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Experimental Analysis of Massive MIMO Antennas in O-RAN Systems

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Dissertation for Doctor of Philosophy in Information and Communication
Engineering

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The Jury committee is composed of:

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systèmes O-RAN**

Présenté par **RIZK Christ**

Thèse de doctorat en Sciences de l'Information et de la Communication

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Abstract

The evolution of wireless networks toward openness, scalability, and intelligence is driven by two key technologies: Open Radio Access Network (ORAN) and massive Multi-Input Multi-Output (MIMO). ORAN introduces a paradigm shift in the design and operation of radio access networks by enabling disaggregation, virtualization, and interoperability through open interfaces and standardized control architectures. Massive MIMO, in turn, scales traditional MIMO systems by orders of magnitude, employing large antenna arrays to serve many users simultaneously and achieve unprecedented levels of spectral and energy efficiency. Together, these technologies form the foundation for future broadband networks that are flexible, energy efficient, and capable of intelligent data-driven optimization.

The integration of massive MIMO within ORAN architectures offers a promising path toward intelligent, adaptive, and energy-efficient networks, where Artificial Intelligence (AI)/Machine Learning (ML)-driven RAN Intelligent Controller (RIC)s dynamically optimize beamforming, resource allocation, and interference management. However, challenges remain in computational complexity, synchronization, distributed learning, and hardware-induced impairments such as Out Of Band (OOB) emissions in Massive MIMO. Addressing these issues will be crucial to realizing the full potential of open, programmable, and virtualized wireless systems. The insights presented here aim to foster innovation toward the development of secure, efficient, and intelligent 6G networks that connect people and things through a unified, software-defined, and data-powered infrastructure.

This work provides an overview of the ORAN specifications, architectures, and operations, highlighting key components such as Radio Unit (RU), Distributed Unit (DU), and Centralized Unit (CU). It examines the 7.2 Fronthaul Interface (FHI) that supports beamforming and Massive MIMO operations supported by experimental testbeds that validate theoretical models in realistic settings.

A particular focus of this thesis is the study of OOB emissions in massive MIMO systems, which arise from nonlinear hardware behavior and can impact spectral efficiency and adjacent channel performance. Understanding their spatial characteristics is essential for improving system design, ensuring regulatory compliance, and guiding future standardization efforts, especially as networks evolve toward higher-frequency 6G bands where such effects become more prominent. In addition, OOB emissions play a critical role in band coexistence, as they can cause harmful interference to neighboring systems operating in adjacent or nearby frequency bands, such as legacy cellular networks, satellite services, or unlicensed technologies. Therefore, a characterization of these emissions is necessary to enable efficient spectrum sharing, protect incumbent services, and support the reliable coexistence of heterogeneous wireless systems in increasingly congested spectral environments.

Finally, we highlight that most of our work has been validated using OpenAirInterface (OAI).

Résumé

L'évolution des réseaux sans fil vers l'ouverture, la scalabilité et l'intelligence est impulsée par deux technologies clés : l'ORAN et le Massive MIMO. ORAN introduit un changement de paradigme dans la conception et l'exploitation des réseaux d'accès radio en permettant la désagrégation, la virtualisation et l'interopérabilité grâce à des interfaces ouvertes et des architectures de contrôle standardisées. Le Massive MIMO, quant à lui, étend les systèmes MIMO traditionnels par ordres de grandeur, en utilisant de grandes matrices d'antennes pour desservir simultanément de nombreux utilisateurs et atteindre des niveaux de performance spectrale et énergétique sans précédent. Ensemble, ces technologies constituent la base des futurs réseaux haut débit flexibles, économes en énergie et capables d'optimisations intelligentes et basées sur les données.

L'intégration du Massive MIMO au sein des architectures ORAN offre une voie prometteuse vers des réseaux intelligents, adaptatifs et économes en énergie, où des RICs pilotés par l'IA/ML optimisent dynamiquement le beamforming, l'allocation des ressources et la gestion des interférences. Cependant, des défis persistent en termes de complexité computationnelle, de synchronisation, d'apprentissage distribué et d'imperfections matérielles telles que les émissions hors bande (OOB) dans le Massive MIMO. La résolution de ces problèmes sera cruciale pour réaliser pleinement le potentiel des systèmes sans fil ouverts, programmables et virtualisés. Les analyses présentées ici visent à encourager l'innovation pour le développement de réseaux 6G sécurisés, efficaces et intelligents, connectant les personnes et les objets via une infrastructure unifiée, définie par logiciel et alimentée par les données.

Ce travail fournit un aperçu des spécifications, architectures et opérations ORAN, en mettant en lumière des composants clés tels que les RU, DU et CU. Il examine l'interface fronthaul 7.2 (FHI) qui supporte le beamforming et les opérations Massive MIMO, appuyées par des bancs d'essai expérimentaux validant les modèles théoriques dans des conditions réalistes.

Un accent particulier de cette thèse est mis sur l'étude des émissions OOB dans les systèmes Massive MIMO, qui résultent du comportement non linéaire du matériel et peuvent affecter l'efficacité spectrale et la performance des canaux adjacents. Comprendre leurs caractéristiques spatiales est essentiel pour améliorer la conception des systèmes, garantir la conformité réglementaire et orienter les futures normalisations, en particulier à mesure que les réseaux évoluent vers les bandes de fréquences plus élevées de la 6G, où ces effets deviennent plus prononcés. De plus, les émissions OOB jouent un rôle crucial dans la coexistence des bandes de fréquences, car elles peuvent engendrer des interférences nuisibles pour les systèmes voisins opérant dans des bandes adjacentes ou proches, tels que les réseaux cellulaires existants, les services satellitaires ou les technologies sans licence. Une caractérisation approfondie de ces émissions est donc nécessaire afin de permettre un partage efficace du spectre, de protéger les services incumbents et d'assurer la coexistence fiable de systèmes de communication hétérogènes dans des environnements spectraux de plus en plus congestionnés.

Enfin, nous soulignons que la majorité de notre travail a été validée en utilisant OAI.

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Glossary

2G	Second-Generation
3GPP	Third Generation Partnership Project
4G	Fourth-Generation
5G	Fifth-Generation
5G NR	Fifth-Generation New Radio
5GC	5G Core
6G	Sixth-Generation
AAS	Active Antenna Systems
ACLR	Adjacent Channel Leakage Ratio
AI	Artificial Intelligence
AMF	Access and Mobility management Function
ANFR	Agence Nationale des Fréquences
API	Application Programming Interface
ASIC	Application Specific Integrated Circuits
AUSF	Authentication Server Function
BBU	Base Band Unit
BF	Beamforming
BM	Beam Management
BS	Base Station
C plane	Control Plane
CAT A	Category A
CAT B	Category B
CEPT	Conférence Européenne des administrations des Postes et Télécommunications
CN	Core Network
COTS	Commercial Off-The-Shelf
CPRI	Common Public Radio Interface
CPU	Central Processing Unit
CQI	Channel Quality Indicator
CSI	Channel State Information
CSI-RS	Channel State Information Reference Signal
CU	Centralized Unit
CU-CP	Centralized Unit- Control Plane
CU-UP	Centralized Unit- User Plane
DB	Digital Beamforming
DBT	Digital Beamforming Table
DC	Direct Current

DCI	Downlink Control Information
DFE	Digital Front End
DL	Downlink
DPD	Digital Predistortion
DPDK	Data Plane Development Kit
DU	Distributed Unit
eAxC	Antenna Carrier
ECC	Electronic Communications Committee
eCPRI	Enhanced Common Public Radio Interface
EIRP	Effective Isometric Radiated Power
eMBB	enhanced Mobile Broadband
eNB	evolved Node B
EVM	Error Vector Magnitude
FAPI	Functional Application Platform Interface
FDD	Frequency Division Duplexing
FFT	Fast Fourier transform
FH	Fronthaul
FHI	Fronthaul Interface
FIR	Finite Impulse Response
FPGA	Field Programmable Gate Arrays
GM	Grand Master
GMP	Generalized Memory Polynomial
gNB	Base-Station
GNSS	Global Navigation Satellite System
HARQ	Hybrid Automatic Repeat Request
IFFT	Inverse Fast Fourier transform
IID	Identically Distributed
IOT	Internet Of Things
IQ	In Phase and Quadrature
KPM	Key Performance Measurements
LLS	Lower Layer Split
LoS	line-of-sight
LTE	Long-Term Evolution
MAC	Medium Access Control
MCS	Modulation and Coding Scheme
MIMO	Multi-Input Multi-Output
ML	Machine Learning
MMSE	Minimum Mean Squared Error
mmWave	millimeter Wave
MRC	Maximum Ratio Combining
MRT	Maximum Ratio Transmission
ms	milliseconds
MSE	Mean Squared Error
MU MIMO	Multi User MIMO
near RT RIC	near Real Time RAN Intelligent Controller
NFV	Network Function Virtualization
NIC	Network Intelligent Controller

non RT RIC	non Real Time RAN Intelligent Controller
NR	New Radio
NRF	Network Repository Function
nrUE	NR User Equipment
OAI	OpenAirInterface
OBUE	Operating Band Unwanted Emissions
OFDM	Orthogonal Frequency Division Multiplexing
OOB	Out Of Band
ORAN	Open Radio Access Network
OSC	ORAN Software Community
OTA	Over The Air
PA	Power Amplifier
PAPR	Peak to Average Power Ratio
PBCH	Physical Broadcast Channel
PDCCH	Physical Downlink Control Channel
PDCP	Packet Data Convergence Protocol
PDSCH	Physical Downlink Shared Channel
PDU	Packet Data Unit
PHY	Physical
PM	Precoding Matrix
PMI	Precoding Matrix Indicator
PRACH	Physical Random Access Channel
PRB	Physical Ressource Blocks
PRTC	Primary Reference Time Clock
PSD	Power Spectral Density
PTP	Precision Time Protocol
PUCCH	Physical Uplink Control Channel
PUSCH	Physical Uplink Shared Channel
Qoe	Quality of experience
Qos	Quality of service
QPSK	Quadrature Phase Shift Keying
RACH	Random Access Channel
RAN	Radio Access Network
RB	Resource Block
RBW	Resolution Bandwidth
RE	Resource Element
RF	Radio Frequency
RI	Rank Indicator
RIC	RAN Intelligent Controller
RLC	Radio Link Control
RMS	Root Mean Square
RRC	Radio Resource Control
RRH	Remote Radio Head
RRU	Remote Radio Unit
RSRP	Reference Signal Received Power
RU	Radio Unit
SA	Standalone

SCS	Subcarrier Spacing
SDAP	Service Data Adaption Protocol
SDN	Software Defined Networking
SDNR	Signal to Distortion Noise Ratio
SDR	Software-Defined Radio
SEM	Spectrum Emission Mask
SIB	System Information Block
SINR	Signal-to-Interference-Noise-Ratio
SMF	Session Management Function
SMO	Service Management and Orchestration
SNR	Signal-to-Noise Ratio
spStream	Spatial Stream
SRS	Sounding Reference Signals
SSB	Synchronization Signal Block
SU MIMO	Single User MIMO
T-GM	Telecom Grand Master
TCI	Transmission Configuration Indicator
TCO	Total cost of Ownership
TDD	Time Division Duplexing
TIP	Telecom Infra Project
TRBs	Transport Blocks
TRP	Total Radiated Power
TTI	Transmission Time Interval
TX/RX	Transmit-Receive
U plane	User Plane
UDM	Unified Data Management
UE	User Equipment
UPF	User Plane Function
URLLC	Ultra Reliable Low Latency Communications
USRP	Universal Software Radio Peripheral
VBW	Video Bandwidth
ZF	Zero Forcing

Chapter 1

Introduction

Cellular networks have advanced rapidly since the introduction of Second-Generation (2G), each generation bringing significant technical complexity[1]. Traditionally, while operators could source their core network and Radio Access Network (RAN) from different vendors, the interoperability between RAN components from different suppliers was often limited in favor of adding more features. As a result, modern RAN solutions are usually provided by a single vendor, making it difficult to mix and match radio and BBU.

In contrast, multi-vendor setups and standardized interfaces have long been used to connect radios and core network elements from different providers. The Open RAN aims to further extend this concept by standardizing the interfaces within the Base Station (BS) itself, allowing operators greater flexibility to combine components from multiple vendors[2].

1.1 History of ORAN

XRAN FORUM + C-RAN ALLIANCE = O-RAN ALLIANCE

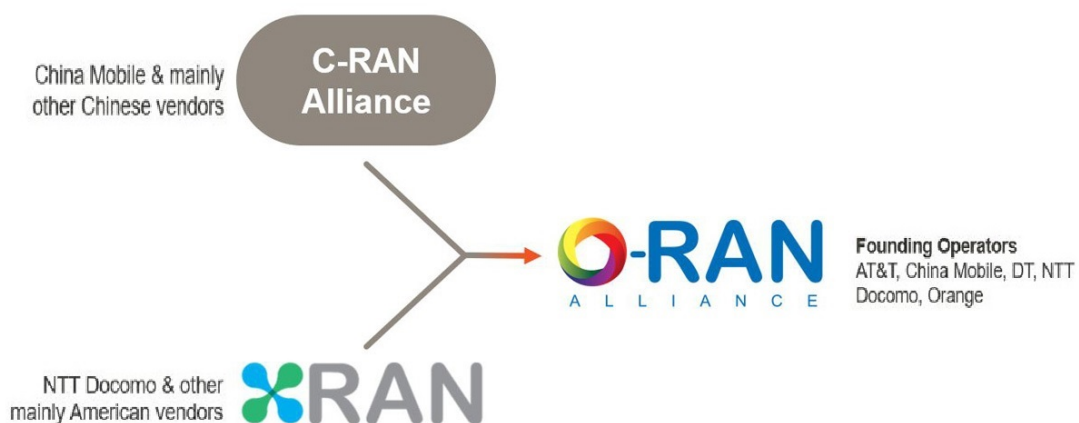


Figure 1.1: History of Open RAN.

Open RAN is still in its early stages, but its origins go back further than many realize. The initial ideas for introducing more interfaces within the base station can be traced to early discussions in the Femtocell (later Small Cell) Forum. Although these early initiatives did not gain significant traction, they

helped lay the groundwork for later developments in Open RAN, particularly as supply chain challenges emerged when certain vendors were restricted from core networks[3].

The concept of open interfaces and open software had existed for some time, but efforts to formalize Open RAN standards began around 2016 with the establishment of the xRAN Forum, which brought together several major mobile network operators. During the same period, the C-RAN Alliance, which focused on virtualizing the radio access network, merged with xRAN in 2018 to form the ORAN Alliance, the main body responsible for developing Open RAN standards[4]. Additionally, the Telecom Infra Project (TIP), also launched in 2016, aimed to accelerate the commercialization of open radio solutions[5][6].

The ORAN Alliance was officially announced at Mobile World Congress in February 2018, with its first working group meetings and symposium held later that year. Its guiding principles focus on driving open, interoperable interfaces and promoting RAN virtualization to create a more flexible and diversified telecom ecosystem.

For that, deploying ORAN provides a clear path to a fully programmable, intelligent, and multi-vendor RAN. Through an intelligent RIC, operators can manage and optimize RAN functions using integrated applications. Specifically, RIC is a software-defined component of the Open RAN architecture that enables service providers, vendors, and third parties to deploy applications. These applications are available via an “app store” and help automate and optimize large-scale RAN operations. They also support innovative use cases that reduce the Total cost of Ownership (TCO) for mobile operators while improving the Quality of experience (Qoe) for end users[7].

1.2 Motivation

1.2.1 Driver factors towards ORAN

The concept of an ORAN is based on two key principles[8]:

- **Disaggregation:** Network functionality is implemented in software and decoupled from specialized hardware. In practice, this means that the software can run on widely available, general purpose hardware platforms, following modern virtualization and containerization approaches.
- **Modular functional blocks with well-defined interfaces:** The network is divided into discrete elements, such as RU, DU, and CU, which communicate through open, standardized interfaces. This design enables multi-vendor interoperability and flexible network deployments[9].

These concepts, disaggregation and modularity, may seem straightforward, but their implications for telecom operators are profound.

Disaggregation allows operators to dynamically scale resources, optimize network performance regionally, and adopt more cost-efficient deployment models.

Modularity reduces dependence on a single supplier and mitigates vendor lock-in. Together, these principles enable operators to build networks that are more adaptable, resilient, and future-ready [8].

The push for Open RAN is driven by several factors[10] and are shown in Figure 1.2:

- **Reducing vendor dependence:** Many mobile operators aim to avoid being locked into proprietary solutions from a small number of equipment vendors. Open RAN allows equipment sourcing and software from multiple suppliers, fostering competition and giving operators greater control over costs and innovation[11].
- **New entrants and ecosystem expansion:** Companies from software and cloud industries see Open RAN as an opportunity to enter the telecom sector. Using expertise in software, cloud computing, and AI, they can provide innovative network solutions, enrich the ecosystem, and accelerate the evolution of mobile networks[12].

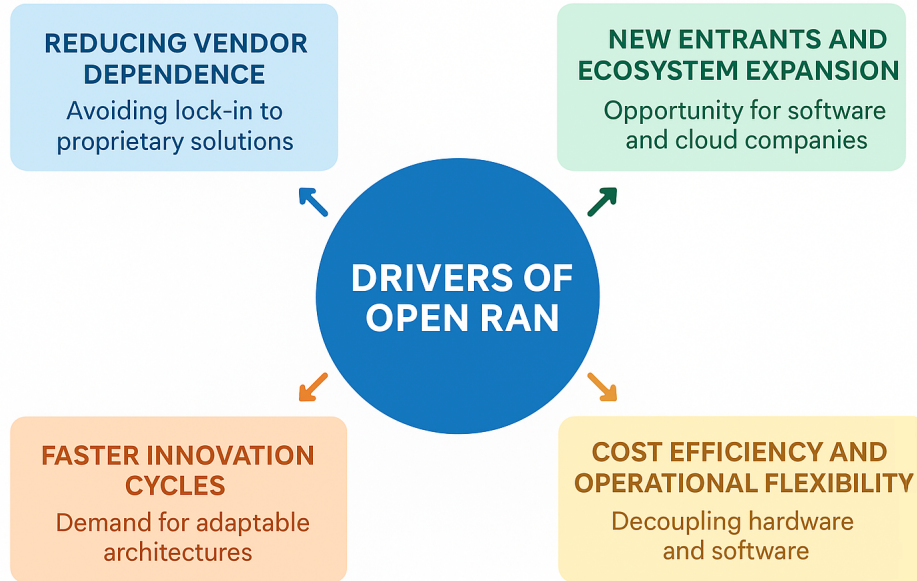


Figure 1.2: Drivers towards ORAN.

- **Faster innovation cycles:** Connectivity, services, and user expectations are evolving rapidly. Traditional network architectures can be slow to adapt, while the flexible and programmable design of Open RAN allows operators to quickly introduce new features, optimize performance, and respond to market demands.
- **Cost efficiency and operational flexibility:** By decoupling hardware and software, operators can mix and match components to optimize costs. They can also virtualize network functions, run them in the cloud, and deploy them where needed, improving network efficiency and reducing total cost of ownership.
- **Support for advanced technologies:** Open RAN provides a foundation for the integration of AI, ML and intelligent automation. Operators can use real-time data to optimize traffic, improve Qoe, and enable new services such as Ultra Reliable Low Latency Communications (URLLC) and massive Internet Of Things (IOT) deployments[13].

1.2.2 Economic advantages of Open RAN

Component	Proprietary RAN	Open RAN (ORAN)
Radio Units (RUs)	Higher cost, limited suppliers	30% lower cost, multiple suppliers
Software	Proprietary and expensive	30% lower cost, open and competitive market
Baseband / Server	Custom hardware (BBUs)	Lower-cost x86 commercial servers

Table 1.1: Comparison of Component Costs: Proprietary RAN vs. Open RAN

The TCO analysis in table 1.1 demonstrates that the ORAN architecture provides substantial financial advantages compared to traditional proprietary RAN. The key areas of cost reduction are outlined below [14][15]:

1.2.2.1 Lower Radio Unit costs

Open RAN radios are expected to be approximately 30% less expensive than proprietary radios over time. This cost reduction is driven by the emergence of multiple manufacturers producing equipment based on standardized radio specifications. In an open ecosystem, increased competition leads to lower prices and reduced profit margins for open, white-box radios compared to proprietary alternatives.

1.2.2.2 Lower software costs

Open RAN software is projected to be around 30% less expensive than proprietary RAN software. This advantage arises from a combination of open-source development and the presence of multiple independent software vendors within the open RAN ecosystem. As a result, operators benefit from lower licensing costs, increased flexibility, and reduced vendor lock-in.[15]

1.2.2.3 Hardware cost advantages through COTS servers

Open RAN uses standard commercial servers based on x86, which are expected to become more cost-efficient than proprietary BBU servers. Continuous improvements in processor performance, following Moore's Law, enable these servers to deliver increasing computational power at declining prices. This shift from specialized to commoditized hardware significantly reduces capital expenditure.

By combining reductions in hardware and software costs, Open RAN reduces significantly overall TCO. This results in a more flexible, competitive, and cost-efficient network infrastructure that better supports long-term scalability and innovation[16].

1.2.3 ORAN architectural principles

The vision of ORAN stems from long standing efforts to make communication networks open, programmable, and interoperable. Drawing on the principles of Software Defined Networking (SDN) and Network Function Virtualization (NFV), Open RAN aims to transform the traditional, closed RAN ecosystem into a flexible, software-driven architecture. Its foundation lies in three key concepts: virtualization, disaggregation, and open interfaces[17] and they are represented in Figure 1.3:

1.2.3.1 Virtualization

Virtualization enables RAN functions to be deployed as software instances on shared, cloud-based infrastructure. Within ORAN, this is realized through the O-Cloud, a virtualized computing environment encompassing hardware resources, hypervisors, orchestration layers, and automation tools.

The O-Cloud provides several key benefits:

- **Decoupling of hardware and software:** RAN functions can run on general purpose hardware without being tied to proprietary equipment.
- **Standardization of infrastructure capabilities:** Ensures interoperability across different vendors and deployment scenarios.
- **Multi-tenancy support:** Multiple operators or network slices can share the same physical resources efficiently.
- **Dynamic resource scaling and energy efficiency:** Computing resources can be allocated or released based on real-time traffic demand, reducing overall power consumption.

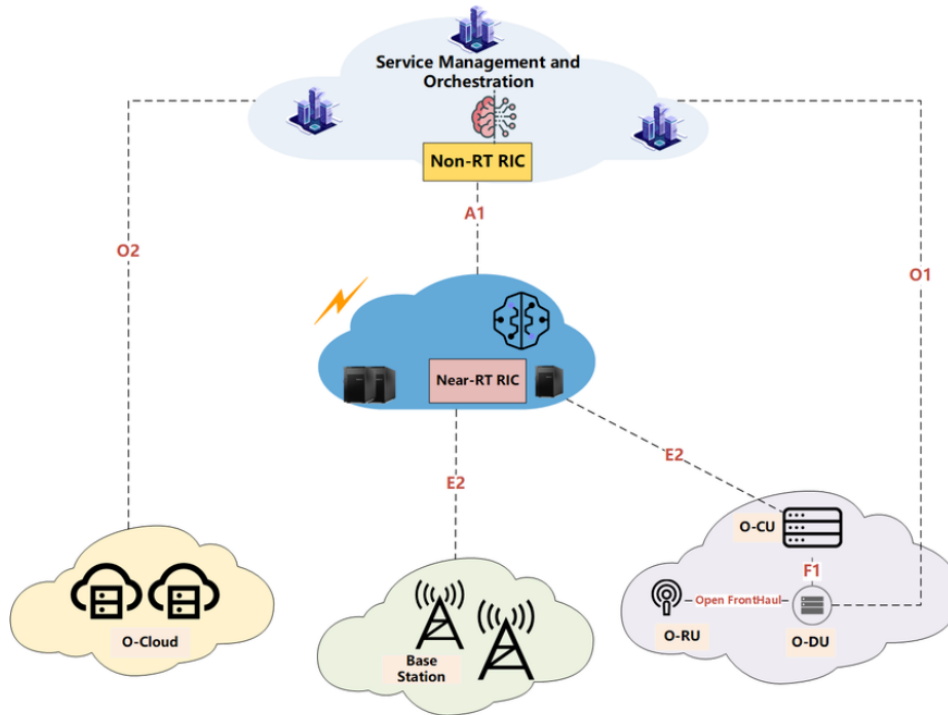


Figure 1.3: ORAN blocks and open interfaces.

1.2.3.2 Disaggregation

One of the main ideas of Open RAN is disaggregation, which means breaking a traditional base station into smaller, separate units that can work together. This makes the network more flexible and easier to manage.

ORAN splits a base station into three parts[18][19]:

- **O-Central Unit (O-CU):** Handles higher-level tasks, like managing connections and data flows.
- **O-Distributed Unit (O-DU):** Takes care of lower-level tasks, like sending and receiving data over the air.
- **O-Radio Unit (O-RU):** Deals with the actual radio signals and antennas.

This approach allows operators to mix and match components from different vendors and place them in different locations, for example putting some parts in a central data center and others near the antennas. Disaggregation makes networks more flexible, easier to upgrade, and ready for intelligent management.

1.2.3.3 Open interfaces

The third key principle of ORAN is the use of open interfaces. These are standard ways for the different parts of the network to communicate, even if they come from different vendors. Open interfaces make the network less like a “black box” and more like a system you can control and improve.

Some key interfaces in ORAN include:

- **E2 Interface:** Connects the near-real-time controller to the base station parts, allowing it to make adjustments quickly.

- **A1 Interface:** Connects the higher-level controller to the near-real-time controller, providing guidance and policies.
- **O1 and O2 Interfaces:** Help manage and orchestrate the network and its virtualized resources.
- **Open Fronthaul Interface:** Connects the radio unit to the distributed unit in a standard way, so hardware from different vendors can work together.

With open interfaces, operators can easily mix equipment, automate network tasks, and introduce new features without being locked into one vendor's system. This helps create flexible, efficient, and future-ready networks.

1.2.4 5G ORAN disaggregated BBU

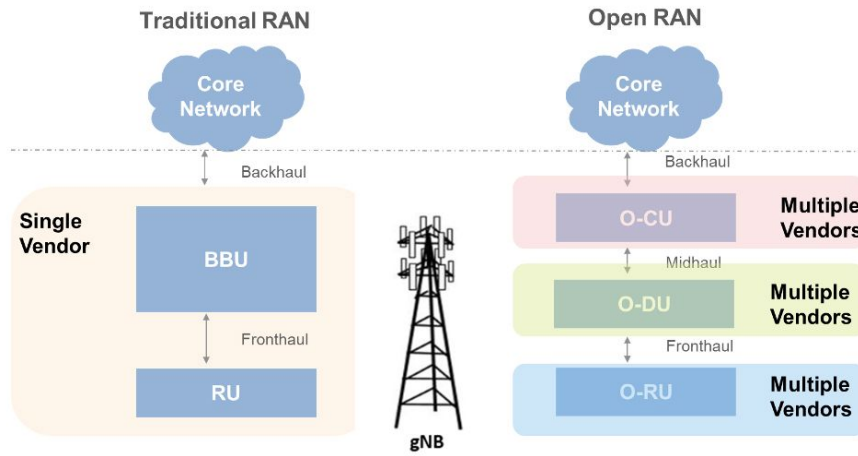


Figure 1.4: Disaggregation splitting traditional BBU into interoperable cloud functionalities connected through open interfaces.

As shown in Figure 1.4, in traditional RAN, the system architecture is vertically integrated, which means that a single vendor typically provides the entire stack, including both BBU and RU. This approach limits flexibility and results in vendor lock-in, as operators depend on proprietary solutions for upgrades, maintenance, and expansion.

The ORAN paradigm addresses these limitations through the disaggregation of network functions and the introduction of standardized open interfaces. This allows multi-vendor interoperability, enabling network operators to mix and match the best hardware and software components from different suppliers, fostering a more competitive and innovative ecosystem.

Furthermore, the adoption of cloud-native virtualization in ORAN architectures improves efficiency through dynamic resource allocation and elastic scaling. Virtualized network functions can be deployed on COTS hardware, reducing both capital and operational expenditures. Maintenance and upgrades are simplified through software-defined updates, improving agility, and reducing downtime. In general, ORAN disaggregation empowers operators with greater flexibility, cost efficiency, and the ability to accelerate innovation within the 5G ecosystem[17].

1.2.5 Beamforming

The rapid growth of mobile data traffic driven by applications such as high definition video, virtual and augmented reality, and IOT is putting heavy pressure on wireless networks. Future communication

systems like Fifth-Generation (5G) and Sixth-Generation (6G) must support very high data rates, low latency, and reliable connections for millions of devices. To achieve these goals, networks need major improvements in spectral efficiency, coverage, signal quality, and capacity[20].

Beamforming is an important technique that helps achieve these improvements. It works by directing radio signals towards desired users and reducing interference from other directions. This increases Signal-to-Interference-Noise-Ratio (SINR) efficiently[21].

At high frequencies, especially in millimeter Wave (mmWave) and massive MIMO systems, signals weaken rapidly and are more easily blocked. Beamforming helps overcome this problem by sending signals in focused beams instead of spreading power in all directions. This makes it easier to maintain strong and stable connections.

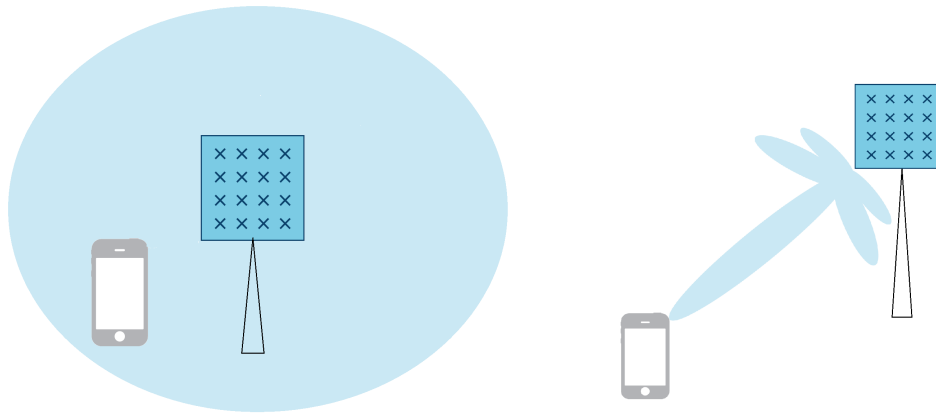


Figure 1.5: Difference between omnidirectional transmission and beamforming

Another important motivation is energy efficiency. Because beamforming concentrates energy where it is needed, as seen in Figure 1.5, the same coverage can be achieved with less transmission power[22]. Beamforming operates by carefully controlling the amplitude and phase of the signals transmitted (or received) by each antenna element in an array. Through constructive and destructive interference of these signals, the antenna array can steer its radiation pattern toward a specific direction. This allows the transmitter to focus energy toward a target receiver, creating a highly directive beam rather than broadcasting signals equally in all directions, as in omnidirectional transmission.

Compared to omnidirectional antennas, which radiate power uniformly and thus waste energy in unwanted directions, beamforming offers several advantages. It enhances signal strength at the intended receiver, reduces interference to and from other users, and improves spectral efficiency by enabling multiple data streams to coexist in the same frequency band. The disaggregation of the traditional BBU station in the CU, DU, and RU allows the beamforming process to be distributed across multiple functional entities, in the selected functional split and deployment scenario.

Beamforming will be discussed further on the basis the ORAN and Third Generation Partnership Project (3GPP) specifications in the State of the Art section in 1.3.

1.2.6 Massive MIMO

Massive MIMO uses a very large number of antennas at the base station to significantly improve the performance of wireless networks. The main idea is to exploit the spatial dimension of the wireless channel to improve spectral efficiency, energy efficiency, and link reliability[21]. Through beamforming, the base station can focus its transmitted energy toward specific users instead of radiating in all directions, strengthening the desired signal, and minimizing interference to others.

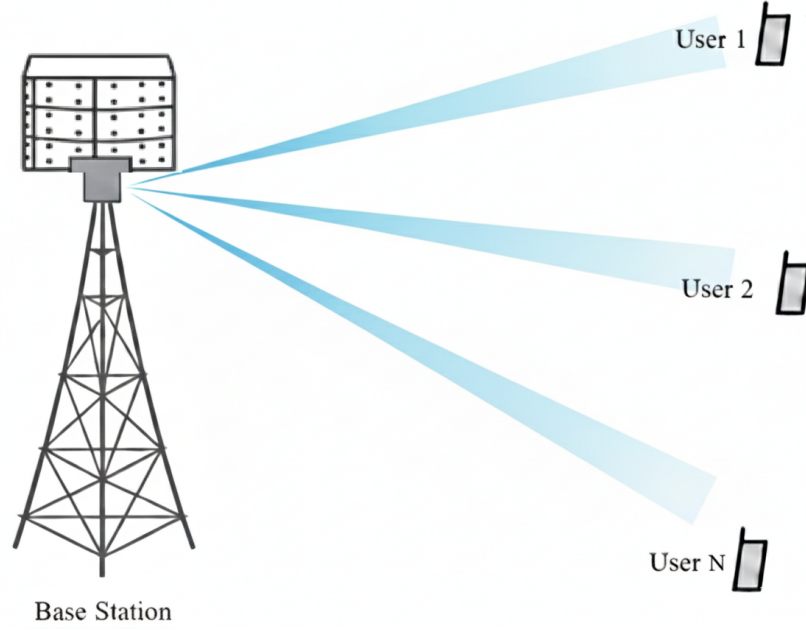


Figure 1.6: Massive MIMO system in downlink and uplink.

This capability enables spatial multiplexing, where multiple users are served simultaneously within the same time and frequency resources by transmitting separate data streams along distinct spatial directions. As a result, Massive MIMO achieves higher throughput, better coverage, and more efficient use of the spectrum compared to conventional MIMO systems[21].

Massive MIMO in Open RAN uses large antenna arrays in the RU to perform beamforming and spatial multiplexing, allowing multiple data streams to be sent to different users simultaneously on the same time-frequency resources. The DU/CU handle baseband and signal processing, while the RU manages the antennas through open interfaces such as ORAN split 7.2x fronthaul interface. This setup increases capacity, coverage, and spectral efficiency, but requires high fronthaul bandwidth, tight synchronization, and coordination often optimized by the near Real Time RAN Intelligent Controller (near RT RIC). Massive MIMO will be discussed further based on the the ORAN and 3GPP specifications in the State of the Art section in 1.3.

1.3 State-of-the-Art and Literature Survey

In this section, the ORAN architecture is discussed in detail. The main functional blocks, including RU, DU, and CU, are described to illustrate the concept of baseband disaggregation. Furthermore, the ORAN 7.2 functional split option is explained to highlight how processing tasks are divided between RU and DU, enabling flexibility, interoperability, and efficient utilization of resources in 5G networks. We also discuss beamforming, the technique that controls the amplitude and phase shift across antenna ports to steer and shape the radiation beam pattern. We examine as well MIMO technology, which uses multiple transmit and receive antennas to exploit spatial multiplexing and transmit several data streams simultaneously.

1.3.1 ORAN architecture

1.3.1.1 The Traditional 4G LTE eNB Split and the Evolution of the Fronthaul Interface

In LTE, RAN was built around the concept of eNB. The eNB is a monolithic base station that integrates both the digital baseband processing and the transmission Radio Frequency (RF) components. To enable flexibility in deployment, vendors and operators commonly divided eNB into two physical entities: the BBU and Remote Radio Unit (RRU), also called the Remote Radio Head (RRH).

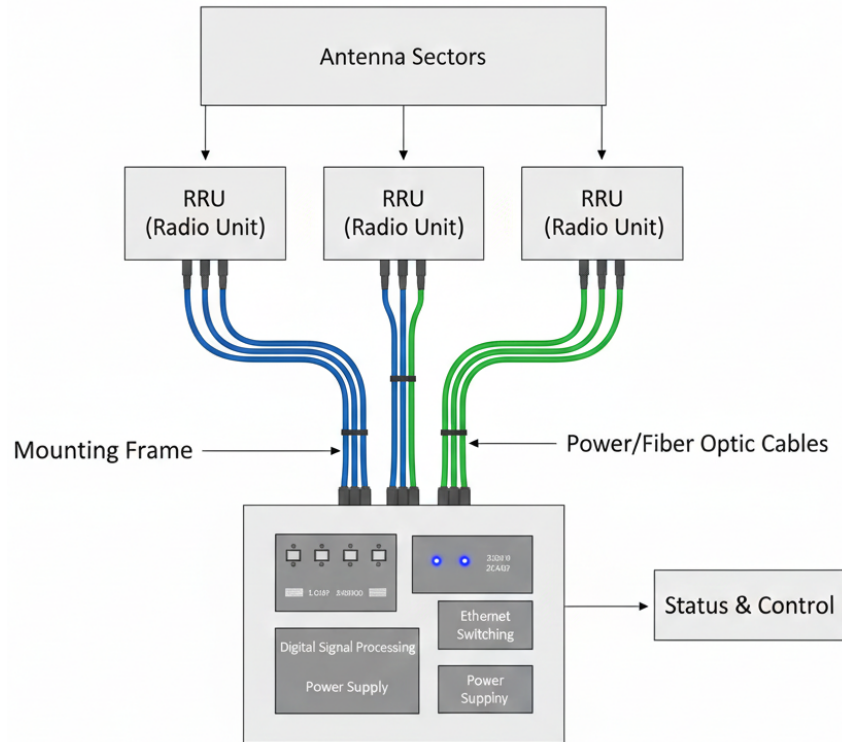


Figure 1.7: eNB block diagram.

As we can see in Figure 1.7, BBU is located in a central site or equipment room. It is in charge of most of the signal processing [23][24]:

- Higher-layer tasks such as Radio Resource Control (RRC) and data handling in Layer 2 such as Packet Data Convergence Protocol (PDCP), Radio Link Control (RLC) and Medium Access Control (MAC)[25].
- Almost all of Physical (PHY), including channel coding, modulation, and the mathematical operations that build the Orthogonal Frequency Division Multiplexing (OFDM) signal, including Fast Fourier transform (FFT) and Inverse Fast Fourier transform (IFFT)[25].

At the end of this chain, the BBU produces digital In Phase and Quadrature (IQ) samples, which are the numerical representation of the radio waveform.

The RRU is installed near the antenna, for example, at the top of a tower in the cell center. Its role is much simpler:

- Converts the digital IQ samples into analog signals (D/A) for transmission and converts the received signals back to digital form (A/D).

- It handles RF-specific tasks such as frequency up-conversion, down-conversion, filtering, and amplification.

This split allowed operators to keep the heavy processing centralized in “BBU hotels”, while only deploying light and cost efficient radio heads in the field[26].

The link between BBU and RRU in LTE is based on Common Public Radio Interface (CPRI). CPRI is a standardized digital interface that continuously sends IQ samples from BBU to RRU over fiber [27].

The main features of CPRI are the following:

- **High bandwidth use:** the required capacity grows directly with the carrier bandwidth and the number of antennas. More MIMO antennas or higher channel bandwidth quickly multiply the needed data rate.
- **Inefficient usage:** CPRI sends IQ data at a constant rate, even when there is little or no user traffic.
- **Strict timing:** the delay and jitter on the fiber must be extremely small (typically less than 100 μ s), so synchronization is very demanding.

For LTE systems with limited bandwidth (e.g., 20 MHz channels and a small number of antennas), this was acceptable. However, with the introduction of 5G features such as massive MIMO and wideband channels (up to 100 MHz or more), fronthaul requirements exploded into the range of tens or even hundreds of gigabits per second. This made the traditional CPRI approach impractical for next generation deployments.

1.3.1.2 Evolution to 5G NR architecture with ORAN

To address these limitations, 3GPP studied multiple split options for the New Radio (NR) architecture [28]. These range from higher layer splits, such as Option 2 (between PDCP and RLC) to lower-layer splits such as Option 7.2x.

The ORAN Alliance, established in 2018, has since defined an open fronthaul specification based on Option 7.2x. This interface adopts the packetized approach of Enhanced Common Public Radio Interface (eCPRI) and standardizes the protocols, control, and synchronization required for multi-vendor interoperability in 5G networks [17].

As shown in Figure 1.8, the concept of RAN disaggregation divides the traditional base station into separate functional units, extending the functional split paradigm introduced by 3GPP for Base-Station (gNB). In this architecture, the gNB is separated into three main components: the Central Unit (CU), the Distributed Unit (DU), and the Radio Unit (RU), known also as O-CU, O-DU, and O-RU in the ORAN specifications [17].

The CU itself is further divided into two logical entities: one responsible for Control Plane (C plane) and another for User Plane (U plane). This logical separation enables different network functions to be deployed in various locations and on different types of hardware. For instance, CUs and DUs can be virtualized on general-purpose (“white box”) servers located at the network edge, often with optional hardware acceleration for specific physical layer tasks. In contrast, RUs are typically implemented on dedicated hardware such as Field Programmable Gate Arrays (FPGA) or Application Specific Integrated Circuits (ASIC) and are deployed close to radio antennas to minimize latency and signal degradation.

The ORAN Alliance has studied multiple functional split options between RU and DU as proposed by 3GPP, with a particular focus on the split of the physical layer. Among these, the so-called 7.2x split has emerged as the most widely adopted, offering a good trade-off between RU simplicity and the performance requirements (in terms of bandwidth and latency) on the fronthaul interface. Under the 7.2x split, RU is responsible for only a small set of physical layer tasks, specifically FFT processing and cyclic prefix addition/removal and RF functionalities. This lightweight design makes the RU cost-effective, power-efficient, and easier to deploy in various environments[18].

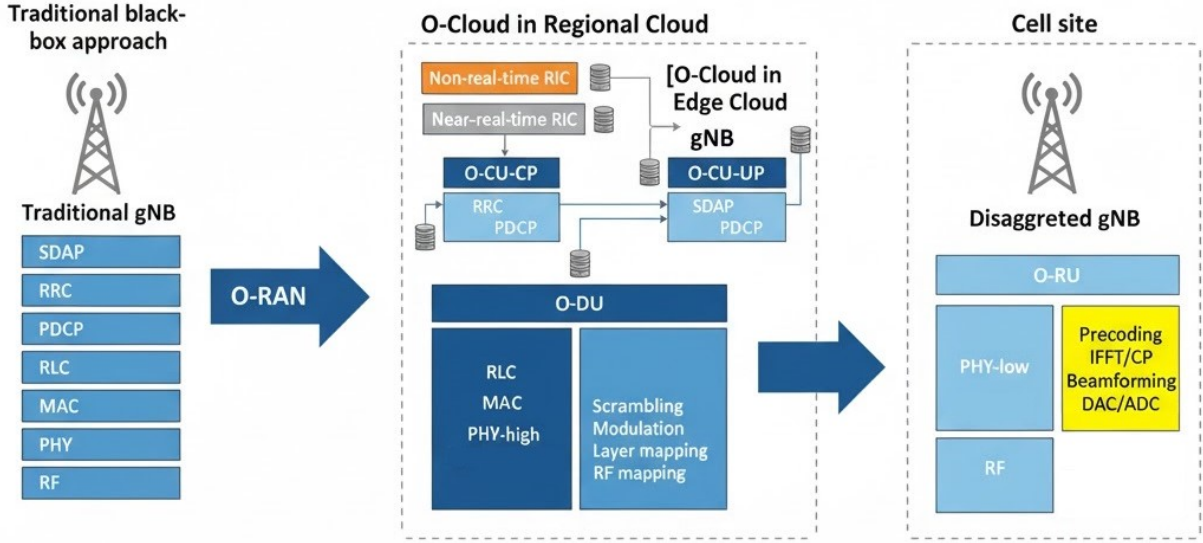


Figure 1.8: Evolution of the traditional black-box base station architecture (left) toward a virtualized gNB with a functional split (right, including the CU and DU at the edge, and the RU at the cell).

The DU handles the remaining functions of the physical layer, as well as the MAC and RLC layers. These three layers must operate in close synchronization: the MAC layer assembles Transport Blocks (TRBs) for transmission over the physical layer using data buffered in the RLC layer, ensuring efficient scheduling and reliable data delivery[29].

Finally, the Centralized Unit- Control Plane (CU-CP) and Centralized Unit- User Plane (CU-UP) components manage the higher layers of the 3GPP protocol stack. The CU-CP implements the RRC layer, which handles connection setup, maintenance, and release procedures. The CU-UP implements the Service Data Adaption Protocol (SDAP) layer that manages Quality of service (Qos) for different traffic flows (or bearers), ensuring appropriate prioritization and resource allocation. The PDCP layer provides additional services such as packet reordering, duplication handling, header compression, and encryption, improving reliability and security over the air interface[30]. In general, this modular, disaggregated design of the ORAN architecture provides greater deployment flexibility, enables scalable cloud-native implementations, and allows operators to mix and match components from different vendors, all while maintaining 3GPP-compliant performance and interoperability. The PHY layer of the 5G NR utilizes OFDM waveform for transmission over the air, structured in a frame format. This frame structure, as defined by NR, is presented in both time and frequency [31].

1.3.1.3 Near Real time and non-Real time intelligent controllers

A key innovation in the ORAN architecture is the introduction of the near RT RIC as seen in Figure 1.9, which provides programmable control and optimization capabilities through closed-loop mechanisms. By collecting and processing hundreds of Key Performance Measurements (KPM) and additional context information, RICs leverage AI and ML algorithms to enforce control policies. This enables automated and data-driven optimization of functions such as slicing, load balancing, handovers, and scheduling[32][17].

The ORAN architecture introduces two types of RICs:

1. **Non-Real-Time RIC:** Part of the Service Management and Orchestration (SMO) framework, it operates on time scales greater than 1 s. Its responsibilities include:

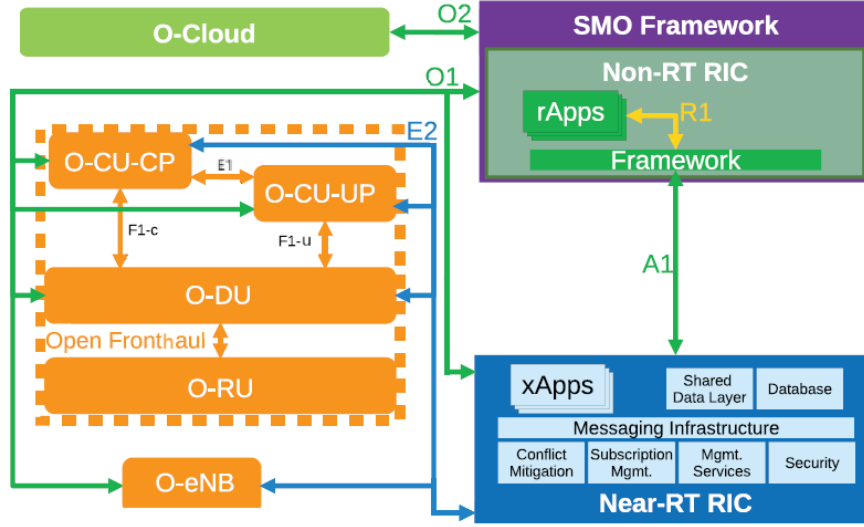


Figure 1.9: Representation of the Near RT RIC and Non RT RIC with their open interfaces A_1 , O_1 and E_2

- providing guidance and enrichment information for the near RT RIC.
- managing and distributing ML models.
- enforcing long-term optimization and policy control across thousands of devices through SMO integration.

Due to its broad scope, the non Real Time RAN Intelligent Controller (non RT RIC) faces scalability challenges that require efficient process and software design.

2. **Near-Real-Time RIC:** Deployed at the network edge, it operates on time scales between 10 ms and 1 s. The near RT RIC interacts directly with DUs, CUs, and ORAN-compliant gNBs, influencing the Qos of large numbers of User Equipment (UE)s[32]. It hosts applications known as xApps, which implement radio resource management logic. Its platform includes:
 - a shared RAN database for xApps.
 - messaging infrastructure and subscription mechanisms.
 - terminations for open interfaces and Application Programming Interface (API)s.
 - conflict resolution mechanisms to manage multiple xApps controlling the same function.

1.3.2 ORAN split 7.2 Fronthaul interface

The ORAN Alliance has standardized the functional split known as Split 7.2 between O-DU and O-RU[17].

In this split as seen in Figure 1.10, the PHY layer is divided such that the High-PHY functions remain in the O-DU, while the Low-PHY functions (e.g., FFT/IFFT, Analog conversion...) are executed within the O-RU. The dotted blocks such as Precoding and Analog Beamforming (BF) blocks are considered optional. This design allows the O-DU to centralize more computationally intensive tasks while the O-RU handles time-critical radio operations close to the antenna[33].

To support this split, the ORAN Open Fronthaul specification defines three main planes for communication between the O-DU and O-RU[34]:

- **User Plane (U-plane):** carries IQ data using packet-based transport, typically following eCPRI encapsulation with ORAN specific message structures.

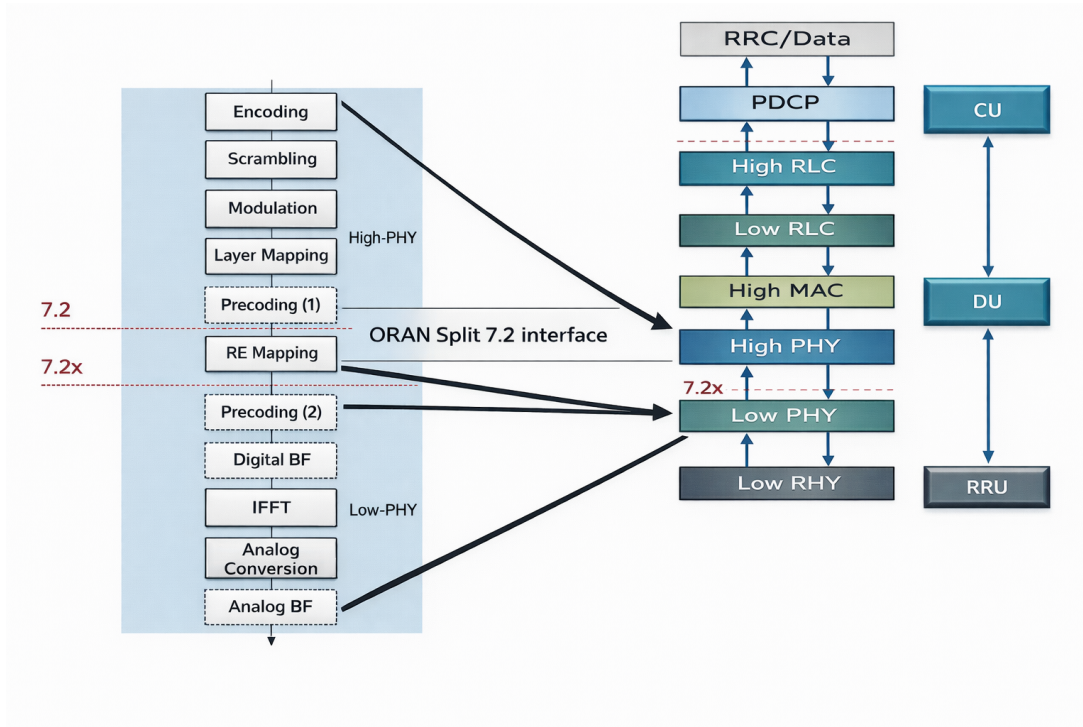


Figure 1.10: ORAN split 7.2 representation.

- **Control Plane (C-plane):** provides signaling for the configuration, scheduling commands, and management of the O-RU behavior.
- **Synchronization Plane (S-plane):** ensures accurate time and frequency alignment between O-DU and O-RU, often relying on IEEE 1588 Precision Time Protocol (PTP) and Synchronous Ethernet.

This architecture replaces traditional constant rate transport CPRI with a packetized, Ethernet-based model, significantly reducing fronthaul bandwidth requirements and enabling multivendor interoperability [34].

The adoption of the ORAN Split 7.2 fronthaul interface offers several technical and operational advantages compared to other split architectures:

1. **Reduced Fronthaul Bandwidth Requirements.** Unlike traditional CPRI which transports raw IQ samples at very high bit rates, Split 7.2 transports partially processed data at the resource block or symbol level. This significantly reduces the demand for fronthaul capacity, reducing transport costs, and making centralized or virtualized architectures feasible.
2. **Multivendor Interoperability.** By standardizing the interface between O-DU and O-RU, ORAN enables interoperability between equipment from different vendors. This fosters an open ecosystem, reduces vendor lock-in, and allows operators to combine best-of-breed components.
3. **Support for Virtualization and Cloud Deployment.** With higher PHY functions centralized in the O-DU, operators can leverage virtualization and cloud-native deployment models at the edge. This supports flexible scaling, efficient resource utilization, and simplified lifecycle management.
4. **Facilitation of Advanced Antenna Systems.** Split 7.2 is designed to support large-scale antenna arrays and massive MIMO by reducing the transport overhead compared to Option 8. This enables a more efficient deployment of advanced antenna technologies without excessive demands for fronthaul bandwidth.

5. **Conformance Testing and Validation.** The ORAN Alliance provides detailed conformance test specifications for the 7.2x interface, allowing operators to validate implementations in a vendor-neutral way. This increases confidence in interoperability and ensures consistent performance across deployments.

1.3.2.1 CAT A and CAT B Radio Unit Splits in Relation to ORAN Split 7.2x

The 7.2x split partitions of the PHY into a high-PHY segment, located in the DU, and a low-PHY segment, located in the RU. The low-PHY functions typically include FFT/IFFT, cyclic prefix addition and removal, digital front-end processing, and digital-to-analog or analog-to-digital conversion. The high-PHY in the DU handles modulation mapping, demapping, precoding, and MAC-PHY interface. This structure reduces the bandwidth requirements for the fronthaul while retaining centralization benefits such as coordinated scheduling and joint processing[34].

Within the 7.2x split framework, ORAN specifies two operational categories, Category A (CAT A) and Category B (CAT B) that determine the distribution of the precoding functions between DU and RU.

Precoding is a key signal processing function in MIMO systems that determines how signals are shaped and distributed across multiple transmit antennas. Its purpose is to ensure that the receiver can correctly separate and decode the different data streams. A major benefit of precoding is enabling spatial multiplexing, where several independent streams are transmitted at the same time while minimizing interference between them, ultimately improving overall system performance.

Feature / Capability	O-RU Category A	O-RU Category B
Precoding Support	Not supported in radio	Supported in radio
Use Case	Non-precoding O-RU	Precoding O-RU
Modulation Compression	Not supported	Supported
Digital Beamforming	Optional	Optional
Analog Beamforming	Optional	Optional
Notes	DU must perform precoding	RU can offload precoding from DU

Table 1.2: Comparison Between O-RU Category A and Category B

As seen in table 1.2, CAT A RU represents a non-precoding radio unit, where precoding is not performed within the radio. As a result, the DU is responsible for all precoding operations, and the RU does not support modulation compression. In contrast, CAT B RU supports precoding within the radio itself, allowing the DU to bypass this function. CAT B additionally enables modulation compression capabilities. Both Category A and Category B radios may optionally support digital and analog beamforming, but their fundamental difference lies in whether precoding is executed in the radio CAT B or CAT A [9].

1.3.3 Massive MIMO

The Massive MIMO concept has recently emerged as one of the most promising technologies for next-generation wireless communication systems. It represents a natural evolution of conventional MIMO, where a base station is equipped with a large number of antennas, typically on the order of several hundreds, to simultaneously serve tens or even hundreds of user terminals within the same time-frequency resources as seen in Figure 1.6. Unlike conventional systems, which operate with roughly the same number of antennas and users, Massive MIMO leverages the principle of spatial multiplexing at an unprecedented scale[21].

The motivation for Massive MIMO is threefold. First, it has the potential to drastically improve the spectral efficiency by serving many users in parallel with high SINR. Second, it can increase energy

efficiency by focusing radiated energy sharply toward the intended terminals, thereby reducing power consumption and interference[35]. Finally, Massive MIMO enables robust and simplified system design: By using the law of large numbers, random channel fluctuations are averaged out, making the effective channel appear nearly deterministic.

Consequently, Massive MIMO is considered a key enabler for future wireless standards (such as 5G and beyond), supporting both mobile broadband and fixed wireless access in dense urban and rural scenarios.

1.3.3.1 Conventional MIMO vs. Massive MIMO

Conventional multi-user MIMO systems are typically constrained by the fact that the number of antennas at the base station is of the same order of magnitude as the number of served users. In such systems, advanced signal processing techniques are required to manage inter-user interference, and the system performance is often limited by scalability issues. Moreover, conventional MIMO solutions are highly dependent on Frequency Division Duplexing (FDD)[36], which requires explicit feedback of Channel State Information (CSI) from users to the base station. This feedback overhead becomes a serious bottleneck as the number of antennas grows [37].

Massive MIMO departs from these limitations in several fundamental ways:

- **Excess antennas:** The base station is equipped with an order of magnitude more antennas than the users. This abundance of spatial degrees of freedom allows simultaneous transmission to many terminals with minimal inter-user interference.
- **Time division duplexing (TDD):** Instead of relying on costly CSI feedback as in FDD, Massive MIMO systems exploit channel reciprocity in Time Division Duplexing (TDD) mode. The uplink pilots sent by the users allow the base station to estimate the channel, which is then reused for uplink and downlink transmission.
- **Simplified processing:** In the large-antenna regime, linear processing techniques such as Maximum Ratio Combining (MRC) or Zero Forcing (ZF) become asymptotically optimal, drastically reducing computational complexity compared to conventional MIMO.
- **Robustness:** With a very large array, small-scale fading effects average out, leading to highly reliable channels and making the system resilient against hardware imperfections and even antenna failures.

1.3.3.2 Advantages of Massive MIMO

Massive MIMO provides several important advantages[37][38]:

1. Beamforming Gain

One of the most important advantages of Massive MIMO is the beamforming gain. By using many antennas, the base station can direct precisely its transmissions to individual users instead of radiating energy in all directions. This focused transmission strengthens the received signal, improves coverage, and reduces interference for users in other directions. As a result, the communication link becomes both more energy-efficient and more reliable, which is particularly valuable in dense urban environments where interference is usually strong.

2. Spatial Diversity

Massive MIMO also improves the reliability of spatial diversity. In a real-world scenario, signals are often reflected, blocked, or scattered, leading to fading and quality fluctuations. With a large antenna array, the system can transmit and receive signals over many different spatial paths. Even if some

of these paths are blocked or weakened, others remain strong, ensuring that the overall connection remains stable. This makes Massive MIMO highly robust in environments where conventional systems would suffer from severe signal degradation, as illustrated in Figure 1.6.

3. Spatial Multiplexing

Another major benefit of Massive MIMO lies in spatial multiplexing. Instead of serving one user at a time or relying on separate frequency bands, the system can transmit several independent data streams to different users simultaneously within the same frequency band. Due to the high number of antennas, the base station is able to separate these users spatially, greatly increasing the total system capacity. In practice, this means that many more users can enjoy high data rates without requiring additional spectrum, which is a scarce and expensive resource.

4. Combatting Fading, Scattering, and Interference

Beyond these gains, Massive MIMO is particularly effective in tackling common impairments of wireless channels such as fading, scattering, and interference. Fading occurs when the strength of a signal fluctuates due to reflections or obstacles, sometimes making the connection unreliable. With many antennas and spatial diversity, Massive MIMO can average out these variations and maintain a more stable link. Scattering, caused by objects in the environment redirecting parts of the signal, can also degrade performance, but the large antenna array is capable of capturing these multiple scattered components and combining them constructively. Furthermore, interference from other users or systems is strongly reduced, since Massive MIMO can form narrow beams that focus only on the intended receiver while minimizing leakage in other directions. This makes the system both more resilient and more efficient in complex and crowded radio environments.

1.3.3.3 Spatial multiplexing in Massive MIMO

Spatial multiplexing in 5G NR is fundamentally achieved through the combined use of precoding and beamforming. These two processing stages map multiple data layers onto the antenna array so that several independent streams can be transmitted simultaneously, either to a single UE in Single User MIMO (SU MIMO) or to multiple UEs in Multi User MIMO (MU MIMO)[39].

The 5G Functional Application Platform Interface (FAPI) framework supports two operational modes for spatial multiplexing[40]:

1. **Semi-static mode**, where the PHY selects precoding. beamforming entries from predefined configuration tables, and
2. **Dynamic mode**, where the PHY computes beam weights in real time based on instantaneous channel conditions.

In the semi-static approach, the PHY layer is preloaded with a set of precoding matrices and digital beamforming vectors. Each Packet Data Unit (PDU) transmitted in a slot carries an index that refers to the selected precoding matrix and/or digital beamforming entry. This reduces signaling overhead and simplifies the implementation PHY while still supporting multi-layer transmission[40][41].

For downlink transmission, spatial processing is implemented as two cascaded linear operations. First, a Precoding Matrix (PM) is selected from a predefined table to perform layer-to-port mapping. Second, Digital Beamforming (DB) is applied through a matrix DB, whose columns are chosen from a Digital Beamforming Table (DBT). Each baseband port thus selects the corresponding beamforming vector by index. These baseband ports are logical transmit entities which may subsequently be processed by the Digital Front End (DFE), and in 5G NR they typically correspond to component carriers mapped to individual RF chains.

To understand the spatial processing chain, we illustrate the different layer mapping steps in the following figure:

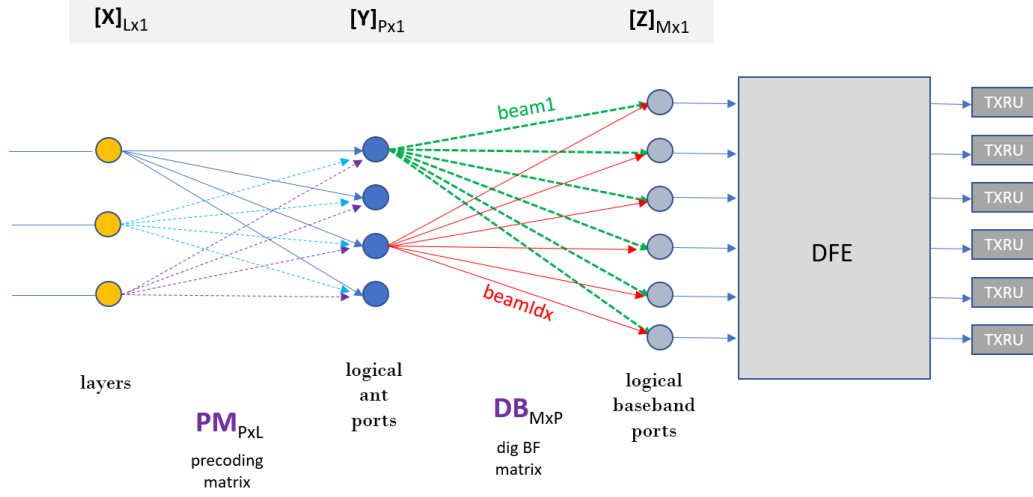


Figure 1.11: Downlink spatial multiplexing[40].

At the input, one or more modulation layers are produced by the MAC and higher PHY functions, we denote the input signal vector as $[X]_{L \times 1}$ where L is the number of layers. These layers are mapped into logical transmit ports by the precoding matrix $PM_{P \times L}$, either selected from a preconfigured table or determined dynamically, where P is the number of logical antenna ports. Precoding controls how multiple layers are spatially separated and is essential for both single-user and multi-user MIMO operation. Logical antenna ports are represented by a vector $[Y]_{P \times 1}$, where P is the number of antenna ports or the number of activated beam IDs. Therefore, the precoding operation can be written as $[Y]_{P \times 1} = PM_{P \times L} \times [X]_{L \times 1}$.

Following precoding, digital beamforming is applied through a beam weight matrix $DB_{M \times P}$, where M is the number of baseband ports. Each column of DB corresponds to a per-antenna complex weight vector, typically selected from a DBT or computed adaptively by the PHY. Beamforming translates the logical ports into antenna-domain signals and directs energy in specific spatial directions. Therefore, the precoding operation can be written as $[Z]_{M \times 1} = DB_{M \times P} \times [Y]_{P \times 1}$.

The output of the DB stage corresponds to antenna-specific samples that are subsequently processed by the DFE and mapped to RF transmit units over the air[40][34].

Precoding techniques In NR, precoding is a fundamental transmission technique that exploits CSI at the transmitter to control how data symbols are spatially mapped onto multiple transmit antennas. By appropriately weighting the antenna signals, precoding improves spectral efficiency, enhances the received signal quality, and mitigates multi-user interference, which is essential to meet the performance requirements of 5G and beyond systems [42]. Let $H \in \mathbb{C}^{M \times K}$ denote the downlink channel matrix between a base station equipped with M transmit antennas and K users. The precoding matrix $PM \in \mathbb{C}^{M \times K}$ determines the spatial weighting applied to the transmitted symbols, and σ^2 denotes the noise power at the receiver. Common linear precoding techniques include:

1. **Maximum Ratio Transmission (MRT):** This technique aims to maximize the received signal power at the intended user by aligning the precoding vector with the conjugate of the channel, i.e.,

$$PM_{\text{MRT}} = H^*.$$

MRT provides the strongest possible desired signal but does not actively suppress inter-user interference, making it suitable mainly for single-user transmission or orthogonal scheduling.

2. **Zero Forcing (ZF) [43]:** ZF precoding eliminates multi-user interference by applying the Moore–Penrose pseudo-inverse of the channel matrix,

$$PM_{ZF} = H^\dagger.$$

While this approach can completely cancel inter-user interference, it may amplify noise and requires the number of transmit antennas to be greater than or equal to the number of users.

3. **Minimum Mean Square Error (MMSE):** MMSE precoding achieves a balance between maximizing the desired signal power and suppressing interference and noise and is given by

$$PM_{MMSE} = H^H (HH^H + \sigma^2 I)^{-1}.$$

This method partially cancels interference while avoiding excessive noise enhancement, making it well suited for practical multi-user NR deployments.

To understand how the full downlink spatial multiplexing work, we will present a modeling of the case in Figure 1.11:

Signal Model Description

Input Vector $\mathbf{X}$ (Layers)

The vector $\mathbf{X}$ contains the transmitted MIMO layers. X_1 , X_2 , and X_3 correspond to layers 1, 2, and 3:

$$\mathbf{x} = \begin{bmatrix} X_1 \\ X_2 \\ X_3 \end{bmatrix}, \quad \text{Dimension: } 3 \times 1.$$

Precoding matrix $\mathbf{PM}$

The precoding matrix $\mathbf{PM}$ maps the $L = 3$ layers to $P = 4$ logical antenna ports. Each column corresponds to one layer, and each row corresponds to one port:

$$\mathbf{PM} = \begin{bmatrix} p_{11} & p_{12} & p_{13} \\ p_{21} & p_{22} & p_{23} \\ p_{31} & p_{32} & p_{33} \\ p_{41} & p_{42} & p_{43} \end{bmatrix}, \quad \text{Dimension: } 4 \times 3.$$

Logical Antenna Port Vector $\mathbf{Y}$

The logical antenna port vector is obtained by applying the precoding matrix to the layer vector:

$$\mathbf{Y} = \mathbf{PMX}.$$

$$\mathbf{Y} = \begin{bmatrix} Y_1 \\ Y_2 \\ Y_3 \\ Y_4 \end{bmatrix}, \quad \text{Dimension: } 4 \times 1.$$

Explicitly:

$$\mathbf{Y} = \begin{bmatrix} p_{11}X_1 + p_{12}X_2 + p_{13}X_3 \\ p_{21}X_1 + p_{22}X_2 + p_{23}X_3 \\ p_{31}X_1 + p_{32}X_2 + p_{33}X_3 \\ p_{41}X_1 + p_{42}X_2 + p_{43}X_3 \end{bmatrix}.$$

Beamforming matrix **DB**

The beamforming matrix **DB** contains the complex weights applied to the logical antenna ports and maps them to $M = 6$ baseband output ports. Each row corresponds to one output port, and each column corresponds to one logical antenna port:

$$\mathbf{DB} = \begin{bmatrix} w_{11} & w_{12} & w_{13} & w_{14} \\ w_{21} & w_{22} & w_{23} & w_{24} \\ w_{31} & w_{32} & w_{33} & w_{34} \\ w_{41} & w_{42} & w_{43} & w_{44} \\ w_{51} & w_{52} & w_{53} & w_{54} \\ w_{61} & w_{62} & w_{63} & w_{64} \end{bmatrix}, \quad \text{Dimension: } 6 \times 4.$$

Output Vector **Z** (Baseband Ports)

The beamformed output signals are obtained as:

$$\mathbf{Z} = \mathbf{DB} \times \mathbf{Y}.$$

$$\mathbf{Z} = \begin{bmatrix} Z_1 \\ Z_2 \\ Z_3 \\ Z_4 \\ Z_5 \\ Z_6 \end{bmatrix}, \quad \text{Dimension: } 6 \times 1.$$

Explicitly:

$$\mathbf{Z} = \begin{bmatrix} w_{11}Y_1 + w_{12}Y_2 + w_{13}Y_3 + w_{14}Y_4 \\ w_{21}Y_1 + w_{22}Y_2 + w_{23}Y_3 + w_{24}Y_4 \\ w_{31}Y_1 + w_{32}Y_2 + w_{33}Y_3 + w_{34}Y_4 \\ w_{41}Y_1 + w_{42}Y_2 + w_{43}Y_3 + w_{44}Y_4 \\ w_{51}Y_1 + w_{52}Y_2 + w_{53}Y_3 + w_{54}Y_4 \\ w_{61}Y_1 + w_{62}Y_2 + w_{63}Y_3 + w_{64}Y_4 \end{bmatrix}.$$

The vector **Z** represents the six baseband output signals that are sent to the antenna ports of the radio unit after both precoding and beamforming have been applied. Each output combines the contributions from the three input layers, weighted by the complex precoding coefficients in **PM** and the beamforming coefficients in **DB** corresponding to the desired spatial beams. Because each column of the beamforming matrix contains the weights for one logical antenna port stacked vertically across the output ports, the resulting signals in **Z** reflect how the three layers are spatially shaped, combined, and directed toward the intended coverage or users. In essence, **Z** is the final, per-antenna baseband signal set that will be transmitted over the air, with each component representing the spatially precoded and beamformed mixture of layers 1, 2, and 3.

Straightwire precoding Spatial Stream (spStream) are the individual subcarrier signals sent over the ORAN FH, determined by the DU's downlink spatial multiplexing capabilities. If a system does not support Split 7.2x, the functions normally handled by the DU, including coding, modulation, and spatial processing such as precoding or beamforming are instead performed by the entire PHY chain up to and including the spatial multiplexing stage[40].

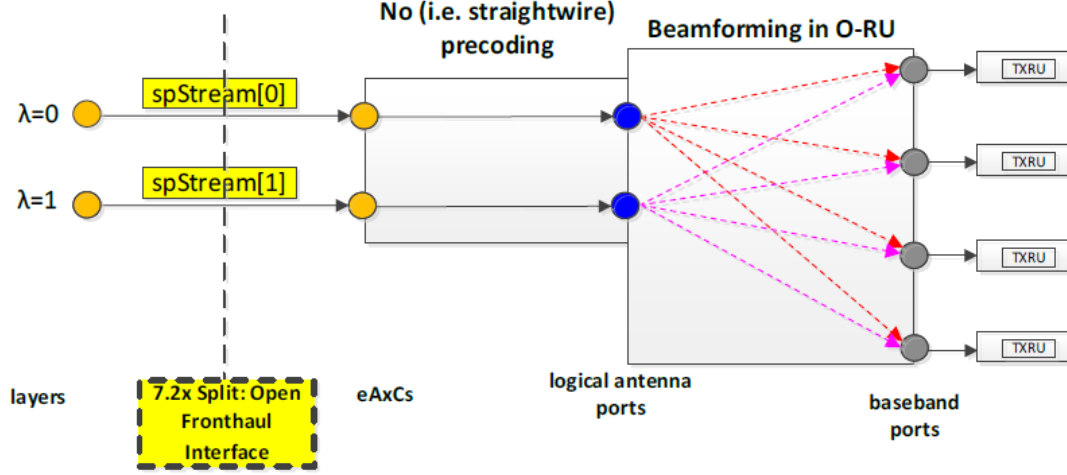


Figure 1.12: Downlink spatial multiplexing with straightwire precoding[40].

Downlink spatial streams can correspond, for example, to downlink data flows called Antenna Carrier (eAxC)s identified by the “c_eAxC_ID”. These data flows, referred to as eAxC, are labeled using eAxC-IDs in [34]. The straightwire precoding is the simplest precoding scheme, where the layers are straightwired to the logical antenna ports. This leads to say that the precoding matrix is an identity matrix ($PM_{P \times L} = I_{P \times L}$)

To understand the logic on how Massive MIMO and spatial multiplexing work, we present a 4×2 MIMO system, where we have 2 layers to be multiplexed to 2 different UEs:

Signal Model Description

Input Vector $\mathbf{X}$ (Layers)

The vector $\mathbf{X}$ contains the transmitted MIMO layers. X_1 corresponds to layer 1 and X_2 corresponds to layer 2:

$$\mathbf{X} = \begin{bmatrix} X_1 \\ X_2 \end{bmatrix}, \quad \text{Dimension: } 2 \times 1.$$

Logical Antenna Port Vector $\mathbf{Y}$

With straight-wire precoding, the antenna-port vector equals the layer vector:

$$\mathbf{Y} = \mathbf{X}.$$

$$\mathbf{Y} = \begin{bmatrix} Y_1 \\ Y_2 \end{bmatrix}, \quad \text{Dimension: } 2 \times 1.$$

Beamforming matrix $\mathbf{DB}$

The beamforming matrix $\mathbf{DB}$ contains the complex weights for two beams (serving two different UEs). Each column represents a beam's weight vector, and each row corresponds to one antenna port of the RU. Weights are filled vertically, i.e., each w_{ij} is the weight applied to a specific antenna element:

$$\mathbf{DB} = \begin{bmatrix} w_{11} & w_{21} \\ w_{12} & w_{22} \\ w_{13} & w_{23} \\ w_{14} & w_{24} \end{bmatrix}, \quad \text{Dimension: } 4 \times 2.$$

Output Vector $\mathbf{Z}$ (Baseband Ports)

The beamformed output signals are obtained as:

$$\mathbf{Z} = \mathbf{DBY}.$$

$$\mathbf{Z} = \begin{bmatrix} Z_1 \\ Z_2 \\ Z_3 \\ Z_4 \end{bmatrix}, \quad \text{Dimension: } 4 \times 1.$$

Explicitly:

$$\mathbf{Z} = \begin{bmatrix} w_{11}Y_1 + w_{21}Y_2 \\ w_{12}Y_1 + w_{22}Y_2 \\ w_{13}Y_1 + w_{23}Y_2 \\ w_{14}Y_1 + w_{24}Y_2 \end{bmatrix}.$$

The vector $\mathbf{Z}$ represents the 4 baseband output signals that are sent to the antenna elements of the radio unit after beamforming has been applied. Each output combines the contributions from the two input layers, weighted by the complex beamforming coefficients corresponding to the beams intended for two different UEs. Because each column of the beamforming matrix contains the weights for one beam stacked vertically across the antenna elements, the resulting signals in $\mathbf{Z}$ reflect how the two layers are spatially shaped and directed. In essence, $\mathbf{Z}$ is the final, per-antenna baseband signal set that will be transmitted over the air, with each component representing the spatially precoded mixture of layer 1 and layer 2.

Dynamic downlink precoding Dynamic precoding is a fundamental mechanism in 5G NR that enables high spectral efficiency, spatial multiplexing, and interference management. It allows the gNB to adapt beamforming weights in real time based on the current wireless channel conditions, optimizing signal delivery to multiple users simultaneously. The process relies on accurate CSI, which is obtained using downlink CSI referencing signal and uplink Sounding Reference Signals (SRS).

Accurate channel knowledge is essential for dynamic precoding. The channel matrix H describes the complex gains between transmit antennas at the base station and receive layers at the UE.

Uplink SRS: Each UE transmits known SRS sequences on the uplink. The base station receives these signals across all antennas and measures amplitude and phase differences to estimate the uplink channel coefficients. In TDD systems, uplink channel measurements can be directly used for downlink precoding due to channel reciprocity.

Downlink CSI-RS: The base station transmits CSI-RS signals to UEs, which measure the downlink channel response and report CSI feedback back to the base station. The feedback typically includes the

Precoding Matrix Indicator (PMI), Channel Quality Indicator (CQI), and Rank Indicator (RI). Using this information, the base station reconstructs an estimate of the downlink channel[44].

Channel matrix: The resulting channel matrix $H \in \mathbb{C}^{N_{\text{UE}} \times N_{\text{TX}}}$ has elements h_{ij} representing the complex gain from transmit antenna j to receive layer i . This matrix serves as the basis for computing the dynamic precoding weights.

Once the channel matrix H has been estimated, the base station computes the precoding vectors used to map transmission layers onto the antenna array. The goal of precoder selection is to exploit the spatial properties of the channel while minimizing inter-user interference, especially in multi-user MIMO (MU-MIMO) scenarios.

Joint MU-MIMO Precoding: For simultaneous transmission to multiple UEs, the base station computes a joint precoding matrix that accounts for all selected layers across all UEs. Techniques such as ZF or regularized ZF may be applied to suppress inter-user interference. The resulting precoding matrix, denoted as **PM**, contains the complex precoding weights.

1.3.3.4 Technical Challenges and Limitations

Although Massive MIMO promises remarkable improvements in spectral efficiency, energy efficiency, and robustness, several fundamental challenges must be addressed before large-scale deployment becomes feasible. These challenges arise primarily from practical system constraints such as hardware limitations, signal processing requirements, and channel estimation issues. The most important challenges are outlined below [37]:

1. **Channel Reciprocity and Calibration:** Massive MIMO systems are generally designed to operate in TDD mode, exploiting the principle of channel reciprocity to obtain downlink CSI from uplink pilot signals. In practice, however, the transmit and receive RF chains of a base station are not perfectly reciprocal due to mismatches in analog components. This hardware asymmetry leads to biased channel estimates and degraded performance. Therefore, reciprocity calibration techniques must be employed, typically involving relative calibration of the antenna array, so that the estimated channels remain valid for both uplink and downlink.
2. **Pilot Contamination:** Channel estimation in Massive MIMO relies on orthogonal pilot sequences transmitted by users. Since the coherence interval of a wireless channel is finite, only a limited number of orthogonal pilots can be assigned within one cell. In multi-cell systems, these pilots must be reused across neighboring cells, leading to the phenomenon known as pilot contamination. When users use the same pilot in different cells, the base station inadvertently estimates a linear combination of the desired and interfering channels. Unlike thermal noise or uncorrelated interference, this effect does not vanish even with an increasing number of antennas, thus representing a fundamental bottleneck for Massive MIMO[45].
3. **Radio Propagation and Orthogonality of Channels:** The performance of Massive MIMO critically depends on favorable propagation, which means that the channel responses of different users should be nearly orthogonal. In practice, this orthogonality is not always guaranteed, especially in environments with limited scattering or when the antenna array has an insufficient aperture. Measurements have shown that very large linear or cylindrical arrays can achieve near-ideal spatial resolution, while compact arrays may suffer from poor separation between users. Thus, careful antenna array design and deployment are necessary to fully exploit the benefits of Massive MIMO.
4. **Computational and Signal Processing Complexity:** Although linear signal processing techniques (e.g., Maximum Ratio Combining or Zero-Forcing) become nearly optimal in the large-antenna regime, the base station must still process vast amounts of baseband data in real time. For example, with hundreds of antennas and wideband channels, the amount of digital signal processing required

for channel estimation, detection, and precoding is immense. Scalable, distributed, and energy-efficient algorithms are therefore essential for practical implementation.

5. **Hardware Limitations and Power Consumption:** A key advantage of Massive MIMO is that it allows the use of many low-power, low-cost amplifiers instead of a few high-power ones. However, this architectural shift introduces new challenges: phase noise, I/Q imbalance, quantization errors, non linear gains[46] giving rise to unwanted emissions that might affect nearby operating band and affect band coexistence, and synchronization issues across the antenna array. Moreover, while the radiated power per antenna is small, the overall baseband processing and RF circuitry consume significant power. Designing cost-effective, energy-efficient hardware capable of supporting hundreds of parallel RF chains remains an open research challenge[45].

1.3.4 Unwanted emissions in NR systems

Unwanted emissions are divided into OOB emissions and spurious emissions. OOB emissions occur immediately outside the channel bandwidth due to the modulation process and transmitter non-linearities, while excluding spurious emissions. Spurious emissions arise from unintended transmitter effects such as harmonic radiation, parasitic emissions, intermodulation products, and frequency conversion products.

For repeater transmitters, OOB emission requirements are specified in terms of Adjacent Channel Leakage Ratio (ACLR) and Operating Band Unwanted Emissions (OBUE)[47]. The OBUE requirement applies within each supported downlink and uplink operating band of repeater type 1-C, including frequency offsets of Δf_{OBUE} above and below the band edges. Unwanted emissions outside this range are subject to spurious emissions limits[47].

Repeater type	Operating band characteristics	Δf_{OBUE} (MHz)
Repeater type 1-C	$F_{\text{DL,high}} - F_{\text{DL,low}} < 200 \text{ MHz}$	10
	$200 \text{ MHz} \leq F_{\text{DL,high}} - F_{\text{DL,low}} \leq 900 \text{ MHz}$	40

Table 1.3: Maximum offset of OBUE outside the downlink operating band of repeater type 1-C DL

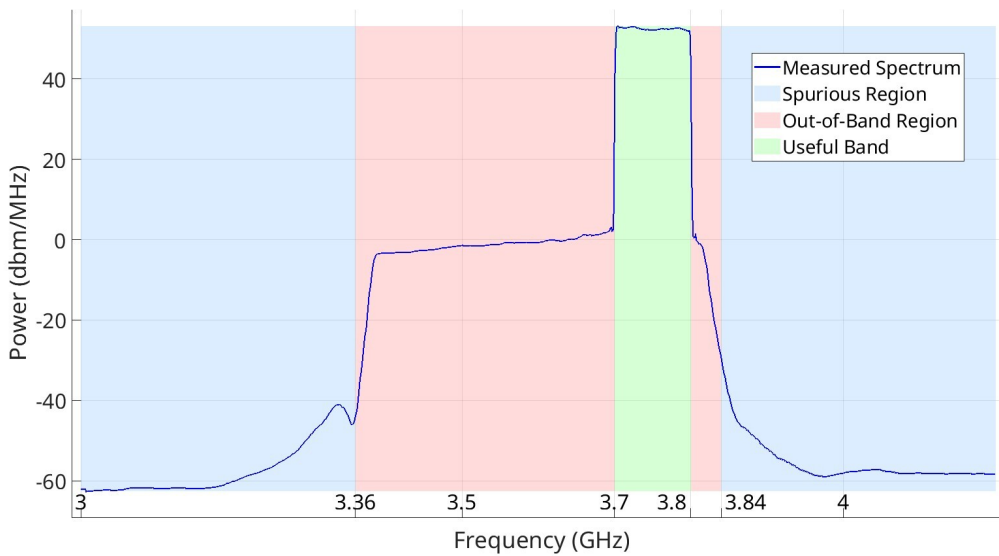


Figure 1.13: 3GPP unwanted delimitations for the n 78 band

For the [3.4–3.8] GHz (n78) band, which has a downlink bandwidth of 400 MHz, the 3GPP specification for repeater type 1-C dictates that the maximum OBUE offset Δf_{OBUE} is 40 MHz outside the operating band. That leads to the observation that OOB emissions are unwanted emissions generated inside the downlink bandwidth of band n78, and spurious emissions are unwanted emissions that arise at less than 3.36 GHz and higher than 3.84 GHz. In this study, we only care about OOB emissions, so we will ignore spurious emissions. This can be understood in Figure 1.13.

1.3.5 Non linearities modeling for power amplifiers

In massive MIMO systems, Power Amplifier (PA) is a key RF component that significantly affects both energy efficiency and signal quality. To achieve high power efficiency, PAs are often driven close to saturation, where their behavior becomes inherently non-linear. This non-linearity leads to signal distortions such as in-band distortion and OOB emissions[47], which can degrade system performance and violate spectral mask requirements. Although the large number of antennas in massive MIMO systems can partially mitigate distortion effects through spatial averaging, the impact of non-linear power amplifiers remains non-negligible. Consequently, accurate modeling of PA non-linearities and the use of compensation techniques, such as digital predistortion, are essential for maintaining acceptable performance while preserving energy efficiency[46].

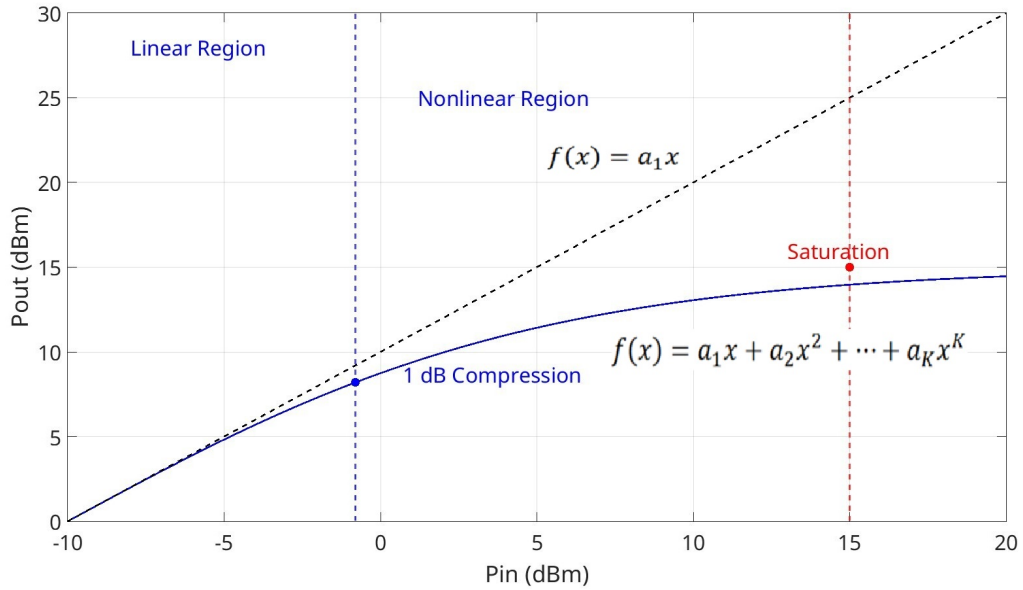


Figure 1.14: The difference between linear and non linear regime of a power amplifier

Power amplifiers can operate in either a linear or a non-linear regime depending on the input signal power level as seen in Figure 1.14. The transition between these regimes and the impact of increasing non-linearity are critical considerations in massive MIMO systems, where signal distortion directly affects overall system performance.

In the linear regime, the PA operates within a range where the output signal is a scaled version of the input signal, preserving both amplitude and phase characteristics. The input–output relationship is approximately proportional, resulting in a constant gain and negligible distortion. Operating in this regime ensures high signal fidelity and minimal spectral spreading. However, linear operation is typically associated with low power efficiency, as the PA must be backed off from saturation to accommodate signals with high Peak to Average Power Ratio (PAPR).

As the input power increases beyond a threshold called "1 dB compression point", the PA enters the non-linear regime. In this region, the proportional relationship between the input and output signals no longer holds, and the PA exhibits gain compression as it approaches saturation. This non-linear behavior introduces distortion effects that degrade system performance. In-band distortion alters the desired signal waveform, reducing error performance, while OOB emissions cause spectral regrowth, leading to interference in adjacent frequency bands and potential violations of spectral mask requirements.

The severity of distortion increases with the order of non-linearity. Lower-order non-linear effects primarily result in moderate gain compression and limited distortion, which may still be tolerable under some operating conditions. As higher-order non-linearities become dominant, the PA generates harmonic and intermodulation products that significantly broaden the transmitted spectrum and further distort the signal. These higher-order effects are more difficult to mitigate and require advanced compensation techniques, such as digital predistortion.

In massive MIMO systems, the large number of antennas can partially average out the effects of non-linear distortion across the array. Nevertheless, as the order of non-linearity increases, the residual distortion remains non-negligible and can limit both spectral efficiency and system reliability. Therefore, a careful trade-off between power efficiency and linearity is required when selecting the PA operating point in massive MIMO transmitters[46].

1.3.5.1 2 carrier analysis power series analysis

In practical massive MIMO transmitters, multicarrier waveforms such as OFDM are widely used due to their high spectral efficiency and robustness to multipath fading. However, analyzing the effects of power amplifier PA non-linearities directly on OFDM signals is challenging because OFDM consists of a large number of closely spaced subcarriers, which interact in complex ways under non-linear amplification. To facilitate tractable analysis while retaining the essential characteristics of intermodulation and spectral regrowth, a two-carrier (two-tone) power series model is commonly employed. In this approach, the OFDM signal is approximated by a two-tone inputs ω_1 and ω_2 as seen in Figure 1.15, which captures the dominant mechanisms of in-band distortion and OOB emissions caused by the PA.

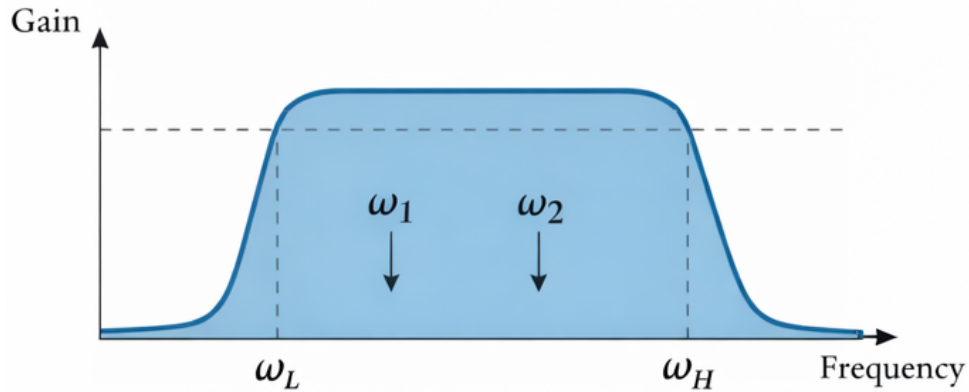


Figure 1.15: Theoretical representation of the RU internal cavity filter

In order to fully understand the effects of PA non-linearities in massive MIMO transmitters, it is instructive to analyze each non-linear order separately. The Massive MIMO RU is equipped with a cavity filter that has cutoff frequencies ω_L and ω_H , as illustrated in Figure 1.15. This filter determines which frequency components generated by non-linearities will pass through the RF chain and which will be attenuated[46].

Among the different non-linear orders, the third-order non-linearity is particularly critical due to its strong correlation with the original signal and its tendency to generate in-band intermodulation products that degrade signal quality.

We consider a simplified two-tone input signal as a proxy for an OFDM waveform[46][48]:

$$x(t) = \cos(\omega_1 t) + \cos(\omega_2 t), \quad (1.1)$$

where ω_1 and ω_2 are the angular frequencies of the two tones. This two-tone model captures the essential mechanisms of intermodulation while keeping the analysis tractable.

Second-order non-linearity: The second-order term of the PA output is

$$y_2(t) = a_2 x^2(t) = a_2 [\cos(\omega_1 t) + \cos(\omega_2 t)]^2. \quad (1.2)$$

Expanding using standard trigonometric identities yields

$$y_2(t) = a_2 \left[\frac{1}{2}(1 + \cos 2\omega_1 t) + \frac{1}{2}(1 + \cos 2\omega_2 t) + (\cos(\omega_1 + \omega_2)t + \cos(\omega_1 - \omega_2)t) \right]. \quad (1.3)$$

The resulting frequency components include Direct Current (DC), second harmonics ($2\omega_1, 2\omega_2$), and sum/difference frequencies ($\omega_1 \pm \omega_2$). In the Massive MIMO RU, these second-order components either fall outside the cavity filter passband or at very low frequencies and are therefore largely filtered. Consequently, second-order distortion is generally negligible.

Third-order non-linearity: The third-order term is

$$y_3(t) = a_3 x^3(t) = a_3 [\cos(\omega_1 t) + \cos(\omega_2 t)]^3. \quad (1.4)$$

with a_3 representing the third-order non-linearity coefficient.

Expanding the cube using $(a + b)^3 = a^3 + b^3 + 3a^2b + 3ab^2$, we have

$$x^3(t) = \cos^3(\omega_1 t) + \cos^3(\omega_2 t) + 3\cos^2(\omega_1 t)\cos(\omega_2 t) + 3\cos(\omega_1 t)\cos^2(\omega_2 t). \quad (1.5)$$

Applying the standard trigonometric identities

$$\cos^3 \theta = \frac{3\cos \theta + \cos 3\theta}{4}, \quad \cos^2 \theta \cos \phi = \frac{1}{2}\cos \phi + \frac{1}{4}(\cos(\phi + 2\theta) + \cos(\phi - 2\theta)),$$

the expansion becomes

$$\begin{aligned} \cos^3(\omega_1 t) &= \frac{3}{4}\cos(\omega_1 t) + \frac{1}{4}\cos(3\omega_1 t), \\ \cos^3(\omega_2 t) &= \frac{3}{4}\cos(\omega_2 t) + \frac{1}{4}\cos(3\omega_2 t), \\ 3\cos^2(\omega_1 t)\cos(\omega_2 t) &= \frac{3}{2}\cos(\omega_2 t) + \frac{3}{4}(\cos(2\omega_1 + \omega_2)t + \cos(2\omega_1 - \omega_2)t), \\ 3\cos(\omega_1 t)\cos^2(\omega_2 t) &= \frac{3}{2}\cos(\omega_1 t) + \frac{3}{4}(\cos(\omega_1 + 2\omega_2)t + \cos(\omega_1 - 2\omega_2)t). \end{aligned} \quad (1.6)$$

Combining like terms, the full third-order output is

$$\begin{aligned}
 y_3(t) = a_3 \bigg[& \frac{9}{4}(\cos \omega_1 t + \cos \omega_2 t) \\
 & + \frac{1}{4}(\cos 3\omega_1 t + \cos 3\omega_2 t) \\
 & + \frac{3}{4} \cos(2\omega_1 + \omega_2) t + \frac{3}{4} \cos(2\omega_1 - \omega_2) t \\
 & + \frac{3}{4} \cos(\omega_1 + 2\omega_2) t + \frac{3}{4} \cos(\omega_1 - 2\omega_2) t \bigg]. \quad (1.7)
 \end{aligned}$$

Effect of the Cavity Filter

In practical Massive MIMO systems, the PA output passes through a cavity filter that suppresses high order frequencies. Specifically:

- The third harmonics ($3\omega_1, 3\omega_2$) lie outside the filter passband and are effectively eliminated.
- The sum-frequency products ($2\omega_1 + \omega_2, \omega_1 + 2\omega_2$) typically fall outside the passband and are largely filtered.
- The difference-frequency products ($2\omega_1 - \omega_2, \omega_1 - 2\omega_2$) often lie within the passband and remain in the signal, causing in-band distortion.

Consequently, the simplified third-order output after filtering is

$$y_3^{\text{filtered}}(t) \approx a_3 \left[\frac{9}{4}(\cos \omega_1 t + \cos \omega_2 t) + \frac{3}{4}(\cos(2\omega_1 - \omega_2) t + \cos(\omega_1 - 2\omega_2) t) \right]. \quad (1.8)$$

This expression highlights the dominant in-band third-order intermodulation products, which are the primary source of signal degradation in OFDM-based massive MIMO transmitters.

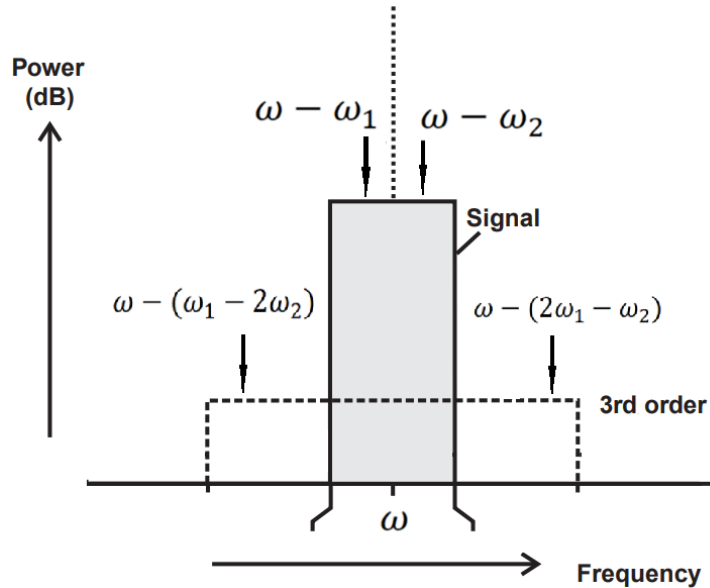


Figure 1.16: Schematic view of the spectral regrowth effect of the 3rd order non linearities

Mathematically, third-order non-linearities in the PA convolve the signal spectrum with itself, generating new frequency components both in- and out-of-band. The Power Spectral Density (PSD) of

the third-order output shows energy spilling into adjacent frequencies, resulting in spectral regrowth as seen in Figure 1.16. This effect is particularly severe in multicarrier OFDM systems, where many subcarriers interact to produce a dense set of intermodulation products, degrading linearity and causing adjacent-channel interference[46].

1.3.5.2 Role of PAPR in the Linear–Nonlinear PA Transition

The PAPR of a transmit signal plays a central role in determining whether a power PA operates in its linear or nonlinear regime. High PAPR signals, such as OFDM waveforms, exhibit occasional large amplitude peaks that significantly exceed the average signal power. While the PA may be biased to operate linearly at the average power level, these instantaneous peaks can drive the amplifier into its nonlinear region.

When the PA enters the nonlinear regime, third-order nonlinearities become dominant, generating in-band intermodulation products and OOB spectral components. As shown in the previous third-order analysis, these nonlinear effects lead to intermodulation distortion and spectral regrowth, which cannot be mitigated by filtering without degrading the desired signal. Therefore, the occurrence of nonlinear distortion is not governed solely by average power, but by the peak excursions of the signal relative to the PA saturation threshold.

In this context, PAPR directly governs the probability and severity of transitions from linear to nonlinear PA operation. High PAPR increases the likelihood of nonlinear behavior, exacerbating third-order distortion and spectral regrowth, whereas PAPR reduction enables operation closer to the linear region while maintaining spectral compliance.

1.3.6 Digital Predistortion

1.3.6.1 Background

Digital predistortion (DPD) is a digital signal processing technique used to improve the linearity of RF power amplifiers, which are inherently nonlinear devices, particularly when operated near saturation. In modern wireless transmitters, power amplifiers must be driven close to their compression region to achieve high power-added efficiency; however, this operating point introduces nonlinear distortion that degrades in-band signal quality and generates spectral regrowth into adjacent channels. DPD addresses this problem by intentionally applying a precisely calculated distortion to the transmit waveform in the digital baseband domain so that the nonlinear effects introduced by the power amplifier are effectively canceled. As a result, the combined response of the predistorter and the power amplifier approximates a linear system[49][50].

The importance of DPD lies in its ability to simultaneously satisfy linearity requirements and maximize transmitter efficiency. Wireless communication standards such as 5G NR and LTE, defined by 3GPP and ETSI, impose strict limits on ACLR and spectral emissions that are verified through Spectrum Emission Mask (SEM). Without DPD, power amplifiers must operate with significant output power back-off to meet these requirements, leading to poor efficiency and increased power consumption. By enabling linear operation at higher output power levels, DPD allows the power amplifier to be driven further into saturation while still meeting ACLR specifications, often resulting in a two fold or greater improvement in effective power amplifier efficiency[51][52][50].

The functional principle of DPD is based on behavioral modeling of the power amplifier in the digital baseband domain. A mathematical model of the amplifier is created using digital samples of the input and output signals passing through the power amplifier. This model captures the in-band nonlinear characteristics of the amplifier, including amplitude and phase distortions and in wideband systems, memory effects. Once the amplifier model is established, its inverse function is implemented as a predistorter. The predistorter modifies the transmit data before it is converted to the analog domain, such

that when the predistorted signal passes through the nonlinear power amplifier, the resulting output signal is highly linear[49][50].

From a theoretical perspective, the Volterra series[53] offers a rigorous mathematical framework to describe nonlinear systems with memory. The series expansion captures both static nonlinearities and memory effects through cross terms of delayed input samples. However, the general Volterra model suffers from exponential growth in the number of coefficients with increasing nonlinear order and memory depth, making it computationally infeasible for real-time implementation. To address this, simplified structures such as memory polynomials, Wiener and Hammerstein models, and dynamic deviation reduction methods were proposed in the literature [53]. These reduced-complexity models preserve essential nonlinear dynamics while drastically lowering the number of required parameters, forming the practical basis for DPD algorithms.

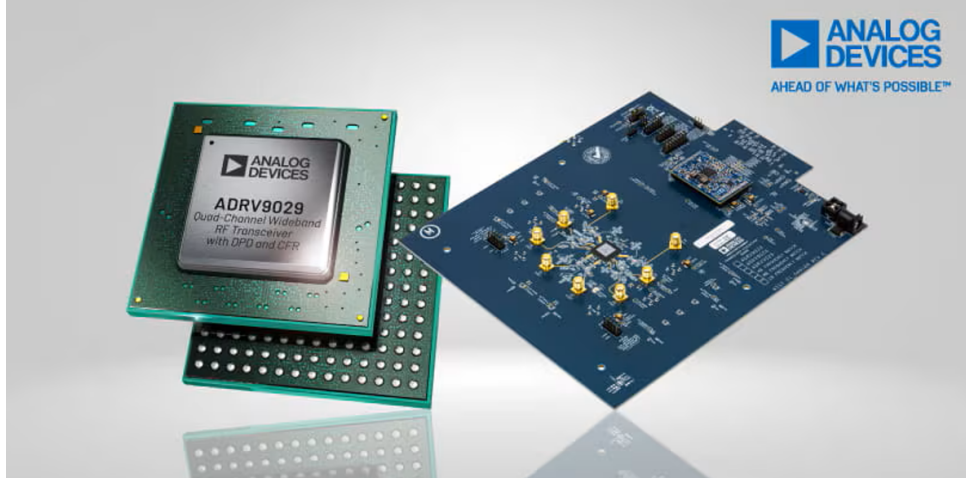


Figure 1.17: ADRV9029 that supports DPD [54]

Commercial implementations reflect these theoretical advances. For example, the Analog Devices ADRV9025 transceiver adopts a Generalized Memory Polynomial (GMP) model, which is an extension of the memory polynomial including cross-terms that capture both in-phase and quadrature distortions. The transceiver integrates an observation receiver that samples the PA output, and an embedded processor that performs adaptive coefficient estimation in real time. This closed-loop adaptation ensures robustness against slow variations in device characteristics, such as temperature drift and aging. The ADRV9025 architecture represents a balance between theoretical generality and hardware efficiency, and it has been further refined in its successor, the ADRV9029 transceiver [54].

The ADRV9025 system is a practical realization of Volterra-based modeling principles.

1.3.6.2 DPD direct learning algorithm

In this work, a direct learning DPD architecture is considered. Unlike indirect learning approaches, which first identify an amplifier model and then invert it, direct learning adapts the predistorter coefficients directly by minimizing the error between the observed output and the original transmit signal[53][50]. This approach is computationally efficient, robust to modeling errors, and well suited for real-time embedded implementations as seen in Figure 1.18.

1. Signal Model and Predistorter Structure

Let $u(n) \in \mathbb{C}$ denote the discrete-time complex baseband transmit signal at sample index n , where $n = 1, 2, \dots, N$, and N is the number of samples used for coefficient estimation. The signal $u(n)$ represents

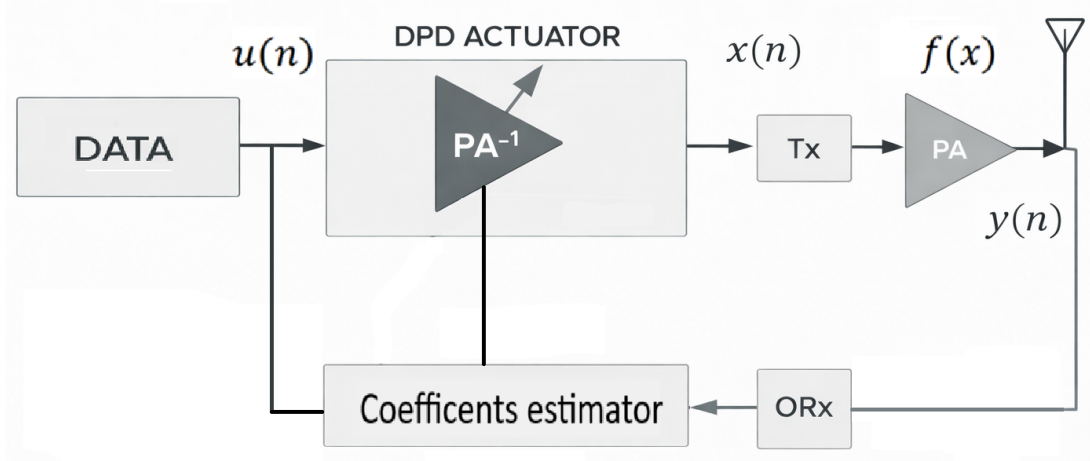


Figure 1.18: Schematic representation of the DPD block diagram in the baseband domain in the ADRV9026 systems

the ideal linear reference that the transmitter aims to reproduce at the output of the amplifier.

The predistorter applies a nonlinear transformation to the input signal prior to digital-to-analog conversion. In this study, a memoryless polynomial predistorter is assumed, without memory effects or cross terms. The predistorted signal $x(n) \in \mathbb{C}$ is defined as

$$x(n) = \sum_{k=1}^K c_k |u(n)|^{k-1} u(n),$$

where K denotes the number of predistortion coefficients (equal to the polynomial order), and $c_k \in \mathbb{C}$ are complex-valued coefficients that control both amplitude and phase predistortion.

2. Nonlinear System and Observation Model

The predistorted signal $x(n)$ is converted to an analog waveform, upconverted to RF, and applied to the nonlinear power amplifier device. The nonlinear system is modeled as an unknown nonlinear function $f(\cdot)$, such that the complex baseband equivalent of the observed output is given by

$$y(n) = f(x(n)).$$

The output is sampled by an observation receiver, downconverted to baseband, and digitally processed to align it with the reference signal $u(n)$. This alignment compensates for timing offset, frequency offset, phase rotation, and gain mismatch.

3. Direct Learning Error Definition

In the direct learning architecture, the error signal is defined as the difference between the observed output and the original transmit signal. The instantaneous error sample $e(n) \in \mathbb{C}$ is defined as

$$e(n) = y(n) - u(n).$$

This definition establishes the fundamental objective of the algorithm: to adapt the predistorter such that the cascade of the predistorter and the nonlinear system behaves as a linear system with unity gain.

4. Basis Functions and Regression matrix

To estimate the predistorter coefficients, nonlinear basis functions are constructed from the input signal. For each sample n , the basis function vector $\phi(n) \in \mathbb{C}^{K \times 1}$ is defined as

$$\phi(n) = \begin{bmatrix} u(n) \\ |u(n)|u(n) \\ |u(n)|^2 u(n) \\ \vdots \\ |u(n)|^{K-1} u(n) \end{bmatrix}.$$

By stacking the basis vectors for all N samples, the regression matrix $\mathbf{F} \in \mathbb{C}^{N \times K}$ is obtained as

$$\mathbf{F} = \begin{bmatrix} \phi(1)^T \\ \phi(2)^T \\ \vdots \\ \phi(N)^T \end{bmatrix}.$$

The error samples are collected into an error vector $\mathbf{E} \in \mathbb{C}^{N \times 1}$,

$$\mathbf{E} = \begin{bmatrix} e(1) \\ e(2) \\ \vdots \\ e(N) \end{bmatrix}.$$

5. Linear Error Model and Coefficient Estimation

Although the overall transmitter system is nonlinear with respect to the signal amplitude, the error can be expressed as a linear function of the predistorter coefficients. This leads to the linear error model

$$\mathbf{E} = \mathbf{F}\mathbf{C},$$

where the coefficient vector $\mathbf{C} \in \mathbb{C}^{K \times 1}$ is

$$\mathbf{C} = \begin{bmatrix} c_1 \\ c_2 \\ \vdots \\ c_K \end{bmatrix}.$$

The coefficients are updated iteratively using a gradient descent-based adaptation rule. Let $\mathbf{C}^{(i)}$ denote the coefficient vector at iteration i - for each sample n . The update equation is

$$\mathbf{C}^{(i+1)} = \mathbf{C}^{(i)} - \mu \mathbf{F}^H \mathbf{E},$$

where $\mu \in [0, 1]$ is the convergence factor and $\mathbf{F}^H$ denotes the Hermitian transpose of the regression matrix, it is usually taken as 0.5. The coefficients of vector $\mathbf{C}^{(i+1)}$ are then injected back again to the DPD actuator where the vector $x(n)$ is computed again for the same sample n . Once convergence is satisfied, the algorithm repeats for sample $n + 1$.

6. Convergence and Stopping Criteria

The adaptation process is executed iteratively until convergence is achieved. Convergence is typically assessed using the Mean Squared Error (MSE),

$$\text{MSE}^{(i)} = \frac{1}{N} \sum_{n=1}^N |e^{(i)}(n)|^2.$$

The algorithm is considered to have converged when the reduction in MSE between successive iterations falls below a predefined threshold. Alternatively, convergence may be determined based on the norm of the coefficient update vector.

The number of coefficients K strongly influences convergence behavior. Smaller values of M lead to faster convergence but limited modeling capability, whereas larger values of M provide improved linearization at the cost of increased convergence time and sensitivity to noise.

In real-time implementations, the algorithm is executed periodically. If convergence criteria are met, the coefficients are retained. If operating conditions change, increased error automatically triggers renewed adaptation.

1.3.7 Z3RO precoder

In large array systems such as Massive MIMO, PAs represent a dominant source of energy consumption, typically accounting for a significant fraction of the total power budget of a base station. Their energy efficiency is maximized when operating close to saturation, yet this regime introduces nonlinear distortion that significantly degrades signal quality at the receiver. More precisely, the combination of multicarrier waveforms such as OFDM and precoding over a large number of antennas leads to high PAPR, which forces the PA to operate over a wide dynamic range and exacerbates nonlinear effects. This creates a fundamental tension between energy efficiency and signal linearity that has long been a central challenge in the design of wireless transmitters.

1.3.7.1 Limitations of the MRT Precoder Under PA Nonlinearity

A central issue in this context is that conventional precoding strategies, and in particular MRT, do not account for nonlinear distortion in their design. MRT is derived by maximizing the received Signal-to-Noise Ratio (SNR) under the assumption that the PA operates in its linear regime, which implicitly sets the nonlinearity coefficient to zero. In this precoder, the signal transmitted from antenna m before PA is given by $x_m = w_m^{\text{MRT}} s$, where s denotes the complex transmit symbol with variance p , and $w_m^{\text{MRT}} = \alpha h_m^*$ is the conjugate beamforming weight corresponding to the channel coefficient h_m . The normalization constant $\alpha = \sqrt{\frac{M}{\sum_{m'} |h_{m'}|^2}}$ ensures a total transmit power of pM , and the resulting SNR scales as M times the power per-antenna channel, reflecting the well-known gain of the MRT precoding matrix.

When the PA is modeled by a memoryless third-order nonlinearity, its output for antenna m is expressed as:

$$y_m = x_m + a_3 x_m |x_m|^2, \quad (1.9)$$

where the complex coefficient a_3 characterizes the distortion of the amplifier. The signal received at the single-user location is then:

$$r = \sum_{m=0}^{M-1} h_m y_m + v = s \sum_{m=0}^{M-1} h_m w_m + a_3 |s|^2 \sum_{m=0}^{M-1} h_m w_m |w_m|^2 + v, \quad (1.10)$$

where v is additive white Gaussian noise with variance σ_v^2 . The second summation constitutes the third-order distortion term that reaches the receiver. Under MRT precoding with a line-of-sight (LoS) channel $h_m = \sqrt{\beta} e^{-j\phi_m}$ and normalization $\alpha = \frac{1}{\sqrt{\beta}}$, this expression simplifies to:

$$r = \sqrt{\beta} M (s + a_3 s |s|^2) + v, \quad (1.11)$$

which makes explicit that both the useful signal and the third-order distortion term are coherently combined at the receiver and benefit from exactly the same array gain M . As a direct consequence, the signal-to-distortion ratio (SDR) in the user is independent of M and does not improve by adding more antennas. When PA approaches saturation and the distortion power becomes non-negligible, the SDR becomes the dominant bottleneck, ultimately limiting Signal to Distortion Noise Ratio (SDNR) regardless of the array size. This phenomenon is not specific to LoS channels: it arises from the fact that the MRT precoder phase-aligns both the linear and nonlinear components of the PA output toward the user, so that both combine coherently in the same spatial direction.

The Z3RO Precoder Design

To address this fundamental limitation, Rottenberg proposed the Zero Third-Order distortion (Z3RO) precoder[55]. The core idea is to incorporate the distortion cancellation condition directly into the precoder optimization, rather than relying on hardware-level linearization or transmit power reduction. Specifically, the precoder weights w_m are designed to jointly maximize the received SNR and enforce the cancellation of the third-order distortion at the user location. The distortion cancellation constraint reads:

$$\sum_{m=0}^{M-1} h_m w_m |w_m|^2 = 0. \quad (1.12)$$

A particularly important property of this constraint is that it does not involve the PA parameter a_3 nor the statistics of the transmit signal s . This means that the precoder can be computed solely based on the estimated channel coefficients h_m , without requiring any characterization of the amplifier model or knowledge of the signal distribution, a significant practical advantage over DPD-based approaches, which must be calibrated and updated for each PA individually.

The full optimization problem is formulated as:

$$\max_{w_0, \dots, w_{M-1}} \left| \sum_{m=0}^{M-1} h_m w_m \right|^2 \quad \text{s.t.} \quad p \sum_{m=0}^{M-1} |w_m|^2 = pM, \quad \sum_{m=0}^{M-1} h_m w_m |w_m|^2 = 0. \quad (1.13)$$

This is a non-convex constrained optimization problem. For a LoS channel, it is solved analytically via a change of variable $g_m = w_m e^{-j\phi_m}$ that separates the phase steering from the amplitude design, reducing the problem to a real-valued constrained optimization. The Lagrangian analysis reveals that at any critical point, the coefficients g_m can only take two distinct real values, one for a subset of M_s antennas and another for the remaining $M - M_s$ antennas. Applying the power and zero-distortion constraints simultaneously yields the closed-form Z3RO precoder:

$$w_m^{\text{Z3RO}, M_s} = \alpha e^{j\phi_m} \begin{cases} -\left(\frac{M - M_s}{M_s}\right)^{1/3} & \text{if } m = 0, \dots, M_s - 1, \\ 1 & \text{otherwise,} \end{cases} \quad (1.14)$$

where $\alpha = \frac{\sqrt{M}}{\sqrt{(M - M_s) + M_s^{\frac{1}{3}} (M - M_s)^{\frac{2}{3}}}}$ is the power normalization constant, and M_s is the number of so-called

saturated antennas, with $0 < M_s < \frac{M}{2}$. The physical interpretation of this design is transparent: a small subset of M_s antennas is deliberately driven with an amplified and phase-inverted weight, such that the third-order distortion they produce at the user exactly cancels the cumulative distortion generated by all remaining $M - M_s$ antennas. The M_s antennas can be arbitrarily chosen from the array, and no specific placement is required. The mechanism can be understood as a spatial distortion null-steering: just as conventional null-steering places a radiation null toward an interfering source, Z3RO places a distortion null at the intended user, using the array's spatial degrees of freedom to cancel, rather than to avoid, the unwanted emission[55].

Array Gain and the M_s Trade-off

The received SNR under the Z3RO precoder is given in closed form as:

$$\text{SNR}_{\text{Z3RO}} = \frac{\beta p}{\sigma_v^2} \cdot M \cdot \frac{((M - M_s)^{2/3} - M_s^{2/3})^2}{(M - M_s)^{1/3} + M_s^{1/3}}. \quad (1.15)$$

Maximizing (1.15) over $M_s \in (0, M/2)$ shows that the array gain is a monotonically decreasing function of M_s , so the globally optimal choice in terms of SNR is $M_s = 1$. The array gain penalty of Z3RO relative to MRT then diminishes asymptotically:

$$\lim_{M \rightarrow +\infty} \frac{\text{SNR}_{\text{Z3RO}}}{\text{SNR}_{\text{MRT}}} = 1, \quad (1.16)$$

confirming that the cost of enforcing distortion cancellation becomes negligible as the array size grows. In practice, for $M = 64$ antennas and $M_s = 1$, Z3RO achieves an array gain of approximately 16.5 dB, compared to 18 dB for MRT, a penalty of only 1.5 dB, while completely nulling the third-order distortion at the location of the user. For $M = 256$, the penalty reduces further to below 0.5 dB.

Although $M_s = 1$ is optimal from a pure SNR point of view, the use of a larger number of saturated antennas offers a meaningful secondary benefit related to the spatial distribution of the residual distortion. When $M_s = 1$, the distortion is canceled by the user, but the total radiated distortion power increases relative to the MRT, spreading diffusely in all other spatial directions. This can increase interference at unintended receivers or contribute to out-of-band emissions in adjacent frequency bands. As M_s increases, two effects occur: the total radiated distortion power decreases (because the distortion benefits from a beamforming gain and is concentrated rather than spread), and its angular pattern becomes progressively more focused in the vicinity of the user beam, null exactly at the user's direction, but concentrated nearby. This spatial focusing is beneficial in coexistence scenarios and reduces the risk of adjacent-channel interference at off-beam locations. The choice of M_s thus offers a tunable trade-off between maximizing array gain on the one hand, and minimizing the total and spatially distributed radiated distortion power on the other[55].

Positioning Relative to Existing Solutions

The Z3RO precoder occupies a distinct and complementary position in the landscape of PA linearization and distortion mitigation techniques. Power back-off is the simplest approach but directly reduces PA efficiency and lowers effective transmit power, negatively impacting coverage — particularly for cell-edge users who already experience large path losses. DPD pre-compensates the PA input signal to cancel the nonlinear response, and can be highly effective when properly calibrated. However, it requires a dedicated feedback path per-antenna, relies on accurate and continuously updated identification of PA models, and involves significant implementation complexity in large arrays with hundreds of RF chains. Furthermore, even a theoretically perfect DPD only linearizes signals whose instantaneous amplitude remains below

the PA saturation threshold; any peak that exceeds this level is irreversibly clipped, producing residual distortion that no pre-distortion scheme can eliminate.

Z3RO operates at a fundamentally different level of the transmitter chain: rather than modifying the signal waveform at each antenna or altering the PA hardware, it redistributes the spatial precoding weights to exploit the large number of degrees of freedom offered by the antenna array. Its implementation complexity is comparable to that of MRT, it requires no PA characterization, no feedback loop, and no knowledge of the signal statistics. Crucially, Z3RO is fully orthogonal to DPD and can be deployed alongside it: DPD reduces the per-antenna nonlinear distortion power, while Z3RO ensures that any residual coherent distortion is spatially canceled at the user location. The combination of both techniques is therefore strictly superior to either alone.

The main limitations of the current Z3RO formulation are its restriction to the single-user case, where a single distortion null is sufficient, and to third-order nonlinearities. In a multi-user MIMO setting, each additional user introduces an additional spatial direction toward which distortion must be canceled, requiring a precoder design with multiple simultaneous constraints. Similarly, accounting for higher-order Volterra terms would require additional cancelation constraints per user, consuming more spatial degrees of freedom [55].

1.3.8 Beamforming

Beamforming is a spatial filtering technique used in antenna array systems to control the directionality of transmitted or received electromagnetic waves. Unlike a single antenna, which radiates power isotropically (in all directions), an antenna array can shape its radiation pattern by adjusting the relative phases and amplitudes of the signals at each antenna element. This allows constructive interference in desired directions and destructive interference elsewhere, effectively steering the beam.

The concept of beamforming originates from the theory of array signal processing, which was historically used in radar and sonar systems, and later adopted in wireless communications to enhance signal quality, coverage, and spectral efficiency.

1.3.8.1 Background

Beamforming is fundamentally about controlling the directionality of antenna radiation. In a wireless communication system, the energy radiated by the antennas can spread in all directions. This omnidirectional propagation is inefficient because much of the transmitted energy may not reach the intended receiver and may cause interference for others. Beamforming solves this by shaping the radiation pattern to focus energy in the desired directions[56].

A planar antenna array consists of multiple antenna elements arranged in a two-dimensional grid. Unlike linear arrays, which steer beams in one plane, planar arrays can steer beams both horizontally and vertically, giving full control over the 3D spatial direction, as illustrated in Figure 1.19. Each antenna element contributes to the overall radiation pattern, and by adjusting the relative phase and amplitude of the signal feeding each element, the array can constructively interfere in the desired direction and destructively interfere elsewhere. This is the essence of beamforming.

Beamforming relies on the principle of superposition. Signals from individual elements combine in space, reinforcing each other in some directions (main lobe) and canceling in others (side lobes). The resulting radiation pattern can be dynamically shaped to track users, reduce interference, or maximize the power of the received signal. In wireless communication, this is particularly useful in environments with multipath propagation, where signals bounce off obstacles. By directing the beam toward the intended receiver, planar matrices can enhance SNR and reduce interference from unwanted directions [57].

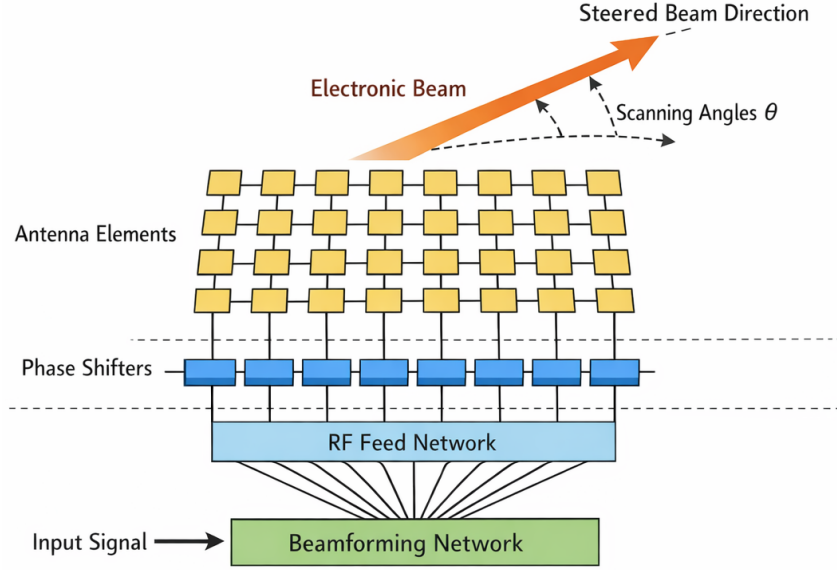


Figure 1.19: Schematic view of a typical planar phased array antenna with phase shifters for beam scanning and steering.

1.3.8.2 Antenna and array beam pattern modeling

Antenna Element Radiation Pattern Model The antenna element radiation pattern $A_E(\theta, \varphi)$ is modeled using the standardized 3GPP directional antenna model, defined in terms of horizontal (azimuth) and vertical (elevation) attenuation functions. This model is widely adopted for Massive MIMO and O-RAN system-level evaluations.

Parameter	Symbol	Unit
Front-to-back ratio	A_m	dB
Vertical side-lobe attenuation	SLA_v	dB
Horizontal half-power beamwidth	φ_{3dB}	degrees
Vertical half-power beamwidth	θ_{3dB}	degrees
Element peak gain	$G_{E,max}$	dBi

Table 1.4: Antenna element parameters

The horizontal (azimuth) attenuation is given by

$$A_H(\varphi) = \min \left[12 \left(\frac{\varphi}{\varphi_{3dB}} \right)^2, A_m \right], \quad (1.17)$$

where φ denotes the azimuth angle in degrees.

The vertical (elevation) attenuation is defined as

$$A_V(\theta) = \min \left[12 \left(\frac{\theta - 90^\circ}{\theta_{3dB}} \right)^2, SLA_v \right], \quad (1.18)$$

where $\theta = 90^\circ$ corresponds to the antenna bore sight direction.

The total antenna element gain is expressed as

$$A_E(\theta, \varphi) = G_{E,\max} - \min[A_H(\varphi) + A_V(\theta), A_m], \quad (1.19)$$

where $G_{E,\max}$ denotes the maximum gain of the antenna element and all corresponding parameters are defined in 1.4. This antenna element model provides a computationally efficient and physically meaningful representation of directional radiation and is suitable for Massive MIMO, O-RAN architectures, and system-level performance and interference analysis.

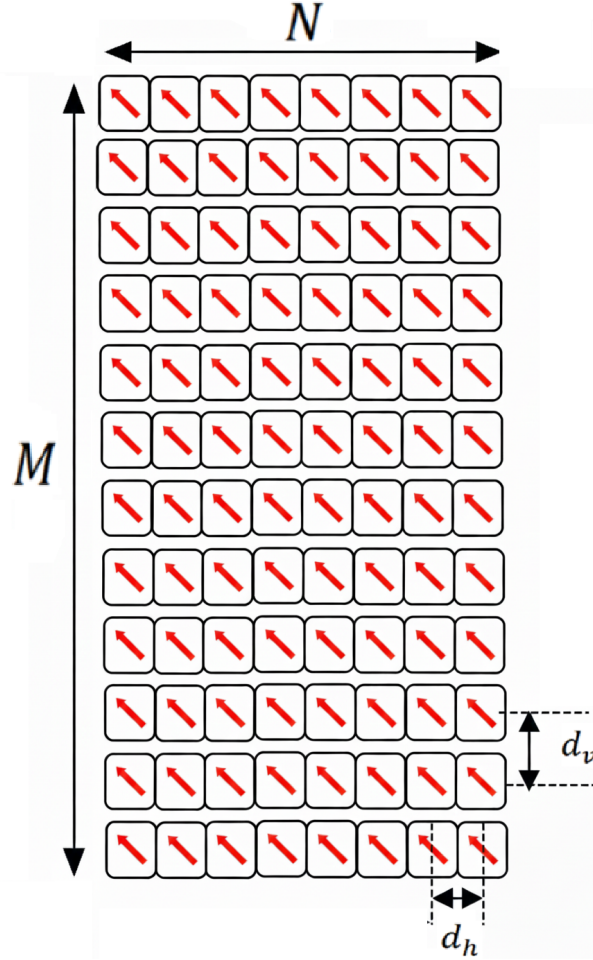


Figure 1.20: General antenna array model.

The radiation pattern of the antenna array for a beamformed antenna array, as an example in Figure 1.20 is computed using all antenna elements. For a specific antenna array, we define[58]:

$$A_A(\theta, \varphi) = A_E(\theta, \varphi) + 10 \log_{10} \left(\left| \sum_{m=1}^M \sum_{n=1}^N w_{m,n} \cdot v_{m,n} \right|^2 \right) \quad (1.20)$$

where the **superposition vector** is:

$$v_{m,n} = \exp \left\{ j 2 \pi \left((m-1) \frac{d_v}{\lambda} \cos(\theta) + (n-1) \frac{d_h}{\lambda} \sin(\theta) \sin(\varphi) \right) \right\}, \quad n = 1, \dots, N, \quad m = 1, \dots, M \quad (1.21)$$

and the **weighting vector** is:

$$w_{m,n} = \frac{1}{\sqrt{MN}} \exp \left\{ j2\pi \left((m-1) \frac{d_v}{\lambda} \sin(\theta_{\text{etilt}}) - (n-1) \frac{d_h}{\lambda} \cos(\theta_{\text{etilt}}) \sin(\varphi_{\text{escan}}) \right) \right\}. \quad (1.22)$$

The definition of the different parameters in the formulas is defined in table 1.5. Note that this formulation gives the correct array radiation pattern; however, the absolute gain is accurate only if the element pattern $A_E(\theta, \varphi)$ is computed for the actual element spacing (d_h, d_v) . For other element spacings, the element pattern must be recalculated to match the spacing. If the element pattern is not linked to the element spacing, the computed absolute gain may deviate from the correct value, and the deviation varies with the beam steering direction.

Parameter	Definition
$A_A(\theta, \varphi)$	Array radiation pattern in dB for a given direction (θ, φ) .
$A_E(\theta, \varphi)$	Single-element radiation pattern (in dB); must match element spacing for absolute gain.
N	Number of antenna elements in the horizontal direction.
M	Number of antenna elements in the vertical direction.
$w_{m,n}$	Complex weighting coefficient for element (m, n) , defining the steering direction.
$v_{m,n}$	Superposition vector for element (m, n) representing the phase contribution to the array factor.
d_H	Horizontal spacing between antenna elements.
d_V	Vertical spacing between antenna elements.
λ	Wavelength of the carrier signal.
θ, φ	Elevation and azimuth angles of observation.
θ_{etilt}	Elevation steering angle.
φ_{escan}	Azimuth steering angle.
m, n	Indices for vertical (m) and horizontal (n) antenna elements.

Table 1.5: Definition of Parameters for the Composite Array Radiation Pattern[58]

Modified AAS system An antenna array system can be modeled hierarchically by grouping antenna elements into subarrays and then combining these subarrays to form the full antenna. Each individual antenna element is characterized by its element radiation pattern $A_E(\theta, \varphi)$, which describes the directional behavior of a single element and does not include any beam steering or tilt.

Within each subarray, the same fixed phase shifts are applied to the signals feeding the individual elements. These phase shifts are chosen to steer the sub-array radiation toward a predefined direction $(\theta_{\text{etilt}}, \varphi_{\text{escan}})$ where θ_{etilt} is downtilt in the positive direction, referred to as the subarray pre-tilt.

Figure 1.21 represents the modified antenna array system, where M is the number of subarrays per column, d_v is the spacing between 2 different subarrays in the same column, we also define M_{sub} as the number of radiating elements per subarray and $d_{v,\text{sub}}$ as the spacing between 2 radiating elements where $d_v = M_{\text{sub}} d_{v,\text{sub}}$, and for that the antenna array pattern $A_A(\theta, \varphi)$ will be written eventually as a function of the subarray beam pattern $A_{\text{sub}}(\theta, \varphi)$ as:

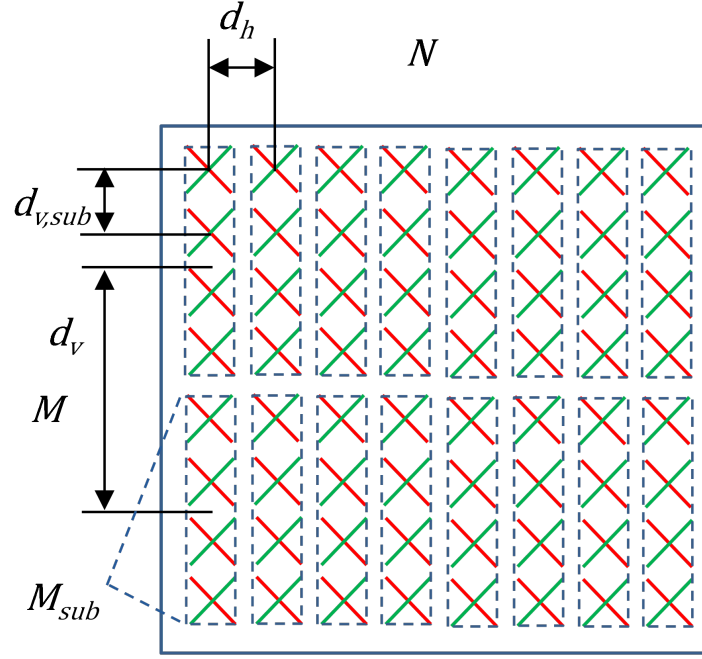


Figure 1.21: 3GPP 37840 extracted theoretical antenna array model [56]

$$A_A(\theta, \varphi) = A_{\text{sub}}(\theta, \varphi) + 10 \log_{10} \left(\left| \sum_{m=1}^M \sum_{n=1}^N w_{m,n} \cdot v_{m,n} \right|^2 \right)$$

In order to understand the relationship between the beam pattern and the assigned beam weights, we will simulate the antenna array used for our experiments. The antenna array follows the same beam pattern representation as defined previously, however, each 6 radiating elements will constitute one subarray, where they will be assigned the same weights. To understand the modification, we refer to Figure 1.21 where each subarray will be treated as a radiating element. The antenna array parameters then become: $d_{v,\text{sub}} = 0.7\lambda$, $d_s = 0.5\lambda$, $M = 2$, $N = 8$, $M_{\text{sub}} = 6$, and then we calculate $d_v = M_{\text{sub}} d_{v,\text{sub}} = 4.2\lambda$. We will simulate 3 different beams in the azimuth and elevation plane respectively:

The radiation patterns plotted in Figures 1.22, 1.23 and 1.24 in the azimuth and elevation planes show how the antenna distributes its radiated power in different directions. The main lobe corresponds to the direction of maximum radiation and represents the intended pointing direction of the antenna. Its angular width indicates how focused the beam is, with a narrower main lobe implying higher directivity. Side lobes are smaller secondary maxima that appear away from the main lobe and represent undesired radiation in other directions; their relative level compared to the main lobe affects interference and overall performance. The azimuth pattern describes the horizontal angular distribution, while the elevation pattern shows the vertical distribution. Together, these cuts provide information on the three-dimensional radiation behavior of the antenna. In general, higher directivity is associated with a narrower main lobe and lower relative side-lobe levels, while wider beams result in lower directivity. 3D representation of the previous radiation patterns are provided for each case, using Orange internal tool for antenna modeling in Figures 1.25, 1.26, 1.27 and 1.28.

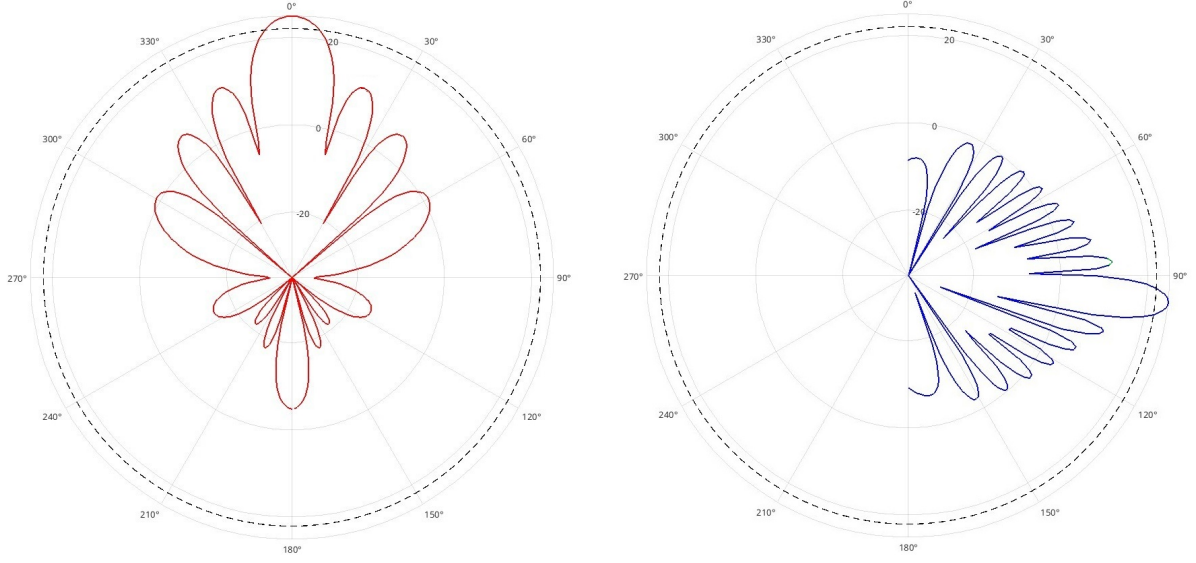


Figure 1.22: Beam pattern in azimuth and elevation plane of a beam pointing at 0° in the azimuth plane and at 96° in the elevation plane with $\theta_{\text{tilt}} = 6^\circ$, $\varphi_{\text{escan}} = 0^\circ$, $\varphi = 0^\circ$ and $\theta = 96^\circ$.

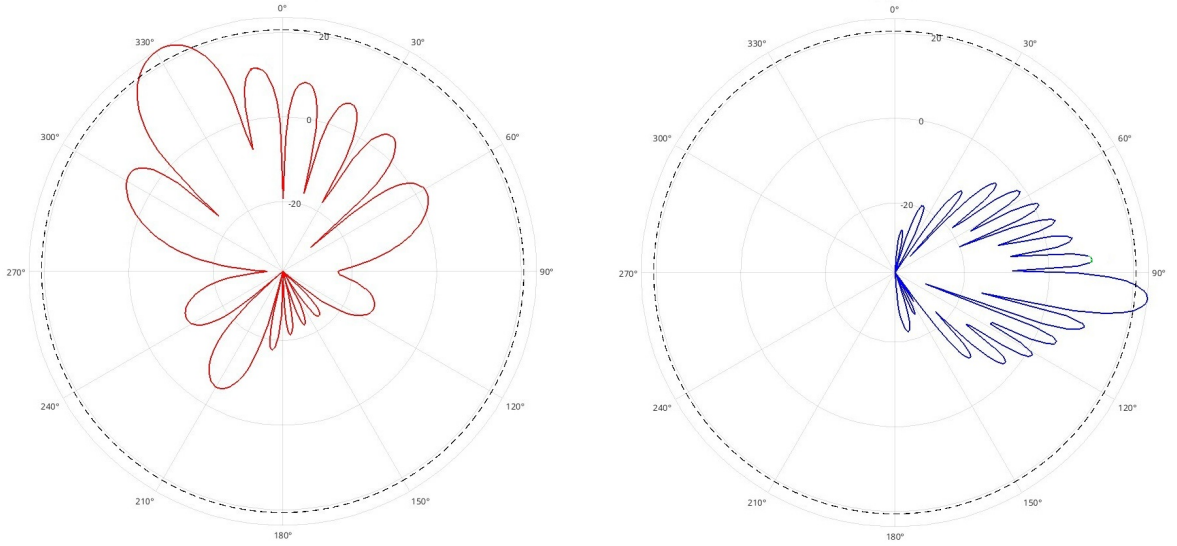


Figure 1.23: Beam pattern in azimuth and elevation plane of a beam pointing at -30° in the azimuth plane and at 96° in the elevation plane with $\theta_{\text{tilt}} = 6^\circ$, $\varphi_{\text{escan}} = -30^\circ$, $\varphi = 0^\circ$ and $\theta = 96^\circ$.

1.3.8.3 Classifications and types of Beamforming

Beamforming techniques can be classified according to their implementation domain, adaptability, and application context. Each category presents specific trade-offs between hardware complexity, flexibility, and overall performance[22].

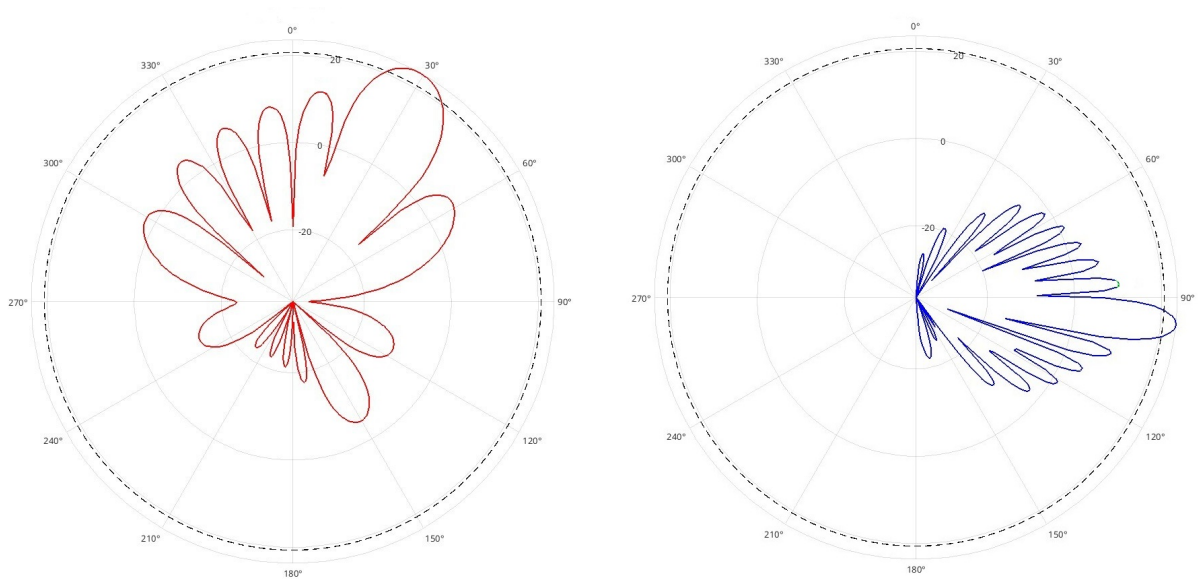


Figure 1.24: Beam pattern in azimuth and elevation plane of a beam pointing at 30° in the azimuth plane and at 96° in the elevation plane with $\theta_{\text{tilt}} = 6^\circ$, $\varphi_{\text{escan}} = 30^\circ$, $\varphi = 0^\circ$ and $\theta = 96^\circ$.

Based on Control Mechanism This classification depends on where the beamforming process occurs in the signal chain (analog, digital, or hybrid domain) as seen in Figure 1.29.

1. **Digital Beamforming:** Performed in the baseband after analog-to-digital conversion, where each antenna has its own RF chain. This allows full control of amplitude and phase, enabling multiple simultaneous beams and adaptive algorithms (e.g. Minimum Mean Squared Error (MMSE)). The main drawbacks are high power consumption and increased hardware complexity.

(a) **Advantages**

- i. High flexibility.
- ii. Supports multi-user MIMO.

(b) **Disadvantages**

- i. Costly to implement.
- ii. Power-intensive operation.

2. **Analog Beamforming:** Implemented in the RF domain using phase shifters and variable gain amplifiers. All antenna elements share a single RF chain, making it hardware-efficient and power-saving. However, it can form only one beam at a time, limiting flexibility. It is commonly applied in mmWave phased-array systems.

(a) **Advantages**

- i. Low complexity.
- ii. Energy-efficient operation.

(b) **Disadvantages**

- i. Limited beam control.

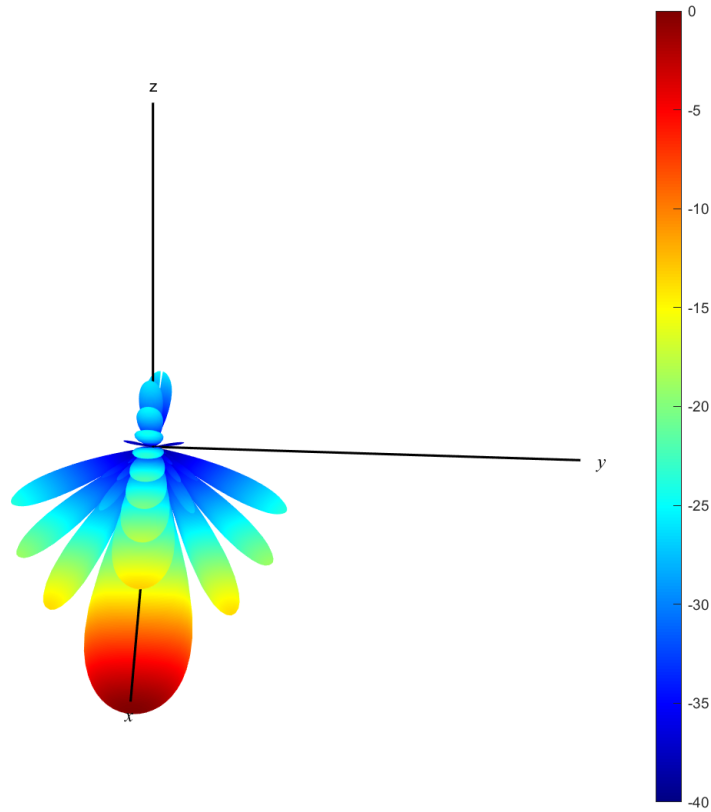


Figure 1.25: 3D beam pattern in azimuth and elevation plane of the beam pointing at 0° in the azimuth plane and at 96° in the elevation plane using Orange internal antenna modeling tool

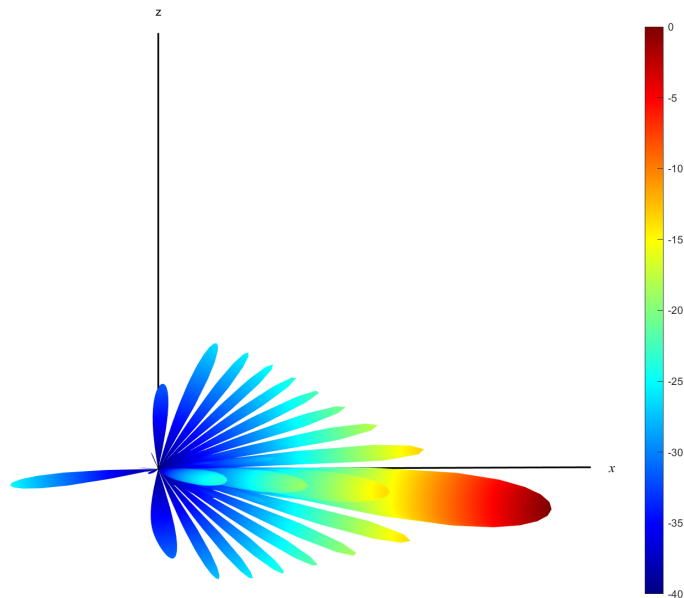


Figure 1.26: 3D beam pattern in azimuth and elevation plane of the beam pointing at 0° in the azimuth plane and at 96° in the elevation plane using Orange internal antenna modeling tool

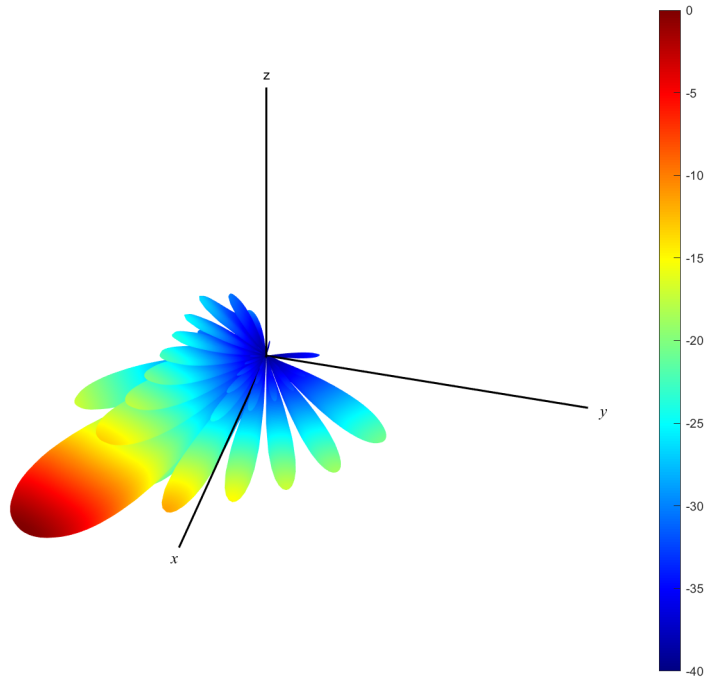


Figure 1.27: 3D beam pattern in azimuth and elevation plane of the beam pointing at -30° in the azimuth plane and at 96° in the elevation plane using Orange internal antenna modeling tool

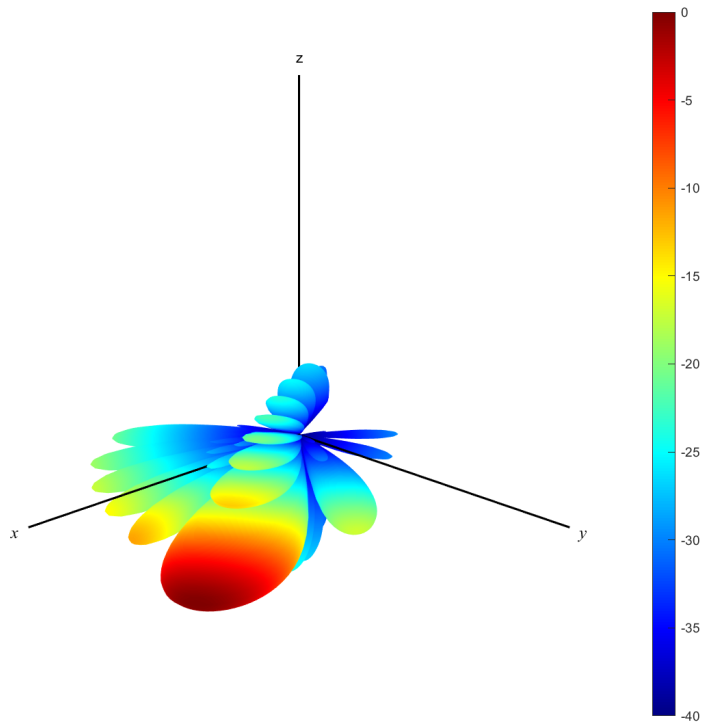


Figure 1.28: 3D beam pattern in azimuth and elevation plane of the beam pointing at 30° in the azimuth plane and at 96° in the elevation plane using Orange internal antenna modeling tool

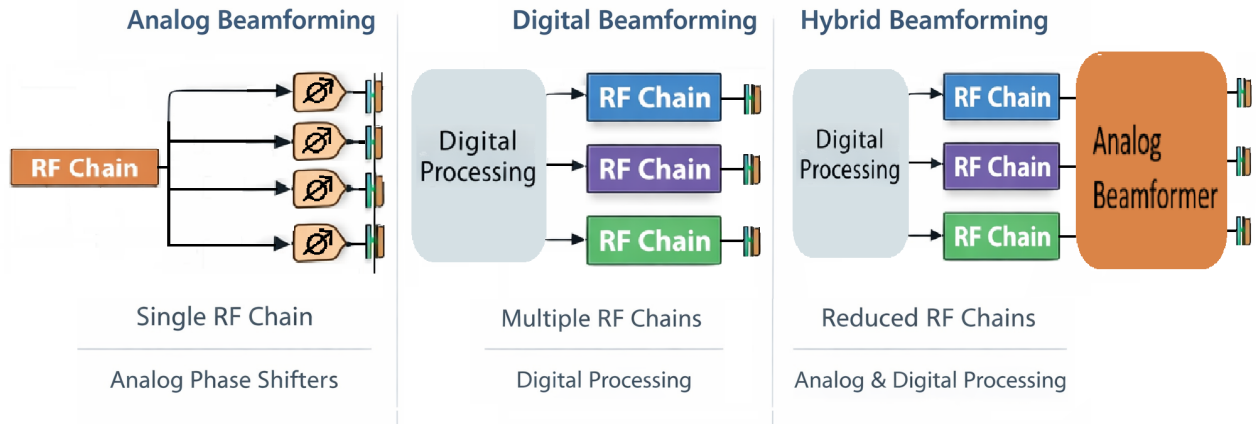


Figure 1.29: Different beamforming classifications[59].

ii. Supports only a single data stream.

3. **Hybrid Beamforming:** Combines analog and digital beamforming to balance flexibility and hardware efficiency. Analog beamforming provides coarse control, while digital beamforming refines beam direction. Hybrid structures are key for 5G and beyond systems, especially for massive MIMO arrays.

(a) **Advantages**

- i. Reduced cost with near-digital performance.

(b) **Disadvantages**

- i. Design complexity.
- ii. Limited resolution due to phase shifters.

Based on Adaptivity Beamforming techniques can be classified according to their ability to adjust to changes in the signal and interference environment. Several beamforming schemes are supported within the ORAN architecture. These methods differ in how beams are defined, how precoding weights are generated, and in what information DU and RU must exchange. The following items summarize and clarify the four main beamforming mechanisms:

1. **Predefined-beam beamforming:** In this scheme, the RU uses a set of beams that are preconfigured and stored locally. The DU selects which beam to activate by sending a beamId. A beamId may refer to a frequency domain beam, a time domain beam, or a hybrid combination of both. Because time domain beams must remain consistent within an OFDM symbol, the DU must ensure correct application of the beamId across PRBs. The internal method used by the O-RU to generate the predefined beams (e.g., per-element gain/phase control, shaped antennas, or other technologies) is not relevant to the DU. At startup, the RU informs the DU via the M-Plane about the beam characteristics, although the DU does not need to understand the implementation details.
2. **Weight-based dynamic beamforming:** In this method, the DU computes real-time beamforming weights that the RU directly applies per antenna element. Because the DU actively generates the weights, it must know the detailed antenna-array configuration of the RU, including the number of vertical and horizontal antenna elements and the spacing between them. The weight vector for each beam is associated with a specific beamId. This approach enables fully dynamic and flexible beam generation driven by channel-dependent weights produced at the DU.

3. **Attribute-based dynamic beamforming:** This method is similar to predefined-beam beamforming in that the DU selects a beamId. However, in this case, the beamId corresponds to a set of beam attributes (e.g., direction, width, or type) rather than a static predefined beam. The O-RU then generates the beam internally according to these attributes, without the DU needing to know how the beam is physically constructed. The DU must ensure correct usage of the beamId, particularly distinguishing whether it represents a frequency-domain or time-domain beam—because different time-domain beams cannot be assigned to different PRBs within the same OFDM symbol. If attribute-based beamforming is used for any layer or spatial stream, then it must be consistently applied to all layers within the same time-frequency resource.
4. **Channel-information-based beamforming:** In this approach, the DU periodically sends channel information for each UE to the O-RU. Based on this information and the per-slot scheduling instructions, the RU computes the appropriate beamforming weights on a slot-by-slot basis. Unlike other methods, no beamId is used; instead, each data section carries a ueId. The RU uses this UE-specific channel information to derive the beamforming weights dynamically for each scheduled slot. If this scheme is applied to one or more layers, the O-DU must ensure that all layers within the affected time-frequency resource use the same beamforming method.

1.4 Research problem and approach

1.4.1 Research problem

The evolution toward open and virtualized radio access networks has placed the ORAN Split 7.2 fronthaul interface at the heart of modern disaggregated RAN architectures. This split distributes lower and higher PHY functionalities between the RU and the DU in a way that enables multi-vendor interoperability, centralized processing, and resource virtualization[17][34]. In parallel, multi-antenna techniques, particularly MIMO and digital or hybrid beamforming, have become essential for enhancing spectral efficiency and delivering high downlink throughput in 5G and beyond[38][21]. Yet, the coexistence of these two developments raises important architectural and signal-processing questions. The fronthaul constraints introduced by Split 7.2 impose new limits and trade-offs on how downlink beamforming is generated, represented, and transmitted across the DU/RU boundary, especially in systems with many antennas[37].

In a downlink-only context, the absence of CSI feedback and uplink measurement signaling fundamentally reshapes this problem. Because the DU cannot rely on instantaneous UE-reported CSI, and because no higher-layer RIC-driven optimization loop is present, all beamforming operations must rely on information already available at the DU or RU at transmission time (e.g., predefined beam codebooks, sector-level beam patterns). This constraint directly impacts where the beamforming and precoding computations should reside. Centralized DU-generated digital precoding can benefit from a unified view of scheduling and resource allocation, but at the cost of significantly higher fronthaul bandwidth, since per-antenna resource-mapped symbols must be transmitted through the 7.2 interface. In contrast, delegating more of the beamforming tasks to the RU reduces fronthaul load but restricts beamforming flexibility to the capabilities and algorithms supported locally, typically with limited per-UE adaptation.

Furthermore, the choice of split and the corresponding mapping of downlink PHY tasks affect not only bandwidth but also fronthaul determinism and latency. Downlink beamforming, particularly in multi-layer MIMO[40], requires precise and time-aligned per-antenna transmissions, yet Split 7.2 introduces stringent timing coordination between DU-generated data and RU-generated lower-PHY operations. These timing requirements become more severe as the number of antennas increases and as the system aims to support fine-grained beam selection or multi-layer transmission strategies[34]. In the absence of CSI feedback loops, the system must rely on robust and stable beam configurations, making the interplay between beam granularity, fronthaul overhead, and RU processing capabilities even more critical.

This leads to the central research problem of the thesis: The first challenge lies in the practical integration of the complete ORAN stack, including the Distributed Unit (DU), the Radio Unit (RU) and the associated transport planes, with a special focus on the synchronization plane, which is critical for coherent massive MIMO operation. Ensuring correct interoperability, timing accuracy, and stable operation across these components remains a non-trivial problem that must be experimentally validated under realistic conditions[17][34].

Building on this integrated ORAN platform, a second challenge concerns the experimental evaluation of beamforming schemes in massive MIMO antenna systems. While beamforming is a key enabler for spatial multiplexing and high spectral efficiency, its implementation in real ORAN systems raises questions regarding performance, stability, and reproducibility when multiple hardware and software components interact.

Finally, this thesis addresses the problem of characterizing and understanding the spatial behavior of OOB emissions in massive MIMO antenna systems[48]. These emissions, primarily generated by nonlinear imperfections in the power amplifiers, may exhibit spatial selectivity due to beamforming and antenna array effects[46]. The interaction between PA non-linearities, beamforming strategies, and massive MIMO architectures in an ORAN context remains insufficiently explored experimentally. Understanding how OOB emissions are distributed in space in the single user and multi-case[48], with and without DPD[54][49] and how they evolve under different beamforming configurations is essential for interference mitigation and regulatory compliance in future wireless networks[51][52].

Addressing these questions is essential to clarify the capabilities and limitations of downlink MIMO transmission in an ORAN Split 7.2 framework, and to guide the design of scalable, interoperable, and high-performance open RAN deployments.

1.4.2 Approach

To address the challenges identified in the problematic and the limitations highlighted by the state of the art, this thesis follows a solution approach that combines practical ORAN integration with controlled experimental evaluation of downlink beamforming procedures. The first component of the approach consists of integrating the ORAN Split 7.2 fronthaul interface using the OAI software framework in order to realize a fully interoperable DU/RU architecture[60][34]. This integration enables fine-grained access to the PHY processing boundary imposed by Split 7.2, ensuring that beamforming operations can be examined precisely at the interface where data and precoding information cross between the digital baseband and the radio unit. By deploying the complete ORAN fronthaul chain on EURECOM's testbed and connecting it to the massive MIMO prototype provided by Andrew, the system provides an ideal platform to analyze how different beamforming strategies behave under the real constraints of an open, disaggregated RAN[17][5].

Given the context of downlink transmission only and the absence of instantaneous CSI feedback or RIC-assisted optimization loops, this thesis adopts a fixed beam ID beamforming approach. Instead of dynamically adapting beams based on real-time UE reports, the RU employs predefined beam patterns that remain stable over time. This choice is consistent with the architectural constraints identified earlier: it avoids the need for tight CSI-dependent updates and aligns with a Split 7.2 pipeline where only limited beam configuration information is transferred from the DU to the RU. Fixed beams also provide a reliable baseline for evaluating directivity, coverage shaping, and beam power distribution without the confounding effects of fast adaptation mechanisms[22].

On the PHY side, the work employs a straight-wire precoding scheme[40], in which the precoder applies minimal spatial processing and essentially forwards the data streams directly to the RU for mapping onto the predefined beam directions. This scheme minimizes fronthaul overhead because it avoids per-antenna, resource-element-level precoding and instead transmits only the necessary data symbols and beam identifiers. It also reflects the design trade-off discussed in the problematic: by pushing most spatial

processing toward the RU, the system reduces fronthaul bandwidth requirements at the cost of reduced adaptability, enabling a controlled investigation of Split 7.2 constraints and their impact on realistic downlink transmission.

In particular, the system setup enables the analysis of OOB emissions in a massive MIMO context[48], where nonlinear power amplifier behavior can generate unwanted spectral components that escape the intended beam shape. Although well-established techniques such as Digital Predistortion (DPD) and cavity filtering exist to mitigate these effects, residual distortions still pose challenges in meeting regulatory spectral masks. Studying these emissions in conjunction with fixed-beam transmission and Split 7.2 processing therefore provides valuable insight into how open, disaggregated, massive MIMO systems behave under real-world constraints, and how their performance can be optimized within the architectural bounds of ORAN.

1.5 Outline of the Dissertation

An outline of the dissertation, along with a brief summary of each chapter and its contributions, is provided below:

Chapter 2 - 5G New Radio using OpenAirInterface: Tools and Methodologies for the Fronthaul interface

This chapter introduces OAI, an open-source platform that implements the 5G NR protocol stack and provides a complete end-to-end testbed for open and virtualized RAN. Compliant with 3GPP standards and extended to support ORAN architecture, OAI enables the deployment and evaluation of disaggregated RAN components, including ORAN split 7.2 functionalities. OAI provides a flexible framework for real time experimentation and validation of advanced techniques such as MIMO and beamforming, enabling researchers to prototype and evaluate performance in standards compliant ORAN architectures.

Chapter 3 -Integration of the ORAN FH Interface

In this chapter, We will provide a comprehensive discussion of the ORAN FHI, covering its functional architecture, protocol stack, and operational planes, followed by a detailed description of its integration and implementation using the OAI framework.

Chapter 4 - Actual Out of Band emissions in Massive MIMO antennas

In this chapter, we study the behavior of OOB emissions in MIMO systems from both theoretical and experimental perspectives. We present and analyze measurement results obtained in an anechoic chamber, focusing on how amplifier nonlinearity and the presence or absence of DPD affect the directionality and strength of these emissions.

The work in this chapter has resulted in the following publication:

1. **Christ Rizk**, Eric Seguenot, Florian Kaltenberger, Dominique Nussbaum, Andrea Moro, Alessandro Sinicco, and Sagar Arora, “Actual Out-of-Band Emissions from Massive MIMO Antennas” in *European Wireless Conference (EWC 2025), Sophia Antipolis, France, 2025*.

The work in this chapter has resulted in the following contribution proposal:

2. **Conférence Européenne des administrations des Postes et Télécommunications (CEPT)/Electronic Communications Committee (ECC) contribution:** Submission of a draft report on the accepted paper “Actual Out-of-Band Emissions from Massive MIMO Antennas” for a possible contribution.

Chapter 5 - Actual Out of Band emissions in multi-user Massive MIMO antennas

In this chapter, we study the behavior of OOB emissions in MIMO systems from both theoretical and experimental perspectives. We present and analyze measurement results obtained in an anechoic chamber for multi-user MIMO system, focusing on how amplifier nonlinearity and the presence or absence of DPD affect the directionality and strength of these emissions, expanding on the previous chapter in terms of multi-beam and multi-UE serving MIMO Antenna systems.

Chapter 6 - Conclusions

Finally, this chapter concludes the dissertation and discusses possible future research problems.

Appendix A - Spatial Upper Bound of Radiated Power in Active Antenna Systems in A

In this appendix, we present a spatial upper bound that characterizes the radiated power behavior of Active Antenna Systems (AAS) employing beamforming and Massive MIMO. The bound establishes that the maximum emission is consistently achieved in the boresight direction and is independent of frequency, signal type, and beamforming configuration. This result provides a compact and robust framework for understanding AAS radiation characteristics. It is included as an appendix since it was developed after the submission of the original manuscript.

Appendix B - Extension to 6G and beyond bands in B

This appendix investigates the spatial behavior of OOB emissions in current and prospective massive MIMO antenna systems through conceptual measurement campaigns at 3.5 GHz and upper-6 GHz (6G). It demonstrates that OOB radiation is mainly driven by subarray-level antenna and RF front-end characteristics rather than full-array digital beamforming. The results support a scalable modeling approach for predicting OOB emissions in future 5G-Advanced and 6G deployments. It is included as an appendix since it was developed after the submission of the original manuscript.

Chapter 2

OpenAirInterface: tools and methodologies

2.1 Introduction

In this chapter, we introduce OAI, an open source software platform that implements the 5G NR protocol stack. OAI provides researchers, industry professionals, and students with an end-to-end implementation of 5G NR, making it one of the most widely used testbeds for experimentation and development in open and virtualized RAN[61]. OAI follows the specifications defined by 3GPP and has been extended to support the ORAN Alliance architecture, enabling deployment and testing of disaggregated and virtualized RAN components. This makes it a powerful tool for studying and testing the ORAN split 7.2 functionalities.

The rest of the chapter is organized as follows: Section 2.2 provides the necessary background to OAI, detailing its components, and operating modes. In Section 2.3, we focus on the ongoing implementation of BF in OAI and the different functionalities. In Section 2.4, we discuss Massive MIMO's ongoing implementation in OAI. Section 2.5 explores Beam Management (BM) in 5G NR, which encompasses procedures such as beam sweeping, measurement, and refinement that allow efficient utilization of directional beams in dynamic environments. Finally, Section 2.6 concludes the chapter.

2.2 Background

This section provides an overview of the various components of OAI software and its operating modes.

2.2.1 OpenAirInterface 5G NR Components

OpenAirInterface (OAI) is an open source project that provides a reference implementation of a 5G gNB, NR User Equipment (nrUE), and 5G Core (5GC) network compliant with 3GPP Release 15 and beyond. It operates on standard x86 computing platforms and uses COTS Software-Defined Radio (SDR) hardware, such as the Universal Software Radio Peripheral (USRP) and ORAN radios. This configuration enables users to deploy a complete, end-to-end 3GPP-compliant 5G network while ensuring interoperability with commercial equipment. By following this approach, OAI not only reduces implementation costs but also increases deployment flexibility. The detailed functionalities of each component in the OAI 5G stack are described below:

2.2.1.1 5G Core Network

The 5G Core Network (CN) network performs several essential functions, including authentication, mobility management, network slicing, and service orchestration.

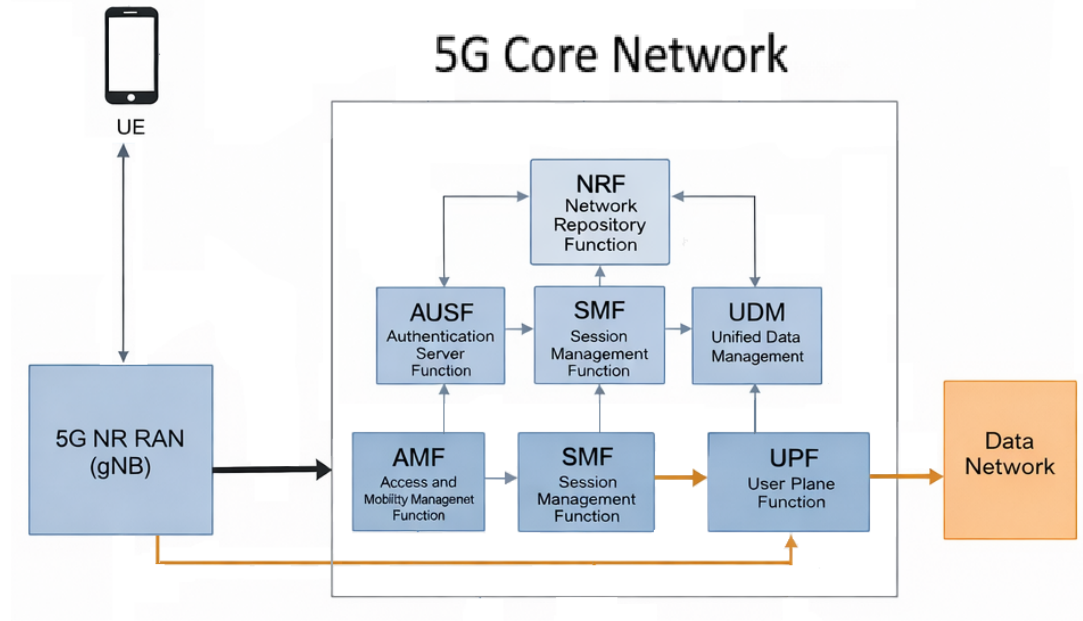


Figure 2.1: 5G Core network architecture

Each of these functions is managed by different network functions, as illustrated in Figure 2.1. The network functions supported by OAI 5GC are [62][63]:

- Access and Mobility management Function (AMF)
- User Plane Function (UPF)
- Network Repository Function (NRF)
- Session Management Function (SMF)
- Unified Data Management (UDM)
- Authentication Server Function (AUSF)

As seen in Figure 2.1, The AMF, SMF, and UPF connect the 5G Core to the RAN. They split the work into a C plane, where the AMF and SMF manage signaling, mobility, and sessions, and a U plane, where UPF handles data traffic and quality of service. In an ORAN system, this split allows more flexible deployment, faster data handling at the edge, and smooth operation across equipment from different vendors. Its tutorial can be found in [64].

2.2.1.2 5G NR Base-station

As shown in Figure 1.8 and referring back to 1.3.1.2, the ORAN architecture disaggregates the traditional gNB into CU, DU, and RU, with the CU further split into control and user plane entities for flexible deployment. RUs handle minimal physical layer tasks (FFT and cyclic prefix) for a cost and power efficient deployment, while DUs manage the remaining PHY, MAC, and RLC layers in close synchronization[17]. The CU-CP handles RRC signaling, and the CU-UP manages SDAP and PDCP for Qos, reliability, and security. The 7.2x functional split is widely adopted, balancing RUs simplicity with fronthaul performance.

The PHY layer of the 5G NR utilizes OFDM waveform for transmission over the air, structured in a frame format. This frame structure, as defined by NR, is presented in both time and frequency [31][65][66].

In the time domain, a 5G NR frame has a duration of 10 ms and is divided into 10 subframes[66], each lasting 1 ms. Furthermore, each subframe is divided into $N_{\text{Slot}}^{\text{Subframe}}$ slots, depending on the Subcarrier Spacing (SCS) used. Each slot consists of 14 OFDM symbols. The number of slots per subframe, $N_{\text{Slot}}^{\text{Subframe}}$, for different SCS values is detailed in Table 2.1.

SCS (KHz)	$N_{\text{Slot}}^{\text{Subframe}}$
15	1
30	2
60	4
120	8
240	16

Table 2.1: Number of slots per subframe for a given SCS

In the frequency domain, the 5G NR frame structure supports various bandwidth configurations based on SCS. Each configuration includes Resource Block (RB)s, where each RB consists of $N_{\text{SC}}^{\text{RB}} = 12$ subcarriers[66], referred to as Resource Element (RE)s. For a given frequency range and a specific SCS (in kHz), the number of resource blocks N_{RB} for bandwidths ranging from 5 MHz to 400 MHz is summarized in Table 2.2. An illustration of the frame structure of the time and frequency domains is shown in Figure 2.2.

FR	SCS (kHz)	5	10	15	20	25	30	40	50	60	70	80	90	100	200	400
1	15	25	52	79	106	133	160	216	270							
1	30	11	24	38	51	65	78	106	133	162	189	217	245	273		
1	60	11	18	24	31	38	51	65	79	93	107	121	135			
2	60														66	132
2	120														32	66

Table 2.2: Number of resource blocks (N_{RB}) for bandwidth configurations in 5G NR

However, SCS support in the OAI gNB is currently limited to 15 kHz and 30 kHz in FR1, and 120 kHz in FR2. The supported bandwidths in FR1 are 10, 20, 40, 50, 60, 80, and 100 MHz, while in FR2 the supported bandwidths are 50, 100, and 200 MHz.

A 5G NR frame in the time domain has a duration of 10 milliseconds (ms) and is divided into 10 subframes, each lasting 1 ms. Furthermore, each subframe is divided into $N_{\text{Slot}}^{\text{Subframe}}$ slots, depending on the SCS used. Each slot comprises 14 OFDM symbols.

In our setup we use numerology $\mu = 1$, which corresponds to a subcarrier spacing of 30 kHz. This gives a slot duration of 0.5 ms, with 2 slots per subframe and 20 slots per frame. Each slot has 14 OFDM symbols. For a 100 MHz channel in FR1, this allows up to 273 resource blocks, i.e., 3276 subcarriers (approximately 98 MHz occupied bandwidth) in the frequency domain[68][31][65].

2.2.1.3 User Equipment

The NR User Equipment in OAI referred to as nrUE is a real-time experimentation platform that implements the functionalities of a 3GPP-compliant UE as open-source software running on standard x86 general computing hardware. Using commercial software-defined radios, such as USRP, for transmission and reception. OAI nrUE works with OAI gNB and OAI 5GC, offering end-to-end functionality from RAN to the Core[69].

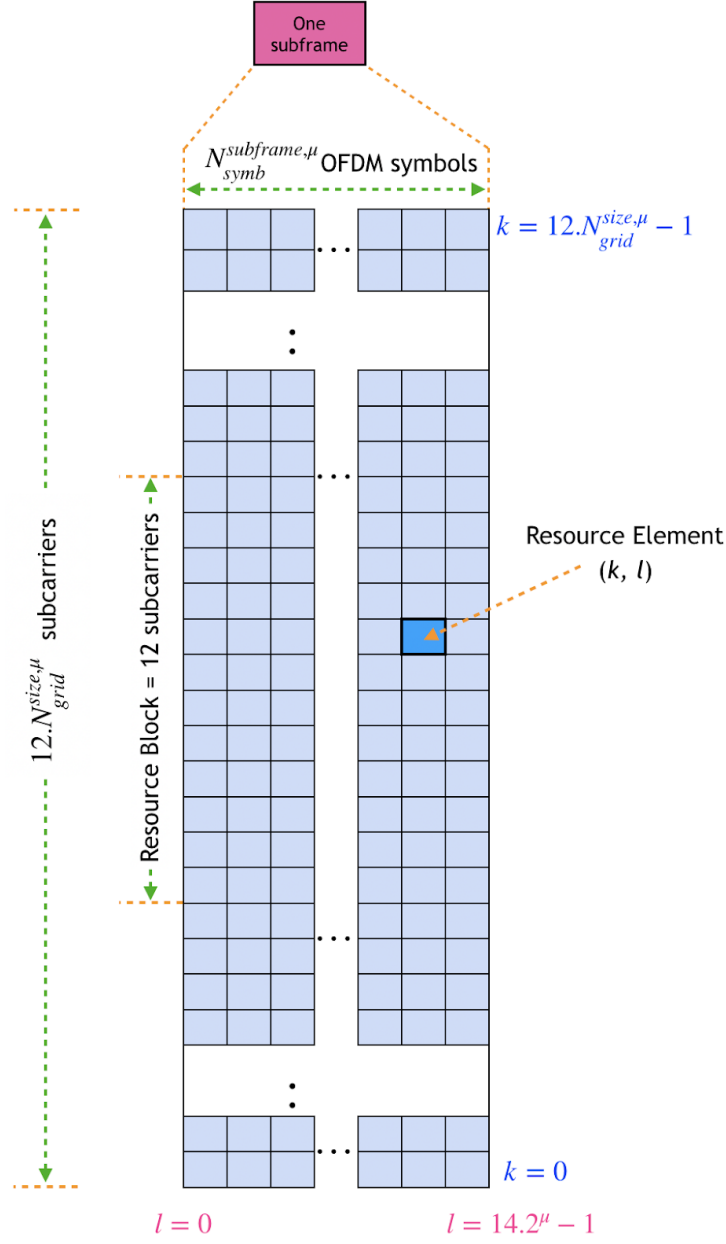


Figure 2.2: 5G OFDM resource grid [67].

2.2.2 Operating Modes

To support the development, debugging and real-time experimentation process, OAI supports several modes of operation. Specifically, the following two modes are widely used in Over The Air (OTA) experiments and are illustrated in Figure 2.3. The OAI softmodem can operate in different modes depending on the intended evaluation or testing scenario. Two important modes are:

1. Standalone (SA) Mode

In this mode, OAI runs an end-to-end 5G network using both the gNB and the core network (either OAI's own core or an external one). This allows a UE (OAI-based or commercial) to connect as in a real 5G Standalone deployment. It is mainly used for validating full-stack behavior, including

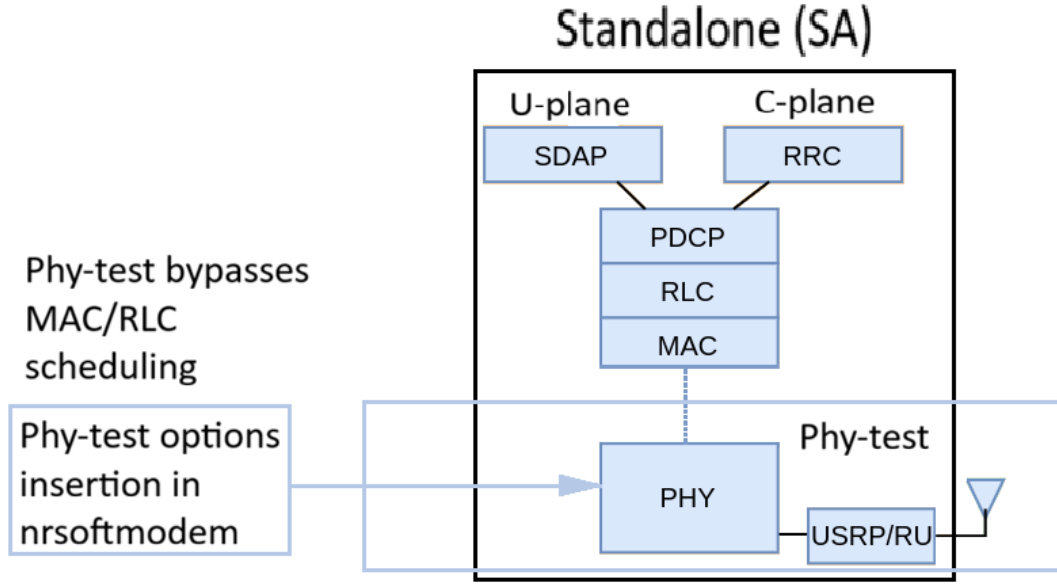


Figure 2.3: OAI Components and Operating modes.

signaling, data throughput, and end-to-end latency.

2. PHY-test Mode

This mode abstracts the higher layers and focuses on the evaluation of the physical layer. The gNB continuously schedules downlink and uplink transmissions even without a complete UE connection. In this setup, the MAC and RLC layers are bypassed, so no higher-layer signaling or bearer management is performed. The PHY-test mode is therefore useful for debugging, calibration, and performance testing of physical layer algorithms such as channel estimation, coding, and synchronization.

2.2.3 OAI's structure

The Openairinterface5g repository contains the software implementation of OAI's 4G/5G RAN and UE. It includes everything needed to build, configure, and run components such as eNB, gNB, and UE on various hardware platforms. The repository also provides tools for automation, testing, deployment, and documentation as shown in table 2.3. In the broader OAI ecosystem, this repository represents the RAN side, while the 4G/5G Core Network is maintained in a separate project[61].

We focus on the `openair1` module because it contains the PHY layer functionality required for beamforming and massive MIMO realization within the ORAN split 7.2 architecture. This layer implements essential signal processing operations such as OFDM modulation/demodulation, RE mapping and scheduling and hardware interfacing that are directly involved in supporting the functional split between the DU and RU.

2.3 Implementation of Beamforming in OAI

OAI supports the implementation of both analog and digital beamforming as defined in the 3GPP and ORAN specifications.

Folder	Description
openair1/	Contains the 4G/5G physical layer (PHY) implementations, including channel estimation, modulation/demodulation, OFDM processing, and interaction with radio hardware (e.g., USRP). This is the lowest layer of the RAN stack.
openair2/	Implements Layer 2 and Layer 3 RAN protocols: MAC, RLC, PDCP, RRC, scheduler, and associated control functions for LTE and NR. It forms the core of the gNB/eNB protocol stack.
openair3/	Implements NAS and EPC/5GC interfaces. Includes S1AP, NGAP, and NAS procedures allowing the gNB/eNB/UE to interact with the Core Network.
common/	Provides shared utilities, data structures, helpers, common libraries, logging mechanisms, and reusable code used across all OpenAirInterface modules.
executables/	Contains the main runnable binaries and entry-point programs such as <code>lte-softmodem</code> , <code>nr-softmodem</code> , <code>oai_ue</code> , and other test applications used to start the eNB, gNB, or UE.
charts/	Contains Kubernetes/Helm templates used to deploy OAI components (gNB, UE, etc.) as containerized services in cloud or clustered environments.
ci-scripts/	Includes scripts for the continuous-integration system, automating build, test, and validation processes.
cmake_targets/	Provides the main CMake-based build system, including build rules and target configurations for compiling OAI executables such as eNB, gNB, and UE.
targets/	Holds project-specific setups and configuration files, including hardware-dependent parameters, runtime configurations, and deployment examples for LTE and NR.
doc/	Contains documentation, tutorials, architecture explanations, and deployment guides for both users and developers.
Repository meta-data (README.md, LICENSE, etc.)	Top-level files providing project description, licensing information, contribution rules, and version history.

Table 2.3: Key folders in the openairinterface5g repository

2.3.1 Digital beamforming

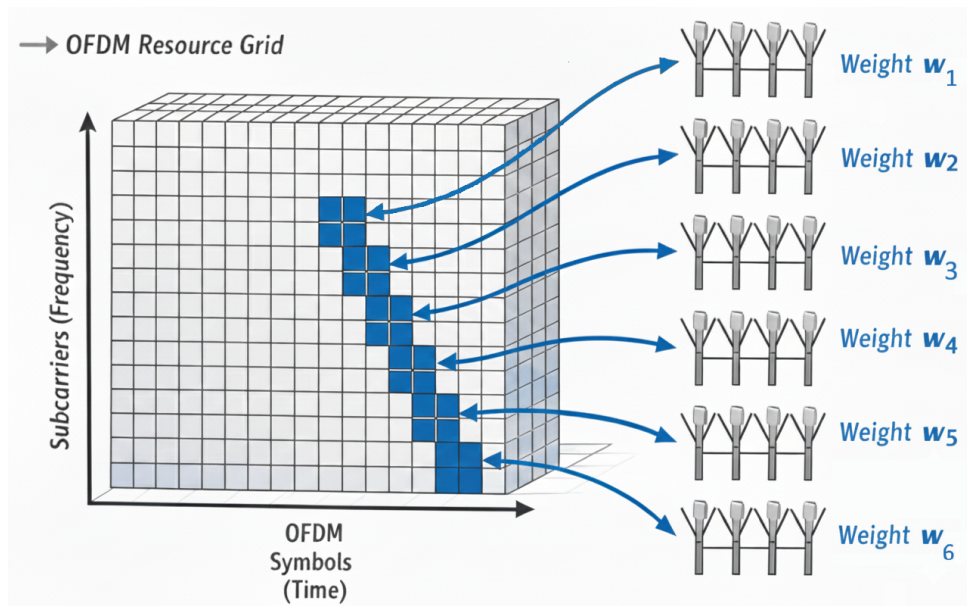


Figure 2.4: Weight multiplication per resource element in the 5G NR OFDM resource grid

As the RU supports digital beamforming, we will explain how digital beamforming is implemented in OAI as a data structure under test. OpenAirInterface implements digital beamforming through several functions that share the same mathematical operation:

$$x_{aa}(k) = \sum_{p=0}^{N_{\text{ports}}-1} w_{p,aa}(k) s_p(k) \quad (2.1)$$

The equation expresses digital transmit beamforming, where contributions from all antenna ports are weighted as seen in Figure 2.4 and combined to produce the signal transmitted on each physical antenna element:

- $x_{aa}(k)$: sample in complex-frequency domain transmitted on antenna element aa at subcarrier k .
- $s_p(k)$: complex-frequency-domain signal for port p at subcarrier k . This corresponds to the per-port OFDM-modulated data (Physical Downlink Shared Channel (PDSCH), Physical Downlink Control Channel (PDCCH), Physical Broadcast Channel (PBCH)/Synchronization Signal Block (SSB), Channel State Information Reference Signal (CSI-RS), etc.).
- $w_{p,aa}(k)$: complex beamforming weight applied for port p on antenna aa at subcarrier k . Controls the amplitude and phase applied to port p before contributing to antenna aa .
- N_{ports} : number of logical antenna ports (e.g., 1, 2, 4, 8 depending on LTE/NR configuration).

In OAI NR, digital beamforming maps the signals transmitted on the logical antenna ports to the physical antenna elements of the RU. The process is implemented in the function `nr_beam_precoding()` where its implementation is present in Appendix E, which applies the precomputed beamforming weights to each OFDM symbol and generates the final antenna-domain signals for transmission as seen in Figure 2.5.

In this code, the input signals are stored in `txdataF[p]`, where p indicates the logical antenna port. Each `txdataF[p]` is a frequency-domain buffer containing all OFDM symbols and subcarriers of a complete subframe or slot. This buffer holds every type of downlink transmission assigned to that port, modulated data, reference signals, and synchronization or broadcast channels, organized so that each array entry corresponds directly to one RE in the OFDM time-frequency grid. Before beamforming, these logical-port signals are identical for all physical antennas because beamforming has not yet been applied.

The beamformed output signals are written in `txdataF_BF[aa]`, where aa indexes the physical antenna element in the BS antenna array. After beamforming is completed in the frequency domain, each `txdataF_BF[aa]` buffer is transformed to the time domain through an IFFT and is ultimately transmitted by the RF chain of the corresponding physical antenna. Thus, `txdataF_BF` contains the final spatially shaped signals that physically leave the gNB.

The beamforming operation proceeds as follows. For every physical antenna aa , the function begins by clearing the output buffer `txdataF_BF[aa]` to zero. Then, for each logical antenna port p , the function multiplies the symbols in the frequency domain of that port by the corresponding complex beamforming weights `beam_weights[p][aa]`. These weights specify how much of the port p should be transmitted from antenna aa , and they apply a complex rotation (phase shift) and amplitude scaling to each resource element. After weighting each resource element of the port, the resulting vector of weighted samples is added to the physical antenna's output buffer. Since each physical antenna may transmit contributions from multiple ports, its final transmitted signal is the sum of all weighted port signals.

Operationally, this accumulation of elements is carried out by the helper function `multadd_cpx_vector()`. This function executes a complex multiply-add across a continuous block of resource elements: it takes one vector containing the port's frequency-domain symbols, multiplies it element-by-element with a second vector containing the corresponding beamforming weights, and adds the resulting

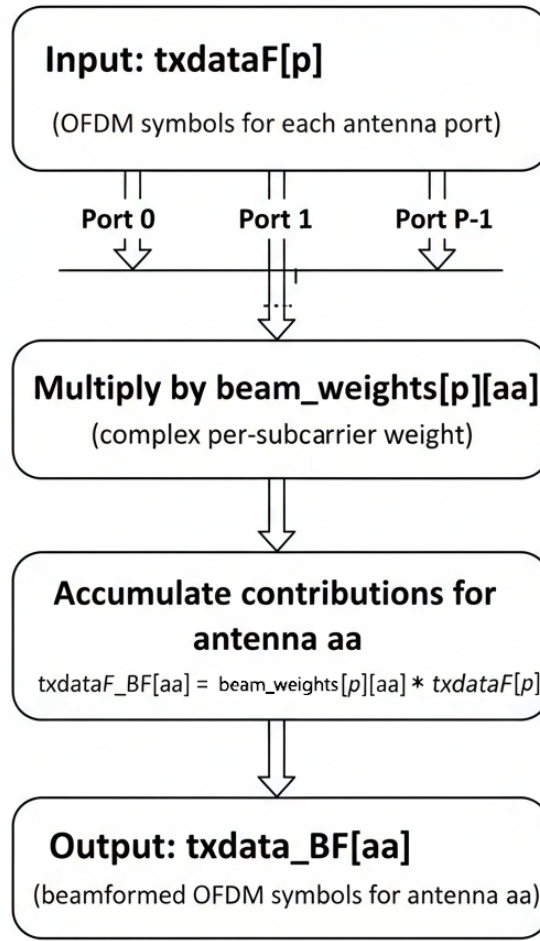


Figure 2.5: Block diagram of nr-beamforming in OAI.

vector into the existing contents of the target physical-antenna buffer. This process is repeated for every OFDM symbol in the subframe and for every active logical antenna port. In effect, the function performs the beamforming in resource-element-wise necessary to produce the fully precoded frequency-domain transmit grid used by the antenna array.

2.3.2 Fixed Beam ID beamforming

`nr-beamforming()` requires the beamforming weights to be accurate, coherent, and correctly matched to the estimated downlink channel conditions. The downstream beamforming stage strictly assumes that these weights are calculated correctly before being applied to OFDM symbols. A set of configuration-file parameters governs how beamforming is implemented and instructs the scheduler on the corresponding behavior:

- **set_beamforming:** Controls the mode used for selecting and signaling the beam index.
 - 0: default behavior,
 - 1: L1 maps the FAPI beam index to a preconfigured Hi-PHY DBM entry,
 - 2: L2 directly signals the beam index to Low-PHY (e.g. via 7.2x fronthaul).

- **beam_duration:** Defines the number of consecutive slots during which the scheduler must remain on the same beam (minimum and default value: 1).
- **beams_per_period:** Indicates the maximum number of beams that the RU can support in parallel within one beam-duration interval (default: 1).
- **beam_weights:** A vector listing the beam indices provided by OAI L1 to the RU. The number of entries must match the number of transmitted SSBs. When `enable_das = 1` is set in the L1 section, since each beam corresponds to a specific logical antenna port, the `beam_index_allocation()` function is simplified in the sense that the beam index corresponds to the antenna port index of the frequency domain buffers; so the field ‘beam_weights’ in ‘MACRLC’ section can be omitted in the configuration file. This concludes to say that the data at any downlink layer will be forwarded to the logical antenna port corresponding to the same id as the layer itself.

The fixed beam ID is sent over the C plane through `beam_weights` to the RU in which the corresponding beam weights will be applied in the RU.

2.4 Mu-MIMO in PHY-test

Although the system architecture natively supports digital beamforming and precoding, these functionalities are deliberately bypassed through configuration. In particular, by enabling direct antenna-stream forwarding (e.g., `enable_das = 1`) and omitting any precoding or C plane based beamforming, the digital beamforming stage is dependent only on the fixed beam ID procedure in the RU configuration file.

When additional downlink eAxC layers are enabled, as described previously, these layers likewise bypass any precoding schemes. Instead of being spatially combined or dynamically beamformed, each layer is transmitted independently and mapped to a fixed logical antenna port. The spatial characteristics of the transmission are therefore determined exclusively by the fixed Beam ID configuration (`force_bid`), which enforces a predefined beam selection at the RU.

This approach ensures deterministic and stable beam behavior across all enabled downlink layers, providing reproducible experimental conditions while intentionally foregoing the adaptive spatial gains and interference mitigation typically associated with digital beamforming and Massive MIMO techniques.

2.5 Beam management and Beamforming in 5G NR

5G cellular networks mark a significant technological shift by extending wireless communication into the mmWave frequency bands, ranging from 24.25 GHz to 52.6 GHz. These bands offer a tremendous amount of unused spectrum compared to the already crowded sub-6 GHz bands used in previous generations of mobile networks. Using this spectrum, 5G can provide extremely high data rates, support enhanced Mobile Broadband (eMBB) services, and enable a wide range of new applications, from immersive augmented and virtual reality experiences to ultra-high-definition streaming.

2.5.1 Challenges of wave Propagation

Although higher frequency bands offer clear benefits in terms of capacity and spectrum availability, they also introduce new challenges related to signal propagation and connection robustness. Unlike lower frequency signals, these bands suffer from several fundamental limitations:

- **Severe path loss:** Radio waves at mmWave frequencies attenuate much faster with distance compared to lower frequencies.

- **Blockage sensitivity:** Everyday objects such as buildings, vehicles, trees, and even the user's own hand can significantly block or attenuate the signal .
- **Limited diffraction and penetration:** Unlike lower frequency signals, mmWave signals do not bend effectively around obstacles and cannot penetrate walls or dense materials well.

These harsh propagation characteristics make the communication process highly directional and fragile in dynamic environments. To overcome these limitations, both BSs and UEs must employ BF, a technique that focuses the signal energy in narrow directional beams rather than radiating it in all directions. Although sub-6 GHz does not require extremely narrow beams such as mmWave, beamforming still improves throughput, reliability, and network efficiency at these lower frequencies. By concentrating energy along a specific spatial direction, beamforming compensates for high path loss and increases the effective communication range [70].

2.5.2 Role of Beam Management

BM refers to the set of procedures that ensure that BS and UE establish and maintain aligned beams throughout a communication session. In wireless communication systems, where directional transmission is essential, BM is critical to maintain reliable connectivity, especially in dynamic environments with mobility or blockage. It encompasses several key functions[70]:

1. Beam Sweeping

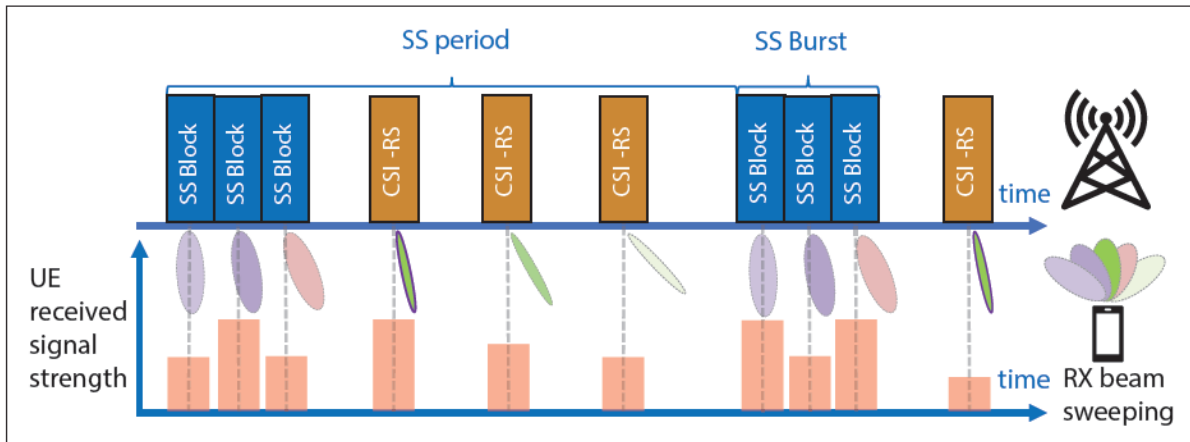


Figure 2.6: Beam management framework based on DL RSs in 5G NR.[70]

Beam sweeping is the process where the base station (gNB/BS) or UE sequentially transmits or receives reference signals using different beams from a pre-defined analog beam codebook as seen in Figure 2.6:

- On Downlink (DL), BS transmits SSB or CSI-RS through different beams.
- On the uplink (UL), UE can transmit SRS for BS to measure.
- This procedure enables both entities to explore spatial directions and identify potential pairs of beams Transmit-Receive (TX/RX).

2. Beam Measurement and Reporting

Once reference signals are received, the UE (or BS) measures their quality:

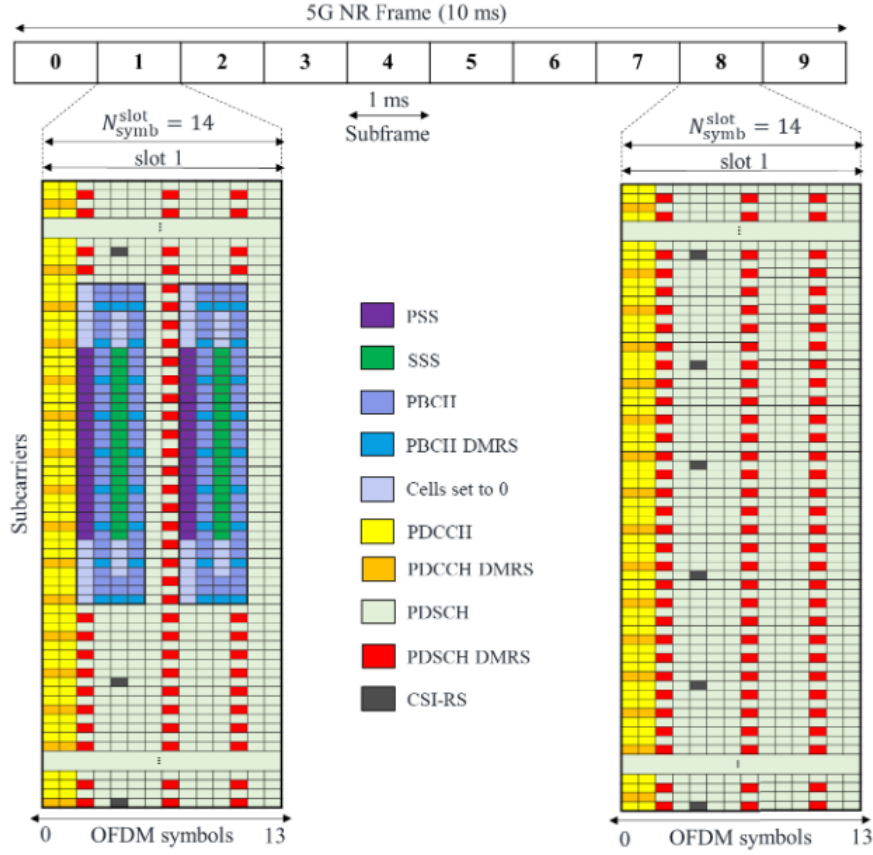


Figure 2.7: Channel allocation in the resource grid in 5G NR[66].

- The most common metric is the Reference Signal Received Power (RSRP) of SSB that is decoded and measured from the PBCH. Another metric also used in TDD is the CSI-RSs.
- Measurement of SINR is also supported.
- The UE reports the quality of the best beams back to the BS. Reporting is performed over Physical Uplink Control Channel (PUCCH) or Physical Uplink Shared Channel (PUSCH).

Depending on the capability, UE may report multiple candidate beams in each reporting instance.

3. Beam Determination and Indication

Based on the reported measurements, BS determines the appropriate beam pair for data transmission:

- For initial access, UE selects the SSB with the highest RSRP and transmits a Random Access Channel (RACH) preamble on the corresponding resource. This implicitly indicates that the selected beam is in the BS. The UE decodes the System Information Block (SIB) that is transmitted over PDSCH that includes Physical Random Access Channel (PRACH) resources.
- For connected mode beam tracking, the BS associates each reference signal with a Transmission Configuration Indicator (TCI) state. By signaling the TCI from the SIB, the BS informs the UE of the chosen TX/RX beam pair.
- A hierarchical signaling structure is used: long-term candidate beams are configured via RRC (Layer 3) messages, while fast beam switching is done using MAC control elements (Layer 2) or Downlink Control Information (DCI)(Layer 1).

The 5G NR BM framework coordinates sweeping, measurement, indication, and recovery to manage highly directional links. BM relies mainly on beam sweeping and reference signal measurements, where BS and UE scan beam directions from a codebook and use metrics like RSRP or SINR for beam selection. Although this approach supports early 5G, higher frequencies, multi-panel devices, and dynamic environments in 5.5G/6G demand more adaptive BM techniques for robust connectivity.

The beam management and beamforming procedures described above have only recently been partially implemented in the OAI 5G NR stack, and their integration is still evolving. In particular, the implementation depends on the availability of configurable beam codebooks at the PHY layer, accurate CSI-RS and SSB-based measurements at the UE, and dynamic beam indication mechanisms via MAC control elements or DCI. Compared to the conceptual framework outlined earlier, the practical realization in OAI currently imposes constraints on beam adaptivity, precoding flexibility, and real-time beam switching, making experimental setups primarily based on dynamic beam configurations. However, these recent developments represent a significant step towards enabling fully functional beam management in open-source 5G NR platforms.

2.6 Conclusion

In this chapter, we introduced OAI, including its components and operating modes that are beneficial to developers and researchers in prototyping. We discussed the fundamental elements of advanced wireless communication systems, focusing on Massive MIMO, beam management in 5G NR, and beamforming. Massive MIMO provides the physical foundation enabling high spatial multiplexing, while beamforming and beam management ensure efficient directional transmission and robust connectivity in dynamic scenarios.

Chapter 3

Experimental setup of the ORAN Fronthaul Interface

In the following chapter, we will provide a detailed discussion and explanation of the ORAN Fronthaul (FH) interface, including its functional architecture, protocol structure, and operational planes, and subsequently describe its integration using the OAI framework.

3.1 Background

In the ORAN architecture, the Lower Layer Split (LLS), commonly referred to as ORAN split 7.2 in Figure 3.1, defines a clear separation of baseband processing responsibilities between the DU and the RU. This split divides the physical layer so that the DU hosts the high-PHY functions, while the RU implements the low-PHY operations and the RF front-end. In modern cloud-native systems such as OAI, this split option is implemented for gNBs and plays a key role in enabling flexible, virtualized, and multi-vendor radio networks.

Within the DU, higher-layer processing begins with Layer 2 (L2) functions, which include the MAC layer, scheduling, and optional RRC tasks. These components manage UE context, allocate resources, and generate the control information required for radio operation, such as timing configuration, reference signal formats (e.g., SSB and PRACH), and beam-related parameters.

Following L2, the High-PHY layer performs computationally intensive baseband processing on transport blocks and physical channels. This includes channel coding and decoding, modulation and scrambling, resource element and resource-block mapping, and downlink layer mapping. Depending on the system configuration, digital precoding may be applied within the High-PHY stage of the DU or delegated to the Radio Unit (RU) for execution.

The outputs of these processing stages are then passed to the ORAN fronthaul library (`libxran`), which packages the control information and IQ sample data according to the ORAN fronthaul specification.

On the receiving side of the fronthaul, the RU interprets the DU's control instructions and performs the required low-PHY processing. This involves FFT and IFFT operations, resource mapping and demapping, cyclic-prefix handling, and IQ sample formatting for both uplink and downlink. In addition to these digital functions, the RU executes beamforming and the complete RF signal chain, including digital-to-analog conversion, up/down conversion, and filtering. The RU also manages timing alignment using PTP and dedicated fronthaul timing profiles so that its internal clocks remain in tight synchronization with the DU.

This architecture is supported by several logical planes within the fronthaul. The synchronization plane ensures that DU and RU share a common stable timing reference, typically through PTP, and the system may employ different topologies such as LLS-C1 through LLS-C4 depending on where the clock source resides and whether intermediate switches are used as explained in subsection 3.1.1.3. The control

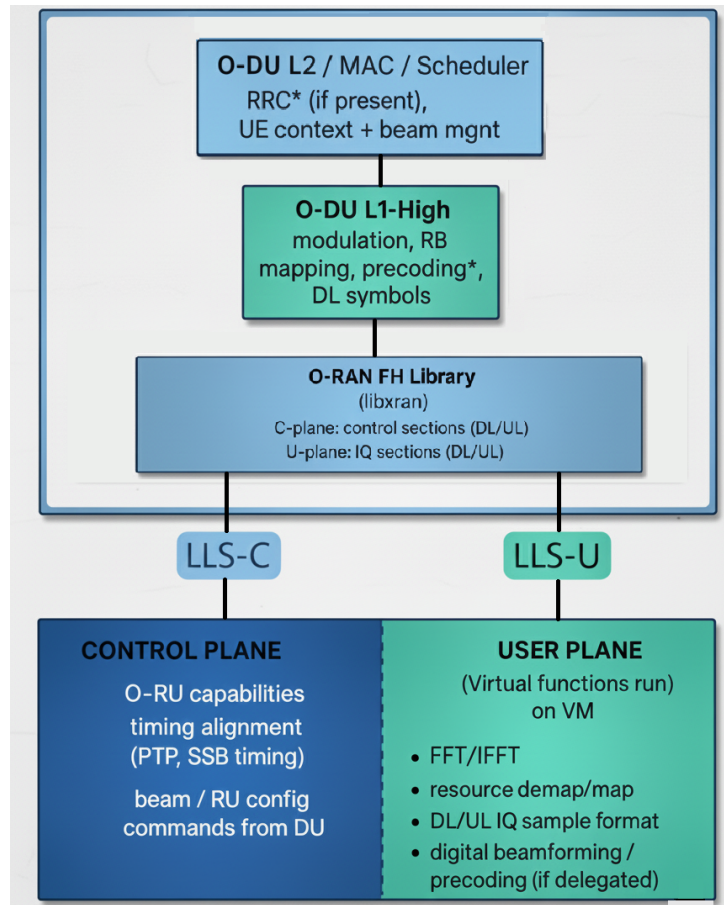


Figure 3.1: ORAN FH interface architecture.

plane carries the scheduling and beam-related commands necessary for the RU to act on incoming IQ samples, while the user plane transports uplink and downlink IQ data, which may be compressed to conserve fronthaul capacity. Alongside these, a management plane supervises configuration, software handling, fault monitoring, and other operational tasks of the O-RU, although this plane lies outside the scope of the present functional split description.

3.1.1 Transport layer

The O-RAN 7.2 Fronthaul Transport Layer provides packet-based connectivity between ORAN network functions (RU, DU, CU, and management entities) across fronthaul domains. It is responsible for ensuring that user, control, synchronization, and management plane traffic are delivered with the performance required by the selected functional split. The Transport Layer supports Ethernet/IP-based transmission, including Qos mechanisms, VLAN tagging and synchronization. It must meet defined latency, jitter, and reliability objectives to ensure proper operation of radio functions. The transport infrastructure may include fronthaul, midhaul, and backhaul segments and must enable interoperability among multi-vendor ORAN components [34].

3.1.1.1 Control plane

The Control Plane (C plane) in the ORAN architecture manages control signaling and coordination between functional entities. It is responsible for procedures such as connection setup and release, handover control,

resource allocation, and policy enforcement. Key components include the DU, responsible for lower-layer control and scheduling; and the RU, which performs radio transmission and receives control instructions as seen in Figure 1.8.

Each C plane packet consists of a common header followed by one or more section headers, which together define how and when user data is transmitted across the fronthaul link.

The C-plane common header includes several fields that describe the operational context of a transmission. These fields can be summarized as follows:

1. **Data Direction:** Indicates whether the data flow is uplink (RU to DU) or downlink (DU to RU).
2. **Payload Version:** Specifies the version of the C plane payload format being used.
3. **Filter Index:** Defines the digital filter configuration applied to the transmitted signal.
4. **Frame, Subframe, and Slot Identifiers:** Indicate the temporal position of the transmission within the radio frame structure.
5. **Start Symbol Identifier:** Marks the OFDM symbol within the slot where data transmission begins.
6. **Number of Sections:** Indicates how many independent transmission sections are contained within the message.
7. **Section Type:** Specifies the type of section being used, such as standard data or special control formats.
8. **Time Offset:** Provides fine-grained synchronization between the DU and RU for precise transmission timing.
9. **Frame Structure:** Describes the frame format and numerology applied to the transmission.
10. **Cyclic Prefix Length:** Defines the cyclic prefix configuration for the transmitted waveform.
11. **Reserved Field:** Reserved for future extensions or additional functionalities.
12. **Beam-Related Parameters:** Each section may include a beam ID or beamforming weight ID, allowing the O-RU to apply specific beamforming configurations for transmission or reception. These parameters support advanced features such as massive MIMO and dynamic beam management.

3.1.1.2 User plane

The User Plane (U plane) carries the actual payload. In the disaggregated ORAN architecture, U plane functions are distributed across the RU, DU. The U plane packet contains essential fields such as data direction, frame and slot identifiers, section and beam IDs, and the IQ data payload. These ensure precise timing, beamforming, and synchronization of user data transmission. The payload of the U plane is composed of IQ samples, which are digitized representations of the radio signals. The IQ samples represent the user's data at the physical layer. When transmitting user traffic, the DU performs baseband processing, including modulation, channel coding, and produces the IQ samples, which are then sent over the U plane to the RU. The RU, in turn, converts these IQ samples into RF signals for OTA transmission. In this way, the U plane acts as a conduit for user data in the form of IQ samples, ensuring that the information is delivered accurately, synchronized, and mapped to the appropriate beams and antenna ports as directed by the C Plane.

3.1.1.3 Synchronization plane

Various methods can be used to distribute time and frequency synchronization to DUs and RUs. In practical deployments, synchronization accuracy is influenced more by implementation aspects, such as

how close timestamping is performed to the network interfaces and the number of intermediate hops than by the specific synchronization technology used.

The following synchronization types are supported:

- **Frequency synchronization:** Ensures that all device clocks operate at the same rate supported by Sync-E.
- **Phase synchronization:** Aligns the phase of clocks across all network elements supported by Sync-E.
- **Time synchronization:** Provides a common absolute time reference to all devices supported by PTP.

Together, these synchronization methods form a network synchronization profile. Such a profile specifies the necessary features and configuration choices for switches and end stations, as well as the compliance requirements for both network equipment and user applications.

Clock Model and Synchronization Topology ORAN defines several synchronization topologies to address a range of deployment requirements. Four topology configurations are considered compliant for providing synchronization to the RU. For reference, each configuration is assigned a specific label:

- **Configuration LLS-C1:** In this topology, the DU is included in the synchronization chain leading to the RU. Timing information is sent from the DU to the RU through a direct connection between the DU site and the RU site.
- **Configuration LLS-C2:** As in LLS-C1, the DU remains part of the synchronization path. In this case, the timing signal is delivered across a fronthaul network that may contain one or more Ethernet switches between DU and RU sites.
- **Configuration LLS-C3:** This topology excludes the DU from the synchronization chain. Instead, timing is supplied directly from the Primary Reference Time Clock (PRTC)/Telecom Grand Master (T-GM) to the RU, typically over links connecting central or aggregation sites to RU sites.
- **Configuration LLS-C4:** In this configuration, the RU obtains its synchronization reference independently of the transport network, typically through a local Global Navigation Satellite System (GNSS) receiver.

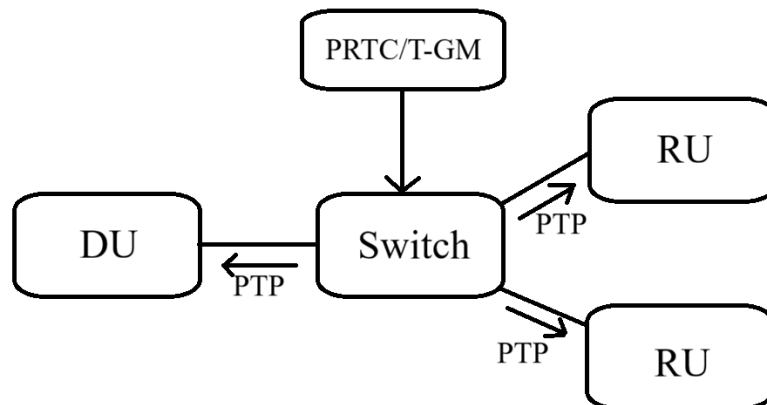


Figure 3.2: LSS-C3 synchronization configuration.

In our setup, we used Configuration LLS-C3. The switch is responsible for delivering time synchronization between the RU and DU as seen in Figure 3.2.

3.2 Use of the ORAN FH Interface from OSC in OAI

The ORAN FH interface is specified and implemented by the ORAN Software Community (OSC) as an open and standardized interface between radio access network components. OAI does not implement the ORAN FH interface itself; instead, it interfaces with the OSC-provided implementation to enable ORAN compliant fronthaul communication. In this setup, the OSC software handles the FH protocol stack, while OAI focuses on 5G NR baseband processing and interacts with the fronthaul interface to connect the RU and DU[71].

The ORAN FH interface enables communication between the DU and the RU and comprises multiple hardware and software components as illustrated in Figure 3.3:

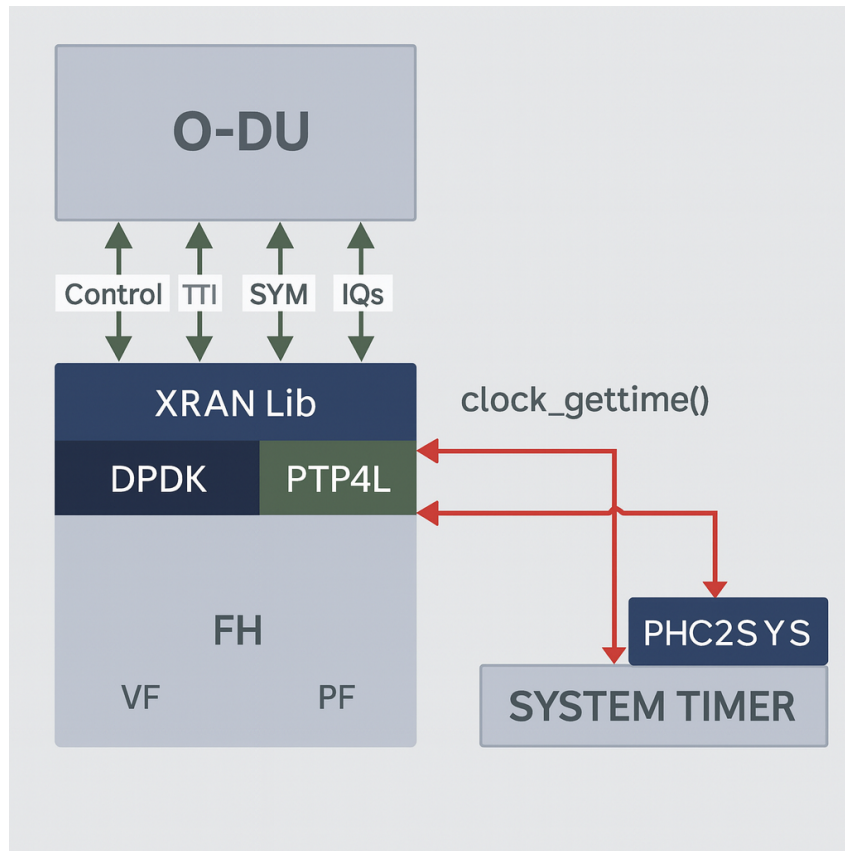


Figure 3.3: Implementation of the ORAN FH interface by OSC

From a hardware perspective, the system employs one networking port used for receiving PTP synchronization. The system timer provides a global time reference to the gNB application, ensuring precise coordination of radio operations.

From a software perspective, several key components contribute to the overall functionality:

- **Synchronization:** The Linux PTP framework ensures that the system timer is synchronized to the Grand Master (GM) clock time. Specifically, PTP4L synchronizes the oscillator on the Network Intelligent Controller (NIC) with the PTP GM, while PHC2SYS aligns the system timer with the NIC oscillator.
- **Networking:** The Data Plane Development Kit (DPDK) provides a high-performance interface to the Ethernet ports, enabling low-latency packet transmission of C/U-plane packets based on VLAN tag.

- **ORAN FH library:** The ORAN library, built on top of DPDK, implements the C plane and U plane functionalities as defined by the ORAN Fronthaul specification. Its main purpose is to encapsulate all ORAN packet processing logic and isolate it from the 5G NR Layer 1 (L1) pipeline, ensuring modularity and clear separation of responsibilities.
- **Physical Layer Interface:** The 5G NR reference PHY layer utilizes the ORAN library to establish communication with the RU. The interface between the PHY and the ORAN library manages Transmission Time Interval (TTI) events, symbol timing, C plane control messages, and IQ sample data exchange.

3.3 Experimental setup description

This section focuses on end to end testing within the ORAN framework, conducted through a collaboration between EURECOM and the LEAT laboratory. The objective is to interconnect both laboratories via high-speed fiber optics, enabling the deployment of a distributed 5G testbed that replicates a realistic ORAN deployment scenario. The ORAN split 7.2 FH was based on the ORAN tutorial in [60].

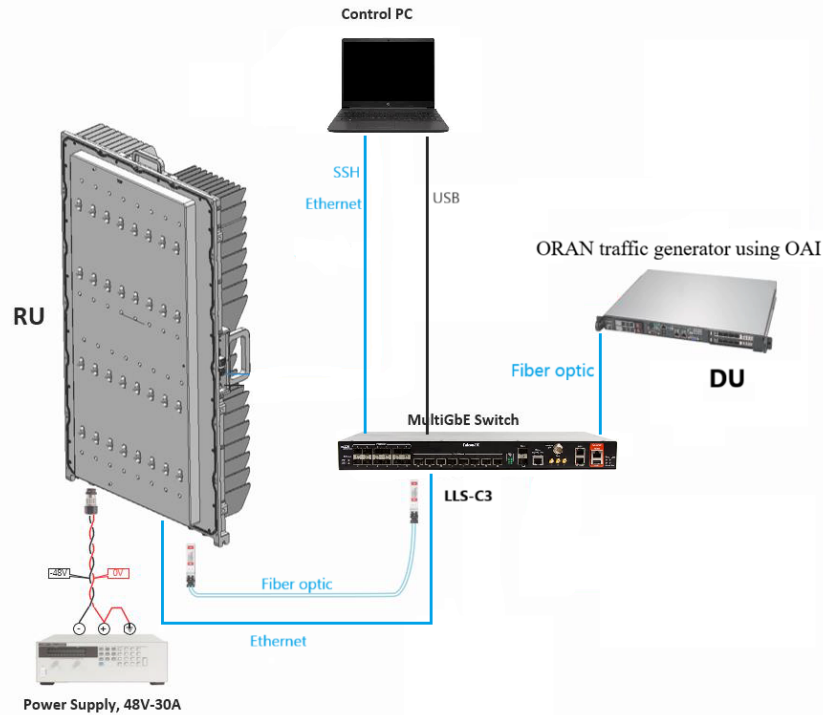


Figure 3.4: Platform architecture

In the setup in Figure 3.4, the DU is hosted in the server room at EURECOM, acting as the computational node responsible for digital baseband and higher physical layer functions. The PTP synchronization switch, essential for ensuring accurate time and frequency alignment across the ORAN components is placed in the LEAT Forum. The RU is located in the anechoic chamber at LEAT, a controlled environment designed to eliminate external interference and multipath effects. This setup provides optimal conditions for evaluating beamforming and radio performance.

The use of optical fiber interconnection ensures real-time communication, minimal latency and high reliability, which are necessary to validate the 7.2 split functional architecture of ORAN between two

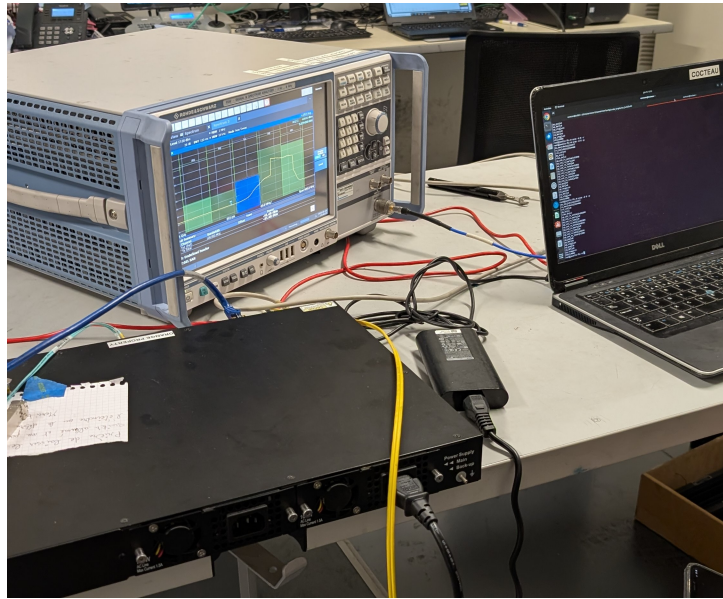


Figure 3.5: Representation of the platform architecture

geographically separated locations, and for that it was beneficial for our setup, since the DU was placed in EURECOM and we had to use a long range optical fiber to reach the anechoic chamber.

3.3.1 Fixed beam ID beamforming with C plane disactivation

```
# [ORAN CONFIGURATION FILE]
oran =
{
    du_mac_addr    = "00:11:22:33:44:77";
    ru_mac_addr    = "00:11:22:33:44:88";
    up_vtag        = {pcp = "0"; dei = "0"; tci = "14"};
    cp_vtag        = {pcp = "0"; dei = "0"; tci = "13"};
    mtu_size       = "9000";
    dl_endpoints = (
        { eAxC = "0"; enabled = "1"; force_bid = "0x5"; cmp_scaling = "0xf";
        cmp_static = "1"; cmp_iq_width="9"; cmp_method = "1";},
        { eAxC = "1"; enabled = "0"; force_bid = "0x8"; cmp_scaling = "0xf";
        cmp_static = "1"; cmp_iq_width="9"; cmp_method = "1";},
        { eAxC = "2"; enabled = "0"; force_bid = "0"; cmp_scaling = "0xf";
        cmp_static = "1"; cmp_iq_width="0"; cmp_method = "0";},
        { eAxC = "3"; enabled = "0"; force_bid = "0"; cmp_scaling = "0xf";
        cmp_static = "1"; cmp_iq_width="0"; cmp_method = "0";},
    );
};
```

```
# [ORAN CONFIGURATION FILE]
oran =
{
    du_mac_addr    = "00:11:22:33:44:77";
    ru_mac_addr    = "00:11:22:33:44:88";
    up_vtag        = {pcp = "0"; dei = "0"; tci = "14"};
    cp_vtag        = {pcp = "0"; dei = "0"; tci = "13"};
    mtu_size       = "9000";
```

```

dl_endpoints = (
    { eAxC = "0"; enabled = "1"; force_bid = "0x5"; cmp_scaling = "0x7";
    cmp_static = "1"; cmp_iq_width="9"; cmp_method = "1";},
    { eAxC = "1"; enabled = "1"; force_bid = "0x8"; cmp_scaling = "0x7";
    cmp_static = "1"; cmp_iq_width="9"; cmp_method = "1";},
    { eAxC = "2"; enabled = "0"; force_bid = "0"; cmp_scaling = "0xf";
    cmp_static = "1"; cmp_iq_width="0"; cmp_method = "0";},
    { eAxC = "3"; enabled = "0"; force_bid = "0"; cmp_scaling = "0xf";
    cmp_static = "1"; cmp_iq_width="0"; cmp_method = "0";},
);
};

```

In the ORAN architecture, the DL configuration defines the parameters for transmitting user data from the RU to the UE via multiple eAxCs. The configuration is specified per endpoint eAxC, each corresponding to a physical antenna port or logical DL stream. The DL configuration in the RU configuration file presented in the previous script can be analyzed in terms of three main components: endpoint activation, fixed beam IDs, and IQ compression parameters.

Each DL endpoint is represented by an eAxC identifier and a set of parameters:

- **eAxC:** Logical identifier for each downlink antenna carrier.
- **enabled:** Binary flag indicating whether this endpoint is active (1) or inactive (0).

For example, in the configuration:

```

{ eAxC = "0"; enabled = "1"; ... }
{ eAxC = "1"; enabled = "1"; ... }
{ eAxC = "2"; enabled = "0"; ... }

```

only eAxC = 0 and eAxC = 1 are active, meaning that the DU schedules downlink data exclusively on these two streams. The remaining endpoints (eAxC = 2–15) are inactive and do not carry downlink traffic.

The `force_bid` parameter specifies the fixed beam ID used by the RU for each endpoint. This is a key parameter in scenarios where deterministic beam selection is required, such as initial access testing, controlled deployments, or static beamforming configurations.

- **force_bid = 0:** Beam selection is dynamic, based on measurements and scheduling decisions.
- **force_bid ≠ 0:** The RU transmits exclusively on the specified beam ID.

In the current configuration:

```

eAxC 0:  force_bid = 0x5
eAxC 1:  force_bid = 0x8

```

This means that DL transmissions on eAxC 0 are constrained to beam 0x5 in hexadecimal from the DBT table, while eAxC 1 uses beam 0x8. All other endpoints have `force_bid = 0`, but are disabled in this configuration.

The DL data is transported over the fronthaul interface, and the configuration includes parameters for IQ data compression:

- **cmp_scaling:** Scaling factor for IQ compression. Here, 0xf corresponds to maximum scaling.

- **cmp_static:** Compression mode. 1 indicates static compression, i.e., fixed quantization without dynamic adaptation.
- **cmp_iq_width:** Number of bits per IQ sample. In this configuration, 9 bits are used for each I and Q sample.
- **cmp_method:** Compression algorithm selection. A value of 1 typically refers to a LUT-based compression scheme.

IQ compression settings directly affect the fronthaul bandwidth and signal fidelity. Static compression with 9-bit IQ width ensures a predictable fronthaul data load and avoids dynamic fluctuations in transport requirements.

In the ORAN architecture, the C plane is responsible for conveying beamforming-related information, including the Beam ID, from the DU to the RU. When dynamic beamforming is enabled, the Beam ID may change over time as a function of scheduling decisions, channel measurements, or beam management procedures.

For the purpose of this work, dynamic beam signaling over the C-plane was intentionally disabled, and fixed Beam IDs were configured using the `force_bid` parameter. This design choice ensures that the RU does not modify the selected beam during operation, thus eliminating any potential beam switching caused by control-plane updates.

By enforcing a fixed Beam ID at the RU, the downlink transmission remains confined to a predetermined spatial direction throughout all experiments. This approach guarantees deterministic beam behavior, improves experimental repeatability, and simplifies performance analysis by eliminating beam dynamics as a confounding factor.

3.3.2 Challenges and difficulties

The integration of the measurement setup within an anechoic chamber involves several practical challenges, which are summarized as follows:

1. **Physical shared space and equipment constraints:** The anechoic chamber and certain measurement equipment are shared resources. As a result, careful coordination, planning, and scheduling were required to manage shared space and equipment availability, making efficient collaboration and time management a significant challenge.
2. **Installation and Cable Management:** Setting up equipment inside an anechoic chamber requires careful planning of RF and fiber-optic cable routing, particularly when connecting to the main server room. Proper cable lengths, bending radii, and terminations are essential to prevent signal loss and interference.
3. **Fiber-Optic Connectivity:** Connecting the setup to the main server room via fiber optics introduces challenges related to attenuation, dispersion, and latency. Appropriate selection of fiber cables, transceivers, and termination techniques is necessary to ensure signal integrity.
4. **Power and Data Management:** Distributing power and managing data communication across components located in different environments (anechoic chamber and server room) require careful planning to ensure reliable operation, sufficient power availability, and proper data synchronization.
5. **Testing and Validation:** Thorough testing and validation are essential to confirm system performance under various conditions, including different beamforming configurations, MIMO scenarios, and fiber-optic link behaviors.
6. **Anechoic Chamber Availability:** The anechoic chamber is a shared facility, requiring strict adherence to allocated time slots. Efficient time management and careful scheduling were therefore necessary to complete all measurements within the assigned time windows.

7. **Safety and Compliance:** Compliance with safety regulations and standards is essential in environments involving RF equipment and fiber-optic connections, in order to protect personnel and equipment from hazards such as electromagnetic radiation, high-voltage power supplies, and physical installation risks.

The main goal of this project is to implement and evaluate a disaggregated 5G Open RAN architecture that separates the DU and the RU through the ORAN split 7.2 functional interface. This setup enables physical layer processing to be divided between two distinct sites, allowing realistic testing of latency, synchronization, and interoperability constraints in a geographically distributed environment.

A tutorial for the integration is explained in detail in the Appendix D based on the OAI tutorial repository [60]

3.4 Conclusion

This chapter briefly explained the different hardware and software components of the ORAN FH interface as well as its functionalities. It detailed the integration and configuration of the fronthaul interface according to the ORAN split 7.2 standard, highlighting the importance of precise PTP synchronization and optimized management of hardware and software resources. The compilation and deployment of the ORAN FH library, along with the construction of the OAI gNB, enabled reliable communication between the DU and RU. The transmission, synchronization, and IQ data flow tests confirmed the system's stability and compliance with the standard. These steps are crucial to ensure effective interoperability within an Open RAN architecture.

Chapter 4

Actual Out Of Band emissions in Massive MIMO antennas

This chapter looks at OOB emissions in 5G Massive MIMO systems. We will describe the theoretical modeling of the OOB emissions and of the full transmission system, we will also describe the measurement setup, and we discuss briefly the obtained measurements.

Larson and Van Der Perre's study[48] says that in single user case the OOB emissions are radiated in the same direction of propagation as the useful signal, "the spatial directivity of the OOB distortion will in the far-field have the same radiation pattern, regardless of the actual implementation of the beamforming", and for that we consider in this chapter a single user case to validate this theory experimentally.

4.1 Introduction

Recent studies, such as those of Larsson and Van der Perre [48], show that the distortion caused by nonlinearities is not completely random. Instead, it is correlated with the actual transmitted signal, meaning that in some cases it can be beamformed just like the useful signal. When PAs are working in the nonlinear regime and DPD is not activated, unwanted OOB components can accumulate across the antenna array and form strong beams in the same direction as the desired signal. This can result in high interference power where it matters most towards users or nearby systems.

However, when DPD is applied, the nonlinear distortions are greatly reduced. In this case, while the useful in-band signals remain focused through beamforming, the remaining OOB emissions become decorrelated, and they do not align across antennas and are radiated incoherently in different directions. This makes the antennas behave like independent noise sources, which spreads the OOB emissions more evenly and reduces the chance of high interference in any specific direction [54]. This effect, sometimes called spatial averaging, is a useful feature of Massive MIMO systems.

The goal of this chapter is to study how OOB emissions behave in Massive MIMO systems, both from a theoretical and experimental point of view. In particular, we look at how the nonlinearity of the amplifier and the use and absence of DPD affect the direction and strength of these emissions.

The chapter is organized as follows:

Section 4.2 introduces the theoretical model we used to understand the spatial behavior of the OOB emissions. It also describes the experimental setup, which includes a 32-transmit/32-receive (32TX/32RX) n78-band Massive MIMO antenna array provided by Andrew-Amphenol. The measurements were performed in an anechoic chamber to ensure a controlled testing environment.

Section 4.4 presents the results of our measurements. We compare OOB emissions in different operating conditions, both with and without DPD and analyze their spatial properties. We evaluate their

impact on performance metrics such as ACLR. The results support the idea that DPD and spatial averaging significantly reduce OOB interference.

Section 4.5 summarizes the key takeaways from the study. We discuss how our findings can help improve system design, support regulatory compliance, and guide future standardization and deployment of Massive MIMO networks.

By combining theoretical insights with real-world experiments, this chapter contributes to a better understanding of how unwanted emissions behave in large antenna systems and how their impact can be reduced through smart design and signal processing techniques.

4.2 Theoretical framework

The theoretical framework and modeling are based on the theoretical aspects used in Larson and Van der Perre's paper[48] and taking into consideration the basband transmission characteristics of the RU.

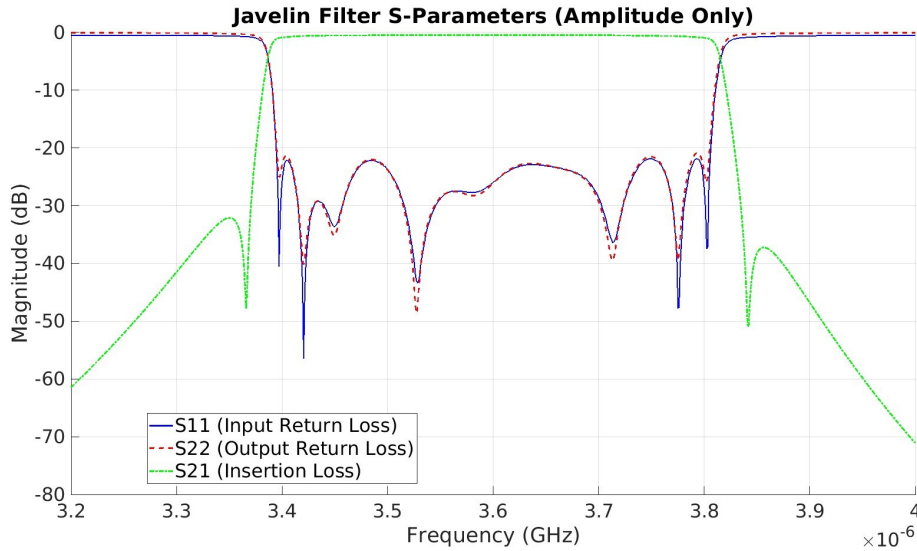


Figure 4.1: Antenna array internal cavity filter S-parameters.

OOB emissions are defined as any unwanted signal components that fall just outside the intended transmission channel band. Our analysis focuses specifically on emissions within the range of [3.4–3.8]GHz, since emissions beyond this range are effectively attenuated and suppressed by the cavity filter and can be considered negligible, as illustrated in Figure 4.1.

We used the commercial band of [3.7–3.8]GHz as the useful channel band, so the most important OOB emissions with higher power levels to study are in the adjacent bands defined by [3.6–3.7]GHz and [3.8–3.9]GHz. The cavity filter is part of the transmission chain in the RU, it filters out the unwanted emissions before directing the signal towards the subarray level.

4.2.1 System Description

Throughout this thesis, we adopt the following notation conventions. Discrete-time indices and samples are denoted by n . Angular frequencies for UE_i are denoted by ω_i , and carrier frequencies by f_i . The beamforming phases applied at antenna m for user l are denoted by ϕ_{ml} . We denote by k the polynomial order with K being the PA's maximum order. These conventions are used consistently throughout the remainder of the manuscript.

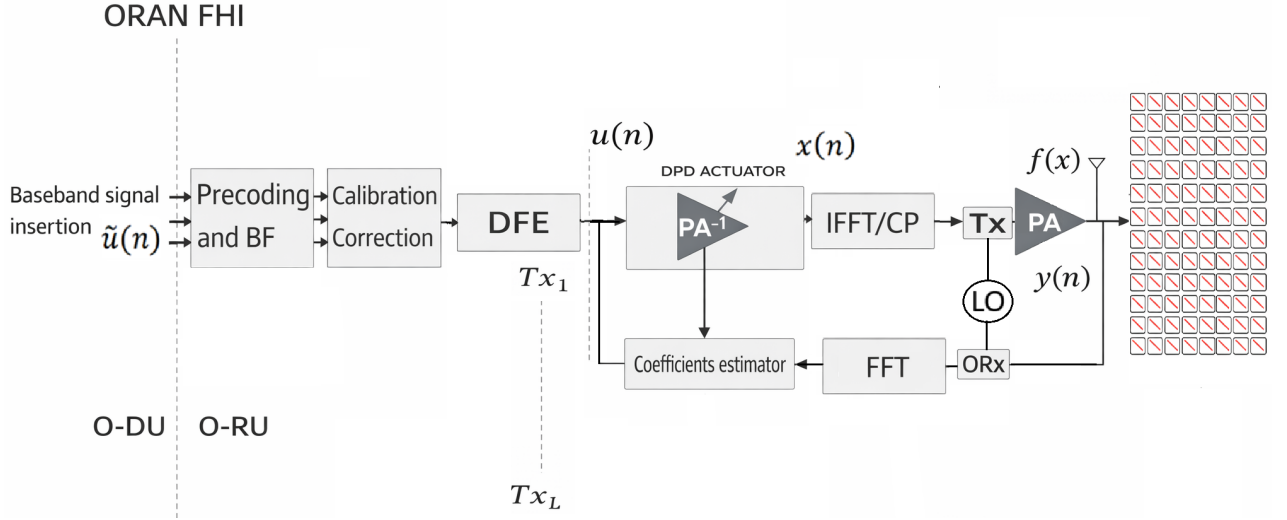


Figure 4.2: Representation of the RF transmission chain, including the fronthaul interface.

The system described in Figure 4.2 is a combination of operations in the frequency and time domain that are delimited by the IFFT/CP block. To simplify calculations and theoretical modeling, all annotated entities are taken in the discrete time domain since the transition from the frequency domain to the time domain does not affect the theoretical analysis of the RF transmission chain.

To support the preceding discussion, Figure 4.2 depicts the FH interface between the O-DU and O-RU as described in 4.3.3, followed by the RF transmission chains in the Spear Andrew antenna array. The transmission begins with a multitone OFDM baseband signal, denoted $\tilde{u}(n)$, which contains the useful information symbols. This baseband signal is processed by the ADRV9029 transceiver, where it is modulated to the required carrier frequency ω_0 . The corresponding passband representation is given by:

$$u(n) = \Re \left\{ \tilde{u}(n) e^{j\omega_0 n} \right\}. \quad (4.1)$$

The modulated signal is then amplified by a PA before being radiated. However, due to the nonlinear characteristics of practical PAs, the output signal is not an exact scaled version of the input but instead contains additional distortion terms, which are responsible for in-band distortion and OOB emissions. To mitigate these effects, a DPD block is employed, but for theoretical modeling, the residual nonlinearities remain relevant. We ignore the DPD in modeling, so we have $x(n) = u(n)$.

4.2.2 OOB Emissions Modeling with Generalized 3rd-Order Volterra PA

Let $y(n)$ denote the passband signal at the PA's output. Its baseband equivalent $\tilde{y}(n)$ can be expressed using a generalized 3rd-order, memoryless Volterra model:

$$\tilde{y}(n) = a_1 \tilde{x}(n) + a_3 |\tilde{x}(n)|^2 \tilde{x}(n) \quad (4.2)$$

where $a_1 \in \mathbb{C}$ is the linear gain coefficient, and $a_3 \in \mathbb{C}$ is the dominant third order nonlinear coefficient. The cubic term $|\tilde{x}(n)|^2 \tilde{x}(n)$ captures amplitude and phase dependent nonlinear distortion. Second order and higher-order terms beyond third order are assumed to be suppressed by the cavity filter.

Additionally, each subarray is affected by additive white Gaussian noise caused by the ADC/DAC quantization noise [72], denoted $w_m(n)$, which is assumed to be zero-mean with variance σ^2 and uncorrelated between subarrays. Taking into account the generalized PA nonlinearity and the presence of

noise, the transmitted signal from the m -th subarray can be approximated as:

$$\tilde{Y}_m(n) = a_1 \tilde{x}(n) + a_3 |\tilde{x}(n)|^2 \tilde{x}(n) + w_m(n) \quad (4.3)$$

The signal $\tilde{u}(n)$ is subject to digital beamforming at the beginning of the transmission process by phase shifting it by a factor of $e^{j\phi_m}$, where ϕ_m denotes the phase shift applied at antenna port m to steer the signal in the desired spatial direction. Therefore, in the baseband domain:

$$\tilde{x}(n) = \tilde{u}(n) e^{j\phi_m} \quad (4.4)$$

and

$$\tilde{Z}_m(n) = a_1 \tilde{u}(n) e^{j\phi_m} + a_3 |\tilde{u}(n)|^2 \tilde{u}(n) e^{j\phi_m} + w_m(n) \quad (4.5)$$

where $\tilde{Z}_m(n)$ is the signal at antenna port m . The antenna array consists of 192 radiating elements, arranged in 8 columns, 12 rows, and 2 polarizations. Each RF chain drives a subarray composed of 6 horizontally aligned, single-polarized elements. In our experiments, only the -45° polarization was utilized, resulting in an effective number of subarrays of $M = 16$ as illustrated in Figure 4.2.

In the far field, the signal and distortion components from each subarray combine coherently because of the constructive interference of the antenna array, while the noise terms add up incoherently. The total received signal is proportional to:

$$\begin{aligned} \tilde{Z}(n) &= \sum_{m=1}^M \tilde{Z}_m(n) \\ &= \sum_{m=1}^M \left(a_1 \tilde{u}(n) e^{j\phi_m} + a_3 |\tilde{u}(n)|^2 \tilde{u}(n) e^{j\phi_m} + w_m(n) \right) \\ &= a_1 \tilde{u}(n) \sum_{m=1}^M e^{j\phi_m} + a_3 |\tilde{u}(n)|^2 \tilde{u}(n) \sum_{m=1}^M e^{j\phi_m} + \sum_{m=1}^M w_m(n) \end{aligned} \quad (4.6)$$

Since the noise terms $w_m(n)$ are independent and Identically Distributed (IID), their sum is also Gaussian with zero mean and variance $M\sigma^2$:

$$w_T(n) = \sum_{m=1}^M w_m(n) \sim \mathcal{N}(0, M\sigma^2) \quad (4.7)$$

As shown in Figure 4.3, the signals transmitted from two vertical subarrays can be written as: $\tilde{Z}_1(n)$ and $\tilde{Z}_2(n)$. The phase shifts are used to steer the beam in a desired direction. For boresight steering, identical phase shifts are applied to all subarrays:

$$\phi_1 = \phi_2 = \dots = \phi_M = \phi \quad (4.8)$$

so that the relative phase difference between any two subarrays m and l is

$$\Delta\phi_{ml} = \phi_m - \phi_l = 0 \quad (4.9)$$

Under this condition, the contributions from all subarrays add coherently:

$$\sum_{m=1}^M e^{j\phi_m} = M e^{j\phi} \quad (4.10)$$

In normal in-band operation, the signals radiated from each subarray combine coherently, producing a beamforming gain of M . The generalized cubic term in the Volterra PA also scales coherently, while the

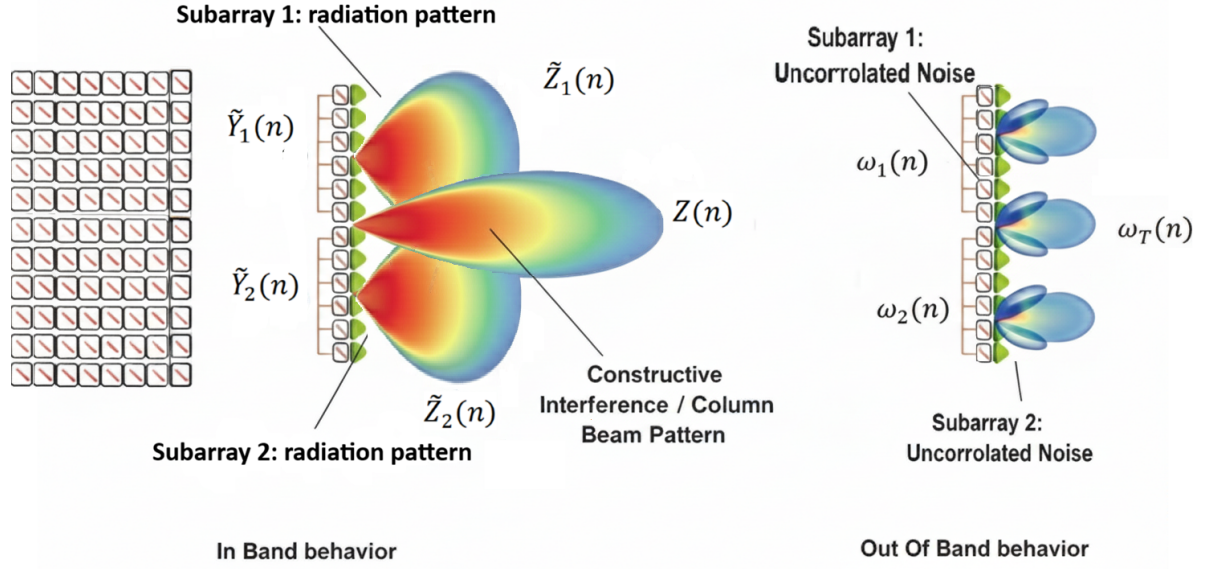


Figure 4.3: Difference between in-band and out-of-band spatial behavior.

noise scales incoherently. The far-field signal in the boresight direction is thus:

$$\tilde{Z}(n) = Ma_1 \tilde{u}(n) e^{j\phi} + Ma_3 |\tilde{u}(n)|^2 \tilde{u}(n) e^{j\phi} + w_T(n) \quad (4.11)$$

This can be decomposed into three components:

- **Useful signal:** $Ma_1 \tilde{u}(n) e^{j\phi}$
- **OOB emissions:** $Ma_3 |\tilde{u}(n)|^2 \tilde{u}(n) e^{j\phi}$
- **Noise:** $w_T(n) \sim \mathcal{N}(0, M\sigma^2)$

Equation (4.11) illustrates three important effects resulting from the application of beamforming gain exclusively to the desired signal:

- **Coherent scaling of the desired signal:** When DPD is active, the desired signal benefits from coherent beamforming, resulting in a total gain of $20\log_{10}(M) = 24$ dB. This 24 dB results from two contributions: a 12 dB beamforming gain (coherent combination) and a 12 dB power gain (from M antennas).
- **Incoherent scaling of noise:** The noise, which is not correlated between subarrays, adds incoherently, producing an array gain of M , corresponding to $10\log_{10}(M) = 12$ dB. Emissions add incoherently across antennas. Hence, only the 12 dB power gain remains, while the 12 dB beamforming gain vanishes due to spatial averaging.

The OOB coherent and incoherent scaling will be discussed apart in a further detail.

Apporximation used: In this study, we consider a system operating at a carrier frequency of $f_c = 3.75$ GHz with a signal bandwidth of $B = 100$ MHz. These values are representative of modern wireless communication systems, such as mid-band 5G deployments.

The first important observation is that the signal bandwidth is small compared to the carrier frequency. This can be seen from the ratio

$$\frac{B}{f_c} = \frac{100 \text{ MHz}}{3.75 \text{ GHz}} \approx 0.027$$

which corresponds to only about 2.7% of the carrier frequency. Because the bandwidth occupies only a small fraction of the carrier frequency, the system can be accurately modeled using a **narrowband approximation**.

Under this narrowband assumption, all frequency components within the signal band are very close to the carrier frequency. In other words, for any two frequencies f_1 and f_2 within the signal bandwidth, we can assume

$$f_1 \approx f_2 \approx f_c$$

This approximation simplifies the analysis because the propagation characteristics and phase shifts across the antenna elements become nearly identical for all frequencies within the band.

As a consequence, the antenna array can apply the same beamforming phase configuration to all antenna elements. Since the variation of frequency within the band is small, this phase configuration effectively steers all frequency components in approximately the same spatial direction.

This allows the use of narrowband amplitude and phase approximations, where the array response does not need to be recalculated for each individual frequency component.

An important implication concerns the behavior of OOB emissions. Because the same beamforming phase shifts are applied across the antenna array, and because the frequencies of the unwanted emissions remain close to the carrier frequency, the spatial radiation pattern of these components is nearly aligned with that of the desired signal.

Consequently, the OOB emissions propagate in almost the same direction as the useful signal rather than being radiated uniformly in space. This means that beamforming influences not only the desired in-band signal but also the unwanted spectral components.

4.2.3 Closed loop DPD for digital MIMO transmitters

Taking into account the input signal $\tilde{u}(n)$, which in the case of OFDM is a multitone waveform composed of many closely spaced subcarriers, nonlinear operations generate new frequency components due to intermodulation. For example, the cubic term $|\tilde{u}(n)|^2 \tilde{u}(n)$ produces third-order intermodulation products that fall both within the useful band and into adjacent frequency bands [48]. Although some of these products can be attenuated by analog cavity filters, others fall inside the filter passband and cannot be removed by hardware alone. If not mitigated, this spectral regrowth would result in adjacent channel interference and violation of regulatory limits, such as the 3GPP ACLR requirement. The role of DPD is to pre-compensate the nonlinear operation $a_3 |\tilde{u}(n)|^2 \tilde{u}(n)$ (and higher-order terms), effectively canceling their contribution before the PA output as described in Figure 1.18. In the array architecture, this predistorted signal is then fed into the beamforming stage, represented mathematically by $e^{j\phi_m}$ in equation (4.11), ensuring that both the linearized spectrum and spatial directivity are preserved.

4.2.4 Simulation expectations and results

We performed a simulation to analyze the behavior of baseband signals extracted from a DU under linear amplification. The input IQ samples were obtained from the DU and represent realistic baseband traffic and to analyze the performance of the DPD algorithm in linearizing the non linearities based on the algorithm defined in 1.3.6.2 and [50].

The simulation simulates the transmission of a complex baseband signal through a nonlinear system and evaluates its spectral behavior in the presence of noise. A recorded IQ signal is first reconstructed in

the time domain and oversampled using a low-pass Finite Impulse Response (FIR) filter to allow accurate observation of OOB effects. A memoryless third-order nonlinear model is applied to emulate transmitter distortion. Then additive white Gaussian noise is introduced to generate multiple noisy realizations of the signal, representing independent receiver branches or antenna elements. These realizations are coherently summed to model an antenna array or subarray combination. Finally, FFT-based spectral analysis is performed to compare the reference transmit signal, an individual noisy realization, and the combined array signal, highlighting the impact of nonlinear distortion and noise averaging on the output spectrum.

4.2.4.1 Simulation approach

The simulation considered a multi-antenna array with 16 elements. The baseband signal was upsampled by a factor of 7 to accurately capture potential spectral effects. A linear memoryless PA was applied, defined as

$$y(n) = a_1 x(n), \quad (4.12)$$

i.e., the output signal is a direct scaled version of the input without any nonlinear distortion.

By assuming a white noise characteristic, the ADC noise power scales linearly with bandwidth. Using the total measured noise, the noise power within a 100 MHz bandwidth was calculated to be approximately -61.4 dBFS as seen in Appendix C. This method provides a reliable way to estimate the expected noise for arbitrary bandwidths based on an initial integrated measurement.

In this work, the SNR is evaluated using multitone excitation rather than a single sinusoidal input. The multitone signal consists of multiple discrete frequency components distributed throughout the band of interest. The total signal power is defined as the sum of the powers of all individual tones.

The multitone waveform is scaled so that its overall Root Mean Square (RMS) voltage is 1 V. This corresponds to a full-scale input condition for the ADC, and therefore the total signal power of the multitone excitation is referenced to 0 dBFS. Although each individual tone occupies only a fraction of the full-scale range, their combined power represents the full-scale signal energy.

The noise power is obtained by integrating the noise floor over the specified bandwidth while excluding the frequency bins occupied by the signal tones and their associated spectral leakage regions.

Because both the total multitone signal power and the integrated noise power are referenced to the same full-scale level, the SNR can be expressed directly as the difference between these two quantities in decibel form. For a noise integration bandwidth of 100 MHz and a full-scale multitone signal with an RMS voltage of 1 V, the resulting SNR is:

$$\text{SNR} = 61.4 \text{ dB} \quad (4.13)$$

The signals from all antennas were then summed to represent the full array output. This results in a scaling of the subarray signal of 24 dB.

Simulation parameters:

- Number of antennas/subarrays: 16
- Oversampling factor: 7
- FFT size: 16384 points
- FFT smoothing window: 30 points
- $a_1 = 1$

Finally, the frequency spectra in Figure 4.4 of both a single subarray and the full antenna array were computed using a high-resolution FFT from the reference signal. The results illustrate the ideal linear

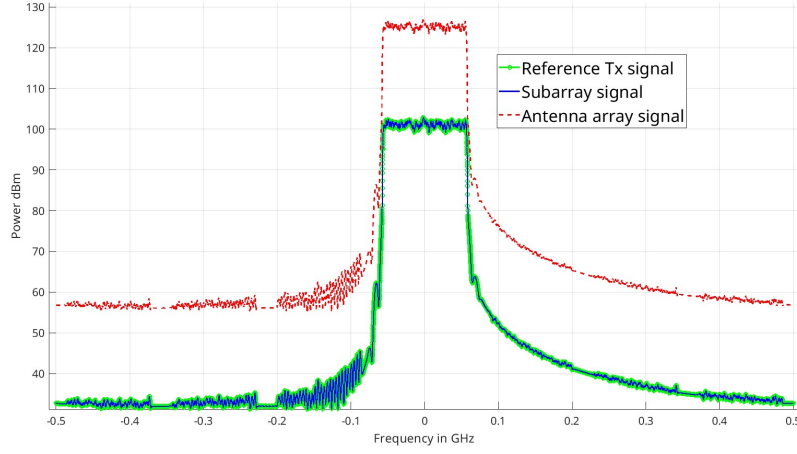


Figure 4.4: Simulation of the power spectrum with linear PA.

amplification scenario, providing a reference for comparison with nonlinear PA behavior. We can clearly observe the 24 dB difference in the useful band and the 12 dB difference in the OOB regions as predicted theoretically in equation (4.11).

We repeated the same simulation setup described for the linear case, this time introducing a nonlinear PA model to evaluate its effect on the transmitted signal. As before, we used the same multi-antenna configuration, upsampling rate, and signal-to-noise ratio that were maintained to ensure a fair comparison.

In this case, each antenna branch includes a memoryless Volterra series PA model of third-order AM/AM type, representing a simplified yet accurate approximation of nonlinear amplification behavior. The PA is modeled as:

$$y(n) = a_1 x(n) + a_3 |\tilde{x}(n)|^2 \tilde{x}(n) \quad (4.14)$$

where $a_1 = 1$ and $a_3 = 0.3 + 0.3j$ controls the strength of the nonlinear component. This third-order model captures the dominant AM/AM distortion, responsible for spectral regrowth and intermodulation in the output signal. The model is memoryless, meaning that each output sample depends only on the instantaneous input value, without any temporal memory or dynamic effects. No DPD was applied in this simulation. The signals from all 16 antennas were coherently summed to represent the combined array output.

The results in Figure 4.5 illustrate how nonlinear distortion affects the transmitted signal and how it accumulates throughout the array, resulting in shoulders on both sides of the useful band due to the leakage of nonlinear intermodulations from the useful band to adjacent bands. In this case the OOB emissions will be beamformed in the intended spatial direction.

To compensate for the PA nonlinearity, a third-order predistorter is trained using a direct inverse learning approach based on the algorithm defined in 1.3.6.2 and [50]. A subset of the oversampled signal is passed through the PA, and a cubic inverse model is estimated by minimizing the mean-squared error between the desired linear signal and the reconstructed estimate. The predistorter coefficient is adapted iteratively using a gradient-descent update rule.

Once the cubic inverse coefficient is learned, the predistorter is applied to the full transmit signal prior to the PA. The predistorted signal is then amplified by the nonlinear PA, and the resulting output spectrum is analyzed

We compare in Figure 4.6, the reference spectral power of one reference OFDM signal frame with a version of the same signal after passing through the non linear amplifier, and a 3rd version which

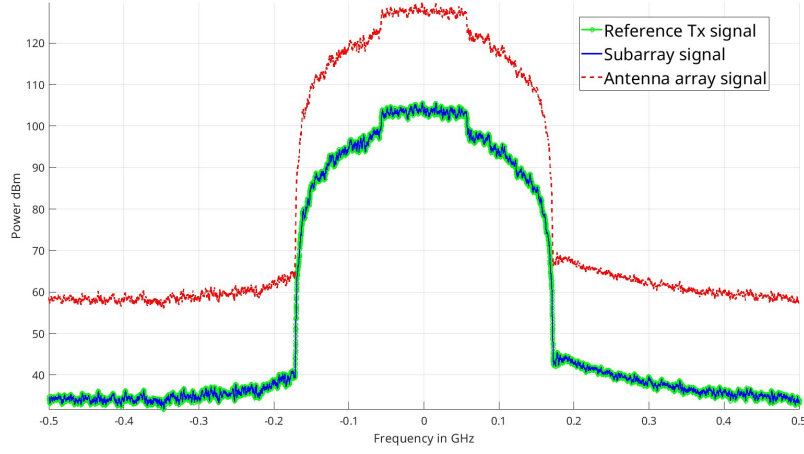


Figure 4.5: Simulation of the power spectrum with non linear PA.

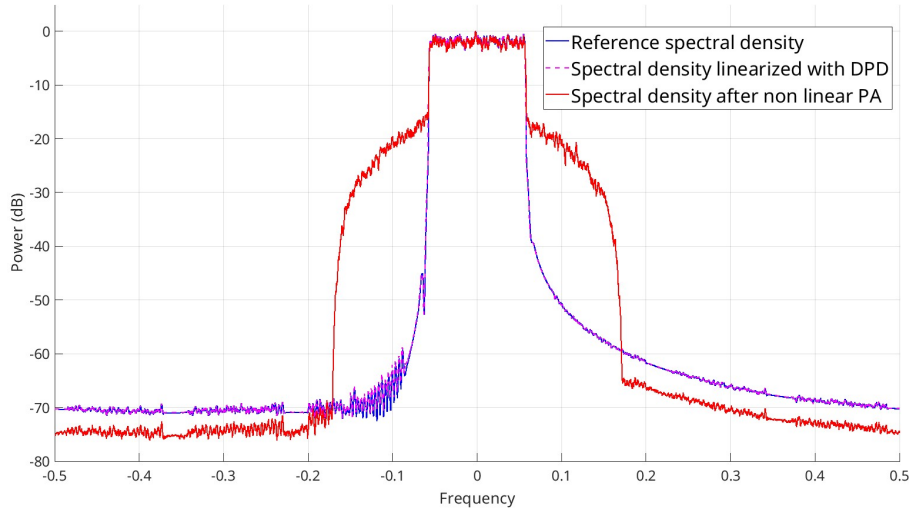


Figure 4.6: Simulation of the DPD algorithm

is linearized with the DPD after being subject to OOB distortions from the PA, and all signals were normalized with respect to the reference subarray signal. The figure shows how the DPD managed to linearize the signal and suppress OOB emissions. The shoulders on both sides of the channel band were suppressed, and the signal resembled the reference signal as both curves were perfectly aligned.

4.2.5 Correlation function theoretical modeling

In multi-path or multi-transceiver systems, unwanted signals such as distortion, interference, or noise are often generated independently within each signal path. However, these unwanted components are not always statistically independent. Depending on their physical origin, the same impairment may appear in multiple transceiver paths with varying degrees of similarity. Understanding and modeling this similarity is essential, because correlated unwanted signals can combine coherently across paths and therefore have a much more severe impact on system performance than uncorrelated ones[56].

The correlation of two antenna subarrays can be expressed by a coefficient $0 \leq \rho \leq 1$, which quantifies the similarity of unwanted signals at the outputs when the same input signal is applied. In some cases, the signals are fully correlated ($\rho = 1$); for example, when the output signal is generated digitally and will appear identically in each path. In other cases, the signals are completely uncorrelated ($\rho = 0$); for instance, thermal noise is inherently random and thus shows no similarity across paths. Since the degree of correlation depends on the implementation and the geometry of the antenna array, different values of ρ must be considered to ensure that worst case conditions are captured properly for specification purposes. [56].

We will deduce the correlation coefficient later once we present the measured subarray and antenna array beam pattern and taking $G = 16$, but we will refer to (C.3) to calculate the correlation coefficient based on the theoretical model in function of G , $A_A(\theta, \varphi)$ and $A_{sub}(\theta, \varphi)$.

Recalling (C.3):

$$\begin{aligned} A_A(\theta, \varphi) &= A_{sub}(\theta, \varphi) + 10 \log_{10} [1 + \rho(G - 1)] \\ A_A(\theta, \varphi) - A_{sub}(\theta, \varphi) &= 10 \log_{10} [1 + \rho(G - 1)] \\ \frac{A_A(\theta, \varphi) - A_{sub}(\theta, \varphi)}{10} &= \log_{10} [1 + \rho(G - 1)] \\ 10^{\frac{A_A(\theta, \varphi) - A_{sub}(\theta, \varphi)}{10}} &= 1 + \rho(G - 1) \\ 10^{\frac{A_A(\theta, \varphi) - A_{sub}(\theta, \varphi)}{10}} - 1 &= \rho(G - 1) \\ \rho &= \frac{10^{\frac{A_A(\theta, \varphi) - A_{sub}(\theta, \varphi)}{10}} - 1}{G - 1} \end{aligned}$$

This expression indicates that the correlation coefficient is directly linked to the difference between the array and subarray radiation patterns in a given observation direction. In practice, however, both the array and subarray radiation patterns depend not only on the spatial direction (θ, φ) but also on the operating frequency f . This frequency dependence arises from the frequency-selective behavior of the antenna elements, the feeding network, and the transceiver impairments that generate the unwanted signals.

To account for this, the beam patterns are more generally written as

$$A_A(f, \theta, \varphi), \quad A_{sub}(f, \theta, \varphi), \quad (4.15)$$

and the pattern difference is defined as

$$\Delta(f, \theta, \varphi) = A_A(f, \theta, \varphi) - A_{sub}(f, \theta, \varphi). \quad (4.16)$$

Substituting this definition into the previous expression yields a correlation coefficient that is, in general, dependent on both frequency and direction:

$$\rho(f, \theta, \varphi) = \frac{10^{\frac{\Delta(f, \theta, \varphi)}{10}} - 1}{G - 1}. \quad (4.17)$$

While this formulation is the most general and rigorous, it is often desirable to characterize the correlation behavior as a function of frequency only. This is particularly relevant for system-level analysis and specification, where a single correlation coefficient per frequency is preferred. In such cases, the angular dependence is removed by applying an averaging or worst-case operation over the spatial domain.

The resulting frequency-dependent correlation coefficient is then given by

$$\rho(f) = \frac{10^{\frac{\Delta(f)}{10}} - 1}{G - 1}. \quad (4.18)$$

This formulation preserves the physical interpretation of the correlation coefficient while ensuring consistency between the spatial and frequency domains. By explicitly distinguishing between angular and frequency dependencies, the model enables a systematic investigation of correlated and uncorrelated unwanted signal contributions across the operating bandwidth. This, in turn, allows worst-case scenarios to be identified and supports robust system design and specification.

The correlation coefficient can be written as demonstrated in Appendix C with the calculation of G , we obtained $G = 16$.

4.3 Measurement setup

To evaluate the performance of the ORAN Split 7.2 architecture, the experimental setup consists of a DU located at EURECOM, connected via fiber optics to a Massive MIMO RU housed inside an anechoic chamber along with the network switch. The setup requires tight time synchronization between DU and RU, and involves complex hardware and software dependencies, including real-time processing constraints, high-throughput data interfaces, and precise coordination between the ORAN software stack and radio hardware. Measurements are performed under controlled laboratory conditions, assuming quasi-far-field behavior after near-field to far-field transformation

4.3.1 RU specifications

Spec	Comments
Operating Band	3400–3800 MHz (Band n78)
Standard	5G NR
Number of TX/RX Paths	32
Output Power	200 W
Instantaneous/Occupied BW	200 MHz / 200 MHz (HW ready, SW supports 1 × 100 MHz)
Number of Carriers	2 × 100 MHz (HW supports multiple bandwidths: 20–100 MHz; SW supports 1 × 100 MHz only)
Number of Layers	16 DL, 8 UL
Supported SCS	30 kHz and 60 kHz (SW supports only 30 kHz currently)
Front Haul Interface	2 × 25Gb (SFP28, ORAN Cat B, SW supports 1 × 25Gb only)
Total EIRP	77 dBm (2 × 74 dBm)
Rx Sensitivity (100 MHz)	-89 dBm (1 Rx path); EIS: -114 dBm/100 MHz
Number of Antenna Elements	192 (12×8×2 configuration with 6×1 sub-arrays)
Polarization mode	-45°, +45°
Maximum Antenna Gain	24.3 dBi
Dynamic Horizontal Steering	±48°
Vert. Digital Steering Range	±3° (@6dB sidelobe suppression, ±1.5° @10dB sidelobe)
Total Vertical Steering	-15° to +3° (includes -12° up via RET motors and mechanical phase shifters)
Power Consumption (W)	750 W (typical: 80% DL, 20% UL TDD ratio)
Temperature Range	-40° to +55°
Dimensions	730 × 430 × 150 mm; Volume 47L; Weight 30 kg (*23 kg with bonded fin heatsink)

Table 4.1: Spear Andrew antenna array specification table

The Spear Andrew RU operates in the [3400–3800] MHz band (n78) with support for 5G NR and a maximum output power of 200 W. It features 32 TX/RX paths, supports 16 downlink and 8 uplink layers, and provides an instantaneous bandwidth of 200 MHz, although current software limits this to 1 × 100 MHz. The RU uses a 2 × 25Gb front-haul interface, with software currently supporting only 1 × 25Gb. It has 192 antenna elements in a 12×8×2 configuration, supports ±48° horizontal and ±3° vertical digital

4.3.2 Anechoic chamber setup

Figure 4.7: OTA measurement setup.

The RU's integrated Massive MIMO antenna array is mounted on a high-precision positioner capable of rotating in the azimuth plane. However, to observe radiation characteristics in the elevation plane, the entire antenna array is physically rotated by 90° relative to the positioner's axis. This reorientation effectively maps the positioner's azimuthal motion into elevation scanning, enabling beam pattern measurements within the angular range of $\theta \in [-60^\circ, 60^\circ]$.

During measurements, the beam emitted by the RU is directed toward a metallic reflector placed at a known distance and orientation. This reflector redirects the signal toward a receiving probe antenna, which is aligned with the target angular direction. The received signal is passed to a high resolution spectrum analyzer, which is used to visualize and analyze the shape, direction, and strength of the radiated beam. The complete physical implementation of this setup is shown in Figure 4.8.

4.3.3 RU-DU Integration and Configuration

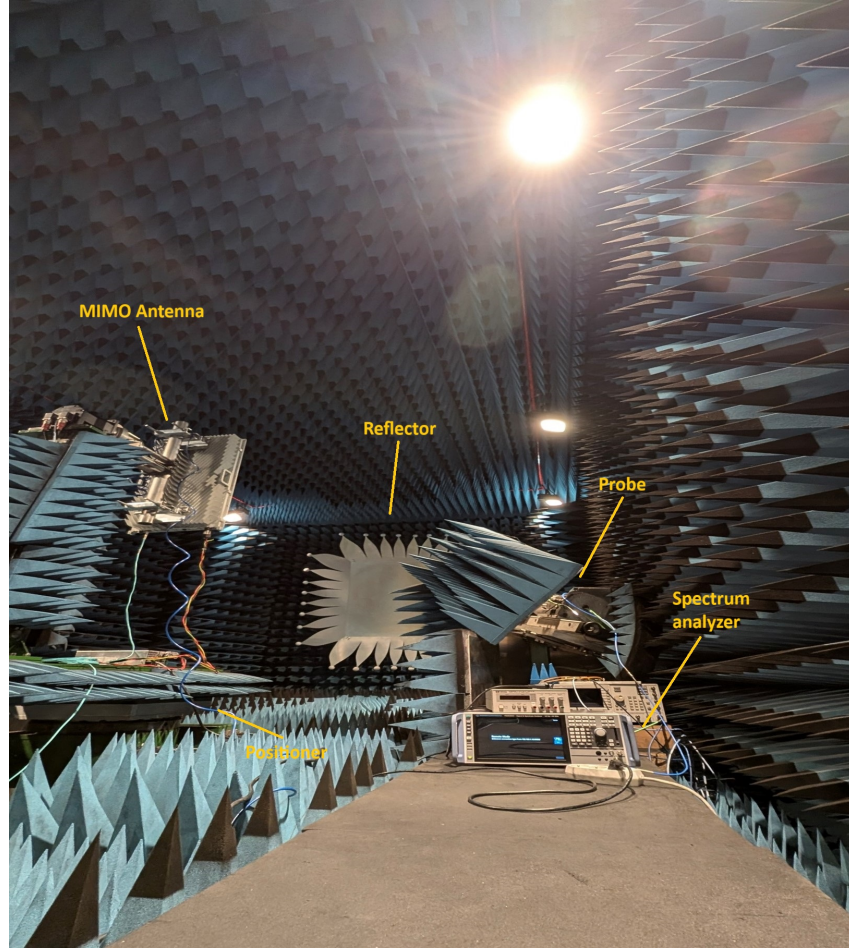


Figure 4.8: Anechoic chamber measurement setup

The RU and DU are both connected to a Fibrolan Falcon RX/G fronthaul switch, which provides both connectivity and time synchronization. Acting as a PTP grandmaster, the switch ensures that the RU and DU remain precisely time-aligned. This timing is essential for proper operation under the split 7.2 architecture, particularly when high-throughput and phase coherent transmissions are required. A step-by-step guide to replicating this setup is available in [60].

The signal transmitted from the RU is a 5G NR compliant OFDM waveform with a 100 MHz bandwidth centered at 3.75 GHz, within the n78 frequency band. To analyze both desired and unintended signal components, spectral measurements were taken across a broader range from 3 GHz to 4.2 GHz, capturing both in-band energy and OOB emissions.

Beamforming in the Spear Andrew RU is based on a fixed Beam ID scheme. This method uses a predefined codebook of beams, each associated with a unique set of beamforming weights. These complex weights are stored in a lookup table and are applied to the antenna elements when a specific Beam ID is selected. This allows the RU to quickly steer the beam in a desired direction with minimal latency and without the need for real-time computation, which makes it ideal for repeatable testing.

For this study, the beam with the highest gain in the boresight direction was selected from the RU's codebook. This beam corresponds to an angular direction of $(\varphi = 0^\circ, \theta = -6^\circ)$, and was chosen to maximize directivity during measurements. Using this beam as a reference helps to evaluate the performance of the antenna array under optimal transmission conditions.

4.4 OTA experimental results and analysis

In what follows, we present a comparative analysis of the maximum power levels measured with and without DPD, followed by an evaluation of coherent beamforming performance under two operating conditions: (i) within the PA's linear regime using DPD, and (ii) in the nonlinear regime without DPD. This comparison aims to highlight the impact of DPD on both the enhancement of radiated power and the preservation of beam coherence under realistic transmission scenarios.

4.4.1 EIRP definition

The radiated measurements presented in this work are performed using an OTA setup in which the antenna array is steered toward a receiving probe antenna located in the far field. The received signal is calibrated to obtain the EIRP in a given spatial direction and for a single polarization. According to 3GPP definitions, EIRP represents the equivalent isotropic power radiated in a specific direction and is distinct from the Total Radiated Power (TRP), which requires integration of the radiated power over the full solid angle and over two orthogonal polarizations [51, 52].

As specified in 3GPP TS 38.141-2 [52], full radiated conformance testing for base stations, including TRP and SEM, requires angular scanning or near-field to far-field transformation across a spherical surface and polarization diversity [52]. Such measurements are outside the scope of this work. Instead, the results reported here focus on directional EIRP based spectral characteristics, which are relevant for evaluating beamformed emissions and the effectiveness of DPD in suppressing OOB radiation along the main beam direction.

Only one polarization is considered in the OTA measurements, as the objective of this study is to analyze the spectral behavior and DPD effectiveness along the dominant beam polarization. Full dual-polarized integration, as required for TRP evaluation, is left for future work in accordance with TS 38.141-2 [52].

4.4.2 Spectral power comparison with DPD in isopower mode

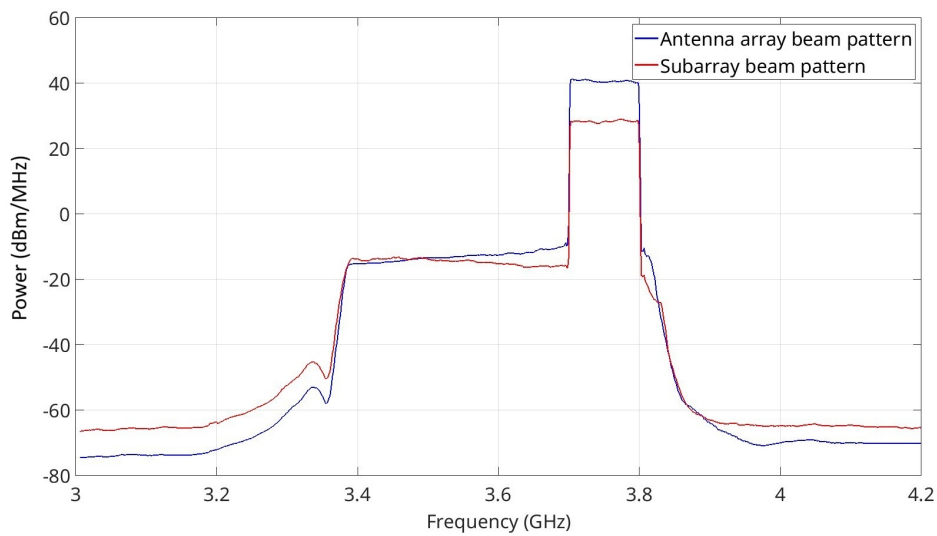


Figure 4.9: Full antenna array spectral power vs subarray spectral power in isopower mode with DPD

Figure 4.9 presents a comparison of the isopower spectrum measured from a single subarray and the full Massive MIMO antenna array in isopower mode, by reducing the spectral power curve of the antenna array by 12 dB, to reference it with respect to the subarray spectral power curve. The in-band performance of the Massive MIMO array demonstrates a substantial improvement over a single subarray, achieving approximately 12 dB of beamforming gain in isopower mode which is relative to the inband useful signal part $Ma_1 \tilde{u}(n)e^{j\phi}$ in equation (4.11). In contrast, OOB emissions from single subarrays are largely uncorrelated, resulting in minimal spatial coherence and no effective beamforming and are relative to the decorrelated noise components $w_T(n)$ in equation (4.11); consequently, the full-array OOB pattern closely resembles that of a single subarray with a nearly flat power distribution and negligible suppression. The application of DPD enables spatial filtering of OOB emissions on the left side of the signal band, while the cavity filter mitigates the right side OOB signals, ensuring that the array maintains high in-band fidelity and meets stringent spectral emission requirements.

In the single user case, we measured the spurious emissions as well for illustrative purposes. The spurious emissions are located in the bands [3-3.36]GHz and [3.84-4.2]GHz. Due to the limited dynamicity of the spectrum analyzer, the spurious emissions were drowned in noise, so we used a band reject filter and a low noise amplifier to measure their power levels. The spurious emissions were then scaled to EIRP by compensating the propagation loss and the the band reject filter's impulse response.

These observations underscore one of the key advantages of Massive MIMO systems: the ability to enhance radiated spectral containment through both digital signal processing and array-level spatial effects.

We can now compute the correlation coefficient defined in 4.2.5. Having the antenna array spectrum and one subarray spectrum, we computed $\Delta(f)$ in Figure 4.10:

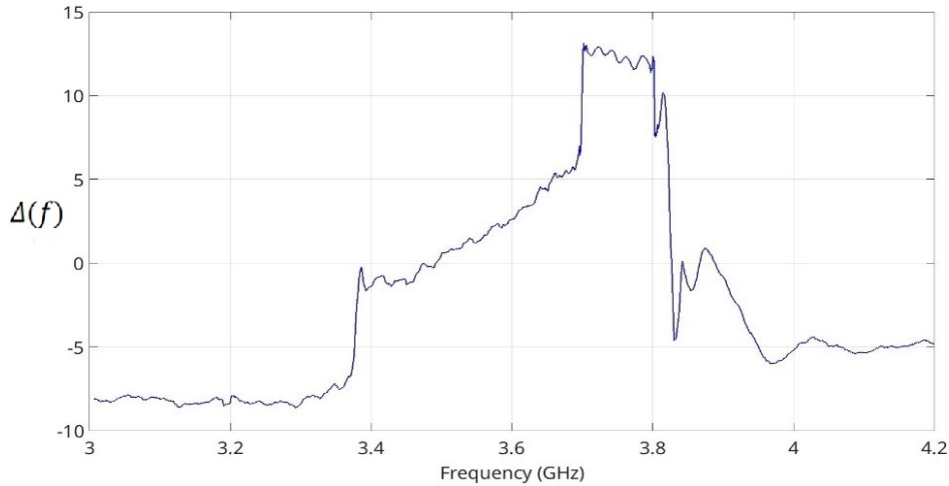


Figure 4.10: Difference between full antenna array spectrum and one subarray spectrum from Figure 4.9 without DPD

Replacing $\Delta(f)$ in the correlation coefficient formula in equation (4.18) leads to have the blue curve in Figure 4.11:

The correlation plot in Figure 4.11 gives an estimate of the experimental correlation coefficient defined in equation (4.18). We can observe that in the in band part spectrum in [3.7-3.8] GHz, the correlation coefficient decreases from 1.3 to 1.1. This behavior is close to the theoretical behavior where $\rho = 1$ in the red curve. We can then observe the jump while decreasing on the right side of the useful band before converging close to 0, this is due to the transition through the extremity of the cavity filter. We also can observe the progressive slow increase with respect to the frequency as the correlation moves close

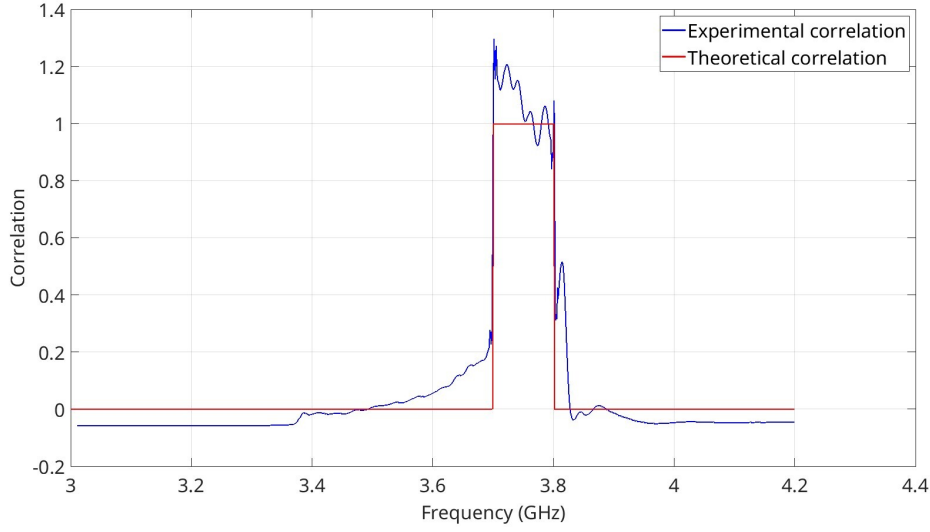


Figure 4.11: Correlation coefficient in function of the frequency without DPD

to the useful band and this phenomenon points at the imperfect effectiveness of the DPD in suppressing the OOB emissions inside the cavity filter. This aspect opens the door to further DPD enhancement and optimization to linearise the PA output signal.

4.4.2.1 3GPP specifications conditions

The ACLR is defined in 3GPP TS 38.104 [51] as the ratio of the transmitted power in the wanted channel to the power leaked into adjacent channels. While ACLR is traditionally defined for conducted measurements, the same ratio can be evaluated in the radiated domain using EIRP-based power measurements, provided that both the wanted and adjacent channel powers are referenced to the same spatial direction and measurement bandwidth [51]. In this work, ACLR is therefore computed from directional EIRP-calibrated spectra, yielding a directional ACLR value representative of the beamformed transmission. High ACLR values indicate cleaner spectral emissions and reduced interference to neighboring users and systems.

In the n78 band (3.3 – 3.8 GHz), one of the most widely deployed 5G NR FR1 TDD bands, ACLR requirements are especially important due to the dense spectrum usage and wide channel bandwidths (up to 100 MHz). Operators often deploy multiple contiguous carriers in this band, so transmitter leakage can directly degrade adjacent cell performance and spectrum efficiency.

gNB must meet strict ACLR requirements, generally ≥ 45 dB, or comply with specified absolute emission limits in dBm/100 MHz for TRP based measurements. These levels ensure coexistence between adjacent 5G carriers and minimize interference into neighboring 4G or fixed services operating near 3.8 GHz.

In Figure 4.13, at the main beam direction, the EIRP-calibrated power integrated over a 100 MHz bandwidth in the wanted channel is 72.5 dBm, while the integrated EIRP in the left adjacent channel is 20.7 dBm. The resulting directional ACLR, defined as the ratio between the wanted and adjacent-channel EIRP, is approximately 52 dB, exceeding the 3GPP requirement of 45 dB.

In practice, achieving ACLR targets in n78 requires careful RF front end design, including PA linearization with DPD, precise filtering, and optimized digital predistortion. Because n78 operates in a mid-band region, designers face a trade-off between high output power and linearity. Insufficient ACLR can cause spectral regrowth, cross-carrier interference, and network capacity degradation in multi-operator or shared spectrum scenarios.

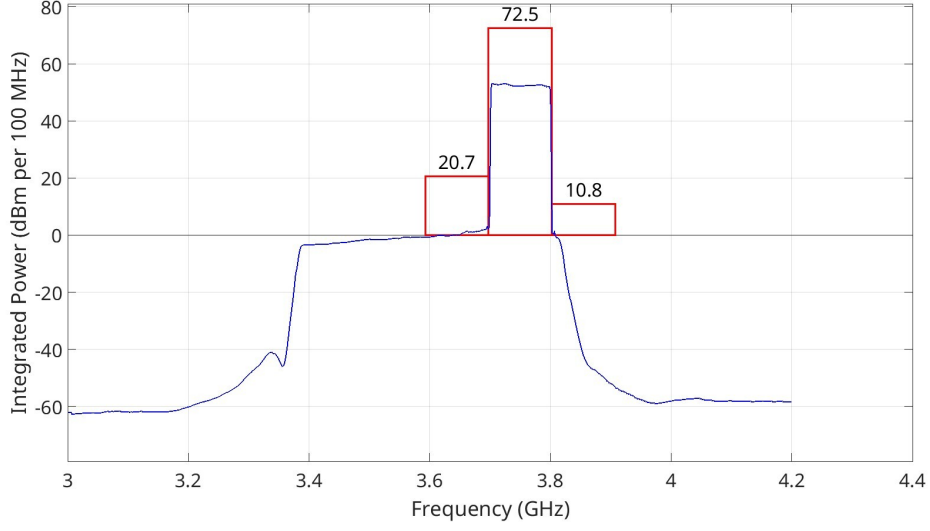


Figure 4.12: Directional OTA spectrum measured at the main beam direction. The plotted values correspond to EIRP-calibrated power integrated over a 100 MHz bandwidth for a single polarization.

This EIRP based ACLR measurement gives an idea and an estimation about TRP based ACLR.

4.4.2.2 3GPP spectrum emission mask

Frequency offset of measurement filter -3 dB point, Δf	Frequency offset of measurement filter centre frequency, f_{offset}	Test requirement bandwidth	Measurement
$0 \text{ MHz} \leq \Delta f < 5 \text{ MHz}$	$0.05 \text{ MHz} \leq f_{\text{offset}} < 5.05 \text{ MHz}$	$4 \text{ dBm} - \frac{7}{5} \left(\frac{f_{\text{offset}}}{\text{MHz}} - 0.05 \right) \text{ dB}$	100 kHz
$5 \text{ MHz} \leq \Delta f < \min(10 \text{ MHz}, \Delta f_{\text{max}})$	$5.05 \text{ MHz} \leq f_{\text{offset}} < \min(10.05 \text{ MHz}, f_{\text{offset,max}})$	-3 dBm	100 kHz
$10 \text{ MHz} \leq \Delta f \leq \Delta f_{\text{max}}$	$10.5 \text{ MHz} \leq f_{\text{offset}} < f_{\text{offset,max}}$	-6 dBm	1 MHz

Table 4.2: Measurement filter frequency offset requirements[52]

The 3GPP emission mask, also referred to SEM, defines the maximum permitted OOB and spurious emissions from a 5G transmitter as a function of frequency offset from the allocated channel band. Its purpose is to ensure that the transmitted power remains confined within the assigned bandwidth and does not cause interference to adjacent channels or other radio services. Unlike the ACLR, which measures the relative leakage power between adjacent channels, the emission mask specifies absolute power limits in dBm/MHz across the frequency spectrum [52] as seen in table 4.3.

Frequency offset of measurement filter -3 dB point, Δf	Frequency offset of measurement filter centre frequency, f_{offset}	Test requirement bandwidth	Measurement
$0 \text{ MHz} \leq \Delta f < 5 \text{ MHz}$	$0.05 \text{ MHz} \leq f_{\text{offset}} < 5.05 \text{ MHz}$	$14 \text{ dBm} - \frac{7}{5} \left(\frac{f_{\text{offset}}}{\text{MHz}} - 0.05 \right) \text{ dB}$	1 MHz
$5 \text{ MHz} \leq \Delta f < \min(10 \text{ MHz}, \Delta f_{\text{max}})$	$5.05 \text{ MHz} \leq f_{\text{offset}} < \min(10.05 \text{ MHz}, f_{\text{offset,max}})$	7 dBm	1 MHz
$10 \text{ MHz} \leq \Delta f \leq \Delta f_{\text{max}}$	$10.5 \text{ MHz} \leq f_{\text{offset}} < f_{\text{offset,max}}$	-6 dBm	1 MHz

Table 4.3: Measurement filter frequency offset requirements scaled to 1 MHz

The 3GPP SEM specifies absolute limits on OOB emissions expressed in dBm/MHz to ensure coexistence with adjacent services [52]. In the context of radiated measurements, SEM compliance is

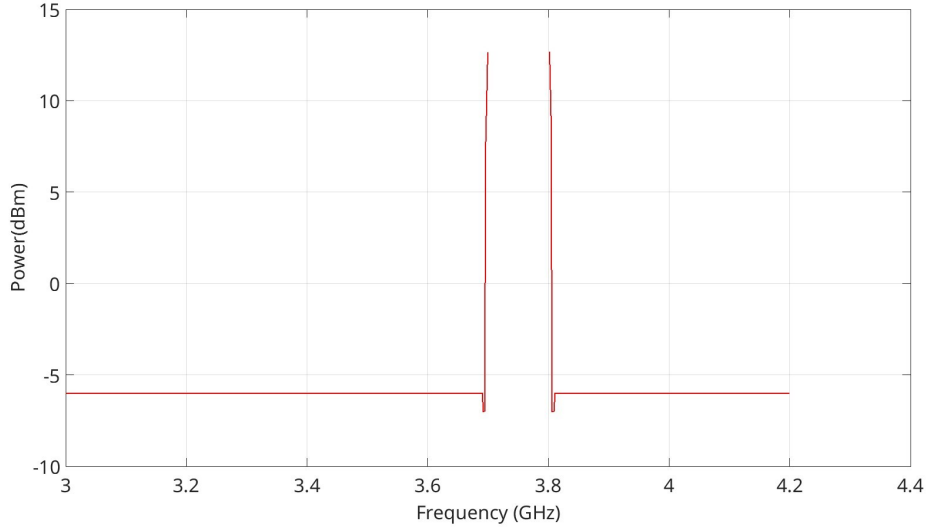


Figure 4.13: The spectrum emission mask plot from Table 4.3.

formally assessed using TRP-based metrics obtained from angular integration over a spherical surface [52] as seen in table 4.3. The SEM results presented in this work are therefore derived from directional EIRP measurements at the main beam direction and should be interpreted as an indication of spectral confinement and relative OOB suppression rather than as a full 3GPP radiated conformance assessment.

The 3GPP emission mask is defined for 100 KHz Resolution Bandwidth (RBW) for the in-band and the transition band edges. To scale it to a RBW of 1 MHz, an extra 10 dB should be added to the power level in the test requirement bandwidth to obtain the modified table previously that will be used for the SEM plotting.

4.4.2.3 Extension toward TRP based ACLR and SEM evaluation

The directional OTA measurements presented in this work provide the spectral power radiated by the antenna array along a single beam direction, calibrated to EIRP. While this approach enables a reliable assessment of beam-level linearity, OOB emissions, and DPD effectiveness, it does not constitute a full radiated compliance evaluation, which requires the TRP according to 3GPP TS 38.141-2 [52].

In the far-field region, TRP is defined as the spatial average of the equivalent isotropically radiated power over the full radiation sphere. In practical radiated conformance testing, this spatial averaging is performed using a finite number of discrete angular samples. Accordingly, the TRP estimate at frequency f is defined as

$$\text{TRP}_{\text{Estimate}}(f) = \frac{1}{N} \sum_{n=1}^N \text{EIRP}(f, \theta_n, \varphi_n), \quad (4.19)$$

where N is the total number of angular sampling points, and each pair (θ_n, φ_n) corresponds to a measurement direction on the radiation sphere as seen in Figure 4.14.

The angular sampling points are defined on a spherical grid with reference angular steps $\Delta\theta_{\text{ref}}$ and $\Delta\varphi_{\text{ref}}$ in elevation and azimuth, respectively. The total number of sampling points is given by

$$N = N_{\theta} N_{\varphi}, \quad (4.20)$$

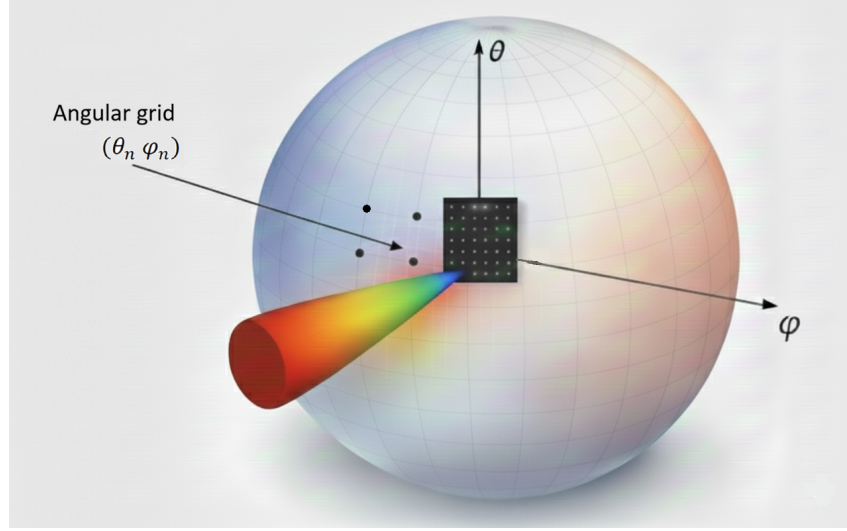


Figure 4.14: TRP measurement with a planar antenna array.

where

$$N_{\theta} = \frac{180^{\circ}}{\Delta\theta_{\text{ref}}}, \quad N_{\varphi} = \frac{360^{\circ}}{\Delta\varphi_{\text{ref}}}. \quad (4.21)$$

The number of sampling points shall satisfy the minimum sampling condition

$$N \geq \frac{4\pi}{\Delta\theta_{\text{ref}}\Delta\varphi_{\text{ref}}}, \quad (4.22)$$

to ensure sufficient coverage of the radiation sphere.

The reference angular steps are defined as

$$\Delta\theta_{\text{ref}} = \min\left(\frac{180^{\circ}}{\pi} \frac{\lambda}{D}, 15^{\circ}\right) \quad (4.23)$$

$$\Delta\varphi_{\text{ref}} = \min\left(\frac{180^{\circ}}{\pi} \frac{\lambda}{D_{\text{cyl}}}, 15^{\circ}\right) \quad (4.24)$$

where $D_{\text{cyl}} = \sqrt{(d^2 + w^2)}$ and $D = \sqrt{(d^2 + w^2 + h^2)}$ and d, w, h represent the depth, width and length of the antenna array respectively.

The sampling grid is determined by the reference angular sampling intervals $\Delta\theta_{\text{ref}}$ and $\Delta\varphi_{\text{ref}}$, as specified in Annex I.2.2 of 3GPP TS 38.141-2[52]. Measurements over both orthogonal polarizations are required at each sampling point, and the corresponding EIRP contributions are combined to obtain the total radiated power per direction.

A practical extension of the present measurement methodology therefore consists in repeating the calibrated directional OTA spectrum measurements over the discrete angular grid $\{(\theta_n, \varphi_n)\}_{n=1}^N$ covering the full radiation sphere, as illustrated in Figure 4.14. For each angular direction, the measured spectrum is converted to directional EIRP using the same calibration procedure as employed for the single beam case. The resulting set of directional EIRP spectra is then spatially averaged according to the expression above to obtain the TRP spectrum estimate.

Once the TRP spectrum estimate is obtained, TRP-based ACLR can be computed as

$$\text{ACLR}_{\text{TRP}} = 10\log_{10}\left(\frac{\text{TRP}_{\text{Estimate,in-band}}}{\text{TRP}_{\text{Estimate,adjacent}}}\right), \quad (4.25)$$

in accordance with 3GPP radiated conformance testing procedures. Similarly, the TRP spectral density can be directly compared against the 3GPP spectrum emission mask (SEM), enabling a complete radiated OOB compliance assessment.

This discrete-sampling-based formulation aligns the proposed methodology with 3GPP-defined TRP estimation procedures, while preserving the same OTA measurement principles and calibration framework used in the present work. It provides a systematic and standards-compliant path toward full TRP-based ACLR and SEM evaluation for radiated massive MIMO transmitters.

4.4.3 Spectral power comparison without DPD in isopower mode

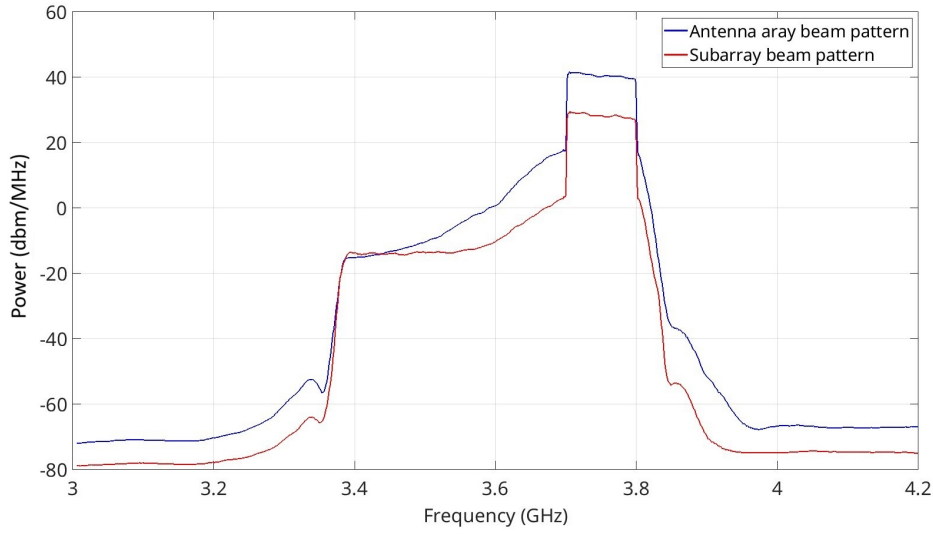


Figure 4.15: Full antenna array spectral power vs subarray spectral power in isopower mode without DPD

Disabling the DPD function leads to the leakage of significant nonlinear distortion components, most notably the third- and fifth-order harmonics (commonly referred to as H3 and H5), where we refer to the H3 components by the term $Ma_3|\tilde{u}(n)|^2\tilde{u}(n)e^{j\phi}$ from equation (4.11). These intermodulation products extend beyond the main signal band, leaking into adjacent frequency bands and causing unwanted spectral emissions. As depicted in Figure 4.15, while a fraction of these harmonics remain within the passband of the cavity filter (3.4–3.8 GHz), they produce pronounced spectral shoulders that adversely impact the quality of the in-band signal. This degradation is particularly evident on the lower frequency side of the band, where the harmonics are insufficiently attenuated, resulting in increased adjacent channel interference. In contrast, the harmonics on the upper side of the spectrum are suppressed more effectively by the cavity filter, thereby reducing their contribution to spectral regrowth in that region.

4.4.4 Spatial behavior comparison with DPD

To check that the measured antenna array pattern at bore sight is correct, we compared it in Figure 4.16 with the array factor of the beam at bore sight direction ($\theta_{etilt} = 6^\circ, \varphi_{escan} = 0^\circ$) using the model described in [56], where both of them were normalized to the maximum power. We clearly realize how it matches perfectly the main lobe, and that it follow the same behavior and variation of the secondary lobes with some slight power level differences.

With the implementation of the full antenna array, a significant increase in beam gain was observed, amounting to 24 dB, as illustrated in Figure 4.17 between the antenna array beam pattern and the subarray

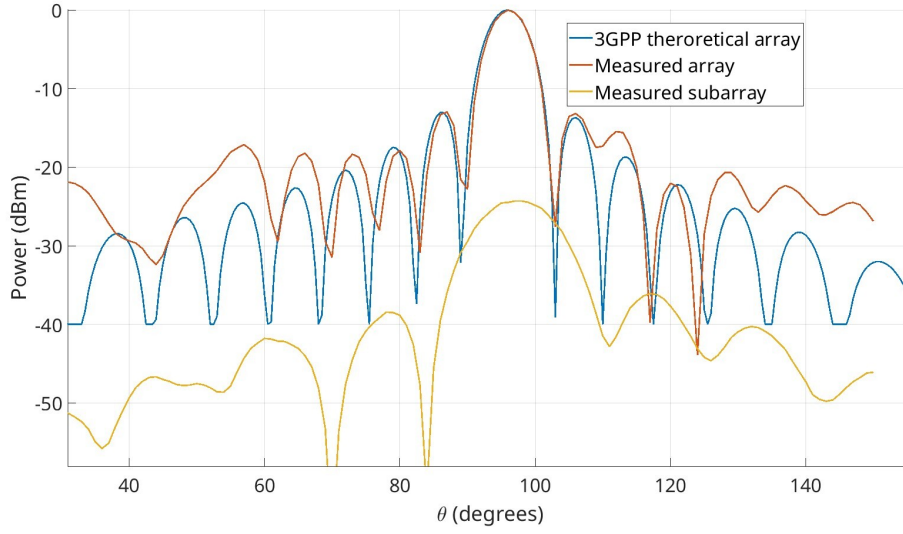


Figure 4.16: Comparison of the measured array radiation pattern with the 3GPP theoretical beam pattern normalized to maximum power in the elevation plane

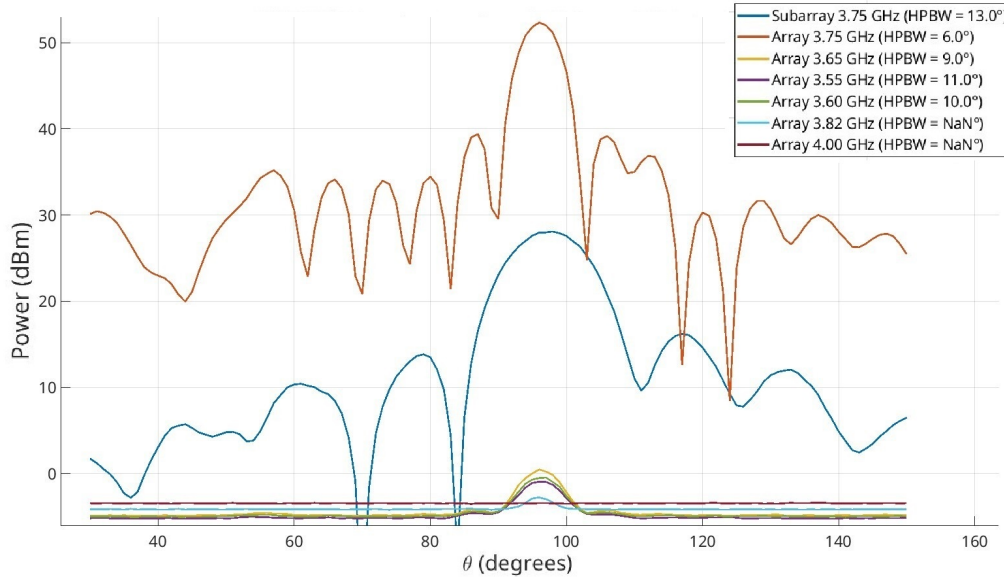


Figure 4.17: Comparison of the array radiation pattern with the measured subarray pattern and multiple OOB array beam pattern at different OOB frequencies with DPD

radiation pattern. This enhancement reflects a marked improvement in the direct performance of the antenna system. The gain increase is primarily the result of the constructive interference of signals from multiple array elements, which effectively concentrates the radiated power in the desired direction when DPD is enabled.

In addition to the gain improvement, the number and amplitude of the side lobes were substantially reduced. This reduction contributes to more effective suppression of unwanted radiation in off-axis directions, thereby enhancing overall beam efficiency and mitigating potential interference. Together, these improvements demonstrate the effectiveness of full-array beamforming implementation and its

ability to achieve the intended radiation characteristics.

The OOB radiation pattern of the full antenna array at different OOB frequencies closely resembles that of a single subarray, as shown in Figure 4.17. This is evident from the Half Power Beamwidth (HPBW), which matches closely between the array and the measured subarray, indicating a lack of spatial coherence in the OOB emissions.

Unlike in-band signals, which are coherently combined across antennas to form a directive beam, OOB components do not benefit from constructive alignment. Instead, they originate from nonlinearities or hardware imperfections and thus lack consistent phase relationships across the array. As a result, the array behaves more like a collection of independent radiators than a coherent beamforming system in the OOB region.

This lack of coherence has a beneficial consequence: it suppresses the overall OOB radiation in massive MIMO systems. Since the OOB contributions from different antennas add non-coherently, the combined radiation averages out spatially, reducing the power projected in any specific direction. Consequently, while each antenna may emit some OOB energy, the aggregated emissions do not scale with the antenna array size in the same way as in-band signals, leading to inherently lower OOB interference.

4.4.5 Spatial behavior comparison with DPD in the azimuth plane

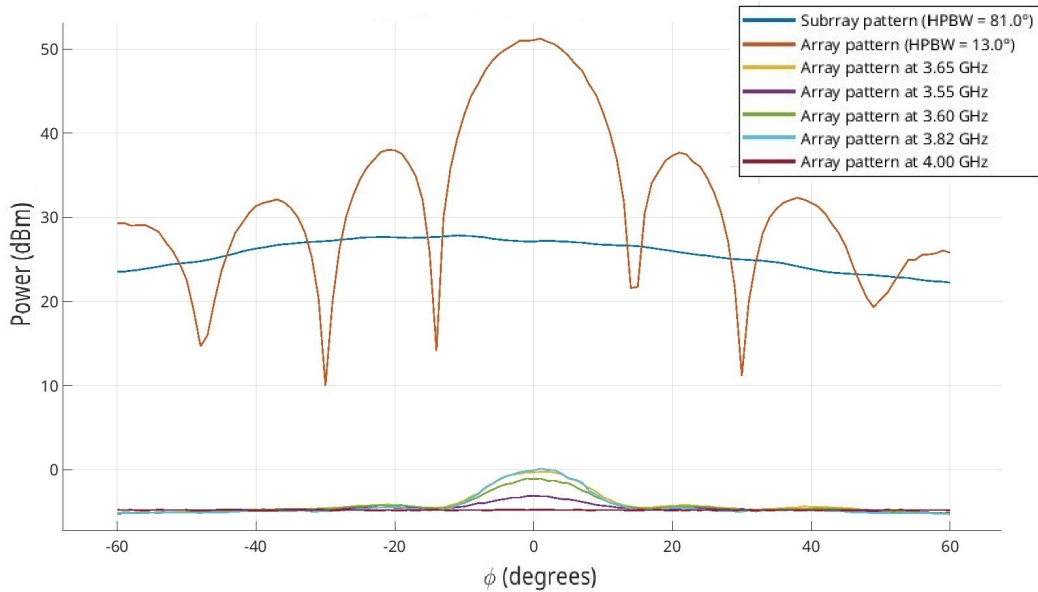


Figure 4.18: Comparison of the array radiation pattern with the measured subarray pattern and multiple OOB array beam pattern at different OOB frequencies with DPD in the azimuth plane

With the complete antenna array, beam gain increased by 24 dB, while sidelobe levels were significantly reduced, improving directivity in the azimuth plane. In contrast, the OOB radiation pattern remained similar to that of a single subarray as shown in Figure 4.18, indicating non-coherent combining of OOB emissions. This means that the OOB energy does not scale with the antenna array size and is naturally suppressed. The importance of multiuser MIMO is that high in-band beam directivity can be achieved in azimuth without increasing OOB interference. This enables better spatial user separation and improved coexistence in dense networks.

4.4.6 Spatial behavior comparison without DPD

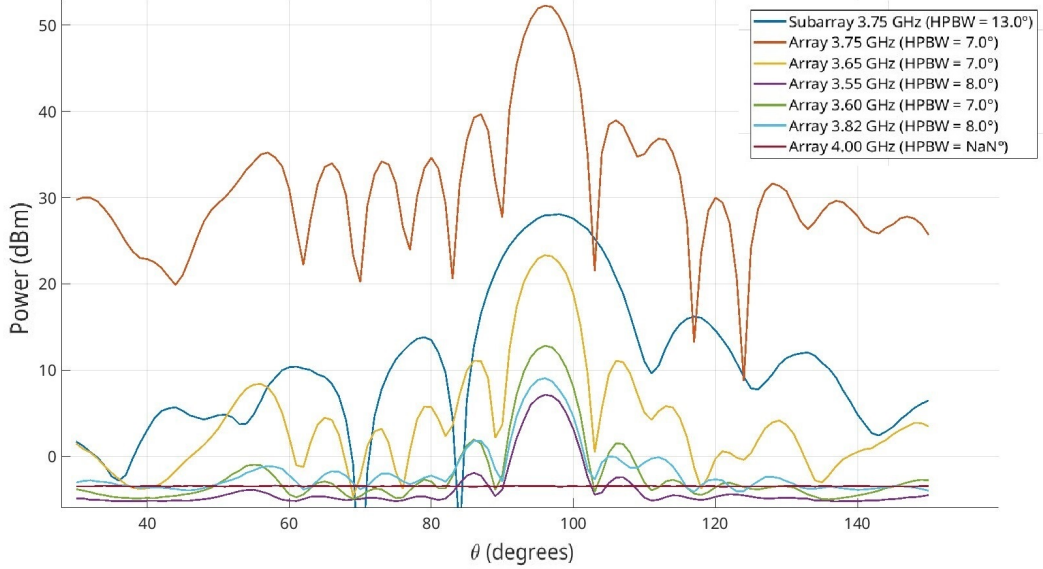


Figure 4.19: Comparison of the array radiation pattern with the measured subarray pattern and multiple OOB array beam pattern at different OOB frequencies without DPD

As shown in Figure 4.19, the spectral measurements reveal that the intermodulation products are not radiated isotropically. Instead, they are beamformed in the same direction as the in-band signal. The OOB radiation pattern of the full array at these distortion frequencies closely tracks the main array pattern, with a Half Power Beamwidth (HPBW) comparable to that at the carrier frequency (3.75GHz). This observation confirms that the distortion components generated by each antenna element combine coherently across the array and are projected into the same spatial beam as the desired transmission.

This behavior stands in sharp contrast to the previously discussed case of non-coherent OOB emissions, where the lack of phase alignment prevented constructive spatial addition. In that regime, the aggregate radiation resembled the response of a single subarray, with energy spreading more uniformly across angles and without strong directional projection. In the non linear case, however, the distortion products are inherently coupled to the beamforming process: since they are generated as nonlinear transformations of the in-band signal, they inherit its spatial characteristics and benefit from the same constructive combining across antennas.

The implications of this difference are significant. While non-coherent OOB emissions are naturally suppressed in massive MIMO systems due to spatial averaging, beamformed distortion products scale with array size in the same way as the desired signal. As a result, the interference caused by such distortions can be highly directive and potentially more harmful to receivers located along the main beam direction. This highlights the critical role of linearization techniques such as DPD in mitigating OOB distortion, especially in large-scale antenna arrays operating at high transmit power levels.

4.5 Conclusion

In this chapter, the analysis has shown two distinct behaviors of OOB emissions in massive MIMO antenna arrays. In the linear regime, OOB components lack spatial coherence and combine non-coherently, leading to natural suppression due to spatial averaging. In contrast, at elevated transmit power without DPD,

intermodulation distortion products emerge coherently and are beamformed in the same direction as the in-band signal, potentially causing harmful directional interference. This confirms that DPD is critical to suppress OOB directionality, while massive MIMO provides spatial filtering benefits only in the linear regime. Furthermore, the theoretical framework by Larsen and Van der Perre holds in the absence of DPD, and the observed phenomena generalize to any frequency band of interest, including 6G and beyond as explained in Appendix B.

Chapter 5

Actual Out Of Band emissions for multi-user Massive MIMO antennas

This chapter examines OOB emissions in 5G Massive MIMO systems in the context of multi-user MIMO with two simultaneously transmitted beams. Larson and Van der Perre's study[48][74], proved theoretically that in multi-user case, the intermodulation emissions are beamformed in directions that are different from the main signal, "Consequently, the distortion does add up constructively in distinct spatial directions, but these directions are different from those of the two intended receivers". For that, we will provide a theoretical and experimental analysis for the modeling of OOB emissions in MU MIMO using the 32TX/32RX n78-band antenna array to validate this behavior.

5.1 Introduction

In the previous chapter, we investigated OOB emissions in Massive MIMO systems for the single user case.

In this chapter, we extend this analysis to the multi-user Massive MIMO scenario. Although realistic deployments may serve many users simultaneously, for the sake of clarity, we focus on a two-user downlink transmission. This simplified setting allows us to examine how nonlinear distortion behaves when multiple spatial beams are formed and how OOB emissions are distributed across directions when several users are served concurrently. The aim is to determine whether the trends observed in the single-user case, coherent distortion buildup in the nonlinear regime, and incoherent radiation under linearized operation hold in the multi-user context as well.

The chapter is organized as follows:

Section 5.2 introduces the theoretical model we used to understand the spatial behavior of OOB emissions.

Section 5.3 describes the experimental setup, which includes the 32-transmit/32-receive (32TX/32RX) n78-band Massive MIMO antenna array that was used to emit 2 simultaneous beams in the azimuth plane inside the anechoic chamber.

Section 5.4 presents the results of our measurements. We compare OOB emissions under different operating conditions, both with and without DPD and analyze their spatial properties. We evaluate their impact on performance metrics such as ACLR for each spatial direction in the azimuth plane.

Section 5.5 summarizes the key takeaways from the study.

5.2 Theoretical framework

Throughout this chapter, we adopt the following notation conventions. Discrete-time indices and samples are denoted by n . Angular frequencies for UE $_i$ are denoted by ω_i , and carrier frequencies by f_i . The beamforming phases applied at antenna m for user l are denoted by ϕ_{ml} . We denote by k the polynomial order with K being the PA's maximum order. These conventions are used consistently throughout the remainder of the manuscript.

The theoretical framework remains the same as described in Section 4.2, with the extension that a second downlink transmission chain, denoted Tx $_2$, is activated and supports the same digital functionalities as the original DL RF chain. Taking into account the two-antenna model, the desired signal is formed by two sinusoids transmitted in two different spatial directions[48], each associated with a distinct receiver and characterized by delays τ_1 and τ_2 . Although the desired signal combines constructively in the intended directions, the distortion components may also add up constructively, but in spatial directions different from those of the intended receivers, as Larson and Van Der Perre states[48].

5.2.1 OOB emissions modeling

We consider a multi-user MIMO system with M transmit antennas and 2 UEs. The baseband signals for the two users are $\tilde{u}_1(n)$ and $\tilde{u}_2(n)$. Each antenna m applies beamforming phases ϕ_{m1}, ϕ_{m2} toward UE $_1$ and UE $_2$, respectively. The transmitted signal at antenna m is

$$\tilde{x}_m(n) = \tilde{u}_1(n)e^{j\phi_{m1}} + \tilde{u}_2(n)e^{j\phi_{m2}}. \quad (5.1)$$

Assuming a memoryless third-order polynomial power amplifier model, the output signal at antenna m can be written as, where $\tilde{x}_m(n)$ denotes the complex baseband input signal. For a multi-user transmission scenario, $\tilde{x}_m(n)$ is expressed as the superposition of the signals intended for each user, weighted by their respective beamforming phases:

$$\tilde{Y}_m(n) = a_1 \tilde{x}_m(n) + a_3 |\tilde{x}_m(n)|^2 \tilde{x}_m(n) + w_m(n), \quad (5.2)$$

For two users, the cubic term expands as

$$|\tilde{x}_m|^2 \tilde{x}_m = \left(|\tilde{u}_1|^2 \tilde{u}_1 + 2|\tilde{u}_2|^2 \tilde{u}_1 \right) e^{j\phi_{m1}} + \left(|\tilde{u}_2|^2 \tilde{u}_2 + 2|\tilde{u}_1|^2 \tilde{u}_2 \right) e^{j\phi_{m2}} \quad (5.3)$$

$$+ \tilde{u}_1^2 \tilde{u}_2^* e^{j(2\phi_{m1}-\phi_{m2})} + \tilde{u}_2^2 \tilde{u}_1^* e^{j(2\phi_{m2}-\phi_{m1})} \quad (5.4)$$

The detailed calculation is presented in Appendix C. Expanding the cubic term $|\tilde{x}_m|^2 \tilde{x}_m$ yields self-distortion terms, which preserve the original beamforming directions, as well as cross-terms corresponding to third-order intermodulation products. These cross-terms are proportional to combinations of the form $\tilde{u}_1^2 \tilde{u}_2^*$ and $\tilde{u}_2^2 \tilde{u}_1^*$, and carry effective spatial phase progressions given by linear combinations of the original beamforming phases.

The far-field transmitted baseband signal across M antennas can be expressed as the superposition of desired beamformed signals, nonlinear distortion components, and noise:

$$\tilde{Z}(n) = \underbrace{\tilde{Z}_{\text{UE}_1}(n) + \tilde{Z}_{\text{UE}_2}(n)}_{\text{Beamformed desired signals}} + \underbrace{\tilde{Z}_{\text{IM}}(n)}_{\text{Intermodulation distortion}} + \underbrace{w_T(n)}_{\text{Noise}}. \quad (5.5)$$

The individual components can be written as

$$\tilde{Z}_{\text{UE}_1}(n) = \tilde{u}_1 \sum_{m=1}^M [a_1 + a_3(|\tilde{u}_1|^2 + 2|\tilde{u}_2|^2)] e^{j\phi_{m1}}, \quad (5.6)$$

$$\tilde{Z}_{\text{UE}_2}(n) = \tilde{u}_2 \sum_{m=1}^M [a_1 + a_3(2|\tilde{u}_1|^2 + |\tilde{u}_2|^2)] e^{j\phi_{m2}}, \quad (5.7)$$

$$\tilde{Z}_{\text{IM}}(n) = \sum_{m=1}^M a_3 \left[\tilde{u}_1^2 \tilde{u}_2^* e^{j(2\phi_{m1} - \phi_{m2})} + \tilde{u}_2^2 \tilde{u}_1^* e^{j(2\phi_{m2} - \phi_{m1})} \right], \quad (5.8)$$

$$w_T(n) = \sum_{m=1}^M w_m(n) \sim \mathcal{N}(0, M\sigma^2). \quad (5.9)$$

Here, a_1 represents the linear gain of the power amplifier, while a_3 denotes the third-order nonlinear coefficient. This third-order term is responsible for generating the dominant nonlinear distortion components, including both beam-aligned distortion and intermodulation products that contribute to OOB emissions.

In the multi-user beamforming case, each antenna applies a superposition of beamforming weights corresponding to the different users. Consequently, the nonlinear distortion generated at each antenna inherits phase terms that are linear combinations of the beamforming phases of the individual users.

The desired signals combine coherently along their respective beamforming directions. The phase difference between two antenna ports m and l for the desired signals can be written as

$$\Delta\phi_{\text{UE},k} = \phi_{mk} - \phi_{lk}, \quad k = 1, 2. \quad (5.10)$$

Because the nonlinear terms inside $\tilde{Z}_{\text{UE}_1}(n)$ and $\tilde{Z}_{\text{UE}_2}(n)$ share the same per-antenna phase factors $e^{j\phi_{m1}}$ and $e^{j\phi_{m2}}$ as the desired signals, these third-order distortion components are also coherently beamformed toward the corresponding user directions. As a result, they experience the same array gain as the useful signals.

In contrast, the explicit intermodulation components contained in $\tilde{Z}_{\text{IM}}(n)$ exhibit different effective phase progressions. For a third-order nonlinearity, the dominant intermodulation terms involve the phase combinations

$$2\phi_{m1} - \phi_{m2} \quad \text{and} \quad 2\phi_{m2} - \phi_{m1}.$$

The corresponding phase differences between antenna elements m and l are therefore given by

$$\Delta\phi_{\text{IM}_1} = (2\phi_{m2} - \phi_{m1}) - (2\phi_{l2} - \phi_{l1}), \quad (5.11)$$

$$\Delta\phi_{\text{IM}_2} = (2\phi_{m1} - \phi_{m2}) - (2\phi_{l1} - \phi_{l2}). \quad (5.12)$$

These phase progressions differ from those of the desired beams and therefore produce distinct spatial radiation patterns. In terms of beam directions, if the desired beams toward UE_1 and UE_2 are steered toward angles θ_1 and θ_2 , the dominant third-order intermodulation beams appear approximately at [74]

$$\theta_{\text{IM},A} \approx \sin^{-1}(2 \sin \theta_1 - \sin \theta_2), \quad (5.13)$$

$$\theta_{\text{IM},B} \approx \sin^{-1}(2 \sin \theta_2 - \sin \theta_1). \quad (5.14)$$

We can summarize all that in the following expectations:

- Desired signals combine coherently along their respective user beam directions due to the applied beamforming phases.
- Third-order distortion components that inherit the same phase progression as a given user are also beamformed toward that user.
- The remaining intermodulation OOB components exhibit different phase progressions and therefore radiate along distinct spatial directions, forming independent distortion beams.

5.2.2 Discussion of Array Output, OOB Emissions, and Noise

The signals received for each UE can be expressed as a combination of useful in-band signal, OOB emissions, and noise. The useful signal is coherently amplified by the antenna array:

$$\begin{aligned}\tilde{Z}_{\text{UE}_1}(n) &= \tilde{u}_1 \left[a_1 + a_3(|\tilde{u}_1|^2 + 2|\tilde{u}_2|^2) \right] G_{\text{UE}_1} \\ \tilde{Z}_{\text{UE}_2}(n) &= \tilde{u}_2 \left[a_1 + a_3(2|\tilde{u}_1|^2 + |\tilde{u}_2|^2) \right] G_{\text{UE}_2}\end{aligned}\quad (5.15)$$

where

$$G_{\text{UE}_i} = \sum_{m=1}^M e^{j\phi_{mi}} \quad (5.16)$$

represents the antenna array gain in the direction of UE_i . The linear term G_{UE_i} is less than the maximum gain of the array in the direction of the optical field, reflecting the reduced off-axis beamforming efficiency. The main beam direction remains largely unchanged.

The OOB emissions arise from cross-products of the user signals due to amplifier nonlinearity:

$$\tilde{Z}_{\text{IM}}(n) = a_3 \left[\tilde{u}_1^2 \tilde{u}_2^* G_{\text{IM},1} + \tilde{u}_2^2 \tilde{u}_1^* G_{\text{IM},2} \right], \quad (5.17)$$

with

$$G_{\text{IM}_1} = \sum_{m=1}^M e^{j(2\phi_{m2} - \phi_{m1})} \quad G_{\text{IM}_2} = \sum_{m=1}^M e^{j(2\phi_{m1} - \phi_{m2})} \quad (5.18)$$

These nonlinear terms are generally steered away from the main beams, producing angular redistribution of distortion. The effective intermodulation gain depends on both the array geometry and the instantaneous signal amplitudes, indicating that distortion is not isotropic.

Finally, the aggregate noise contribution

$$w_T(n) = \sum_{m=1}^M w_m(n) \sim \mathcal{N}(0, M\sigma^2) \quad (5.19)$$

sums incoherently across antennas, scaling linearly with the number of elements M . Unlike the coherent signal and OOB gains, the noise remains isotropic and does not benefit from array beamforming.

5.3 Measurement setup

Figure 5.1 illustrates the experimental setup in top view used to carry out OTA measurements in a controlled radio environment. The objective of this campaign is to characterize the beamforming behavior of the Spear Andrew RU in a 5G NR multi-beam configuration. As before, all measurements are performed inside an anechoic chamber to suppress reflections and external interference, ensuring accurate radiation pattern evaluation. Measurements are performed under controlled laboratory conditions, assuming quasi-far-field behavior after near-field to far-field transformation

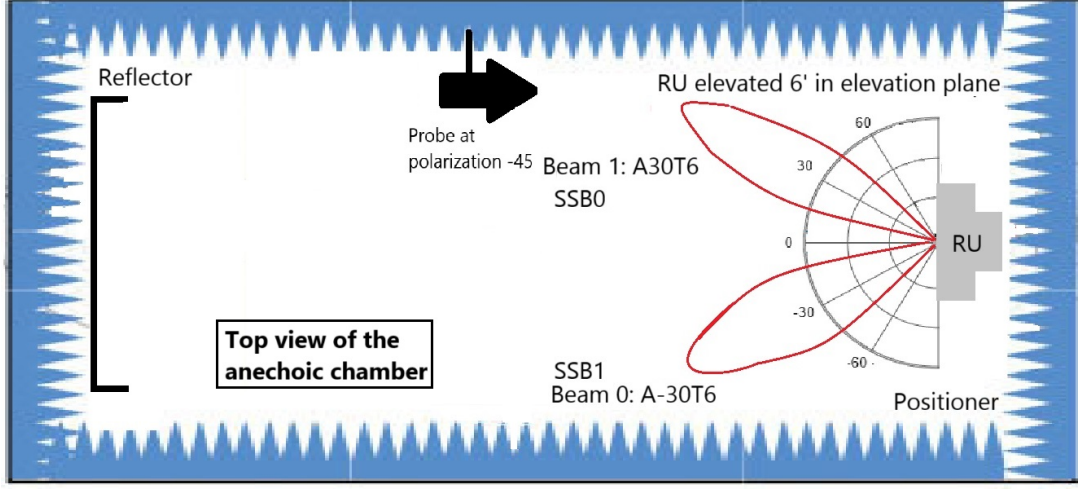


Figure 5.1: Measurement setup in the anechoic chamber in the azimuth plane

Unlike the previous single-beam campaign, the RU is now evaluated while steering two simultaneous beams. The antenna array remains mounted on a high-precision azimuth positioner, and all measurements are performed over an angular sweep of $\varphi \in [-60^\circ, 60^\circ]$. Beam 0 is oriented toward an azimuth of -30° with an elevation tilt of -6° , denoted A-30T6, and is associated with SSB 1. Beam 1 is directed towards $+30^\circ$ with the same elevation tilt, denoted A30T6, and is associated on SSB 0. This configuration enables the study of the azimuthal radiation characteristics of the RU under a multi-beam transmission of 2 users. To compensate for the downward tilt of the RU and obtain maximum directivity, the RU was elevated in the elevation plane of 6° so that both beams point directly into the elevation reference plane 0.

As in the previous setup, the radiated signal is directed toward a metallic reflector placed at a calibrated distance and orientation. The reflector redirects the incident energy to a receiving probe antenna rotated at -45° to receive the -45° polarization signal, which is aligned with the desired observation angle. The received signal is then analyzed using a high-resolution spectrum analyzer to assess beam shape, directionality, and relative power levels. The complete physical implementation is shown in Figure 4.8.

In this measurement campaign, the RU is operated in the PHY Test Mode with two downlink layers activated; therefore, the parameter $-L$ is fixed to 2. Furthermore, by enabling fixed beam-ID beamforming, the RU is configured to transmit using the specific beam identifiers corresponding to the two selected beams described above. These beam IDs remain fixed throughout the experiment during the entire azimuth sweep. For the 2 beam activation process, we refer back to subsection 3.3.1, we set 2 layers to active with corresponding beam ID from the DBT and we scale each beam to half, so that each layer gets allocated the transmit power equally by setting `cmp_scaling = "0x7"`.

5.4 OTA experimental results and analysis

In what follows, we present a comparative analysis of the maximum radiation levels measured with and without DPD, followed by an evaluation of coherent beamforming performance under two operating conditions: (i) when using DPD, and (ii) in the nonlinear regime without DPD for both spatial locations at $\varphi = -30^\circ$ and $\varphi = +30^\circ$. This comparison aims to highlight the impact of DPD on both the enhancement of radiated power and the coherence of the beam in realistic transmission scenarios. However, measurements are compensated for by the S_{21} parameter of the cable loss between the probe and the spectrum analyzer, as well as the path loss in the anechoic chamber. ACLR and SEM were measured OTA at the main lobes of two directional beams, $\varphi = -30^\circ$ and $\varphi = +30^\circ$, calibrated to directional EIRP for a single polarization.

Full TRP compliance is not assessed. We mention that we did not measure the spurious emissions as defined in the previous chapter, so the observed level in power curves in the spurious regions is noise power from the spectrum analyzer due to the limitation of its dynamicity in these regions.

5.4.1 Spectral power comparison with DPD

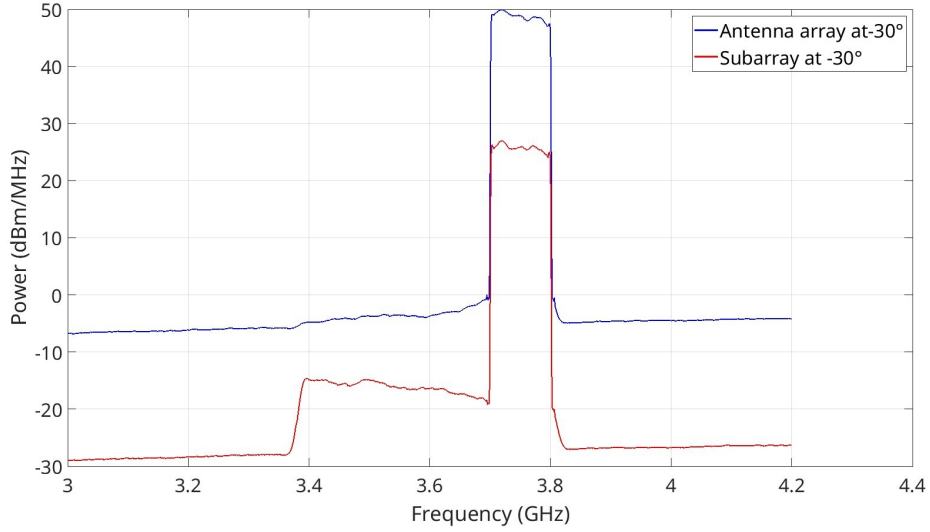


Figure 5.2: Comparison of the spectral power density of the full antenna array with one subarray at -30° with DPD calibrated to EIRP with one polarization

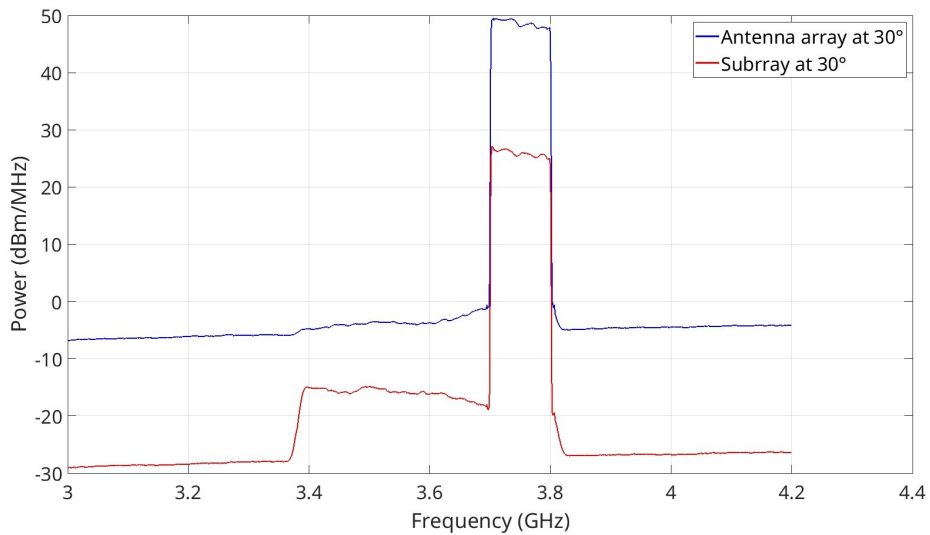


Figure 5.3: Comparison of the spectral power density of the full antenna array with one subarray at $+30^\circ$ with DPD calibrated to EIRP with one polarization

Figure 5.2 presents the measured spectral power for both a single subarray and the full Massive MIMO antenna array at an observation angle of -30° . The in-band signal benefits from coherent amplification across the antenna elements:

$$G_{UE_1} \approx 22 \text{ dB}, \quad (5.20)$$

which is slightly below the maximum array gain of 24 dB in the boresight direction for the single-user scenario. This shows that beamforming effectively concentrates energy toward the desired direction, improving signal strength compared to a single subarray and this represents the theoretical term $a_1 \tilde{u}_1 G_{UE_1}$ for UE_1 and $a_1 \tilde{u}_2 G_{UE_2}$ for UE_2 .

In contrast, the OOB emissions exhibit limited spatial coherence. The signals of individual subarrays are largely uncorrelated, resulting in minimal constructive interference across the array. Consequently, the measured full-array OOB pattern closely resembles that of a single subarray, showing an almost flat angular power distribution:

$$G_{OOB} \approx 11 \text{ dB at the band edge}, \quad (5.21)$$

which remains below the 12 dB level observed in the single-user case, and this represents the theoretical uncorrelated noise component derived theoretically. This confirms that nonlinear distortion is not coherently amplified by the array, in agreement with theoretical expectations.

The application of DPD provides spatial filtering of OOB emissions on the left side of the signal band, while the cavity filter mitigates emissions on the right side. As a result, the array maintains high in-band fidelity and meets the spectral emission requirements. Similar trends are observed in Figure 5.3 for the $+30^\circ$ direction, confirming the consistency of these effects across off-axis angles.

These experimental observations align with the theoretical framework presented in Section 5.2, validating the predicted relationships between coherent in-band gain, spatially distributed OOB emissions, and array geometry.

5.4.1.1 Correlation coefficient

We can now compute the correlation coefficient defined in subsubsection 4.2.5 for both directions. We repeat the same calculation, but we take $\varphi_{escan} = \varphi = -30^\circ$ for the first beam and $\varphi_{escan} = \varphi = 30^\circ$ for the 2nd beam to compute G , and due to symmetry, we get $G = 14.8485$ as calculated in details in Appendix C. By repeating the same calculation, we can now compute the correlation coefficient ρ for both directions:

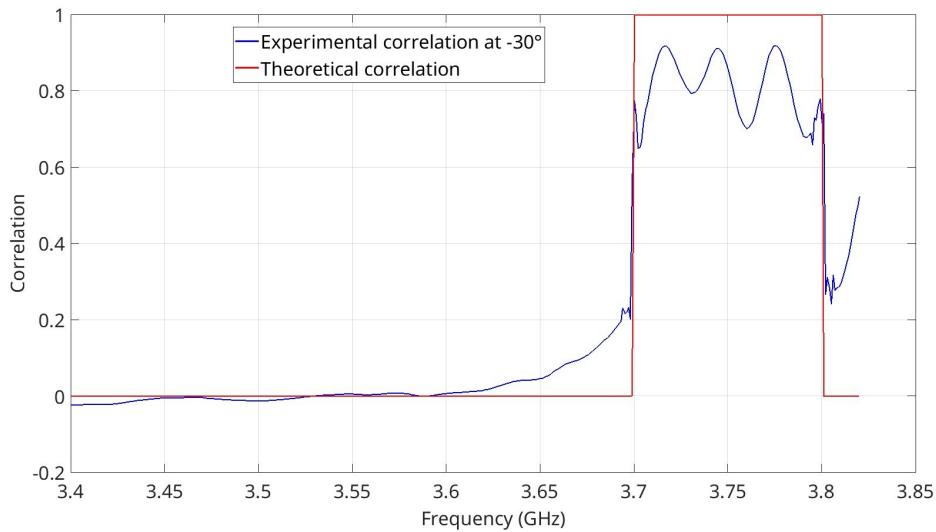


Figure 5.4: Correlation coefficient in function of the frequency at -30°

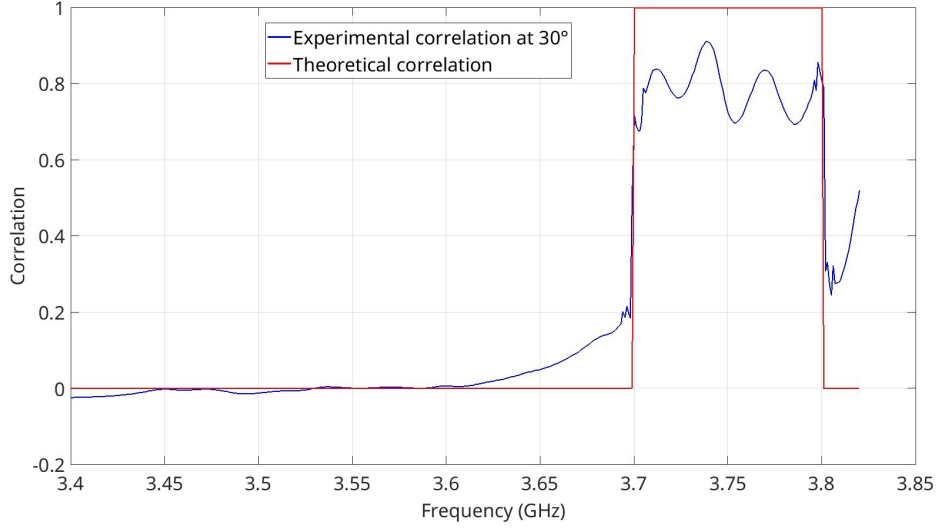


Figure 5.5: Correlation coefficient in function of the frequency at 30°

The correlation plot in Figure 5.4 provides an estimate of the experimental correlation coefficient defined in equation (4.18) for $\varphi = -30^\circ$. Within the useful band [3.7–3.8] GHz, the correlation coefficient remains relatively high, around $\rho \approx 0.8$, indicating that the distortion components retain a noticeable level of similarity across the antennas. This behavior is consistent with the expected theoretical trend. The decreasing behavior in the correlation level is proportional to the decreasing behavior of the direct gain, since getting far away from the bore sight direction reduces the correlation across the subarrays.

Outside the useful band, the correlation progressively decreases and eventually approaches zero, confirming that the remaining OOB components are largely decorrelated across antennas. The correlation level in the OOB region is similar to the correlation behavior in the boresight direction, this implies that the uncorrelated OOB emissions are independent from the main signal's direction, even while moving away from boresight direction. The same conclusion is interpreted for the $\varphi = 30^\circ$ as seen in Figure 5.5. Overall, the results support the same conclusion drawn previously: while DPD significantly reduces nonlinear distortion, some residual correlation persists in-band, and decorrelation dominates in the OOB region.

For a conventional array, the boresight direction is geometrically distinguished because all elements are equidistant from a far-field observation point along the array normal. When the elements are excited with identical phases, this geometry ensures phase coherence at boresight, yielding maximal constructive interference and, consequently, maximal array gain as seen in the single user case.

Radiation in directions away from boresight like $\varphi = -30^\circ$ and $\varphi = +30^\circ$ alters this geometric relationship. The unequal path lengths from different elements to an off-axis observation point introduce relative phase delays, leading to partial destructive interference and a reduction in the radiated field strength in that direction if no compensation is applied. Beam steering mitigates this effect by imposing deliberate phase or time delays across the array, restoring phase coherence in the desired direction and thereby translating the direction of maximum gain away from boresight.

Nevertheless, even under ideal steering, the achievable beamforming gain generally decreases as the scan angle increases. This reduction is fundamentally physical rather than algorithmic: the effective electromagnetic aperture of a planar array diminishes with scan angle due to projection effects, individual element radiation patterns typically attenuate at oblique angles, and power is increasingly redistributed into sidelobes. Consequently, while beamforming enables directional control of array gain, the maximum attainable gain is intrinsically constrained by array geometry and element characteristics, leading to

unavoidable scan loss when operating away from boresight, which is why the correlation coefficient and the beamforming gains in band are lower than in the boresight single user case, $G_{\text{in-band}} < 20\log(M)$ and $G_{\text{OOB}_i} < 10\log(M)$.

5.4.1.2 3GPP specifications conditions

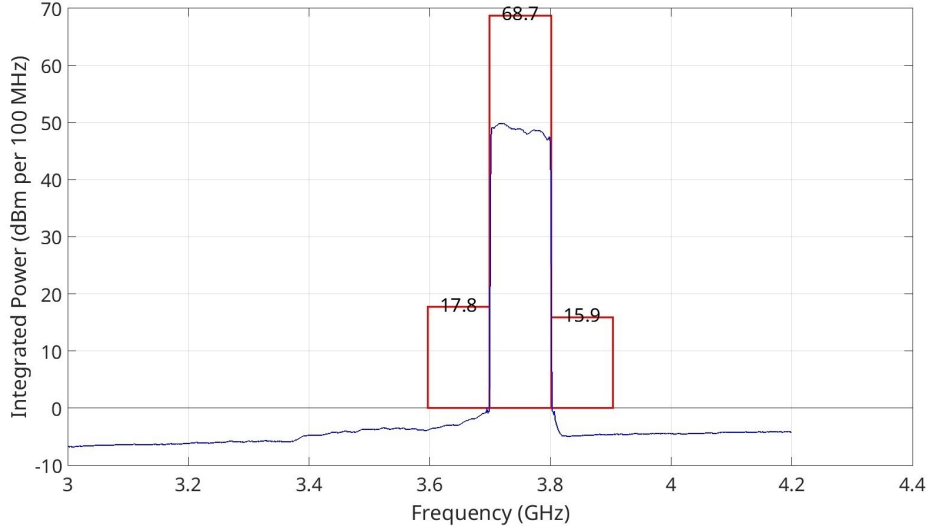


Figure 5.6: Integrated power per 100 MHz bandwidth for the antenna array beam pattern spectrum curve at -30°

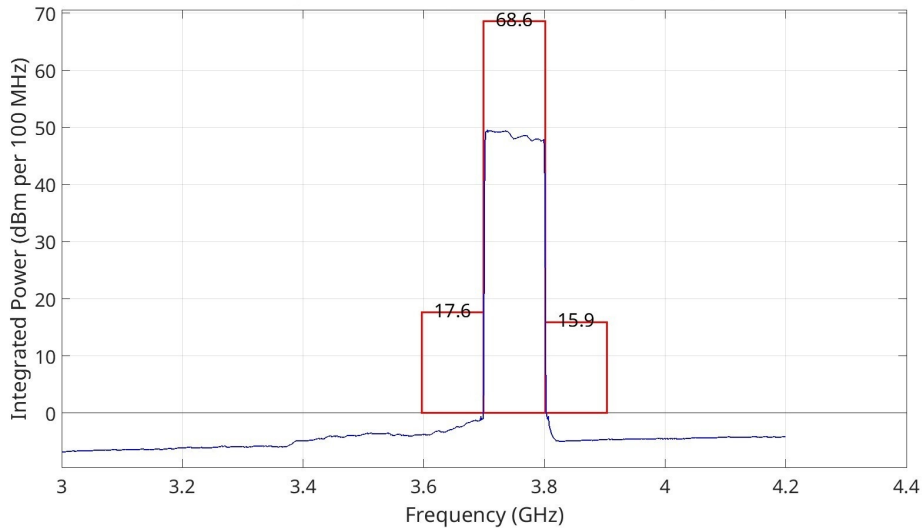


Figure 5.7: Integrated power per 100 MHz bandwidth for the antenna array beam pattern spectrum curve at $+30^\circ$

In Figure 5.6, the power integrated in a 100 MHz bandwidth within the allocated channel is 68.7 dBm, while the power integrated in the left adjacent channel is 17.8 dBm. These values correspond to EIRP-based measurements obtained in a given observation direction. The resulting ACLR, evaluated in the

adjacent channel with the highest measured power, is approximately 51 dBc, which exceeds the 45 dB requirement.

Similarly, in Figure 5.7, the integrated EIRP in the 100 MHz channel bandwidth is measured to be 68.6 dBm, while the corresponding EIRP adjacent channel on the left is 17.6 dBm. The resulting EIRP-based ACLR is again close to 51 dBc, providing a comfortable margin relative to the specification.

It is important to note that these results correspond to directional EIRP-based ACLR measurements, evaluated in specific angular directions, and therefore do not represent a full TRP-based ACLR assessment, which would require integration of the radiated power over the entire solid angle. Nevertheless, the consistently high EIRP-based ACLR values indicate excellent spectral confinement in the measured directions, suggesting strong transmitter linearity and effective suppression of adjacent-channel emissions at the beam level as described in 4.4.2.3.

5.4.2 Spectral power comparison without DPD

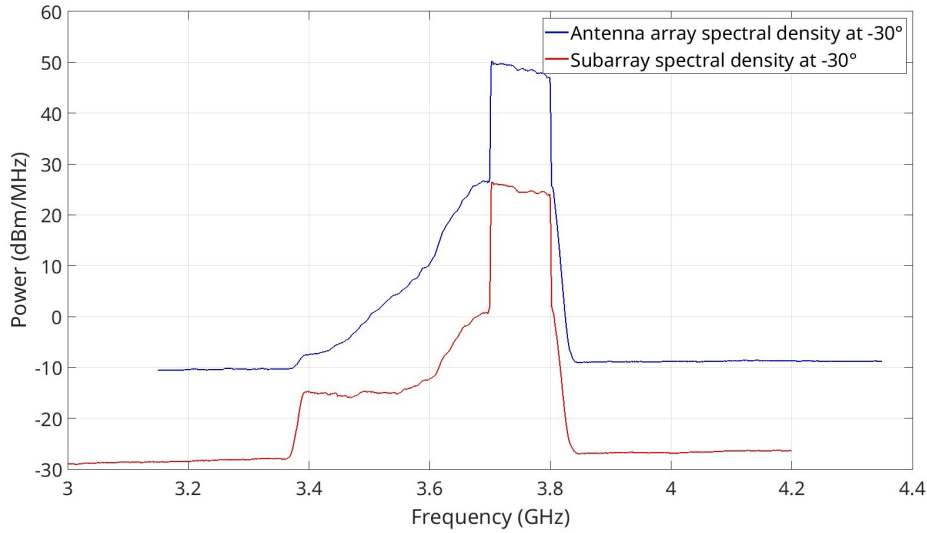


Figure 5.8: Comparison of the antenna array and subarray spectral power at -30° without DPD in the non linear PA regime

Disabling the DPD function results in significant nonlinear distortion, dominated by third- and fifth-order harmonics (H3 and H5). These intermodulation products extend into adjacent bands, causing unwanted spectral emissions. As shown in Figure 5.8 and 5.9, part of the distortion falls within the cavity filter passband (3.4–3.8 GHz), creating strong spectral shoulders and degrading in-band signal quality for both positions -30° and $+30^\circ$ and these emissions are represented by the theoretical components $\left[a_3(|\tilde{u}_1|^2 + 2|\tilde{u}_2|^2) \right] G_{UE_1}$ for UE_1 and $a_3(2|\tilde{u}_1|^2 + |\tilde{u}_2|^2) G_{UE_2}$ for UE_2 . The effect is more pronounced on the lower side of the band due to weaker filter attenuation, while the upper side is better suppressed.

5.4.3 Spatial behavior comparison with DPD

To verify that the measured antenna array pattern with the 2 activated beams simultaneously is correct, we compared it in Figure 5.10 with the array factor of the 2 beams separately using the model described in [56]. We clearly realize how it matches perfectly both beams in the 2 intended directions in the main lobes, and that it follows the same behavior and variation of the secondary lobes with some slight power level differences due to the inter-user interference caused by the side lobes constructive interference.

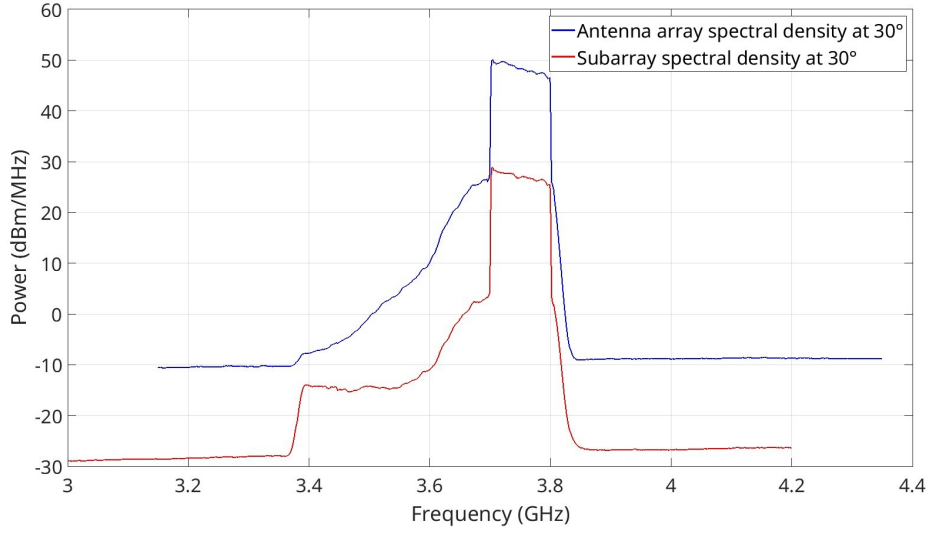


Figure 5.9: Comparison of the antenna array and subarray spectral power at $+30^\circ$ without DPD in the non linear PA regime

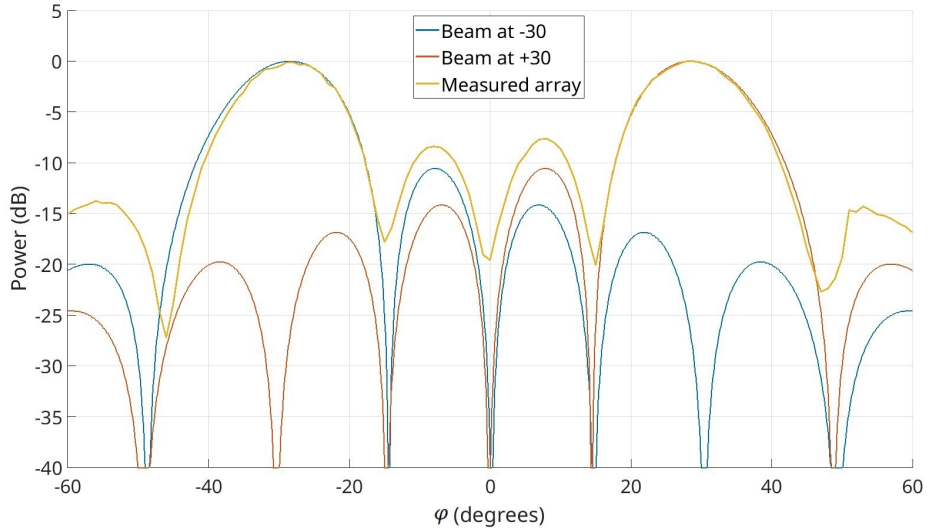


Figure 5.10: Comparison of the array radiation pattern with both beams activated with the theoretical beam pattern of both singular beams at -30° and 30°

The use of the full antenna array results in a substantial increase in beam gain, as shown in Figure 5.11 when comparing the full-array at both positions and subarray radiation patterns. This improvement demonstrates a significant enhancement in directional performance and is mainly due to the constructive superposition of signals from multiple antenna elements when DPD is enabled.

Furthermore, the full array exhibits a notable reduction in the number of sidelobes, which improves the suppression of undesired radiation in off-axis directions. This leads to higher beam efficiency and lower potential for interference, confirming the effectiveness of full-array beamforming in achieving the intended radiation characteristics.

The OOB radiation pattern of the entire array at different OOB frequencies closely matches that of a

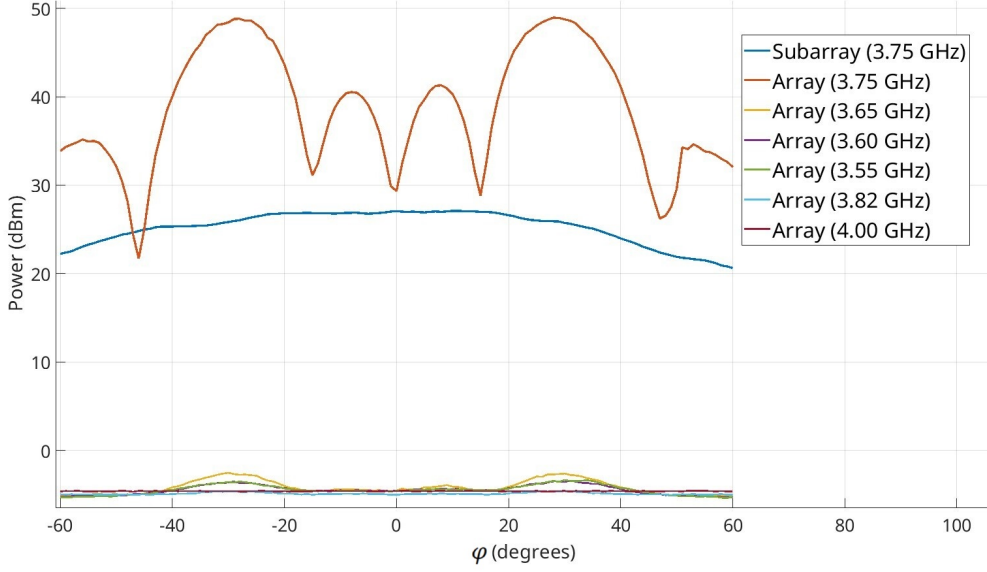


Figure 5.11: Comparison of the array radiation pattern with both beams activated with the measured subarray pattern and multiple OOB array beam pattern at different OOB frequencies with DPD

single subarray, as illustrated in Figure 4.17. For the in-band signal, the array beam pattern clearly shows two dominant main lobes at $+30^\circ$ and -30° , indicating strong directional focusing for the relative phase differences relative to the phase differences $\Delta\phi_{UE,1}$ and $\Delta\phi_{UE,2}$.

In contrast to in-band signals, which are coherently combined across antennas, the OOB components do not exhibit coherent behavior. These components mainly arise from nonlinearities and hardware imperfections, leading to random phase relationships across the array elements. As a result, the array behaves more like a collection of independent radiators in the OOB region.

This spatial incoherence has a beneficial effect by naturally suppressing OOB radiation in massive MIMO systems. Since the OOB emissions from different antennas add non-coherently, they do not concentrate in specific directions and do not scale with array size in the same manner as in-band signals, thereby reducing overall OOB interference.

5.4.4 Spatial behavior comparison without DPD

As shown in Figure 5.12, the spectral measurements for the two-beam transmission scenario reveal that the OOB distortion components are spatially beamformed and projected in the same directions as the corresponding in-band beams, namely at -30° and $+30^\circ$. This reciprocal propagation behavior originates from the nonlinear mixing of the two useful signals. The measured OOB radiation patterns of the entire array at the distortion frequencies closely follow the main array patterns associated with each beam.

From an analytical perspective, this observation is consistent with the signal decomposition, where a subset of the third-order nonlinear distortion terms appears within $\tilde{Z}_{UE_1}(n)$ and $\tilde{Z}_{UE_2}(n)$. Specifically, the distortion components proportional to $a_3\tilde{u}_1(|\tilde{u}_1|^2 + 2|\tilde{u}_2|^2)$ and $a_3\tilde{u}_2(2|\tilde{u}_1|^2 + |\tilde{u}_2|^2)$ share the same beamforming phase progressions $e^{j\phi_{m1}}$ and $e^{j\phi_{m2}}$ as the corresponding linear terms. As a result, these nonlinear distortion components are inherently beam-aligned and are radiated in the same directions as the intended transmissions toward UE_1 and UE_2 , respectively.

These results indicate that the distortion components generated by each antenna element combine coherently across the array for each beam and are simultaneously steered toward both beam directions.

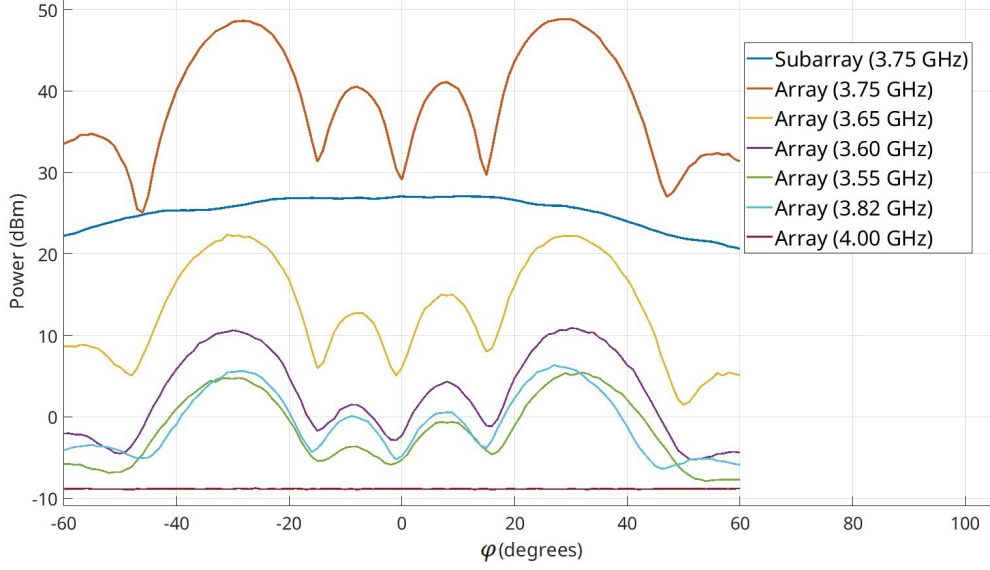


Figure 5.12: Comparison of the array radiation pattern with both beams activated with the measured subarray pattern and multiple OOB array beam pattern at different OOB frequencies without DPD in PA non linear regime

Consequently, the nonlinear distortion inherits the spatial characteristics of the underlying beamforming weights, leading to two distinct directive OOB lobes aligned with the desired in-band transmissions. Consequently, the present results do not contradict the findings of the reference work of Larson and Van Der Perre [48], but rather extend them by demonstrating that the spatial directivity of OOB emissions depends on the distortion mechanism.

In contrast, the cross intermodulation terms arising from the nonlinear mixing of the two users' signals do not, in general, share the same phase progression as either useful beam. As shown in the analytical development, the inter-element phase differences of these cross terms are proportional to those of the useful signals but scaled by frequency-dependent factors. Since these scaling factors differ from unity, the resulting phase progressions correspond to equivalent propagation delays that are distinct from those of both in-band beams. Consequently, these cross terms are beamformed toward different angular directions and do not align with the main lobes of the useful signals.

For the considered case with $\varphi_1 = 30^\circ$ and $\varphi_2 = -30^\circ$, the predicted intermodulation directions are approximately $\varphi_{IM,A} \approx 90^\circ$ and $\varphi_{IM,B} \approx -90^\circ$. Since the azimuth measurements were limited to the range from -60° to 60° , these directions fall outside the angular span covered in the measurement and therefore could not be observed experimentally.

After the submission of the manuscript, additional measurements were carried out using a more appropriate selection of beam angles to ensure that the predicted intermodulation directions fall within the measured angular range. These additional experiments were performed to validate the analytical predictions and confirm the beamforming behavior of the intermodulation products.

5.4.5 Enhanced spatial measurement

5.4.5.1 Measurement setup

To investigate OOB emissions under multi-user transmission, the RU is operated in a dual-beam configuration where two users are simultaneously served. The measurement focuses on the azimuthal radiation

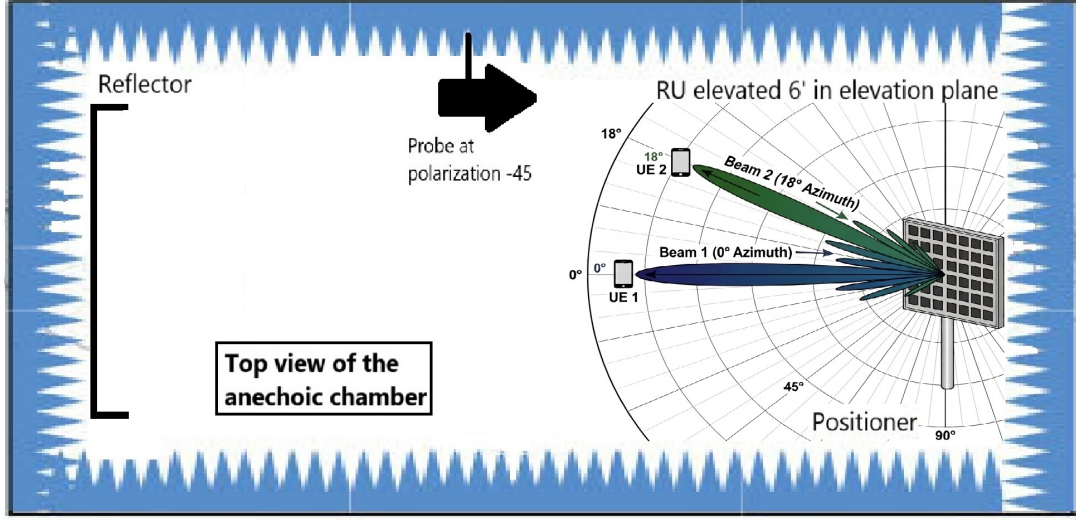


Figure 5.13: Top-view OTA measurement setup for dual-beam azimuth characterization.

characteristics while both beams are active.

In this configuration, Beam 1 is steered towards 0° azimuth and associated with SSB 1, denoted as A0T6. Beam 2 is directed towards $+18^\circ$ azimuth and associated with SSB 0, denoted as A18T6. The azimuth radiation patterns are measured over an angular range of $\varphi \in [-85^\circ, 85^\circ]$ as seen in Figure 5.13.

The RU operates in PHY Test Mode with two active downlink layers ($-L = 2$). Fixed beam-ID beamforming is enabled so that the two predefined beam directions remain constant throughout the measurement. The transmit power is equally distributed between the two beams to ensure identical power allocation for both users during the azimuth sweep.

5.4.5.2 Theoretical expectations

In the considered scenario, two users are in Line-of-Sight (LoS) with main beam directions:

$$\varphi_1 = 0^\circ, \quad \varphi_2 = 18^\circ,$$

and the approximation $f_1 \approx f_2$ is assumed.

According to Van Der Perre's modeling, the intermodulation beam angles are given by

$$\varphi_{IM,1} \approx \sin^{-1}(2 \sin \varphi_1 - \sin \varphi_2),$$

$$\varphi_{IM,2} \approx \sin^{-1}(2 \sin \varphi_2 - \sin \varphi_1).$$

For this configuration, the predicted intermodulation directions are

$$\varphi_{IM,1} \approx -18^\circ, \quad \varphi_{IM,2} \approx 38.2^\circ.$$

Based on the analytical model, the intermodulation products are expected to appear at angles distinct from the main in-band beams. Each cross-term beam is predicted to steer away from φ_1 and φ_2 due to the frequency-dependent scaling of the inter-element phase differences. Consequently, these intermodulation beams should be observable at the predicted angles $\varphi_{IM,1}$ and $\varphi_{IM,2}$, with lower relative power than the main beams but following the angular positions predicted by the model.

5.4.5.3 Spatial behavior with DPD

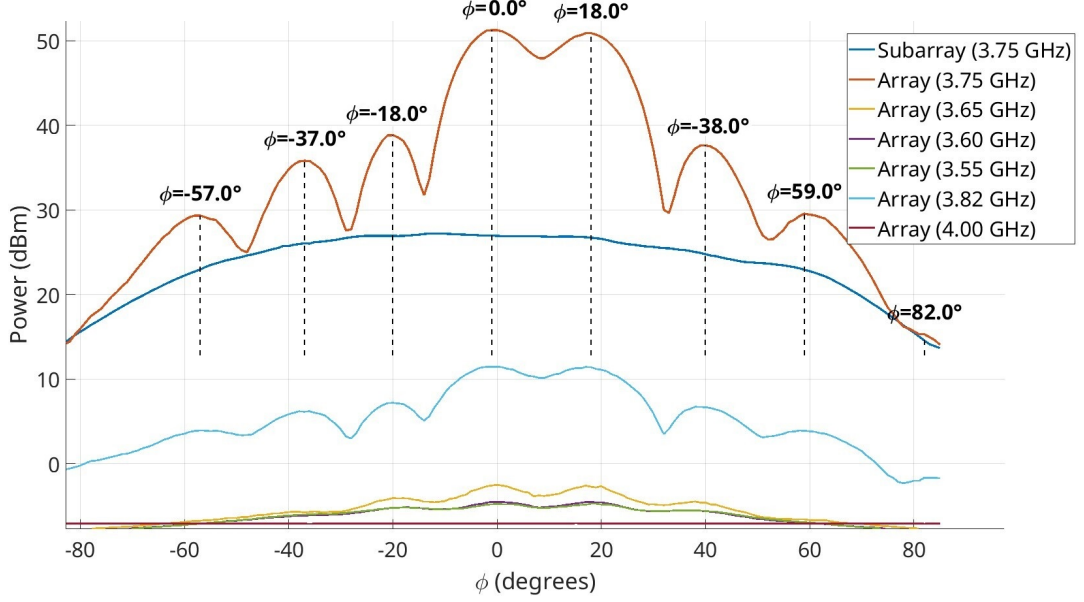


Figure 5.14: Comparison of the array radiation pattern with both beams activated with the measured subarray pattern and multiple OOB array beam pattern at different OOB frequencies with DPD

When DPD is applied to the RU, the in-band signals are coherently beamformed, resulting in two dominant main lobes at $\phi = 0^\circ$ and $\phi = +18^\circ$. These angles correspond to the intended directions of the users and clearly indicate strong directional focusing due to the constructive combination of signals across all antenna elements.

Third-order intermodulation products are observed at angles corresponding to the theoretical predictions, approximately $\phi_{IM,1} \approx -18^\circ$ and $\phi_{IM,2} \approx +38^\circ$. These intermodulation beams are spatially separated from the main lobes and appear at lower relative power compared to the in-band signals as shown in Figure 5.14 that are relative to the theoretical terms $\tilde{Z}_{IM}(n) = a_3 [\tilde{u}_1^2 \tilde{u}_2^* G_{IM,1} + \tilde{u}_2^2 \tilde{u}_1^* G_{IM,2}]$.

The OOB emissions are broadly distributed and spatially decorrelated. Unlike the coherently combined in-band signals, the OOB components do not exhibit constructive interference, as they originate from nonlinearities and hardware imperfections. This spatial incoherence naturally suppresses OOB radiation in massive MIMO systems, preventing concentration in specific directions and reducing the potential for interference.

5.4.5.4 Spatial behavior without DPD

When the power amplifiers operate in the nonlinear regime without DPD, intermodulation distortion is generated across the array. The OOB emissions inherit the beamforming phase of the in-band signals, forming directive lobes aligned with the main beam directions, approximately at $\phi = 0^\circ$ and $\phi = +18^\circ$.

These distortion components combine coherently across the full antenna array, producing unintended array gain. Consequently, the nonlinear distortion exhibits strong directivity and mirrors the spatial characteristics of the intended in-band beams.

Third-order intermodulation products appear at angles distinct from the main beams, approximately $\phi_{IM,1} \approx -18^\circ$ and $\phi_{IM,2} \approx 38.2^\circ$ as seen in Figure 5.15. These cross terms do not share the same phase progression as the in-band signals and are steered toward different angular directions. In the initial

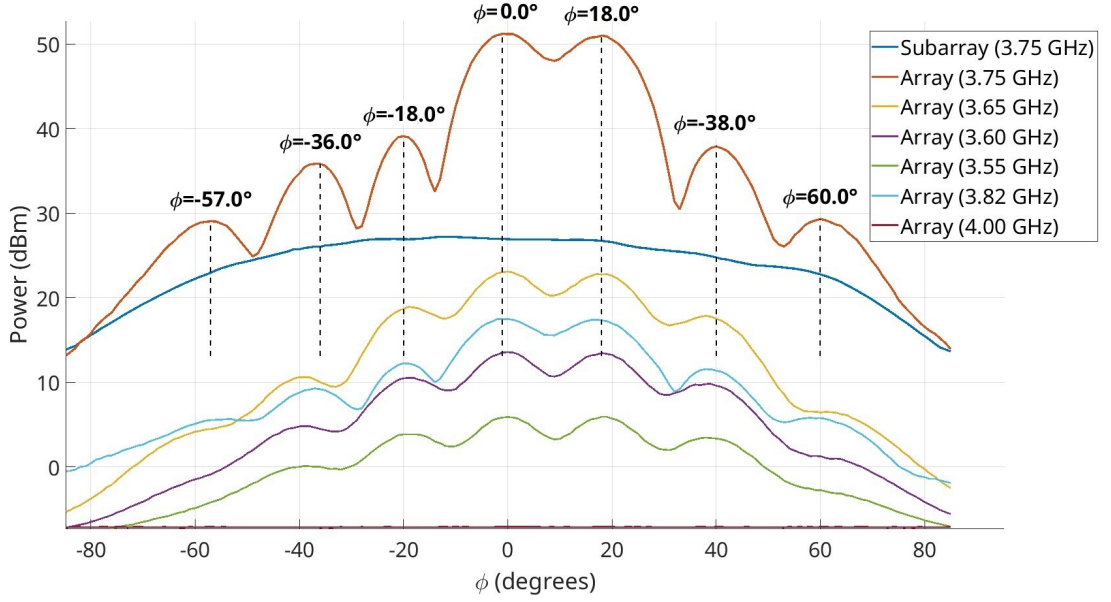


Figure 5.15: Comparison of the array radiation pattern with both beams activated with the measured subarray pattern and multiple OOB array beam pattern at different OOB frequencies without DPD in PA non linear regime

measurements, the azimuth scan did not fully capture these intermodulation directions. Additional measurements with an extended azimuth span were conducted to validate these predictions and confirm the array beamforming behavior of the intermodulation products.

5.5 Conclusion

In this chapter, the analysis has revealed two distinct behaviors of OOB emissions in massive MIMO antenna arrays under multi-beam operation. In the linear regime, OOB components associated with multiple beams lack spatial coherence and therefore combine non-coherently across the array, resulting in natural suppression through spatial averaging. In contrast, at elevated transmit power levels and in the absence of DPD, intermodulation distortion products become spatially coherent and are simultaneously beamformed in the same directions as in-band signals. As a consequence, the resulting OOB radiation exhibits pronounced directional lobes aligned with the active beams, potentially leading to harmful and beam-directed interference.

These results confirm that DPD is essential for suppressing the spatial directionality of OOB emissions in massive MIMO systems, while the inherent spatial filtering benefits of large antenna arrays are effective only in the linear operating regime. Furthermore, the experimental observations are consistent with the theoretical framework proposed by Larsen and Van Der Perre[48][74] in the absence of DPD for single user MIMO. Consequently, the results presented in this chapter should be interpreted as complementary to the findings of the reference work by Larsen and Van Der Perre[48][74] for multi-user case, indicating that the spatial directivity of OOB emissions is influenced by the underlying distortion mechanisms and system architecture.

Chapter 6

Conclusions and Future Perspectives

6.1 Conclusions

This thesis adopts a solution approach grounded in the constraints and capabilities defined by the O-RAN specifications, with a focus on end-to-end 5G transmission in massive MIMO systems. Split 7.2 establishes a precise functional boundary between the DU and RU, specifying where multi-antenna processing occurs, how IQ samples and beamforming parameters are transported, and the phase-coherence requirements across large antenna arrays. In massive MIMO, where spatially coordinated transmission across tens or hundreds of antennas is essential, the fronthaul interface becomes both a potential bottleneck and a critical enabler. By implementing a fully compliant Split 7.2 chain using the OpenAirInterface (OAI) framework, this work evaluates massive MIMO under realistic timing, signaling, and interoperability constraints.

The architecture is deployed end-to-end on Eurecom's testbed, interfaced with a large-array RU prototype, providing a standard-aligned environment to study Multi-user MIMO and hardware impairments throughout the entire 5G physical layer. The hardware and software setup, including the compilation and deployment of the O-RAN FH library, PTP-based synchronization, and optimized resource allocation, ensures reliable, standards-compliant DU–RU communication. Transmission, synchronization, and IQ data-flow tests confirm the stability of the system and validate the functional behaviors of the massive MIMO deployment that is expected of a Split 7.2.

The platform enables controlled investigation of OOB emissions in massive MIMO systems. Analysis reveals two operating regimes: in the linear regime, OOB components combine non-coherently across the array, yielding natural spatial suppression; at elevated transmit powers without DPD, nonlinear intermodulation products combine coherently, beamforming in the same direction as the in-band signal and potentially causing harmful directional interference. These effects persist under multi-beam MU-MIMO transmission, confirming that DPD is essential to mitigate directional OOB emissions. Observations can be generalized across frequency bands relevant to future 6G deployments.

A summary of the contributions in each chapter is as follows:

- In Chapter 2, we introduced OAI, including its components and operating modes that are beneficial to developers and researchers in prototyping. We presented its ongoing implementation of Massive MIMO transmission and beamforming procedure. We highlighted the limitations of these implementations and proposed different enhancements in terms of precoding schemes and beam management procedures.
- In Chapter 3, we explained the implementation of the ORAN split 7.2 interface to disaggregate the functionalities between DU and RU. We presented a tutorial that was used to integrate the ORAN FH interface into the Massive MIMO RU, ensuring timing synchronization and appropriate IQ samples transmission and system functionality.

- In Chapter 4, we can conclude that the experimental setup allows a detailed investigation of OOB emissions in massive MIMO systems. In the linear regime, OOB components combine non-coherently across antennas, yielding natural spatial suppression. At elevated transmit powers without DPD, nonlinear intermodulation products emerge coherently, beamforming in the same direction as the intended signal. This effect can generate harmful directional interference and violates spectral constraints. The study confirms that DPD is essential for controlling directional OOB emissions. The contribution provides quantitative insights into the interaction between the array size, transmit power, and spectral behavior in massive MIMO systems.
- In Chapter 5, The analysis is extended to multi-user, two-beam MU-MIMO transmission, showing that OOB behaviors persist in multi-beam scenarios. Coherent nonlinear distortion still aligns with in-band beams at high transmit powers, while linear-regime spatial averaging continues to provide suppression. This demonstrates that multi-beam scaling does not fundamentally change the underlying emission characteristics. The study provides practical guidance for designing multi-user massive MIMO systems under fronthaul and regulatory constraints. It confirms the generality of the findings across single and multi-beam setups. The contribution lies in extending OOB emission insights to realistic multi-beam deployments relevant for 5G and beyond networks.

We mention that the findings concluded in chapters 4 and 5, are relevant for straightwire precoding for model simplicity reasons, and we have an interest to investigate the spatial and frequency behavior of the OOB emissions in Massive MIMO systems for other precoding schemes such as NMSE, ZF, MRT and Z3RO precoders.

Growing industrial interests from Verizon: The following contributors in the industry reached out for questions, clarifications, and proposals for contributions based on the findings of the published paper. Verizon aims to study the Assessment of C-Band Mobile Telecommunications Interference Impact on Low Range Radar Altimeter Operations[75].

OOB emissions study by NOKIA and Agence Nationale des Fréquences (ANFR): This work was a reference for OOB emissions studies that were held by NOKIA and ANFR under the title 'C-band AAS antenna pattern/gain OOB Lab tests - ITU-R (WP5D) contributions on beamforming measurements'.

Beyond its academic contributions, this work provides practical and actionable insights for regulators, mobile network operators, and other telecommunications stakeholders involved in the deployment and operation of massive MIMO systems. In particular, the proposed framework contributes to a clearer understanding of OOB emissions in realistic multi-antenna transmission scenarios.

For mobile operators, the results of this thesis support the interpretation and management of OOB emissions under regulatory radiated emission measurements, allowing a more accurate assessment of compliance margins in operational networks. By explicitly accounting for beamforming, spatial power distribution, and user-dependent transmission conditions, the proposed methodology helps operators identify scenarios in which OOB emissions may become spatially concentrated or, conversely, mitigated through antenna array effects. This knowledge is essential to ensure compliance with regulatory limits while maintaining efficient spectrum utilization.

From a network performance and Qos perspective, the framework also provides insight into the relationship between OOB emissions, adjacent-channel interference, and system level performance. Understanding how OOB emissions evolve with traffic load, beamforming strategies, and antenna configurations allows operators to better anticipate potential interference impacts on neighboring bands and services, thus supporting more robust network planning and optimization decisions.

For regulators and standardization bodies, this work offers a technical basis for the evolution of radiated emission assessment methodologies tailored to massive MIMO and other advanced antenna systems. The proposed framework highlights limitations of existing measurement paradigms and suggests directions for measurement procedures that better reflect realistic network operation, without compromising the enforceability and reproducibility required for regulatory compliance.

6.2 Future Perspectives

The future prospects of our work in experimental analysis of Massive MIMO in antenna systems:

6.2.1 Massive MIMO 5G system in standalone mode

Performing massive MIMO experiments in SA mode significantly alters the system context compared to PHY-Test mode due to the activation of the full 5G protocol stack. In SA mode, the gNB operates in conjunction with a core network, either the OAI core or an external 5G Core enabling real end-to-end UE connections. Consequently, the system behavior is no longer restricted to PHY-layer transmissions but also includes MAC scheduling, RLC/PDCP processing, bearer management, and signaling interactions. This integration allows for the evaluation of massive MIMO beamforming in a fully operational 5G environment, where precoding and beam selection can be influenced by actual UE feedback, including CSI reports, mobility patterns, and traffic demands. Unlike PHY-Test mode, which typically relies on fixed beams and pre-configured transmission patterns, SA mode permits dynamic adaptation of spatial streams in response to real-time UE measurements, thereby enabling a more realistic assessment of beamforming efficiency, multi-user interference management, and spatial multiplexing capabilities.

In addition, the end-to-end operation of the SA mode introduces higher-layer effects that impact the observable performance of massive MIMO. For example, Hybrid Automatic Repeat Request (HARQ) retransmissions, MAC-layer scheduling delays, and core-network processing contribute to the overall latency and throughput experienced by the UE, which must be considered along with PHY-layer metrics such as SINR, beam orthogonality, and spatial correlation. While the PHY-test mode provides precise information on isolated beamforming and signal propagation characteristics, the SA mode captures the interaction between PHY performance and system-level constraints, offering a comprehensive view of network behavior. Moreover, OOB emission patterns, beamforming accuracy, and precoder efficiency are now affected not only by array geometry and transmit power but also by the timing and scheduling decisions made at the DU and core network, as well as feedback granularity from UEs.

6.2.2 Extension for OOB and band coexistence behavior

In this work, OOB emissions were analyzed for both single-user and multi-user MIMO transmission scenarios. Although the present study provides information on the spectral and spatial characteristics of OOB radiation under these conditions, further investigations are required to fully characterize its spatial behavior in more general and dynamic settings. Future work will therefore focus on an in-depth analysis of OOB spatial propagation and beamforming effects, with the objective of improving DPD through joint software and hardware optimization. Such efforts aim to ensure compliance with 3GPP regulatory requirements under realistic deployment conditions, including diverse propagation environments and multi-user scenarios where inter-user interference and coexistence constraints play a critical role.

6.2.3 Conduct TRP studies to meet 3GPP regulations SEM and ACLR

OOB emissions were experimentally investigated for both single-user and multi-user MIMO transmission scenarios, with the measurements conducted in terms of EIRP. Although EIRP-based characterization provides valuable information on directional OOB radiation and worst-case emission levels, regulatory compliance in practical deployments is increasingly assessed using TRP, which accounts for the aggregate radiated energy over the full spatial domain. Consequently, future work will focus on the estimation and modeling of OOB emissions in terms of TRP, including accurate spatial integration of nonlinear radiation patterns. This extension will enable a more comprehensive evaluation of the SEM and ACLR requirements defined by the 3GPP standards. Furthermore, joint software–hardware optimization of DPD

techniques will be investigated to mitigate OOB emissions under realistic multi-user and coexistence scenarios, where inter-user interference, beam dynamics, and spatial power distribution play a critical role for TRP-based measurements.

6.2.4 Extending the work towards 6G

The framework developed in this thesis can be extended to 6G by addressing higher frequency bands, such as mmWave and THz, which require ultra-narrow beams and precise phase synchronization in massive MIMO arrays. AI-driven and RIC-assisted beam management can replace fixed-beam strategies, enabling real-time adaptation, predictive CSI usage, and enhanced multi-user precoding. End-to-end evaluation of latency, reliability, and OOB emissions becomes critical to support 6G use cases like holographic communications and industrial automation. Additionally, the platform can be adapted to study cell-free massive MIMO and distributed antenna systems, investigating cooperative beamforming, dynamic resource allocation, and multi-beam interference management. Overall, the current O-RAN compliant massive MIMO testbed provides a scalable and standards-aligned foundation for prototyping and evaluating 6G networks.

Appendix A

Spatial Upper Bound of Radiated Power in Active Antenna Systems

A.1 Introduction

5G BS increasingly rely on Massive MIMO technology combined with AAS and digital beamforming to achieve high spectral efficiency and flexible spatial multiplexing. Although these architectures enable highly directive in-band transmissions, they also raise important questions about the spatial behavior of unwanted radiated emissions, particularly in adjacent-band coexistence scenarios [48] [76] with safety-critical systems such as aeronautical radio altimeters. Conventional interference assessment methods, originally developed for passive or weakly directive antennas, are not well suited to dynamically beamformed arrays and may lead to overly complex or insufficiently conservative evaluations. This appendix addresses a fundamental regulatory and engineering question: Can a simple and deterministic upper bound be established for the maximum radiated emission in any spatial direction, independently of beamforming configuration, signal characteristics, or frequency? To answer this question, we develop an analytical framework showing that the maximum radiated emission always occurs at boresight, thereby defining a spatial envelope whose angular dependence is determined solely by the element or sub-array radiation pattern. The methodology first considers a two-element antenna array to derive the spatial upper bound in an intuitive manner, and then extends the result to realistic Massive MIMO AAS architectures. The theoretical analysis is further supported by over-the-air measurements performed on a 3.75 GHz Massive MIMO antenna, providing experimental validation of the proposed spatial bound.

A.2 Theoretical framework

A.2.1 Single user transmission case

A.2.1.1 Two-Element Array Model

To introduce the concept of a spatial upper bound in a simple and intuitive manner, we first consider the elementary case of a two-element active antenna array.

The array is composed of two identical radiating elements positioned along the y -axis and separated by an inter-element distance $d = \frac{\lambda}{2}$, where λ denotes the wavelength at the carrier frequency. A schematic representation of the configuration considered is illustrated in Fig. A.1. This half-wavelength spacing is widely adopted in practical antenna arrays as it provides an appropriate compromise between spatial sampling and physical compactness, while preventing the appearance of grating lobes within the visible angular region.

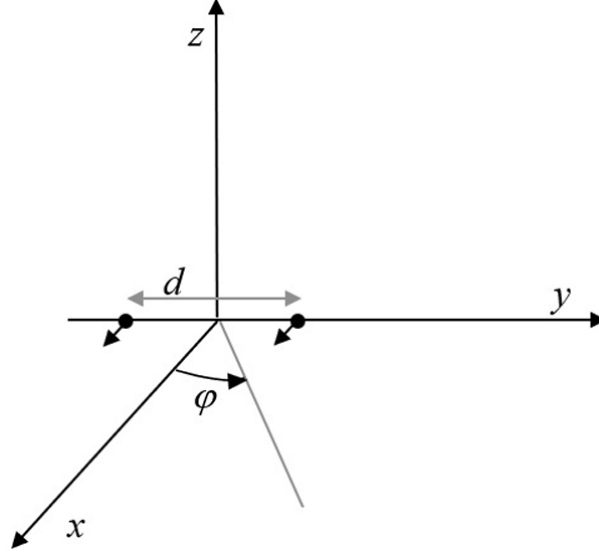


Figure A.1: 2 element antenna array aligned at the y axis

In this analysis, mutual coupling between antenna elements is neglected. This assumption is reasonable for arrays with $\frac{\lambda}{2}$ spacing and well-designed antenna structures, where the inter-element coupling is typically lower than -15 dB in practical Massive MIMO implementations. Furthermore, ideal amplitude and phase control of excitation signals is assumed[57]. This assumption is consistent with modern digital beamforming architectures, where calibration procedures enable accurate control of the excitation coefficients applied to each antenna branch.

The analysis is restricted to the horizontal plane defined by $\theta = 90^\circ$. This configuration is particularly relevant for terrestrial coexistence scenarios involving ground-based systems such as 5G base stations and aeronautical radio altimeters operating in adjacent frequency bands. Radiation characteristics are therefore studied as a function of the azimuth angle $\varphi \in [0, 2\pi]$.

The radiation pattern of the elementary radiator follows the 3GPP antenna modeling framework[58]. In the horizontal plane, the element gain is expressed as:

$$A_E(\varphi) = G_{E,\max} + A(\varphi), \quad (\text{A.1})$$

where $G_{E,\max}$ denotes the peak gain of the antenna element and $A(\varphi)$ represents the angular attenuation function defined as

$$A(\varphi) = -\min\left(12\left(\frac{\varphi}{\varphi_{3\text{dB}}}\right)^2, A_m\right), \quad (\text{A.2})$$

where A_m denotes the front-to-back ratio and $\varphi_{3\text{dB}}$ is the horizontal half-power beamwidth.

For a coherent frequency component, the radiation pattern of the array in the horizontal plane can be written as:

$$G(\varphi, \Delta\Phi) = A_E(\varphi) + 20\log_{10}|AF(\varphi, \Delta\Phi)|, \quad (\text{A.3})$$

where $\Delta\Phi$ denotes the relative excitation phase between the two radiating elements and $AF(\varphi, \Delta\Phi)$ is the array factor. For the considered geometry, the array factor can be expressed as:

$$AF(\varphi, \Delta\Phi) = 1 + e^{j(\psi(\varphi) + \Delta\Phi)}, \quad (\text{A.4})$$

where $\psi(\varphi)$ represents the geometrical phase difference arising from the propagation path difference between the two elements in the observation direction φ . For an inter-element spacing d and wave number $k = \frac{2\pi}{\lambda}$, this phase term is given by:

$$\psi(\varphi) = kd \sin \varphi. \quad (\text{A.5})$$

The magnitude of the array factor therefore becomes

$$|AF(\varphi, \Delta\Phi)| = 2 \left| \cos \left(\frac{\pi d}{\lambda} \sin \varphi + \frac{\Delta\Phi}{2} \right) \right|. \quad (\text{A.6})$$

We now examine the maximum radiated power that can be achieved in any spatial direction when the beamforming phase is varied. In practical beamforming systems, the relative phase $\Delta\Phi$ between antenna branches can be continuously adjusted in order to steer the main beam toward different directions. From an interference assessment perspective, it is therefore important to determine the maximum radiated power that can occur in each direction when $\Delta\Phi$ is allowed to vary over its full range.

When the excitation phase is swept over the interval $\Delta\Phi \in [0, 2\pi]$, the cosine term in the array factor spans its entire range. Since the absolute cosine function $|\cos(\cdot)|$ reaches a maximum value of unity, the maximum achievable magnitude of the array factor in any spatial direction satisfies

$$\max_{\Delta\Phi} |AF(\varphi, \Delta\Phi)| = 2, \quad \forall \varphi. \quad (\text{A.7})$$

This result leads directly to a spatial envelope describing the maximum possible radiation level that can occur in each azimuth direction when the beamforming phase is varied. The envelope of the array radiation pattern can therefore be expressed as

$$G_{\text{env}}(\varphi) = A_E(\varphi) + 20 \log_{10}(2) = A_E(\varphi) + 6 \text{ dB}. \quad (\text{A.8})$$

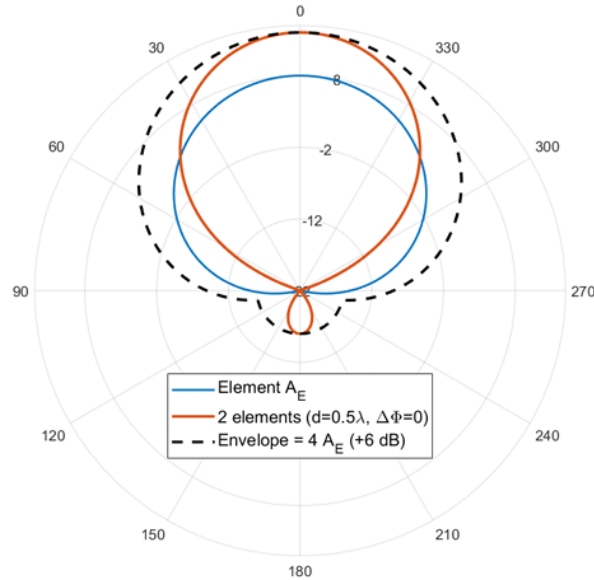


Figure A.2: Two-element array pattern envelope compared with a single element radiation pattern

The resulting spatial envelope is illustrated in Fig. A.2. This result provides an important insight: although beamforming allows the main lobe of the array to be steered toward different directions, the maximum radiated power achievable in each direction remains fundamentally limited by the radiation

characteristics of the elementary antenna element. Consequently, the envelope of the array radiation pattern is independent of the beamforming phase and is entirely governed by the element radiation pattern.

A.2.1.2 Spatial upper bound for radiated power

We now analyze the spatial distribution of the radiated power resulting from the two-element array model introduced previously. Using the third-order nonlinear signal model, three distinct spectral regions can be identified depending on the dominant contribution to the radiated signal. These regions correspond to: (i) the useful-signal-dominated region (in-band), (ii) the third-order distortion-dominated region (adjacent-band), and (iii) the noise-dominated region[76].

Independently of the spectral region considered, the objective is to determine the maximum radiated power that can occur in any spatial direction when the relative excitation phase $\Delta\Phi$ between the two antenna branches is swept over its entire range $\Delta\Phi \in [0, 2\pi]$. This maximum defines a spatial envelope which, as will be shown, constitutes a deterministic upper bound on the radiated power.

We first consider a single-user transmission scenario. The discrete-time complex baseband signal $x(n)$ represents the useful signal to be transmitted. The signal is assumed to be zero-mean and identical at the input of the two RF chains, except for a controllable excitation phase shift introduced by the beamforming process.

The output of the PA feeding the first radiating element is modeled by a memoryless third-order nonlinearity and can be expressed as

$$y_1(n) = x(n) + \alpha|x(n)|^2x(n) + w_1(n), \quad (\text{A.9})$$

where α denotes the third-order nonlinear coefficient and $w_1(n)$ represents an additive noise term accounting for thermal noise and residual impairments.

By convention, the second RF chain applies a relative excitation phase $\Delta\Phi$ to the useful signal. The corresponding PA output is therefore written as

$$y_2(n) = x(n)e^{j\Delta\Phi} + \alpha|x(n)|^2x(n)e^{j\Delta\Phi} + w_2(n), \quad (\text{A.10})$$

where $w_2(n)$ is assumed to be uncorrelated with $w_1(n)$.

This signal model explicitly separates three distinct contributions at the antenna inputs: the useful in-band component proportional to $x(n)$, the third-order OOB component proportional to $|x(n)|^2x(n)$, and additive noise terms. These three components exhibit different spatial behaviors when radiated by the antenna array, which motivates the analysis of the corresponding spectral regions.

Useful Signal Dominated Region In the in-band region, the useful signal $x(n)$ dominates the radiated spectrum. The signals feeding the two radiating elements are therefore phase-shifted versions of the same waveform and remain fully correlated.

As a result, the radiated electromagnetic fields combine coherently in space. The directional EIRP can be expressed as

$$EIRP_{\text{sig}}(\varphi, \Delta\Phi) = P_e + A_E(\varphi) + 20\log_{10}|AF(\varphi, \Delta\Phi)|, \quad (\text{A.11})$$

where P_e denotes the conducted power per antenna element and $AF(\varphi, \Delta\Phi)$ is the array factor.

When the relative phase $\Delta\Phi$ is swept over its full range, the maximum achievable array-factor magnitude is obtained. The corresponding spatial envelope becomes

$$EIRP_{\text{sig,env}}(\varphi) = P_e + A_E(\varphi) + 6\text{dB}. \quad (\text{A.12})$$

The envelope therefore has exactly the same angular dependence as the single-element radiation pattern, scaled by a factor of four in power.

Third-Order Distortion-Dominated Region We now consider the adjacent-band spectral region where the third-order distortion component $\alpha|x(n)|^2x(n)$ becomes dominant.

In the considered two-element configuration, the input signals applied to the two RF chains differ only by the deterministic phase shift $\Delta\Phi$. Consequently, the third-order distortion term experiences the same inter-element phase difference as the useful signal. The distortion components generated by the two power amplifiers therefore remain fully correlated.

As a result, the third-order intermodulation products are beamformed in exactly the same spatial direction as the useful signal. The corresponding directional EIRP can be written as

$$EIRP_{IM3}(\varphi, \Delta\Phi) = P_{IM3} + A_E(\varphi) + 20\log_{10}|AF(\varphi, \Delta\Phi)|, \quad (\text{A.13})$$

where P_{IM3} denotes the conducted power of the third-order distortion component at the output of a single element.

By sweeping $\Delta\Phi$ over the interval $[0, 2\pi]$, the spatial envelope becomes

$$EIRP_{IM3,env}(\varphi) = P_{IM3} + A_E(\varphi) + 6\text{dB}. \quad (\text{A.14})$$

Therefore, even in the distortion-dominated region, the maximum radiated power is obtained when the fields radiated by the two elements combine coherently. The spatial envelope is again determined solely by the radiation pattern of the elementary radiator.

Noise-Dominated Region Finally, we consider spectral regions where the additive noise terms dominate the radiated signal. In this case, the noise contributions $w_1(n)$ and $w_2(n)$ applied to the two antenna branches are assumed to be statistically independent and uncorrelated.

Unlike the useful signal and distortion components, which combine coherently in the electromagnetic field domain, uncorrelated noise contributions combine incoherently in power. The resulting directional EIRP can therefore be expressed as

$$EIRP_{noise}(\varphi) = P_{noise} + A_E(\varphi) + 3\text{dB}, \quad (\text{A.15})$$

where P_{noise} denotes the noise power per branch. The additional 3 dB term results from the incoherent power addition of two independent noise sources.

A.2.1.3 Spatial Upper Bound

Considering the three spectral regions together, a key conclusion emerges. Regardless of whether the radiated power is dominated by the useful signal, by third-order distortion components, or by noise, the maximum radiated power that can occur in any spatial direction is bounded by a deterministic spatial envelope.

For the two-element array considered here, this envelope has the same angular shape as the elementary radiator pattern and reaches its maximum in the boresight direction. The corresponding maximum level results either from the coherent addition of the radiated fields (for the useful signal and distortion components) or from incoherent power addition in the noise regime.

This establishes the existence of a spatial upper bound for the radiated power which is independent of frequency and of the nature of the radiated signal component. Although derived here using a third-order nonlinear model, the result can be naturally extended to higher-order nonlinearities described by Volterra models.

A.2.2 Multi-user transmission case

We now extend the previous analysis to a multi-user transmission scenario, which is representative of practical Massive MIMO systems where several users are served simultaneously through spatial multiplexing. For clarity, the discussion is restricted to the case of two users, which is sufficient to illustrate the fundamental spatial mechanisms governing the radiation of useful signals and nonlinear distortion products. The extension to a larger number of users ($K > 2$) follows similar principles, although a complete formal proof is left for future work. Nevertheless, the two-user case already reveals the essential effects of power sharing and spatial dispersion of nonlinear distortion, which play a key role in the resulting radiation patterns.

A.2.2.1 Signal Model

Let $u_1(n)$ and $u_2(n)$ denote the discrete-time complex baseband signals associated with user 1 and user 2, respectively. These signals are assumed to be zero-mean and mutually uncorrelated. At the input of the first radiating element, the composite transmit signal is formed by the superposition of the two user signals:

$$x_1(n) = u_1(n) + u_2(n). \quad (\text{A.16})$$

This signal is applied to a power amplifier modeled by a memoryless third-order nonlinearity. The corresponding output signal can be written as

$$y_1(n) = x_1(n) + \alpha |x_1(n)|^2 x_1(n) + w_1(n), \quad (\text{A.17})$$

where α denotes the third-order nonlinear coefficient and $w_1(n)$ represents an additive noise contribution accounting for thermal noise and residual RF impairments.

For the second radiating element, independent beamforming phase shifts are applied to the user signals in order to steer two independent beams toward the intended user directions. The input signal to the second RF chain is therefore

$$x_2(n) = u_1(n)e^{j\Delta\Phi_1} + u_2(n)e^{j\Delta\Phi_2}, \quad (\text{A.18})$$

where $\Delta\Phi_1$ and $\Delta\Phi_2$ denote the beamforming phase shifts associated with user 1 and user 2, respectively.

After amplification by the nonlinear power amplifier, the output signal of the second branch becomes

$$y_2(n) = x_2(n) + \alpha |x_2(n)|^2 x_2(n) + w_2(n), \quad (\text{A.19})$$

where $w_2(n)$ denotes an additive noise term assumed to be independent of $w_1(n)$.

Substituting the beamformed input signals into the nonlinear model allows the output signals of the two branches to be expressed explicitly. For branch 1 we obtain

$$y_1(n) = u_1(n) + u_2(n) \quad (\text{A.20})$$

$$\begin{aligned} &+ \alpha \left(|u_1(n)|^2 u_1(n) + |u_2(n)|^2 u_2(n) \right. \\ &+ |u_1(n)|^2 u_2(n) + |u_2(n)|^2 u_1(n) \\ &\left. + u_1^2(n) u_2^*(n) + u_2^2(n) u_1^*(n) \right) + w_1(n). \end{aligned} \quad (\text{A.21})$$

For branch 2 the output becomes

$$y_2(n) = u_1(n)e^{j\Delta\Phi_1} + u_2(n)e^{j\Delta\Phi_2} \quad (\text{A.22})$$

$$+ \alpha (\Gamma_{SD}(n) + \Gamma_{XA}(n) + \Gamma_{XB}(n)) + w_2(n), \quad (\text{A.23})$$

where the nonlinear distortion terms can be grouped according to their physical origin. The *self-distortion* terms are defined as

$$\Gamma_{SD}(n) = |u_1(n)|^2 u_1(n)e^{j\Delta\Phi_1} + |u_2(n)|^2 u_2(n)e^{j\Delta\Phi_2}, \quad (\text{A.24})$$

while the *Type-A cross-intermodulation* terms are

$$\Gamma_{XA}(n) = 2|u_2(n)|^2 u_1(n)e^{j\Delta\Phi_1} + 2|u_1(n)|^2 u_2(n)e^{j\Delta\Phi_2}. \quad (\text{A.25})$$

Finally, the *Type-B cross-intermodulation* terms are given by

$$\Gamma_{XB}(n) = u_1^2(n)u_2^*(n)e^{j(2\Delta\Phi_1 - \Delta\Phi_2)} + u_2^2(n)u_1^*(n)e^{j(2\Delta\Phi_2 - \Delta\Phi_1)}. \quad (\text{A.26})$$

These expressions reveal that the nonlinear distortion components generated by the power amplifier can be classified into three categories:

- Self-distortion terms proportional to $|u_k(n)|^2 u_k(n)$,
- Type-A cross-intermodulation terms proportional to $|u_i(n)|^2 u_j(n)$,
- Type-B cross-intermodulation terms proportional to $u_i^2(n)u_j^*(n)$.

A.2.2.2 Spatial Behavior of Nonlinear Components

The useful signal components $u_1(n)$ and $u_2(n)$ are beamformed toward two independent spatial directions determined by the phase shifts $\Delta\Phi_1$ and $\Delta\Phi_2$, respectively.

An important observation is that the self-distortion terms preserve the phase structure of their corresponding user signal. Consequently, the component $|u_k(n)|^2 u_k(n)$ radiates in the same spatial direction as the useful signal $u_k(n)$. In other words, self-distortion remains aligned with the beam associated with its originating user.

A similar behavior is observed for the Type-A cross-intermodulation terms. Specifically,

- The term $|u_2|^2 u_1$ carries the phase of u_1 and is therefore beamformed in the direction of user 1.
- The term $|u_1|^2 u_2$ carries the phase of u_2 and is therefore beamformed in the direction of user 2.

In contrast, the Type-B cross-intermodulation terms involve quadratic phase combinations that do not correspond to either user beam. For example,

- The term $u_1^2 u_2^*$ radiates in a direction determined by the phase combination $2\Delta\Phi_1 - \Delta\Phi_2$,
- The term $u_2^2 u_1^*$ radiates in a direction determined by $2\Delta\Phi_2 - \Delta\Phi_1$.

As a result, the total radiated distortion power is dispersed over four distinct spatial directions:

1. User 1 beam direction ($\Delta\Phi_1$): useful signal u_1 , self-distortion $|u_1|^2 u_1$, and Type-A cross-IM $2|u_2|^2 u_1$.
2. User 2 beam direction ($\Delta\Phi_2$): useful signal u_2 , self-distortion $|u_2|^2 u_2$, and Type-A cross-IM $2|u_1|^2 u_2$.
3. Cross-IM direction $2\Delta\Phi_1 - \Delta\Phi_2$: Type-B term $u_1^2 u_2^*$.
4. Cross-IM direction $2\Delta\Phi_2 - \Delta\Phi_1$: Type-B term $u_2^2 u_1^*$.

A.2.2.3 Implications for Radiated Power Distribution

From a spatial-power perspective, the multi-user case is intrinsically more favorable than the single-user case for two fundamental reasons.

First, power sharing occurs between the different beams. Since the total transmit power is distributed among multiple users, the peak EIRP in any given direction is reduced compared to the single-user case, where all the power is concentrated in a single beam steered toward boresight.

Second, the nonlinear distortion components experience spatial dispersion. As demonstrated above, the third-order intermodulation products are radiated in four different spatial directions rather than being concentrated in a single beam. This dispersion further reduces the radiated power density in any individual direction.

Consequently, for any observation direction ϕ , the radiated EIRP in the multi-user case satisfies the inequality

$$EIRP_{MU}(\phi) \leq EIRP_{SU,max}(\phi), \quad (A.27)$$

where $EIRP_{SU,max}(\phi)$ denotes the maximum EIRP obtained in the single-user boresight configuration.

This inequality holds for all signal components, including the useful signal, nonlinear distortion products, and noise contributions. Therefore, the spatial upper bound derived for the single-user remains valid and conservative in a multi-user transmission scenario. The worst-case radiated power is still achieved in the single-user boresight configuration, while multi-user operation can only reduce the maximum radiated power in any given spatial direction.

A.2.3 Extension to Active Antenna Systems

The spatial upper-bound concept derived for the two-element array is extended to a realistic AAS. The objective is to demonstrate that the key result obtained in the simplified configuration, namely the existence of a deterministic spatial envelope independent of beamforming direction or signal nature, remains valid for practical large-scale arrays. The theoretical findings are further supported by OTA measurements performed on a commercial Massive MIMO radio unit, previously used in the measurement setup described in 4.3.2. The AAS under test is the SPEAR radio unit from Andrew, driven by OAI platform, enabling flexible and controlled beamforming configurations.

In practical AAS implementations, individual antenna elements are grouped into sub-arrays, each driven by an independent RF chain. Let M and N denote the number of sub-arrays along the vertical and horizontal dimensions, respectively. Considering dual-polarized elements (typically $\pm 45^\circ$), the total number of RF chains is equal to $2MN$ as seen in Figure 1.20.

The EIRP in a given direction (θ, ϕ) can be expressed as [77]

$$EIRP_{sig}(\theta, \phi, \Delta\Phi) = P_{sub} + 3 \text{ dB} + A_{sub}(\theta, \phi) + 20 \log_{10} |AF_A(\theta, \phi, \Delta\Phi)|, \quad (A.28)$$

where P_{sub} denotes the conducted power per sub-array, the +3 dB term accounts for the contribution of the two polarizations, $A_{sub}(\theta, \phi)$ represents the sub-array radiation pattern, and $AF_A(\theta, \phi, \Delta\Phi)$ is the array factor of the AAS.

The array factor can be written as

$$AF_A(\theta, \phi) = \sum_{m=0}^{M-1} \sum_{n=0}^{N-1} w_{m,n} v_{m,n}(\theta, \phi), \quad (A.29)$$

where the geometrical phase term is given by

$$v_{m,n}(\theta, \varphi) = \exp \left(j2\pi \left[\frac{md_v}{\lambda} \cos \theta + \frac{nd_h}{\lambda} \sin \theta \sin \varphi \right] \right), \quad (\text{A.30})$$

and the excitation weights are defined as

$$w_{m,n} = \exp(j[m\Delta\Phi_V + n\Delta\Phi_H]), \quad (\text{A.31})$$

with $\Delta\Phi_V$ and $\Delta\Phi_H$ denoting the controllable phase gradients in the vertical and horizontal directions, respectively.

The key idea underlying the spatial upper-bound extension is that, for any given observation direction (θ_0, φ_0) , the geometrical phase progression across the array is fully deterministic. Since the excitation phase gradients $\Delta\Phi_V$ and $\Delta\Phi_H$ are controllable through beamforming, they can always be selected to compensate exactly for this geometrical phase.

Specifically, by choosing

$$\Delta\Phi_V = -\frac{2\pi d_v}{\lambda} \cos \theta_0, \quad \Delta\Phi_H = -\frac{2\pi d_h}{\lambda} \sin \theta_0 \sin \varphi_0, \quad (\text{A.32})$$

the excitation phase applied to each sub-array becomes

$$m\Delta\Phi_V + n\Delta\Phi_H = -2\pi \left[\frac{md_v}{\lambda} \cos \theta_0 + \frac{nd_h}{\lambda} \sin \theta_0 \sin \varphi_0 \right], \quad (\text{A.33})$$

which exactly cancels the geometrical phase term in $v_{m,n}(\theta_0, \varphi_0)$.

As a consequence, all the contributions in the double summation become co-phased in the considered direction. The array factor therefore reaches its maximum value:

$$AF_A(\theta_0, \varphi_0) = MN. \quad (\text{A.34})$$

This result generalizes the two-element case, where the maximum array-factor magnitude was equal to the number of elements. Here, the maximum field addition corresponds to the coherent summation of all MN sub-arrays.

Following the reasoning developed in Sections II and III, the fundamental interpretation of the spatial upper bound naturally extends to the AAS case. The role of the elementary radiator is no longer played by a single antenna element, but by the sub-array itself. Consequently, the spatial envelope that bounds the maximum radiated power over all possible beamforming configurations has the same angular dependence as the sub-array radiation pattern $A_{\text{sub}}(\theta, \varphi)$.

This result demonstrates that, even in large-scale beamforming systems with many degrees of freedom, the maximum radiated power in any spatial direction remains governed by a deterministic envelope. This envelope is independent of the specific beamforming configuration, the transmitted signal, and the frequency, and is solely determined by the radiation characteristics of the sub-array.

A.3 Experimental validation

A.3.1 Measurement Environment and Methodology

The experimental validation is carried out through OTA measurements with the setup described previously in Chapter 4. The antenna system is first configured in a single-user transmission mode with the fixed boresight beam.

The received signal is captured using a calibrated probe antenna over an azimuthal angular sweep $\varphi \in [-60^\circ, 60^\circ]$. For each spatial direction, the received PSD is estimated by integrating the received signal

power over a 1 MHz RBW. This procedure is repeated over the full frequency range from 3.4 GHz to 4 GHz, yielding a joint spatial–spectral characterization of the radiated field.

A.3.2 Boresight Frequency-Domain Results

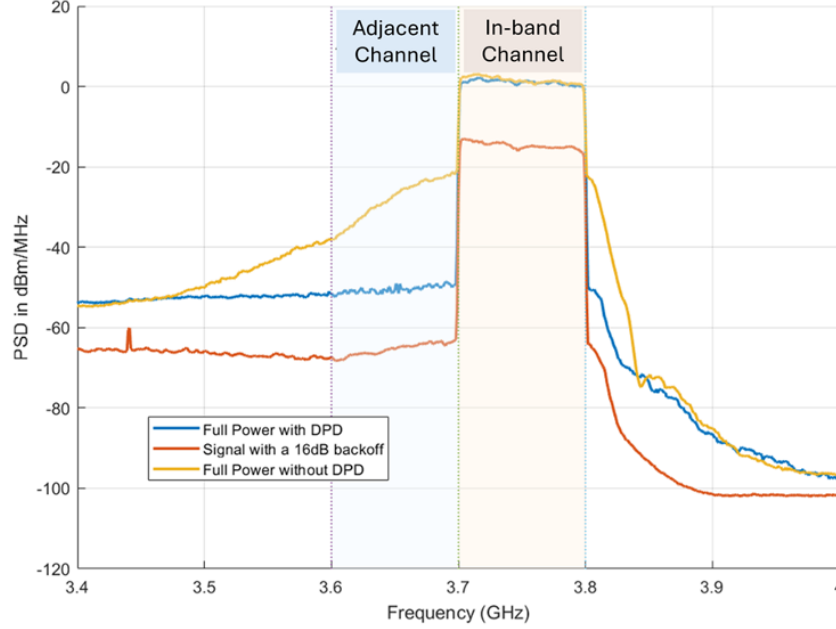


Figure A.3: Measured boresight EIRP spectral density (1 MHz RBW) from 3.4 to 4.0 GHz for three configurations (Full power with and without DPD, 16 dB backoff). Vertical dashed lines indicate measurement frequencies for angular analysis: 3.75 GHz (in-band region), 3.65 GHz (IM3 and noise dominated region).

Figure A.3 reports the measured boresight equivalent isotropic radiated power (EIRP) spectral density for three operating conditions:

- full-power 5G NR transmission with digital predistortion (DPD),
- full-power transmission without DPD,
- 16 dB backoff operation.

Full-power transmission corresponds to approximately 72 dBm per polarization (75 dBm total). Considering the single-polarization probe configuration and a total attenuation of 51 dB between the antenna array and the probe, the measured in-band PSD is approximately 1 dBm/MHz.

The results show that:

- the in-band emission remains stable across the 3.7–3.8 GHz range,
- DPD significantly reduces out-of-band emissions,
- power backoff provides an additional reduction in nonlinear distortion products.

These measurements provide a direct experimental upper bound of the radiated spectral envelope at boresight, which is later used as a reference for spatial analysis.

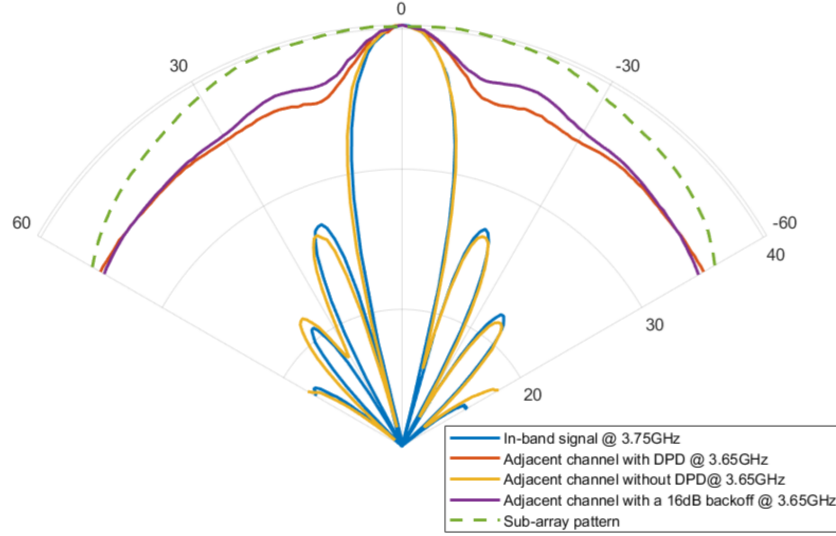


Figure A.4: Azimuth cuts for boresight beam, signals integrated on 100 MHz (in-band, adjacent with and without DPD, adjacent with a 16 dB backoff)

A.3.3 Angular Dispersion of Spectral Components

Figure A.4 presents azimuth cuts for the boresight beam under different spectral configurations, including in-band and adjacent-band components with and without DPD.

The main observations are:

- The in-band radiation exhibits a highly directive pattern centered at boresight.
- Without DPD, adjacent-channel emissions closely follow the same spatial distribution as the useful signal, indicating that nonlinear distortion products are beamformed coherently.
- With DPD, adjacent-band components are significantly reduced and exhibit increased spatial dispersion.
- Under 16 dB backoff, nonlinear components are inherently low and also show reduced spatial coherence.

All radiation patterns are normalized to the boresight level to enable direct comparison of spatial shapes independently of absolute power levels.

A.3.4 Beam Steering and Spatial Upper Bound Validation

A second experimental campaign evaluates the antenna under multiple beam steering directions (-42° , -30° , -18° , 0° , 18° , 30° , 42°) using full-power transmission with DPD.

Figure A.5 and A.6 compares measured radiation patterns with a theoretical spatial upper bound derived from the boresight measurement and a 3GPP-compliant sub-array model.

The main findings are:

- The theoretical spatial envelope accurately predicts the maximum observed EIRP across all steering angles.
- The 3GPP sub-array model with $\varphi_{3dB} = 85^\circ$ provides a consistent representation of the elementary radiator behavior.

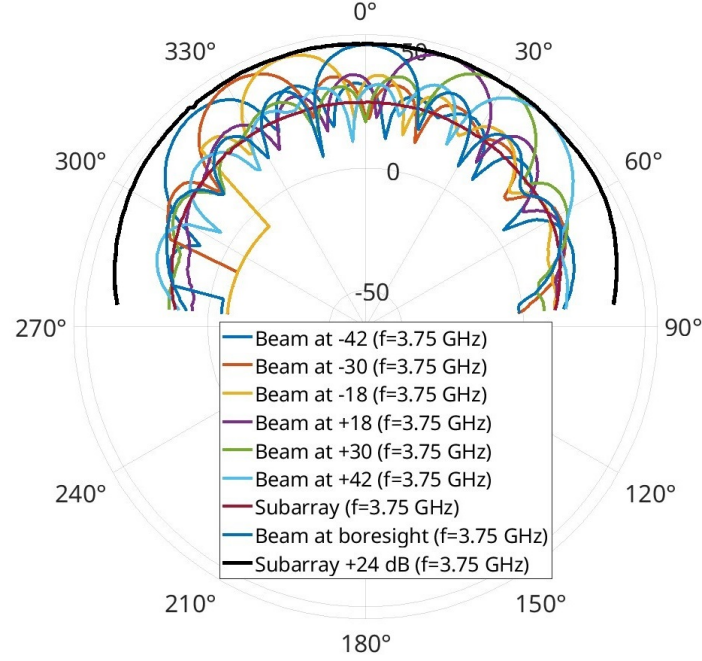


Figure A.5: Validation of spatial upper bound for seven beam steering angles ($-42^\circ, -30^\circ, -18^\circ, 0^\circ, 18^\circ, 30^\circ, 42^\circ$) in-band region. The maximum deviation between the bound and measurements is below 1 dB.

- The maximum deviation between measured and theoretical envelopes remains below 1 dB across all configurations.

This excellent agreement confirms that the boresight-based envelope constitutes a valid conservative predictor of spatial radiated power for arbitrary beam directions.

A.3.5 Multi-User MIMO Experimental Analysis

The multi-user (MU) configuration is evaluated by simultaneously steering two beams toward $\phi_1 = 0^\circ$ and $\phi_2 = 18^\circ$. For each direction, in-band and adjacent-band radiated powers are computed by integrating the PSD over the corresponding frequency bands.

From the nonlinear signal model, third-order distortion terms generate two categories of interference components:

- **Type A terms:** proportional to $|u_k|^2 u_k$ and $|u_\ell|^2 u_k$, preserving the phase gradient of each user and thus radiating in the original beam directions.
- **Type B terms:** proportional to $u_1^2 u_2^*$ and $u_2^2 u_1^*$, generating additional phase gradients:

$$\Delta\Phi_{B1} = 2\Delta\Phi_1 - \Delta\Phi_2, \quad \Delta\Phi_{B2} = 2\Delta\Phi_2 - \Delta\Phi_1. \quad (\text{A.35})$$

For a phased array with $\Delta\Phi(\phi) \propto \sin \phi$, the corresponding angular locations satisfy:

$$\sin \phi_{B1} = 2 \sin \phi_1 - \sin \phi_2, \quad \sin \phi_{B2} = 2 \sin \phi_2 - \sin \phi_1. \quad (\text{A.36})$$

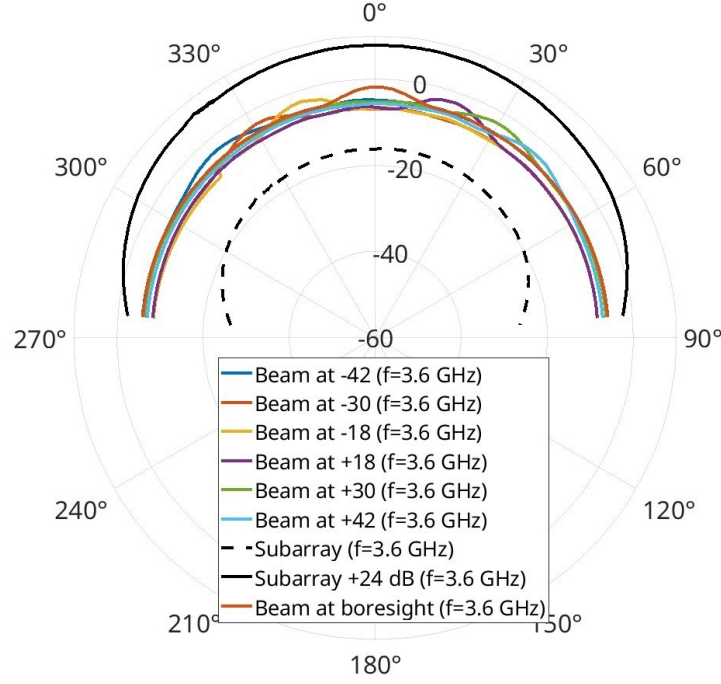


Figure A.6: Validation of spatial upper bound for seven beam steering angles ($-42^\circ, -30^\circ, -18^\circ, 0^\circ, 18^\circ, 30^\circ, 42^\circ$) in OOB region with DPD. The maximum deviation between the bound and measurements is below 1 dB.

For $\phi_1 = 0^\circ$ and $\phi_2 = 18^\circ$, this leads to:

$$\phi_{B1} \approx -18^\circ, \quad \phi_{B2} \approx 38.2^\circ. \quad (\text{A.37})$$

A.3.6 Multi-User Spatial Results

Figure A.7 illustrates the spatial distributions measured for the MU scenario.

The key observations are:

- The in-band radiation forms two dominant lobes aligned with the intended user directions.
- Adjacent-band emissions without DPD exhibit structured spatial patterns partially aligned with the useful beams, consistent with Type A distortion mechanisms.
- Additional angular spreading is observed, consistent with the presence of Type B intermodulation components.
- DPD significantly reduces both structured and dispersed out-of-band emissions.

These results confirm that multi-user transmission inherently increases spatial complexity of nonlinear distortion products, leading to additional radiation directions not present in single-user operation.

The experimental results validate a conservative spatial upper bound framework derived from boresight measurements. In particular:

- The boresight EIRP provides a reliable worst-case reference for spatial radiation envelopes.
- Beam steering does not increase the maximum radiated envelope beyond the boresight-determined bound.

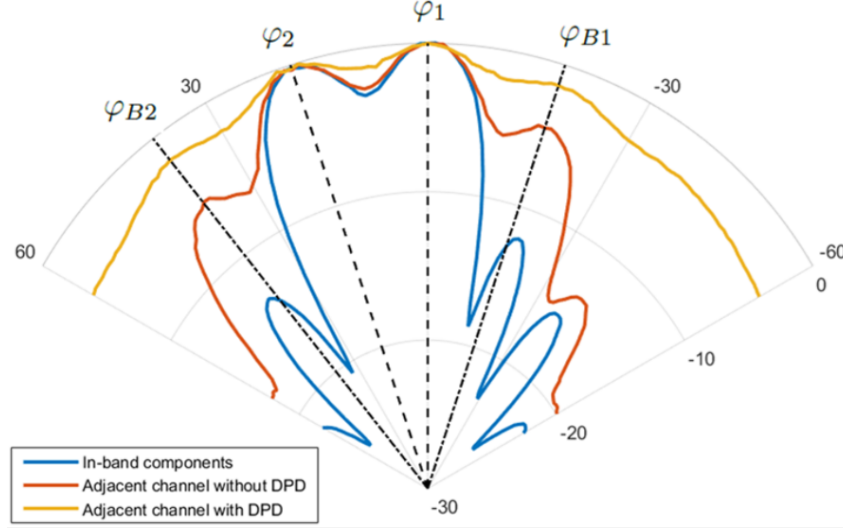


Figure A.7: Multi-user spatial dispersion. All radiation patterns (in-band and adjacent band) are normalized to the in-band boresight level (0 dB reference), enabling direct comparison of spatial distribution shapes.

- Multi-user operation introduces additional spatial components but remains bounded by the same envelope when uncertainties are included.

The proposed framework incorporates measurement and modeling uncertainties, including OTA measurement accuracy (± 1.3 dB for in-band power, 95% confidence per 3GPP TS 38.141-2), sub-array characterization errors, and manufacturing variability.

A conservative bound is defined as:

$$\text{EIRP}_{\text{bound}}(f) = \text{EIRP}_{\text{meas}}(f) + 1.3 \text{ dB}. \quad (\text{A.38})$$

For out-of-band emissions, a margin up to 3 dB may be applied when a uniform worst-case treatment is required.

Overall, the results demonstrate that a simple boresight-based characterization, combined with well-defined uncertainty margins, is sufficient to predict a conservative spatial envelope for both single-user and multi-user Massive MIMO transmissions.

A.4 Conclusions

This work demonstrates that AAS using beamforming and Massive MIMO are subject to a deterministic spatial upper bound on their radiated power. Beginning with a simple two-element array exhibiting third-order nonlinearities, it is shown that varying the excitation phase produces a spatial envelope whose angular profile matches that of a single radiating element. This principle extends naturally to practical AAS implementations, where sub-arrays function as the fundamental radiating units. In every case, the highest emission occurs in the boresight direction.

A key result is that this upper bound does not depend on operating frequency, signal type (whether desired signal, nonlinear distortion, or noise), beamforming strategy, or explicit modeling of correlations between RF chains. Instead, all such effects are inherently reflected in the measured spectral density at boresight.

Experimental validation conducted on a 3.5 GHz Massive MIMO antenna supports the theoretical findings in three spectral regions and seven beam steering angles, with deviations remaining within typical over-the-air measurement uncertainty. In multi-user scenarios, the peak EIRP is naturally reduced due to power distribution and spatial spread of intermodulation products, consistent with the predicted four-lobe radiation pattern.

Overall, this framework offers a conservative and practical approach for coexistence studies involving beamformed systems, with particular relevance to adjacent-band interference cases such as interactions between 5G systems and radio altimeters.

Appendix B

Extension to 6G and beyond bands

B.1 OOB behavior at 3.5 GHz

To investigate the spatial characteristics of OOB emissions in practical 5G massive MIMO deployments, measurement campaigns were conducted on a commercial Nokia AirScale AAS operating in the n78 band at a carrier frequency of $f_c = 3.5$ GHz. The evaluated base station is a 64TX64RX massive MIMO panel comprising 192 radiating elements and delivering a total output power of 320 W. The antenna architecture is based on 3-element subarrays, yielding an effective configuration of 32 subarrays per polarization.

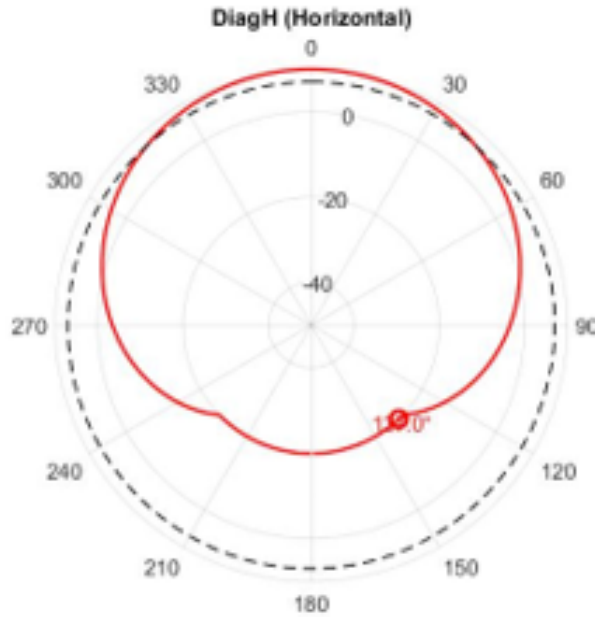


Figure B.1: Subarray radiation pattern in the azimuth plane based on the 3GPP model

Radiation measurements were performed using the ANFR drone-based three-dimensional electromagnetic field measurement platform, which allows for full spatial characterization of the OOB emission pattern in both the azimuth and elevation planes. The expected measured patterns are depicted in Figures B.1 and B.2.

Since the carrier frequency remains fixed during the measurements, the wavelength λ is constant. Consequently, the principal array-scaling parameter that influences radiation behavior is the effective vertical aperture, expressed as $M \times d_v$, where M denotes the number of vertically stacked radiating units

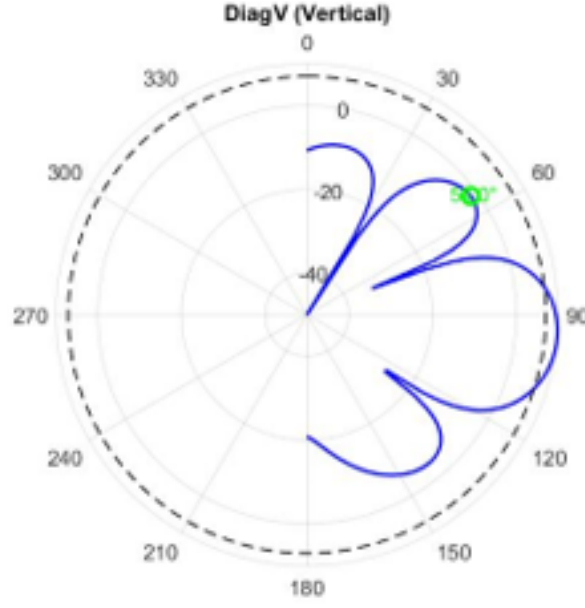


Figure B.2: Subarray radiation pattern in the elevation plane based on the 3GPP model

and d_v the vertical inter-element spacing. Variations in this effective aperture directly affect the achievable gain and directivity, with the considered configuration corresponding to an approximate gain factor of $G = 32$.

A key observation from the measurement campaign is that the measured OOB radiation pattern does not replicate the highly directive in-band beamformed pattern of the full massive MIMO array. Instead, the OOB spatial distribution closely resembles the radiation pattern of an individual 3-element subarray, while preserving a residual electrical downtilt characteristic. This behavior suggests that OOB emissions are primarily governed by subarray-level analog combining and RF front-end nonlinearities rather than by the full digital beamforming process applied in-band.

These preliminary findings indicate that the spatial selectivity of OOB emissions may be significantly lower than that of in-band transmissions, which has important implications for electromagnetic exposure assessment and interference analysis in future 5G-Advanced and 6G deployments. Nevertheless, additional measurements and broader statistical validation are required before definitive conclusions can be drawn.

B.2 OOB behavior in 6G and beyond bands

A second measurement campaign was conducted in mid-2025 in Châtillon using a proof-of-concept antenna platform representative of prospective upper-6 GHz / early 6G deployments. The investigated system operates in the frequency range of [6.889–7.289] GHz with an effective transmission bandwidth of 200 MHz, corresponding either to a contiguous 200 MHz allocation or two aggregated NR carriers of 100 MHz each.

The considered active antenna system employs a 128TX/128RX architecture comprising 64 subarrays, each formed by 6 radiating elements. The array is organized in a dual-polarized configuration 24×16 , with four vertically stacked subarrays per column and per polarization, resulting in an effective array gain of approximately $G = 64$.

Compared with the 3.5 GHz deployment, operation in the upper-6 GHz band leads to an approximate halving of the wavelength λ . To preserve similar electromagnetic aperture characteristics and beamforming

performance, the vertical inter-element spacing d_v is correspondingly reduced by approximately a factor of two. This geometrical scaling maintains comparable normalized array behavior despite the higher operating frequency.

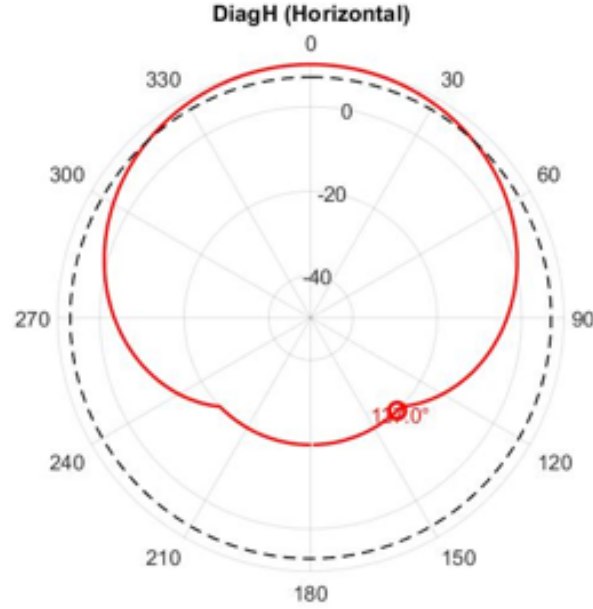


Figure B.3: Subarray radiation pattern in the azimuth plane based on the 3GPP model

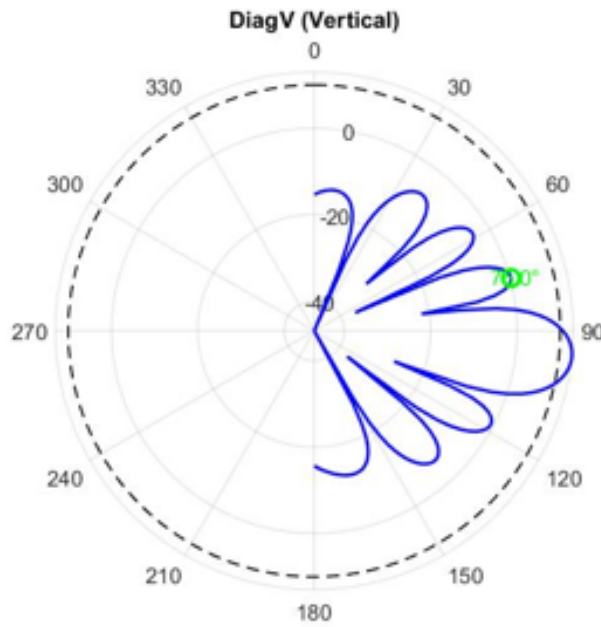


Figure B.4: Subarray radiation pattern in the elevation plane based on the 3GPP model

The expected OOB radiation patterns in the azimuth and elevation planes are shown in Figures B.3 and B.4. Analysis of the measurements reveals that, similar to the case 3.5 GHz, the full-array OOB

radiation pattern can be accurately predicted from the radiation pattern of a single 6-element subarray when considering both horizontal and vertical cuts.

This observation strongly supports the hypothesis that OOB emissions are primarily determined by subarray-level RF behavior and front-end architecture, rather than by the overall array dimensions or the in-band beamforming configuration. Consequently, the spatial behavior of the OOB appears to remain largely invariant with respect to the size of the antenna array and the operating frequency band, provided that the underlying subarray architecture is preserved.

These results suggest the existence of a scalable and frequency-independent modeling approach for predicting OOB emissions in future massive MIMO systems, including prospective 6G deployments. Such predictability could significantly simplify compliance assessment, interference modeling, and exposure evaluation for next-generation antenna arrays.

Appendix C

Correlation function and antenna gain

C.1 Calculation of G at boresight

We define the relation between one array beam pattern $A_A(\theta, \varphi)$ and one subarray beam pattern $A_{sub}(\theta, \varphi)$ and the correlation coefficient ρ as [56]:

$$A_A(\theta, \varphi) = A_{sub}(\theta, \varphi) + 10 \log_{10} \left[1 + \rho \cdot \left(\left| \sum_{m=1}^M \sum_{n=1}^N w_{m,n} \cdot v_{m,n} \right|^2 - 1 \right) \right] \quad (C.1)$$

where N is the number of columns of radiating elements per row, M is the number of subarrays per column, $w_{m,n}$ is the weight vector to be fixed for every radiating element and $v_{m,n}$ is the phase shift related to the position of the radiating element. We note that all the radiating elements in each separate subarray, will have the same complex weight.

We define G to be the antenna array gain:

$$G = \left| \sum_{m=1}^M \sum_{n=1}^N w_{m,n} \cdot v_{m,n} \right|^2 \quad (C.2)$$

so that the composite array radiation pattern can be expressed as

$$A_A(\theta, \varphi) = A_{sub}(\theta, \varphi) + 10 \log_{10} \left[1 + \rho(G - 1) \right] \quad (C.3)$$

Our target is to find the correlation coefficient between the subarray and the full array beam pattern relative to the measured beam patterns from the Andrew radio unit.

Referring to Table 5.4.4.2-3 [56] Composite array pattern for multiple column, we can write $w_{m,n}$ and $v_{m,n}$ with their respective formulas:

$$w_{m,n} = \frac{1}{\sqrt{MN}} e^{i 2\pi \left[\frac{(m-1)d_v}{\lambda} \sin(\theta_{etilt}) - \frac{(n-1)d_h}{\lambda} \cos(\theta_{etilt}) \sin(\varphi_{escan}) \right]} \quad (C.4)$$

$$v_{m,n} = e^{i 2\pi \left[\frac{(m-1)d_v}{\lambda} \cos(\theta) + \frac{(n-1)d_h}{\lambda} \sin(\theta) \sin(\varphi) \right]} \quad (C.5)$$

where d_v is the vertical distance between 2 subarrays, d_h is the horizontal distance between 2 radiating elements, θ_{etilt} and φ_{escan} are the elevation and azimuth directions of the selected beam direction [58].

We reflect these formulas on our case with the Andrew-Amphenol antenna array: $d_v = 4.2\lambda$ which is the sum of 6 spaced radiating elements of 0.7λ , $d_h = 0.5\lambda$, $N = 8$, $M = 2$, $\theta_{etilt} = 6^\circ$, $\varphi_{escan} = 0^\circ$, $\varphi = 0^\circ$ and $\theta = 96^\circ$.

We will calculate first the parameter G :

Combine $w_{m,n}$ and $v_{m,n}$:

$$w_{m,n}v_{m,n} = \frac{1}{\sqrt{MN}} e^{j2\pi(m-1)(\frac{d_v}{\lambda} \sin\theta_{etilt} + \frac{d_v}{\lambda} \cos\theta)} e^{j2\pi(n-1)(-\frac{d_h}{\lambda} \cos\theta_{etilt} \sin\varphi_{escan} + \frac{d_h}{\lambda} \sin\theta \sin\varphi)}. \quad (C.6)$$

Since $\varphi_{escan} = \varphi = 0^\circ$, the n -dependent exponential is 1. Thus, the sum over columns simplifies to

$$\sum_{n=1}^N w_{m,n}v_{m,n} = N \cdot \frac{1}{\sqrt{MN}} e^{j2\pi(m-1)(\frac{d_v}{\lambda} \sin\theta_{etilt} + \frac{d_v}{\lambda} \cos\theta)}. \quad (C.7)$$

Finally, the sum over subarrays is

$$\sum_{m=1}^M \sum_{n=1}^N w_{m,n}v_{m,n} = \sum_{m=1}^M \frac{N}{\sqrt{MN}} e^{j2\pi(m-1)(\frac{d_v}{\lambda} \sin\theta_{etilt} + \frac{d_v}{\lambda} \cos\theta)}. \quad (C.8)$$

Given: $M = 2$, $N = 8$, $d_v = 4.2\lambda$, $d_h = 0.5\lambda$, $\theta_{etilt} = 6^\circ$, $\theta = 96^\circ$, $\varphi_{escan} = \varphi = 0^\circ$.

$$\frac{N}{\sqrt{MN}} = \frac{8}{\sqrt{16}} = 2$$

$$\frac{d_v}{\lambda} (\sin\theta_{etilt} + \cos\theta) = 4.2(\sin 6^\circ + \cos 96^\circ)$$

$$\sin 6^\circ \approx 0.1045, \quad \cos 96^\circ \approx -0.1045 \quad \Rightarrow \quad \sin 6^\circ + \cos 96^\circ \approx 0$$

Thus, the phase factor becomes

$$e^{j2\pi(m-1)4.2(\sin 6^\circ + \cos 96^\circ)} = e^{j0} = 1.$$

Sum over subarrays:

$$\sum_{m=1}^2 \sum_{n=1}^8 w_{m,n}v_{m,n} = \sum_{m=1}^2 2 \cdot 1 = 2 + 2 = 4$$

$$G = \left| \sum_{m=1}^2 \sum_{n=1}^8 w_{m,n}v_{m,n} \right|^2 = |4|^2 = 16$$

The array gain $G = 16$ confirms that all subarrays and elements add coherently in this beam direction, achieving maximum constructive interference along the desired direction.

C.2 Calculation of G at $\varphi = 30^\circ$

The array gain is defined as

$$G \triangleq \left| \sum_{m=1}^M \sum_{n=1}^N w_{m,n}v_{m,n} \right|^2. \quad (C.9)$$

The complex excitation weights are given by

$$w_{m,n} = \frac{1}{\sqrt{MN}} \exp \left\{ j2\pi \left[\frac{(m-1)d_v}{\lambda} \sin \theta_{etilt} - \frac{(n-1)d_h}{\lambda} \cos \theta_{etilt} \sin \varphi_{escan} \right] \right\}, \quad (C.10)$$

and the phase response of the array elements by

$$v_{m,n} = \exp \left\{ j2\pi \left[\frac{(m-1)d_v}{\lambda} \cos \theta + \frac{(n-1)d_h}{\lambda} \sin \theta \sin \varphi \right] \right\}. \quad (C.11)$$

The product of the excitation weight and phase response can be written as

$$w_{m,n} v_{m,n} = \frac{1}{\sqrt{MN}} \exp(j2\pi(m-1)A) \exp(j2\pi(n-1)B), \quad (C.12)$$

where

$$A \triangleq \frac{d_v}{\lambda} (\sin \theta_{etilt} + \cos \theta), \quad (C.13)$$

$$B \triangleq \frac{d_h}{\lambda} (-\cos \theta_{etilt} \sin \varphi_{escan} + \sin \theta \sin \varphi). \quad (C.14)$$

The numerical parameters used in the evaluation are

$$M = 2, \quad N = 8, \quad (C.15)$$

$$d_v = 4.2\lambda, \quad d_h = 0.5\lambda, \quad (C.16)$$

$$\theta_{etilt} = 6^\circ, \quad \theta = 96^\circ, \quad (C.17)$$

$$\varphi_{escan} = 30^\circ, \quad \varphi = 30^\circ. \quad (C.18)$$

Substituting the parameters into the vertical phase coefficient yields

$$A = 4.2 (\sin 6^\circ + \cos 96^\circ). \quad (C.19)$$

Using the trigonometric identity

$$\cos(90^\circ + \alpha) = -\sin \alpha, \quad (C.20)$$

it follows that

$$\cos 96^\circ = -\sin 6^\circ, \quad (C.21)$$

and therefore

$$A = 0. \quad (C.22)$$

Consequently,

$$\exp(j2\pi(m-1)A) = 1, \quad (C.23)$$

and the vertical summation is fully coherent.

Substituting the parameters into the horizontal phase coefficient gives

$$B = 0.5 (-\cos 6^\circ \sin 30^\circ + \sin 96^\circ \sin 30^\circ). \quad (C.24)$$

This expression is retained in exact form and evaluated numerically without algebraic simplification:

$$B = \frac{1}{2} \left(-\cos 6^\circ \cdot \frac{1}{2} + \sin 96^\circ \cdot \frac{1}{2} \right). \quad (\text{C.25})$$

The horizontal array factor is

$$S_N = \sum_{n=1}^N \exp(j2\pi(n-1)B), \quad (\text{C.26})$$

which admits the closed-form magnitude

$$|S_N| = \left| \frac{\sin(\pi NB)}{\sin(\pi B)} \right|. \quad (\text{C.27})$$

For $N = 8$, direct numerical evaluation yields

$$|S_8| = 7.699. \quad (\text{C.28})$$

Including the normalization factor $1/\sqrt{MN} = 1/4$ and the vertical summation over $M = 2$ elements, the magnitude of the total array response is

$$\left| \sum_{m=1}^2 \sum_{n=1}^8 w_{m,n} v_{m,n} \right| = 2 \cdot \frac{1}{4} \cdot 7.699 = 3.852. \quad (\text{C.29})$$

The array gain at $\varphi = 30^\circ$ is therefore

$$G = \left| \sum_{m=1}^2 \sum_{n=1}^8 w_{m,n} v_{m,n} \right|^2 = (3.852)^2 = 14.8485. \quad (\text{C.30})$$

C.3 Calculation of G at $\varphi = -30^\circ$

The array gain is defined as

$$G = \left| \sum_{m=1}^M \sum_{n=1}^N w_{m,n} v_{m,n} \right|^2. \quad (\text{C.31})$$

The complex excitation weights are given by

$$w_{m,n} = \frac{1}{\sqrt{MN}} \exp \left\{ j2\pi \left[\frac{(m-1)d_v}{\lambda} \sin \theta_{etilt} - \frac{(n-1)d_h}{\lambda} \cos \theta_{etilt} \sin \varphi_{escan} \right] \right\}, \quad (\text{C.32})$$

and the phase response of the array elements by

$$v_{m,n} = \exp \left\{ j2\pi \left[\frac{(m-1)d_v}{\lambda} \cos \theta + \frac{(n-1)d_h}{\lambda} \sin \theta \sin \varphi \right] \right\}. \quad (\text{C.33})$$

The product of the excitation weight and phase response can be written as

$$w_{m,n} v_{m,n} = \frac{1}{\sqrt{MN}} \exp(j2\pi(m-1)A) \exp(j2\pi(n-1)B), \quad (\text{C.34})$$

where

$$A = \frac{d_v}{\lambda} (\sin \theta_{etilt} + \cos \theta), \quad (C.35)$$

$$B = \frac{d_h}{\lambda} (-\cos \theta_{etilt} \sin \varphi_{escan} + \sin \theta \sin \varphi). \quad (C.36)$$

The numerical parameters used in the evaluation are

$$M = 2, \quad N = 8, \quad (C.37)$$

$$d_v = 4.2\lambda, \quad d_h = 0.5\lambda, \quad (C.38)$$

$$\theta_{etilt} = 6^\circ, \quad \theta = 96^\circ, \quad (C.39)$$

$$\varphi_{escan} = -30^\circ, \quad \varphi = -30^\circ. \quad (C.40)$$

Substituting the parameters into the vertical phase coefficient yields

$$A = 4.2 (\sin 6^\circ + \cos 96^\circ). \quad (C.41)$$

Using the trigonometric identity

$$\cos(90^\circ + \alpha) = -\sin \alpha, \quad (C.42)$$

it follows that

$$\cos 96^\circ = -\sin 6^\circ, \quad (C.43)$$

and therefore

$$A = 0. \quad (C.44)$$

Consequently,

$$\exp(j2\pi(m-1)A) = 1, \quad (C.45)$$

and the vertical summation is fully coherent.

Substituting the parameters into the horizontal phase coefficient gives

$$B = 0.5 (-\cos 6^\circ \sin(-30^\circ) + \sin 96^\circ \sin(-30^\circ)). \quad (C.46)$$

Using the odd symmetry of the sine function,

$$\sin(-\alpha) = -\sin \alpha, \quad (C.47)$$

the above expression becomes

$$B = 0.5 (\cos 6^\circ \sin 30^\circ - \sin 96^\circ \sin 30^\circ), \quad (C.48)$$

which is the negative of the corresponding coefficient obtained for $\varphi_{escan} = \varphi = 30^\circ$.

The horizontal array factor is

$$S_N = \sum_{n=1}^N \exp(j2\pi(n-1)B). \quad (C.49)$$

Its magnitude is given by

$$|S_N| = \left| \frac{\sin(\pi NB)}{\sin(\pi B)} \right|. \quad (C.50)$$

Since this expression depends only on the absolute value of B ,

$$|S_N(B)| = |S_N(-B)|, \quad (\text{C.51})$$

and therefore, for $N = 8$,

$$|S_8| = 7.699. \quad (\text{C.52})$$

Including the normalization factor $1/\sqrt{MN} = 1/4$ and the vertical summation over $M = 2$ elements, the magnitude of the total array response is

$$\left| \sum_{m=1}^2 \sum_{n=1}^8 w_{m,n} v_{m,n} \right| = 2 \cdot \frac{1}{4} \cdot 7.699 = 3.852. \quad (\text{C.53})$$

The array gain at $\varphi = -30^\circ$ is therefore

$$G = \left| \sum_{m=1}^2 \sum_{n=1}^8 w_{m,n} v_{m,n} \right|^2 = (3.852)^2 = 14.8485. \quad (\text{C.54})$$

C.4 Calculation of $|\tilde{x}_m|^2 \tilde{x}_m$

$$\tilde{x}_m = \tilde{u}_1 e^{j\phi_{m1}} + \tilde{u}_2 e^{j\phi_{m2}}$$

$$\tilde{x}_m^* = \tilde{u}_1^* e^{-j\phi_{m1}} + \tilde{u}_2^* e^{-j\phi_{m2}}$$

$$|\tilde{x}_m|^2 = \tilde{x}_m \tilde{x}_m^*$$

$$= \left(\tilde{u}_1 e^{j\phi_{m1}} + \tilde{u}_2 e^{j\phi_{m2}} \right) \left(\tilde{u}_1^* e^{-j\phi_{m1}} + \tilde{u}_2^* e^{-j\phi_{m2}} \right)$$

$$= \tilde{u}_1 \tilde{u}_1^* + \tilde{u}_1 \tilde{u}_2^* e^{j(\phi_{m1} - \phi_{m2})} + \tilde{u}_2 \tilde{u}_1^* e^{j(\phi_{m2} - \phi_{m1})} + \tilde{u}_2 \tilde{u}_2^*$$

$$= |\tilde{u}_1|^2 + |\tilde{u}_2|^2 + \tilde{u}_1 \tilde{u}_2^* e^{j(\phi_{m1} - \phi_{m2})} + \tilde{u}_2 \tilde{u}_1^* e^{j(\phi_{m2} - \phi_{m1})}$$

$$|\tilde{x}_m|^2 \tilde{x}_m = \left(|\tilde{u}_1|^2 + |\tilde{u}_2|^2 + \tilde{u}_1 \tilde{u}_2^* e^{j(\phi_{m1} - \phi_{m2})} + \tilde{u}_2 \tilde{u}_1^* e^{j(\phi_{m2} - \phi_{m1})} \right) \times \left(\tilde{u}_1 e^{j\phi_{m1}} + \tilde{u}_2 e^{j\phi_{m2}} \right)$$

$$= |\tilde{u}_1|^2 \tilde{u}_1 e^{j\phi_{m1}} + |\tilde{u}_2|^2 \tilde{u}_1 e^{j\phi_{m1}} + \tilde{u}_1^2 \tilde{u}_2^* e^{j(2\phi_{m1} - \phi_{m2})}$$

$$+ |\tilde{u}_1|^2 \tilde{u}_2 e^{j\phi_{m2}} + |\tilde{u}_1|^2 \tilde{u}_2 e^{j\phi_{m2}} + |\tilde{u}_2|^2 \tilde{u}_2 e^{j\phi_{m2}}$$

$$+ |\tilde{u}_2|^2 \tilde{u}_1 e^{j\phi_{m1}} + \tilde{u}_2^2 \tilde{u}_1^* e^{j(2\phi_{m2} - \phi_{m1})}$$

$$= \left(|\tilde{u}_1|^2 \tilde{u}_1 + 2|\tilde{u}_2|^2 \tilde{u}_1 \right) e^{j\phi_{m1}} + \left(|\tilde{u}_2|^2 \tilde{u}_2 + 2|\tilde{u}_1|^2 \tilde{u}_2 \right) e^{j\phi_{m2}}$$

$$+ \tilde{u}_1^2 \tilde{u}_2^* e^{j(2\phi_{m1} - \phi_{m2})} + \tilde{u}_2^2 \tilde{u}_1^* e^{j(2\phi_{m2} - \phi_{m1})}$$

Appendix D

ORAN FH integration tutorial

D.1 ORAN FH integration

D.1.1 OAI gNB configuration file

O-RU Parameter	Configured Value	Description
pdsch_AntennaPorts_XP	1	number of enabled DL layers
pusch_AntennaPorts	1	number of enabled UL layers
absoluteFrequencySSB	650016	ARFCN parameter 3750.24 MHz
dl_frequencyBand	78	band n78
dl_absoluteFrequencyPointA	646724	reference point A 3700.86 MHz
dl_subcarrierSpacing	1	DL subcarrier spacing 30 KHz
dl_carrierBandwidth	273	number of PRBs
nb_tx	1	number of DL ports
nb_rx	1	number of UL ports
ru_addr	("00:11:22:33:44:88", "00:11:22:33:44:88")	C plane and U plane MAC addresses
mtu	9600	Maximum transmission unit
iq_width	9	IQ samples compression of size 9 bits
dpdk_devices	("0000:18:0a.0", "0000:18:0a.1")	C plane and U plane VF bus IDs

Table D.1: O-RU Configuration Parameters

D.1.2 CPU allocation

The server used as a DU is an Intel (R) Gold 6354 and an operating Central Processing Unit (CPU) at 3 GHz. The operating system is Ubuntu Debian 22.04.4. It has only one NUMA node that includes 17 cores.

DPDK, OAI, and kernel threads must be properly allocated to extract maximum real-time performance for your use case.

Note: Based on the Current Configuration

- The default OAI 7.2 configuration file requires isolated CPUs **0, 1, 2** for DPDK/libXRAN, CPU **5** for ru_thread, CPU **3** for L1_rx_thread, and CPU **4** for L1_tx_thread.
- Allocating CPUs to the OAI nr-softmodem is done using the `-thread-pool` option. Allocating **4 CPUs** is the minimal configuration, but we recommend allocating at least **8 CPUs**. These threads can be on a different socket than the DPDK threads.
- To avoid the kernel preempting these allocated CPUs, it is better to force the kernel to use unallocated CPUs (e.g., CPU **16,17** for K_thread).

Applicative Threads	Allocated CPUs
XRAN DPDK usage	0,1,2
OAI L1_rx_thread	3
OAI L1_tx_thread	4
OAI ru_thread	5
OAI nr-softmodem	6,7,8,9
K_thread	16,17

Table D.2: CPU Allocation for Applicative Threads

Modifying the linux command line usually requires editing string GRUB_CMDLINE_LINUX in `/etc/default/grub`, we ran a grub command and rebooted the server:

```
GRUB_CMDLINE_LINUX="skew_tick=1 isolcpus=0-15 nohz_full=0-15 rcu_nocbs=0-15
kthread_cpus=16,17 default_hugepagesz=1GB hugepagesz=1G hugepages=20
rcu_nocb_poll nosoftlockup intel_iommu=on iommu=pt"
```

We apply the same changes in `/proc/cmdline`.

For that, we apply these changes as well to the OAI gNB config file in the L1s section for L1_rx_thread_core and L1_tx_thread_core:

```
L1s = (
{
  num_cc = 1;
  tr_n_preference = "local_mac";
  prach_dtx_threshold = 130;
  pucch0_dtx_threshold = 80;
  pusch_dtx_threshold = 10;
  tx_amp_backoff_dB = 9; # needs to match O-RU configuration
  L1_rx_thread_core = 3;
  L1_tx_thread_core = 4;
  phase_compensation = 0; # needs to match O-RU configuration
}
```

and in the fhi_72 we apply the changes for system_core, io_core and worker_cores that represent the 3 different cores of the XRAN library:

```
fhi_72 = {
  dpdk_devices = ("0000:18:0a.0", "0000:18:0a.1");
  system_core = 0;
  io_core = 1;
  worker_cores = (2);
  ru_addr = ("00:11:22:33:44:88", "00:11:22:33:44:88");
  mtu = 9600;
```

```
fh_config = ({
    T1a_cp_dl = (0,200) #(380, 400); #(380,400) #(419, 470);
    T1a_cp_ul = (250,250); # (285, 336);
    T1a_up = (125, 300); #(294, 345);
    Ta4 = (0, 150);
    # T1a_cp_dl = (250, 500); # (100, 350);
    # T1a_cp_ul = (250, 500); # (250, 450); # (150, 200); # (min, max)
    # T1a_up = (125, 300); # (200, 500); #(200, 250); # (200, 250); # (min, max)
    # Ta4 = (0, 150); # (0, 100); # (0, 100); # (min, max)
    ru_config = {
        iq_width = 9;
        iq_width_prach = 9;
    };
    prach_config = {
        eAxC_offset = 32;
        #kbar = 0;
    };
});
};
```

D.1.3 PTP configuration

With the O-RU, PTP is required for accurate synchronization.

The `linuxptp` package, which includes the utilities `ptp4l` and `phc2sys`, are installed:

```
sudo apt install linuxptp
```

PTP configuration is done via the file `/etc/ptp4l.conf` to adjust the clock domain. For this setup, the configuration followed the tutorial, with the network interface parameter `[your_PTP_ENABLED_NIC]` replaced by `[ens1f2]` to match the system's interface.

This ensures that the O-RU system is properly synchronized using PTP according to the recommended tutorial settings.

D.1.4 DPDK compilation and installation

DPDK provides libraries and drivers to accelerate packet processing by bypassing the kernel, enabling high-performance, low-latency networking. It accelerates U plane packets in user space while C plane packets are typically handled by the kernel. Additionally, DPDK manages VFs by binding them to its drivers, allowing fast, dedicated NIC access for virtualized workloads.

we installed the DPDK using:

```
sudo apt install meson
meson build
ninja -C build
sudo ninja install -C build
```

We wrote the following script to setup the sriov and to bind the DPDK to the NIC:

```
set -x
# Setting up the server to performance mode
sudo cpupower idle-set -D 0
sudo ethtool -G ens1f2 rx 4096
sudo ethtool -G ens1f2 tx 4096
sudo ethtool -g ens1f2
sudo ifconfig ens1f2 mtu 9000
sudo modprobe iavf
sudo sh -c 'echo "0" > /sys/class/net/ens1f2/device/sriov_numvfs'
```

```

sudo sh -c 'echo "2" > /sys/class/net/ens1f2/device/sriov_numvfs'
sudo ip link set ens1f2 vf 0 mac 00:11:22:33:44:77 vlan 14 spoofchk off mtu 9000
sudo ip link set ens1f2 vf 1 mac 00:11:22:33:44:77 vlan 13 spoofchk off mtu 9000
sudo /usr/local/bin/dpdk-devbind.py --unbind 0000:18:0a.0
sudo /usr/local/bin/dpdk-devbind.py --unbind 0000:18:0a.1
sudo modprobe vfio_pci
sudo /usr/local/bin/dpdk-devbind.py --bind vfio-pci 0000:18:0a.0
sudo /usr/local/bin/dpdk-devbind.py --bind vfio-pci 0000:18:0a.1

```

- **SR-IOV Setup:** The script creates 2 VFs on the NIC `ens1f2` by first resetting the number of VFs to 0, then setting it to 2. VF 0 is assigned MAC `00:11:22:33:44:77`, **VLAN 14** for U plane packets, and MTU 9000, while VF 1 is assigned the same MAC, **VLAN 13** for C plane packets, and MTU 9000. This allows each VF to act as an independent virtual NIC with its own network configuration.
- **DPDK Binding:** The physical functions of the NIC (`0000:18:0a.0` and `0000:18:0a.1`) are unbound from the kernel and bound to the `vfio-pci` driver using `dpdk-devbind.py`, enabling DPDK to directly access these NIC resources for high-speed, low-latency packet processing.

D.1.5 Build the ORAN FH library

1. Clone the O-RAN O-DU repository:

```

git clone https://gitlab.eurecom.fr/oai/openairinterface5g.git ~/
  openairinterface5g
cd ~/openairinterface5g/

```

2. F release

```

git clone https://gerrit.o-ran-sc.org/r/o-du/phy.git ~/phy
cd ~/phy
git checkout oran_f_release_v1.0
git apply ~/openairinterface5g/cmake_targets/tools/oran_fhi_integration_patches
  /F/oaioran_F.patch

```

3. Build the O-DU library using XTRAN shared objects with the correct DPDK path for the F release:

```

WIRELESS_SDK_TOOLCHAIN=gcc \
RTE_SDK=~/.dpdk-stable-20.11.9/ \
XTRAN_DIR=~/.phy/fhi_lib \
make XTRAN_LIB_SO=1

```

D.1.6 Build OAI gNB

The tutorial provides a script `build_oai` but it requires needed to input the fronthaul lib location:

```

./build_oai --gNB --ninja -t oran_fhlib_5g --cmake-opt -Dxran_LOCATION=$HOME/phy/
  fhi_lib/lib

```

D.1.7 Run OAI gNB

We run OAI in Phy-test mode by running `nr-softmodem` and by assigning the corresponding options enabling full downlink transmission:

```

sudo ./nr-softmodem -O <configuration file> --thread-pool 6,7,8,9 --phy-test -M 273
  -T 273 -l 1 -L 1 -D 524287 -m 0 -t 0

```

phy-test options	Value
Downlink PRBs (-M)	273
Uplink PRBs (-T)	273
Number UL layers (_l)	1
Number DL layers (-L)	1
Downlink slots configuration (-D)	524287
Uplink slots configuration (-U)	0
DL MCS (-m)	0 (QPSK)
UL MCS (-t)	0 (QPSK)

Config File	Value
TDD periodicity	10 ms
TDD configuration pattern	19D1S
Subcarrier spacing	30kHz

Table D.3: PHY Test and gNB Configuration

The definition of the options are listed in table D.3.

In phy-test mode, the DU sends downlink signals using PDSCH, with data generated randomly. The main parameters used during the test are summarized in Table D.3. Specifically, all 273 Physical Resource Blocks (PRB)s are used for downlink, and a slot bitmap value of 524287 ensures that all 19 downlink slots in each 20 TDD slot frame are active. Uplink transmission is disabled. Both the downlink and uplink modulation settings are fixed to Quadrature Phase Shift Keying (QPSK) (Modulation and Coding Scheme (MCS) 0), which ensures consistent and predictable signal generation throughout the experiments.

The radio system is configured for TDD operation with a 10 ms frame duration and a slot pattern of 19 downlink slots followed by one special slot (19D1S). A subcarrier spacing of 30 kHz is used. For simplicity, the O-RAN C plane functionality is disabled on the RU side, meaning the RU does not process control messages from the DU. Instead, the beamforming behavior, specifically, the Beam ID is statically defined in the RU's configuration file. This setup allows for fixed, repeatable beam selection without requiring dynamic signaling during operation.

D.2 Integration testing

Testing the ORAN split 7.2 fronthaul interface is essential to ensure seamless interoperability between the DU and the Radio Unit RU. It verifies correct synchronization, data transport, and protocol compliance, ensuring that IQ samples and control information are exchanged reliably and within strict latency and timing budgets. Proper testing not only confirms functional integration but also guarantees performance, stability, and conformance to ORAN standards in a multi-vendor environment. In this section, we present a few testing policies that allowed us to validate the successful integration of the FH interface:

D.2.1 PTP logs

The commands `systemctl status ptp4l.service` and `systemctl status phc2sys` display the running state and timing statistics of the PTP synchronization services. The rms and max

values show how far your clock drifts from the master; lower numbers mean better synchronization. The freq value shows how much your local clock is adjusted to stay aligned, and the delay value indicates the network timing stability.

```
eurecom@colibri:~$ sudo systemctl status ptp41.service
    ptp41.service - Precision Time Protocol (PTP) service
Loaded: loaded (/lib/systemd/system/ptp41.service; enabled; vendor preset: enabled)
Active: active (running) since Mon 2025-11-17 15:58:58 CET; 2min 2s ago
Tasks: 3 (limit: 23299)
Memory: 284.0K
CPU: 331ms
CGroup: /system.slice/ptp41.service
        196210    /usr/sbin/ptp41 -f /etc/ptp41.conf

Nov 17 16:00:59 colibri.boot.cs.eurecom.fr ptp41[196210]: ptp41[72608106.485]: rms 4
        max 6 freq +10233 +/- 0 delay 1606 +/- 0
Nov 17 16:00:58 colibri.boot.cs.eurecom.fr ptp41[196210]: ptp41[72608106.485]: rms 4
        max 6 freq +10213 +/- 0 delay 1606 +/- 0
Nov 17 16:00:57 colibri.boot.cs.eurecom.fr ptp41[196210]: ptp41[72610707.416]: rms 4
        max 6 freq +10213 +/- 0 delay 1607 +/- 0
Nov 17 16:00:58 colibri.boot.cs.eurecom.fr ptp41[196210]: ptp41[72610707.416]: rms 4
        max 6 freq +10213 +/- 0 delay 1607 +/- 0
Nov 17 16:00:59 colibri.boot.cs.eurecom.fr ptp41[196210]: ptp41[72610707.416]: rms 4
        max 6 freq +10213 +/- 0 delay 1607 +/- 0
Nov 17 16:00:58 colibri.boot.cs.eurecom.fr ptp41[196210]: ptp41[72610707.416]: rms 4
        max 6 freq +10214 +/- 0 delay 1607 +/- 0
Nov 17 16:00:59 colibri.boot.cs.eurecom.fr ptp41[196210]: ptp41[72610816.436]: rms 4
        max 5 freq +10214 +/- 0 delay 1607 +/- 0
Nov 17 16:00:58 colibri.boot.cs.eurecom.fr ptp41[196210]: ptp41[72610816.436]: rms 4
        max 5 freq +10215 +/- 0 delay 1607 +/- 0
Nov 17 16:00:59 colibri.boot.cs.eurecom.fr ptp41[196210]: ptp41[72610816.436]: rms 4
        max 5 freq +10215 +/- 0 delay 1607 +/- 0
Nov 17 16:01:00 colibri.boot.cs.eurecom.fr ptp41[196210]: ptp41[72611018.686]: rms 4
        max 8 freq +10217 +/- 0 delay 1606 +/- 1
Nov 17 16:01:01 colibri.boot.cs.eurecom.fr ptp41[196210]: ptp41[72611018.686]: rms 4
        max 8 freq +10217 +/- 0 delay 1606 +/- 1
eurecom@colibri:~$ sudo systemctl status ptp41.service
    ptp41.service - Precision Time Protocol (PTP) service
Loaded: loaded (/lib/systemd/system/ptp41.service; enabled; vendor preset: enabled)
Active: active (running) since Mon 2025-11-17 15:58:58 CET; 2min 10s ago
Tasks: 3 (limit: 23299)
Memory: 284.0K
CPU: 331ms
CGroup: /system.slice/ptp41.service
        196210    /usr/sbin/ptp41 -f /etc/ptp41.conf

Nov 17 16:01:03 colibri.boot.cs.eurecom.fr ptp41[196210]: ptp41[726115.499]: rms 4
        max 7 freq +10227 +/- 4 delay 1607 +/- 0
Nov 17 16:01:04 colibri.boot.cs.eurecom.fr ptp41[196210]: ptp41[726115.499]: rms 4
        max 7 freq +10227 +/- 4 delay 1607 +/- 0
Nov 17 16:01:05 colibri.boot.cs.eurecom.fr ptp41[196210]: ptp41[726116.391]: rms 4
        max 7 freq +10228 +/- 0 delay 1607 +/- 0
Nov 17 16:01:06 colibri.boot.cs.eurecom.fr ptp41[196210]: ptp41[726116.391]: rms 3
        max 7 freq +10228 +/- 0 delay 1607 +/- 1
Nov 17 16:01:07 colibri.boot.cs.eurecom.fr ptp41[196210]: ptp41[726117.528]: rms 3
        max 5 freq +10229 +/- 0 delay 1606 +/- 1
Nov 17 16:01:08 colibri.boot.cs.eurecom.fr ptp41[196210]: ptp41[726117.528]: rms 3
        max 5 freq +10229 +/- 0 delay 1606 +/- 1
```



```

Nov 17 16:01:09 colibri.boot.cs.eurecom.fr ptp4l[196210]: ptp4l[726118.531]: rms 3
max 6 freq +10229 +/- 0 delay 1606 +/- 0
Nov 17 16:01:09 colibri.boot.cs.eurecom.fr ptp4l[196210]: ptp4l[726118.531]: rms 3
max 6 freq +10230 +/- 0 delay 1606 +/- 0
Nov 17 16:01:10 colibri.boot.cs.eurecom.fr ptp4l[196210]: ptp4l[726118.531]: rms 3
max 6 freq +10228 +/- 0 delay 1607 +/- 0
Nov 17 16:01:11 colibri.boot.cs.eurecom.fr ptp4l[196210]: ptp4l[726119.53]: rms 3
max 6 freq +10227 +/- 0 delay 1607 +/- 0

```

D.2.2 OAI transmission logs

After starting the gNB with the selected configuration file, if everything went well, we should see the RU (rx) and DU (tx) counters for go up with time, meaning that there is exchange of packets over time between the RU and the DU:

```

[NR_PHY] [o-du 0][rx 7168 pps 7168 kbps 205716][tx 31919 pps 31919 kbps
708187][Total Msgs_Rcvd 7168]
[NR_PHY] [o-du0][pusch0 7168 prach0 0]
[NR_PHY] [o-du 0][rx 14336 pps 7168 kbps 343168][tx 63791 pps 31872 kbps
1179492][Total Msgs_Rcvd 14336]
[NR_PHY] [o-du0][pusch0 14336 prach0 0]
[NR_PHY] [o-du 0][rx 21504 pps 7168 kbps 343168][tx 95663 pps 31872 kbps
1179492][Total Msgs_Rcvd 21504]
[NR_PHY] [o-du0][pusch0 21504 prach0 0]
[NR_PHY] [o-du 0][rx 28672 pps 7168 kbps 343168][tx 127535 pps 31872 kbps
1179492][Total Msgs_Rcvd 28672]
[NR_PHY] [o-du0][pusch0 28672 prach0 0]
[NR_PHY] [o-du 0][rx 35840 pps 7168 kbps 343168][tx 159407 pps 31872 kbps
1179492][Total Msgs_Rcvd 35840]
[NR_PHY] [o-du0][pusch0 35840 prach0 0]

```

Listing D.1: O-DU PHY Layer log output

D.2.3 RU PTP state

In the PTP **LOCKED** state, the RU is synchronized to a valid Grandmaster clock provided by the switch, ensuring precise timing and alignment with the DU. This state indicates active synchronization, but accuracy is still monitored, as a **LOCKED** node may occasionally drift beyond required limits. Multiple Grandmasters and short **HOLDOVER** periods help maintain stable operation during timing source changes[34]. By checking the PTP status on the RU level, we successfully had the PTP in **LOCKED** state:

```

[PTP Status]
[MasterOffset]=-13 [port_state]=SLAVE [ptp_locked_state]=LOCKED [synch_state]=LOCKED
CURRENT_DATA_SET
  stepsRemoved      2
  offsetFromMaster -13
  meanPathDelay     532
DEFAULT_DATA_SET
  twoStepFlag       1
  slaveOnly          1
  numberPorts       1
  priority1         128
  clockClass        255
  clockAccuracy      0xfe
  offsetScaledLogVariance 0xffff
  priority2         128

```

```

    clockIdentity      000dee.ffffe.128006
    domainNumber      24
PARENT_DATA_SET
    parentPortIdentity      4006d5.ffffe.dba6b3-165
    parentStats              0
    observedParentOffsetScaledLogVariance 0xffff
    observedParentClockPhaseChangeRate    0x7fffffff
    grandmasterPriority1      128
    gm.ClockClass              6
    gm.ClockAccuracy           0x21
    gm.OffsetScaledLogVariance 0x00ff
    grandmasterPriority2      128
    grandmasterIdentity       fcac6a.ffffe.02bc86
TIME_PROPERTIES_DATA_SET
    currentUtcOffset      37
    leap61                 0
    leap59                 0
    currentUtcOffsetValid 1
    ptpTimescale           1
    timeTraceable          1
    frequencyTraceable     1
    timeSource              0x20
PORT_DATA_SET
    portIdentity           000dee.ffffe.128006-1
    portState               SLAVE
    logMinDelayReqInterval -4
    peerMeanPathDelay       0
    logAnnounceInterval     1
    announceReceiptTimeout  3
    logSyncInterval         0
    delayMechanism           1
    logMinPdelayReqInterval 0
    versionNumber           2

```

Listing D.2: PTP Status output from O-RU

D.2.4 RU ORAN packet counters

Monitoring the RU RAN counters is essential to verify proper data flow, synchronization accuracy, and protocol compliance on the fronthaul interface. These counters help detect packet losses, timing errors, or mismatched configurations between the RU and DU:

```

root@javelin-fh:~# oran -d 0 -p
[du_mac]=0x1122334456
[ru_mac]=0xdee1280e0
[rst_cnt]=0
[da_en]=1
[vlan_en]=2
[up_cnt_rx_ontime]=27171
[up_cnt_rx_early]=0
[up_cnt_rx_late]=0
[up_cnt_rx_total]=27177
[cp_cnt_rx_ontime]=2264
[cp_cnt_rx_early]=0
[cp_cnt_rx_late]=0
[cp_cnt_rx_total]=2265
[up_cnt_tx_total]=1692
[up_pcp]=0
[up_dei]=0

```

```
[up_vtag]=14
[cp_pcp]=0
[cp_dei]=0
[cp_vtag]=13
[mtu]=9000
```

Listing D.3: O-RU O-RAN counter output

D.2.5 IQ samples buffer with respect to time

The **Capture Buffer** displays the full range of data recorded during the most recent capture. These RF data are visualized on the 5GNR FSVA spectrum analyzer application. The capture buffer processes the data using an auto peak detector, which evaluates each measurement point by determining both the maximum and minimum signal levels. These two values are then combined into a single sample point for display [78].

- The **x-axis** corresponds to time, with its maximum value equal to the Capture Time.
- The **y-axis** shows the amplitude of the captured I/Q data in dBm (for RF input).

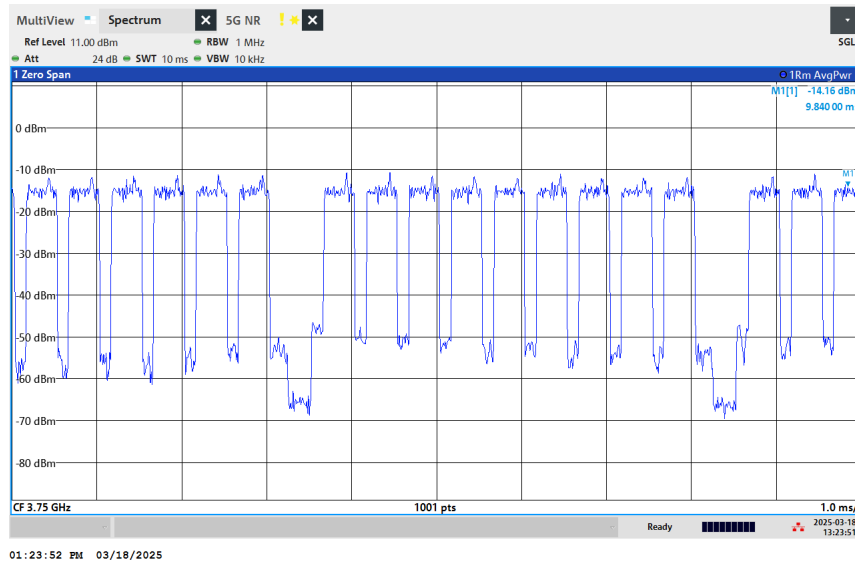


Figure D.1: Representation of the platform architecture

As we can see in the buffer captured in Figure D.1, we see the IQ samples grouped into consecutive time slots. We can observe 20 different time slots that align perfectly with the selected numerology and the chosen TDD periodicity 19D1S, 18 of these slots represent DL transmitted IQ samples, one is a special slot and there is one that is left unallocated due to the fact that OAI schedules only the SSB in slot 0 and leaves it empty of any PDSCH data.

D.2.6 Signal spectrum in function of the frequency

The "Power Spectrum" shows the power density of the complete capture buffer in dBm/Hz. The displayed bandwidth depends on the channel bandwidth. The x-axis represents the frequency. On the y-axis, the power level is plotted [78].

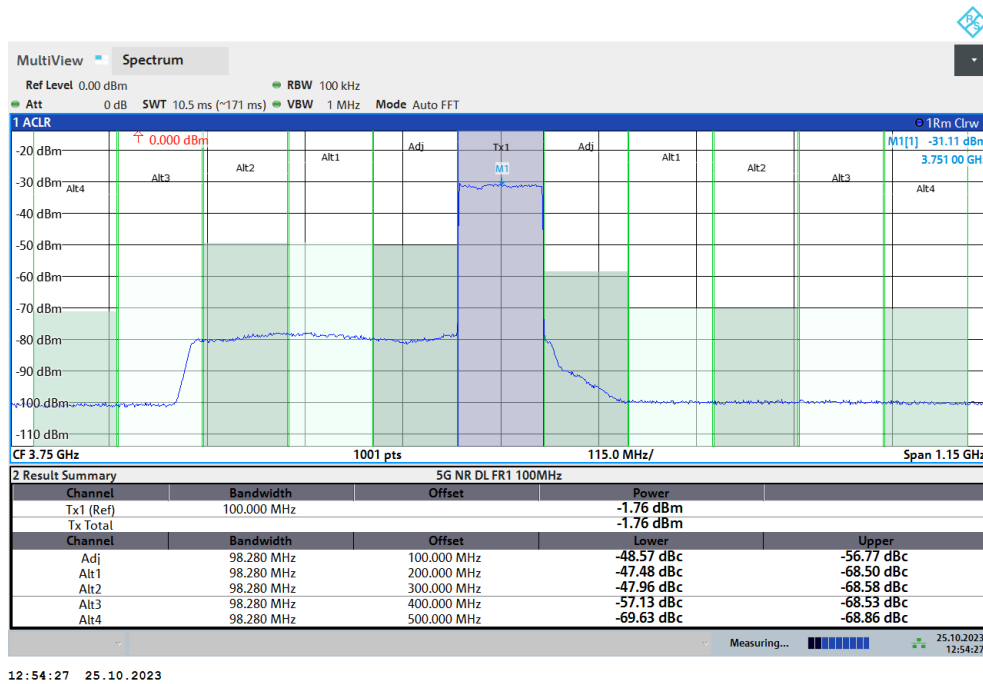


Figure D.2: Representation of the platform architecture

Figure D.2 shows The spectrum analysis that illustrates a prominent signal centered around 3.75 GHz, which falls within the n78 band limits of 3.4 GHz to 3.8 GHz. The peak power of this signal is approximately -31.11 dBm, indicating a strong transmission at this frequency. The signal's bandwidth is roughly 98.28 MHz, consistent with a typical 5 GHz channel and as defined in the OAI config file. The spectrum does not extend beyond the upper limit of the n78 band, which confirms that the signal remains within the authorized frequency range. Further tuning in the RBW, Video Bandwidth (VBW) and RU calibration allowed one to have more transmit power that was found to be consistent with the true RU's transmit power; however, this test was done only to check the coherence of the FH integration from the RF perspective.

Appendix E

Implementations in OAI

E.1 nr-beamforming()

```
/* NR Beamforming: Apply per-port precoding to per-antenna signals */
void nr_beam_precoding(cl6_t **txdataF,
                      cl6_t **txdataF_BF,
                      NR_DL_FRAME_PARMS *frame_parms,
                      int32_t ***beam_weights,
                      int slot,
                      int symbol,
                      int aa,
                      int nb_antenna_ports,
                      int offset)
{
    // Clear the output buffer for this antenna and symbol
    memset(&txdataF_BF[aa][symbol*frame_parms->ofdm_symbol_size], 0,
          sizeof(cl6_t) *(frame_parms->ofdm_symbol_size));

    // Loop over all logical antenna ports
    for (int p = 0; p < nb_antenna_ports; p++) {
        // Complex multiply-and-accumulate: x_aa += w_{p,aa} * s_p
        multadd_cpx_vector(
            (cl6_t *)&txdataF[p][(symbol * frame_parms->ofdm_symbol_size) + offset],
            (cl6_t *)beam_weights[p][aa],
            (cl6_t *)&txdataF_BF[aa][symbol * frame_parms->ofdm_symbol_size],
            frame_parms->ofdm_symbol_size,
            15);
    }
}
```

Appendix F

5G-NR application

The 5G NR application is a firmware-based extension for the FSW spectrum analyzer that enables downlink signal measurements in compliance with the 3GPP 5G New Radio NR standard. The application supports features defined up to 3GPP Release 15[78].

It provides comprehensive functionality for the analysis of 5G NR signals, including:

- Measurement of fundamental 5G NR signal characteristics, such as multiple component carriers, supported frequency ranges, and channel bandwidths.
- Automatic demodulation and configuration of the PDSCH and SSB/PBCH.
- Configuration and analysis of multiple frames and bandwidth parts, supporting multi-numerology operation.
- Configuration and analysis of dedicated downlink channels and reference signals, including PDCCH and CSI-RS.
- Configuration and analysis of various demodulation reference signals.
- Mapping and analysis of channels across different antenna ports.
- Analysis of LTE–NR coexistence scenarios.
- Combined evaluation of key performance metrics, including Error Vector Magnitude (EVM), ACLR, and SEM, within a single integrated application.

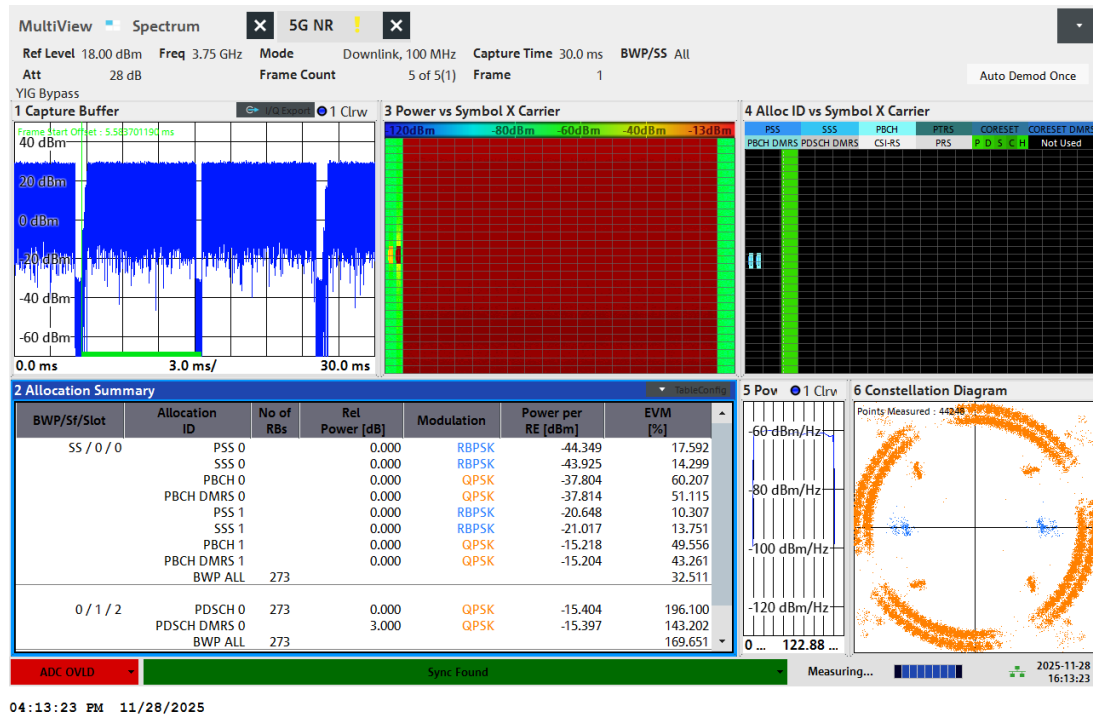
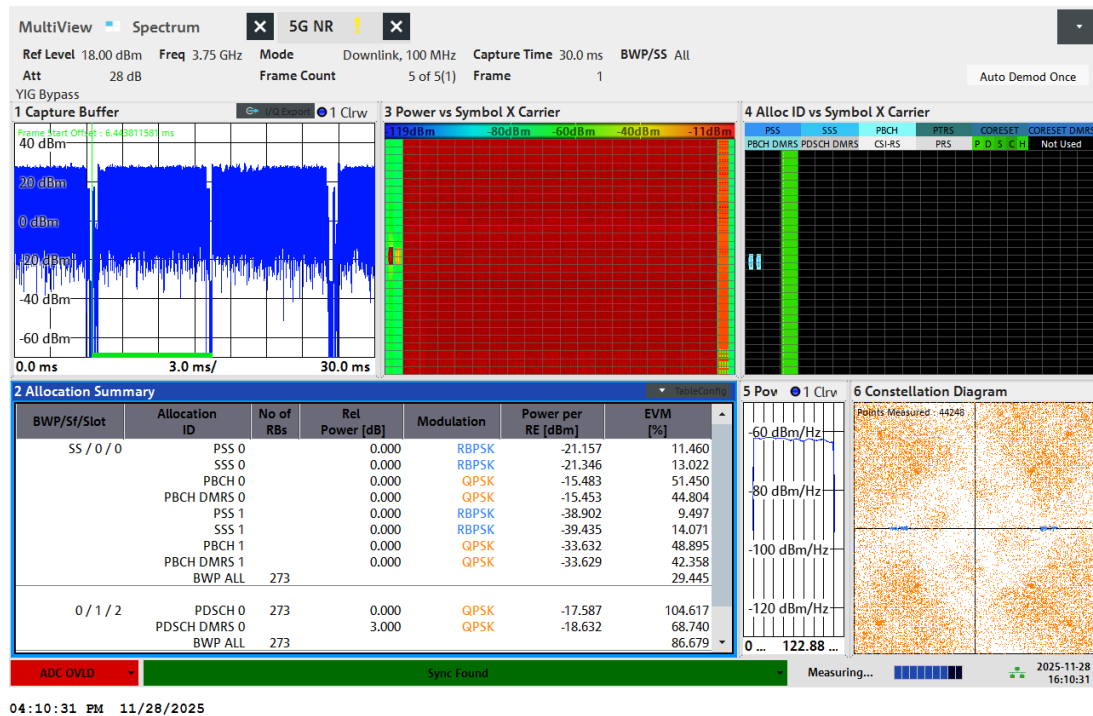
The different components of the 5G-NR application are illustrated in Figures F.1 and F.2 and are the following:

F.1 Capture Buffer

The **Capture Buffer** displays the complete time range of data recorded during the most recent capture. The x-axis represents time, with its maximum value corresponding to the configured *Capture Time*. The y-axis shows the amplitude of the captured I/Q data, expressed in dBm for RF input signals.

The measurement data is evaluated using an *auto peak detector*. This detector determines the maximum and minimum measured levels at each measurement point and combines both values into a single sample point.

At the bottom of the diagram, a green bar indicates the frame that is currently under analysis. A green vertical line at the beginning of this bar marks the frame start. In addition, the diagram displays the *Start Offset*, which represents the time difference between the start of the capture buffer and the start of the frame.

Figure F.1: 5G-NR application extracting 5G-NR data at -30° Figure F.2: 5G-NR application extracting 5G-NR data at 30°

In our case in both figures, we can see how each frame of durations 10 ms, allocates high power IQ samples for 18 consecutive DL slots, the first slot is left empty and has no allocated DL IQ samples since

it is reserved for SSB transmission, and the last DL slot is left empty which represents the mixed slot in the TDD configuration used 19D1S.

F.2 Power vs. Symbol $\times$ Carrier

The Power vs. Symbol $\times$ Carrier display presents the power level of each carrier for every symbol. The x-axis represents the symbols, while the y-axis represents the subcarriers.

Power levels are visualized using different colors within the diagram area. A color map located in the diagram header indicates the corresponding power values associated with each color. As we can see in both figures, the first slot only has high powers where both SSB subcarriers were detected in the resource grid and the rest of the slot the subcarriers are empty, and the next consecutive 18 DL slots we can observe the high power indication due to the PDSCH subcarrier allocation in the slots and finally the special slot is powerless since it is empty.

F.3 Allocation Summary

The **Allocation Summary** provides a tabular overview of various parameters related to the measured allocations. Each row in the table corresponds to a single allocation. Multiple allocations together form a slot, and a horizontal line marks the beginning of a new slot.

Special summary allocations are included to represent aggregated characteristics of all allocations within a bandwidth part (**BWP ALL**) and across the entire radio frame (**TOTAL ALL**).

The table columns provide the following information for each allocation:

- Allocation location (slot, subframe, and bandwidth part number)
- Allocation ID (channel type)
- Number of resource blocks used
- Relative allocation power in dB
- Modulation scheme
- Power per resource element in dBm
- EVM of the allocation (unit depends on the selected EVM unit)

F.4 Constellation Diagram

The Constellation Diagram displays the IQ components of the signal and serves as an indicator of modulation quality.

The colors in the constellation diagram represent either the modulation type or the allocation type, depending on the selected configuration. The color mapping for modulation types is defined as follows, while the color mapping for allocation types matches that used in the *Allocation ID vs. Symbol vs. Carrier* display:

- RBPSK: Orange
- QPSK: Blue

The results displayed can be filtered by adjusting the evaluation range. The constellation diagram also indicates the total number of constellation points currently displayed.

We can observe in the constellation, the 2 binary points in blue at -1 and 1 which represent the RBSK relative to both SSBs. The main QPSK signal forms the primary constellation with its ideal symbol points in the I-Q plane, and in the presence of inter-user interference, each symbol of the main QPSK constellation is affected by additional QPSK signals from other users. As a result, each main constellation point is surrounded by a small cluster of points that represent the possible interference-induced shifts from the second user. The received IQ samples therefore form a “constellation of constellations” where the central circle corresponds to the main signal of higher power in the intended direction, and the surrounding mini-QPSK points reflect the inter-user interference. This illustrates how interference spreads the ideal symbol locations into distinct clusters around the original QPSK points. Observing Figure F.1, we can see a circle of IQ samples, this circle is a result of phase noise, leading to a phase shift in the affected IQ samples. Observing Figure F.2, we can see a cloud of IQ samples, this cloud is a result of noise.

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