

# Design and holistic evaluation of dual-band circularly polarized smartphone antenna for direct-to-device non-terrestrial networks

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MASTER'S THESIS

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Design and holistic evaluation of dual-band  
circularly polarized smartphone antenna for  
direct-to-device non-terrestrial networks

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# Abstract

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Direct-to-Device (D2D) communication in Non-Terrestrial Networks (NTNs) enables smartphones to connect directly to satellites, extending wireless coverage beyond terrestrial networks. This thesis presents the design of a smartphone antenna optimized for NTN operation, followed by a link budget analysis and Electromagnetic Field (EMF) exposure assessment.

The proposed solution is a dual-band circularly polarized antenna operating in the S-band NTN frequency ranges. The antenna is specifically optimized to provide radiation towards the upper hemisphere, while satisfying the size constraints of a smartphone. In addition, a linearly polarized C-band antenna is obtained as an extension of the S-band structure, enabling additional functionality. The antenna performance is evaluated in terms of impedance matching, radiation characteristics and efficiency, and its impact on system performance is evaluated by a link budget analysis based on a Low Earth Orbit (LEO) satellite scenario. The results demonstrate that the optimized S-band antenna achieves data rates ranging from several kbps to several Mbps, depending on transmit power and elevation angle, demonstrating its suitability to enable NTN communication with significant data rates. In the C-band, the system can provide additional connectivity, supporting low data rate services, due to higher propagation losses and reduced antenna gain in the skywards direction. Electromagnetic exposure is assessed through Specific Absorption Rate (SAR) simulations. The results show that SAR levels remain within international limits when using certain duty cycle transmission period, needed at certain phone positions relative to the body and higher transmit power levels.

The findings provide insights into the trade-offs between antenna design, link performance, and user exposure in the context of smartphone-based NTN communication.



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## Popular Science Summary

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Today's mobile networks allow people to stay connected in many places around the globe, but some regions still lack reliable communication. This can be particularly problematic in emergency situations, where a connection may be critical to safety.

A promising solution to this challenge is to use satellites that provide communication services directly to smartphones. Instead of relying on ground base stations, a phone could connect directly to a satellite orbiting the Earth, enabling coverage in any location on the planet. However, in order to make this possible, improved antennas in the smartphones are required, as the current systems are not designed to communicate over such large distances between the user and a satellite.

In this thesis, a specific type of antenna was designed to enable connectivity between the smartphone and the satellite. The antenna is designed to fit within the limited space of a modern smartphone, while directing the signal towards the sky, where the satellites are located. In addition, the amount of information that can be transmitted over this link was analyzed, and electromagnetic field exposure is evaluated as it is needed to ensure compliance with limit values.

The results show that it is possible for a smartphone to establish a connection with a satellite, particularly at certain frequencies that are specifically reserved for this use. However, the large distance between the phone and the satellite leads to a very weak signal, making a robust antenna design essential.

Overall, this work demonstrates that direct satellite communication with modern smartphones is technically feasible, but it requires a careful balance between performance and reliability.



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## Declaration

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I declare that this thesis is the result of my own work. Except where specific references are made, the content is original and my own. In order to improve the clarity and quality of the language, I have used AI-based tools for text refinement and proofreading. All analysis, results, and conclusions presented in this thesis are my own, and I take full responsibility for the work submitted.



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## List of Acronyms

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|               |                                                               |
|---------------|---------------------------------------------------------------|
| <b>FSS</b>    | Fixed-Satellite Services                                      |
| <b>BSS</b>    | Broadcasting-Satellite Services                               |
| <b>GNSS</b>   | Global Navigation Satellite Systems                           |
| <b>MSS</b>    | Mobile-Satellite Services                                     |
| <b>NTN</b>    | Non-Terrestrial Network                                       |
| <b>LEO</b>    | Low Earth Orbit                                               |
| <b>MEO</b>    | Medium Earth Orbit                                            |
| <b>GEO</b>    | Geostationary Earth Orbit                                     |
| <b>HEO</b>    | Highly Elliptical Orbit                                       |
| <b>HAPS</b>   | High-Altitude Platform Stations                               |
| <b>UE</b>     | User Equipment                                                |
| <b>D2D</b>    | Direct-to-Device                                              |
| <b>IoT</b>    | Internet of Things                                            |
| <b>EMF</b>    | Electromagnetic Field                                         |
| <b>SAR</b>    | Specific Absorption Rate                                      |
| <b>ICNIRP</b> | International Commission on Non-Ionizing Radiation Protection |
| <b>FCC</b>    | Federal Communications Commission                             |
| <b>3GPP</b>   | Third Generation Partnership Project                          |
| <b>ETSI</b>   | European Telecommunication Standards Institute                |
| <b>NR</b>     | New Radio                                                     |
| <b>FR</b>     | Frequency Range                                               |
| <b>6G</b>     | Sixth Generation                                              |
| <b>LB</b>     | Low Band                                                      |

|             |                                                   |
|-------------|---------------------------------------------------|
| <b>MHB</b>  | Mid and High Band                                 |
| <b>WiFi</b> | Wireless Fidelity                                 |
| <b>NFC</b>  | Near Field Communications                         |
| <b>UWB</b>  | Ultra Wide Band                                   |
| <b>IFA</b>  | Inverted F Antenna                                |
| <b>PIFA</b> | Planar Inverted F Antenna                         |
| <b>UL</b>   | Uplink                                            |
| <b>DL</b>   | Downlink                                          |
| <b>AR</b>   | Axial Ratio                                       |
| <b>CP</b>   | Circular Polarization                             |
| <b>Tx</b>   | Transmitter                                       |
| <b>Rx</b>   | Receiver                                          |
| <b>gNB</b>  | gNodeB                                            |
| <b>RF</b>   | Radio Frequency                                   |
| <b>IEEE</b> | Institute of Electrical and Electronics Engineers |
| <b>LoS</b>  | Line-of-Sight                                     |
| <b>FSPL</b> | Free-Space Path Loss                              |
| <b>HPBW</b> | Half Power Beam Width                             |
| <b>TN</b>   | Terrestrial Network                               |
| <b>PCB</b>  | Printed Circuit Board                             |
| <b>RTT</b>  | Round-Trip Time                                   |
| <b>BER</b>  | Bit Error Rate                                    |
| <b>SAM</b>  | Specific Anthropomorphic Mannequin                |
| <b>CDF</b>  | Cumulative Distribution Function                  |
| <b>DC</b>   | Duty Cycle                                        |
| <b>AWGN</b> | Additive White Gaussian Noise                     |
| <b>MCS</b>  | Modulation and Coding Scheme                      |
| <b>LHCP</b> | Left-Hand Circular Polarization                   |
| <b>RHCP</b> | Right-Hand Circular Polarization                  |
| <b>PSK</b>  | Phase-Shift Keying                                |
| <b>BPSK</b> | Binary Phase-Shift Keying                         |
| <b>QPSK</b> | Quadrature Phase-Shift Keying                     |

|             |                                           |
|-------------|-------------------------------------------|
| <b>IEC</b>  | International Electrotechnical Commission |
| <b>TRP</b>  | Total Radiated Power                      |
| <b>EIRP</b> | Effective Isotropic Radiated Power        |
| <b>FEC</b>  | Forward Error Correction                  |
| <b>LP</b>   | Linear Polarization                       |
| <b>SNR</b>  | Signal-to-Noise Ratio                     |



---

# Introduction

---

## 1.1 Background

Wireless communication systems have become essential for enabling global connectivity and supporting a rapidly expanding ecosystem of digital services. Modern networks support applications ranging from autonomous transportation and smart manufacturing to remote healthcare, immersive media and distributed edge computing [1]. As society and industries become increasingly dependent on continuous and seamless connectivity, the demand for truly ubiquitous coverage continues to grow. However, despite their maturity and widespread deployment, terrestrial communication networks remain mainly limited by geographical, economic and environmental constraints. Large parts of the planet, such as rural regions, sparsely populated areas, oceans and mountainous regions still lack reliable or affordable wireless coverage.

To address this coverage limitation, satellite communication systems have served as a complementary solution to terrestrial networks and traditionally support a wide range of use cases. These include Fixed-Satellite Services (FSS), which provide long distance backhaul and international internet connectivity, Broadcasting-Satellite Services (BSS) which deliver television and radio content directly to end users, Global Navigation Satellite Systems (GNSS), which offer precise positioning for vehicles and mobile devices, and Mobile-Satellite Services (MSS), which enable communications for mobile users equipped with dedicated satellite terminals, including applications in emergency response, maritime and aviation communications, and government or military use [2]. Although these traditional services are essential, they were not designed to integrate with cellular networks or support direct communication with standard mobile devices.

Recent technological advancements have enabled a new class of satellite communications systems, known as Non-Terrestrial Networks (NTNs), capable of supporting direct communication between satellites and standard mobile phones [3]. The deployment of LEO satellite constellations combined with advances in mobile phone antenna systems now make it possible for the User Equipment (UE) terminals to establish communication links from remote or under-served areas. This emerging capability extends the coverage of terrestrial mobile networks and aligns with the growing demand for truly global and reliable connectivity.

The introduction of D2D NTN capabilities unlocks a diverse set of use cases

that extend far beyond traditional satellite communications [4]. One of the most significant applications is the provision of basic connectivity directly on smartphones, without the need for a dedicated terminal, in areas where terrestrial networks are not available. NTN can also improve coverage in rural and remote areas, enabling mobile operators to reach regions where terrestrial networks are economically unfeasible. Beyond consumer use, NTN offers significant benefits to sectors such as maritime, aeronautical and railway transportation, where reliable connectivity across oceans, air routes, and sparsely populated regions is essential for safety and operational efficiency [3]. Additionally, NTN supports large-scale Internet of Things (IoT) deployments by connecting sensors and devices in environments where terrestrial infrastructure is lacking, including agriculture or environmental monitoring. These emerging use cases highlight the potential of NTN to complement terrestrial networks and serve as an essential enabler for ubiquitous global coverage [5].

Despite being a key technology, D2D communication presents various challenges. The large distance between the UE and the satellite results in significant path losses, leading to a very tight link budget. Establishing a reliable link is further complicated by the limitations of standard smartphones, which typically have low-gain, linearly polarized antennas, while satellite links generally rely on Circular Polarization (CP) to mitigate ionospheric polarization rotation. Further considerations include the Doppler shift, caused by fast moving LEO satellites, and increased latency caused by the long propagation distances compared to terrestrial networks. These effects introduce synchronization challenges and delay in the transmission, making it harder to achieve reliable, real-time communication comparable to terrestrial networks.

Beyond performance considerations, EMF exposure must be assessed after designing smartphones intended for satellite communication. All wireless devices placed on the market must comply with established EMF exposure limits. The limit values for EMF exposure are set in international guidelines by the International Commission on Non-Ionizing Radiation Protection (ICNIRP) and the Institute of Electrical and Electronics Engineers (IEEE) or by national authorities, as the Federal Communications Commission (FCC).

To support this evolution, standardization entities like the Third Generation Partnership Project (3GPP) and the European Telecommunication Standards Institute (ETSI) have incorporated NTN-specific requirements into 5G specifications, beginning with 3GPP Release 17 [6], which defined the foundational architecture for New Radio (NR)-NTN and introduced dedicated frequency bands in Frequency Range (FR) 1 and FR2 to enable satellite-based NR services. Releases 18 [7] and 19 [8] further developed the specification by expanding the supported frequency bands, improving link performance requirements, and addressing mobility, power constraints, and waveform adaptations. Ongoing work continues to expand these features, with the broader objective of enabling seamless integration between non-terrestrial and terrestrial systems, setting the foundation for future fully unified 6G networks.

## 1.2 Motivation

The realization of D2D NTN connectivity depends on mitigating the severe physical and practical constraints associated with satellite to smartphone communication links with the use of directive antennas. Compared to terrestrial links, NTN links suffer from significantly higher path losses due to the long propagation distances, power limitations at the UE, and propagation effects such as Doppler shift or polarization mismatch. These factors result in a tight link budget which limits achievable data rates, coverage and service reliability. Improving link capacity and robustness is therefore a key factor for the deployment of D2D services.

Among all system components, the UE antenna plays a decisive role in determining end-to-end link performance. Currently, TN antennas in a UE typically are omni-directional and have low gain and Linear Polarization (LP). In contrast, satellite communication benefits from higher gain toward the upper hemisphere and CP to mitigate polarization rotation and alignment losses. As a result, antenna design emerges as one of the critical components for enhancing NTN link capacity.

Motivated by these considerations, this work focuses on the design and evaluation of a smartphone-integrated antenna specifically designed for NTN operation. In parallel, EMF exposure is assessed. Ensuring compliance with international SAR limits is compulsory, requiring a joint consideration of antenna performance, radiation characteristics, transmit power and transmission duration. Together, these aspects aim to enable reliable NTN connectivity for future mobile devices.

## 1.3 Objectives

The overall objective of this project is to investigate and evaluate a smartphone antenna solution that enables reliable D2D communication. To achieve this goal, the project is structured following three main objectives:

1. UE antenna design:

Design and integrate a compact multi-antenna system suitable for smartphones, including both a dual-band antenna specifically optimized for NTN operation, and a wideband antenna covering terrestrial bands and potential future NTN bands. The specific NTN antenna should provide enhanced radiation towards the upper hemisphere, increased gain compared to conventional smartphone antennas, and CP to reduce polarization mismatch losses in satellite links. Design constraints related to device form factor, operational frequency bands, and practical implementation aspects are taken into account.

2. Link budget analysis:

Evaluate the impact of the proposed antenna designs on the NTN communication link through the link budget analysis. This includes assessing key parameters such as antenna gain, path loss, and transmission power limitations in representative LEO satellite scenarios.

3. EMF exposure evaluation:

Evaluate the electromagnetic field exposure associated with the proposed antenna, using the SAR metric. The evaluation will assess whether the antenna performance enhancements required for NTN communication comply with relevant international guidelines on EMF exposure, considering transmit power levels and duty cycles required for compliance.

## 1.4 Thesis structure

This thesis is structured as follows. Chapter 2 presents the theoretical background relevant to NTN communication systems, including antenna fundamentals, satellite link characteristics, and electromagnetic exposure metrics. Chapter 3 describes the methods, simulations, and calculations used in this study. Chapter 4 presents the results obtained. Chapter 5 discusses the results. Finally, Chapter 6 summarizes the main findings and outlines possible directions for future work.

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# Theoretical background

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## 2.1 Antenna theory

This section introduces the antenna concepts relevant to this thesis, including the main radiation mechanisms, impedance behaviour, and performance parameters used in the analysis.

### 2.1.1 Basic antenna concepts

An antenna is defined as the part of a transmitting or receiving system that is designed to radiate or receive electromagnetic waves [9]. From a circuit perspective, the antenna can be modelled as a load connected to a transmission line and characterized by a complex impedance

$$Z_A = [(R_{rad} + R_{loss}) + jX_A], \quad (2.1)$$

where  $R_{rad}$  is the radiation resistance, representing the conversion of accepted power into radiated electromagnetic waves,  $R_{loss}$  accounts for conductive and dielectric losses within the antenna structure, and  $X_A$  is the reactive part of the antenna impedance associated with stored electric and magnetic energy [10].

In addition to the intrinsic losses present in the antenna structure, impedance mismatch between the feed line and the antenna leads to reflection losses, reducing the power delivered to the antenna. Maximum power transfer is achieved under conjugate matching conditions, where the antenna impedance is the complex conjugate of the source impedance.

While the impedance description captures how power is delivered to the antenna, it does not describe how the radiated power is distributed in space. This spatial distribution is characterized by the radiation pattern of the antenna, which represents the radiation properties of the antenna as a function of space coordinates. These properties include power flux density, radiation intensity, field strength, directivity and polarization.

In the far field, radiation intensity is defined as the power radiated by an antenna per unit of solid angle [9], and is expressed as [10]

$$U(\theta, \phi) = \frac{r^2}{2\eta} |\mathbf{E}(r, \theta, \phi)|^2, \quad (2.2)$$

where  $\eta$  is the intrinsic impedance of the surrounding medium,  $r$  is the radial distance from the antenna to the observation point and  $\mathbf{E}(r, \theta, \phi)$  is the magnitude of the electric field at position  $(r, \theta, \phi)$ .

To quantify the directional concentration of power, directivity is defined as the ratio of the radiation intensity in a given direction to the radiation intensity averaged over all directions [10], i.e.,

$$D(\theta, \phi) = \frac{U(\theta, \phi)}{U_0} = 4\pi \frac{U(\theta, \phi)}{P_{rad}}. \quad (2.3)$$

Directivity accounts only for the spatial distribution of the radiated power, but does not include dissipation losses. To incorporate the losses, the antenna gain is defined as the ratio of the radiation intensity in a given direction to the radiation intensity that would be obtained if the power accepted to the antenna were radiated isotropically [9, 10]

$$G(\theta, \phi) = 4\pi \frac{U(\theta, \phi)}{P_{in}} = \eta_{rad} D, \quad (2.4)$$

where  $\eta_{rad}$  is the radiation efficiency [10]:

$$\eta_{rad} = \frac{R_{rad}}{R_{rad} + R_{loss}}. \quad (2.5)$$

The impedance matching between the antenna and the feed line is commonly quantified using the reflection coefficient. The reflection coefficient at the antenna input is defined as [10]

$$\Gamma = \frac{Z_A - Z_0}{Z_A + Z_0}, \quad (2.6)$$

where  $Z_0$  is the characteristic impedance of the feed line. A portion of the incident power is reflected whenever  $|\Gamma| \neq 0$ , reducing the power accepted by the antenna.

Considering mismatch losses between the antenna and the feeding network leads to the definition of the realized gain, which accounts for both radiation and mismatch losses

$$G_{realized}(\theta, \phi) = G(\theta, \phi) (1 - |\Gamma|^2), \quad (2.7)$$

where  $\Gamma$  is the reflection coefficient at the antenna input. The realized gain therefore represents the effective radiated performance of the antenna under actual operating conditions.

### 2.1.2 User Equipment antenna theory

Modern smartphones integrate a large number of antennas to meet increasingly demanding communication requirements. A typical device incorporates multiple antennas for the Low Band (LB), Mid and High Band (MHB), and 5G NR, in addition to two or more millimeter-wave antennas, one or more satellite communication antenna, two GNSS antennas, several Bluetooth and WiFi antennas, one Near Field Communications (NFC) and multiple Ultra Wide Band (UWB) antennas [11]. The design of these antennas is primarily constrained by the limited

space available within the smartphone. They have to occupy only a small portion of the device's total volume, while smartphones continue to become thinner and more compact [12].

Generally, smartphone antennas for TN have low gain and a quasi-omnidirectional radiation pattern, particularly at frequencies below 1 GHz, as they are designed to minimize the sensitivity to device orientation. Some common antenna types include patch antennas, the Planar Inverted F Antenna (PIFA), the Inverted F Antenna (IFA), slot antennas, and loop antennas.

### 2.1.2.1 Microstrip patch antennas

Microstrip patch antennas consist of a conducting surface of very small thickness  $t$  compared to the free-space wavelength ( $t \ll \lambda_0$ ), printed on one side of a dielectric substrate and backed by a conductive ground plane on the opposite side. The substrate has thickness  $h$ , typically in the range  $0.003\lambda_0 \leq h \leq 0.05\lambda_0$ , and relative permittivity  $2.2 < \epsilon_r < 12$  [10]. A schematic illustration of a rectangular microstrip patch antenna is shown in Figure 2.1.

The patch can be of several shapes such as rectangular, square, circular or other more complex geometries. For a rectangular or square patch operating in the dominant transverse magnetic mode ( $\text{TM}_{10}$ ), the length of the side  $L$  usually satisfies  $\lambda_0/3 < L < \lambda_0/2$ . The resonant frequency is primarily determined by the effective electrical length of the patch and the effective dielectric constant  $\epsilon_{\text{eff}}$ , which accounts for fringing fields at the patch edges.

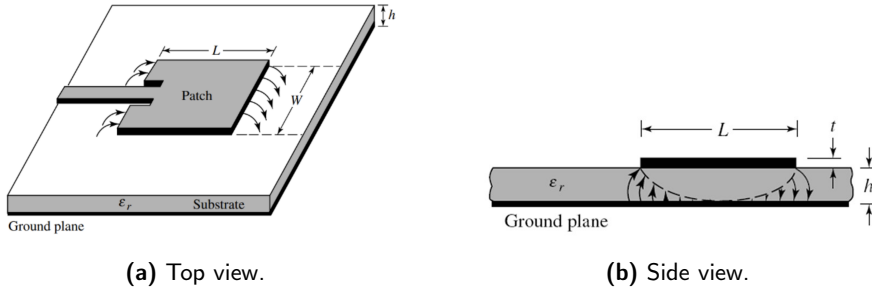


**Figure 2.1:** Patch antenna [10].

For rectangular or square patches, radiation is commonly associated with operation in the dominant transverse magnetic mode  $\text{TM}_{10}$ . In this mode, the electric field varies sinusoidally along the patch length and is approximately uniform along its width. The resonant frequency  $f_r$  of the  $\text{TM}_{10}$  mode can be approximated as

$$f \approx \frac{c}{2L_{\text{eff}}\sqrt{\epsilon_{\text{eff}}}}, \quad (2.8)$$

where  $c$  is the speed of light in free space,  $L_{\text{eff}}$  is the effective electrical length of the patch, and  $\epsilon_{\text{eff}}$  is the effective relative permittivity, which accounts for the fact that the electromagnetic fields are partly confined within the substrate and partly extend into free space due to fringing effects at the open edges.

Radiation from a microstrip patch antenna is primarily caused by the equivalent magnetic currents at the open-circuited sides of the patch, resulting in a broadside radiation pattern. The input impedance is determined by the feed method and location; impedance matching is achieved by selecting an appropriate feed offset from the patch edge.

Microstrip patch antennas are inherently resonant structures and therefore exhibit relatively narrow impedance bandwidth. Typical relative bandwidth values are commonly in the range of approximately 2–5%, depending on the substrate properties and antenna geometry [10]. In general, thicker substrates and lower dielectric constants lead to increased bandwidth, while higher permittivity substrates result in more compact antennas at the expense of reduced bandwidth. Despite their relatively narrow bandwidth, microstrip patch antennas are widely used due to their low profile, planar geometry, mechanical robustness, and ease of integration with printed circuit boards.

### 2.1.2.2 Planar Inverted-F Antenna (PIFA)

Planar Inverted-F Antennas (PIFAs) are compact, low-profile resonant antennas derived from the microstrip patch antenna by introducing a shorting element between the radiating patch and the ground plane. A typical PIFA, as shown in Figure 2.2, consists of a planar metallic patch placed parallel to a ground plane, a vertical shorting pin or shorting wall that connects the patch to the ground plane, and a feeding point.

The presence of the shorting element alters the boundary conditions of the resonator, transforming the half-wavelength behaviour of a conventional microstrip patch into a quarter-wavelength resonator. The fundamental resonance of a PIFA is therefore approximately achieved when the effective electrical length of the current path satisfies

$$L_{\text{eff}} \approx \frac{\lambda_0}{4\sqrt{\epsilon_{\text{eff}}}}, \quad (2.9)$$

where  $L_{\text{eff}}$  is determined by the combined length of the patch and the shorting path, and  $\epsilon_{\text{eff}}$  accounts for the substrate and air regions.

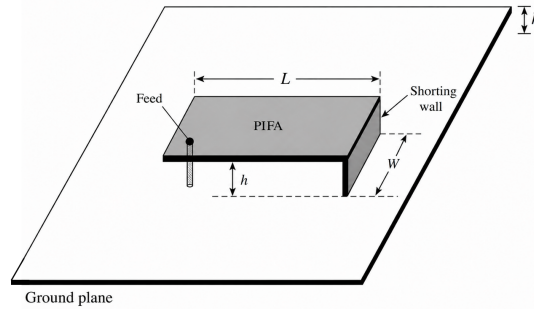


**Figure 2.2:** PIFA.

Electromagnetically, the shorting element enforces a current maximum at its

location, while the open edge of the patch corresponds to a current minimum. As a result, the current distribution along the patch is similar to a quarter-wave monopole folded parallel to the ground plane. Compared to microstrip patch antennas, this configuration enables significant size reduction while maintaining acceptable radiation efficiency.

The input impedance of a PIFA is primarily controlled by the relative positions of the feed point and the shorting element. By placing the feed between the short and the open edge, impedance values near  $50\ \Omega$  can be obtained without additional matching circuitry. The resonant frequency is influenced by the patch dimensions, height above the ground plane, substrate properties, and the geometry of the shorting element. As with microstrip patch antennas, PIFAs are resonant structures and exhibit limited fractional bandwidth.

Due to their compact size, ease of impedance matching, and favourable integration with device chassis, PIFAs are widely used for low-band and mid-band cellular applications [10].

### 2.1.2.3 Rectangular loop antennas

Loop antennas are wire or printed antennas in which radiation is primarily produced by a closed conducting loop. Among the various loop geometries, rectangular loop antennas are of particular practical interest due to their ease of fabrication, compatibility with planar technologies, and convenient integration into compact devices and PCBs.

A rectangular loop antenna typically consists of a thin conducting trace forming a closed rectangle of perimeter  $P = 2(L + W)$ , where  $L$  and  $W$  denote the loop length and width, respectively. The conductor radius or trace width is assumed to be electrically small compared to the wavelength, such that the current can be reasonably approximated as flowing along the loop contour. Depending on the electrical size of the loop, its radiation characteristics differ significantly [10].

For electrically small loops, defined by  $P \ll \lambda_0$ , the current distribution around the loop is approximately uniform. In this regime, the loop antenna behaves as a magnetic dipole. As the loop dimensions increase and the loop perimeter approaches a significant fraction of the wavelength, the current distribution becomes non-uniform and standing-wave effects develop. For rectangular loop antennas with perimeter approximately equal to one wavelength,  $P \approx \lambda_0$ , a resonant condition is achieved, and the loop can be viewed as a folded dipole-like structure supporting a standing-wave current distribution. In this resonant regime, radiation efficiency increases substantially, and the antenna no longer behaves as a purely magnetic dipole. Instead, the radiation pattern typically exhibits broad-side maxima normal to the plane of the loop [10].

The resonant frequency of a rectangular loop antenna is primarily determined by its perimeter and the effective wavelength supported by the structure. For planar or printed loops on a dielectric substrate, the resonance can be approximately expressed as

$$f_r \approx \frac{c}{P_{\text{eff}} \sqrt{\epsilon_{\text{eff}}}}, \quad (2.10)$$

where  $P_{\text{eff}}$  is the effective electrical perimeter of the loop, which is the total current path length corrected for fringing-field effects, making the resonant length slightly larger than the physical perimeter; and  $\epsilon_{\text{eff}}$  denotes the effective relative permittivity. As with other planar antennas, fringing fields and substrate loading cause the electrical length to differ from the physical dimensions.

Due to their simplicity, robustness, and favourable coupling properties, rectangular loop antennas are widely used in compact wireless devices. With appropriate scaling and layout optimization, planar rectangular loops can also be adapted for higher frequency or multi-band operation, including integration into modern wireless and non-terrestrial communication platforms [13, 14].

#### 2.1.2.4 Monopole-like antennas

A conventional monopole consists of a single radiating conductor mounted above a conductive ground plane, which acts as an image plane and effectively forms the equivalent of a dipole antenna. The most common example is the quarter-wavelength monopole, whose physical length is approximately  $\lambda/4$ , where  $\lambda$  is the wavelength of operation.

Monopole-like antennas include practical antenna structures that resemble or extend the monopole concept while deviating from the ideal thin-wire geometry. Examples include planar monopoles, printed monopoles, and other compact wide-band designs. A key distinguishing feature of these antennas is that the main radiating element is not placed above a ground plane; instead, a ground-cleared region is required around or beneath the radiator to support monopole-like operation. These antennas are widely used because of their simple structure, low profile, ease of fabrication, and good impedance matching over a relatively broad bandwidth.

The radiation pattern of monopole-like antennas is typically omnidirectional in the azimuth plane, which makes them suitable for applications requiring wide coverage. Their polarization is usually linear. The performance of a monopole-like antenna is strongly influenced by the size and shape of the ground plane, as well as by the geometry of the radiating element, which affect parameters such as input impedance, bandwidth, radiation efficiency, and gain.

Due to their compactness and versatility, monopole-like antennas are commonly employed in wireless communication systems, including mobile terminals, and satellite-related user equipment.

## 2.2 Matching networks

The simplest type of matching network is the L-network, which is composed of two reactive elements to match any load impedance to a transmission line or source with specific  $Z_0$  or  $Z_S$ , respectively. Two configurations are possible for this kind of network, as shown in Figure 2.3. The circuit in Figure 2.3a is used if the load resistance is greater than the transmission line resistance. On the other hand, if the load resistance is lower than the transmission line resistance, the circuit configuration in Figure 2.3b is used.

To synthesize the L-match, a series reactance  $X_s$  is cascaded with the load

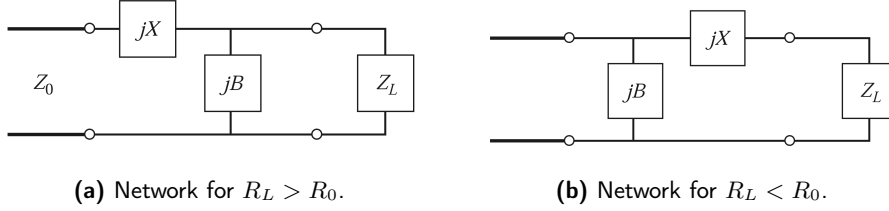


**Figure 2.3:** L-section matching network configurations [15].

and a shunt susceptance  $B_p$  is applied at the appropriate node. In normalized form, the series element transforms  $z_L = Z_L/Z_0$  along a constant resistance circle on the impedance Smith chart, whereas the shunt element (implemented as a susceptance in the admittance domain) transforms  $y = Y/Y_0$  along a constant conductance circle until the input impedance is driven to  $z_{in} = 1 + j0$ . The L-match is synthesized in terms of a required series reactance  $X_s$  and a required shunt susceptance  $B_p$ . Once  $X_s$  and  $B_p$  are obtained, the component type and value follow from their sign

$$X_s = \begin{cases} \omega L_s, & X_s > 0 \\ -\frac{1}{\omega C_s}, & X_s < 0 \end{cases}, \quad B_p = \begin{cases} \omega C_p, & B_p > 0 \\ -\frac{1}{\omega L_p}, & B_p < 0 \end{cases}. \quad (2.11)$$

## 2.3 Propagation and NTN channel characteristics

NTNs are a type of communication network in which radio access is provided by non-terrestrial platforms, such as satellites, high altitude platforms or unmanned aerial vehicles. The non-terrestrial platforms considered include satellite systems operating in LEO, Medium Earth Orbit (MEO), Geostationary Earth Orbit (GEO) or Highly Elliptical Orbit (HEO) orbits, as well as aerial platforms such as High-Altitude Platform Stations (HAPS), including balloons or solar powered aircraft.

Two main payload architectures are defined for NTN operation: transparent (bent-pipe) and regenerative. In a transparent payload architecture, no baseband processing is performed on the platform; instead, the satellite or aerial platform acts as a Radio Frequency (RF) repeater, with all gNodeB (gNB) functions located in the ground segment. In contrast, a regenerative payload architecture offers on-board processing, allowing some or all gNB functions to be implemented directly on the satellite or aerial platform [3].

### 2.3.1 NTN system geometry

The geometry of the NTN system describes the relative positioning between the UE on the Earth's surface and the non-terrestrial platform, including distances, angles and orientation. Unlike TNs, NTNs are characterized by large separation distances between Transmitter (Tx) and Receiver (Rx), which significantly influ-

ence propagation losses.

Some key parameters are the platform altitude, the slant range between the user and the platform, and the elevation angle at the receiver. In NTN, LEO satellites are generally used as the radio access platform [16].

The slant range is defined as the Line-of-Sight (LoS) distance between both ends of the link

$$d = \sqrt{R_E^2 \sin^2 \alpha + h_0^2 + 2R_E h_0 - R_E \sin \alpha}, \quad (2.12)$$

where  $R_E$  is the mean Earth radius  $R_E = 6371$  km,  $h_0$  is the satellite altitude above Earth's surface and  $\alpha$  is the elevation angle, defined as the angle between the horizontal plane at the ground terminal and the LoS toward the satellite [17].

Figure 2.4 presents the typical geometry between the UE and satellite, showing the elevation angle  $\alpha$ , the slant range  $d$ , the nadir position of the satellite, its height  $h_0$  and the Earth radius  $R_E$ .

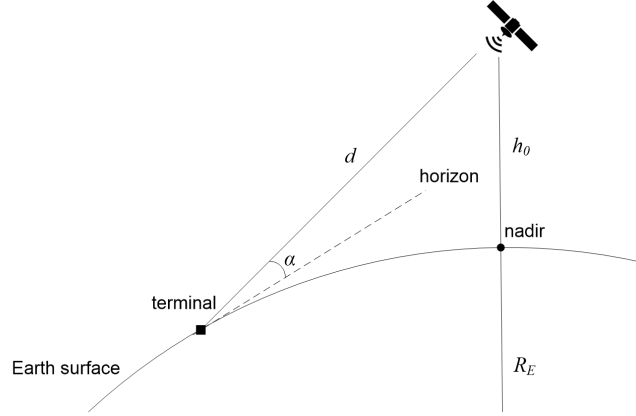


**Figure 2.4:** Slant range  $d$  between a satellite and a ground terminal [3].

### 2.3.2 Free-space attenuation

Free-Space Path Loss (FSPL) is a physical mechanism caused due to the geometric spreading of electromagnetic waves as they propagate away from the transmitter and depends only on the transmission distance and the operating frequency, assuming an ideal, lossless medium [18].

From the principle of energy conservation, the total radiated power  $p_t$  by a transmitting antenna must equal the integral of the radiated power flux over any closed surface that surrounds the transmitting antenna. Considering a spherical surface of radius  $r$  centered at the transmitter and assuming isotropic radiation, the power flux density is uniformly distributed over the sphere and can be expressed as

$$\phi = \frac{p_t}{4\pi r^2}. \quad (2.13)$$

The receiving antenna is characterized by its effective aperture  $A_{\text{eff}}$ , which represents the effective area over which the incident power is captured. Under the assumption that all power incident on this effective area is received, the received power at distance  $r$  can be expressed as

$$p_r = \phi \cdot A_{\text{eff}}. \quad (2.14)$$

By substituting the expression for the power flux density, the FSPL can be expressed as

$$l_{fs} = \left( \frac{4\pi d}{\lambda_0} \right)^2, \quad (2.15)$$

where  $\lambda_0$  is the free-space wavelength and  $d$  is the distance between the transmitting and receiving antennas.

For practical link budget calculations, the FSPL is commonly expressed in decibels. Depending on the chosen units, it can be written as [19]

$$L_{fs}[\text{dB}] = 32.44 + 20 \log_{10}(r_{\text{km}}) + 20 \log_{10}(f_{\text{MHz}}), \quad (2.16)$$

where  $r_{\text{km}}$  is the distance expressed in km, and  $f_{\text{MHz}}$  is the frequency expressed in MHz.

For a directional transmitting antenna, the power flux density in the direction of the receiver is increased by the antenna gain  $G_t$ . In this case, the received power becomes

$$p_r = p_t G_t \frac{A_{\text{eff}}}{4\pi d^2}. \quad (2.17)$$

The effective aperture of the receiving antenna is related to its gain  $G_r$  by

$$A_{\text{eff}} = \frac{\lambda^2}{4\pi} G_r, \quad (2.18)$$

where  $\lambda$  denotes the wavelength. Substituting this expression into the previous equation yields:

$$p_r = p_t G_t G_r \left( \frac{\lambda}{4\pi d} \right)^2, \quad (2.19)$$

which is known as the Friis transmission equation [18, 20].

The total losses can be expressed as  $l_{\text{total}}$ , that adds different loss components, typically including FSPL, atmospheric attenuation, rain and clouds attenuation, scintillation fading, polarization effects, and shadowing and blockage.

### 2.3.3 System noise

Noise is introduced into a satellite communication link at multiple points along the signal path. Hardware components such as amplifiers, mixers, converters, filters, switches, and interconnecting elements all contribute to adding noise within the system bandwidth. In addition to hardware-induced noise, the radiowave propagation path introduces noise through atmospheric effects, such as thermal

emission from atmospheric gases, absorption by oxygen and water vapour, rain attenuation and scattering, and scintillation. Among all sources, noise introduced at the receiver front end is the most critical, since the received carrier power at the antenna terminals is extremely low. Therefore, even small noise contributions at this stage can significantly degrade link performance, making the design of the receiver components essential to maintain acceptable signal quality.

At radio frequencies, thermal noise is the dominant noise mechanism and arises from the random thermal motion of electrons in both active and passive devices. This noise is commonly modelled using an equivalent noise temperature, which represents the temperature of an ideal resistor that would produce the same noise power per unit bandwidth. Noise power is characterized by the Nyquist relation and is independent of frequency. For a system noise temperature  $T_{\text{sys}}$  and receiver noise-equivalent bandwidth  $B$ , the available thermal noise power can be written as

$$p_n \equiv n = k T_{\text{sys}} B, \quad (2.20)$$

where  $k$  is Boltzmann's constant. Noise contributions from cascaded components, along with antenna and sky noise, are combined and referenced to the receiver antenna terminals to obtain the system noise temperature. This system-level metric provides a direct measure of the overall noise impact on the satellite link and is fundamental to the evaluation of link performance [20].

### 2.3.4 Link performance parameter

Satellite link performance is usually measured by the ratio between the average carrier power and the noise power, defined as the carrier-to-noise ratio (C/N) [20]. The received power can be expressed as

$$p_r = p_t G_t G_r \left( \frac{1}{l_{\text{total}}} \right), \quad (2.21)$$

where  $l_{\text{total}}$  is the total link losses.

Considering (2.20), the carrier-to-noise ratio at the receiver terminals is therefore [20]

$$\frac{C}{N} = \frac{p_r}{n_r} = \frac{p_t G_t G_r}{l_{\text{total}} k T_s B}. \quad (2.22)$$

Alternatively, expressing the result in terms of the Effective Isotropic Radiated Power (EIRP) and the receive gain over system temperature  $G_r$ , the carrier-to-noise ratio can be written as

$$\frac{C}{N} = \frac{\text{EIRP}}{kB} \left( \frac{G_r}{T_{\text{sys}}} \right) \frac{1}{l_{\text{total}}}. \quad (2.23)$$

For digital communication systems, link performance can also be characterized by the ratio of energy per information bit to noise power spectral density, denoted as  $E_b/N_0$ . This parameter is related to the carrier-to-noise ratio through the bit rate  $R_b$  as [20]

$$\frac{E_b}{N_0} = \frac{C}{N} \frac{B}{R_b}. \quad (2.24)$$

### 2.3.5 Propagation effects in NTN links

Propagation in Non-Terrestrial Networks (NTNs) differ from terrestrial links because, with respect to TN, signals travel a longer distance.

Atmospheric gases introduce attenuation primarily through molecular resonance absorption of oxygen and water vapour. This effect is typically modelled via a specific attenuation  $\gamma_g$  (dB/km) integrated along the slant path, and strongly depends on frequency [21].

Rain causes both absorption and scattering, leading to attenuation and depolarization. The rain attenuation for a given time percentage can be obtained from ITU-R models using

$$A_r = \gamma_R r_{\text{eff}}, \quad \gamma_R = k R^\alpha, \quad (2.25)$$

where  $R$  is the rain rate, and  $k$  and  $\alpha$  depend on frequency and polarization, and  $r_{\text{eff}}$  is an effective path length that depends on elevation angle and the rain height. It has a strong dependence on elevation  $\theta$ , which implies that low-elevation ( $\alpha \approx 0$ ) links may require substantial fade margins or adaptive mechanisms [22, 23].

Small-scale refractive index fluctuations caused by atmospheric turbulence introduce rapid amplitude and phase variations, known as tropospheric scintillation. Scintillation effects are more pronounced at higher frequencies and low elevation angles. Scintillation is usually treated statistically and added as a fading margin in the link budget [22].

At carrier frequencies below a few gigahertz, ionospheric effects are significant, leading to group delay and phase shift, and Faraday polarization rotation [24].

In addition to atmospheric phenomena, NTN links often experience terrestrial-like impairments at the UE location. Trees can intermittently block the LoS, producing excess loss and outage events. This effect strongly depends on the specific scenario and is commonly modelled with the LoS probability [3].

For system design, the overall propagation contribution is typically represented as a total loss term

$$l_{\text{total}} = l_{fs} l_g l_r l_c l_{\text{misc}}, \quad (2.26)$$

where  $l_g$ ,  $l_r$ , and  $l_c$  correspond to gaseous, rain, and cloud losses, and  $l_{\text{misc}}$  may include scintillation margin, polarization mismatch, pointing loss which accounts for misalignment between transmit and receive antennas, and shadowing, which accounts for losses due to large obstacles obstructing the signal path. The relative importance of each term depends primarily on frequency band, elevation angle, climate zone, and terminal environment which defines the physical surroundings and conditions around the UE.

### 2.3.6 Doppler shift and propagation delay

Doppler shift arises from the relative radial motion between the transmitter and the receiver, producing a frequency offset proportional to the LoS velocity. For a carrier frequency  $f_c$ , the Doppler shift can be approximated as

$$f_D = \frac{v_r}{\lambda_0} = \frac{v_r}{c} f_c, \quad (2.27)$$

where  $v_r$  is the radial relative velocity,  $\lambda_0$  is the free-space wavelength, and  $c$  is the speed of light in vacuum. In NTN links,  $v_r$  varies over time with respect to the user due to satellite motion, resulting in a time-varying Doppler offset that must be tracked by the receiver.

Propagation  $\tau$  delay corresponds to the time taken by the electromagnetic wave to travel over the slant range  $d$

$$\tau = \frac{d}{c}. \quad (2.28)$$

The Round-Trip Time (RTT) is the time required for the signal to propagate both ways

$$RTT \approx \frac{2d}{c}. \quad (2.29)$$

Compared to terrestrial systems, satellite links exhibit significantly larger and time-varying delays as  $d$  changes with the satellite position, which impacts synchronization, timing alignment, and latency.

## 2.4 EMF exposure

Smartphones are typically used in proximity to the human body. During normal operation, the device antenna transmits electromagnetic fields, part of which is absorbed into the human body and converted to heat [25]. At frequencies below 6 GHz, international organizations such as the International Commission on Non-Ionizing Radiation Protection (ICNIRP), the IEEE, and national regulators like FCC establish exposure limits, with wide margins to avoid excessive heating, in terms of Specific Absorption Rate (SAR).

SAR is a metric that quantifies the absorbed power per unit mass of human tissue. It can be calculated as

$$\text{SAR} \left[ \frac{\text{W}}{\text{kg}} \right] = \frac{\sigma |E(\mathbf{r})|^2}{\rho}, \quad (2.30)$$

where  $\sigma$  denotes the local electrical conductivity of the tissue,  $\rho$  is the local mass density and  $E(\mathbf{r})$  is the rms local electrical field.

From (2.30), it can be observed that SAR is directly proportional to the square of the rms electric field magnitude inside human tissue. This makes the transmit power, antenna placement, field distribution, and device orientation critical factors affecting SAR.

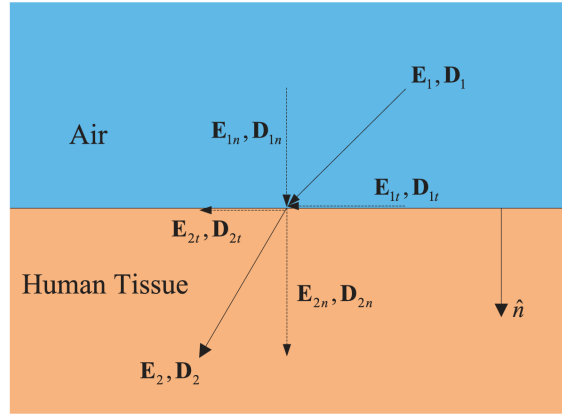


**Figure 2.5:** Boundary conditions of electric field in the human body [26].

Figure 2.5 illustrates the boundary conditions of the electric field at the interface between free space and biological tissue. Part of the incident electromagnetic field propagates into the dielectric tissue, where the dielectric losses cause the absorbed energy to be dissipated as heat.

ICNIRP specifies a localized SAR limit of 2.0 W/kg, averaged over a cube of 10 g of tissue and over a 6 minute period, for exposure in the head and trunk. FCC defines a maximum localized SAR of 1.6 W/kg, averaged over 1 g of tissue and over a 6 minute period for the same regions.

For exposure in the limbs, both ICNIRP and FCC standards set the limit for localized SAR of 4.0 W/kg averaged over 10 g of tissue. These limits are summarized in Table 2.1.

**Table 2.1:** Maximum localized SAR exposure limits specified by different international bodies and regulatory organizations for different body parts [27, 28].

| Org.   | Exposure region | Averaging mass | SAR limit | Averaging time |
|--------|-----------------|----------------|-----------|----------------|
| FCC    | Head and trunk  | 1 g            | 1.6 W/kg  | 6 min          |
|        | Limbs           | 10 g           | 4.0 W/kg  |                |
| ICNIRP | Head and trunk  | 10 g           | 2.0 W/kg  | 6 min          |
|        | Limbs           | 10 g           | 4.0 W/kg  |                |



In this chapter, the main methods used for the study of the different parts of the project are explained. First, the antenna requirements and design targets are presented. Then, the link budget model assumptions and calculations are explained. Finally, the EMF exposure evaluation simulation methods and configurations are discussed. The antenna is modelled and simulated in the time-domain solver of CST Studio Suite 2026, including the SAR simulations [29].

### 3.1 Antenna requirements

In this section, the antenna design is explained and the main steps of the design process are outlined.

#### 3.1.1 Frequency band selection

NTNs are an emerging type of network that is currently transitioning from standardization to initial deployment [30, 31]. In this context, 3GPP has already defined dedicated frequency bands to support non-terrestrial operation. In particular, satellite access for mobile devices is specified in the S-band, while larger and higher capacity devices are specified in higher frequency bands in the 10–15 GHz range [1].

Based on these specifications, the antenna design in this project targets two of the standardized 3GPP frequency bands for NTN operation. Specifically, a dual-band configuration covering the n252 and n256 bands is proposed. These bands correspond to UL frequency ranges of 2000–2020 MHz and 1980–2010 MHz, respectively, and to Downlink (DL) frequency ranges of 2180–2200 MHz and 2170–2200 MHz, respectively [8]. The full list of introduced frequencies for NTN in Release 19 is shown in Figure 3.1.

In addition to the standardized n252 and n256 bands for NTN, the n78 band between 3.3 and 3.8 GHz is also considered in the antenna design [32]. This frequency band is used for terrestrial 5G NR, making it relevant from an integration perspective. Moreover, this band provides a good trade-off between available bandwidth and FSPL, which makes it attractive for exploring non-terrestrial communications beyond the currently standardized bands, including cases where antenna configurations are not specifically optimized for NTN operation. Its suitability for

satellite-to-ground propagation at mid-band frequencies has been investigated in both analytical and experimental studies, including C-band NTN channel measurements and elevation-dependent path-loss modelling [33, 34]. Furthermore, the reuse of already standardized terrestrial spectrum could simplify spectrum utilization and enable closer convergence between Terrestrial Networks (TNs) and NTNs. In this sense, recent system-level studies have demonstrated the feasibility of operating both TN and NTN links in the same frequency band [35]. From now on, the antenna covering the n252 and n256 bands will be referred to as S-band antenna, and the antenna covering the n78 band will be referred to as C-band antenna.

| NTN satellite operating band <sup>1</sup>                                                                                                                                                                                                                                               | Uplink (UL) operating band<br>Satellite Access Node receive / UE transmit<br>$F_{UL,low} - F_{UL,high}$ | Downlink (DL) operating band<br>Satellite Access Node transmit / UE receive<br>$F_{DL,low} - F_{DL,high}$ | Duplex mode |
|-----------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------|---------------------------------------------------------------------------------------------------------|-----------------------------------------------------------------------------------------------------------|-------------|
| n256                                                                                                                                                                                                                                                                                    | 1980 MHz – 2010 MHz                                                                                     | 2170 MHz – 2200 MHz                                                                                       | FDD         |
| n255                                                                                                                                                                                                                                                                                    | 1626.5 MHz – 1660.5 MHz                                                                                 | 1525 MHz – 1559 MHz                                                                                       | FDD         |
| n254                                                                                                                                                                                                                                                                                    | 1610 – 1626.5 MHz                                                                                       | 2483.5 – 2500 MHz                                                                                         | FDD         |
| n253                                                                                                                                                                                                                                                                                    | 1668 MHz – 1675 MHz                                                                                     | 1518 MHz – 1525 MHz                                                                                       | FDD         |
| n252                                                                                                                                                                                                                                                                                    | 2000 – 2020 MHz                                                                                         | 2180 – 2200 MHz                                                                                           | FDD         |
| n251                                                                                                                                                                                                                                                                                    | 1626.5 MHz – 1660.5 MHz                                                                                 | 1518 MHz – 1559 MHz                                                                                       | FDD         |
| n250                                                                                                                                                                                                                                                                                    | 1668 MHz – 1675 MHz                                                                                     | 1518 MHz – 1559 MHz                                                                                       | FDD         |
| n248                                                                                                                                                                                                                                                                                    | 14000 MHz – 14500 MHz                                                                                   | 10700 MHz – 12750 <sup>2,3</sup> MHz                                                                      | FDD         |
| n247                                                                                                                                                                                                                                                                                    | 13750 MHz – 14000 MHz                                                                                   | 10700 MHz – 12750 <sup>2,3</sup> MHz                                                                      | FDD         |
| NOTE 1: NTN satellite bands are numbered in descending order from n256.<br>NOTE 2: For Region 2, the downlink is limited to 10700 – 12700 MHz.<br>NOTE 3: For the US, the downlink is limited to 10700 – 12200 MHz for Mobile VSAT.<br>NOTE 4: Operating band n200 is a reserved value. |                                                                                                         |                                                                                                           |             |

**Figure 3.1:** Frequency bands for NTN in 3GPP Release 19, indicating the bands of interest (Table 5.2.2-1: NTN satellite bands in FR1-NTN) [1].

### 3.1.2 Design targets

In addition to frequency coverage, some performance targets have been defined for the S-band NTN antenna. These targets aim to provide reliable and robust satellite operation. One of the key design choices concerns polarization. Due to ionospheric effects, linearly polarized signals experience Faraday rotation when propagating through the ionosphere [24], which can lead to severe polarization mismatch and associated link degradation. To mitigate this effect and improve link robustness, Circular Polarization (CP) is selected for both the uplink and downlink frequency bands of the S-band antenna (n252 and n256 bands). The quality of CP is quantified by the Axial Ratio (AR), which represents the ratio between the major and minor axes of the polarization ellipse. Another critical parameter is the antenna gain. Higher gain in the region of interest is targeted to help mitigate the large free-space path loss and to maintain a reliable communication link. Finally, impedance matching is assessed through the reflection coefficient  $S_{11}$ . A maximum reflection coefficient below -6 dB is set as a target across the operating bands, which corresponds to 25% of the input power being reflected back from the antenna. This criterion represents a reasonable trade-off between bandwidth and design complexity, and is commonly adopted in practical antenna designs for UEs, where strict volume constraints must be satisfied. The resulting

S-band antenna performance requirements adopted in this work are summarized in Table 3.1.

**Table 3.1:** S-band NTN antenna performance targets.

| Parameter                           | Target                                                          |
|-------------------------------------|-----------------------------------------------------------------|
| Axial Ratio (AR)                    | $< 3$ dB within operating band and radiating region of interest |
| Reflection coefficient ( $S_{11}$ ) | $< -6$ dB                                                       |
| Radiation region of interest        | Upper hemisphere ( $30^\circ \leq \alpha \leq 90^\circ$ )       |

For the C-band antenna, only a reflection coefficient below -6 dB is required, as the aim is not to optimize performance for this band, but to investigate the behaviour of a more general antenna configuration in the context of NTN operation. LP and a broad radiation pattern are therefore considered for this purpose.

## 3.2 Antenna design

The antenna design process is divided into two stages. First, the S-band antenna is developed. In a second stage, the design is extended to include the C-band antenna.

To meet the design targets defined in Section 3.1, different state-of-the-art antenna solutions for satellite and skyward communication applications have been reviewed, e.g., [13, 14, 36–40]. These include designs suitable for smartphones with both metallic and plastic frames.

Different solutions have been proposed in the literature to achieve high gain and CP under practical design constraints. In [36], a rectangular loop antenna is proposed to obtain a single-band CP region with radiation in the endfire direction. In [37], two narrowed crossed dipoles are designed to obtain a single band CP operation. In [38], an aperture structure combining a PIFA with parasitic elements is employed to generate CP. A metallic bezel solution is presented in [39] where LP high gain radiation in the endfire direction is achieved. In [14], a loop antenna combined with a parasitic metallic element is used to achieve two CP bands. Similarly, [13] proposes a loop antenna together with a meandered monopole, while [40] proposes a shared-aperture configuration with perpendicular antennas to achieve CP.

From this review, several conclusions can be drawn. The main challenge is to achieve high gain and CP within the limited volume available in a smartphone. Dipole and aperture-based structures can provide CP and directional radiation, but they may present limitations in terms of compactness and ease of integration into the smartphone. Metallic bezel solutions are attractive because they use the chassis to enhance radiation performance, but they may complicate the generation of CP or dual-band operation. In contrast, loop-based configurations appear particularly promising for smartphone integration due to their compact geometry, their flexibility in achieving endfire radiation, and their compatibility with parasitic elements for bandwidth extension, resonance control, and polarization

shaping.

Based on these conclusions, a loop antenna combined with a parasitic element approach has been chosen for this work [14]. This topology enables dual-band operation and dual CP in the endfire direction, making it a suitable candidate for the integration within the smartphone covering the n252 and n256 NTN frequency bands.

For the C-band antenna, the available space in the metallic corner of the phone is utilized to realize a resonant structure, where excitation is achieved through a coupled feeding mechanism. This approach allows for the use of the chassis and enables resonance in the desired frequency range.

### 3.2.1 Design Procedure

The antenna design follows a sequential approach consisting of four main steps:

1. The S-band antenna is first designed and optimized to achieve low AR over the frequency bands of interest, ensuring CP performance in the target operating range.
2. The C-band antenna is then designed and integrated with the S-band structure. This step focuses on achieving the desired resonance in the higher frequency band.
3. A coupling reduction technique is introduced to mitigate the interaction between the two antenna structures and improve port isolation.
4. Finally, an external matching network is designed to compensate for the impedance mismatch resulting from the S-band optimization, ensuring proper impedance alignment at the target frequencies.

### 3.2.2 Antenna model

Figure 3.2 shows the geometry of the proposed antennas. The S-band antenna is located along the top edge of the smartphone in order to improve radiation in the skywards direction, which is desirable for satellite-oriented links. The C-band antenna is integrated into the top right metallic corner, seen from the front.

The structure of the S-band antenna consists of a rectangular loop located at the top of the device (Figure 3.2b) and an additional parasitic branch placed above it (Figure 3.2c), separated by a dielectric substrate. Each of these elements is designed for a specific resonant frequency and optimized to achieve low AR in both the UL and DL bands. The S-band antenna is excited through a coupled feeding mechanism, where the feeding line is not directly connected to the radiating loop. Instead, energy is electromagnetically coupled from the feed to the loop and parasitic branch. The width, length and offset of the feeding line provide several degrees of freedom to optimize the phase difference between the horizontal and vertical field components.

For the C-band antenna, the metallic bezel of the smartphone is used as the radiating element (Figure 3.2d). To isolate the radiating element from the main ground plane of the device, cutouts are introduced along the side and back of

the phone, forming a dedicated metallic region. The antenna is excited through a coupled feeding mechanism, which is optimized to achieve impedance matching within the desired frequency range.

An unfolded view of the structure is shown in Figure 3.3. The rectangular loop is highlighted in green, the parasitic branch in yellow, and the corner antenna in purple. The notch filters at each port are also shown, along with the matching network at Port 1. For clarity, the substrates are hidden in this representation. The central blue region corresponds to the back of the smartphone, while the orange regions represent the sides of the smartphone.

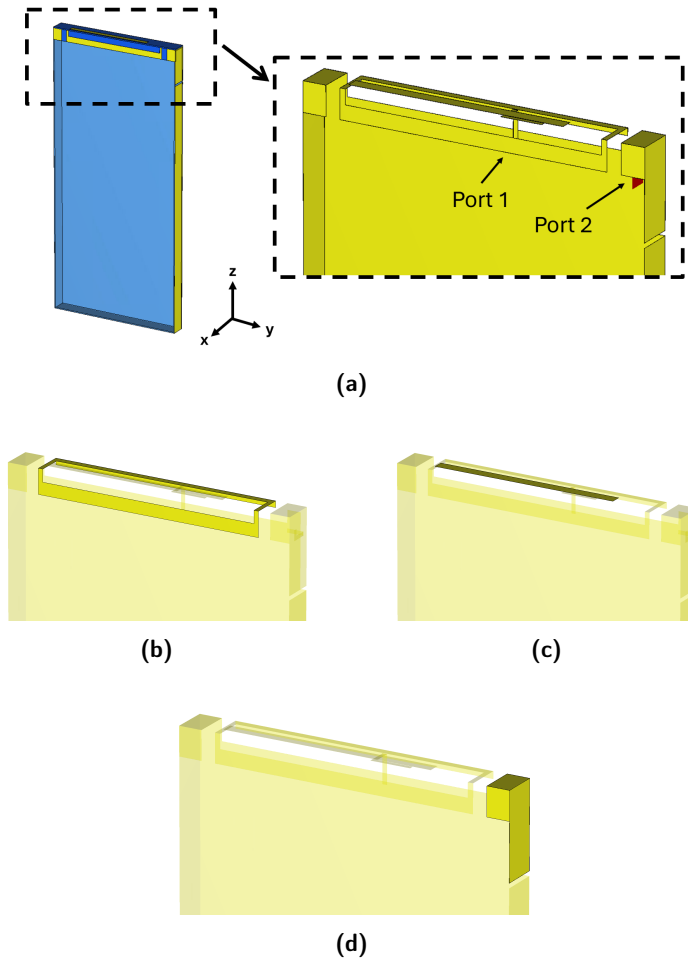


**Figure 3.2:** Geometry of the proposed S-band and C-band smartphone antennas. (a) Overall view of the device and zoomed-in antennas region. (b) Rectangular loop. (c) Parasitic branch. (d) Corner bezel antenna.

Two different dielectric materials are used in the design in order to balance

efficiency and economy. A lower permittivity substrate (ROGERS 5880 with  $\varepsilon_r = 2.2$ ) is placed in the top region, where the antenna is located, to reduce field confinement and thus improve radiation efficiency [41]. On the left, right, bottom and back sides of the phone a higher permittivity (FR4 with  $\varepsilon_r = 4.4$ ) dielectric has been used [42]. A detailed view of the location of the substrates is found in Appendix A. FR4 is widely used in consumer electronic devices due to its low cost and mechanical robustness, and it is the most common PCB material in user equipment. However, for the NTN antenna, where link performance is more critical, a lower-loss material is preferred in order to improve radiation efficiency. Therefore, a higher-performance dielectric is employed in the radiating region, while FR4 is retained in less critical areas of the device. Table 3.2 shows the properties of the materials used.

**Table 3.2:** Materials used in the antenna model.

| Material          | Properties                                                |
|-------------------|-----------------------------------------------------------|
| FR4               | $\varepsilon_r = 4.4$ , $\tan \delta = 0.02$ (@ 2 GHz)    |
| Rogers 5880       | $\varepsilon_r = 2.2$ , $\tan \delta = 0.0009$ (@ 10 GHz) |
| Copper (annealed) | Conductivity $\sigma = 5.8 \times 10^7$ S/m               |

The antennas are fed through electromagnetic coupling using a  $50 \Omega$  microstrip line implemented on the FR4 substrate. The line width is calculated from standard microstrip design equations. For a substrate thickness of  $h = 0.5$  mm, the corresponding normalized width is approximately  $W/h \approx 1.91$ , resulting in a feed width of  $W \approx 0.96$  mm. This value is used as an initial estimate and is later refined in simulation.

Finally, all the antenna dimensions are presented in Figure 3.4, where top, side, front and back views are provided; and the feeding line dimensions are presented in Figure 3.5. In addition, the optimized values of the dimensions corresponding to the final proposed antenna structure are summarized in Table 3.3.

**Table 3.3:** Main antenna dimensions.

| Parameter         | Description                                                   | Value (mm) |
|-------------------|---------------------------------------------------------------|------------|
| $W$               | Overall antenna width                                         | 70         |
| $L$               | Overall antenna length                                        | 150        |
| $l_1$             | Length of loop                                                | 52.5       |
| $l_2$             | Length of parasitic branch                                    | 44         |
| $l_3$             | Length of top corner branch                                   | 18         |
| $l\_gap$          | Gap between ground and top corner branch                      | 1.5        |
| $l\_gap\_2$       | Gap between ground and loop                                   | 3          |
| $l\_gap\_3$       | Gap between side ground and top corner branch                 | 1          |
| $w\_loop$         | Width of the loop                                             | 1          |
| $l_2\_offset\_x$  | Horizontal offset of parasitic branch                         | 8.85       |
| $l_2\_offset\_y$  | Vertical offset of parasitic branch                           | 26.15      |
| $w\_front\_loop$  | Width of loop in the front part                               | 3          |
| $l\_sep$          | Side separation between loop and ground                       | 2.9        |
| $l\_sep\_2$       | Vertical separation between loop front side and the substrate | 3          |
| $feed\_offset\_1$ | Port 1 offset from the center axis                            | 4.75       |
| $feed\_offset\_2$ | Port 2 offset from the top edge                               | 6          |
| $h\_sub\_1$       | ROGERS dielectric height                                      | 0.508      |
| $h\_sub\_2$       | FR4 dielectric height                                         | 0.5        |
| $w\_feed\_1$      | Width of feeding line 1                                       | 0.94       |
| $l\_feed\_1a$     | Length of feeding line 1 section a                            | 6          |
| $l\_feed\_1b$     | Length of feeding line 1 section b                            | 4          |
| $l\_feed\_1c$     | Length of feeding line 1 section c                            | 8.4        |
| $w\_feed\_2$      | Width of feeding line 2                                       | 0.94       |
| $l\_feed\_2a$     | Length of feeding line 2 section a                            | 5          |
| $l\_feed\_2b$     | Length of feeding line 2 section b                            | 4          |
| $l\_feed\_2c$     | Length of feeding line 2 section c                            | 2.5        |

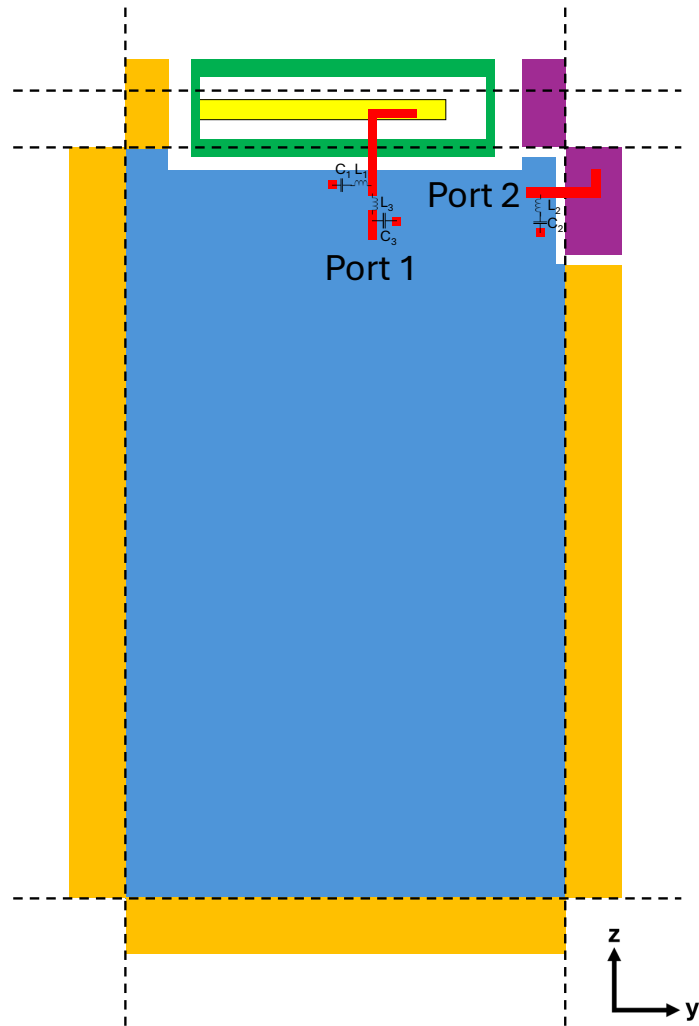


**Figure 3.3:** Proposed antenna design, including the S-band (green and yellow regions) and C-band antenna (purple region).

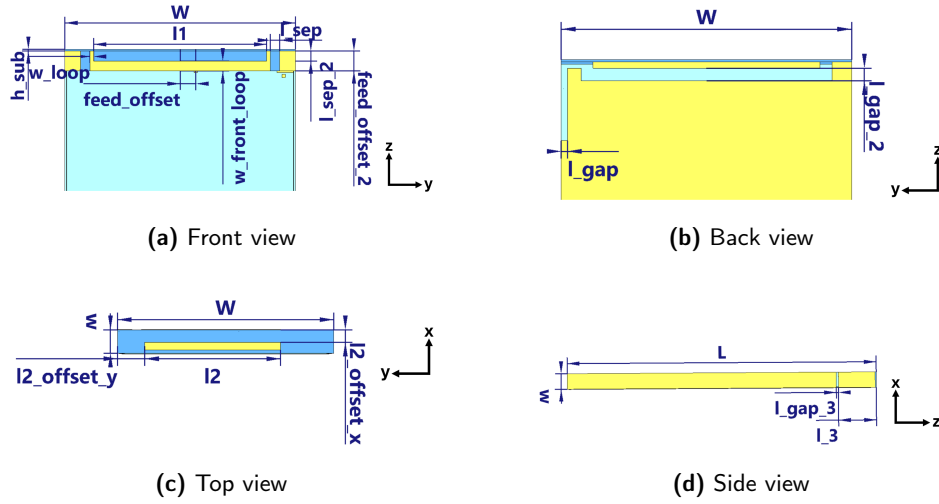


**Figure 3.4:** Antenna dimensions shown in the front, back, top, and side views.

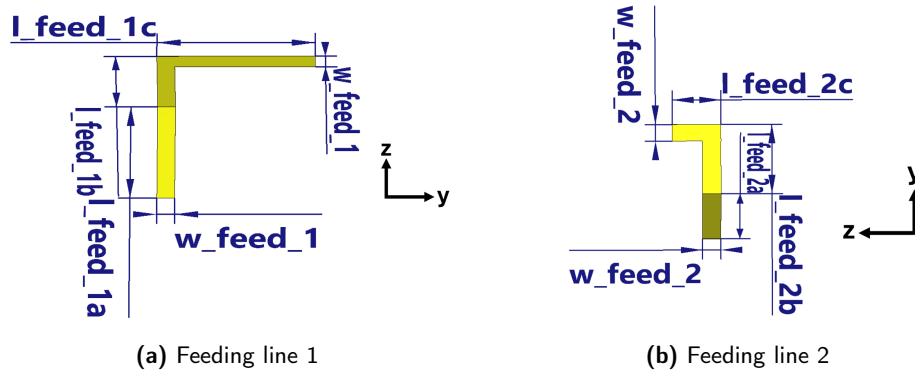


**Figure 3.5:** Dimensions of feeding lines for ports 1 and 2.

### 3.2.2.1 Design of notch filters

To improve the isolation between the two antenna ports, notch filters have been designed. A notch filter is a frequency-selective circuit that strongly attenuates signals within a frequency range while having a smaller effect outside that range. This filter can be used to suppress unwanted currents at specific frequencies and thereby reduce mutual coupling between radiating elements situated close to each other.

Each notch filter is implemented as a series LC resonator connected in shunt to ground, as shown in Figure 3.6. The input impedance of the series branch is given by

$$Z_{\text{series}} = R_s + j\omega L + \frac{1}{j\omega C}, \quad (3.1)$$

where  $R_s$  represents the equivalent series losses of the inductor, capacitor and connections. At the resonance frequency, the reactive terms cancel and the series LC presents a low impedance, providing a path to ground for currents at the specific frequency. Each filter, one for each port, is tuned to the operating frequency of the opposite port, reducing the mutual coupling between the antennas. Neglecting losses and parasitics, the series resonance frequency is

$$f_0 = \frac{1}{2\pi\sqrt{LC}}. \quad (3.2)$$

Given a target frequency  $f_0$  and selecting a capacitor value  $C$ , the corresponding inductance can be estimated as

$$L \approx \frac{1}{(2\pi f_0)^2 C}. \quad (3.3)$$

For a series-loss model  $Z = R_s + jX$ , the unloaded quality factor is defined as the ratio between the imaginary and the real part of the impedance. A higher  $Q$  results in a narrower and deeper attenuation notch, whereas lower  $Q$  increases the notch bandwidth but reduces the maximum achievable isolation.

The analytical estimates in (3.2)–(3.3) provide a first design iteration. In the final design, the resonance frequencies were refined in simulation. The component values were iteratively tuned to place the attenuation notches at the desired frequencies in the complete structure. The final implemented values are summarized in Table 3.4, and correspond to commercially available component values [43–45].

**Table 3.4:** Notch filter component values for the two operating frequencies.

| Port  | Inductor               | Capacitor               |
|-------|------------------------|-------------------------|
| $P_1$ | $L_1 = 3.9 \text{ nH}$ | $C_1 = 1.3 \text{ pF}$  |
| $P_2$ | $L_2 = 4.1 \text{ nH}$ | $C_2 = 0.43 \text{ pF}$ |

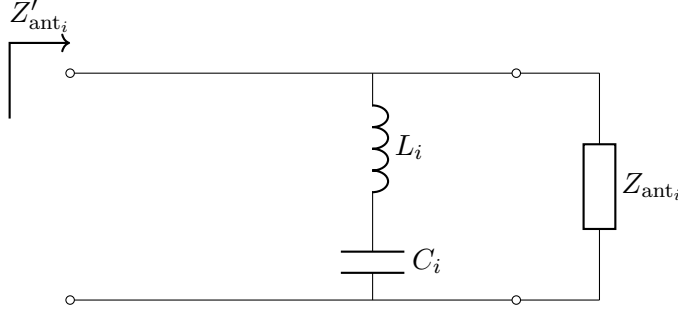


**Figure 3.6:** Series LC notch filter in parallel with the antenna impedance  $Z_{ant_i}$ , where  $(L_i, C_i)$  correspond to port  $P_i$ , with  $i = 1$  and  $i = 2$  representing ports  $P_1$  and  $P_2$ , respectively.

### 3.2.2.2 Design of matching network

As described in Subsection 3.2.1, the design of the S-band antenna is primarily optimized to achieve low AR values in the two bands of interest. As a result, its input impedance is not initially matched to the target frequency bands. Therefore, an external matching network is required to match the resonant frequencies with the desired operating frequencies.

The antenna impedances at  $f_1 = 2$  GHz and  $f_2 = 2.185$  GHz, denoted by  $Z_{ant}(f_1)$  and  $Z_{ant}(f_2)$ , are extracted from the results presented in Section 4.1. These impedances are located in the capacitive region of the Smith chart and exhibit resistive parts below  $50 \Omega$ . This impedance behaviour indicates that the antenna presents both insufficient resistance and negative reactance at the frequencies of interest. Following the matching theory introduced in Section 2.2, a first-order matching network composed of a series inductor and shunt capacitor is therefore selected, as shown in Figure 3.7.

The role of the series inductor is to compensate for part of the capacitive reactance of the antenna. In impedance terms, adding a series inductance contributes a positive reactance

$$X_L = \omega L, \quad (3.4)$$

which shifts the impedance point upwards along a constant resistance circle in the Smith chart. After this transformation, a shunt capacitor is introduced to further move the impedance towards the center of the chart. Since a shunt element is more conveniently interpreted in admittance form, the transformed impedance is first viewed in the admittance Smith chart, where the shunt capacitor adds a positive susceptance,

$$B_C = \omega C. \quad (3.5)$$

The incorporation of the shunt capacitor modifies the conductance and susceptance of the input, enabling the overall impedance seen at the input to approach  $50 \Omega$ .

The use of this topology is motivated by the initial position of the impedance points on the Smith chart, as presented in Section 4.1. Because both frequencies lie on the capacitive side and are below the normalized resistance of unity, the series inductor provides the required compensating reactance, while the shunt capacitor offers an additional degree of freedom to bring the transformed impedance closer to the matching condition. However, since only two reactive elements are used, the network provides limited control over the impedance behaviour across frequency. As a result, it is not generally possible to simultaneously obtain a perfect match at both  $f_1$  and  $f_2$ . Therefore, the component values are chosen as a compromise that provides good impedance matching in both bands.

The final component values selected for the matching network are  $L_3 = 2.7$  nH and  $C_3 = 3.2$  pF, and correspond to commercially available component values [46, 47]. These values were obtained through tuning to achieve the best trade-off between the two operating frequencies. At the two frequencies of interest, the input impedance becomes  $Z_{\text{in,ant}}(f_1) = 26.95 - j29 \Omega$  and  $Z_{\text{in,ant}}(f_2) = 41.3 + j40.3 \Omega$  for  $f_1 = 2$  GHz and  $f_2 = 2.185$  GHz, respectively.

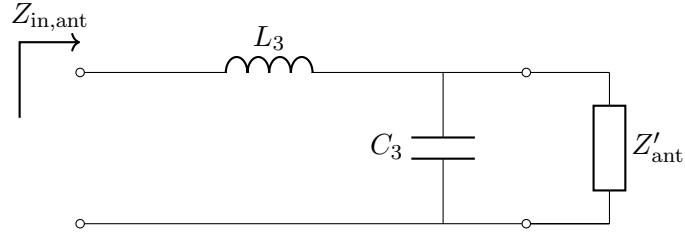


**Figure 3.7:** L-section matching network with a series inductor and a shunt capacitor, used to match the antenna impedance  $Z'_{\text{ant}}$ .

### 3.3 Link budget analysis methodology

To study the NTN link, this section evaluates the smartphone to satellite communication scenario through a link budget analysis. The link is modelled and simulated in MATLAB [48], accounting for the main propagation and system contributions.

The method followed to evaluate the link viability is summarized as follows. First, a smartphone-to-satellite geometry is defined. The satellite elevation angles  $\alpha$  are swept from  $30^\circ$  to  $90^\circ$ , as shown in the red region in Figure 3.8. This range is selected as lower elevation angles are less likely to be used in practice and typically result in reduced link performance [22]. The choice is also consistent with 3GPP NTN evaluation and test examples in which a minimum elevation angle of  $30^\circ$  is adopted as a reference assumption [8]. The work evaluates the UL under outdoor and Line-of-Sight conditions. The UE is assumed to be located in Lund, Sweden, at  $55.704^\circ\text{N}$ ,  $13.195^\circ\text{E}$  [49]. However, since the analysis is based on elevation angle, the specific geographical location would only affect specific rain attenuation results, which are not considered in this study as rain attenuation is relatively small compared to other propagation losses in the frequency bands considered [22, 23]. The selected satellite constellation type for this study is LEO, with an approximate altitude above the Earth's surface of 550 km, which lies within the typical LEO range considered for NTN [50]. This choice is motivated by ongoing NTN and direct-to-device developments that predominantly rely on LEO constellations due to their comparatively lower propagation delay and more favourable link budget than higher orbits [50–53]. In addition, recent industry activity shows a rapid growth in NTN satellite operator partnerships and commercial trials, with leading initiatives focused on LEO-based connectivity [54].

For each scenario, the received power  $P_r$  at the satellite is computed from the transmitted power  $P_t$ , transmit antenna gain  $G_t$ , and the propagation losses. The noise power is derived from the receiver bandwidth and system temperature, which is implicit in the  $G_r/T$  term. The resulting  $C/N_0$ ,  $E_b/N_0$  and  $R_b$  are calculated. Finally, these metrics are compared to the required thresholds to determine the quality of the link.

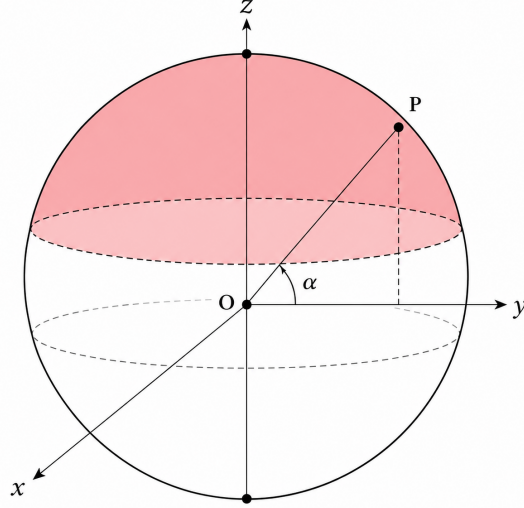


**Figure 3.8:** Elevation diagram.

### 3.3.1 Link budget calculations

The link budget analysis is first carried out for the 2 GHz UL case, which corresponds to signal transmission from UE's S-band antenna to the LEO satellite. The main simulation parameters are summarized in Table 3.5. The UL bandwidth corresponds to  $B = 20$  MHz [1], and the receiver  $G_r/T$  value of a LEO satellite has been chosen following 3GPP recommendation for these systems [16]. The transmit power  $P_t$  is swept over 23, 26 and 29 dBm in order to compare the link performance with a possible power increase from the standard Power Class 3 of 23 dBm [32].

In the simulations, only user geometries with elevation angle  $\alpha \geq 30^\circ$  are calculated, which avoids excessively low-elevation links and focuses the analysis on the practically usable part of the visible satellite pass.

For the 3.5 GHz UL case corresponding to transmission from the C-band antenna, the same orbital altitude and receiver  $G_r/T$  are maintained, since the same satellite system is assumed. Thus, the main differences between the two carrier frequencies are the wavelength, which directly affects the FSPL, and the transmit antenna pattern.

The link-budget model intentionally includes only FSPL, while additional losses such as atmospheric attenuation, ionospheric losses, rain attenuation, receiver losses, and transmitter losses are neglected to isolate the effect of link geometry and antenna patterns. At S-band and C-band frequencies, these effects are negligible compared with FSPL, supporting the use of a simplified propagation model for a baseline analysis. These additional propagation losses are evaluated in Appendix E, where the maximum total loss is shown to be approximately 0.4 dB. Polarization loss is implicitly accounted for through the simulated antenna

**Table 3.5:** Main parameters used in the link-budget simulations.

| Parameter                | Value             |
|--------------------------|-------------------|
| Carrier frequency $f_0$  | 2.0 GHz / 3.5 GHz |
| Satellite altitude $h$   | 550 km            |
| Earth radius $R_E$       | 6371 km           |
| Receiver $G_r/T$         | 1.1 dB/K          |
| Uplink bandwidth $B$     | 20 MHz            |
| Transmit power $P_t$     | 23, 26, 29 dBm    |
| Elevation angle $\alpha$ | $\geq 30^\circ$   |
| Polarization             | LHCP              |

radiation patterns. Therefore, the results should be interpreted as a baseline propagation case [22].

Since the angular region of interest is not uniformly represented by the elevation angle values, area-weighted statistics are used in the simulation. In spherical coordinates, equal increments of elevation angle do not correspond to equal surface areas on the sphere. As a result, directly sampling angles would over-represent directions close to the horizon and under-represent directions near the zenith. To compensate this, each sample is weighted proportionally to  $\cos \alpha$ , which corresponds to the differential surface of a sphere. This ensures a more uniform and physically meaningful representation of the results.

To evaluate the link performance under these conditions, in addition to  $C/N_0$ , the link is also evaluated in terms of energy per bit over noise power spectral density,  $E_b/N_0$ , since satellite uplinks are typically power-limited rather than bandwidth-limited [55, 56]. The relationship used to convert the simulated carrier-to-noise-density ratio into an information-rate metric was introduced in (2.24). Rearranging (2.24) gives the maximum feasible bitrate for a selected threshold on  $E_b/N_0$

$$R_{b,\max} = 10^{\frac{(\frac{C}{N_0}) - (\frac{E_b}{N_0})_{\text{req}}}{10}}. \quad (3.6)$$

This expression is used in the thesis as a link-budget level relation used to map the simulated  $C/N_0$  to the maximum supportable bitrate.

The considered uplink can be characterized as operating in a power-limited regime, meaning that the achievable bitrate is primarily constrained by the available received power, not by the system bandwidth. This can be seen from the relationship between  $C/N_0$ ,  $E_b/N_0$  and the achievable bitrate. Since  $C/N_0$  depends on the received carrier power and receiver noise spectral density, but not on the communication bandwidth, the achievable bitrate is fundamentally constrained by the available received power. This behaviour can be further understood through the Shannon capacity expression

$$C = B \log_2 \left( 1 + \frac{S}{N_0 B} \right), \quad (3.7)$$

where the term  $\frac{S}{N_0 B}$  represents the signal-to-noise ratio (SNR) over the bandwidth  $B$  [56]. In the considered uplink scenario, the received signal power is relatively low, such that  $\frac{S}{N_0 B} \ll 1$ , corresponding to a low-SNR regime.

Under this condition, the logarithmic term can be approximated using the first-order expansion  $\log_2(1+x) \approx \frac{x}{\ln 2}$  for  $x \ll 1$ , which reduces to

$$C \approx B \frac{1}{\ln 2} \frac{S}{N_0 B} = \frac{1}{\ln 2} \frac{S}{N_0}. \quad (3.8)$$

In this regime, the capacity becomes independent of bandwidth and is determined by  $S/N_0$ , or equivalently on  $C/N_0$  in the adopted link-budget notation [20].

A target Bit Error Rate (BER) of  $10^{-5}$  is selected as a representative performance criterion for uncoded digital communication systems. This value provides a reasonable compromise between reliability and required Signal-to-Noise Ratio (SNR), and is commonly used as a reference point in link-budget analyses. While lower error rates can be achieved in practice through Forward Error Correction (FEC), a BER of the order of  $10^{-5}$  is typically considered acceptable for baseline performance evaluation in the absence of coding. The use of this threshold therefore enables a conservative and technology-independent comparison of achievable data rates [56].

Consistent with this assumption, no coding gain is introduced in the throughput mapping. Instead, a conservative uncoded low-order PSK reference is adopted. The corresponding required threshold is obtained from the analytical error probability of coherent BPSK/QPSK in Additive White Gaussian Noise (AWGN), given by:

$$\text{BER} = Q\left(\sqrt{\frac{2E_b}{N_0}}\right), \quad (3.9)$$

where  $Q(\cdot)$  denotes the Gaussian  $Q$ -function. Solving this expression for  $\text{BER} = 10^{-5}$  yields a required ratio of energy per bit over noise power density of

$$\left(\frac{E_b}{N_0}\right)_{\text{req}} = 9.55 \text{ dB}, \quad (3.10)$$

which corresponds to the uncoded coherent BPSK/QPSK performance in an AWGN channel.

Figure 3.9 presents reference uncoded BER curves in the AWGN channel for different modulation orders. These curves are used as a theoretical benchmark to relate the available  $E_b/N_0$  values to the corresponding error-rate requirement and explain the choice of the uncoded threshold (3.10). As expected, lower-order modulation schemes such as BPSK and QPSK exhibit superior noise robustness, achieving low error rates at comparatively low  $E_b/N_0$  values. Higher-order modulation schemes require progressively larger  $E_b/N_0$  values to achieve the same error performance, due to the reduced Euclidean distance between constellation points.

The curves in the figure can also be interpreted in terms of practical service capabilities. Low-order modulations such as BPSK or QPSK provide highly reliable links that are well suited for low-rate but robust applications such as text messaging, or basic voice communications. As the available  $E_b/N_0$  increases, transitioning to higher-order modulation schemes allows for increased throughput, enabling

more demanding services such as voice calls with higher quality or low-resolution video streaming.

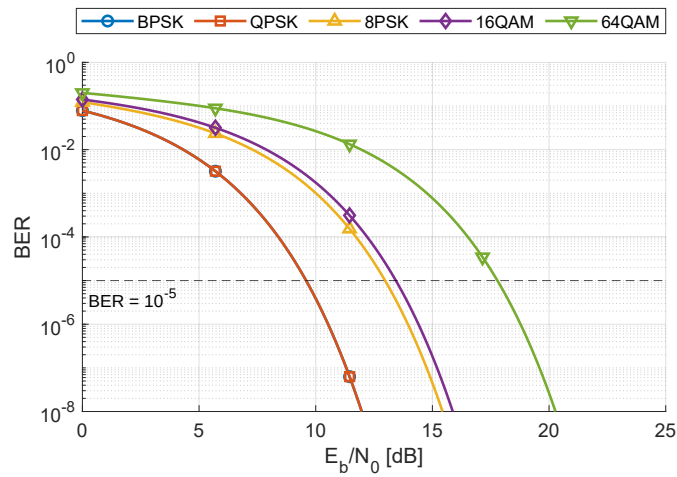


**Figure 3.9:** BER for different  $E_b/N_0$  values for different modulation schemes.

### 3.4 Exposure scenarios and phantom models

This section presents the EMF exposure evaluation methodology and scenarios. The simulations are carried out in the time-domain solver of CST Studio Suite [29]. The test cases are those specified for handheld devices in [57].

EMF exposure is evaluated for head, torso and extremities. Extremities and trunk exposure is evaluated using a flat phantom, whereas head exposure is evaluated using the SAM phantom [57]. The exposure metric used in this work is the SAR, averaged over 1g or 10g of tissue and six minutes. The limit values for 1g and 10g SAR are given in Table 2.1.

The phantom consists of two layers: a tissue equivalent medium and an outer shell, as shown in Figure 3.11. The dielectric properties assigned to the tissue equivalent medium are shown in Figure 3.10, defined by the IEC/IEEE 62209-1528 standard [57].

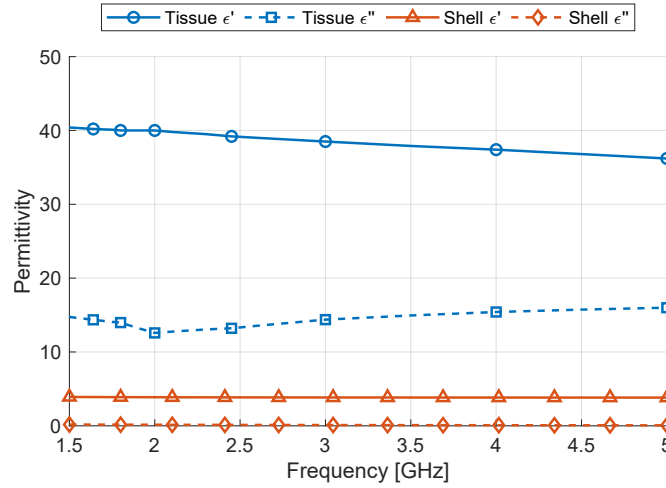


**Figure 3.10:** Tissue-simulating liquid and shell material properties.

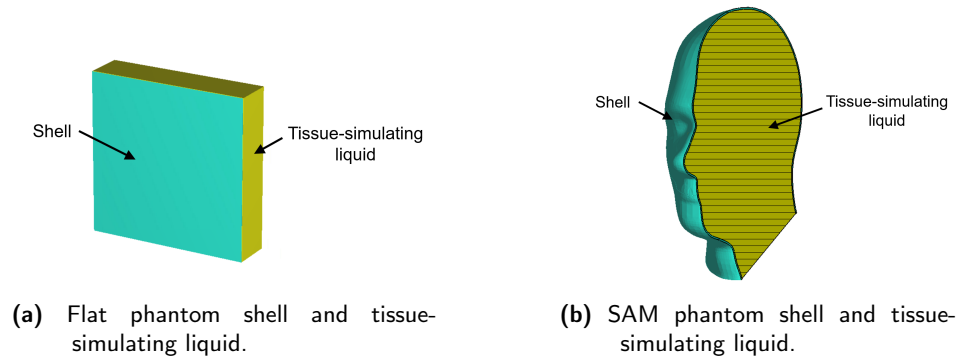


**Figure 3.11:** Detailed views of flat phantom and SAM phantom.

### 3.4.1 Extremity and trunk exposure scenario

Extremity and trunk exposure is evaluated, and the geometry of the flat phantom is shown in Figure 3.12.

Five different device orientations are considered: front, back, left side, right side and top orientations, as presented in Figure 3.13. These configurations are intended to capture different realistic placements and to assess the exposure for different orientations between the device and the body.

For the left side, right side, and top configurations, the SAR assessment is performed using the limits defined for extremities, as these positions are considered to primarily expose localized parts of the body. In contrast, for the front and back configurations, the SAR limits applicable to the trunk are used for evaluation in these scenarios.

The flat phantom thickness is chosen to be at least three times the penetration depth,  $\delta$ , in order to ensure sufficient absorption of the power penetrating the phantom model. The penetration depths at 2 GHz and 3.5 GHz are 24.3 mm and 11.4 mm, respectively [57]. Therefore, the phantom thickness is set to 72.9 mm. The height and width of the phantom are selected to be at least one wavelength of the lowest considered frequency of the device, to ensure enough attenuation before reaching the boundary; in this case, both are set to 300 mm.

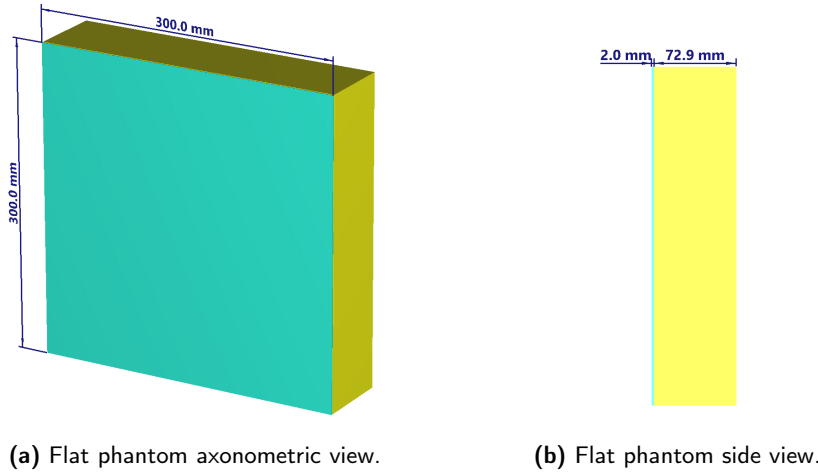


**Figure 3.12:** Geometry and dimensions of the flat phantom used in the EMF exposure assessment.

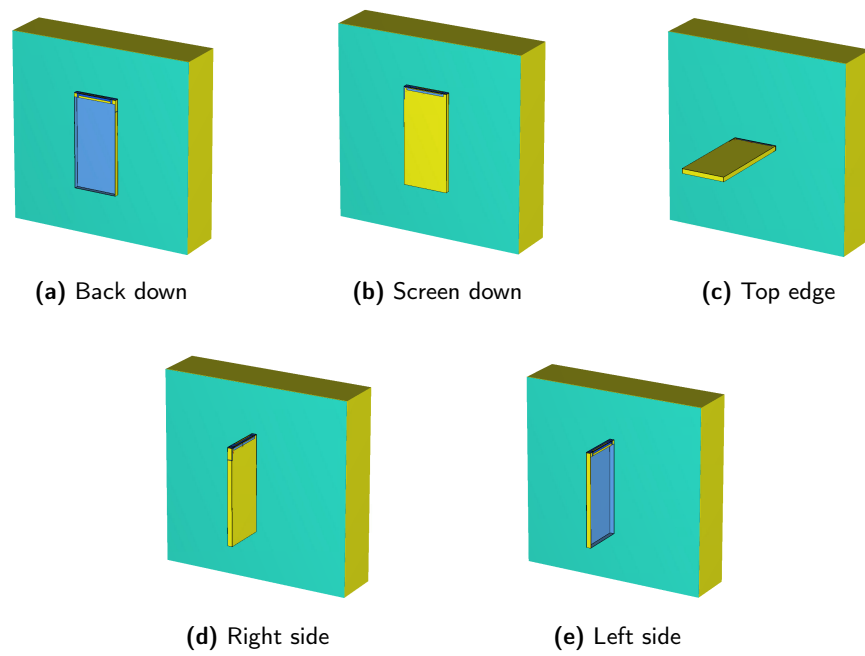


**Figure 3.13:** Phone placement configurations used in the flat phantom exposure scenario.

### 3.4.2 Head exposure scenario

The device is positioned at both the left and right ear using the two standardized placements defined for head SAR assessment: the cheek position, corresponding to  $0^\circ$  tilt relative to the cheek, and the tilt position, corresponding to  $15^\circ$  tilt relative to the cheek [57].

These configurations are intended to represent realistic phone usage conditions and to ensure that the maximum SAR value is assessed. Figure 3.14 illustrates the different configurations.

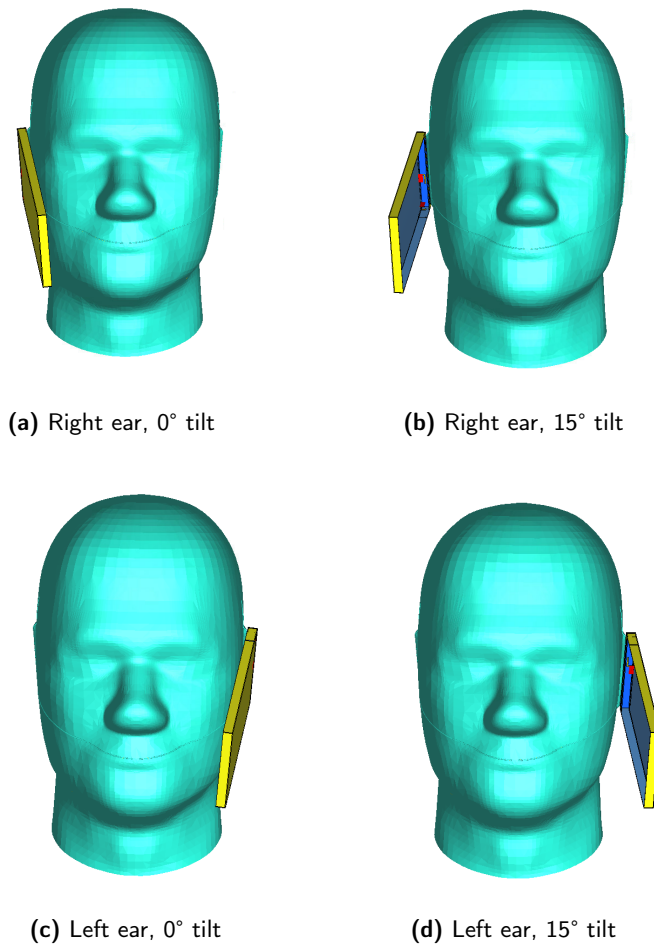


**Figure 3.14:** Phone placement configurations used for the head exposure scenario with the SAM phantom.

### 3.4.3 EMF simulation setup

The EMF exposure simulations are carried out for two operating frequencies, 2 GHz and 3.5 GHz, in order to evaluate SAR for both considered UL bands. For each frequency, three transmit power levels are analyzed: 23 dBm, 26 dBm, and 29 dBm. These values are selected to represent different operating conditions suitable for the NTN link scenario.

**Table 3.6:** Summary of the EMF exposure simulation configurations.

| Phantom      | Frequency (GHz) | Transmit power (dBm) |
|--------------|-----------------|----------------------|
| Flat phantom | 2.0, 3.5        | 23, 26, 29           |
| SAM phantom  |                 |                      |

The reported SAR results are normalized to the accepted power at the antenna port. The accepted power corresponds to the portion of the input power that is effectively delivered to the antenna after accounting for reflection losses caused by impedance mismatch, and coupling. The use of this power assumes that a dynamic tuner is available before the antenna ensuring that it is perfectly matched, which add some conservativeness to the results. In all EMF studies, the UE was positioned 0.5 mm further away from the phantom to account for any plastic covers not part of the simulated structure.

The SAR results are extracted in terms of peak spatial-average SAR over 1 g and 10 g of tissue, and then compared with the corresponding exposure limits. A separation distance of 0 mm is used for the 10g SAR, while 5 mm is applied for the 1g SAR. These quantities are used to assess the dependence of the frequency on the exposure level under the considered standardized usage scenarios.

The SAR results are obtained and compared with the exposure limits, and a maximum duty cycle to ensure compliance is calculated if the exposure is above the limit. This duty cycle represents the fraction of the time, over a six-minute averaging period, during which the UE can transmit while remaining below the limits. Since the simulated SAR values differ among the different configurations, the resulting duty cycle constraint also varies from case to case.

## 4.1 Antenna design results

### 4.1.1 Working principle

To achieve dual-band CP, the rectangular loop and the parasitic element are designed to produce different radiation modes.

Figure 4.1 shows the surface current distributions at the three main resonant frequencies. Based on the estimated resonant lengths, mode 1, appearing around 1.9 GHz, is shown in Figure 4.1a. This mode corresponds to the  $\lambda/2$  folded-dipole mode supported by the rectangular loop, shown in Figure 3.2b, with resonant frequency  $f_{\text{folded}}(l = \lambda/2) \approx 1.9$  GHz. It produces a horizontal current distribution. Mode 2 is associated with the loop resonance of the rectangular loop, shown in Figure 3.2b. It appears at around  $f_{\text{loop}}(l = 1\lambda) \approx 2.1$  GHz, and produces a vertical current distribution, as shown in Figure 4.1b. Mode 3, located near 2.40 GHz, is attributed to the  $\lambda/2$  mode of the parasitic branch, shown in Figure 3.2c, with resonant frequency  $f_{\text{parasitic}}(l = \lambda/2) \approx 2.4$  GHz. This mode produces a horizontal current distribution, as shown in Figure 4.1c.

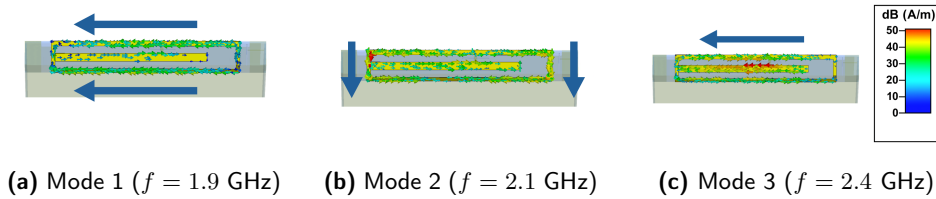


**Figure 4.1:** Current distribution of the three main modes.

The surface current direction of mode 2 is orthogonal to those of mode 1 and mode 3, so that two orthogonal field components with  $90^\circ$  phase difference are generated around the two operating regions. As a consequence, two AR minima are produced at  $f_1 = 2$  GHz, from the combination of Modes 1 and 2, and  $f_2 = 2.185$  GHz, from the combination of Modes 2 and 3, obtaining two CP bands that satisfy

$$|E_{\perp 1}| \approx |E_{\perp 2}| \quad \text{and} \quad \angle E_{\perp 2} - \angle E_{\perp 1} \approx \pm 90^\circ. \quad (4.1)$$

### 4.1.2 Parametric analysis

To better understand the influence of the main geometrical parameters on antenna performance, a parametric analysis was carried out. The study focuses on the dimensions that most strongly affect impedance matching, resonance positioning, and CP. The dimensions studied are the length of the rectangular loop  $l_1$ , the length of the rectangular parasitic branch  $l_2$ , the horizontal offset of the parasitic branch  $l_2\_offset\_x$  abbreviated as  $l_o$ , and the offset of the feeding port 1 from the center axis  $feed\_offset$  abbreviated as  $f_o$ , previously shown in Figure 3.4.

Figure 4.2 shows the  $S_{11}$  and AR result of the sweep of values for  $l_1$  between 50.5 and 54.5 mm. This parameter affects both the loop resonances as well as the parasitic branch resonance, consequently both AR minima are affected.

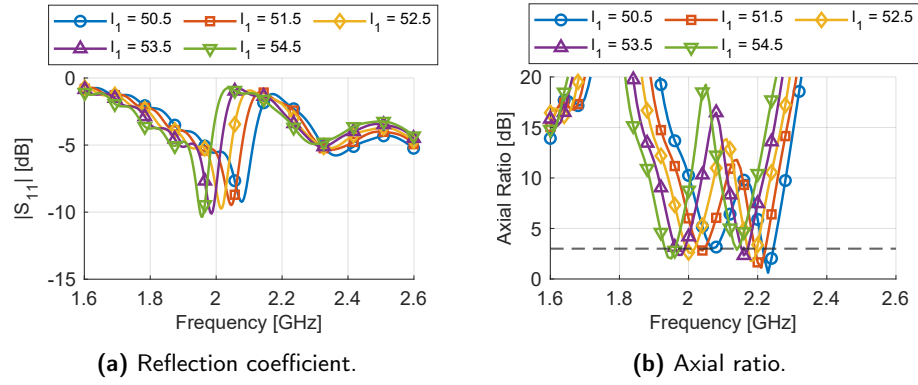


**Figure 4.2:** Parametric analysis of  $l_1$ .

Figure 4.3 shows the results of the sweep of values for  $l_2$  between 42 and 46 mm. This parameter mainly affects the parasitic branch resonance, providing a good way to set the upper frequency AR minima to the frequency of interest  $f_2$  without having an effect on the lower frequency.

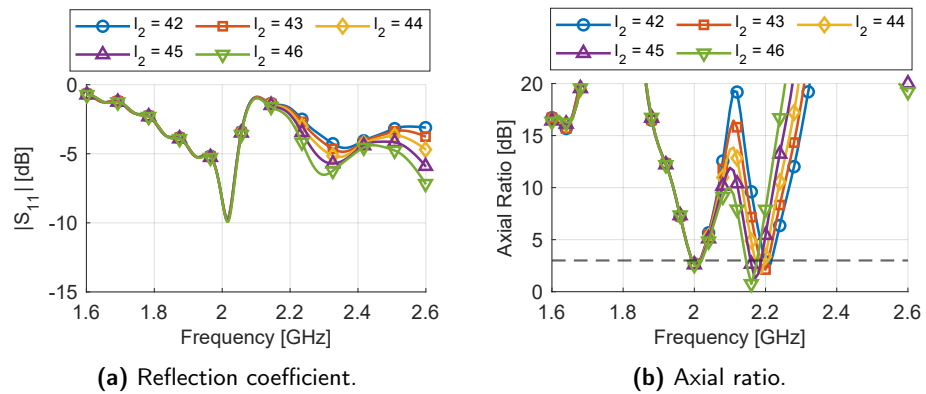


**Figure 4.3:** Parametric analysis of  $l_2$ .

Figure 4.4 shows the results of the sweep of values for the horizontal offset

of the parasitic branch  $l_o$ , between 6.85 and 10.85 mm. This parameter mainly affects the parasitic branch resonance. It also strongly affects the higher frequency AR minima, being a key parameter for optimization. From Figure 4.4b it can be seen that for some dimensions, the 3 dB target AR is not satisfied.

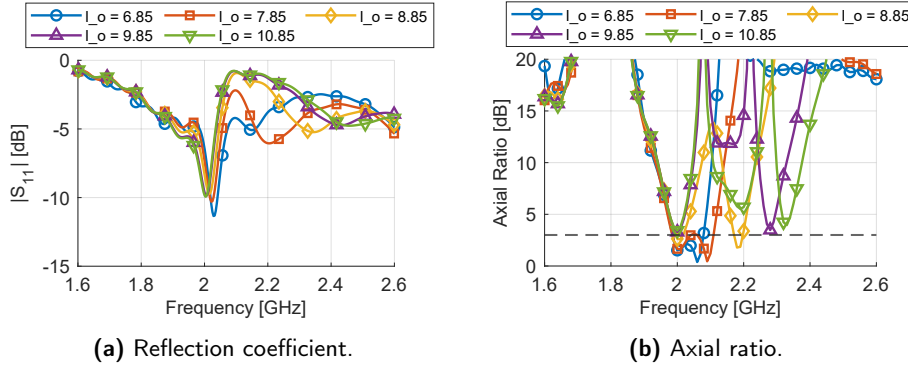


**Figure 4.4:** Parametric analysis of  $l_{2\_offset}$  ( $l_o$ ).

Finally, Figure 4.5 shows the results of the sweep of values for the feed offset from the center axis  $f_o$ , between 2.75 and 6.75 mm. This parameter mainly affects the parasitic branch resonance. It also slightly affects both AR minima. This parameter is used for fine tuning the results.

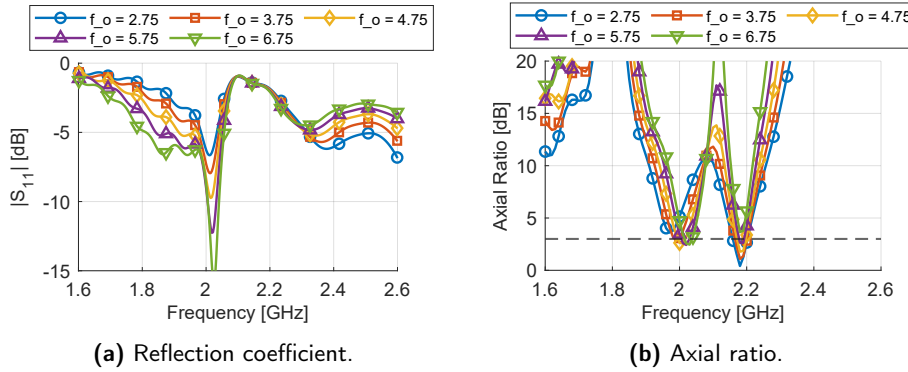


**Figure 4.5:** Parametric analysis of  $feed\_offset$  ( $f_o$ ).

These dimensions have been key to optimize the results of the antenna to meet the specified targets in Table 3.1. A compromise between the lower and higher frequency bands has been chosen, balancing AR performance between  $f_1$  and  $f_2$  to achieve values below 3 dB at both frequencies. Impedance matching is not explicitly optimized at this stage, as the primary objective is to achieve low AR; it will be addressed subsequently through the design of an external matching network.

In the case of the C-band antenna, the main tuning parameter  $l_3$  is analyzed. As shown in Figure 4.6, this parameter directly controls the resonant frequency

of the antenna. This provides a direct tuning adjustment of the resonance to the desired operating frequency.

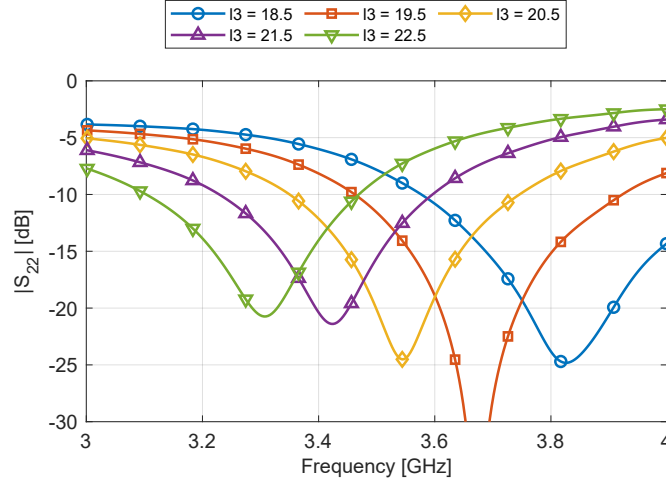


**Figure 4.6:** Parametric analysis of  $l_3$ .

#### 4.1.3 Performance of S-band antenna

In Step 1 of the design procedure, the S-band antenna is designed and optimized, as described in Subsection 3.2.1. The reflection coefficient and AR results are shown in Figures 4.7 and 4.8, respectively. These two metrics provide insights of whether the antenna satisfies the two main targets of the design, impedance matching and CP. The antenna is optimized to achieve low AR within the frequency bands of interest, while good impedance matching will be achieved in Step 4 of the design process.

The AR results are evaluated in the endfire direction, which provides a reference for the optimization process. Although the endfire AR does not fully represent the whole CP behaviour over the entire angular region of interest (elevation  $> 30^\circ$ ), it is a useful metric for optimization during the design process, as it describes the performance in the direction of the intended NTN operation.

To better show the polarization performance over the full angular region of interest, highlighted in red in Figure 3.8, the 2D AR distributions are shown in Figures 4.9 and 4.10, while the corresponding Cumulative Distribution Function (CDF) is shown in Figure 4.11. These results provide a more complete view of how the CP behaviour evolves with elevation and azimuth, and therefore allow the antenna performance to be better assessed.

The 2D AR maps show that the lowest AR values are obtained at higher elevation angles. This means that the antenna exhibits better CP performance in directions closer to the endfire direction, while the polarization quality gradually degrades towards lower elevation angles. This is consistent with the fact that the antenna has been optimized to radiate most effectively towards the upper hemisphere, particularly in directions close to the endfire direction.

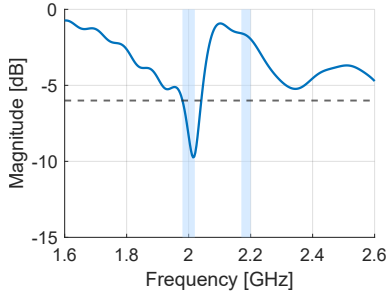


**Figure 4.7:** Reflection coefficient.

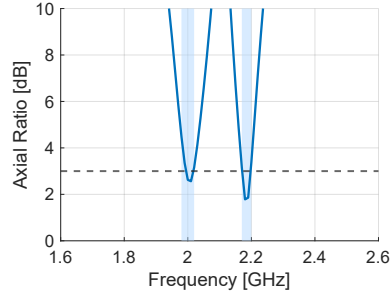


**Figure 4.8:** Axial ratio at bore-sight angle.

The CDF provides a compact way of summarizing the polarization performance over the considered angular region, and facilitates the comparison between both links. It represents the probability that a given metric falls below a certain value across the considered angles. From Figure 4.11, where the 3 dB target is marked with a dashed line, it can be seen that the 50th percentile is 3.2 dB at 2 GHz and 12.9 dB at 2.185 GHz. This indicates that the antenna provides better CP performance over a larger angular region at 2 GHz than at 2.185 GHz.

These results represent a design choice compromise. The antenna has been optimized primarily to favour the UL frequency band at 2 GHz, which is more constrained from a link-budget perspective and therefore requires enhanced antenna performance. As a consequence, the polarization quality at 2.185 GHz is not as strong over the full angular region.

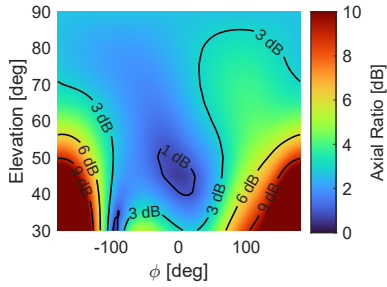


**Figure 4.9:** 2D AR pattern of Antenna Structure 1 across the angles of interest (Elevation  $\alpha > 30^\circ$ ,  $-180 < \phi < 180$ ) at 2 GHz.

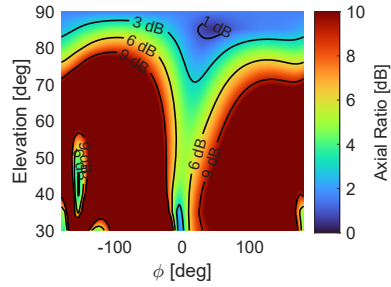


**Figure 4.10:** 2D AR pattern of Antenna Structure 1 across the angles of interest (Elevation  $\alpha > 30^\circ$ ,  $-180 < \phi < 180$ ) at 2.185 GHz.

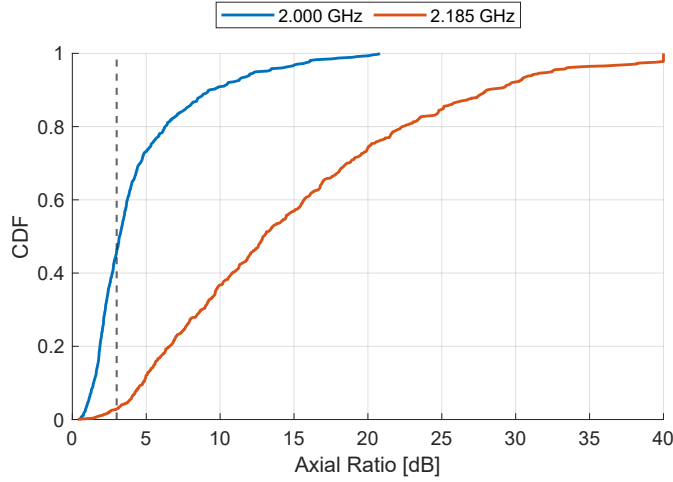


**Figure 4.11:** CDF of AR of Antenna Structure 1 across the angles of interest ( $\theta > 30^\circ$ ,  $-180 < \phi < 180$ ).

#### 4.1.4 Performance of the integrated S-band and C-band antennas

Once the S-band antenna is designed, the C-band antenna is integrated in the top right corner of the structure, as shown in Figure 3.3, corresponding to Step 2 of the design procedure. Port 1 feeds the S-band antenna, while Port 2 feeds the C-band antenna.

The integration of the second antenna introduces electromagnetic interaction between the two ports. As a result, the impedance response of the S-band antenna is slightly modified, and the mutual coupling between the two radiating structures becomes a critical design constraint. The following results therefore focus not only on the matching of each port individually, but also on the isolation between the two ports.

The reflection coefficient and mutual coupling of both antennas are shown in Figure 4.12. This figure shows the results of both structures, where  $S_{11}$  corresponds to the reflection coefficient of Port 1, and  $S_{22}$  corresponds to the reflection coefficient of Port 2. The matching values of port 1 are slightly affected compared to the previous result in Figure 4.7 of the isolated S-band antenna structure, with frequency dips now obtained at 2.01 and 2.34 GHz. In the case of the new integrated C-band antenna, the matching results show a pronounced resonance with  $S_{22}$  values below -6 dB covering 3.13–3.92 GHz.

A limitation of this intermediate configuration is the mutual coupling between the two ports, which reaches values of -13 dB at 2.01 GHz. While this level of coupling indicates some interaction between the ports, it remains within the range typically observed in compact multi-antenna designs. Nevertheless, further reduction of the coupling would be beneficial to minimize undesired power transfer and to improve the independence of the two antennas [11].

A simple way to reduce the coupling between the antennas could be to increase the separation distance between them; however, this is not preferred due to the

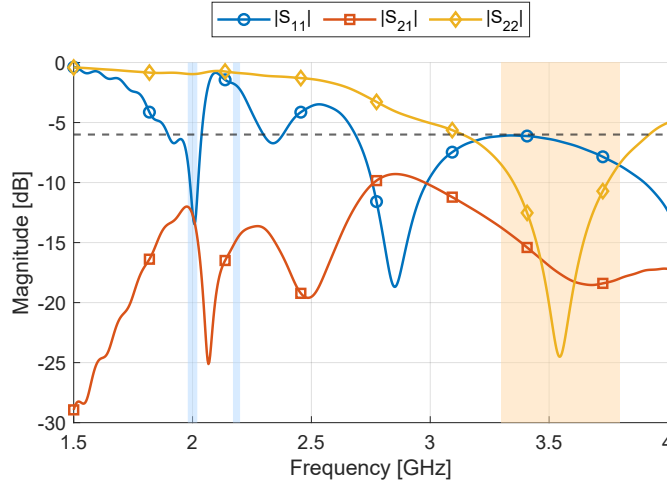


**Figure 4.12:** Scattering parameters of the first iteration of the integrated antenna structure.

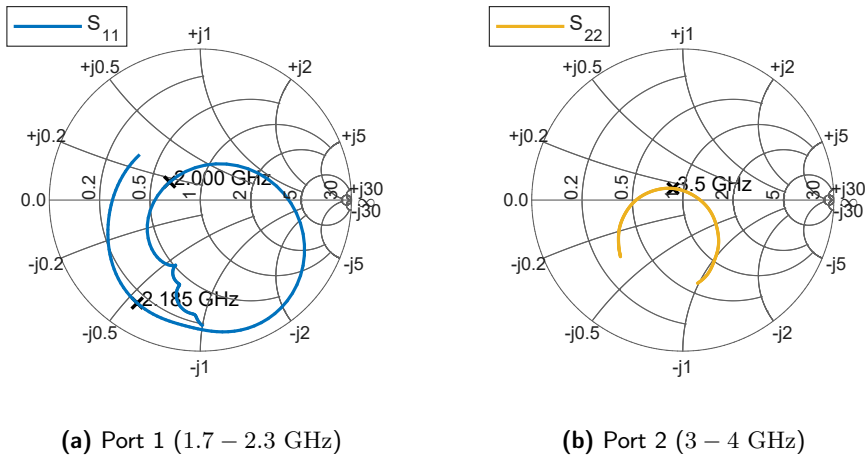


**Figure 4.13:** Smith chart of the first iteration of the integrated antenna structure.

volume and size limitations of the smartphone. An alternative way of reducing coupling without strongly affecting impedance matching is to introduce a notch filter, also called a band reject filter, in each of the ports. These filters are designed to suppress the undesired coupling, while preserving the desired impedance response [15].

Once the filters have been designed following the methodology described in Subsection 3.2.2.1, they are integrated into each port of the antenna structure, corresponding to Step 3 of the design procedure, as illustrated in Figure 3.3. In the integrated antenna model, the effective resonance frequencies of the notch filters

are shifted by the layout and component parasitics. After tuning the values of the capacitor and inductor, the simulated notches appear around the frequencies of interest (2.1 GHz and 3.5 GHz), as shown in Figure 4.14, improving the isolation between the two ports. The results now show a higher isolation between both ports, with coupling values  $|S_{21}| < -15$  dB.

Comparing Figure 4.13a with 4.15a, and Figure 4.13b with 4.15b, it can be seen that the impedance at the frequencies of interest is only slightly affected by the effect of the filters.

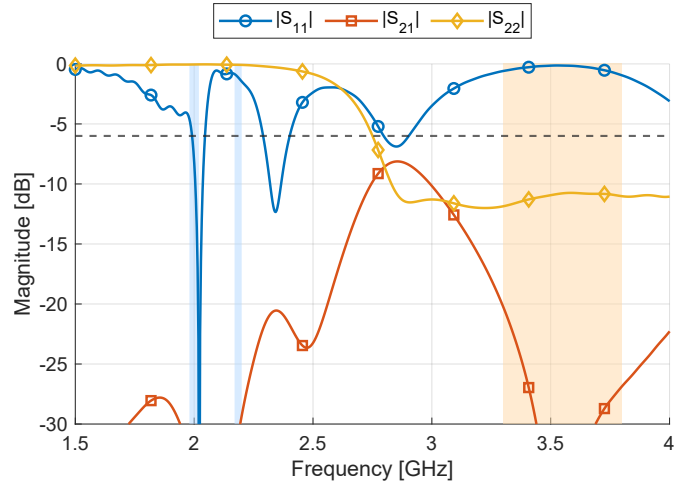


**Figure 4.14:** Scattering parameters of the integrated antenna structure after the introduction of notch filters.

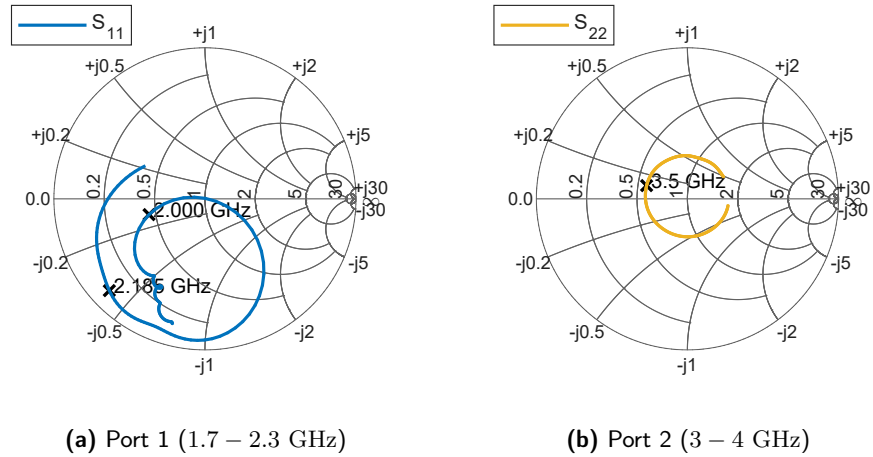


**Figure 4.15:** Smith chart of the integrated antenna structure after the introduction of notch filters.

After introducing the matching network, previously explained in Subsection 3.2.2.2, the final antenna results are presented in Figure 4.16. It can be observed that both resonant frequencies of  $S_{11}$  are located within the frequency bands of interest. The obtained -6 dB impedance bandwidth is 80 MHz (1.92–2 GHz) for the 2 GHz UL band and 40 MHz (2.18–2.22 GHz) for the 2.185 GHz DL band.

At the same time, the C-band antenna covers the n78 band with a reflection coefficient  $S_{22}$  of -6 dB across the entire band, while the isolation between port 1 and port 2 along the band is improved, with obtained coupling values below -25 dB. These results indicate that the combined use of notch filters and matching elements is effective for this structure. The final impedances of port 1 are  $Z_{in,ant}(f_1) = 26.95 - j29 \Omega$  and  $Z_{in,ant}(f_2) = 41.3 + j40.3 \Omega$  for  $f_1 = 2$  GHz and  $f_2 = 2.185$  GHz, respectively, and at port 2 at the center frequency of 3.55 GHz it is  $Z_{ant}(f = 3.55 \text{ GHz}) = 29.3 + j10.3 \Omega$ , as shown in Figure 4.17b.

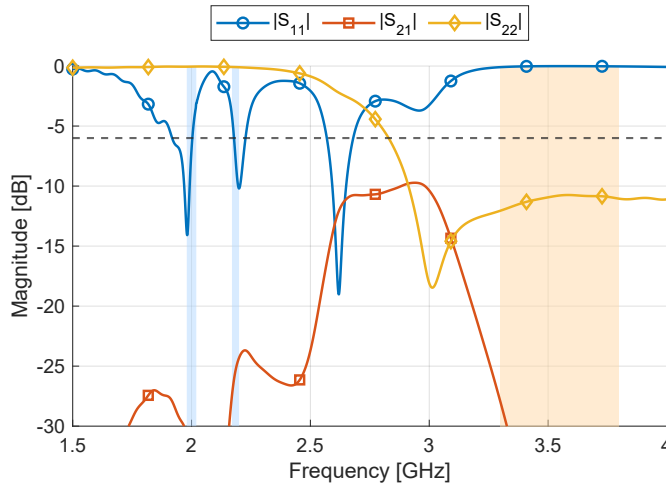


**Figure 4.16:** Scattering parameters of the final antenna structure after the introduction of notch filters and matching network.

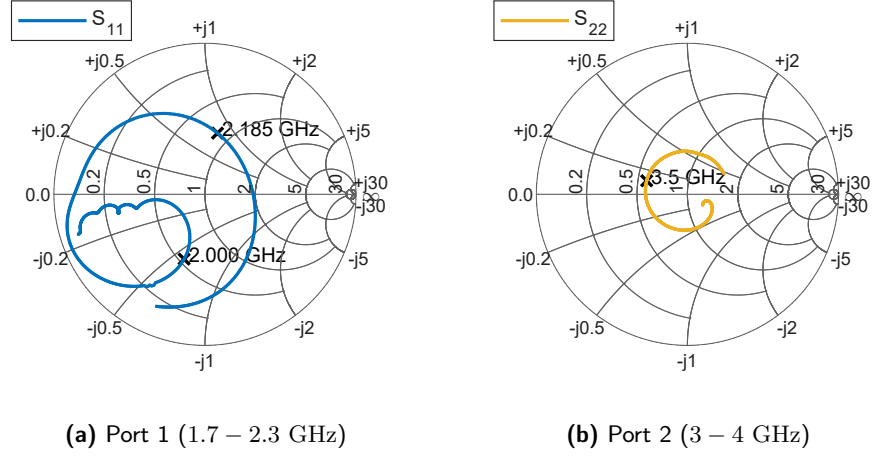


**Figure 4.17:** Smith chart of the final antenna structure after the introduction of notch filters and matching network.

#### 4.1.4.1 Radiation patterns

The IEEE gain radiation patterns for the S-band and C-band antennas at the frequencies of interest are presented in this section. For reference, Figure 4.18 shows the coordinate system used. For the S-band antenna, the results are presented at 2.0 GHz and 2.185 GHz in Figures 4.19 and 4.20, respectively. For each frequency, two principal cuts are included: the  $\phi = 0^\circ$  cut, corresponding to the front-to-back plane, and  $\phi = 90^\circ$ , corresponding to the side-to-side plane of the smartphone. As this antenna is specifically targeted for NTN communications, the angles with elevation higher than  $30^\circ$  are shaded in blue. The corresponding gain patterns for Antenna Structure 2 at the lower, center and upper frequencies of the n78 band are presented in Figures 4.21, 4.22 and 4.23, respectively.

The CP of the S-band antenna at the UL frequency of 2 GHz exhibits a wide beam with Left-Hand Circular Polarization (LHCP) directed toward the zenith. The cross-polarized component, which in this case is the Right-Hand Circular Polarization (RHCP), remains low within the angular region of interest shaded in blue. The maximum gain is 3.5 dBi and is slightly offset from the zenith direction.

At the DL frequency of 2.185 GHz, the highest gain is obtained for the RHCP in the lower hemisphere, reaching 4.47 dBi. Since this region is not relevant for the intended use case, this component is not of high interest. The LHCP component, which is relevant for the analysis, reaches a value of 1.13 dBi. Although this represents a reduction in useful gain compared to the UL case at 2 GHz, the performance can still be considered acceptable. The desired polarization is maintained in the relevant angular region, and the radiation pattern provides coverage toward the upper hemisphere. A good compromise is achieved between the two frequencies. The UL frequency provides higher gain performance, which helps compensate for the low transmit power of the smartphone resulting in a tight link budget. In contrast, the performance at the DL frequency is less favourable as the

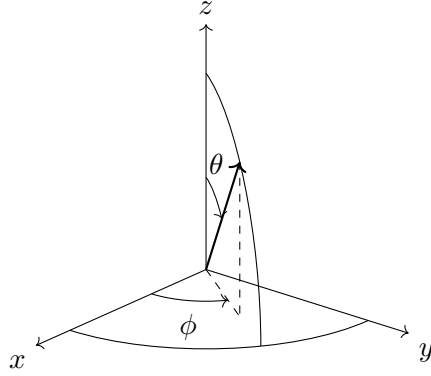


**Figure 4.18:** Spherical coordinate system used, where  $\theta$  is the polar angle measured from the positive  $z$ -axis, and  $\phi$  is the azimuth angle.

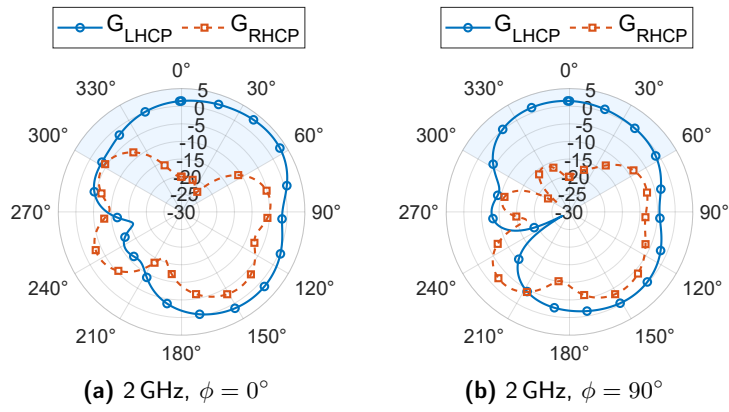


**Figure 4.19:** Gain patterns of Antenna Structure 1 at 2 GHz for the  $\phi = 0^\circ$  and  $\phi = 90^\circ$  cuts.

main beam achieves lower gain. This is acceptable, as the downlink is generally less constrained due to the higher transmit power available at the satellite [58].

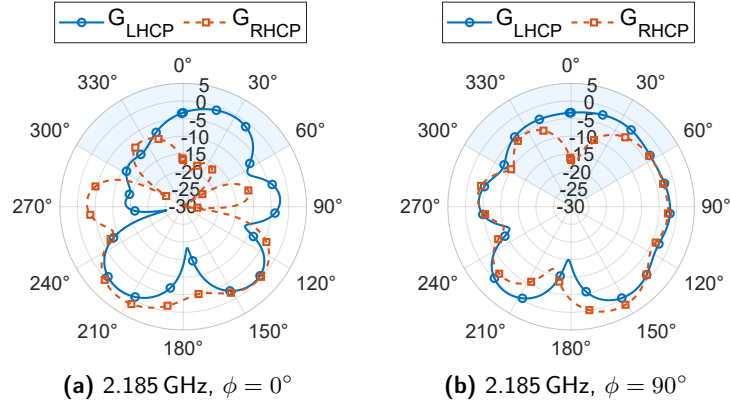


**Figure 4.20:** Gain patterns of Antenna Structure 1 at 2.185 GHz for the  $\phi = 0^\circ$  and  $\phi = 90^\circ$  cuts.

In the case of the C-band antenna, since the antenna is linearly polarized, the two linear field components,  $\theta$  and  $\phi$ , are presented in the radiation pattern plots (i.e., Figures 4.21-4.23) instead of circularly polarized components.

The design of the C-band antenna was aimed for simplicity, compact size and ease of integration into the smartphone bezel, therefore the requirements of the radiation pattern are relaxed. As a result, the radiation patterns at the different frequencies do not exhibit a clearly defined main lobe. Unlike NTN scenarios, terrestrial communication environments are typically characterized by multipath propagation, reflections, scattering and polarization changes [18]. Therefore, a highly directive radiation pattern is not required in this case. Moreover, in a general scenario, the antenna is not intended to be oriented towards the base station. The implications of this radiation behaviour for a satellite communication scenario will later be analyzed.

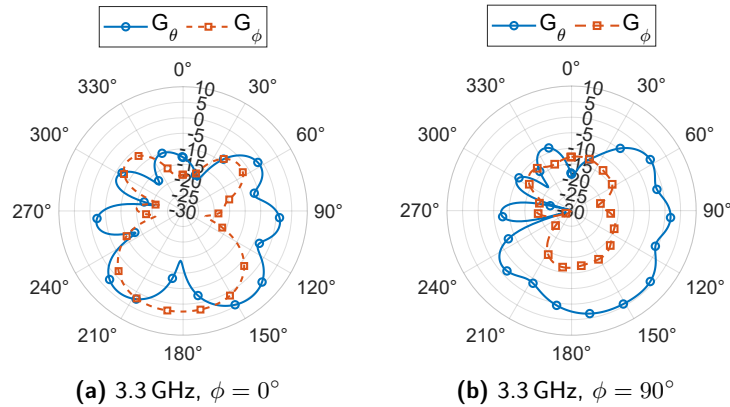


**Figure 4.21:** Gain patterns of Antenna Structure 2 at 3.3 GHz for the  $\phi = 0^\circ$  and  $\phi = 90^\circ$  cuts.

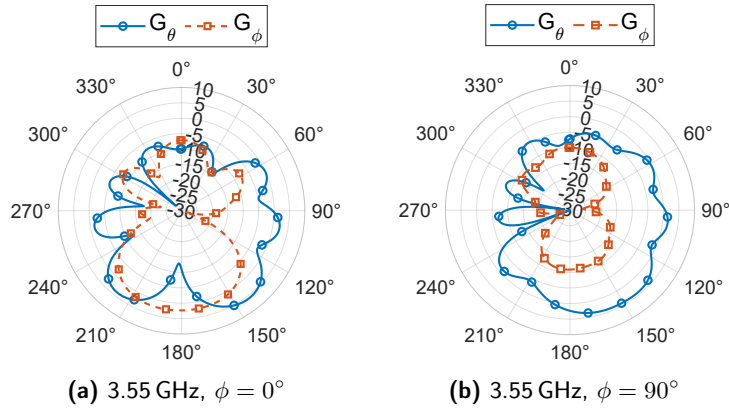


**Figure 4.22:** Gain patterns of Antenna Structure 2 at 3.55 GHz for the  $\phi = 0^\circ$  and  $\phi = 90^\circ$  cuts.

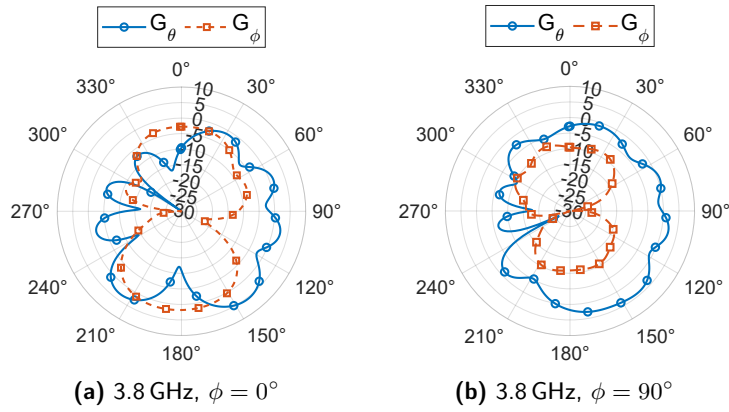


**Figure 4.23:** Gain patterns of Antenna Structure 2 at 3.8 GHz for the  $\phi = 0^\circ$  and  $\phi = 90^\circ$  cuts.

#### 4.1.4.2 Directivity and gain values across frequency

To further evaluate the performance of the antennas, the maximum directivity, IEEE gain and realized gain are analyzed as a function of frequency within the angular region of interest (elevation angles  $\alpha > 30^\circ$ ). The results are presented in Figure 4.24a for the S-band antenna and in Figure 4.24b for the C-band antenna.

For the S-band antenna, in the frequency region around the UL band, the antenna exhibits good performance, with a maximum directivity of 4 dBi, maximum gain of 3.5 dBi, and maximum realized gain of 2.5 dBi at 2 GHz. The difference between gain and realized gain reflects the impact of impedance matching. At 2.185 GHz, corresponding to the DL frequency, a degradation is observed, where the maximum directivity for the angular region of interest drops to 0.7 dBi, the maximum gain to 0 dBi, and the maximum realized gain to -0.8 dBi.

In the case of the C-band antenna, with the results considering the maximum values over the angular region of interest (elevation angles  $\alpha > 30^\circ$ ), the directivity, gain and realized gain remain relatively constant. The overall performance is limited compared to the S-band antenna UL frequency, as the maximum directivity, gain and realized gain values are 1.5 dBi, 1.1 dBi and 0.7 dBi, respectively. Compared to the DL frequency of the S-band antenna, slightly higher values are observed. However, it is important to note that the S-band antenna employs CP, while the C-band antenna uses LP, which limits the direct comparison of the results.

In both figures, the difference between directivity, IEEE gain, and realized gain provides additional insight into the antenna performance. The relatively small gap between directivity and IEEE gain in the S-band antenna indicates limited conductive and dielectric losses, whereas the larger deviation between IEEE gain and realized gain at some frequencies points to mismatch losses.

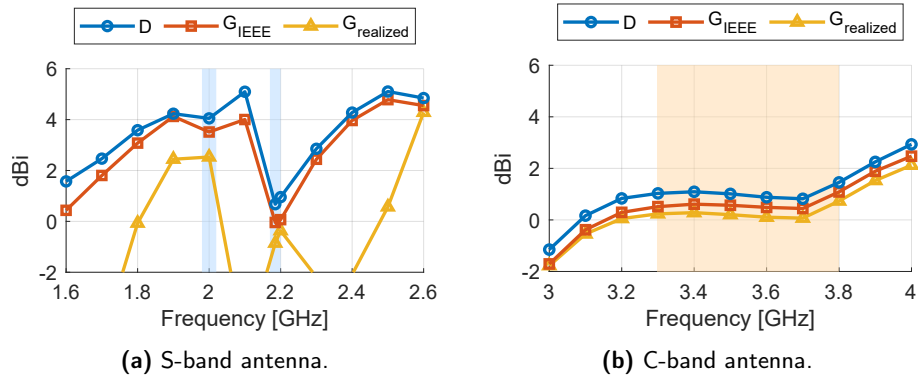


**Figure 4.24:** Maximum directivity, gain and realized gain as a function of frequency for  $\alpha > 30^\circ$ .

#### 4.1.4.3 Efficiency results

Finally, the efficiency values as a function of frequency are presented in Figures 4.25a and 4.25b for S-band and C-band antenna, respectively.

For the S-band antenna, the radiation efficiencies at the UL and DL frequencies are 88.4% and 84.8%, respectively, which indicate that the antenna radiates efficiently at both operating frequencies and conductive and dielectric losses remain limited. The corresponding total efficiencies are 70.5% at 2 GHz and 70.3% at 2.185 GHz. The difference between the total efficiency and the radiation efficiency is due to mismatch losses.

For the C-band antenna, the radiation efficiency and total efficiency at the center frequency are 90.3% and 83%, respectively, and remain relatively stable across the operating bands. These values show that the antenna has good radiation performance and better matching compared to the S-band antenna. The small difference between radiation and total efficiency is due to mismatch losses. Therefore, although the C-band antenna does not provide strong radiation toward the zenith, its overall efficiency is high.

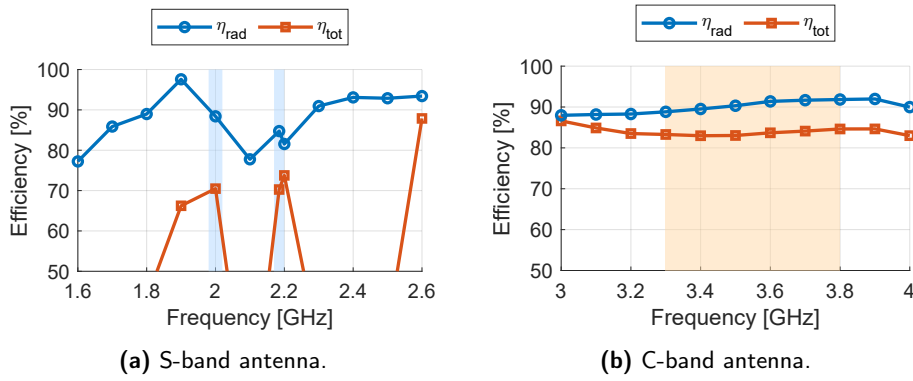


**Figure 4.25:** Efficiency as a function of frequency.

## 4.2 Link budget analysis results

First, the FSPL is evaluated for both the 2 GHz and 3.5 GHz UL scenarios, as shown in Figure 4.26. The higher frequency link experiences a larger free-space path loss. In addition, the figure shows that lower elevation angles suffer from higher losses, since the slant path  $d$  is longer. This will have a direct effect on the received power at the satellite, which depends strongly on the relative geometry between the user and the satellite. These FSPL values are used as the propagation loss component in the following link-budget results.

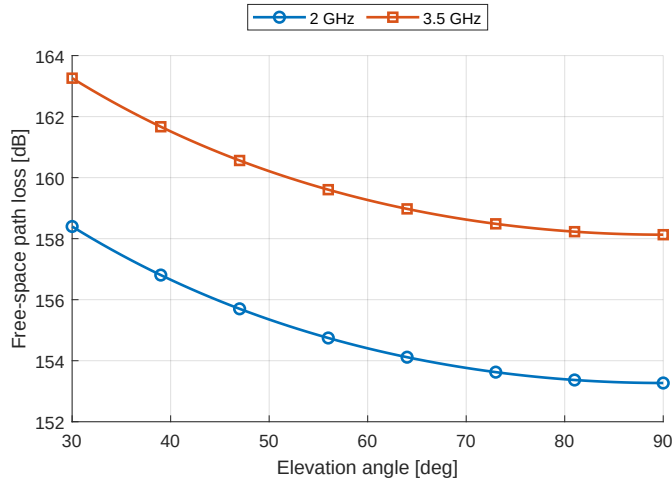


**Figure 4.26:** FSPL along different elevation angles for frequencies  $f = 2$  GHz and  $f = 3.5$  GHz.

### 4.2.1 Link budget results at 2 GHz

The link-budget results for the 2 GHz uplink are first analyzed in terms of the carrier-to-noise-density ratio,  $C/N_0$ , since this metric provides a description of the received link quality independently of the selected bitrate. Figure 4.27 shows the CDF of  $C/N_0$  over elevation angles from  $30^\circ$  to  $90^\circ$ , including the effect of the designed UE's antenna pattern. This representation is more informative than a value for a single direction, as it captures the variation of the link budget over the full useful angular region.

As expected, increasing the transmit power shifts the  $C/N_0$  distribution towards higher values. A 3 dB increase in transmit power produces a 3 dB improvement in  $C/N_0$ , so the three curves exhibit the same overall shape but are shifted horizontally, as shown in Figure 4.27. The slope of the CDF curves reflects the variability of the link conditions across the considered angular region. In the present case, the total spread is of the order of 10 dB, indicating a significant dependence on the antenna gain pattern and by the increase in FSPL at lower elevation angles, where the slant range is longer.

To interpret the uplink under different bitrate assumptions, the available

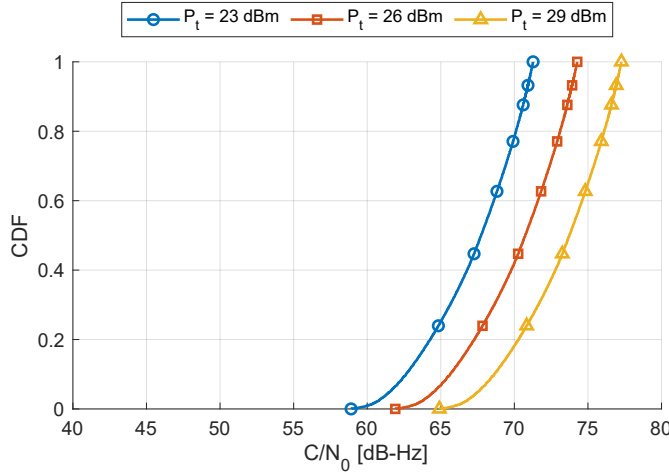


**Figure 4.27:** Cumulative distribution function of carrier power over noise power density for different  $P_t$  for the UL at 2 GHz.

$E_b/N_0$  is evaluated for three representative bitrates: 10 kbps, 100 kbps, and 1000 kbps. These values were selected to represent a range of possible services, from low-rate messaging to voice-like communication and higher data-rate applications like video calls [59], [60].

Figure 4.28 shows the CDF curves of the available  $E_b/N_0$  for a transmit power of  $P_t = 23$  dBm under these three bitrate assumptions. The relative position of the curves reflects the physical meaning of  $E_b/N_0$ : for a given received  $C/N_0$ , a lower bitrate implies that the available energy is distributed over fewer information bits, resulting in a larger  $E_b/N_0$ . Consequently, the 10 kbps curve appears at the highest values, while the 1 Mbps curve is shifted towards the lowest values. This figure therefore provides a direct interpretation of which bitrate ranges are supported by the available UL received power  $P_r$ .

Finally, the specific maximum bitrate values that can be achieved for each  $E_b/N_0$  value are presented in Figure 4.29. A +3 dB increase in transmit power results in a doubling of the achievable bitrate, leading to a consistent factor of two difference across all CDF values.

To estimate the maximum feasible bitrate, the available  $C/N_0$  at the receiver is compared against a fixed required  $E_b/N_0$  threshold corresponding to an uncoded coherent low-order Phase-Shift Keying (PSK) reference at a target bit error rate of  $\text{BER} = 10^{-5}$ . Since no coding gain is introduced in this work, the resulting bitrate should be interpreted as a conservative energy-limited benchmark. The corresponding CDF of the maximum feasible bitrate,  $R_{b,\text{max}}$ , is shown in Figure 4.29.

The feasible bitrate doubles with a +3 dB increase in transmit power  $P_t$ , since the selected bitrate metric is energy-limited and therefore strongly dependent on the available  $C/N_0$ . A convenient way to summarize the findings through the distributions is through representative percentiles. In particular, the 10th and 90th percentiles are used to characterize poor and good operating conditions,

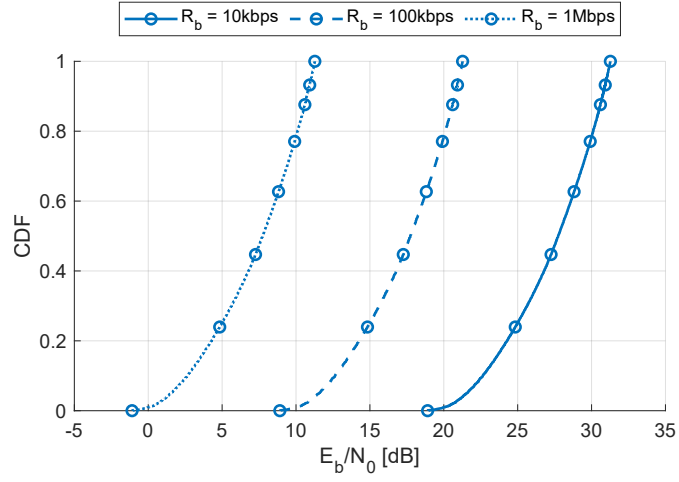


**Figure 4.28:** Cumulative distribution function of energy per bit over noise power density for  $P_t = 23$  dBm and  $\text{BER} = 10^{-5}$  for the UL at 2 GHz.

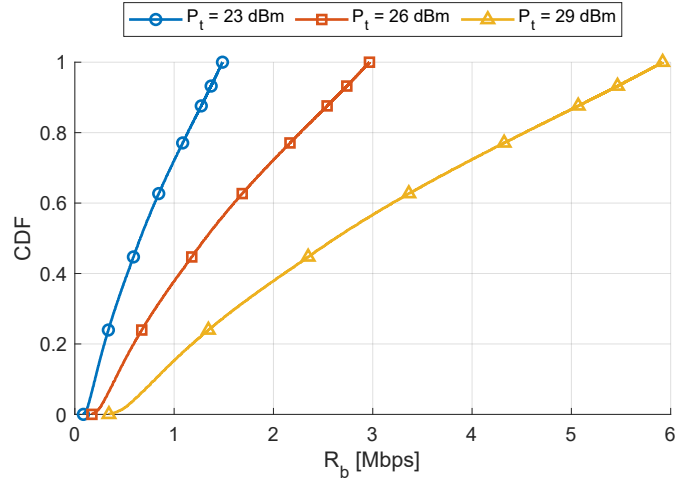


**Figure 4.29:** Cumulative distribution function of maximum bitrate  $R_b$  for different  $P_t$  and  $\text{BER} = 10^{-5}$  for the UL at 2 GHz.

respectively, and the obtained values are reported in Table 4.1.

These values indicate that, even under the conservative uncoded assumption adopted here, the 2 GHz uplink is capable of providing non-negligible data rates over a substantial fraction of the considered coverage region ( $30 \leq \alpha \leq 90$ ). However, it must be emphasized that  $R_{b,\max}$  in this form is a link-budget-level bitrate estimate only. It does not include coding gain, adaptive modulation and coding, protocol overhead, retransmissions, or multi-user resource sharing. Therefore, it should be interpreted as a reference for comparing the impact of antenna radiation

**Table 4.1:** Representative values of the maximum feasible bitrate  $R_{b,\max}$  at the 10th and 90th percentiles for the 2 GHz uplink case.

| Transmit power $P_t$ | 10th percentile | 90th percentile |
|----------------------|-----------------|-----------------|
| 23 dBm               | 205 kbps        | 1.32 Mbps       |
| 26 dBm               | 410 kbps        | 2.64 Mbps       |
| 29 dBm               | 820 kbps        | 5.30 Mbps       |

pattern and transmit power effects, rather than as the final user throughput of a complete 5G NR NTN system.

#### 4.2.2 Link budget results at 3.5 GHz

The 3.5 GHz case is analyzed as a comparative case study.

Figure 4.30 shows the CDF of the carrier-to-noise-density ratio,  $C/N_0$ , for the 3.5 GHz UL case. Compared with the 2 GHz results, the distribution is shifted to lower values. This degradation comes from the higher carrier frequency, which leads to larger FSPL for the same propagation distance. The difference also comes from the antenna pattern used in this case, which is not optimized for the upper hemisphere region. Therefore, the available gain over the angular region of interest is less favourable.

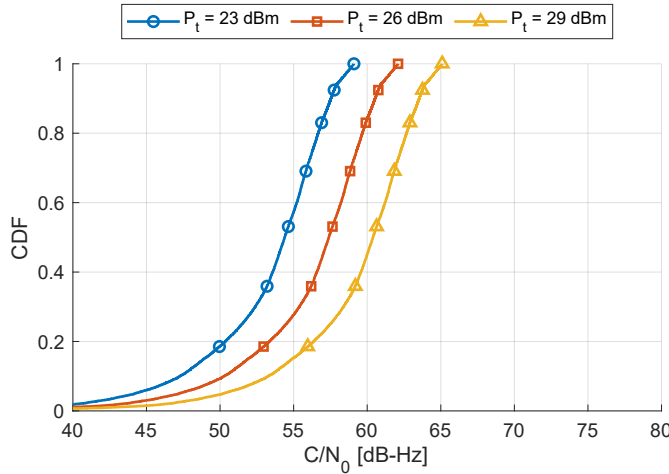


**Figure 4.30:** Cumulative distribution function of carrier power over noise power density for different  $P_t$  for the UL at 3.5 GHz.

This behaviour is also seen in the available  $E_b/N_0$  distributions shown in Figure 4.31. For the same transmit power and same bitrate, the lower  $C/N_0$  values translate directly into lower available  $E_b/N_0$ . Consequently, the C-band antenna is less capable of supporting the same bitrates over the considered range of elevation angles. This confirms that the antenna radiation pattern has a clear impact on

the energy available per information bit.

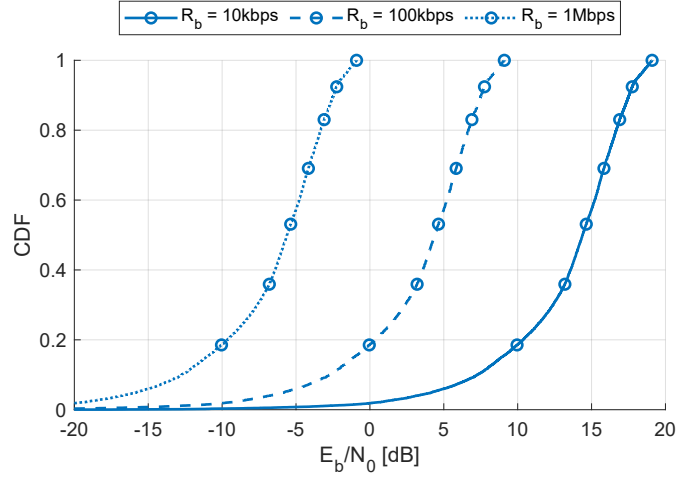


**Figure 4.31:** Cumulative distribution function of energy per bit over noise power density for  $P_t = 23$  dBm and  $\text{BER} = 10^{-5}$  for the UL at 3.55 GHz.

The effect is also clear when considering the maximum feasible bitrate, shown in Figure 4.32. The achievable rates are lower than in the 2 GHz case, which is a direct consequence of the reduced  $C/N_0$ . This result demonstrates that antenna design is a critical element in NTN terminal performance. It shows that using an antenna for both TN and NTN in a satellite link leads to a substantial deterioration of the results, even if the transmit power remains the same.

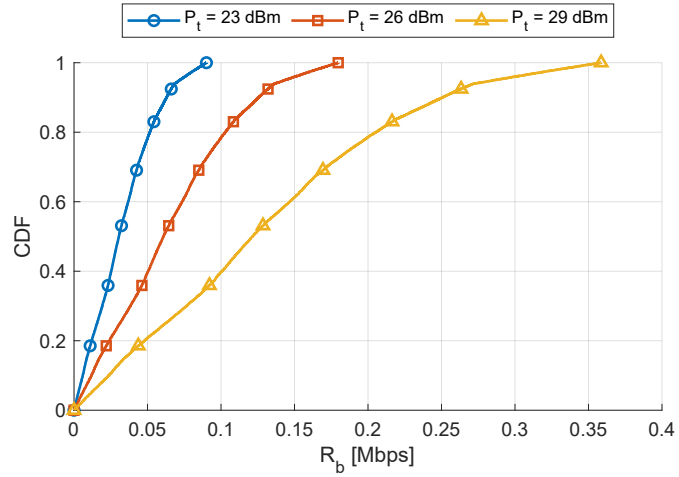


**Figure 4.32:** Cumulative distribution function of maximum bitrate  $R_b$  for different  $P_t$  and  $\text{BER} = 10^{-5}$  for the UL at 3.55 GHz.

The specific values for the 10% and 90% percentiles are summarized in Table 4.2. Compared to the results obtained at the 2 GHz UL, the bitrate suffers a degradation of 97.1% and 95.3%, for the 10th and 90th percentile, respectively.

**Table 4.2:** Representative values of the maximum feasible bitrate  $R_{b,\max}$  at the 10th and 90th percentiles for the 3.5 GHz uplink case.

| Transmit power $P_t$ | 10th percentile | 90th percentile |
|----------------------|-----------------|-----------------|
| 23 dBm               | 6 kbps          | 62 kbps         |
| 26 dBm               | 12 kbps         | 125 kbps        |
| 29 dBm               | 24 kbps         | 250 kbps        |

### 4.3 EMF exposure results

This section presents the EMF exposure results obtained for the flat phantom and SAM phantom models. The purpose is to evaluate the exposure behaviour in representative body and head scenarios, and assess the duty-cycle required for compliance at any distance from the device. A complete overview of all simulated SAR and duty-cycle results for the flat and SAM phantoms at all investigated transmit powers is provided in Appendix D. For clarity, only a few of the cases are summarized in the main text.

#### 4.3.1 Flat phantom results

The results for the flat phantom configurations described in Section 3.4 are presented in this subsection.

SAR depends on the relative orientation between the device and the phantom surface, as well as on the separation distance.

For the top and side configurations the limits applicable for limbs are used, and for the front and back configurations the head and trunk limit is used.

For the 2 GHz antenna, the highest SAR is observed when the smartphone is positioned with its back side facing the flat phantom. For this case, the ICNIRP defines a limit of 2 W/kg for averaged over 10g of tissue for a separation distance of 0 mm, and the obtained value is 5.9 W/kg. The FCC defines a limit of 1.6 W/kg averaged over 1g of tissue for a separation distance of 5 mm, and the obtained value is 4.57 W/kg. The corresponding 10g SAR distribution is shown in Figure 4.33. As expected, the peak values are concentrated in the region closest to the location of the 2 GHz antenna. To ensure compliance, a duty cycle of 34% over a six minutes transmission interval must be applied to comply with the ICNIRP limit, and of 35% to comply with the FCC.

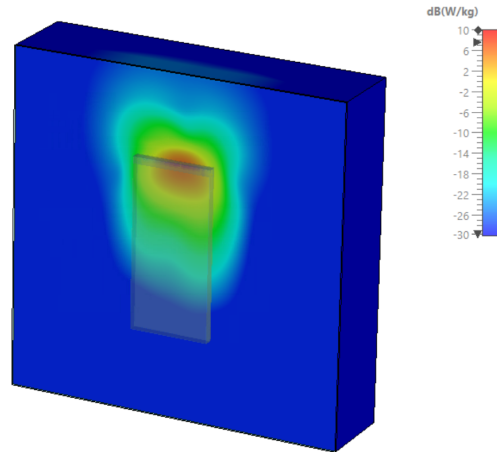


**Figure 4.33:** Simulated 10g spatial-averaged SAR distribution for the flat phantom configuration with back down at 2 GHz, 23 dBm and separation distance of 0 mm.

For the 3.5 GHz antenna, the configuration with the back side of the phone facing the flat phantom at 5 mm separation results in the highest 1g SAR with an obtained value of 4.35 W/kg, whereas the highest 10g SAR values are obtained for the configuration with the front side of the phone facing the flat phantom at 0 mm, with an obtained value of 7.09 W/kg. Due to the antenna placement at the top right corner of the phone, seen from the front, the region of maximum absorption is located in proximity to that area, as shown in Figure 4.34. A temporal duty cycle of 34% must be applied for EMF compliance with ICNIRP limits and of 37% to comply with FCC limits. The simulated SAR values and corresponding duty cycle for compliance for all configurations are summarized in Tables 4.3 and 4.4, for both considered frequencies. The tables include both 1g and 10g results, which are applicable depending on the regulatory body. The previously discussed configurations leading to the highest exposure and thus lowest duty cycle are shown in the tables, as well as the rest of the configurations. A duty cycle of 100% indicates that no reduction of duty cycle is needed to ensure compliance with the limits.

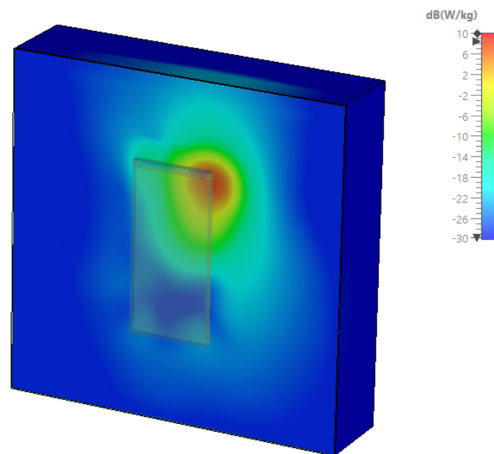


**Figure 4.34:** Simulated 10g spatial-averaged SAR distribution for the flat phantom configuration with back down at 3.5 GHz, 23 dBm and separation distance of 0 mm.

**Table 4.3:** 1g and 10g SAR (W/kg) and DC results for the flat phantom at 2 GHz ( $P_t = 23$  dBm) and different separation distances (mm).

| Configuration              | Sep. | 1g SAR | DC  | 10g SAR | DC   |
|----------------------------|------|--------|-----|---------|------|
| Screen against phantom     | 0    | -      | -   | 2.79    | 72%  |
|                            | 5    | 4.07   | 39% | -       | -    |
| Back against phantom       | 0    | -      | -   | 5.90    | 34%  |
|                            | 5    | 4.57   | 35% | -       | -    |
| Top against phantom        | 0    | -      | -   | 3.04    | 100% |
| Right side against phantom | 0    | -      | -   | 1.87    | 100% |

**Table 4.4:** 1g and 10g SAR (W/kg) and DC results for the flat phantom at 3.5 GHz ( $P_t = 23$  dBm) and different separation distances (mm).

| Configuration              | Sep. | 1g SAR | DC  | 10g SAR | DC   |
|----------------------------|------|--------|-----|---------|------|
| Screen against phantom     | 0    | -      | -   | 4.80    | 42%  |
|                            | 5    | 4.35   | 37% | -       | -    |
| Back against phantom       | 0    | -      | -   | 7.09    | 28%  |
|                            | 5    | 3.64   | 44% | -       | -    |
| Top against phantom        | 0    | -      | -   | 1.15    | 100% |
| Right side against phantom | 0    | -      | -   | 7.08    | 57%  |

### 4.3.2 SAM phantom results

All the results for the different configurations at 2 GHz and 3.5 GHz are shown in Tables 4.5 and 4.6, respectively.

At 2 GHz, the configuration that presents the highest peak spatially averaged 1g SAR corresponds to the left-ear position at a tilt angle of  $0^\circ$ , as shown in Figure 4.35. The distribution indicates that the most localized absorption occurs when the antenna is placed closest to the head, which corresponds to a tilt angle of  $0^\circ$ . In contrast, the highest 10g spatially averaged SAR is obtained for the left-ear position with a tilt angle of  $15^\circ$ , as shown in Figure 4.36. Although in general both 1g and 10g SAR can be expected to reach their peak values under the same configuration, the present results show that this is not always the case. Therefore, the maximum SAR configuration not only depends on the antenna location, but also on the averaging mass used in the SAR computation.

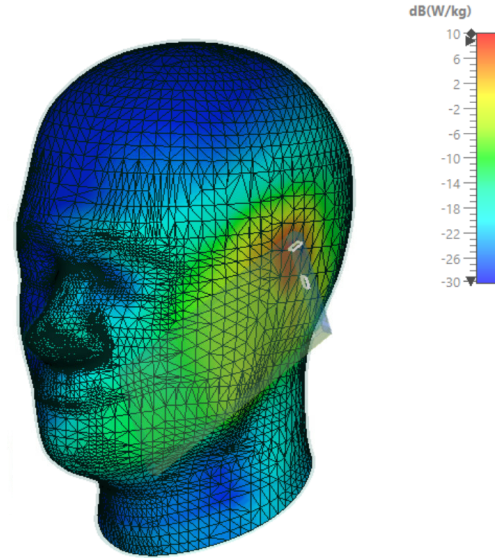


**Figure 4.35:** Simulated spatial-average 1g SAR distribution for the SAM phantom configuration leading to highest SAR, at 2 GHz and 23 dBm (Left ear,  $0^\circ$  tilt angle).

At 3.5 GHz, the configuration with the highest values for both the 1g and 10g SAR corresponds to the left ear position at a tilt angle of  $0^\circ$ , as shown in Figures 4.37 and 4.38, respectively. This indicates that, at this frequency, the highest SAR for this antenna occurs when the smartphone is closest to the cheek.

Figure 4.39 illustrates the farfield gain radiation pattern of the antenna in the presence of the SAM phantom for left-ear configuration. In both frequency bands, the phantom perturbs the radiation pattern, as part of the transmitted energy is reflected and part is absorbed by the tissue-equivalent material, instead of being transmitted into free space.

At 2 GHz, this effect leads to reduced radiation toward the upper hemisphere

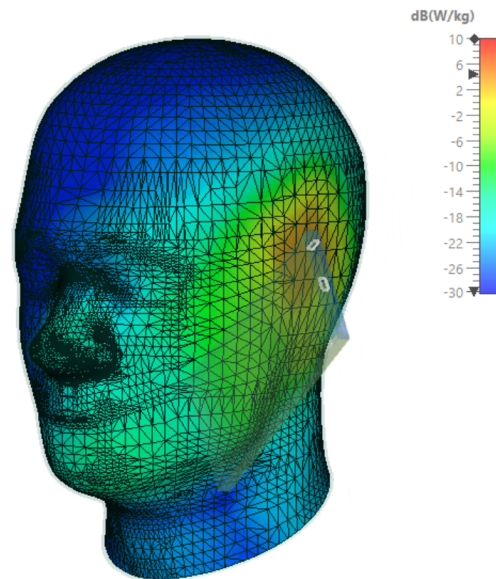


**Figure 4.36:** Simulated spatial-average 10g SAR distribution for the SAM phantom configuration leading to highest SAR, at 2 GHz and 23 dBm (Left ear 15° tilt).

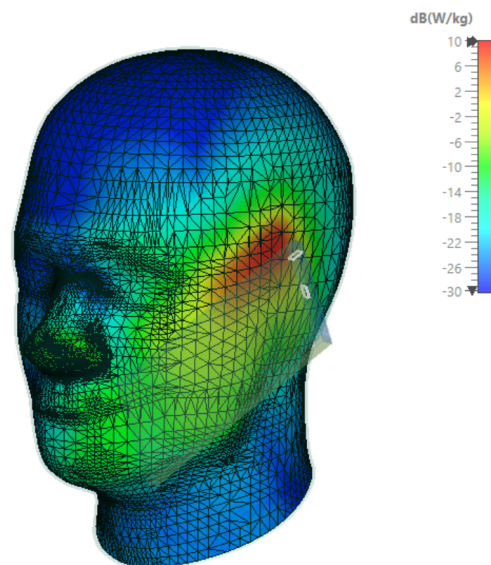


**Figure 4.37:** Simulated spatial-average 1g SAR distribution for the SAM phantom configuration leading to highest SAR, at 3.5 GHz and 23 dBm (Left ear 0° tilt).

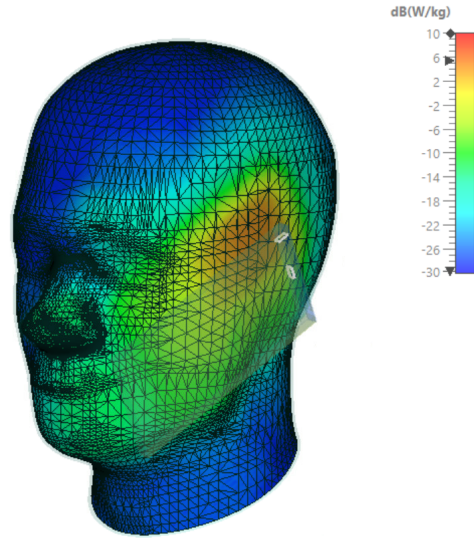


**Figure 4.38:** Simulated spatial-average 10g SAR distribution for the SAM phantom configuration leading to highest SAR, at 3.55 GHz and 23 dBm (Left ear 0° tilt).

**Table 4.5:** SAR and DC results for the SAM phantom at 2 GHz ( $P_t = 23$  dBm).

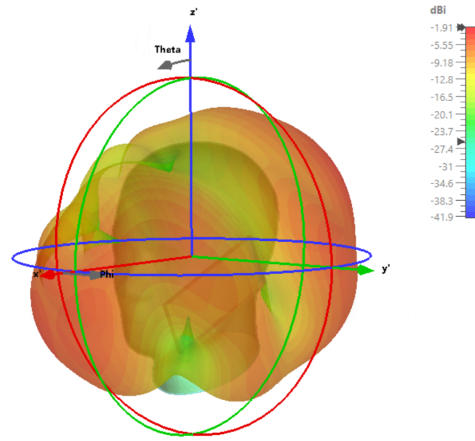
| Simulation  | Configuration      | $P_t = 23$ dBm, $f = 2$ GHz |     |         |      |
|-------------|--------------------|-----------------------------|-----|---------|------|
|             |                    | 1g SAR                      | DC  | 10g SAR | DC   |
| SAM phantom | Tilt 0° left ear   | 7.53                        | 21% | 2.71    | 74%  |
|             | Tilt 0° right ear  | 5.32                        | 30% | 1.98    | 100% |
|             | Tilt 15° left ear  | 7.39                        | 22% | 2.80    | 71%  |
|             | Tilt 15° right ear | 6.35                        | 25% | 2.45    | 82%  |

**Table 4.6:** SAR and DC results for the SAM phantom at 3.5 GHz ( $P_t = 23$  dBm).

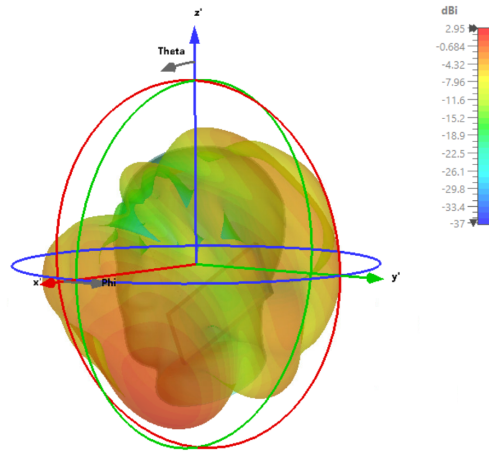
| Simulation  | Configuration      | $P_t = 23$ dBm, $f = 3.5$ GHz |     |         |      |
|-------------|--------------------|-------------------------------|-----|---------|------|
|             |                    | 1g SAR                        | DC  | 10g SAR | DC   |
| SAM phantom | Tilt 0° left ear   | 9.89                          | 16% | 3.58    | 56%  |
|             | Tilt 0° right ear  | 5.08                          | 32% | 1.89    | 100% |
|             | Tilt 15° left ear  | 5.91                          | 27% | 2.17    | 92%  |
|             | Tilt 15° right ear | 1.98                          | 81% | 0.76    | 100% |

and redistributes the transmitted power towards other directions. This effect is very relevant for NTN operation, since the intended link relies on efficient radiation towards the sky. Any reduction in gain in this region directly impacts the available link budget and may therefore reduce the achievable uplink performance.

A similar perturbation is observed at 3.5 GHz, although the resulting pattern distortion is more irregular. In this case, the presence of the SAM phantom has a strong influence on the field distribution. The results therefore confirm that the human head modifies the radiation characteristics of the antenna.



(a) 2 GHz from Antenna Structure 1.



(b) 3.5 GHz from Antenna Structure 2.

**Figure 4.39:** Gain radiation pattern of the antennas in the presence of the SAM phantom.

### 4.3.3 EMF exposure discussion

Every wireless device going to market has to be tested for compliance with the EMF regulations and must meet the EMF limits.

The results indicate that SAR levels are strongly influenced by antenna placement. In particular, the highest SAR at 3.5 GHz is obtained at the left-ear,  $0^\circ$  tilt configuration. For  $P_t = 23$  dBm, the peak spatially averaged 1g SAR is 9.89 W/kg. A duty cycle of 16% is therefore required to ensure compliance with EMF exposure limits at any distance from the device, including direct contact (i.e., touch compliance). When the accepted power is increased to 26 dBm, the 1g SAR rises to 19.55 W/kg and the duty cycle is reduced to 8%. For  $P_t = 29$  dBm, the 1g SAR reaches 39.09 W/kg, and the corresponding duty cycle drops to 4%, which is the lowest value obtained in all considered scenarios. For the 10g SAR, the resulting duty-cycle of 14% is higher than that for the 1g case.

At 2 GHz, the highest SAR case is also obtained for the left-ear,  $0^\circ$  tilt configuration. For  $P_t = 23$  dBm, the peak spatially averaged 1g SAR is 7.53 W/kg, which corresponds to a duty cycle of 21%. When the accepted power is increased to 26 dBm, the 1g SAR rises to 15.06 W/kg and the duty cycle is reduced to 11%. For  $P_t = 29$  dBm, the 1g SAR reaches 30.12 W/kg, and the corresponding duty cycle decreases to 5%, which is the lowest value obtained at 2 GHz.

The antenna assumptions also influence the obtained results. Due to the simplified UE model, additional efficiency losses present in practical implementations caused by internal components are not considered. As a result, the simulated antennas exhibit relatively high total efficiencies, reaching -1.5 dB (70.5%) for the 2 GHz antenna and -0.8 dB (83%) for the 3.5 GHz antenna. In comparison, smartphones operating in FR1 are typically associated with the nominal UE Power Class 3 limit of 23 dBm, although their measured free-space Total Radiated Power (TRP) at 3.5 GHz is generally lower due to antenna and implementation losses; for example, values ranging from 9.53 dBm to 18.55 dBm were reported in [61]. For a transmit power of 23 dBm, a lossless antenna would transmit 23 dBm; the difference between the transmit power and the TRP reflects the total antenna efficiency. For the reported devices, this corresponds to an efficiency ranging from -13.5 dB to -4.5 dB. Compared to these values, the antennas considered in this work exhibit higher efficiency, and the obtained results can therefore be regarded as conservative. In practice, higher losses would mean that higher transmit power or duty cycle are achieved while maintaining touch compliance. Additionally, advanced power control techniques [62, 63] can further increase instantaneous power over short intervals while ensuring that time-averaged SAR remains within limits.

If data transmission is only allowed during a fraction of the time while keeping a fixed bitrate during active periods, the overall amount of data transmitted over time is reduced proportionally to the fraction of time the transmitter is active. However, these duty-cycle levels may still be sufficient for certain applications, such as data exchanges or calls. In this context, intelligent proximity sensing could be used to relax duty-cycle constraints while maintaining EMF compliance, by enabling dynamic adjustment of the duty cycle according to the use position instead of relying on a fixed value.

The differences observed between the flat and SAM phantom results highlight

the influence of geometry and curvature on the exposure scenario.

The presented SAR results are based on the highest exposure scenarios. However, in practice, a wide range of operating scenarios lead to significantly lower SAR levels. For instance, changing the device location in the head position from one ear to the other. As a result, the duty-cycle limitations derived from the highest SAR configurations should be interpreted as conservative bounds. In many situations, higher duty cycles, and therefore more continuous transmission may still be possible while remaining compliant with exposure limits, especially if the human body proximity and its orientation can be detected.

This chapter discusses the main findings of this thesis in relation to the initial objectives. The main goal of this work was to investigate the feasibility of Direct-to-Device Non-Terrestrial Network communications by using a smartphone integrated antenna, while ensuring compliance with electromagnetic exposure limits. To achieve this, three main topics were addressed: antenna design, link budget analysis and EMF exposure evaluation.

The combined analysis of the three main topics is essential to understand the overall viability of the proposed solution. The antenna performance directly affects the link budget through the achievable gain and polarization, while the required transmit power puts limitations on the transmit time in order to ensure compliance with EMF limits. Therefore, the discussion focuses on the interaction between antenna characteristics, link budget, and EMF exposure, highlighting the main trade-offs and practical implications of the design.

One of the proposed antennas successfully achieves dual-band CP in the targeted NTN frequency bands through the interaction of different resonant modes. The results show a better CP performance in the UL band at 2 GHz than in the DL band at 2.185 GHz. This behaviour is explained by the interaction of different modes, which change their behaviour depending on the main tuned geometrical parameters, better optimized for the lower frequency band. Specifically tuned parameters help improve the UL performance, which is limited by UE transmit power, at the expense of DL performance.

A key observation is that the AR has lower values at higher elevation angles, which means that the CP performance is better in directions close to the zenith. This characteristic is desired for NTN links, due to the relative position of the satellite to the user. Greater advantage of this behaviour could be achieved if position awareness were to be introduced to the user, where the user could point to the satellite direction to attain the best CP performance.

The gain results show a difference between the UL and DL performances. At 2 GHz, the antenna achieves a maximum gain of 3.5 dBi in the upper hemisphere, with a radiation pattern directed toward the zenith. In contrast, at the DL frequency of 2.185 GHz, the gain in the upper hemisphere is reduced to around 1 dBi. This difference reflects an implicit design trade-off that prioritizes uplink performance. In NTN scenarios the downlink benefits from the significantly higher transmit power at the satellite. Therefore, the resulting gain difference between UL and DL are not necessarily a limitation but a reasonable design choice that

aligns with system-level constraints.

The antenna exhibits high radiation efficiency, exceeding 80% across the bands of interest, while the total efficiency is reduced to around 70%, which reflects the effect of mismatch losses. These losses can be attributed to the compromise in the design of the matching network. Due to the size constraints and the cost efficiency needed in smartphones, a simple LC network was chosen.

The integration of the C-band antenna introduces additional challenges, especially due to the mutual coupling between the two antenna elements. The initial coupling levels were relatively high, but were successfully reduced through the introduction of notch filters specifically designed to suppress radiation from the opposite antenna.

The link budget analysis demonstrates that an NTN at 2 GHz is feasible under the considered conditions. The results show that the link is power-limited, which is consistent with the large propagation distances and low transmit power of the user equipment. The elevation angle has a strong impact on the results, with higher elevation angles resulting in lower path losses and improved link performance. This leads to variations in the achievable link quality. In a real implementation of this system, this variability would need to be addressed through adaptive modulation and coding schemes.

The analysis of the achievable bitrates shows that at 2 GHz the system can support data rates ranging from hundreds of kbps to several Mbps, depending mainly on transmit power and elevation angle. This indicates that the proposed antenna is capable of supporting several NTN use cases, including emergency messaging, voice communication, and potentially low rate data services such as online browsing or standard resolution video calls, depending on system-level implementations. In contrast, the results at 3.5 GHz show reduced performance. The higher path losses, combined with the lower antenna gain in the endfire direction result in a substantial degradation of the link budget. Nevertheless, the results show that emergency messaging could still be supported. Improved antenna designs, specifically optimized for NTN requirements, could enhance the link performance at 3.5 GHz and support a broader range of services.

The results clearly demonstrate that antenna performance plays a critical role in determining the feasibility of NTN communications. Improvements in antenna gain and use of CP directly translate in higher carrier-to-noise ratio and thus increased data rates. This confirms the importance of an NTN specific antenna design, which clearly improves the system performance and provides a feasible solution for future implementation in 6G systems.

The EMF analysis shows that SAR values depend on the device position and proximity to the human body. The highest SAR values are observed in head exposure scenarios. The comparison with international exposure limits shows that compliance can be achieved, but may require the implementation of duty cycle limitations, especially at higher transmit power levels. This finding is particularly relevant for the studied system, as higher transmit power significantly improves the link quality, and it might be beneficial to maximize the transmit power while ensuring EMF compliance.

Overall, the results demonstrate that direct-to-device communication using a smartphone-integrated antenna is technically possible. The proposed S-band

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antenna provides suitable radiation characteristics that enable good link performance, while the C-band antenna achieves a lower level of performance for low-rate applications. However, the analysis also shows that the performance depends on transmit power while ensuring EMF compliance. These findings reveal that the viability of NTN links is not determined by a single stage, but by the joint interaction of the multiple system parameters.



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## Conclusions and future work

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### 6.1 Conclusions

This thesis has demonstrated the feasibility and limitations of enabling direct-to-device NTN communication using a smartphone-integrated antenna, considering both link performance and electromagnetic exposure. A circularly polarized dual-band antenna and a wideband linearly polarized antenna were designed and evaluated, and their impact on system performance was studied through link budget analysis and EMF exposure analysis.

The proposed S-band antenna successfully achieves the dual-band operation in the 2 GHz NTN bands defined by 3GPP, and the C-band antenna provides an additional wideband operation with LP. The results show that the S-band antenna has a radiation pattern directed to the upper hemisphere, enough gain and CP to support NTN communication. There is a performance difference between the UL and DL bands. However, from a system-level perspective, UL performance is typically more affected by the lower transmit power available at the UE, and, therefore, is prioritized.

The link budget analysis demonstrates that D2D NTN communication at 2 GHz is feasible under the considered assumptions. System performance is dependent on elevation angle and transmit power. The achieved data rates are in the range of hundreds of kbps to several Mbps, indicating the potential to support voice calls and medium data rate services such as web browsing or video calls. On the other hand, the 3.5 GHz band exhibits reduced performance, mainly due to higher propagation losses, lower antenna gain in the relevant directions, and polarization loss, although still being able to support text messaging or emergency services.

The EMF exposure analysis shows that the proposed antenna meets international SAR limits, but the use of duty cycle constraints in some smartphone positions relative to the human body and for higher transmit power might be needed.

### 6.2 Future work

Although this thesis has addressed the design of an NTN antenna, the analysis of the link budget, and the evaluation of EMF exposure, several directions for future

work could be conceived.

First, the proposed antenna design has only been validated through simulations. Fabricating and measuring a prototype would therefore be an important next step in order to verify its performance under real-world conditions. In addition, the influence of internal smartphone components and the display on antenna behaviour should be investigated, since these elements may affect impedance matching, radiation characteristics, and overall efficiency.

Second, the link budget analysis could be extended by considering more realistic and dynamic conditions, such as satellite movement, Doppler effects, and additional propagation losses. Furthermore, a more in-depth bitrate evaluation could be carried out through the analysis of specific 5G packet structures and protocols, to get more precise values of the achievable throughput. A worst-case analysis including the combined effect of all relevant losses would also provide valuable insight into system performance under more limiting conditions. Additionally, the DL case could be computed to get the whole-system performance results.

Finally, the radiation pattern obtained using the SAM phantom should be incorporated into the link budget analysis to quantify how human-body proximity affects the achievable bitrate due to its impact on the antenna performance. Additional investigations could also examine different hand configurations and usage scenarios, as these may influence antenna performance.

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## Antennas substrates layout

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Appendix A illustrates the substrate configuration in the smartphone structure. ROGERS 5880 substrate is highlighted in green, and FR4 substrate is highlighted in blue.

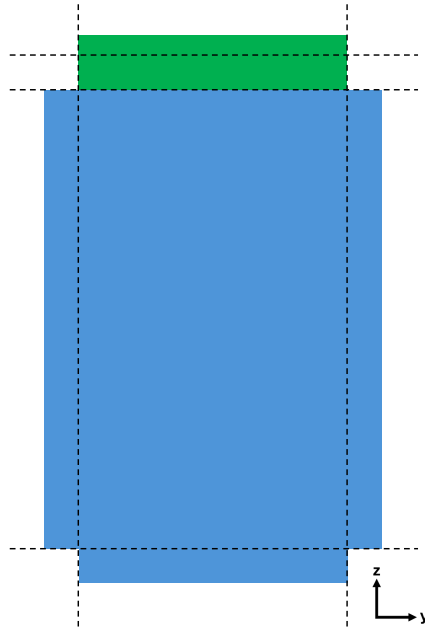


**Figure A.1:** Substrates layout of the proposed structure.



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## Surface currents of S-band antenna

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This Appendix presents the surface currents of the S-band antenna for instantaneous phase values  $0^\circ$ ,  $90^\circ$ ,  $180^\circ$  and  $270^\circ$  for the relevant resonant modes.

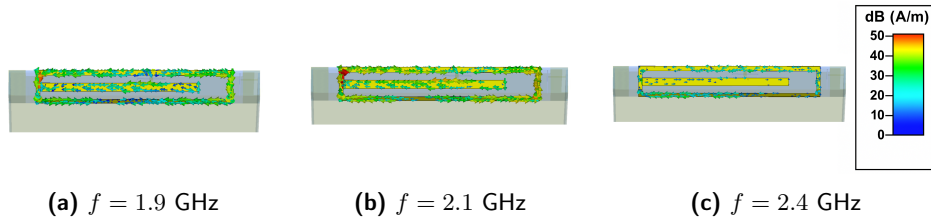


**Figure B.1:** Surface current distribution for excitation phase  $0^\circ$  at the three main mode frequencies.

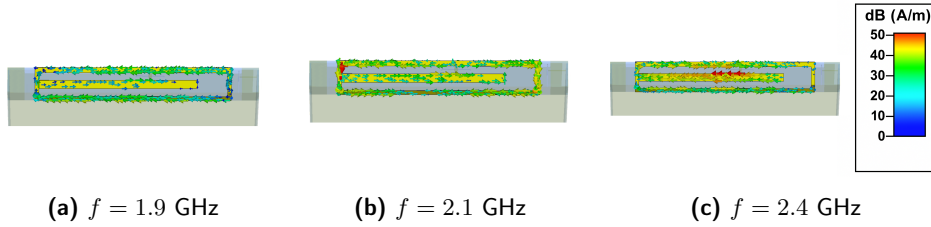


**Figure B.2:** Surface current distribution for excitation phase  $90^\circ$  at the three main mode frequencies.

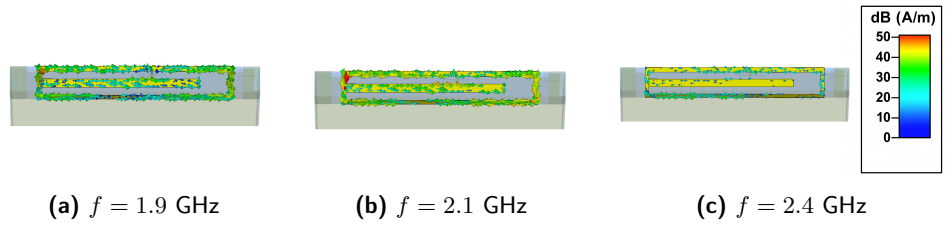


**Figure B.3:** Surface current distribution for excitation phase  $180^\circ$  at the three main mode frequencies.

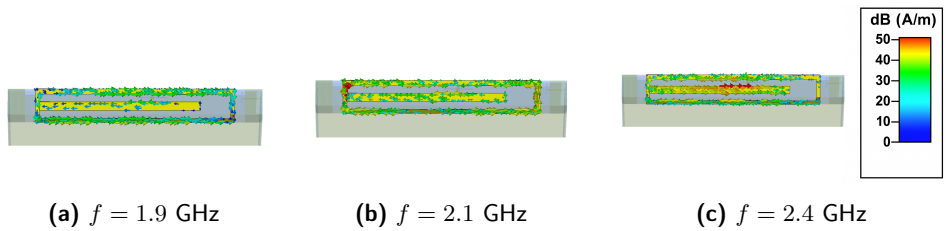


**Figure B.4:** Surface current distribution for excitation phase  $270^\circ$  at the three main mode frequencies.

## 3D gain radiation patterns

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This Appendix presents the 3D IEEE gain radiation patterns of both S-band and C-band antenna at their working frequencies.

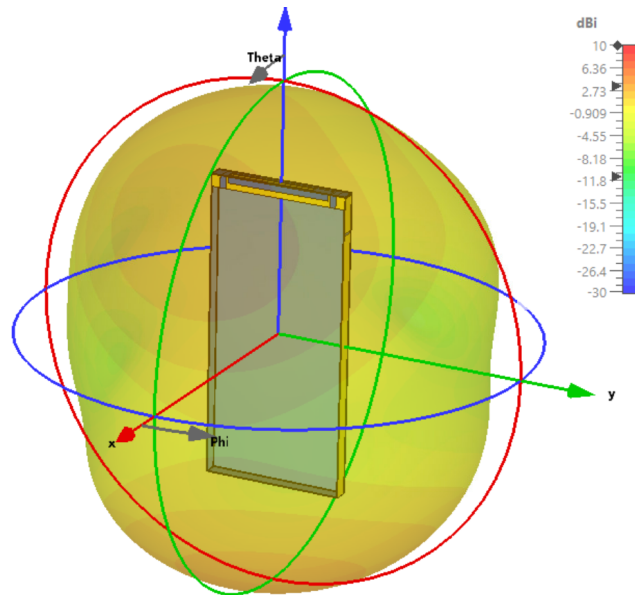


**Figure C.1:** 3D far-field gain pattern of Antenna Structure 1 at  $f = 2.0$  GHz.

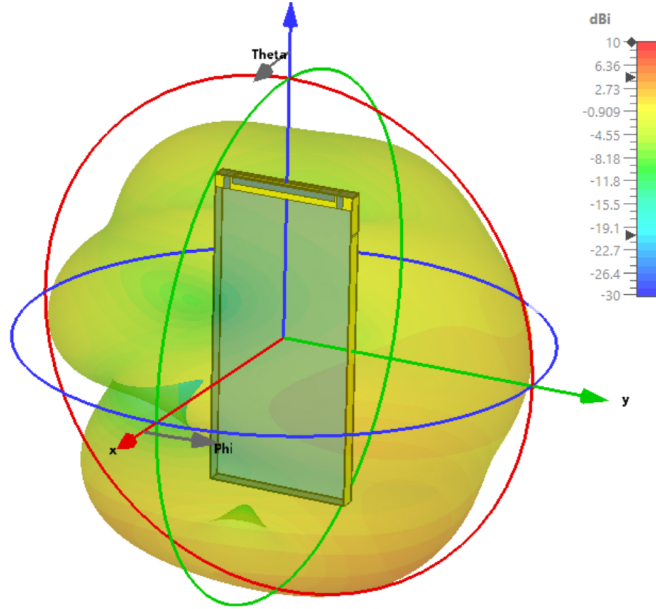


**Figure C.2:** 3D far-field gain pattern of Antenna Structure 1 at  $f = 2.185$  GHz.

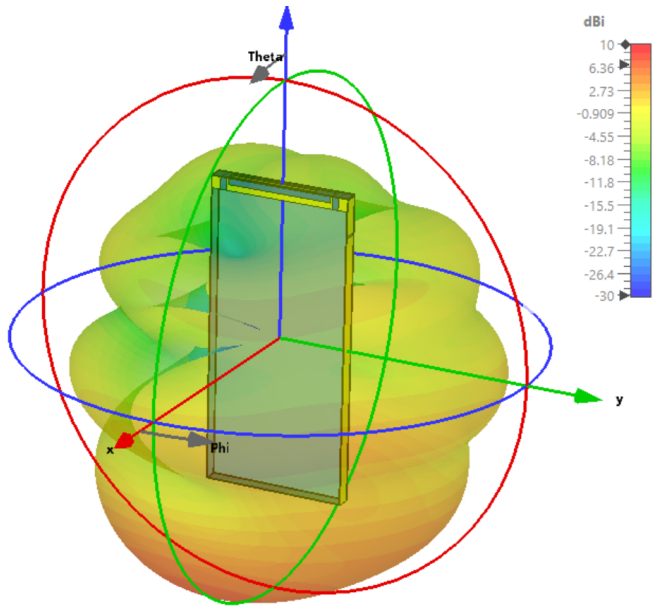


**Figure C.3:** 3D far-field gain pattern of Antenna Structure 2 at  $f = 3.5$  GHz.

## SAR simulation results

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| Simulation       | Configuration              | Separation | P=-43 dBm |     |         |           |        |     | P=-29 dBm |      |        |           |         |       | P=-26 dBm |     |         |           |        |    | FCC          |               | ICNIRP       |               |         |      |   |   |      |   |      |
|------------------|----------------------------|------------|-----------|-----|---------|-----------|--------|-----|-----------|------|--------|-----------|---------|-------|-----------|-----|---------|-----------|--------|----|--------------|---------------|--------------|---------------|---------|------|---|---|------|---|------|
|                  |                            |            | f=2 GHz   |     |         | f=3.5 GHz |        |     | f=2 GHz   |      |        | f=3.5 GHz |         |       | f=2 GHz   |     |         | f=3.5 GHz |        |    | SAR 1g limit | SAR 10g limit | SAR 1g limit | SAR 10g limit |         |      |   |   |      |   |      |
|                  |                            |            | SAR 1g    | DC  | SAR 10g | DC        | SAR 1g | DC  | SAR 10g   | DC   | SAR 1g | DC        | SAR 10g | DC    | SAR 1g    | DC  | SAR 10g | DC        | SAR 1g | DC |              |               |              |               | SAR 10g | DC   |   |   |      |   |      |
|                  | Screen against phantom     | 0          | -         | -   | 2.79    | 72%       | -      | -   | 4.80      | 42%  | -      | -         | 5.99    | 36%   | -         | -   | 9.61    | 21%       | -      | -  | 11.17        | 18%           | -            | -             | 19.22   | 10%  | - | - | 1.6  | - | 2.00 |
|                  | Back against phantom       | 5          | 4.07      | 39% | -       | -         | 4.35   | 37% | -         | -    | 8.13   | 20%       | -       | 16.27 | 10%       | -   | 16.27   | 10%       | -      | -  | 23.58        | 8%            | -            | -             | 28.57   | 7%   | - | - | -    | - |      |
|                  |                            | 0          | -         | -   | 5.90    | 34%       | -      | -   | 7.09      | 28%  | -      | -         | 11.79   | 17%   | -         | -   | 14.19   | 14%       | -      | -  | 18.29        | 9%            | -            | -             | -       | -    | - | - | -    | - |      |
|                  |                            | 5          | 4.57      | 35% | -       | -         | 3.64   | 44% | -         | -    | 9.15   | 17%       | -       | 7.28  | 22%       | -   | -       | 14.95     | 11%    | -  | -            | -             | -            | -             | -       | -    | - | - | -    | - |      |
|                  | Top against phantom        | 0          | -         | -   | 3.04    | 100%      | -      | -   | 1.15      | 100% | -      | -         | 6.08    | 66%   | -         | -   | 2.30    | 100%      | -      | -  | 12.15        | 33%           | -            | -             | 4.60    | 87%  | - | - | -    | - | -    |
| Flat phantom     | Right side against phantom | 0          | -         | -   | 1.87    | 100%      | -      | -   | 7.08      | 57%  | -      | -         | 3.74    | 100%  | -         | -   | 14.15   | 28%       | -      | -  | 7.49         | 53%           | -            | -             | 28.30   | 14%  | - | - | 4.00 | - | 4.00 |
|                  | Left side against phantom  | 0          | -         | -   | 0.33    | 100%      | -      | -   | 0.03      | 100% | -      | -         | 0.67    | 100%  | -         | -   | 0.07    | 100%      | -      | -  | 1.34         | 100%          | -            | -             | 0.14    | 100% | - | - | -    | - | -    |
|                  |                            | 0          | -         | -   | 0.33    | 100%      | -      | -   | 0.03      | 100% | -      | -         | 0.67    | 100%  | -         | -   | 0.07    | 100%      | -      | -  | 1.34         | 100%          | -            | -             | 0.14    | 100% | - | - | -    | - | -    |
|                  | ear                        | 0          | 7.53      | 21% | 2.71    | 74%       | 9.89   | 16% | 3.58      | 56%  | 15.06  | 11%       | 5.42    | 37%   | 19.78     | 8%  | 7.16    | 28%       | 30.12  | 5% | 10.83        | 18%           | 39.56        | 4%            | 14.31   | 14%  | - | - | -    | - | -    |
|                  | ear                        | 0          | 5.32      | 30% | 1.98    | 100%      | 5.08   | 32% | 1.89      | 100% | 10.65  | 15%       | 3.96    | 50%   | 10.15     | 16% | 3.78    | 53%       | 21.29  | 8% | 7.92         | 25%           | 20.31        | 8%            | 7.55    | 28%  | - | - | -    | - | -    |
| Head SAM phantom | Titl 0 deg, right ear      | 0          | 7.39      | 22% | 2.80    | 71%       | 5.91   | 27% | 2.17      | 92%  | 14.78  | 11%       | 5.60    | 36%   | 11.81     | 14% | 4.34    | 46%       | 29.57  | 5% | 11.20        | 18%           | 23.62        | 7%            | 8.69    | 23%  | - | - | -    | - | 1.6  |
|                  | Titl 15 deg, left ear      | 0          | 7.39      | 22% | 2.80    | 71%       | 5.91   | 27% | 2.17      | 92%  | 14.78  | 11%       | 5.60    | 36%   | 11.81     | 14% | 4.34    | 46%       | 29.57  | 5% | 11.20        | 18%           | 23.62        | 7%            | 8.69    | 23%  | - | - | -    | - | 2.00 |
|                  | Titl 15 deg, right ear     | 0          | 6.35      | 25% | 2.45    | 82%       | 1.98   | 81% | 0.76      | 100% | 12.69  | 13%       | 4.89    | 41%   | 3.96      | 40% | 1.52    | 100%      | 25.39  | 6% | 9.78         | 20%           | 7.93         | 20%           | 3.03    | 66%  | - | - | -    | - | -    |

**Figure D.1:** Complete SAR and DC results for the flat phantom and head SAM phantom simulations at  $P_t = 23, 26$ , and  $29$  dBm, including FCC and ICNIRP limits.

## Atmospheric effects calculations

The propagation losses shown in Figure E.1 were calculated as the sum of rain attenuation, atmospheric gaseous absorption, and tropospheric scintillation

$$A_{\text{tot}} = A_{\text{rain}} + A_{\text{gas}} + A_{\text{scint}}, \quad (\text{E.1})$$

for the 0.01% time percentage for which the scintillation fade depth and rain attenuation are exceeded.

Rain attenuation was estimated following the ITU-R recommendation. The specific rain attenuation is

$$\gamma_R = kR_{0.01}^\alpha, \quad (\text{E.2})$$

where  $R_{0.01}$  is the rainfall rate exceeded for 0.01% of an average year, and  $k$  and  $\alpha$  are the frequency- and polarization-dependent coefficients from ITU-R P.838 [23]. The rain attenuation exceeded for 0.01% of the time is then

$$A_{\text{rain},0.01} = \gamma_R L_{\text{eff}}, \quad (\text{E.3})$$

where  $L_{\text{eff}}$  is the effective slant path length through the rain region, calculated according to ITU-R P.618 [22].

Atmospheric gaseous absorption was calculated according to ITU-R P.676 [21]. The total gaseous specific attenuation is

$$\gamma_{\text{gas}}(f) = \gamma_o(f) + \gamma_w(f), \quad (\text{E.4})$$

where  $\gamma_o$  and  $\gamma_w$  are the oxygen and water-vapour contributions, respectively. The slant-path gaseous attenuation is approximated by

$$A_{\text{gas}} \approx \frac{A_{\text{gas,zenith}}}{\sin \alpha}. \quad (\text{E.5})$$

Tropospheric scintillation was calculated using the ITU-R P.618 model [22]. The scintillation fade depth exceeded for  $p\%$  of the time is:

$$A_{\text{scint}}(p) = a(p)\sigma_{\text{scint}}, \quad (\text{E.6})$$

where  $a(p)$  is the time-percentage factor and  $\sigma_{\text{scint}}$  is the scintillation standard deviation.

The parameters used in the calculation were  $D_{\text{ant}} = 0.05$  m,  $\eta_{\text{ant}} = 0.7$ ,  $p = 0.01\%$ . Here,  $D_{\text{ant}}$  is the assumed effective smartphone antenna,  $\eta_{\text{ant}}$  is the antenna efficiency, and  $p$  is the scintillation time percentage. The surface meteorological parameters were temperature  $T_0 = 283.15$  K, atmospheric pressure  $P_0 = 1013.25$  hPa, and relative humidity  $RH_0 = 80\%$ . These correspond to a temperature of approximately  $10^\circ\text{C}$ , standard atmospheric pressure, and a humid-climate assumption [64]. These values were used as representative surface conditions in Lund, Sweden for the propagation-loss results in Figure E.1.

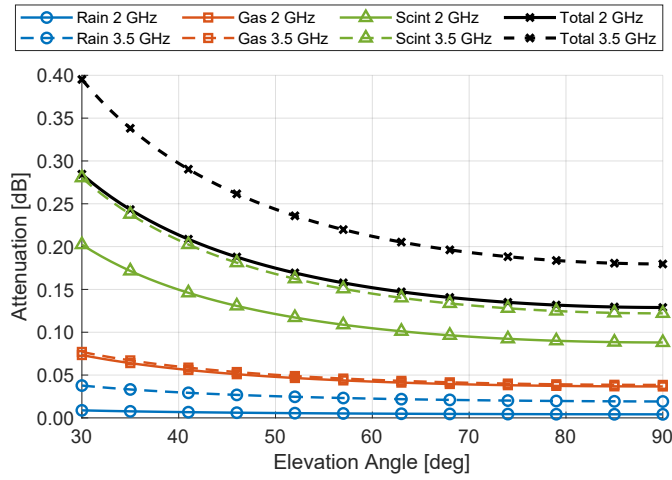


**Figure E.1:** Rain, gas and scintillation caused attenuation for the 2 GHz and 3.5 GHz links, for a UE located in Lund, Sweden.



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