loss. Frequency flat fading is observed when signal bandwidth is less than the coherence bandwidth of the channel. Frequency-flat fading is suitable for applications that require less complex receiver design.

3.9.10 Frequency selective fading

It's bandwidth over which spectrum experiences more attenuation at some frequencies and less attenuation at other frequencies. This condition of selectivity is observed when signal bandwidth is larger than the coherence bandwidth. Frequency selective fading creates inter-symbol-interference. OFDM is used as a solution to mitigate frequency-selectivity in the channel.

3.10 Second Order Statistics

Second order statistics is important to understand wireless channel. Second order statistics of wireless channel are level crossing rate and average fade duration.

3.10.1 Level Crossing Rate (LCR)

It's a rate at which the envelope passes a threshold ρ_{th} with positive or negative slope [21]. LCR for Rayleigh fading with isotropic scattering

$$N(r)_{Ray} = \sqrt{2\pi} f_d \rho_{th} e^{-\rho_{th}^2} \quad (3.47)$$

Where f_d is Doppler shift. LCR for Rician fading with isotropic scattering

$$N(r)_{Ric} = \sqrt{2\pi(K+1)} f_d \rho_{th} e^{-K-(K+1)\rho_{th}^2} I_0 \left(2\rho_{th} \sqrt{K(K+1)} \right) \quad (3.48)$$

Where I_0 is modified Bessel function of first kind $I_0(x) = \frac{1}{2\pi} \int_0^{2\pi} e^{-x \cos \theta} d\theta$

3.10.2 Average Fade Duration (AFD)

Average fade duration is ratio of CDF of fading $F(r)$ to level crossing rate $N(r)$.

$$T(r) = \frac{F(r)}{N(r)} \quad (3.49)$$

AFD is best to determine an estimate of the average length of burst errors. The average period of time normalized envelope is below a threshold ρ_{th} . AFD for Rayleigh fading with isotropic scattering

$$\tau(r)_{Ray} = \frac{e^{\rho_{th}^2 - 1}}{\sqrt{2\pi} f_d \rho_{th}} \quad (3.50)$$

AFD for Rician fading with isotropic scattering

$$\tau(r)_{Ric} = \frac{1 - Q \left(\sqrt{2K}, \sqrt{2(K+1)\rho_{th}^2} \right)}{\sqrt{2\pi(K+1)} f_d \rho_{th} e^{-K-(K+1)\rho_{th}^2} I_0 \left(2\rho_{th} \sqrt{K(K+1)} \right)} \quad (3.51)$$

Where $Q(a, b)$ is Marcum Q-function. LCR and AFD for diversity techniques such as SC, EGC and MRC is given in [21][22].

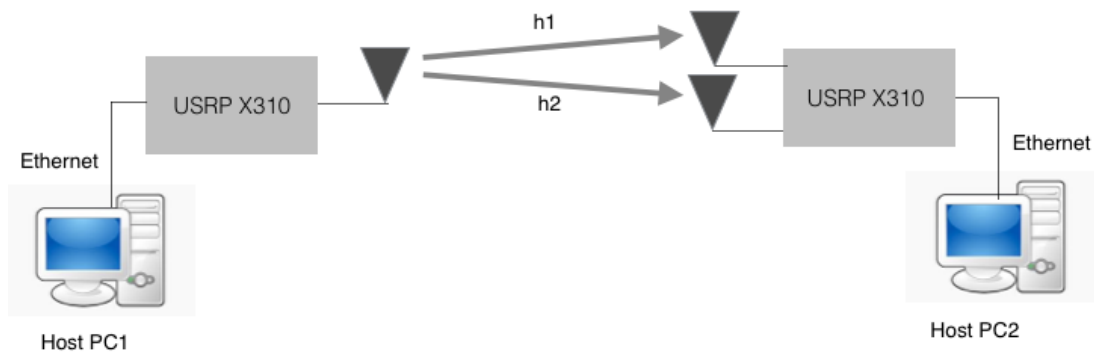


Figure 3.3: Sketch of USRP based Single-Input Multiple-Output.

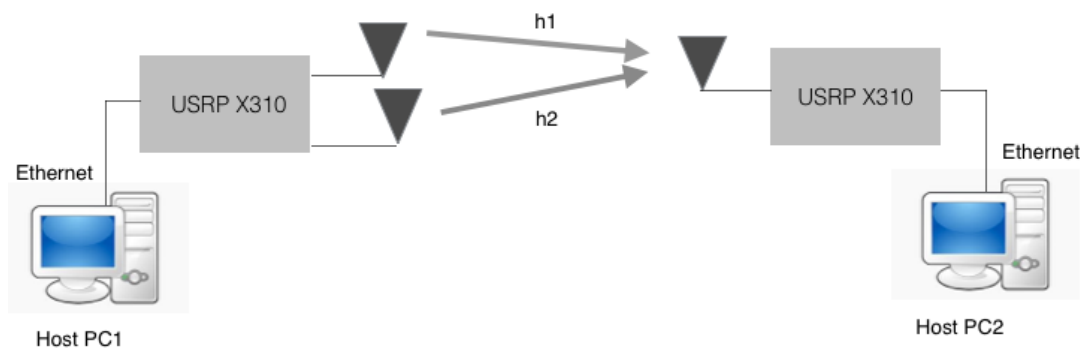


Figure 3.4: Sketch of USRP based Multiple-Input Single-output.

3.11 Diversity

Spatial diversity helps counteract multipath fading effects in wireless communication. It is used to increase the performance of wireless communication system. Diversity increases signal quality because of random nature of multipath fading. In multi-branch receive diversity when one antenna is in deep fade other antenna is not. Therefore, diversity combiner can increase instantaneous and average SNR. In this way diversity is useful to avoid deep fades and increase signal quality. By adopting transmit or receive diversity BER can be reduced. Diversity can be achieved using multiple-antennas at the transmitter and/or at the receiver. Fig. 3.3 shows multiple antennas at the receiver hence known as receive diversity and Fig. 3.4 shows multiple antennas at the transmitter hence known as transmit diversity. Transmit diversity is also known as Alamouti tech-

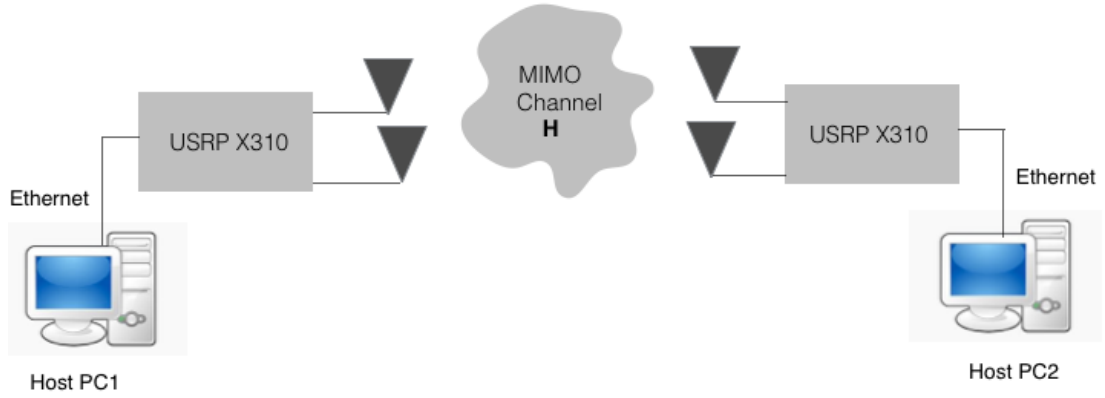


Figure 3.5: Sketch of USRP based Multiple-Input Multiple-Output.

niques where two symbols transmitted by two antennas in time interval t_1 are complex conjugated and switched for transmission in time interval t_2 . Fig. 3.5 which shows multiple antennas at the transmitter and at the receiver. This setup shows benefit of introducing both diversity and spatial multiplexing. Bit Error Rate App is used to plot BER of QPSK modulation in different channels and it is easy to demonstrate that diversity improves BER. Fig. 3.6 shows advantage of using diversity. Fig. shows BER of QPSK in AWGN and fading channels and Fig. 3.6b shows QPSK BER in presence of fading. It can be noticed that the diversity fills the gap between Rayleigh fading case and AWGN case on BER-EbNo plot. QPSK BER approaches AWGN case as the diversity increases. This is a powerful example of diversity enhancing quality of wireless signal and negating the effect of fading. The curves were obtained using MATLAB BER simulation app.

3.12 Selection Combining (SC)

In multi-branch antenna system the selection combining diversity technique selects the branch with highest instantaneous SNR γ_i . The probability that all L -branch received signals fall below threshold ρ_{th} :

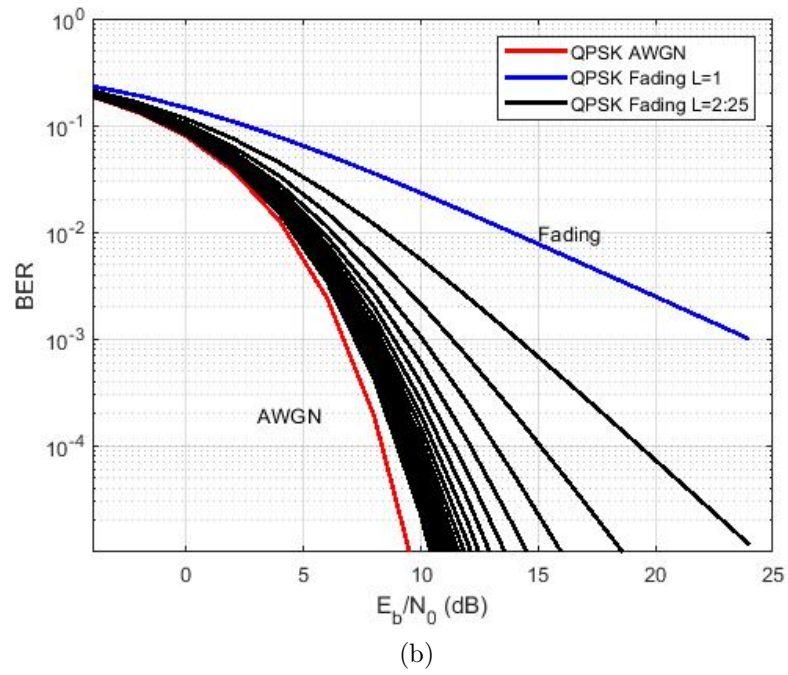
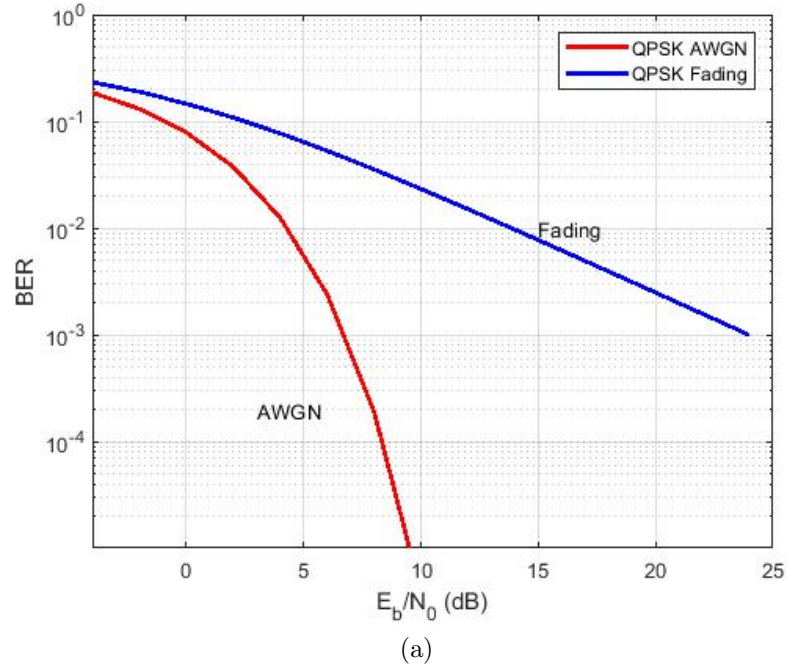


Figure 3.6: Theoretical BER of QPSK (a) AWGN and Rayleigh fading (b) AWGN, Rayleigh fading and Diversity $L=2$ to 25.

$$P_L(\rho_{th}) = Pr[\gamma_{sc} \leq \rho_{th}] = Pr. \left[\max_i \{\gamma_i\} \leq \rho_{th} \right] \quad (3.52)$$

$$= (1 - e^{-\rho_{th}/\rho})^L \quad (3.53)$$

Detailed derivation of average SNR of selection combining is present in the literature. Average SNR of selection combining is:

$$\gamma_{sc} = \mathcal{E}[\max(\gamma_i)] = \Gamma \sum_{i=1}^L (-1)^{i-1} \left[\frac{M!}{i!(M-i)!} \right] \frac{1}{i} = \Gamma \sum_{i=1}^L \frac{1}{i} \quad (3.54)$$

Selection combining is easy to implement and it does not require sophisticated power management or additional communication circuitry. However, this technique is required to monitor wireless channel faster the fading process. Faster scanning allows receivers to select branch with maximum SNR.

3.13 Equal Gain Combining (EGC)

In equal gain combining, received signals from all the branches are coherently combined. Branch weights are set to unity and signals from each branch are co-phased. Received SNR at the output of L -branch equal gain combiner is:

$$\gamma_{egc} = \frac{E_b \left(\sum_{i=1}^L \alpha_i \right)^2}{\sigma^2 L} = \frac{E_b \sum_{i=1}^L \sum_{j=1}^L \alpha_i \alpha_j}{\sigma^2 L} \quad (3.55)$$

Average SNR of equal gain combiner:

$$\bar{\gamma}_{EGC} = \Gamma \left[1 + \frac{\pi}{4}(L-1) \right] \quad (3.56)$$

3.14 Maximum Ratio Combining (MRC)

In L-branch MRC diversity received signal is combined with beamforming weights w . Therefore, the received MRC signal:

$$\mathbf{y} = \mathbf{w}^H(\mathbf{h}\mathbf{x} + \mathbf{n}) = \mathbf{w}^H\mathbf{h}\mathbf{x} + \mathbf{w}^H\mathbf{n} \quad (3.57)$$

In MRC diversity combining the SNR at output of beamformer:

$$\gamma = \frac{|\mathbf{w}^H\mathbf{h}|^2}{\sigma_n^2(\mathbf{w}^H\mathbf{w})} \quad (3.58)$$

Where $E\{n_i\} = \sigma_n^2$ is noise power and $\sigma_n^2\mathbf{w}^H\mathbf{w}$ is scaled noise power or noise power at the output of beamformer. For optimal beamformer that maximises the SNR $\mathbf{w}^H\mathbf{w} = 1$.

3.14.1 BER with MRC Diversity

Received SNR at the output of MRC combiner with two receive side independent fading branches.

$$\gamma = \sum_{i=0}^{L-1} |h_i|^2 P / \sigma_n^2 \quad (3.59)$$

Hence, the instantaneous BER is $Q(\sqrt{\gamma})$. BER with single receive antenna:

$$\text{BER} = \frac{1}{2} \left(1 - \sqrt{\frac{\gamma}{2 + \gamma}} \right), \quad \text{and at high SNR} \quad \text{BER} \approx \frac{1}{2\gamma} \quad (3.60)$$

BER of L-branch MRC at high SNR is given as:

$$\text{BER} = C_L^{2L-1} \frac{1}{2^L} \left(\frac{1}{\gamma} \right)^L \quad (3.61)$$

For two branch MRC, BER is given as:

$$\text{BER} = C_L^{2L-1} \frac{1}{2^L} \left(\frac{1}{\gamma} \right)^L = C_3^2 \frac{1}{2^2} \left(\frac{1}{\gamma} \right)^2 = \frac{3}{4} \left(\frac{1}{\gamma} \right) \quad (3.62)$$

Probability of deep fade decreases with diversity. Probability of deep fade for L-branch diversity:

$$P\left(\sum_{i=0}^{L-1} < 1/\gamma\right) = \left(\frac{1}{\gamma}\right)^L \quad (3.63)$$

The probability of deep with L-branch diversity is small.

3.14.2 Diversity capacity analysis

Information-theoretic understanding of capacity of diversity combined signal can be obtained from the measured results. The capacity and mean of the capacity of EGC,SC and MRC combined signal is obtained as

$$C_{SC} = \log_2(1 + \gamma_{div}) \quad (3.64)$$

Similarly, Ergodic capacity of highly random γ_{div} combined signal is

$$\bar{C}_{SC} = \mathcal{E}[C_{SC}] = \log_2(1 + \gamma_{div}) \quad (3.65)$$

There is a direct influence of power delay profile on the capacity of diversity combined signals. This aspect of the effect of PDP on the capacity of diversity combined signal has been theoretically investigated by authors in [23].

3.15 Conclusion

Wireless channel measurement and modeling is an important stage before a wireless communication system is deployed. This chapter is a summary of wireless channel concepts and related terminology. The most common distribution techniques are Rayleigh, Rician, Weibull distribution, and Nakagami-m distribution have been explained. These distribution techniques can be classified into three categories which encompass all classical fading distributions. Small-scale fading, large-scale fading and path loss equations are reviewed. Additionally, RMS delay spread and Doppler spread are explained. Coherence bandwidth, coherence

time and coherence distance concepts are discussed and how these quantities are inter-related in frequency, time, and space. Second order statistics such as level crossing rate (LCR) and average fade duration (AFD) are important to understand the severity of fading. LCR and AFD relationship in different propagation environment is discussed. Transmit and receive diversity techniques which are vital for the signal quality improvement and BER enhancements are also discussed with the help of diversity expressions.

Bibliography

- [1] Marvin K Simon and Mohamed-Slim Alouini. *Digital communication over fading channels*, volume 95. John Wiley & Sons, 2005.
- [2] John G Proakis and Masoud Salehi. *Digital communications*, volume 4. McGraw-hill New York, 2001.
- [3] Bernard Sklar and Fredric J Harris. *Digital communications: fundamentals and applications*, volume 2001. Prentice-hall Englewood Cliffs, NJ, 1988.
- [4] Athanasios Papoulis and H Saunders. Probability, random variables and stochastic processes. 1989.
- [5] Bernard Sklar. Rayleigh fading channels in mobile digital communication systems. i. characterization. *IEEE Communications magazine*, 35(7):90–100, 1997.
- [6] David Tse and Pramod Viswanath. *Fundamentals of wireless communication*. Cambridge university press, 2005.
- [7] Michel Daoud Yacoub, JE Vargas Bautista, and L Guerra de Rezende Guedes. On higher order statistics of the nakagami-m distribution. *IEEE Transactions on Vehicular Technology*, 48(3):790–794, 1999.
- [8] Hirofumi Suzuki. A statistical model for urban radio propogation. *IEEE Transactions on communications*, 25(7):673–680, 1977.
- [9] George Tzeremes and CG Christodoulou. Use of weibull distribution for describing outdoor multipath fading. In *IEEE Antennas and Propagation Society International Symposium (IEEE Cat. No. 02CH37313)*, volume 1, pages 232–235. IEEE, 2002.

- [10] Nikos C Sagias, Dimitris A Zogas, George K Karagiannidis, and George S Tombras. Channel capacity and second-order statistics in weibull fading. *IEEE Communications Letters*, 8(6):377–379, 2004.
- [11] Theodore S Rappaport et al. *Wireless communications: principles and practice*, volume 2. prentice hall PTR New Jersey, 1996.
- [12] Niklas Jaldén, Per Zetterberg, Bjorn Ottersten, Aihua Hong, and Reiner Thoma. Correlation properties of large scale fading based on indoor measurements. In *2007 IEEE Wireless Communications and Networking Conference*, pages 1894–1899. IEEE, 2007.
- [13] Andreas F Molisch. *Wireless communications*, volume 34. John Wiley & Sons, 2012.
- [14] Tapan K Sarkar, Zhong Ji, Kyungjung Kim, Abdellatif Medouri, and Magdalena Salazar-Palma. A survey of various propagation models for mobile communication. *IEEE Antennas and propagation Magazine*, 45(3):51–82, 2003.
- [15] Robert JC Bultitude. A study of coherence bandwidth measurements for frequency selective radio channels. In *33rd IEEE Vehicular Technology Conference*, volume 33, pages 269–278. IEEE, 1983.
- [16] MO Al-Nuaimi and Andreas G Siamarou. Coherence bandwidth characterisation and estimation for indoor rician multipath wireless channels using measurements at 62.4 ghz. *IEE Proceedings-Microwaves, Antennas and Propagation*, 149(3):181–187, 2002.
- [17] Mercedes Sanchez Varela and Manuel García Sánchez. Rms delay and coherence bandwidth measurements in indoor radio channels in the uhf band. *IEEE Transactions on Vehicular Technology*, 50(2):515–525, 2001.
- [18] Xiaoming Chen, Per-Simon Kildal, Charlie Orlenius, and Jan Carlsson. Channel sounding of loaded reverberation chamber for over-the-air testing of

wireless devices: Coherence bandwidth versus average mode bandwidth and delay spread. *IEEE Antennas and Wireless Propagation Letters*, 8:678–681, 2009.

- [19] Lin Cheng, Benjamin E Henty, Daniel D Stancil, Fan Bai, and Priyantha Mudalige. Mobile vehicle-to-vehicle narrow-band channel measurement and characterization of the 5.9 ghz dedicated short range communication (dsrc) frequency band. *IEEE Journal on Selected Areas in Communications*, 25(8):1501–1516, 2007.
- [20] Nektarios Moraitis and Philip Constantinou. Measurements and characterization of wideband indoor radio channel at 60 ghz. *IEEE Transactions on Wireless Communications*, 5(4):880–889, 2006.
- [21] Xiaofei Dong and Norman C Beaulieu. Average level crossing rate and average fade duration of selection diversity. *IEEE Communications Letters*, 5(10):396–398, 2001.
- [22] Norman C Beaulieu and Xiaofei Dong. Level crossing rate and average fade duration of mrc and egc diversity in ricean fading. *IEEE transactions on communications*, 51(5):722–726, 2003.
- [23] Dejan V Djonin and Vijay K Bhargava. On the influence of the power profile on diversity combining schemes. *IEEE Transactions on Wireless Communications*, 3(5):1854–1861, 2004.

Chapter 4

Near-field MIMO BER and EVM Measurements in Metal Enclosure

4.1 Reverberation Chamber

A reverberation chamber is a large metallic cavity which when excited gives rise to many modes. RC has been used for electromagnetic compatibility (EMC) and electromagnetic interference (EMI) analysis [1],[2], measurement and characterisation of antenna radiation and total efficiency of antennas [3],[4],[5]. Multipath fading environment of the chamber can be controlled by loading the RC with objects whose function is to absorb signal energy such as radiation absorbent material (RAM). The type of absorbing material, location of absorbing material, and the quantity of absorbing material have a huge impact on the field distribution inside the reverberation chamber. Increased absorbing material placement in RC absorbs more energy and field distribution changes from isotropic to non-isotropic. This may cause channel parameters to change drastically. For example, an increased number of absorbers shall absorb more energy and reduce the intensity of fading and reduce RMS delay spread. As a consequence of reduced RMS delay spread coherence bandwidth is increased. This means wideband signals can be used for testing of wireless communication link without frequency-selective fading and inter-symbol interference. OTA testing of wireless devices in loaded reverberation chamber is performed in [6] and an approach to calculate coher-

ence bandwidth using average mode bandwidth has been introduced. Authors have concluded that coherence bandwidth is same as average mode bandwidth. Average mode bandwidth is easier to measure than coherence bandwidth.

4.2 Q-factor and Rician Factor Measurement

Reverberation chamber is often characterized using Q-factor which can be used to estimate important wireless channel parameters such as Rician factor, RMS delay spread and coherence bandwidth. Q-factor of the chamber can be obtained as a function of chamber volume, operating frequency and transmission power [7]:

$$Q = 16\pi^2 V c^{-3} f^3 P \quad (4.1)$$

Where V is volume of the chamber, c is speed of light, f is center frequency, P is transmitted power. K-factor can be derived from Q-factor [8].

$$K = \frac{3VD}{2\lambda Qd^2} \quad (4.2)$$

Where $D = d/\lambda$ is spatial diversity and d is antenna inter-element distance. Substituting for D and Q in above equation Rician factor can be rewritten as:

$$K = \frac{3D\lambda^2}{32\pi^2 P d^2} \quad (4.3)$$

It is shown in [3] that in the chamber the delay spread (τ_{RMS}) is directly related to the Quality factor of the chamber given by.

$$\tau_{RMS} = \frac{Q_{total}}{\omega}$$

RMS delay spread is inversely proportional to average mode bandwidth or coherence bandwidth. This implies Q factor is also inversely related to coherence bandwidth. Q factor is reduced by introducing the losses in the chamber i.e. in-

creasing the number of absorbers and hence increasing the coherence bandwidth.

4.3 Near-field Measurements using QPSK Receiver

QPSK receiver was implemented using both GNURadio as well as using MATLAB Simulink. Key elements of digital receiver design were realized using signal processing libraries from both MATLAB/Simulink and GNU Radio. In the following section, Simulink based QPSK digital receiver design stages are discussed. This receiver model has been used for 2x2 MIMO BER and EVM measurements. QPSK receiver implementation of SDR based on MATLAB/Simulink was used for the BER and EVM measurements in a mode-stirred metal enclosure. QPSK receiver mainly consists of the following stages:

4.3.1 Automatic Gain Control

This stage is responsible for stabilizing the received signal amplitude. Its tunable parameters are the step size (slow and fast attack) and maximum desired output power. The step size of AGC can have a significant effect on the input signal. Received signal variation may have significant impact on the end-to-end BER performance. AGC uses fast attack in case of higher signal variation to obtain lock to desired maximum power and slow attack in case of small signal variation. This stage affects end-to-end BER performance of the wireless communication system for changes in input power values. Fig. 4.1 shows estimate of the input power available at AGC when the Tx gain values were tuned from 2 dB to 12 dB . It can be seen that estimate of input power of AGC gain increases with the increase in USRP Tx gain.

Fig. 4.2 shows AGC output signal magnitude as a function of AGC gain values. It can be seen that the signal magnitude increases with the increase in AGC gain values from 20 dB to 60 dB .

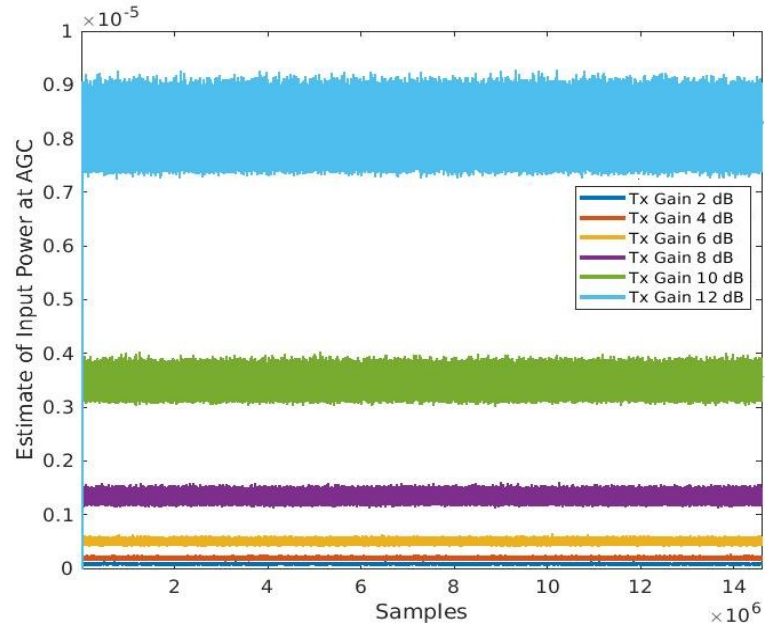


Figure 4.1: Estimate of input power at AGC when Tx gain of the USRP was swept from 2 dB to 12 dB and Rx gain of the USRP was set to 0 dB. AGC step size was 0.01. The measured estimate of input power is in the metal enclosure in presence of NLOS metal sheet between transmitter and receiver.

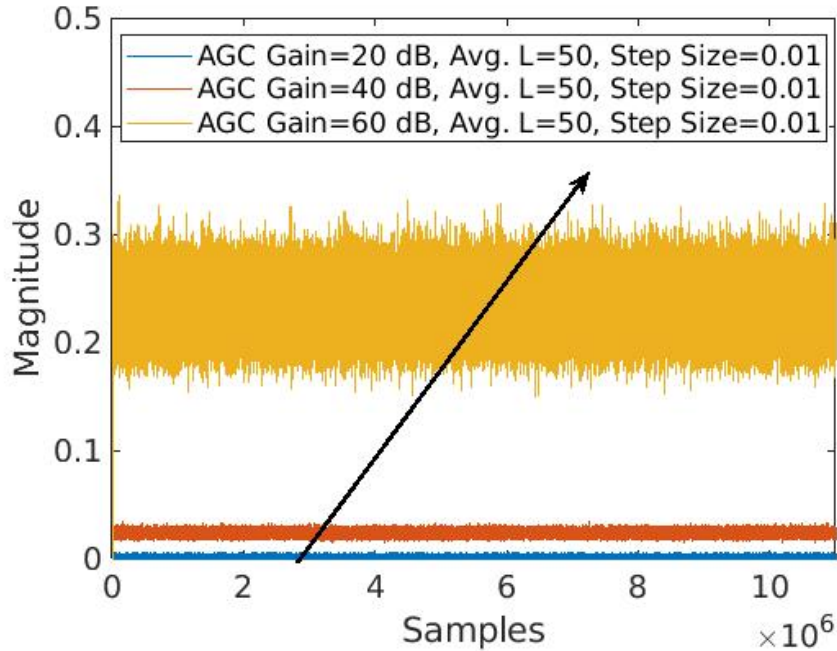


Figure 4.2: Output signal magnitude with AGC gain values of 20 dB, 40 dB, and 60 dB respectively. The measurements were performed in metal enclosure with Tx gain of USRP equal to 2 dB and Rx gain of USRP equal to 0 dB.

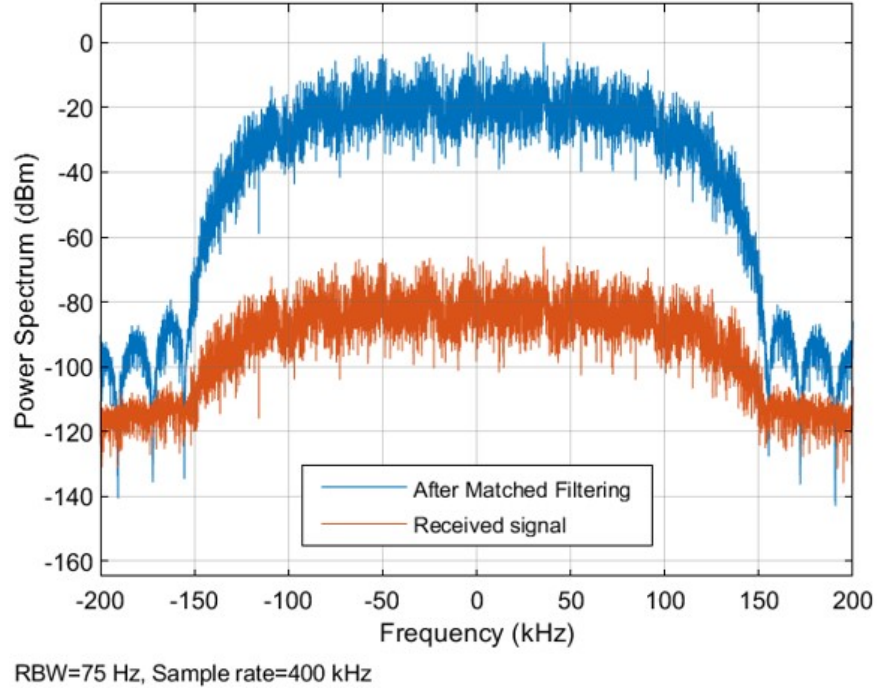


Figure 4.3: PSD of received signal before and after the matched signal.

4.3.2 Transmit and Receive RC Filter

RC filtering is a pulse shaping operation to suppress out-of-band radiation and therefore more channels can be accommodated in a crowded spectrum. This filter is implemented in two stages one half is implemented at the transmitter and another half is implemented at the receiver[9, 10]. This operation is also known as matched filtering. Matched filter is a linear filter that increases signal-to-noise ratio (SNR) in presence of stochastic noise. The performance of this filter depends on the roll-off factor and the filter order. Higher filter order requires more computational resources and it may easily clog computer when implemented as part of the communication system on host PC with limited resources. Fig 4.3 shows PSD of received signal before and after performing matched filtering with root cosine receive filter. It can be clearly seen that matched filtering increases signal strength hence SNR of the received signal.

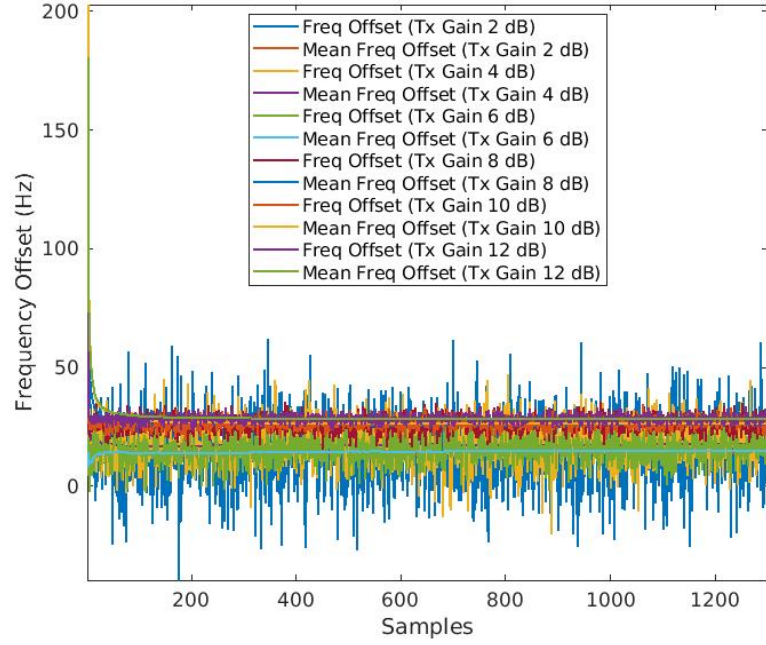
4.3.3 Coarse Frequency Compensation

The coarse frequency compensation stage roughly estimates the frequency offset present in the signal. It corrects for a rough estimate of frequency offset. Residual frequency offset compensation is performed by a separate stage designed for the purpose of the residue of frequency offset compensation and it's known as fine-frequency compensation. Coarse-frequency compensation uses an FFT-based or correlated-based algorithm to correct the rough estimate of frequency offset. Using FFT-based algorithm frequency deviation $\Delta\hat{f}$ can be given as[11]:

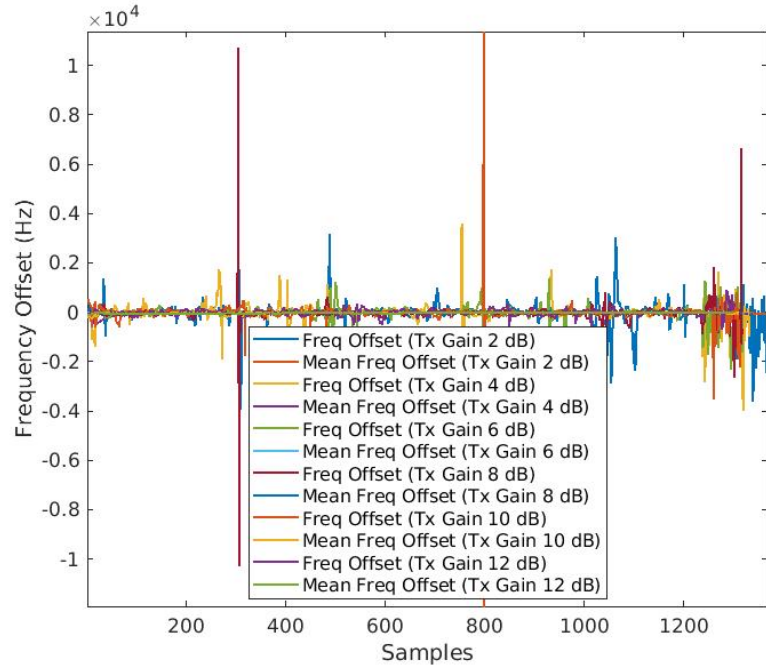
$$\Delta\hat{f} = \frac{f_{\text{samp}}}{N \cdot m} \arg \max_f \left| \sum_{k=0}^{N-1} r^m(k) e^{-j2\pi k/N} \right|, \quad \left(-\frac{R_{\text{sym}}}{2} \leq f \leq \frac{R_{\text{sym}}}{2} \right), \quad (4.4)$$

Where R_{sym} is symbol rate, $r^m(k)$ is m^{th} power of received signal. Frequency deviation is estimated by periodogram of the m^{th} power of received signal. QPSK constellation continuously rotates clock-wise or anti-clockwise this depends on whether the frequency offset is positive or negative.

Fig. 4.4 shows frequency offset and mean frequency offset when the stirrer was off and stirrer was on. It can be seen from Fig. 4.4a that there is a frequency offset with its mean falling below 50 Hz. Frequency offset and mean frequency offset is shown for six Tx gains values of X310 USRP. Similarly, Fig. 4.4b shows frequency offset and mean frequency offset when the stirrer was turned on. It can be seen that frequency offset is time-varying which is evident because of stirrer pedal rotation. Low and high maximum gain values can be observed as the position of the stirrer changes. In response to stirrer pedal position and received signal strength the response of AGC changes and as a result coarse frequency compensation block which relies on AGC output also changes and produces frequency offset measurement values. It can be noticed there are instances of high frequency offsets which can be tracked by coarse frequency compensation block. The compensation depends on tracking range of the coarse frequency compensation block which is 12.5 KHz in the experiments. It's sufficient value to track the



(a) Frequency offset and mean frequency offset when stirrer is off



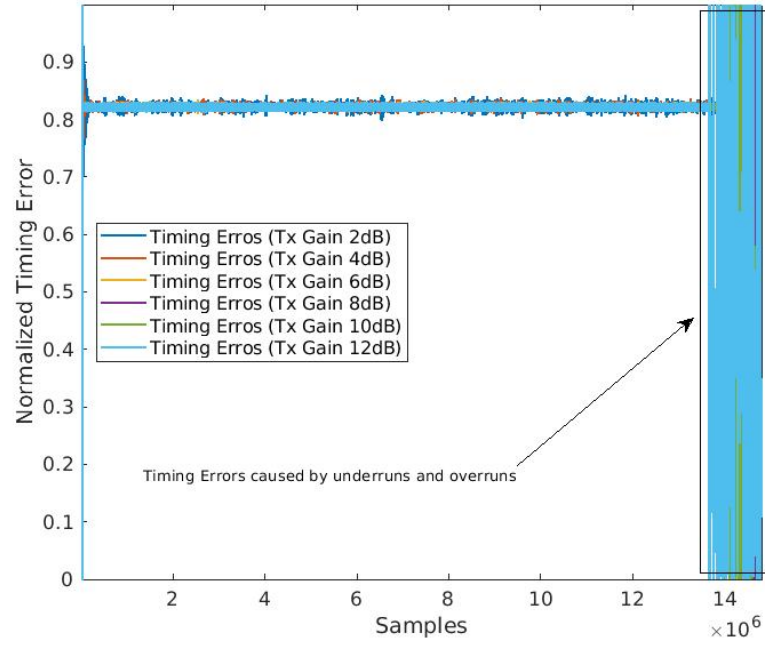
(b) Frequency offset and mean frequency offset when stirrer is on

Figure 4.4: Frequency offset and mean frequency offset in metal enclosure when the stirrer was off and when it was on. The measurements were performed in metal enclosure with NLOS metal sheet and the Tx gain was swept from 2 dB to 12 dB and the RX gain of the Rx USRP was kept 0 dB.

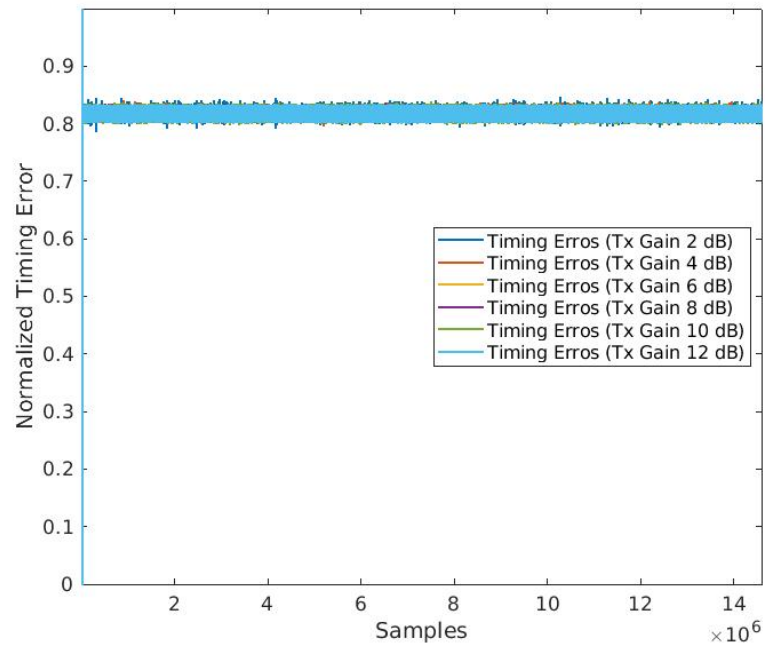
frequency offset. Three instants of maximum frequency offset are within tracking range of coarse frequency compensation block. If any instant of frequency offset is higher than the tracking range of the coarse frequency compensation stage then it may go uncorrected. It has a direct effect on the performance of the carrier frequency or fine-frequency compensation stage. It hurts the BER of two USRP channels.

4.3.4 Symbol Synchronization

The symbol synchronization stage is one of the crucial steps in the design of the digital receiver. There are different algorithms used for symbol synchronization. This stage is the PLL-based stage. It corrects the timing errors i.e. groups the received samples around ideal sampling points. The tunable parameters of this stage are the damping factor and normalized loop bandwidth. This stage uses a non-data aided timing recovery algorithm in a timing error detector (TED). Symbol synchronization immediately follows the coarse-frequency compensation stage. This is the reason that QPSK constellation rotates because there is still uncorrected residual frequency offset present. The effect of symbol synchronization or the timing recovery is seen after the fine-frequency compensation or carrier synchronization stage. Fig. 4.5 shows normalized symbol timing errors. It can be seen from Fig. 4.5a that there is initial timing error transient and later it settles down when the PLL has converged. There are also high values of transients at the end of Simulation for each Tx Gain values. It was observed that these timing errors are caused by underruns and overruns at transmitter and receiver respectively. It's clear from these measurements that when there are underruns and overruns data is lost. The receiver is not running in real-time. There are timing errors and hence the BER and EVM may not be tractable and reliable for end-to-end performance analysis of communication system. In contrast to Fig. 4.5a which shows underruns and overruns at the very end. After following the guidelines to overcome underflow and overflow as suggested in previous chapter,



(a) Symbol timing error in presence of underruns and overruns



(b) Normalized symbol timing error without underruns and overruns

Figure 4.5: Normalized symbol timing errors in metal enclosure with NLOS metal sheet. The Tx gain swept from 2 dB to 12 dB and the Rx gain of the Rx USRP was kept 0 dB.

the Fig. 4.5b shows no underflow except initial transient period. The BER and EVM are reliable in this case. There was no inconsistency observed in the measurements. In this case in absence of underflows and overflows if there is data loss then it should be attributed to reduced signal strength or strong in-band interference. The QPSK receiver and SDR based measurement setup is tested extensively for BER measurements. End-to-end BER measurement was tested as a function of each block in the receiver independently also the chain effect of the tunable parameters in each block was considered. This resulted in hundreds and thousands of experiments to reach at the point of confidence after which measurements are included in thesis. BER measurement were performed for longer duration to check the stability of testbed and reliability of QPSK receiver stages. It was tested in metal enclosure so that no outer RF interference affected the measurements. Since end-to-end BER measurement were conducted reliably as a function of Tx USRP gain and after carefully considering the QPSK receiver parameters. Only problem was when the samples were lost because of the underflow and overflow. Measurements were conditioned to stop on any instance of underflow and/or overflow.

4.3.5 Fine Frequency Compensation

Residual frequency offset that is not corrected by coarse frequency compensation stage is corrected at this stage. This stage compensates residual frequency offset and the effect can be seen in the form of no rotation of QPSK constellation. Received samples are grouped and are closer to ideal sampling point. This stage is also PLL-based and its tunable parameters are the damping factor and the normalized loop bandwidth. This block removes rotation and establishes lock to a particular constellation state and hence it may or may not introduce phase ambiguity. Fig.4.6 shows measurement setup where USRP X310 is connected to host PC via Ethernet cable. The output and input port of the USRP X310 is connected with the transmitting and receiving vertically polarized monopoles

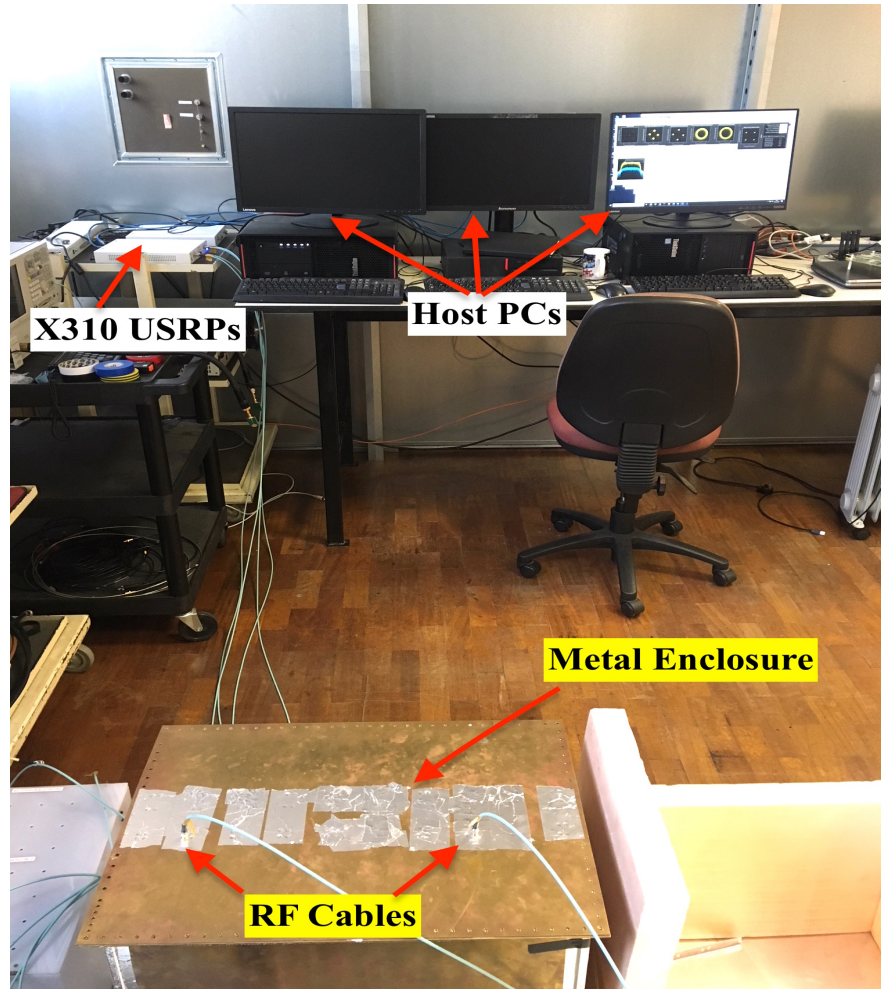


Figure 4.6: Measurement setup for QPSK receiver and BER measurement in the metal enclosure. The measurements are performed when the Tx gain and Rx gain of the USRP X310 are set to 10 dB and 0 dB respectively.

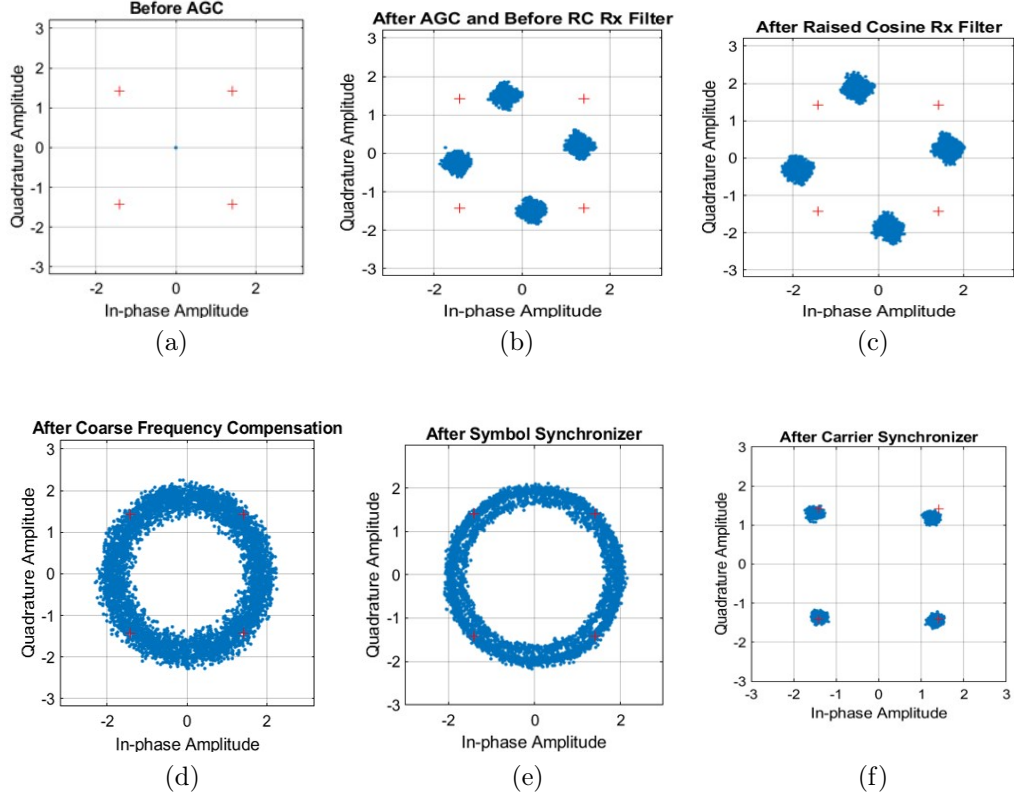


Figure 4.7: QPSK constellation diagram at different stages of the digital receiver (a) before AGC (b) after AGC (c) After Rx RC (d) after coarse frequency compensation (e) after symbol synchronisation (f) after fine frequency compensation. The constellation plots are recorded from an experiment at 5.6 GHz in the metal enclosure.

operating at 5.6 GHz.

4.3.6 Preamble Detector

One of the important stage for data detection is frame detection. This is done by adding preamble to the data frame. Preamble can be sequence of 1s and 0s. However, well-structured preamble are already investigated. Hence, QPSK receiver uses Barker-13 code as a preamble for frame detection. Tunable parameter for this stage is detection threshold. Therefore, the measurements are sensitive to detection threshold. There is trade-off in selection of threshold being too tight and too relaxed. Preamble detector generates too much symbol index when the threshold is too low and frame detection is missed if the threshold set too high. Fig.4.8 shows preamble detection metric where from we can choose threshold.

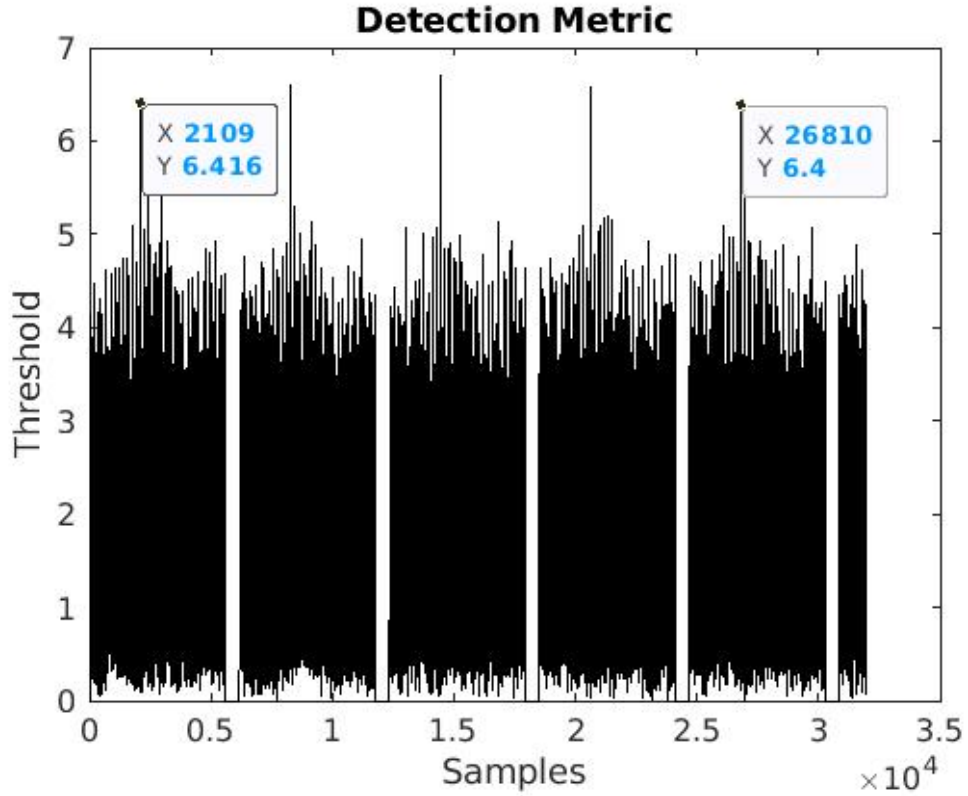


Figure 4.8: Preamble detection metric.

For too low threshold such as 0.6 there are many preamble locations as shown in Fig.4.9a. Ideally there should be one preamble location per frame. Therefore, from Fig. 4.8, a threshold of 6 can give only one preamble location per frame. Fig.4.9b shows this was achieved by setting threshold metric to 6.

4.3.7 Frame Synchronization

Barker-13 code is useful for frame synchronisation. Barker code is over sampled by factor of 2 and used as frame header. Therefore, auto-correlation of received frame results in peaks at every 13th bit. After performing frame synchronization frame header is removed and rest of the frame samples are down streamed for data decoding and BER measurement.

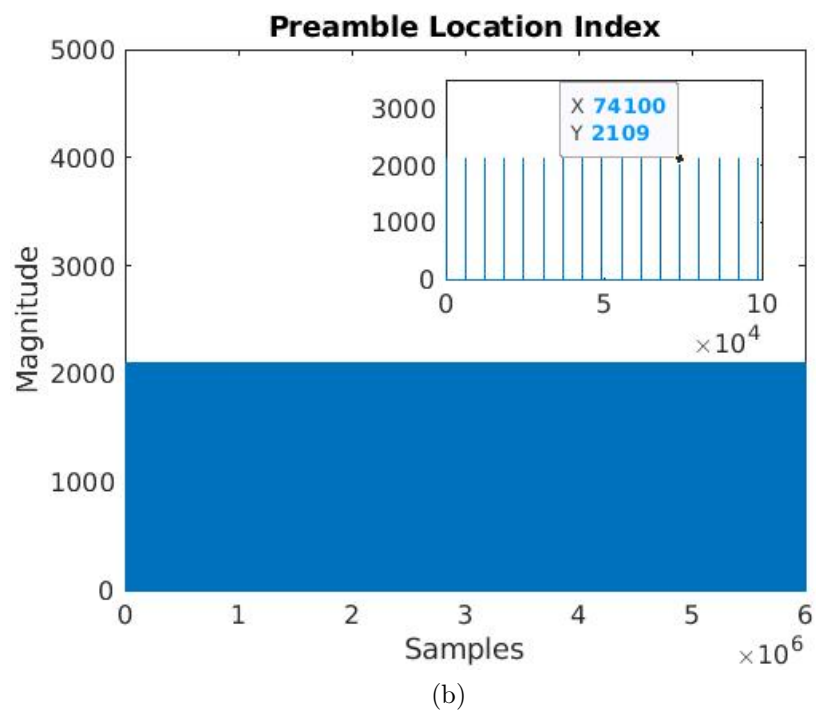
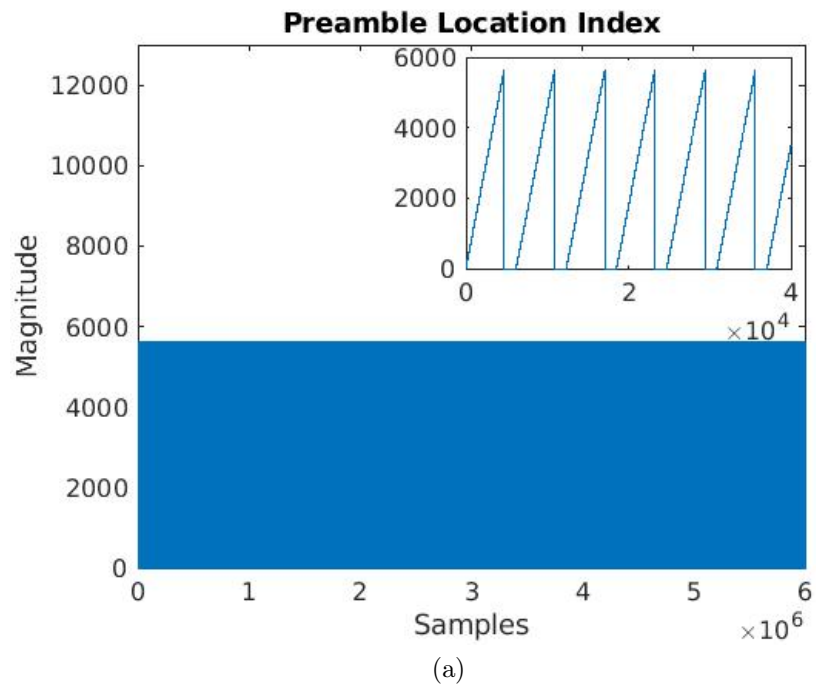


Figure 4.9: Preamble detection (a) Preamble index with threshold=0.6 (b) Preamble index with threshold=6.

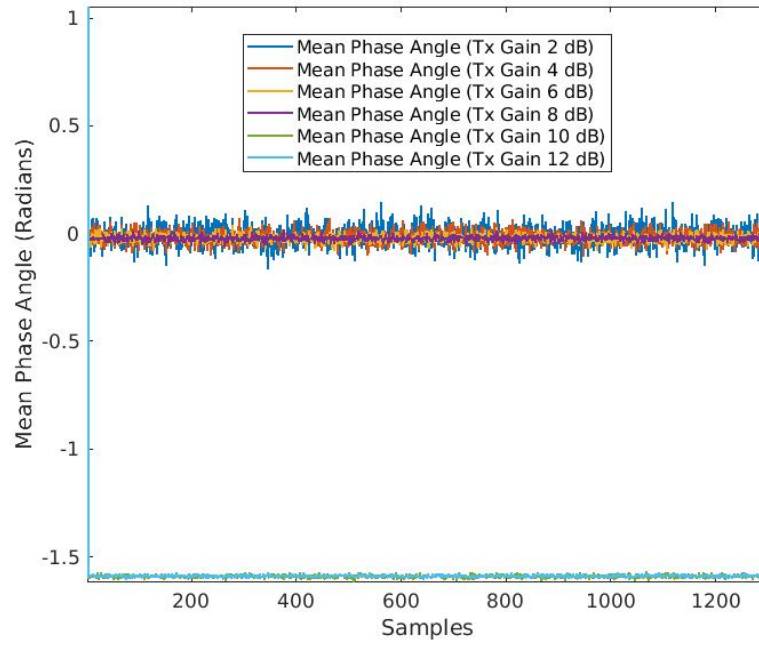


Figure 4.10: Mean phase angle in radians input to phase ambiguity resolution block.

4.3.8 Phase Ambiguity Resolution

After the frame synchronisation data decoding is performed where the transmitted text or image is recovered. Phase ambiguity is realized as part of data decoding stage. Phase ambiguity resolution is used to correct phase ambiguity introduced in the constellation. Received constellation diagram after different stages of synchronization may visually look locked to ideal constellation points but it might have incurred phase shift. Carrier synchronizer or fine-frequency compensation block in the digital receiver locks to phase which can be compensated by phase ambiguity block shown in Fig. 4.10. It can be seen that the carrier synchronizer locks to a different phase every time Simulink restarts for Tx gains values of the USRP.

The mean estimate phase angle is input to phase ambiguity resolution block and the phase ambiguity for QPSK is resolved by the following code.

```
1 function Out = phasedecision(In)
2 % Map the estimated phase to one of -pi, -pi/2, 0, pi/2,
3 %#codegen
4 Out = round(In.*2./pi)./2.*pi;
```

```

5 %% OR
6 % The above processing is equivalent to the following ...
   processing using
7 % if/else statements
8 if In<pi/4 && In>-pi/4
9     Out = 0;
10 elseif In>=pi/4 && In<pi*3/4
11     Out = pi/2;
12 elseif In>=pi*3/4 || In<=-pi*3/4
13     Out = -pi;
14 else
15     Out = -pi/2;
16 end

```

The phase ambiguity resolution for BPSK modulation which has only two possible phases i.e 0 or π is resolved using following MATLAB code.

```

1 function Out = phasedecision(In)
2 if In<pi/2 && In>-pi/2
3     Out=0;
4 else
5     Out=pi;
6 end

```

Alternatively, differential encoded modulation techniques are suitable for transmission such as DBPSK, DQPSK, DMPSK etc where differential phase between adjacent symbols is transmitted instead of the absolute phases.

4.4 BER and EVM Measurements

Error vector magnitude (EVM) and bit error rate (BER) are two important communication system metrics. EVM is useful quantity to measure algorithmic performance of communication systems. EVM is a difference between ideal constellation point and the received constellation samples around ideal constellation point[12].

$$EVM_{RMS} = 100 \times \sqrt{\sum_{k=0}^{N-1} e_{const}(k) \sum_{k=0}^{N-1} (\text{Re}(\bar{y}(k))^2 + \text{Im}(\bar{y}(k))^2)} \quad (4.5)$$

where

$$e_{const}(k) = (\text{Re}(y(k)) - \text{Re}(\bar{y}(k)))^2 + (\text{Im}(y(k)) - \text{Im}(\bar{y}(k)))^2 \quad (4.6)$$

Where $\bar{y}(k)$ is reference symbol for $y(k)$. EVM in dB can be written as:

$$EVM_{dB} = 20 \log_{10} \left(\frac{EVM_{RMS}}{100} \right) \quad (4.7)$$

Tx gain (dB)	P_{Rx} (dBm)	BER	EVM (%)
2	-75.49	0.1519	85
4	-73.93	0.05579	34.6
6	-72.17	0.01217	24.5
8	-70.23	0.00056	30.8

Table 4.1: Measured received power, BER and EVM. Averaging length of AGC and USRP Rx gain is set to 60 and 0 dB respectively.

Averaging Length	BER	EVM (%)
50	0.01217	30.8
60	0.0126	29.8
80	0.00987	30
100	0.00811	28.9
120	0.0098	29.1

Table 4.2: BER and EVM measurement with transmitter gain Tx=6 dB and receiver gain Rx=0 dB.

Max. Avg. Power Gain (dB)	BER	EVM (%)
50	0.0095	95.7
60	0.01044	86.7
70	0.01054	57.8
80	0.01042	29.9

Table 4.3: BER and EVM measurement as a function of maximum average power gain of AGC. AGC average length=60, AGC step size=0.01, Tx gain=6 dB and Rx gain=0 dB.

4.5 MIMO BER and EVM Measurements

Near-field MIMO BER and EVM measurements are performed using wideband antennas at frequency of 5.6 GHz in the metal enclosure. For this four wide-band antennas have been used. The distance between transmitter and receiver

Step Size	BER	EVM (%)
0.01	0.01217	30
0.001	0.01117	29
0.0001	0.01099	28

Table 4.4: BER and EVM measurement as a function of AGC step size. AGC averaging length is set to 60 and AGC maximum power gain is 80.

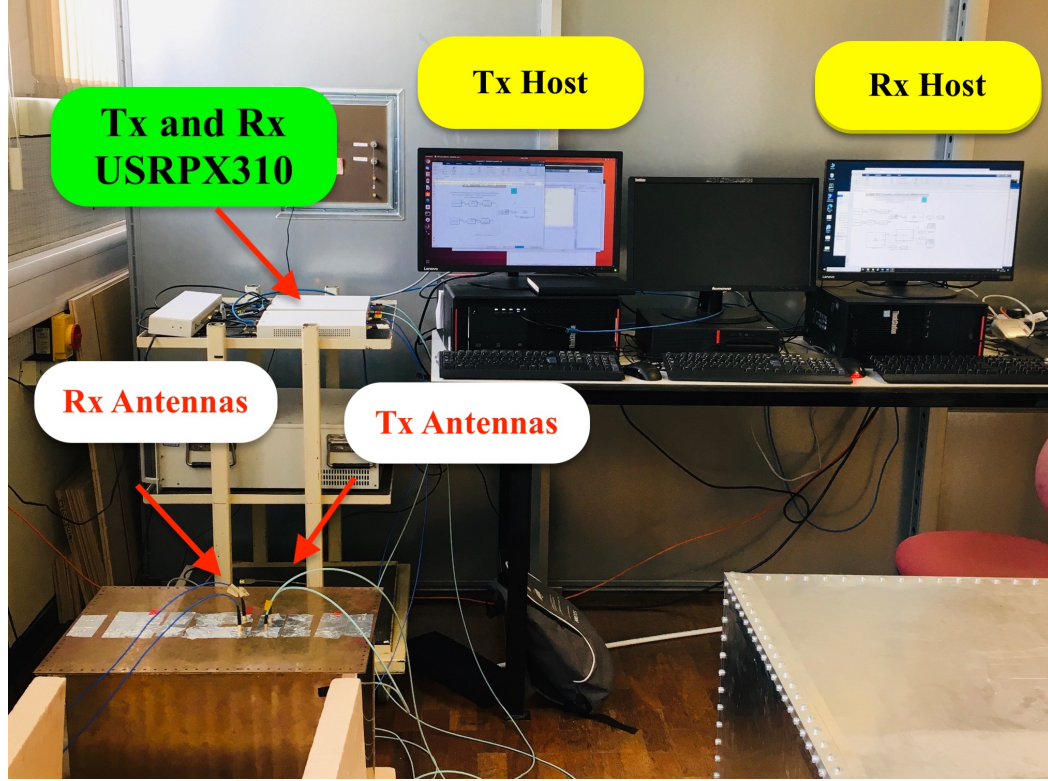
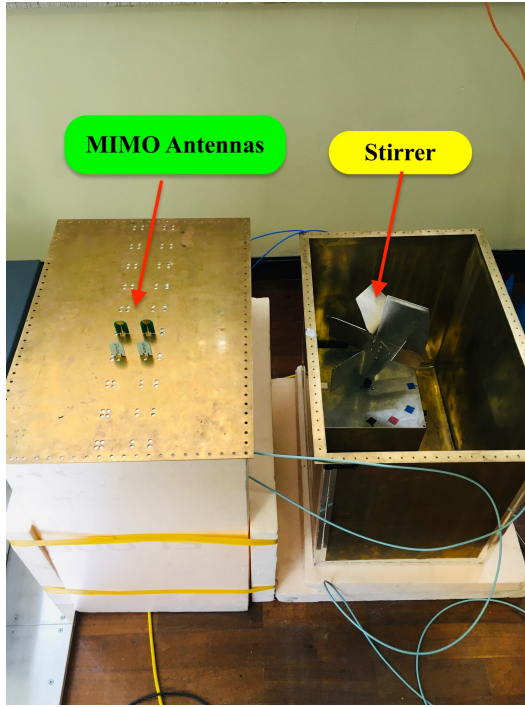


Figure 4.11: Overall measurement setup for Spatial BER and EVM measurement.

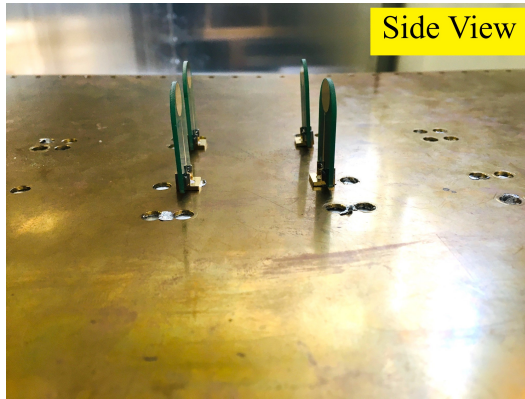
is kept fixed and the distance between transmit elements and receive antenna elements is also kept fixed for all RF frequencies. The overall measurement setup for MIMO BER and EVM measurement is shown in Fig. 4.11 where vertically polarized UWB transmit and receive antennas are mounted such that Tx and Rx antennas face each other. Tx and Rx UWB antennas are connected to Tx and Rx ports of X310 USRPs. Data transfer between host PC and USRPs takes place over 1 Gigabit Ethernet cable. There is no underflow and overflow during the measurements. The measurements are performed in stationary channel conditions. Fig. 4.11 shows overall measurement setup which consists of two host PCs, two USRPs, Octoclock-G, metal enclosure and broadband antennas. Fig.



(a)



(b)



(c)



(d)

Figure 4.12: 2x2 MIMO BER and EVM measurement setup inside cavity (a) MIMO antennas and inside view of cavity (b) front view of antennas (c) side view of antennas (d) top view of antennas.

4.12 shows internal view of the metal enclosure including stirrer, broadband antennas attached to metal enclosure panel. This also shows front view, side view and top view of broadband MIMO antenna configurations. BER measurement is performed using the QPSK receiver. Before running the model USRPs were frequency calibrated. Frequency offset of 17.19 Hz was found using frequency calibration transmitter and receiver baseband models which is tractable by coarse frequency compensation block. The baseband QPSK receiver is configured to operate in spatial multiplexing with two decodable streams. The 2x2 MIMO QPSK baseband transmitter and receiver is shown in Fig.4.13. The baseband model show two streams are being reshaped into two column vectors for digital upconversion and transmission over the channel using two USRP channels. Additionally, it can be seen that the SDRuTransmitter and SDRuReceiver blocks are set to count total number of underruns and overruns. BER display shows BER, the total number of received bits and the number of bits that are corrupted. The flowchart of the measurement process is shown in Fig. 4.14 which shows critical design and measurement steps. The flowchart is has multiple conditions one of the critical is underflow and overflow. The flowchart summarizes all the steps from baseband models and USRP initialization to loading of the metal enclosure.

There are many tunable parameters in the QPSK transmitter and receiver baseband model. At the baseband receiver side the parameters can be adjusted to meet the requirement for understanding the effect of channel on key performance indicators such as BER and EVM. The USRP configuration parameters are summarized in the table 4.5. Additionally, the detailed parameter configuration for SDRuTransmitter and SDRuReceiver block are shown in appendix-B. SDRuTransmitter and SDRuReceiver block contains fields for USRP configurations such as channel mapping, center frequency, LO offset, number of samples, master clock rates, and interpolation/decimation factors. Earlier version of the MATLAB also required sampling time field which was calculated based on master clock rates and the decimation factor at the receiver.

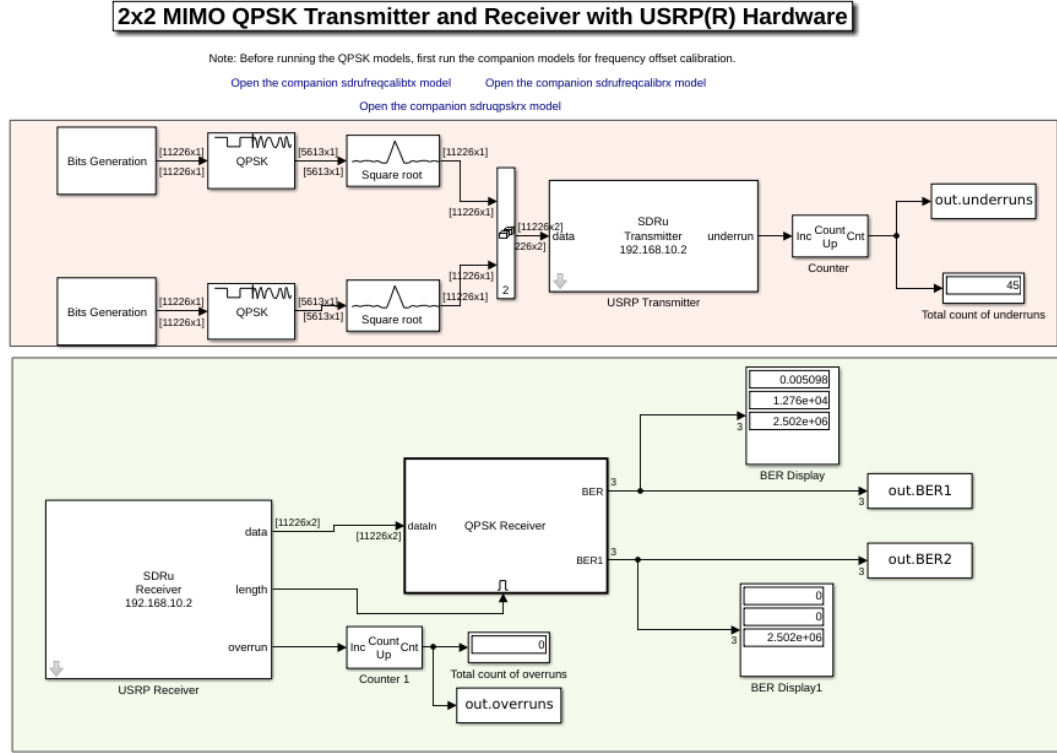


Figure 4.13: 2x2 MIMO QPSK transmitter and receiver Simulink baseband model. The Tx streams are shaped to produce two column vectors for each of the two USRP channels. In GNU Radio the two streams can be fed directly to UHD transmitter block.

Parameters	Values
Channel Configuration	MIMO
Channel Mapping	[1 2]
Frequency	5.6 GHz
Sampling Rate	400 ksps
Clock Rate	200 Mbps
Interpolation/Decimation	500

Table 4.5: USRP configuration and QPSK baseband model parameters.

Tx Gain (dB) Rx Gain=0 dB	Channel 1			Channel 2		
	Prx(dBm)	EVM (%)	BER	Prx(dBm)	EVM (%)	BER
2	-71	87.2	0.03344	-68	81.8	0.007452
4	-69	84.2	0.00787	-66	77.2	0.003448
6	-67	79	0.002508	-64	70	0.002497
8	-61	58	0.003508	-61	61	0.004007

Table 4.6: BER and EVM measurement in the metal enclosure using 2x2 MIMO. Maximum power gain of AGC is set to 60 and RF frequency is 5.6 GHz.

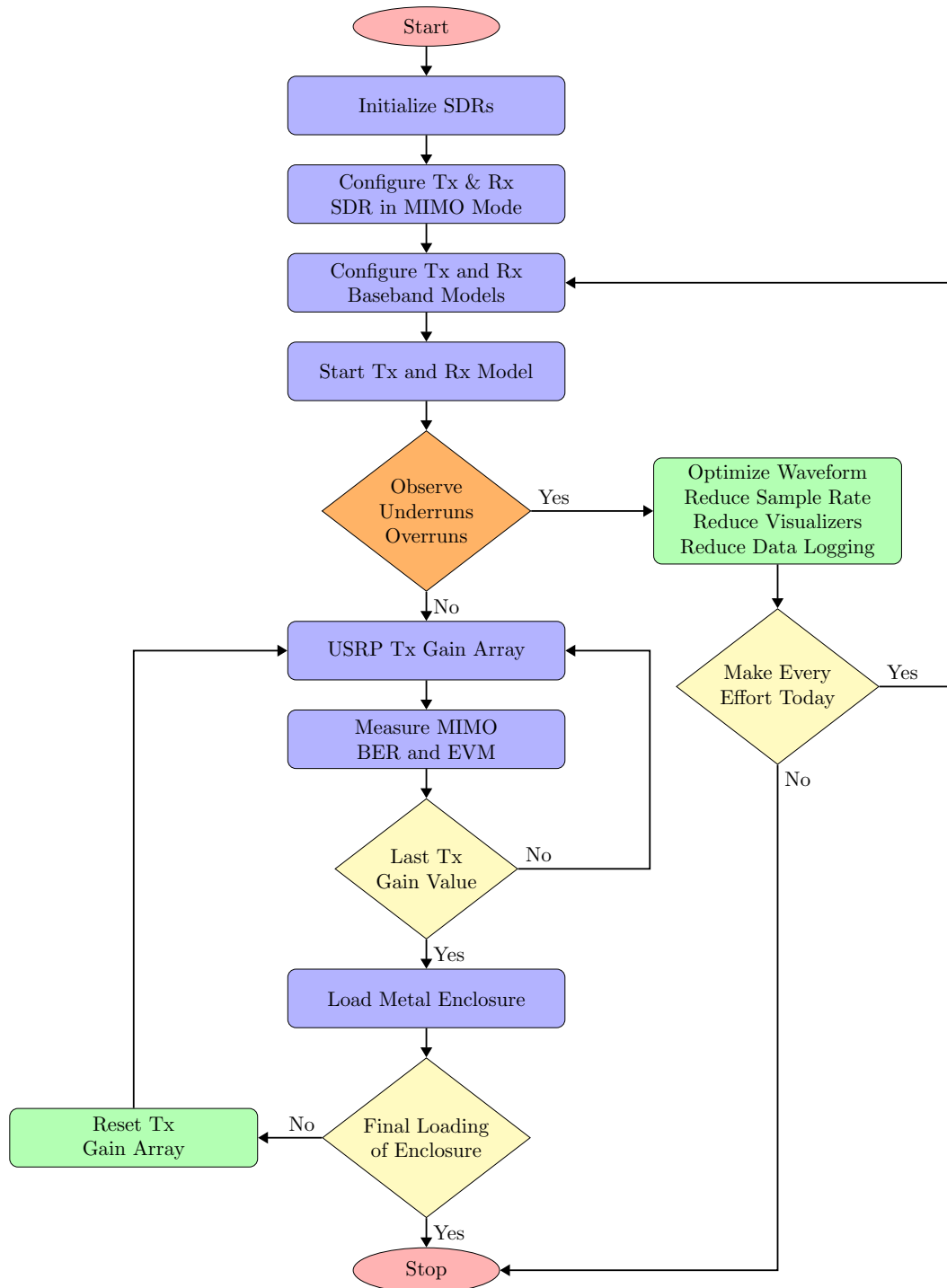


Figure 4.14: Flowchart of MIMO BER and EVM measurement process.

Tx Gain (dB) Rx Gain=0 dB	Channel 1			Channel 2		
	Prx(dBm)	EVM (%)	BER	Prx(dBm)	EVM (%)	BER
2	-67.23	22.3	0.03473	-68.08	20.5	0.005305
4	-65.31	17.3	0.005122	-66.1	18.4	0.004709
6	-63.19	14.6	0.009308	-64	15.5	0.009397
8	-61.12	13.7	0.005441	-61.94	13.3	0.005141

Table 4.7: BER and EVM measurement in the metal enclosure using 2x2 MIMO. Maximum power gain of AGC is set to 80 and RF frequency is 5.6 GHz.

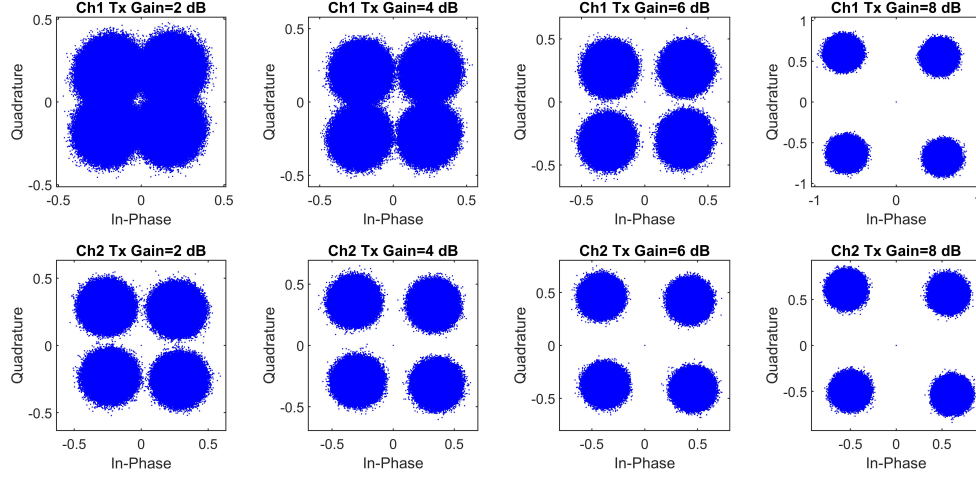


Figure 4.15: Channel 1 and channel 2 QPSK constellation diagram at 5.6 GHz. AGC gain=60, Rx Gain=0 dB and Tx Gain varied from 2 dB to 8 dB with step of 2 dB.

Table. 4.6 shows BER and EVM measurements of two USRP channels at frequency of 5.6 GHz and AGC gain of 60. It can be seen that the BER and EVM of two SDR channels improves as the TX gain of the USRP is increased. Table 4.7 shows BER and EVM measurement of two USRP channels at frequency of 5.6 GHz and AGC gain of 80. It can be seen that the BER and EVM of two SDR channels improves as the Tx gain of the USRP is increased. A noticeable difference in the measured BER and EVM values can be observed for AGC gain of 60 and 80 respectively. For both the cases BER and EVM values improve however, the BER values decrease faster for AGC gain value of 80. This is because of the extra increase in the AGC gain.

Fig, 4.15 shows QPSK constellation diagram of two USRP channels. The QPSK constellation diagram is recorded after carrier synchronization for different Tx gain values of the USRPs. It can be seen that the USRP channel 2 is better

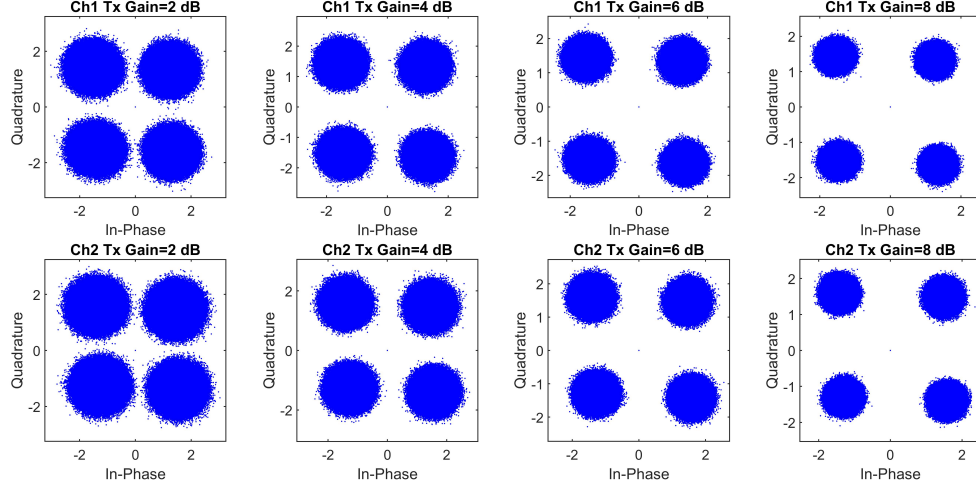


Figure 4.16: Channel 1 and channel 2 QPSK constellation diagram at 5.6 GHz. AGC gain=80, Rx Gain=0 dB and Tx Gain varied from 2 dB to 8 dB with step of 2 dB.

than USRP channel 1. Additionally, the two USRP channel QPSK constellation diagram improves as the Tx gain of the USRP increases which corresponds to improved BER and EVM measurements. Fig. 4.16 shows QPSK constellation diagram of two USRP channels for AGC gain of 80 and different Tx gains. The QPSK constellation diagram is recorded after carrier synchronization for different Tx gain values of the USRPs. It can be seen the two USRP channels have better performance at low Tx gain values in contrast to when AGC gain was 60. Similar to the theoretical predictions and previous measurements, the two USRP channel QPSK constellation diagrams improve as the Tx gain of the USRP increases which corresponds to improved BER and EVM measurements. The measurements are performed at stationary channel conditions when the mode-stirrer was off. Mode-stirrer rotation has direct implications on the measured BER and EVM values because when mode-stirrer rotates channel becomes non-stationary. The channel power gain of the MIMO communication link varies drastically as a result AGC at the receiver tries to compensate low and high gain values to stabilise received signal for carrier frequency offset correct. Hence EVM measurement of two USRP channels is not stationary because of time-varying channel power gain and time varying frequency offset. The same measurements were repeated when the mode-stirrer was turned on. In presence of mode-stirrer rotation the MIMO channel

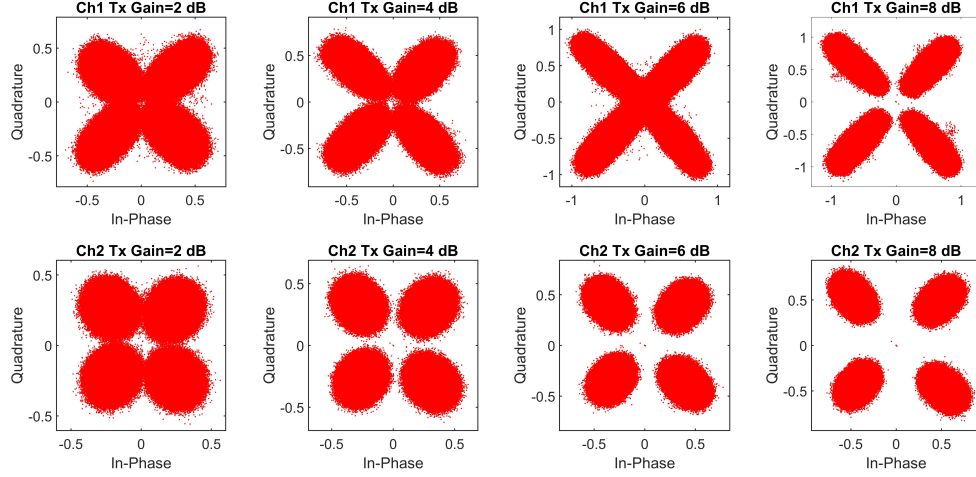


Figure 4.17: Channel 1 and channel 2 QPSK constellation diagram at 5.6 GHz in mode-stirred metal enclosure. AGC gain=60 dB, Rx Gain=0 dB and Tx Gain varied from 2 dB to 8 dB with step of 2 dB.

is time-varying, therefore the constellation diagram and MIMO digital receiver KPIs are also time-varying. Figure 4.17 shows QPSK constellation diagram of two SDR channels. It can be seen that the QPSK constellation diagrams are also time-varying as the mode-stirrer rotates the received power values changes whose effect is visualized in the form of constellation diagram with low and high gain values. This means the BER and EVM are also time-varying. Average BER and EVM of two SDR channels is summarized in table 4.8.

Fig. 4.18 shows time-varying EVM of two SDR channels for Tx gain values from 2 dB to 8 dB respectively. It can be seen that the SDR channel 1 EVM has high variation in max and min values of EVM. SDR channel 2 is more stable. Note that the possible cable faults were ruled out by exchanging the two USRP channels which resulted in exactly similar performance. In other words, SDR channel 1 EVM was better than SDR channel 2 EVM.

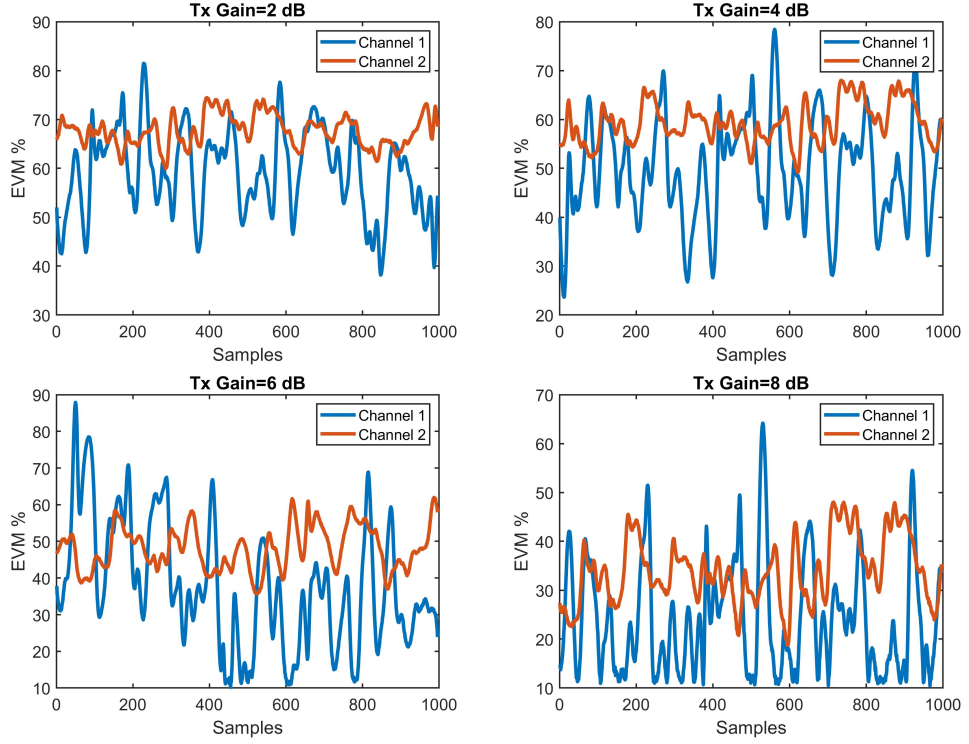


Figure 4.18: Channel 1 and channel 2 instantaneous EVM measurement at 5.6 GHz in mode-stirred metal enclosure. AGC gain=60 dB, Rx Gain=0 dB and Tx Gain varied from 2 dB to 8 dB with step of 2 dB.

Tx Gain (dB) Rx Gain=0 dB	Channel 1			Channel 2		
	Avg. Prx(dBm)	Avg. EVM (%)	BER	Avg. Prx(dBm)	Avg. EVM (%)	BER
2	-69.31	59.15	0.01453	-69.18	67.63	0.005529
4	-65.62	49.61	0.007342	-67.30	59.27	0.000832
6	-63.92	37.84	0.005759	-65.27	48.08	0
8	-61.50	23.9	0.001682	-63.16	33.99	0

Table 4.8: Two Channel BER and EVM measurement in mode-stirred metal enclosure at 5.6 GHz. Maximum AGC=60, Rx Gain=0 dB and Tx Gain is varied from 2 dB to 8 dB with step of 2 dB.

4.6 MIMO BER and EVM Measurements in Presence of MIMO Interference

The EVM and BER of the QPSK receiver is tested in presence of MIMO interference generated by B210 USRP. The measurement setup is shown in Fig. 4.19 which shows MIMO antenna configuration and MIMO antenna interference antennas attached to the enclosure lid. B210 USRPs which generated MIMO interference can also be seen on the enclosure lid. The Tx and Rx antennas are placed at a distance of 50 mm and inter-element distance between transmit and receive elements is 45 mm . The MIMO interference signal which is white Gaussian noise is generated using B210 USRP. GNU Radio based flowgraph as shown in Fig. 4.20 which has been used to generate white Gaussian noise. Fig. 4.21 shows IQ data and histogram of the generated noise signal. The noise parameters such as its mean and variance can be controlled using QT GUI slider. Additionally, the Tx gain of B210 USRP can be changed dynamically. The noise samples can be generated for the required time period using Head block of GNU Radio. The head and skip head block can be used to receive a fixed number of samples. This operations is quite helpful for repeatable measurements and monitoring the channels effects of key performance indicators. The embedded python block can also be used for specialized signal processing operations such as diversity and MIMO decoders when there are no such blocks in GNU Radio library.

Fig. 4.22 shows QPSK constellation diagrams in the presence of interference. It can be seen that the SDR channel 1 and SDR channel 2 constellation diagram degrades as the interference increases. The degradation can be observed in the increase of noise cloud variance. However, the QPSK constellation diagram can be improved by increasing the Tx gain of the communication link USRP (the noise cloud around ideal QPSK symbol decreases). Fig. 4.23 shows QPSK SDR channel 1 and SDR channel 2 when the Tx gain value is increased from 8 dB to 10 dB. The difference is clear noise cloud variance has reduced because of the increased Tx

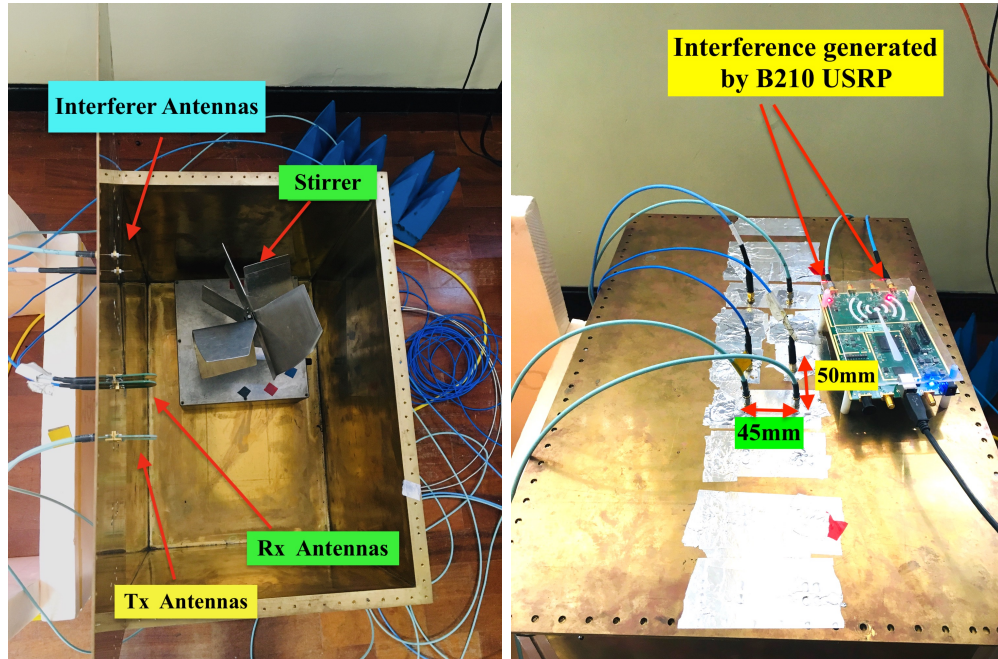


Figure 4.19: 2x2 MIMO BER and EVM measurement setup in presence of MIMO interference. The measurements are conducted in an empty enclosure. Interference signal is transmitted using two monopoles located closer to Rx antennas and the mode-stirrer.

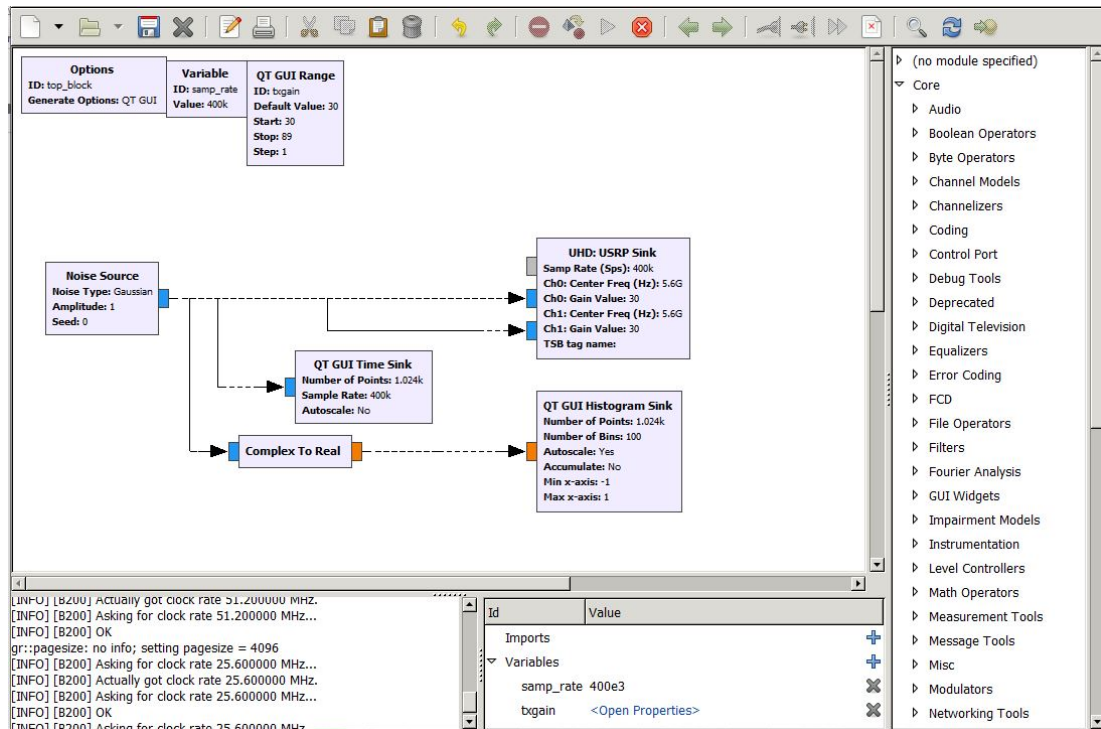


Figure 4.20: GNU Radio flowgraph for white Gaussian noise. GNU radio has searchable library of signal processing blocks. Python code can be generated from the flowgraph and more operations can be done in generated python file.

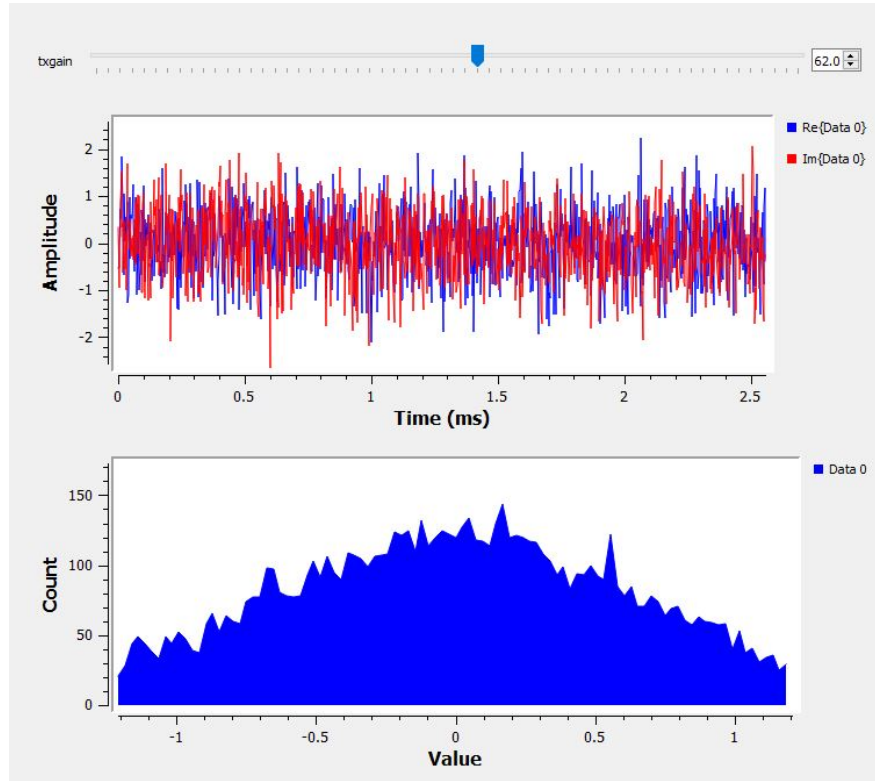


Figure 4.21: IQ data and histogram of generated white Gaussian noise. The magnitude and Tx gain of the USRP can be changed using QT GUI range block.

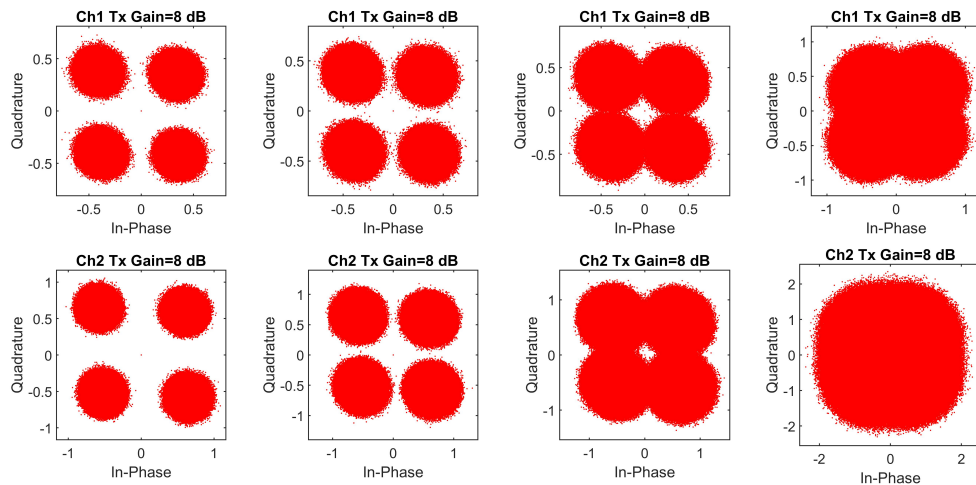


Figure 4.22: Channel 1 and channel 2 QPSK constellation diagram at 5.6 GHz in the metal enclosure in presence of MIMO interference. AGC gain=60 dB, Rx Gain=0 dB and Tx Gain= 8 dB.

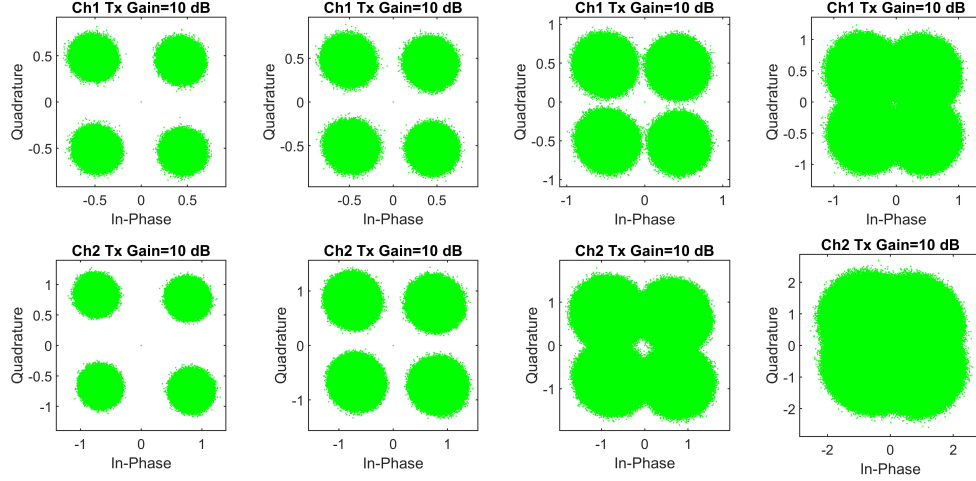


Figure 4.23: Channel 1 and channel 2 QPSK constellation diagram at 5.6 GHz in the metal enclosure in presence of MIMO interference. AGC gain=60 dB, Rx Gain=0 dB and Tx Gain= 10 dB.

gain, however it can also be seen that the two SDR channel performance degrades when the interference power is increased. For higher interference power values the state of QPSK constellation diagram is same. This implies that the BER and EVM performance can not be improved by only spending on communication link power. For high interference power there is clear evidence of BER floor beyond which performance can not be improved.

Fig. 4.24 and Fig. 4.25 shows measured EVM of SDR channel 1 and channel 2 for the gain values interfering B210 USRP are increased from 60 dB to 75 dB. The EVM of two SDR channels is constant except the jump in EVM values when interfering USRP gain values are increased. The constant EVM shows that the measurements are conducted in an stationary environment. It's clear that EVM degrades as interference signal level increases from 60 dB to 75 dB. However, from these figures it can be observed that there is EVM gain as the Tx gain of the communication USRP increases from 8 dB to 10 dB. Similarly, the BER of two SDR channels is measured as a function of interferer Tx gain values as shown in Fig. 4.26. The measured BER is as low as 10^{-9} at SDR channel 1 for Tx gain values of 10 dB. The BER at same SDR channel degrades to 1 when the interfering USRP gain value is set to 75 dB. The best BER values for SDR channel 1 at 8 dB and 10 dB are 10^{-9} and 10^{-3} which degrades as the gain of interfering B210

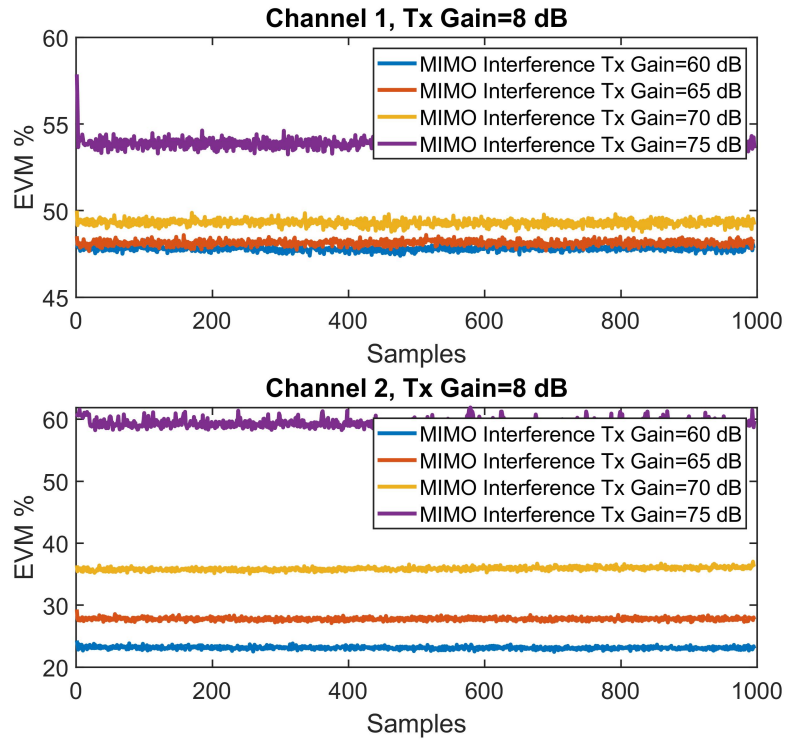


Figure 4.24: Measured EVM of channel 1 and channel 2 in presence of different levels of MIMO interference. Tx gain of the USRP is 8 dB.

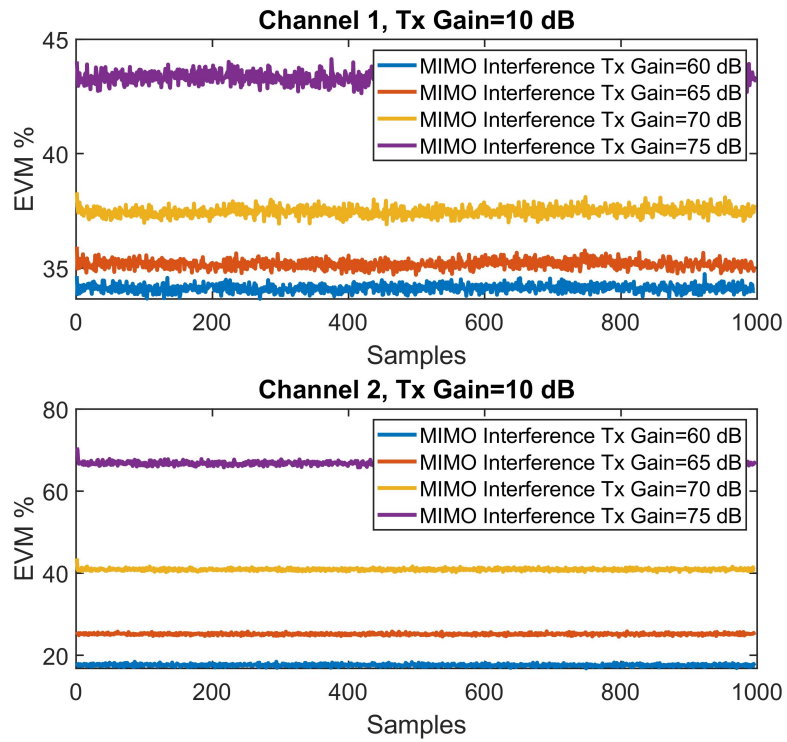


Figure 4.25: Measured EVM of channel 1 and channel 2 in presence of different levels of MIMO interference. Tx gain of the USRP is 10 dB.

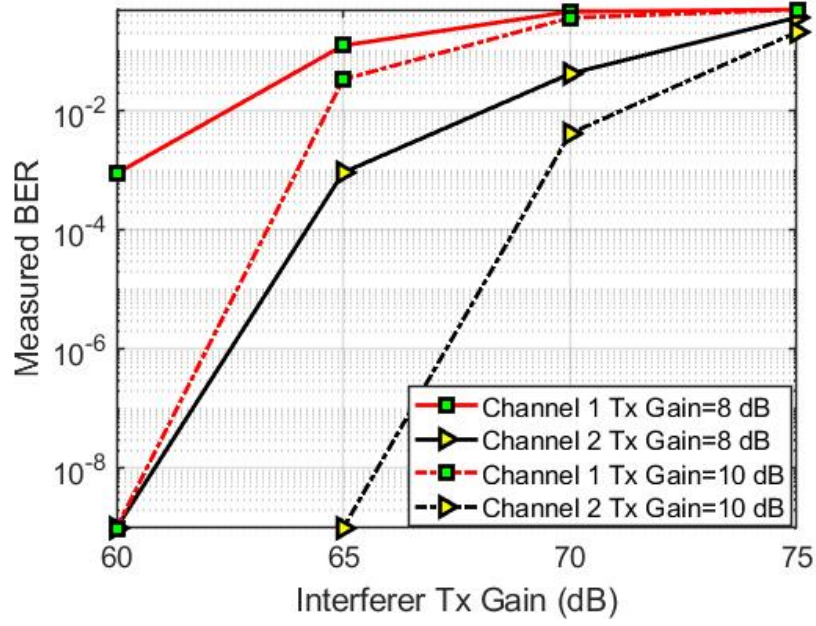


Figure 4.26: Measured BER of SDR channel 1 and SDR channel 2 in presence of different MIMO interference levels.

USRP increases. The worst case BER are observed when at interference gain values of 75 dB. The text information at the main communication link becomes undecodable at the interference gain values of 75 dB.

4.7 BER Measurement in Empty and Loaded Metal Enclosure in Presence of MIMO Interference

In previous section BER and EVM were measured in presence of interference in an empty metal enclosure. Here, the BER performance is measured in an empty enclosure and when the enclosure is loaded with the losses. Here BER performance of near-field MIMO communication link is measured as a function of transmit gain, enclosure losses and interference level. The measurement setup two channel BER and EVM is shown in Fig. 4.27. The mode-stirrer has been removed to avoid the effect of clustering when the measurements are done in stationary channel conditions. The measurements are performed in both empty enclosure and when the enclosure is loaded with RF absorbers.

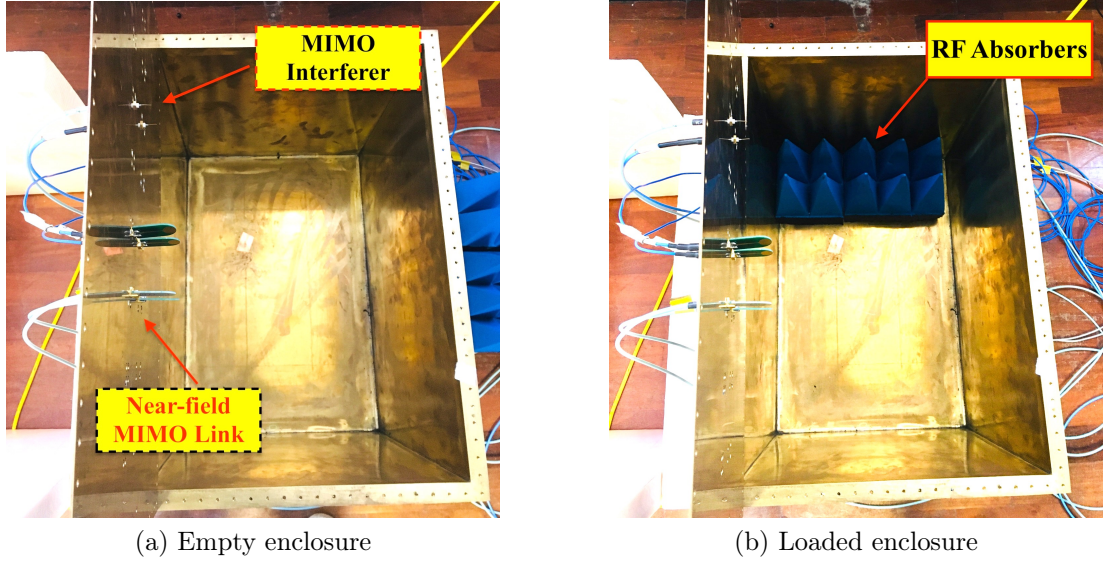


Figure 4.27: 2x2 MIMO BER measurement setup in presence of MIMO interference when metal enclosure is empty and loaded with ten piece of absorbers.

Fig. 4.28 shows measured BER of channel 1 and channel 2 in an empty metal enclosure at different interference gain values. The two SDR channel values were extracted by configuring the QPSK receiver in sweep mode. This was useful to avoid data management problems that would have resulted when changing the gain values manually for each BER values. The BER values for channel 1 at interference gain of 55 dB and the Tx gain of USRP of 6 dB is less than 10^{-5} and there are no errors for Tx gain of 8 dB and 10 dB respectively. There are only three data points and BER plots follow standard waterfall curves. It can be seen that there are higher BER values at lower Tx gain values as the interference is dominant at these values for all MIMO interference values. As the interference increases the BER curves remain flat (there are only errors) for all the Tx gain values. Fig. 4.29 shows measured BER of channel 1 and channel 2 when the metal enclosure was loaded with ten pieces of absorber cones. It can be seen that the measured BER degrades with the increase in MIMO interference power. However, it requires higher MIMO interference power to get the same BER value of channel 1 and channel 2. After the addition of 10 RF absorber cones, the propagation becomes favourable for main MIMO communication link as seen from the BER values (see Fig. 4.29). It has given additional communication link

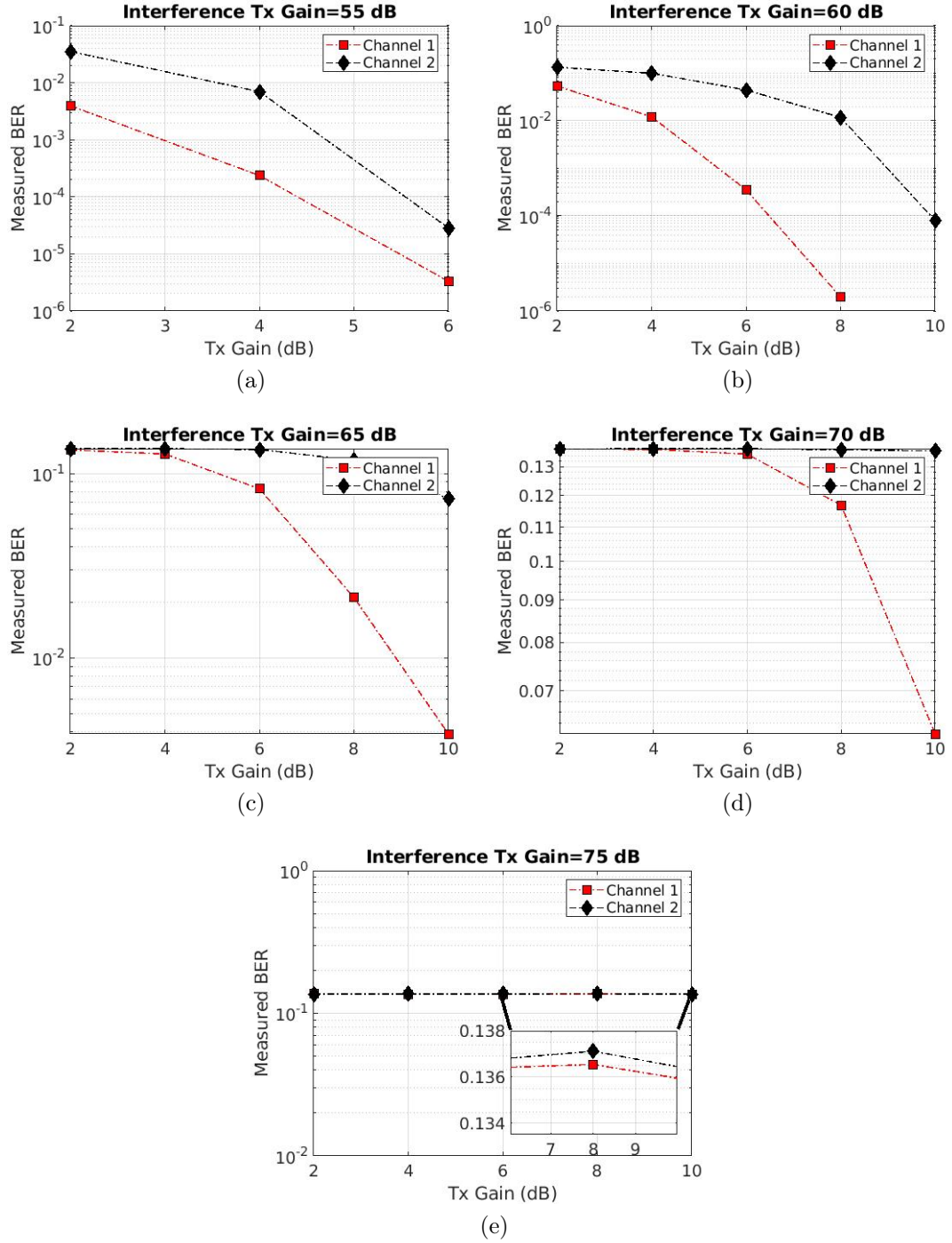


Figure 4.28: Measured MIMO BER of channel 1 and channel 2 in presence of MIMO interference when the metal enclosure was empty.

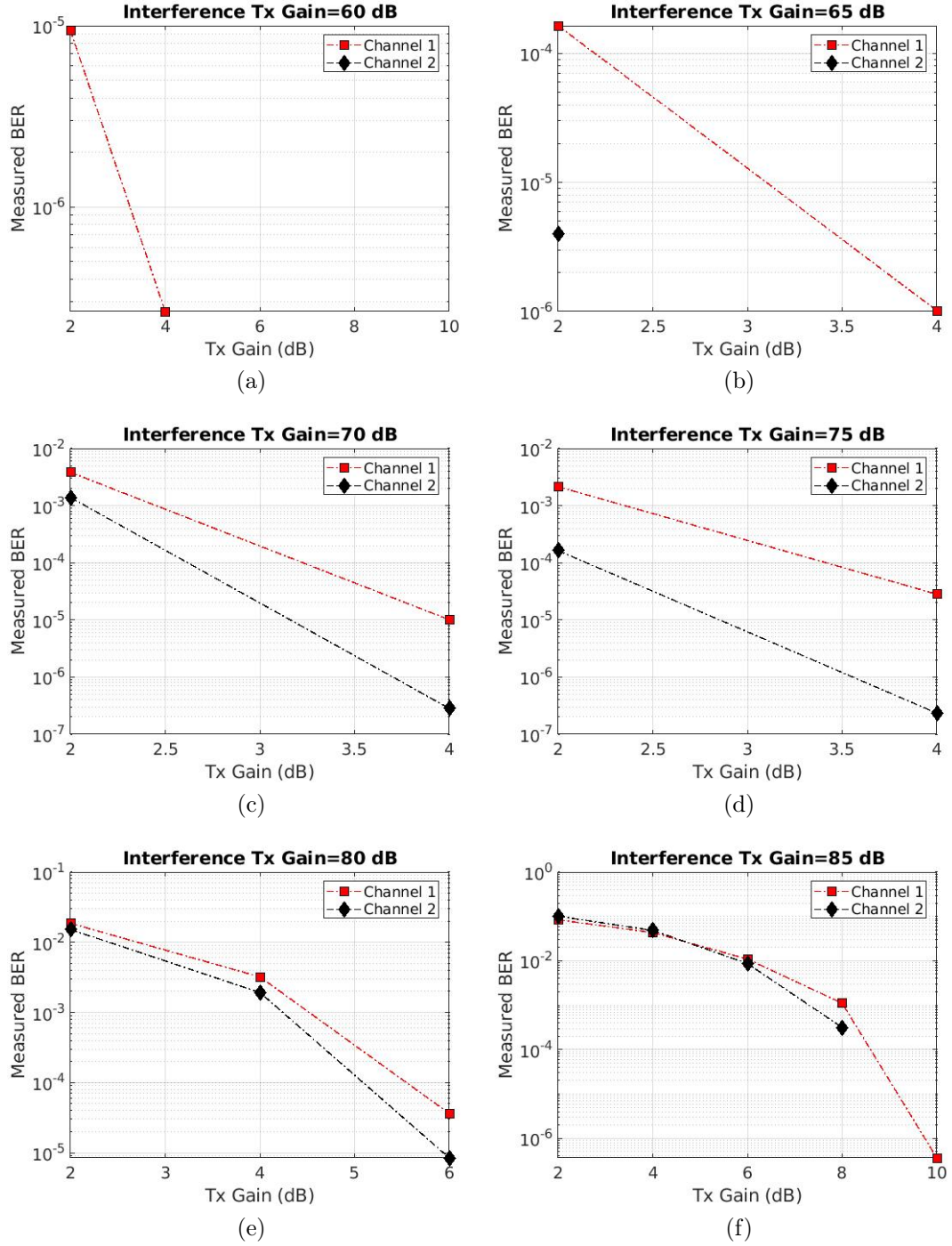
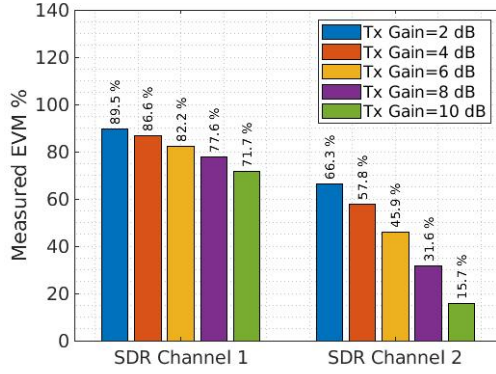


Figure 4.29: Measured MIMO BER of channel 1 and channel 2 in presence of MIMO interference when the metal enclosure was loaded with ten pieces of RF absorber cones.

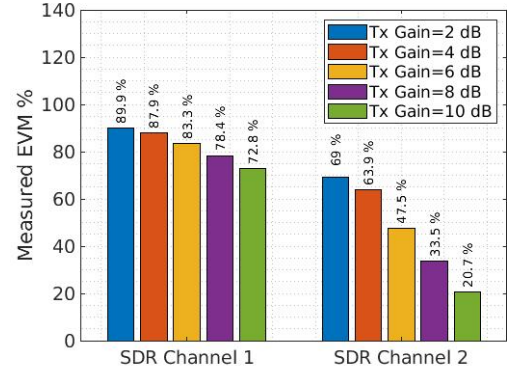
gain from 15 dB to 20 dB. The measured BER values at 60 dB of interference gain are better than that at 55 dB. The one reason is the addition of RF absorbers increases the coherence bandwidth of the channel and another potential reason is the significant attenuation of interfering signals as the interfering antennas are located closer to RF absorbers. Therefore, there is not enough reflected energy that would contribute to the disruption of MIMO communication link. For 60 dB of interference gain channel 2 is error-free and channel 1 has BER of roughly 10^{-7} at 4 dB of Tx gain and no errors at high Tx gains. The two SDR channel BER starts to degrade as the interference gain increases. For interference gain of 85 dB value, the channel 2 is still error free at 10 dB of Tx gain whereas channel 1 BER is around 10^{-6} . Therefore, even more dBs of interference Tx gain may be required to break the MIMO communication link.

Fig.4.30 shows EVM measurement of SDR channel 1 and SDR channel 2. It can be seen that with no interference the SDR channel 1 and SDR channel 2 EVM decreases. However, as the magnitude of interference increases the EVM of channel 1 and channel 2 decreases. Channel 2 EVM has degraded instead of improving at high interference B210 USRP gain of 75 dB. The prime reason is because channel 2 had already higher gain than channel 1. In addition to increase in magnitude of channel 2, there was increase in received noise interference which has caused EVM of channel 2 to degrade further. This is also known as error floor in terms of BER measurements that's BER doesn't improve in presence of interfering noise irrespective of increase in Tx power of the communication link.

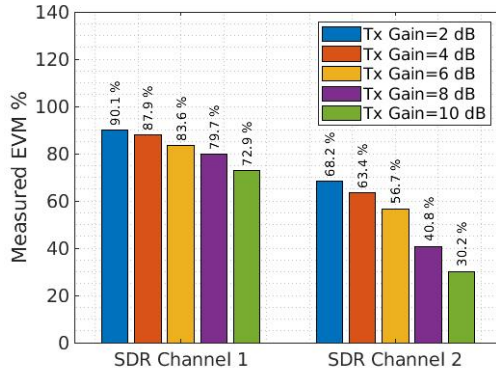
Fig. 4.31 shows EVM measurement of SDR channel 1 and SDR channel 2 when the empty chamber was loaded with ten pieces of absorbers and the AGC gain remained at 50. It can be seen that both SDR channels have comparatively stable EVM. This is because of absorption effect of the absorbers and AGC gain. In the next experiment, effect of change in AGC gain is demonstrated on EVM measurements of two SDR channels. EVM measurements are recorded by keeping the same environment and USRP configuration parameters and only changing the



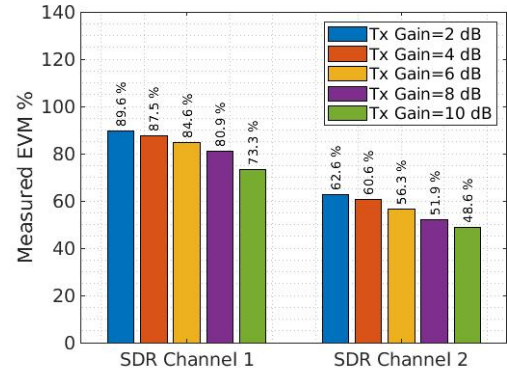
(a) No Interference



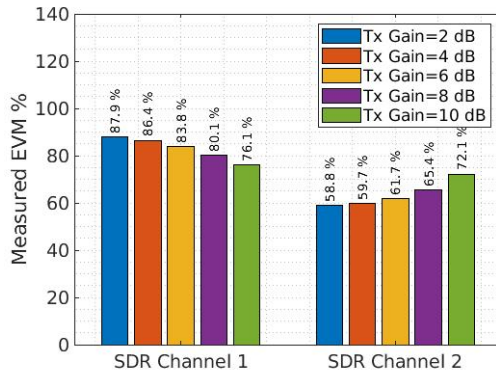
(b) Interference Gain=55 dB



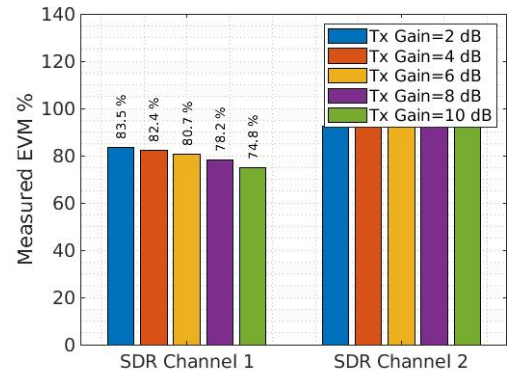
(c) interference Gain=60 dB



(d) Interference Gain=65 dB

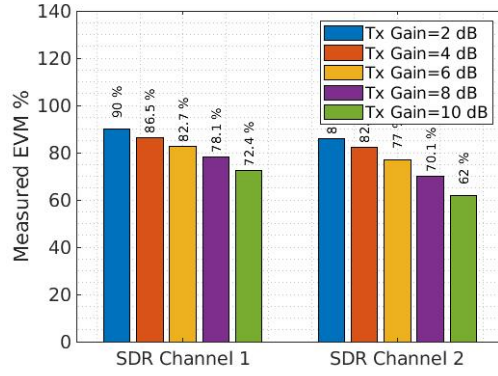


(e) Interference Gain=70 dB

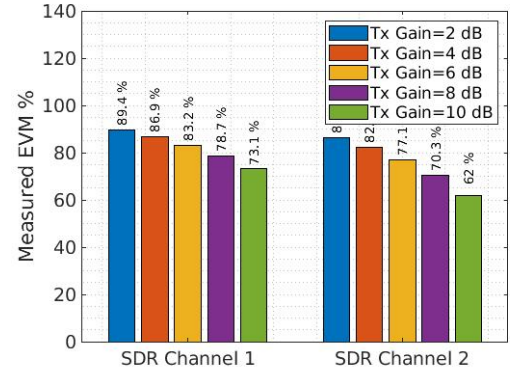


(f) Interference Gain=75 dB

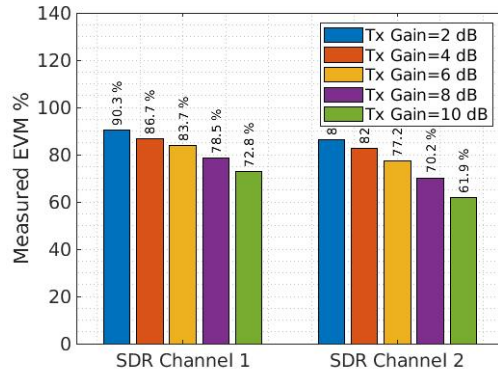
Figure 4.30: Measured MIMO EVM of SDR channel 1 and SDR channel 2 in presence of MIMO interference when the metal enclosure was empty. The Tx gain was swept from 2 dB to 10 dB. AGC gain was set to 50 dB.



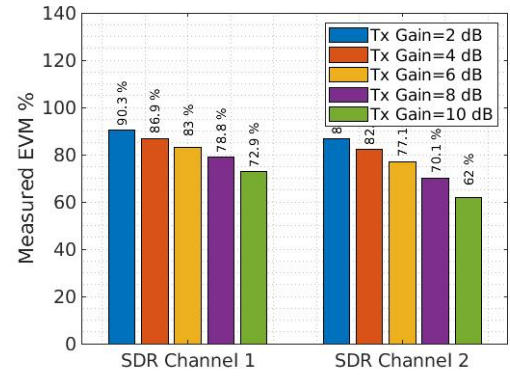
(a) No Interference



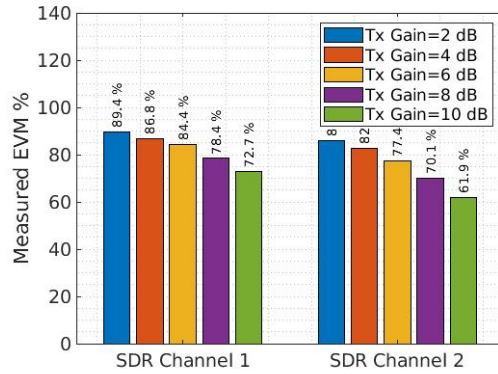
(b) Interference Gain=55 dB



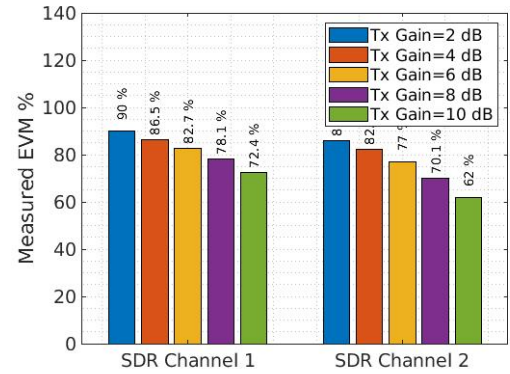
(c) Interference Gain=60 dB



(d) Interference Gain=65 dB



(e) Interference Gain=70 dB



(f) Interference Gain=75 dB

Figure 4.31: Measured MIMO EVM of SDR channel 1 and SDR channel 2 in presence of MIMO interference when the metal enclosure was loaded with ten pieces of RF absorber cones. The Tx gain was swept from 2 dB to 10 dB. AGC gain was set to 50 dB.

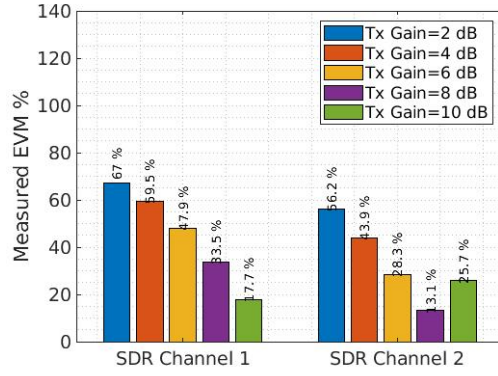
AGC gain from 50 dB to 60 dB.

Fig. 4.32 shows EVM measurement of SDR channel 1 and SDR channel 2 when the metal enclosure was loaded with 10 pieces of RF absorber cones. It can be seen that the EVM has improved because the RF absorbers have increased the coherence bandwidth and also reduce the effect of interference.

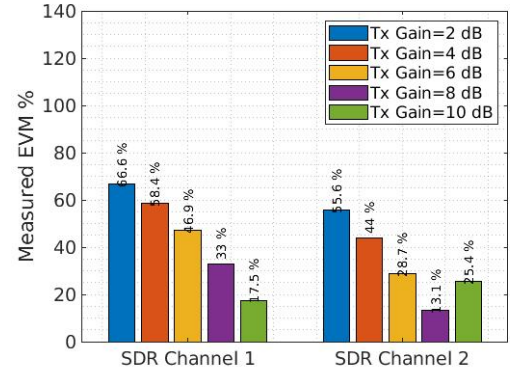
Fig. 4.33 shows plot of measured EVM of two SDR channels. It can be seen that the EVM of SDR channel 2 starts to degrade as the Tx gain increases. This is because of received samples shifting away from ideal sampling point on the reference constellation diagram. For comparison QPSK constellation diagrams show this difference clearly. At Tx gain of 8 dB, the received samples are around reference of ideal constellation point which is shown in red star, however it not true when Tx gain is 10 dB. The gain of the received samples can be dynamically adjusted to keep the received constellation around ideal sampling points. The dynamic adjustment of the Tx gain of the main communication link can done via feedback control channel which asks the transmitted to transmit the symbols at low-power if the path loss of the medium is small.

4.8 2x2 MIMO BER and EVM Measurements in NLOS

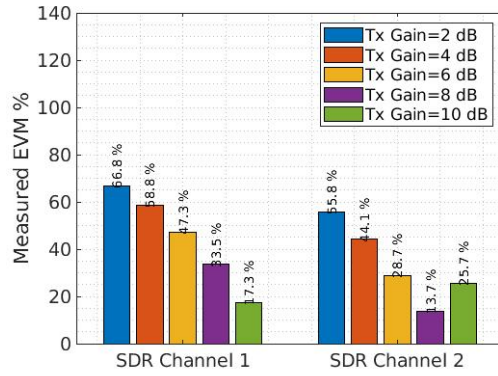
In this section, 2x2 MIMO BER and EVM measurements in NLOS are presented. MIMO communication links may exists in both LOS and NLOS in stationary and dynamic channel conditions. MIMO capacity is increased in NLOS situation when there is rich scattering because the channels become uncorrelated. In contrast, the capacity of LOS channel may increase because of the presence of dominant link and hence increase in SNR. Fig. 4.34 shows MIMO EVM measurement setup in NLOS when the metal enclosure is empty and when the metal enclosure is loaded with 10 RF absorbing cones. NLOS sheet is introduced between transmitting and receiving antennas which are $D = 225\text{ mm}$ apart. The transmitting and



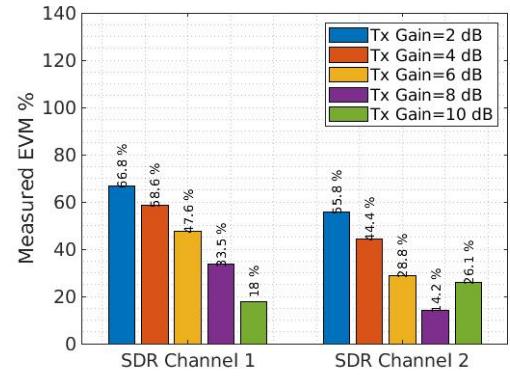
(a) No Interference



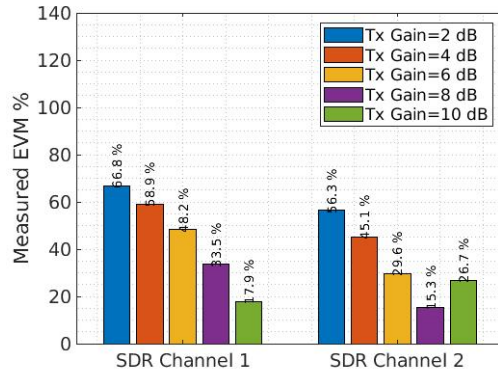
(b) Interference Gain=55 dB



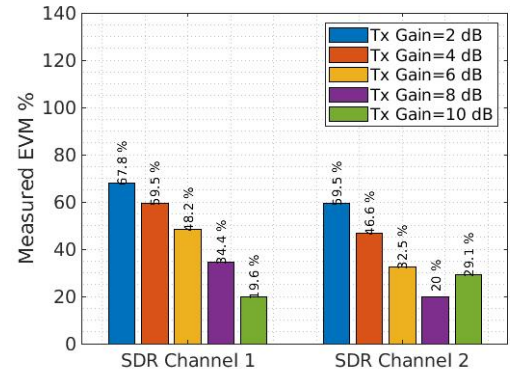
(c) Interference Gain=60 dB



(d) Interference Gain=65 dB



(e) Interference Gain=70 dB



(f) Interference Gain=75 dB

Figure 4.32: Measured MIMO EVM of SDR channel 1 and SDR channel 2 in presence of MIMO interference when the metal enclosure was loaded with ten piece of RF absorber cones. The Tx gain was swept from 2 dB to 10 dB. AGC gain was set to 60 dB.

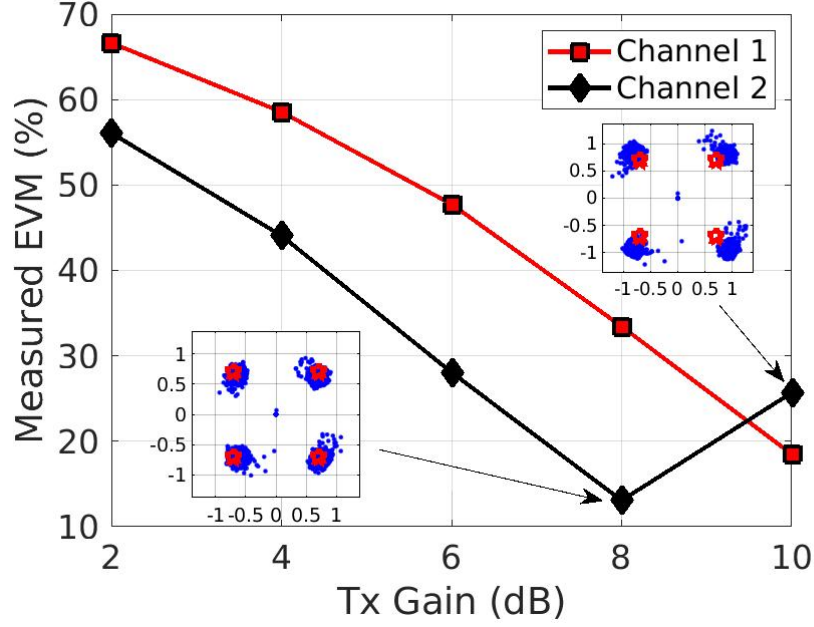


Figure 4.33: Measured MIMO EVM of SDR channel 1 and SDR channel 2 in presence of MIMO interference when the metal enclosure was loaded with ten pieces of RF absorber cones. The Tx gain was swept from 2 dB to 10 dB. AGC gain was set to 60 dB.

receiving antennas are vertically monopole antennas resonating at 5.6 GHz. The distance between transmit antenna elements is $d_t = 70\text{ mm}$, distance between receive antenna elements is $d_r = 70\text{ mm}$, the distance between NLOS sheet and the transmitter is $D_t = 115\text{ mm}$, the distance between receive antennas and the NLOS sheet is $D_r = 110\text{ mm}$. It can be seen that Tx monopole antennas are located closer to stirrer and Rx monopoles are located closer to RF absorber cones. The Tx monopole and Rx monopole antennas are separated by thick metal sheet creating NLOS scenario and the partial propagation through sheet is negligible. Hence, received signals are only through NLOS scattering. Tx and Rx antennas could be switched such that the Tx antennas are more closely located to RF absorbers. In this scenario more power is spent to meet the same performance criteria because Tx waves will interact with the highly attenuating factors in the first instant one with NLOS sheet and the other with RF absorbers. Additionally, it depends on stirrer pedal positions and it may or may not support effective forward propagation.

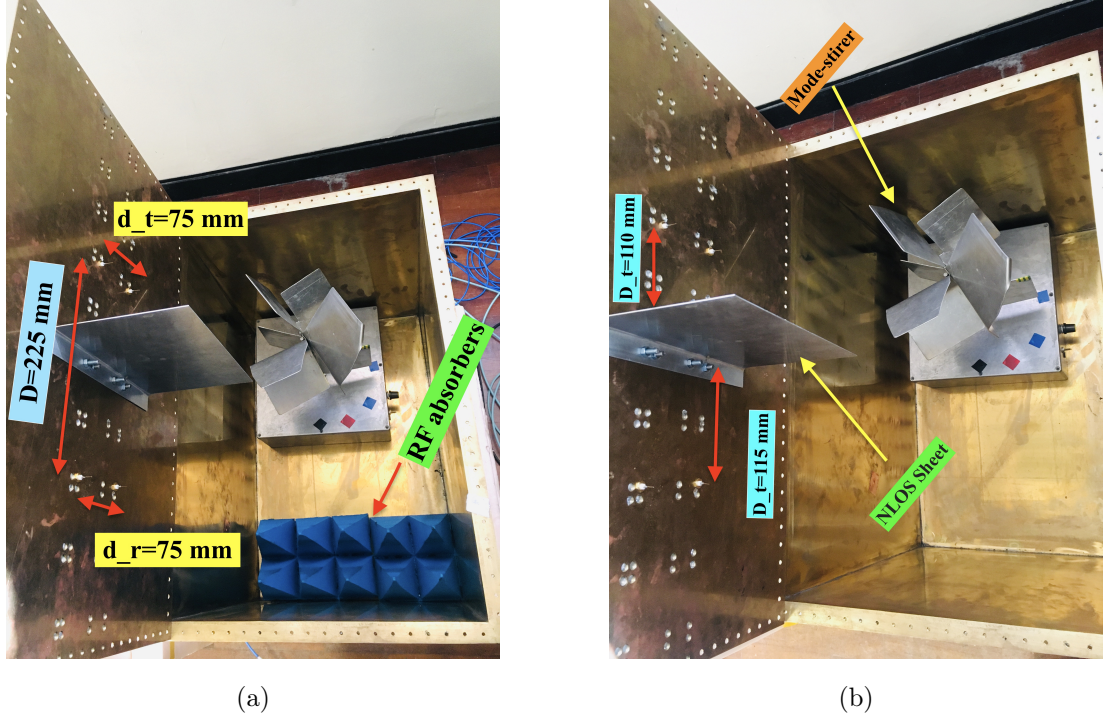
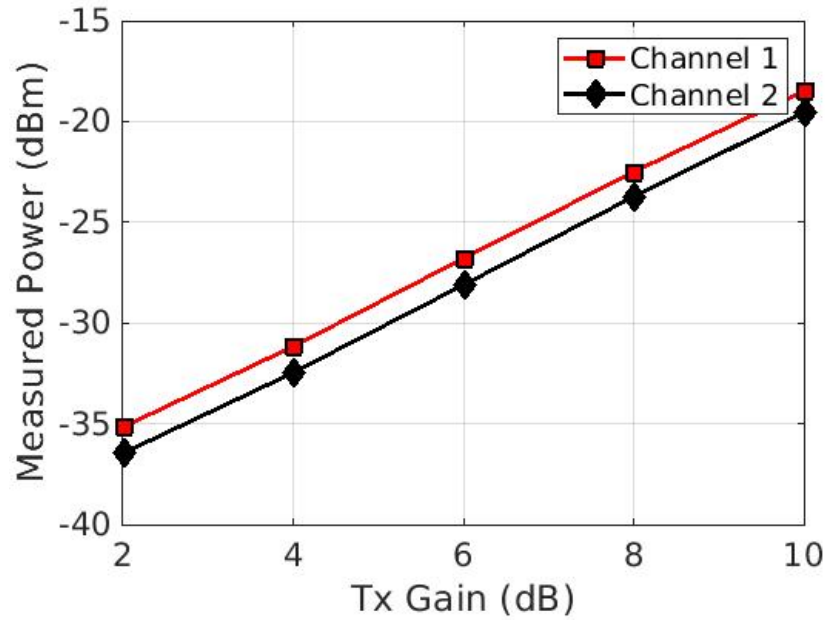


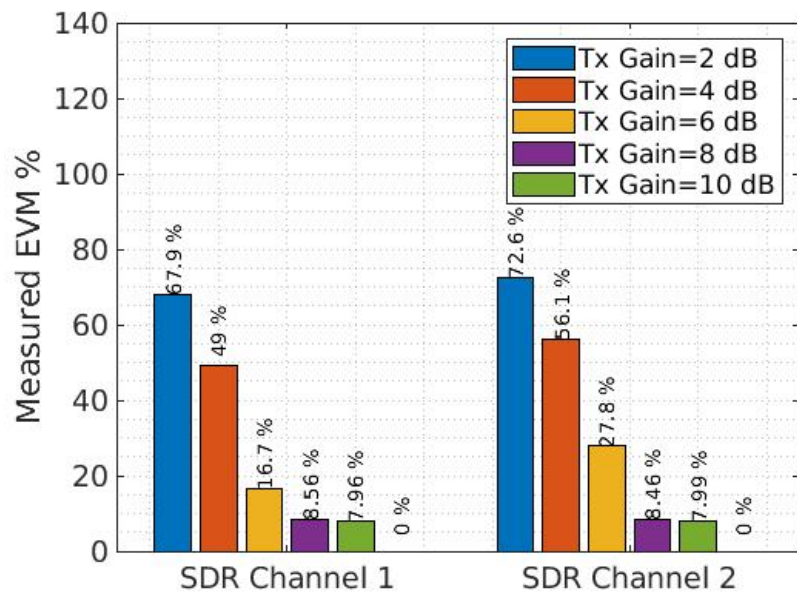
Figure 4.34: MIMO EVM measurement setup in metal enclosure (a) NLOS measurement setup with absorbers (b) NLOS measurement setup without absorbers.

Fig. 4.35 shows measured EVM and received power of two USRP channels in the presence of 10 absorbing cones. It can be seen in Fig. 4.36a that the power increases linearly with the increase in gains of USRP channel 1 and USRP channel 2. This is a firm confirmation of proper measurement setup operation, a linear increase in Tx USRP gains results in linear increase in average receive power. It is important to operated USRPs in the linearity range. Similarly, Fig. 4.35b shows QPSK EVM of both the USRP channels and the QPSK EVM increases with the increase in USRP channels.

Fig. 4.36 shows measured EVM and received power of two USRP channels in an empty chamber. QPSK EVM for both the channels also decreases with the increase in TX power when there are no RF absorbers in the metal enclosure. The received power all gain values is slightly higher in empty enclosure than loaded metal enclosure. Additionally, the EVM in both the cases is decreases and the EVM is higher in case of loaded metal enclosure. The EVM of USRP channel 1 decreases gradually for first Tx gain values and it satys constant for rest of the Tx

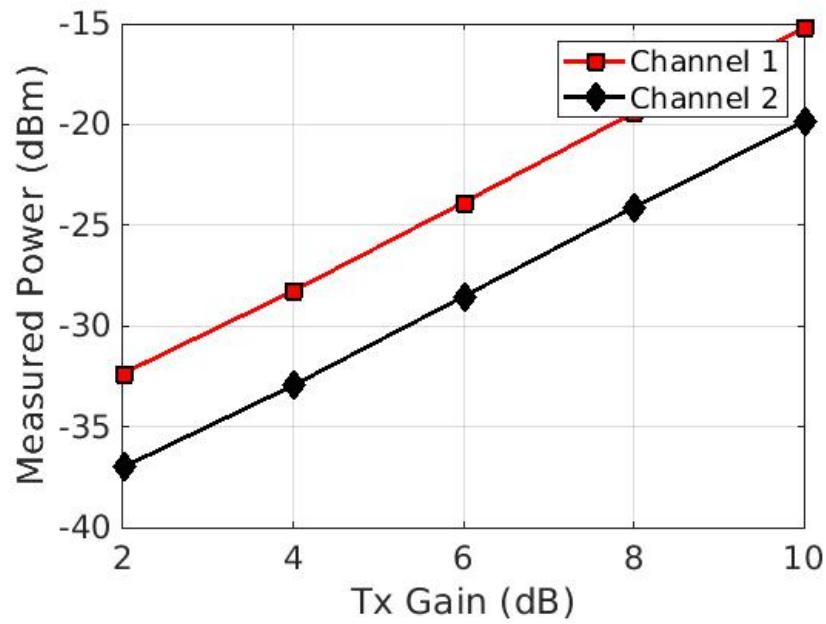


(a) Received power of two USRP channels

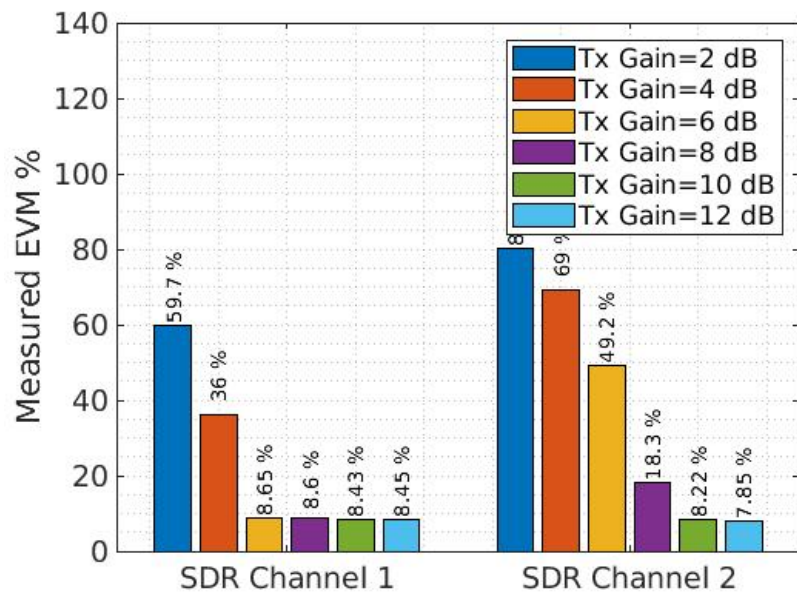


(b) EVM of two USRP channels at different Tx gains

Figure 4.35: 2x2 MIMO NLOS received power measurement and EVM in the presence of 10 RF absorber cones. Received power increases linearly as the Tx gain increases. However, the EVM of two SDR channels decreases with the increase in Tx gain.



(a) Received power of two USRP channels



(b) EVM of two USRP channels at different Tx gains

Figure 4.36: NLOS 2x2 MIMO received power measurement and EVM of the two USRP channels in an empty enclosure. Received power increases linearly as the Tx gain increases. However, the EVM of two SDR channels decreases with the increase in Tx gain.

gain values. For first two gain values the EVN of USRP channel 1 has decreased from 67 % and 49 % to 59 % and 36 % respectively. Hence, for the first two Tx gain values of the USRP channel 1, the EVM decreased by 11.9 % and 26.53 %.

4.9 Conclusion

This chapter is about the design stages of SDR based QPSK digital receiver and the relevant spatial MIMO near-field BER and EVM measurement in the metal enclosure. Each stage of the digital receiver is explained with the direct implication of sub-stages of QPSK receiver on BER and EVM measurements of two USRP channels. BER and EVM measurements are performed as function of transmit and receive antennas distances and as function of transmit antenna gains. BER and EVM measurements improve with the increase in transmit antenna gains. MIMO BER and EVM measurements are performed in presence of MIMO interference. MIMO BER and EVM performance degrades as the Tx gain of MIMO interference increases. It's imperative to understand underflow and overflow while the measurements are in progress. Underflow causes data loss which affects specially BER measurements. MIMO BER and EVM measurements are also carried out in an empty metal enclosure and when it's loaded with carbon absorbers. EVM of two USRP channels have variations when the mode-stirrer is turned on which is obvious because the channel is not stationary and the EVM is now function of stirrer pedal potions.

Bibliography

- [1] Paolo Corona, John Ladbury, and Gaetano Latmiral. Reverberation-chamber research-then and now: a review of early work and comparison with current understanding. *IEEE transactions on Electromagnetic Compatibility*, 44(1):87–94, 2002.
- [2] Vignesh Rajamani, Charles F Bunting, and James C West. Stirred-mode operation of reverberation chambers for emc testing. *IEEE Transactions on Instrumentation and Measurement*, 61(10):2759–2764, 2012.
- [3] Kent Rosengren, Per-Simon Kildal, Charlie Carlsson, and Jan Carlsson. Characterization of antennas for mobile and wireless terminals in reverberation chambers: Improved accuracy by platform stirring. *Microwave and Optical Technology Letters*, 30(6):391–397, 2001.
- [4] Christian S Lötbäck Patané, Magnus Franzén, and Charlie Orlenius. Analysis and improvement of reverberation chamber method for characterization of small and terminal antennas. In *Proceedings of the 5th European Conference on Antennas and Propagation (EUCAP)*, pages 1832–1836. IEEE, 2011.
- [5] Christopher L Holloway, Haider A Shah, Ryan J Pirkel, William F Young, David A Hill, and John Ladbury. Reverberation chamber techniques for determining the radiation and total efficiency of antennas. *IEEE transactions on antennas and propagation*, 60(4):1758–1770, 2012.
- [6] Xiaoming Chen, Per-Simon Kildal, Charlie Orlenius, and Jan Carlsson. Channel sounding of loaded reverberation chamber for over-the-air testing of wireless devices: Coherence bandwidth versus average mode bandwidth and delay spread. *IEEE Antennas and Wireless Propagation Letters*, 8:678–681, 2009.

- [7] David A Hill, Mark T Ma, Arthur R Ondrejka, Bill F Riddle, Myron L Crawford, and Robert T Johnk. Aperture excitation of electrically large, lossy cavities. *IEEE transactions on Electromagnetic Compatibility*, 36(3):169–178, 1994.
- [8] Christopher L Holloway, David A Hill, John M Ladbury, Perry F Wilson, Galen Koepke, and Jason Coder. On the use of reverberation chambers to simulate a rician radio environment for the testing of wireless devices. *IEEE transactions on antennas and propagation*, 54(11):3167–3177, 2006.
- [9] Nader Sheikholeslami Alagha and Peter Kabal. Generalized raised-cosine filters. *IEEE transactions on Communications*, 47(7):989–997, 1999.
- [10] Xiang-Gen Xia. A family of pulse-shaping filters with isi-free matched and unmatched filter properties. *IEEE Transactions on Communications*, 45(10):1157–1158, 1997.
- [11] Yan Wang, Kai Shi, and Erchin Serpedin. Non-data-aided feedforward carrier frequency offset estimators for qam constellations: A nonlinear least-squares approach. *EURASIP Journal on Advances in Signal Processing*, 2004(13):1–9, 2004.
- [12] Alexander M Wyglinski, Robin Getz, Travis Collins, and Di Pu. *Software-Defined Radio for Engineers*. Artech House, 2018.

Chapter 5

Near-field MIMO Wireless Communication

5.1 MIMO System Model

The narrowband MIMO input-output relationship with N_T transmit antennas and N_R receive antennas can be given as:

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \mathbf{n} \quad (5.1)$$

Where $\mathbf{y} \in \mathbb{C}^{N_R \times 1}$ is complex received signal vector. $\mathbf{x} \in \mathbb{C}^{N_T \times 1}$ is complex transmitted signal vector. $\mathbf{H} \in \mathbb{C}^{N_R \times N_T}$ is complex MIMO channel matrix with $N_T \times N_R$ complex channel elements. Channel elements of $\mathbf{H}$ are IID Rayleigh.

The narrowband MIMO input-output relationship in eq.(5.1) at time instant k can be given as:

$$\mathbf{Y} = \mathbf{H}\mathbf{X} + \mathbf{N} \quad (5.2)$$

Where $\mathbf{Y} \in \mathbb{C}^{N_R \times k}$ is complex received signal vector. $\mathbf{X} \in \mathbb{C}^{N_T \times k}$ is complex transmitted signal vector. $\mathbf{H} \in \mathbb{C}^{N_R \times N_T \times k}$ is complex MIMO channel matrix with $N_R \times N_T$ complex channel elements at time k . 2×2 MIMO channel configuration is shown in Fig. 5.1a where the complex channel gains between each transmit and receive antenna form MIMO channel matrix $\mathbf{H}$. Similarly, Fig. 5.1b shows MIMO configuration with MIMO channel elements where the Tx antenna elements are separated by distance d_t and Rx antenna elements are separated by distance

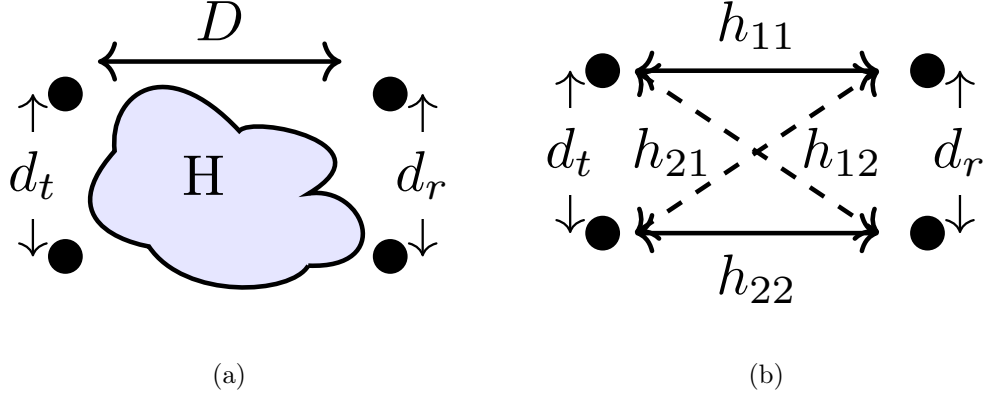


Figure 5.1: MIMO Configuration (a) MIMO channel matrix $\mathbf{H}$ (b) MIMO channel elements.

d_r . Transmitter and receiver are separated by a distance D . Characteristics of MIMO system depends upon type of propagation environment, transmit and receive antenna correlation and whether MIMO communication takes place in LOS or NLOS environment.

5.2 MIMO Channel Capacity with Unknown CSI at Transmitter

MIMO channel capacity with N_T transmit and N_R receive antennas and with no channel state information at the transmitter and CSI only available at the receiver and under the transmit power constraints is represented by eq. (5.3). When there is no CSI at the transmitter available the optimal power allocation strategy is to divide transmit power equally among all transmit antennas hence known as equal power allocation strategy. When CSI is also available at the transmitter water-filling algorithm proposed by Gallagar[1] is used. Water-filling algorithm is known to perform at low SNR. i.e. if there is low SNR then it's better to allocate higher power to stronger channel.

$$C = \log_2 \left(\det \left(\mathbf{I}_{N_R} + \frac{P_T}{N_T \sigma_n^2} \mathbf{H} \mathbf{H}^H \right) \right) \quad \text{bits/s/Hz} \quad (5.3)$$

Here $\rho = \frac{P_T}{\sigma_n^2}$ represents SNR, $\mathbf{I}_{N_R}$ is identity matrix of size $N_R \times N_R$.

5.2.1 Ergodic Capacity

MIMO channel capacity is a function of channel state information which is a random quantity and the mean or average of MIMO channel capacity is known as Ergodic Capacity.

$$C = E \left[\log_2 \det \left(\mathbf{I}_{N_R} + \frac{P_T}{N_T \sigma_n^2} \mathbf{H} \mathbf{H}^H \right) \right] \quad \text{bits/s/Hz} \quad (5.4)$$

Where $\rho = \frac{P_T}{\sigma_n^2}$ is SNR.

5.2.2 Wishart Matrix

Wishart matrix is multiplication of channel matrix $\mathbf{H}$ and Hermitian of complex channel matrix $\mathbf{H}^H$. From MIMO channel capacity:

$$C = \log_2 \left(\det \left(\mathbf{I}_{N_R} + \frac{P_T}{N_T \sigma_n^2} \mathbf{H} \mathbf{H}^H \right) \right) \quad \text{bits/s/Hz} \quad (5.5)$$

The Wishart matrix $\mathbf{W}$ is given by:

$$\mathbf{W} = \begin{cases} \mathbf{H}^H \mathbf{H} & N_R > N_T \\ \mathbf{H} \mathbf{H}^H & N_R \leq N_T \end{cases} \quad (5.6)$$

if $\lambda = [\lambda_1, \lambda_2, \dots, N_{max}]^T$ denote non-zero eigenvalues of $N_{min} \times N_{min}$ Wishart matrix, then the capacity can be written as:

$$C = \sum_{i=1}^{N_{min}} \log_2 \left(\mathbf{I}_{N_R} + \frac{P_T}{N_T \sigma_n^2} \lambda_i \right) \quad \text{bits/s/Hz} \quad (5.7)$$

Where $N_{min} = \min(N_T, N_R)$. From eq. (5.7) it can be seen that the MIMO capacity increases linearly as a function of the number of eigenvalues. This means

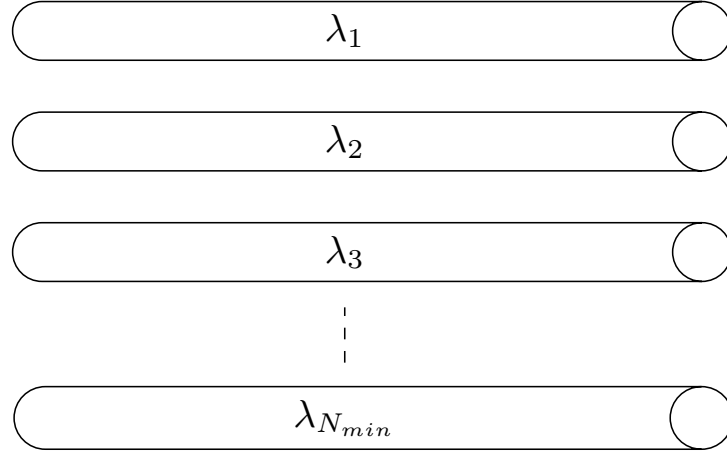


Figure 5.2: MIMO channel $\mathbf{H}$ eigenvalue decomposition. There are N_{min} parallel channels of MIMO channel matrix. Each pipe is represented by its eigenvalue λ_1 to $\lambda_{N_{min}}$.

in MIMO communication spectral efficiency linearly increases with the minimum of number of transmit and receive antennas. Ergodic Capacity in terms of eigenvalues λ_i can be written as:

$$E[C] = E \left[\sum_{i=1}^{N_{min}} \log_2 \left(\mathbf{I}_{N_R} + \frac{P_T}{N_T \sigma_n^2} \lambda_i \right) \right] \quad \text{bits/s/Hz} \quad (5.8)$$

Fig.5.2 shows eigenvalues of MIMO channel matrix which corresponds to N_{min} parallel pipes.

Fig.5.3 shows instantaneous IID capacity of 2x2 MIMO and 4x4 MIMO using equal power allocation strategy. It can be seen that the instantaneous IID capacity of 2x2 MIMO and 4x4 MIMO increases with the increase in SNR. It can also be seen that the MIMO capacity increases with MIMO antenna order. Fig.5.4 shows capacity of doubly-correlated 2x2 MIMO with equal power allocation strategy. It can be seen that the MIMO capacity increases with the increase in SNR and the capacity degrades when the transmit and receive correlation increases. Fig.5.5 shows 2x2 MIMO capacity when CSI is known at the transmitter and the correlation is present at both transmitter and receiver side. When the CSI is known

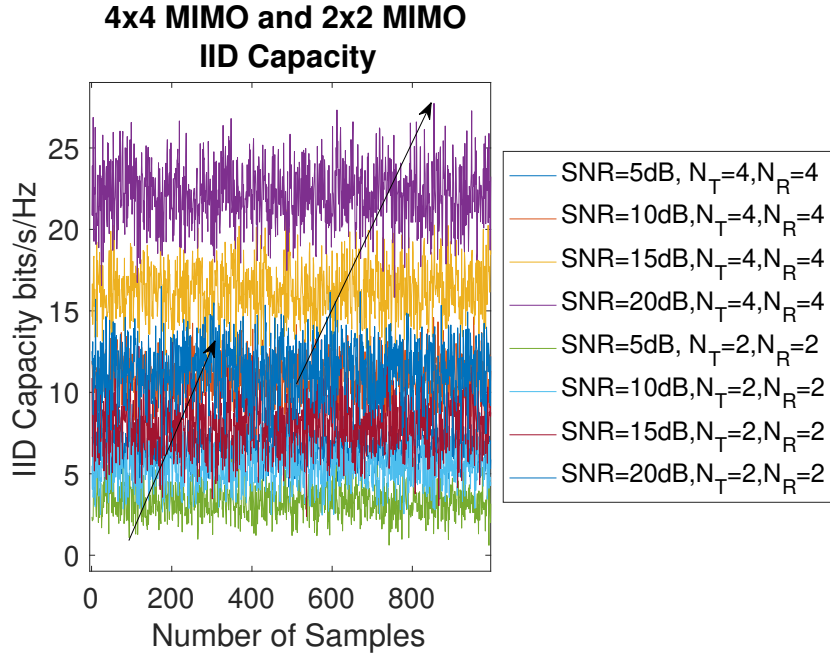


Figure 5.3: Instantaneous 4x4 MIMO and 2x2 MIMO IID capacity at different SNRs when CSI is not known at the transmitter and best power allocation is equal power allocation strategy.

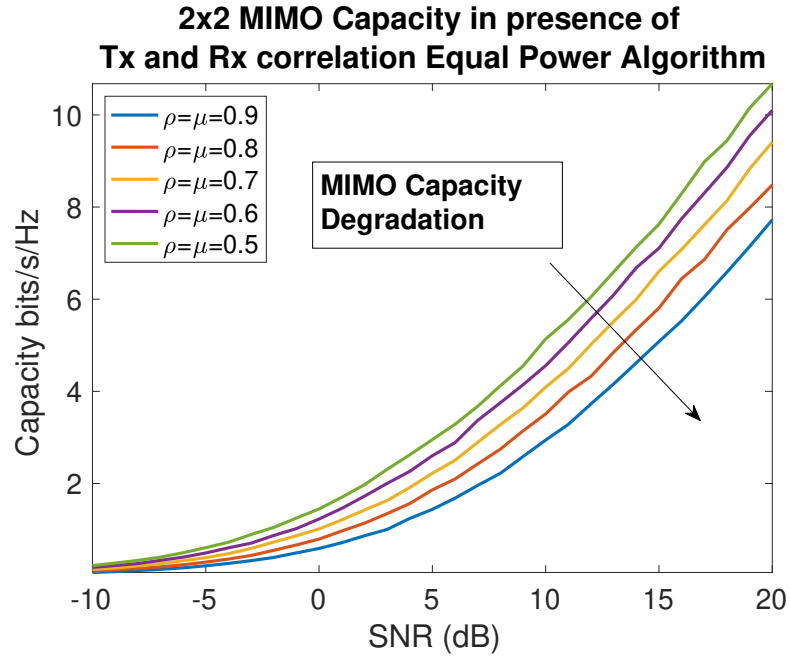


Figure 5.4: 2x2 MIMO capacity as a function of SNR when CSI is only available at the receiver only and transmitter remains uninformed. When CSI is not known at transmitter then best power allocation strategy is equal power allocation strategy.

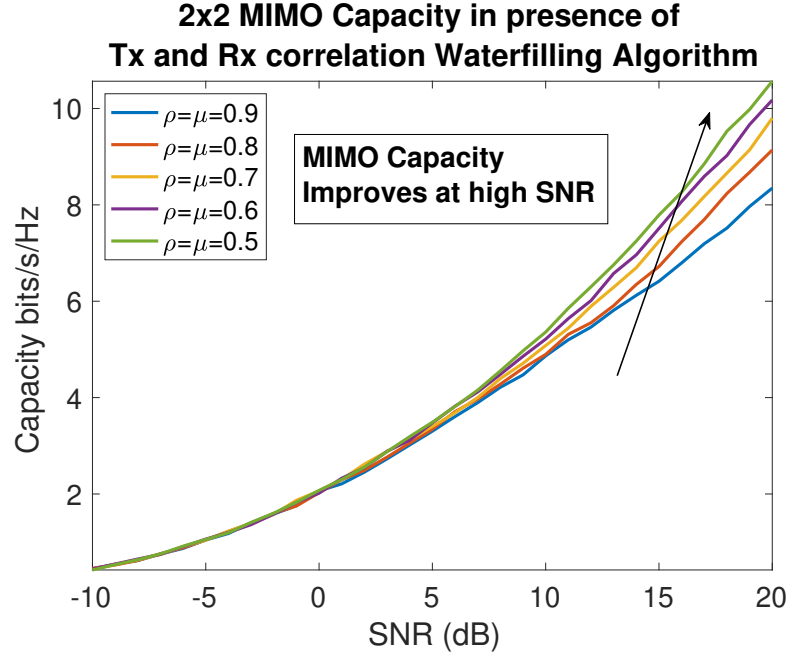


Figure 5.5: 2x2 MIMO capacity as function of SNR in presence of transmit and receive correlation. Best strategy to allocate transmit power when CSI is also known at transmitter side is by using Water-filling algorithm.

at the transmitter then optimal power allocation strategy is by water-filling algorithm where stronger channel with high SNR is assigned more power. MIMO capacity increases with increase in SNR, however for different correlation values the performance of water-filling algorithm increases. Fig.5.6 shows 2x2 MIMO capacity using equal power allocation and water-filling algorithm in presence of transmit and receive correlation of $\rho = \mu = 0.5$. It's shown that there is capacity loss when there is transmit and receive correlation. IID capacity is higher than non-IID capacity i.e. correlated capacity in both scenarios when CSI is unknown and known at the transmitter side under equal power allocation and water-filling algorithm strategy for transmit power distribution.

5.3 MIMO System Correlation Matrix

MIMO channel capacity depends on MIMO channel matrix $\mathbf{H}$ and received SNR ρ . MIMO channel capacity is also affected by correlation at the transmitter side and at the receiver side. It also gets affected by the AoA, AoD, angular spectrum

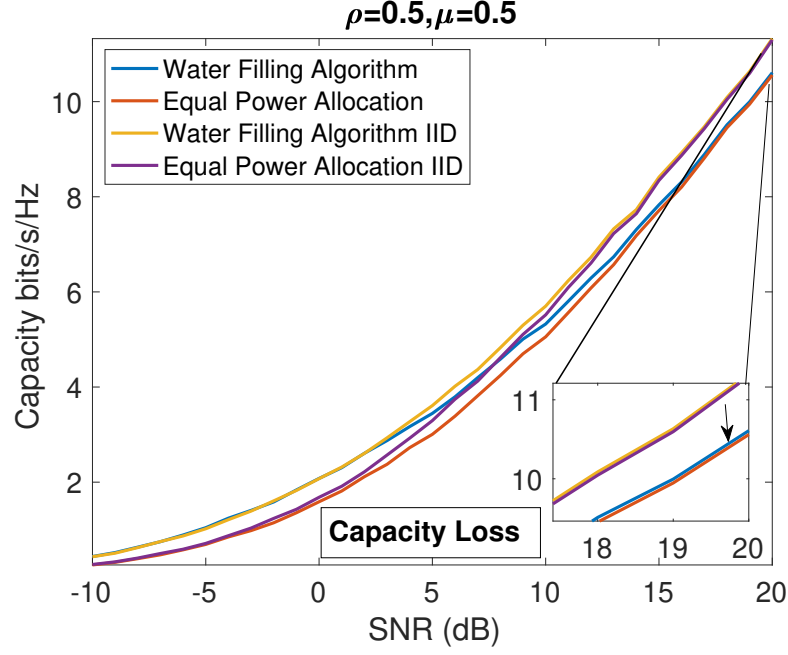


Figure 5.6: 2x2 MIMO capacity using equal power allocation strategy and water-filling strategy in presence of transmit and receive correlation.

and distribution of AoA and AoDs. MIMO system correlation is captured by MIMO system correlation matrix. Transmit correlation and receive correlation is measured based on MIMO channel matrix $\mathbf{H}(\tau, t)$

$$\mathbf{H}(\tau, t) = \begin{bmatrix} h_{11}(\tau, t) & h_{12}(\tau, t) \\ h_{21}(\tau, t) & h_{22}(\tau, t) \end{bmatrix} \quad (5.9)$$

MIMO channel capacity is affected in presence of correlation at the transmit side and/or presence of correlation at receive side. MIMO Tx and Rx correlation is captured by MIMO transmit correlation matrix and MIMO receive correlation matrix. Overall MIMO system correlation is captured by MIMO system correlation matrix

$$\mathbf{R}_H = E \left(\text{vec}(\mathbf{H}(\tau, t)) \text{vec}(\mathbf{H}(\tau, t))^H \right) \quad (5.10)$$

$$\text{vec}(\mathbf{H}) = \begin{bmatrix} h_{11}(\tau, t) \\ h_{12}(\tau, t) \\ h_{21}(\tau, t) \\ h_{22}(\tau, t) \end{bmatrix} \quad (\text{vec}(\mathbf{H}))^H = \begin{bmatrix} h_{11}^*(\tau, t) & h_{12}^*(\tau, t) & h_{21}^*(\tau, t) & h_{22}^*(\tau, t) \end{bmatrix}$$

Transmit correlation matrix and receive correlation matrix can be obtained from measured MIMO channel matrix $\mathbf{H}$ as $\mathbf{R}_{Tx} = E\{\mathbf{H}^H\mathbf{H}\}$ and $\mathbf{R}_{Rx} = E\{\mathbf{H}\mathbf{H}^H\}$ respectively. Similarly, the trace of transmit and receive correlation matrix can be obtained from square of Frobenius norm of MIMO channel matrix $\mathbf{H}$. Trace $tr(\mathbf{H})$ or square of Frobenius norm represent MIMO channel power.

$$\text{tr}\{\mathbf{R}_T\} = \text{tr}\{\mathbf{R}_R\} = E\{\|\mathbf{H}\|_F^2\} = \sum_{i=1}^{N_T} \sum_{j=1}^{N_R} E\{|H_{i,j}|^2\} \quad (5.11)$$

MIMO system correlation matrix is mean of the product of vectorized MIMO channel matrix $\mathbf{H}$.

$$\mathbf{R}_H = E\left(\text{vec}(\mathbf{H}(\tau, t))\text{vec}(\mathbf{H}(\tau, t))^H\right) \quad (5.12)$$

$$\mathbf{R}_H = \begin{bmatrix} E\{h_{11}h_{11}^*\} & E\{h_{11}h_{21}^*\} & E\{h_{11}h_{12}^*\} & E\{h_{11}h_{22}^*\} \\ E\{h_{21}h_{11}^*\} & E\{h_{21}h_{21}^*\} & E\{h_{21}h_{12}^*\} & E\{h_{21}h_{22}^*\} \\ E\{h_{12}h_{11}^*\} & E\{h_{12}h_{21}^*\} & E\{h_{12}h_{12}^*\} & E\{h_{21}h_{22}^*\} \\ E\{h_{22}h_{11}^*\} & E\{h_{22}h_{21}^*\} & E\{h_{22}h_{12}^*\} & E\{h_{22}h_{22}^*\} \end{bmatrix} \quad (5.13)$$

Where main diagonal represents antenna powers $E\{h_{ij}h_{ij}^*\}$. Transmit correlation matrix is represented by $\mathbf{R}_{Tx} = E\{\mathbf{H}^H\mathbf{H}\}$ and receive correlation matrix is represented by $\mathbf{R}_{Rx} = E\{\mathbf{H}\mathbf{H}^H\}$.

5.4 Kronecker Model

MIMO system channel matrix $\mathbf{H}$ can be factorized in terms of transmit correlation matrix $\mathbf{R}_T$ and in terms of receive correlation matrix $\mathbf{R}_R$. Kronecker product model says that transmit and receive correlation matrices can be separated by kronecker product.:

$$\mathbf{R}_H = \mathbf{R}_T \otimes \mathbf{R}_R \quad (5.14)$$

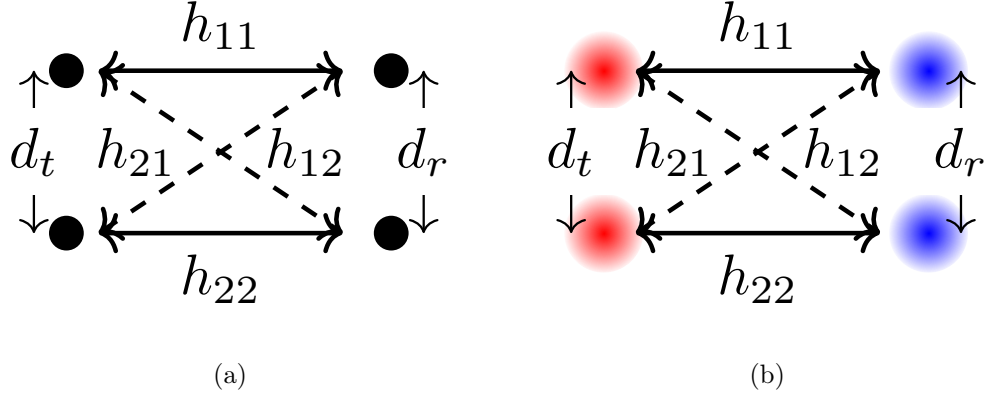


Figure 5.7: MIMO Configuration (a) MIMO (b) Near-field MIMO. Red and blue clouds show near-field around antenna elements. Near-field scenario can arise when Tx and Rx elements are in the near-field of each other or metal objects, which can act as a source secondary waves, in the near-field of receiving antennas.

Hence MIMO channel matrix $\mathbf{H}$ in presence of transmit correlation and receive correlation can be generated as follows:

$$\mathbf{H} = \mathbf{R}_R^{1/2} \mathbf{H}_w \left(\mathbf{R}_T^{1/2} \right)^T \quad (5.15)$$

Where the elements of $\mathbf{H}_w$ are i.i.d CSCG with zero mean and unit variance. i.i.d matrix maximizes MIMO channel capacity, however diagonally correlated channels also known to maximize MIMO channel capacity slightly higher than IID MIMO channels. According to Kronecker correlation model MIMO system correlation matrix $\mathbf{R}_H$ can be obtained by Kronecker product of transmit and receive correlation matrices. It can be rewritten as:

$$\mathbf{H} \sim \mathcal{CN}(\mathbf{0}, \mathbf{R}_T \otimes \mathbf{R}_R) \quad (5.16)$$

Where $\otimes$ denotes Kronecker product.

5.5 Semi-correlated MIMO Channel Capacity

In this section, MIMO channel capacity of semi-correlated channels is discussed. Semi-correlated MIMO channels means the correlation is present at the Tx antenna elements or Rx antenna elements. This depends upon the type of propagation environment. Increased antenna element spacing at the transmitter and at the receiver or the large number of scatterers around transmitter and receiver are required to get decorrelated MIMO channels.

5.5.1 MIMO Channel Capacity in Presence of Tx Correlation

In MIMO channel there are physical scenarios which generate correlation at the transmitter and/or at the receiver. When the correlation at the transmitter is present MIMO channel capacity eq.(5.1) from MIMO signal model with transmit power constraint P_T can be given as [2][3]:

$$C = \log_2 \left(\det \left(\mathbf{I}_{N_R} + \frac{P_T}{N_T \sigma_n^2} \mathbf{H} \mathbf{H}^H \right) \right) \quad \text{bits/s/Hz} \quad (5.17)$$

in case of correlated fading and the fading introduced by separate physical scenario. Let $\mathbf{H} = \mathbf{R}_R \mathbf{H} \mathbf{R}_T$ and using this in Wishart matrix $\mathbf{W} = \mathbf{H} \mathbf{H}^H$ then the MIMO channel capacity eq.(5.17) can be rewritten as [4][5]:

$$C = \log_2 \left(\det \left(\mathbf{I}_{N_R} + \frac{P_T}{N_T \sigma_n^2} \mathbf{R}_R \mathbf{H} \mathbf{R}_T \mathbf{R}_T^H \mathbf{H}^H \mathbf{R}_R^H \right) \right) \quad \text{bits/s/Hz} \quad (5.18)$$

Now by performing eigenvalue decomposition of term $\mathbf{R}_T \mathbf{R}_T^H = \mathbf{\Phi}_{R_T} \mathbf{\Gamma}_{R_T} \mathbf{\Phi}_{R_T}^H$ in eq. (5.19), the MIMO channel capacity in terms of presence of transmit correlation can be written as:

$$C = \log_2 \left(\det \left(\mathbf{I}_{N_R} + \frac{P_T}{N_T \sigma_n^2} \mathbf{\Phi}_{R_T} \mathbf{\Gamma}_{R_T} \mathbf{\Phi}_{R_T}^H \mathbf{R}_T^H \mathbf{H}^H \mathbf{R}_R^H \right) \right) \quad \text{bits/s/Hz} \quad (5.19)$$

where Φ_{R_T} is unitary matrix and Γ_{R_T} is diagonal matrix of eigenvalues of $\mathbf{R}_T \mathbf{R}_T^H$.

$$C = \log_2 \left(\det \left(\mathbf{I}_{N_R} + \frac{P_T}{N_T \sigma_n^2} \mathbf{R}_R \mathbf{H} \Gamma_{R_T} \mathbf{H}^H \mathbf{R}_R^H \right) \right) \quad \text{bits/s/Hz} \quad (5.20)$$

5.5.2 MIMO Channel Capacity in Presence of Rx Correlation

Similar to MIMO channel capacity expression in presence of correlation only at the transmitter, MIMO channel capacity in presence of correlation at the receiver side can be given as function of diagonal eigenvalue matrix $\mathbf{R}_R \mathbf{R}_R^H$. Following same procedure in subsection (5.5.1) MIMO channel capacity with Rx correlation can be obtained.

$$C = \log_2 \left(\det \left(\mathbf{I}_{N_R} + \frac{P_T}{N_T \sigma_n^2} \mathbf{H} \mathbf{H}^H \right) \right) \quad \text{bits/s/Hz} \quad (5.21)$$

$$C = \log_2 \left(\det \left(\mathbf{I}_{N_R} + \frac{P_T}{N_T \sigma_n^2} \mathbf{R}_R \mathbf{H} \mathbf{H}^H \mathbf{R}_R^H \right) \right) \quad \text{bits/s/Hz} \quad (5.22)$$

MIMO channel capacity equation with Rx correlation in eq. (5.22) can be rearranged:

$$C = \log_2 \left(\det \left(\mathbf{I}_{N_R} + \frac{P_T}{N_T \sigma_n^2} \mathbf{H}^H \mathbf{R}_R^H \mathbf{R}_R \mathbf{H} \right) \right) \quad \text{bits/s/Hz} \quad (5.23)$$

Product of Rx correlation matrix $\mathbf{R}_R \mathbf{R}_R^H = \Phi_{R_R} \Gamma_{R_R} \Phi_{R_R}^H$ into unitary matrix Φ_{R_R} and eigenvalue matrix Γ_{R_R} .

$$C = \log_2 \left(\det \left(\mathbf{I}_{N_R} + \frac{P_T}{N_T \sigma_n^2} \mathbf{H}^H \Gamma_{R_R} \mathbf{H} \right) \right) \quad \text{bits/s/Hz} \quad (5.24)$$

Generalized MIMO channel capacity formula in presence of Tx and Rx correlation.

$$C = \log_2 \left(\det \left(\mathbf{I}_{N_R} + \mathbf{H} \Gamma \mathbf{H}^H \right) \right) \quad \text{bits/s/Hz} \quad (5.25)$$

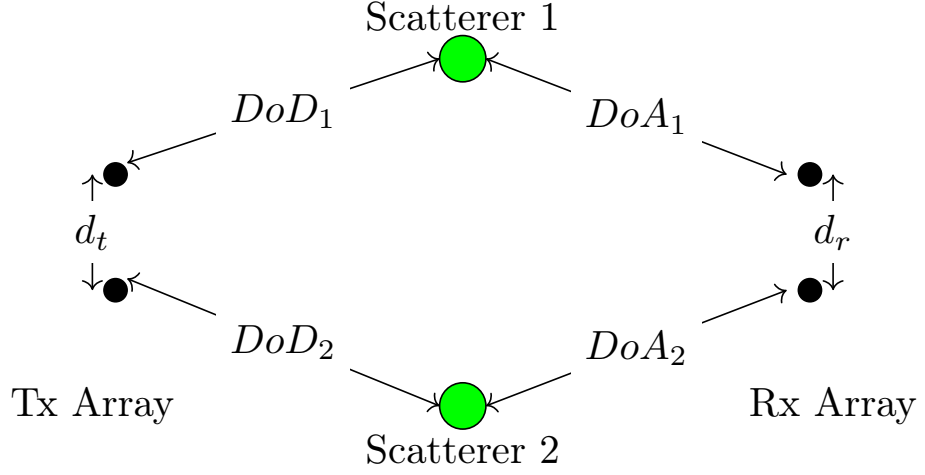


Figure 5.8: Diagonally-correlated 2x2 MIMO channel scenario.

Where $\mathbf{\Gamma} = \frac{P_T}{N_T \sigma_n^2} \mathbf{\Gamma}_{R_R}$ when the correlation is present at the receiver or $\mathbf{\Gamma} = \frac{P_T}{N_T \sigma_n^2} \mathbf{\Gamma}_{R_T}$ when the correlation is present only at the transmitter.

5.6 Diagonally-Correlated MIMO Channels

Diagonally-correlated MIMO channel are type of MIMO correlated channels which exhibit higher MIMO channel capacity than IID MIMO Rayleigh fading channels. Diagonally-correlated channels are realized when power from specific DoD couples into specific DoAs [6][7]. Fig. 5.8 shows physical scenario of diagonally-correlated channels where two equally-powered independent Rayleigh fading components with orthogonal DoDs and AoAs are shown. There is specific symmetry in diagonally-correlated MIMO channel matrix elements. The rows of channels matrix are cyclically left-shifted or right-shifted versions of each other. In IID MIMO channels capacity grows asymptotically with minimum number of transmit and receive antennas in rich multipath scattering environment. IID MIMO Rayleigh fading channels yield maximum capacity and MIMO channels correlation reduces

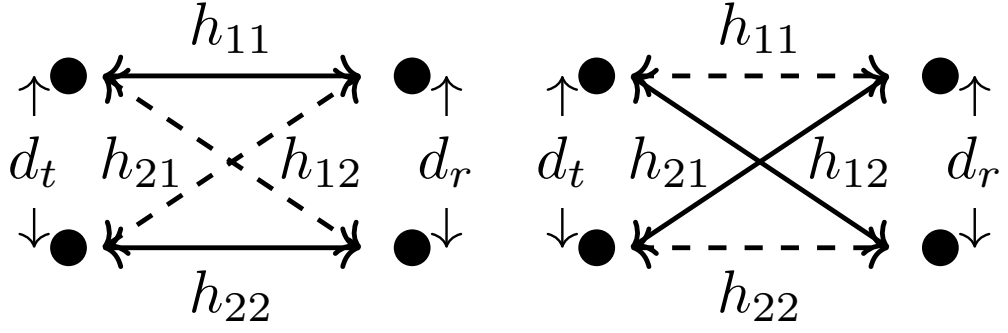


Figure 5.9: Diagonally-correlated 2x2 MIMO channel showing $|h_{11}| - |h_{22}|$ and $|h_{12}| - |h_{21}|$ fading pairs.

MIMO capacity. Channel matrix of 2x2 MIMO channels:

$$\mathbf{H} = \begin{bmatrix} h_{11}(\tau, t) & h_{12}(\tau, t) \\ h_{21}(\tau, t) & h_{22}(\tau, t) \end{bmatrix} \quad (5.26)$$

For diagonally-correlated 2x2 MIMO channels can be given as:

$$\mathbf{H}(\tau, t) = \begin{bmatrix} h_{11}(\tau, t) & h_{12}(\tau, t) \\ h_{12}(\tau, t) & h_{11}(\tau, t) \end{bmatrix} \quad (5.27)$$

Full-MIMO channel correlation matrix can be obtained from complex correlations between all matrix element pairs. Hence, full-MIMO system correlation matrix is obtained as:

$$\mathbf{R}_{\mathbf{H}} = \mathbf{E} \{ \text{vec}(\mathbf{H}(\tau, t)) \text{vec}(\mathbf{H}^H(\tau, t)) \} \quad (5.28)$$

Which can be expressed as mean of the multiplication of MIMO channel matrix elements.

$$\mathbf{R}_{\mathbf{H}} = \begin{bmatrix} E\{h_{11}h_{11}^*\} & E\{h_{11}h_{21}^*\} & E\{h_{11}h_{12}^*\} & E\{h_{11}h_{22}^*\} \\ E\{h_{21}h_{11}^*\} & E\{h_{21}h_{21}^*\} & E\{h_{21}h_{12}^*\} & E\{h_{21}h_{22}^*\} \\ E\{h_{12}h_{11}^*\} & E\{h_{12}h_{21}^*\} & E\{h_{12}h_{12}^*\} & E\{h_{12}h_{22}^*\} \\ E\{h_{22}h_{11}^*\} & E\{h_{22}h_{21}^*\} & E\{h_{22}h_{12}^*\} & E\{h_{22}h_{22}^*\} \end{bmatrix} \quad (5.29)$$

Now, absolute values of correlation matrix of 2x2 diagonally-correlated MIMO channels can be written as:

$$|\mathbf{R}_H| = \begin{bmatrix} 1 & 0 & 0 & 1 \\ 0 & 1 & 1 & 0 \\ 0 & 1 & 1 & 0 \\ 1 & 0 & 0 & 1 \end{bmatrix} \quad (5.30)$$

Ergodic capacity of MIMO system with N_T transmit and N_R receive antennas is given as [4] [3][8]:

$$C_{IID}^- = \mathcal{E}_{\mathbf{H}} \left(\log_2 \left(\det \left(\mathbf{I}_{N_R} + \frac{P_T}{N_T \sigma_n^2} \mathbf{H} \mathbf{H}^H \right) \right) \right) \quad \text{bits/s/Hz} \quad (5.31)$$

According to Jensen's inequality expectation and $\log_2$ can be exchanged, hence MIMO capacity under equal-power allocation can be rewritten as [4]:

$$C_{IID}^- = \log_2 \left(\mathcal{E}_{\mathbf{H}} \left(\det \left(\mathbf{I}_{N_R} + \frac{P_T}{N_T \sigma_n^2} \mathbf{H} \mathbf{H}^H \right) \right) \right) \quad \text{bits/s/Hz} \quad (5.32)$$

Diagonally-correlated MIMO channel $\mathbf{H}$ can be expressed in terms of IID fading matrix $\mathbf{H}_w$ whose entries are ZMCSCG and a coupling matrix $\mathbf{D}$, left unitary matrix $\mathbf{A}_{Rx}$ and right unitary matrix $\mathbf{A}_{Tx}$.

$$\mathbf{H} = \mathbf{A}_{Rx} (\mathbf{D} \odot \mathbf{H}_w) \mathbf{A}_{Tx}^T \quad (5.33)$$

Unitary matrices $\mathbf{A}_{Tx}$ and $\mathbf{A}_{Rx}$ contain orthogonal steering and response vectors of DoDs and DoAs respectively. Note that $\mathbf{D}$ is either diagonal or anti-diagonal matrix with identical entries which means one-to-one coupling between DoDs and DoAs. Additionally, the Kronecker model is not applicable for diagonally-correlated MIMO channels. Kronecker model assumes independent DoDs and DoAs hence any DoD can couple to all DoAs and any DoA can couple to all DoDs.

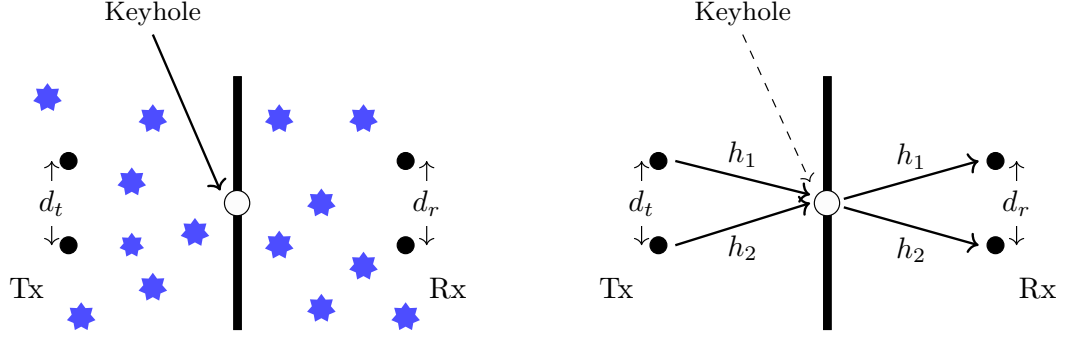


Figure 5.10: Schematic of Keyhole MIMO channel.

5.7 Keyhole or Degenerate MIMO Channels

Keyhole or pinhole channels are observed when Tx antenna array and Rx antenna array is separated by a pinhole or keyhole[9][10][11][12][13][14]. Fig. 5.10 shows a keyhole or pinhole channel where transmitter and receiver are separated by screen with a pinhole. The dimension of keyhole is much smaller than signal dimension. Keyhole channels are predicted in theory and likely to observed in wireless chip-to-chip communication and board-to-board communication. Keyhole channels are also known as rank-deficient or degenerate channels. The capacity of MIMO keyhole channel degenerates to SISO channels. Keyhole MIMO channels are studied as single keyhole channel or multikeyhole channels. Similarly, second order statistics such as level crossing rate and average fade duration of Nakagami-m keyhole channels is given in [15].

5.7.1 Single Keyhole Channel

Single keyhole channel scenarios is achieved when Tx and Rx antenna arrays are separated by a screen with a pinhole and no forward transmission through screen except the keyhole. Channel gain matrix of single Keyhole channel is given by:

$$\mathbf{H} = \mathbf{h}_R \mathbf{h}_T^H \quad (5.34)$$

Where $\mathbf{h}_T \in \mathbb{C}^{N_T \times 1}$ and $\mathbf{h}_R \in \mathbb{C}^{N_R \times 1}$ represents complex gains from N_T transmit antennas to a keyhole and complex gains from a single keyhole to N_R receive antennas. $\mathbf{h}_T$ and $\mathbf{h}_R$ are complex circularly symmetric Gaussian vectors with transmit correlation matrix $\mathbf{R}_T = \mathbf{E}\{\mathbf{h}_T \mathbf{h}_T^H\}$ and receive correlation matrix $\mathbf{R}_R = \mathbf{E}\{\mathbf{h}_R \mathbf{h}_R^H\}$. MIMO channel matrix $\mathbf{H}$ is normalized such that $\mathbf{E}\{\|\mathbf{H}\|^2\} = N_T N_R$ and

$$\frac{E\{\|\mathbf{h}_T\|^2\}}{N_T} = \frac{E\{\|\mathbf{h}_R\|^2\}}{N_R} = 1 \quad (5.35)$$

which is same as

$$\frac{\text{trace}\{\mathbf{R}_T\}}{N_T} = \frac{\text{trace}\{\mathbf{R}_R\}}{N_R} = 1 \quad (5.36)$$

MIMO channel capacity when CSI is known only at the receiver:

$$C = \log_2 \left(\det \left(\mathbf{I}_{N_R} + \frac{P_T}{N_T \sigma_n^2} \mathbf{H} \mathbf{H}^H \right) \right) \quad \text{bits/s/Hz} \quad (5.37)$$

It's known that when N_T and N_R tend to infinity, the distribution of MIMO channel capacity tend to be Gaussian if

$$\frac{\text{trace}\{\mathbf{R}_T\}}{N_T} < \infty, \quad \frac{\|\mathbf{R}_T\|^2}{N_T^2} \rightarrow 0 \quad \text{as} \quad N_T \rightarrow \infty \quad (5.38)$$

and

$$\frac{\text{trace}\{\mathbf{R}_R\}}{N_R} < \infty, \quad \frac{\|\mathbf{R}_R\|^2}{N_R^2} \rightarrow 0 \quad \text{as} \quad N_R \rightarrow \infty \quad (5.39)$$

the asymptotic mean and variance of Keyhole MIMO channel capacity

$$\mu = \log_2 \left(1 + N_R \frac{P_T}{N_T \sigma_n^2} \right), \quad \sigma_c^2 = \frac{\|\mathbf{R}_T\|^2}{N_T^2} + \frac{\|\mathbf{R}_R\|^2}{N_R^2} \quad (5.40)$$

5.7.2 Multi-keyhole MIMO Channel

Multi-keyhole channel may have number of keyholes[11]. Ergodic mutual information of multi-keyhole channels is studied in [14]. Similarly, outage capacity distribution is assessed by authors in [10]. Theoretical and experimental ver-

ification of MIMO keyhole channels is presented in [9][12]. Channel matrix of multi-keyhole channel:

$$\mathbf{H} = \sum_{k=1}^M a_k \mathbf{h}_{Rk} \mathbf{h}_{Tk} = \mathbf{H}_R \mathbf{A} \mathbf{H}_T^T \quad (5.41)$$

Where M is the number of keyholes and a_k is complex gain of the k th keyhole. $\mathbf{h}_{Tk} \in \mathbb{C}^{N_T \times 1}$ is random vector or channel gains from N_T transmit antennas to k th keyhole channel and $\mathbf{h}_{Rk} \in \mathbb{C}^{N_R \times 1}$ is a random vector or channel gains from k th keyhole to N_R receive antennas. Similarly, $\mathbf{H}_T \in \mathbb{C}^{N_T \times M}$ or $\mathbf{H} = \{\mathbf{h}_{T1}, \mathbf{h}_{T2}, \dots, \mathbf{h}_{TM}\}$ is complex channel matrix of complex gains from N_T transmit antennas to M th keyhole channel and $\mathbf{H}_R \in \mathbb{C}^{N_R \times M}$ or $\mathbf{H}_R = \{\mathbf{h}_{R1}, \mathbf{h}_{R2}, \dots, \mathbf{h}_{RM}\}$ is a complex matrix of complex gains from M keyhole to N_R receive antennas. For every k , $\mathbf{h}_{Tk}$ and $\mathbf{h}_{Rk}$ are mutually independent complex circular symmetric Gaussian vectors with correlation matrices $\mathbf{T}_{Tk} = E\{\mathbf{h}_{Tk} \mathbf{h}_{Tk}^H\}$ and $\mathbf{R}_{Rk} = E\{\mathbf{h}_{Rk} \mathbf{h}_{Rk}^H\}$. For independent keyholes $E\{\mathbf{h}_{Tk} \mathbf{h}_{Tk}^H\} = \{\mathbf{h}_{Rk} \mathbf{h}_{Rk}^H\} = 0$. For comparison $\mathbf{H}$ is normalized so that $E\{||\mathbf{H}||^2\} = N_T N_R$ and for every k $\sum_{k=1}^M |a_k|^2 = 1$.

$$\frac{\{||\mathbf{h}_{Tk}||^2\}}{N_T} = \frac{\{||\mathbf{h}_{Rk}||^2\}}{N_R} = 1 \quad (5.42)$$

Instantaneous capacity of frequency-flat quasi-static multi-keyhole MIMO channel when CSI is available at Rx only [9][12].

$$C = \log_2 \left(\det \left(\mathbf{I}_{N_R} + \frac{P_T}{N_T \sigma_n^2} N_R \mathbf{B}_R \mathbf{A} \mathbf{B}_T \mathbf{A}^H \right) \right) \quad \text{bits/s/Hz} \quad (5.43)$$

Where $\mathbf{B}_T = \frac{\mathbf{H}_T^H \mathbf{H}_T}{N_T}$ and $\mathbf{B}_R = \frac{\mathbf{H}_R^H \mathbf{H}_R}{N_R}$. Multi-keyhole channels can be of two types: Full-rank multi-keyhole MIMO channel ($M \geq \min(N_T, N_R)$) and rank-deficient multi-keyhole MIMO channel ($M < \min(N_T, N_R)$).

5.8 Antenna Correlation

Antenna correlation is combined effect of the channel and antennas [16]. Three correlation types are used in the literature which are signal correlation, envelope correlation and power correlation. Mathematically, signal correlation is [17]:

$$\rho_{m,n} = \frac{E[(X_m - E[X_m])(X_n - E[X_n])^*]}{\sqrt{E[|X_m - E[X_m]|^2]E[|X_n - E[X_n]|^2]}} \quad (5.44)$$

From above equation envelope and power correlation can be derived by replacing X_i $i \in \{n, m\}$ with $|X_i|$ and $|X_i|^2$ respectively. Complex correlation coefficient is useful for channel modeling and characterization of MIMO channels. Antenna perspective of calculating antenna correlation is based on embedded far-field radiation functions [18]:

$$\rho_{mn} = \frac{\int \int_{4\pi} g_m^H(\Omega) P_{inc}(\Omega) g_n(\Omega) d\Omega}{\sqrt{\int \int_{4\pi} g_m^H(\Omega) P_{inc}(\Omega) g_m(\Omega) d\Omega \cdot \int \int_{4\pi} g_n^H(\Omega) P_{inc}(\Omega) g_n(\Omega) d\Omega}} \quad (5.45)$$

Where g_i is far-field function, Ω is solid angle and P_{inc} is dyadic power angular spectrum. For isotropic environment in RC $\mathbf{P}_{inc} = \mathbf{I}$. Correlation matrix at N-port antenna can be constructed once antenna correlations are known:

$$\Phi = \begin{bmatrix} 1 & \rho_{12} & \rho_{13} & \cdots & \rho_{1N} \\ \rho_{21}^* & 1 & \rho_{23} & \cdots & \rho_{2N} \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ \rho_{1N}^* & \rho_{2N}^* & \rho_{3N} & \cdots & 1 \end{bmatrix} \quad (5.46)$$

Spatial correlation without antenna effect is deterministic sinc function of antenna separation. There are many correlation models.

5.8.1 Uniform Correlation

Uniform correlation is simplest correlation model. Practical multi-port-antenna have uniform correlation under certain circumstances [19].

$$\rho_{mn} = \rho \quad (5.47)$$

5.8.2 Exponential Correlation

Exponential correlation as a function of distance x_{mn} between m th and n th antenna elements and de-correlation distance [20][21]:

$$\rho_{mn} = \exp\left(-\frac{x_{mn}}{x_0}\right) \quad (5.48)$$

Exponential correlation depends on both antenna elements separation and de-correlation distance x_0 .

5.8.3 Bessel Correlation

Bessel correlation model is function of antenna elements separation x_{mn} and wave number k [22]:

$$\rho_{mn} = J_0(kx_{mn}) \quad (5.49)$$

Where $J_0()$ is zeroth order Bessel function of the first kind.

5.9 MIMO Performance Metrics

There are many parameters based on which MIMO system is evaluated. Main MIMO system performance evaluation parameters are:

5.9.1 Diversity Gain

Correlation matrix of N-port antenna in multipath Rayleigh fading environment is:

$$\mathbf{R} = E[\mathbf{h}\mathbf{h}^H] \quad (5.50)$$

Where $\mathbf{h}$ is multipath fading gain of the diversity channel and $\mathbf{h}^H$ is Hermitian of $\mathbf{h}$. The output power of MRC combiner $P_{MRC} = \mathbf{h}^H \mathbf{h}$. Diversity gain is ratio of output SNR of diversity antenna to output SNR of ideal antenna at 1% outage level [23].

$$DG = \frac{F^{-1}(\gamma)}{F_{ideal}^{-1}(\gamma)}|_{1\%} \quad (5.51)$$

CDF of the MRC output SNR for two-port antenna [18]:

$$F(\gamma) = 1 - \frac{\lambda_1 \exp(-\gamma/\lambda_1) - \lambda_2 \exp(-\gamma/\lambda_2)}{\lambda_1 - \lambda_2} \quad (5.52)$$

eigenvalues of the co-variance matrix:

$$\mathbf{R} = \begin{bmatrix} 1 & \rho \\ \rho^* & 1 \end{bmatrix} \quad (5.53)$$

are $\lambda_1 = 1 + |\rho|$ and $\lambda_2 = 1 - |\rho|$.

5.9.2 Diversity Measure

Easy way for MIMO characterization as compared to diversity gain is diversity measure. Hence, independent samples denoting diversity measure [24]. It has been used for the estimation of number of independent samples in RC [25, 26, 27]

$$N_{ind} = \left(\frac{tr(\mathbf{R})}{\|\mathbf{R}\|_F} \right)^2 \quad (5.54)$$

The effect of polarization on diversity performance in reverberation chamber is shown in [28]. Authors have also measured apparent diversity gain and effective

diversity gain in reverberation chamber at power gain CDF threshold of 0.001 or 10^{-2} . Authors in [29] have proposed diversity measure without extra antennas in reverberation chamber. Apparent and effective diversity has been measured.

5.9.3 K-factor

K-factor is defined as ratio of LOS component to NLOS scattered components. There are various ways to measure K-factor that are available in the literature. In reverberation chamber it is ratio of un-stirred field to stirred field. K-factor measurements and parameters estimation has been presented in [30]. K-factor can be defined as:

$$K = \frac{A^2}{2\sigma^2} = \frac{(|\langle S_{21} \rangle|)^2}{\langle S_{21} - \langle S_{21} \rangle \rangle^2} \quad (5.55)$$

Where $\langle S_{21} \rangle$ is complex average of frequency transfer function. Narrowband K-factor

$$f_R(r) = \frac{2r(K+1)}{\Omega} \exp\left(-K - \frac{(K+1)r^2}{\Omega}\right) \times I_0\left(2r\sqrt{\frac{K(K+1)}{\Omega}}\right), r \geq 0, K \geq 0, \Omega \geq 0 \quad (5.56)$$

$\Omega = E[R^2]$ is scale parameter. The Rician K-factor can be estimated by a moment-based method [31][32].

$$K = \frac{\sqrt{1 - \text{Var}[P(d)]/(E[P(d)])^2}}{1 - \sqrt{1 - \text{Var}[P(d)]/(E[P(d)])^2}} \quad (5.57)$$

$$K = \frac{\sqrt{1 - \gamma}}{1 - \sqrt{1 - \gamma}} \quad (5.58)$$

Where $\gamma = \sigma_r^2/P_r^2$, σ_r is the variance of the receive signal power about its mean P_r . Rician factor in terms of MIMO channel matrix $\mathbf{H}$ is given as:

$$K(\text{dB}) = \left(\frac{\varepsilon[|H|]^2}{2 \text{var}(|H|)} \right) \quad (5.59)$$

Parameters	Values
USRP Type	X310
Tx Gain	4 dB
Rx Gain	4 dB
Sample Rate	400 ksps
Interpolation	500
Decimation	500
Sampling Time	$2.5 \mu s$
Channel Mapping	MIMO
USRP Clock	200 Mbps

Table 5.1: USRP configuration parameters.

5.10 Experimental Setup I

2x2 MIMO empirical measurements are presented. Experimental setup is shown in Fig. 5.12 which shows overview of measurement setup and inside view of metal enclosure. Additionally, NLOS metal sheet attached to panel and mode-stirrer is shown in Fig. 5.13. In this experimental setup 2x2 MIMO measurements are performed in NLOS propagation environment. NLOS between Tx and Rx antennas is created using metal sheet of dimensions. Measurement results are presented in the form of measured MIMO channel elements i.e. transfer function of each sub-channel elements. This is followed by MIMO channel capacity, eigenvalues, Rician K-factor, Doppler spread, coherence bandwidth etc. MIMO measurement results are conducted with the following Tx and Rx MIMO settings:

- Tx inter-element spacing $d_t = 45 \text{ mm}$
- Rx inter-element spacing $d_r = 42 \text{ mm}$
- Distance between Tx and Rx $D = 45 \text{ mm}$

Table 5.1 shows near-field MIMO testbed configuration parameters. USRP is configured to operate at sample rate of 400 ksps with interpolation and decimation factor of 500. Sampling rate of 400 ksps is obtained when USRP clock rate is set to 200 Msps and hence sampling rate of the USRP is:

$$\text{Sample Rate} = \frac{\text{Master Clock Rate}}{\text{Interpolation/Decimation}} \quad (5.60)$$

Data rate for complex samples each of I samples and Q samples:

$$\text{Data Rate} = \frac{\text{Master Clock Rate}}{\text{Interpolation/Decimation}} * 2 * 16 \quad (5.61)$$

5.11 Measurement Results

Complex MIMO channel transfer function is measured in NLOS scenario in metal enclosure. Fig.5.14 shows complex frequency transfer function of sub-channel elements of 2x2 MIMO channel matrix obtained in NLOS measurement scenario. Measured complex MIMO channel matrix is used in standard MIMO capacity formula and empirical 2x2 MIMO channel capacity has been measured. Fig.5.15 shows capacity of NLOS 2x2 MIMO system in mode-stirred metal enclosure. It can be seen that channel 2 has highest attainable channel capacity of 8.26 bit-s/s/Hz. Fig.5.16 shows eigenvalues of MIMO channel matrix. It can be seen that there are two distinct channels and the power distribution in each channel is different. Information about eigenvalues helps to determine whether MIMO channel is properly conditioned and is not rank-deficient. It also helps to optimally distribute transmit power when CSI is not available at the transmitter and when CSI is available at the transmitter. General strategies for power distribution are equal power allocation in case of CSI not known at the Tx and water-filling strategy when CSI is also known at the transmitter. However, inverse power allocation has also been used as a strategy to distribute total transmit power. Throughput of SVD-MIMO with inverse power allocation in reverberation chamber is measured in [33] where higher power is allocated to more weaker channels contrary to water-filling algorithm which assigns more power to stronger channels. Throughput is measured using ideal-threshold receiver based model [34]. Fig.5.17 shows complex cloud of subchannel elements of channel matrix $\mathbf{H}$. It can be seen that the S21 parameters of each sub-channel form complex cloud centered at origin

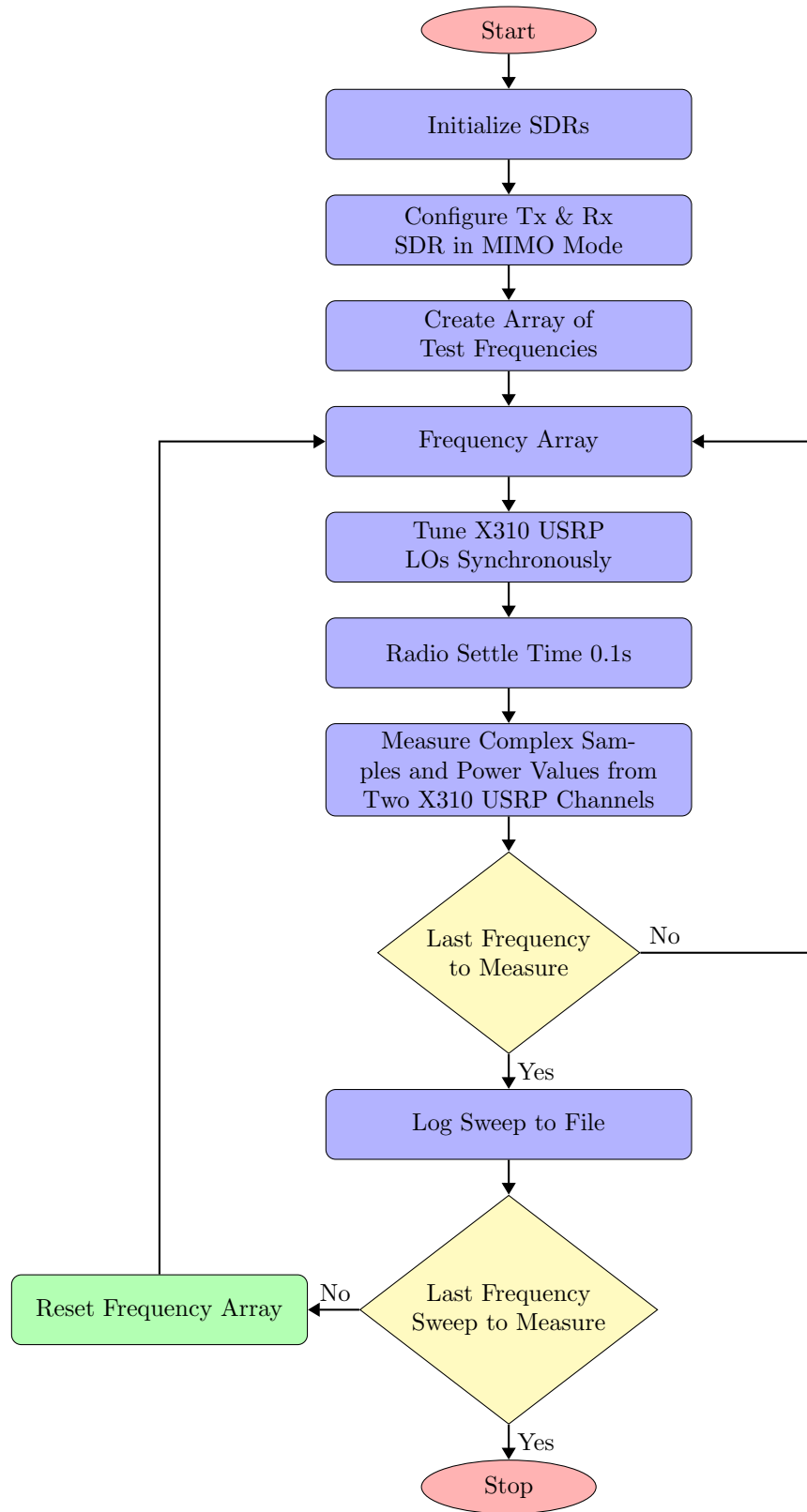


Figure 5.11: Frequency transfer function measurement flowchart.

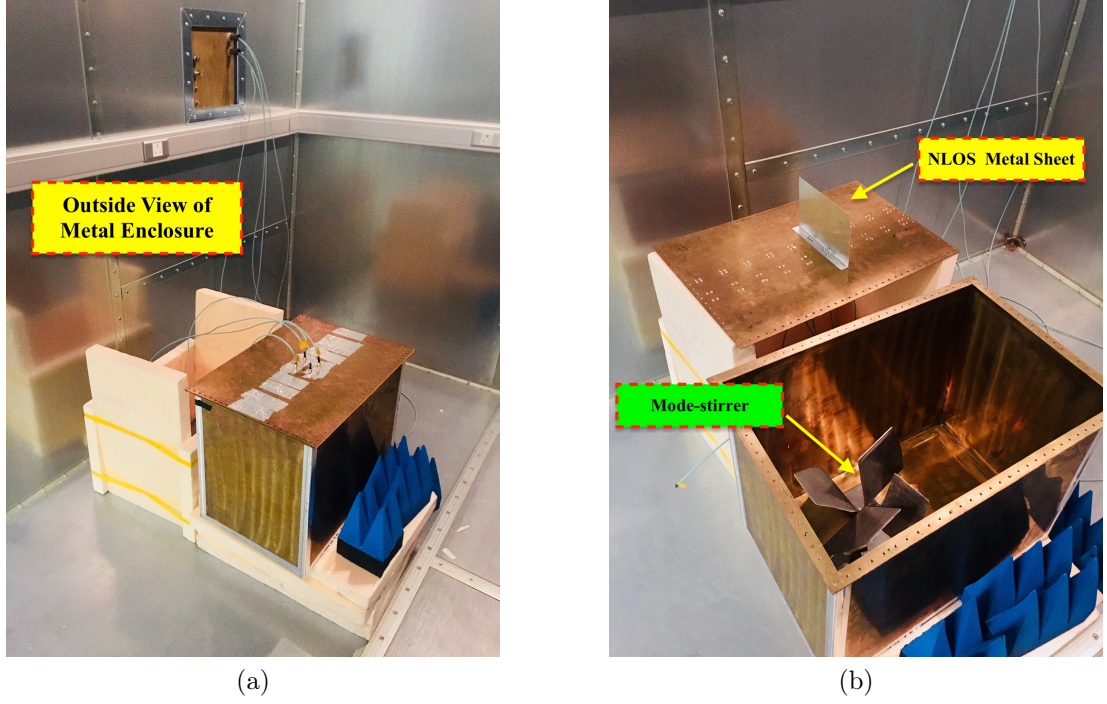


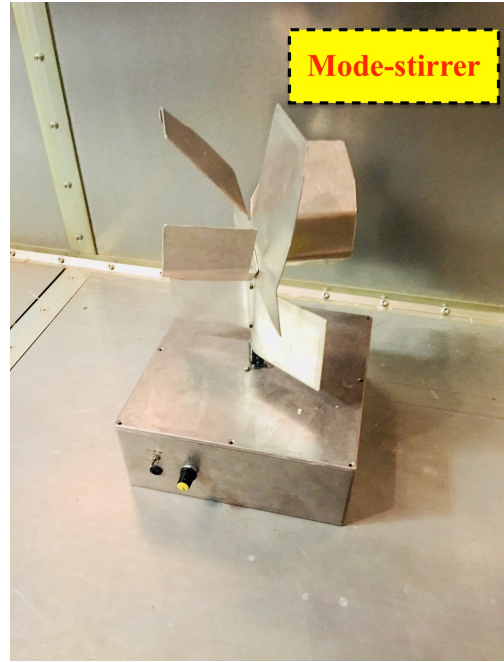
Figure 5.12: Measurement setup (a) overview of measurement setup (b) inside view of measurement setup showing panel and mode-stirrer.

in complex cloud. Loading of chamber has implication on the shape of complex cloud which means for different amount of loading it develops distinct clusters and decreases its spread [35]. Loading the chamber means coherence bandwidth is increased and field distribution deviates from isotropic to non-isotropic.

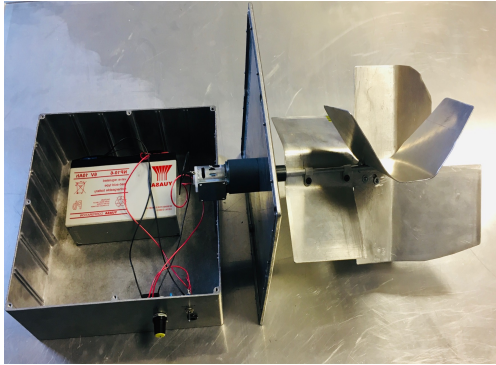
Fig.5.19 shows Rician factor or K-factor. It can be seen that the K-factor is different for different frequencies which depends on the mode-stirrer positions. Mode-stirrer in this setup is set to highest attainable speed and is continuous rotation mode. Uncertainty evaluation of low Rician K-factor in reverberation chamber is presented in [30]. Angle of arrival in rich-scattered multipath fading channel is expected to be uniformly distributed which is shown in Fig.5.22. It can be seen from the histogram of angle-of-arrival that almost an equal number of multipath samples are received from all directions.



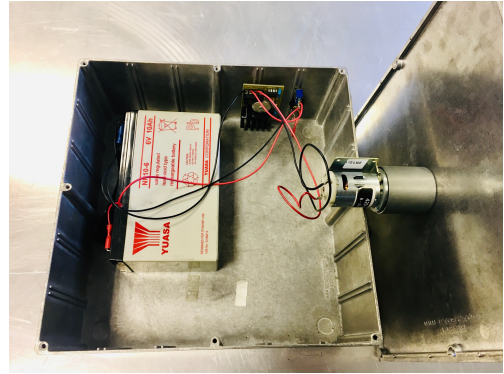
(a)



(b)



(c) Mode-stirrer Base



(d) Mode-stirrer Control

Figure 5.13: Measurement setup (a) metal sheet connected with the panel for NLOS measurements (b) mode-stirrer.

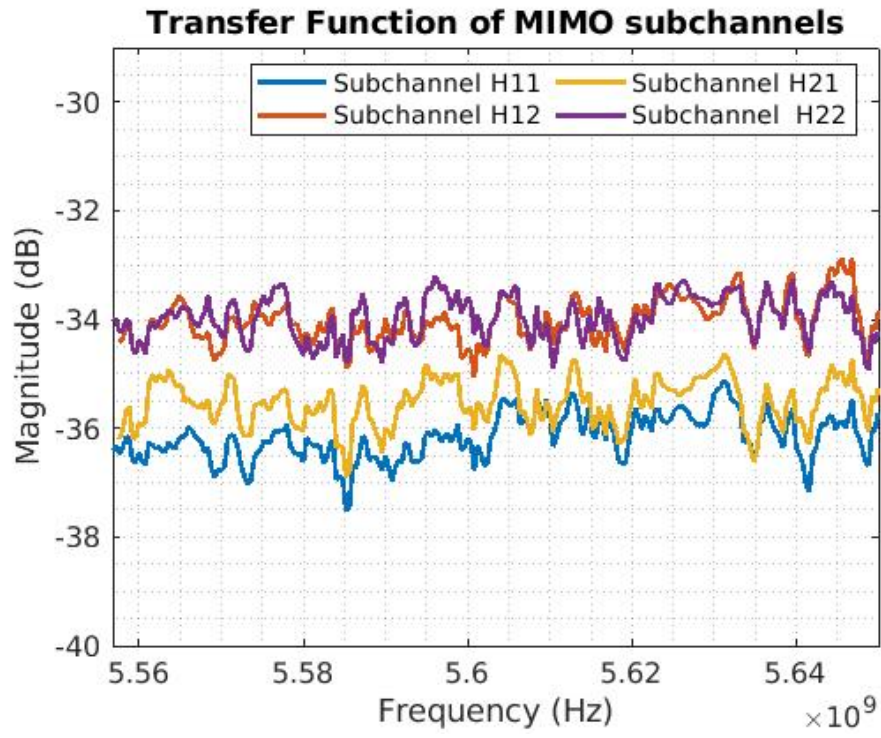


Figure 5.14: Subchannel elements of MIMO channel matrix: after sliding window of width=5.

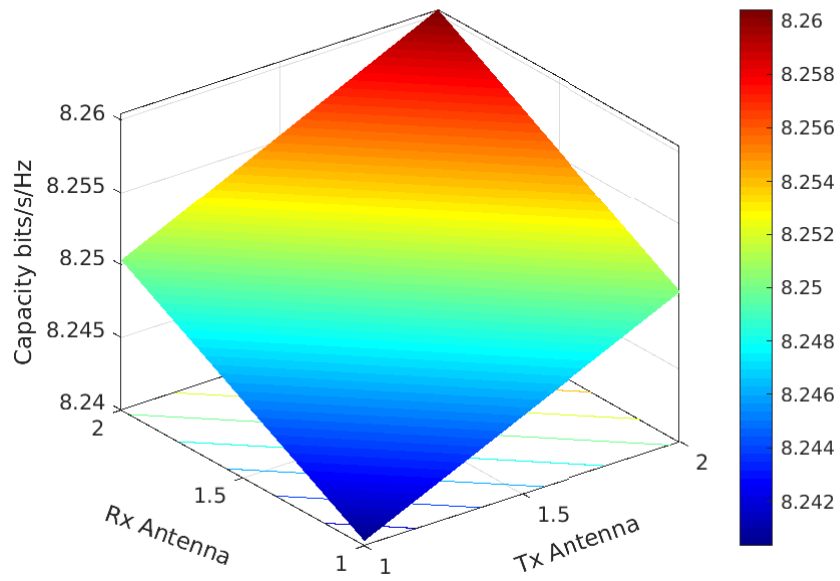


Figure 5.15: 2x2 NLOS MIMO channel capacity.

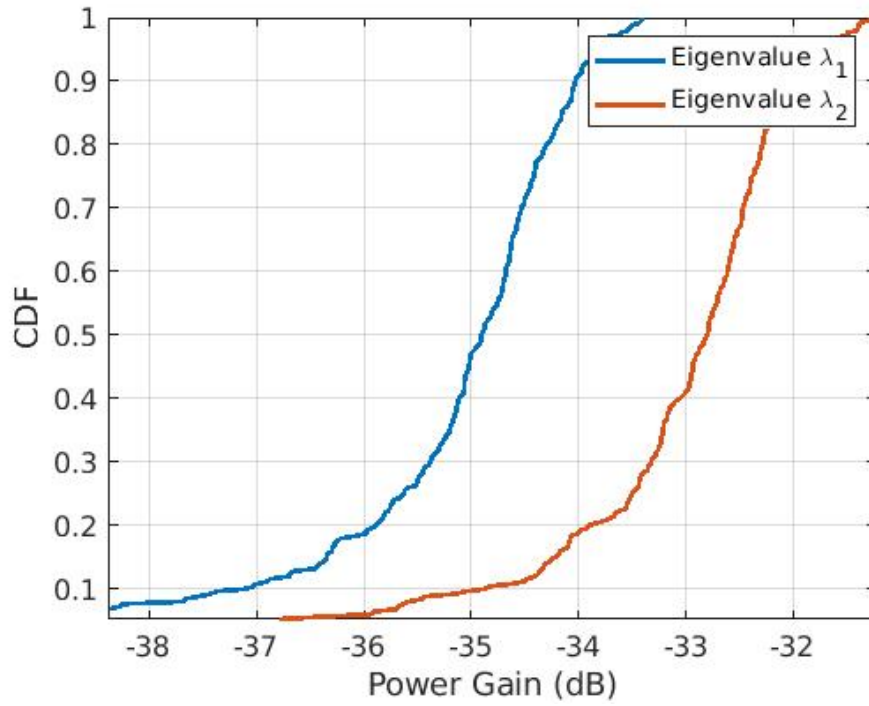


Figure 5.16: CDF of Eigenvalues of MIMO channel matrix $\mathbf{H}$.

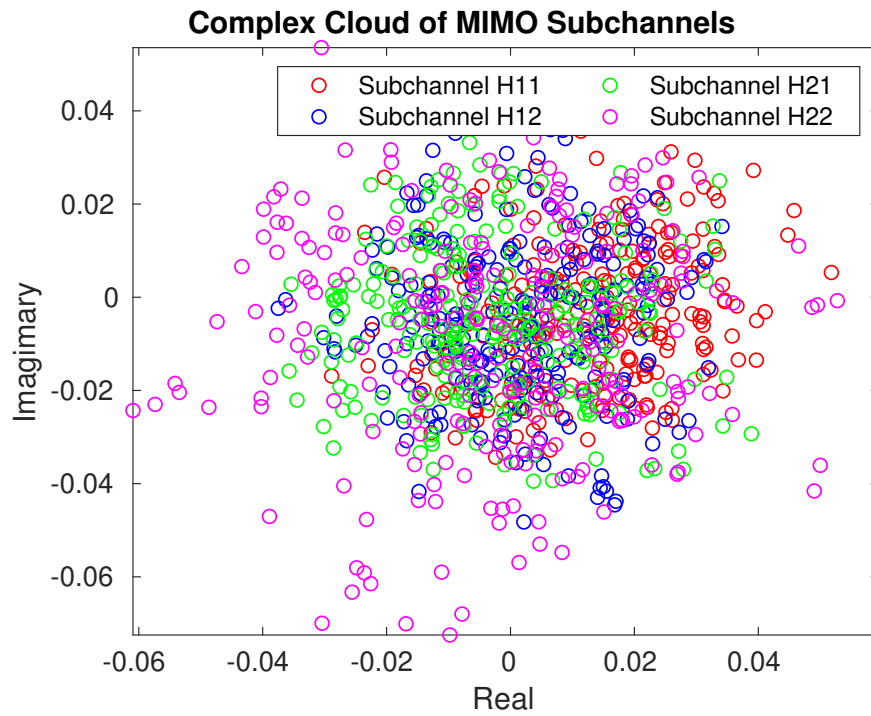


Figure 5.17: Complex cloud of MIMO sub-channels.

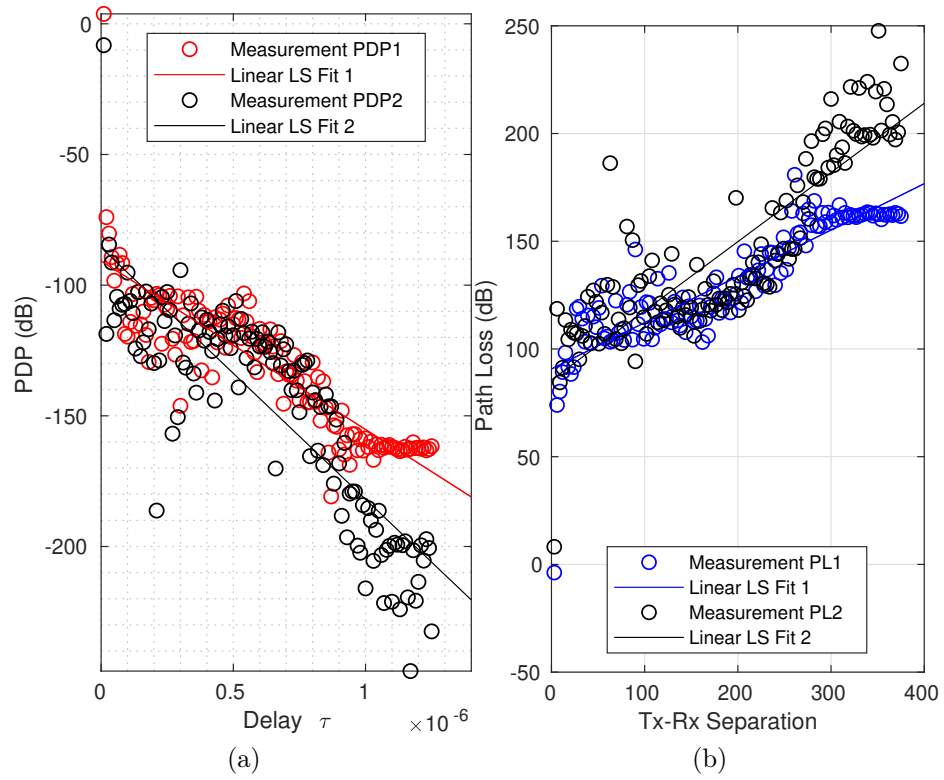


Figure 5.18: PDP and Path loss with least square fit.

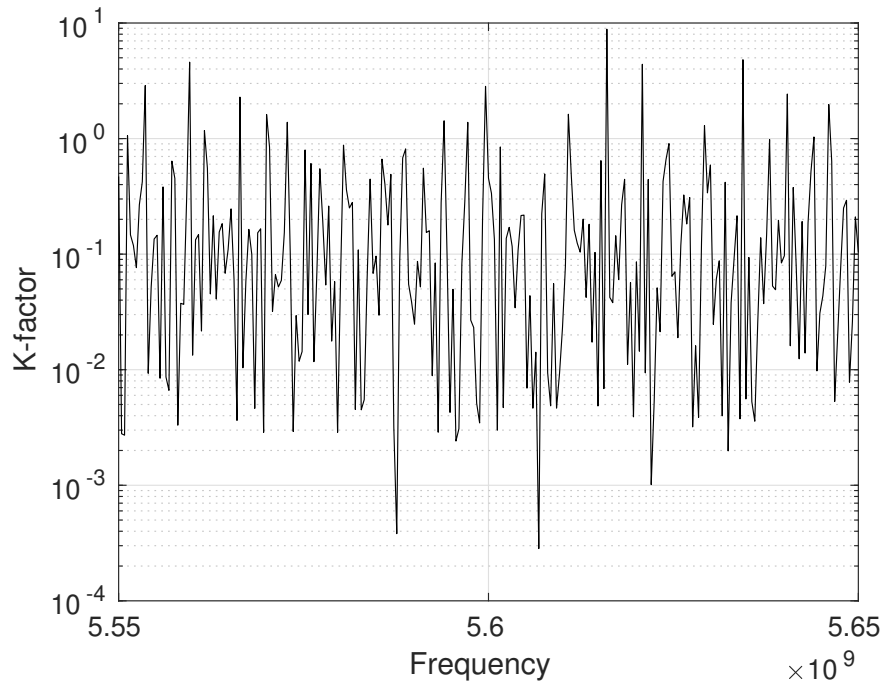


Figure 5.19: Rician Factor.

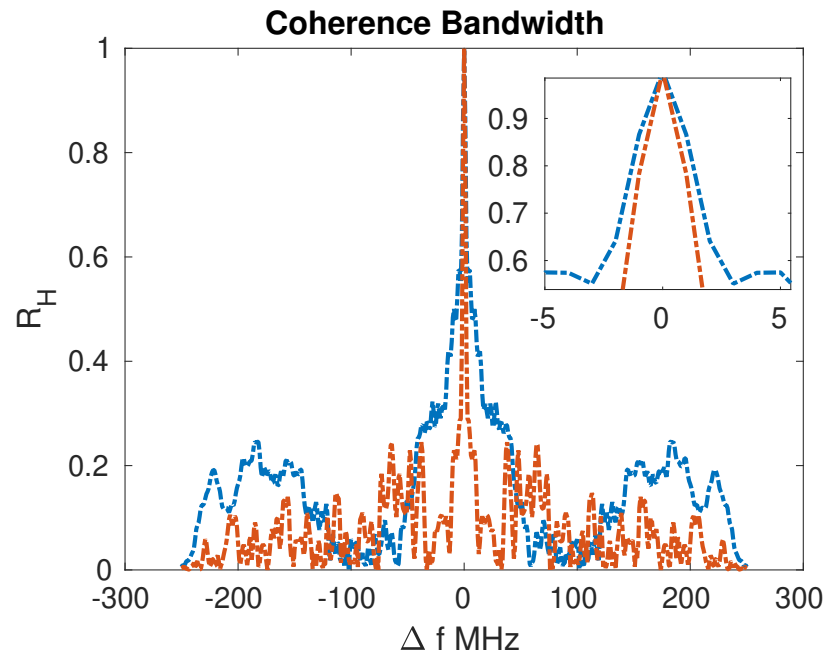


Figure 5.20: Coherence bandwidth.

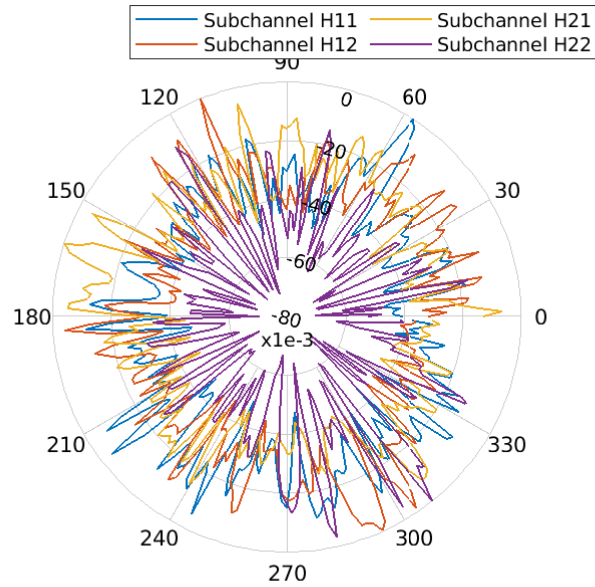


Figure 5.21: Polarplot of MIMO subchannels after sliding window.

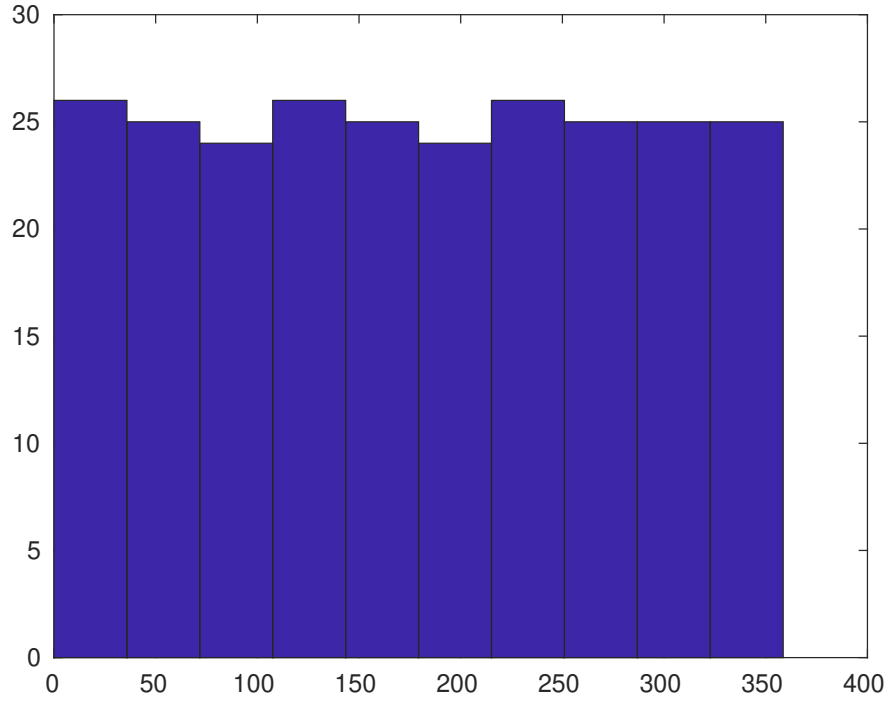


Figure 5.22: Histogram of AoA is uniformly distributed.

5.12 Conclusion

In this chapter MIMO communication is investigated in detail. Mathematical system model for MIMO system is studied and expression for the capacity are explained under transmit power constraints and transmit and receive correlation. MIMO system increases spectral efficiency of communication system, but the capacity of MIMO system degrades in presence of transmit and receive correlation i.e. when the MIMO channel is doubly correlated. Additionally, MIMO channel capacity can be degraded under keyhole or pinhole MIMO channels which are also known as degenerate channels or rank-deficient channels. It's known that IID MIMO channel maximize MIMO channel capacity but there are still exceptions like diagonally-correlated MIMO channels which can give MIMO capacity higher than the IID MIMO channels, however the physical scenarios that can give rise to diagonally-correlated channels are considered to be rare and hence rare experimental measurements. MIMO channel capacity is maximized under transmit power constraints, therefore channel state information at the transmitter and at the receiver side is critical component. When CSI is only available at

the receiver side and no CSI is available at the transmitter side then equal power allocation is best strategy to maximize the MIMO capacity. However, the best strategy for power allocation is power allocation using water-filling algorithm. According to water-filling algorithm higher power is allocated to channels with higher SNRs. It's applied when CSI is also known at the transmitter side. Finally, 2x2 MIMO measurements are also conducted using software-defined-radio testbed in the metal enclosure. MIMO channel transfer function is measured in the metal enclosure from which 2x2 MIMO channel capacity is determined. The measurements are performed at different transmit and receive antenna elements spacing and the distance between transmitter and receiver. Important parameters like Rician factor, coherence bandwidth, angular spread of the mutlipath is discussed.

Bibliography

- [1] Robert G Gallager. *Information theory and reliable communication*, volume 2. Springer, 1968.
- [2] Niklas Jaldén, Per Zetterberg, Bjorn Ottersten, Aihua Hong, and Reiner Thoma. Correlation properties of large scale fading based on indoor measurements. In *2007 IEEE Wireless Communications and Networking Conference*, pages 1894–1899. IEEE, 2007.
- [3] Andrea Goldsmith. *Wireless communications*. Cambridge university press, 2005.
- [4] Andreas F Molisch. *Wireless communications*, volume 34. John Wiley & Sons, 2012.
- [5] Peter J Smith, Sumit Roy, and Mansoor Shafi. Capacity of mimo systems with semicorrelated flat fading. *IEEE Transactions on Information Theory*, 49(10):2781–2788, 2003.
- [6] Hüseyin Ozelik and Claude Oestges. Capacity of diagonally correlated mimo channels. In *2005 IEEE 61st Vehicular Technology Conference*, volume 1, pages 116–120. IEEE, 2005.
- [7] Hüseyin Ozelik and Claude Oestges. Some remarkable properties of diagonally correlated mimo channels. *IEEE transactions on vehicular technology*, 54(6):2143–2145, 2005.
- [8] Arogyaswami Paulraj, Arogyaswami Paulraj Rohit, Rohit Nabar, and Dhananjay Gore. *Introduction to space-time wireless communications*. Cambridge university press, 2003.
- [9] Peter Almers, Fredrik Tufvesson, and Andreas F Molisch. Keyhole effect in mimo wireless channels: Measurements and theory. *IEEE Transactions on Wireless Communications*, 5(12):3596–3604, 2006.

- [10] Georgy Levin and Sergey Loyka. On the outage capacity distribution of correlated keyhole mimo channels. *IEEE Transactions on Information Theory*, 54(7):3232–3245, 2008.
- [11] George Levin and Sergey Loyka. Multi-keyhole mimo channels: Asymptotic analysis of outage capacity. In *2006 IEEE International Symposium on Information Theory*, pages 1305–1309. IEEE, 2006.
- [12] Peter Almers, Fredrik Tufvesson, and Andreas F Molisch. Measurement of keyhole effect in a wireless multiple-input multiple-output (mimo) channel. *IEEE communications letters*, 7(8):373–375, 2003.
- [13] Georgy Levin and Sergey Loyka. From multi-keyholes to measure of correlation and power imbalance in mimo channels: Outage capacity analysis. *IEEE transactions on information theory*, 57(6):3515–3529, 2011.
- [14] Caijun Zhong, Shi Jin, Kai-Kit Wong, and Matthew R McKay. Ergodic mutual information analysis for multi-keyhole mimo channels. *IEEE transactions on wireless communications*, 10(6):1754–1763, 2011.
- [15] Nikola Zlatanov, Zoran Hadzi-Velkov, and George K Karagiannidis. Level crossing rate and average fade duration of the double nakagami-m random process and application in mimo keyhole fading channels. *IEEE Communications Letters*, 12(11):822–824, 2008.
- [16] Xiaoming Chen. Antenna correlation and its impact on multi-antenna system. *Progress In Electromagnetics Research*, 62:241–253, 2015.
- [17] Jean-Philippe Kermoal, Laurent Schumacher, Klaus I Pedersen, Preben E Mogensen, and Frank Frederiksen. A stochastic mimo radio channel model with experimental validation. *IEEE Journal on selected areas in Communications*, 20(6):1211–1226, 2002.
- [18] Xiaoming Chen, Per-Simon Kildal, Jan Carlsson, and Jian Yang. Mrc diversity and mimo capacity evaluations of multi-port antennas using reverbera-

- tion chamber and anechoic chamber. *IEEE Transactions on Antennas and Propagation*, 61(2):917–926, 2012.
- [19] Per-Simon Kildal, Xiaoming Chen, Mattias Gustafsson, and Zhengzhao Shen. Mimo characterization on system level of 5g microbase stations subject to randomness in los. *IEEE Access*, 2:1062–1075, 2014.
- [20] Marco Chiani, Moe Z Win, and Alberto Zanella. On the capacity of spatially correlated mimo rayleigh-fading channels. *IEEE Transactions on Information Theory*, 49(10):2363–2371, 2003.
- [21] Dmitry Chizhik, Farrokh Rashid-Farrokhi, Jonathan Ling, and Angel Lozano. Effect of antenna separation on the capacity of blast in correlated channels. *IEEE Communications letters*, 4(11):337–339, 2000.
- [22] Theodore S Rappaport et al. *Wireless communications: principles and practice*, volume 2. prentice hall PTR New Jersey, 1996.
- [23] Mischa Schwartz, William R Bennett, and Seymour Stein. *Communication systems and techniques*. John Wiley & Sons, 1995.
- [24] Eduard Jorswieck, Holger Boche, et al. Majorization and matrix-monotone functions in wireless communications. *Foundations and Trends® in Communications and Information Theory*, 3(6):553–701, 2007.
- [25] Xiaoming Chen. On independent platform sample number for reverberation chamber measurements. *IEEE transactions on electromagnetic compatibility*, 54(6):1306–1309, 2012.
- [26] Xiaoming Chen. Experimental investigation of the number of independent samples and the measurement uncertainty in a reverberation chamber. *IEEE transactions on electromagnetic compatibility*, 55(5):816–824, 2013.

- [27] Ryan J Pirkel, Kate A Remley, and Christian S Lötbäck Patané. Reverberation chamber measurement correlation. *IEEE Transactions on Electromagnetic Compatibility*, 54(3):533–545, 2011.
- [28] Mohsen Koohestani, Ahmed Hussain, Antonio A Moreira, and Anja K Skrivervik. Diversity gain influenced by polarization and spatial diversity techniques in ultrawideband. *IEEE Access*, 3:281–286, 2015.
- [29] Qian Xu, Yi Huang, Xu Zhu, Saqer S Alja’afreh, Lei Xing, and Zhihao Tian. Diversity gain measurement in a reverberation chamber without extra antennas. *IEEE Antennas and Wireless Propagation Letters*, 14:1666–1669, 2015.
- [30] CM Jack Wang, Kate A Remley, Ansgar T Kirk, Ryan J Pirkel, Christopher L Holloway, Dylan F Williams, and Paul D Hale. Parameter estimation and uncertainty evaluation in a low rician k-factor reverberation-chamber environment. *IEEE Transactions on Electromagnetic Compatibility*, 56(5):1002–1012, 2014.
- [31] Larry J Greenstein, David G Michelson, and Vinko Erceg. Moment-method estimation of the rician k-factor. *IEEE Communications Letters*, 3(6):175–176, 1999.
- [32] Ali Abdi, Cihan Tepedelenlioglu, Mostafa Kaveh, and Georgios Giannakis. On the estimation of the k parameter for the rice fading distribution. *IEEE Communications letters*, 5(3):92–94, 2001.
- [33] Xiaoming Chen, Bjarni Þór Einarsson, and Per-Simon Kildal. Improved mimo throughput with inverse power allocation—study using usrp measurement in reverberation chamber. *IEEE Antennas and Wireless Propagation Letters*, 13:1494–1496, 2014.
- [34] Per-Simon Kildal, Ahmed Hussain, Xiaoming Chen, Charlie Orlenius, Anton Skarbratt, John Asberg, Tommy Svensson, and Thomas Eriksson. Threshold

receiver model for throughput of wireless devices with mimo and frequency diversity measured in reverberation chamber. *IEEE Antennas and Wireless Propagation Letters*, 10:1201–1204, 2011.

- [35] Xiaoming Chen, Per-Simon Kildal, and Jan Carlsson. Estimation of average rician k-factor in reverberation chamber. In *International Symposium on Antennas and Propagation*, 2011.

Chapter 6

Near-field MIMO in Metal Enclosure

6.1 MIMO Channel Sounding

Narrowband MIMO channel can be measured using PN-sequences modulated by BPSK, QPSK, and MSK signals. Authors [1] have performed narrowband MIMO channel measurement using the PN sequence modulated by MSK signals. BPSK modulated random sequences were used for channel sounding. Similar channel sounding has been proposed by NIST for channel sounding of the smart industry environment [2]. Un-tethered clock synchronization is provided using a rubidium clock for both transmitter and receiver. The authors have used double-ridged horn antenna and omnidirectional dipole for channel impulse response measurements at sub-6 GHz frequencies. The important design parameters of channel sounder are delay resolution, maximum detectable distance, and the dynamic range. Authors in [3] have presented large scale channel sounding system with RFNoC using SDR platforms. Authors have used open source testbed consisted of 20 x 20 grid USRPs. This testbed highlights critical issues that may arise in high order MIMO channel sounder using USRPs such as high sampling requirement at the host side and the dealing the sampling rate issue with signal processing using RF network on chip (RFNoC) on FPGAs of the USRPs. Critical signal processing functions are offloaded to FPGAs of USRPs to avoid any underflow or overflow for real-time communication system testing.

6.2 Channel Impulse Response Measurement

The channel impulse response of the radio channel environment is measured by using random channel sounding sequences with good auto-correlation and cross-correlation properties. Some of the notable channel sounding sequences are m-sequences which are generated using linear-feedback shift register (LFSR), Gold code sequences which are generated by XORing the output of two PN codes with different generator polynomials. The most popular sounding sequence is a Zadoff-Chu sounding sequence. Zadoff-Chu sounding sequence is popular because of its excellent cross-correlation and constant amplitude zero autocorrelation (CAZAC) properties. Zadoff-Chu sequences were used in 4G LTE standard [4]. Constant amplitude zero autocorrelation (CAZAC) Zadoff-Chu sequences are used for channel estimation and time synchronization [5]. This code has unique properties which include zero cyclic autocorrelation, low cross-correlation, and constant magnitude before and after FFT and IFFT. Recently Zadoff-Chu codes are considered for coarse timing synchronization by utilizing their symmetric properties [6]. This code has unique properties which include zero cyclic autocorrelation, low cross-correlation, and constant magnitude before and after FFT and IFFT. Recently Zadoff-Chu codes are considered for coarse timing synchronization by utilizing their symmetric properties [6]. Its applications are diverse Zadoff-Chu sequences are used for time and frequency synchronization in OFDM downlink transmission [7], ultrasound ranging [8], ultrasonic underwater communication [9], aerial wireless communication [10], random access (RA) in machine-type communication (MTC) as enhanced Physical RACH (PRACH) preamble detector [11] and also excellent choice for MIMO channel sounding and all latest wireless communication systems [12][13][14]. BPSK modulated random sequences were used for channel sounding. Similar channel sounding has been proposed by NIST for channel sounding of the smart industry environment [2]. Un-tethered clock synchronization is provided using a rubidium clock for both transmitter and receiver. MIMO channel sounding can also be performed by

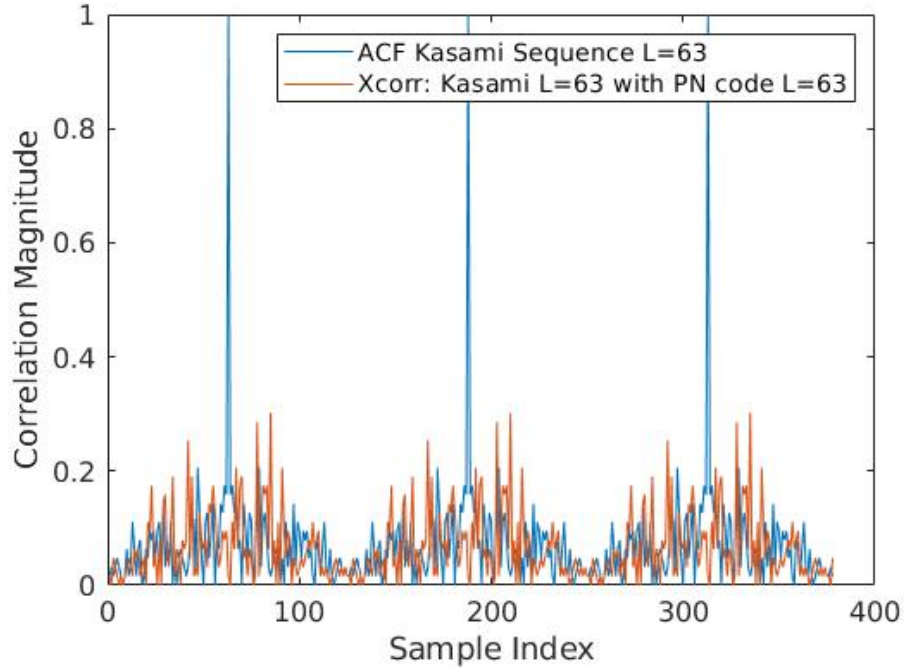


Figure 6.1: Autocorrelation of Kasami sequence of length $L=63$ and cross correlation of Kasami sequence with PN sequence of length $L=63$.

transmitting two different code sequences with good cross-correlation properties. Figure 6.1 shows auto-correlation of Kasami sequence of length $L=63$ and cross correlation of Kasami sequence with PN sequence of length $L=63$. Autocorrelation of Kasami sequence results in high peaks as its cross correlation with PN sequence reflects the suitability of using PN-sequence and Kasami sequence for MIMO channel sounding. Figure 6.2 shows auto-correlation of PN sequence of length $L=63$ and cross correlation of PN sequence with Kasami sequence. It's shown that the two sequences have good cross correlation properties, hence are suitable for MIMO channel sounding. All the elements of MIMO channel can be measured by cross-correlating Tx sequences at each channel of the USRP. However, the time resolution is limited by the sampling rate of the USRP.

6.3 2x2 Near-field MIMO Channel Matrix $\mathbf{H}$

Time-domain transfer function of MIMO channel matrix can be obtained using orthogonal sequences such as PN and Zadoff-Chu sequences as discussed in pre-

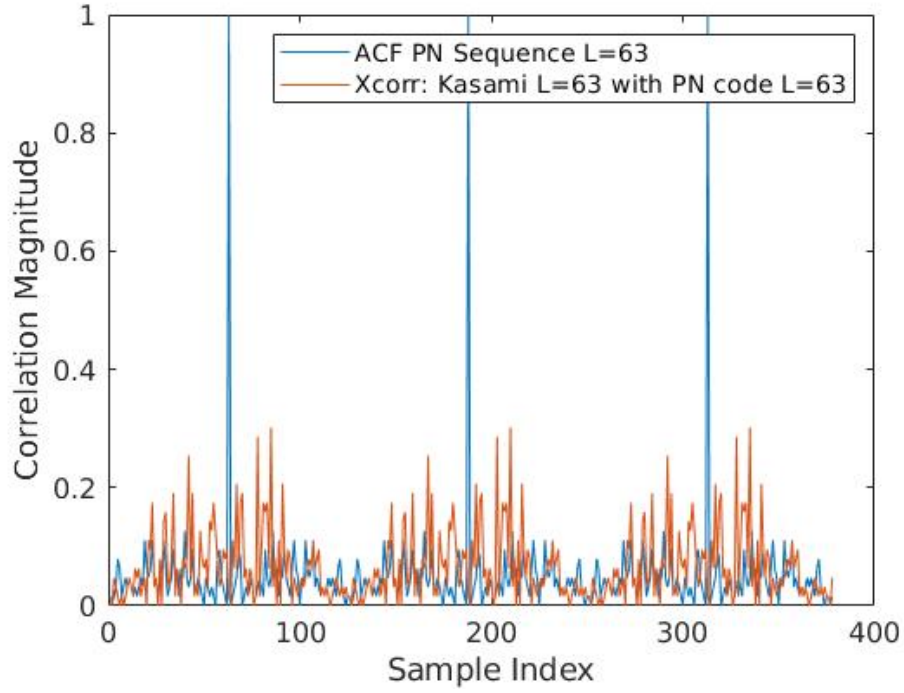


Figure 6.2: Auto-correlation of PN sequence of length $L=63$ and cross correlation of PN sequence of length $L=63$ with Kasami sequence of length $L=63$.

vious section. Baseband IQ samples of orthogonal signals are sent to both RF boards of USRP X310 for transmission over two antennas for 2x2 MIMO channel sounding and the received baseband IQ samples are cross-correlated with original copies of orthogonal sequences such as PN sequence and Kasami sequence. Baseband Simulink model to generate orthogonal sequences is shown in Fig 6.3 and Fig.6.4. The orthogonal sequences PN sequence and Kasami sequences are generated in the baseband model. The data from two basedband blocks is shaped into two column vector. The data in each columns of the reshaped matrix is fed to SDRuTransmitter for upconversion and transmission over the channel. The USRPs are configured to transmit and receive using two channels. At the receiver side as shown in Fig.6.4 the down-converted complex baseband signals are divided into two parallel streams. After the AGC, coarse frequency synchronization and fine synchronization blocks two orthogonal sequences are cross-correlated to get the individual MIMO channel matrix. The main drawback of this method of extracting MIMO channels is limited sample rates of the USRPs and even if we

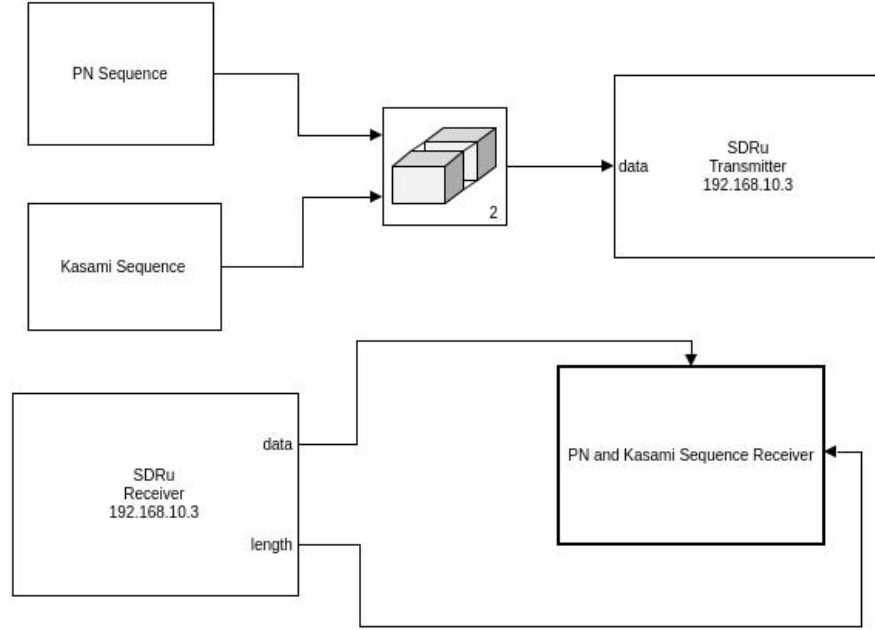


Figure 6.3: Top-level diagram of Tx and Rx baseband Simulink models showing orthogonal sequences configuration for 2x2 MIMO channel sounding.

exploit the maximum sample rates of the USRP there is another bottleneck of samples rates introduced by the USB or Ethernet cable connecting USRP with the host PC. Therefore, it is a good practice to obtain IQ sample at high rates and post-process the received IQ samples to avert the computational intensive operations in real-time.

Figure 6.5 shows one snapshot of near-field channel impulse response of each element of near-field channel matrix $\mathbf{H}$ at operating frequency of 5.8 GHz. The impulse response is obtained using PN sequence of length 1023 and Kasami sequence of length 1023. Similarly, 20 snapshots of CIR of near-field channel matrix are shown in Fig.6.6. Figure 6.7 shows power delay profile of the near-field channel matrix. In each channel element PDF there are low gain values which suggest that there was data loss at the start and at the end of measurement process. The data loss could have been underflow/overflow problem and the random start up times of the two USRPs.

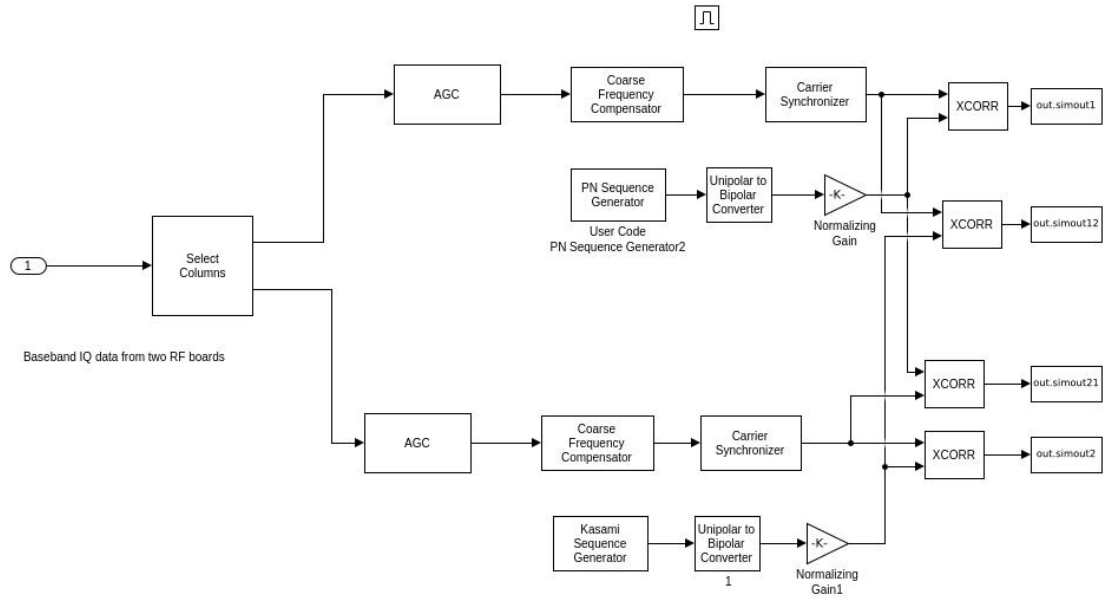


Figure 6.4: Baseband receiver subsystem with per channel cross-correlation operations for 2x2 MIMO channel estimation. The receiver subsystem can be reconfigured in accordance with source of external timing and frequency synchronization. The AGC, coarse frequency compensation and carrier synchronization blocks can be removed when the source of timing and frequency synchronization is set to external.

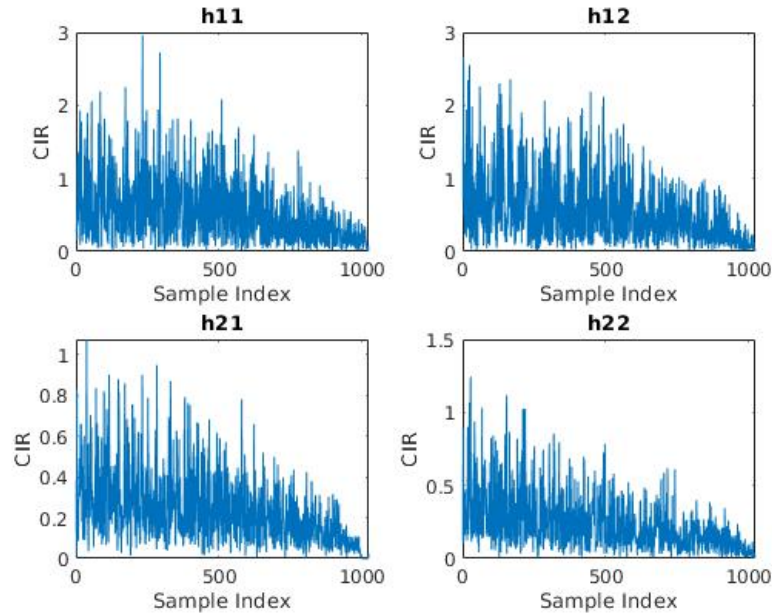


Figure 6.5: 2x2 MIMO Channel impulse response using PN sequence of $L=1023$ and Kasami sequence of $L=1023$ operating frequency=5.8 GHz, Tx/Rx gains=[10 10] dB. There is no power imbalance.

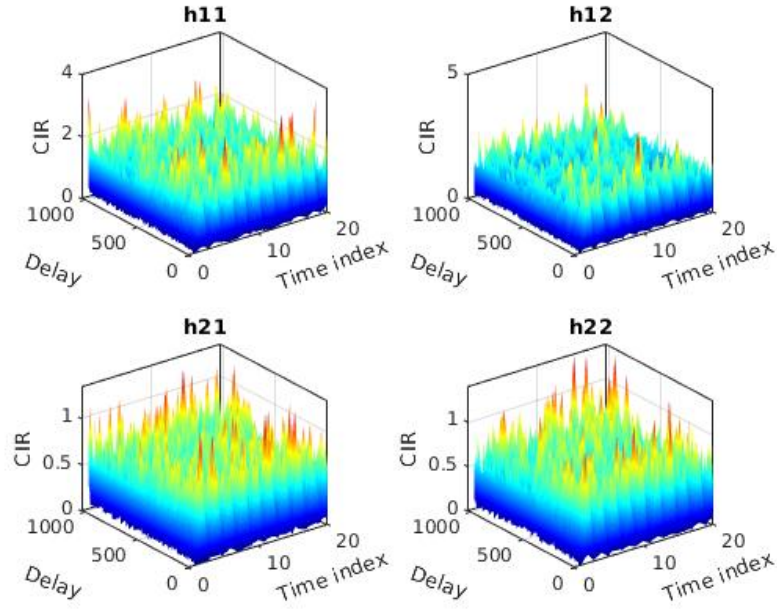


Figure 6.6: 2x2 MIMO Channel impulse response using PN sequence of $L=1023$ and Kasami sequence of $L=1023$ operating frequency=5.8 GHz, Tx/Rx gains=[10 10] dB. There is no power imbalance.

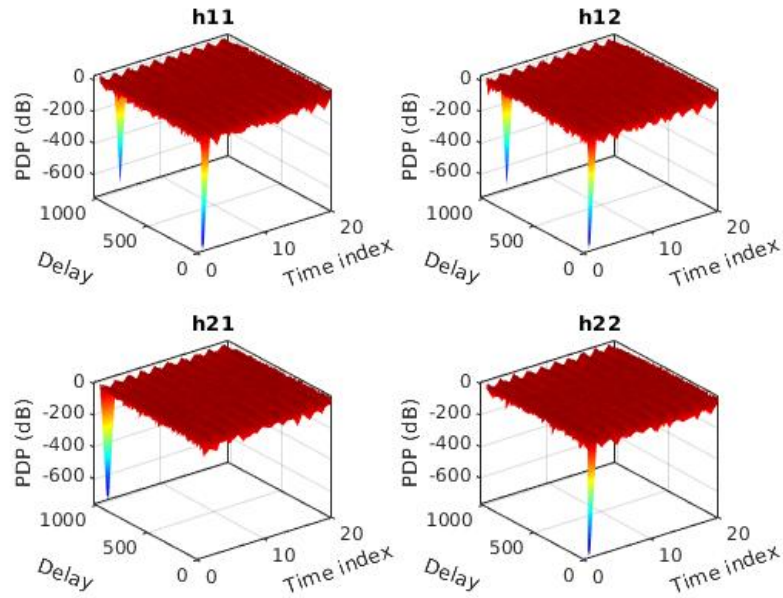


Figure 6.7: 2x2 MIMO power delay profile using PN sequence of $L=1023$ and Kasami sequence of $L=1023$ operating frequency=5.8 GHz, Tx/Rx gains=[10 10] dB. There is no power imbalance.

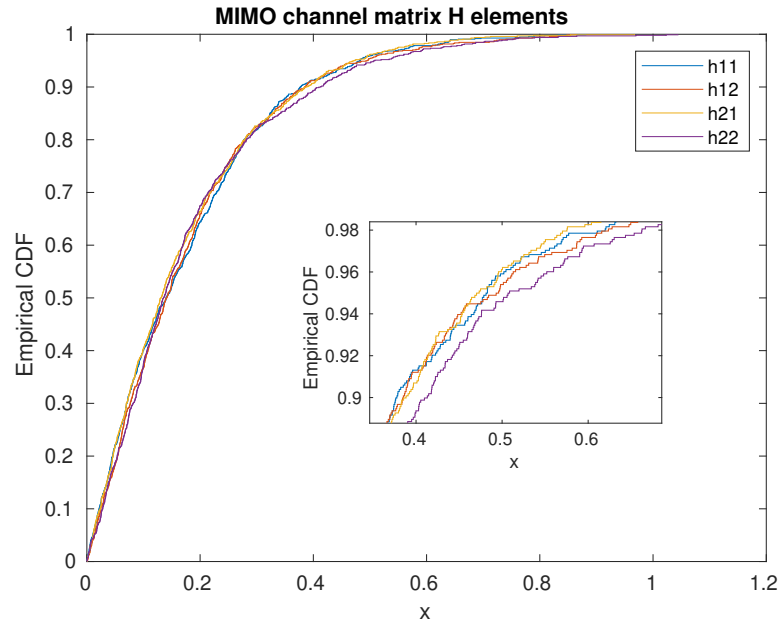


Figure 6.8: Empirical CDF of the 2x2 MIMO channel matrix elements at 5.8 GHz.

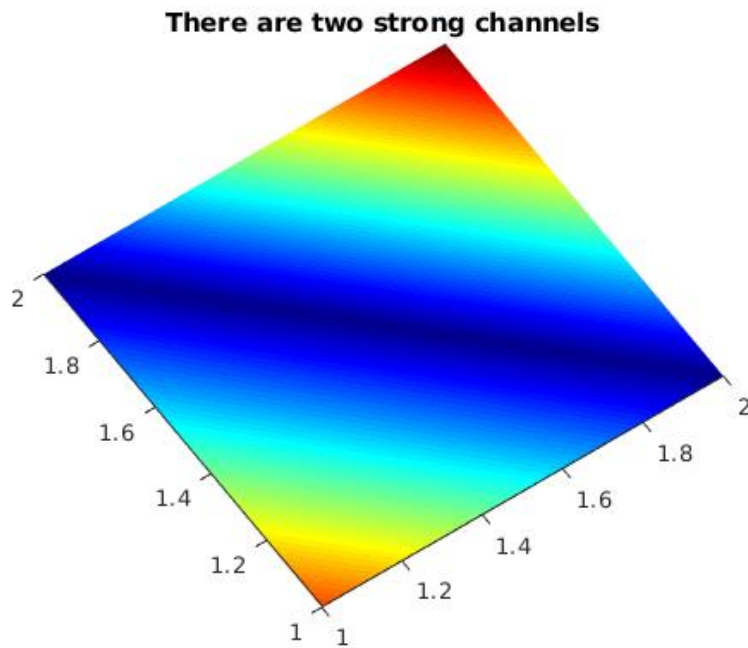


Figure 6.9: Two strong orthogonal channels are achieved. However channel 1 is stronger than channel 2.

6.4 MIMO System Model

Input-output relationship of narrowband MIMO flat fading channel with N_T transmit and N_R receive antennas is given as:

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \mathbf{n} \quad (6.1)$$

Where $\mathbf{H}$ is MIMO channel matrix whose elements are complex gains between N_T transmit and N_R receive antennas, $\mathbf{y}$ is received signal vector, $\mathbf{x}$ is transmitted signal vector and $\mathbf{n}$ is additive white Gaussian noise vector. MIMO channel is affected by transmit and receive antennas. Therefore, MIMO channel with transmit and receive antenna effects [15][16]:

$$\mathbf{H} = \mathbf{R}_R^{1/2} \mathbf{H}_w \mathbf{R}_T^{1/2} \quad (6.2)$$

Where $\mathbf{R}_T^{1/2}$ and $\mathbf{R}_R^{1/2}$ are transmit and receive co-variance matrices. $\mathbf{R}_T^{1/2}$ is Hermitian square root of the transmit covariance matrix. $\mathbf{H}_w$ is spatially MIMO channel matrix whose elements are complex Gaussian numbers with zero mean and unit variance.

6.5 Rank and Condition Number

The rank of the channel matrix $\mathbf{H}$ is the number of non-zero singular values. It indicates the number of decodable data streams that can be spatially multiplexed. For proper MIMO spatial multiplexing the rank of MIMO system should be at least 2. Rank of MIMO system is given as $\min(N_T, N_R)$ which also corresponds to number of decodable parallel pipes without a cross-talk. There is no MIMO capacity gain for rank-deficient MIMO systems. Condition number of MIMO channel matrix is defined as ratio of strongest singular value σ_{max} to the weakest singular value σ_{min} :

$$K(\mathbf{H}) = \frac{\sigma_{max}}{\sigma_{min}} \geq 1 \quad (6.3)$$

Condition number in logarithmic form can be rewritten as:

$$K(\mathbf{H})_{dB} = 20 \log_{10}(K(\mathbf{H})) = 20 \log_{10} \frac{\sigma_{max}}{\sigma_{min}} \geq 0 \text{ dB} \quad (6.4)$$

Condition number $K(\mathbf{H}) \leq 10 \text{ dB}$ is considered suitable for MIMO spatial multiplexing. MIMO capacity is a function of condition number. MIMO channel capacity degrades as condition number increases [17]. In terms of Eigenvalues of channel correlation matrix $\mathbf{H}^H \mathbf{H}$:

$$K(\mathbf{H})_{dB} = 10 \log_{10} \frac{\lambda_{max}}{\lambda_{min}} \geq 0 \text{ dB} \quad (6.5)$$

Where singular values σ_i of channel matrix $\mathbf{H}$ are related to Eigenvalues λ_i as $\sigma_i^2 = \lambda_i$.

6.6 Near-field Capacity with Polarization Diversity

MIMO system with polarization diversity is often used when there are form-factor or design limitations and wireless chip-to-chip communication channel is not an exception. However, using polarization diversity there is MIMO capacity loss. MIMO channel measurement of co-polar and cross-polar subchannels is conducted under LOS and NLOS channel conditions. Parameters under discussions are Rician K-factor, XPD factor, and correlation of subchannels. Subchannel correlation is reduced by using polarization diversity in compact form-factor devices. In the MIMO system with polarization diversity, the subchannels of the MIMO channel matrix are not identically distributed [18][19]. Utilization of polarization diversity with MIMO hybrid array have been reported in [20][21][22]. However,

all these measurements are performed in an indoor radio propagation channel environment. System model for narrowband MIMO channel using N_T transmit and N_R receive antennas:

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \mathbf{n} \quad (6.6)$$

Where $\mathbf{y}$ is complex received signal vector, $\mathbf{x}$ is transmitted complex signal vector and $\mathbf{n}$ is additive complex Gaussian noise vector with autocorrelation $\mathbf{R}_{nn} = \sigma^2 \mathbf{I}_{N_R}$

6.6.1 Cross-polarized and co-polarized channels

Near-field MIMO capacity can be assessed in presence of transmit and receive polarization diversity i.e. in terms of cross-polarized or co-polarized antenna elements. Polarization diversity MIMO channel can be measured by measuring cross polarization and co-polarization complex channel matrix gains in presence of losses or in highly reflective multi-path rich scattering environment in metal enclosure.

6.6.2 Spatial MIMO

Spatial MIMO is a normal MIMO multiplexing condition two co-polarized antenna array elements are used at the transmitter as well at the receiver. Therefore, its parameters are same it has been discussed in the previous chapters.

6.6.3 Cross-polar Discrimination (XPD) factor

Transmitted signal undergoes reflection, diffraction and scattering which results in the coupling of one form of polarization into another. For example, the horizontally polarized signal is received as a vertically polarized signal because of depolarization in the multipath fading channel. This phenomenon is known as depolarization which is a characteristic property of radio channel[23]. Because of this effect power received in subchannels of the MIMO channel matrix is different.

The channel matrix of 2x2 MIMO with polarization diversity is given as:

$$\mathbf{H} = \begin{bmatrix} h^{VV} & h^{VH} \\ h^{HV} & h^{HH} \end{bmatrix} \quad (6.7)$$

Cross-polar discrimination is a measure of depolarization in the wireless channel.

$$X_V = \frac{E\{|h^{VV}|^2\}}{E\{|h^{HV}|^2\}}, X_H = \frac{E\{|h^{HH}|^2\}}{E\{|h^{VH}|^2\}} \quad (6.8)$$

Depolarization depends on both environment as well as radiation pattern of V and H elements employed for the channel measurements [20]. Hence, $X_V \not\approx X_H$ in contrast to popular assumption in the literature.

	X_V	X_H
LOS	12 dB	22 dB
NLOS	10 dB	21 dB

Table 6.1: Measured XPD Values.

Because of Brewster angle phenomenon in horizontally polarized waves [24] power in vertically polarized co-channels is higher than horizontally polarized co-channels i.e. $E\{|h^{VV}|^2\} > E\{|h^{HH}|^2\}$.

6.6.4 LOS/NLOS MIMO capacity

Path loss difference comparison of LOS and NLOS MIMO capacity. MIMO channel capacity measurement in LOS and NLOS is important. It's known in the literature that MIMO capacity is higher in NLOS because of rich scattering which reduces the amount of correlation present at the transmitter and receiver. There are also some references where it's shown that MIMO capacity in LOS is higher than the MIMO capacity in NLOS. The potential reasons for higher LOS MIMO capacity is presence of dominant component which increases the SNR of the MIMO communication link.

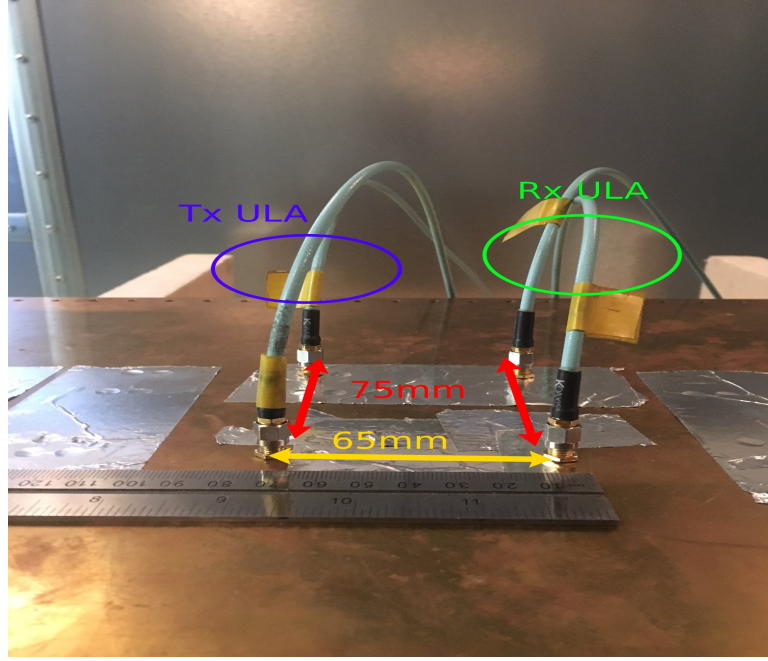


Figure 6.10: 2x2 MIMO configuration. Tx and Rx are separated by a distance of 65mm and inter-element distance at both sides is 75 mm. Tx and Rx USRP gains were set to 0 dB with no power imbalance.

6.7 MIMO Measurements in Metal Enclosure

In this section, the MIMO channel is measured in a metal enclosure with Tx and Rx antenna separations. All antennas used are monopoles operating at 5.6 GHz. All antenna elements used at the transmitter and receiver side are linearly polarized and isotropic radiation patterns. Depending on the orientation of antenna elements in the stirred cavity their polarization can be classified as horizontal or vertical.

Measurement setup is shown in the following Fig.6.10 where the transmitter and receiver are separated by distance $D = 65\text{ mm}$ and inter-element distance at the transmitter $d_t = 65\text{ mm}$ and the inter-element distance at the receive $d_r = 65\text{ mm}$.

After performing MIMO channel measurements at 5.6 GHz with equal Tx and Rx inter-element spacing, the Rx inter-element was reduced from 75 mm to 45 mm to check if there is difference in channel distribution or correlation matrix. The measurement scenario for 45 mm inter-element distance at Rx side is shown

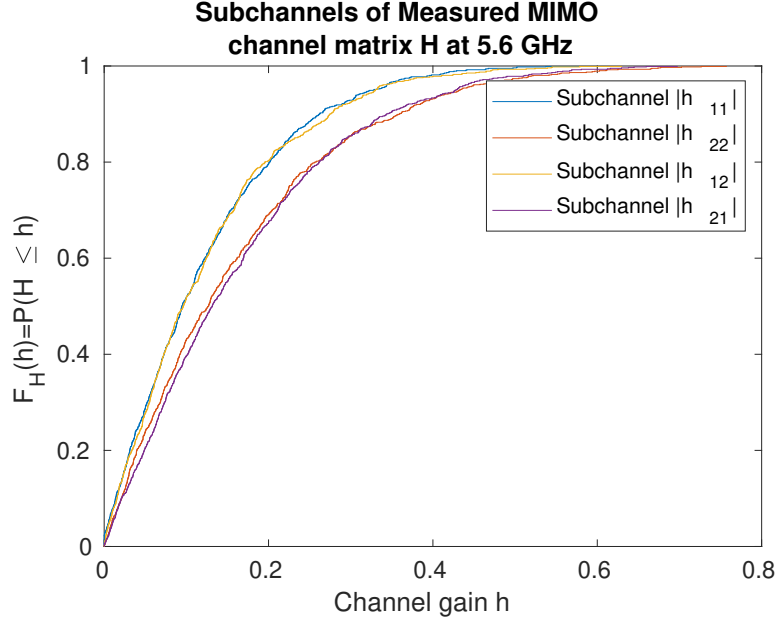


Figure 6.11: CDF of channel elements of measured MIMO channel matrix of open-loop MIMO system under equal power allocation strategy at operating frequency of 5.6 GHz at receive inter-element distance of 75 mm.

in Fig.6.16. Figure 6.11 shows the CDF of the each element of the MIMO channel matrix when the Tx and Rx gains of the USRPs were set to 4 dB. Similarly, Fig.6.12 shows local mean of the measured channel impulse response of each element of MIMO channel matrix. It can be seen that there is some form clustering. The clustering of the impulse response can be resolved by increasing the sample rates of the USRP testbed. The clustering of the channel can be modelled using Saleh-Valenzuela (SV) model. In SV model each cluster is modelled with its inter-arrival time of the rays in the cluster and decay time of the cluster. Usually, each cluster follow exponential decay of the ray energy in each cluster and Poisson distributed ray inter-arrival times and cluster inter-arrival times.

Fig.6.13 and Fig.6.14 show PDF and CDF of each elements of the measured MIMO channel matrix. distribution fitting of the measured channel elements has been performed with well-know statistical distributions i.e. Rician, Rayleigh, lognormal and Weibull distributions. It can be seen that the best distribution that fits the measured channel is Weibull distribution. Weibull is generic measurement based distribution and the Rayleigh fading is special case of the Weibull distribution.

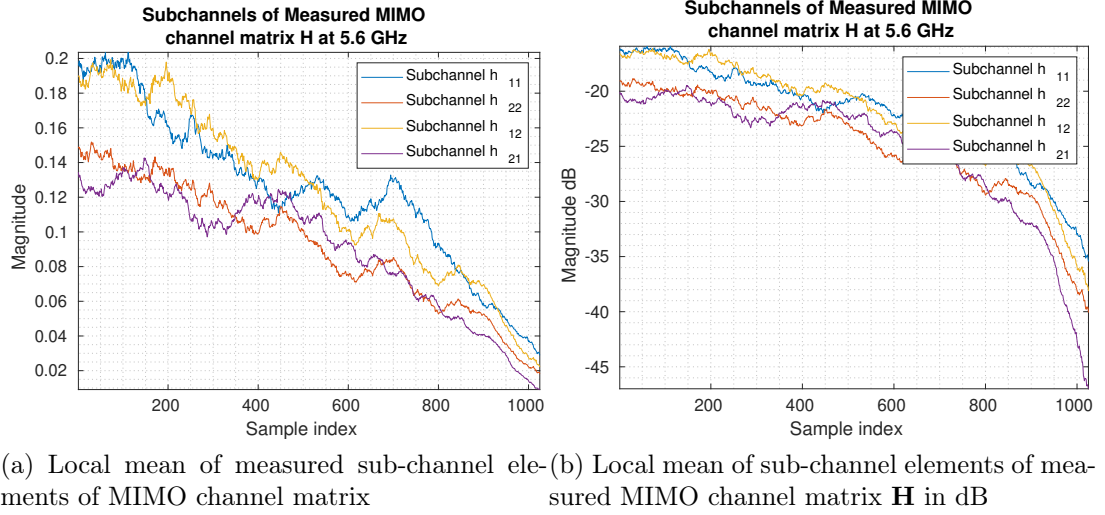


Figure 6.12: Local mean of sub-channel elements of measured MIMO channel matrix $\mathbf{H}$ of open-loop MIMO system under equal power allocation strategy at frequency of 5.6 GHz and Tx and Rx USRP gain of 4 dB respectively.

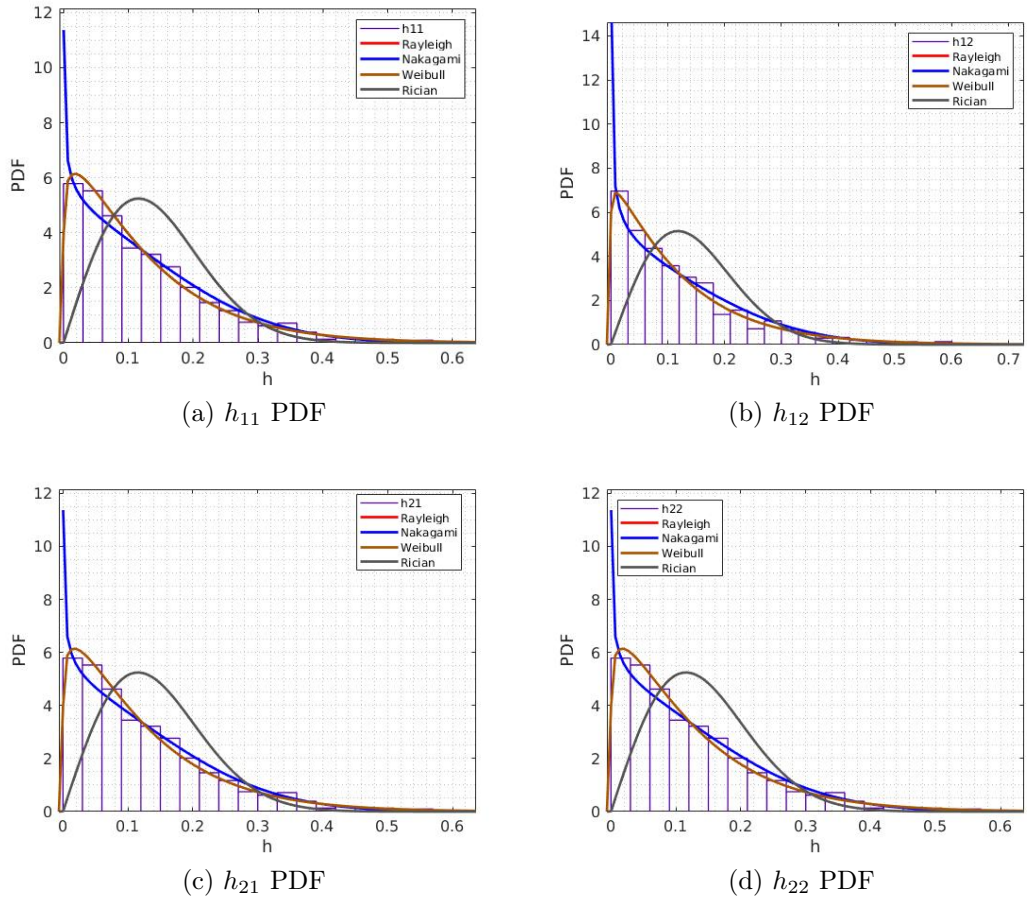


Figure 6.13: PDF distribution of sub-channels of measured MIMO channel matrix $\mathbf{H}$ at frequency of 5.6 GHz and inter-element distance of 75 mm. Empirical results of channel elements of MIMO channel matrix show an excellent fit with Weibull distribution.

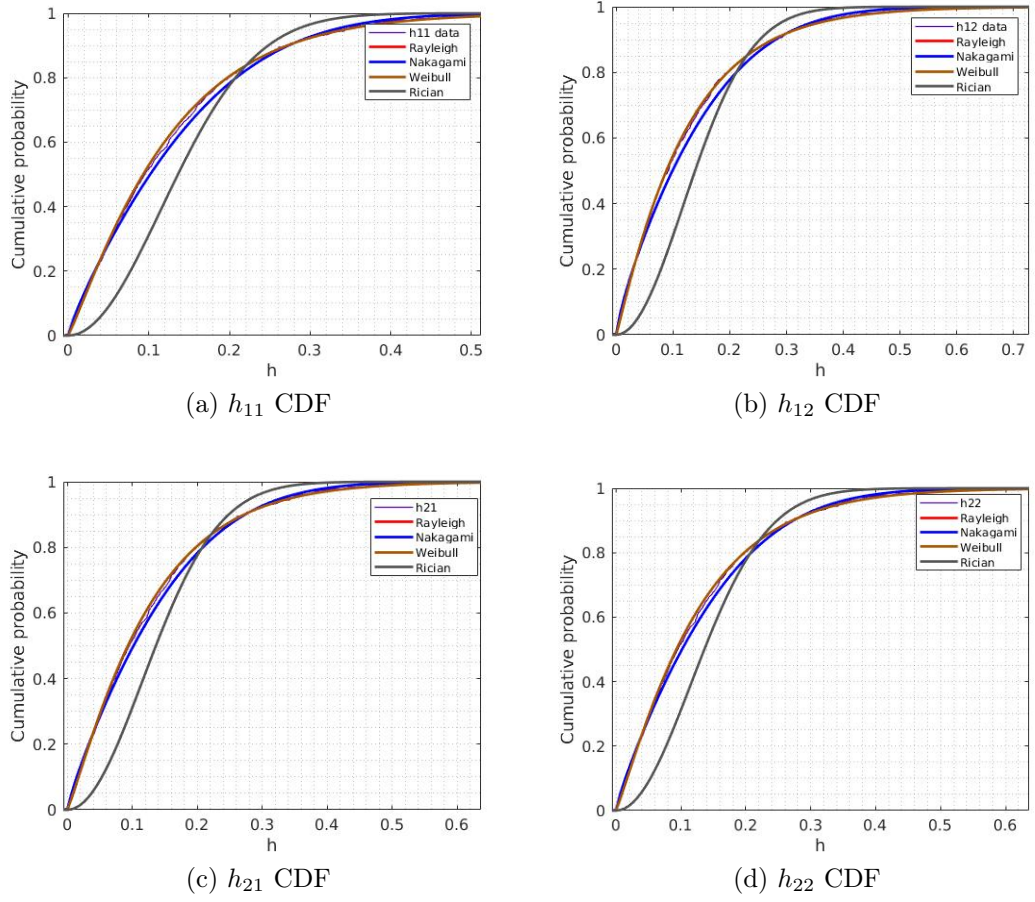
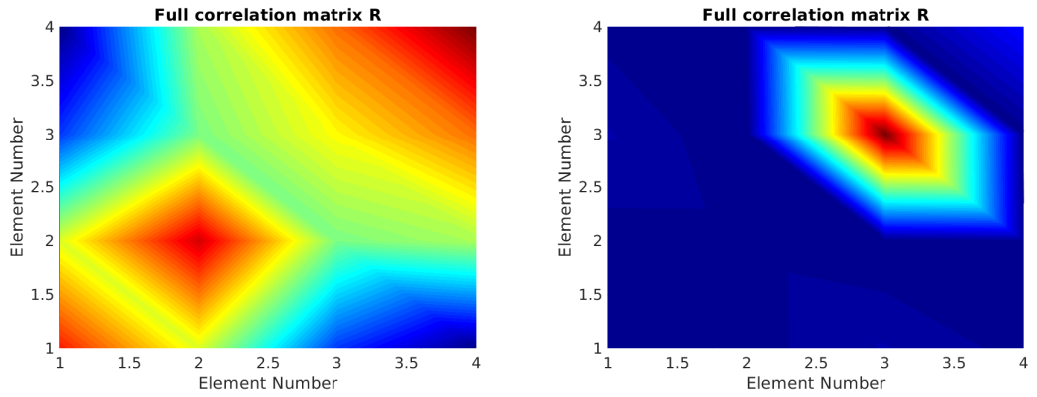


Figure 6.14: CDF distribution of sub-channels of measured MIMO channel matrix $\mathbf{H}$ at frequency of 5.6 GHz and inter-element distance of 75 mm. Empirical results of channel elements of MIMO channel matrix show an excellent fit with Weibull distribution.



(a) Full correlation matrix $\mathbf{R}_{\mathbf{H}}$ of measured open-loop MIMO system (b) Full correlation matrix $\mathbf{R}_{\mathbf{H}}$ of measured open-loop MIMO system.

Figure 6.15: Full-correlation matrix $\mathbf{R}_{\mathbf{H}}$ of measured open-loop MIMO channel matrix in metal enclosure at operating frequency of 5.6 GHz. (a) MIMO channel elements when no averaging is performed (b) MIMO channel elements are average of 100 CIRs under equal power allocation strategy.

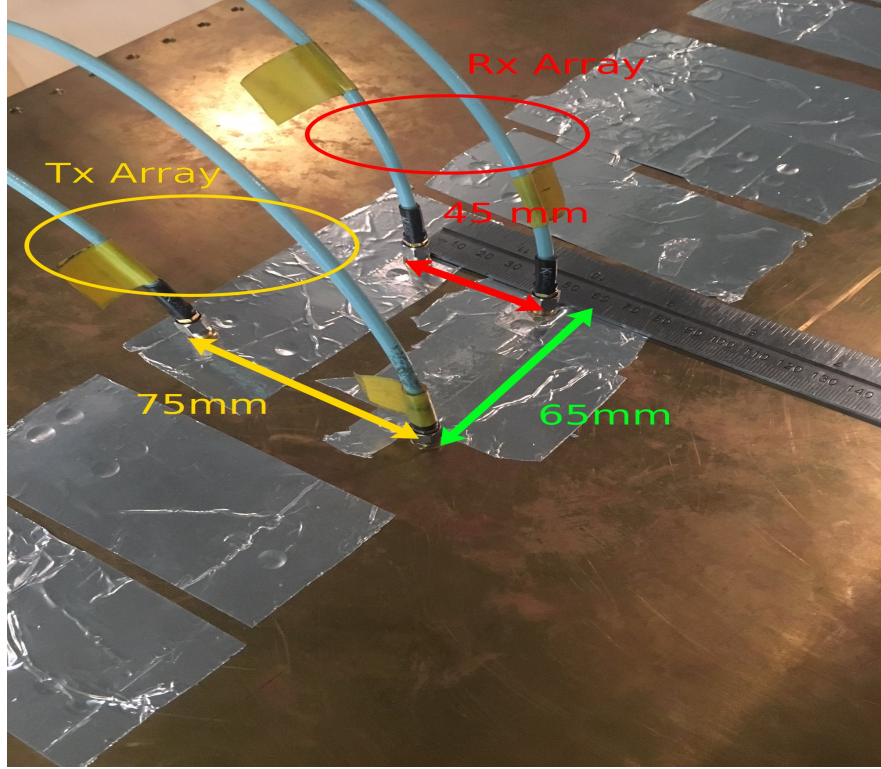
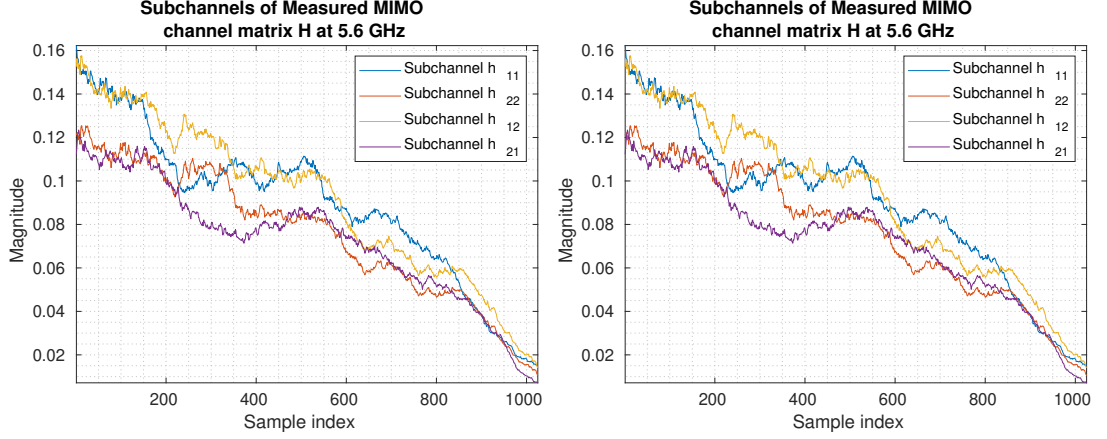


Figure 6.16: 2x2 MIMO configuration. Tx and Rx are separated by a distance of 65 mm and inter-element distance at Tx is 75 mm and inter-element distance for Rx is reduced from 75 mm to 45 mm. Tx and Rx USRP gains were set to 4 dB with no power imbalance.

Figure 6.16 shows 2x2 MIMO measurement setup where the receive inter-element distance is reduced from 75 mm to 45 mm. Tx and Rx antennas are separated by a distance of 65 mm. Figure 6.17 shows measured impulse response of the each element of the MIMO channel matrix. It can be seen that there is no effect on the clustering despite the Rx inter-element distance is reduced. It clear that the clusters are coming from mode-stirrer pedal shapes where each pedal in the mode-stirrer creates unique reflections. In the measurement setup mode-stirrer was kept stationary. Clustering shape and the statistical properties can be studied using SV model.

Fig.6.18 and Fig.6.19 show distribution PDFs and CDFs of the measured MIMO channel matrix elements and the fitting of the MIMO channel elements in metal enclosure at RF frequency of 5.6 GHz and at Tx and Rx inter-element distance of 75 mm and 45 mm. Distribution fitting of the measured MIMO channel elements using well-known statistical distributions is performed. All four distri-



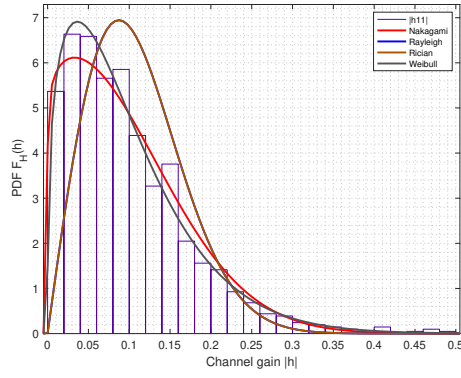
(a) Local mean of measured sub-channel elements of MIMO channel matrix $\mathbf{H}$ (b) Local mean of sub-channel elements of measured MIMO channel matrix $\mathbf{H}$

Figure 6.17: Local mean of of sub-channel elements of measured MIMO channel matrix $\mathbf{H}$ of open-loop MIMO system under equal power allocation strategy at frequency of 5.6 GHz and Tx and Rx USRP gain of 4 dB respectively. Tx inter-element distance of 75 mm, rx inter-element distance of 45 mm and distance between Tx and Rx is kept 65 mm.

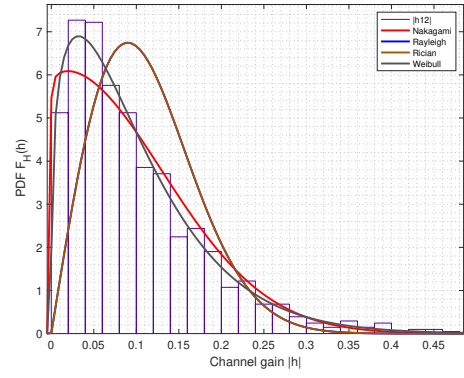
butions seems to be close apart from Rician distribution. Rayleigh and Weibull distribution fits the measured MIMO channel elements. It is obvious since the Weibull distribution is special type of Rayleigh distribution with shape factor of 2. It can be seen that similar to previous scenario the best distribution that fits the PDF and CDFs of MIMO channel elements is Weibull distribution fits the measured MIMO channel matrix. Probability plot of the measured channel elements shows clear picture of the distribution fitting. Fig.6.20 shows probability plot of the fitted distribution where it is shown that Rayleigh and Weibull distribution fits MIMO sub-channel elements.

6.8 Conclusion

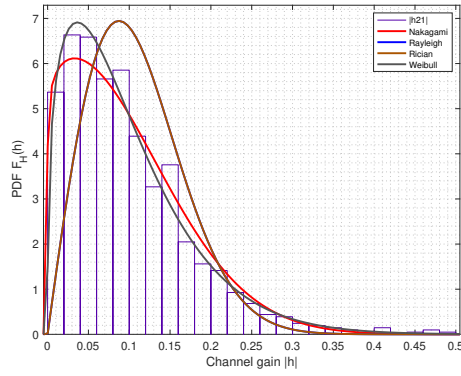
MIMO channel measurement has been performed in the metal enclosure using SDR based testbed. Orthogonal PN sequences have been transmitted from two channels of USRP and with the help of cross-correlation at each channel of the USRP with corresponding PN sequence, the channel elements of MIMO channel matrix are measured when the distance between Tx and Rx elements is set to



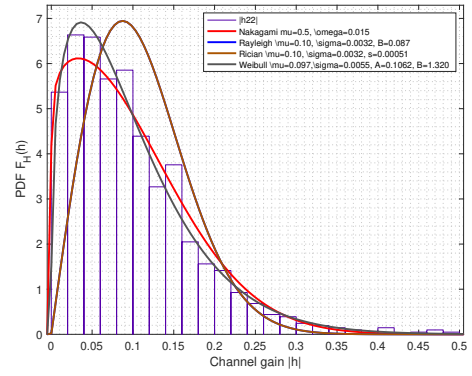
(a) h_{11} PDF



(b) h_{12} PDF

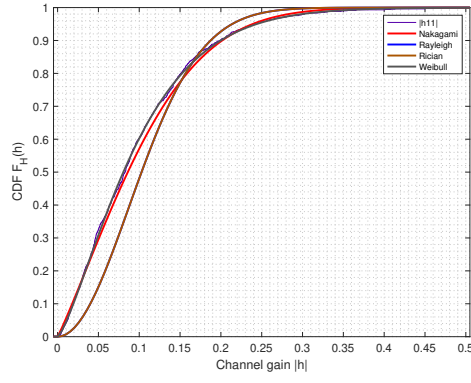


(c) h_{21} PDF

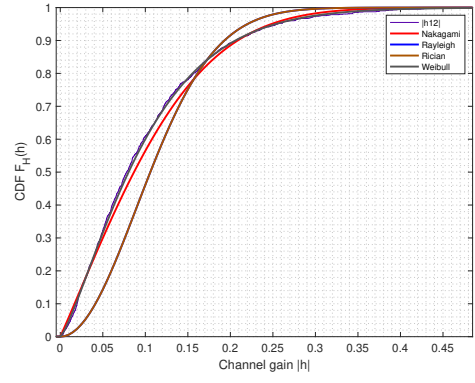


(d) h_{22} PDF

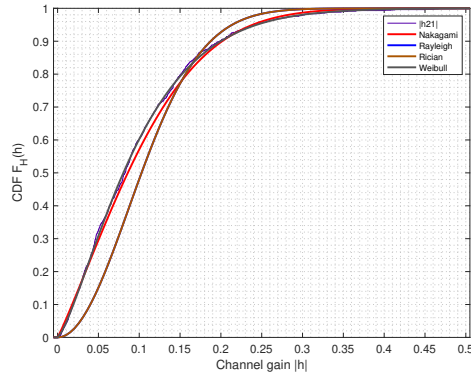
Figure 6.18: PDF distribution of sub-channels of measured MIMO channel matrix $\mathbf{H}$ at frequency of 5.6 GHz and inter-element distance of 75 mm. Empirical results of channel elements of MIMO channel matrix show an excellent fit with Weibull distribution.



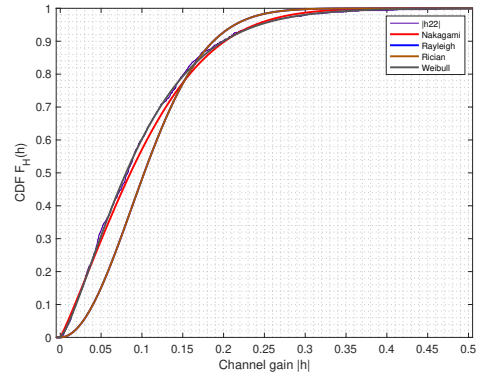
(a) h_{11} CDF



(b) h_{12} CDF



(c) h_{21} CDF



(d) h_{22} CDF

Figure 6.19: CDF distribution of sub-channels of measured MIMO channel matrix $\mathbf{H}$ at frequency of 5.6 GHz and inter-element distance of 75 mm. Empirical results of channel elements of MIMO channel matrix show an excellent fit with Weibull distribution.

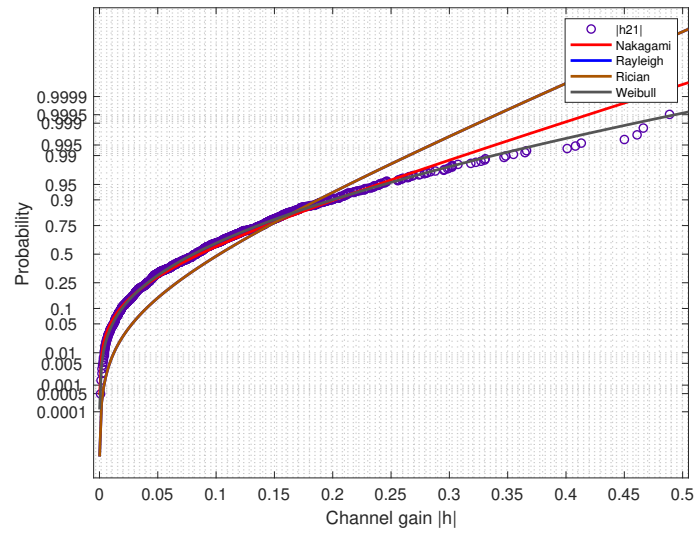
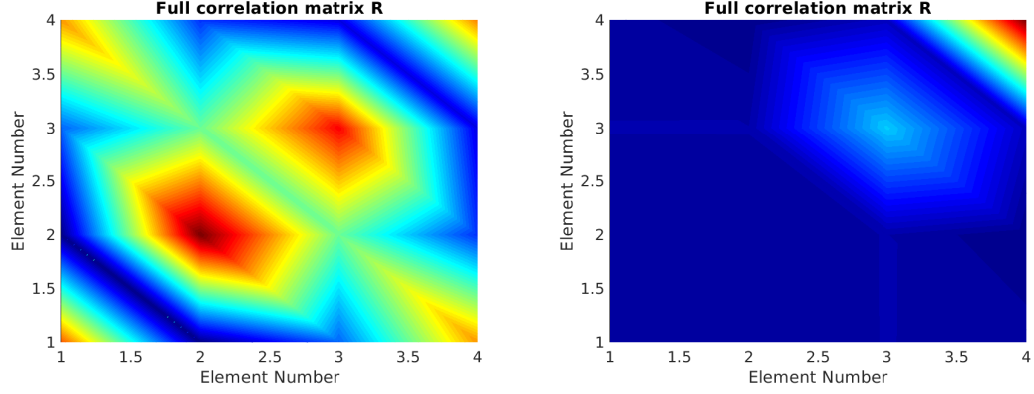


Figure 6.20: Probability plot of h_{21} channel of measured MIMO channel matrix.



(a) Full correlation matrix $\mathbf{R}_H$ of measured open-loop MIMO system (b) Full correlation matrix $\mathbf{R}_H$ of measured open-loop MIMO system.

Figure 6.21: Full-correlation matrix $\mathbf{R}_H$ of measured open-loop MIMO channel matrix in metal enclosure at operating frequency of 5.6 GHz. Tx and Rx inter-element distances are 75 mm and 45 mm respectively. Tx and Rx separation was kept to 65 mm (a) MIMO channel elements when no averaging is performed (b) MIMO channel elements are average of 100 CIRs under equal power allocation strategy.

65 mm. For fixed distance between Tx and Rx, the Rx inter-element distance is varied from 75 mm to 45 mm. In both the case, clustering of the channel elements can be observed which is evident because of the stirrer shape. The measurements were performed at single stirrer position for Tx and Rx gains of USRPs set to 4 dB for each channel. Additionally, the measured channel elements are fitted to well-know statistical distributions such as Rayleigh, Rician, Nakagami, Lognormal and Weibull distribution. It was observed that there was clustering for both the Rx inter-element spacing. The best distribution that fits the measured MIMO channel elements was Rayleigh and Weibull distribution. Weibull distribution is a special case of Rayleigh distribution for shape factor of 2.

Bibliography

- [1] Jean-Philippe Kermoal, Laurent Schumacher, Klaus I Pedersen, Preben E Mogensen, and Frank Frederiksen. A stochastic mimo radio channel model with experimental validation. *IEEE Journal on selected areas in Communications*, 20(6):1211–1226, 2002.
- [2] Jeanne T Quimby, Richard Candell, Catherine A Remley, David R Novotny, Joseph Diener, Peter B Papazian, Alexandra Curtin, and Galen H Koepke. Nist channel sounder overview and channel measurements in manufacturing facilities. Technical report, 2017.
- [3] Bhargav Gokalgandhi, Prasanthi Maddala, and Ivan Seskar. Implementation of fgpa based channel sounder for large scale antenna systems using rfno on usrp platform. In *Proceedings of the GNU Radio Conference*, volume 2, pages 9–9, 2017.
- [4] Mashud Hyder and Kaushik Mahata. Zadoff–chu sequence design for random access initial uplink synchronization in lte-like systems. *IEEE Transactions on Wireless Communications*, 16(1):503–511, 2016.
- [5] Yang Wen, Wei Huang, and Zhongpei Zhang. Czac sequence and its application in lte random access. In *2006 IEEE Information Theory Workshop-ITW’06 Chengdu*, pages 544–547. IEEE, 2006.
- [6] Jiri Blumenstein and Marek Bobula. Coarse time synchronization utilizing symmetric properties of zadoff–chu sequences. *IEEE Communications Letters*, 22(5):1006–1009, 2018.
- [7] Malik Muhammad Usman Gul, Xiaoli Ma, and Sungeun Lee. Timing and frequency synchronization for ofdm downlink transmissions using zadoff–chu sequences. *IEEE Transactions on Wireless Communications*, 14(3):1716–1729, 2015.

- [8] Mohammed H AlSharif, Mohamed Saad, Mohamed Siala, Tarig Ballal, Hatem Boujemaa, and Tareq Y Al-Naffouri. Zadoff-chu coded ultrasonic signal for accurate range estimation. In *2017 25th European Signal Processing Conference (EUSIPCO)*, pages 1250–1254. IEEE, 2017.
- [9] Jesús López-Fernández, Unai Fernández-Plazaola, José F Paris, Luis Díez, and Eduardo Martos-Naya. Wideband ultrasonic acoustic underwater channels: Measurements and characterization. *IEEE Transactions on Vehicular Technology*, 69(4):4019–4032, 2020.
- [10] Dennis Ogbe, David J Love, Matthew Rebholz, and T Patrick Bidigare. Efficient channel estimation for aerial wireless communications. *IEEE Transactions on Aerospace and Electronic Systems*, 55(6):2774–2785, 2019.
- [11] Taehoon Kim, Inkyu Bang, and Dan Keun Sung. An enhanced prach preamble detector for cellular iot communications. *IEEE Communications Letters*, 21(12):2678–2681, 2017.
- [12] Felipe AP De Figueiredo, Fabbryccio ACM Cardoso, Ingrid Moerman, and Gustavo Fraidenraich. Channel estimation for massive mimo tdd systems assuming pilot contamination and frequency selective fading. *IEEE Access*, 5:17733–17741, 2017.
- [13] Nitin Jonathan Myers, Amine Mezghani, and Robert W Heath. Spatial zadoff-chu modulation for rapid beam alignment in mmwave phased arrays. In *2018 IEEE Globecom Workshops (GC Wkshps)*, pages 1–6. IEEE, 2018.
- [14] Jens Elofsson and Peter Seimar. Software defined radio based mimo channel sounding. 2016.
- [15] Xiaoming Chen, Per-Simon Kildal, Jan Carlsson, and Jian Yang. Mrc diversity and mimo capacity evaluations of multi-port antennas using reverberation chamber and anechoic chamber. *IEEE Transactions on Antennas and Propagation*, 61(2):917–926, 2012.

- [16] Ruiyuan Tian, Buon Kiong Lau, and Zhinong Ying. Multiplexing efficiency of mimo antennas. *IEEE Antennas and Wireless Propagation Letters*, 10: 183–186, 2011.
- [17] Leilei Liu, Wei Hong, Haiming Wang, Guangqi Yang, Nianzu Zhang, Hongxin Zhao, Jing Chang, Cheng Yu, Xutao Yu, Hongjun Tang, et al. Characterization of line-of-sight mimo channel for fixed wireless communications. *IEEE Antennas and Wireless Propagation Letters*, 6:36–39, 2007.
- [18] Vinko Erceg, Pitchaiah Soma, Daniel S Baum, and Severine Catreux. Multiple-input multiple-output fixed wireless radio channel measurements and modeling using dual-polarized antennas at 2.5 ghz. *IEEE transactions on Wireless Communications*, 3(6):2288–2298, 2004.
- [19] Vikram R Anreddy and Mary Ann Ingram. Capacity of measured ricean and rayleigh indoor mimo channels at 2.4 ghz with polarization and spatial diversity. In *IEEE Wireless Communications and Networking Conference, 2006. WCNC 2006.*, volume 2, pages 946–951. Citeseer, 2006.
- [20] Xiongwen Zhao, Suiyan Geng, Lasse Vuokko, Jarmo Kivinen, and Pertti Vainikainen. Polarization behaviours at 2, 5 and 60 ghz for indoor mobile communications. *Wireless Personal Communications*, 27(2):99–115, 2003.
- [21] Persefoni Kyritsi, Donald C Cox, Reinaldo A Valenzuela, and Peter W Wolniansky. Effect of antenna polarization on the capacity of a multiple element system in an indoor environment. *IEEE Journal on Selected areas in Communications*, 20(6):1227–1239, 2002.
- [22] Jon W Wallace, Michael A Jensen, A Lee Swindlehurst, and Brian D Jeffs. Experimental characterization of the mimo wireless channel: Data acquisition and analysis. *IEEE Transactions on Wireless Communications*, 2(2): 335–343, 2003.

- [23] Ana Vazquez Alejos, Manuel Garcia Sanchez, and Iñigo Cuiñas. Performance analysis of polarization diversity for indoor scenarios at 41.4 ghz and 61.5 ghz. *International Journal of Antennas and Propagation*, 2012, 2012.
- [24] Persefoni Kyritsi and Donald C Cox. Propagation characteristics of horizontally and vertically polarized electric fields in an indoor environment: simple model and results. In *IEEE 54th Vehicular Technology Conference. VTC Fall 2001. Proceedings (Cat. No. 01CH37211)*, volume 3, pages 1422–1426. IEEE, 2001.

Chapter 7

Near-field MIMO Capacity Measurements

7.1 Introduction

MIMO capacity in rich-scattering environment is important to understand in NLOS and LOS scenarios. The measurements in previous chapters highlight the frequency transfer functions measurements using 2x2 MIMO USRP testbed. Extensive measurements of MIMO channel distribution, 2x2 MIMO BER and EVM measurements were presented in previous chapters. The measurements were conducted in different environment loaded enclosure, empty enclosure, LOS and NLOS, and at different stirrer positions in the presence of MIMO interference. Rich-scattering environment can be obtained in reverberation chambers and metal enclosure. MIMO capacity measurements in reverberation chamber are conducted in [1] [2]. Authors in [2] have studied the maximum ratio combining diversity and MIMO capacity in reverberation chamber. Authors in [3] have performed diversity and MIMO capacity measurements in non-isotropic environment in reverberation chamber. The study has investigated 3x3 MIMO and 3x4 MIMO in non-isotropic environment. authors in [4] have introduced ideal threshold receiver for diversity and MIMO capacity analysis. The ideal threshold receiver works on receiver power in each channel of the MIMO systems and the designed threshold. Measured MIMO throughput has step transition from 0 to maximum value.

7.2 Wideband Channel capacity

High data rate transmission over MIMO wireless channel is important to understand for a low complex receiver design. For high bandwidth signals, the wireless channel may exhibit frequency selectivity and inter-symbol interference at the receiver. It may require more efforts to design a sophisticated receiver that can give low BER in presence of frequency selectivity. In such case MIMO-OFDM receivers have remained promising. However, OFDM modulation which has noise-like properties can exhibit higher peak-to-average power ratio thus over-driving transceiver. However, it counteracts ISI at the cost of receiver complexity but it is useful in rich-scattering. Also OFDM provide frequency diversity by dividing transmission bandwidth into narrowband subcarriers. MIMO-OFDM system can introduce spatial diversity and frequency diversity. For this it is important to understand wideband MIMO channel capacity. Therefore, for a MIMO system with N_T transmit and N_R receive antenna elements, the narrowband MIMO capacity is given as [5]:

$$C_{nb}(f) = B \log \left(1 + \frac{\rho}{N_T} H(f) H(f) \right) \quad \text{bits/s/Hz} \quad (7.1)$$

Where B is bandwidth, ρ is MIMO SNR and $H(f)$ MIMO channel transfer function. Hence, the wideband MIMO channel capacity can be expressed as sum of narrowband MIMO capacities [5][6]:

$$C_{wb} = \frac{1}{K} \sum_{f=1}^K C_{nb} \quad \text{bits/s/Hz} \quad (7.2)$$

MIMO channel transfer $H(f)$ function is measured using procedure as defined in previous chapters. MIMO transfer function is ratio of received power to the transmitted power. frequency transfer function of each elements of MIMO channel matrix is measured using USRP as outlined in chapter 2. Fig.7.4 shows S21 parameters sweep of MIMO channel in metal enclosure. The measurements are

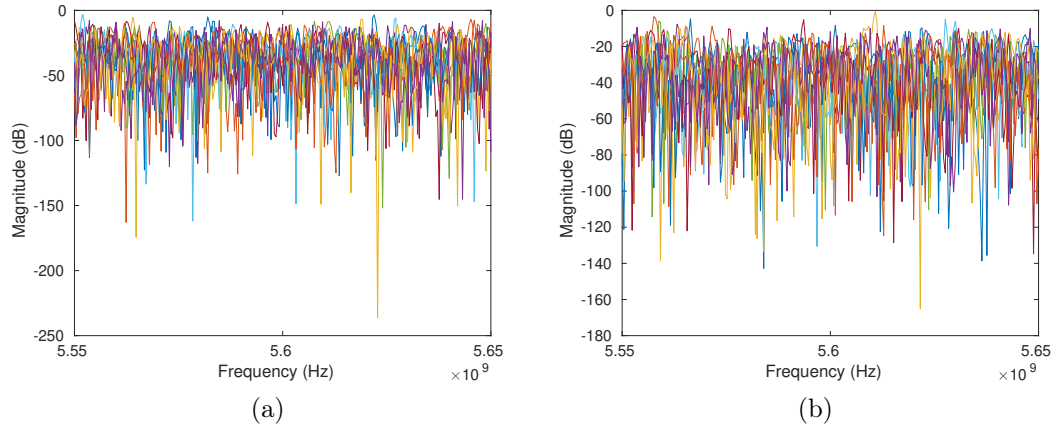


Figure 7.1: Channel transfer function of MIMO channel in metal enclosure. The frequency is swept from 5.55 GHz to 5.65 GHz with step size of 400 KHz. Total of 251 complex S21 parameters are measured per sweep (a) Channel 1 S21 parameters (b) Channel 2 S21 parameters.

performed when the Tx inter-element is 75 mm, Rx inter-element is 45 mm and distance between Tx antenna elements and Rx antenna elements is set to 65 mm.

Fig.7.2 shows channel transfer function of two orthogonal channels. It can be seen from Fig.7.2 that signal energy is received at all frequency values because of highly reflective environment and omnidirectional antenna. This is because in enclosed metal environment all MPCs are received and there is no substantial difference in channel transfer function despite the change in antenna positions. In 2D representation overflow is observed in the form of low-gain at around 35 sec interval.

Fig.7.3 shows that the K-factor is smaller than 1 which means the fading is more like Rayleigh fading than the Rician. This is consistent with previous measurements in which channel coefficients of measured MIMO channel matrix follow Weibull distribution which is special case of Rayleigh fading. Peak values are observed when there is direct un-stirred component received at receiving antenna. This is consistent with previous studies. This occurs because of ineffective shape of the mode-stirrer.

Fig.7.4 shows average magnitude of channel 1 and channel 2. It can be seen that the two channels are not frequency selective and have only small scale-fading

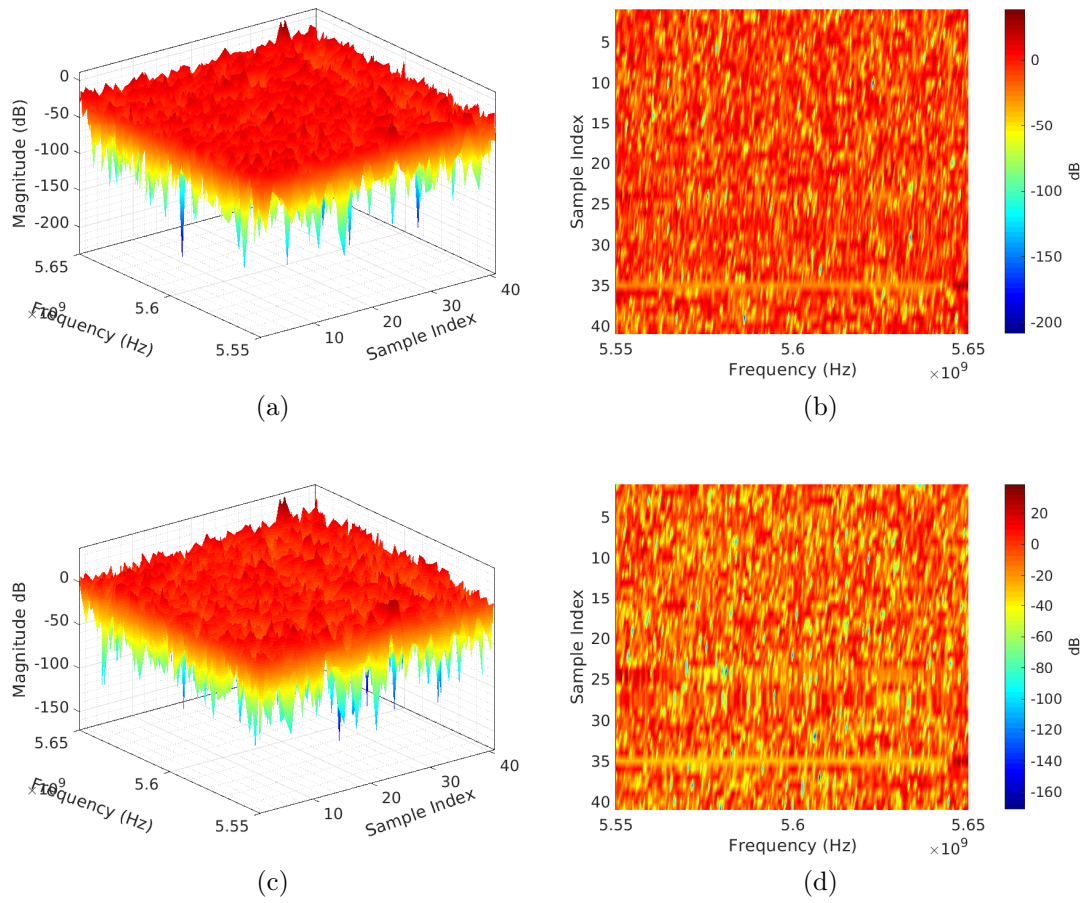


Figure 7.2: Channel transfer function of open loop MIMO channel in mode-stirred metal enclosure. The measurements are performed under equal power allocation strategy with Tx and Rx USRP antenna gains of 4 dB at both Tx and Rx side.

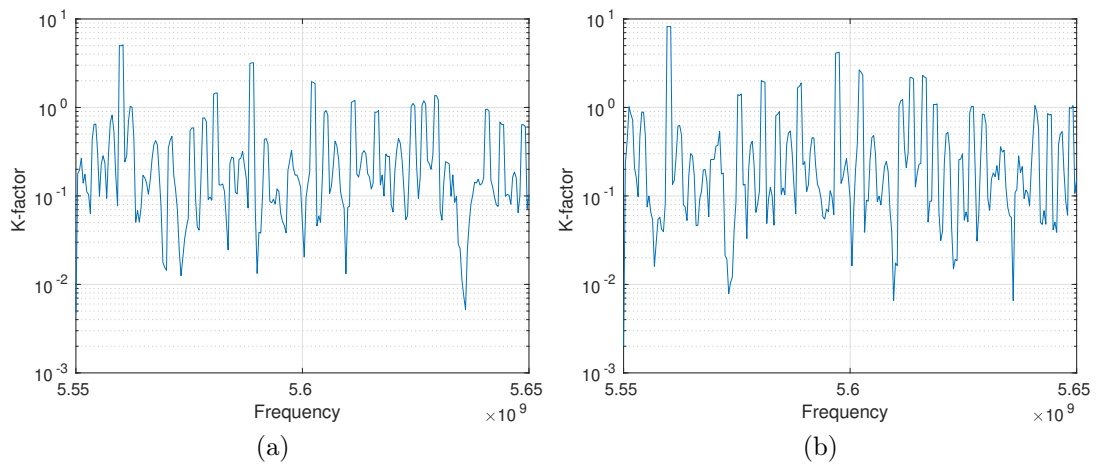


Figure 7.3: K-factor of channel 1 and channel 2.

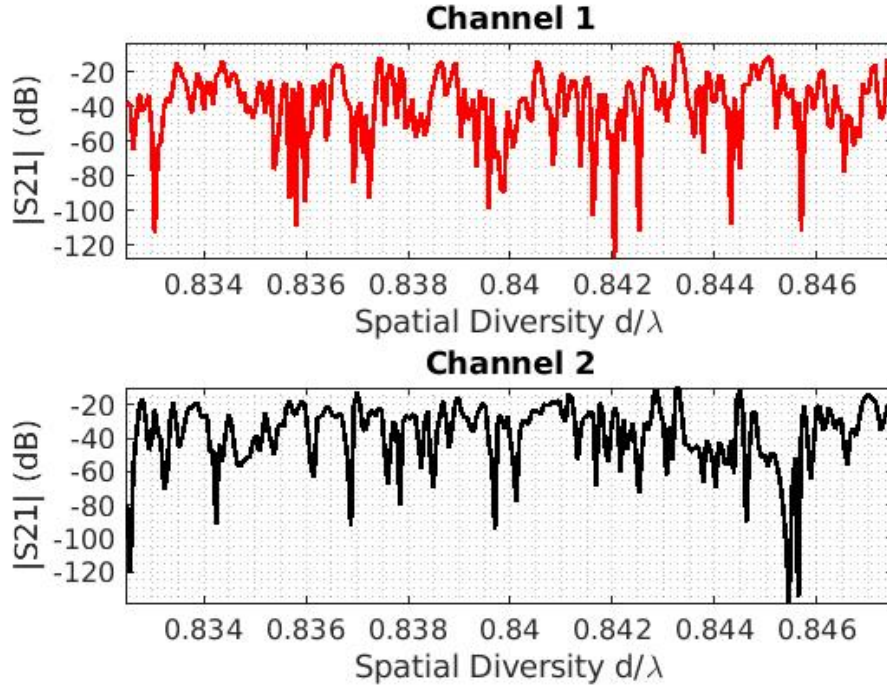


Figure 7.4: Average path gain of MIMO channel in mode-stirred metal enclosure.

because of the reflective metal enclosure environment. Fig.7.5 shows the amount of Doppler introduced in the channel because of moving mode-stirrer. Doppler is relative change in frequency of received signal when the transmitter and/or receiver are moving. Doppler is also introduced because of moving scatterers. Continuously rotating mode-stirrer is analogous to moving scatterers. Hence, the Doppler is introduced because of rotating mode-stirrer. It's important to statistical characterize Doppler spread. As this effect can introduce frequency offset and timing offset in the signal. The offset values are very important for the design of wireless communication system. The importance of these two parameters is presented in the form of measured results in previous chapters. Fig.7.6 shows diversity in 2x2 MIMO channel. Three diversity techniques are implemented namely selection combining (SC), equal gain combining (EGC), and maximum ratio combining (MRC). Diversity gain can be measured using empirical definition as given by Turkmani in [7] and is measured as the difference in signal level of the branch with the higher mean and that of the output of the diversity combiner for a given signal level [8]. MRC yields the highest diversity gain followed

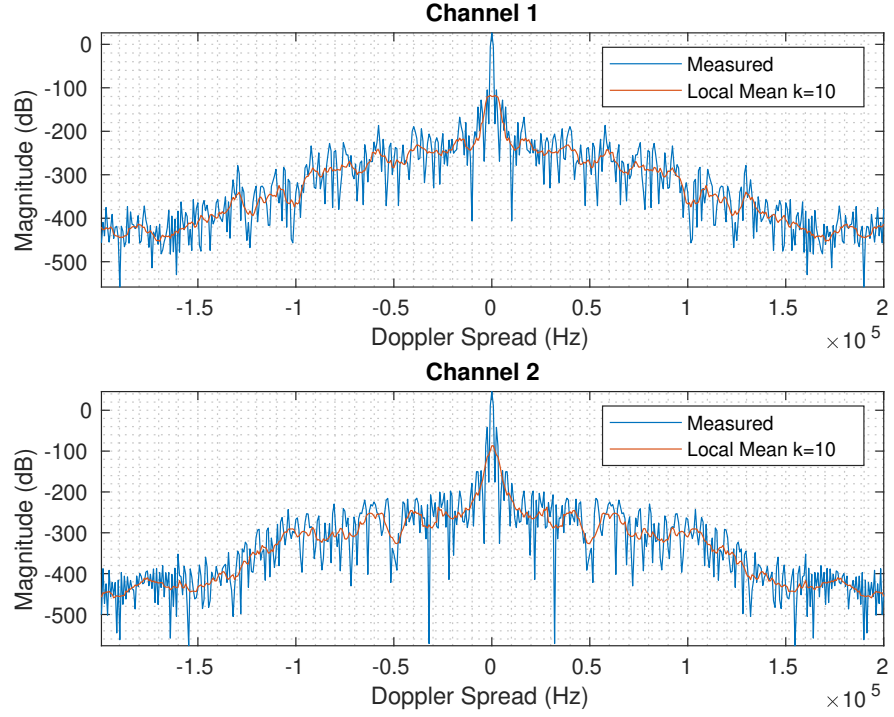


Figure 7.5: Doppler spread experienced by two channels in mode-stirred metal enclosure.

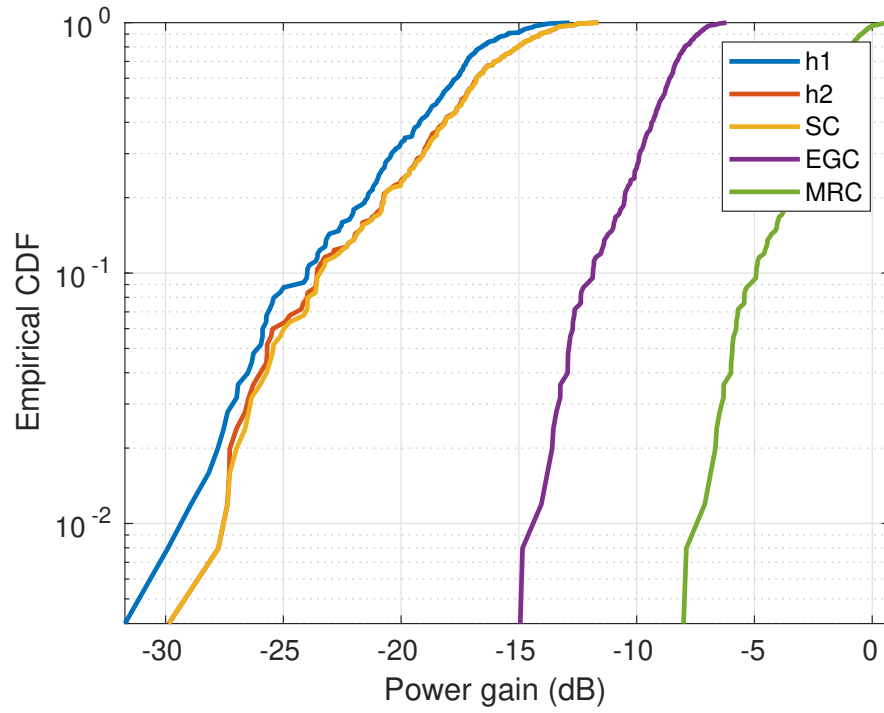


Figure 7.6: Empirical CDFs of two MIMO channels in mode-stirred metal enclosures. SC, EGC and MRC are three diversity techniques. Rx inter-element distance is 45 mm and Tx and Rx USRP gains were set to 4 dB respectively.

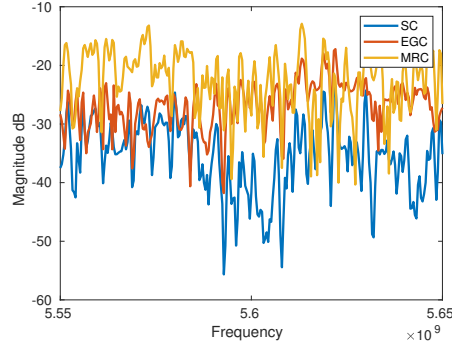
by EGC. SC is maximum of the two channels hence it lower diversity gain and appears closer to stronger channel. For two-branch diversity channel expression for SC, EGC and MRC can be given as follows:

$$\begin{aligned} y_{sc} &= \max \{y_1, y_2\} \\ y_{egc} &= \frac{y_1 + y_2}{\sqrt{2}} \\ y_{mrc} &= \sqrt{y_1^2 + y_2^2} \end{aligned} \tag{7.3}$$

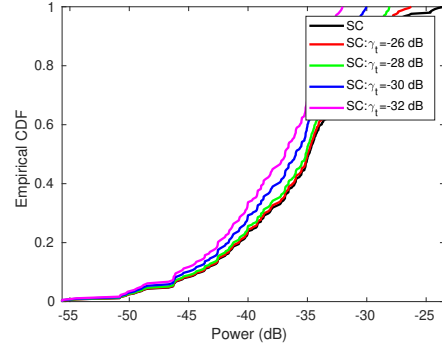
Fig.7.7 shows measured MIMO diversity techniques in mode-stirred metal enclosure. As expected MRC outperforms SC and EGC. This figure also shows that number of SC,EGC and MRC diversity samples received at threshold values of -26 dB to 32 dB. It's again clear that for all the threshold values MRC has more number of received samples at all threshold values followed by EGC.

7.3 Conclusion

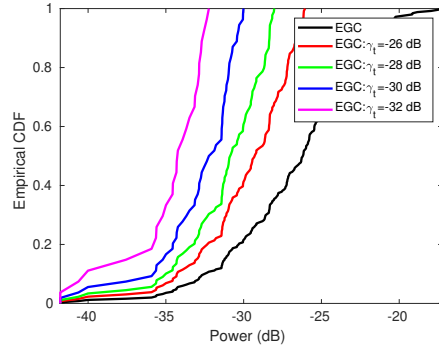
Wideband MIMO channel capacity measurements are performed in the metal enclosure for different distance between transmitter and receiver. A total of 100 MHz bandwidth is swept. The measured parameters are swept for the both SDR channels. It's shown that most of the energy is contained in the metal enclosure and the received signal components are strong. Additionally, diversity measurement is performed at different thresholds and based on the empirical definition of diversity gain the results show that the MRC technique outperform other diversity techniques i.e. EGC and SC.



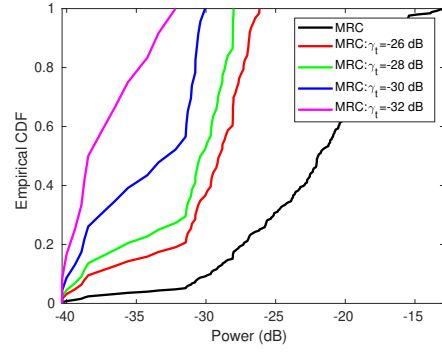
(a) SC, EGC and MRC



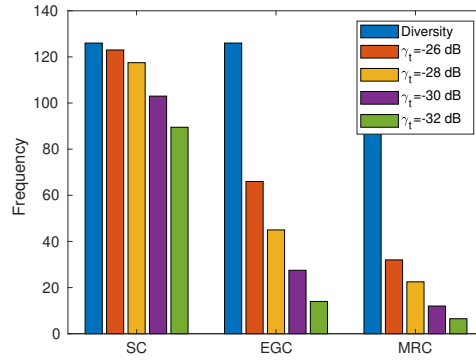
(b) SC diversity empirical CDF at different thresholds (γ_{th})



(c) EGC diversity empirical CDF at different thresholds (γ_{th})



(d) MRC diversity empirical CDF at different thresholds (γ_{th})



(e) Comparison of number of samples of three measured diversity techniques at different thresholds

Figure 7.7: Diversity measurement of open-loop 2x2 MIMO with equal power allocation strategy in mode-stirred metal enclosure. The inter-element distance at Tx and Rx was set to $d_t = 75\text{ mm}$ and $d_r = 45\text{ mm}$ respectively. The distance between Tx and Rx was set to $D = 65\text{ mm}$. The frequency was swept from 5.55 GHz to 5.65 GHz.

Bibliography

- [1] Kent Rosengren and P-S Kildal. Radiation efficiency, correlation, diversity gain and capacity of a six-monopole antenna array for a mimo system: theory, simulation and measurement in reverberation chamber. *IEE Proceedings-Microwaves, Antennas and Propagation*, 152(1):7–16, 2005.
- [2] Xiaoming Chen, Per-Simon Kildal, Jan Carlsson, and Jian Yang. Mrc diversity and mimo capacity evaluations of multi-port antennas using reverberation chamber and anechoic chamber. *IEEE Transactions on Antennas and Propagation*, 61(2):917–926, 2012.
- [3] Juan F Valenzuela-Valdés, Antonio M Martínez-González, and David A Sánchez-Hernández. Diversity gain and mimo capacity for nonisotropic environments using a reverberation chamber. *IEEE Antennas and Wireless Propagation Letters*, 8:112–115, 2009.
- [4] Per-Simon Kildal, Ahmed Hussain, Xiaoming Chen, Charlie Orlenius, Anton Skarbratt, John Asberg, Tommy Svensson, and Thomas Eriksson. Threshold receiver model for throughput of wireless devices with mimo and frequency diversity measured in reverberation chamber. *IEEE Antennas and Wireless Propagation Letters*, 10:1201–1204, 2011.
- [5] Yaoqing Yang, Guanghan Xu, and Hao Ling. An experimental investigation of wideband mimo channel characteristics based on outdoor non-los measurements at 1.8 ghz. *IEEE transactions on antennas and propagation*, 54(11):3274–3284, 2006.
- [6] Arogyaswami Paulraj, Arogyaswami Paulraj Rohit, Rohit Nabar, and Dhananjay Gore. *Introduction to space-time wireless communications*. Cambridge university press, 2003.
- [7] Adel MD Turkmani, Andrew A Arowojolu, PA Jefford, and CJ Kellett. An experimental evaluation of the performance of two-branch space and polarization

diversity schemes at 1800 mhz. *IEEE Transactions on Vehicular Technology*, 44(2):318–326, 1995.

- [8] Carl B Dietrich, Kai Dietze, J Randall Nealy, and Warren L Stutzman. Spatial, polarization, and pattern diversity for wireless handheld terminals. *IEEE transactions on antennas and propagation*, 49(9):1271–1281, 2001.

Chapter 8

Conclusion and Future Work

8.1 Conclusion

Near-field MIMO communication for wireless chip-to-chip communication has been investigated in details. It has wide aspects of open-research problems. The legacy interconnects for chip-to-chip communication may not be suitable for higher data transmission because of the increased transistor downscaling and chip-to-chip system scalability. Wireless chip-to-chip communication can be helpful for increased scalability and reduced routing delays in wireless network on chip. chip-to-chip communication is a complex propagation environment and has the possibility of exhibiting rich-scattering environment, therefore wireless transfer of information at high data rates can be leveraged by using higher number of transmit and receive antennas. MIMO increases spectral efficiency under rich-scattering environment. However, increased MIMO efficiency can be degraded by presence of transmit and receive correlation. Independent and identically distributed MIMO channels assure higher channel capacity. However, diagonally correlated MIMO channel can offer even higher channel capacity than IID MIMO channels. MIMO channel capacity is constraint on distribution of transmit power. Distribution of transmit power depends on availability of channel state information at the transmitter. Knowledge of channel state information at MIMO transmitter is important in order to optimally distribute power in MIMO chan-

nel. Therefore, two predominant power allocation strategies are used known as equal power allocation and water-filling algorithm. Equal power allocation in MIMO system is used when there is no MIMO channel state information at the transmitter side and water-filling algorithm is used as an optimal power allocation strategy in closed-loop MIMO system i.e. when MIMO CSI is available at the transmitter in addition to the receiver. In order to maximize the MIMO channel capacity, MIMO channel should be well-conditioned i.e. it shouldn't be rank-deficient. Rank-deficient channels are also known as keyhole or pin-hole channels in which SISO capacity is obtained despite multiple antennas at the transmitter and receiver. MIMO channel measurements are performed in metal enclosure and the channel characterization of MIMO channel has been performed. Generally, MIMO subchannel elements are best fitted by Weibull distribution. MIMO channels also offer diversity. Therefore, from the measurements in different propagation environment diversity techniques are selection combining, equal gain combining and maximum ratio combining are extracted at different threshold levels. Dual polarized and hybrid polarized MIMO channels are beneficial in terms of form-factor over spatial MIMO. Polarization diversity can be helpful to pack more number of antennas and orientation of antennas can be adjusted considering the architectures and layout of the electronic chips and boards within a package. Cross-polar discrimination factor is measured to check the orthogonality of the two polarization. However, it's inherent property of wireless channel to depolarize the signal such that vertically polarized signal is received by horizontally polarized antenna. Despite all of this, there is still gain loss in MIMO subchannel elements. Hence, reduced form-factor can be obtained by using dual-polarization and hybrid polarization MIMO system but the capacity gain is less than the spatial MIMO, because of power loss in subchannel elements in dual and hybrid-polarized MIMO systems. A 2x2 MIMO SDR based testbed has been created to measure different aspects of complex MIMO communication in a metal enclosure environment. Metal enclosure environment generates higher number of

multipath component i.e. rich-scattering environment which can be exploited by using MIMO communication for improved BER, EVM and MIMO capacity. Two channel USRP BER and EVM measurements are presented in different scenarios i.e LOS and NLOS in metal enclosure. Additionally, the two channel BER and EVM measurements were also performed in presence of RF absorbers i.e. loading the enclosure, in presence of RF interference and in presence of mode-stirrer movement. Mode-stirrer performed one revolution roughly in 60 seconds. The result of moving mode-stirrer in metal enclosure is a non-stationary environment and the effect of moving stirrer on two channel EVM was varying instantaneous EVM. It is observed that the two channel EVM increases in presence of different levels of interference. For main communication link gain of 10 dB, the best channel 1 EVM was 34% and the best EVM value for channel 2 was 20%. The worst channel 1 and channel 2 EVM was 43% and 65% respectively. Similarly for two-channel BER, the best channel 1 BER was 10^{-9} and channel 2 BER was 0. The worst channel 1 and channel 2 BER values were around 1 and 0.5 respectively.

8.2 Future Work

There are many future aspects of this research. This experimental research generated many ideas in terms of channel modeling, scalability of SDR based testbed, operating frequency, types of antennas, type of radio environment, choice of modulation techniques, and communication systems implementation aspects. A similar SDR based testbed for higher frequencies such as mmWave can be more helpful. Additionally, the MIMO channel sounding can be performed with SDRs that offer higher bandwidth such as Cyan high bandwidth SDR by Per Vices which offers RF bandwidth of 3 GHz.

mmWave MIMO links

The near-field MIMO testbed can be extended to work at mmWave frequencies and with increased number of transmit and receive antennas. This testbed then can be used for measurements in the metal enclosure or computer casings. Although, there are measurement present in the literature but are used in SISO communication. It shall be useful to understand type of mmWave MIMO channel in confined environment with realistic scenario where computer architectural components are present.

MIMO-OFDM Receiver

Full-stack of MIMO-OFDM receiver can be used for end-to-end communication system measurements such as throughput and BER measurements. The state-of-the-art MIMO-OFDM receiver can be developed that meets the requirements of a high data rate wireless chip-to-chip communication link.

Channel Sounding

Channel sounding in time-domain using orthogonal sequences should be performed at higher sampling rates which can provide descent time-resolution. This means SDRs with increased bandwidth can be used for channel sounding of the metal enclosure and computer casings. Increased resolution can help resolve multipath clusters which can be modeled using Saleh-Valenzuela multi-clustering approach. Additionally, adaptive power control mechanism should be applied when the there's likelihood that the channel profile will change. However, for short-range propagation environment within metal enclosure stationary channels can be expected.

Reconfigurable Intelligent Surfaces

Reconfigurable intelligent surfaces are a passive reflectors with huge number of low-cost elements that can tune properties of wireless channels. RIS is energy-efficient and spectral-efficient solution. It can be a game-changer technology for the realization of wireless chip-to-chip communication as it can be designed at most of electromagnetic spectrum bands. RIS can perform passive beamforming and it can create virtual LOS links. This can relax chip fabrication and electronic circuitry related issues. Since RIS can create multiple links using passive beamforming, therefore, RIS-assisted multi-chip wireless links can potentially be created. At the time of writing, some aspects of this work were exploited to create robust RIS-assisted links in metal enclosure. The RIS-assisted measurements were performed at 5 GHz. EVM optimization was performed in presence of different levels of noise. The series of experiments generated plenty of ideas which can be materialized for multi-chip wireless communication.

Appendix A

- USRP Frequency Sweep for Received Power Measurements

```
1 %% Frequency Sweep for received power measurements
2 f_values=[linspace(5.55e9,5.65e9,100)];
3 for i = 1:length(f_values)
4     simIn(i) = Simulink.SimulationInput('sdrupsktxrx_2x2');
5     simIn(i) = setVariable(simIn(i),'f',f_values(i));
6 end
7 simOutputs(1,:) = sim(simIn);
8 figure
9 BER1=zeros(1,length(f_values));
10 BER2=zeros(1,length(f_values));
11 %BER3=zeros(1,length(SNR_values));
12 for i=1:length(f_values)
13     BER1(1,i)=mean(simOutputs(1,i).BER1.signals.values(:,1));
14     BER2(1,i)=mean(simOutputs(1,i).BER2.signals.values(:,1));
15 end
16 semilogy(f_values,BER1,...
17     '-.sr','MarkerSize',10,...
18     'MarkerEdgecolor','k','MarkerFaceColor','r',...
19     'linewidth',1.15)
20 hold on
21 semilogy(f_values,BER2,...
22     '-.dk','MarkerSize',10,...
23     'MarkerEdgecolor','k','MarkerFaceColor','k',...
24     'linewidth',1.15)
25 xlabel('Tx Gain (dB)')
26 ylabel('Measured BER')
27 legend('Channel 1','Channel 2')
28 set(gca,'fontsize',14)
29 grid on
30 set(gca,'linewidth',1.25)
31 title('Interference Tx Gain=70dB')
32 %saveas(gcf,'empty_gainsweep_85int_6cones.jpg')
```

- USRP Gain Sweep for BER Measurements

```
1 %% USRP Gain Sweep for BER measurements
2 gain_values=2:2:10;
3 for i = 1:length(gain_values)
4     simIn(i) = Simulink.SimulationInput('sdrupsktxrx_2x2');
5     simIn(i) = setVariable(simIn(i),'gain',gain_values(i));
6 end
```

```

7  simOutputs(1,:) = sim(simIn);
8  figure
9  BER1=zeros(1,length(gain_values));
10 BER2=zeros(1,length(gain_values));
11 %BER3=zeros(1,length(SNR_values));
12 for i=1:length(gain_values)
13     BER1(1,i)=mean(simOutputs(1,i).BER1.signals.values(:,1));
14     BER2(1,i)=mean(simOutputs(1,i).BER2.signals.values(:,1));
15 end
16 semilogy(gain_values,BER1,...
17     '-.sr','MarkerSize',10,...
18     'MarkerEdgecolor','k','MarkerFaceColor','r',...
19     'linewidth',1.15)
20     hold on
21 semilogy(gain_values,BER2,...
22     '-.dk','MarkerSize',10,...
23     'MarkerEdgecolor','k','MarkerFaceColor','k',...
24     'linewidth',1.15)
25 xlabel('Tx Gain (dB)')
26 ylabel('Measured BER')
27 legend('Channel 1','Channel 2')
28 set(gca,'fontsize',14)
29 grid on
30 set(gca,'linewidth',1.25)
31 title('Interference Tx Gain=85dB')
32 %saveas(gcf,'empty_gainsweep_85int_6cones.jpg')

```

- USRP Gain Sweep for EVM Measurements

```

1  %% USRP Gain Sweep for EVM Measurements
2  gain_values=2:2:10;
3  for i = 1:length(gain_values)
4      simIn(i) = Simulink.SimulationInput('sdrugpsktxrx_2x2');
5      simIn(i) = setVariable(simIn(i),'gain',gain_values(i));
6  end
7  simOutputs(1,:) = sim(simIn);
8  figure
9  EVM1=zeros(1,length(gain_values));
10 EVM2=zeros(1,length(gain_values));
11 %BER3=zeros(1,length(SNR_values));
12 for i=1:length(gain_values)
13     EVM1(1,i)=mean(simOutputs(1,i).EVM1.signals.values(:,1));
14     EVM2(1,i)=mean(simOutputs(1,i).EVM2.signals.values(:,1));
15 end
16 for i=1:length(gain_values)
17     Pdbm1(1,i)=mean(simOutputs(1,i).Pdbm1.signals.values(:,1));
18     Pdbm2(1,i)=mean(simOutputs(1,i).Pdbm2.signals.values(:,1));
19 end
20 plot(gain_values,EVM1,...
21     '-rs','MarkerSize',10,...
22     'MarkerEdgecolor','k','MarkerFaceColor','r',...
23     'linewidth',2)
24     hold on
25 plot(gain_values,EVM2,...
26     '-dk','MarkerSize',10,...
27     'MarkerEdgecolor','k','MarkerFaceColor','k',...
28     'linewidth',2)

```

```

29 xlabel('Tx Gain (dB)')
30 ylabel('Measured EVM (%)')
31 legend('Channel 1','Channel 2')
32 set(gca,'fontsize',16)
33 grid on
34 %saveas(gcf,'empty_gainsweep_6cones_60_int75_EVMbar.jpg')
35 %%
36 figure
37 bar_data=[EVM1;EVM2];
38 Pdbm=[Pdbm1;Pdbm2];
39 % Get the max value of data (used ot det the YLIM)
40 mx=max(bar_data(:))
41 % Draw the grouped bars
42 hbar=bar(bar_data)
43 % Get the XDATA
44 XDATA=get(hbar(1),'XData')';
45 % Define the vertical offset of the labels
46 ygap=mx*0.13;
47 ygap_pow=mx*0.25;
48 % Loop over the bar's group
49 for i=1:length(hbar)
50     % Get the YDATA of the i-th bar in each group
51     YDATA=get(hbar(i),'YData')';
52     % Loop over the groups
53     for j=1:length(XDATA)
54         % Get the XPOS of the j-th bar
55         xpos=XDATA(j);
56         % Get the height of the bar and increment it with ...
           the offset
57         ypos=YDATA(j)+ygap;
58         ypos1=YDATA(j)+ygap_pow;
59         % Define the labels
60         labels=[num2str(YDATA(j),3),' %'];
61         % Add the labels
62         t = ...
           text(xpos+hbar(i).XOffset,ypos,labels,'Color','k',...
           'HorizontalAlignment','center','Rotation',90)
63     end
64 end
65 str={'SDR Channel 1';'SDR Channel 2'};
66 set(gca,'XTickLabel',str,'XTick',1:numel(str))
67 ylabel('Measured EVM %')
68 set(gca,'fontsize',14)
69 legend('Tx Gain=2dB','Tx Gain=4dB','Tx Gain=6dB','Tx ...
           Gain=8dB','Tx Gain=10dB')
70
71 grid minor
72 shg
73 ylim([0 140])
74 %saveas(gcf,'empty_gainsweep_6cones_60_int75_EVMbar.jpg')

```

- Parseval's Theorem: Power content of a signal in time domain is equal to power content in frequency domain.

```

1 function [pow_f,pow_t,error] = fcn(u)
2 power_time=10*log10(sum(abs(u).^2)); % Power in time-domain
3 power_frequency=10*log10(sum(abs(fft(u)).^2)/length(u)); ...

```

```

        % Power in frequency-domain
4  pow_t =max(power_time);
5  pow_f=max(power_frequency);
6  error=(pow_t-pow_f);

```

- Programmatically received power measurements. This may take a lot of time, therefore, it's better to resort on frequency sweep using orthogonal sinusoids for MIMO frequency response measurements.

```

1  %% Frequency Sweep for received power measurements
2  tic;
3  f_values=[linspace(5.55e9,5.65e9,251)];
4  for i = 1:length(f_values)
5      simIn(i) = Simulink.SimulationInput('S21_sinesweep');
6      simIn(i) = setVariable(simIn(i), 'f', f_values(i));
7  end
8  simOutputs(1,:) = sim(simIn);
9  toc
10 %% Received Power in dB
11 figure
12 Prx1_dB=zeros(1,length(f_values));
13 Prx2_dB=zeros(1,length(f_values));
14 for i=1:length(f_values)
15     Prx1_dB(1,i)=mean(simOutputs(1,i).Pdb1.signals.values(:,1));
16     Prx2_dB(1,i)=mean(simOutputs(1,i).Pdb2.signals.values(:,1));
17 end
18 semilogy(f_values,Prx1_dB,'linewidth',1.15)
19 hold on
20 semilogy(f_values,Prx2_dB,'linewidth',1.15)
21 xlabel('Frequency (Hz)')
22 ylabel('Received Power (dB)')
23 legend('Channel 1','Channel 2')
24 set(gca,'fontsize',14)
25 grid on
26 %% Received Power in dBm
27 figure
28 Prx1_dBm=zeros(1,length(f_values));
29 Prx2_dBm=zeros(1,length(f_values));
30 for i=1:length(f_values)
31     Prx1_dBm(1,i)=mean(simOutputs(1,i).Pdbm1.signals.values(:,1));
32     Prx2_dBm(1,i)=mean(simOutputs(1,i).Pdbm2.signals.values(:,1));
33 end
34 semilogy(f_values,Prx1_dBm,'linewidth',1.15)
35 hold on
36 semilogy(f_values,Prx2_dBm,'linewidth',1.15)
37 xlabel('Frequency (Hz)')
38 ylabel('Received Power (dBm)')
39 legend('Channel 1','Channel 2')
40 set(gca,'fontsize',14)
41 grid on

```

Appendix B

- AD9371 RFIC Schematic:

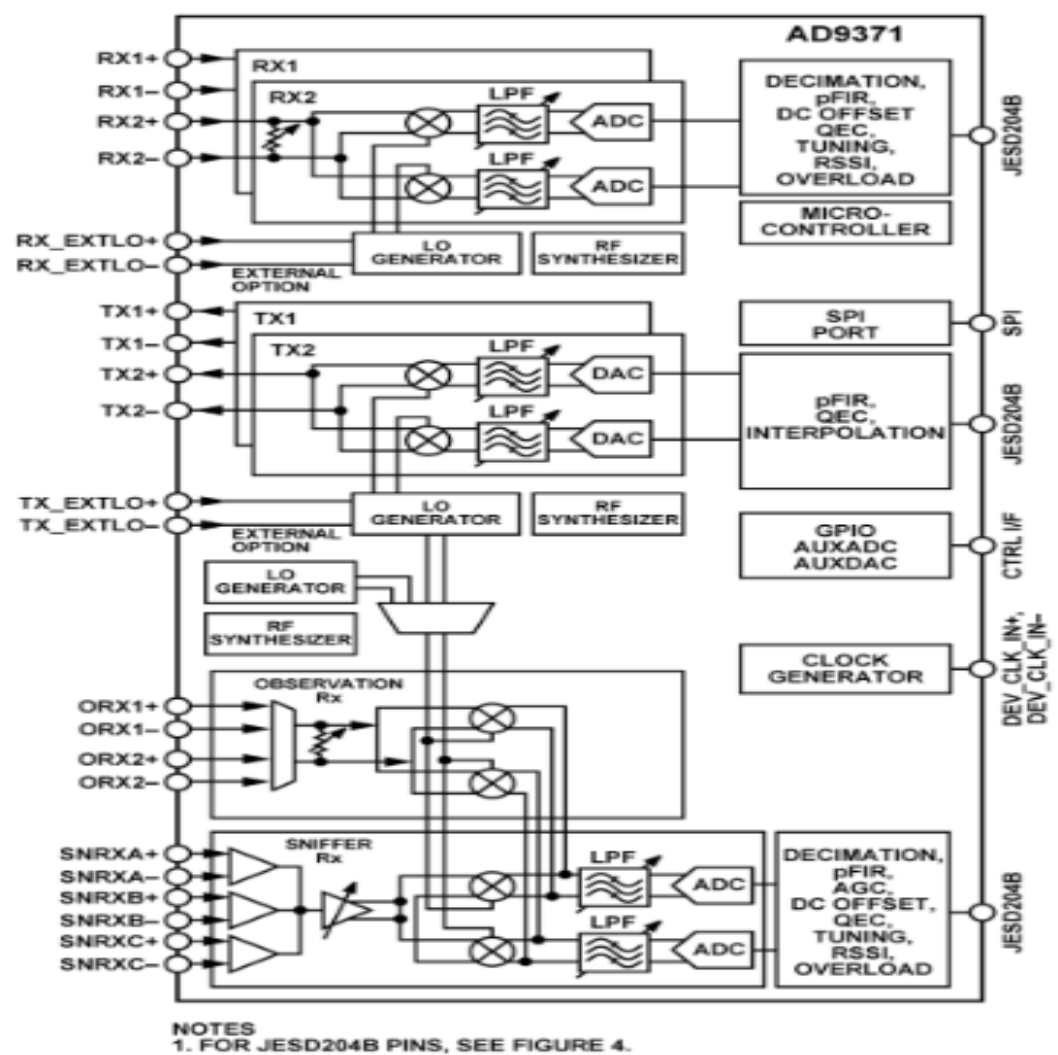
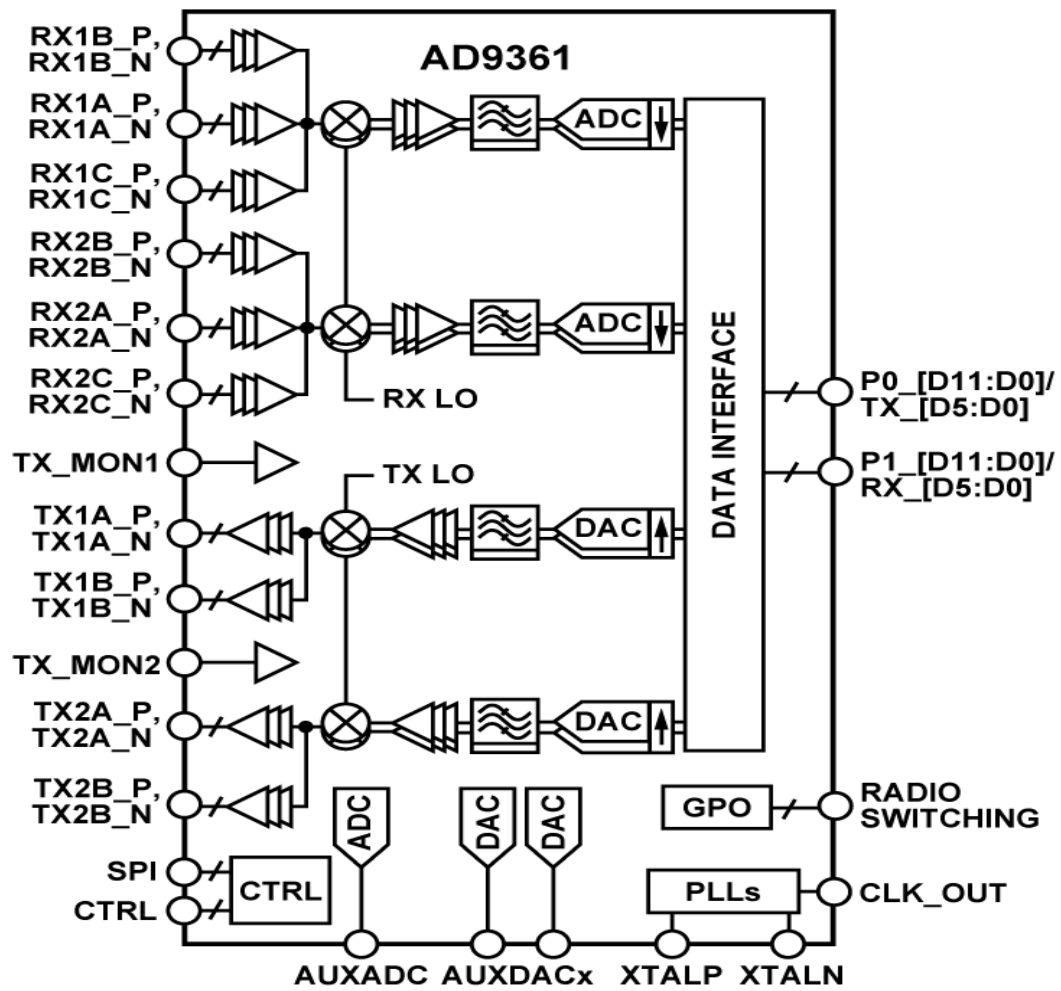


Figure 8.1: AD9371 Schematic

- AD9361 RFIC Schematic:



NOTES

1. SPI, CTRL, P0_[D11:D0]/TX_[D5:D0], P1_[D11:D0]/RX_[D5:D0], AND RADIO SWITCHING CONTAIN MULTIPLE PINS.

10453-001

Figure 8.2: AD9361 Schematic

- SDRu MIMO Transmitter Settings

Block Parameters: USRP Transmitter

SDRu Transmitter (mask) (link)

Send data to the Universal Software Radio Peripheral (USRP).

Radio Connection

Platform: X310
IP address: 192.168.10.2
Refresh Device List
Info

Radio Properties

Channel mapping: [1 2]
Source of center frequency: Dialog
Center frequency (Hz): 5.6e9
Source of LO offset: Dialog
LO offset (Hz): 0
Source of gain: Dialog
Gain (dB): 2
PPS source: Internal
Clock source: Internal
Master clock rate (Hz): 200e6
Source of interpolation factor: Dialog
Interpolation factor: 500

Data

Baseband sample rate (Hz): 400000
Transport data type: int16
☒ Enable underrun output port
☐ Enable burst mode

OK
Cancel
Help
Apply

Figure 8.3: Transmitter Settings

- SDRu MIMO Receiver Setting

Block Parameters: USRP Receiver

SDRu Receiver (mask) (link)

Receive data from the Universal Software Radio Peripheral (USRP).

Radio Connection

Platform: X310

IP address: 192.168.10.2 Refresh Device List

Info

Radio Properties

Channel mapping: [1 2]

Source of center frequency: Dialog

Center frequency (Hz): 5.6e9

Source of LO offset: Dialog

LO offset (Hz): 0

Source of gain: Dialog

Gain (dB): 0

PPS source: Internal

Clock source: Internal

Master clock rate (Hz): 200e6

Source of decimation factor: Dialog

Decimation factor: 500

Data

Baseband sample rate (Hz): 400000

Transport data type: int16

☒ Enable overrun output port

Output data type: double

Samples per frame: sdrqpskrx.USRPFramLength

☐ Enable burst mode

OK Cancel Help Apply

Figure 8.4: Receiver Settings

- `uhd_find_devices`

```
mir@mirlodro:~$ uhd_find_devices
[INFO] [UHD] linux; GNU C++ version 7.5.0; Boost_106501; UHD_3.13.1.HEAD-0-g4a356623
-----
-- UHD Device 0
-----
Device Address:
  serial: 30C098D
  addr: 192.168.10.2
  fpga: HG
  name:
  product: X310
  type: x300
```

Figure 8.5: UHD probe command result. It is used to know number of UHD devices connected to host PC.

- `uhd_usrp_probe --args addr=192.168.10.2`

```
mir@mirlodro:~$ uhd_usrp_probe --args addr=192.168.10.2
[INFO] [UHD] linux; GNU C++ version 7.5.0; Boost_106501; UHD_3.13.1.HEAD-0-g4a356623
[INFO] [X300] X300 initialization sequence...
[INFO] [X300] Maximum frame size: 1472 bytes.
[INFO] [X300] Radio 1x clock: 200 MHz
[INFO] [GPS] No GPSDO found
[INFO] [0/DmaFIFO_0] Initializing block control (NOC ID: 0xF1F0D00000000000)
[INFO] [0/DmaFIFO_0] BIST passed (Throughput: 1300 MB/s)
[INFO] [0/DmaFIFO_0] BIST passed (Throughput: 1306 MB/s)
[INFO] [0/Radio_0] Initializing block control (NOC ID: 0x12AD100000000001)
[INFO] [0/Radio_1] Initializing block control (NOC ID: 0x12AD100000000001)
[INFO] [0/DDC_0] Initializing block control (NOC ID: 0xDDC0000000000000)
[INFO] [0/DDC_1] Initializing block control (NOC ID: 0xDDC0000000000000)
[INFO] [0/DUC_0] Initializing block control (NOC ID: 0xD0C0000000000000)
[INFO] [0/DUC_1] Initializing block control (NOC ID: 0xD0C0000000000000)

Device: X-Series Device
-----
Mboard: X310
revision: 8
revision_compat: 7
product: 30818
mac-addr0: 00:80:2f:23:7b:67
mac-addr1: 00:80:2f:23:7b:68
gateway: 192.168.10.1
ip-addr0: 192.168.10.2
subnet0: 255.255.255.0
ip-addr1: 192.168.20.2
subnet1: 255.255.255.0
ip-addr2: 192.168.30.2
subnet2: 255.255.255.0
ip-addr3: 192.168.40.2
subnet3: 255.255.255.0
serial: 30C098D
FW Version: 6.0
FPGA Version: 35.1
FPGA git hash: d0360f7
RFNoC capable: Yes

Time sources: internal, external, gpsdo
Clock sources: internal, external, gpsdo
Sensors: ref_locked

RX Dboard: A
ID: UBX-160 v1 (0x007a)
Serial: 30C2A76

RX Frontend: 0
Name: UBX_PV
```

Figure 8.6: Output of the command. It populates important information about connect USRP and the number of RF cards, the number of channels and the type of RF card.

- Continue

```

RX Dboard: A
ID: UBX-160 v1 (0x007a)
Serial: 30C2A76

RX Frontend: 0
Name: UBX RX
Antennas: TX/RX, RX2, CAL
Sensors: lo_locked
Freq range: 10.000 to 6000.000 MHz
Gain range PGA0: 0.0 to 31.5 step 0.5 dB
Bandwidth range: 160000000.0 to 160000000.0 step 0.0 Hz
Connection Type: IQ
Uses LO offset: No

RX Codec: A
Name: ads62p48
Gain range digital: 0.0 to 6.0 step 0.5 dB

RX Dboard: B
ID: UBX-160 v1 (0x007a)
Serial: 30C0DD3

RX Frontend: 0
Name: UBX RX
Antennas: TX/RX, RX2, CAL
Sensors: lo_locked
Freq range: 10.000 to 6000.000 MHz
Gain range PGA0: 0.0 to 31.5 step 0.5 dB
Bandwidth range: 160000000.0 to 160000000.0 step 0.0 Hz
Connection Type: IQ
Uses LO offset: No

RX Codec: B
Name: ads62p48
Gain range digital: 0.0 to 6.0 step 0.5 dB

TX Dboard: A
ID: UBX-160 v1 (0x0079)
Serial: 30C2A76

TX Frontend: 0
Name: UBX TX
Antennas: TX/RX, CAL
Sensors: lo_locked
Freq range: 10.000 to 6000.000 MHz

```

Figure 8.7: *continued*...Output of the command. It populates important information about connect USRP and the number of RF cards, the number of channels and the type of RF card.

- *uhd_image_loader*

```

eexnm10@duip67448:~$ uhd_image_loader --args="type=x300,addr=192.168.10.3" --fpga-path="/usr/share/uhd/images/usrp_x310_fpga_HG.bit"
[INFO] [UHD] linux; GNU C++ version 4.8.4; Boost_105400; UHD_3.11.0.0-release
Unit: USRP X310 (30C0991, 192.168.10.3)
FPGA Image: /usr/share/uhd/images/usrp_x310_fpga_HG.bit
-- Initializing FPGA loading...successful.
-- Loading HG FPGA image: 100% (121/121 sectors)
-- Finalizing image load...successful.
Power-cycle the USRP X310 to use the new image.

```

Figure 8.8: Image update of the USRP using *uhd_image_loader* command. It targets the connected USRP using IP address and specifies path to FPGA image.

- `uhd_image_loader` success

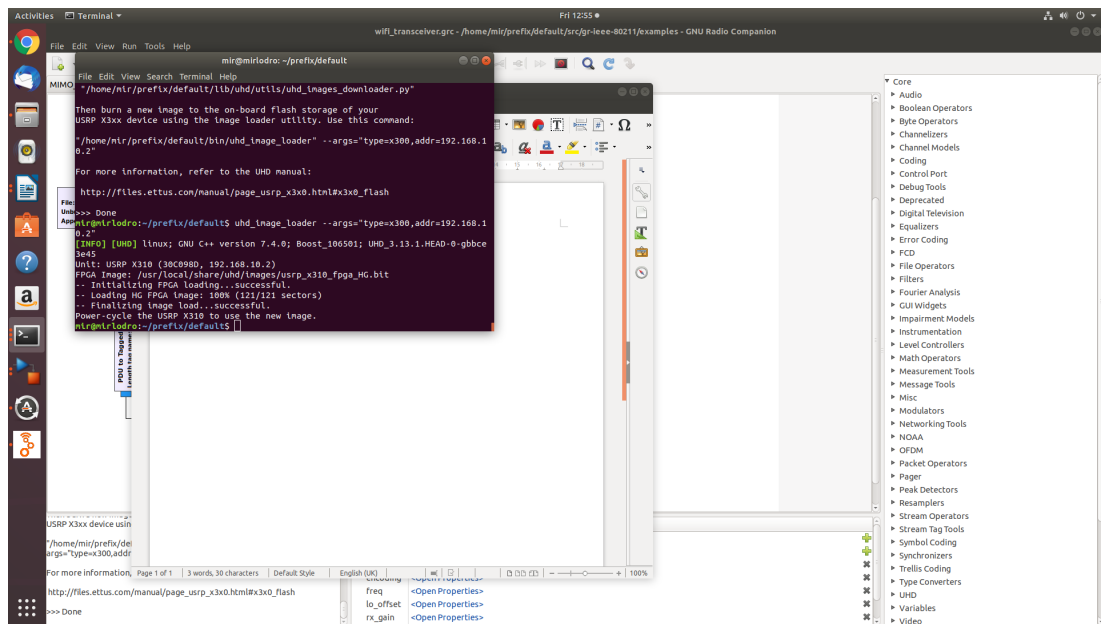


Figure 8.9: UHD image update success. Power-cycle is necessary once image has been updated. It's mandatory not to disconnect power supply of USRP when the image update is being performed otherwise this could brick the USRP.

Bibliography

- [1] Jinbang Fu, Prateek Juyal, and Alenka Zajić. Modeling of 300 ghz chip-to-chip wireless channels in metal enclosures. *IEEE Transactions on Wireless Communications*, 19(5):3214–3227, 2020.
- [2] Jinbang Fu, Prateek Juyal, and Alenka Zajić. Thz channel characterization of chip-to-chip communication in desktop size metal enclosure. *IEEE Transactions on Antennas and Propagation*, 67(12):7550–7560, 2019.
- [3] Bahaa Radi, Ajaypal Singh Dhillon, and Odile Liboiron-Ladouceur. Demonstration of inter-chip rf data transmission using on-chip antennas in silicon photonics. *IEEE Photonics Technology Letters*, 32(11):659–662, 2020.
- [4] Siavash Moghadami, Fleura Hajilou, Priyanka Agrawal, and Shahab Ardalani. A 210 ghz fully-integrated ook transceiver for short-range wireless chip-to-chip communication in 40 nm cmos technology. *IEEE Transactions on Terahertz Science and Technology*, 5(5):737–741, 2015.
- [5] Mohammadreza F Imani, Sergi Abadal, and Philipp del Hougne. Smart on-chip electromagnetic environment. *arXiv preprint arXiv:2109.03284*, 2021.
- [6] Marco Di Renzo, Alessio Zappone, Merouane Debbah, Mohamed-Slim Alouini, Chau Yuen, Julien De Rosny, and Sergei Tretyakov. Smart radio environments empowered by reconfigurable intelligent surfaces: How it works, state of research, and the road ahead. *IEEE Journal on Selected Areas in Communications*, 38(11):2450–2525, 2020.
- [7] Linglong Dai, Bichai Wang, Min Wang, Xue Yang, Jingbo Tan, Shuangkaisheng Bi, Shenheng Xu, Fan Yang, Zhi Chen, Marco Di Renzo, et al. Reconfigurable intelligent surface-based wireless communications: Antenna design, prototyping, and experimental results. *IEEE Access*, 8: 45913–45923, 2020.

- [8] Karsten König and Andreas Ostendorf. *Optically induced nanostructures: biomedical and technical applications*. Walter de Gruyter GmbH & Co KG, 2015.
- [9] John L Volakis. *Antenna engineering handbook*. McGraw-Hill Education, 2007.
- [10] Yen-Sheng Chen, Shih-Yuan Chen, and Hsueh-Jyh Li. Analysis of antenna coupling in near-field communication systems. *IEEE Transactions on antennas and propagation*, 58(10):3327–3335, 2010.
- [11] Hristomir H Yordanov and Peter Russer. Integrated on-chip antennas for chip-to-chip communication. In *2008 IEEE Antennas and Propagation Society International Symposium*, pages 1–4. IEEE, 2008.
- [12] Seahee Hwangbo, Yong-Kyu Yoon, and Aric B Shorey. Millimeter-wave wireless chip-to-chip (c2c) communications in 3d system-in-packaging (sip) using compact through glass via (tgv)-integrated antennas. In *2018 IEEE 68th Electronic Components and Technology Conference (ECTC)*, pages 2074–2079. IEEE, 2018.
- [13] Xavier Timoneda, Albert Cabellos-Aparicio, Dionysios Manassis, Eduard Alarcón, and Sergi Abadal. Channel characterization for chip-scale wireless communications within computing packages. In *2018 Twelfth IEEE/ACM International Symposium on Networks-on-Chip (NOCS)*, pages 1–8. IEEE, 2018.
- [14] Yazan Al-Alem, Ahmed A Kishk, and Raed Shubair. Wireless chip to chip communication link budget enhancement using hard/soft surfaces. In *2018 IEEE Global Conference on Signal and Information Processing (GlobalSIP)*, pages 1013–1014. IEEE, 2018.
- [15] Ihsan El Masri, Thierry Le Gouguec, Pierre-Marie Martin, Rozenn Allanic, and Cedric Quendo. Electromagnetic characterization of the intrachip prop-

- agation channel in *ka*-and *v*-bands. *IEEE Transactions on Components, Packaging and Manufacturing Technology*, 9(10):1931–1941, 2019.
- [16] Constantine A Balanis. *Antenna theory: analysis and design*. John wiley & sons, 2015.
- [17] Christoph Degen. Inductive coupling for wireless power transfer and near-field communication. *EURASIP Journal on Wireless Communications and Networking*, 2021(1):1–20, 2021.
- [18] Sebastian Stoecklin, Elias Rosch, Adnan Yousaf, and Leonhard Reindl. Very high bit rate near-field communication with low-interference coils and digital single-bit sampling transceivers for biomedical sensor systems. *Sensors*, 20(21):6025, 2020.
- [19] Han-Joon Kim, Hiroshi Hirayama, Sanghoek Kim, Ki Jin Han, Rui Zhang, and Ji-Woong Choi. Review of near-field wireless power and communication for biomedical applications. *IEEE Access*, 5:21264–21285, 2017.
- [20] Werner Meskens, Erika J Van Baelen, and Tony M Nygard. Magnetic induction communication system for an implantable medical device, October 20 2015. US Patent 9,166,655.
- [21] Xin Tan, Zhi Sun, and Ian F Akyildiz. Wireless underground sensor networks: Mi-based communication systems for underground applications. *IEEE Antennas and Propagation Magazine*, 57(4):74–87, 2015.
- [22] Shih-Chun Lin, Ian F Akyildiz, Pu Wang, and Zhi Sun. Distributed cross-layer protocol design for magnetic induction communication in wireless underground sensor networks. *IEEE Transactions on Wireless Communications*, 14(7):4006–4019, 2015.
- [23] Zhi Sun and Ian F Akyildiz. Magnetic induction communications for wireless underground sensor networks. *IEEE transactions on antennas and propagation*, 58(7):2426–2435, 2010.

- [24] Hongzhi Guo, Zhi Sun, and Pu Wang. Multiple frequency band channel modeling and analysis for magnetic induction communication in practical underwater environments. *IEEE Transactions on Vehicular Technology*, 66(8):6619–6632, 2017.
- [25] Ian F Akyildiz, Pu Wang, and Zhi Sun. Realizing underwater communication through magnetic induction. *IEEE Communications Magazine*, 53(11):42–48, 2015.
- [26] Yuzhou Li, Shengnan Wang, Cheng Jin, Yu Zhang, and Tao Jiang. A survey of underwater magnetic induction communications: Fundamental issues, recent advances, and challenges. *IEEE Communications Surveys & Tutorials*, 21(3):2466–2487, 2019.
- [27] Vedat Coskun, Busra Ozdenizci, and Kerem Ok. A survey on near field communication (nfc) technology. *Wireless personal communications*, 71(3):2259–2294, 2013.
- [28] Vedat Coskun, Busra Ozdenizci, and Kerem Ok. The survey on near field communication. *Sensors*, 15(6):13348–13405, 2015.
- [29] Han-Joon Kim, Jinho Park, Kyoung-Sub Oh, Jihwan P Choi, Jae Eun Jang, and Ji-Woong Choi. Near-field magnetic induction mimo communication using heterogeneous multipole loop antenna array for higher data rate transmission. *IEEE Transactions on Antennas and Propagation*, 64(5):1952–1962, 2016.
- [30] Yue Ping Zhang. Bit-error-rate performance of intra-chip wireless interconnect systems. *IEEE Communications letters*, 8(1):39–41, 2004.
- [31] Brian A Floyd, Chih-Ming Hung, et al. Intra-chip wireless interconnect for clock distribution implemented with integrated antennas, receivers, and transmitters. *IEEE Journal of Solid-State Circuits*, 37(5):543–552, 2002.

- [32] Ramanathan Subramanian, Benjamin Drozdenko, Eric Doyle, Rameez Ahmed, Miriam Leeser, and Kaushik Roy Chowdhury. High-level system design of ieee 802.11 b standard-compliant link layer for matlab-based sdr. *IEEE Access*, 4:1494–1509, 2016.
- [33] Hussein Kwasme and Sabit Ekin. Rssi-based localization using lorawan technology. *IEEE Access*, 7:99856–99866, 2019.
- [34] Luca Catarinucci, Danilo De Donno, Riccardo Colella, Fabio Ricciato, and Luciano Tarricone. A cost-effective sdr platform for performance characterization of rfid tags. *IEEE Transactions on Instrumentation and Measurement*, 61(4):903–911, 2011.
- [35] Han Ding, Chen Qian, Jinsong Han, Jian Xiao, Xingjun Zhang, Ge Wang, Wei Xi, and Jizhong Zhao. Close-proximity detection for hand approaching using backscatter communication. *IEEE Transactions on Mobile Computing*, 18(10):2285–2297, 2018.
- [36] Florian Galler, Stefan Grebien, Thomas Faseth, Klaus Witrisal, Gottfried Magerl, and Holger Arthaber. Extension of an sdr uhf rfid testbed for mimo and monostatic time of flight based ranging. *IEEE Journal of Radio Frequency Identification*, 1(1):32–38, 2017.
- [37] Wahab Khawaja, Ozgur Ozdemir, and Ismail Guvenc. Uav air-to-ground channel characterization for mmwave systems. In *2017 IEEE 86th Vehicular Technology Conference (VTC-Fall)*, pages 1–5. IEEE, 2017.
- [38] Guojin Zhang, Xuesong Cai, Wei Fan, and Gert Frølund Pedersen. A usrp-based channel sounder for uav communications. *Eucap 2020*, 2019.
- [39] Xuesong Cai, Nanxin Wang, José Rodríguez-Piñeiro, Xuefeng Yin, Antonio Pérez Yuste, Wei Fan, Guojin Zhang, Gert Frølund Pedersen, and Li Tian. Low altitude air-to-ground channel characterization in lte net-

- work. In *2019 13th European Conference on Antennas and Propagation (EuCAP)*, pages 1–5. IEEE, 2019.
- [40] Rickard Nilsson and Jaap van de Beek. Channel measurements in an open-pit mine using usrps: 5g—expect the unexpected. In *2016 IEEE Wireless Communications and Networking Conference*, pages 1–6. IEEE, 2016.
- [41] Sukhyun Hwang, Han-Joon Kim, Kyung Tae Kim, and Ji-Woong Choi. Testbed implementation of near-field magnetic mimo communication system using sdr. In *2016 International Symposium on Antennas and Propagation (ISAP)*, pages 610–611. IEEE, 2016.
- [42] MinKeun Chung, Min Soo Sim, Jaeweon Kim, Dong Ku Kim, and Chan-Byoung Chae. Prototyping real-time full duplex radios. *IEEE Communications Magazine*, 53(9):56–63, 2015.
- [43] Ahmed Hussain, Andrés Alayón Glazunov, Bjarni Þór Einarsson, and Per-Simon Kildal. Antenna measurements in reverberation chamber using usrp. *IEEE Transactions on Antennas and Propagation*, 64(3):1152–1157, 2016.
- [44] Jeremy Hershberger, Thomas Pratt, and Robert Kossler. Automated calibration of a dual-polarized sdr radar. In *2016 IEEE Conference on Antenna Measurements & Applications (CAMA)*, pages 1–4. IEEE, 2016.
- [45] Alexander M Wyglinski, Robin Getz, Travis Collins, and Di Pu. *Software-Defined Radio for Engineers*. Artech House, 2018.
- [46] Robert W Stewart, Kenneth W Barlee, Dale SW Atkinson, and Louise H Crockett. *Software defined radio using MATLAB & Simulink and the RTL-SDR*. Number 1. Strathclyde Academic Media, 2015.
- [47] Adnan Quadri, Huacheng Zeng, and Y Thomas Hou. A real-time mmwave communication testbed with phase noise cancellation. In *IEEE INFOCOM 2019-IEEE Conference on Computer Communications Workshops (INFOCOM WKSHPS)*, pages 455–460. IEEE, 2019.

- [48] Adrian Loch, Arash Asadi, Gek Hong Sim, Joerg Widmer, and Matthias Hollick. mm-wave on wheels: Practical 60 ghz vehicular communication without beam training. In *2017 9th International Conference on Communication Systems and Networks (COMSNETS)*, pages 1–8. IEEE, 2017.
- [49] Adriana Moreno, Jesus Omar Lacruz, and Joerg Widmer. Open source rfnoc-based testbed for millimeter-wave experimentation using usrp software defined radios. In *2020 IEEE International Symposium on Circuits and Systems (ISCAS)*, pages 1–5. IEEE, 2020.
- [50] Martin Danneberg, Roberto Bomfin, Ahmad Nimr, Zhongju Li, and Gerhard Fettweis. Usrc-based platform for 26/28 ghz mmwave experimentation. In *2020 IEEE Wireless Communications and Networking Conference Workshops (WCNCW)*, pages 1–6. IEEE, 2020.
- [51] Marvin K Simon and Mohamed-Slim Alouini. *Digital communication over fading channels*, volume 95. John Wiley & Sons, 2005.
- [52] John G Proakis and Masoud Salehi. *Digital communications*, volume 4. McGraw-hill New York, 2001.
- [53] Bernard Sklar and Fredric J Harris. *Digital communications: fundamentals and applications*, volume 2001. Prentice-hall Englewood Cliffs, NJ, 1988.
- [54] Athanasios Papoulis and H Saunders. Probability, random variables and stochastic processes. 1989.
- [55] Bernard Sklar. Rayleigh fading channels in mobile digital communication systems. i. characterization. *IEEE Communications magazine*, 35(7):90–100, 1997.
- [56] David Tse and Pramod Viswanath. *Fundamentals of wireless communication*. Cambridge university press, 2005.

- [57] Michel Daoud Yacoub, JE Vargas Bautista, and L Guerra de Rezende Guedes. On higher order statistics of the nakagami-m distribution. *IEEE Transactions on Vehicular Technology*, 48(3):790–794, 1999.
- [58] Hirofumi Suzuki. A statistical model for urban radio propagation. *IEEE Transactions on communications*, 25(7):673–680, 1977.
- [59] George Tzeremes and CG Christodoulou. Use of weibull distribution for describing outdoor multipath fading. In *IEEE Antennas and Propagation Society International Symposium (IEEE Cat. No. 02CH37313)*, volume 1, pages 232–235. IEEE, 2002.
- [60] Nikos C Sagias, Dimitris A Zogas, George K Karagiannidis, and George S Tombras. Channel capacity and second-order statistics in weibull fading. *IEEE Communications Letters*, 8(6):377–379, 2004.
- [61] Theodore S Rappaport et al. *Wireless communications: principles and practice*, volume 2. prentice hall PTR New Jersey, 1996.
- [62] Niklas Jaldén, Per Zetterberg, Bjorn Ottersten, Aihua Hong, and Reiner Thoma. Correlation properties of large scale fading based on indoor measurements. In *2007 IEEE Wireless Communications and Networking Conference*, pages 1894–1899. IEEE, 2007.
- [63] Andreas F Molisch. *Wireless communications*, volume 34. John Wiley & Sons, 2012.
- [64] Tapan K Sarkar, Zhong Ji, Kyungjung Kim, Abdellatif Medouri, and Magdalena Salazar-Palma. A survey of various propagation models for mobile communication. *IEEE Antennas and propagation Magazine*, 45(3):51–82, 2003.
- [65] Robert JC Bultitude. A study of coherence bandwidth measurements for

- frequency selective radio channels. In *33rd IEEE Vehicular Technology Conference*, volume 33, pages 269–278. IEEE, 1983.
- [66] MO Al-Nuaimi and Andreas G Siamarou. Coherence bandwidth characterisation and estimation for indoor rician multipath wireless channels using measurements at 62.4 ghz. *IEE Proceedings-Microwaves, Antennas and Propagation*, 149(3):181–187, 2002.
- [67] Mercedes Sanchez Varela and Manuel García Sánchez. Rms delay and coherence bandwidth measurements in indoor radio channels in the uhf band. *IEEE Transactions on Vehicular Technology*, 50(2):515–525, 2001.
- [68] Xiaoming Chen, Per-Simon Kildal, Charlie Orlenius, and Jan Carlsson. Channel sounding of loaded reverberation chamber for over-the-air testing of wireless devices: Coherence bandwidth versus average mode bandwidth and delay spread. *IEEE Antennas and Wireless Propagation Letters*, 8: 678–681, 2009.
- [69] Lin Cheng, Benjamin E Henty, Daniel D Stancil, Fan Bai, and Priyantha Mudalige. Mobile vehicle-to-vehicle narrow-band channel measurement and characterization of the 5.9 ghz dedicated short range communication (dsrc) frequency band. *IEEE Journal on Selected Areas in Communications*, 25(8):1501–1516, 2007.
- [70] Nektarios Moraitis and Philip Constantinou. Measurements and characterization of wideband indoor radio channel at 60 ghz. *IEEE Transactions on Wireless Communications*, 5(4):880–889, 2006.
- [71] Xiaofei Dong and Norman C Beaulieu. Average level crossing rate and average fade duration of selection diversity. *IEEE Communications Letters*, 5(10):396–398, 2001.
- [72] Norman C Beaulieu and Xiaofei Dong. Level crossing rate and average fade

- duration of mrc and egc diversity in ricean fading. *IEEE transactions on communications*, 51(5):722–726, 2003.
- [73] Dejan V Djonin and Vijay K Bhargava. On the influence of the power profile on diversity combining schemes. *IEEE Transactions on Wireless Communications*, 3(5):1854–1861, 2004.
- [74] Paolo Corona, John Ladbury, and Gaetano Latmiral. Reverberation-chamber research-then and now: a review of early work and comparison with current understanding. *IEEE transactions on Electromagnetic Compatibility*, 44(1):87–94, 2002.
- [75] Vignesh Rajamani, Charles F Bunting, and James C West. Stirred-mode operation of reverberation chambers for emc testing. *IEEE Transactions on Instrumentation and Measurement*, 61(10):2759–2764, 2012.
- [76] Kent Rosengren, Per-Simon Kildal, Charlie Carlsson, and Jan Carlsson. Characterization of antennas for mobile and wireless terminals in reverberation chambers: Improved accuracy by platform stirring. *Microwave and Optical Technology Letters*, 30(6):391–397, 2001.
- [77] Christian S Lötbäck Patané, Magnus Franzén, and Charlie Orlenius. Analysis and improvement of reverberation chamber method for characterization of small and terminal antennas. In *Proceedings of the 5th European Conference on Antennas and Propagation (EUCAP)*, pages 1832–1836. IEEE, 2011.
- [78] Christopher L Holloway, Haider A Shah, Ryan J Pirkel, William F Young, David A Hill, and John Ladbury. Reverberation chamber techniques for determining the radiation and total efficiency of antennas. *IEEE transactions on antennas and propagation*, 60(4):1758–1770, 2012.
- [79] David A Hill, Mark T Ma, Arthur R Ondrejka, Bill F Riddle, Myron L Crawford, and Robert T Johnk. Aperture excitation of electrically large,

- lossy cavities. *IEEE transactions on Electromagnetic Compatibility*, 36(3):169–178, 1994.
- [80] Christopher L Holloway, David A Hill, John M Ladbury, Perry F Wilson, Galen Koepke, and Jason Coder. On the use of reverberation chambers to simulate a rician radio environment for the testing of wireless devices. *IEEE transactions on antennas and propagation*, 54(11):3167–3177, 2006.
- [81] Nader Sheikholeslami Alagha and Peter Kabal. Generalized raised-cosine filters. *IEEE transactions on Communications*, 47(7):989–997, 1999.
- [82] Xiang-Gen Xia. A family of pulse-shaping filters with isi-free matched and unmatched filter properties. *IEEE Transactions on Communications*, 45(10):1157–1158, 1997.
- [83] Yan Wang, Kai Shi, and Erchin Serpedin. Non-data-aided feedforward carrier frequency offset estimators for qam constellations: A nonlinear least-squares approach. *EURASIP Journal on Advances in Signal Processing*, 2004(13):1–9, 2004.
- [84] Robert G Gallager. *Information theory and reliable communication*, volume 2. Springer, 1968.
- [85] Andrea Goldsmith. *Wireless communications*. Cambridge university press, 2005.
- [86] Peter J Smith, Sumit Roy, and Mansoor Shafi. Capacity of mimo systems with semicorrelated flat fading. *IEEE Transactions on Information Theory*, 49(10):2781–2788, 2003.
- [87] Hüseyin Ozelik and Claude Oestges. Capacity of diagonally correlated mimo channels. In *2005 IEEE 61st Vehicular Technology Conference*, volume 1, pages 116–120. IEEE, 2005.

- [88] Hüseyin Ozelik and Claude Oestges. Some remarkable properties of diagonally correlated mimo channels. *IEEE transactions on vehicular technology*, 54(6):2143–2145, 2005.
- [89] Arogyaswami Paulraj, Arogyaswami Paulraj Rohit, Rohit Nabar, and Dhananjay Gore. *Introduction to space-time wireless communications*. Cambridge university press, 2003.
- [90] Peter Almers, Fredrik Tufvesson, and Andreas F Molisch. Keyhole effect in mimo wireless channels: Measurements and theory. *IEEE Transactions on Wireless Communications*, 5(12):3596–3604, 2006.
- [91] Georgy Levin and Sergey Loyka. On the outage capacity distribution of correlated keyhole mimo channels. *IEEE Transactions on Information Theory*, 54(7):3232–3245, 2008.
- [92] George Levin and Sergey Loyka. Multi-keyhole mimo channels: Asymptotic analysis of outage capacity. In *2006 IEEE International Symposium on Information Theory*, pages 1305–1309. IEEE, 2006.
- [93] Peter Almers, Fredrik Tufvesson, and Andreas F Molisch. Measurement of keyhole effect in a wireless multiple-input multiple-output (mimo) channel. *IEEE communications letters*, 7(8):373–375, 2003.
- [94] Georgy Levin and Sergey Loyka. From multi-keyholes to measure of correlation and power imbalance in mimo channels: Outage capacity analysis. *IEEE transactions on information theory*, 57(6):3515–3529, 2011.
- [95] Caijun Zhong, Shi Jin, Kai-Kit Wong, and Matthew R McKay. Ergodic mutual information analysis for multi-keyhole mimo channels. *IEEE transactions on wireless communications*, 10(6):1754–1763, 2011.
- [96] Nikola Zlatanov, Zoran Hadzi-Velkov, and George K Karagiannidis. Level crossing rate and average fade duration of the double nakagami-m random

- process and application in mimo keyhole fading channels. *IEEE Communications Letters*, 12(11):822–824, 2008.
- [97] Xiaoming Chen. Antenna correlation and its impact on multi-antenna system. *Progress In Electromagnetics Research*, 62:241–253, 2015.
- [98] Jean-Philippe Kermoal, Laurent Schumacher, Klaus I Pedersen, Preben E Mogensen, and Frank Frederiksen. A stochastic mimo radio channel model with experimental validation. *IEEE Journal on selected areas in Communications*, 20(6):1211–1226, 2002.
- [99] Xiaoming Chen, Per-Simon Kildal, Jan Carlsson, and Jian Yang. Mrc diversity and mimo capacity evaluations of multi-port antennas using reverberation chamber and anechoic chamber. *IEEE Transactions on Antennas and Propagation*, 61(2):917–926, 2012.
- [100] Per-Simon Kildal, Xiaoming Chen, Mattias Gustafsson, and Zhengzhao Shen. Mimo characterization on system level of 5g microbase stations subject to randomness in los. *IEEE Access*, 2:1062–1075, 2014.
- [101] Marco Chiani, Moe Z Win, and Alberto Zanella. On the capacity of spatially correlated mimo rayleigh-fading channels. *IEEE Transactions on Information Theory*, 49(10):2363–2371, 2003.
- [102] Dmitry Chizhik, Farrokh Rashid-Farrokhi, Jonathan Ling, and Angel Lozano. Effect of antenna separation on the capacity of blast in correlated channels. *IEEE Communications letters*, 4(11):337–339, 2000.
- [103] Mischa Schwartz, William R Bennett, and Seymour Stein. *Communication systems and techniques*. John Wiley & Sons, 1995.
- [104] Eduard Jorswieck, Holger Boche, et al. Majorization and matrix-monotone functions in wireless communications. *Foundations and Trends® in Communications and Information Theory*, 3(6):553–701, 2007.

- [105] Xiaoming Chen. On independent platform sample number for reverberation chamber measurements. *IEEE transactions on electromagnetic compatibility*, 54(6):1306–1309, 2012.
- [106] Xiaoming Chen. Experimental investigation of the number of independent samples and the measurement uncertainty in a reverberation chamber. *IEEE transactions on electromagnetic compatibility*, 55(5):816–824, 2013.
- [107] Ryan J Pirkel, Kate A Remley, and Christian S Lötbäck Patané. Reverberation chamber measurement correlation. *IEEE Transactions on Electromagnetic Compatibility*, 54(3):533–545, 2011.
- [108] Mohsen Koohestani, Ahmed Hussain, Antonio A Moreira, and Anja K Skrivervik. Diversity gain influenced by polarization and spatial diversity techniques in ultrawideband. *IEEE Access*, 3:281–286, 2015.
- [109] Qian Xu, Yi Huang, Xu Zhu, Saqer S Alja’afreh, Lei Xing, and Zhihao Tian. Diversity gain measurement in a reverberation chamber without extra antennas. *IEEE Antennas and Wireless Propagation Letters*, 14:1666–1669, 2015.
- [110] CM Jack Wang, Kate A Remley, Ansgar T Kirk, Ryan J Pirkel, Christopher L Holloway, Dylan F Williams, and Paul D Hale. Parameter estimation and uncertainty evaluation in a low rician k-factor reverberation-chamber environment. *IEEE Transactions on Electromagnetic Compatibility*, 56(5):1002–1012, 2014.
- [111] Larry J Greenstein, David G Michelson, and Vinko Erceg. Moment-method estimation of the rician k-factor. *IEEE Communications Letters*, 3(6):175–176, 1999.
- [112] Ali Abdi, Cihan Tepedelenlioglu, Mostafa Kaveh, and Georgios Giannakis. On the estimation of the k parameter for the rice fading distribution. *IEEE Communications letters*, 5(3):92–94, 2001.

- [113] Xiaoming Chen, Bjarni Þór Einarsson, and Per-Simon Kildal. Improved mimo throughput with inverse power allocation—study using usrp measurement in reverberation chamber. *IEEE Antennas and Wireless Propagation Letters*, 13:1494–1496, 2014.
- [114] Per-Simon Kildal, Ahmed Hussain, Xiaoming Chen, Charlie Orlenius, Anton Skarbratt, John Asberg, Tommy Svensson, and Thomas Eriksson. Threshold receiver model for throughput of wireless devices with mimo and frequency diversity measured in reverberation chamber. *IEEE Antennas and Wireless Propagation Letters*, 10:1201–1204, 2011.
- [115] Xiaoming Chen, Per-Simon Kildal, and Jan Carlsson. Estimation of average rician k-factor in reverberation chamber. In *International Symposium on Antennas and Propagation*, 2011.
- [116] Jeanne T Quimby, Richard Candell, Catherine A Remley, David R Novotny, Joseph Diener, Peter B Papazian, Alexandra Curtin, and Galen H Koepke. Nist channel sounder overview and channel measurements in manufacturing facilities. Technical report, 2017.
- [117] Bhargav Gokalgandhi, Prasanthi Maddala, and Ivan Seskar. Implementation of fgpa based channel sounder for large scale antenna systems using rfnoc on usrp platform. In *Proceedings of the GNU Radio Conference*, volume 2, pages 9–9, 2017.
- [118] Mashud Hyder and Kaushik Mahata. Zadoff–chu sequence design for random access initial uplink synchronization in lte-like systems. *IEEE Transactions on Wireless Communications*, 16(1):503–511, 2016.
- [119] Yang Wen, Wei Huang, and Zhongpei Zhang. Czac sequence and its application in lte random access. In *2006 IEEE Information Theory Workshop-ITW’06 Chengdu*, pages 544–547. IEEE, 2006.

- [120] Jiri Blumenstein and Marek Bobula. Coarse time synchronization utilizing symmetric properties of zadoff–chu sequences. *IEEE Communications Letters*, 22(5):1006–1009, 2018.
- [121] Malik Muhammad Usman Gul, Xiaoli Ma, and Sungeun Lee. Timing and frequency synchronization for ofdm downlink transmissions using zadoff–chu sequences. *IEEE Transactions on Wireless Communications*, 14(3):1716–1729, 2015.
- [122] Mohammed H AlSharif, Mohamed Saad, Mohamed Siala, Tarig Ballal, Hatem Boujemaa, and Tareq Y Al-Naffouri. Zadoff-chu coded ultrasonic signal for accurate range estimation. In *2017 25th European Signal Processing Conference (EUSIPCO)*, pages 1250–1254. IEEE, 2017.
- [123] Jesús López-Fernández, Unai Fernández-Plazaola, José F Paris, Luis Díez, and Eduardo Martos-Naya. Wideband ultrasonic acoustic underwater channels: Measurements and characterization. *IEEE Transactions on Vehicular Technology*, 69(4):4019–4032, 2020.
- [124] Dennis Ogbe, David J Love, Matthew Rebholz, and T Patrick Bidigare. Efficient channel estimation for aerial wireless communications. *IEEE Transactions on Aerospace and Electronic Systems*, 55(6):2774–2785, 2019.
- [125] Taehoon Kim, Inkyu Bang, and Dan Keun Sung. An enhanced prach preamble detector for cellular iot communications. *IEEE Communications Letters*, 21(12):2678–2681, 2017.
- [126] Felipe AP De Figueiredo, Fabbryccio ACM Cardoso, Ingrid Moerman, and Gustavo Fraidenraich. Channel estimation for massive mimo tdd systems assuming pilot contamination and frequency selective fading. *IEEE Access*, 5:17733–17741, 2017.
- [127] Nitin Jonathan Myers, Amine Mezghani, and Robert W Heath. Spatial

- zadoff-chu modulation for rapid beam alignment in mmwave phased arrays. In *2018 IEEE Globecom Workshops (GC Wkshps)*, pages 1–6. IEEE, 2018.
- [128] Jens Elofsson and Peter Seimar. Software defined radio based mimo channel sounding. 2016.
- [129] Ruiyuan Tian, Buon Kiong Lau, and Zhinong Ying. Multiplexing efficiency of mimo antennas. *IEEE Antennas and Wireless Propagation Letters*, 10: 183–186, 2011.
- [130] Leilei Liu, Wei Hong, Haiming Wang, Guangqi Yang, Nianzu Zhang, Hongxin Zhao, Jing Chang, Cheng Yu, Xutao Yu, Hongjun Tang, et al. Characterization of line-of-sight mimo channel for fixed wireless communications. *IEEE Antennas and Wireless Propagation Letters*, 6:36–39, 2007.
- [131] Vinko Erceg, Pitschaiah Soma, Daniel S Baum, and Severine Catreux. Multiple-input multiple-output fixed wireless radio channel measurements and modeling using dual-polarized antennas at 2.5 ghz. *IEEE transactions on Wireless Communications*, 3(6):2288–2298, 2004.
- [132] Vikram R Anreddy and Mary Ann Ingram. Capacity of measured ricean and rayleigh indoor mimo channels at 2.4 ghz with polarization and spatial diversity. In *IEEE Wireless Communications and Networking Conference, 2006. WCNC 2006.*, volume 2, pages 946–951. Citeseer, 2006.
- [133] Xiongwen Zhao, Suiyan Geng, Lasse Vuokko, Jarmo Kivinen, and Pertti Vainikainen. Polarization behaviours at 2, 5 and 60 ghz for indoor mobile communications. *Wireless Personal Communications*, 27(2):99–115, 2003.
- [134] Persefoni Kyritsi, Donald C Cox, Reinaldo A Valenzuela, and Peter W Wolniansky. Effect of antenna polarization on the capacity of a multiple element system in an indoor environment. *IEEE Journal on Selected areas in Communications*, 20(6):1227–1239, 2002.

- [135] Jon W Wallace, Michael A Jensen, A Lee Swindlehurst, and Brian D Jeffs. Experimental characterization of the mimo wireless channel: Data acquisition and analysis. *IEEE Transactions on Wireless Communications*, 2(2): 335–343, 2003.
- [136] Ana Vazquez Alejos, Manuel Garcia Sanchez, and Iñigo Cuiñas. Performance analysis of polarization diversity for indoor scenarios at 41.4 ghz and 61.5 ghz. *International Journal of Antennas and Propagation*, 2012, 2012.
- [137] Persefoni Kyritsi and Donald C Cox. Propagation characteristics of horizontally and vertically polarized electric fields in an indoor environment: simple model and results. In *IEEE 54th Vehicular Technology Conference. VTC Fall 2001. Proceedings (Cat. No. 01CH37211)*, volume 3, pages 1422–1426. IEEE, 2001.
- [138] Kent Rosengren and P-S Kildal. Radiation efficiency, correlation, diversity gain and capacity of a six-monopole antenna array for a mimo system: theory, simulation and measurement in reverberation chamber. *IEE Proceedings-Microwaves, Antennas and Propagation*, 152(1):7–16, 2005.
- [139] Juan F Valenzuela-Valdés, Antonio M Martínez-González, and David A Sánchez-Hernández. Diversity gain and mimo capacity for nonisotropic environments using a reverberation chamber. *IEEE Antennas and Wireless Propagation Letters*, 8:112–115, 2009.
- [140] Yaoqing Yang, Guanghan Xu, and Hao Ling. An experimental investigation of wideband mimo channel characteristics based on outdoor non-los measurements at 1.8 ghz. *IEEE transactions on antennas and propagation*, 54(11):3274–3284, 2006.
- [141] Adel MD Turkmani, Andrew A Arowojolu, PA Jefford, and CJ Kellett. An experimental evaluation of the performance of two-branch space and po-

larization diversity schemes at 1800 mhz. *IEEE Transactions on Vehicular Technology*, 44(2):318–326, 1995.

- [142] Carl B Dietrich, Kai Dietze, J Randall Nealy, and Warren L Stutzman. Spatial, polarization, and pattern diversity for wireless handheld terminals. *IEEE transactions on antennas and propagation*, 49(9):1271–1281, 2001.