

**UNIVERSITÀ DEGLI STUDI DI CASSINO
E DEL LAZIO MERIDIONALE**

CORSO DI DOTTORATO IN METODI, MODELLI E TECNOLOGIE PER L'INGEGNERIA

DIPARTIMENTO DI INGEGNERIA ELETTRICA E DELL'INFORMAZIONE
"MAURIZIO SCARANO"



**DESIGN OF WEARABLE SENSORS FOR MEASURING 5G
AND BEYOND SIGNALS FOR HEALTH RISK ASSESSMENT**

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In Partial Fulfillment of the Requirements for the Degree of
PHILOSOPHIAE DOCTOR
in Electrical and Information Engineering

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XXXVIII Cycle

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Date: 12/05/2026

Author: Gaetano Chirico

Title: Design of wearable sensors for measuring 5G and beyond signals for health risk assessment

Department: DIPARTIMENTO DI INGEGNERIA ELETTRICA E DELL'INFORMAZIONE
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*A mia moglie Giusi, per la pazienza infinita ed il sostegno incrollabile,
Alla piccola Carol, il mio capolavoro, con la speranza che tu possa volare sempre più in alto*

Acknowledgements

I primi ringraziamenti che sento di fare vanno, senza ombra di dubbio, ai Supervisor del mio progetto di ricerca, Professori Fulvio Schettino e Marco Donald Migliore. Entrambi sono stati per me molto più che dei semplici tutor; ho avuto la fortuna di lavorare con loro su diverse tematiche estremamente interessanti e di accrescere non solo il mio bagaglio culturale ma anche, e soprattutto quello umano. Il Professore Schettino mi è stato più volte vicino nei diversi momenti di difficoltà che hanno caratterizzato il mio percorso, mostrandosi sempre pronto a strapparmi un sorriso ed attivandosi sempre e comunque, anche quando non era tenuto a farlo, per tendermi una mano nel momento del bisogno.

Considerazioni analoghe valgono anche per il Professore Migliore e per gli altri componenti del gruppo di ricerca di cui fanno parte i Professori Schettino e Migliore, vale a dire il Professore Daniele Pinchera ed il Professore Mario Lucido. Il Professore Migliore è dotato di un'energia inesauribile e riesce a rallegrare anche le situazioni più complicate; lavorare con lui è un'esperienza che difficilmente si può descrivere a parole in quanto, nonostante si lavori duramente, il clima in cui si opera è sempre gioviale e, quindi, non si avverte il peso di ciò che si sta facendo. Il Professore Pinchera, ovviamente, non è da meno: è una fonte inesauribile di conoscenza, in diversi campi applicativi, e le ore spese con lui mi hanno arricchito in modo impagabile dal punto di vista culturale. In ultimo, ma non per importanza, voglio ringraziare il Professore Mario Lucido: analogamente agli altri docenti prima menzionati, è una fonte inesauribile di cultura, da cui ho cercato di attingere quanto più possibile, e al contempo, è una persona squisita sul lato umano. A tutti Voi dico un immenso grazie: avete creato un ambiente di lavoro sereno e stimolante che difficilmente si trova altrove; come ripeto ogni qualvolta ne ho l'occasione: nella vita ho fatto molti sbagli; una scelta di cui sicuramente non mi pentirò è quella di aver deciso di trascorrere con Voi gli anni che hanno caratterizzato questo percorso di studio.

Voglio anche ringraziare la Professoressa Rita Massa, afferente al dipartimento di Fisica dell'Università degli Studi di Napoli "Federico II" ed il Professore Gaetano Panariello: siete stati coloro con i quali ho mosso i primi passi nell'ambito della ricerca e ve ne sarò sempre grato. La Professoressa Massa è per me una persona speciale e lo stesso vale per il Professore Panariello: grazie di cuore per tutto ciò che mi avete dato e per le occasioni che mi avete fornito.

Voglio inoltre ringraziare mia moglie Giusi, per essermi sempre stata vicino e per avermi donato il regalo più bello che si potesse desiderare: la mia adorata piccolina Carol. La ringrazio un po' meno per tutti i traslochi che mi ha costretto a fare e per l'infinità di chilometri che mi ha fatto macinare, mi fa macinare e sicuramente mi farà macinare in futuro: per questo e tutto quello che è da venire (traslochi compresi), grazie.

Infine, un ringraziamento speciale a tutti i miei familiari: nei momenti di difficoltà mi siete sempre stati vicino. È una grande fortuna avervi nella mia vita. Grazie a tutti

Abstract

This thesis investigates the feasibility of a compact, low-cost wearable device for the assessment of electromagnetic field (EMF) exposure levels in prospective 6G scenarios in the 14.8–15.35 GHz frequency range. The work is motivated by the intention to anticipate future measurement requirements by developing a solution specifically conceived for frequency bands and propagation conditions that are expected to play a significant role in next-generation wireless systems, thus contributing, at least in part, to the mitigation of one of the main issues that may hinder the early deployment of new communication technologies, namely the lack of adequate instruments for measuring the EMF associated with the signals of interest.

To this end, the main characteristics of previous generations of mobile communication systems were first examined in order to gain a deeper understanding of RF-EMF exposure and of the challenges that newer generations, such as 5G and prospective 6G systems, pose to researchers involved in the development of new measurement and modeling techniques. Standardized and non-standardized measurement protocols for the assessment of RF-EMF exposure levels were then presented, together with the main characteristics of the different types of devices that may be employed for this purpose, so as to identify useful insights supporting the achievement of the main objectives of the present work. Furthermore, since the characterization of 6G signal levels is expected to raise challenges similar to those already posed by 5G signals, two real-environment measurement campaigns were discussed in order to examine the solutions currently adopted for the characterization of exposure levels associated with 5G signals.

The considerations drawn from this analysis led to the definition of a proof-of-concept architecture based on a planar radiating element, an RMS RF power detector, and a digital acquisition section. Particular attention was devoted to the design and validation of the antenna, which represents the most critical and original element of the system, since its radiation characteristics, physical footprint, and fabrication cost directly affect both the measurement performance and the overall compactness and economic viability of the proposed device. More precisely, in order to satisfy the main requirements of the target application in terms of compactness, cost-effective manufacturability, polarization behavior, and radiation performance, a single-microstrip-fed circularly polarized patch antenna, operating at 15 GHz on a 0.508 mm-thick Rogers RO4003C substrate and occupying an overall area of 22.00 mm × 12.70 mm, was designed and optimized through extensive numerical simulations carried out with CST Studio Suite.

On the basis of the obtained design, the antenna prototype was fabricated and experimentally characterized in a semi-anechoic chamber in order to verify whether it was capable of reproducing the behavior predicted by the electromagnetic simulations performed on the numerical model of the antenna. Overall, the obtained results validate the proposed proof-of-concept design and demonstrate the practical feasibility of a compact, low-cost wearable personal exposimeter for prospective 6G applications. In this sense, the work contributes to the development of devices specifically aimed for long-term personal EMF exposure assessment in realistic scenarios.

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Glossary

Abbreviation / Term	Description
1024-QAM	1024-Quadrature Amplitude Modulation
16-QAM	16-Quadrature Amplitude Modulation
1G-6G	First-Sixth Generation Technology Standard For Broadband Cellular Networks
256-QAM	256-Quadrature Amplitude Modulation
32-QAM	32-Quadrature Amplitude Modulation
3GPP	3rd Generation Partnership Project
64-QAM	64-Quadrature Amplitude Modulation
8-PSK	8-Phase Shift Keying
AAS	Active Antenna System
ADC	Analog To Digital Converter
AFM	Analogue Frequency Modulation
AGC	Automatic Gain Control
AL	Action Levels
AM	Amplitude Modulation
BER	Bit Error Rate
BPSK	Binary Phase Shift Keying
CDMA2000	Code Division Multiple Access 2000
CORESET	Control Resource Set
CP	Channel Power
CP-OFDM	Cyclic Prefix Orthogonal Frequency Division Multiplex
CQI	Channel Quality Information
CSI	Channel State Information
CSI-RS	Channel-State Information Reference Signal
DANL	Displayed Average Noise Level
dBFS	Decibels Relative To Full Scale
DC	Direct Current
DCI	Downlink Control Information
DDC	Digital Downconverter
DFT	Discrete Fourier Transform
DFT-S-OFDM	Dft-Spread Ofdm
DL	Downlink
DMRS	Demodulation Reference Signal

DPCM	Decree Of The President Of The Council Of Ministers
DSP	Digital Signal Processor
DSS	Dynamic Spectrum Sharing
DVB-T	Digital Video Broadcasting - Terrestrial
EC	European Council
EDGE	Enhanced Data Rates for Gsm Evolution
EMF	Electromagnetic Fields
EN	European Norm
ENBW	Equivalent Noise Bandwidth
EPRE	Energy Per Resource Element
EU	European Union
E-UTRA	Evolved Universal Terrestrial Radio Access
FDD	Frequency Division Duplex
FDM	Frequency Division Multiplexing
FDMA	Frequency Division Multiple Access
FEC	Forward Error Correction
FM	Frequency Modulation
FR1	Frequency Range 1
FR2	Frequency Range 2
FR3	Frequency Range 3
FWA	Fixed Wireless Access
GERAN	Gsm/Edge Radio Access Network
GMSK	Gaussian Minimum Shift Keying
GPRS	General Packet Radio Service
GPS	Global Positioning System
GSM	Global System for Mobile Communications
HSCSD	High-Speed Circuit-Switched Data
HSPA+	Evolved High Speed Packet Access
IBW	Integration Bandwidth
ICNIRP	International Commission On Non-Ionizing Radiation Protection
IF	Intermediate Frequency
IMT-2000	International Mobile Telecommunications-2000
IMT-2030	International Mobile Telecommunications For 2030
IP	Internet Protocol
IQ	In-Phase And Quadrature
IR	Ionizing Radiations
ITU-R	International Telecommunication Union - Radiocommunication Sector
LAN	Local Area Network
LNA	Low Noise Amplifier
LOS	Line Of Sight
LTE	Long Term Evolution
MAC	Medium Access Control

MBMS	Multimedia Broadcast And Multicast Service
MBSFN	Multicast/Broadcast Over A Single Frequency Network
MIB	Master Information Block
MSK	Minimum Shift Keying
NBW	Noise Bandwidth
NIR	Non-Ionizing Radiations
NR	New-Radio
NSA/SA	Non-Standalone / Standalone
NZP	Non-Zero Power
OFDM	Orthogonal Frequency Division Multiplexing
OFDMA	Orthogonal Frequency Division Multiplexing Access
PAS	Passive Antenna System
PBCH	Physical Broadcast Channel
PBCH-DMRS	Physical Broadcast Channel Demodulation Reference Signal
PCB	Printed Circuit Board
PCI	Physical Layer Cell Identity / Identities
PDCCH	Physical Downlink Control Channel
PDCCH-DMRS	Physical Downlink Control Channel Demodulation Reference Signal
PDE	Personal Distributed Exposimeter
PDSCH	Physical Downlink Shared Channel
PDSCH-DMRS	Physical Downlink Shared Channel Demodulation Reference Signal
PMI	Precoding Matrix Index
POI	Probability Of Intercept
PSK	Phase Shift Keying
PSS	Primary Synchronization Signal
PTRS	Phase Tracking Reference Signal
PUCCH	Physical Uplink Control Channel
PUSCH	Physical Uplink Shared Channel
QPSK	Quadrature Phase Shift Keying
RB	Resource Block
RBW	Resolution Bandwidth
RF/IF	Radio Frequency / Intermediate Frequency
RF-EMF	Radiofrequency Electromagnetic Fields
RI	Rank Indicator
RMS	Root Mean Square
RSS	Root Sum Of The Square
RTBW	Real-Time Bandwidth
RTSA	Real-Time Spectrum Analyzer / Analyzers
SA	Spectrum Analyzer
SAR	Specific Energy Absorption Rate
SC-FDMA	Single-Carrier Fdma
SCHEER	Scientific Committee On Health, Environmental, And Emerging Risks

SDR	Software-Defined Radio / Radios
SNR	Signal To Noise Ratio
SS	Synchronization Signals
SS/PBCH	Synchronization Signal / Physical Broadcast Channel
SSB	Ss Block
SSS	Secondary Synchronization Signal
T-DAB	Terrestrial Digital Audio Broadcasting
TDD	Time Division Duplex
TDM	Time Division Multiplexing
TDMA	Time Division Multiple Access
TRS	Tracking Reference Signal
UE	User Equipment
UHF	Ultra High Frequency
UL	Uplink
UMTS	Universal Mobile Telecommunication System
URLLC	Ultra-Reliable Low Latency Communication
VHF	Very High Frequency
W-CDMA	Wideband Code Division Multiple Access
WHO	World Health Organization
WRC-23	World Radiocommunication Conference 2023
WRC-27	World Radiocommunication Conference 2027
ZP	Zero Power

1 Chapter 1: Overview of mobile communications technologies

Mobile technologies have dramatically evolved over the past decades, leading to a fully connected society and empowering socio-economic transformations, including those for productivity, sustainability, and well-being [1]. More specifically, over the last 30 years, a remarkable growth in connectivity and in traffic density/volume has been observed, driven by the continuously increasing demand for mobile services. In this context, an unprecedented step forward was represented by the introduction of the 5G New Radio (NR) standard, which provides significantly higher throughput, substantially reduced latency, ultra-high reliability, much higher connectivity density, and an extended mobility range. Furthermore, additional advancements in this direction are expected with the introduction of the sixth generation (6G) of mobile communications, presumably starting from 2030. Indeed, it is expected that 6G networks will enable data rates on the order of 100 Gbit/s (approximately ten times the theoretical maximum speed achievable with 5G technology), with latency values below 100 μ s. However, it is clear that the adoption of new networking technologies, such as 6G, cannot disregard the assessment of potential causal relationships between the duration and levels of exposure to radiofrequency electromagnetic fields (RF-EMF) and possible effects on human health, in order to ensure the safety of individuals who, consciously or not, will inevitably be exposed to such fields. It is precisely within this framework that the present research project is positioned. Its main objective, in fact, is to provide a proof of concept for the realization, through the use of low-cost components, of a **wearable device** suitable for assessing exposure levels to upcoming 6G signals.

This choice offers the clear advantage of anticipating future needs and, therefore, making available a device capable of characterizing exposure levels to 6G signals at the time when 6G communication networks will be deployed. This aspect should not be underestimated. Indeed, one of the factors that initially limited the widespread adoption of 5G technologies was the lack of adequate tools (both from a mathematical and a hardware perspective) for the measurement of 5G signals in air. In fact, as will be better explained in the following sections, 5G NR introduced new solutions in terms of standards (frame duration, bandwidth, power allocation, etc.) and a completely new concept for managing radiation distribution (beamforming, power control). Consequently, previously adopted approaches for exposure monitoring and assessment, as well as for coverage optimization, proved to be no longer adequate making it necessary to introduce entirely new paradigms [1].

On the other hand, the intention to operate with signals for which standards have not yet been defined requires making assumptions about their specific characteristics and does not currently allow for a full in-field testing of the device under investigation. From this perspective, it is therefore useful to provide a brief overview of the evolution of mobile communications technologies, with particular

attention to the main characteristics of the different generations. An overview of the previous generations of mobile communications systems is, indeed, useful to take advantage of the developed methods and models in order to better understand the impact and the characteristics of EMF exposure and the challenges that new generations (5G and 6G) pose to researchers for the development of new measurements and modelling techniques.

In fact, it is clear that different types of signals and networking technologies cause antennas and mobile devices to generate profoundly different emission and reception power levels. Mobile phones, in particular, employ low-power transmitters, which can reach peak transmitted powers below two Watts, and are designed to automatically transmit at the minimum power required to maintain connection quality (this feature is known as adaptive power control). Also, mobile network technologies affect radiation exposure to the same extent as the device design itself. In fact, users of the same mobile device may absorb different amounts of radiofrequency emissions depending on which carrier serves them [2]. For instance, when operators shifted from 2G to 3G technologies, mobile devices operating on 3G networks used lower power levels for a larger amount of the transmit time, and their users experienced lower radiation exposure levels [3]. Therefore, the signal standards adopted in any telecommunication scheme represent one of the most critical factors not only for network performance and enhanced services, but also for compliance with health and safety requirements. Thus, as stated before, to facilitate a thorough outlook of the progress of the signal standards, a brief overview of the evolution of the technologies used will be given in the following sections.

1.1 1G, 2G and 3G technologies

The earliest form of mobile networking technology, which is now referred as **0G** (zero generation) mobile technology, was developed in the previous century and was mostly used for military purposes. Usually vehicle mounted, huge transceivers and dials were utilised, so as to achieve communication among moving units.

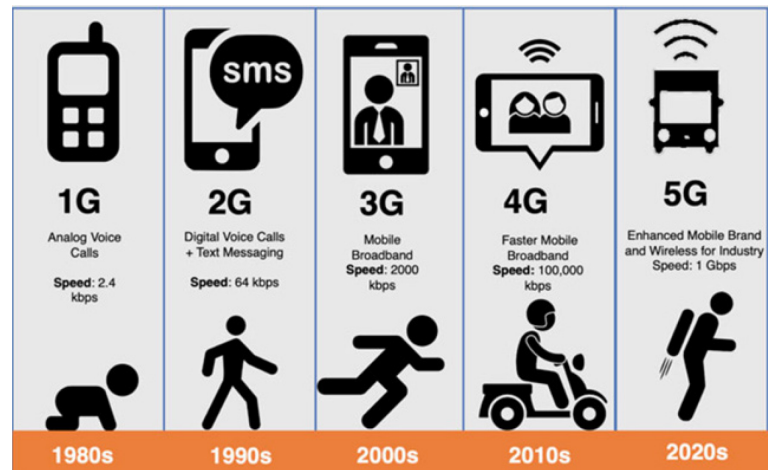


Figure 1.1 – Evolution of Mobile Networks.

1G (first generation) was later developed (Figure 1.1) and mobile phones were realized for commercial use. The key features of 1G, in terms of technology and services were briefly the following:

- the **Cell** concept developed by the Bell Labs in the 70's was utilized and the first commercial use of cellularly structured networks for mobile communications was introduced. Each cell is an area where a base station serves all mobile users in range. When a user moves from one cell to another their service is handed over to the next base station without interruption of communication (handover or handoff);
- voice was the only type of information transferred in the network;
- the information (sound) signal transfer was made possible with Analogue Frequency Modulation (AFM) with carrier frequencies in the Very High Frequency (VHF) band;
- the modulation nature defined a single analogue channel for signal transmission and reception;
- the multiplexing technique used was Frequency Division Multiplexing (FDM), in which the entire bandwidth is divided in several sections and each sub band is assigned to a certain device [4]. It is feasible to apply Time Division Multiplexing (TDM) on each of the sub bands separately.

The channel access technique used in 1G was the Frequency Division Multiple Access (FDMA), as it was the earliest and most sophisticated channel access method at the time, for transmission of multiple channels of information through a single communication medium. The FDMA scheme is a technique built on the FDM method (Figure 1.2, left). This technique can be considered the oldest and the simplest form of multiplexing. In this technique the available frequency band is divided into channels of equal bandwidth so that each conversation is carried on a different frequency sub-band [4]

and each sub-band is assigned to a certain device. Users can send data through a subchannel by modulating it on a carrier wave at the subchannel's frequency. It is also feasible to apply TDM on each sub-band separately. Guard bands are used to minimise the crosstalk between the channels, which leads to a waste of capacity. Moreover, if an FDMA channel is not in use, then it is idle as it cannot be used by other users to increase or share capacity.

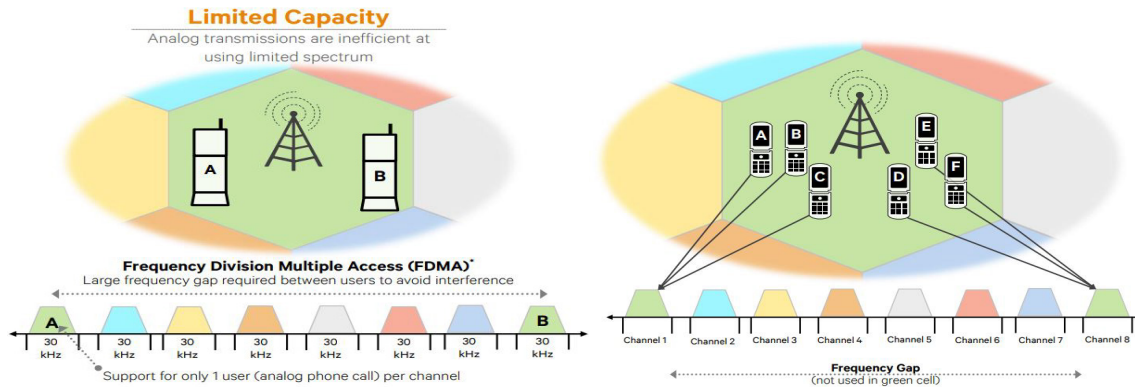


Figure 1.2 – FDMA (left) channel distribution (1G) vs TDMA (right) channel distribution (2G) [1].

The advantages of 1G systems were numerous and unprecedented. However, the capacity limitations, the quality issues, and the limitations of national standards of the first analogue mobile communication systems led to the development of a 2nd generation of cellular mobile communication systems.

The second generation of cellular network (2G) was commercially launched in 1991 (Figure 1.1) and 2G was the first full digital standard introduced, enabling a significantly more efficient use of the frequency spectrum. There were several systems deployed as part of the second generation of cellular network but the most successful and universally deployed 2G system, for over three decades, was GSM (Global System for Mobile communications), upon which all others were applied as enhancements. The main features of GSM technology are listed below [1]:

- **Gaussian Minimum Shift Keying (GMSK)** modulation. This modulation scheme is a filtered version of Minimum Shift Keying (MSK), which is itself a spectrally efficient modulation scheme for use in mobile communication systems because of its properties such as constant envelope, spectral efficiency, good Bit Error Rate (BER) performance, and self-synchronizing capability, used to overcome one of the problems encountered with standard forms of Phase Shift Keying (PSK) that is the sidebands that extend out from the carrier [5]. Later on, higher order modulation schemes were used.
- **Time Division Multiple Access (TDMA)** signalling. This allows several users to share the same frequency channel by dividing the signal into different time slots. The users transmit in rapid succession, one after the other, each using its own time slot (Figure 1.2, right). Slots from all subscribers are combined into cyclically repeated super-slots. Synchronization can be performed by periodically sending known service sequences within one or more super-slots;

- **Frequency Division Duplex (FDD)** carriers, used for establishing a full-duplex communications link by using two different radio frequencies for transmit and receive operations. More precisely, the transmission and reception frequencies are separated by a defined frequency offset; this is a solution to resolve the transmit-receive signal ratio situation, just as power control is a solution to resolve the near-far signal strength ratio problem;
- channel spacing of 200 kHz;
- designed for operation in the **900 MHz** band but was adapted also for **1800 MHz** and upon its later introduction in North America, GSM incurred further adaptation to the 800 and 1900 MHz bands, as well as other bands over the years.

GSM was designed for digital voice telephony. Later on, the High-Speed Circuit-Switched Data (HSCSD) service was introduced in the GSM Phase 2+. A significant increase in speed could then be obtained by aggregating up to four channels into a single, grouped, circuit-switched connection. HSCSD is an example of multi-slot configuration, in which all channels have the same channel mode; it has later been replaced by General Packet Radio Service (GPRS) and Enhanced Data Rates for GSM Evolution (EDGE), which offer higher data rates.

GPRS adds packet-switched functionality to GSM, which is essentially circuit switched. It offers faster data rates than plain GSM by aggregating several GSM timeslots into a single bearer, potentially up to eight. The access scheme is still TDMA, with eight basic physical channels per carrier. Therefore, one physical channel (Packet Data Channel-PDCH) is employed for data transfer, upon which several logical channels are mapped, and is defined by a sequence of TDMA frames, a time slot number, and a frequency hopping sequence. The basic radio resource is a time slot, whose physical content is called burst, and is lasting $\approx 576,9 \mu\text{s}$ and transmitting information at a modulation rate of $\approx 270,833$ ksymbol/s. Each symbol is one to five bits depending on the actual modulation used: GMSK, Quadrature Phase Shift Keying (QPSK), 8-PSK, 16QAM (Quadrature Amplitude Modulation) and 32QAM respectively. The overall slot structure in the GPRS physical channel is the same as in GSM and therefore shares the same power profile. The RF channel spacing is 200 kHz, allowing for 41 (T-GSM 380), 41 (T-GSM 410), 35 (GSM 450), 35 (GSM 480), 89 (GSM 710), 74 (GSM 750), 74 (T-GSM 810), 124 (GSM 850), 209 (ER-GSM 900), 194 (GSM 900), 374 (DCS 1 800) and 299 (PCS 1900) radio frequency channels.

EDGE is an enhancement of the GSM radio access technology to provide faster bit rates for data applications, both circuit and packet switched. As such, EDGE is realized via modifications of the existing GSM physical layer, rather than by separate, stand-alone specifications. The increased data rate was accomplished by the 8-PSK modulation technique coupled with new channel coding resulting in improved spectral efficiency. In fact, EDGE uses a combination of 8-PSK and GMSK, which improves the bit rate under virtually all radio conditions. The four coding schemes of GPRS increased to nine in EDGE, and new segmentation techniques can radically improve throughput by permitting the coding scheme to be changed on the fly in case of retransmission of a segment in rapidly changing radio conditions. EDGE is transparent to the upper layers, so it is possible to apply EDGE on top of GPRS (Enhanced GPRS), thus increasing the theoretical data rate to 384 kbit/s which is comparable with the early 3G rates. For this reason, EDGE is considered to be the bridge between the 2nd and the 3rd generation of mobile communication systems. GSM network enhanced with

HSCSD, GPRS, and EDGE is referred under the umbrella acronym GERAN (GSM/EDGE Radio Access Network).

	GSM / GMSK	EDGE / 8PSK
Symbol Rate	270.833 kbps	270.833 kbps
bits/symbol	1	3
Bit rate/time slot	22.8kbps	68.4kbps

Table 1.1 – GMSK, 8PSK Comparison.

The capacity limitations of second-generation mobile communication systems, together with an increasing demand for supporting multimedia services (not primarily voice) and higher data transmission rates, as well as the demand for a worldwide mobile communication system, triggered the introduction of 3rd generation systems. All different proposals were included under the International Telecommunications Union – Radio (ITU-R) umbrella called International Mobile Telecommunications (IMT-2000), and their harmonization led to the foundation of the 3rd Generation Partnership Project (3GPP) and, finally, to the establishment of the Universal Mobile Telecommunication System (UMTS). One of the most important mobile standards included, was the CDMA2000 standard, which operated in a paired 2x1.25 MHz FDD channel, instead of 2x5 MHz.

This narrower channel band allowed for greater spectrum assignment flexibility and cost efficiency.

UMTS offers greater spectral efficiency than GSM and has 2 modes [6]:

- in FDD mode UpLink (UL) and DownLink (DL) streams are separated in the frequency domain via different frequency bands. This mode is also called Wideband Code Division Multiple Access (W-CDMA) or Direct Spread CDMA. W-CDMA, as an air interface standard, has been designed for always-on packet-based wireless service, so that computers and entertainment devices may all share the same wireless network and connect to the Internet simultaneously;
- in Time Division Duplex (TDD) mode UL and DL streams are separated in the time domain via different time slots. Both modes use direct sequence to separate the different users, i.e., each symbol of one user is multiplied by a user specific spreading code. With this CDMA technique multiple users can transmit in the same (larger) band and the decoder, knowing one user's spreading code, can pick up the data of this specific user. The data of other users appear as noise in this decoding process. Using a wide frequency band makes the system inherently resistant to many drawbacks of narrow band systems such as interference from other transmissions, multipath reflections, etc.;
- W-CDMA supports data rates of up to 2.048 Mbps if the user is stationary, thereby allowing high-quality data, multimedia, streaming, and broadcast services to consumers. With W-CDMA, data rates from as low as 8 kbps to as high as 2 Mbps can be carried simultaneously on a single W-CDMA 5 MHz radio channel, with each channel supporting between 100 and 350 simultaneous voice calls at once, depending on antenna sectoring, propagation conditions, user velocity and antenna polarisation;

- both modes were originally (in the first release: Release 99) defined with a 10 ms radio frame divided into 15 slots, a chip rate of 3,84 Mchips/s and a channel spacing of 5 MHz. A physical channel is defined as a code or number of codes and complementarily as a sequence of time slots in TDD mode;
- UMTS was originally specified for operation in bands in the 2 GHz range. Subsequently, it was extended to operate in a number of other bands, including those originally reserved for 2G services;
- UMTS was continuously enhanced over several releases with new techniques such as 64QAM, and Multiple Input Multiple Output (MIMO), and offered the potential capability for Active Antenna Systems, Indoor Positioning, Self-Optimising Networks, and others.

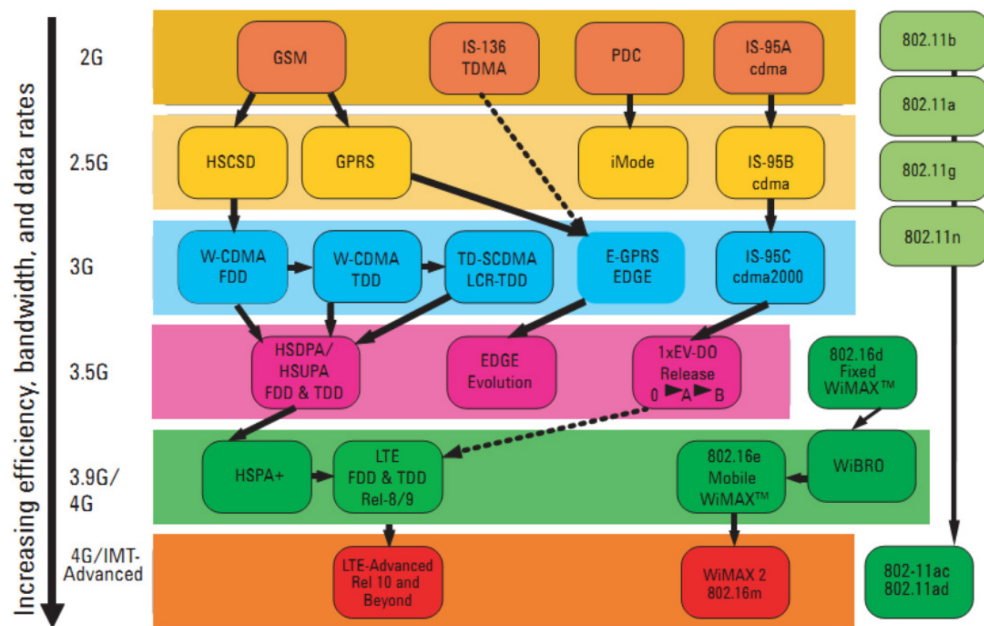


Figure 1.3 – Evolution of Mobile Technologies [1], [7].

Essentially, 3G systems integrated voice and data applications and offered greater end to end security than their predecessors. More precisely, 3G telecommunication networks support services that provide an information transfer rate of at least 384 kbit/s (moving vehicles) up to 2 Mbit/s (stationary or walking). Later 3G releases, often denoted 3.5G and 3.75G, also provide mobile broadband access of several Mbit/s to smartphones and mobile modems in laptop computers. This ensures it can be applied to wireless voice calls, mobile internet access, fixed wireless internet access, video calls and mobile TV technologies. The latest UMTS release, Evolved High Speed Packet Access (HSPA+), can provide, in theory, peak data rates up to 56 Mbit/s in the downlink (28 Mbit/s in existing services) and 22 Mbit/s in the uplink. However, as shown in Figure 1.3, the evolution of mobile telecommunication systems has been rapid during the last three decades and the older generations, such as 2G and 3G, have long been considered obsolete; for this reason, 2G and 3G will be gradually phased out by mobile network operators in order to use the available frequencies more efficiently, specifically for the expansion of 4G and 5G that represent the today standards. An overview of fourth and fifth generation mobile technologies will be provided in the following sections.

1.2 4G technology

The fourth generation of cellular network technology was introduced around 2010, according to the standard IMT Advanced, which consists of an IP-based packet-switched network and uses a broadband technique called Long Term Evolution (LTE), with increased spectral efficiency, offering a significant increase in speed as compared to 3G, up to 100 Mbit/s for high mobility use and 1 Gbit/s for low mobility use (LTE Advanced). A comparison between 3G and 4G is shown in Table 1.2:

3G Technology	4G Technology
maximum upload rate: 5 Mbps.	maximum upload rate: 500 Mbps.
maximum download rate: 21 Mbps.	maximum download rate: 1 Gbps.
It uses a packet switching technique.	It uses the packet switching technique as well as the message switching technique.
frequency range: from 1.8 GHz to 2.5 GHz.	frequency range: from 2 GHz to 8 GHz.
wide area cell-based network architecture.	integration of Wireless LAN as well as Wide Area cell-based network architecture.
turbo codes are used for error correction.	concatenated codes are used for error correction.
Internet Service is broadband.	Internet Service is ultra-broadband.
data bandwidth: 2 Mbps – 21 Mbps.	data bandwidth: 2 Mbps – 1 Gbps.

Table 1.2 – Comparison between 3G and 4G technology.

More precisely, in December 2004 the industry that has developed the 3GPP technologies launched a project (called Long Term Evolution) to study requirements for a new air interface called E-UTRA (Evolved Universal Terrestrial Radio Access). LTE is the upgrade to both GSM/UMTS networks and CDMA2000 networks and it was developed to increase the speed and capacity of mobile communication networks using digital signal processing techniques and modulations. The new characteristics of 4G included [1]:

- significantly increased peak data rates e.g., 100 Mbps in downlink/50 Mbps in uplink;
- increased bitrates at the edge of cells assuming current site locations;
- improved spectrum efficiency and reduced latency;
- scalable bandwidth for a greater flexibility in frequency allocations;
- reduced capital and operational expenditure including backhaul;
- acceptable system and terminal complexity;
- ad hoc and multi-hop networks;
- cost and power consumption;
- support for inter-working with existing 3G systems and non-3GPP specified systems;
- efficient support of the various types of services (e.g., Voice over IP, Presence);
- optimisation for low data rates but supporting high mobile speed;

Some of the features of the **LTE** systems are the following [1]:

- while UMTS started with a focus on circuit-switched data that was then more and more enhanced via shared channels and HSPA into the direction of a packet switched system, LTE is a pure **packet-switched system**;
- the multiplexing technique used is the **Orthogonal Frequency Division Multiplexing (OFDM)** and the access method is OFDMA (Orthogonal Frequency Division Multiplexing Access). In brief, OFDMA works by dividing a wireless channel into smaller sub-carriers and separating those sub-carriers into even smaller resource units. The transmitter then can divide those very small parts of bandwidth among multiple users by bundling separate packets into a single simultaneous transmission. OFDM makes use of the concept of orthogonality of signals, as shown in Figure 1.4, and the properties of Discrete Fourier Transform (DFT) to transmit the signals whose spectrum can overlap with each other avoiding the use of guard band which was needed in traditional frequency division multiplexing technique. Since DFT is a linear combination of orthogonal sinusoids, it essentially correlates its input signal with each sinusoid basis function. So, if the signal has some energy at certain frequency there will be a peak in the correlation of the input signal and the basis sinusoid that is at that corresponding frequency. OFDM with a cyclic prefix is used in the downlink, and Single-Carrier FDMA (SC-FDMA) with a cyclic prefix in the uplink. SC-FDMA is also called DFT-S-OFDM which indicates that it can be understood as a precoding (by Discrete Fourier transform) plus the same OFDMA that is used in downlink;

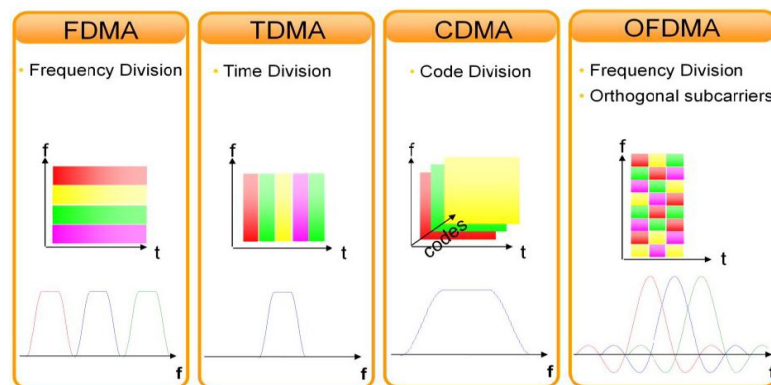


Figure 1.4 – Multiple Access Methods [1].

- to support transmission in paired and unpaired spectrum, two duplex modes are supported: FDD, supporting full duplex and half duplex operation, and TDD. Frequency Division Multiplexing extends the concept of single carrier modulation by using multiple subcarriers within the same single channel. The total data rate is divided between various subcarriers. The data is not needed to be divided evenly nor they have to originate from the same information source. Different modulation schemes can be used for each data, and this is the advantage of FDD over single carrier modulation. The FDD LTE frequency bands are paired to allow simultaneous transmission on two frequencies. The TDD LTE bands are unpaired because the uplink and downlink share the same frequency, being time multiplexed (Figure 1.5);

- while UMTS (at least FDD and 3,84Mbps TDD) used a channel bandwidth of 5MHz, LTE allows for six different channel bandwidths (1,4 / 3 / 5 / 10 / 15 / 20 MHz). UMTS/UTRA as well as LTE/E-UTRA use both a 10 ms radio frame, both have FDD and TDD modes and LTE/E-UTRA supports full interoperability with UMTS/UTRA and GSM/GERAN/EDGE.
- **signals** (logical) are defined as reference signals, primary and secondary synchronization signals, resynchronization signals, wake-up signals, and discovery signals. The modulation schemes supported in the uplink, depending on the type of operation are, $\pi/2$ BPSK, $\pi/4$ QPSK, QPSK, 16QAM, 64QAM and 256QAM, in the downlink, QPSK, 16QAM, 64QAM, 256QAM and 1024QAM, in the sidelink, QPSK, 16QAM and 64QAM;
- **MIMO** was introduced into 4G LTE to improve the signal performance. LTE MIMO adds complexity to the system, but it is able to provide significant improvements in performance and spectral efficiency.

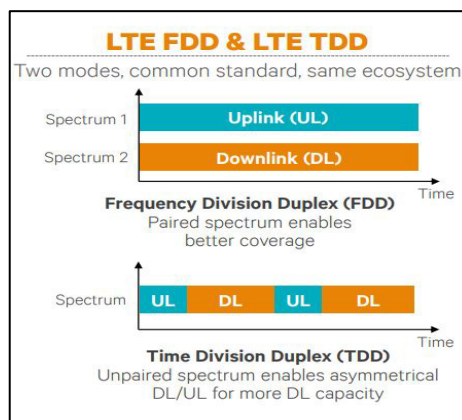


Figure 1.5 – LTE FDD and LTE TDD [8].

Generally, the radio interface is layered into three protocol layers:

- the physical layer;
- the data link layer;
- network layer.

The physical layer performs the following main functions:

- macrodiversity distribution/combining and soft handover execution;
- error detection on transport channels and indication to higher layers;
- FEC encoding/decoding and interleaving/deinterleaving of transport channels;
- multiplexing of transport channels and demultiplexing of coded composite transport channels;
- rate matching;
- mapping of coded composite transport channels on physical channels;
- power weighting and combining of physical channels;
- modulation and spreading/demodulation and despread of physical channels;

- frequency and time (chip, bit, slot, frame) synchronization;
- measurements and indication to higher layers (interference power, transmit power, etc.);
- closed-loop power control;
- RF processing;
- support of timing advance on uplink channels (TDD only);
- support of Uplink Synchronisation (TDD only).

The physical layer in 4G, as shown in Figure 1.6, is defined in a bandwidth agnostic way based on resource blocks (RBs). This allows adaptation to various spectrum allocations, that are articulately described in 3GPP specifications: “A resource block spans either 12 sub-carriers with a sub-carrier bandwidth of 15 kHz or 24 sub-carriers with a sub-carrier bandwidth of 7.5 kHz or 72 sub-carriers with a sub-carrier bandwidth of 2.5 kHz, each over a slot duration of 0.5 ms, or 144 sub-carriers with a sub-carrier bandwidth of 1.25 kHz over a slot duration of 1 ms, or 486 sub-carriers with a sub-carrier bandwidth of approximately 0.37 kHz over a slot duration of 3 ms. Narrowband operation is also defined, whereby certain User Equipment (UE) may operate using a maximum transmission and reception bandwidth of 6 contiguous RBs within the total system bandwidth; for narrowband operation, sub-resource-block operation may also be used in the uplink, using 2, 3 or 6 sub-carriers”.

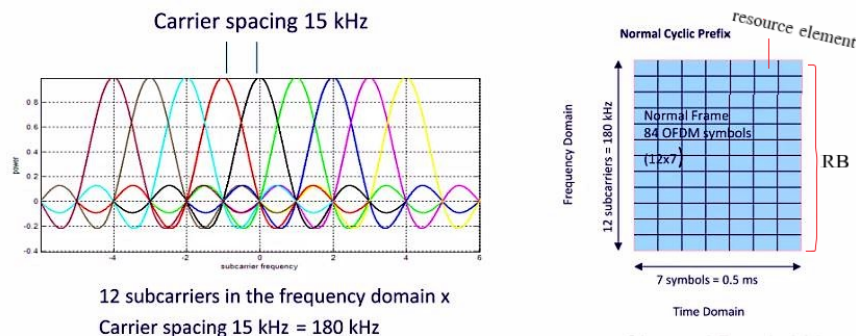


Figure 1.6 – Resource blocks [9].

The proposals for additional spectrum for IMT systems (in 450 MHz band, in Ultra High Frequency (UHF) band (698 ÷ 960 MHz), in the 2.3 ÷ 2.4 GHz band, in C-band (3400 ÷ 4200 MHz)) and the general request for an advanced radio interface triggered further developments of the fourth generation of mobile communication systems. To support a Multimedia Broadcast and Multicast Service (MBMS), LTE offers the possibility to transmit Multicast/Broadcast over a Single Frequency Network (MBSFN), where a time-synchronized common waveform is transmitted from multiple cells for a given duration.

Finally, transmission with MIMO antennas is supported with configurations which allow for multi-layer downlink transmissions and beamforming in both horizontal and vertical dimensions, where a single code word is transmitted over a single spatial layer. Moreover, multi-layer uplink transmissions with up to four streams are supported in the uplink and Multi-user MIMO, i.e., allocation of different spatial streams to different users is supported in both UL and DL.

1.3 5G technology

The fifth generation of cellular network technology represent the latest evolution of communication systems and was introduced late 2018, according to the 5G NR technology standard set by the 3GPP consortium. Like 4G, it is an IP-based technology offering increased bandwidth (download speeds up to 10 Gbit/s), lower latency, ability to handle a larger number of simultaneous connected devices, and improved machine-to-machine support. Moreover, 5G NR is applied for FDD in the 700 MHz band (uplink: 703 MHz ÷ 733 MHz; downlink: 758 MHz ÷ 788 MHz) and for TDD in the 3.5 GHz band (3.3 GHz ÷ 3.8 GHz) and the 26 GHz band (26.5 GHz ÷ 29.5 GHz). Frequencies below 7.125 GHz are collectively called Frequency Range 1 (FR1), whereas above 24.250 GHz are named Frequency Range 2 (FR2).

However, in addition to what has just been said, it is important to clarify that the 5G NR technology standard does not merely ensure an increase in bandwidth or a reduction in latency compared to 4G technology. Indeed, 5G technology has introduced new solutions in terms of standards (frame duration, power allocation, etc.) and a completely new concept for managing radiation distribution (beamforming, power control). More precisely, one of the major innovations introduced by 5G NR technology is the use of Active Antenna System (AAS), instead of Passive Antenna System (PAS) as done in previous technologies, that generates one or more specific beams capable of directing flows of information towards service requests, manages the interference by spatially decoupling beams towards specific users and/or generates nulls of the radiation pattern towards specific directions, optimizes power management and therefore can provide very high performance even on cell border. More specifically, AAS can use two different classes of radiation patterns:

- broadcast radiation pattern(s), that define the coverage area where UEs can attach;
- traffic radiation pattern(s), providing data connection over the coverage area once the UEs has completed the attachment procedure.

In general, broadcast beam performance is different from traffic beam performance. Moreover, unlike PAS, broadcast radiation pattern can change over time. In fact, by using the sweeping broadcast beams approach, the coverage area is divided into sub-areas each one covered by a specific beam and the set of beams is periodically swept to scan the whole coverage area. The radiating system is then capable of knowing the position of users and then can provide service by generating dedicated traffic beams in the specific direction. The ability to generate service beams based on user's distribution makes the traffic coverage time variant, with variable spatial characteristics and time duration based on service requirements. Therefore, radio traffic coverage is no longer static and stable over time, as for technologies prior 5G, but changes with randomness as users' distribution is.

The high performances that 5G services aims to achieve, in addition to the use of AAS, has also required changes to the signal standards in terms of frame structure, physical channel and physical signal. A detailed description of 5G signal standards with respect to frame structure, physical channel and physical signal will be provided in the following section.

1.4 5G signal standard

5G NR specifies a new air interface to enable higher data throughput and low-latency use cases. Key to enabling higher data throughput is the addition of mmWave spectrum up to 52.6 GHz. At these higher frequencies, more contiguous spectrum is available to send more data through the channel. Release 15 of standard by 3GPP specifies a maximum carrier bandwidth up to 400 MHz and up to 16 component carriers that, when aggregated, add up to 800 MHz of bandwidth. Also, flexibility and scalability in the slot structure help support the many new and diverse use cases expected in 5G.

In 5G NR, cyclic prefix orthogonal frequency division multiplex (CP-OFDM) is the modulation format (or waveform) in the DL and UL. CP-OFDM use is well-known for DL transmissions, but it is new for UL transmissions in mobile. Having the same waveform in both UL and DL enables easier device-to-device communication.

Unlike 4G, NR allows for scalable OFDM numerology (μ) where the subcarrier spacings are no longer fixed to 15 kHz. With NR, subcarrier spacing scales by $2^\mu \times 15$ kHz to cover different services. Lower frequency bands use 15, 30, and 60 kHz subcarrier spacings, and higher frequency bands use 60, 120, and 240 kHz subcarrier spacings. Scalable numerology enables scalable slot duration to optimize for different service levels in throughput, latency, or reliability. Larger subcarrier spacing at higher frequencies also helps with the robustness of the waveform since integrated phase noise is an issue in mmWave designs. Figure 1.7 shows how the different subcarrier spacings and the transmission time interval associated with each scales the size of the slot.

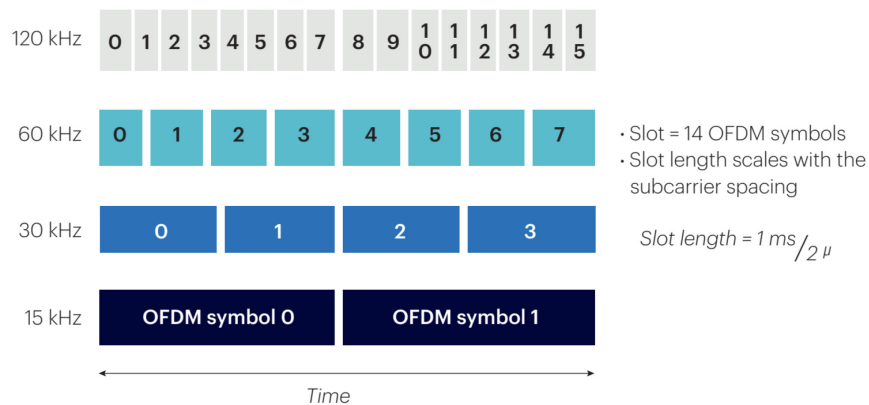


Figure 1.7 – Relationship between subcarrier spacings and time durations [1].

Ultra-Reliable Low Latency Communication (URLLC) is one of three primary 5G use cases and is achieved partially through mini-slots. In LTE, transmissions adhere to the standard slot boundaries, but they are not optimized for minimal latency. A standard slot has 14 OFDM symbols, shown in dark blue in Figure 1.7, and as the subcarrier spacing increases, the slot duration decreases, as shown in light blue. A mini-slot is shorter in duration than a standard slot and located anywhere within the slot. A mini-slot is 2, 4, or 7 OFDM symbols long. Mini-slots provide low-latency payloads with an immediate start time without the need to wait for the start of a slot boundary (Figure 1.8).

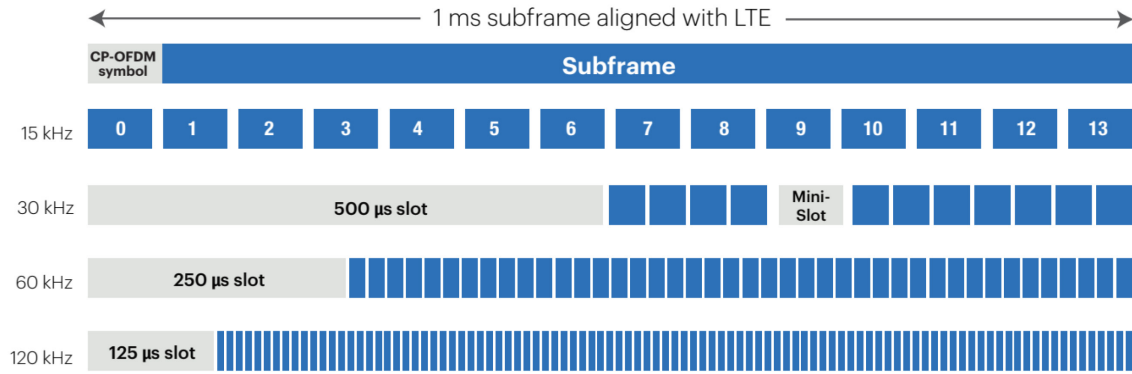


Figure 1.8 – Slots and mini-slots within a subframe and their associated slot duration time [1].

NR subframe structure also allows for dynamic assignments of the OFDM symbol link direction and control within the same subframe. By using this dynamic TDD mechanism, the network dynamically balances UL and DL traffic requirements and includes control and acknowledgment all in the same subframe. The slot format indicator denotes whether a given OFDM symbol in a slot is used for UL or DL, or if it is flexible.

In LTE, carriers are narrower in bandwidth (up to 20 MHz maximum) and when aggregated they create a wider channel bandwidth of up to 100 MHz. In 5G NR, the maximum carrier bandwidth is up to 100 MHz in FR1, and up to 400 MHz in FR2. New in 5G NR are bandwidth parts where the carrier is subdivided for different purposes. Each bandwidth part has its own numerology and is signalled independently. One carrier can have mixed numerologies to support a mixed level of services, such as power saving or multiplexing of numerologies and services in unlicensed bands. However, only one bandwidth part in the UL and one in the DL are active at a given time. Bandwidth parts will support legacy 4G devices with new 5G devices on the same carrier. With 4G, 5G, and potentially Wi-Fi multiplexing services, it is necessary to minimize both in-band and out-of-band emissions.

1.4.1 5G frame structure

The frame structure has been conceived to obtain high energy efficiency, and NR avoids as much as possible “always on air” signals. Consequently, the fundamental pieces of information that are required by the UE for initial connection to the cell are “packed” in a very compact structure, called the Synchronization Signal / Physical Broadcast Channel (SS/PBCH), or “SS Block” (SSB), that includes the Synchronization Signals (SS), the PBCH and the Physical Broadcast Channel Demodulation Reference Signal (PBCH-DMRS). SSB is mapped into 4 OFDM symbols in the time domain and 240 contiguous subcarriers (20 RBs) in the frequency domain (Figure 1.9):

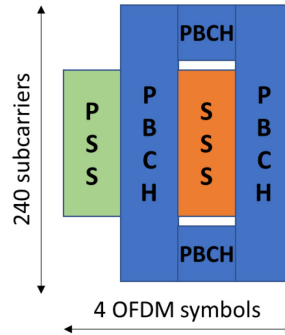


Figure 1.9 – Synchronization Signal Block (SSB) in 5G NR [1].

SSBs are grouped in block patterns called SS bursts. There are 5 block patterns which have different subcarrier spacings and are applicable for different carrier frequencies:

- Case A (15 kHz subcarrier spacing);
- Case B (30 kHz subcarrier spacing);
- Case C (30 kHz subcarrier spacing);
- Case D (120 kHz subcarrier spacing);
- Case E (240 kHz subcarrier spacing).

The maximum number of SSBs in a single burst is frequency dependent, and ranges from 4 to 8 in FR1 to up to 64 blocks per burst in FR2. Each SSB in a SS Burst is associated to a different beam of the grid of broadcast beams. The other signals are transmitted when required. They are usually transmitted on traffic beams and are “user-dependent”. 5G uses a sophisticated signalling structure. In the following, the discussion will be limited on the physical channels and signals associated to downlink connection. The choice of the beam (broadcast or traffic) on which the signals are transmitted is generally left to the vendor decision. In the following the most common choices are indicated.

1.4.2 Physical channels

The Physical Channels defined in downlink are (Figure 1.10):

- the Physical Broadcast Channel (PBCH);
- the Physical Downlink Shared Channel (PDSCH);
- the Physical Downlink Control Channel (PDCCH).

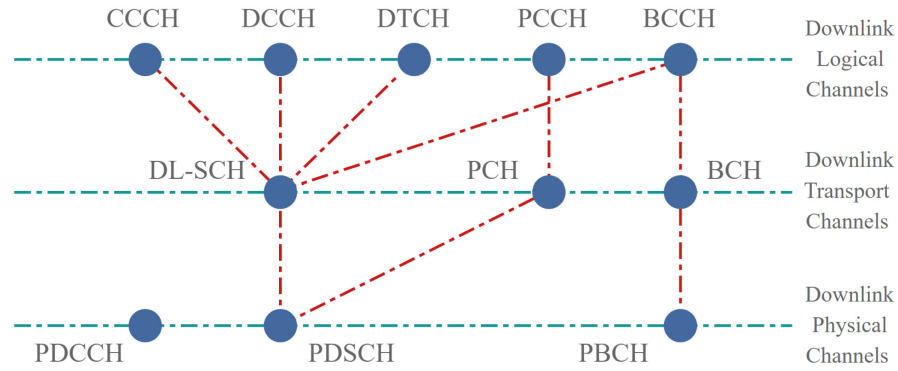


Figure 1.10 – 5G NR downlink channel mapping.

Each Physical Channel has its own Demodulation Reference Signal (DMRS) for channel estimation and equalization, as discussed in the following.

Physical Broadcast Channel (PBCH): PBCH carries basic system information required to access the network, and in particular the Master Information Block (MIB). The PBCH is transmitted using QPSK in the SS-PBCH blocks with its own demodulation reference signal using broadcast beams. NR supports only one antenna port-based transmission for PBCH data in order to avoid blind decoding.

Physical Downlink Control Channel (PDCCH): The PDCCH provides scheduling decisions necessary for the reception of PDSCH, as well as special purposes such as slot format indication and power control. The information carried by the PDCCH is referred to as Downlink Control Information (DCI). PDCCH is transmitted in a CORESET (COnTrol REsource SET) using QPSK modulation. A CORESET is a set of RBs which has varying time and frequency domain length. In time domain it may occupy 1, 2 or 3 OFDM symbols whereas in frequency domain it is a multiple of one RB (12 subcarriers). A CORESET defines a resource where UE specific scheduling information can occur. PDCCH has its own PDCCH-DMRS signal and is transmitted on traffic beams.

Physical Downlink Shared Channel (PDSCH): This is the main channel used for data transmission, paging information, some parts of system information and also random-access response message. It decodes DCI from PDCCH which provides the necessary information to decode data in PDSCH. The frequency resource allocation is organized in RBs, with flexible modulation schemes selected on the basis of the Signal to Noise Ratio (SNR). PDSCH has its own PDSCH-DMRS signal and is transmitted on traffic beams.

1.4.3 Physical signals

Physical channels carry information originating from the higher layers. In addition, 5G NR defines additional physical signals that do not carry information from higher layers, but are used for functionalities like synchronization, channel estimation, tracking and beam identification. The physical signals defined in downlink are:

- the Primary Synchronization Signal (PSS) and Secondary Synchronization Signal (SSS);
- the Physical Broadcast Channel Demodulation Reference Signal (PBCH-DMRS);
- the Physical Downlink Control Channel Demodulation Reference Signal (PDCCH-DMRS);

- the Physical Downlink Shared Channel Demodulation Reference Signal (PDSCH-DMRS);
- Phase Tracking Reference Signal (PTRS);
- the Channel-state Information reference signal (CSI-RS).

Primary Synchronization Signal and Secondary Synchronization Signal (PSS, SSS): PSS and SSS play a key role in synchronization and cell search. PSS helps to achieve subframe, slot and symbol synchronization in the time domain, identify the centre of the channel bandwidth in the frequency domain and deduce a pointer to 1 of 3 Physical layer Cell Identities (PCI). SSS helps to achieve radio frame synchronization and deduce a pointer to 1 of 168 Physical layer Cell Identity (PCI) groups. PSS and SSS are transmitted in the SS-PBCH block using BPSK constellation on broadcast beams with constant power, so they are suitable as reference signals. SSS has been proposed as reference signal in [10].

Physical Broadcast Channel-Demodulation Reference Signal (PBCH-DMRS): PBCH-DMRS is a special type of physical layer signal used as reference signal for decoding PBCH. PBCH-DMRS is transmitted in the SS-PBCH block using broadcast beams with constant power and is transmitted also in absence of users. It is one of the signals proposed as reference in MPE techniques [11].

Physical Downlink Control Channel-Demodulation Reference Signal (PDCCH-DMRS): PDCCH-DMRS is a special type of physical layer signal used as reference signal for decoding PDCCH. It is transmitted together with the PDCCH in a CORESET using traffic beams. NR considers the possibility of a power boosting factor for the PDCCH-DMRS. In particular 3GPP defines the $\beta_{\text{PDCCH DMRS}}$ factor as the ratio between PDCCH Energy per Resource Element (EPRE) and PDCCH-DMRS EPRE.

Physical Downlink Shared Channel-Demodulation Reference Signal (PDSCH-DMRS): PDSCH-DMRS has a 'double role' in PDSCH decoding. In fact, NR assumes that precoding is used for data. PDSCH-DMRS is transmitted on traffic beams using the same precoding of the data, allowing to estimate both the response of the propagation channel and the precoding. The position of the PDSCH-DMRS is characterized by a high degree of flexibility in order to match the characteristics of the communication channel and of the user application. In particular, PDSCH-DMRS is mapped in PDSCH RB according to Mapping Type A or B and Configuration Type 1 or 2. The Mapping Type A or B defines the mapping location in the slot. In Type A the DMRS allocation is in the symbol 2 or 3, while in Type B the DMRS is allocated in the first symbol of PDSCH allocation and is used when the PDSCH covers only a part of the slot, allowing a fast demodulation of the data required for low latency applications. The DMRS Configuration Type 1 or 2 specifies the density of the DMRS. Type 1 is denser in frequency, while Type 2 supports a larger number of orthogonal DMRS sequences, required to handle a large number of MIMO layers as in MU-MIMO.

The minimum number of DMRS symbols per slot is 1, but it is possible to add up to 3 additional DMRS symbols per slots to handle critical synchronization scenarios as high-speed applications. In case of multiple layers transmission, DMRS uses different locations in frequency domain to allocate the ports. Two orthogonal pseudo-noise sequences are used to map two antenna ports on the same frequency, allowing to half the number of frequencies. Each symbol can be associated to two different codes, and DMRS can be allocated in a single symbol or in a double symbols' configuration. Consequently, Type 1 can support only up to 4 orthogonal signals for single symbol DMRS, or 8

orthogonal symbols for double symbols, while Type 2 supports up to 6 orthogonal signals for each DMRS symbol, or 16 in double symbols configuration.

Finally, NR considers the possibility of a power boosting factor associated to the EPRE. In particular 3GPP defines the β_{DMRS} factor as the ratio between PDSCH EPRE and DMRS EPRE.

Phase Tracking Reference Signals (PTRS): PTRS can be optionally used in PDSCH channel for compensating the phase noise and is particularly important in FR2, where phase noise is an important problem. It is sparser in frequency and denser in time as compared to DMRS. 5G introduces a power boosting factor associated to the EPRE. In particular 3GPP defines the β_{PTRS} factor as the ratio between PTRS EPRE and PDSCH EPRE.

Channel State Information-Reference Signal (CSI-RS): CSI-RS is used by the device to acquire information about the downlink channel. It is decoded by the device and sent as feedback to the base station as part of reports via PUSCH and PUCCH uplink channels. In particular, CSI reports consist of three parameters: RI (Rank Indicator), PMI (Precoding Matrix Index) and CQI (Channel Quality Information). The Rank Indicator indicates how many independent layers can be transmitted by the base station. The Precoding Matrix Indicator indicates what precoding matrix should be used by the transmitter for beam-forming on the base of a codebook of precoded matrices. Finally, CQI gives information on the quality of the communication channel. Based on the report, the base station optimizes various parameters like the number of transmission layers, modulation and coding scheme, and the pre-coding matrix to be used. The CSI report only provides suggestions, and the transmitter is free to decide whether to follow the recommendation included in the report. The CSI-RS resources can be allocated in three ways, namely in a periodic, semi-persistent or aperiodic basis. Semi-persistent is equivalent to periodic configuration, the only difference being that the activation and deactivation of CSI-RS transmission are controlled by MAC. Not all kinds of combinations among periodicity types for measurement and reporting are allowed. CSI-RS can have a frequency density equal to 1 in which case it will be transmitted in every RB for the full bandwidth, or density of 1/2 in which case it will be transmitted in alternate RBs. In a multi-port, the separation is achieved by modulating with different orthogonal codes, by separating in frequency, or by separating in time. The CSI-RS can be Non-Zero Power (NZP) or Zero Power (ZP, or Muted). The NZP REs are associated to a no-transmission condition. This allows to use the frequency-time resources associated to a ZP CSI-RS to estimate the interference level due for example to the presence of other cells active on the same bandwidth.

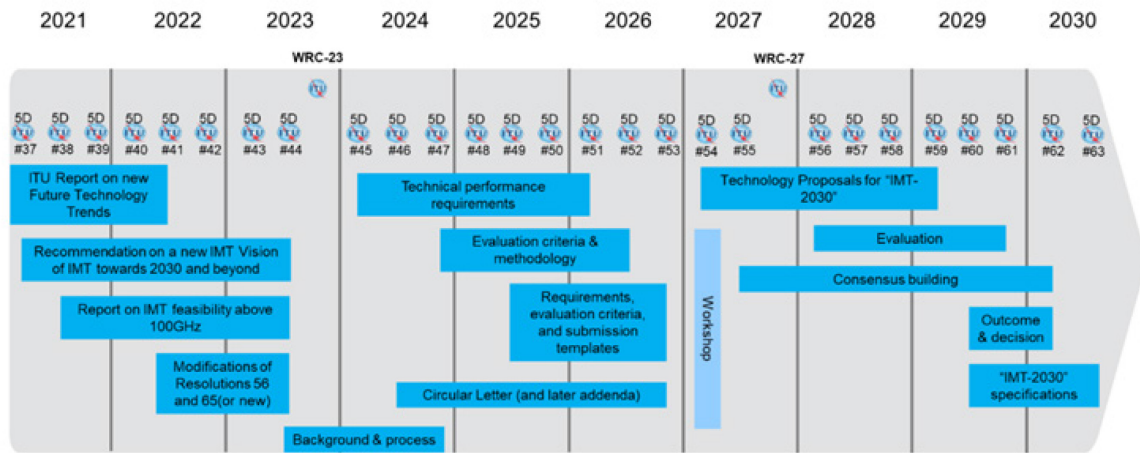
Finally, there is also a third kind of CSI-RS with 3 subcarriers for each RB. This kind of CSI-RS is very dense in frequency and is used for the Tracking Reference Signal (TRS). On the contrary of the other CSI-RS transmissions, the TRS is single port. In practice, the high flexibility of CSI-RS allows to configure it for multiple purposes, as TRS, CSI acquisition, and beam management.

1.5 6G technology

Although the rollout of 5G technologies has not yet been fully completed — especially with regard to technologies operating in FR2, which at present have virtually no deployment — research activities have already started on the sixth generation of mobile communications technologies aiming at a fully standardized set of specifications around year 2030.

Obviously, in line with what has been observed in past transitions, the shift from 5G networks to 6G ones should be accompanied by significant technical advancements. More precisely, it is expected that 6G networks will enable data rates on the order of 100 Gbit/s (approximately ten times the theoretical maximum speed achievable with 5G technology), with latency values below 100 μ s. In addition, 6G will not simply be faster than 5G; in fact, even though it is still too early to clearly identify the complete set of new technologies for 6G, the ambition is to transform the network into an intelligent infrastructure, capable of integrating artificial intelligence, satellites, and terrestrial systems into a single digital architecture.

In terms of frequency band allocations for 6G, with respect to some preliminary assessments carried out in accordance with the ITU-R IMT-2030 (International Telecommunication Union-Radiocommunication sector for International Mobile Telecommunications for 2030) process (Figure 1.11), the 7 GHz ÷ 24 GHz frequency range 3 (FR3) band is viewed as the most promising new spectrum [12]. More specifically, the 2023 World Radiocommunication Conference (WRC-23) designated the 6 GHz band in all ITU regions and set the agenda for the WRC-27. The new WRC cycle will prioritize the following frequency bands for IMT: 4400 MHz ÷ 4800 MHz, 7125 MHz ÷ 8400 MHz, and 14.8 GHz ÷ 15.35 GHz [12].



Note 1: Meeting 5D#59 will additionally organize a workshop involving the Proponents and registered IEGs to support the evaluation process
 Note 2: While not expected to change, details may be adjusted if warranted. Content of deliverables to be defined by responsible WP 5D groups

Note by the ITU-R Radiocommunication Bureau: This document is taken from Attachment 2.12 to Chapter 2 of Document 5D/1361 (Meeting report WP 5D #41, June 2022) and adjustments could be made in the future. ITU holds copyright in the information – when used, reference to the source shall be done.

Figure 1.11 – Overview timeline for IMT towards the year 2030 and beyond [15].

Obviously, considering that 6G will need to comply with, among others, fundamental requirements on sustainability and human-centricity, the EMF exposure topic is a key element to be considered in developing the next generation of mobile communication system. In that context, it is envisaged that the results obtained in this research project could be really promising in addressing EMF exposure assessment issues. In fact, the primary objective of this work is to provide a proof of concept for the development—using low-cost components—of a wearable device capable of assessing exposure levels to forthcoming 6G signals operating in the FR3 band. Based on these considerations, in the following chapters will therefore be described the theoretical aspects required to meet the primary objectives of this work, as well as the technical solutions adopted to design a wearable device capable of assessing exposure levels to EMF signals operating in the 14.8 GHz ÷ 15.35 GHz frequency band.

2 Chapter 2: Basics of EMF measurements

The distinctive characteristics of the EMF signals employed in the various cellular communication technologies that have emerged over the past 30 years were analyzed in the previous chapter, with particular attention given to the description of the standards and signal characteristics used in the latest-generation 5G communication networks. In this chapter, instead, the main theoretical aspects related to EMF field measurements will be described.

In fact, as already mentioned in Chapter 1, the primary objective of this research project is to provide a proof of concept regarding the design and the development, through the use of low-cost components, of a wearable device capable of assessing exposure levels to upcoming 6G signals operating in the FR3 frequency band. Clearly, in order to achieve the intended outcome and, in particular, to properly design the device of interest, the first step consists of conducting a careful review of the relevant literature in order to analyze the distinctive characteristics and the operating principles of the devices that are typically used to perform such measurements. This type of analysis is, in fact, useful for two main reasons: on the one hand, it allows the identification of interesting insights that may contribute to achieving the primary objectives of the present work; on the other hand, it provides the theoretical knowledge necessary for the correct use of the equipment normally employed to carry out the measurements of interest.

The ability to properly use suitable instrumentation for performing measurements aimed at characterizing exposure levels to EMF signals will represent a crucial aspect for verifying the correct functionality of the device developed within the present work. Indeed, in order to validate the results obtained through the use of the proposed device it will be necessary to compare them with those obtained using appropriately calibrated commercial instrumentation, once the 6G network will be effectively deployed. Moreover, in order to ensure the achievement of the primary objectives of the present work, it is useful to analyze the guidelines developed by the International Commission on Non-Ionizing Radiation Protection (ICNIRP) committee concerning human exposure to non-ionizing electromagnetic radiation. In fact, as will be discussed in greater detail in the following section, the ICNIRP recommendations define the procedures according to which measurements must be carried out in order to assess individuals' exposure levels to the EMF signals of interest within the scope of this work. Furthermore, the aforementioned recommendations constitute the basis of the regulatory framework adopted, both at the National and at the European level, for the definition of the limit values for the physical quantities characterizing human exposure to EMFs at frequencies compatible with those of the signals considered in this research project. Therefore, as will be specified more clearly in the following sections, this analysis is of fundamental importance for the proper design of the device that is intended to be developed.

Accordingly, because of the considerations introduced above, in order to allow the reader to fully understand the reasons behind the design choices that have been made (that will be discussed in the Chapter 4), this chapter will be structured as follows:

- The first two sections will focus on the description of the ICNIRP guidelines and the main regulatory references adopted, both at the National and European levels, concerning the protection of individuals from exposure to non-ionizing EMF radiation. This analysis, will make it possible to define the minimum field level that the device of interest must be able to detect and, consequently, as will be discussed in more detail in Chapter 4, it is essential for the proper design of the field sensors to be used in order to achieve the primary objectives of the present work, as well as for the selection of the different electronic components to be adopted for this purpose;
- subsequently, the main types of devices that can be used to perform EMF level measurements will be described. Thanks to this analysis, in fact, it will be possible to identify, providing adequate justification, the device type that has been chosen to meet the primary objectives of this research project.

2.1 ICNIRP exposure guidelines

As is well known, the different interaction mechanisms that may occur between EM radiations and material systems — such as polarization phenomena of atoms and/or molecules, induction of currents, and conversion of the kinetic energy of ions and particles into heat — essentially arise from energy transfer processes established between the physical agent (the EM field) and the material system with which it interacts. Clearly, the energy transfer phenomena just mentioned also occur when EM waves interact with biological systems, such as the human body. However, the effects produced by the inevitable interactions between EM radiations and biological systems strongly depend on the amount of energy that is effectively transferred from the EM field to the biological system itself. In other words, exposure of individuals to EM radiations does not necessarily lead to the occurrence of harmful health effects; for such effects to arise, in fact, the amount of energy absorbed as a result of the interaction with electromagnetic radiation must be sufficient to induce significant alterations in the individuals' physiological condition. From this perspective, the first parameter to be considered when assessing the potential effects induced in individuals by exposure to EMFs is the EM waves' frequency. Indeed, as is well known, according to the quantum interpretation of electromagnetic fields, the energy carried by the photons constituting the radiation, although quantized, is directly proportional to the frequency in accordance with the well-known *Planck–Einstein relation* [16]:

$$E = hf \quad (2.1)$$

where f is the photons' frequency (measured in Hz) and $h = 6.626 \times 10^{-34} \text{ J} \cdot \text{Hz}^{-1}$ is a fundamental physical constant known as *Planck's constant*.

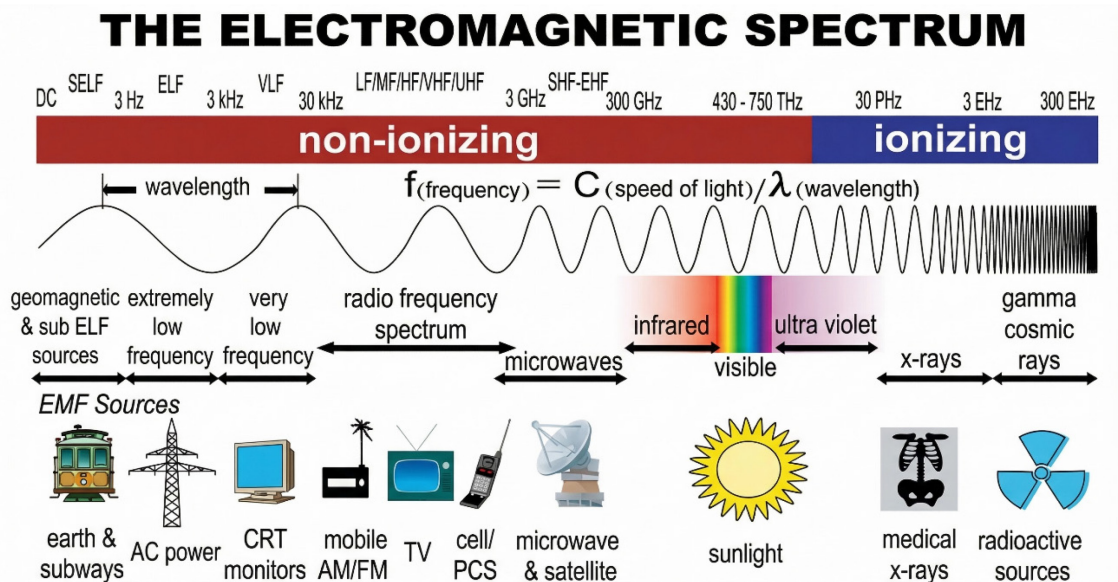


Figure 2.1 – Non-Ionizing and Ionizing radiation spectrum [17].

More precisely, with reference to the electromagnetic spectrum (Figure 2.1), depending on whether or not the waves possess sufficient energy to ionize the atoms of the matter with which they interact, two different categories of electromagnetic radiations can be identified:

- **Non-Ionizing Radiations (NIR)**, which includes radio waves, microwaves, infrared radiation, visible light, and part of the ultraviolet spectrum;
- **Ionizing Radiations (IR)**, comprising the higher energy ultraviolet part of the electromagnetic spectrum, gamma rays and X-rays.

In the following discussion, attention will be limited to the description of the recommendations given to prevent human exposure to NIR from causing harmful health effects. Indeed, as discussed in Chapter 1, the frequencies of the EMF signals considered within the scope of this research project are well below the frequency values required to induce ionization phenomena, which generally require energies greater than several tens of electronvolts (eV), corresponding to frequencies on the order of several hundreds of THz [18].

As previously stated, the guidelines concerning protection from human exposure to NIR are developed by a non-governmental organization, formally recognized by the World Health Organization (WHO), known as ICNIRP. More precisely, ICNIRP identifies, through a careful and periodic review of the relevant scientific literature, the harmful and scientifically substantiated effects resulting from exposure to RF-EMF; for each of these effects, the corresponding threshold exposure levels (i.e., the minimum exposure level capable of causing them) are then determined, and starting from these values, by adopting a conservative approach - i.e., by defining **reduction factors** that take into account the biological variability of individuals (for example, relating to age and sex), variations in basic conditions (such as, for example, tissues temperature), the dosimetric uncertainty associated with the determination of exposure values, and variations in environmental factors (for example, relating to humidity and temperature) - the so-called **basic restrictions** are defined, referring to physical quantities directly related to the adverse effects induced by EM radiation.

The following section provides a detailed description of the adverse effects resulting from RF-EMF exposure and of the corresponding threshold exposure levels used by ICNIRP to establish the basic restrictions for EMF exposure in the frequency range of interest for this research project.

2.1.1 Basic restrictions

Figures 2.2 and 2.3 show the basic restrictions for EMF exposure from 100 kHz to 300 GHz defined in the latest edition of the guidelines provided by ICNIRP in 2020 [19].

Exposure scenario	Frequency range	Whole-body average SAR (W kg ⁻¹)	Local Head/Torso SAR (W kg ⁻¹)	Local Limb SAR (W kg ⁻¹)	Local S _{ab} (W m ⁻²)
Occupational	100 kHz to 6 GHz	0.4	10	20	NA
	>6 to 300 GHz	0.4	NA	NA	100
General public	100 kHz to 6 GHz	0.08	2	4	NA
	>6 to 300 GHz	0.08	NA	NA	20

^aNote:

1. “NA” signifies “not applicable” and does not need to be taken into account when determining compliance.
2. Whole-body average SAR is to be averaged over 30 min.
3. Local SAR and S_{ab} exposures are to be averaged over 6 min.
4. Local SAR is to be averaged over a 10-g cubic mass.
5. Local S_{ab} is to be averaged over a square 4-cm² surface area of the body. Above 30 GHz, an additional constraint is imposed, such that exposure averaged over a square 1-cm² surface area of the body is restricted to two times that of the 4-cm² restriction.

Figure 2.2 – Basic restrictions for EMF exposure from 100 kHz to 300 GHz, for averaging intervals ≥6 min [19]

Exposure scenario	Frequency range	Induced electric field; E _{ind} (V m ⁻¹)
Occupational	100 kHz to 10 MHz	$2.70 \times 10^{-4}f$
General public	100 kHz to 10 MHz	$1.35 \times 10^{-4}f$

^aNote:

1. f is frequency in Hz.
2. Restriction values relate to any region of the body, and are to be averaged as root mean square (rms) values over 2 mm × 2 mm × 2 mm contiguous tissue (as specified in ICNIRP 2010).

Figure 2.3 – Basic restrictions for EMF exposure from 100 kHz to 10 MHz, for peak spatial values [19].

It is important to note that the guidelines differentiate between occupationally-exposed individuals and members of the general public. Occupationally-exposed individuals are defined as “adults who are exposed under controlled conditions associated with their occupational duties, trained to be aware of potential radiofrequency EMF risks and to employ appropriate harm-mitigation measures, and who have the sensory and behavioral capacity for such awareness and harm-mitigation response” [19]. The general public is defined as “individuals of all ages and of differing health statuses, which includes more vulnerable groups or individuals, and who may have no knowledge of or control over their exposure to EMFs” [19]. These differences suggest the need to include more stringent restrictions for the general public, as members of the general public would not be suitably trained to mitigate harm, or may not have the capacity to do so. This is exactly what happens in the ICNIRP guidelines; in fact, as shown in Figures 2.2 and 2.3, by applying different reduction factors on the threshold exposure levels for occupationally-exposed individuals and members of the general public, the generic public's basic restrictions are lower than the ones defined for the occupational exposed individuals.

Furthermore, as shown in Figure 2.2, the physical quantities on which the basic restrictions are defined change depending on the frequency range examined. The reason for this is that as the

frequency varies, the interaction mechanisms between EM radiation and the biological systems exposed to it vary. It is well known, indeed, that when a biological body is exposed to RF-EMFs, some of the EMF's power is reflected away from the body, and some is absorbed by it; this results in complex patterns of electromagnetic fields inside the body that are heavily dependent on the EMF characteristics as well as the physical properties and dimensions of the body itself. More precisely, the *penetration depth* of the field inside the body is given by the following equation [22]:

$$\delta = \frac{c}{2\pi f \sqrt{\frac{\epsilon_r}{2} \left(\sqrt{1 + \left(\frac{\sigma}{2\pi f \epsilon_0 \epsilon_r} \right)^2} - 1 \right)}} \quad (2.2)$$

where $c \cong 3 \times 10^8 \text{ m} \cdot \text{s}^{-1}$ is the speed of light, f is the frequency (Hz), $\epsilon_0 = 8.854 \times 10^{-12} \text{ F} \cdot \text{m}^{-1}$ is the vacuum permittivity and ϵ_r and σ are the relative permittivity and the conductivity ($\text{S} \cdot \text{m}^{-1}$) of the biological tissues through which the absorbed field propagates, respectively.

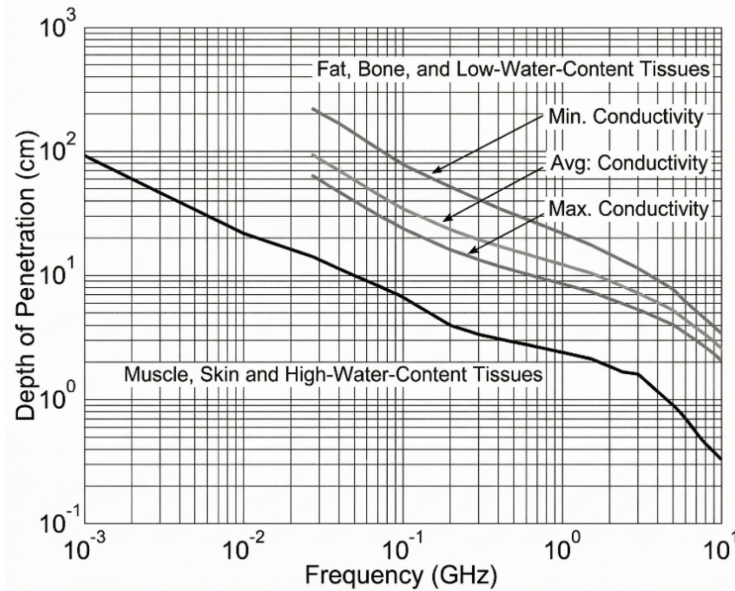


Figure 2.4 – Penetration depth in biological tissues [23]. The curves for the “fat, bone and tissues with low water content” correspond to the minimum, average and maximum values of the conductivity ranges listed in [24].

As shown by Equation 2.2 and schematically illustrated in Figure 2.4, for any given values of ϵ_r and σ , the penetration depth of the EMF inside the body increases as the frequency decreases. In particular, at sufficiently low frequencies, e.g. for frequencies between 100 kHz and 10 MHz, the penetration depth exceeds 10 cm (Figure 2.4) and the Electric Field inside the body can exert electrical forces that are sufficient to stimulate nerves; moreover, if it is strong and brief enough (as can be the case for pulsed low frequency EMFs), it can exert electrical forces that are sufficient to cause dielectric breakdown of biological membranes [25]. This explain why the ICNIRP basic restrictions for EMF exposure from 100 kHz to 10 MHz are given in terms of the amplitude of the Electric Field inside the body (referred to as induced Electric Field E_{ind} in Figure 2.3).

In addition to the possible nerve stimulation and the potential biological membranes' dielectric breakdown, the induced Electric Field in the body exerts a force on both polar molecules (mainly water molecules) and free moving charged particles such as electrons and ions. In both cases a portion of the EMF energy is converted to kinetic energy, forcing the polar molecules to rotate and charged particles to move as a current. As the polar molecules rotate and charged particles move, they typically interact with other polar molecules and charged particles, causing the kinetic energy to be converted to heat that can adversely affect health [19]. Indeed, if the body core temperature rises of more than 1 °C – 2 °C adverse health effects such as heat exhaustion and/or heat stroke could happen [26]. In addition to body core temperature rises, excessive localized heating (i.e., tissue temperatures greater than 41 °C – 43 °C) can cause pain and thermal damage with damage likelihood and severity increasing as a function of time [28][29], [30]. For these reasons, from a health risk perspective, it is important to quantify the amount of EMF energy absorbed by biological tissues, since this is largely responsible for the thermal effects just mentioned.

To do what has just been said, it is possible to refer to the **Specific energy Absorption Rate (SAR)** that is defined as the power absorbed per mass of tissue [W·kg⁻¹]:

$$SAR = \frac{1}{V} \int_{sample} \frac{\sigma(r)|\mathbf{E}_{ind}(r)|^2}{\rho(r)} dr \quad (2.3)$$

where σ is the sample electric conductivity [S·m⁻¹], ρ is the mass density of the sample [kg·m⁻³], and V is the volume of the considered sample [m³] respectively.

Indeed, according to the definition given above (equation 2.3), the SAR is a measure of the rate at which EMF energy is absorbed per unit mass by a human body exposed to RF-EMF and, therefore, it can be used to predict the exposure conditions required to induce a given temperature in it. For example, recent theoretical modeling and generalization from experimental research across a range of species predicts that exposures resulting in a whole-body average SAR of approximately 6 W·kg⁻¹, within the 100 kHz to 6 GHz range, over at least a 1-hour interval under thermoneutral conditions is required to induce a 1 °C body core temperature rise in human adults [19].

Based on the above considerations, the ICNIRP has thus adopted a conservative temperature rise value of 1 °C as the operational adverse health effect threshold and, always for conservative purposes, within the frequency range 100 kHz ÷ 300 GHz, uses 4 W kg⁻¹ averaged over 30 min (to take into account the time it takes to reach a steady-state temperature), instead of 6 W·kg⁻¹, as the radiofrequency EMF exposure level corresponding to a body core temperature rise of 1 °C; by applying on the conservative value of 4 W kg⁻¹ a reduction factor equal to 10 for the occupationally-exposed individuals and equal to 50 for members of the general public, within the 100 kHz to 300 GHz range the basic restrictions for the whole-body SAR averaged over 30 minutes become equal to 0.4 W kg⁻¹ and 0.08 W kg⁻¹ for the occupational and the general public exposure scenario, respectively (Figure 2.2).

However, from a health risk perspective, in addition to body core temperature rise, excessive localized heating must also be avoided. With respect to the possible harmful effects on health resulting from localized temperature rise, it is important to point out that the ICNIRP treat as potentially harmful the RF-EMF exposure that results in local temperatures of 41 °C or greater [19].

Based on this assumption, since body temperature varies as a function of body regions, in order to avoid possible harmful effects on health resulting from localized temperature rise induced by RF-EMF exposure, it would be necessary to identify operational thresholds for adverse health effects for all the different kind of body tissues and set exposure restrictions for each of them. Clearly, such an approach would be extremely complex and is not feasible in practice. Therefore, in order to identify operational thresholds for local heat-induced health effects, ICNIRP has firstly classified tissues of major body regions into two different categories, based on their temperature under normal conditions [19]:

- “Type-1” tissue, including all tissues in the upper arm, forearm, hand, thigh, leg, foot, pinna and the cornea, anterior chamber and iris of the eye, epidermal, dermal, fat, muscle, and bone tissue, for which the normothermal temperature is typically lower than 36 °C;
- “Type-2” tissue, including all tissues in the head, eye, abdomen, back, thorax, and pelvis, except those defined as Type-1 tissue, for which the normothermal temperature is typically lower than 38.5 °C.

Then, by adopting a value of 41 °C, for tissue temperature, as potentially harmful, ICNIRP decided to take a conservative approach and consider as operational thresholds for adverse health effects for local exposure RF-EMF induced temperature rises of 5 °C and 2 °C in Type-1 and Type-2 tissues, respectively. Finally, due to the difficulties of setting exposure restrictions as a function of the above tissue-type classification, as shown in Figure 2.2, ICNIRP opted to define two different “regions” and to sets separate exposure restrictions, where relevant, for these regions [19]:

- **Head and Torso**, comprising the head, eye, pinna, abdomen, back, thorax and pelvis, which includes both Type-1 and Type-2 tissue;
- **Limbs**, comprising the upper arm, forearm, hand, thigh, leg and foot, which only includes Type-1 tissue.

Exposure levels have thus been determined for each of these regions such that they do not result in temperature rises of more than 5 °C and 2 °C, in Type-1 and Type-2 tissue, respectively. More specifically, within the 100 kHz to 6 GHz range, the guidelines take a SAR of 20 W kg⁻¹, averaged over a 10 g cubic mass and 6-min interval, as the local exposure level corresponding to the operational adverse health effect threshold for the Head and Torso (5 °C in Type-1 tissue and 2 °C in Type-2 tissue), and a SAR of 40 W·kg⁻¹ averaged over a 10-g cubic mass and 6-min interval, as the local exposure level corresponding to the operational adverse health effect threshold for the Limbs of a 5°C rise in local temperature. A 10 g mass is used because, although there can initially be EMF-induced temperature heterogeneity within that mass, heat diffusion rapidly distributes the thermal energy to a much larger volume that is well-represented by a 10-g cubic mass [31]. Moreover, similarly to what has been discussed for whole-body exposure, in order to account for scientific uncertainty and differences in thermal physiology across the population, and in order to include more stringent restrictions for the general public (as members of the general public would not be suitably trained to mitigate harm), reduction factors of 2 and 10 are applied to the aforementioned values of 20 W kg⁻¹ and 40 W·kg⁻¹ to obtain the basic restrictions for local exposure of Head and Torso and Limbs for occupational and general public exposure, respectively (Figure 2.2). It is useful to note that reduction factors for local exposure are smaller than ones for whole-body exposure; this is mainly because the associated health effect is less serious medically [19].

Moreover, as shown in Figure 2.2, within the frequency range from 6 GHz to 300 GHz the basic restrictions for local exposure are no longer expressed in terms of SAR, but are instead provided in terms of another physical quantity denoted as S_{ab} . This happens because within the 6 GHz to 300 GHz range, as EMF frequency increases, exposure of the body and the resultant heating becomes more superficial, and above about 6 GHz this heating occurs predominantly within the skin; therefore, the SAR averaged over a 10 g cubic mass, which includes deeper tissues, become less relevant to this frequency range. Conversely, the **absorbed power density** (S_{ab}) provides a measure of the power absorbed in tissue that closely approximates the superficial temperature rise [32]. More precisely, S_{ab} is defined in terms of the Poynting vector ($\mathbf{S}$) by means of the following equation [19]:

$$S_{ab} = \frac{1}{A} \iint_A \text{Re}(\mathbf{S}) \cdot d\mathbf{s} = \frac{1}{2A} \iint_A \text{Re} [\mathbf{E} \times \mathbf{H}^*] \cdot d\mathbf{s} \quad (2.4)$$

where $\text{Re}(X)$ and X^* are the real part and complex conjugate of a complex value X respectively, and $d\mathbf{s}$ is the integral variable vector with its direction normal to the integral area A on the body surface.

As can be noticed from the equation (2.4), S_{ab} represents the Poynting vector flux per unit area through the body surface or, in other words, the electromagnetic energy absorbed by the body per unit surface and per unit time. Therefore, since the temperature rise resulting from RF-EMF exposure depends on the conversion of EM energy into heat, with the help of numerical simulations and appropriate thermal models (e.g. bioheat transfer equation) it is possible to analyze, for example, the relationship between the size of the area over which incident power density is averaged and the local peak temperature elevation [33], or the temperature increase associated with certain values of power density. More precisely, thanks to Monte Carlo simulations, it was found that an absorbed power density of approximately 200 W m^{-2} is required to produce the Type-1 tissue operational adverse health effect threshold of a 5°C local temperature rise for frequencies between 6 GHz and 300 GHz [34]. Based on this result, the guidelines take an absorbed power density of 200 W m^{-2} , averaged over 6 min and a square 4 cm^2 surface area of the body, as the local exposure corresponding to the operational adverse health effect threshold for both the Head and Torso and Limb regions. As with the local SAR restrictions, a reduction factor of 2 was applied to this exposure level for occupational exposure while a reduction factor of 10 was used for the general population exposure; as shown in Figure 2.2, this results in basic restrictions of 100 W m^{-2} and 20 W m^{-2} , both averaged over 6 min and a square 4 cm^2 surface area of the body, for occupational exposure and the general population exposure, respectively.

After understanding the rationale underlying the basic restrictions values provided to ensure the protection of individuals from exposure to NIR, to conclude the analysis of the guidelines provided by ICNIRP it is necessary to give a brief description of the so-called **reference level**; this will be the focus of the following section.

2.1.2 Reference level

In the previous section, the ICNIRP basic restrictions for EMF exposure from 100 kHz to 300 GHz were examined. As has been seen, these restrictions are expressed in terms of physical quantities, such as SAR and absorbed power density, which depend on the distribution and the amplitude of the induced Electric Field inside the body as a result of exposure to RF-EMF.

The purpose of the basic restrictions, in fact, is to limit tissues heating, caused by the conversion into heat of the EM energy absorbed following exposure to RF-EMF, in order to prevent harmful effects on human health. For this reason, the recommendations provided by ICNIRP are usually used by the governments of European Union member states (such as Italy) as the basis for defining national legislation regulating individuals' exposure to RF-EMF. Therefore, the supervision of the electromagnetic fields generated by users of the frequency spectrum is often done by checking compliance with the basic restrictions provided by ICNIRP guidelines.

Assessing compliance with the basic restrictions, however, is not at all simple in practice; the direct verification of compliance with the basic restrictions, indeed, would require a high-precision spatial measurement of the induced field in the body. For this reason, in order to provide a practical means for assessing compliance with the basic restrictions using quantities that are easier to measure, ICNIRP derived, from a combination of computational and measurement studies, the so-called **reference levels** for the unperturbed Root Mean Square (RMS) values of E-Field strength, H-Field strength and incident power density (S), that ensure compliance with the basic restrictions.

Figures 2.5 and 2.6 shows, respectively, the reference levels defined in the latest release of ICNIRP guidelines (2020) for whole-body and local exposure to EMF from 100 kHz to 300 GHz [19]:

Exposure scenario	Frequency range	Incident E-field strength; E_{inc} ($V m^{-1}$)	Incident H-field strength; H_{inc} ($A m^{-1}$)	Incident power density; S_{inc} ($W m^{-2}$)
Occupational	0.1 – 30 MHz	$660/f_M^{0.7}$	$4.9/f_M$	NA
	>30 – 400 MHz	61	0.16	10
	>400 – 2000 MHz	$3f_M^{0.5}$	$0.008f_M^{0.5}$	$f_M/40$
	>2 – 300 GHz	NA	NA	50
General public	0.1 – 30 MHz	$300/f_M^{0.7}$	$2.2/f_M$	NA
	>30 – 400 MHz	27.7	0.073	2
	>400 – 2000 MHz	$1.375f_M^{0.5}$	$0.0037f_M^{0.5}$	$f_M/200$
	>2 – 300 GHz	NA	NA	10

^aNote:

1. “NA” signifies “not applicable” and does not need to be taken into account when determining compliance.
2. f_M is frequency in MHz.
3. S_{inc} , E_{inc} , and H_{inc} are to be averaged over 30 min, over the whole-body space. Temporal and spatial averaging of each of E_{inc} and H_{inc} must be conducted by averaging over the relevant square values (see eqn 8 in Appendix A for details).
4. For frequencies of 100 kHz to 30 MHz, regardless of the far-field/near-field zone distinctions, compliance is demonstrated if neither E_{inc} or H_{inc} exceeds the above reference level values.
5. For frequencies of >30 MHz to 2 GHz: (a) within the far-field zone: compliance is demonstrated if either S_{inc} , E_{inc} or H_{inc} , does not exceed the above reference level values (only one is required); S_{eq} may be substituted for S_{inc} ; (b) within the radiative near-field zone, compliance is demonstrated if either S_{inc} , or both E_{inc} and H_{inc} , does not exceed the above reference level values; and (c) within the reactive near-field zone: compliance is demonstrated if both E_{inc} and H_{inc} do not exceed the above reference level values; S_{inc} cannot be used to demonstrate compliance, and so basic restrictions must be assessed.
6. For frequencies of >2 GHz to 300 GHz: (a) within the far-field zone: compliance is demonstrated if S_{inc} does not exceed the above reference level values; S_{eq} may be substituted for S_{inc} ; (b) within the radiative near-field zone, compliance is demonstrated if S_{inc} does not exceed the above reference level values; and (c) within the reactive near-field zone, reference levels cannot be used to determine compliance, and so basic restrictions must be assessed.

Figure 2.5 – Reference levels for exposure, averaged over 30 min and the whole-body, to EMF from 100 kHz to 300 GHz (unperturbed rms values) [19].

2. Basics of EMF measurements

Exposure scenario	Frequency range	Incident E-field strength; E_{inc} ($V m^{-1}$)	Incident H-field strength; H_{inc} ($A m^{-1}$)	Incident power density; S_{inc} ($W m^{-2}$)
Occupational	0.1 – 30 MHz	$1504/f_M^{0.7}$	$10.8/f_M$	NA
	>30 – 400 MHz	139	0.36	50
	>400 – 2000 MHz	$10.58f_M^{0.43}$	$0.0274f_M^{0.43}$	$0.29f_M^{0.86}$
	>2 – 6 GHz	NA	NA	200
	>6 – <300 GHz	NA	NA	$275/f_G^{0.177}$
	300 GHz	NA	NA	100
General public	0.1 – 30 MHz	$671/f_M^{0.7}$	$4.9/f_M$	NA
	>30 – 400 MHz	62	0.163	10
	>400 – 2000 MHz	$4.72f_M^{0.43}$	$0.0123f_M^{0.43}$	$0.058f_M^{0.86}$
	>2 – 6 GHz	NA	NA	40
	>6 – 300 GHz	NA	NA	$55/f_G^{0.177}$
	300 GHz	NA	NA	20

^a Note:

1. “NA” signifies “not applicable” and does not need to be taken into account when determining compliance.

2. f_M is frequency in MHz; f_G is frequency in GHz.

3. S_{inc} , E_{inc} , and H_{inc} are to be averaged over 6 min, and where spatial averaging is specified in Notes 6–7, over the relevant projected body space. Temporal and spatial averaging of each of E_{inc} and H_{inc} must be conducted by averaging over the relevant square values (see eqn 8 in Appendix A for details).

4. For frequencies of 100 kHz to 30 MHz, regardless of the far-field/near-field zone distinctions, compliance is demonstrated if neither peak spatial E_{inc} or peak spatial H_{inc} , over the projected whole-body space, exceeds the above reference level values.

5. For frequencies of >30 MHz to 6 GHz: (a) within the far-field zone, compliance is demonstrated if one of peak spatial S_{inc} , E_{inc} or H_{inc} , over the projected whole-body space, does not exceed the above reference level values (only one is required); S_{eq} may be substituted for S_{inc} ; (b) within the radiative near-field zone, compliance is demonstrated if either peak spatial S_{inc} , or both peak spatial E_{inc} and H_{inc} , over the projected whole-body space, does not exceed the above reference level values; and (c) within the reactive near-field zone: compliance is demonstrated if both E_{inc} and H_{inc} do not exceed the above reference level values; S_{inc} cannot be used to demonstrate compliance; for frequencies >2 GHz, reference levels cannot be used to determine compliance, and so basic restrictions must be assessed.

6. For frequencies of >6 GHz to 300 GHz: (a) within the far-field zone, compliance is demonstrated if S_{inc} , averaged over a square 4-cm² projected body surface space, does not exceed the above reference level values; S_{eq} may be substituted for S_{inc} ; (b) within the radiative near-field zone, compliance is demonstrated if S_{inc} , averaged over a square 4-cm² projected body surface space, does not exceed the above reference level values; and (c) within the reactive near-field zone reference levels cannot be used to determine compliance, and so basic restrictions must be assessed.

7. For frequencies of >30 GHz to 300 GHz, exposure averaged over a square 1-cm² projected body surface space must not exceed twice that of the square 4-cm² restrictions.

Figure 2.6 – Reference levels for local exposure, averaged over 6 min, to EMF from 100 kHz to 300 GHz (unperturbed rms values) [19].

As already mentioned, the reference levels defined in the ICNIRP guidelines form the basis of the regulatory standards governing individuals’ exposure to RF-EMF, both at the National and European levels. The reason for this is quite simple and lies in the way they are defined: the reference levels, indeed, as afore mentioned, are nothing other than the maximum permissible values of E-field strength, H-field strength, and incident power density outside the human body that ensure that, even under worst-case exposure conditions (e.g. whole-body exposure to a uniform field distribution), the field induced within the body does not produce SAR or S_{ab} values exceeding the basic restrictions; therefore, given the highly conservative nature of the reference levels, if they are not exceeded one can be confident that exposure to RF-EMF does not produce biological effects capable of harming human health.

With regard to the above, however, an important clarification should be made from the outset: although the latest revision of the ICNIRP guidelines was published six years ago, and despite the Scientific Committee on Health, Environmental, and Emerging Risks (SCHEER) has recommended the European Council (EC) to revise the relevant directives in light of the latest available scientific evidence [35], the legislation currently in force in many European countries, including Italy, still refers to the recommendations set out in the previous edition of the ICNIRP guidelines, published in 1998. It

is, therefore, useful to examine the main differences between the reference levels defined in the ICNIRP (1998) and the ones defined in the ICNIRP (2020).

Figure 2.7 provides a concise comparison between the whole-body average reference levels values for the general public defined in the ICNIRP (1998) guidelines [20] and those set out in the ICNIRP (2020) guidelines (Figure 2.5) [19].

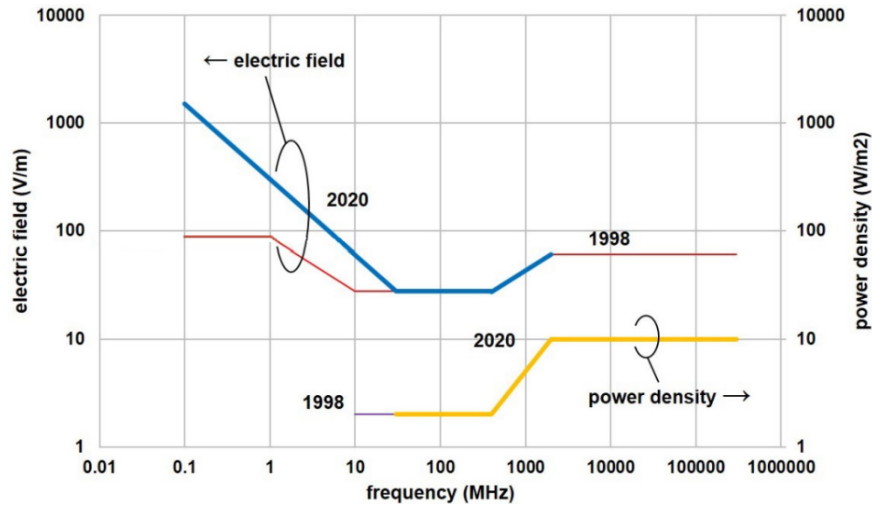


Figure 2.7 – Whole-body average reference levels for the general public for the ICNIRP (1998) and ICNIRP (2020) guidelines, for the 100 kHz to 300 GHz frequency range [36]; It should be noted that, for the sake of simplicity, only the reference levels values for the E-field and power density are shown. The reference levels for the H-field exhibit a frequency dependence analogous to that of the E-field and, therefore, all the considerations made for the reference levels values for E-field also apply to those for H-Field.

As shown in Figure 2.7, in the low frequency range (i.e., from 100 kHz to 30 MHz) the reference level values defined in the ICNIRP (1998) are very different from the ones defined in the ICNIRP (2020). This is due to the fact that for ICNIRP (1998) there was limited research available below 30 MHz to specify the reference levels and, accordingly, the reference levels were set very conservatively [36]. However, research has now clarified more accurately the relationships between the basic restrictions and the reference levels for both the E-Field and H-Fields, and, as it is known that higher reference level values are needed to reach the basic restrictions, ICNIRP (2020) has increased those reference levels accordingly. Moreover, while in the ICNIRP (1998) the reference levels for the E-Field and H-Fields began to increase as the frequency decreased from 20 MHz onward, subsequent studies showed that, in order to ensure consistency with the basic restrictions for whole-body exposure, this increase should begin at 30 MHz [36]. Consequently, in the ICNIRP (2020), the reference level values for both the E-Field and H-Fields increase monotonically as the frequency decreases, starting from 30 MHz. Further, from Figure 2.7 can also be seen that ICNIRP (1998) allowed E-field and H-field to be used for whole-body average reference levels across the entire 100 kHz to 300 GHz frequency range. However, as E-field and H-field values do not always provide a good estimate of the basic restriction values above approximately 2 GHz, E-field and H-field values are not used in ICNIRP (2020) for whole-body reference levels above 2 GHz [19][36]. There are no differences, instead, between the ICNIRP (1998) and ICNIRP (2020) whole-body average reference level values for power density above 30 MHz.

Additionally, the rules for applying the reference levels differ between the two guidelines. ICNIRP (1998), indeed, did not specify separate reference level values for exposures in the far-field and near-field zones, but instead allowed the far-field zone reference level values to be used for fields within the near-field zone. ICNIRP (2020), instead, as indicated by the annotations in Figure 2.5 and Figure 2.6, set rules for the application of reference levels in the near-field, radiative near-field and far-field zones separately, in order to ensure that exposures within the near-field zone will not result in over-exposure. Finally, the averaging time for the whole-body average reference levels has also been changed from 6 minutes in ICNIRP (1998) to 30 minutes in ICNIRP (2020), to better match the time taken for body core temperature to rise, and no local exposure reference levels were given in the ICNIRP (1998).

Considerations analogous to those set out above, also apply to the reference levels for the occupational exposure scenario (Figure 2.8).

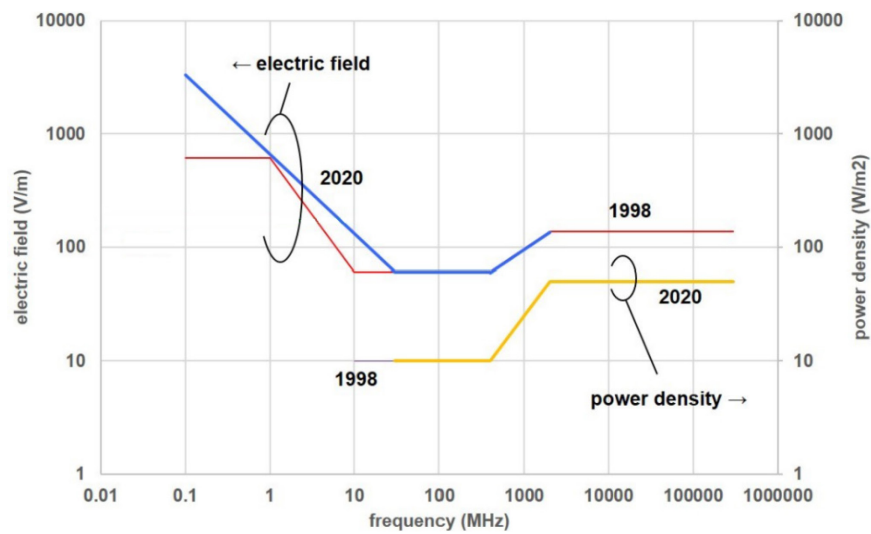


Figure 2.8 – Whole-body average reference levels for workers for the ICNIRP (1998) and ICNIRP (2020) guidelines, for the 100 kHz to 300 GHz frequency range [36].

Obviously, at present it is not possible to predict when the EC will revise the relevant directives on the limitation of exposure to electromagnetic fields in light of the new recommendations issued by ICNIRP (2020), nor to foresee what impact these recommendations will have on the definition of the exposure limits to be adopted in the future national regulations, or on the procedures to be followed when measuring RF-EMF exposure levels for the purpose of verifying compliance with the relevant legal provisions; Indeed, the ICNIRP (2020) guidelines themselves state, for example, that, depending on inputs from technical standards bodies, the actual measurement times used to obtain an appropriate estimate of exposure quantities may be shorter than the averaging time suggested in the notes to Figures 2.5 and 2.6. For these reasons, therefore, the primary objectives of the present research project were pursued with reference to the legislation currently in force at the national level. The exposure limits laid down by the legislation adopted in Italy, will be presented in the following section.

2.2 Italian regulation on limiting public exposure to EMF

As stated in the previous section, the regulatory framework currently in force within the European Union for the protection of individuals from exposure to RF-EMF in the frequency range from 100 kHz to 300 GHz, based on the *Council Recommendation on the limitation of exposure of the general public to electromagnetic fields* (1999/519/EC) and the Directive 2013/35/EU (concerning workers), relies on the recommendations set out in the ICNIRP (1998) guidelines. Both of them, in fact, fully adopt the ICNIRP (1998) recommendations and establish exposure limits (Table 2.1 and Table 2.2, respectively) which, within their respective areas of application, coincide with the reference levels suggested by ICNIRP (1998).

Frequency Range	E-Field Strength ($\text{V}\cdot\text{m}^{-1}$)	H-Field Strength ($\text{A}\cdot\text{m}^{-1}$)	B-Field (μT)	Equivalent Plane Wave Power Density S_{eq} ($\text{W}\cdot\text{m}^{-2}$)
0 ÷ 1 Hz	—	3.2×10^4	4×10^4	—
1 ÷ 8 Hz	10000	$3.2 \times 10^4/f^2$	$4 \times 10^4/f^2$	—
8 ÷ 25 Hz	10000	$4000/f$	$5000/f$	—
0.025 ÷ 0.8 kHz	$250/f$	$4/f$	$5/f$	—
0.8 ÷ 3 kHz	$250/f$	5	6.25	—
3 ÷ 150 kHz	87	5	6.25	—
0.15 ÷ 1 MHz	87	$0.73/f$	$0.92/f$	—
1 ÷ 10 MHz	$87/f^{1/2}$	$0.73/f$	$0.92/f$	—
10 ÷ 400 MHz	28	0.073	0.092	2
400 ÷ 2000 MHz	$1.375f^{1/2}$	$0.0037f^{1/2}$	$0.0046f^{1/2}$	$f/200$
2 ÷ 300 GHz	61	0.16	0.20	10

Table 2.1 – European Recommendation 1999/519/EC—Reference levels for electric, magnetic and electromagnetic fields (0 Hz to 300 GHz, unperturbed RMS values) [37]; note: f as indicated in the frequency column.

Frequency Range	E-Field Strength ($\text{V}\cdot\text{m}^{-1}$)	B-Field (μT)	Power Density ($\text{W}\cdot\text{m}^{-2}$)
100 kHz ÷ 1 MHz	6.1×10^2	$2 \times 10^6/f$	—
1 MHz ÷ 10 MHz	$6.1 \times 10^8/f$	$2 \times 10^6/f$	—
10 MHz ÷ 400 MHz	61	0.2	—
400 MHz ÷ 2 GHz	$3 \times 10^{-3}f^{1/2}$	$1 \times 10^{-5}f^{1/2}$	—
2 GHz ÷ 6 GHz	1.4×10^2	4.5×10^{-1}	—
6 GHz ÷ 300 GHz	1.4×10^2	4.5×10^{-1}	50

Table 2.2 – Action Levels (AL) for exposure to electric and magnetic fields from 100 kHz to 300 GHz [38]; note: f is the frequency expressed in hertz (Hz).

In Italy, by contrast, the RF-EMF exposure limits (Table 2.3) established by the legislation currently in force are significantly more precautionary than those just shown, which, as noted, constitute the international reference framework on the subject.

	Frequency Range	E-Field Strength ($\text{V}\cdot\text{m}^{-1}$)	H-Field Strength ($\text{A}\cdot\text{m}^{-1}$)	Power Density ($\text{W}\cdot\text{m}^{-2}$)
Exposure Limits	100 kHz ÷ 3 MHz	60	0.2	—
	3 MHz ÷ 3 GHz	20	0.05	1
	3 GHz ÷ 300 GHz	40	0.01	4
Attention Values	100 kHz ÷ 300 GHz	15	0.039	0.59
Quality Objectives	100 kHz ÷ 300 GHz	15	0.039	0.59

Table 2.3 – Exposure Limits, Attention Values and Quality Objectives for the protection of the public from exposure to electric, magnetic, and electromagnetic fields at frequencies between 100 kHz and 300 GHz [39, 40].

Indeed, as shown in Table 2.3, the Italian legislature, through the Decree of the President of the Council of Ministers (DPCM) of 8th July 2003 [39] and its subsequent amendment [40], in addition to the **Exposure Limits** — which represent the values of electric, magnetic, and electromagnetic fields that **must not be exceeded under any conditions of public exposure** — has also established:

- **Attention Values**, namely the values of electric, magnetic, and electromagnetic fields that **must not be exceeded** inside buildings intended for **occupancy of no less than four hours** per day, and in their associated outdoor areas usable as residential environments. These limits constitute an **additional precautionary measure** aimed at protecting against possible **long-term effects** that may be associated with exposure to RF-EMFs;
- **Quality Objectives**, which apply outdoors in intensively frequented areas, that aim to ensure the progressive minimization of exposure to electromagnetic fields.

Since, as already emphasized, the exposure limits commonly adopted in most European countries are significantly higher than those laid down by Italian legislation, it was decided to refer to the latter for the achievement of the primary objectives of this research project. Indeed, by designing the device of interest so as to enable the detection of field levels consistent with the Quality Objectives (Table 2.3) established by Italian legislation, one can be confident that it will also be capable of fulfilling its function in contexts where less stringent exposure limits apply.

However, identifying the minimum field levels to be detected is not, in itself, sufficient to fully achieve the intended objectives. Indeed, in order to properly design the device of interest, it is also necessary to consider the methodological framework in which it will have to operate. It is therefore useful to examine the protocols generally used to carry out EMF measurements, and the different types of instruments that may be employed for this purpose. This type of analysis, in fact, will make it possible to identify the type of device best suited to achieving the primary objectives of this research project and, consequentially, will be useful in selecting the hardware architecture to be employed for that purpose.

2.3 Standardized measurement protocols

Measurements intended to verify compliance of exposure levels with the limits laid down by the regulations must necessarily be carried out using standardized procedures, defined by specific technical committees, capable of ensuring the reliability of the obtained results.

With respect to EMF measurements, the European harmonized Standard (EN – European Norm, from the German name “Europäische Norm”) that defines the procedures for carrying out measurements aimed at verifying compliance with the exposure limits established by the applicable regulations is the **EN:50413**. Specifically, according to EN 50413, the measurement protocol to be adopted in order to perform EMF compliance measurements, may be outlined as follows [41]:

- **Identifications before the measurements:**
 - determine the relevant EMF sources and their characteristics (e.g. frequency, power, modulation, radiation characteristics, operating and worst-case conditions);
 - determine possible relevant influences on the field, such as ground conductivity, presence of reflecting objects, etc.;
 - determine the location of the exposed persons and accessible areas;
 - determine the time of exposure (high frequency only);
 - make a rough assessment of the exposure to prevent overexposure of the person performing the measurement;
 - choose type of measurement (broadband or narrowband);
 - select the adequate measurement system in compliance with the characteristics of the relevant EMF source;
 - check the proper operation of the probe and measurement equipment;
- **Preliminary scan:**
 - a preliminary scan may be performed to determine the distribution of the field;
 - the overall result of the scan shall be documented;
- **Detailed measurement:**
 - selection of the measuring procedure;
 - the locations for a detailed scan should be chosen taking into account the preliminary scan results;
 - positioning the probes in accordance to the activities and the locations of the exposed persons (head, trunk, sit, stand);
 - define the conditions of the relevant EMF sources during the measuring;
 - perform a plausibility check on the measurements.

In order to better clarify the rationale behind the standardized protocol just presented, it is useful to make a few observations. First of all, it is useful to emphasize that EMF compliance measurements must be done by **qualified personnel**; to make meaningful measurements, indeed, the behavior and the characteristics of the source of exposure shall be determined and the operation of the measurement equipment must be well known. Moreover, the expected emission characteristics of the source should be understood, and it should be confirmed that these matches the measurements' results.

In addition, the selection of the measurement points is crucial to the validity of the measurements themselves. Indeed, the selected points must be significant for a comprehensive assessment of the

electromagnetic impact produced by the installation under examination. Typically, a number of points are selected within an area of approximately 200 m radius around the installation site (beyond the field is considered negligible) on the basis of an analysis of the area map and an on-site inspection aimed at verifying the actual field conditions. Moreover, these points must be those most affected by the antenna lobes and normally occupied by, or accessible to, the general public (houses, rooftops and terraces, etc.). They must also be located away from reflective elements (such as railings or metal window frames), as well as from the operator, in order to obtain measurements of the theoretically **unperturbed field**. In fact, a high-frequency EMFs can induce surface currents on metallic objects, resulting in scattered wave that superimposes on the primary radiated field and reduces the overall accuracy of the measurement. Similarly, proximity effects of the observer, as well as others that may be in the vicinity of the probe, can influence the EMF being measured and, therefore, should be avoided; to prevent these effects, the personnel making the measurement and anyone else shall be sufficiently far from the meter and the measuring system (at least ten wavelengths) or the probe shall be set up on a non-conductive stand (e.g. wooden tripod) or supported using a non-conductive pole. Moreover, it is important to position the probe at least at three different heights, corresponding to the average height of children (1.1 m) and to the average trunk (1.5 m) and head (1.8 m) heights of adults.

Another important observation concerns the type of measurement to be carried out. Indeed, there are two different types of measurements that may be performed:

- **Broadband measurements**, carried out over a wide frequency range, and intended to estimate the overall EMF intensity, i.e., the Root Sum of the Square (RSS) of all contributions generated, at the measurement point, by all sources operating within the frequency range relevant to radioprotection regulations, such as broadcast stations (AM, FM, T-DAB, DVB-T) and mobile phone base stations (2G, 3G, 4G, and 5G);
- **Narrowband (or frequency-selective) measurements**, conducted, at specific bands, in order to isolate and measure the individual contribution produced by a specific EMF source.

It should be noted that broadband measurement is the appropriate method for identifying an actual exceedance of the exposure limits. Indeed, if such a measurement shows that, at a given point and over the time interval prescribed by the applicable regulations, the field value exceeds the prescribed limits, it may be concluded with certainty that the overall emission levels are not compliant with what established by law. The frequency-selective measurement, on the other hand, serve as a supplementary procedure, when broadband measurements exceed a predefined threshold (e.g. $0.1 \text{ W}\cdot\text{m}^{-2}$), to identify a **potential** exceedance of the limits [43]. Indeed, thanks to frequency-selective measurement it is possible to collect the incident power density within a designated frequency range. Using the measured value of specific, reference signals per technology, therefore, the **maximum potential** EMF level (i.e., the field level that would be produced under conditions in which all nearby sources are operating simultaneously, each at its maximum radiated power) at a particular location can be calculated based on extrapolation.

Finally, in order to obtain reliable and repeatable results, the equipment (field meter and probe) used for the measurements must be certified, compliant with the relevant harmonized standards [42][44], and subject to periodic calibration procedures. More specifically, the measurement equipment should be designed for practical use in rough industrial environment and for outdoor use; in addition, the measurement device should employ at least an RMS detector, and should be designed.

2.4 Non-standardized measurement protocol in exposure assessment studies

In the previous section, the standardized protocol to adopt for performing EMF level measurements aimed at verifying compliance with the exposure limits established by the legislation in force was presented. As is well known, however, EMF measurements are not used solely for compliance assessments; in fact, one may be interested in evaluating EMF levels in order to satisfy others highly interesting objectives in the research fields related to the assessment of the human exposure to EMFs, such as validating numerical models [45], tracking the evolution of exposure levels over time and comparing average exposure levels across different geographical regions [46], evaluating exposure levels among specific population segments [47], creating exposure heat maps for specific areas [48], and estimating realistic exposure levels [49]. For many of these purposes, the measurement techniques and the measurement devices typically employed for compliance testing are not the most suitable choices. This is due to a variety of factors, including the time required to perform the measurements of interest, the physical size of the high-performance measurement equipment typically used for compliance testing, and the very high expense needed to buy specialized measurement equipment. In many cases, in fact, measurements need to be conducted at multiple locations, individuals being tested must carry measuring devices without significant disruption to their daily routines, and/or measurements have to be performed continuously for extended periods in locations where delicate equipment could be at risk of damage or theft. Therefore, in order to address these issues, and also thanks to the widespread availability of low-cost electronic components with very good performance, custom devices have been developed over the last two decades whose primary purpose is to collect large amounts of data from which scientifically relevant conclusions can be drawn [50].

An analysis of the main characteristics of the custom devices mentioned above is undoubtedly valuable for the achievement of the primary objectives of the present work. As emphasized several times in the preceding sections, indeed, the main aim of the research project under consideration is to provide a proof of concept for the development of a low-cost wearable device for assessing exposure levels to future 6G signals operating in the FR3 band. In this respect, examining the main features of the aforementioned devices may provide useful insights for achieving the intended objective. This analysis, however, will be provided in the following section. Attention is now devoted to the brief description of the most common non-standardized measurement protocols made possible by the development and the adoption of the aforementioned low-cost devices [50]:

2.4.1.1 *Microenvironmental studies*

The objective of microenvironmental studies is to assess the electric field strength or exposure by studying multiple locations of a similar type [51] ÷ [57]. These types of locations or microenvironments (MEs), more precisely, are classified, according to the activities carried out within them, in categories such as home, workplace, indoor public spaces, shopping areas, and outdoor settings, and within these main categories, depending on the number of samples, further subdivisions can be defined (for instance for transport: waiting at railway station or waiting at bus stop). Other important levels defining the exposure level are time of day (night, morning, afternoon, evening, rush hour), season and type of area (rural, suburban, urban) [58][59]. For each microenvironment, repeated

measurements are carried out at multiple points or along predefined routes; such measurements are generally performed using wearable devices capable of assessing exposure, across different frequency bands, at a high rate (typically recording data every 3 s across all frequency bands). To obtain a representative measurement of exposure, a minimum of 15 min of data acquisition is required [60]. The fundamental idea underlying this type of study, therefore, is that individuals typically spend their daily routine across a sequence of microenvironments (e.g., home, street, workplace, supermarket, etc.). By characterizing the average exposure level associated with each of these environments, it is possible to estimate the individual's overall average exposure throughout the course of a typical day.

2.4.1.2 *Survey studies*

The objective of survey studies is to evaluate individual exposure in specific population groups, defined according to characteristics such as age, residential area, and occupational category, with the aim of obtaining a representative estimate of personal exposure across the different population subgroups. [47]. From each subset, a select number of participants are chosen and provided with a portable and/or wearable measuring device. Participants are required to carry the measurement device with them for several days and maintain a diary of their activities [61]. Additionally, the measurement device often records GPS data [61]. Key requirements include the random selection of participants who are representative of their respective population subsets and ensuring that the participant sample size is sufficiently large [47].

2.4.1.3 *Distributed sensor networks*

Distributed networks consisting of multiple sensor nodes (typically ranging from tens to hundreds of nodes) allow the monitoring of EMF exposure over a relatively large geographic area [48], [62] ÷ [69]. Unlike approaches focused on individual subjects or specific microenvironments, the distributed networks methodology is designed to characterize the spatial and temporal variability of exposure at the scale of neighborhoods, urban districts, or entire sections of a city. The sensor nodes, indeed, may be deployed either in fixed locations, including both indoor and outdoor public spaces, or mounted on moving platforms such as postal service cars [62], thereby forming either fixed or mobile monitoring networks [50]. Therefore, by collecting data over extended periods of time (ranging from weeks to years), these sensor networks make it possible to investigate the temporal progression of exposure levels, either within specific frequency bands or over the entire spectrum. Obviously, given the large number of nodes required for the deployment of distributed monitoring networks, together with the risk of damage or vandalism when such nodes are installed in public areas, they must be designed cost-effectively.

As noted previously, the different kinds of studies described above have been enabled by the development of custom devices whose physical sizes and implementation costs are substantially lower than those of the high-end commercial equipment commonly used for RF-EMF measurements. For the sake of completeness, therefore, in the following section a brief description of the different types of instruments that may be employed for RF-EMF measurements will be given.

2.5 State-of-the-art RF-EMF measurement equipment

In this section, in order to provide the reader with a clearer understanding of the design choices adopted to fulfil the primary objectives of the present work, which will be outlined in Chapter 4, the main classes of devices that can be employed for RF-EMF exposure assessment are briefly described. More specifically, on the basis of a careful review of the relevant literature, a concise overview of the main characteristics of high-end laboratory instrumentation, as well as user-developed devices and sensors, that have been employed in measurement campaigns documented in peer-reviewed publications is presented.

2.5.1 High-end commercial instruments

2.5.1.1 Spectrum Analyzers



Figure 2.9 – Handheld Spectrum Analyzers (i) Rohde Swartz FPH series (ii) Keysight FieldFox series (iii) Aaronia HF series.

The reference class of instruments employed for signal detection, demodulation and RF-EMF levels measurement across a large frequency spectrum and bandwidths is known as spectrum/signal analyzers (SAs). More specifically, as will be discussed in greater detail in Chapter 3, which presents the results of two different EMF measurement campaigns carried out using a handheld SA, a SA is a superheterodyne receiver architecture which is capable of analyzing the spectrum of the signals applied to its input by measuring the amplitude of the input signal as a function of frequency over the entire operating frequency range of the instrument. SAs can be broadly classified into two different categories:

- **swept-tuned SAs**, which operate in the frequency domain, in which the spectrum is scanned by mixing the input signal with a variable-frequency sinusoidal signal, while a single fixed-frequency filter progressively analyzes the signal one frequency at a time. Reducing the bandwidth of this filter allows a very high frequency resolution to be achieved. However, this comes at the expense of analysis speed, since the complete scanning of the signal requires a significantly longer time;

- **real-time** SAs (RTSA), which operate in the time domain, in which the input signal is partitioned into n sub-bands by means of n narrowband filters; within each sub-band, the signal is periodicized, and the discrete Fourier transform (DFT) is subsequently applied. These analyzers are particularly suitable for the analysis of modern 5G NR signals, whose time-varying behavior, burst-like transmissions, and dynamic resource allocation make real-time observation especially advantageous.

SAs provide best-in-class performance enabling frequency sweeping coverage up to 85 GHz, a sensitivity level, displayed average noise level (DANL), exceeding $-160 \text{ dBm}\cdot\text{Hz}^{-1}$ and provide analysis bandwidths larger than 8 GHz [50]. While an SA is typically designed by test and measurement equipment manufacturers as a versatile laboratory benchtop instrument, vendors often integrate a similar RF front-end and Digital Signal Processor (DSP) into specialized units which can be used for RF-EMF exposure assessments, known as **network scanners** [70] which often incorporates multiple channels to scan different frequency bands in parallel.

2.5.1.2 Broadband Field meters

A portable device for measuring the total electric-field strength or power density across a wide frequency range (ranging from hundreds of kHz up to a hundred GHz) is established by combining a broadband field meter with a triaxial electric field probe. Typically, the field meter display directly indicates the field measurement outcomes, such as the average level E_{avg} over a specific averaging time, and occasionally the minimum and maximum values during that period [50]. Furthermore, advanced broadband field meters, like the Narda FieldMan [75], incorporate additional functionalities such as a built-in distance meter, compatibility with a smartphone application, and integration with a software platform [71].



Figure 2.10 – Narda FieldMan (left), Narda Probe EFD-0691 (center) and Wavecontrol Probe WPF6(right).

Many probes, such as the Narda EFD-0691 [76], employ diode-based sensors characterized by a non-linear response depends on the strength of the measured signal. In this scenario, the sensor acts as RMS detector at lower field levels and transitions to a peak detector at higher field levels [50].

In contrast, other probes utilize (t)RMS (true-RMS) thermocouple sensors, which provide a (t)RMS response [71]. However, this type of detector comes with a trade-off: it reduces sensitivity to $8 \text{ V}\cdot\text{m}^{-1}$, and the dynamic range within the range of 30–40 dB [71]. It is worth noting that, for diode-based

probes, the dynamic range within the RMS domain is also typically limited to approximately 40 dB [50].

The sensitivity of broadband probes varies based on their type and the frequency range they cover. For diode-based probes used in the measurement of FR1 (such as the Narda EFD-0691 and the WaveControl WPF6 [77], both designed for the 100 kHz to 6 GHz frequency range), the sensitivity is typically at least $0.2 \text{ V}\cdot\text{m}^{-1}$. In older systems with a range of 100 kHz to 3 GHz, the minimum sensitivity was $0.3 \text{ V}\cdot\text{m}^{-1}$ [67]. For FR2-capable probes designed for very wide frequency ranges (e.g., the Narda EFD 9091 [78], which measures electric-field strength from 100 MHz to 90 GHz), the sensitivity is found to be at least $0.7 \text{ V}\cdot\text{m}^{-1}$ [50].

Broadband probes have also been designed in-lab, mostly as small dipoles loaded with a Schottky diode [79] with similar characteristics as commercial probes (i.e., in terms of sensitivity, dynamic range, frequency range). Often, these probes were developed with the goal to implement them into low-cost, small, autonomous measurement devices.

2.5.1.3 Area Monitors

Closely related to broadband field meters are devices known as **area monitors**, such as the Narda AMB series and the Wavecontrol MonitEM system (Figure 2.11).

Area monitors are designed for fully autonomous operation and are often equipped with features like solar panels, internal batteries, wireless connectivity, and automatic data transfer. In addition to a processing unit featuring a data logger, they include interchangeable field probes, which can be either broadband or tri-band/quad-band to distinguish between mobile telephone services.



Figure 2.11 – Example of area monitors: Narda AMB 8059 (left) and Wavecontrol MonitEM (right).

It is worth mentioning that Narda also offers a selective area monitor capable of monitoring up to 20 individually programmable frequency bands. Furthermore, both Narda and WaveControl offer versions of their area monitors designed for installation on vehicles. The broadband probes specially designed for area monitors exhibit similarities to those discussed for field meters. However, the tri-

band/quad-band probes, as well as those used in selective area monitors, demonstrate significantly improved sensitivity, falling within the range of $0.01\text{--}0.05\text{ V}\cdot\text{m}^{-1}$ [50].

2.5.1.4 Commercial Exposimeter

Exposimeters, also referred to as personal exposure meters (PEMs), are compact wearable devices specifically intended to assess personal exposure that can be used either by trained operators or, in some cases, by untrained volunteers [52]. Typically, these devices operate as passive objects, requiring no interaction with the body during measurements. Prior to measurement, the device is configured, and data is collected afterward.

PEMs are usually designed to measure the incident E-field strength or power density, with a sampling rate that is often customizable by the user (with a typical range of 1 to 3 s), within specific RF bands which can range in FR1 bands from 1 MHz to 6 GHz, corresponding to various technologies. This approach, in particular, allows for the segregation of exposure sources from different technologies (e.g., Wi-Fi, 2G ÷ 5G) and different sources within these technologies (e.g., UL from user devices and DL from base stations, which may use different frequency bands) [50]. In addition to their band-selective capabilities, PEMs are commonly battery-powered and are therefore suitable for prolonged measurement sessions. Depending on the specific device and application, these sessions may last from a few hours to several days and, in some cases, even up to a week. The antennas employed in this type of device, moreover, may include monopole antennas or sets of multiple monopoles oriented in different directions to approach isotropy.

Two prominent lines of commercial devices are commonly found: the ExpoM-RF (from Fields at Work) series and the EME-Spy series (designed by Microwave Vision Group). The most recent models in these lines are the ExpoM-RF 4 and the EME Spy Evolution, respectively (Figure 2.12).



Figure 2.12 – Example of commercial Exposimeters: ExpoM-RF4 (left) and EME Spy Evolution (right).

The ExpoM-RF 4 is capable of measuring at up to 25 customizable center frequencies, ranging from 50 MHz to 6 GHz, with bandwidth options of 35, 75, or 100 MHz. It allows users to set the maximum measurable field strength, though this affects sensitivity. The sensitivity for $6\text{ V}\cdot\text{m}^{-1}$ measurements fall within the range of $0.005\text{ to }0.010\text{ V}\cdot\text{m}^{-1}$, with variations across different frequencies. Users can also define the sample interval, with the shortest interval being 2 s for 10

frequency bands and up to 4 s for 25 bands. Notably, the ExpoM-RF 4 introduces a spectrum analyzer mode, distinguishing it from previous versions. Moreover, an important distinction of this latest version lies in the fact that the ExpoM-RF 3 only measured using (t)RMS with an integration time of 0.3 s [50], while the ExpoM-RF 4 performs both (t)RMS and envelope peak field strength measurements simultaneously within a 50 ms time interval. The specific post-processing method to combine these two measurements into a single E-Field value is not specified, but this information may be available from the manufacturer. The device employs a triaxial isotropic antenna.

The EME Spy Evolution, on the other hand, empowers users to monitor their choice of up to 20 frequency bands from a selection of 74 fixed options, between 80 MHz and 6 GHz, covering broadcast, cellular (2G, 3G, 4G, 5G), Wi-Fi, & ISM frequency bands. The exact sensitivity is contingent on the specific frequency; more precisely, the sensitivity is equal to $0.05 \text{ V}\cdot\text{m}^{-1}$ for frequency from 80 MHz to 700 MHz and from 3 GHz to 6 GHz, and equal to $0.02 \text{ V}\cdot\text{m}^{-1}$ for frequency between 700 MHz and 3 GHz. Users can also define the sampling interval, with a minimum setting of 2 s.

Finally, it is worth to point out that there are presently no commercially available PEMs designed for FR2. Additionally, it is worth noting that both of the commercial devices mentioned earlier are equipped with a GPS logger.

2.5.2 Custom-developed instrument

2.5.2.1 SDR-based Sensor Nodes

Spectrum analyzers, which operate in the frequency domain, may encounter limitations due to the inherent trade-off between frequency and time resolution. These limitations can make it challenging to differentiate between various signals that share the same frequency band but occur at different time intervals, like 5G-TDD UL and DL signals [80], albeit it has been demonstrated that a trade-off is achievable in some cases [81].

In contrast, time domain instruments, such as RTSA and Software-Defined Radios (SDR), can overcome this frequency-related constraint. Between these two, SDRs are notable for their compactness, portability, and relatively lower cost in comparison to SAs and RTASs. As a result, SDRs could find application in large-scale deployments, such as sensor networks.

An SDR is a flexible RF communication system that can be employed for both transmitting and receiving RF signals. It enables the implementation of physical components like filters, attenuators, amplifiers, synchronizers, modulators, demodulators, and detectors in a digital form, positioned as close as possible to the antenna [82],[83]. In this manner, the SDRs functionality is primarily dictated by software (e.g., utilizing GNU Radio), rather than hardware. This software-driven approach allows for dynamic adjustments to the SDRs operation and characteristics [82].

Examples of SDR-based measurement nodes can be found in e.g., [82] (development of a digital Fast Fourier Transform SA), [84], [85] (broadband field meter), [83] (measurement node for specific frequency bands), and [80], [86] (isotropic broadband field meter).

2. Basics of EMF measurements

SDR	Frequency Range	Resolution (bit)	Sample Rate (MSPS)	Bandwidth (MHz)	Open Source?	Cost (€)	Chipset
Adalm Pluto	325 MHz – 3.8 GHz (70 MHz – 6 GHz with software modification)	12	61.44	20	yes	250	AD9363
USRP E312	70 MHz – 6 GHz	12	61.44	56	no	6000	AD9361
RTL-SDR Blog v3	500 kHz – 1766 MHz 22 MHz – 2.2 GHz	8	3.2	3.2	no	6000	RTL2832U
HackRF One	1 MHz – 6 GHz	8	20	20	yes	370	MAX5864
FreeSRP One	70 MHz – 6 GHz	12	61.44	56	yes	420	AD9364
LimeSDR	100 kHz – 3.8 GHz	12	61.44	56	yes	300	LMS7002M
BladeRF xA44	47 MHz – 6 GHz	12	61.44	56	no	540	

Table 2.4 – Technical specifications of commercially available SDRs.

A variety of commercially available SDRs offer diverse performance levels and price ranges, as summarized in Table 2.4. While many of these devices are primarily designed for GPS or satellite transmission bands, some also cover the 5G FR1 frequency bands. Utilizing SDRs in measurement nodes offers the advantage of readily available hardware, simplifying field deployment. Additionally, certain compact SDRs, like the Adalm-Pluto (Figure 2.13), are pocket-sized and can be controlled via MATLAB or Python environments, making them well-suited for deployment as measurement nodes.

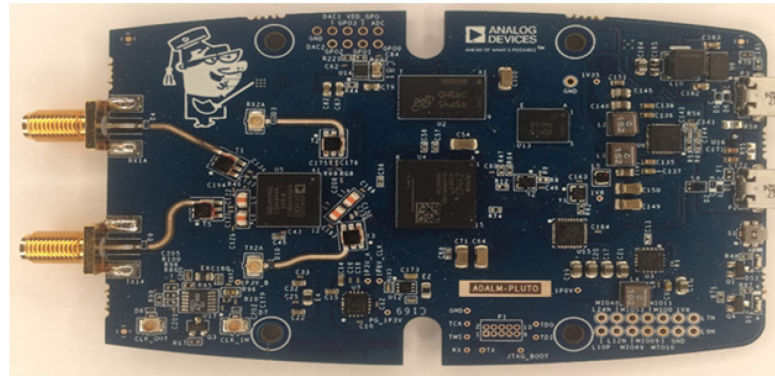


Figure 2.13 – PCB of the Adalm-Pluto SDR from Analogue Devices.

The Adalm-Pluto is an intriguing choice due to its appealing features [50]. Like all SDRs, it serves both as a transmitter (Tx) and receiver (Rx). As shown in Table 2.4, it features an AD9363 chipset with an initial frequency range of 325 MHz to 3.8 GHz, but it can be expanded to cover 70 MHz to 6 GHz with a simple adjustment. This adjustment essentially tricks the device into recognizing the AD9363 as if it was a AD9364 with extended capabilities.

The Adalm-Pluto has a 12-bit Analog to Digital Converter (ADC) resolution, a detection limit of -98 dBm, and an impressive dynamic range exceeding 80 dB [83]. It supports the connection of two antennas, such as the example in [83] where a dual-band JCG401 antenna (covering frequencies from

828 MHz to 984 MHz and 1710 MHz to 2170 MHz) and a wideband W5150 antenna (ranging from 617 MHz to 6000 MHz) were used. Furthermore, it has the capability to store in-phase and quadrature (IQ) samples (“in-phase” and “quadrature” refer to two sinusoids that have the same frequency and are 90° out of phase). Alternatively, the software on the device allows for the conversion of IQ samples into another unit, specifically dBFS (decibels relative to full scale). Through software control, gain settings of up to 74.5 dB can be configured, either manually or using automatic gain control (AGC). When employing AGC, it is important to note that changes in the full-scale range necessitate device calibration with a known source to obtain meaningful physical values [80].

2.5.2.2 Hardware-based Compact Measurements Node

To achieve accurate spatiotemporal RF-EMF exposure maps, a significant number of densely-distributed sensors are necessary [62]. However, such an extensive sensor network is economically unfeasible with high-end, costly, and relatively bulky devices. As a solution, researchers are developing low-cost (priced below 300 €), compact, and energy-efficient sensor units tailored for specific FR1 frequency bands. These units are integral in establishing a continuous RF-EMF exposure monitoring system [62], [87] ÷ [89][90].

For instance, a dedicated measurement node optimized for 5G applications, enabling measurements across the four frequency bands intended for use in the Netherlands, has been presented in [91].

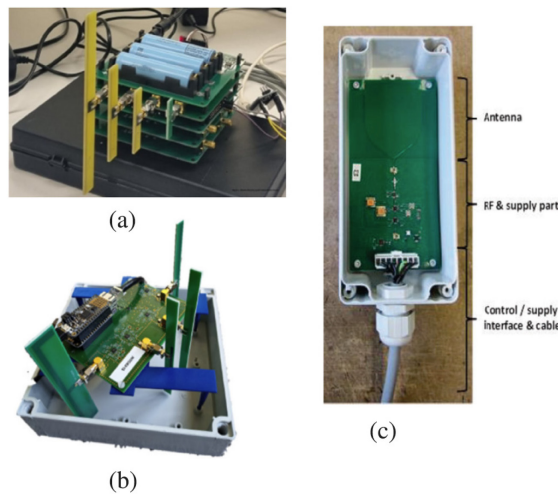


Figure 2.14 – Low-cost RF-EMF sensor nodes [50]: (a) S³R sensor node [83], WAVES sensor node [83], (c) low-complexity dosimeter [87].

Figure 2.14 displays some examples of laboratory-built, budget-friendly RF-EMF sensor nodes. The S³R and WAVES sensor nodes utilize in-house designed narrowband planar half-wavelength dipole antennas for each frequency band [83], while the low-complexity dosimeter employs a printed PCB monopole antenna probe [87]. It is important to note that all low-cost sensor node antennas are linearly polarized and exhibit omnidirectional radiation patterns.

Typically, these sensors can measure up to four frequency bands used by GSM, UMTS, LTE, and/or 5G NR for DL communications. In the design of these sensor nodes, band-pass filters are commonly employed for each band to suppress signals from neighboring and interfering bands.

Depending on the desired sensitivity level, a low noise amplifier (LNA) may be utilized to adjust the RF power level within the power detector's dynamic range. There are two primary types of power detectors: logarithmic and (t)RMS. Logarithmic detectors convert the input RF power into a DC voltage that is directly proportional to the logarithm of the input, providing an output in dB. RMS detectors produce a DC output that is proportional to the RMS value of the signal. It is worth noting that using a (t)RMS detector is more accurate, as it takes into account the instantaneous values of the signal without assuming a specific waveform shape [50]. The dynamic range of power detectors typically falls within the range of 56 – 70 dB for FR1 frequencies. The DC output is then sampled with an ADC and is converted into a power level (dBm) using a lookup table within a standard microcontroller.

Similarly, in [92] ÷ [94] cost-effective sensors for distributed EMF exposure assessment have been proposed. The sensors consist of three orthogonal dipole antennas, each featuring a diode detector to transform field values into a DC signal. Subsequently, with the use of an ADC and a microcontroller, these signals are converted into power or field values. It is worth noting that the sensors measure accurately within the frequency range of 200 – 5000 MHz [92].

All of the suggested EMF sensors are compact, budget-friendly, and user-friendly, requiring no specialized expertise for operation.

2.5.2.3 Custom-build Exposimeters

In response to the inherent measurement uncertainties associated with commercial exposimeters, an on-body Personal Distributed Exposimeter (PDE) has been proposed, featuring multiple measurement nodes [95]. Each node is seamlessly integrated onto a textile antenna feed plane equipped with patch antennas. The PDE takes the form of a wearable vest, featuring multiple nodes distributed across various pockets. This particular PDE was tailored for the GSM DL frequency band of 925 MHz – 960 MHz, employing a logarithmic RF power detector with an impressive dynamic range of 80 dB. The PDE was designed with modularity in mind, accommodating up to 11 frequency bands [95], [96]. Much like the sensors of the previous section, nodes consist of a band-pass filter, RF power detector, ADC, and a microcontroller. Figure 2.15 depicts the PDE antenna and the processing unit [50].

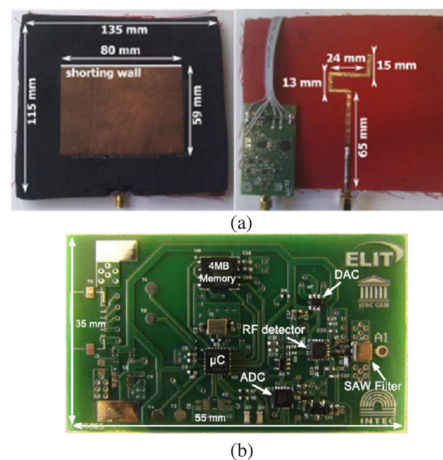


Figure 2.15 – The PDE consists of a number of distributed nodes with each (a) a textile patch antenna and (b) a processing unit [50].

A similar design is employed for a drone-based RF exposure measurement system in [97].

In a related development, a PDE-helmet was introduced as part of a pilot project [51]. Based on simulations, it was determined that the human head experiences the least variation in E-fields and, consequently, the lowest measurement uncertainty. To capitalize on this, the PDE-helmet featured four monopoles positioned within a bicycle helmet in various orientations. This prototype also employed a diode-based RMS detector and was specifically designed for the 900 MHz GSM band.

A system resembling a spectrum analyzer while retaining the key benefits of conventional exposimeters was proposed in [98]. This system operates by assessing the highest received power level through the use of a logarithmic detector. It operates across a spectrum spanning from 78 MHz to 6 GHz, employing 300 kHz resolution bandwidths within numerous narrow bands. This configuration enables the system to identify various sources of electromagnetic fields. Each measurement takes place at one-second intervals, corresponding to 46.22 μ s per frequency point. The frequency range is covered by an array of five fractal antennas sourced from Fractus Antennas, based in Barcelona, Spain. These antennas possess a wide bandwidth and are comparatively compact in relation to other types of commercial antennas. Notably, the system has a dynamic range of 90 dB with RF input power ranging from -70 to 20 dBm with a 0.04 dB resolution.

To date, only two research studies have been identified that detail the creation of custom-built measurement devices for evaluating RF-EMF exposure at frequencies within the FR2 range. These studies, documented in [99][100], introduced two distinct versions of a mm-Wave personal exposure meter (mm-PEM) designed to operate at 60 GHz. The initial version featured a limited set of wearable antennas designed for placement on a forearm [96]. These wearable antennas were based on a microstrip-fed four-patch single-layer antenna array originally developed for applications in body area networks (BANs) [101]. In the second study, the mm-PEM was specifically evaluated for its effectiveness in assessing exposure within indoor diffuse fields [100].

Based on the foregoing discussion, exposimeters may be regarded as ideal candidates for the implementation of the device of interest in the present work. Indeed, as shown above, their underlying architecture is relatively simple, as it essentially comprises a field sensor, a detector (preferably an RMS detector), and a DSP/microcontroller for the digitization of the acquired signals. Furthermore, at the frequencies relevant to the present research project, the overall dimensions of such devices can be made very compact. In fact, the physical size of the antennas remains relatively limited at these frequencies; consequently, by properly engineering the PCB forming the signal-processing unit of the device of interest, it is possible to develop wearable devices with dimensions small enough not to interfere with the users' normal daily activities. This would allow the device to be worn for extended periods, which is a key requirement for performing exposure measurements that are truly representative of actual exposure conditions for signals which exhibit highly dynamic and time-varying behavior, such as 5G NR and, probably, the forthcoming 6G signals.

3 Chapter 3: Reference measurement campaigns

In the previous chapter, the measurement protocols for assessing RF-EMF exposure levels, together with the main characteristics of the different types of devices that can be employed for this purpose, were presented. In this chapter, attention is instead focused on a more detailed discussion of the operating principles of the SAs, that represent the main instrument class used for signal characterization, and on the main difficulties encountered in the measurement of 5G signals. The rationale for this analysis lies in the fact that, once 6G become effectively available, a calibration phase of the exposimeter discussed in this work will be required. This calibration procedure will need to be carried out both in the laboratory and in the real environment, using an SA as the reference instrument. Moreover, since it is reasonable to expect that the characterization of 6G signal levels will entail challenges similar to those already posed by 5G signals, examining the solutions currently adopted for the characterization of exposure levels associated with 5G signals is particularly useful, as such solutions may also be applied, at least in part, to the future characterization of 6G signals.

3.1 Operating principles of spectrum analyzers

The reference instruments for signal detection and RF-EMF level measurements over wide frequency ranges and bandwidths are SAs, i.e., superheterodyne receivers that analyze an input signal by measuring the amplitude of its spectral components as a function of frequency over the instrument operating range. SAs can be broadly classified into two main categories:

- **Swept-Tuned** SAs, which operate in the frequency domain by progressively analyzing the signal one frequency at a time;
- **Real-Time** SAs, Real-Time SAs, which operate in the time domain by processing the entire acquired signal bandwidth simultaneously.

At present, high-end instruments marketed by the leading manufacturers (e.g., Keysight and Rohde & Schwarz) generally support both types of analysis within a single instrument. Therefore, before describing the two measurement campaigns presented in this chapter, it is useful, for the sake of completeness, to briefly review the operating principles of the two operating modes mentioned above. Indeed, although the measurement campaigns discussed in the following sections were carried out exclusively using conventional SA, it is still useful to recall the operating principles of RTSAs since they represent an important methodological reference for the characterization of radio signals with marked temporal dynamics, such as 5G and, in all likelihood, future 6G signals. This description will also help the reader to better understand the limitations of conventional SAs for the measurement of 5G EMF levels, as well as the impact that the instrument's operating parameters may have on the measurement results.

3.1.1 Superheterodyne swept-tuned SA

To understand the operating principle of a superheterodyne swept-tuned SA, let us refer to the simplified block diagram shown in Figure 3.1 and assume that a signal with spectrum $X(f)$ is applied to the instrument input.

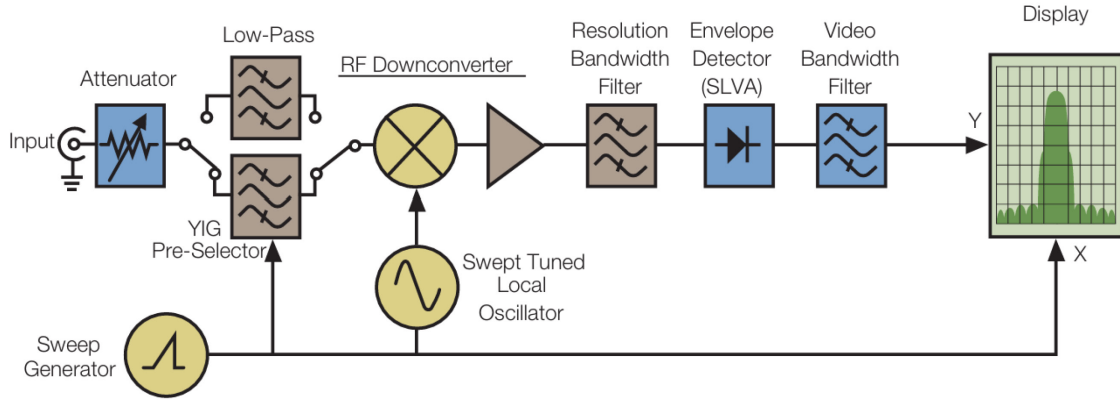


Figure 3.1 – Typical swept-tuned SA architecture [102].

As shown in Figure 3.1, after passing through a preconditioning stage consisting of a variable attenuator and a filtering section, the input signal is **mixed** with a variable-frequency sinusoidal signal generated by a voltage-controlled local oscillator. In particular, as illustrated in Figure 3.2, this operation shifts the spectrum of the input signal $X(f)$, assumed to be upper-bounded by the frequency f_m , by an amount equal to the frequency f_{LO} of the signal provided by the local oscillator, thus yielding at the mixer output a signal with spectrum $Y(f) = X(f \pm f_{LO})$.

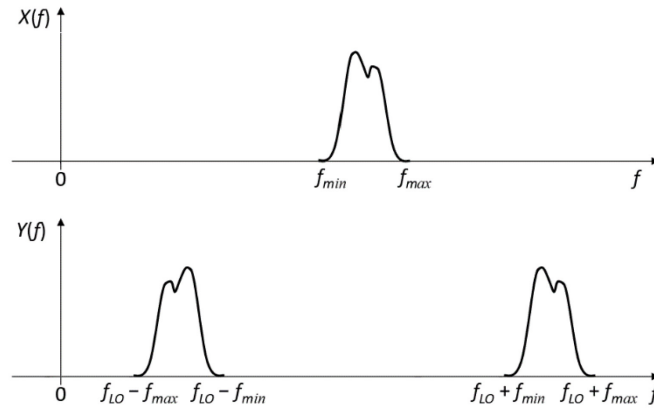


Figure 3.2 – Spectrum of the input signal, $X(f)$, and of the mixer output signal, $Y(f)$.

The converted signal $Y(f)$ is then amplified and fed to a **fixed** intermediate-frequency (IF) band-pass filter. Amplification of the signal at the mixer output is necessary to restore the original amplitude level of the signal present at the instrument input. Indeed, as previously stated, the input signal is attenuated and filtered in order to adjust the amplitude level and bandwidth of the signal applied to the mixer. This is necessary, in particular, to prevent damage to the mixer and to avoid distortion phenomena that may arise in the presence of high-level or wideband input signals, since the

mixer is a nonlinear device. However, in order to correctly display the amplitude of the spectral components of the signal at the device input, this attenuation must be compensated for. For this reason, a functional relationship is established between the RF attenuation block and the IF gain stage, so that the gain of the latter is automatically adjusted to compensate for any attenuation applied at the input.

As already mentioned, after amplification, the mixer output is fed to a band-pass filter centered at the fixed intermediate frequency f_{IF} . If f_{IN} is a frequency component present in the spectrum of the signal under analysis, the mixer output will contain the frequency components $f_{LO} \pm f_{IN}$. Since f_{IF} is always lower than the minimum frequency of the signal generated by the local oscillator, f_{LO_min} , assuming high-side injection (i.e., $f_{LO} > f_{IN}$) the only components of the input signal that can pass through the IF filter are those whose frequency f_{IN} is such that the difference $f_{LO} - f_{IN}$ falls within the IF filter passband. Moreover, as shown in Figure 3.3, by varying linearly over time f_{LO} , the aforementioned condition is satisfied sequentially for the different spectral components of the input signal. In this way, the output of the IF filter is a signal whose amplitude, as extracted by an envelope detector, represents the spectrum of the input signal.

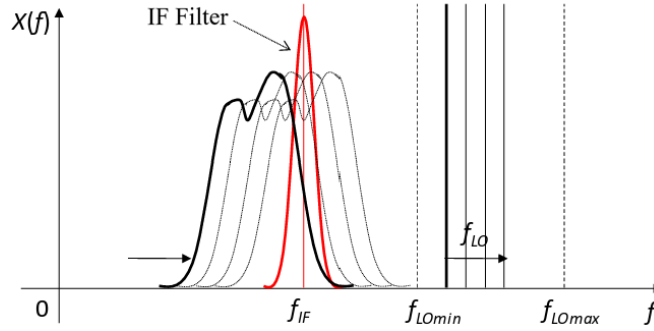


Figure 3.3 – Sweeping of the mixer output signal spectrum through the IF filter.

The linear variation over time of the local oscillator frequency f_{LO} , which, as previously discussed, makes it possible to shift the various spectral components of the input signal sequentially to the intermediate frequency f_{IF} , is achieved by means of a **sweep generator**, i.e., a ramp-voltage generator (Figure 3.4) capable of producing a periodic voltage signal, referred to as a sawtooth signal because of its resemblance to the teeth of a saw blade, that increases linearly with time and then drops abruptly. Indeed, the local oscillator used in superheterodyne SA is a voltage-controlled oscillator, capable of generating at its output a sinusoidal signal whose frequency $f_{LO} \in [f_{LO_min}, f_{LO_max}]$ is proportional to the voltage of the signal applied to its input.

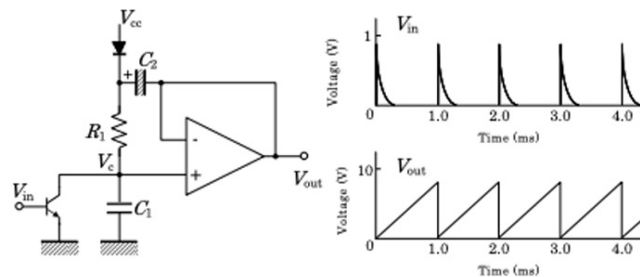


Figure 3.4 – Example of ramp-voltage generator.

The considerations outlined above further emphasize the importance of properly conditioning the signal fed to the mixer in order to achieve accurate spectral analysis and to prevent unwanted input spectral components from giving rise to the so-called image-frequency problem. Indeed, any unwanted component at a frequency f_{IN} such that $|f_{IN} - f_{LO}| = f_{IF}$ is converted to the same intermediate frequency as the signal of interest and may therefore pass through the IF filter, appearing on the instrument display as if it belonged to the portion of the spectrum under analysis. However, the filtering stage at the instrument input (Figure 3.1) limits the bandwidth of the signal applied to the mixer and thus prevents unwanted image-frequency components from being converted to the same intermediate frequency. A further disadvantage of the superheterodyne approach is that it measures the amplitude at only one frequency point at a time. This operating principle therefore relies on the assumption that at least one sweep can be completed without significant changes occurring in the signal under test. Consequently, the measurement is reliable only for relatively stable input signals. If the signal undergoes rapid variations, transient spectral events may occur and disappear before the sweep reaches the corresponding frequency region, and may therefore remain undetected (Figure 3.5). Therefore, in order to minimize the probability of missing transient events of this kind, the time required to complete a sweep over the desired frequency span should be kept as short as possible.

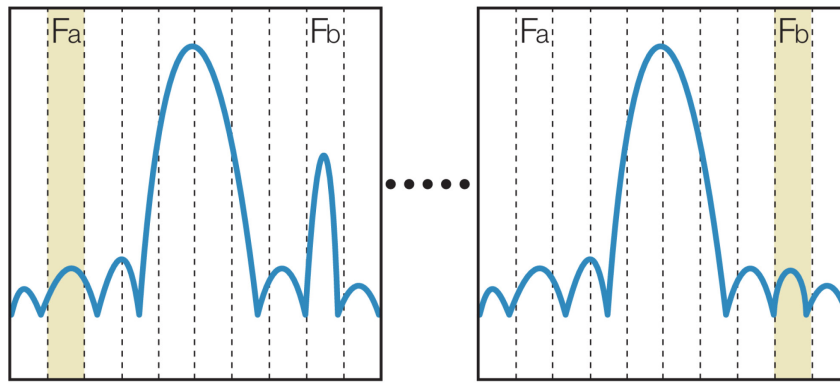


Figure 3.5 – The Swept SA tunes across a series of frequency segments, often missing important transient events that occur outside the current sweep band highlighted in tan segment Fb on the right [102].

In this context, it is important to note that, as mentioned earlier, the IF filter has a fixed center frequency. Its bandwidth, referred to as the **Resolution Bandwidth** (RBW) because it determines the frequency resolution of the instrument, i.e., its ability to distinguish between two tones that are closely spaced in frequency (Figure 3.6), is instead adjustable within a certain range in order to achieve the best trade-off between frequency resolution and sweep time. Indeed, a reduction in the RBW leads to an increase in the sweep time because a narrower IF filter, although providing better frequency resolution, exhibits a slower time-domain response. Since each spectral component remains within the filter passband only for a limited time interval during the sweep, this interval must be sufficiently long for the filter response to settle to its correct value in order for the measured amplitude to be accurate. Consequently, as the RBW decreases, the sweep speed must be reduced, resulting in a longer sweep time.

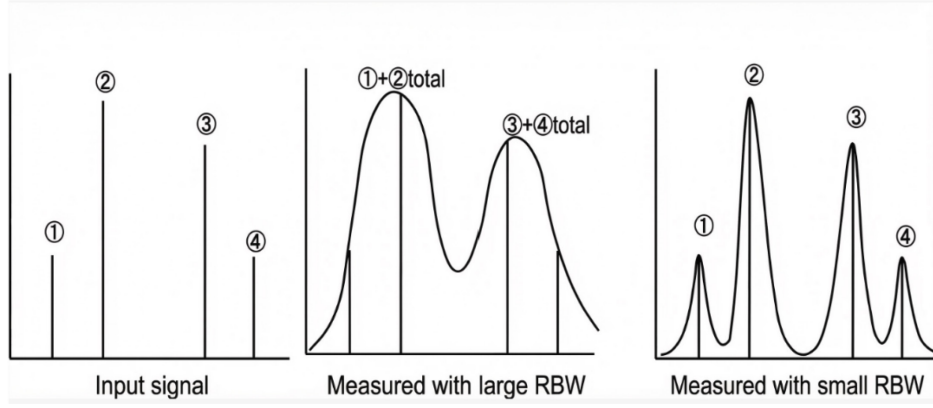


Figure 3.6 – Effect of RBW settings on the resolution of adjacent spectral components.

In addition to the RBW, the other key parameters that define the performance of a swept-tuned spectrum analyzer are:

- **frequency range**, defined as the range of input frequencies that can be analyzed by the instrument. This parameter is directly related to the characteristics of the local oscillator. In fact, the lower limit is constrained by the fact that, in order to enable operation at low frequencies, f_{LO_min} must approach f_{IF} . The upper limit, on the other hand, is determined by the maximum value that f_{LO} can reach; indeed, in the case of high-side injection, it is given by $f_{IN_max} = f_{LO_max} - f_{IF}$. In practice, this limit is also constrained by the operating bandwidth of the RF front-end and the mixer;
- **dynamic range**, defined as the difference between the maximum and minimum signal levels that the instrument can measure simultaneously. It is mainly determined by the internal noise floor, the linearity of the RF front-end and mixer, and the level of internally generated distortion products;
- **sensitivity**, defined as the minimum amplitude level that the input signal must have in order to be detected. It is mainly determined by the internal noise of the receiver, the noise figure of the input stages and the minimum achievable RBW.

Finally, in order to complete the description of the operating principle of a swept-tuned spectrum analyzer, it is useful to provide a brief overview of the different measurement modes offered by the instrument:

- **Channel Power (CP)**: this measurement mode is based on the evaluation of the total power associated with a specific channel and, in mathematical terms, it can be interpreted as the integral of the acquired spectrum over a given frequency interval, commonly referred to as the **integration bandwidth**. The value of the CP can be calculated from the acquired spectral traces by applying the following formula [103]:

$$CP = 10 \log_{10} \left[\frac{IBW}{NBW} \cdot \left(\frac{1}{N} \sum_{i=1}^N 10^{\frac{P_i - \alpha_i}{10}} \right) \right] [dBm] \quad (3.1)$$

where (Figure 3.7Figure 1.1) IBW is the integration bandwidth, P_i (dBm) is the power value associated with the i -th of the N pixels falling within the IBW , α_i is a correction parameter that

3. Reference measurement campaigns

varies from pixel to pixel and accounts for the weighting introduced by any digital filters implemented in the instrument and employed during the measurement, and *NBW*, sometimes also referred to as the Equivalent Noise Bandwidth (ENBW), denotes the noise bandwidth of the IF filter used, i.e., the bandwidth of an ideal rectangular filter having the same area as the actual frequency response of the IF filter (in modern spectrum analyzers the ratio between *NBW* and *RBW* is typically close to unity).

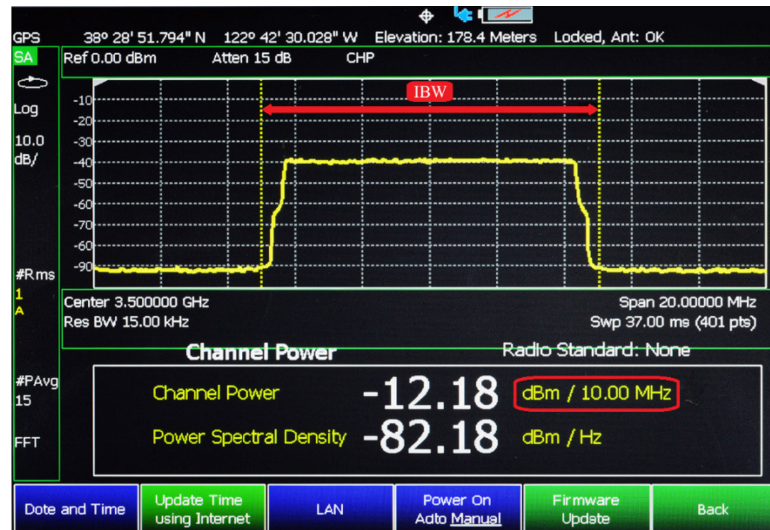


Figure 3.7 – Example of CP measurement.

- **Zero-span:** this is a particular operating setting of the SA in which the instrument is effectively used as an oscilloscope. In this mode, indeed, the instrument provides the time evolution of the received power at a fixed frequency, integrated over the bandwidth identified by the selected RBW value (Figure 3.8). This capability is useful for analyzing modulation characteristics, such as pulsed measurements.

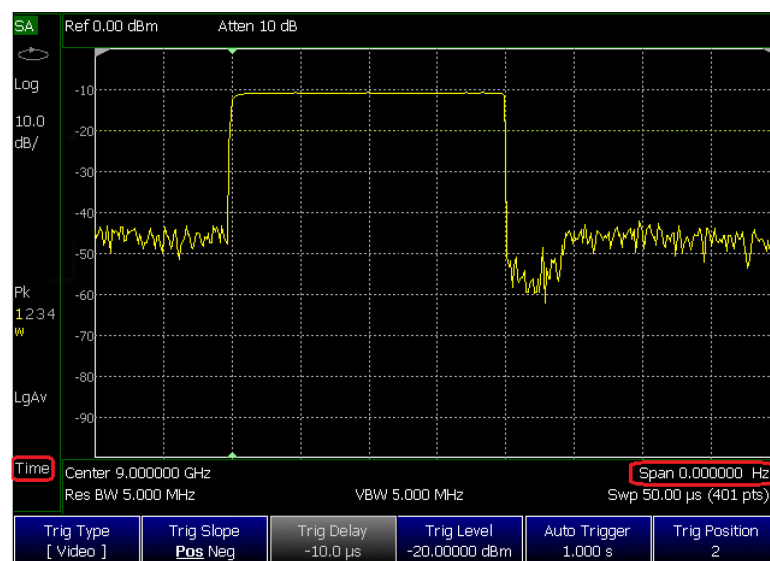


Figure 3.8 – Example of zero-span measurement.

3.1.2 Real-time SA

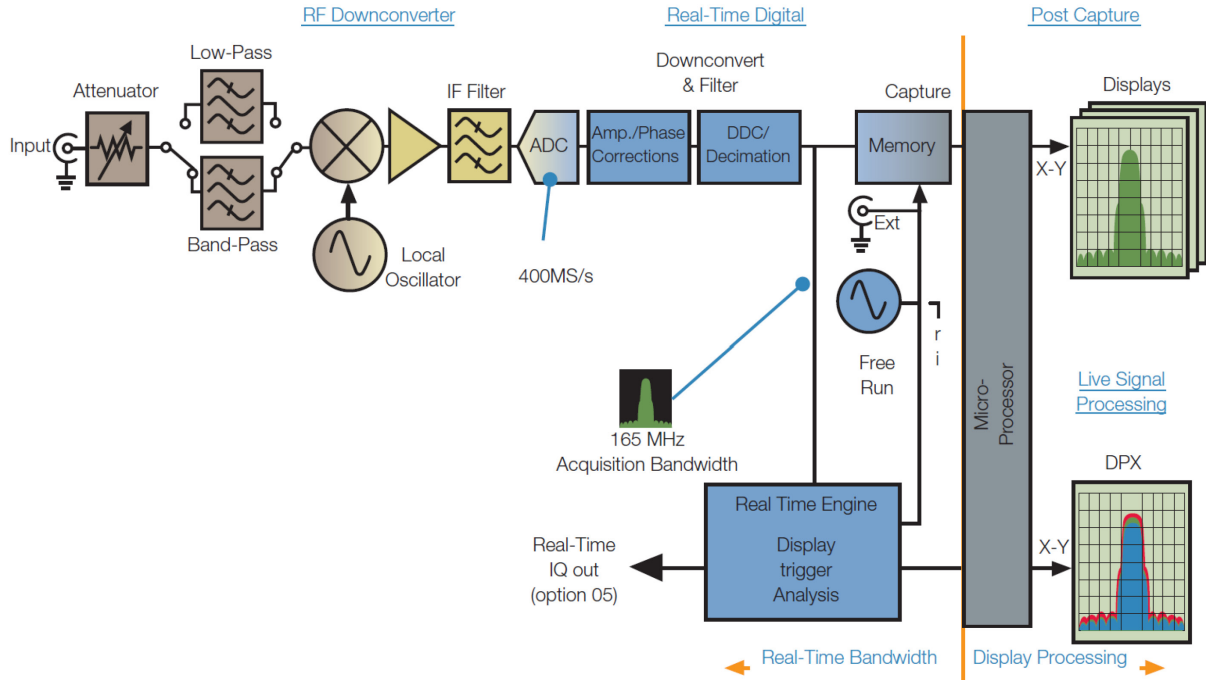


Figure 3.9 – Simplified Block Diagram of Real-Time SA [102].

As shown in the simplified block diagram of Figure 3.9, the basic architecture of an RTSA does not differ radically from that of the conventional swept-tuned SA discussed in the previous section, since both share the general superheterodyne receiver architecture. The fundamental difference between the two architectures lies in the way the spectrum of the signals applied at the instrument input is reconstructed. Indeed, as discussed in the previous section, a swept-tuned SA performs power-versus-frequency measurements (one frequency point at a time) by downconverting the signal of interest and sweeping it through the passband of a RBW filter; an RTSA, on the other hand, as will be discussed in the following, instead of sweeping across a range of frequencies, acquires at once the entire RF content within the band of interest and can therefore detect and analyze any signal occurring within that range. To better clarify what has just been mentioned, it is useful to examine how the functional blocks shown in Figure 3.9 process the signals applied at the instrument input.

As in a conventional swept-tuned SA, the first functional stage of an RTSA is the signal-conditioning stage, which consists of a variable attenuator followed by a filtering section. Since the analyzer must receive signals with widely different amplitudes without overloading the internal circuitry or compromising measurement accuracy, the variable attenuator reduces the input signal to a level suitable for downstream processing while also providing a broadband impedance match, which is essential for accurate RF measurements. The filtering section, instead, acts on the spectral content of the input signal: the low-pass and band-pass filters suppress unwanted input spectral components and help ensure image-free frequency conversion from the RF and microwave input to the final IF [102].

After the signal-conditioning stage, the input signal reaches the RF downconversion stage, where the first substantial differences with respect to the architecture of a swept-tuned SA become apparent. Indeed, a comparison of Figure 3.9 with Figure 3.1 shows that the RTSA architecture does not include a sweep-generator to linearly vary the frequency of the local oscillator that drives the mixer. This reflects the fact that, during the real-time measurement, the local oscillator frequency is kept constant. Under these conditions, the band of interest is shifted at the mixer output by a constant frequency offset and, through a subsequent wideband IF filtering stage, is brought to a fixed IF so as to simplify the subsequent processing stages. In practice, in actual RTSA implementations, as shown in Figure 3.10, tuning is achieved through a multi-stage frequency-conversion scheme employing a combination of tunable and fixed local oscillators within a multiple-conversion superheterodyne architecture. Each frequency-conversion stage consists of a mixer followed by IF filtering and amplification, and the IF frequencies, filter shapes, gains, and signal levels are chosen so as to optimize several critical performance parameters, including spurious responses due to mixer and filter imperfections, dynamic range, amplitude flatness across the real-time bandwidth (namely, the widest frequency range over which real-time analysis can be performed), phase linearity, and delay match between the signal and trigger paths [102][103].

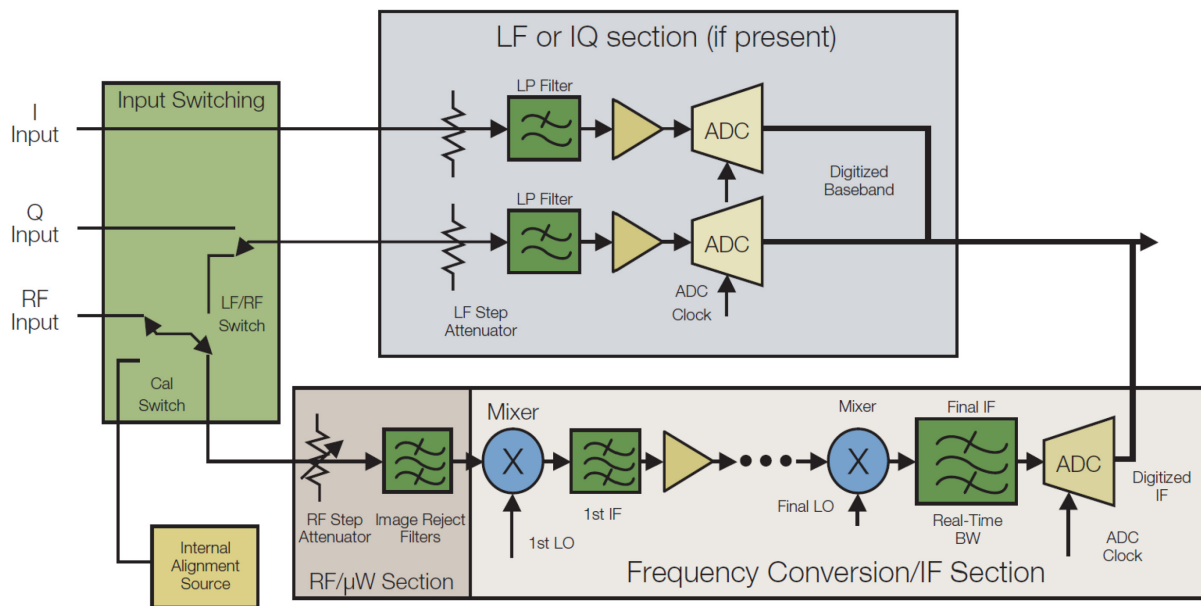


Figure 3.10 – RTSA RF/IF block diagram [102].

After RF/IF conversion, the output signal from the final IF filter (Figure 3.10) is digitized by means of an ADC whose sampling rate must be high enough to ensure that the entire real-time bandwidth of the instrument can be digitized correctly without aliasing. This step marks the transition from analog RF/IF conditioning to digital signal processing; from this point onward, all subsequent operations are carried out digitally.

The first digital operation is performed by the amplitude/phase correction block. The need for this stage arises from the fact that the amplitude flatness, phase linearity, and level accuracy required for real-time spectrum analysis cannot be guaranteed by the analog RF/IF chain alone [102]. Indeed, mixers, filters, and amplifiers introduce amplitude and phase ripple over the wide bandwidths

acquired by RTSAs. For this reason, the RF response measured at the factory (stored in a non-volatile memory) and the IF response measured during internal alignments, are used to characterize the overall response of the signal path up to the ADC and to generate digital correction filters that compensate for these nonidealities.

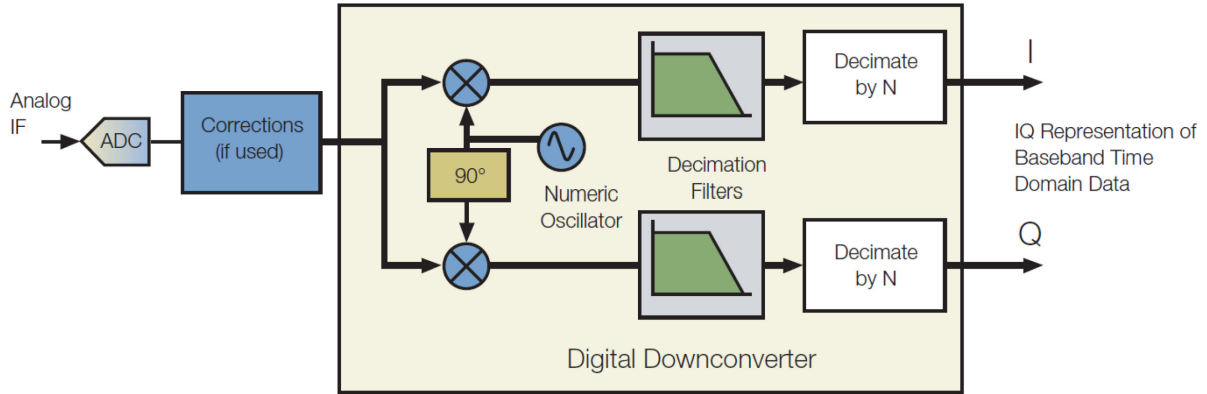


Figure 3.11 – IF to IQ conversion in a RTSA [102].

As shown in Figure 3.11, once corrected, the digitized IF signal is processed by the digital downconverter/decimation block, whose purpose is to represent the time-sampled data in terms of the I (in-phase) and Q (quadrature) baseband components of the signal by means of a digital downconverter (DDC). In general, a DDC contains a numeric oscillator that generates a sine and a cosine at the center frequency of the band of interest, which are numerically multiplied with the digitized IF signal, thereby generating streams of I and Q baseband samples that contain all of the information present in the original IF signal. Since all of the information contained in the original IF signal is preserved in the two baseband signals, the minimum required sampling frequency for each of them is now half of the original one, so that decimation by a factor of two becomes possible [102]. However, the Nyquist theorem must also be satisfied when decimating. If the data rate is reduced by a factor of two, then the bandwidth of the digital signal also must be reduced by a factor of two. This must be achieved by means of a digital filter applied prior to the reduction in sample rate in order to prevent aliasing (Figure 3.11)

It is worth noting that, up to this point, the instrument has not yet reconstructed the spectrum of the input signal. Rather, the processing chain described so far has only translated, corrected, and reformulated the band of interest into a digital time-domain representation suitable for analysis. The actual transition from the time domain to the frequency domain takes place only in the **real-time DSP engine** (Figure 3.9). Here, successive DFT/FFT operations are applied to consecutive segments of the I/Q record prior to memory storage, thereby generating a time-correlated description of the signal spectral occupancy. It is important to note that the real-time capability of the RTSA does not arise merely from wideband acquisition and the use of DFT/FFT operations, but from the fact that the incoming data is processed by a dedicated DSP engine prior to memory storage and at a rate high enough to avoid time gaps between consecutive analysis frames; in this way, all computations can be performed continuously and fast enough such that the output of the analysis keeps up with the changes in the input signal.

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From a mathematical point of view, applying sequential DFTs at equally spaced time intervals is equivalent to passing the input signal through a bank of parallel bandpass filters with equal frequency resolution and bandwidth and then sampling the magnitude and phase at the output of each filter (Figure 3.12).

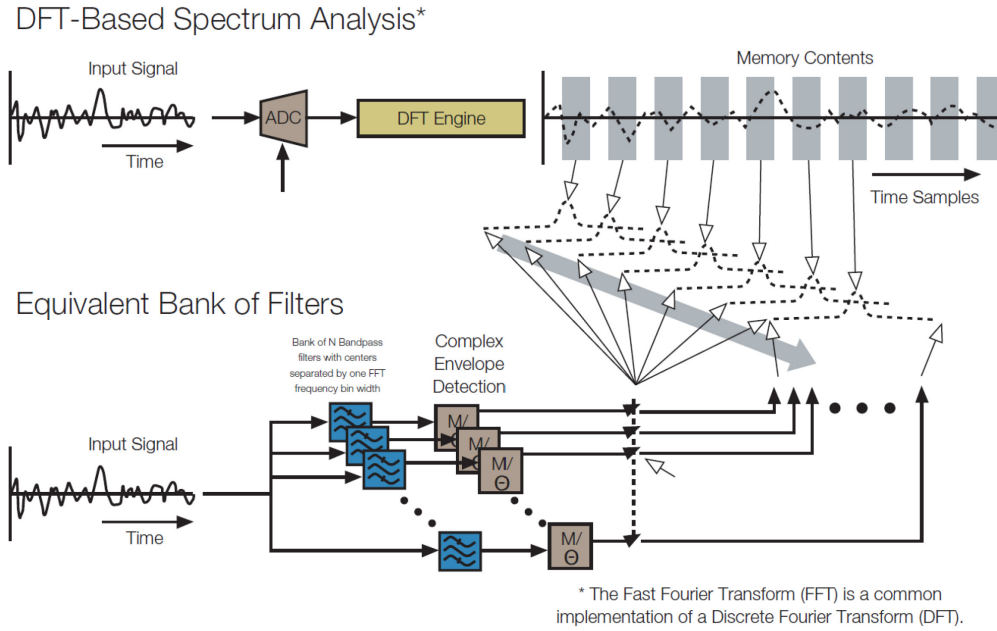


Figure 3.12 – A DFT-based SA and an equivalent implementation using a bank of bandpass filters [102].

Finally, once the signal has been processed by the real-time engine, the relevant waveform segments can be stored in memory and passed to the subsequent post-capture processing and display blocks. The processed data can then be represented in different time-correlated forms. In particular, a spectrogram may be used to display complete spectra as a function of time, with frequency plotted along the horizontal axis, time along the vertical axis, and amplitude represented by color (Figure 3.13).

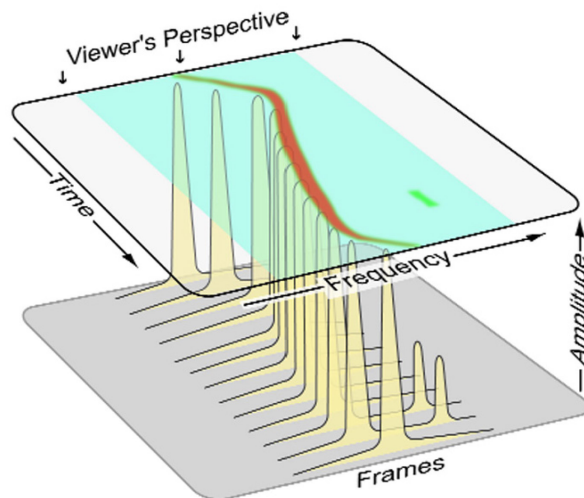


Figure 3.13 – Spectrogram.

In conclusion, a brief description of the parameters that characterize the performance of an RTSA is provided below:

- **Frequency range:** the frequency range defines the interval of input frequencies over which the instrument can operate. It identifies the lowest and highest RF frequencies that can be analyzed by the RTSA;
- **Real-time bandwidth (RTBW):** it is the maximum bandwidth that the instrument can acquire and analyze simultaneously in real time. It is one of the most distinctive parameters of an RTSA, since it determines the portion of spectrum that can be observed continuously without sweeping;
- **Probability of Intercept (POI):** this parameter quantifies the ability of the instrument to detect short-duration transient events. It is usually specified as the minimum event duration that can be captured with 100% probability within the real-time bandwidth;
- **RBW:** similarly to what has been discussed for swept-tuned SAs, the RBW defines the frequency resolution of the analysis, that is, the ability to distinguish two spectral components that are close in frequency. For an RTSA, the RBW is inversely proportional to the acquisition time;
- **Displayed Average Noise Level (DANL):** DANL indicates the average noise level displayed by the instrument when no input signal is present. It defines the minimum signal level that can be distinguished from the analyzer internal noise;
- **Dynamic range:** expresses the ability of the instrument to measure weak and strong signals simultaneously without excessive distortion, compression, or masking effects caused by internal spurious responses or noise;
- **Streaming capability:** refers to the ability of the instrument to transfer or record acquired data continuously for long observation times. This is important when long-duration events or signal statistics must be analyzed;
- **Amplitude accuracy:** Amplitude accuracy specifies how close the displayed signal level is to the true signal amplitude. It is a key parameter whenever quantitative power measurements are required;
- **Amplitude flatness:** Amplitude flatness describes the uniformity of the amplitude response across the analyzed bandwidth. Good flatness ensures that signals at different frequencies within the same acquisition bandwidth are measured consistently;
- **Phase linearity:** phase linearity indicates how uniformly phase is preserved across the acquisition bandwidth. It is important whenever accurate time-domain reconstruction or modulation analysis is required;

3.2 FR1 measurement campaign

The aim of this section is to describe the measurement protocol adopted for carrying out a broadband measurement campaign aimed at assessing the EMF levels in the FR1 band that can be detected in the urban areas of small- to medium-sized cities.

3.2.1 Case study in Cassino (FR)

The measurements campaign under consideration were carried out in Cassino [104], a small-to-medium city with about 35,000 inhabitants located in Italy, halfway between Rome and Naples. Figure 3.14 shows the map of the city. In the same map, the locations of the base stations (BSs) are indicated by blue markers and labeled as BS_n, whereas the measurement points are marked by green stars and identified by labels denoted as P_n. All the measurement points were chosen in Line of Sight (LOS) condition, and were generally distributed around the BSs. The study was limited to urbanized areas, while external areas with low urbanization have not been included in the study.

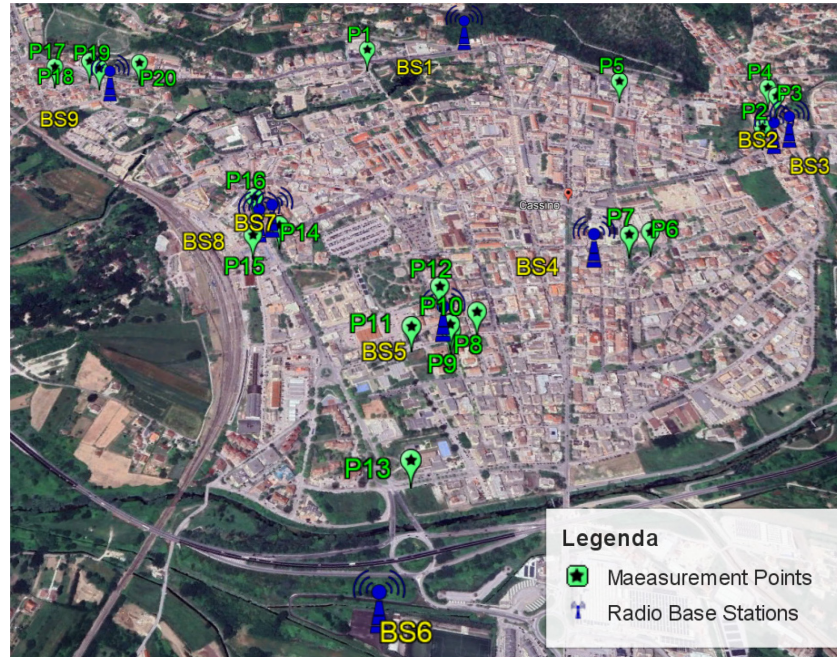


Figure 3.14 – Measurement points (P_n) and base stations (BS_n) locations [104].



Figure 3.15 – Measurement setup [104].

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The measurements were carried out during three different working days, approximately between 9-11am and 3-5pm, during which cellular traffic is generally higher, using the following equipment (Figure 3.15):

- FieldFox N9960B SA;
- ARIA-6000 isotropic triaxial antenna;
- Antenna cable model SDCF02;
- GPS antenna.

The Keysight FieldFox N9960B SA used during the measurement campaign covers a frequency range from 9 kHz to 32 GHz. This band includes all frequencies used by cellular systems, including millimeter waves. However, the ARIA-6000 antenna covers a smaller band, from 30 MHz to 6 GHz. Therefore, the system in the described configuration allowed the measurement of the EMF level in the entire band covered by cellular systems in Italy (Figure 3.16), excluding the mmWave frequency band. This does not limit the analysis since no mmWave base stations are present in the town.

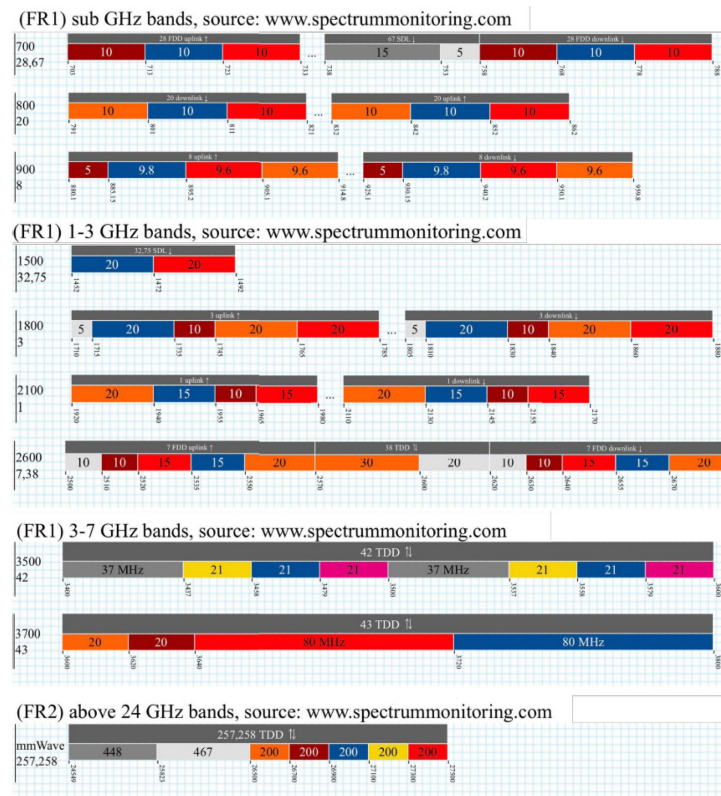


Figure 3.16 – Bands allocation in Italy [105]. Different colors correspond to different operators. Light grey slots are not allocated.

To carry out the measurements, the FieldFox N9960B was set to SA mode with an RMS detector and, in every measurement location, two different recording modes were used: MAX-HOLD mode and AVERAGE mode, considering 15 traces. Moreover, in each measurement position (Figure 3.14), the probe was placed at a 1.8 m above the ground and the EMF level was measured, over 3501 frequency points, in the 500 MHz ÷ 4000 MHz frequency range using an RBW of 100 kHz. These settings were chosen to have a good trade-off between measurement accuracy and measurement time.

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As an example, in Figure 3.17 the power spectrum ($\text{dB } \mu\text{V}\cdot\text{m}^{-1}$) measured in AVERAGE recording mode in the $n = 4$ and $n = 10$ measurement positions is shown with different colors for the three measurement sessions, with the respective CP calculated over the full measurement band. Uplink and downlink frequencies are indicated with light-blue and light-yellow background respectively, whereas light green background represents Time Division Duplex operating mode.

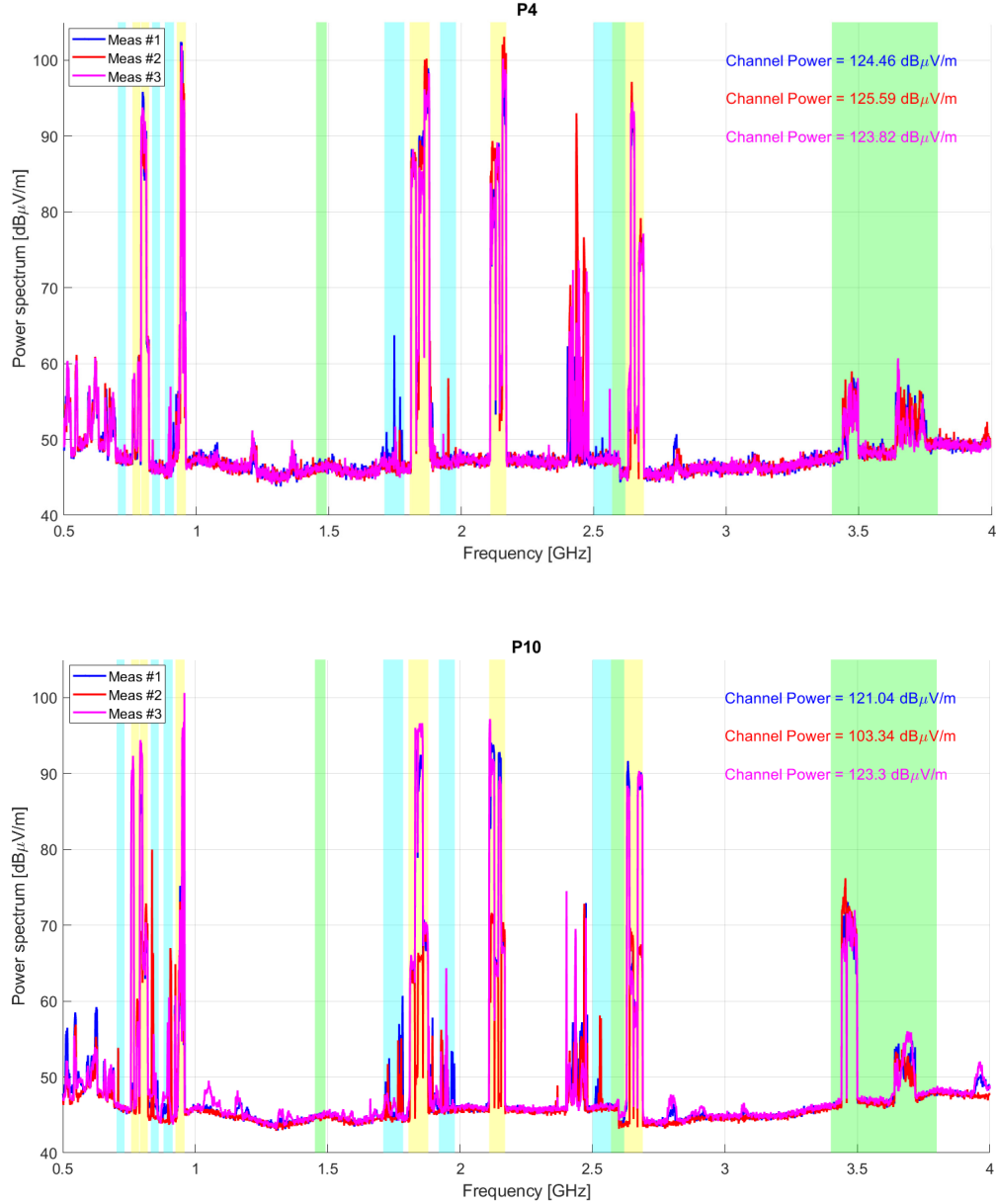


Figure 3.17 – Measured Power Spectrum in AVERAGE mode at locations P4 (top) and P10 (bottom), in different colors for the three measurement sessions. Uplink and downlink frequencies are indicated with light-blue and light-yellow background, respectively, whereas light green background represents TDD [104].

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For each measurement position, the E-field level was obtained from the CP values measured over the entire measurement frequency range after properly compensating the results for cable attenuation (the antenna gain correction was applied directly by the instrument). The E-Field levels at the different measurement positions in MAX-HOLD mode are shown in Figure 3.18, whereas Figure 3.19 shows the E-Field levels measured at each position in AVERAGE mode.

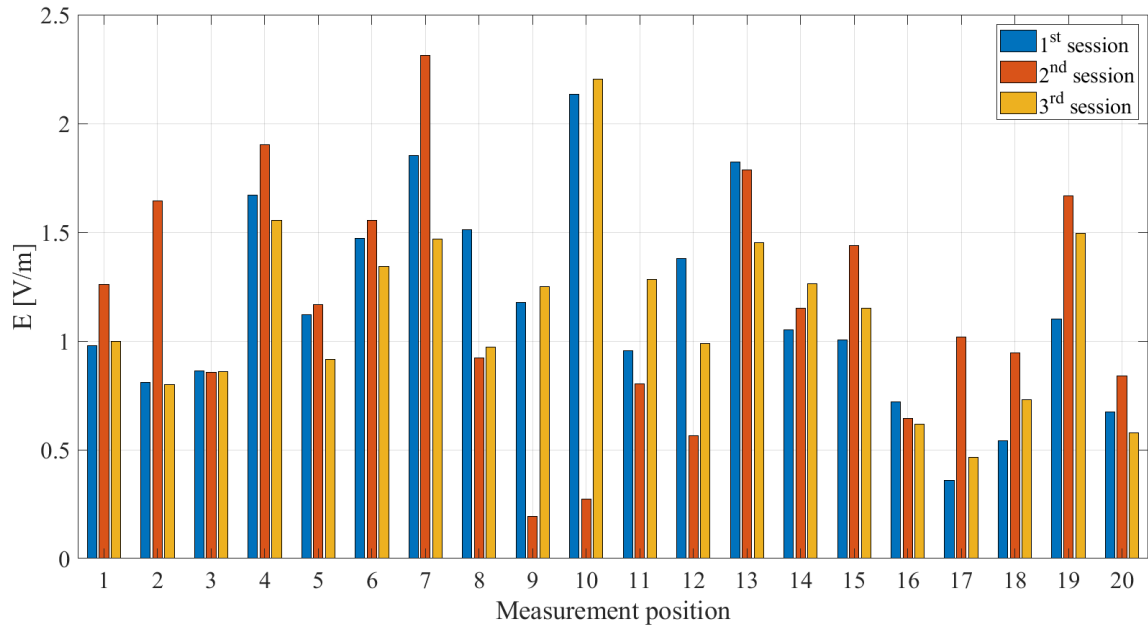


Figure 3.18 – E-Field level in the 20 measurement positions: MAXHOLD recording mode trace. The three colors represent different measurement sessions [104].

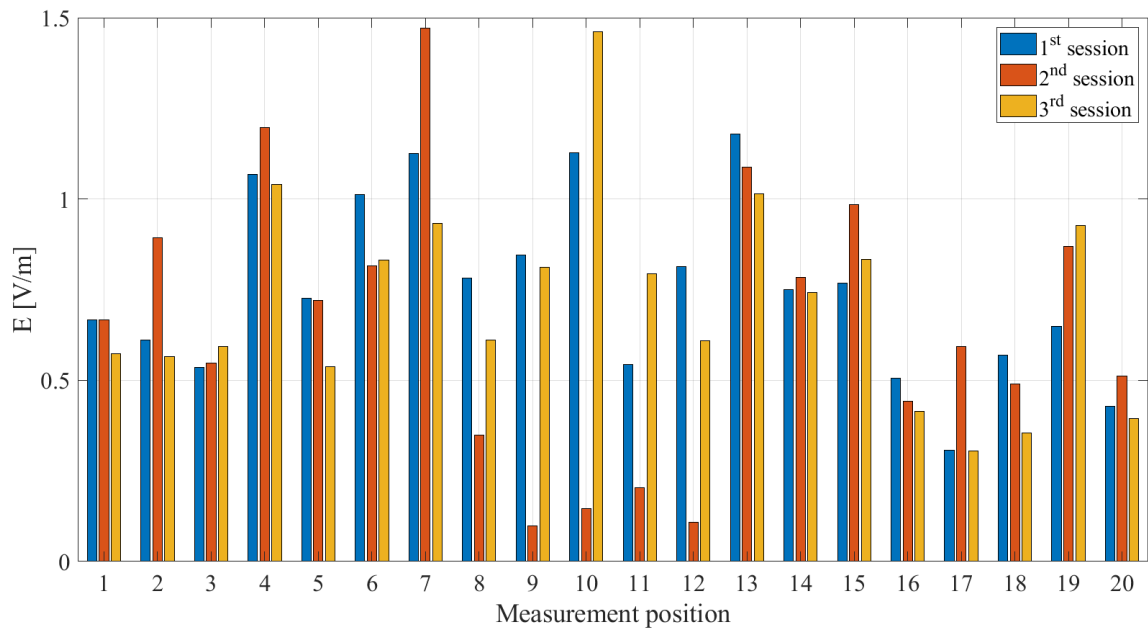


Figure 3.19 – E-Field level in the 20 measurement positions: AVERAGE recording mode trace. The three colors represent different measurement sessions [104].

3. Reference measurement campaigns

The above figures show that the EMF level exhibits limited variation. The mean field value measured at the different positions and over all measurement sessions is $1.14 \text{ V}\cdot\text{m}^{-1}$ in MAX-HOLD mode, with a standard deviation of $0.47 \text{ V}\cdot\text{m}^{-1}$. In AVERAGE mode, the mean value decreases to $0.71 \text{ V}\cdot\text{m}^{-1}$, with a standard deviation of $0.30 \text{ V}\cdot\text{m}^{-1}$. The maximum field value was recorded at position 7 and was equal to $2.31 \text{ V}\cdot\text{m}^{-1}$ in MAX-HOLD mode and $1.47 \text{ V}\cdot\text{m}^{-1}$ in AVERAGE mode. These values are summarized in Table 3.1. It is worth mentioning that at the time all the measures were carried out, Italian law provided for a maximum value of the electric field equal to 20 V/m and a more restrictive “attention value” equal to 6 V/m (rms values, averaged over 6 minutes and 24 hours).

E-Field	MAX-HOLD	AVERAGE
Max	$2.31 \text{ V}\cdot\text{m}^{-1}$	$1.47 \text{ V}\cdot\text{m}^{-1}$
Mean	$1.14 \text{ V}\cdot\text{m}^{-1}$	$0.71 \text{ V}\cdot\text{m}^{-1}$
Standard deviation	$0.47 \text{ V}\cdot\text{m}^{-1}$	$0.30 \text{ V}\cdot\text{m}^{-1}$

Table 3.1 – E-Field statistics over all locations and sessions.

It is also interesting to observe that the field values of the first and third measurement campaign are very close, whereas the campaign #2 exhibits very low values in locations P8 ÷ P12. Such points, shown in Figure 3.14, are all around RB5, clearly indicating that during the 2nd measurement campaign that particular BS was not fully operative. A possible explanation for this result is that the measurement campaign #2 has been carried out on a working day close to Easter, when lessons in schools and universities were suspended: consequently, less traffic demand is expected.

The distribution of the measured EMF level was also used to have a rough estimation of the degree of deployment of the different generations of the cellular systems in the city by evaluating the field level within the frequency bands associated with each generation. In this regard, with reference to Figure 3.16, it is useful to note that in Italy the different generations are primarily allocated in the following bands:

- GSM is primarily allocated at 900 MHz (Band 8), whereas 1800 MHz (Band 3) has been almost completely reallocated to 4G;
- 3G (UMTS) is allocated in 2100 MHz (Band 1); however, since 3G cells are being shut down the spectrum is gradually reassigned to 4G;
- 4G (LTE) primarily uses 1800 MHz (Band 3), 2600 MHz (Band 7), 2100 MHz (Band 1) and 800 MHz (Band 20). 4G frequencies are also used for 5G Dynamic Spectrum Sharing (DSS) [12], that insert the 5G into the 4G frame;
- 5G is allocated in $3.6 \div 3.8 \text{ GHz}$ (Band 43 or n78), 700 MHz (Band 28) and 26 GHz (mmWave, n258), that is used mainly for Fixed Wireless Access (FWA) connections.

These bands can therefore be gathered according to the corresponding technologies to give 2G/3G, 4G and 5G band. It is useful to note that, in addition to the aforementioned bands, frequencies below 700 MHz (labeled as the UHF band), as well as those between 2.4 and 2.5 GHz (ISM band), also fall within the frequency range over which the measurements were carried out. In Figure 3.20, the contribution in each band is reported, both in MAX-HOLD and AVERAGE mode.

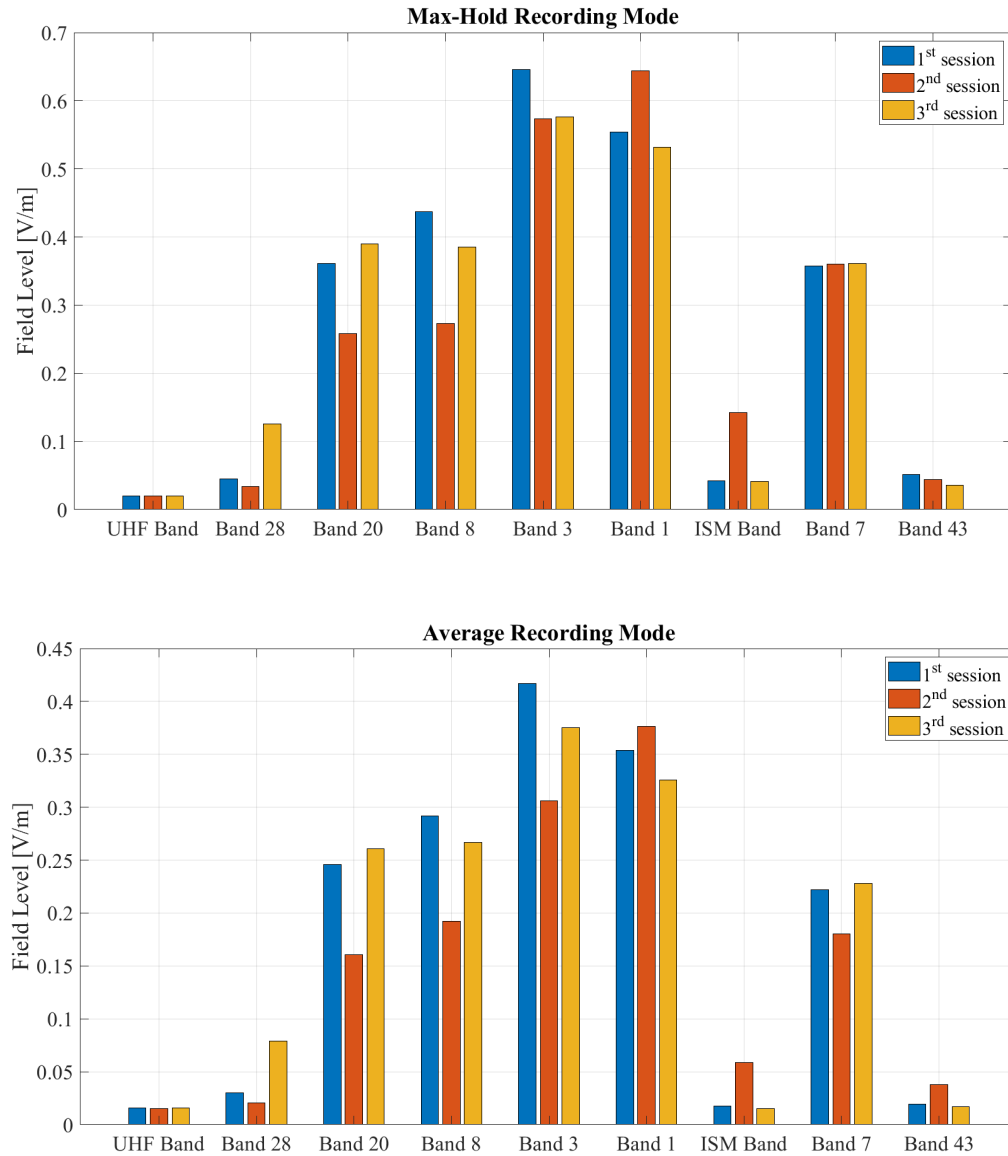


Figure 3.20 – Histogram charts of the field levels in the sub-bands falling within the measurement frequency range for the three measurement sessions. Top: MAX-HOLD mode; bottom: AVERAGE mode [104].

The histograms above show that the difference observed for measurement campaign #2, as outlined above, is limited to bands 3, 8 and 20. By contrast, the contribution of the ISM band was larger in measurement campaign #2: this could again be related to the particular day on which the measurements were carried out, when students and many other people were probably at home and using Wi-Fi more intensively than on other days. Moreover, as mentioned before, the measurement band was divided into three sub-bands, according to the different generations of cellular systems. The field level in these three sub-bands is mainly used respectively by 2G/3G, 4G and 5G NSA/SA. Figure 3.21 shows the percentage of the field level in the three sub-bands in the 20 measurement points, both in MAX-HOLD and AVERAGE modes.

3. Reference measurement campaigns

Figure 3.21 shows that 4G provides the most important contribution at the field level, while the impact of 5G is limited, showing that the implementation of 5G is significantly slower in smaller than in larger cities.

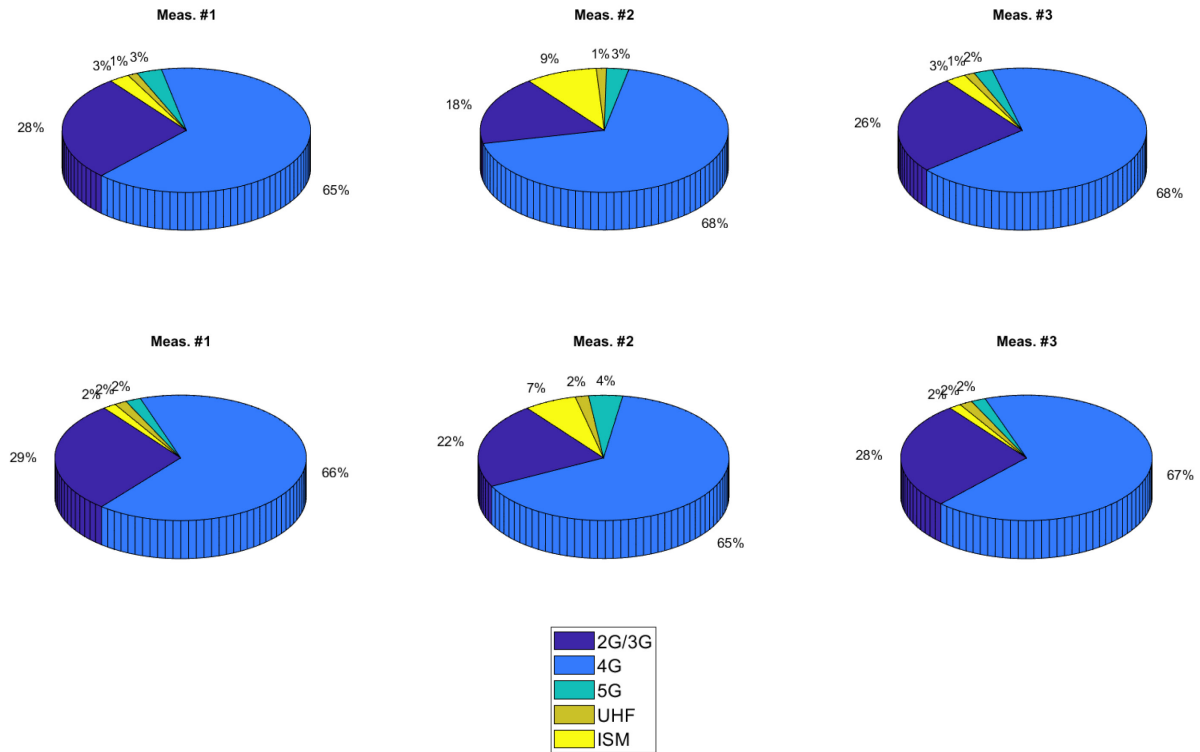


Figure 3.21 – Pie charts showing the percentage contributions of the different technologies. Top: MAX-HOLD mode; bottom: AVERAGE mode [104].

It should be stressed that the results discussed in this section refer only to the sub-6 GHz frequency range because no FR2/mmWave BSs were present in Cassino at the time of the measurement campaign. Nevertheless, carrying out measurements in the FR2 band remains relevant to this work, since the measurement of EMF levels associated with 5G FR2 signals poses specific challenges related to beamforming, traffic beam activation, and the strong dependence of the measured field level on the operating conditions of the network. Therefore, a dedicated FR2 measurement campaign was carried out at a different site where a suitable 5G NR FR2 BS was available, as discussed in the following section.

3.3 FR2 Measurement campaign

The aim of this section is to describe the measurement protocol adopted for carrying out a frequency-selective measurement campaign aimed at assessing the EMF levels related to 5G NR signals in the FR2 band under controlled operating conditions. The same protocol is expected to be suitable for in-situ calibration of the final device, once the 6G network will become operational.

3.3.1 Case study in Front (TO)

The measurements campaign under consideration were carried out in Front–Turin, Italy, in the vicinity of a commercial pilot 5G NR FR2 BS operating (Figure 3.22b) at 27 GHz. Figure 3.22a shows the measurement positions and the location of the BS within the measurement site. All measurements were conducted under LOS conditions. Furthermore, the baseline position, located in the parking area behind the Municipality of Front, was at a straight-line distance of 343 m from the BS, whereas positions A, B, C, and D were located at straight-line distances of 232 m, 212 m, 176 m, and 97 m from the BS, respectively.

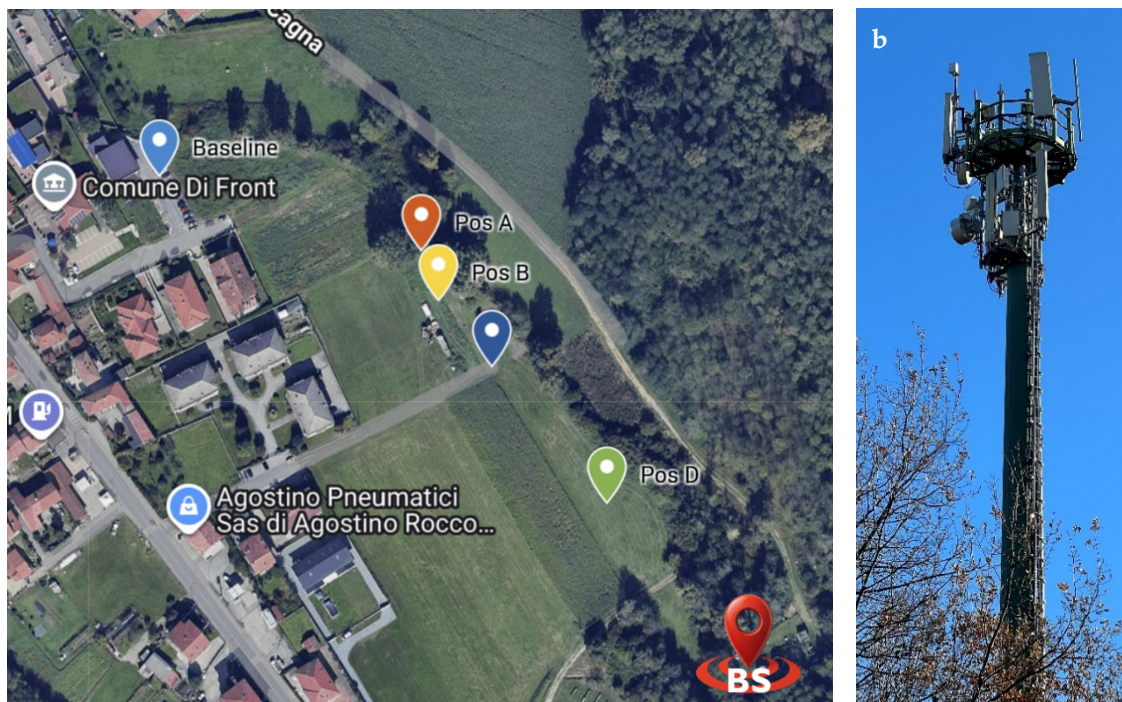


Figure 3.22 – Google maps screenshot of the measurement area (a) and a picture of the FR2 Base station (b).

The measurements were carried out using the following equipment (Figure 3.23):

- Keysight FieldFox N9960B SA;
- Steatite QOM-26-40-K-SG-R vertically polarized omnidirectional antenna, with a gain of 2.8 dBi at 27 GHz (from datasheet);
- Keysight N9927-60025 2.92 mm RF-cable, exhibiting an insertion loss of 2 dB at 27GHz;
- non-conductive tripod;
- two customer premises equipment (CPE) devices
- GPS antenna.

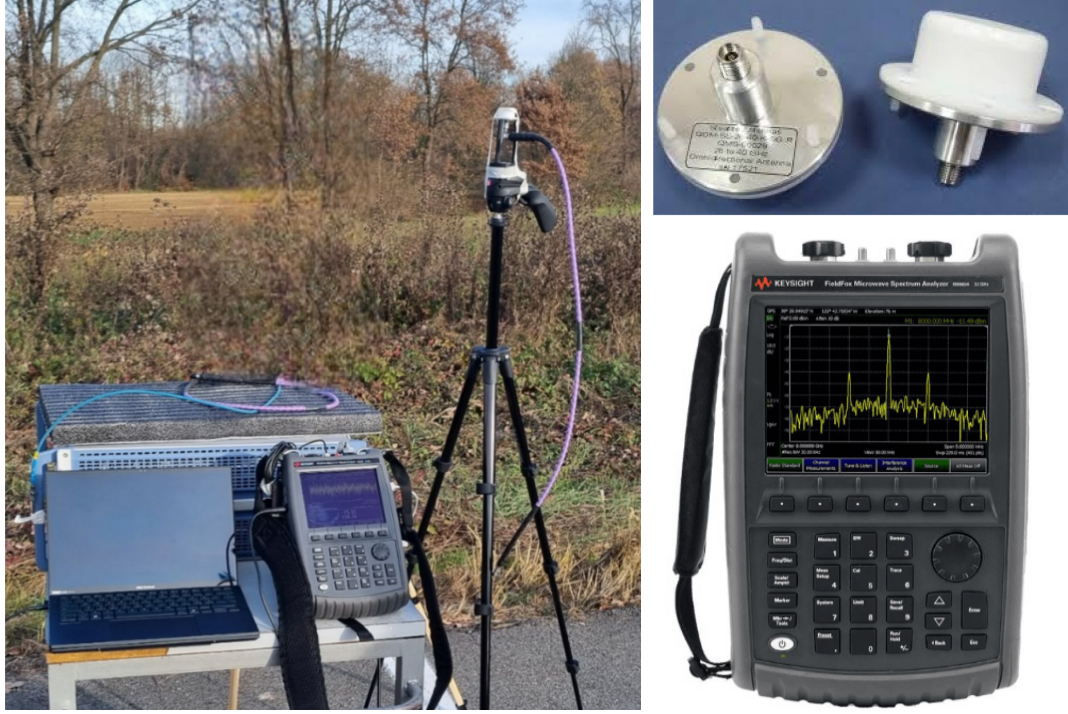


Figure 3.23 – Measurement setup used in the FR2 measurement campaign.

The following measurement protocol was adopted:

- the Keysight FieldFox N9960B was operated in SA mode with an RMS detector;
- at each measurement position, the antenna was mounted on the tripod and kept at a height of 1.8 m above the ground;
- the CPEs have been used to emulate active user connections by generating iPerf traffic, so as to force sustained high data throughput and thereby force the traffic toward themselves;
- one of the two CPEs (namely, CPE₁) was fixed at the baseline measurement position;
- the other CPE device (namely, CPE₂) was placed at each measurement location, approximately 2 m behind the measurement setup;
- at each measurement position, a predefined activation sequence was applied: first, only CPE₂ was activated while CPE₁ remained inactive; subsequently, both CPE₁ and CPE₂ were activated simultaneously; finally, only CPE₁ remained active. Each activation stage lasted 30 s;
- at each measurement position and for each activation stage, the power spectrum (dBm) was continuously acquired over 1001 frequency points in the 26.75 GHz ÷ 27.25 GHz frequency range, using an RBW of 100 kHz. In addition, for each sweep, the CP over a 200 MHz bandwidth (26.9 GHz ÷ 27.1 GHz) was automatically evaluated by the instrument.

The choice to adopt the protocol described above to carry out the measurement campaign, and in particular to use CPEs in order to emulate an active user connection, was motivated by the distinctive characteristics of 5G networks. Indeed, as discussed in Chapter 1, 5G systems, particularly those operating in FR2, rely on AAS capable of radiating two distinct types of beams, namely the broadcast beam and the traffic beam, which differ in their directivity [106] ÷ [108]. More specifically, the traffic

beam is used to transmit payload data to the UE and exhibits greater directivity than the broadcast beam, which instead carries information intended for all users within the cell, such as control signals. Consequently, even if the control signals and payload data are transmitted at the same power, they are received at different power levels, resulting in different field levels [106]. Furthermore, most 5G BS employ directive antennas whose beams can be steered towards the user through beamforming techniques [109]. In the most widely adopted approach, a grid of beams with different directions and gains is used to cover the sector of interest. These characteristics pose significant challenges for the measurement of RF-EMF levels associated with 5G signals in the FR2 band. In particular, in the absence of active UEs, user-specific traffic beams are generally not transmitted, and only cell-defining common signals are radiated, primarily through SSB beam sweeping. In these conditions, passive CP measurements, i.e., measurements performed without any interaction with the BS, would not provide useful results for estimating either the power of the individual SSBs or the RF-EMF levels associated with active-user operation. Rather, such measurements would only provide a time- and frequency-averaged estimate of the channel power under idle conditions, when only always-on/common signals are present. In particular, if the whole NR channel were used as the integration bandwidth for CP calculation, the power associated with the individual SSBs would be averaged over the entire channel and could therefore not be isolated from the CP result. Moreover, the repeatability of the measurements might be limited by the bursty nature of SSB transmissions. Better results for SSB characterization could be obtained by means of zero-span measurements or by using a real-time RTSA. In fact, as discussed in the previous sections, unlike conventional CP measurements performed over the entire channel bandwidth, zero-span measurements make it possible to observe the received power as a function of time at a fixed frequency corresponding to the SSB, thereby enabling the identification of the individual SSBs within the SS burst and of the strongest one among them. Similarly, an RTSA can provide a more reliable characterization of SSB transmissions thanks to its real-time acquisition capability, which makes it possible to detect intermittent signals. However, although these approaches are better suited for SSB characterization, the resulting measurements would still be only partially representative of realistic operating conditions with active UEs, since they would mainly characterize cell-defining common-signal emissions rather than user-specific traffic beams.

Passive CP measurements would also provide results of limited representativeness in the presence of multiple active users. This is because, under these conditions, the BS would dynamically distribute the available resources among the active users, steering the beams according to the spatial distribution of the users requiring traffic at a given time. As a result, the measured EMF levels would strongly depend on the specific traffic and scheduling conditions occurring during the measurement session and would therefore exhibit limited repeatability.

In order to address the issues described above, a possible approach is to perform active measurements, in accordance with the measurement protocols adopted for the campaign described in this section, by forcing the 5G BS to allocate downlink traffic to a properly configured device located close to the measurement position [106][108]. In this way, the BS can be forced to steer a traffic beam towards the measurement positions and the measured field can be regarded as a close estimate of the maximum exposure achievable at the measurement point when all the resource are given to only one user [106]. Consistently with the previous discussion on resource sharing in the presence of multiple active users, the field value obtained in this way should be regarded as a conservative upper bound

3. Reference measurement campaigns

rather than as a representative estimate of the exposure conditions occurring in a realistic user scenarios. Nevertheless, this type of measurement may provide the basis for maximum-power extrapolation procedures, in which suitable correction factors are introduced to account for the statistical variability of traffic and resource allocation, thereby enabling a plausible estimate of the exposure levels expected under realistic operating conditions [106]–[109].

Moreover, the different directivities of the broadcast beam and the traffic beam radiated by a 5G BS should also be taken into account when selecting the receiving antenna whenever, unlike in the measurement campaign described here, an omnidirectional antenna cannot be used. Because the traffic beam is highly directive and therefore characterized by a narrow angular width, low- to moderate-gain receiving antennas are generally preferable in such cases, as their wider beamwidth makes the measurement results less sensitive to the exact antenna pointing. By contrast, the use of high-gain antennas, such as horn antennas, would make the alignment much more critical and could reduce the repeatability of the measurements, even under controlled traffic conditions, such as those achievable by following the measurement protocol described above.

Finally, it should be observed that LOS conditions are highly desirable for this type of measurement. At FR2 frequencies, propagation is highly sensitive to shadowing effects, and NLOS conditions may cause the received field to be dominated by reflected or alternative propagation paths. LOS conditions, on the other hand, make it possible to relate the measured field more directly to the traffic beam directed towards the measurement point.

Having briefly outlined the rationale behind the protocol adopted for carrying out the measurements discussed in this work, it is now useful to briefly analyse the results obtained.

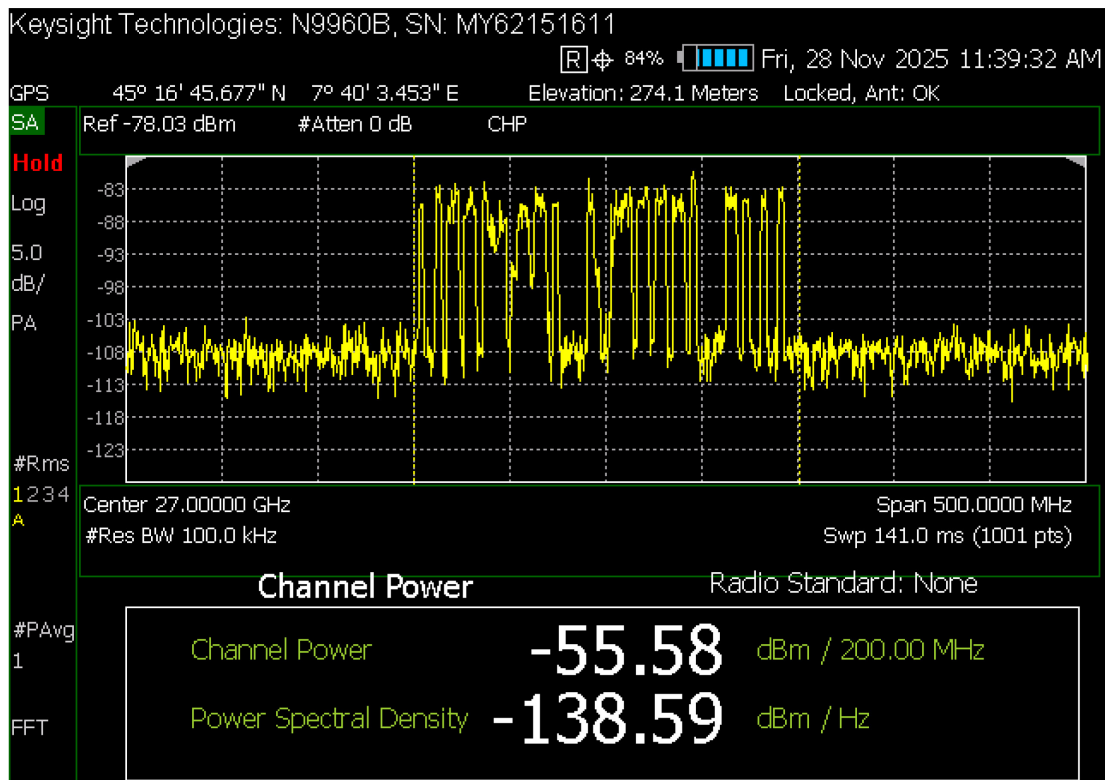


Figure 3.24 – Example of a sweep acquired in the FR2 measurement campaign.

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As discussed above, at each measurement position the power spectrum (dBm) was continuously acquired over 1001 frequency points within the 26.75 GHz ÷ 27.25 GHz frequency range, using an RBW of 100 kHz and following a predefined activation sequence. First, only CPE₂ was activated while CPE₁ remained inactive; subsequently, both CPE₁ and CPE₂ were activated simultaneously; finally, only CPE₁ remained active. Each activation stage lasted 30 s. In addition, for each sweep, the CP over a 200 MHz bandwidth (26.9 GHz ÷ 27.1 GHz) was automatically evaluated by the instrument. Figure 3.24 shows, as an example, a screenshot of the FieldFox display corresponding to one of the multiple sweep acquisitions performed at measurement position A during the 30 s stage in which both CPEs were active.

The power spectra measured at each measurement position during each acquisition performed by the instrument were stored in the instrument memory for subsequent post-processing. In particular, the obtained results were appropriately compensated to account for the antenna gain and the attenuation introduced by the cable used for the measurements. The results thus obtained were then processed in order to evaluate, for each activation stage, the average and max-hold values of the CP measured over a 200 MHz bandwidth centered at 27 GHz across the multiple sweeps performed during each 30 s stage of the activation sequence (the average sweep time was 145 ms). Figure 3.25, shows, by way of example, the max-hold power spectra and the corresponding CP, evaluated over a 200 MHz bandwidth centered at 27 GHz, obtained at position A for the three different CPE activation stages.



Figure 3.25 – Max-hold results obtained at position A for each activation stage in the FR2 measurement campaign: only CPE₂ activated (blue line), both CPEs active (red line), only CPE₁ active (orange line).

The power measured at the instrument input connector, after correction to account for cable attenuation and antenna gain, was also used to calculate, for each stage of the activation sequence, the average and max-hold values of the E-field strength at the measurement point by means of the following relations:

$$S = \frac{4\pi P_r}{\lambda^2 G} \quad (3.2)$$

$$E = \sqrt{\eta_0 S} \quad (3.3)$$

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where S ($\text{W}\cdot\text{m}^{-2}$) is the incident power density, P_r (W) is the power received from the antenna, G is the antenna gain, λ (m) is the wavelength, E ($\text{V}\cdot\text{m}^{-1}$) is the E-Field strength, and $\eta_0 \cong 377 \Omega$ is the characteristic impedance of free space.

The results obtained at the different measurement positions are summarized in Table 3.2:

Position	Coordinates		CPE	CP (dBm)		E-Field ($\text{V}\cdot\text{m}^{-1}$)	
	Latitude	Longitude		Average	Max-Hold	Average	Max-Hold
Baseline	45°16'46.800"N	7°39'57.989"E	CPE ₁	-59.20	-53.53	0.20	0.38
			both	-59.80	-54.02	0.18	0.36
			CPE ₂	-59.57	-53.69	0.19	0.37
A	45°16'45.698"N	7°40'3.447"E	CPE ₁	-74.62	-70.36	0.03	0.05
			both	-57.86	-50.08	0.23	0.56
			CPE ₂	-55.42	-49.68	0.30	0.59
B	45°16'44.980"N	7°40'3.775"E	CPE ₁	-73.54	-68.62	0.04	0.07
			both	-57.39	-49.05	0.24	0.63
			CPE ₂	-54.37	-47.73	0.34	0.73
C	45°16'44.024"N	7°40'4.904"E	CPE ₁	-75.55	-70.33	0.03	0.05
			both	-57.57	-49.73	0.24	0.58
			CPE ₂	-55.28	-49.33	0.31	0.61
D	45°16'42.052"N	7°40'7.276"E	CPE ₁	-67.64	-61.98	0.07	0.14
			both	-58.74	-45.99	0.21	0.90
			CPE ₂	-60.15	-49.88	0.18	0.57

Table 3.2 – Results of the FR2 measurement campaign conducted in Front (TO).

As shown in Table 3.2, at the baseline position the CP values measured across the three stages of the CPE activation sequence, both average and max-hold, are very similar. This was expected since, according to the protocol previously described, at the baseline position the measurements were carried out with both CPEs placed in the same area as the measurement device. For this reason, throughout the entire CPE activation sequence, at the baseline position the traffic beams were always directed towards the measurement device. Moreover, at positions A to D, the measured CP values, both average and max-hold, were very low and close to the background noise level when only CPE₁ was active. This was because, under this condition, the beam was directed towards the baseline position and was therefore not detected by the antenna located at positions A, B, C, and D.

By contrast, when only CPE₂ was active, the average CP values measured at positions A, B, and C were unexpectedly very similar. Indeed, as the distance from the antenna decreased, an increase in the average power level would have been expected. One possible explanation for the observed behavior is that the beam used by the antenna to serve positions A, B, and C was not always the same. In this case, even small variations in the gain of the selected beam would have been sufficient to offset the reduction in free-space path loss expected when moving from a distance of 232 m (position A) to 176 m (position C). At measurement position D, when only CPE₂ was active, the measured CP value was approximately 5-6 dB lower than those recorded at the other positions. In this case, the observed behavior may be explained by the antenna's limited ability to steer the beam towards the

measurement equipment. Indeed, point D was the closest to the antenna. Given the considerable height of the antenna tower, the downtilt required to direct the beam towards the measurement position may have exceeded the steering capability of the antenna.

To conclude the analysis of the results, it is interesting to observe that, at all measurement positions, when both CPEs were active, the measured CP average values were approximately 2.5-3 dB lower than those measured at the same positions when only CPE₂ was active. This is consistent with the expected beam behavior when both CPEs were active. Under these conditions, the antenna had to share the available resources approximately equally between the CPE located in the measurement area and the one placed at the baseline position, which, as observed, resulted in a reduction of the average CP value of about 3 dB.

The measurement protocol discussed in this section therefore proved to be suitable for the assessment of EMF levels associated with 5G FR2 signals under controlled operating conditions. However, the EMF levels measured according to the discussed procedure represent a conservative upper bound for the evaluation of human exposure to RF-EMF in the FR2 band and cannot be regarded as representative of the exposure conditions that could occur in a realistic scenario for the users.

This further highlights the usefulness of having a device such as the one discussed in the present research project. Indeed, it is reasonable to assume that the assessment of field levels associated with forthcoming 6G signals will pose challenges similar to, if not greater than, those already encountered by researchers in the evaluation of field levels associated with 5G signals. From this perspective, the availability of a wearable exposimeter, characterized by a low purchase cost and capable of being used for long measurement sessions without interfering with users' normal daily activities, could enable the execution of unconventional studies aimed at determining long-term average exposure levels under realistic conditions. In this context, the measurement technique just presented, with the necessary modifications that may be required by the specific features of future 6G networks, could prove to be highly valuable for carrying out an in-situ calibration of the device of interest in this work.

4 Chapter 4: Proof of concept design of a wearable device for 6G FR3 signals level assessments

This chapter describes the design choices adopted to fulfil the primary objective of the present research project. As repeatedly emphasized in the preceding sections, this objective is to provide a proof of concept for the realization, through the use of low-cost components, of a wearable device capable of assessing exposure levels to forthcoming 6G signals operating in the FR3 band.

This choice is motivated by the intention to anticipate future needs by making available a device capable of characterizing exposure to 6G signals once 6G communication networks are effectively deployed. This is particularly desirable in order to avoid issues similar to those that, at least in the early stages, hindered the diffusion of 5G technologies. Indeed, one of the factors that initially limited the widespread adoption of 5G technologies was the lack of adequate tools, both from a mathematical and from a hardware standpoint, for measuring 5G signals in real propagation environments.

It is worth stressing that the device investigated in the present work, being required to satisfy constraints in terms of compactness and cost-effectiveness, is not intended to represent an alternative to certified instrumentation for EMF level measurements. Rather, it should be regarded as an additional tool that may be employed, for example, in non-conventional studies aimed at collecting, on a large scale and over extended time periods, data concerning the levels of 6G signals to which users will be exposed under ordinary daily-life conditions. Studies of this kind, especially when combined with statistical techniques for the analysis of the acquired data, could, for instance, be used to investigate how the average exposure level to 6G signals varies as a function of the age of the individuals involved, their geographical area of residence, or their occupational category. Such studies can realistically be conducted only by means of devices that are both unobtrusive, given the need to perform analyses over prolonged periods, and inexpensive, in view of the high likelihood of damage or failure resulting from possible user negligence.

Anticipating future needs, however, also entails certain drawbacks. Among these, most significant factor affecting the achievement of the project's primary objective is that 6G signal standards have not yet been fully defined. This has necessarily required the formulation of a priori assumptions regarding the possible characteristics of 6G signals, in order to identify the minimum technical specifications that the device of interest must satisfy.

For this reason, before describing the antenna design phase and the hardware components that could be employed for the realization of the device under investigation, it is appropriate to briefly recall the characteristics that the device must possess in order to fulfil its intended function.

4.1 Main design requirements and constraints of the proposed device

In Chapter 2 the main types of custom devices commonly employed in research for EMF level measurements were discussed. Among these, the class of devices of greatest interest for pursuing the primary objectives of this project is that of personal exposimeters.



Figure 4.1 – Simplified Block Diagram of a personal exposimeter.

Indeed, as shown in Figure 4.1, these devices are characterized by a relatively simple architecture, which is well suited to meeting the low-cost requirement of the device considered in the present work. This is because the remarkable development of consumer electronics over the last few decades has made it possible to find microcontrollers and ADCs on the market that are characterized by good, if not excellent, performance at very low cost. Clearly, the low cost of microcontrollers and ADCs alone is not sufficient to ensure that the overall cost of the device of interest remains low; to achieve this objective, the other key components, such as antennas and RF detectors, must also exhibit purchase or fabrication costs that are as low as possible, while still providing performance levels consistent with the fulfilment of the project's primary objectives. For this reason, as will be better clarified in the following, the design choices adopted in designing the antennas to be integrated into the device of interest were guided not only by the performance targets to be achieved, but also on the manufacturing technology required for their fabrication, since the latter has a significant impact on the final cost of the device.

Low implementation cost, however, is not the only requirement that the device under consideration must satisfy. Another important requirement is compactness. Indeed, for the device of interest to be truly wearable, its dimensions must be such as not to interfere excessively with the normal daily activities of potential users.

Furthermore, still with a view to fulfilling the primary objectives of the present work, the characteristics of the antennas and of the hardware components to be employed in the realization of the device of interest must be such as to enable the detection of 6G signals with operating frequencies between 14.8 GHz and 15.35 GHz. As discussed in Section 1.5, this frequency range is indeed one of the candidate bands currently under study in preparation for WRC-27. With respect to the latter point, however, it should be emphasized that the operating frequency is only one of the aspects that must be taken into account in order to properly design the device under investigation. Another very important parameter, for example, is the channel bandwidth that will be used for the transmission of the signals of interest. This parameter affects the choice of antenna type, since effective detection requires adequate gain and impedance matching over a bandwidth at least as wide as that of the signals to be detected. Moreover, as discussed in the previous sections, the functional characteristics of the BSs that will be used for the transmission of 6G signals must also be taken into account when defining the characteristics of the antennas to be employed in order to fulfil the primary objectives of this research

project. For example, as discussed in Section 3.3.1, the distinctive characteristics of the antenna systems used for the transmission of 5G signals make low- to medium-gain antennas the most suitable option for measuring the EMF levels associated with 5G signals

With reference to the above, however, it should be pointed out that, at present, standards concerning the characteristics of 6G signals and the technologies that will be used to implement the new generation of cellular communication networks have not yet been fully defined. In order to proceed with the definition of the minimum technical specifications of the antennas to be used for the development of the device under investigation, it was therefore necessary to formulate a priori assumptions regarding the possible characteristics of future 6G signals. In particular, assuming that the main technological innovations introduced with 5G — e.g., beamforming, beam steering, and use of highly directive traffic beams in order to maximize the signal-to-noise ratio — will also be retained in the future generation of cellular communication networks, it was decided to refer to the distinctive characteristics of 5G FR2 signals in defining the minimum technical specifications, in terms of gain and bandwidth, of the antennas to be used for the realization of the device of interest. For this reason, in defining the design specifications, it was established that the antennas with which the device of interest is to be equipped should exhibit low- to medium- gain values (namely, gain below 10 dBi [110]) and should be capable of ensuring adequate reception of signals with a bandwidth of at least 200 MHz.

A further design specification concerning the functional characteristics of the antennas to be employed in the realization of the device of interest relates to their polarization. It is therefore useful to briefly review the main concepts related to antenna polarization.

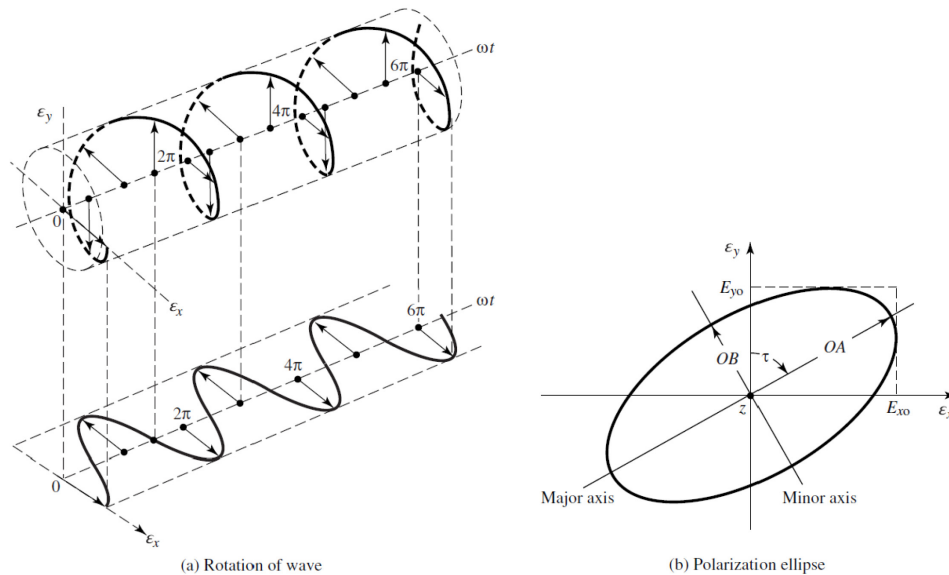


Figure 4.2 – Rotation of a plane EM wave and its polarization ellipse at $z = 0$ as a function of time [111].

The polarization of an antenna is usually defined in terms of the polarization of the wave transmitted or received by the antenna in a given direction, generally taken to be the direction of maximum gain [111].

More precisely, the polarization of an EM wave radiated by an antenna is defined as the property that describes the time-varying direction and relative magnitude of the E-field component of the EMF [111]. From a conceptual point of view, therefore, the polarization of an EM wave may be regarded as the figure traced, as a function of time and at a fixed point in space, by the projection of the tip of the vector representing the E-field onto the plane perpendicular to the direction of wave propagation.

In general, as can be observed from Figure 4.2(a) and (b), the polarization of a wave is elliptical. There are, however, two particular cases of elliptical polarization: linear polarization and circular polarization. Linear polarization occurs when the magnitude of the vector representing the E-field varies over time, while its direction remains constant; in this case, the projection of the tip of the vector representing the E-field onto the plane perpendicular to the direction of propagation traces a straight line. In other words, a linearly polarized wave is an EM wave in which the amplitude of the electric component of the EMF (and, consequently, also that of the magnetic component) varies sinusoidally with time, whereas the direction along which this variation occurs remains constant. Circular polarization, on the other hand, occurs when the magnitude of the vector representing the E-field at a given point in space remains constant over time, while its direction varies in such a way that the projection of its tip onto the plane perpendicular to the direction of propagation traces a circle. For a circularly polarized field, therefore, the E-field and the H-field rotate over time at each point in space while maintaining constant magnitude. The polarization is said to be right-hand circular polarization (RHCP) if the rotation is clockwise when the wave is observed in the direction of propagation; otherwise, it is defined as left-hand circular polarization (LHCP).

The polarization of a wave received by an antenna, on the other hand, is defined as “the polarization of a plane wave, incident from a given direction and having a given power flux density, which results in maximum available power at the antenna terminals” [111]. It should be noted that the two definitions just given are actually entirely equivalent. Indeed, as is well known, according to the reciprocity theorem, the behavior of an antenna in transmission is analogous to that in reception; therefore, the polarization of the incident wave which results in maximum available power at the antenna terminals is that of the field radiated by the antenna itself.

Clearly, the above discussion does not necessarily imply that a mismatch between the polarization of the incident field and that of the antenna precludes signal reception. In fact, with respect to a given receiving antenna, the incident field can always be decomposed into a **co-polarized** component, namely the component compatible with the polarization state of the antenna, and a **cross-polarized** component, namely the one orthogonal to it. In the ideal case, only the former contributes effectively to the power available at the antenna terminals, whereas the latter is not coupled to the antenna. It follows that the power received by the antenna depends on the magnitude of the co-polarized component of the incident field: the smaller this component, the lower the power available at the antenna terminals. Only in the case in which the incident field is entirely cross-polarized with respect to the antenna does the received power vanish.

This aspect is particularly relevant for the device considered in the present work. Since the device is intended to be wearable, its orientation in space cannot be assumed to remain constant during operation. As a result, the relative polarization between the incident field and the receiving antenna may vary significantly under actual exposure conditions. Under these circumstances, the use of circularly polarized antennas is preferable, since it makes the reception behavior less sensitive to

changes in device orientation than would be the case, for example, for linearly polarized antennas. This advantage becomes even more significant if the device is equipped with multiple antennas arranged, for example, to form a bracelet, since such a configuration can further improve the robustness of the received signal under different exposure conditions.

In light of the above discussion, the main design requirements and constraints of the proposed device, and in particular of its antenna and hardware subsystems, can be summarized as follows:

- the device must be characterized by **low implementation cost**;
- the device must satisfy **compactness requirements** consistent with its intended wearable use;
- the antenna and hardware subsystems must enable the detection of signals with operating frequencies in the **14.8 GHz ÷ 15.35 GHz range**;
- the antennas must guarantee adequate reception of signals with a **bandwidth of at least 200 MHz**;
- the antennas must exhibit **low- to medium-gain values**, namely **below 10 dBi**;
- the antennas should be **circularly polarized**, in order to reduce the sensitivity of the received power to variations in device orientation under actual exposure conditions;
- the design choices adopted for the antennas must take into account not only the required electromagnetic performance, but also the manufacturing technology needed for their fabrication, since this has a significant impact on the final cost of the device;
- the other key hardware components, i.e., RF detectors, ADCs and microcontrollers, must be selected among commercially available solutions so as to keep the overall purchase cost as low as possible, while still ensuring performance levels consistent with the primary objectives of the present research project.

In the following sections, the design choices adopted for the antennas to be employed in the realization of the proposed device will be discussed, and the commercially available hardware components selected for the same purpose will be presented. Although these activities were carried out in parallel, they are discussed in separate sections to ensure greater clarity and readability.

4.2 Antenna design

In order to meet the design constraints defined in the previous section, the design of the antennas intended for the device under study was carried out with reference to the class of **microstrip patch antennas** following the workflow summarized in Figure 4.3:

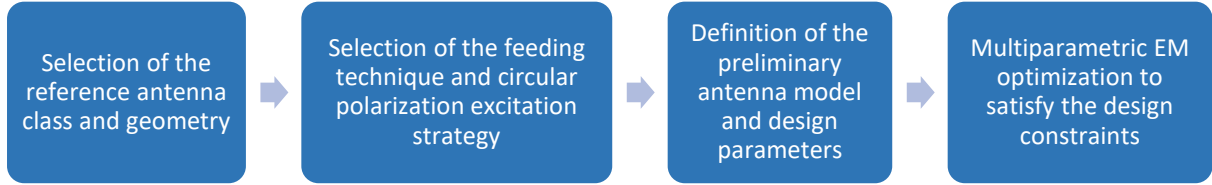


Figure 4.3 – Simplified block diagram of the antenna design workflow.

As schematically shown in Figure 4.4(a) and (b), this class of microstrip patch antennas is characterized by a relatively simple structure consisting of a very thin metallic patch ($t \ll \lambda_0$, where λ_0 denotes the free-space wavelength at the antenna resonant frequency) placed a small fraction of λ_0 ($h \ll \lambda_0$, usually $0.003\lambda_0 \leq h \leq 0.05\lambda_0$) above a metallic ground plane [111]. The strip (patch) and the ground plane are separated by a dielectric layer, referred to as the substrate.

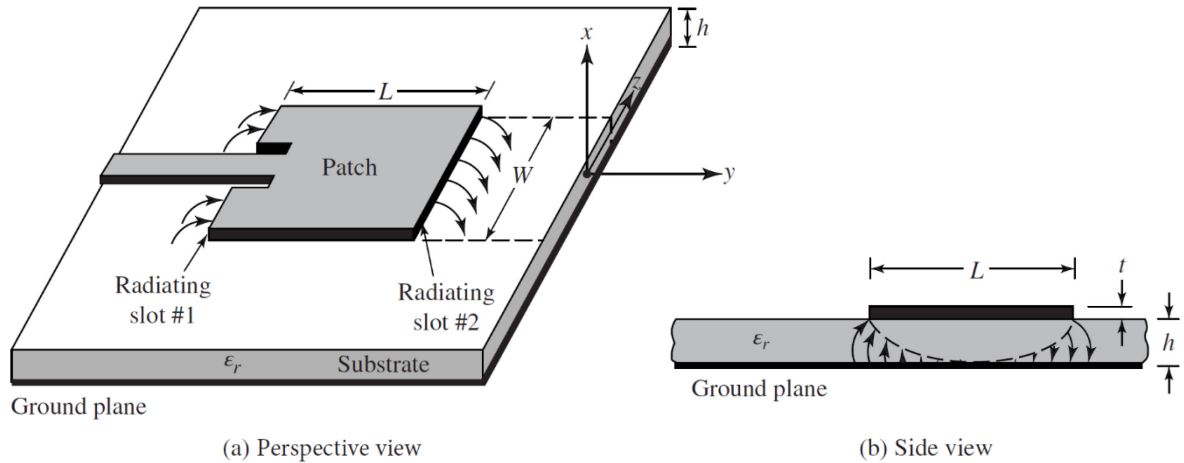


Figure 4.4 – Structure of a microstrip patch antenna [111]: (a) perspective view; (b) side view.

Owing to their simple structure, microstrip patch antennas do not require particularly complex or expensive manufacturing processes. Indeed, they can be fabricated by means of processes closely analogous to those commonly employed for printed circuit boards. More specifically, starting from a copper-clad dielectric laminate, the geometry of the patch and the feeding network is defined by patterning techniques such as photolithography, or equivalent photo-imaging processes applied to a photosensitive resist layer, followed by chemical etching, i.e., the selective removal of the excess copper not protected by the patterned photosensitive layer. Any required drilling, metallization, and surface-finishing steps can then be carried out according to standard PCB manufacturing workflows. This compatibility with printed-circuit technology makes patch antennas manufacturable even by commercial PCB suppliers specialized in prototyping and small-batch production, with a consequent

reduction in production costs. Moreover, owing to the relatively low complexity of the manufacturing processes required for the realization of patch antennas, this class of antennas can be fabricated in a variety of geometric configurations (Figure 4.5).

Therefore, given the degree of freedom in selecting the geometry of the patch antennas to be considered for the future realization of the device of interest, the first step in the antenna design process consisted in the choice of the geometry to adopt as the reference. More specifically, the analysis initially focused on two different antenna configurations: square patch antennas and circular patch antennas.

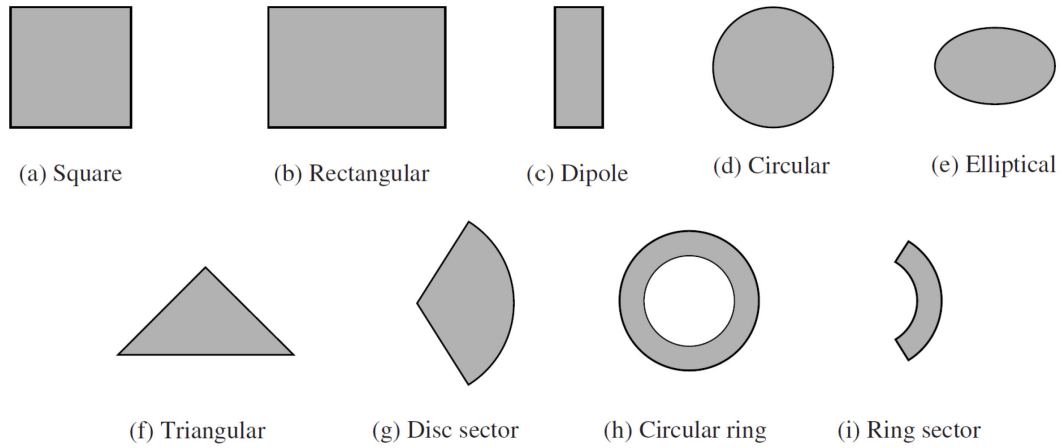


Figure 4.5 – Representative shapes of microstrip patch elements [111].

The attention initially devoted to the two geometric configurations mentioned above is mainly motivated by the fact that these antenna types are among the most widely used in practice and are well characterized from a theoretical standpoint. Consequently, their design can rely on well-established mathematical relations that make it possible, at least as a first approximation, to determine, for example, the characteristic dimensions required for the radiating element to resonate at a given frequency on a substrate with assigned electromagnetic properties. Based on the analyses carried out, a circular geometry was selected as the reference for the design of the antennas to be employed for the future realization of the device of interest. In keeping with the design specifications discussed in the previous section, this choice was motivated by the intention to excite circular polarization by means of a single feed line. Indeed, as shown in Figure 4.5, circular polarization in patch antennas is commonly achieved by means of two orthogonal feed lines excited by equal-amplitude signals in phase quadrature, obtained, for example, through suitable four-port passive reciprocal networks, such as 90° hybrid couplers, which provide equal power division between the two output ports with a prescribed phase difference of $\pi/2$. This design approach essentially stems from the fact that circular polarization can be obtained through the superposition of two orthogonal linearly polarized resonant modes, namely those excited when conventional single-feed configurations are used without modification, provided that they have equal amplitude and a time-phase difference equal to an odd multiple of $\pi/2$ [111].

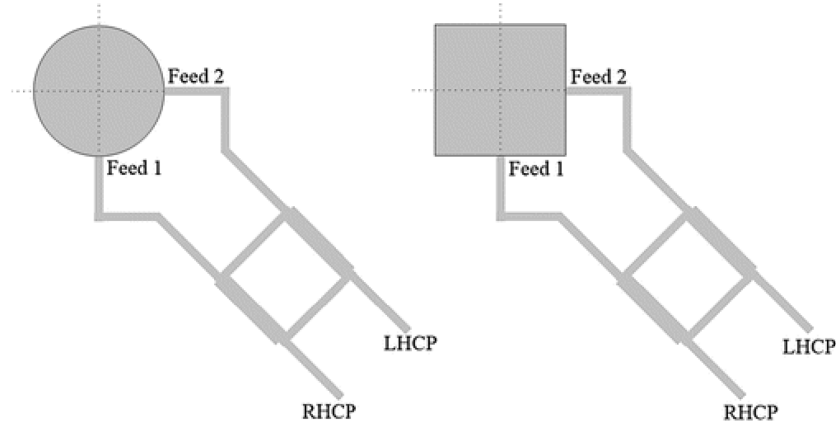


Figure 4.6 – Hybrid coupler used to excite a circular polarization on patch antennas.

However, as mentioned above, for the design of the antennas to be used for the realization of the device of interest, it was decided not to adopt this solution. This choice was motivated both by the need to minimize the overall footprint of the antennas and by the need to avoid additional losses, such as conductor, dielectric, and mismatch losses, that may be introduced by feed networks more complex than those commonly adopted in single-feed configurations. Indeed, the magnitude of such losses depends, among other factors, on the type of substrate used and on the length of the transmission lines required to realize such networks and, in general, increases with frequency. Therefore, especially at high frequencies such as those of interest in this research project, they may ultimately result in a reduction in the overall sensitivity of the measurement chain.

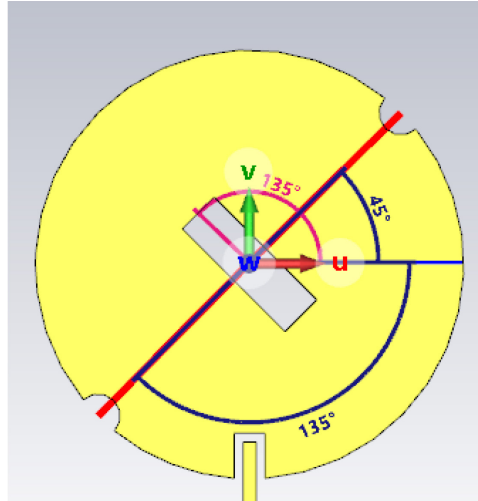


Figure 4.7 – Geometrical perturbations introduced to excite circular polarization in a single-feed circular patch antenna.

More precisely, in designing the antennas to be employed in the device of interest, it was decided to adopt a single microstrip feed line configuration, entrusting the excitation of circular polarization to the introduction of suitable geometric perturbations on the patch. These perturbations are intended to partially break the degeneracy of the pair of orthogonal modes associated with the fundamental TM_{11} mode of a circular patch, thereby promoting, over a bandwidth of at least 200 MHz centered around 15 GHz, the simultaneous excitation of two orthogonal modal components characterized by

comparable amplitudes and a phase difference close to $\pi/2$. In particular, as shown in Figure 4.7, in order to achieve this objective, a rectangular slot was introduced at the patch center, tilted by $3\pi/4$ with respect to the horizontal axis of a Cartesian reference system centered at that point, together with two semicircular notches etched along its edge and symmetrically arranged along directions tilted by $\pm \pi/4$ with respect to the u -axis shown in the figure. Obviously, a solution of this kind is not feasible only with circular patch antennas. A similar approach could also be applied to other types of patch antennas, such as square patches. However, a circular patch is inherently characterized by continuous rotational symmetry, which may make the achievement of circular polarization by means of the proposed solution less constrained. Indeed, for an unperturbed circular patch there are no privileged azimuthal directions and, consequently, unlike what occurs in other geometric configurations, the fundamental resonant mode TM_{11} is associated with a degenerate modal family with no privileged azimuthal directions. In this context, the synthesis of circular polarization through suitably dimensioned geometric perturbations is less constrained by geometry-imposed privileged directions and may therefore be potentially less critical.

After defining the basic geometry, the next step required to proceed with the design of the antennas with which the device of interest is to be equipped concerned the choice of the substrate to be used for their realization. Indeed, the substrate characteristics have a considerable impact on the dimensioning and overall performance of patch antennas. This becomes clear when reference is made to the first-order design relations commonly used for the preliminary dimensioning of an unperturbed circular patch operating in its fundamental TM_{11} resonant mode [111]:

$$f_r = \frac{8.791 \times 10^9}{2\pi a_{eff} \sqrt{\epsilon_r}} \quad (4.1)$$

$$a_{eff} = a \left\{ 1 + \frac{2h}{\pi \epsilon_r a} \left[\ln \left(\frac{\pi a}{2h} \right) + 1.7726 \right] \right\}^{1/2} \quad (4.2)$$

$$a = \frac{F}{\left\{ 1 + \frac{2h}{\pi \epsilon_r F} \left[\ln \left(\frac{\pi F}{2h} \right) + 1.7726 \right] \right\}^{1/2}} \quad (4.3)$$

$$F = \frac{8.791 \times 10^9}{f_r \sqrt{\epsilon_r}} \quad (4.4)$$

where f_r (Hz) is the resonant frequency of the fundamental TM_{11} mode, a (cm) is the physical radius of the patch, a_{eff} (cm) is the effective radius of the patch, i.e., an equivalent radius slightly larger than a introduced to account for fringing-field effects which make the patch appear electrically larger than its physical dimensions, h (cm) is the substrate thickness, and ϵ_r is the relative permittivity of the substrate material.

As can be inferred from the above relations, increasing ϵ_r reduces the resonant frequency of a circular patch characterized by a given effective radius a_{eff} . Therefore, the use of substrates with high relative permittivity makes it possible, for a given resonant frequency f_r , to reduce the overall dimensions of the patch. A similar effect is also observed, as a first approximation, when the substrate thickness increases. Indeed, as h increases, the effective radius of the patch tends to increase and, consequently, the resonant frequency of the fundamental mode decreases; therefore, in order to keep the resonant frequency unchanged, the physical dimensions of the patch must be reduced. The choice of the substrate to be used for the fabrication of patch antennas, however, cannot be made solely on the basis of reducing the overall footprint; indeed, ϵ_r and h also significantly affect the antenna

performance. More precisely, high values of ϵ_r end to reduce the antenna bandwidth and gain, while also promoting the excitation of surface waves which, by giving rise to spurious radiation, may degrade the polarization purity, radiation pattern, and radiation efficiency of the antenna [111]. An increase in substrate thickness, on the other hand, may enlarge the antenna bandwidth; however, excessively high values of h may enhance the excitation of surface waves and make impedance matching more critical. Conversely, excessively small values of h tend to confine the field more strongly within the substrate, thereby penalizing radiation efficiency [111]. The choice of the substrate must therefore be regarded as a compromise between miniaturization requirements and the overall performance of the antenna. With reference to this aspect, it is worth observing that, at the frequencies of interest within the present research project, antenna miniaturization does not represent a particularly critical issue. Indeed, according to the equations reported above, the high values of the operating frequencies of interest make it possible to keep the overall antenna footprint limited even when substrates characterized by relatively low values of ϵ_r are used. Therefore, in order to meet the primary objectives of the present work, it was decided to guide the choice of the substrate material for the fabrication of the antennas of interest primarily on the basis of performance and cost considerations. More precisely, in the preliminary stages of the design of the antennas to be employed in the device of interest, a market survey was carried out in order to identify, among the materials commonly used for PCB fabrication, a substrate characterized by low values of relative permittivity and loss tangent at the frequencies relevant to this work. The need to employ materials with low loss tangent value stems from the fact that, as frequency increases, dissipative effects in materials tend to become progressively more significant; consequently, in order to ensure good antenna performance at the frequencies of interest, it is preferable to use a substrate material with loss tangent values as low as possible. Table 4.1 summarizes the EM properties, typical thickness, maximum available thickness, and cost of the main substrate types used in RF applications for which publicly available price quotations can be found.

Substrate	Marketed as	Dk (ϵ_r) @10 GHz	Df ($\tan \delta$) @10 GHz	Common / Standard Thickness (mm)	Maximum Standard Thickness (mm)	Reference Format Dimensions (mm×mm)	Reference Cost (€)
Rogers	RT/duroid 5880	2.20	0.0009	0.508	1.575	305×457	319.12
Rogers	RO3003	3.00	0.0010	0.508	1.524	305×457	280.15
Rogers	RO4003C	3.38	0.0027	0.508	1.524	305×457	54.46
Rogers	RO4350B	3.48	0.0037	0.508	1.524	305×457	88.02

Table 4.1 – Main characteristics and cost of substrate types commonly used in RF applications.

As shown in Table 4.1, among the analyzed materials, Rogers RT/duroid 5880 exhibits the best electromagnetic properties. Indeed, this material is characterized by very low values of relative permittivity and loss tangent and, in accordance with the considerations discussed above, is especially suitable for high-frequency and broadband microstrip/stripline implementations. In particular, it represents an excellent choice for low-loss patch antennas and microwave feeding networks, since it helps limit dielectric losses and preserve high radiation efficiency, although at the expense of larger physical dimensions compared with higher-permittivity substrates. This material, however, is very

expensive. Therefore, in order to comply with the cost constraints of this work, Rogers RO4003C with a thickness of 0.508 mm was selected as the substrate for the fabrication of the antennas to be employed in the device of interest, as it represents, according to Table 4.1, the best compromise in terms of cost and performance. The choice to adopt a substrate thickness of 0.508 mm was guided by the fact that this value is among those commonly used in practice for the fabrication of PCBs based on Rogers RO4003C, since it corresponds to one of the standard laminate thickness values typically marketed by the manufacturer. This choice therefore makes it possible to avoid resorting to additional processing steps or to particular construction solutions aimed exclusively at obtaining different thickness values, such as, for example, the coupling or lamination of multiple dielectric sheets, possibly through the use of bonding films, i.e., dielectric adhesive films used to laminate different layers of a structure together, or prepreg materials, i.e., dielectric sheets pre-impregnated with partially cured resin and commonly employed in PCB lamination processes. Indeed, such solutions could increase the complexity of the manufacturing process and, consequently, the fabrication costs. At the same time, the choice of a widely used thickness makes it possible to increase the number of PCB manufacturers to which it will be possible to turn for the fabrication of the antennas of interest and, therefore, to have greater flexibility in selecting the most economically convenient solution, in accordance with the cost constraints of the final device.

After defining the type of substrate on which to focus attention, several preliminary EM simulations were carried out using CST Studio Suite in order to determine the values of the design parameters of the patch antenna shown in Figure 4.8 on the following page that are required for the antenna to resonate at 15 GHz and, in accordance with the design specifications defined in the previous section, to exhibit circular polarization over a bandwidth of at least 200 MHz centered around the resonant frequency. This preliminary analysis was necessary to verify whether the values of the design parameters required to implement the choices made, namely the excitation of circular polarization by means of geometric perturbations and the use of a Rogers RO4003C substrate with a thickness of 0.508 mm, were compatible with the manufacturing capabilities of the PCB producers to whom it will be necessary to turn for the fabrication of the antennas intended to equip the device of interest. The results of this analysis are discussed in the following section.

4.2.1 Preliminary antenna design

Figure 4.8 shows the model used for the EM simulations carried out to verify whether the optimal values of the design parameters of the proposed solution were compatible with the fabrication constraints imposed by PCB manufacturers operating in the consumer market.

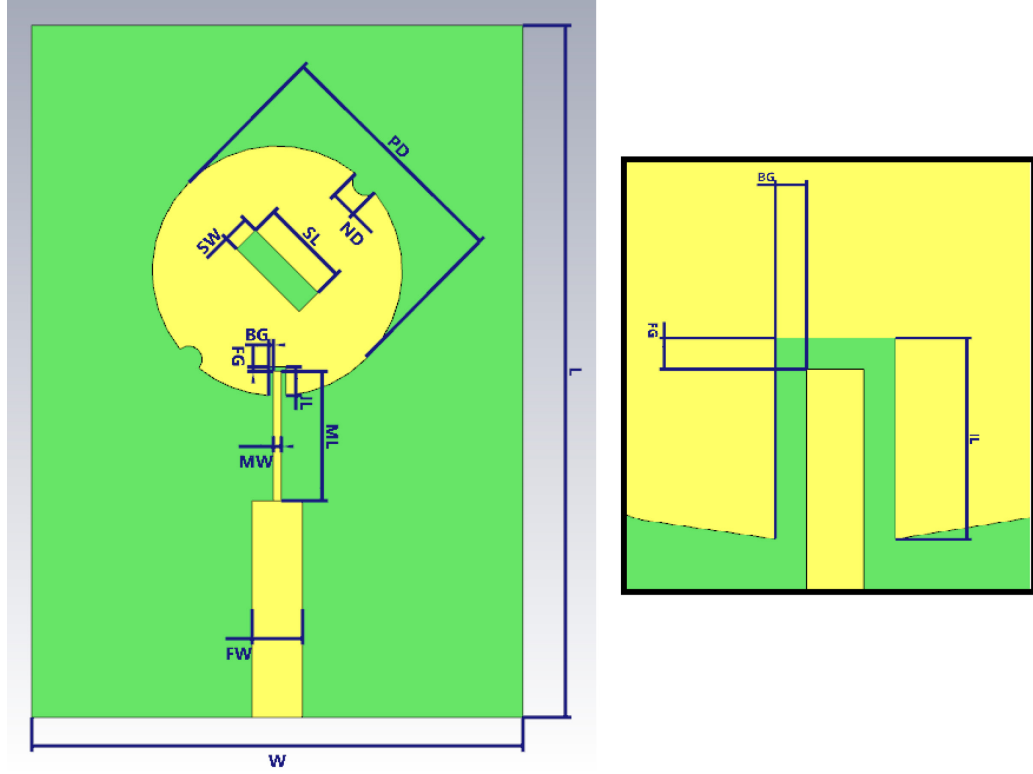


Figure 4.8 – Preliminary antenna model used to verify the feasibility of the proposed solution (left), with a magnified view of the feeding section (right).

For the initial dimensioning of the design parameters of the model shown in the figure, reference was made to the equations discussed in the previous section and to some preliminary considerations concerning the expected electromagnetic behavior of the antenna. More precisely, on the basis of Equations (4.1) ÷ (4.4), the diameter of the unperturbed circular patch (namely, PD in Figure 4.8) was first determined so as to correspond, on a Rogers RO4003C substrate ($\epsilon_r = 3.38$) with a thickness of 0.508 mm, to a resonant frequency of the fundamental TM_{11} mode slightly higher than the design value of 15 GHz. The reason for this choice lies in the fact that the central slot and the two notches on the edge, required to excite circular polarization by means of a single feed line, modifies the current paths on the patch and may, overall, lead to an increase in the effective electrical length of the radiating element, which in turn results, for unchanged physical dimensions, in a reduction of the resonant frequency.

On the basis of the value thus obtained, the first approximation values for the dimensions of the central slot (namely, SW and SL in Figure 4.8) and of the diameter of the two edge notches (namely, ND in Figure 4.8) were then defined. Indeed, as discussed extensively in the previous section, the purpose of these elements is to introduce controlled geometric perturbations, so as to enable the excitation of two quasi-degenerate resonant modes with comparable amplitudes and a phase

difference approximately equal to $\pi/2$; excessively strong perturbations could therefore lead to an excessive separation of the two resonances or to an imbalance in the modal amplitudes. For this reason, following a conservative approach, the preliminary values of the parameters SL , SW , and ND were defined as, $0.30PD$, $0.10PD$, and $0.05PD$, respectively.

Moreover, as shown in Figure 4.8, it was decided to feed the patch antenna by means of the feeding technique known as a coupled recessed microstrip feed, in conjunction with the use of a $\lambda/4$ transformer, in order to obtain, at the frequencies of interest, a good impedance match to the $50\ \Omega$ feed line. The choice of using a $\lambda/4$ transformer stems from the fact that, as is well known from the patch antenna theory, at the resonant frequency the input impedance of a circular patch is real and decreases, along the radial direction, from the edge towards the center, where it vanishes [111]. Moreover, at resonance, the input resistance at the edge of a circular patch is typically much higher than $50\ \Omega$. Consequently, the radial position at which the input resistance becomes equal to $50\ \Omega$ is generally located well inside the patch edge. Therefore, if the patch were to be fed directly by a $50\ \Omega$ line, a relatively deep inset would be required in order to reach the feed point corresponding to an input resistance of $50\ \Omega$. In addition, as is well known from the transmission-line theory, for a given substrate the characteristic impedance of a microstrip line decreases as the line width increases. Consequently, if the patch were to be fed directly by a microstrip line with a characteristic impedance of $50\ \Omega$, not only would the inset depth need to increase, but its width would also have to increase in order to accommodate the wider microstrip line. This, however, could severely limit the feasibility of exciting circular polarization through the proposed approach, since the geometric perturbation introduced by an inset feed whose dimensions are not negligible with respect to the patch radius may compete with the intentionally introduced perturbations, thereby making more difficult to achieve the amplitude and phase balance between two quasi-degenerate resonant modes required to generate the desired circular polarization. Therefore, in order to limit as much as possible both the depth and the width of the inset feed, it is advisable to use, for feeding the antenna, a high-impedance line section, which is characterized by a smaller width, properly dimensioned so that its input impedance is equal to $50\ \Omega$. In the case of a $\lambda/4$ transformer, as is well known, this result is obtained by using a line section of fixed length, namely $\lambda/4$ at the frequency of interest, with a characteristic impedance equal to the geometric mean of the load impedance and the desired input impedance. A $\lambda/4$ transformer, however, can provide impedance matching only for purely real loads. For a circular patch, as mentioned above, this condition is satisfied only at the resonant frequency; indeed, away from resonance, the antenna input impedance is generally complex and therefore includes a reactive component. From this standpoint, in order to increase the frequency range over which a good impedance match can be achieved, as mentioned above, it was decided to adopt the coupled recessed microstrip feeding technique. Indeed, this feeding technique provides an additional degree of freedom, beyond the sole variation of the characteristic impedance of the $\lambda/4$ line section, that can be exploited to improve impedance matching; more specifically, by properly varying the distance between the edges of the patch and the high-impedance microstrip section used for feeding, it is possible to introduce an additional capacitive reactance which can help compensate for the reactive part of the patch input impedance in the vicinity of the resonant frequency. Therefore, in accordance with the considerations discussed above, in order to avoid selecting excessively large initial values for the depth and width of the inset feed, the preliminary value of the parameter IL (inset length) was set equal to $0.10 * PD$, whereas the values of the parameters (Figure 4.8) FG (front gap), BG (borders gap),

and *MW* (matching width) were defined with reference to the average values of the *minimum copper trace width* and the *minimum spacing between two traces* specified in the manufacturing capabilities of the main PCB manufacturers serving the consumer market.

Finally, the value of the parameter *FL* (feed line) in Figure 4.8 was chosen such that, for the selected substrate, a microstrip line of width *FL* had a characteristic impedance of 50 Ω . Regarding the parameters *L* and *W*, which represent the substrate dimensions, preliminary conservative values of 15.5 mm and 11 mm, respectively, were selected based on the empirical rule prescribing a minimum clearance between the patch and the substrate edges equal to three times the substrate thickness, in order to minimize edge effects.

After defining the initial values of the design parameters for the antenna shown in Figure 4.8, several EM simulations were carried out, using the EM simulation software CST Studio Suite, in order to identify the optimal values of the aforementioned parameters that allow the design objectives to be achieved, namely to make the antenna resonate at 15 GHz and exhibit circular polarization over a bandwidth of at least 200 MHz around the resonance frequency. More precisely, the procedure followed to determine the optimal values of the design parameters of the antenna shown in Figure 4.7 is as follows:

- initially, several EM simulations were carried out to identify the values of the parameters *PD*, *IL*, *FG*, *BG*, *MW*, and *ML* that allow the antenna shown in Figure 4.8 to resonate at the target frequency of 15 GHz and to exhibit good impedance matching with the 50 Ω feed line over a bandwidth of at least 200 MHz around 15 GHz, while keeping the parameters *SL*, *SW*, and *ND* fixed at the preliminary values determined, as a first approximation, according to the considerations discussed previously.
- once the values of *PD*, *IL*, *FG*, *BG*, *MW*, and *ML* ensuring resonance at 15 GHz and good impedance matching had been identified, further simulations were carried out iteratively in order to optimize the parameters affecting the achievement of the desired circular polarization;
- more specifically, in each subsequent simulation, only one of the parameters *SL*, *SW*, and *ND* was varied by small amounts, while the others were kept fixed. At the end of each simulation, it was verified whether the resonance frequency and the impedance matching achieved in the previous step were still preserved. In addition, the effect of the applied variation of the parameter under consideration on the antenna axial ratio (*AR*) was evaluated in order to verify whether it was converging towards values below 3 dB. The *AR*, defined as the ratio of the major axis to the minor axis of the polarization ellipse associated to the radiated field (Figure 4.2b), is the parameter commonly used to assess the polarization type of an antenna. Indeed, according to the above definition, linear polarization corresponds to $AR \rightarrow \infty$, whereas circular polarization corresponds to $AR = 1$ (i.e., 0 dB). In practice, the condition $AR \leq 3$ dB is commonly adopted as a practical criterion to identify circular polarization;
- if the variation applied to the parameter under analysis caused an excessive shift in the resonance frequency or a significant degradation of the impedance matching obtained in the previous step, the parameters *PD*, *IL*, *FG*, *BG* and *MW* were retuned through additional EM simulations in order to restore resonance at 15 GHz and good impedance matching before proceeding further with the one-parameter-at-a-time optimization of the

parameters SL , SW , and ND . This step was necessary because modifying the slot and notches dimensions also alters the antenna input impedance and its effective electrical length, and may therefore compromise the optimization achieved in the previous stage;

- furthermore, if the variation applied to the parameter under examination proved to be detrimental in terms of AR , that parameter was restored to the best value obtained in the previous step, and the same procedure was then repeated for the remaining parameters;
- finally, once the values of SL , SW , and ND yielding the lowest AR had been identified, a final fine-tuning of the parameters IL , FG , BG , and MW was carried out in order to further improve the antenna impedance bandwidth.

The multiparametric optimization procedure just described is schematically summarized in Figure 4.9:

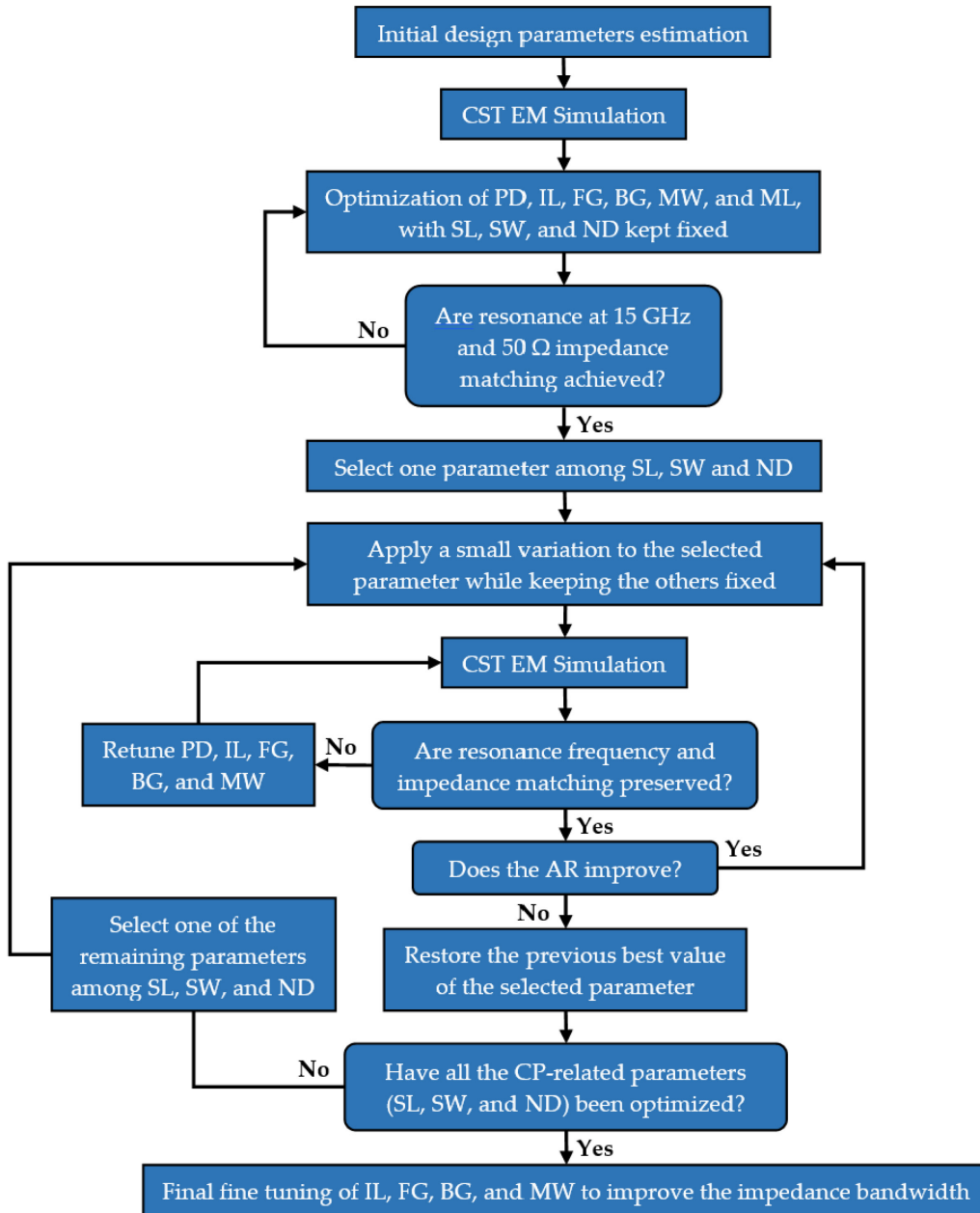


Figure 4.9 – Flowchart of the multiparametric EM optimization procedure adopted to determine the optimal values of the design parameters of the antenna shown in Figure 4.8.

Table 4.2 lists the optimal values of the design parameters obtained through the procedure described above for the antenna shown in Figure 4.8. The obtained values are consistent with the manufacturing capabilities of the main PCB manufacturers serving the consumer market. The obtained results are therefore consistent with the objective of developing a low-cost solution.

Parameter	Optimal Value	Parameter	Optimal Value
PD	5.60 mm	MW	0.18 mm
FG	0.10 mm	FW	1.13 mm
BG	0.10 mm	SL	1.98 mm
IL	0.65 mm	SW	0.60 mm
ML	2.90 mm	ND	0.60 mm

Table 4.2 – Optimal design parameter values for the antenna shown in Figure 4.8.

The results of the simulations performed using the values listed in Table 4.2 are shown in Figure 4.10 ÷ Figure 4.14.

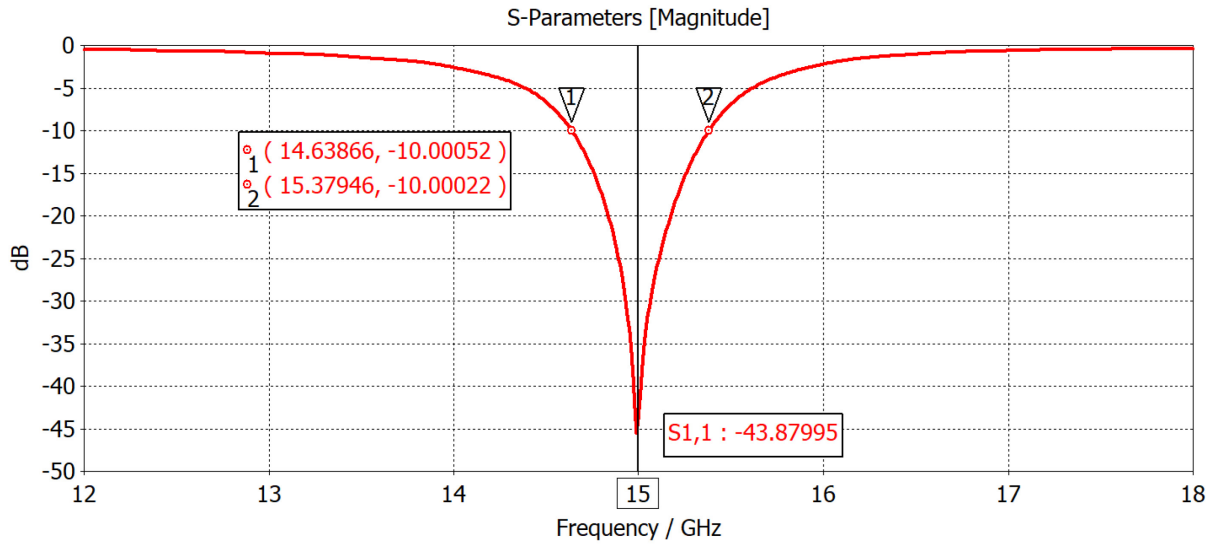


Figure 4.10 – Simulated magnitude of S11 versus frequency for the patch antenna shown in Figure 4.8.

As can be observed from Figure 4.10, the adopted design solutions allow the antenna under study to resonate at 15 GHz and to exhibit very good impedance matching over a bandwidth of approximately 740 MHz around the resonance frequency. Therefore, the antenna shown in Figure 4.8 is characterized by an impedance bandwidth that covers the entire frequency range of interest in this study.

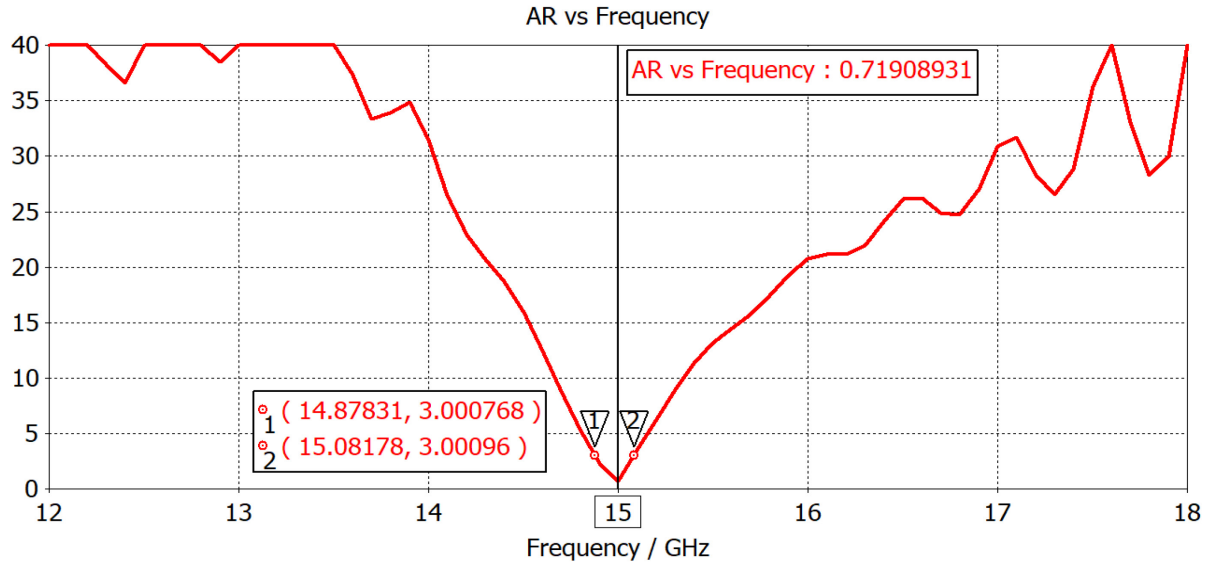


Figure 4.11 – Simulated AR versus frequency for the patch antenna shown in Figure 4.8.

Moreover, as shown in Figure 4.11, the proposed solution effectively enables the excitation of circular polarization by means of a single feed line. Indeed, the *AR* evaluated at the resonance frequency is lower than 1 dB. In addition, the *AR* remains below the 3 dB threshold, which, as previously discussed, is commonly adopted in practice as a criterion for identifying circular polarization, over a bandwidth of approximately 210 MHz around the antenna resonance frequency. This value is consistent with the design specifications defined in Section 4.1.

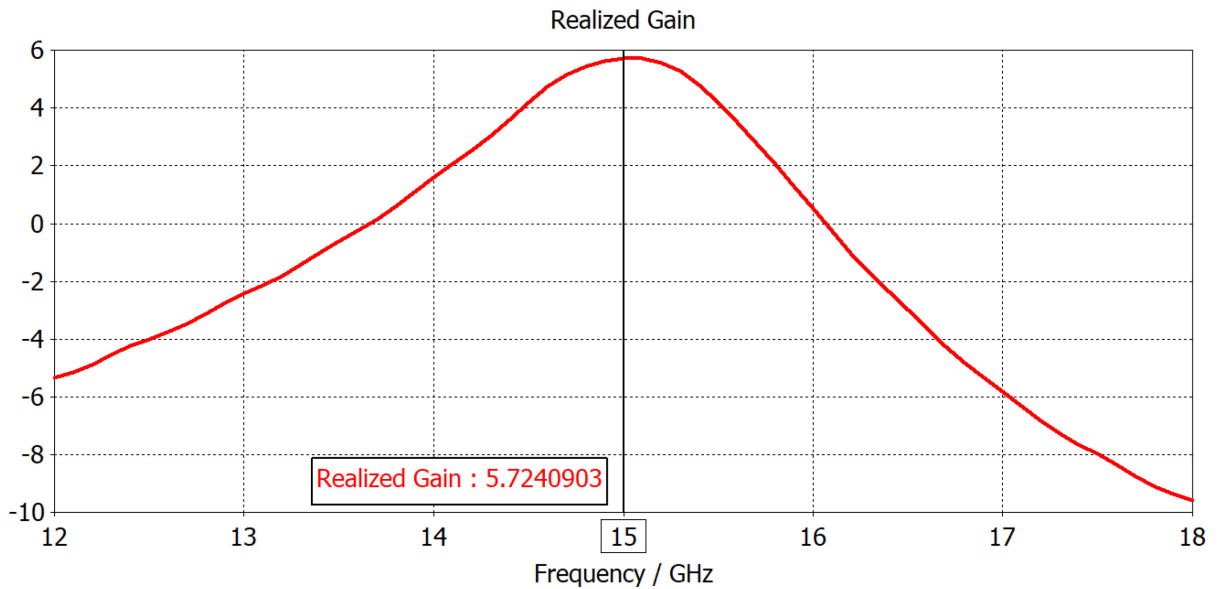


Figure 4.12 – Simulated broadside realized gain vs frequency for the patch antenna shown in Figure 4.8.

Figure 4.12 shows the broadside realized gain as a function of frequency. It is worth noting that the maximum broadside realized gain, achieved at the resonance frequency, is slightly lower than 6 dBi. This value is in very good agreement with the gain values typically achieved by patch antennas.

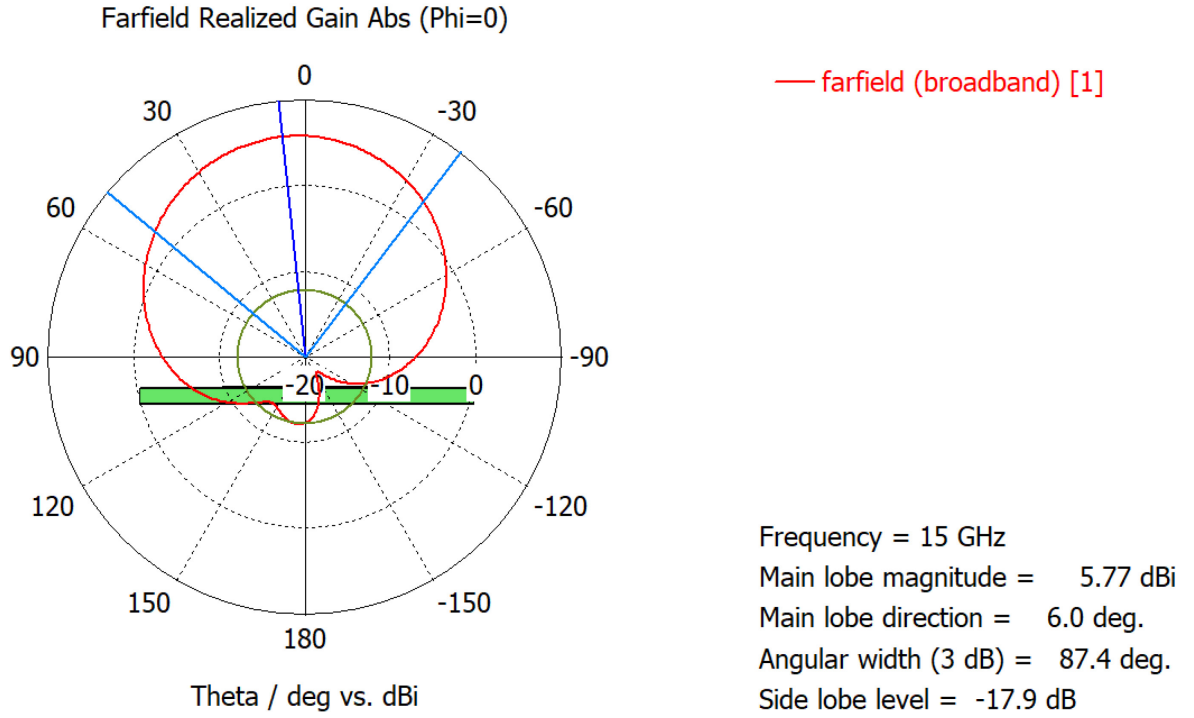


Figure 4.13 – Simulated radiation pattern at 15 GHz in the $\phi=0^\circ$ plane ($x-z$ plane) for the antenna shown in Figure 4.8.

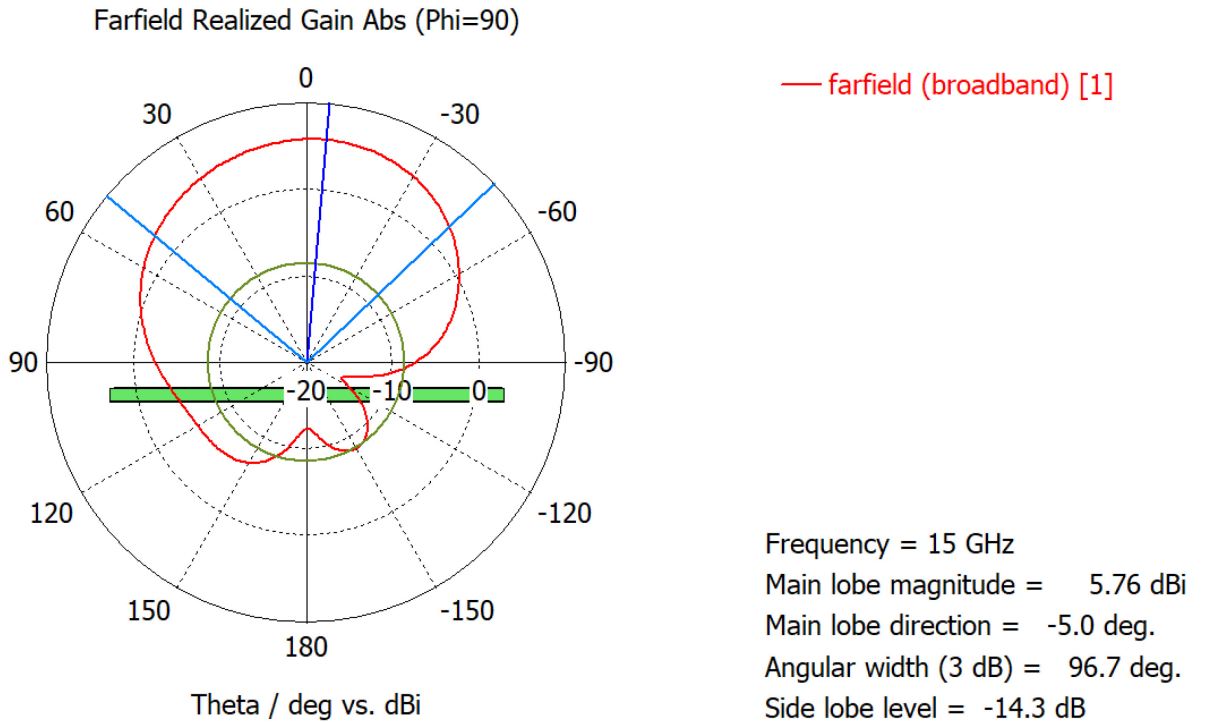


Figure 4.14 – Simulated radiation pattern at 15 GHz in the $\phi=90^\circ$ plane ($y-z$ plane) for the antenna shown in Figure 4.8.

As for the radiation behavior, Figures 4.13 and 4.14 report the radiation patterns in the two principal vertical planes, namely the θ -cuts at $\phi=0^\circ$ and $\phi=90^\circ$, corresponding to the $x-z$ and $y-z$ planes, respectively. According to the adopted coordinate system, the z -axis is normal to the patch

plane, whereas the y - and x -axes are parallel to the longer and shorter sides of the patch, respectively. These results further confirm that the antenna exhibits a broadside radiation behavior at 15 GHz. The main lobe is directed close to the broadside direction, although a slight angular deviation is observed, with the main beam pointing at 6.0° for $\phi=0^\circ$ and at -5.0° for $\phi=90^\circ$. The corresponding 3 dB beamwidths are 87.4° and 96.7° , respectively, indicating a relatively broad main beam in both principal planes. Moreover, the side lobe levels are equal to -17.9 dB for $\phi=0^\circ$ and -14.3 dB for $\phi=90^\circ$, thus confirming that secondary radiation remains sufficiently limited. Overall, the two cuts indicate similar radiation characteristics in the principal planes, although slight differences can be observed, likely due to the geometrical asymmetries introduced by the feeding structure and by the perturbations adopted to achieve circular polarization.

Although the presented solution proved sufficient to satisfy the antenna design requirements defined in Section 4.1, a slightly different antenna configuration was ultimately adopted in order to meet the primary objectives of the research project under consideration. Indeed, as mentioned earlier, the solution just presented had been primarily conceived to verify the practical feasibility of the proposed antenna configuration for achieving circular polarization by means of a single feed line. At the same time, it had been developed while also taking into account the possibility of connecting the antennas to the measurement electronics by means of coaxial connectors and RF cables, as is commonly done in practice. However, since a commercially available RF detector characterized by very compact dimensions and low cost was identified as suitable for meeting the primary objectives of the project, it was decided to investigate the possibility of integrating it directly onto the same PCB used for the antenna implementation, so as to improve the overall wearability of the device of interest. This alternative solution is discussed in the following section.

4.2.2 Final antenna design

Having verified that, at the frequencies of interest for the research project under consideration, the excitation of circular polarization in a circular patch antenna by means of a single feed line and controlled geometrical perturbations is compatible with the manufacturing capabilities of the main PCB manufacturers serving the consumer market, and having further verified that this solution satisfies the design constraints defined in Section 4.1, it was decided to adopt, as the basis for the design of the antennas intended to equip the device of interest, a slightly modified version of the configuration presented in the previous section. In particular, as shown in Figure 4.15, the antenna layout was modified so as to allow the installation of the RF detector, intended for measuring the levels of the signals of interest within the scope of the present research project, directly on the same PCB used for the antenna implementation. This solution is particularly attractive, since it avoids the use of coaxial connectors and RF cables, which, in addition to increasing the overall cost of the device, could compromise its wearability by increasing the overall bulk of the measurement system.

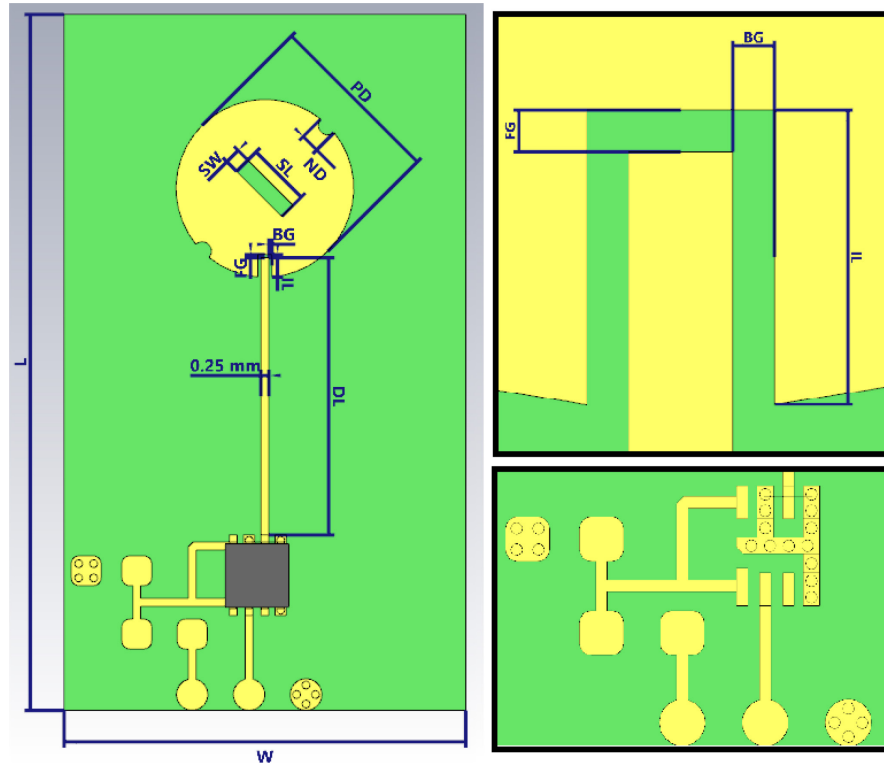


Figure 4.15 – Final antenna design (left), with a magnified view of the feeding section (top right) and the RF detector footprint (bottom right).

More precisely, as can be observed from the comparison between Figure 4.8 and Figure 4.15, in order to achieve the intended objective, it was decided to exploit part of the results obtained from the analysis carried out in the previous section and to modify the numerical model to be used for the optimization of the design parameters of the antennas of interest by replacing the feeding section of the antenna shown in Figure 4.8, consisting of the $50\ \Omega$ microstrip line and the $\lambda/4$ transformer, with a microstrip line having a fixed width of 0.25 mm, corresponding to the width of the input pin of the RF detector to be used for the implementation of the exposimeter of interest, and connected to the input pin pad of the footprint designed to host the detector.

Before presenting the results of the simulations carried out on the model shown in Figure 4.15 in order to achieve the intended objective, it is useful to clarify the rationale underlying their execution. Indeed, the need to use, for the direct connection to the detector, a microstrip section having a fixed width of 0.25 mm introduces additional issues from the impedance-matching standpoint. More precisely, as will be discussed in the following section, the input pin of the detector to be used to satisfy the primary objectives of this work is internally matched to 50 Ω . A 0.25 mm-wide microstrip line implemented on an RO4003C substrate with thickness equal to 0.508 mm, however, exhibits a characteristic impedance significantly higher than this value, being on the order of 100 Ω . Under these conditions, the direct connection of the detector cannot be achieved simply by replacing the feeding section of the previous design with a 0.25 mm-wide line section. In fact, even if the patch were properly matched to such a line, in the absence of suitable design measures the detector would see, at the connection point, an impedance different from that for which its input was designed, thus resulting in impedance mismatch.

To overcome the issue described above, and in order to introduce as few modifications as possible to the model described in the previous section, it was decided to adopt a solution based on the impedance transformation that occurs along a transmission line terminated with a mismatched load. Indeed, it is well known that the input impedance of a transmission line terminated with a mismatched load varies as the length of the considered line section changes. Therefore, if, by varying the length of the line section to be used to connect the patch to the detector, it were possible to identify a configuration for which the impedance seen at the input of such a line had a real part as close as possible to 50 Ω then, by connecting the detector at that point and acting on the inset parameters of the patch, it would be possible to achieve good impedance matching without the need to resort to more complex and more demanding solutions, also from a computational standpoint.

Therefore, in order to assess the validity of the proposed solution, several EM simulations were carried out, starting from the optimal values of the design parameters related to the geometry of the radiating element reported in Table 4.2, with the aim of identifying the minimum length that a 0.25 mm-wide microstrip line, implemented on an RO4003C substrate with thickness equal to 0.508 mm and coupled to the patch as shown in Figure 4.8 and Figure 4.15, must have so that its input impedance exhibits, at the design frequency of 15 GHz, a real part equal to 50 Ω . Once the desired result had been achieved, recalling that the impedance seen at two sections of a transmission line separated by half a guided wavelength is the same, the value of the parameter DL , shown in Figure 4.15, was determined by extending the minimum microstrip length required for the line to exhibit, under the design conditions, an input impedance with a real part equal to 50 Ω by an amount equal to $\lambda g/2$. In particular, this was done to avoid placing the detector too close to the radiating element, in order to limit the possibility of it excessively perturbing the antenna radiation pattern.

After defining the initial value of the parameter DL , EM simulations were carried out to finely tune the design parameters shown in Figure 4.15 by following a procedure analogous to that described in the previous section. This operation was necessary to compensate for the changes in the patch AR caused by the slight increase in the width of the line coupled to the patch compared with that considered in the model shown in Figure 4.8. The results obtained following the optimization procedure are summarized in Table 4.3:

Parameter	Optimal Value	Parameter	Optimal Value
PD	5.62 mm	DL	8.99mm
FG	0.10 mm	SL	2.02 mm
BG	0.10 mm	SW	0.50 mm
IL	0.72 mm	ND	0.60 mm
W	12.70 mm	L	22.00 mm

Table 4.3 – Optimal design parameter values for the antenna shown in Figure 4.15.

By comparing the data reported in Table 4.3 with those obtained for the antenna model described in the previous section, it can be observed that the optimal design values for the antenna shown in Figure 4.13 do not differ significantly from those derived for the antenna shown in Figure 4.7. These results therefore indicate that the antennas of interest can be fabricated using the standard manufacturing processes commonly adopted in printed circuit board (PCB) production. Indeed, the values reported in Table 4.3 fall well within the manufacturing capabilities of PCB manufacturers serving the consumer market and, consequently, further support the feasibility of producing the antennas of interest through low-cost manufacturing processes.

Figure 4.16 ÷ Figure 4.20 show the simulation results obtained for the antenna model under consideration when the optimal values of the design parameters reported in Table 4.3 are used.

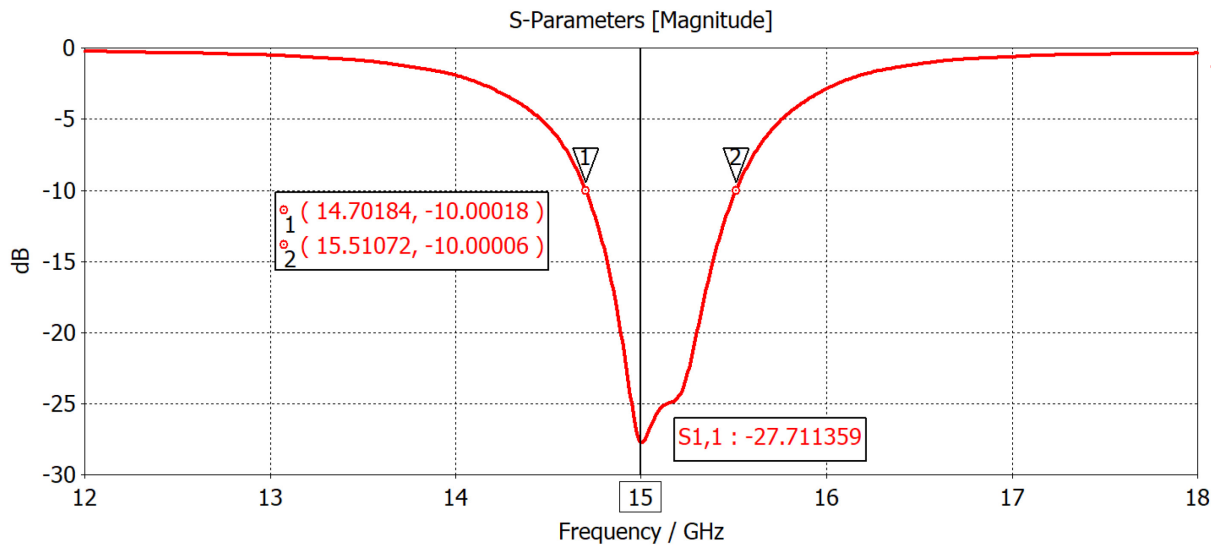


Figure 4.16 – Simulated magnitude of S11 versus frequency for the patch antenna shown in Figure 4.15.

As shown in Figure 4.16, the frequency response of the magnitude of S11 confirms the validity of the solution adopted to interface the patch antenna under consideration with the RF detector to be employed in the implementation of the target device. Indeed, the same figure shows that the antenna resonates at the design frequency of 15 GHz and exhibits good impedance matching over the entire frequency range of interest for this research project. In particular, the antenna impedance bandwidth extends from 14.7 GHz to 15.51 GHz, corresponding to approximately 810 MHz.

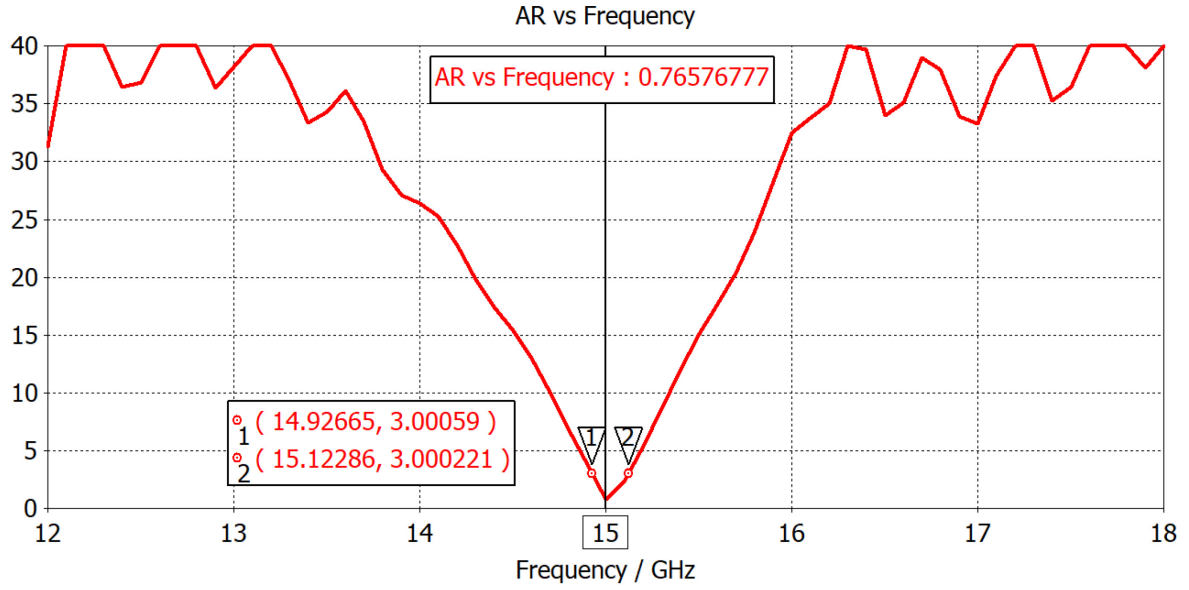


Figure 4.17 – Simulated AR versus frequency for the patch antenna shown in Figure 4.15.

Figure 4.17 shows the AR response of the antenna as a function of frequency. It can be observed that the AR value obtained at the resonant frequency is equal to 0.77 dB, thus confirming that circular polarization is effectively achieved at the design frequency. Moreover, as shown in the same figure, the 3-dB AR bandwidth is approximately 200 MHz, in agreement with the design specifications defined in Section 4.1.

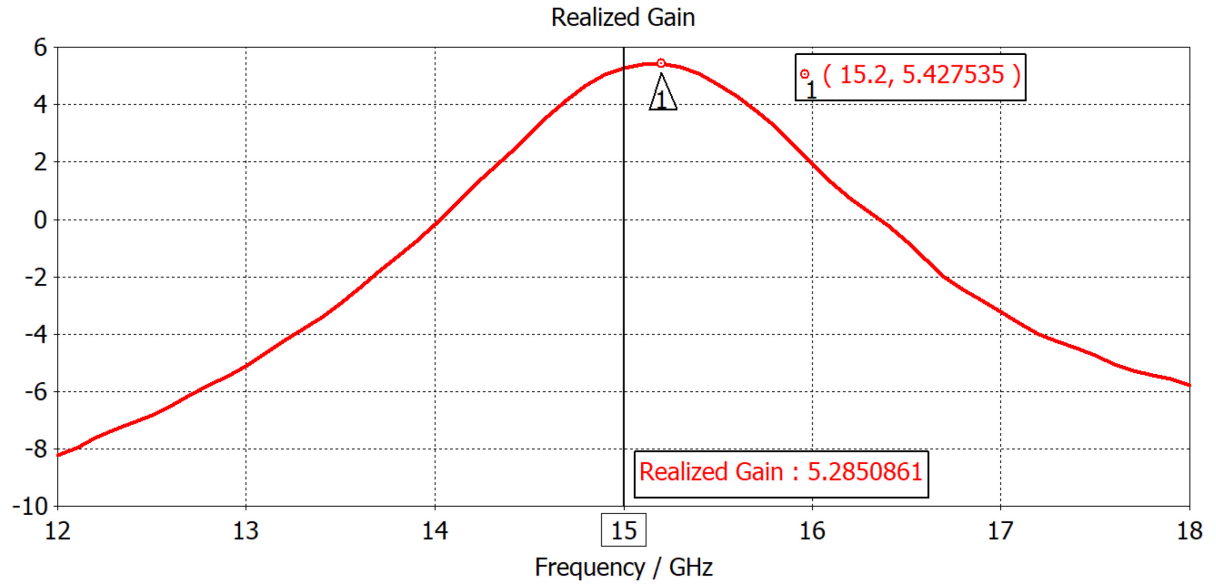


Figure 4.18 – Simulated broadside realized gain vs frequency for the patch antenna shown in Figure 4.15.

As shown in Figure 4.18, the maximum broadside realized gain occurs at approximately 15.2 GHz, with a value of about 5.43 dBi. This value is only slightly higher than the realized gain obtained at the antenna resonant frequency, which is approximately 5.29 dBi. The difference between the two values is therefore limited to 0.14 dB and may be considered negligible from a practical standpoint.

The obtained realized-gain values are consistent with those typically associated with patch antennas and therefore further confirm the soundness of the proposed design.

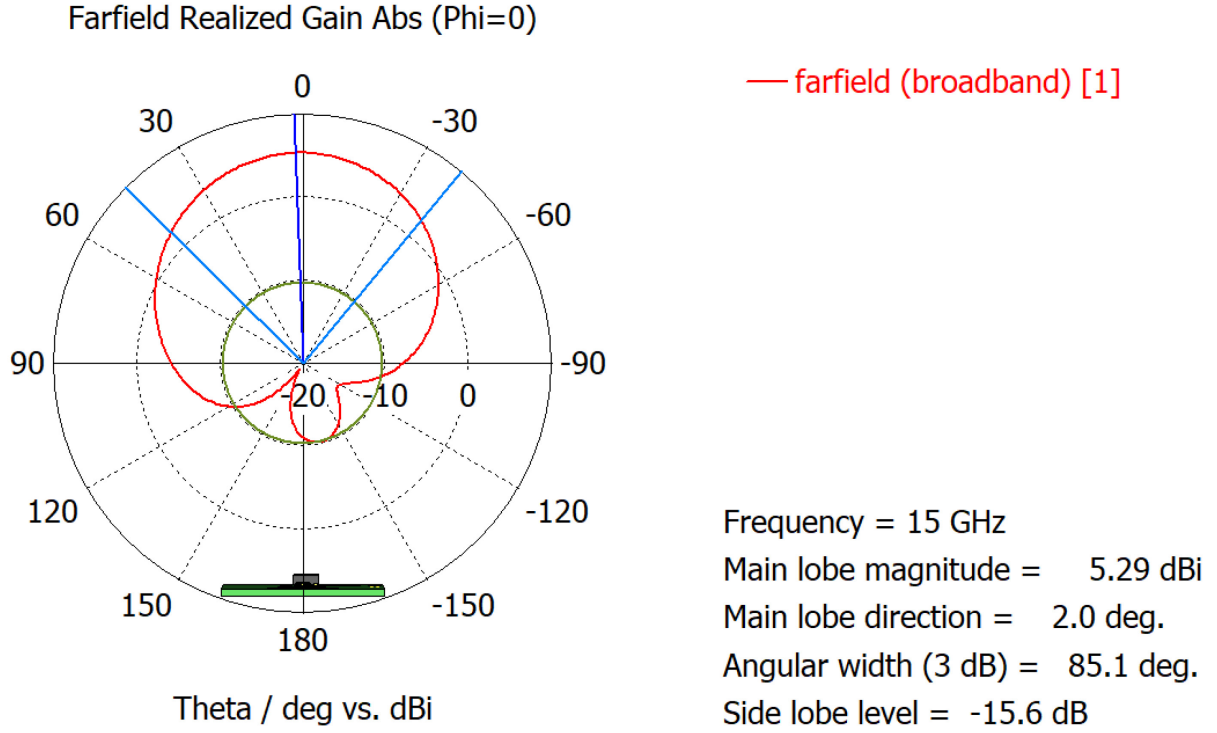


Figure 4.19 – Simulated radiation pattern at 15 GHz in the $\phi=0^\circ$ plane ($x-z$ plane) for the antenna shown in Figure 4.15.

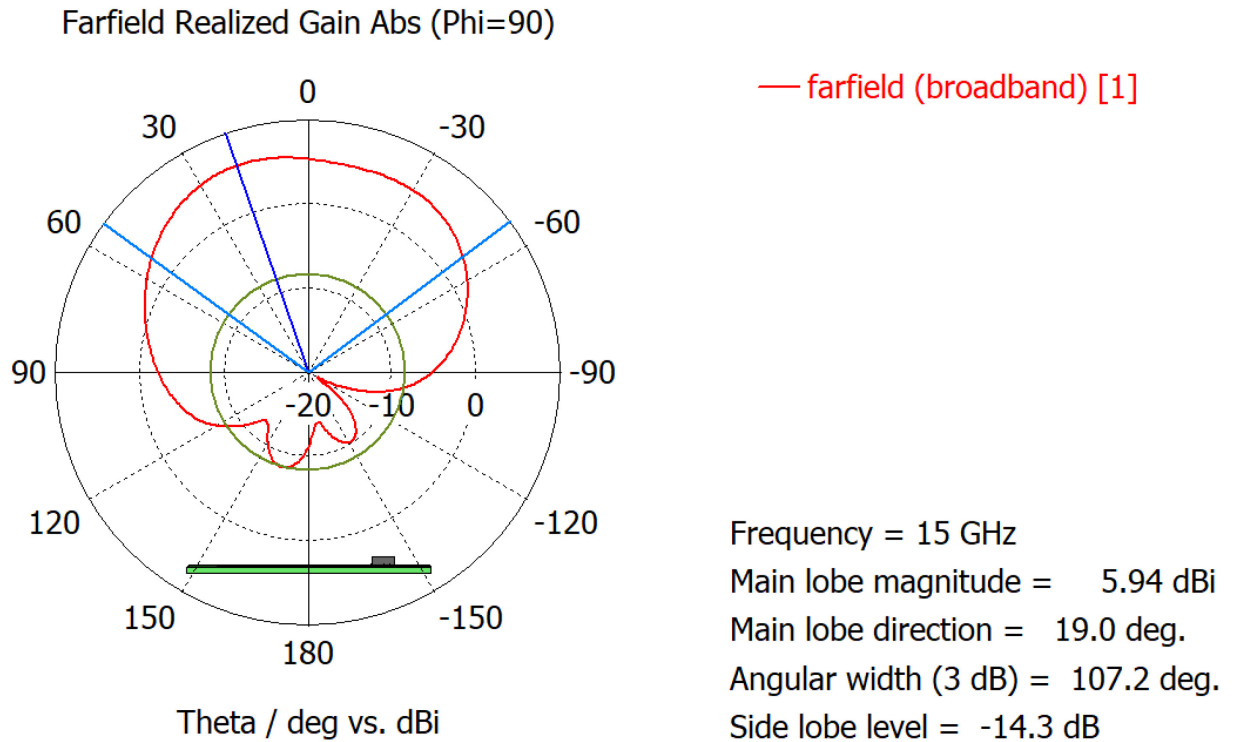


Figure 4.20 – Simulated radiation pattern at 15 GHz in the $\phi=90^\circ$ plane ($y-z$ plane) for the antenna shown in Figure 4.15.

Figure 4.19 and Figure 4.20 show the far-field cuts of the proposed antenna evaluated at the design frequency of 15 GHz in the two principal vertical planes. According to the adopted coordinate system, the z -axis is normal to the patch plane, whereas the y - and x -axes are parallel to the longer and shorter sides of the patch, respectively. As can be observed, in the $\phi=0^\circ$ plane (corresponding to the $x-z$ plane) the maximum realized gain is equal to 5.29 dBi and is achieved close to the broadside direction, with a main lobe tilt of approximately 2° . In the $\phi=90^\circ$ plane (corresponding to the $y-z$ plane), instead, the maximum realized gain is equal to 5.94 dBi and is observed at an angle of approximately 19° , thus indicating a non-negligible off-broadside displacement of the main lobe in the $y-z$ plane. The obtained results therefore indicate that the antenna does not exhibit a perfectly symmetric broadside radiation pattern. This behavior can reasonably be attributed to the geometry of the final integrated structure. More precisely, although the patch itself is correctly tuned at the design frequency, the presence of the elongated microstrip interconnection and of the detector circuitry arranged along the y -direction introduces a structural asymmetry which may perturb the current distribution on the radiating element. As a consequence, the distortion of the radiation pattern mainly appears in the $y-z$ plane, i.e., in the plane containing the dominant geometrical asymmetry of the layout. Nevertheless, such a behavior should not be regarded as particularly critical for the intended application. Indeed, as shown in Figure 4.19 and Figure 4.20, the 3-dB beamwidth values are approximately 85.1° and 107.2° for the $\phi=0^\circ$ and $\phi=90^\circ$ cuts, respectively. The relatively broad main beam observed in both principal planes, together with the moderate realized-gain values obtained, ensures a sufficiently wide angular response and, consequently, limited sensitivity to antenna pointing. In particular, although the maximum realized gain in the $\phi=90^\circ$ plane is achieved at an off-broadside angle of about 19° , the broadside realized gain at 15 GHz is equal to 5.29 dBi, i.e., only 0.65 dB lower than the maximum value obtained in the same plane. This means that, despite the observed tilt, the antenna response along the broadside direction remains close to the maximum one and can therefore be considered suitable for the desired application because it provides a good balance between radiation performance, compactness, integration level, and manufacturability within the scope of the proposed proof of concept design.

To conclude the description of the proposed antenna, it is useful to clarify the reasons why no filtering stage has been included between the antenna and the detector at the present stage of the design. The main reason lies in the fact that no off-the-shelf SMT band-pass filter has been identified with a passband fully compatible with the bandwidth requirements of the target application. In particular, the commercially available solutions found for the frequency range of interest generally exhibit passbands wider than the impedance bandwidth of the proposed antenna and, therefore, would not provide any substantial advantage in terms of selectivity. Furthermore, once the relevant 6G standards have been defined and compatible signal generators become available, the target device will be assembled and subsequently tested in an anechoic chamber under controlled conditions. In such a scenario, no interfering signals are expected to be present, and the introduction of a filtering stage is therefore not necessary. For this reason, it was considered preferable not to further increase the complexity of the overall structure at the present stage of the design. Finally, the proposed solution remains well suited to subsequent integrations. Therefore, if SMT band-pass filters meeting the design specifications become commercially available in the future, they may be integrated on the antenna PCB, in a manner analogous to that envisaged for the RF detector.

4.3 RF detector and microcontroller specifications

This section presents the main features of the hardware components selected, together with the antenna configuration described in the previous section, to meet the primary objectives of the research project under consideration.

Indeed, for the design of a wearable exposimeter capable of detecting the levels of future 6G signals operating in the FR3 band, it is necessary not only to properly design the antennas with which the device will be equipped, but also to carefully select the hardware components to be used in conjunction with them, so as to enable the proper detection and digitization of the signal levels of interest. For this reason, in parallel with the activities carried out to properly design the antennas presented in the previous section, a market survey was conducted in order to identify an RF detector suitable for achieving the intended objectives. In carrying out this search, the attention was focused on the class of RMS detectors. Indeed, this category of detectors is particularly suitable for the development of devices aimed at assessing EMF exposure levels, since it provides an output that can be directly related to the average power of the received signal. Among the various commercially available options, one solution that proved particularly attractive for meeting the objectives of this work is the true RMS RF detector manufactured by Analog Devices and marketed under the name **LTC5596**. This device, in fact, combines a very wide operating frequency range with a simple output interface and a compact footprint. More precisely, the LTC5596 is housed in a compact $2\text{ mm} \times 2\text{ mm}$ dual flat no-lead package (Figure 4.21), operates from 100 MHz up to 40 GHz and is designed as a high-accuracy RMS power detector that provide an output DC voltage related to the average RF input power through a linear-in-dB transfer characteristic.

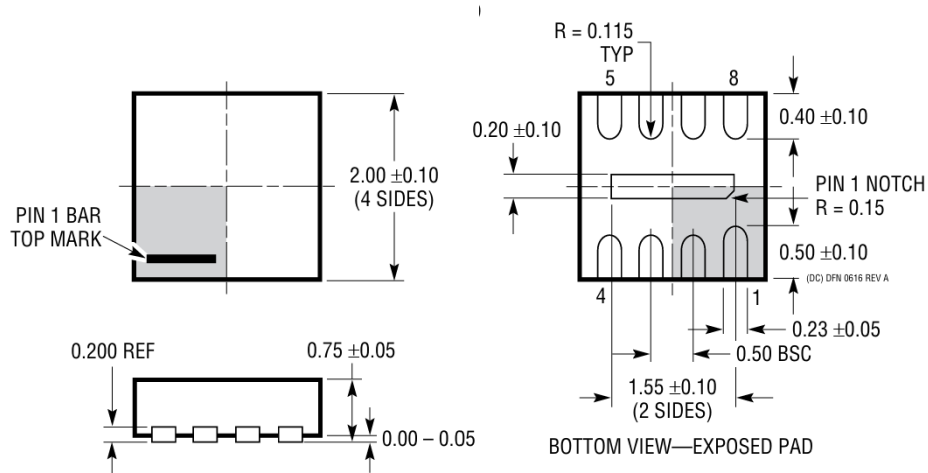


Figure 4.21 – Package outline and main dimensions of the LTC5596 RF detector [112].

As mentioned above, these features are particularly relevant for the development of a wearable device such as that envisaged in the present research project. In particular, the reduced footprint, as shown in the previous section, makes it possible to integrate the LTC5596 directly onto the PCBs used for the implementation of the patch antennas intended to equip the target device, even when the antennas themselves are characterized by very compact dimensions. Furthermore, the simple RF to DC conversion provided by the device make it possible to use low-cost MCU in order to digitize and process its outputs.

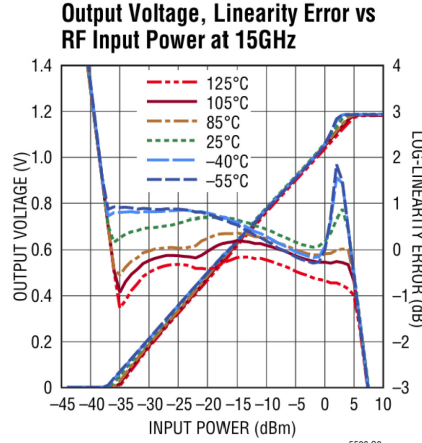


Figure 4.22 – Output voltage and linearity error versus RF input power for the LTC5596 at 15 GHz [112].

Another aspect that makes the LTC5596 an attractive solution for meeting the primary objectives of the present research project is that it provides a sufficiently wide dynamic range for the intended applications of the target device and, moreover, as shown in Figure 4.22, it exhibits an approximately linear-in-dB transfer characteristic, with a specified slope of $29 \text{ mV} \cdot \text{dB}^{-1}$ over its dynamic range. In particular, as shown in Figure 4.22, at 15 GHz the RF input-power range corresponding to a $\pm 1 \text{ dB}$ log-linearity error extends from approximately -37.5 dBm to $+5.5 \text{ dBm}$ at 25°C . These values allow a first-order estimate of the minimum and maximum incident power level that could be detected by the LTC5596 when equipped with the antennas described in Section 4.2.2. Indeed, assuming that the minimum input-power level reliably detectable by the detector at 15 GHz is approximately -37 dBm , whereas the maximum input-power level is approximately $+5.5 \text{ dBm}$, and considering that the antennas presented in Section 4.2.2 exhibit a broadside realized gain of 5.29 dBi , the corresponding minimum and maximum incident power-density levels can be evaluated by means of the following equation:

$$S = \frac{4\pi P_r}{\lambda^2 G} \quad (4.5)$$

By applying the equation introduced above, using the maximum and minimum power levels detectable by the LTC5596, and considering that the wavelength at 15 GHz is 0.02 m , the following expressions are obtained when the antenna gain is expressed on a linear scale:

$$\begin{aligned} S_{\max} &\approx 33 \text{ W} \cdot \text{m}^{-2} & E_{\max} &= \sqrt{377 S_{\max}} & E_{\max} &= 112 \text{ V} \cdot \text{m}^{-1} \\ S_{\min} &\approx 1.65 \times 10^{-3} \text{ W} \cdot \text{m}^{-2} & E_{\min} &= \sqrt{377 S_{\min}} & E_{\min} &= 0.79 \text{ V} \cdot \text{m}^{-1} \end{aligned} \quad (4.6)$$

In the case of a linearly polarized incident field, since the antennas analyzed in Section 4.2.2 are circularly polarized, the evaluation of the minimum and maximum incident power-density levels detectable by the device requires taking the polarization mismatch factor into account. More precisely, for an ideally circularly polarized antenna illuminated by a linearly polarized incident field, the polarization mismatch factor is equal to $1/2$, corresponding to a 3 dB polarization mismatch loss. Under these conditions, in order to achieve the same received power, the required incident power density doubles, while the electric-field strength increases by a factor of $\sqrt{2}$. Even under these conditions, the minimum detectable field levels achievable through the combined use of the antennas

described in the previous section and the LTC5596 remain well suited to the applications for which the device considered in the present research project has been conceived.

Another important advantage of the LTC5596 lies in its compatibility with wideband and highly dynamic waveforms. Indeed, the device is capable of providing accurate RMS power measurements for signals with crest factors up to 12 dB and remains suitable even when the crest factor varies over time. [112]. This feature is of particular importance for ensuring that the proposed device can properly fulfill the function for which it has been designed. Indeed, similarly to present 5G signals, future 6G signals are expected to be broadband and potentially characterized by high crest factors. In such a scenario, if the detector were not sufficiently robust with respect to this aspect, its output could depend too strongly on signal peaks and would therefore fail to correctly represent the average received power. Furthermore, since 6G standards have not yet been fully defined, the adoption of a detector that remains reliable even in the presence of strongly modulated signals and time-varying crest factors makes the proposed architecture more flexible and more robust with respect to future technological uncertainty.

Finally, another aspect that makes the LTC5596 particularly attractive with respect to the objectives of the present research project is its cost. Indeed, the device is commercially available at a price on the order of several tens of euros, which makes it particularly interesting for the development of a low-cost exposimeter intended for the measurement of future 6G signals.

In addition to the detector, the realization of the target device, in agreement with the simplified model shown in Figure 4.1, also requires the adoption of an ADC and a microcontroller, which are needed to digitize and process, respectively, the signals provided at the detector output. From this point of view, a wide range of hardware solutions may in principle be considered. Indeed, since the output of the LTC5596 is a slowly varying DC signal whose fluctuations are only related to variations in the power of the input RF signal, the digitization of the detector output does not require the use of ADCs characterized by particularly demanding specifications. For this reason, the ADCs currently integrated in modern microcontroller architectures can suitably perform the required function without particular difficulties. As an example, the microcontrollers belonging to the STM32 family integrate ADC stages whose performance largely exceeds the actual sampling requirements associated with the signals available at the LTC5596 output. A possible solution that may be adopted in this respect is represented by the STM32L433RTP6 device. According to the manufacturer datasheet, this MCU is based on an Arm Cortex-M4 core operating at up to 80 MHz and integrates a 12-bit ADC with a maximum conversion rate of 5 Msps, hardware oversampling, and up to 16 external input channels. Such characteristics are well beyond the actual acquisition requirements of the detector output. Indeed, the LTC5596 is characterized by rise time equal to 2.9 μ s and fall time equal to 8.1. These values correspond to variation of the output signals in the order of a few hundreds of kilohertz. Therefore, the use of an MCU integrating a multi-megasample ADC should be regarded as more than sufficient for the correct acquisition of the detector output, while also providing ample margin for signal averaging, oversampling, and possible multi-channel implementations.

4.4 Experimental validation of the simulation

The aim of this section is to present the results of the measurements carried out on patch-antenna prototypes representative of those discussed in Section 4.2.2, in order to validate the numerical-simulation results obtained for the optimal design of the antennas to be used for the realization of the target device.

Before presenting the results of the measurements under consideration, however, it is useful to make a brief preliminary remark. Indeed, the antennas discussed in Section 4.2.2 were designed so as to ensure direct interfacing with the LTC5596 RMS detector; therefore, in their original configuration, the antennas analyzed in Section 4.2.2 cannot be directly connected to the laboratory instrumentation normally used, in a controlled environment, to perform the measurements required for their experimental characterization. For this reason, in order to carry out an experimental evaluation of the characteristics of the antennas to be employed in the realization of the target device, it was necessary to slightly modify the model discussed in Section 4.2.2 before proceeding with the physical fabrication of the prototypes on which the measurements under consideration were performed. More precisely, as shown in Figure 4.23, in order to enable the connection of the antennas under consideration to laboratory instrumentation, the section intended to accommodate the RF detector was removed from the original model, and a GCPW-to-microstrip transition was designed to allow the use of end-launch coaxial connectors.

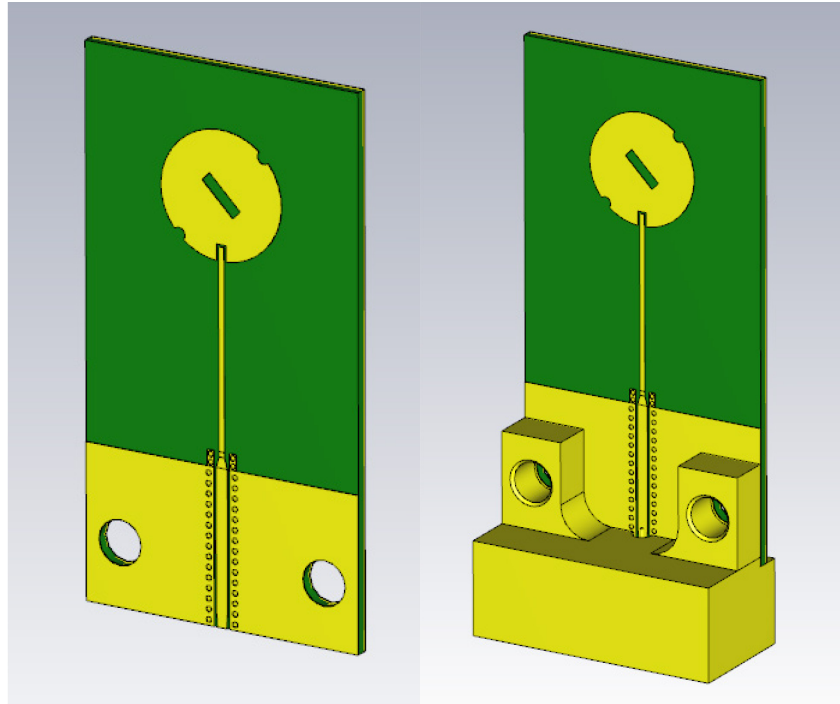


Figure 4.23 – CST EM model of the connectorized circular patch antenna.

It is worth emphasizing that the design parameters of the connectorized patch shown in Figure 4.23 are fully equivalent to those reported in Table 4.3. Indeed, the numerical models shown in the above figure were used exclusively for the design of the GCPW-to-microstrip transition required to enable the use of end-launch coaxial connectors.

Figure 4.24 shows one of the antenna prototypes fabricated according to the model shown in Figure 4.23.

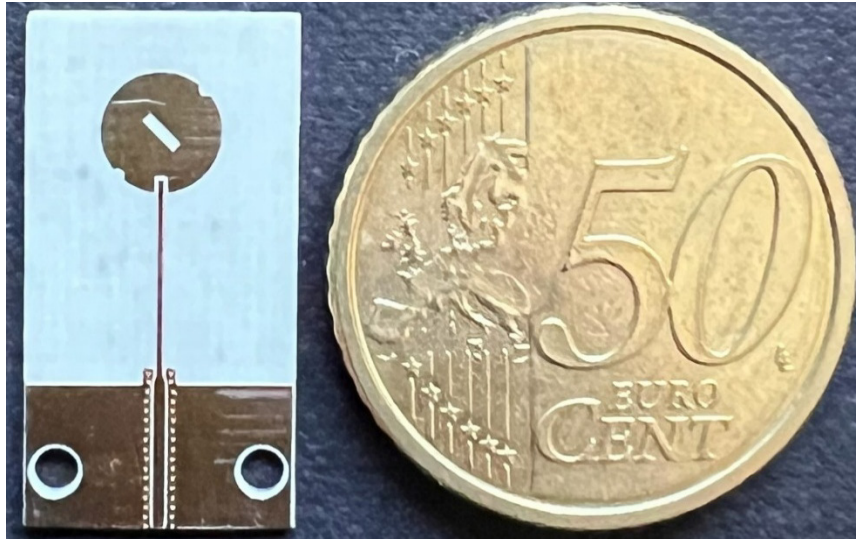


Figure 4.24 – Prototype of the circular patch antenna used to carry out the simulations validation measurements.

After the fabrication of the patch-antenna prototypes to be used for the realization of the target device, laboratory measurements were carried out in order to verify the agreement between the experimental results and the numerical ones obtained from the simulations performed on the model shown in Figure 4.23 by using the optimal values of the design parameters reported in Table 4.3. More precisely, before carrying out the antenna-characterization measurements in the semi-anechoic chamber available at the University of Cassino and Southern Lazio, some preliminary measurements were performed in the μ LAB microwave laboratory of the Department of Electrical and Information Engineering “Maurizio Scarano” at the University of Cassino. The preliminary measurements were carried out using a Rohde & Schwarz ZNA43 vector network analyzer and a ZV-Z9x test cable (Figure 4.25).



Figure 4.25 – Measurement equipments.

Figure 4.26 shows a comparison between the measured magnitude response of S11 and the corresponding response obtained from the numerical analyses carried out on the model shown in Figure 4.23.

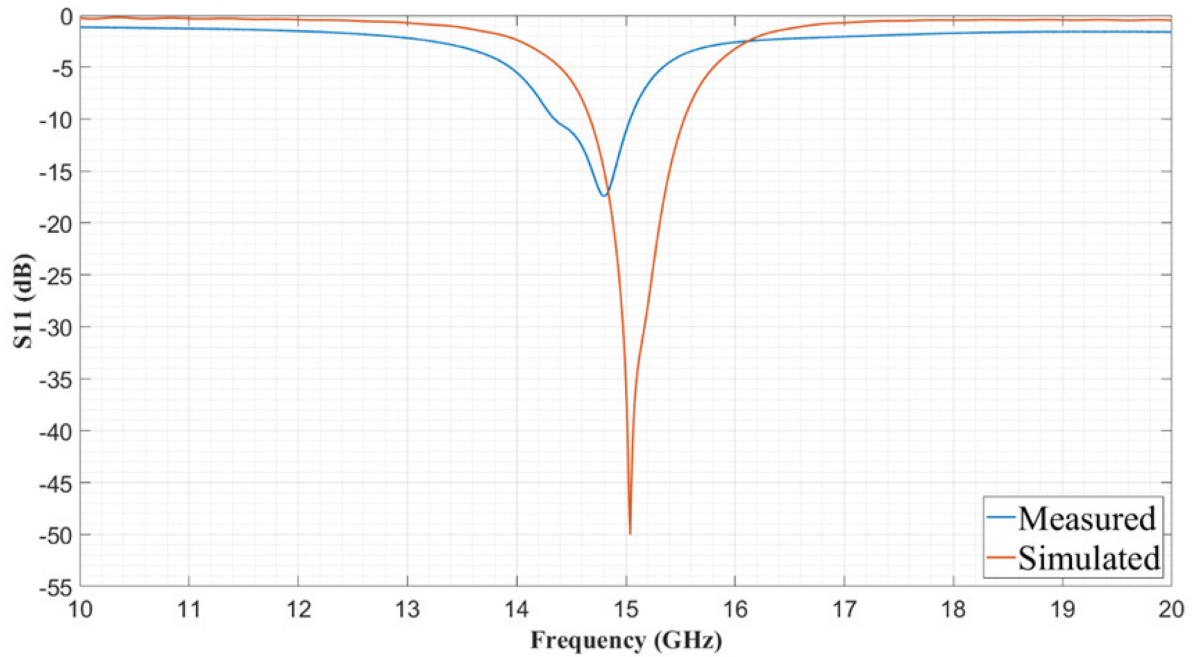


Figure 4.26 – Comparison between the magnitude response of the measured S11 and the one simulated with optimal values for design parameters.

As shown in Figure 4.26, the fabricated prototypes resonate at a frequency lower than the design frequency of 15 GHz. In particular, the resonance-frequency shift observed in the antenna prototypes under consideration is approximately 200 MHz.

In order to investigate the possible causes of the evident discrepancy between the simulated and measured S11 magnitude responses, it was decided to carefully examine the accuracy of the fabrication processes carried out by the manufacturer on the antenna prototypes. Indeed, the optimization procedures performed during the design phase of the antennas of interest, discussed in Sections 4.2.1 and 4.2.2, proved highly instructive in understanding the effects resulting from variations in the values of the different parameters governing the overall antenna behavior. To carry out these analyses, a visual inspection of the antenna prototypes was performed using a microscope. The analysis revealed the presence of several fabrication-related issues affecting the prototypes under consideration. For this reason, in order to assess the extent of the geometric deviations introduced during the antenna fabrication process, high-definition scans of the patch feeding region and of the GCPW-to-microstrip transition were acquired. The dimensions of the design parameters governing the electromagnetic behavior of the overall structure were then evaluated on the basis of these scans by means of imaging software. The results of the analyses carried out are shown in Figures 4.27 and 4.28.

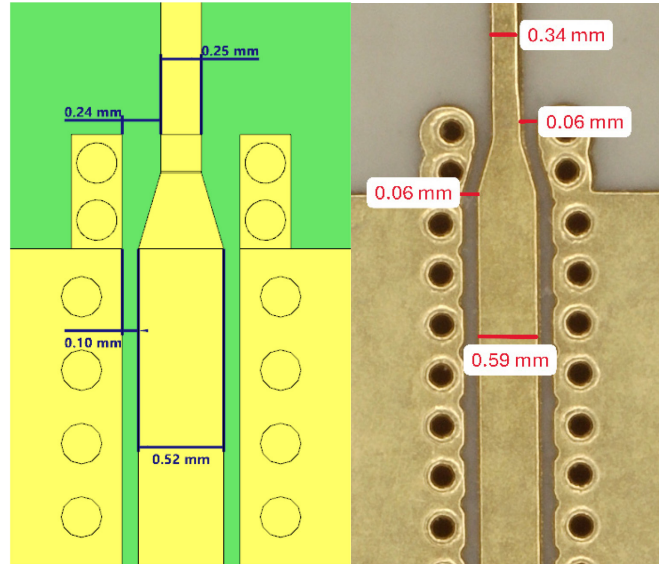


Figure 4.27 – Comparison between the simulated GCPW and the manufactured GCPW.

Figure 4.27 shows that all the design parameters of the GCPW-to-microstrip transition were altered during the fabrication process. In particular, with reference to the numerical values reported in the figure, it can be observed that all the design-parameter values of the GCPW-to-microstrip transition differ from the intended ones. It can also be noted that the width of the line section used to feed the patch is approximately 40 % larger than the corresponding design value. Similarly, the gap between the central strip of the GCPW and the two lateral ground planes was reduced by approximately 40%. Similar deviations can also be observed in the antenna feeding region. In particular, the deviations affecting the width of the feeding microstrip undoubtedly have an impact on the impedance matching, since the radial position of the feeding point is selected as a function of the characteristic impedance of the microstrip section actually connected to the patch. A similar consideration also applies to the lateral and frontal gaps in the adopted feeding technique, since a variation of approximately 20% in these parameters cannot be regarded as negligible.

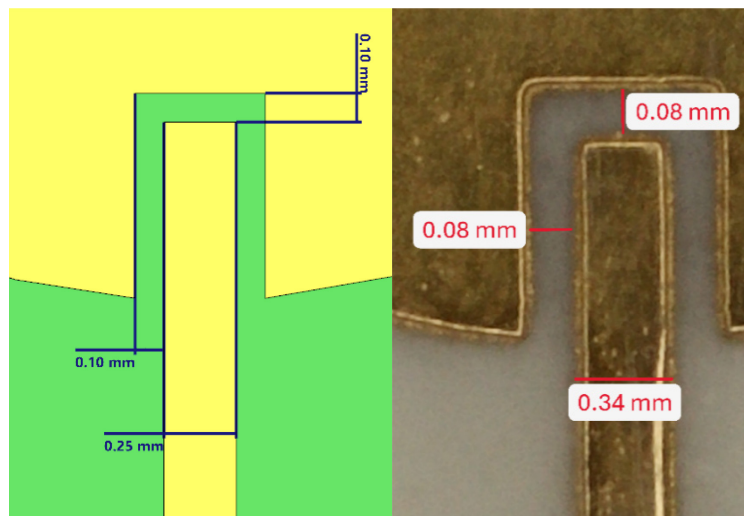


Figure 4.28 – Comparison between the simulated inset and the manufactured one.

In order to assess whether the observed dimensional deviations were responsible for the discrepancy between the numerical simulations and the experimental measurements, the model shown in Figure 4.23 was simulated again by using, as design-parameter values, those obtained from the dimensional analysis carried out on the antenna prototype shown in Figure 4.27 and Figure 4.28.

Figure 4.29 shows a comparison between the magnitude response of the measured S11 and the corresponding simulated one obtained for the model shown in Figure 4.23 after updating its design parameters as mentioned above.

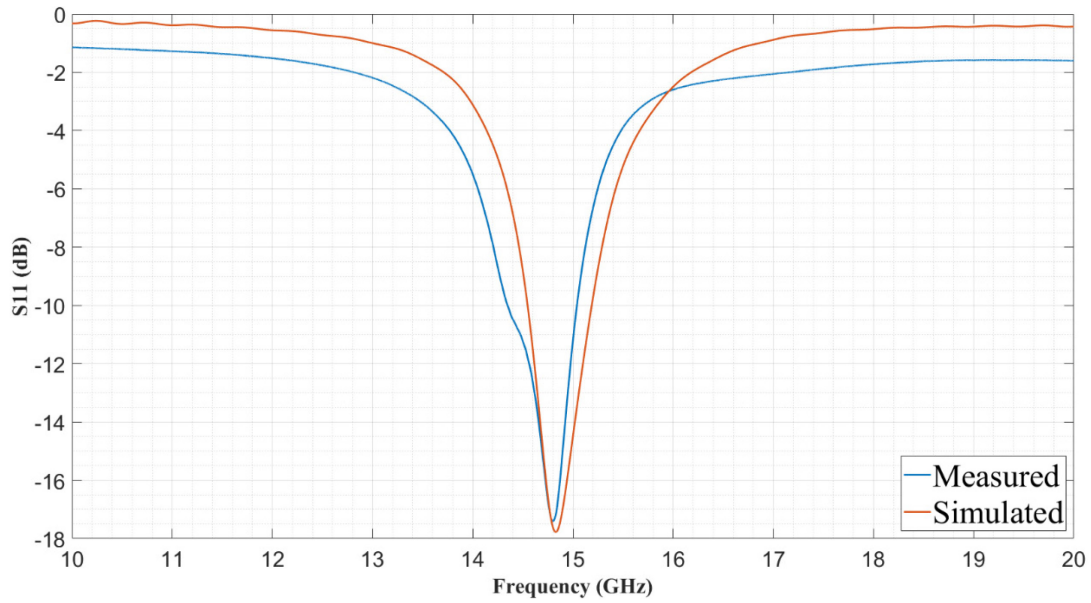


Figure 4.29 – Comparison between the magnitude response of the measured S11 and the one simulated with design parameter values obtained from the dimensional analysis on the antenna prototype.

The results obtained after updating the design parameter values of the reference model in accordance with the dimensional analysis described above confirm that the discrepancy between the experimental and numerical results highlighted in Figure 4.26 is to be attributed to the non-optimal fabrication of the antenna prototype on which the measurements were carried out. However, since a numerical model that appears to be capable of adequately reproducing the actual behavior of the fabricated prototypes was available, it was decided to perform further measurements in a controlled environment in order to assess their radiation pattern.

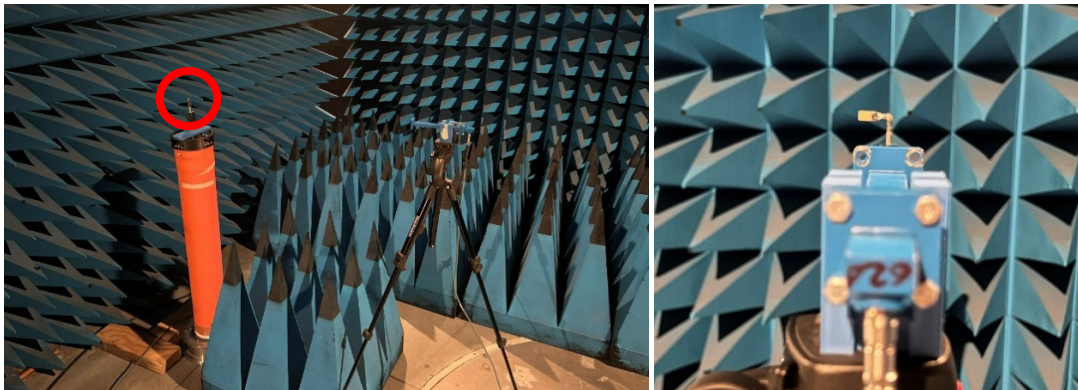


Figure 4.30 – Measurements inside the semi-anechoic chamber: $\phi=0^\circ$ (left) and $\phi=90^\circ$ (right)

More specifically, the measurements were carried out inside the semi-anechoic chamber (Figure 4.30) located in the Engineering building of the University of Cassino and Southern Lazio, using, in addition to the VNA shown in Figure 4.25, a cylindrical scanning system controlled by a Newport Universal ESP300 motion controller and a probe consisting of a WR62 waveguide operating over the $12.4 \div 18$ GHz frequency range. More precisely, by using the just mentioned instrumentation, varying the azimuth angle of the antenna under test with a discrete step of 3° , and performing two separate sets of measurements, one with the probe in vertical polarization and the other with the probe in horizontal polarization, the θ -cuts of the radiation pattern were evaluated in the two principal vertical planes, namely $\phi=0^\circ$ and $\phi=90^\circ$, according to a reference system in which the z -axis is perpendicular to the patch plane, whereas the x - and y -axes are parallel to its shorter and longer sides, respectively.

Figures 4.31 ÷ 4.34 show the comparison between the measured and simulated normalized-gain radiation patterns in the $\phi=0^\circ$ and $\phi=90^\circ$ planes at 14.6 GHz, 14.8 GHz, 15.0 GHz, and 15.2 GHz, respectively.

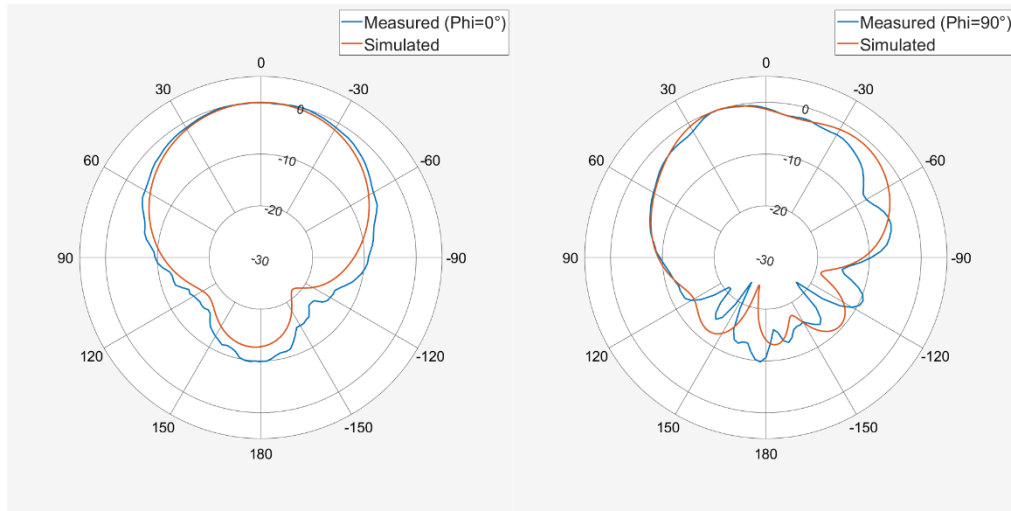


Figure 4.31 – Comparison of the measured and simulated normalized-gain radiation patterns in the $\phi=0^\circ$ (left) and $\phi=90^\circ$ (right) planes at 14.6 GHz.

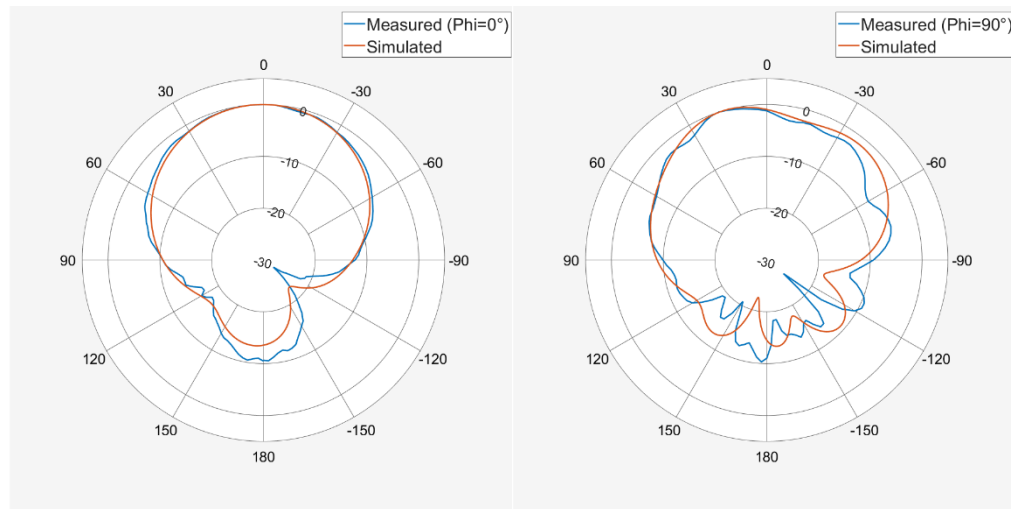


Figure 4.32 – Comparison of the measured and simulated normalized-gain radiation patterns in the $\phi=0^\circ$ (left) and $\phi=90^\circ$ (right) planes at 14.8 GHz.

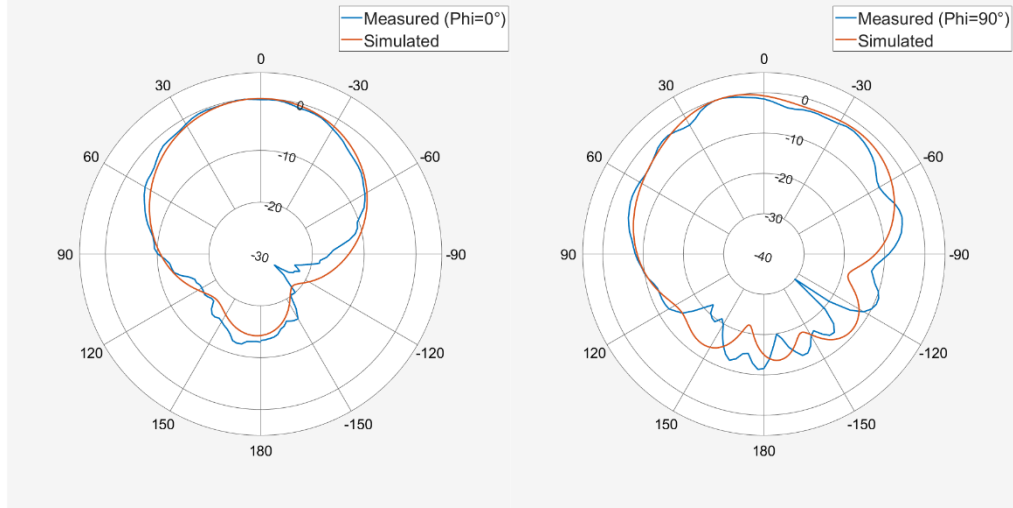


Figure 4.33 – Comparison of the measured and simulated normalized-gain radiation patterns in the $\phi=0^\circ$ (left) and $\phi=90^\circ$ (right) planes at 15.0 GHz.

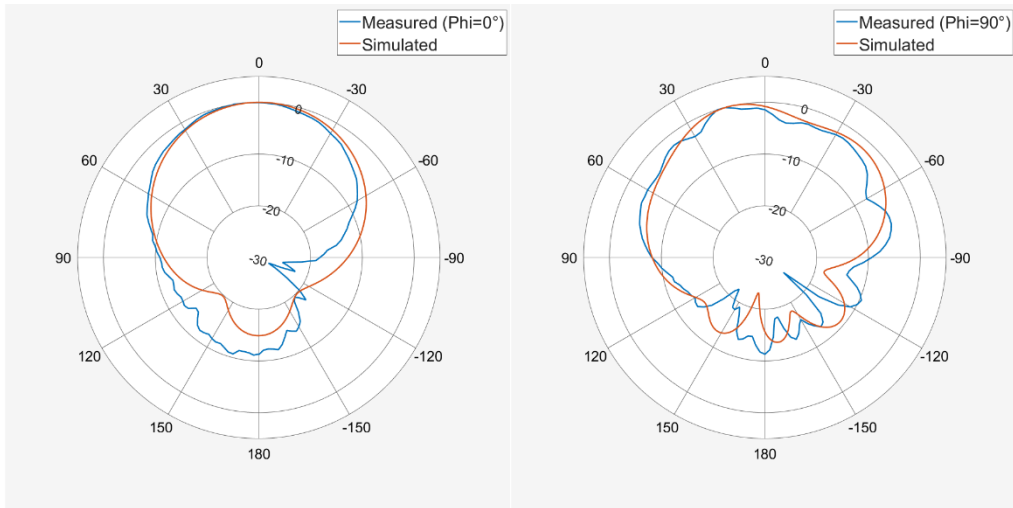


Figure 4.34 – Comparison of the measured and simulated normalized-gain radiation patterns in the $\phi=0^\circ$ (left) and $\phi=90^\circ$ (right) planes at 15.2 GHz.

The figures above demonstrate good overall agreement between the measured results and those obtained from the simulations. Indeed, the general shape of the radiation patterns is preserved in all the examined cases, the width of the main beam is fairly similar in both measurement and simulation, and the overall behavior remains reasonably stable in the vicinity of the resonant frequency of the antennas under test.

It is, however, worth observing that the comparison between the measured and simulated results is generally more regular in the $\phi=0^\circ$ plane than in the $\phi=90^\circ$ plane. Indeed, the radiation patterns evaluated in the $\phi=0^\circ$ plane exhibit, on the whole, a smoother trend and a better agreement with the corresponding numerical results, whereas those evaluated in the $\phi=90^\circ$ plane are characterized by more pronounced ripples, deeper local minima and, especially in the back-radiation region, a lower regularity of the measured pattern. This behavior can reasonably be attributed primarily to the fabrication-induced imperfections of the real prototype with respect to the geometry considered in the

simulations. Indeed, at the frequencies of interest, even small dimensional deviations that cannot be directly assessed, or fully quantified, through the dimensional analysis carried out such as the actual morphology of the copper edges, the extent and uniformity of the rounding at corners and terminations, possible local variations in the microstrip width along its path, and the trapezoidal cross-sectional profile of the metallization resulting from the etching process may locally perturb the current distribution on both the radiating structure and the feeding network, thus generating parasitic contributions that affect the measured radiation pattern. The effect of such non-idealities tends to manifest itself more clearly in the back-radiation region, where the field radiated by the patch is inherently weaker and, therefore, any residual contributions are less effectively masked by the main lobe radiation. In the present case, this effect may also be more evident in the $\phi=90^\circ$ plane, since, according to the adopted reference system, both the patch feeding line and the central conductor of the GCPW-to-microstrip transition mainly develop along the y -axis, i.e., along the direction parallel to the longer side of the patch.

It is also worth observing that, in the $\phi=0^\circ$ plane, the comparison between the measured and simulated radiation patterns is not perfectly symmetric with respect to the $\theta=0^\circ$ direction. Indeed, as shown in the figures above, for θ values between 60° and 120° a generally good agreement between experimental and numerical results is observed at all the frequencies considered. In the opposite angular sector, namely for θ values between -120° and -60° , the agreement is slightly less good, with more evident discrepancies between the two patterns, which tend to become more pronounced as the frequency increases. Such a behavior can reasonably be interpreted as indicative of the presence, in the fabricated prototype, of asymmetries with respect to the y - z plane of the adopted reference system. These asymmetries may be associated, for instance, with differences between the right- and left-hand sides in the three-dimensional profile of the metallization resulting from the fabrication process, as well as with slight out-of-plane deformations of the board and/or other prototype non-idealities that cannot be directly identified through the top-view microscopic analysis discussed above. Indeed, effects of this kind may alter the symmetry of the structure with respect to the y - z plane and, consequently, affect the radiation pattern measured in the $\phi=0^\circ$ plane.

In light of the considerations discussed above, it can therefore be concluded that the experimental results substantially confirm the radiation behavior predicted by the EM model of the antenna under examination. Indeed, in both principal planes considered, the measured radiation patterns reproduce with satisfactory accuracy the general behavior of the corresponding simulated patterns, thereby confirming the main-lobe configuration and the directivity characteristics of the antenna over the analyzed frequency range. It may therefore be stated that the measurements performed provide, overall, a significant confirmation of both the validity of the developed antenna design and the reliability of the numerical model of the antenna adopted to predict its radiative behavior.

5 Summary and conclusions

In this thesis, a proof-of-concept design for a compact, low-cost wearable device for the assessment of EMF exposure levels in prospective 6G scenarios in the $14.8 \div 15.35$ GHz frequency range was developed. This work was motivated by the intention to anticipate future needs by providing, already in the early stages of the development of new 6G communication networks, a measurement solution specifically conceived for frequency bands and propagation conditions that are expected to play a significant role in future wireless communication systems, thereby contributing, at least in part, to overcoming one of the main critical issues that may hinder the development of new communication technologies, especially in their initial stages, namely the lack of adequate instruments for measuring the levels of the EMF associated with the signals of interest. It is worth highlighting that the frequency range was selected as one of the leading candidates indicated by WRC-27, but the proposed architecture can be adapted to different frequency ranges, even in a multi-frequency configuration.

Within this framework, an architecture based on a planar radiating element, an RMS RF power detector, and a digital acquisition section was defined, with particular attention devoted to the design and validation of the antenna, which represents the most critical and original element of the proposed system, since its radiation characteristics, physical footprint, and fabrication cost directly affect both the measurement performance of the target device and the overall compactness and economic viability of the proposed solution. For this reason, a significant portion of the work was devoted to the design and optimization of the radiating element. In particular, through an extensive numerical-simulation activity, a single-microstrip-fed circularly polarized patch antenna, with overall dimensions of $22.00 \text{ mm} \times 12.70 \text{ mm}$ and operating at 15 GHz on a 0.508 mm-thick Rogers RO4003C substrate, was developed and optimized in order to meet the main requirements of the target application in terms of compactness, cost-effective manufacturability, polarization behavior, and radiation performance.

On the basis of the obtained design, the antenna prototype was then fabricated and experimentally characterized in order to verify whether its radiative behavior was consistent with that predicted by the EM simulations performed on the numerical model of the antenna. To this end, the radiation patterns of the fabricated prototype were measured in a semi-anechoic chamber and subsequently compared with the corresponding simulated ones in the two principal planes, namely $\phi=0^\circ$ and $\phi=90^\circ$, defined for the considered antenna by the broadside direction and, respectively, by the directions along the minor and major dimensions of the antenna layout.

The comparison showed good overall agreement between numerical and experimental results. More specifically, in both principal planes the measured radiation patterns reproduced the general shape of the corresponding simulated ones, including the main-lobe configuration and the main directivity features of the antenna. In particular, the measured results confirmed that, in the $\phi=0^\circ$

plane, the main lobe is oriented close to the broadside direction, while in the $\phi=90^\circ$ plane a slight angular deviation is observed as predicted by the numerical analysis. The observed discrepancies are mainly confined to limited angular regions, especially in the back-radiation region, and therefore do not affect the overall agreement between measurements and simulations. On this basis, it may be stated that the measurements performed provide, overall, a significant confirmation of both the validity of the developed antenna design and the reliability of the numerical model of the antenna adopted to predict its radiative properties.

This result is particularly relevant in the overall perspective of the proposed device. Indeed, the antenna experimentally characterized in this work preserves the same patch geometry, feeding microstrip, and substrate dimensions as the antenna intended for use in the proposed wearable personal exposimeter, namely the antenna directly connected to the RMS RF power detector integrated on the same PCB. Therefore, the experimental validation of the numerical simulations carried out for the design of the analyzed antenna prototypes provides a sound basis for considering as realistic also the numerical results obtained for the integrated antenna-detector structure. On this basis, the simulated response of the integrated antenna-detector system can be combined with the input-output characteristics of the adopted detector in order to derive first quantitative estimates of the expected performance of the measurement device. More specifically, by referring to the lower end of the useful input range of the adopted RMS RF power detector and to the simulated realized gain of the integrated antenna-detector system, the minimum detectable field strength is found to be approximately $0.80 \text{ V}\cdot\text{m}^{-1}$ at 15 GHz. If a larger dynamic range is required, a LNA can be easily integrated in the device. In addition, the combination of the slope of the linear-in-dB response of the adopted detector and the resolution of an ADC commonly integrated into low-cost microcontrollers (12 bit with 3.3 V reference voltage) makes it possible to estimate a minimum distinguishable input-level difference of about 0.028 dB, corresponding to a minimum distinguishable relative variation of the incident field of approximately 0.32%. These values further support the suitability of the proposed device for the assessment of EMF exposure levels associated with the signals of interest within the scope of the present work.

On the basis of the results discussed above, the proof-of-concept design developed in this work may therefore be considered as fully validated within the scope of the present thesis. Indeed, the integrated antenna-detector structure is characterized by an overall footprint of only $22.00 \text{ mm} \times 12.70 \text{ mm}$ and by an estimated total cost of about 90 €. Owing to its limited size and low cost, the proposed integrated antenna-detector unit can be adopted as the elementary sensing block of a multi-antenna exposimeter, for instance in a bracelet-like arrangement, in order to improve the robustness of the received signal under different exposure conditions. Furthermore, since low-cost microcontrollers (f.i. the STM32L433RTCP6 identified in Section 4.3) typically provide several ADC input channels, a single unit can, in principle, be used to process the outputs of multiple antenna-detector structures, while contributing only a few euros to the overall cost. As a consequence, the total cost of a multi-antenna implementation can still be kept well below 1000 €. This conclusion is further reinforced by the fact that the estimated cost of about 90 € refers to a single integrated antenna-detector unit and can already be expected to decrease for moderate procurement volumes, owing to the significant reduction in the unit price of the adopted RMS RF power detector and to the lower PCB fabrication cost achievable in larger-scale production. Overall, these considerations confirm the feasibility of a compact and relatively low-cost solution for the intended application, strengthen the practical relevance of the

proposed architecture, and support its future development toward a fully integrated wearable personal exposimeter.

The full development of such an architecture, however, will require the complete definition of 6G signal specifications, as well as the availability of signal generators capable of producing waveforms compliant with those specifications. Indeed, once these conditions are met an adequate calibration of the proposed device may be carried out, both in controlled environments and in open-field scenarios, using high-end instrumentation as a benchmark for the assessment of the overall performance of the system. This activity will constitute the natural next step toward the complete experimental validation of the proposed architecture and its future implementation as a wearable personal exposimeter for prospective 6G scenarios

List of publications

- D. Pinchera, F. Schettino, M. Lucido, G. CHIRICO, and M. D. Migliore, *“Design and Realization of an Ultra-Wideband, Pattern-Stable Antenna for Ground Sensing Applications with UAVs”*, Applied Sciences, Vol. 16, Issue 3, 1159, Jan. 2026;
- M. L. Iványi, G. CHIRICO, Y. Aslan, A. Yarovoy, and M. Spirito, *“Performance Characterization of an Active Phased Array Antenna by Simultaneously Measuring the Radiation Pattern and the Error Vector Magnitude”*, 2025 55th European Microwave Conference (EuMC), Utrecht, Sept. 2025;
- R. Otin, E. Soudah, G. CHIRICO, F. Schettino, and N. Betzalel, *“Computational Human Skin Model for 5G Electromagnetic Field Exposure Studies”*, 2025 IEEE International Workshop on Electromagnetics: Applications and Student Innovation Competition (iWEM), Hong Kong, Aug. 2025 (honorable mention award for the contribution of this paper in the best paper competition);
- A. Sannino, M. Allocca, M. R. Scarfi, S. Romeo, V. Peluso, G. Panariello, F. Schettino, G. CHIRICO, and O. Zeni, *“Exposure to 26.5 GHz, 5G modulated and unmodulated signal, does not affect key cellular endpoints of human neuroblastoma cells”*, Scientific Reports, Vol. 15, Issue 1, 20614, July 2025;
- N. D’Ambrosio, F. Di Sio, A. Esposito, F. Lodato, R. Massa, G. CHIRICO, and F. Schettino, *“Microwave-Induced Inhibition of Germination in Portulaca oleracea L. Seeds”*, Agronomy, Vol. 15, Issue 10, 2418, Oct. 2025;
- D. Pinchera, F. Schettino, M. Lucido, G. CHIRICO, and M. D. Migliore, *“Numerical Investigation and Design Curves for Thinned Planar Antenna Arrays for 5G and 6G”*, Electronics, Vol. 13, Issue 23, 4711, Nov. 2024;
- G. CHIRICO, C. J. D’Silva, D. Pinchera and F. Schettino, *“Analysis of the Contribution of Different Generations of Cellular Communication Systems to the EMF Levels in a Small City”*, 2024 IEEE International Symposium on Measurements & Networking (M&N), pp. 1-5, Rome, Italy, July 2024;
- M. Lucido, G. A. Casula, G. CHIRICO, M. D. Migliore, D. Pinchera, and Fulvio Schettino, *“Helmholtz–Galerkin Technique in Dipole Field Scattering from Buried Zero-Thickness Perfectly Electrically Conducting Disk”*, Applied Sciences 2024, Vol. 14, Issue 13, 5544, June 2024;

- R. Massa, F. Lodato, F. Carraturo, A. Nappo, M. Annunziata, M. Guida, G. CHIRICO et al., *"Preliminary Results on Fungicidal Efficacy of Microwave Treatment on Fusarium Oxysporum"*, 2024 4th URSI Atlantic Radio Science Meeting, AT-RASC 2024, Gran Canaria, Spain, 19 – 24 May 2024;
- M. Lucido, G. CHIRICO, M. D. Migliore, D. Pinchera, F. Schettino, *"Plane-Wave Diffraction from Resistive-Filled Circular Hole in Infinite Resistive Plane: an Analytically Regularizing Approach"*, Applied Sciences (Switzerland), vol. 13, Issue 13, June 2023;
- R. Massa, F. Schettino, G. Panariello, M.D. Migliore, D. Pinchera, G. CHIRICO, C. J. D'Silva, R. Griffio, T. Yaseen, *"Using Microwave Technology against Red Palm Weevil: an Innovative Sustainable Strategy to Contrast a Lethal Pest of Date Palms"*, Acta Horticulturae, June 2023;
- G. CHIRICO, C. D'Elia, N. D'Ambrosio, R. Massa, *"Design and Evaluation of an Applicator for Magnetopriming Treatments"*, IEEE Journal of Electromagnetics, RF and Microwaves in Medicine and Biology, Apr. 2023;
- D. Pinchera, M. Lucido, G. CHIRICO, F. Schettino, M. D. Migliore, *"Controllable Local Propagation Environment to Maximize the Multiplexing Capability of Massive MIMO Systems"* - Electronics (Switzerland), 12(9), Apr. 2023;
- F. Schettino, G. CHIRICO, C. D'Elia, M. Lucido, D. Pinchera, M. D. Migliore, *"A Simple and Low-Cost Technique for 5g Conservative Human Exposure Assessment"*, Applied Sciences (Switzerland), 13(6), Mar. 2023;

Bibliography

- [1] F. Schettino et. al, NextGEM deliverable 2.2;
URL: <https://www.nextgem.eu/dissemination-communication/public-deliverables/>
- [2] M. Vrijheid et al., "*Determinants of mobile phone output power in a multinational study: implications for exposure assessment*", Occupational and environmental medicine, 66(10), pp. 664-671, June 2009;
- [3] M. Kelsh et al., "*Measured radiofrequency exposure during various mobile-phone use scenarios*", Journal of Exposure Science & Environmental Epidemiology, 21(4), pp. 343-354, July 2011;
- [4] V. K. Garg, and Y. C. Wang, "*Wireless network access technologies*", in The Electrical Engineering Handbook, Academic Press, 2005, pp. 1005-1009;
- [5] K. Asatani, "*Trends and Issues in 5G Networking and Beyond*", Journal of ICT Standardization, 5(3), pp. 203-224, Jan. 2018;
- [6] ETSI Technical Specifications, "*Universal Mobile Telecommunications System (UMTS), Physical layer - general description*", 3GPP TS 25.201, version 17.0.0, Release 17, May 2022;
- [7] B. Yang, "*Equipment in the LTE Network*", China Academy of Information and Communications Technology, ITU, CAICT 2016;
- [8] The Evolution of Mobile Technologies: 1G -> 2G -> 3G -> 4G LTE, Qualcomm, June 2014;
URL: https://www.qualcomm.com/content/dam/qcomm-martech/dm-assets/documents/the_evolution_of_mobile_technologies-wireless-networks.pdf
- [9] LTE Air-Interface, AIRCOM LTE Webinar Series;
URL: <https://www.slideshare.net/AIRCOMmarketing/aircom-lte-webinar-series-lte-air-interface>
- [10] A. K. Lee, S. B. Jeon, and H. D. Choi, "*EMF levels in 5G new radio environment in Seoul, Korea*", IEEE Access, 9, pp. 19716-19722, Jan. 2021;
- [11] S. Adda et al., "*A theoretical and experimental investigation on the measurement of the electromagnetic field level radiated by 5G base stations*", IEEE Access, vol. 8, pp. 101448-101463, May 2020;
- [12] H. Miao et al., "*6G New Mid-Band/FR3 (6–24 GHz): Channel Measurement, Characteristics and Modeling*", IEEE Open Journal of the Communications Society, vol. 6, pp. 9942-9960, Nov. 2025;
- [13] J. Zhang et al., "*New mid-band for 6G: several considerations from the channel propagation characteristics perspective*", IEEE Communication Magazine, vol. 63, pp. 175–180, Jan. 2025;
- [14] P. Tang et al., "*Preliminary perspectives on 3GPP standardization of the propagation channel model for FR3 bands for NR*", Science. China Information Sciences, vol. 68, 137301, Feb. 2025;

- [15] ITU Radiocommunication Sector (ITU-R), Working Party 5D (WP 5D) - IMT Systems;
- [16] A. P. French, *"An Introduction to Quantum Physics"*, 1st ed., London: Routledge, 1978;
- [17] C. Y. Yong, T. A. Rahman, and K. M. Chew, *"Wireless industry emission: Electromagnetic field monitoring and analysis"*, ARPN Journal of Engineering and Applied Sciences, Vol. 10, Issue 20, pp. 9800–9807, Nov. 2015;
- [18] C. R. Arumainayagam et al., *"Extraterrestrial prebiotic molecules: photochemistry vs. radiation chemistry of interstellar ices"*, Chemical Society Reviews, Vol. 48, Issue 8, pp. 2293–2314, Feb. 2019;
- [19] ICNIRP, *"Guidelines for limiting exposure to electromagnetic fields (100 kHz to 300 GHz)"*, Health Physics, Vol. 118, Issue 5, pp. 483–524, 2020;
- [20] ICNIRP, *"Guidelines for limiting exposure to time-varying electric, magnetic, and electromagnetic fields (up to 300 GHz)"*, Health Physics, Vol. 74, Issue 4, pp. 494–522, 1998;
- [21] ICNIRP, *"Guidelines for limiting exposure to time-varying electric and magnetic fields (1 Hz to 100 kHz)"*, Health Physics, Vol. 99, Issue 6, pp. 818–836, 2010;
- [22] D. Andreuccetti, et al., *"Protezione dai campi elettromagnetici non ionizzanti"*, 3rd ed., Firenze, Istituto di ricerca sulle onde elettromagnetiche "Nello Carrara", 2001;
- [23] A. Nguyen, *"The Potential Implementation of Radio-Frequency Identification Technology for Personal Health Examination and Monitoring"*, McGill Journal of Medicine, Vol. 12, Issue 2, pp. 67–72, Aug. 2020;
- [24] C. C. Johnson, and A. W. Guy, *"Nonionizing Electromagnetic Wave Effects in Biological Materials and Systems"*, Proceedings of the IEEE, Vol. 60, Issue 6, pp. 692–718, June 1972;
- [25] L. M. Mir, *"Application of Electroporation Gene Therapy: Past, Current, and Future"*, Methods in molecular biology, Vol. 423, pp. 3–17, Feb. 2008;
- [26] American Conference of Governmental Industrial Hygienists, *"TLVs and BEIs: based on the documentation of the threshold limit values for chemical substances and physical agents and biological exposure indices"*, Cincinnati, Ohio: ACGIH; 2022;
- [27] W. P. Cheshire Jr., *"Thermoregulatory disorders and illness related to heat and cold stress"*, Autonomic Neuroscience, Vol. 196, pp. 91–104, Apr. 2016;
- [28] P. S. Yarmolenko et al., *"Thresholds for thermal damage to normal tissues: an update"*, International Journal of Hyperthermia, Vol. 27, Issue 4, pp. 320–343, May 2011;
- [29] G. C Van Rhoon et al., *"CEM43°C thermal dose thresholds: a potential guide for magnetic resonance radiofrequency exposure levels?"*, European Radiology, Vol. 23, Issue 8, pp. 2215–2227, Aug. 2013;
- [30] M. W. Dewhirst et al., *"Basic principles of thermal dosimetry and thermal thresholds for tissue damage from hyperthermia"*, International Journal of Hyperthermia, Vol. 19, Issue 3, pp. 267–294, May–June 2003;
- [31] A. Hirata, and O. Fujiwara, *"The correlation between mass-averaged SAR and temperature elevation in the human head model exposed to RF near-fields from 1 to 6 GHz"*, Physics in Medicine & Biology, Vol. 54, Issue 23, pp. 7171–7182, Nov. 2009;
- [32] D. Funahashi, A. Hirata, S. Kodera, and K. R. Foster, *"Area-Averaged Transmitted Power Density at Skin Surface as Metric to Estimate Surface Temperature Elevation"*, IEEE Access, Vol. 6, pp. 77665–77674, Dec. 2018;

- [33] Y. Hashimoto et al., "On the averaging area for incident power density for human exposure limits at frequencies over 6 GHz", *Physics in Medicine & Biology*, Vol. 62, Issue 8, pp. 3124-3138, Mar. 2017;
- [34] K. Sasaky et al., "Monte Carlo simulations of skin exposure to electromagnetic field from 10 GHz to 1 THz", *Physics in Medicine & Biology*, Vol. 62, Issue 17, pp. 6693-7010, Aug. 2017;
- [35] Scientific Committee on Health, Environmental, and Emerging Risks (SCHEER), "The need of a revision of the annexes in the Council Recommendation 1999/519/EC and Directive 2013/35/EU, in view of the latest scientific evidence available with regard to radiofrequency (100kHz - 300GHz)", Sept. 2022; URL:https://health.ec.europa.eu/consultations/scheer-public-consultation-preliminary-opinion-scientific-evidence-radiofrequency_en
- [36] International Commission on Non-Ionizing Radiation Protection (ICNIRP), "Differences between the ICNIRP (2020) and previous guidelines"; URL:<https://www.icnirp.org/en/differences.html>
- [37] Publications Office of the European Union, CELEX1, 1999/519/EC: "Council Recommendation of 12 July 1999 on the limitation of exposure of the general public to electromagnetic fields (0 Hz to 300 GHz)", 1999; URL:<https://op.europa.eu/en/publication-detail/-/publication/9509b04f-1df0-4221-bfa2-c7af77975556/language-en>
- [38] Publications Office of the European Union, "Directive 2013/35/EU of the European Parliament and of the Council of 26 June 2013 on the minimum health and safety requirements regarding the exposure of workers to the risks arising from physical agents (electromagnetic fields)", 2013; URL:<https://eur-lex.europa.eu/eli/dir/2013/35/oj>
- [39] Decreto del Presidente del Consiglio dei Ministri (DPCM) 8 luglio 2003, "Fissazione dei limiti di esposizione, dei valori di attenzione e degli obiettivi di qualità per la protezione della popolazione dalle esposizioni a campi elettrici, magnetici ed elettromagnetici generati a frequenze comprese tra 100 kHz e 300 GHz", GU Serie Generale Anno 144° - Numero 199; Aug. 2003; URL:<https://www.gazzettaufficiale.it/eli/id/2003/08/28/03A09711/sg>
- [40] LEGGE 30 dicembre 2023, n. 214, "Legge annuale per il mercato e la concorrenza 2022", GU Serie Generale Anno 164° - Numero 303, Dec. 2023; URL:<https://www.normattiva.it/uri-res/N2Ls?urn:nir:stato:legge:2023-12-30;214>
- [41] EN 50413, "Basic standard on measurement and calculation procedures for human exposure to electric, magnetic and electromagnetic fields (0 Hz – 300 GHz)", 2021; URL:<https://standards.iteh.ai/catalog/standards/clc/52672ee1-158b-442c-9982-6dad5f991869/en-50413-2019>
- [42] EN 50383, "Basic standard for the calculation and measurement of electromagnetic field strength and SAR related to human exposure from radio base stations and fixed terminal stations for wireless telecommunication systems (110 MHz - 40 GHz)", 2013; URL:<https://standards.iteh.ai/catalog/standards/clc/50044370-081e-4453-98a8-13f97b304e74/en-50383-2010-ac-2013>
- [43] IEC 62232, "Determination of RF field strength, power density and SAR in the vicinity of base stations for the purpose of evaluating human exposure", International Electrotechnical Commission, 2022; URL:<https://webstore.iec.ch/en/publication/64934>

- [44] EN 50401, “Product standard to demonstrate the compliance of base station equipment with radiofrequency electromagnetic field exposure limits (110 MHz – 100 GHz), when put into service”, 2017; URL:<https://standards.iteh.ai/catalog/standards/clc/2f86cba5-4e47-43e8-bb68-d039e1557623/en-50401-2017>
- [45] J. Beekhuizen et al., “Geospatial modelling of electromagnetic fields from mobile phone base stations”, *Science of The Total Environment*, Vol. 445–446, pp. 202–209, Feb. 2013;
- [46] M. Velghe et al., “Characterisation of spatial and temporal variability of RF-EMF exposure levels in urban environments in Flanders, Belgium”, *Environmental Research*, Vol. 175, pp. 351–366, Aug. 2019;
- [47] M. Eeftens et al., “Personal exposure to radio-frequency electromagnetic fields in Europe: Is there a generation gap?”, *Environment International*, Vol. 121, pp. 216–226, Dec. 2018;
- [48] S. Aerts, J. Wiart, L. Martens, and W. Joseph, “Assessment of long-term spatio-temporal radiofrequency electromagnetic field exposure”, *Environmental Research*, Vol. 161, pp. 136–143, Feb. 2018;
- [49] I. Liorni et al., “Evaluation of specific absorption rate in the farfield, near-to-far field and near-field regions for integrative radiofrequency exposure assessment”, *Radiation Protection. Dosimetry*, Vol. 190, Issue 4, pp. 459–472, Oct. 2020;
- [50] E. Korkmaz et al., “A comprehensive review of 5G NR RF-EMF exposure assessment technologies: fundamentals, advancements, challenges, niches, and implications”, *Environmental Research*, Vol. 260, Nov. 2024;
- [51] A. Thielens et al., “Representativeness and repeatability of microenvironmental personal and head exposures to radio-frequency electromagnetic fields”, *Environmental Research*, Vol. 162, pp. 81–96; Apr. 2018;
- [52] M. Rösli et al., “Conduct of a personal radiofrequency electromagnetic field measurement study: proposed study protocol”, *Environmental Health*, Vol. 9, Issue 1, May 2010;
- [53] C. R. Bhatt et al., “Assessment of personal exposure from radiofrequency-electromagnetic fields in Australia and Belgium using on-body calibrated exposimeters” *Environmental Research*, Vol. 151, pp. 547–563, Nov. 2016;
- [54] H. Jalilian, M. Eeftens, M. Ziaei, M. Rösli, “Public exposure to radiofrequency electromagnetic fields in everyday microenvironments: An updated systematic review for Europe”, *Environmental Research*, Vol. 176, Sep. 2019;
- [55] S. Sagar et al., “Comparison of radiofrequency electromagnetic field exposure levels in different everyday microenvironments in an international context”. *Environment International*, Vol. 114, pp. 297–306, May 2018;
- [56] G. Vermeeren et al., “Spatial and temporal RF electromagnetic field exposure of children and adults in indoor micro environments in Belgium and Greece”, *Progress in Biophysics and Molecular Biology*, Vol. 113, Issue 2, pp. 254–263, Nov. 2013;
- [57] C. R. Bhatt et al., “Measuring personal exposure from 900MHz mobile phone base stations in Australia and Belgium using a novel personal distributed exposimeter”, *Environment International*, Vol. 92–93, pp. 388–397; Aug. 2016;

- [58] J. F. B. Bolte, and T. Eikelboom, “Personal radiofrequency electromagnetic field measurements in the Netherlands: Exposure level and variability for everyday activities, times of day and types of area”, *Environment International*, Vol. 48, pp. 133–142, Nov. 2012;
- [59] P. Frei et al., “Temporal and spatial variability of personal exposure to radio frequency electromagnetic fields”, *Environmental Research*, Vol. 109, Issue 6, pp. 779–785, Aug. 2009;
- [60] D. Urbinello et al., “Use of portable exposure meters for comparing mobile phone base station radiation in different types of areas in the cities of Basel and Amsterdam”, *Science of The Total Environment*, Vol. 468–469, pp. 1028–1033, Jan. 2014;
- [61] L. Van Wel et al., “Context-sensitive ecological momentary assessments; Integrating real-time exposure measurements, data-analytics and health assessment using a smartphone application”, *Environment International*, Vol. 103, pp. 8–12; June 2017;
- [62] S. Aerts et al., “Lessons learned from a distributed RF-EMF sensor network”, *Sensors*, Vol. 22, Issue 5, 1715; Feb. 2022;
- [63] S. Iakovidis, C. Apostolidis, A. Manassas, and T. Samaras, “Electromagnetic fields exposure assessment in Europe utilizing publicly available data”, *Sensors*, Vol. 22, Issue 21, 8481, Nov. 2022;
- [64] A. Gotsis et al., “Non-ionizing electromagnetic radiation monitoring in Greece”. *Annals of Telecommunications*, Vol. 63, pp. 109–123; Jan. 2008;
- [65] A. Manassas, A. Boursianis, T. Samaras, and J. N. Sahalos, “Continuous electromagnetic radiation monitoring in the environment: Analysis of the results in Greece”, *Radiation Protection Dosimetry*, Vol. 151, Issue 3, pp. 437–442, Sep. 2012;
- [66] C. Oliveira et al., “The moniT Project: Electromagnetic radiation exposure assessment in mobile communications” *IEEE Antennas and Propagation Magazine*, Vol. 49, Issue 1, pp. 44–53, Mar. 2007;
- [67] J. T. Rowley, and K. H. Joyner, K.H., “Observations from national Italian fixed radiofrequency monitoring network”, *Bioelectromagnetics*, Vol. 37, Issue 2, pp. 136–139, Feb. 2016;
- [68] L. Díez, R. Agüero, and L. Muñoz, “Electromagnetic field assessment as a smart city service: The SmartSantander use-case”, *Sensors*, Vol. 17, Issue 6, 1250, May 2017;
- [69] Djurić et al., “The EMF exposure monitoring in cellular networks by Serbian EMF RATEL system”, 2022 IEEE International Symposium on Measurements and Networking (M&N), Padua, Italy, pp. 1-6, July 2022;
- [70] G. Betta et al., “Human exposure to 5G systems: Experimental analysis and measurement issues in FR1 and FR2 operating bands”, 2023 IEEE International Instrumentation and Measurement Technology Conference (I2MTC), pp. 01–06, May 2023;
- [71] S. Liu et al., “E-Field Strength Measurements of a 5G Base Station in 28 GHz Band for EMF Exposure Assessment”, 2021 IEEE USNC-URSI Radio Science Meeting (Joint with AP-S Symposium), Singapore, Singapore, pp. 49-50; Dec. 2021;
- [72] The Rohde & Schwarz signal and spectrum analyzer portfolio, URL:https://www.rohde-schwarz.com/us/products/test-and-measurement/signal-and-spectrum-analyzers_63665.html
- [73] FieldFox Handheld RF and Microwave Analyzer, URL:<https://www.keysight.com/us/en/catalog/key-34752/fieldfox-handheld-rf-microwave-analyzers.html>
- [74] Aaronia SPECTRAN V5X, Datasheet, <https://downloads.aaronia.com/datasheets/analyzers/V5/V5-X.pdf>

- [75] Narda FieldMan datasheet, URL:https://www.narda-sts.com/fileadmin/DE/Downloads_D_EN/Produkt_Dokumente_Datenblaetter_AN_TN_Flyer_D_EN/FieldMan/Datenblaetter_CE/Datasheet_FieldMan.pdf
- [76] Narda Probe EFD-0691 datasheet, URL:https://www.gruppompb.com/public/upload/EFD-0691_DATASHEET.pdf
- [77] Wavecontrol WPF6 Probe datasheet, URL:https://users.wavecontrol.com/rfsafety/images/data-sheets/en/WPF6_Datasheet_EN.pdf
- [78] Narda Probe EFD-9091 datasheet, URL:https://www.narda-sts.com/fileadmin/DE/Downloads_D_EN/Produkt_Dokumente_Datenblaetter_AN_TN_Flyer_D_EN/FieldMan/Datenblaetter_CE/Datasheet_EFD-9091.pdf
- [79] F. Mavromatis et al., "A Broadband Monitoring System for Electromagnetic-Radiation Assessment", IEEE Antennas and Propagation Magazine, Vol. 51, Issue 1, pp. 71–79, May 2009;
- [80] F. Minucci et al., "Measuring 5G electric fields strength with software defined radios", IEEE Open Journal of the Communications Society, Vol. 3, pp. 2258–2271, Nov. 2022;
- [81] S. Aerts et al., "In-situ measurement methodology for the assessment of 5G NR massive MIMO base station exposure at sub-6 GHz frequencies". IEEE Access, Vol. 7, pp. 184658–184667; Dec. 2019;
- [82] I. Santiago Rivera, S. Vidal Beltrán, and F. Martínez-Piñón, "Spectrum analyzer by software defined radio", 2018 International Conference on Mechatronics, Electronics and Automotive Engineering (ICMEAE), pp. 93–97, Nov. 2018;
- [83] K. Deprez et al., "Comparison of low-cost 5G electromagnetic field sensors", Sensors, Vol. 23, Issue 6, 3312, Mar. 2023;
- [84] A.C. Bechet et al., "Low cost solution based on software defined radio for the RF exposure assessment: A performance analysis", 2019 11th International Symposium on Advanced Topics in Electrical Engineering (ATEE), pp. 1–4, May. 2019;
- [85] R. Helbet, P. Bechet, S. Miclaus, and A. Sarbu, "Real time broadband electromagnetic spectrum monitoring system based on software defined radio technology", 2021 9th International Conference on Modern Power Systems (MPS), pp. 1–6, June 2021;
- [86] A. Sârbu et al., "SDR-Enabled multichannel real-time measurement system for in situ EMF exposure evaluation", Electronics, Vol. 11, Issue 17, 2670, Aug. 2022;
- [87] L. F. Diez et al., "Design and integration of a low-complexity dosimeter into the smart city for EMF assessment", 2014 European Conference on Networks and Communications (EuCNC), pp. 1–5, Aug. 2014;
- [88] K. Deprez et al., "Design of a low-cost modular 5G RF-EMF exposure sensor", Joint Annual Meeting of the Bioelectromagnetics Society and the European BioElectromagnetics Association, pp. 421–426; Sept. 2021;
- [89] J. Van der Straeten et al, "Design and Validation of a Low-Cost Triaxial 5G RF-EMF Exposure Sensor", IEEE Sensors Journal, Vol. 25, Issue 9, pp. 16050-16060, May 2025;
- [90] D. Kwon et al., "Loop-Type field probe to measure human body exposure to 5G Millimeter-Wave base stations", Applied Science, Vol. 13, Issue 21, 11777, Oct. 2023;
- [91] E. Korkmaz, S. Littel, M. Spirito, and J. Bolte, "A low-cost EMF field sensor for 5G EMF exposure measurements", Joint Annual Meeting of the Bioelectromagnetics Society and the European BioElectromagnetics Association, June 2023;

- [92] F. Viani et al., "A WSN-based system for real-time electromagnetic monitoring", 2011 IEEE International Symposium on Antennas and Propagation (APSURSI), pp. 3129–3132, July 2011;
- [93] F. Viani, A. Polo, M. Donelli, and E. Giarola, "A relocable and resilient distributed measurement system for electromagnetic exposure assessment", IEEE Sensors Journal, Vol. 16, Issue 11, pp. 4595–4604, Mar. 2016;
- [94] L. Ioriatti et al., "Real-time distributed monitoring of electromagnetic pollution in urban environments". 2009 IEEE International Geoscience and Remote Sensing Symposium, pp. V–100–V–103, July 2009;
- [95] P. Vanveerdeghem et al., "Compact personal distributed wearable exposimeter", IEEE Sensors Journal, Vol. 15, Issue 8, pp. 4393–4401, Apr. 2015;
- [96] A. Thielens et al., "A personal, distributed exposimeter: Procedure for design, calibration, validation, and application", Sensors, Vol. 16, Issue 2, 180, Feb. 2016;
- [97] W. Joseph et al., "Drone based measurement system for radiofrequency exposure assessment", Bioelectromagnetics, Vol. 37, Issue 3, pp. 195–199, Apr. 2016;
- [98] M. X. Rivera González et al., "Compact exposimeter device for the characterization and recording of electromagnetic fields from 78 MHz to 6 GHz with several narrow bands (300 kHz)", Sensors, Vol. 21, Issue 21, 7395, Nov. 2021;
- [99] R. Aminzadeh et al., "Personal exposimeter for radiation assessment in real environments in the 60-GHz band", Radiation Protection Dosimetry, Vol. 176, Issue 3, pp. 316–321, Nov. 2017;
- [100] R. Aminzadeh et al., "A multi-band body-worn distributed radio-frequency exposure meter: Design, on-body calibration and study of body morphology", Sensors, Vol. 18, Issue 1, 272, Jan. 2018;
- [101] Chahat et al., "Characterization of the interactions between a 60-GHz antenna and the human body in an offbody scenario", IEEE Transactions on Antennas and Propagation, Vol. 60, Issue 12, pp. 5958–5965, Aug. 2012;
- [102] Tektronix, "Fundamentals of Real-Time Spectrum Analysis", URL:https://download.tek.com/document/37W_17249_5_HR_Letter.pdf
- [103] Sistema Nazionale per la Protezione dell’Ambiente (SNPA), "Linee guida per la misura dei campi elettromagnetici generati da sistemi di telecomunicazione in tecnologia 5g", URL:https://www.snambiente.it/wp-content/uploads/2025/03/LG_SNPA_Misure_5G.pdf
- [104] G. Chirico, C. J. D’Silva, D. Pinchera, and F. Schettino, "Analysis of the Contribution of Different Generations of Cellular Communication Systems to the EMF Levels in a Small City", 2024 IEEE International Symposium on Measurements & Networking (M&N), July 2024;
- [105] Global mobile frequency database, URL:<https://www.spectrummonitoring.com/frequencies.php?market=I>
- [106] M. D. Migliore et al., "A new Paradigm in 5G Maximum Power Extrapolation for Human Exposure Assessment: forcing gNB traffic toward the measurement equipment", IEEE Access, Vol. 9, pp. 101946–101958, June 2021;
- [107] M. D. Migliore et al., "Application of the Maximum Power Extrapolation Procedure for Human Exposure Assessment to 5G Millimeter Waves: Challenges and Possible Solutions", IEEE Access, Vol. 10, pp. 103438–103446, Sep. 2022;
- [108] M. D. Migliore, "5G Field level measurement for human exposure assessment: A lesson for 6G", IOP Conference Series: Materials Science and Engineering, Vol. 1254, Issue 1, 012001, Sep. 2022;

- [109] D. Pinchera, M. D. Migliore, and F. Schettino, "*Compliance Boundaries of 5G Massive MIMO Radio Base Stations: A Statistical Approach*", IEEE Access, Vol. 8, pp. 182787-182800, Oct. 2020;
- [110] H. Arai, "*Measurement of Mobile Antenna Systems*", 2nd ed., Artech House, Norwood, Massachusetts, 2013;
- [111] C. A. Balanis, "*Antenna Theory Analysis and Design*", 4th ed., John Wiley & Sons, Hoboken, New Jersey, 2016;
- [112] LTC5596 datasheet, URL:<https://www.analog.com/media/en/technical-documentation/data-sheets/5596f.pdf>