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RF Front-End for RADAR

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Resumo

RADAR de abertura sintética (SAR) é uma tecnologia usada para obter imagens a partir de radiação RF que são relativamente insensíveis a efeitos como nuvens, poeiras, luz solar... A partir da junção entre teoria de RADAR e técnicas de processamento de sinal, é possível obter-se imagens deste tipo com resoluções apreciáveis. Ao longo dos anos, avanços nestas áreas têm permitido obter imagens com cada vez maior qualidade.

Um destes avanços foi a introdução do Software Defined Radio (SDR) como uma forma de substituir muito do hardware de rádio convencional por software eficiente e flexível. Mas, apesar de substituir parte do hardware necessário para comunicações rádio, ainda não é capaz de fazer uma transição total para software. Desta forma, é ainda necessária alguma eletrônica de interface entre o SDR e as antenas de rádio. Este tipo de módulos designam-se de "RF front-ends".

A experiência SARIA é um projeto académico que procura fazer SAR a partir de uma arquitetura baseada em SDR. Esta experiência será albergada num balão estratosférico no contexto do programa BEXUS. Duas antenas, uma insuflável e um refletor estático, serão usadas para sessões de SAR.

Com esta configuração em mente, um "RF front-end" será necessário. Espera-se que este módulo transmita e receba sinais de banda C com considerações de baixo custo, baixo consumo e ganho ajustável. Esta dissertação descreve o desenvolvimento de um "RF front-end" que seja adequado a aplicações semelhantes à da experiência SARIA. O módulo contém uma cadeia de transmissão (TX) com ganho ajustável até uma potência de saída de 30 dBm; uma cadeia de receção (RX) adaptada para frequências na ordem dos 5.8 GHz e uma comutação entre RX e TX baseada em deteção de potência RMS com uma frequência de repetição de pulsos de 1 kHz.

Abstract

Synthetic Aperture RADAR (SAR) is a technology used to record RF images that are significantly insensitive to effects like dust, clouds, daylight... By coupling RADAR theory with signal processing techniques, it is possible to obtain images with a good level of resolution. Over the years, advances in the RADAR and signal processing domains have made it possible to obtain increasingly better quality images.

One such advance was the introduction of Software Defined Radio (SDR), as a way to replace a lot of the conventional radio hardware with flexible and effective software. But, even though it substitutes a lot of the previously used hardware, there is still the need for some electronics to interface the SDR with radio antennas. These modules are commonly referred to as RF front-ends.

The SARIA experiment is a student project that intends to perform SAR, using a SDR based architecture. This experiment will be housed in a stratospheric balloon in the context of the BEXUS program. Within the context of this experiment, SAR sessions will be done by two antennas, an inflatable antenna, and a static reflector.

With this setup in mind, a RF front-end will be required. This module is intended to adequately transmit and receive C-band signals, with a concern for low-power, low-cost and adjustable transmission gain. This thesis work describes the development of a suitable RF front-end for applications with similar specifications to the ones exhibited in the SARIA experiment. The module contains a Transmission chain with adjustable gain and up to 30 dBm of output power; a Reception chain adapted to frequencies around 5.8 GHz and switching between RX and TX contexts based on RMS power detection with a pulse repetition frequency of 1 kHz.

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“If you can’t explain it simply, you don’t understand it well enough.”

Albert Einstein

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List of Acronyms

ADC Analog to Digital Converter. 2, 10, 11

ADS Advanced Design System. xi, xii, xiii, 31, 32, 33, 34, 35, 36, 37, 38, 39, 40, 41, 42, 43, 45, 46, 50, 51, 52, 53, 54, 56, 57, 59, 63, 64, 65

BEXUS Balloon Experiments for University Students. i, iii, xi, 3, 4

BPF Band Pass Filter. xii, 20, 29, 35, 36, 37, 39, 49, 51, 64, 73

CAD Computer Aided Design. 30, 33

COTS Commercial off the Shelf. 4, 12

DAC Digital to Analog Converter. 2, 11

DLR German Aerospace Center. 3

DPDT Double Pole Double Throw. 26, 42

DUT Device Under Test. 17

ENR Excess Noise Ratio. 17, 18

ESA European Space Agency. 3

FEUP Faculty of Engineering of the University of Porto. 4

FPGA Field Programmable Gate Array. 11

FSPL Free Space Losses. 26, 72

GPIO General Purpose Input Output. 11, 28, 45

HASL Hot Air Solder Leveling. 63

ISM Industrial, Scientific and Medical. 4, 71

LNA Low Noise Amplifier. xi, xii, xiii, 11, 13, 15, 16, 19, 20, 26, 29, 35, 36, 37, 46, 51, 52, 58, 59, 64, 73

LPF Low Pass Filter. 20, 26, 46, 58, 73

MS Monostable. 26, 44, 45, 58

OBC Onboard Computer. 11

PA Power Amplifier. xii, xiii, 11, 15, 20, 26, 29, 36, 38, 41, 42, 49, 53, 54, 55, 58, 64

PCB Printed Circuit Board. 5, 30, 31, 32, 36, 45, 50

PRF Pulse Repetition Frequency. 9, 21, 24, 73

RADAR Radio Detection and Ranging. i, iii, ix, xiii, 1, 3, 4, 7, 8, 9, 23, 25, 68, 69, 71

RAR Real Aperture Radar. xv, 9, 69, 71

RCS Radar Cross Section. 26, 67, 72

REXUS Rocket Experiments for University Students. 3

RF Radio Frequency. i, iii, xi, xiii, xv, 2, 3, 4, 5, 7, 8, 10, 11, 12, 13, 15, 16, 18, 19, 20, 21, 22, 23, 24, 25, 26, 27, 28, 29, 31, 32, 33, 36, 37, 38, 39, 40, 41, 42, 44, 45, 46, 49, 53, 54, 56, 57, 58, 59, 64, 65, 67, 72, 73

RFIC Radio Frequency Integrated Circuit. 11

RMS Root Mean Square. i, iii, 24, 38, 40, 44

RX Reception. i, iii, xiii, 3, 5, 11, 16, 21, 22, 24, 26, 28, 31, 35, 38, 39, 42, 44, 45, 49, 51, 52, 54, 56, 57, 58, 59, 60, 65, 71

SAR Synthetic Aperture RADAR. i, iii, xiii, xv, 1, 2, 3, 4, 5, 7, 8, 9, 23, 24, 25, 28, 68, 69, 71, 72, 73

SARIA Synthetic-Aperture Radar using an Inflatable Antenna. i, iii, xi, 3, 4, 5, 7, 8, 9, 10, 11, 20, 21, 23, 24, 25, 45, 58, 64, 73

SDR Software Defined Radio. i, iii, xi, 1, 2, 3, 4, 5, 7, 10, 11, 12, 24, 26, 28, 36, 38, 58, 73

SED Student Experiment Documentation. xi, 23, 24, 25, 45

SMA SubMiniature version A. 28, 49, 50, 63, 65

SNR Signal to Noise Ratio. 13, 16, 26, 73

SNSA Swedish National Space Agency. 3

SSC Swedish Space Corporation. 3

TX Transmission. i, iii, xiii, 3, 5, 11, 21, 22, 24, 26, 28, 31, 36, 38, 39, 41, 42, 44, 45, 49, 52, 53, 54, 56, 57, 58, 59, 60, 61, 65, 71

USB Universal Serial Bus. 11

VNA Vector Network Analyzer. xii, xiii, 12, 49, 50, 51, 52, 53, 55, 63, 64

WPD Wilkinson Power Divider. xi, xii, 19, 20, 39, 40, 52

ZARM Center of Applied Space Technology and Microgravity. 3

Chapter 1

Introduction

This chapter will introduce the Dissertation proposal: RF Front-End for RADAR. For that, it will frame topics such as RADAR (specifically Synthetic Aperture RADAR) and Software Defined Radio (SDR) in a way that provides a starting point for the development of this thesis. It will then explain the motivation and objectives of this proposal, relating to the application that the expected final product will have.

1.1 Context

Synthetic Aperture RADAR (SAR) is a subset of RADAR technology used to produce high-resolution images of a given region under a broad spectrum of environmental conditions. With that in mind, a SAR system actively transmits radio/microwave energy and collects the respective echoes. [6] Depending on the surface, the received data will differ. Through the use of signal processing techniques, it is possible to represent the analyzed area.

The great advantage of using SAR imaging comes from the fact that, unlike optical imaging, it is possible to obtain precise images in darkness, clouds, rain... [8] At the same time, interpreting SAR images is not particularly intuitive when compared to the optical images. Even though SAR technology is not a replacement for optical imaging, it provides great complementary information that is sometimes impossible to gather with just optical signals. [9]

This technology is not particularly new as it dates back to the 1950s. However, over the years, advances in signal processing and microwave engineering have improved the resolution and overall quality of SAR images. [5]

The resolution of the obtained SAR images is directly related to the wavelength used in the signals for transmission. To obtain useful images, a specific range of wavelengths (and corresponding frequencies) is of interest. Table 1.1 shows the frequency bands associated with their respective

wavelengths, as well as the applications related to the SAR domain. [6]

Band	Frequency	Wavelength	Typical Application
Ka	27 – 40 GHz	1.1 – 0.8 cm	Rarely used for SAR (airport surveillance)
K	18 – 27 GHz	1.7 – 1.1 cm	Rarely used (H ₂ O absorption)
Ku	12 – 18 GHz	2.4 – 1.7 cm	Rarely used for SAR (satellite altimetry)
X	8 – 12 GHz	3.8 – 2.4 cm	High resolution SAR (urban monitoring,; ice and snow, little penetration into vegetation cover; fast coherence decay in vegetated areas)
C	4 – 8 GHz	7.5 – 3.8 cm	SAR Workhorse (global mapping; change detection; monitoring of areas with low to moderate penetration; higher coherence); ice, ocean maritime navigation
S	2 – 4 GHz	15 – 7.5 cm	Little but increasing use for SAR-based Earth observation; agriculture monitoring (NISAR will carry an S-band channel; expands C-band applications to higher vegetation density)
L	1 – 2 GHz	30 – 15 cm	Medium resolution SAR (geophysical monitoring; biomass and vegetation mapping; high penetration, InSAR)
P	0.3 – 1 GHz	100 – 30 cm	Biomass. First P-band spaceborne SAR was launched in 2020; vegetation mapping and assessment. Experimental SAR.

Table 1.1: SAR frequency bands and respective applications. Credit: NASA Earth Data [6]

Looking to the C-band (because it is the most used frequency band for SAR systems and the one of interest for this thesis work), the wavelength range is in the order of 7.5 – 3.8 cm. For these values, to obtain a resolution of about 10 m, one would require an antenna with a real aperture of around 4250 m! Physically this is naturally unfeasible to build. It is at this point that a synthetic aperture comes into play. Rather than focusing on obtaining such an impractical real aperture, the idea is to combine the readings of a shorter (and more feasible) antenna during an integration interval to simulate the larger antenna. With this clever technique, it is possible to use C-band signals and obtain relevant precision in the resulting SAR images. [6]

While the SAR technology was evolving to obtain better and more precise images, in the early 90s, a new design paradigm for wireless communications was being created - the Software Defined Radio (SDR). Its creator, Joseph Mitola, envisioned a model where the RF architecture would be based around only 3 physical components: an antenna, the SDR module and Digital to Analog Converter (DAC) for transmission or Analog to Digital Converter (ADC) for reception chains. The remaining conventional functions (performed by amplifiers, filter, mixers...) would all be handled by reprogrammable processors. Even though the idea of the SDR paradigm is not yet fully achieved, the term SDR is associated with a viable system where most of the signal creation

and processing is defined by software, but it also includes significant hardware components. [10]

With the introduction of software development to systems previously built around defined hardware, higher levels of flexibility became possible to achieve in radio systems design. Due to this factor, there is a great incentive to use a SDR, to obtain high performance with the possibility to reconfigure and update RF systems simply by modifying the respective software. [11]

When working with SAR systems coupled with SDR technology, there is a need to transmit high power signals coming from a wave generator and receive faint reflections (of the transmitted signal) at the reception stage. In a pulse RADAR system, both transmissions and receptions use the same port (the antenna sending and receiving signals is the same). In these cases, there is a need to develop an interface between the antenna and the SDR. Such an interface is responsible for transmitting the adequate signal and receiving the correspondent echoes for subsequent processing.

The systems that carry out the description above are identified as "RF front-end modules". These kinds of modules usually encompass the electronics for transmission of high power signals, reception of very faint echoes, as well as control signals that allow the switching between Transmission (TX) and Reception (RX).

1.2 Motivation

The Synthetic-Aperture Radar using an Inflatable Antenna (SARIA) experiment is a student project that plans to develop a C-band SAR system using a SDR. This project is to be included aboard a stratospheric balloon in the context of the BEXUS program. The thesis work described in this document will have direct application in the SARIA experiment, particularly in the RF engineering section of the project.

1.2.1 REXUS/BEXUS Program

The REXUS/BEXUS program is an initiative of a group of European space agencies that intends to allow university students to implement an experiment and carry out a research mission in a stratospheric balloon (for BEXUS) or a rocket (for REXUS). This group includes: the European Space Agency (ESA), the German Aerospace Center (DLR), the Swedish National Space Agency (SNSA), the Swedish Space Corporation (SSC) and the Center of Applied Space Technology and Microgravity (ZARM). Experts from these organizations assist university students in the development of their projects that are launched, at the end of each cycle, from SSC, Esrange Space Center in northern Sweden. In particular, for BEXUS, experiments are housed in a stratospheric balloon with a volume of 12000 m^3 , going to altitudes of around 30 km. The flight duration for these missions is between 2 to 5 hours. [12]



(a) SARIA logo



(b) BEXUS logo

Figure 1.1: SARIA and BEXUS logos

1.2.2 SARIA Experiment

The SARIA experiment is a student project from a Portuguese group of students from the Faculty of Engineering of the University of Porto (FEUP). The experiment had its inception in May 2019 and, since then, it has been approved to be a part of Cycle 13 of the BEXUS program, scheduled to launch in October 2021. Figures 1.1a and 1.1b show the logos for the SARIA project and the BEXUS program respectively.

This experiment has the goal of performing low-power SAR to obtain images of the ground overflown by the BEXUS balloon, using a Commercial off the Shelf (COTS) SDR. To tackle the problem of mass and portability (associated with traditional large metallic reflectors), it also plans to test the functionality of an inflatable antenna to obtain the referred SAR images. [4]

Though the experiment will test the viability of a light and portable inflatable antenna, the main objective is to obtain low-power SAR images with decent quality. To guarantee an operational antenna for SAR sessions and to assess the quality of the inflatable antenna, a COTS static reflector will be used with the same functionality.

The experiment also has a cost concern for the development of the inflatable antenna and SAR operation. In that sense, there is a requirement for the frequency of SAR signals to be within a license-free ISM band¹, such as C-band. [13]

From the description above, there is a clear need for the development of a module that enables the interaction between the COTS SDR with the two antennas to be used. Motivated by this necessity, the thesis work presented here will elaborate on the development of such a module.

¹Industrial, Scientific and Medical (ISM) bands are a set of frequency bands used, amongst other purposes, for low-power, license-free and error-tolerant RF/microwave RADAR systems

1.3 Objectives

With the context and motivation for this work presented, the objectives can be established. As such, the objectives of this thesis are to design, simulate, implement and qualify a RF front-end module. This system shall effectively transmit and receive C-band RF signals, in half-duplex mode to 2 independent antennas. By properly switching between TX and RX mode, as well as alternating between the two considered antennas, the module must take into account aspects like isolation between transmit and receive signals, attenuation of RF signals, and global noise figure.

1.4 Document Structure

The document presented here will be divided into five parts: the State Art, System Design, the Integration and Manufacturing, Testing and Validation, and Final Remarks. In chapter 2, a literature review will be made regarding the topics of Radar Theory and SAR technology, specifically in integration with SDR architectures, along with some key RF theory concepts. The chapter will finish with a Market Survey to present commercial solutions related to the RF front-end module domain. The following chapter 3, will discuss the Scientific Background of the SARIA experiment, as well as provide a System Overview of the module to be developed. Chapter 4 will detail the RF and PCB design as well as the entire integration and assembly process of the implemented boards. After that, chapter 5 will elaborate on the testing phase of the manufactured boards. As such, it will first present the obtained results at each stage and then evaluate their quality. The final chapter of the document is reserved for final remarks on the work that lead to the final product of this thesis. It will talk about some of the difficulties encountered in the execution of this project, share conclusions that were realized, and propose future work to do in the context of this dissertation.

Chapter 2

State of the Art

The content of this chapter will focus on a brief presentation of some key concepts of RADAR theory and then specify to the subset of Synthetic Aperture RADAR implemented with SDR. As these topics will serve as the basis for the SARIA experiment, an introduction to this theme seems appropriate for the adequate development of a RF front-end meant for this application. The chapter then goes on to elaborate on some more specific subjects of RF and microwave engineering that are paramount for the implementation of a front-end system. After presenting the aforementioned literature review, the chapter closes with a Market Survey that will denote competing products and overall explain why it makes sense to devote effort to the development of a RF front-end, within the constraints and specifications of the SARIA experiment.

2.1 RADAR Theory

The topics of RADAR theory are rather vast and can get relatively complex. Yet, not all are necessary for the basic knowledge required in this application. That said, the sections below will briefly discuss important concepts related to RADAR theory that will support the eventual development of the RF front-end module.

2.1.1 Wireless Communications

By exchanging sound, infrared, optical, or radio waves between two points physically separated in space, it is possible to achieve wireless communications. Within the electromagnetic spectrum, RF and microwave signals offer wide bandwidths while being able to penetrate clouds, fog, rain, dust... as well as correctly function either in day or night time. This versatility has made these frequency bands the most desirable for wireless communications. [2]

Such communications imply transmissions and receptions of signals between at least two points, depending on the directionality of communication. In a *simplex* system, communication only occurs in one direction, meaning that one end can only transmit information and the other can only receive it. On the other hand, *duplex* systems allow for bidirectional communication. When discussing the *duplex* mode, it is necessary to further detail if the system is *half-duplex* (where transmissions and receptions can happen but not at the same time) or if it is *full-duplex* (where communication can take place in both directions at the same time). [2]

2.1.2 RADAR Equation

Radio Detection and Ranging (RADAR) is a widely recognized technique that enables the detection of objects at a relatively long distance and uses the knowledge of wireless communications theory. SAR is a type of RADAR so, it shares many of its fundamentals. It comes with no surprise that, to understand SAR, one must understand the basics of RADAR theory. [5]

For the thesis work presented in this document, there should be a solid understanding of the so-called *radar equation*, as it is the base for the development of a link budget that proves the feasibility of any RF system. In section A.1 a further explanation of the RADAR equation is made, explaining what the respective parameters represent. This description is not fundamental for the development of the front-end module, but it does help to clarify some of the aspects of the SARIA experiment Link Budget that prove the respective RF feasibility.

2.1.3 Pulse RADAR systems

In general, RADAR systems can be classified in two main types: continuous wave RADAR or pulse RADAR. In the first one, the transmission and reception happen continuously in a permanent regime (the transmitting and receiving antennas are usually separated). In pulse RADAR systems, the transmission takes place using short finite pulses, and it is possible to receive the echoes using the same antenna. [14]

Even though there are relatively complex continuous-wave RADAR systems, because of their characteristics, these types of systems are usually found in simple applications such as police speed-timing radars, altimeters, and proximity fuses. [14] With that said, the work presented in this thesis will focus on the other type of RADAR systems - pulse RADAR - with a few considerations for power consumption.

When it comes to pulse RADAR systems, it is important to think about the switching between transmission and reception mode. Since the system will only send short pulses and wait for the detection of the correspondent echoes, the duty cycle of transmission will be relatively low. Consequently, in between pulses, the system will be at receiving mode. Because of this, proper switching between transmission and reception is required. In these scenarios, a pulse generator controls the

transmit/receive switch. The produced pulse lasts for a signal duration of τ and, during that time, the transmission chain active. The Pulse Repetition Frequency (PRF) corresponds to the frequency of emission of the pulses. [2]

The switching mechanism has two main functions: controlling when and for how long the transmission of pulses happens and switching the system between transmission and reception mode. [2]

To make use of the pulse RADAR capabilities, but also have a concern for power consumption, one can prolong the duration of the pulse and send it over a longer signal duration, $\tau > \frac{1}{\beta}$ (with β being the bandwidth). For SAR systems, this is useful to send the same amount of energy at transmission but with lower power consumption. The drawback is that, because each transmission will not have as much power, the range resolution of the obtained SAR images will decrease. It is possible to improve this factor by employing signal processing techniques that correlate the obtained echoes over the signal duration. This is a form of processing gain designated Range Compression Gain. The SARIA experiment will use this technique to improve the quality of the received SAR images. [4]

2.2 SAR with SDR

When using Synthetic Aperture RADAR technology, the RADAR part illuminates the area that the antenna points to by transmission of microwave pulses. These pulses are then reflected by the overflown targets and gathered at the RADAR receiver. After reception, it is possible to determine the distance of the reflecting object by measuring the time difference between transmitted and received pulse. These differences, along with signal processing techniques, will allow to create SAR images along two axes: the slant range direction and the azimuth/along-track direction. [5] For each dimension, there will be an associated resolution that states the minimum distance for which it is possible to distinguish two different objects. Table 2.1 compares the spatial resolutions between Real Aperture Radar and Synthetic Aperture RADAR. Section A.2 goes into further detail about how to determine these resolutions.

Spatial Direction	Synthetic Aperture RADAR	Real Aperture Radar
Slant Range	$\frac{c}{2\beta}$	$\frac{c}{2\beta}$
Along-track	$\frac{D}{2}$	$\frac{\lambda R}{D}$

Table 2.1: Spatial resolutions for SAR and RAR

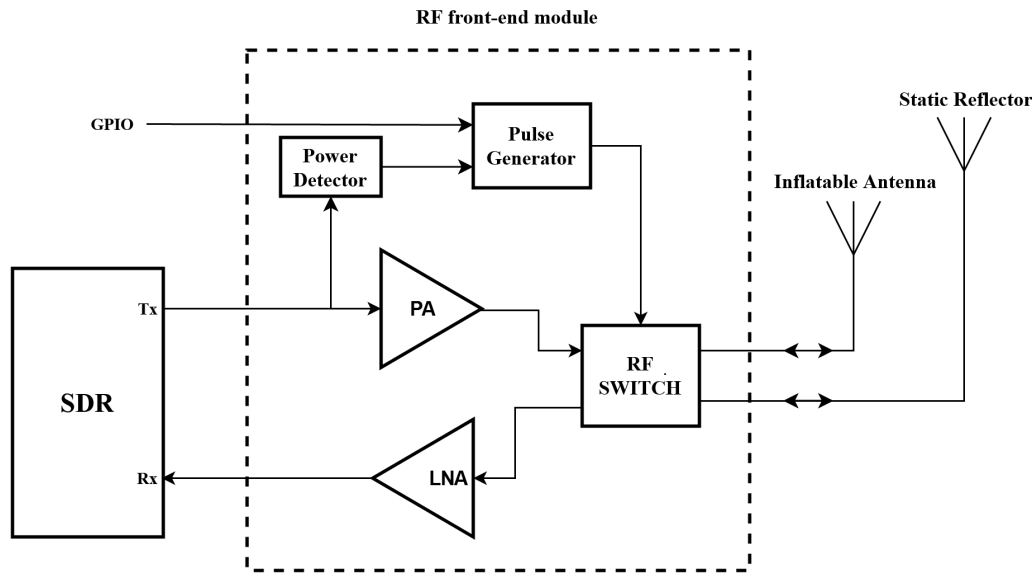


Figure 2.1: High level block diagram of a RF front-end

2.2.1 SDR Receiver Architectures

The Software Defined Radio (SDR) receiver uses software to perform many of the basic functions of a RF/Microwave receiver. With this technology, it is possible to achieve good performance while there is the flexibility to reconfigure radio systems simply by updating software. In these configurations, it is possible to employ the same hardware platform in different scenarios. Although its many advantages, some drawbacks might constrain the use of this kind of equipment. For simple radio systems, the SDR platform might be too expensive. The effort associated with software development might also be non-negligible, depending on the programming skills of the designer. And for more complex systems, the Analog to Digital Converter might limit the sampling frequency to be considered in the digital part of the SDR. [11]

All in all, the SDR platform brings good advantages to the RF design if the system is complex enough for it to be worth the acquisition, but not too much demanding that the platform cannot meet the requirements of the experiment. The SARIA project is one of these cases, where it makes sense to use a SDR platform since its use will aid the transmission and reception of RF signals, and there is the availability of compatible equipment. [4] In fact, the considered SDR to be used in the SARIA experiment is the Ettus USRP B200. [1]

As stated in section 1.1, even though the SDR platform introduces software development for tasks that were previously carried out by hardware, there still needs to be a layer of hardware to interface the SDR with the antennas in the SARIA experiment. The development of such an interface is the main motivator of this thesis work. Figure 2.1 shows a high-level block diagram of the RF front-end to be developed.

In the block diagram of Figure 2.1 there is a representation of the the SDR module, the two

antennas to be on board the SARIA experiment (the Inflatable Antenna and the Static Reflector) and the RF front-end module. The front-end module is essentially composed of a RF transmission path that connects the TX port of the SDR to the antennas; a reception path that connects the RX port of the SDR to the antennas and a RF switching block that controls which antenna is either at transmission or reception mode. A pulse generator controls RF switching, which in turn, is triggered by a GPIO pin connected to the OBC of the SARIA experiment and a Power Detector module. The GPIO pin controls which antenna is to be selected. The Power Detector module is used to create a pulse with enough duration for a full transmission. This module allows for the half-duplex operation to be triggered by the TX signal.

Taking into consideration the block diagram of Figure 2.1, it is noticeable the 3 main components that constitute the RF front-end module: the Power Amplifier (PA), the Low Noise Amplifier (LNA) and the RF switch. The PA is a component whose goal is to amplify a signal for transmission with adequate levels of gain, stability, and linearity. At the reception chain, a LNA amplifies the received signals. These types of amplifiers are capable of amplifying the incoming signals with not only interesting levels of gain, linearity, and stability but also with a concern for noise figure. This last parameter indicates how much noise the LNA adds to a signal during amplification. Generally, it is not possible to minimize the noise figure and maximize the amplifier gain simultaneously, so a compromise must be made between these two parameters. [2] Moreover, the RF switch needs to be a device with as low as possible insertion loss when conducting RF signals. Conversely, when the switch is not conducting, it is important to have good levels of isolation. For the specific case of the RF front-end, it is also crucial to have good isolation between the TX and RX paths. [2]

With this front-end circuitry connected to the SDR, it is possible to perform the desired RF transmission and reception. The remaining functions, not included in the RF front-end module, are carried out by the elements inside the SDR platform. Figure 2.2 shows the block diagram for the considered SDR taken from the manufacturer website. [1]

In the block diagram of Figure 2.2 it is noticeable two main blocks: a Field Programmable Gate Array (FPGA) and a Radio Frequency Integrated Circuit (RFIC). The RFIC module transmits/receives the analog signals from the RF front-end, performing operations of amplification, filtering, up/down conversion and converting the signals from analog to digital (with an ADC module) or from digital to analog (with a DAC module). The converted signals will then be processed, at the digital domain, in the FPGA module where operations of digital up/down conversion are also performed. Lastly, a USB interface allows for the signal data to flow between the SDR and the OBC of the SARIA experiment.

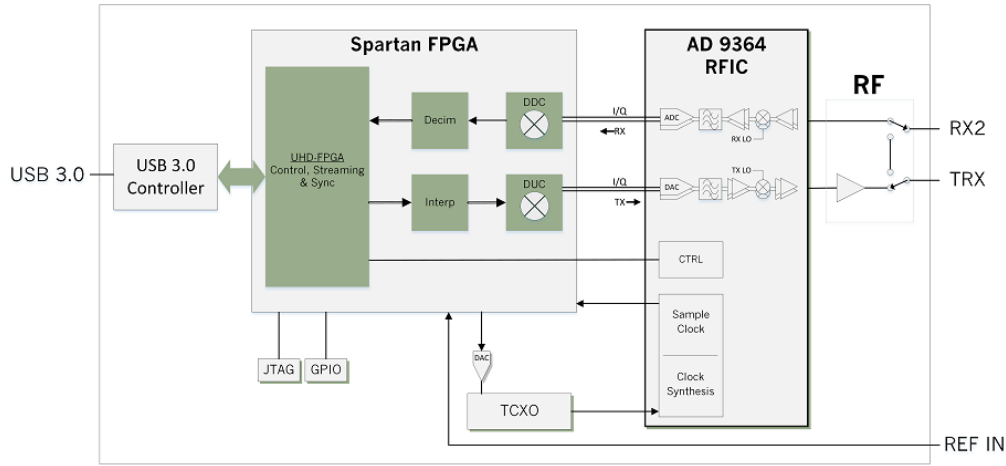


Figure 2.2: SDR block diagram. Credit: Ettus Research [1]

2.3 RF Theory

Following up on the high-level block diagram of Figure 2.1, this section will discuss fundamental RF theory for microwave circuit design. It will focus essentially on: scattering parameters, impedance matching, and RF parameters. These topics are key for the eventual design of the RF front-end board since they will elaborate on important aspects of attenuation and reflection of signals in RF domain, as well as important parameters that manufacturers use to characterize their COTS components.

2.3.1 Scattering Parameters

In a general circuit usually, impedance or admittance matrices are used to characterize its performance. These parameters are obtained from measurements of voltage and current. However, for the RF/microwave domain, voltages and currents are not as easily measured due to the increased frequency, so measurements of power and energy are used. In these conditions, impedance and admittance matrices become somewhat of an abstraction, and a different (and more useful) alternative is required for circuit characterization. The scattering matrix allows for such an alternative. [2]

The impedance and admittance matrices are dependent on voltage and current measurements. However, the scattering matrix uses the voltage waves incident and reflected on the circuit ports to characterize its performance. These parameters can be calculated for some circuits, but often they are measured directly using a Vector Network Analyzer (VNA). [2]

Considering an n -port network with incident voltage on port n , V_n^+ , and reflected voltage on port n , V_n^- , the scattering matrix can be defined as:

$$\begin{bmatrix} V_1^- \\ V_2^- \\ \dots \\ V_n^- \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} & \dots & S_{1n} \\ S_{21} & \dots & \dots & \dots \\ \dots & \dots & \dots & \dots \\ S_{n1} & \dots & \dots & S_{nn} \end{bmatrix} \begin{bmatrix} V_1^+ \\ V_2^+ \\ \dots \\ V_n^+ \end{bmatrix}$$

Each element of the S-matrix, S_{ij} , can be determined by driving a port j with an incident wave voltage, V_j^+ and measure the reflected wave voltage, V_i^- while all the other $k \neq j$ ports are set to zero. When this process is taking place, all k ports must be terminated by matched loads. Therefore, in practice, the S_{ii} element of the S-matrix corresponds to the reflection coefficient seen looking into i , when all other ports are terminated in matched loads. Simultaneously, the S_{ij} parameter will state the transmission coefficient from j to i when all other ports are matched to the characteristic impedance. [2]

2.3.2 Impedance Matching

The section above introduced the topic of S-matrix and S-parameters as an important tool to characterize a RF circuit in terms of incident and reflected power in its ports. But, more than measure and characterize the circuit, it is usually important to adjust the S-parameters of the overall RF circuit, to minimize its reflections. Impedance matching networks fulfill this purpose. These networks are located between the load and the transmission line to obtain an impedance looking into the matching network equal to the characteristic impedance, Z_0 . This procedure is important for several reasons: it allows for maximum power to be delivered (when the load is matched to the line); it may improve SNR in impedance matching sensitive components (e.g. antennas, LNAs...) and it may reduce amplitude and phase errors in a power distribution network. [2]

That said, there is a great variety of topologies and types of networks that can be applicable at a given circuit section. To assist the network design there are a few factors to be considered. Starting with the **complexity** of the design it is usually preferable to select the simplest impedance matching network possible. This will typically result in a cheaper, smaller, and more reliable solution. Then there is the concern for **bandwidth**: depending on the functionality of the circuit, it may be desirable to have a wider band of frequencies for which the matching is useful. Looking into the **implementation** of the matching network it may be useful to select the best solution depending on the type of transmission line or waveguide used. And finally, the aspect of **adjustability** might also be of interest to obtain a matching network variable with the load impedance. [2]

Consequently, there is a considerable amount of points to keep in mind in the design of matching networks, and they usually imply tradeoffs. At the same time, the different possibilities of achievable networks also come from the fact that there are different types of matching networks. The

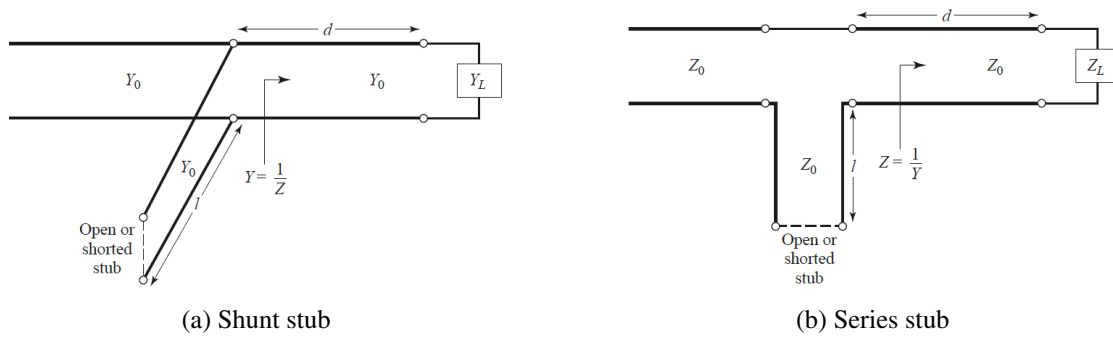


Figure 2.3: Single stub tuning. Credit: *Microwave Engineering* (page 235), David M. Pozar [2]

following list will describe four particular ways of achieving impedance matching networks, each with advantages and drawbacks. [2]

1. *Matching with Lumped Elements (L-Networks)* - It is the simplest type of matching network, and it uses either inductors or capacitors (depending on the load impedance) for impedance matching. If the frequency or the circuit size is low enough, actual capacitors and inductors can be used. However, if this is not the case, this type of matching is not applicable. [2]
2. *Single Stub Tuning* - With this technique, an open or short-circuited length of the transmission line (stub) is connected in series or parallel (shunt) to the transmission line connected to the load at a specific distance. It is also very convenient since the stub can be fabricated as just another part of the transmission line (avoiding extra components in the design). Figures 2.3a and 2.3b show a representation of this method. [2]
3. *Double Stub Tuning* - Although the single stub tuning method is convenient, it has the drawback of not being particularly adjustable. This problem is addressed by the double stub tuning method, in which two fixed tuning stubs are used to have the convenience of using transmission lines with adjustable impedance matching networks. The drawback here is that it is not possible to be used at all frequencies. Figure 2.4 illustrates the double stub tuning method. [2]
4. *Quarter-wave Transformer* - The quarter-wave transformer is a simple circuit that can be used for impedance matching using a transmission line with a length of a quarter wave-length, $\lambda/4$. This method can be extended to multisection designs to provide a broader bandwidth in the matching network. One disadvantage of this technique is that it only allows the matching of resistive loads. [2]

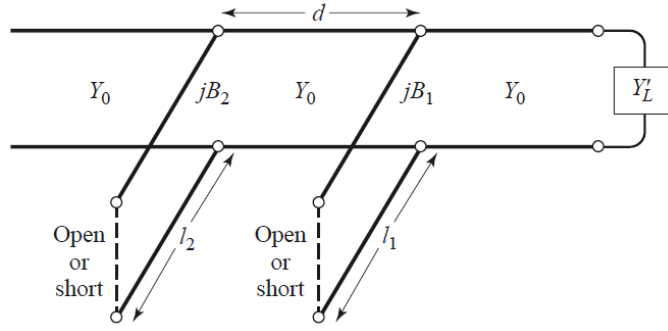


Figure 2.4: Double stub tuning. Credit: *Microwave Engineering* (page 242), David M. Pozar [2]

2.3.3 RF Parameters

Besides the concepts of S-parameters and impedance matching, for reflection characterization and adjustments, another important consideration in RF design has to do with RF parameters of the selected components. Namely, for the more crucial RF components of the front-end board, like the LNAs, PAs and switches, it is necessary to evaluate their performance in terms of gain, losses, and linearity. In that sense, this section will present commonly seen RF parameters, useful to aid the selection process of RF components.

Return Loss

The Return Loss is associated with the common situation when a load is mismatched to the characteristic impedance. If this is the case, then not all available power from the generator will be delivered to the load, and there will be some loss. This loss can be expressed in dB by the Return Loss (RL) factor as: [2]

$$RL = -20 \log |\Gamma| \quad (2.1)$$

In equation 2.1, the Γ term corresponds to the reflection coefficient. The RL parameter is usually provided by the manufacturer for input and output RF ports of the respective component, and it is an important measure for RF power loss. For total reflection, $RL = 0$ dB and for no reflection, $RL = -\infty$ dB. [2]

Insertion Loss

Correspondently to the Return Loss, the Insertion Loss represents the transmission coefficient between 2 ports in a circuit. The transmission coefficient relates to the reflection coefficient as $T = \Gamma + 1$ and is usually represented in dB, being then referred to as Insertion Loss (IL). This value is usually relevant for components where there is some attenuation of signals (e.g. filters or

switches) and it can be expressed as: [2]

$$IL = -20 \log |T| \quad (2.2)$$

Noise Figure

The noise figure of a component measures the degradation of the SNR between input and output. When gain or attenuation is applied to a RF input, both the signal and the noise get amplified or attenuated, so the SNR remains the same. However, in practical systems, besides this effect, some noise is introduced, and so the SNR decreases. The noise factor, F , measures how much is this decrease at the output, and it is given by: [2]

$$F = \frac{S_i/N_i}{S_o/N_o} \quad (2.3)$$

Considering that the noise is given as $N_i = k \cdot B \cdot T_0$ and that an equivalent noise temperature, T_e , is introduced by the RF component, the noise factor can be rewritten as:

$$F = 1 + \frac{T_e}{T_0} \geq 1 \quad (2.4)$$

Usually, a microwave system contains many different cascaded components, so it is relevant to understand how the degradation of signal-to-noise ratio happens between the input and output of the cascaded elements. By knowing the noise temperature of every component in the cascaded network, it is possible to determine the overall noise factor, F , using the Friis Formula for noise factor, which is given as: [2]

$$F_{cas} = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \dots \quad (2.5)$$

This relation is particularly important for receiving chains of RF front-end modules where Low Noise Amplifiers are used. These types of amplifiers are purposely built to have low values of noise figure. Also, from equation 2.5, it is noticeable that the cascaded noise factor, F_{cas} , is much more affected by the first term (the first component in the chain) than the remaining elements, so it is worth to make an effort to minimize the noise factors of all cascaded components, but especially the first one! The noise figure, NF , corresponds to the noise factor expressed in dB. [2]

Since this parameter is of such importance in a RX chain of a front-end system, it is important to have an understanding of how it could be measured. Essentially, there are two main ways to make noise figure measurements: Y-factor and the Gain method. Besides using these techniques, it is also possible to test this parameter using Noise Figure Meters, which are specially designed equipment to measure noise figure based on the Y-factor process. [15]

Starting with the latter alternative, using Noise Figure Meters/Analyzers is the most straightforward and, in most cases, the most accurate way of measuring noise figure. With this equipment, the test engineer can measure the system gain together with the noise figure. However, this method does not come without disadvantages. Usually, the frequency range possible to measure is limited and measurements exceeding 10 dB start to become very inaccurate. Also relevant is that this kind of equipment is rather expensive. [16]

A second approach to measure this parameter is the Y-factor method. This is a popular one that requires the use of the Y-factor and an Excess Noise Ratio (ENR) source. The idea consists in connecting the ENR noise source to the input of the Device Under Test (DUT) and measure the output power density of the component with the assistance of a Spectrum Analyzer. The ENR usually works at a relatively wide frequency band and has a known noise figure parameter at different specified frequencies. The Y-factor corresponds to the ratio of the output noise power when the noise source is turned ON and OFF. Once the test engineer is in conditions to use the ENR noise source and determine the Y-factor, the procedure of this method consists in turning ENR source ON and OFF and measure the output power density using a Spectrum Analyzer. Once those results are taken, and considering Y and ENR as linear (unitless) values, the noise figure can be calculated as: [7]

$$NF = 10 \log_{10} \frac{ENR}{Y - 1} \quad (2.6)$$

Another alternative way to measure the noise figure parameter is to use the Gain method. This method has a more convenient setup compared to the Y-factor approach since the procedure consists in connecting a 50Ω load at the input of the DUT and measuring the output noise power using a Spectrum Analyzer. This method also uses the known noise power density at ambient temperature, $P_{N,amb}$, ($T = 290K$) given as: [7]

$$P_{N,amb} = k \cdot T \cdot BW = -174 \text{ dBm Hz}^{-1} \quad (2.7)$$

Where k is the Boltzmann's constant, T is the noise temperature, and BW is the noise bandwidth in Hz. If the gain of the DUT is known, then the noise figure measurement is given as:

$$NF = P_{N,out} - P_{N,amb} - \text{Gain} \quad (2.8)$$

In short, as long as the gain of the DUT is known and the Spectrum Analyzer allows, it is possible to determine the noise figure. Normally, the biggest constrain of this technique is related to the noise floor of the Spectrum Analyzer that prevents making proper noise power density measurements when the gain is relatively low. [7]

Overall, each method to determine the noise figure parameter for a given RF component offers different advantages and disadvantages depending on the type of device to test. That said, it is important to realize the pros and cons of each method to select the most suitable one for each application. Table 2.2 presents the three described approaches for comparison.

Method	Applications	Advantages	Disadvantages
Noise Figure Meter	Very Low NF	Convenient and useful for very low NF measures (≈ 0.2 dB)	Expensive and with limited frequency range
Y-factor	Wide Range of NF	Independent of Gain and frequency range.	For high values of NF, the error can be large and requires an ENR source
Gain Method	Very High Gain and / or NF	Easy to setup and accurate for very high NF measurements at any frequency range	Limited by the Spectrum Analyzer noise floor. Not suitable for systems with low gain and/or NF

Table 2.2: Comparison between different methods to measure NF. Credit: Maxim Integrated [7]

P1dB

Any real amplifier has a zone of saturation where it no longer maintains its linearity. At this point, for increasing input power, the output power remains at the same saturated value. Naturally, this is not desirable, and so, in the RF domain, it is important to look for amplifiers with good levels of linearity so that their nominal gain is achievable. For that, there is a parameter, often given by manufacturers, that quantifies the level of the amplifier linearity by stating the point where the power level of the output decreases by 1 dB, in comparison to the ideal linear characteristic of the amplifier. This point is called the 1 dB compression point, and it is used as a way to measure how linear the amplifier is. Figure 2.5 shows a representation of the 1 dB compression point of an amplifier. [2]

IP3

When a single frequency input is applied to a non-linear network (like an amplifier), the initial frequency is converted to other frequency components. Most of the created frequency components will appear outside the passband region of the amplifier, so they will not interfere with the output signal. The situation differs when the input signal is composed of more than one closely spaced frequencies. In this case, not only harmonic components but also intermodulation products get created. These products are linear combinations of the initial frequencies, and some of them can appear inside the passband region of the amplifier. When that happens, there is a problem because they will also be amplified and possibly create disruptive distortion in the output signal. The Third-order Intercept Point (IP3) corresponds to a theoretical intersection point where the first-order and third-order powers would be equal. This value is often used as a way to characterize the linearity

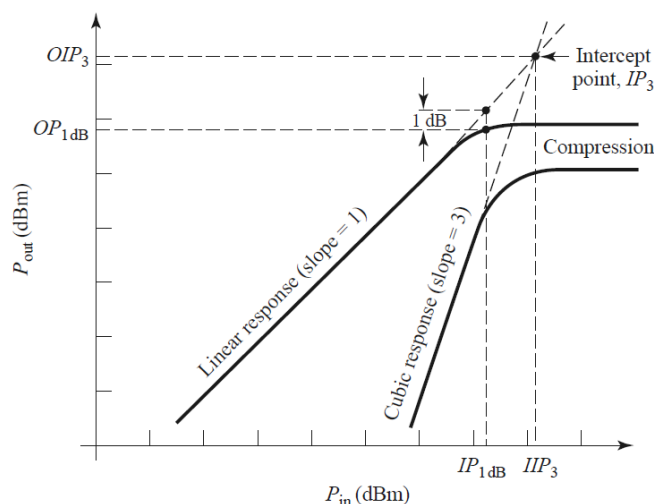


Figure 2.5: 1 dB compression point and third-order intercept diagram. Credit: *Microwave Engineering* (page 515), David M. Pozar [2]

of a non-linear device, like a RF amplifier. Figure 2.5 illustrates how to locate the IP3 point in an amplifier device. [2]

The presented RF parameters are relevant measurements for a correct selection of components. For amplifiers, it is interesting to have good values of linearity, so one should look for high values of P1dB and IP3 coupled with adequate levels of gain. For LNAs in particular, besides linearity and gain, noise figure is also a crucial value to take into account. For other components like switches and filters, it is essential to look for the insertion loss characteristic to assess the attenuation that will take place between RF input and output. Lastly, the return loss will be present in all RF components, and it is desirable to keep it as low as possible.

2.3.4 RF Power Division

Power dividers are passive components whose goal is to divide input power into equal or unequal parts at their respective output. These devices may have three, four, or more ports and are ideally lossless. Due to the laws of conservation of energy, a three-port network cannot be simultaneously lossless, reciprocal, and matched at all ports, so a compromise between these aspects must be made when designing power dividers. [2]

This compromise originates different types of power dividers like the resistive divider, the T-junction divider, or the Wilkinson Power Divider (WPD). The resistive divider is a network that is allowed to be lossy to be matched at all ports and reciprocal. On the other hand, the T-junction power divider is another type of power divider. It allows for lossless and reciprocal power division, but it is not matched at all ports. [2]

Even though it is not possible to achieve the ideal scenario of lossless, reciprocal, and matched at

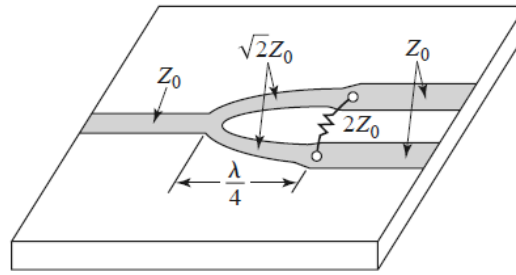


Figure 2.6: WPD topology Credit: *Microwave Engineering* (page 328), David M. Pozar [2]

all ports power division, the Wilkinson Power Divider is a topology that achieves a very interesting compromise. Like the resistive divider, it allows for matched and reciprocal power division, and it is also lossy, but it has the useful property of appearing lossless when the output ports are matched (in this case, only reflected power from output ports is dissipated). [2]

The WPD is a robust power divider that uses a quarter-wave transformer (described in section 2.3.2) and a resistive component with value $2Z_0$ (for equal power division). It can be implemented in a stripline or microstrip transmission lines. Figure 2.6 shows the topology of a WPD. [17]

2.4 Market Survey

As Figure 2.1 hints, RF front-end modules usually encompass PAs, LNAs and RF switches. Not shown in that figure, but also relevant for good RF design are RF filters (e.g. Low Pass Filters, Band Pass Filters...). These are the key components of interest for this thesis work.

Considering the market of RF front-ends, even though there is a wide variety of applications associated with it (e.g. automotive, military, consumer electronics...), it is at the wireless communications domain where significant development has been made in recent years. This is justified by the increasing development of mobile devices (such as smartphones or tablets) as well as the rise of low-cost nanosatellites. [3].

Looking into Figure 2.7, it is visible that the market concentration for RF front-end modules is rather fragmented, in the sense that there are a variety of competitors offering viable alternatives for RF design. From that group, the following were considered relevant manufacturers for this thesis work: Qorvo, Broadcom, Infineon Technologies AG, Murata Manufacturing, NXP Semiconductors, Skyworks Solutions, Qualcomm Technologies, TDK Corporation, Analog Devices, and Mini-Circuits. [3]

At this point, it is important to understand the specifications of the front-end to be developed in the context of the SARIA experiment, to survey the market and check if there is a suitable

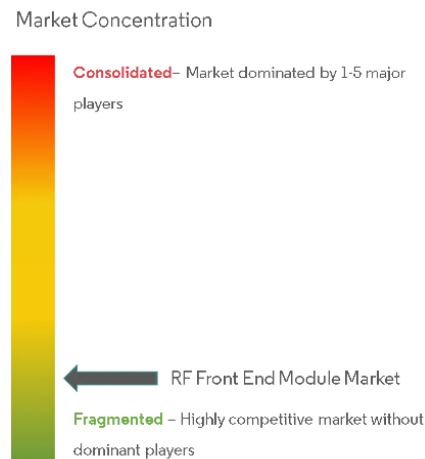


Figure 2.7: Concentration of the front-end module market. Credit: Mordor Intelligence [3]

solution. As explained in section 2.2.1, the SARIA RF front-end needs to properly amplify the transmitting signals, amplify the receiving signals without introducing much noise, and switch between transmission and reception modes without excessively influencing the passing RF signals. The front-end shall also work at the 5.8 GHz C-band frequency. Other than that, it should also minimize the overall cost of manufacturing the module. [4]

Considering the available manufacturers and the specification of the SARIA experiment, it made sense to survey the market for suitable solutions. The search looked for front-end modules with transmission and reception chains with a T/R switching method. These modules had to be compatible with the C-band frequency of 5.8 GHz. Aspects like the switching insertion loss and attenuation, TX/RX gain and RX noise figure were also of concern. In these conditions, some alternatives were found worth mentioning. Table 2.3 presents a block diagram of these solutions with a complementary specification.

From Table 2.3, it is possible to see some market solutions similar to what is described in Section 2.2.1. The main problem, common to all alternatives, was that they condition the flexibility of the front-end module to be selected. Even though this module has a set of defined specifications, it is also meant to be adjustable in terms of TX gain, PRF used for half-duplex operation, duty-cycle in TX mode...

With that in mind, this study concluded that the development of a custom-made RF module makes sense to obtain a product where it is possible to tune the parameters of gain, attenuation, and losses. This approach will allow more control over the final module parameters and, consequently, its performance.

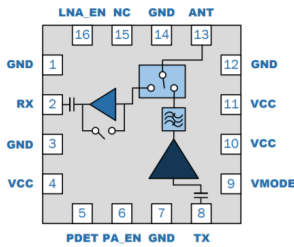
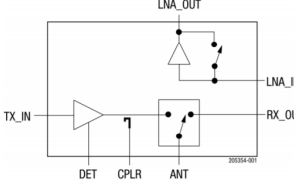
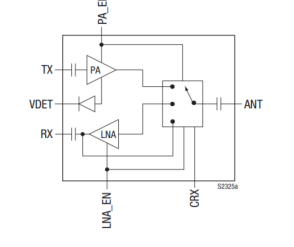
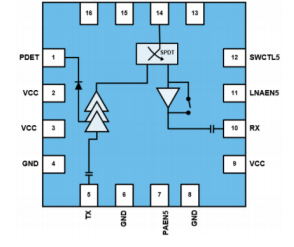
Reference	Functional Diagram	Specifications
Qorvo-RFFM8528P https://www.qorvo.com/products/p/RFFM8528P		• TX gain of 28 dB • RX gain of 13 dB • RX noise figure of 2.5 dB • Switch logic controlled by PA_EN, LNA_EN and VMODE pins
Skyworks-SKY85747 https://www.skyworksinc.com/en/Products/Front-end-Modules/SKY85747-11		• TX gain of 34.5 dB • RX gain of 16 dB • RX noise figure of 2.0 dB • Switch logic controlled by PA_EN and LNA_EN pins
Skyworks-SKY85712 https://www.skyworksinc.com/Products/Front-end-Modules/SKY85712-11		• TX gain of 27 dB • RX gain of 12 dB • RX noise figure of 2.7 dB • Switch logic controlled by PA_EN, LNA_EN and CRX pins
Qorvo-RFFM8516 https://www.qorvo.com/products/p/RFFM8516		• TX gain of 29 dB • RX gain of 12.5 dB • RX noise figure of 2.5 dB • Switch logic controlled by PAEN, LNAEN and SWCTL pins

Table 2.3: Market Survey for RF front-end modules

Chapter 3

System Design

The following chapter will elaborate on the RF front-end system development. It will start by describing what has already been defined in the context of the SARIA experiment. Following that, it will detail what will this thesis work add to the overall SARIA experiment. As it happens, the SARIA project has already had an iteration of the RF front-end module development. Furthermore, a lot of the scientific background and RF viability analyses, is done and documented in the SARIA Student Experiment Documentation (SED). [4] The thesis presented here does not plan to repeat this work but instead use it to hopefully develop a fully functional RF front-end module for this experiment. With that in mind, this chapter will begin with a presentation of the SARIA scientific background. It will then move on to exhibit a preliminary system overview of the RF front-end module to be built. The chapter will end by enumerating the requirements of the module and the selected components to be used.

3.1 SARIA Scientific Background

To perform low-power SAR, there is a need for high gain antennas so that images with significant quality can be obtained. At the same time, large metallic reflectors are rather large and heavy, reducing their portability and constraining their usage in aerial vehicles. With that in mind, the SARIA experiment plans to perform SAR in a high altitude balloon while tackling the problem of mass and portability by using an inflatable antenna. To test the performance of this antenna the experiment will also include a conventional static reflector, also performing SAR at similar areas of operation. [4]

A general RADAR system works by precisely measuring the time difference between the transmitted and received signals to determine the distance from the antenna to the reflecting object. Particularly for SAR, a collection system stores the phase history of the RADAR response at each position, as the vehicle carrying the antenna moves during a given SAR integration period. Pro-

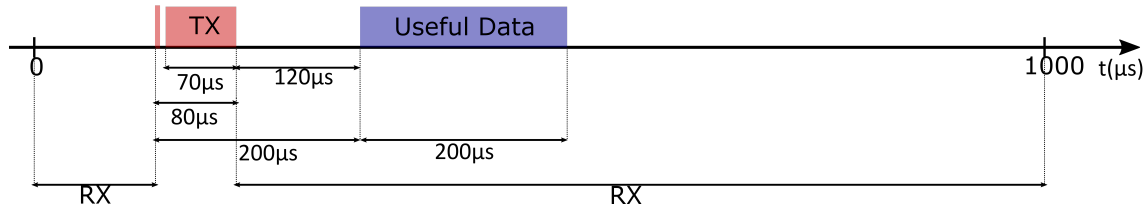


Figure 3.1: Window for signal transmission and reception. Credit: SARIA SED [4]

cessing of these results is made later by compressing the data obtained in two dimensions (slant and along-track) to obtain a final SAR image. [4]

A SDR takes care of the transmission and reception of signals, while a RF front-end module interfaces this block with the two antennas of operation. This module has a RF switch implemented that allows the commutation between transmit and receive mode (time duplexing). The switching is triggered by Root Mean Square (RMS) power detection from the SDR TX port. [4]

3.1.1 Transmitted Signal

The Pulse Repetition Frequency (PRF) considered for SAR operation is 1 kHz. This means that, during SAR sessions, the SDR will be continuously saving data during cycles of 1000 μs. During this period, the transmission of a chirp signal will be preceded by a 6 μs preamble, whose role is to switch the RF front-end from the default RX mode to TX. Following this preamble, there will be 4 μs of radio silence. After the initial trigger time of 10 μs, the useful signal will be sent for a duration of 70 μs. All things considered, the RF front-end is expected to be in TX mode for at least 80 μs. For the remaining time of the 1000 μs cycle, the front-end should be at RX mode. Furthermore, assuming a low altitude stable flight of 21 km with the antenna at a 45° NADIR¹ orientation, the distance to the ground will be 30 km. The time for the signal to travel to the target and back can be given by: [4]

$$\Delta t = \frac{2 \cdot 30 \text{ km}}{c} = 200 \mu\text{s} \quad (3.1)$$

The result from equation 3.1 means that only after 200 μs of the first transmission, useful data will arrive at the SDR. This amounts for 120 μs of radio silence after each transmission is completed. Figure 3.1 shows the windows for TX and useful RX in each cycle.

Figure 3.2 shows how the complete TX signal will look like considering the dummy triggering + dead time + chirp signal.

¹Satellite/balloon/airplane subpoint on the earth's surface that is centered directly below the satellite/balloon/airplane

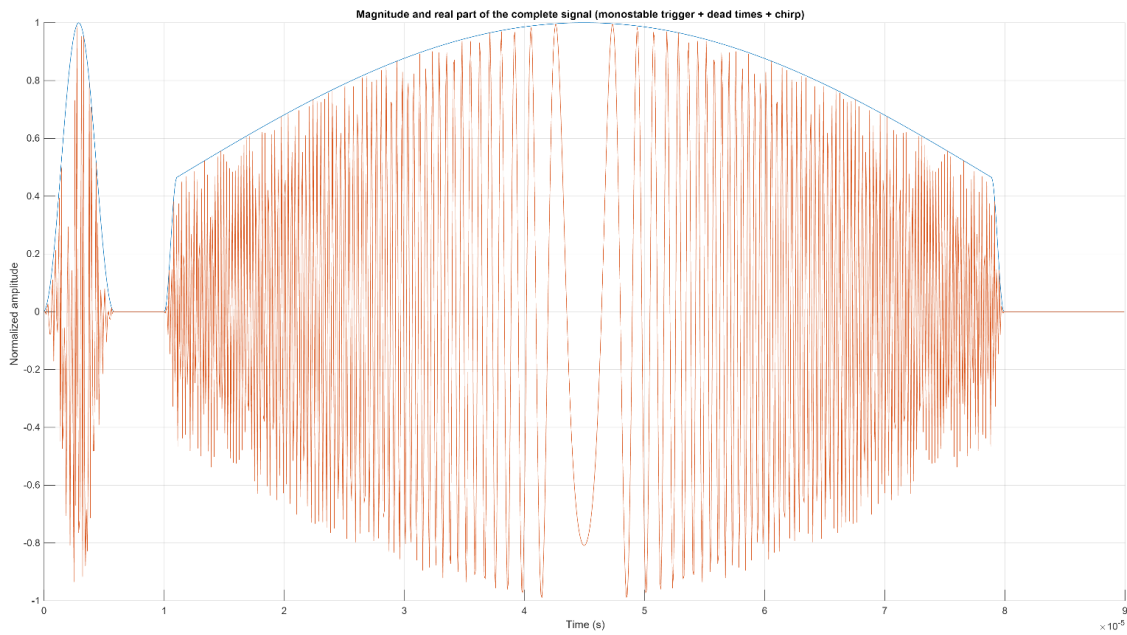


Figure 3.2: Signal to be sent, including the trigger (first part), chirp (second part), and dead times. Credit: SARIA SED [4]

3.1.2 Link Budget

As a way to confirm the RF feasibility of the SARIA experiment, a Link Budget was made. Considering a distance to RADAR targets of 45 km, a positive link margin of 8 dB was obtained. [4] Table 3.1 shows the SARIA Link Budget. Because the SAR technology results from a coupling of RADAR and signal processing technologies [5], the Link Budget will exhibit the effects of both domains to achieve RF viability. As such, the theoretical background for the RADAR part comes from the literature described in section 2.1. Likewise, the fundamentals of the signal processing for improving spatial resolution are presented in section 2.2. Section A.3 details how each parameter was obtained and the considerations made along the way, to achieve the positive link margin.

3.2 System Overview

Regarding the RF front-end block diagram, since the last iteration of development in the SARIA experiment, there has been a major design change in the project that directly affects the front-end module: the introduction of a static reflector to work in alternation with the inflatable antenna. It is up to this module to make the proper alternation between antennas and still allow for half-duplex operation. With that in mind, the system block diagram represented in Figure 3.3 looks to tackle the challenge of inserting a new antenna. Essentially, this block diagram is a more detailed version of the one presented in Figure 2.2, explaining further the switching method, as well as the pulse generation block for this particular application.

Designation	Value	
TX Power	20	dBm
TX Antenna Gain	30	dB
Forward Free Space Losses (@ 45 km)	−141	dB
Projected Radar Cross Section	47	dB
Minimum Distributed Target Reflectivity	−25	dB
Backward Free Space Losses	−141	dB
TX Antenna Gain	30	dB
Diverse System Losses	−6	dB
Total	−186	dBm
Noise	−105	dBm
SNR	−81	dB
Range Compression Gain	27	dB
Azimuth Compression Gain	46	dB
Total	−8	dB
Azimuth Image LPF (40×)	16	dB
Link Margin	8	dB

Table 3.1: Link budget calculation

From Figure 3.3, it is noticeable the PA, LNA and SDR modules as well as the two antennas still maintain the same functionality as the one described in section 2.2.1. The novelty blocks relate to the switching method and the pulse generation to trigger the RF switch. Concerning the switch, it is a DPDT type switch, meaning that with two control signals (VC1 and VC2), it is possible to make two connections from two inputs to two outputs. It is not possible to connect both inputs or both outputs. In this case, there are two states of switching: if the control input CTR_A is at logical 0, the connections PA to Inflatable Antenna and LNA to Static Reflector will be established; if the same control input is at logical 1, the PA will connect to the static reflector, and the LNA will connect to the Inflatable Antenna.

This type of switch was selected because it makes possible the switching between TX-RX, and commutation of the two antennas to happen using only one component. Regarding the triggering of the switch, a Power Detector block was employed to detect the transmitted signal preamble described in section 3.1.1. Afterward, a timing circuit in a Monostable (MS) mode is used to create a pulse of around 100 μ s, keeping in mind (from section 3.1.1) that it should be higher than 80 μ s but lower than 200 μ s. Then, depending on the CTR_A input (that selects the antenna of operation), an XOR operation is performed to properly switch the circuit. An inverted block is also needed to obtain the second control signal of the RF switch. Essentially this is simply a

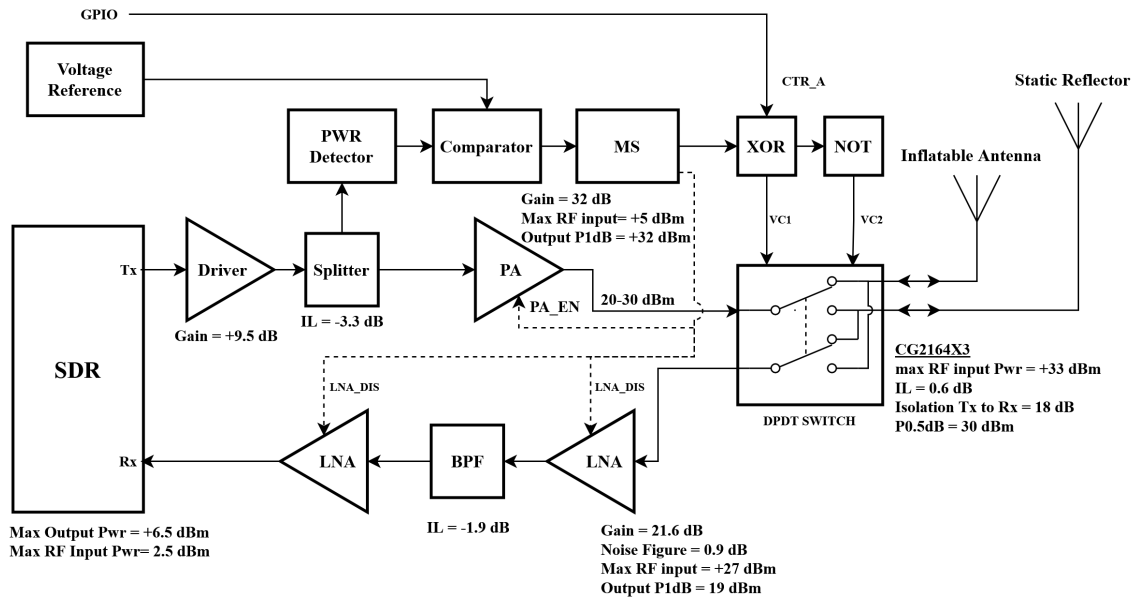
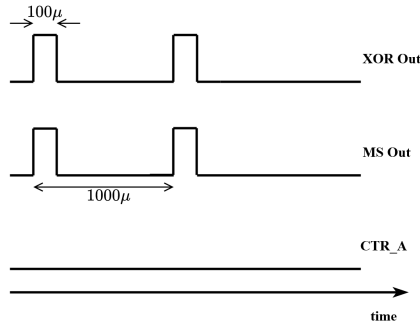
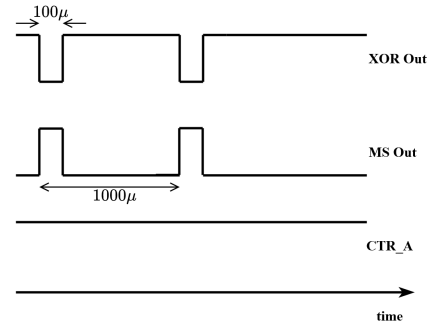


Figure 3.3: Proposed block diagram of RF front-end

detail of the selected switch. Figures 3.4a and 3.4b show the different waveforms used in pulse generation depending on the antenna selection. Figures 3.5a and 3.5b show the switching state of the system depending on the value at the output of the XOR block.

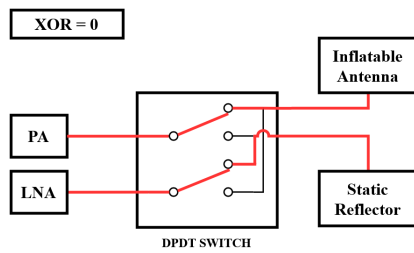


(a) Waveforms of switching signals with CTR_A = 0

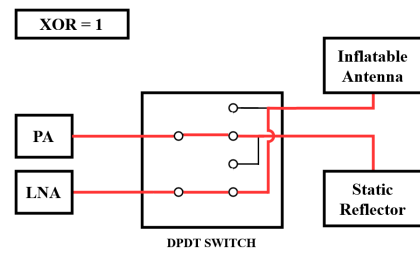


(b) Waveforms of switching signals with CTR_A = 1

Figure 3.4: Waveforms of switching related signals



(a) Switching state with XOR = 0



(b) Switching state with XOR = 1

Figure 3.5: Switching states

3.3 System Requirements

The following section will list a compilation of the functional and non-functional (performance, design, and operational) requirements regarding the RF front-end module.

Functional Requirements

- F.1 The RF front-end shall transmit and receive C-band SAR signals.
- F.2 The RF front-end shall switch between TX and RX modes (half-duplexing).
- F.3 The RF front-end shall select which antenna is performing SAR at a given instant.

Performance Requirements

- P.1-1 The overall TX chain shall have variable gain, sufficient to transmit the stipulated TX power of 20 dBm, but with possibility to increase it to 30 dBm.
- P.1-2 The RX chain shall not introduce excessive noise; all things considered, the overall system losses shall be less than 6 dB.
- P.1-3 The low loss cables used to interface the RF front-end with the antennas shall not introduce an attenuation higher than 4 dB.
- P.2-1 For a given antenna, the RF front-end shall maintain itself in TX mode for an interval of time between 80 μ s and 200 μ s.
- P.2-2 The switching time of the RF switch shall not affect the transmitted signal. Therefore it shall have a duration of only a few μ s.
- P.3-1 The RF front-end shall perform SAR operations of 60 s with each antenna.

Design Requirements

- D.1 The RF front-end, while in TX mode for a given antenna, shall turn off the respective RX chain. Conversely, during RX mode for a given antenna, the TX shall be inactive.
- D.2 The RF front-end shall use SMA connectors for interface with the antennas and SDR.
- D.3 The RF front-end shall use low loss RF cables for interface with antenna.
- D.4 The RF front-end shall have a connection to a GPIO pin for antenna selection.

Operational Requirements

- O.1 The RF front-end shall receive commands through the CTR_A GPIO pin to switch between the antenna of operation.

3.4 Selected Components

Following the establishment of requirements and the description of the system overview, this section will detail the selected components to implement the proposed front-end architecture. Table 3.2 shows a compilation of the chosen materials.

Component	Reference	Datasheet
Driver	GVA-81+	https://www.minicircuits.com/WebStore/dashboard.html?model=GVA-81%2B
Power Amplifier (PA)	TQP5525	https://www.qorvo.com/products/p/TQP5525
Low Noise Amplifier (LNA)	QPL9503	https://www.qorvo.com/products/p/QPL9503
Switch	CG2164X3	https://www.cel.com/documents/datasheets/CG2164X3.pdf
Band Pass Filter (BPF)	BFCN-5750+	https://www.minicircuits.com/pdfs/BFCN-5750+.pdf
Power Detector	ADL5903	https://www.analog.com/media/en/technical-documentation/data-sheets/ADL5903.pdf
Comparator	TLV7031	https://www.ti.com/product/TLV7031
Monostable	NE555	https://www.ti.com/lit/ds/symlink/ne555.pdf
XOR Block	SN74LVC1G86-Q1	https://www.ti.com/store/ti/en/p/product/?p=SN74LVC1G86QDCKRQ1

Table 3.2: Components selected

At this stage, the approach to select the specific references for each component was to consider the following selection criteria presented in order of priority (the first is the most important one, and the last is the least significant aspect):

- How adequate are the specifications of the component in relation to the specification of the RF front-end module.
- Price and delivery time of the components (sample components are preferable).

- Type of packaging of the component (this will greatly impact the easiness of subsequent PCB manufacturing).
- Availability of CAD models for the components.

Chapter 4

Integration and Manufacturing

Following up on the system design, this next chapter will detail the integration process of the front-end module. To do so it will first identify the tools used in the execution of the proposed design. It will then describe how they were used as well as the overall manufacturing process. Over the course of this dissertation, one of the main design choices was to develop test boards for individual key and sensitive components. With this strategy, it is possible to tune the parameters of said components before implementing the actual final front-end PCB. As such, this chapter will display the resulting test boards for both the RX and TX chains. A detailed explanation of the control loop implemented in the final board will follow. The chapter ends by presenting the implemented PCB of the front-end module.

4.1 Tools and Support

Regarding the tools for the development of the front-end board, the RF simulations and design were aided by the Advanced Design System (ADS) software from Keysight, and the KiCAD software facilitated the PCB board design.

Another important design decision was to produce the PCB boards from an external entity - JL-CPCB¹. Since this company is capable of producing multi-layer controlled impedance PCBs for an affordable cost, as long as the controlled impedance stackup option is selected, it was selected for the fabrication of every PCB.

PCB design for RF applications are usually done in 2 or 4 layer boards. When using 2-layer boards, the top layer is used for component placement, RF and signal routing, while the bottom layer is meant mostly for ground connections to allow the shortest path for return currents. In a 2-layer PCB the cost of production is lower, and it is even possible to achieve similar RF performance than in a 4-layer design. But for that, the signal routing and component placing must be more carefully

¹<https://jlcpcb.com/>

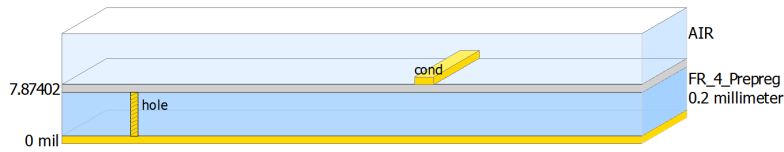


Figure 4.1: Considered substrate for ADS simulations

thought. Another important detail of these designs is that, to achieve feasible $50\,\Omega$ trace widths, the thickness of the board is generally limited to 0.8-1.0 mm. [18]

On the other hand, 4-layer boards allow for simple routing between ground and power planes and more relaxed routing considerations than a correspondent 2-layer design. In a typical RF 4-layer PCB, the first (and top) layer is used for component placement and signal routing, the second layer is used as a ground plane, the third is used as a power plane, and the fourth (and bottom) layer is used both for signal routing and as a ground plane. Even though the cost of production rises, the thickness of such a board also increases (to around 1.5 mm), providing a more robust mechanical strength, compared to a 2-layer board. [18]

That said, and since there is the possibility to do so, a 4-layer design was made for this application, considering the controlled impedance stackup provided by JLCPCB. The definition of the substrate is crucial for ADS RF simulations, mainly between the top layer (with RF signal routing) and the second layer (ground plane). For this design, conductors with $35\,\mu\text{m}$ of thickness were considered, along with an FR-4 Prepreg substrate of 0.2 mm with a dielectric constant of 4.25 and loss tangent of 0.015. Figure 4.1 shows the considered substrate in all the RF simulations performed.

4.2 ADS Design Flow

The RF design of every board implemented in this thesis work was aided by the ADS software. This section will describe how was the software used to achieve that goal. The methodology below was maintained for all intermediate boards. The final design was obtained from a combination of all the intermediate designs. As such, the sections following this one will not repeat this information, but it will be implied that the respective ADS design was made according to the same procedure.

The ADS design started with an analysis of the component datasheet. In particular, it was relevant to look for an application circuit of a given component to understand how the respective final board should look.

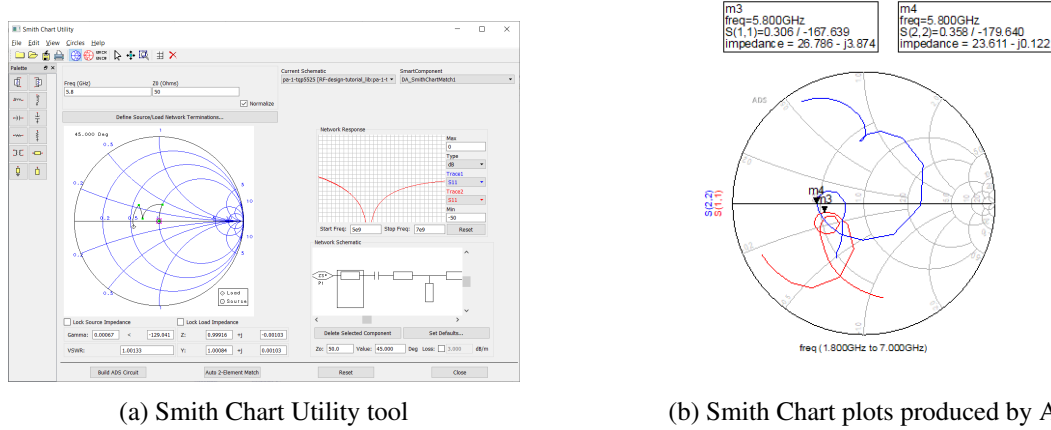


Figure 4.2: Design flow the baseline matching networks

From the initial inspection of the datasheet, the next step was to obtain the de-embedded S-parameters for each component and test them in ADS to obtain a set of preliminary results that would serve as a baseline for the following adaptation. As stated in section 3.4, selecting components with available CAD models was a concern, exactly to avoid setbacks at this stage. All of the de-embedded information was obtained from each of the manufacturers of the individual RF components. No further de-embedding measurements were done during testing.

After this initial test, the actual adaptation at the input and output took place. ADS has different built-in tools to assist the implementation of matching networks. In this thesis, the Smith Chart Utility was extensively used for this purpose, as shown in Figure 4.2a.

The way to obtain the initial matching networks was to measure the impedance looking into the de-embedded component using the S11 Smith Chart plot available in ADS, as shown in Figure 4.2b. From the input impedance measurement, the Smith Chart Utility tool generated an appropriate input matching network. With both the newly created input network and the RF component, the output impedance of the overall circuit changed. Again, the process was to measure the output impedance using the S22 Smith Chart plot produced by ADS, also shown in Figure 4.2b. Using the conjugate output impedance measurement, the ADS Smith Chart tool produced the output matching network. In these conditions, the initial input and output matching networks were created to adapt a given RF device.

Once the Smith Chart Utility tool yield the baseline input and output matching networks, it was necessary to include the substrate, previously shown in Figure 4.1, for the stubs of the adaptation circuits. On the other hand, it was also required to account for the effect of the pads and spacing of the passive and active components. In all the implemented boards, 0603 size resistors, capacitors and inductors were considered. After these changes, the schematic would look like the one presented in Figure 4.3 with the respective substrate being defined in the highlighted red square. These alterations usually meant that the performance of the circuit became a bit degraded so it was necessary to perform some optimization in the lengths and widths of the stubs, as well as the

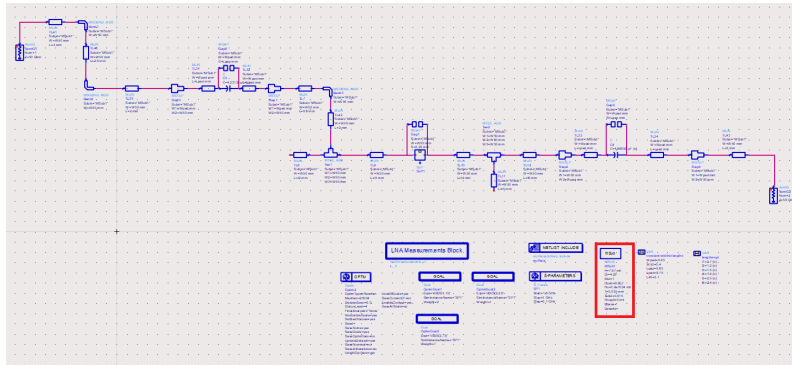


Figure 4.3: Example of the ADS schematic for optimization of stub parameters

values of the passive devices. For that, the ADS optimization cockpit, presented in Figure 4.4, came in handy.

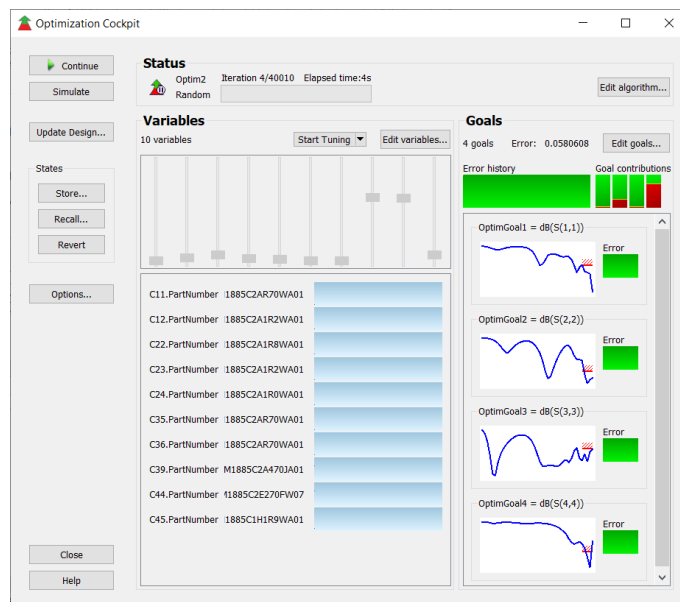


Figure 4.4: Example of the ADS optimization cockpit

After establishing the schematic with the adequate input and output matching networks, the ADS layout for electromagnetic simulation followed. Since ADS allows for electromagnetic simulations, this feature was used a lot to incorporate variables that are not considered with just schematic simulations. At this point, a proper setup was required in terms of selected substrate, frequency plan, configuration of ports... Figure 4.5 shows an example of this kind of setups.

From the electromagnetic setup, the correspondent electromagnetic model emerged. In these conditions, it was finally possible to perform a co-simulation using the obtained model in conjunction with the active and passive devices to be used.

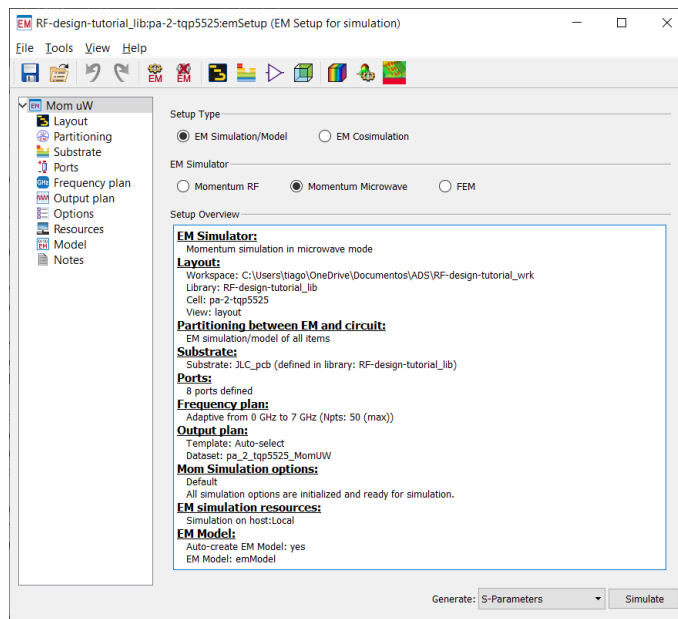


Figure 4.5: Example of the setup for electromagnetic simulation

Once a proper ADS design was achieved, the migration between ADS and KiCAD (to implement the actual board layout) was relatively straightforward. The transmission lines designed in ADS were exported to KiCAD and the PCB layout continued from that point.

4.3 RX Chain Design

Taking into account the ADS design flow described above, the actual design process of the front-end module started. At this point, it was relevant to break down the final system into smaller subsystems and design them individually before joining everything together. With that in mind, the following section will describe the design flow of the RX chain of the front-end board, including the design of the LNA and the BPF.

4.3.1 Low Noise Amplifier (LNA)

Starting with the design of the QPL9503 LNA, the first step was to analyze the respective datasheet (present in table 3.2). This LNA offers a noise figure of 0.95 dB at 5.5 GHz, operation between 0.6-6 GHz with an integrated shutdown control pin. Mainly because of these specifications, this device was the selected LNA for this particular application.

This LNA was used in the respective Time Division Duplex configuration to disable it during Transmission of the front-end module. The LNA (and the overall RX chain) is meant to be turned on by default. However, it must be switched off when a transmission is taking place to avoid

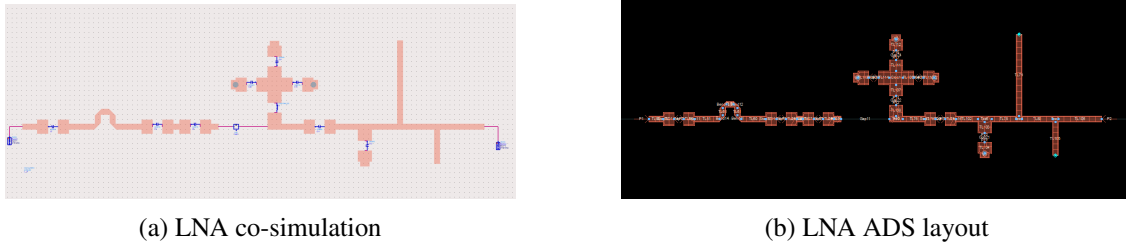


Figure 4.6: LNA ADS design

saturation of the LNAs and of the SDR at reception. Once a given transmission is completed, the LNA must be re-enabled while the PA (of the TX chain) must be disabled. At this point, it is crucial to guarantee that there is no possibility for the LNA and PA to be enabled at the same time, so an RC circuit was used at the Shut-Down pin of the QPL9503 to delay the re-enabling of the LNA just long enough to ensure that the PA is already disabled once the LNA turns back on.

The ADS software was used to design the layout of the RF lines of this part of the overall circuit. This circuit was then tested individually in a preliminary PCB containing the QPL9503 as the only active component. Figure 4.6a shows the ADS simulation of the RF part of this circuit, while figure 4.6b shows the layout considered in electromagnetic simulation. Figure 4.7 shows the test board schematic with the design considerations described above, and Figure 4.8a shows the respective test board layout also taking into account the ADS layout design. Figure 4.8b shows this board after assembly.

4.3.2 Band Pass Filter (BPF)

Regarding the development of the Band Pass Filter (BPF), the BFCN5750+ was the selected device. This device has the simple but key role of filtering the received RF signals after the first LNA. That said, it was important to find a filter with a not too wide passband around the region of 5.8 GHz. That was the case for the BPFCN5750+ by Mini-Circuits. A bonus in its acquisition

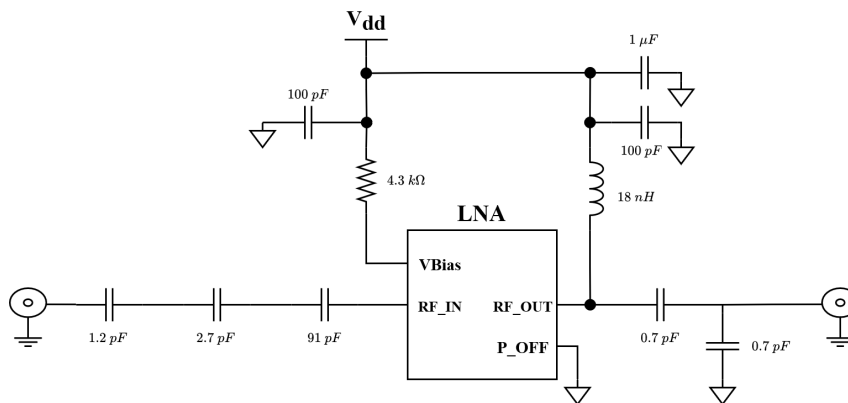
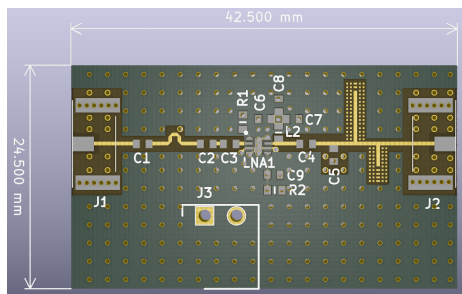
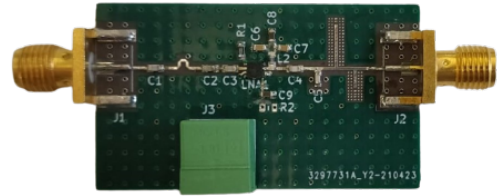


Figure 4.7: LNA test board schematic



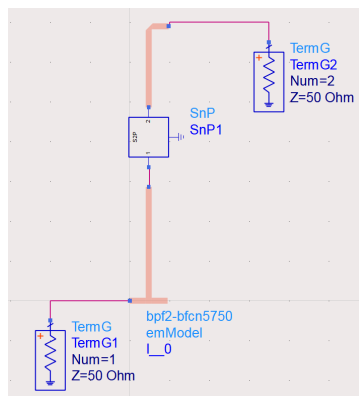
(a) LNA test board layout



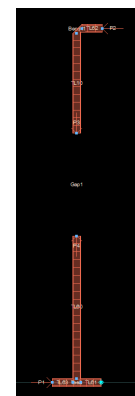
(b) LNA test board view

Figure 4.8: LNA test board design

was that it was possible to obtain enough samples for this device with no added cost. Considering the nature of this component, a test board was also implemented to fully characterize the device. Figure 4.9a shows the RF simulation performed with the electromagnetic simulated layout present in figure 4.9b. Figures 4.10a and 4.10b show the test board design of the implemented board.

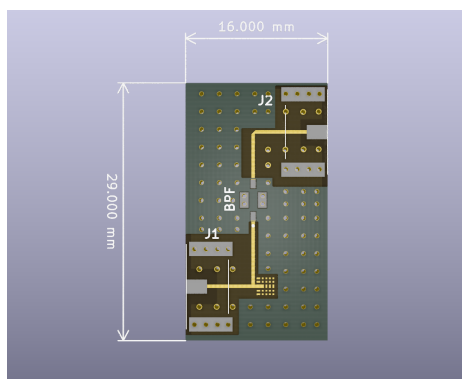


(a) BPF co-simulation

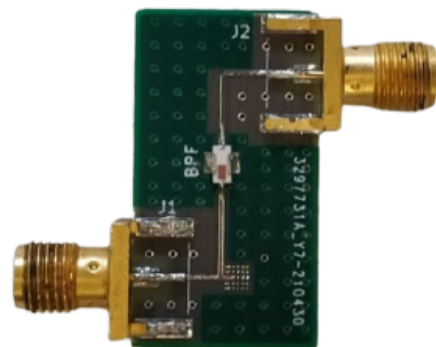


(b) BPF ADS layout

Figure 4.9: BPF ADS design



(a) BPF test board layout



(b) BPF test board view

Figure 4.10: BPF test board design

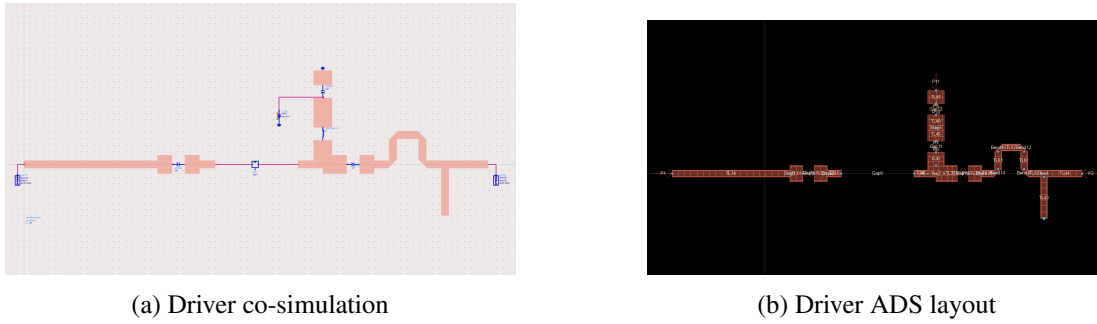


Figure 4.11: Driver ADS design

4.4 TX Chain Design

Following the RX chain design, this next section will describe the development of the individual blocks of the TX chain. Similar to the RX chain, due to the nature of each component used and the respective design goals, it made sense to develop individual boards for testing and validation of the RF design. This was the case for the Driver and the PA blocks.

4.4.1 Driver

Regarding the design of the Driver, it is crucial to understand its role in the front-end system, to adapt its behavior accordingly. This driver block has the pivotal function of establishing an interface between the SDR TX port and the remaining of the TX chain. As such, it is essential to optimize the respective input return loss to achieve an appropriate impedance matching at the beginning of the TX chain. This was the main design goal at this point.

With that in mind, the GVA-81+ by Mini-Circuits was selected. From the datasheet of this device present in Table 3.2, it has an interesting input and output return loss over a relatively wide frequency range. It was also possible to acquire samples of this amplifier, further reducing the overall cost of the board fabrication. Figures 4.11a and 4.11b show the RF design of the component, whereas Figures 4.12 and 4.13a illustrate the test board schematic and final layout. Figure 4.13b shows the Driver board after assembly.

4.4.2 Wilkinson Power Divider (WPD)

This front-end module is expected to switch between TX and RF contexts based on RMS power detection. To do so, and as referred to in section 3.1.1, in the signal to be transmitted, there is a first $6\ \mu\text{s}$ part that is used for power detection. This detection has to be performed by the front-end board and starts by splitting the signal into two parts. One part follows the RF TX chain to be eventually transmitted while the other goes on to be detected by the Power Detector module. After

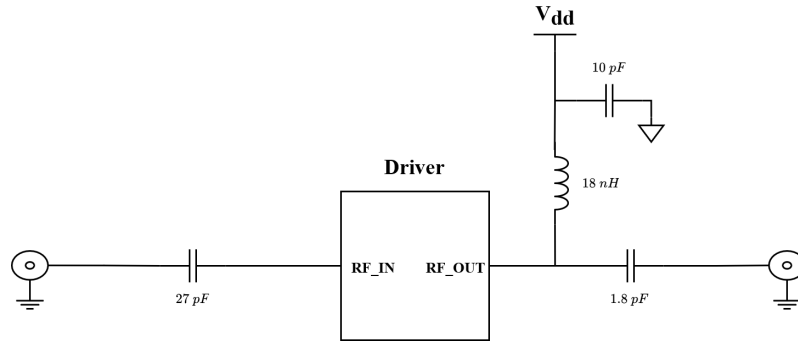


Figure 4.12: Driver test board schematic

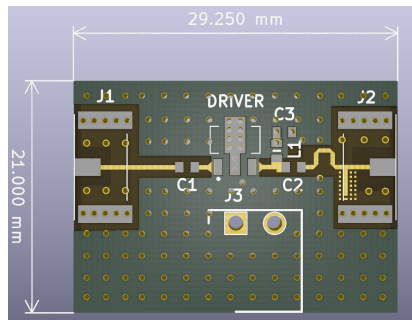
that, there is a control loop that creates a $100\ \mu\text{s}$ window for the module to switch from the default state of RX to TX.

Given the need for this block to equally divide the RF power in 2 parts, and following up on the RF Power Division study carried out in section 2.3.4, a Wilkinson Power Divider (WPD) type circuit was designed to perform this specification following the ADS design guide present in [17]. Figures 4.14a and 4.14b show the RF design of the developed WPD.

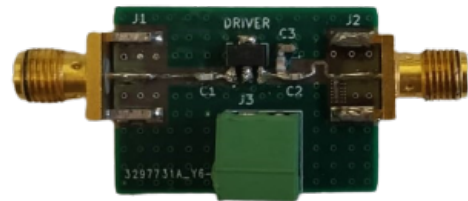
4.4.3 Power Detector

The power detector module is a key part of the overall front-end system since it interfaces the RF signal part with the control loop that allows for switching between TX and RX contexts. As such, it is crucial to have both an adequate RF design at the input of this device and appropriate handling over the control voltage that will be present at the output. This module also corresponds to the first step of the control loop that leads to the RX to TX switching while containing an RF input that requires careful design. This section will elaborate on the RF design. Section 4.6 further details the control logic of the overall system.

Considering the frequency of operation, input power, switching times, and output voltage levels,



(a) Driver test board layout



(b) BPF test board view

Figure 4.13: Driver test board design

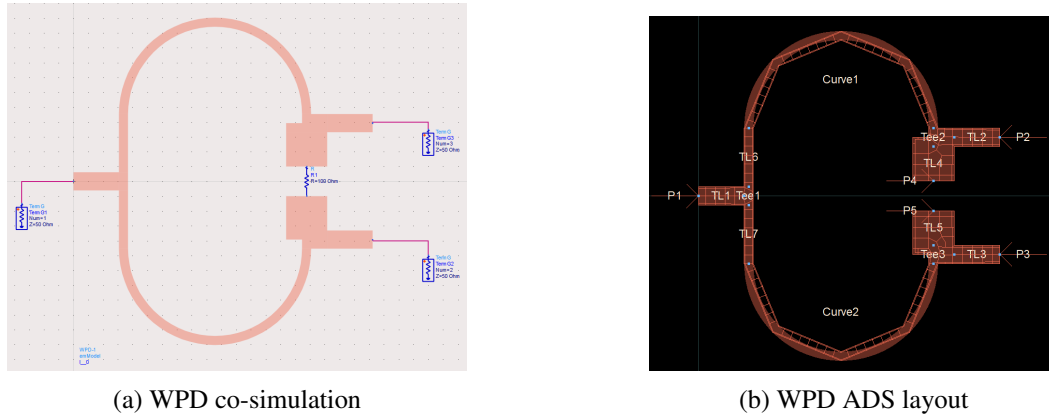


Figure 4.14: WPD ADS design

the ADL5903 by Analog Devices was selected. An analysis of the datasheet of this device (present in Table 3.2) exposed that some passive devices needed to be appropriately selected depending on the circuit application. As it turns out, this component allows for a slightly different RF input interface with the addition of a $127\ \Omega$ resistor at the RF input pin. After studying the changes in adding this resistor, it was concluded that, for the frequency of 5.8 GHz, there was a better input return loss with the placement of said resistor. That said, Figures 4.15a and 4.15b show the ADS RF design for this block.

Another design consideration had to do with one of the external capacitors to be placed at the device C_{RMS} pin. From the ADL5903 datasheet, the selection of this component influences the averaging function for the internal RMS computation. Using a smaller value for the C_{RMS} affects the output voltage signal by having a quicker response time but with a higher level of noise. Increasing the C_{RMS} value worsens the response time but reduces the output noise level. With this detail in mind, and considering the design goal of creating a $100\ \mu s$ control signal within the control loop, a value of $3.3\ nF$ was selected to aim for an output rise time of around $1\ \mu s$.

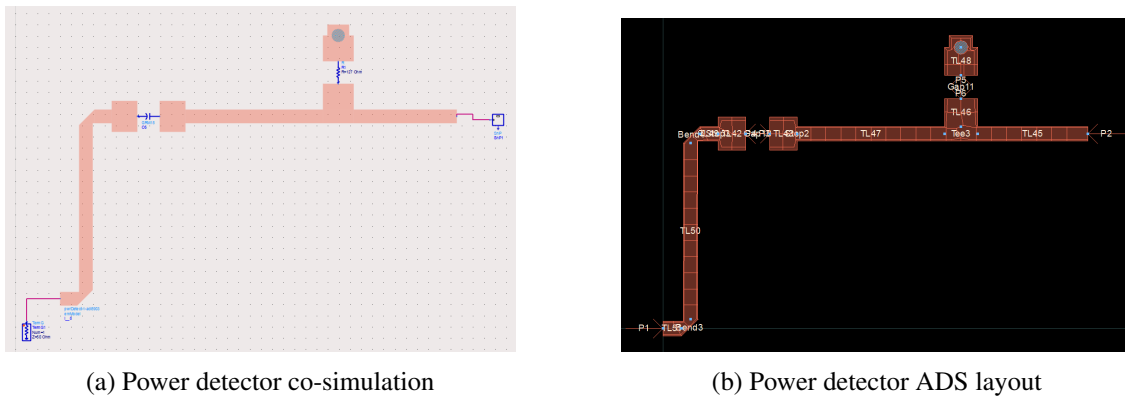


Figure 4.15: Power detector ADS design

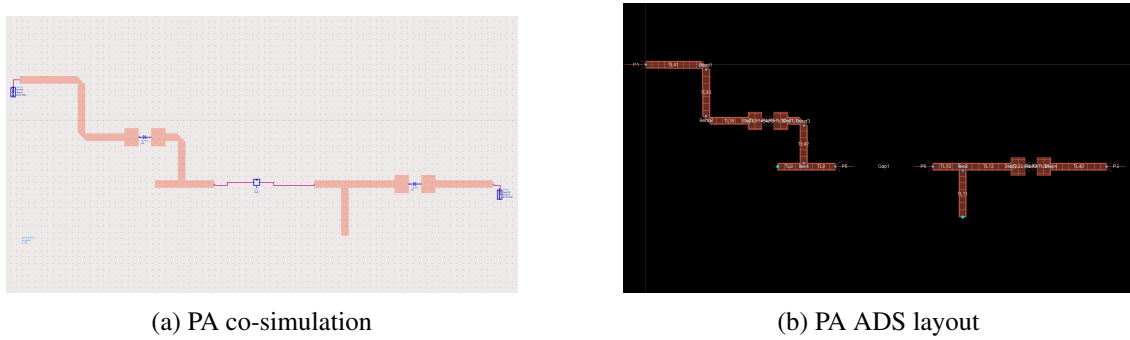


Figure 4.16: PA ADS design

4.4.4 Power Amplifier (PA)

The last element in the TX chain is the PA module. This component has the function of amplifying the input RF signal to sufficient levels for transmission. After this block, the RF signal will have no further amplification, so it has the responsibility of leaving the RF signal with enough energy for a full transmission.

When selecting such a device, besides gain, frequency of operation, input and output return loss, and power consumption, a critical specification to look out for, was the output P1dB, as well as the output saturation point of the amplifier, keeping in mind the requirement P.1-1 for a Transmission power of 30 dBm. That said, the TQP5525 by Qorvo was selected to perform this specification.

Being one of the most important and sensitive blocks in the whole front-end system, it made sense to develop a test board to ensure the functionality of the device and the respective RF design. Figures 4.16a and 4.16b illustrate the RF ADS design while Figures 4.17 and 4.18a display the design of the test board. Figure 4.18b shows the board after fabrication and soldering.

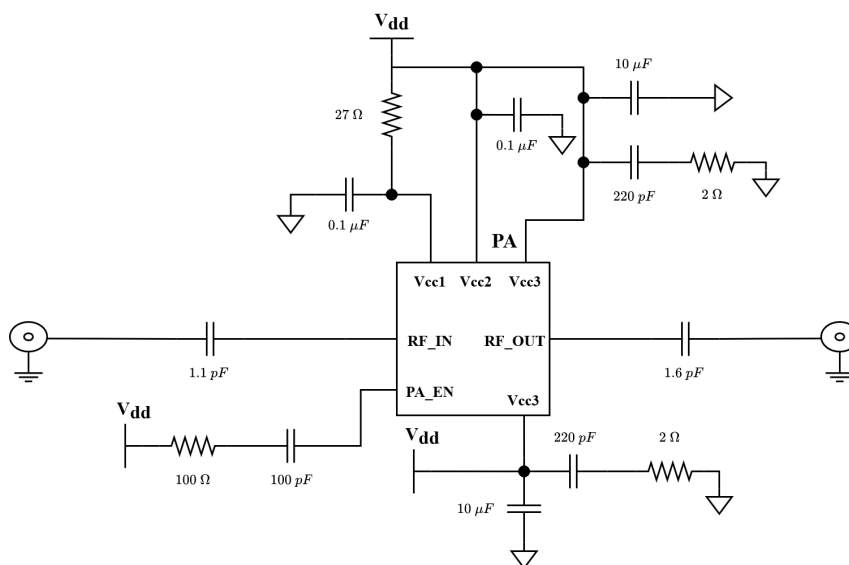


Figure 4.17: PA test board schematic

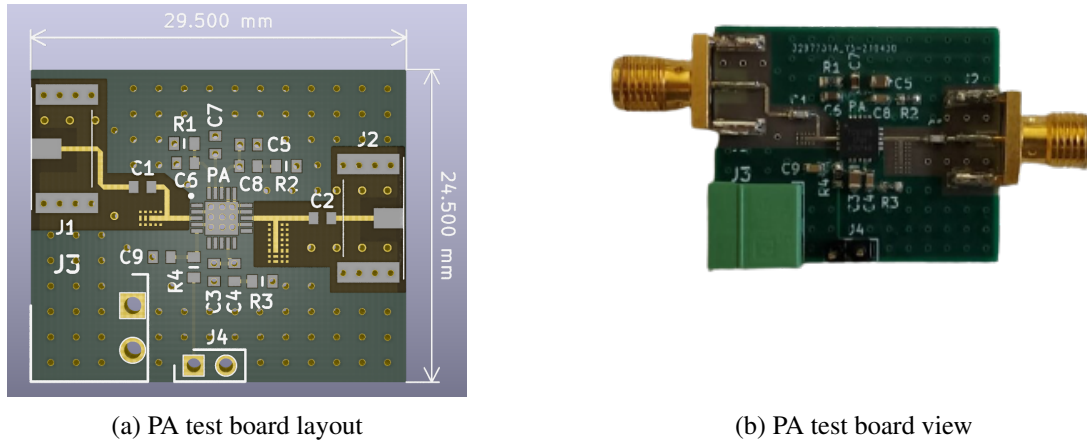


Figure 4.18: PA test board design

4.5 RF Switching Design

Regarding the RF switching part of the overall front-end module, as stated previously in section 3.2, a DPDT type switch was thought of as a way to accommodate the interface between TX and RX paths with both the inflatable antenna and the static reflector. After selecting the type of switch, it was necessary to survey the available market for a component with a relatively low insertion loss and appropriate isolation between the respective TX and RX ports. With these two key aspects in mind, the CG2164X3 by CEL was selected (datasheet present in Table 3.2).

According to its datasheet, this RF switch allows for an insertion loss of 0.6 dB at 4.9-6.0 GHz, with an isolation between TX and RX of 18 dB for the same frequency band. It also offers a fairly good return loss at the 5.8 GHz frequency and allows up to 33 dBm of input power. Mainly due to these specifications this component was chosen.

After the selection of the switch, the next step was to initiate the respective ADS design. Here, it became clear that this component is a rather sensitive part of the whole system, and it was worth developing a test board to evaluate the individual performance of this component. Because of its nature, it was also important to carefully think about the positioning of the transmission lines to adequately fit with the TX and RX chains, as well as the two antenna ports. Another important detail mentioned in the device datasheet was to include DC blocking capacitors at all RF ports. Figures 4.19a and 4.19b show the ADS design of this part. Figures 4.20 and 4.21a show the test board design of the RF switch board. After assembly, the board showed by Figure 4.21b was attained.

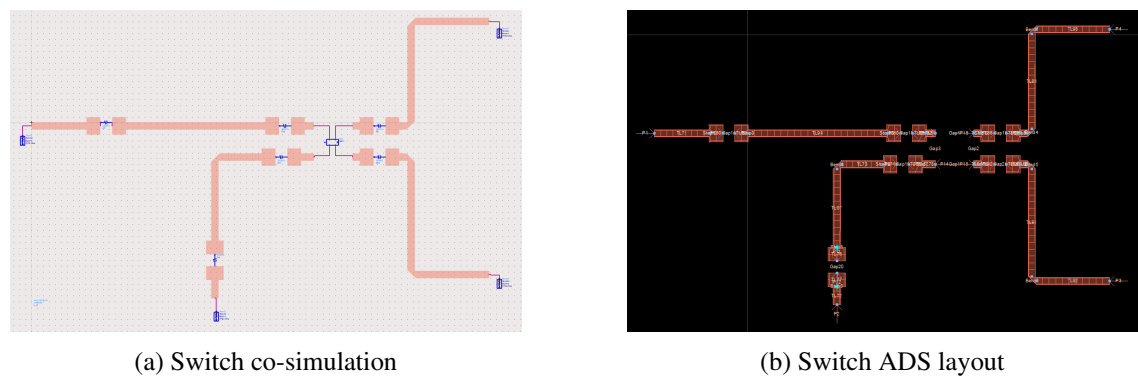


Figure 4.19: Switch ADS design

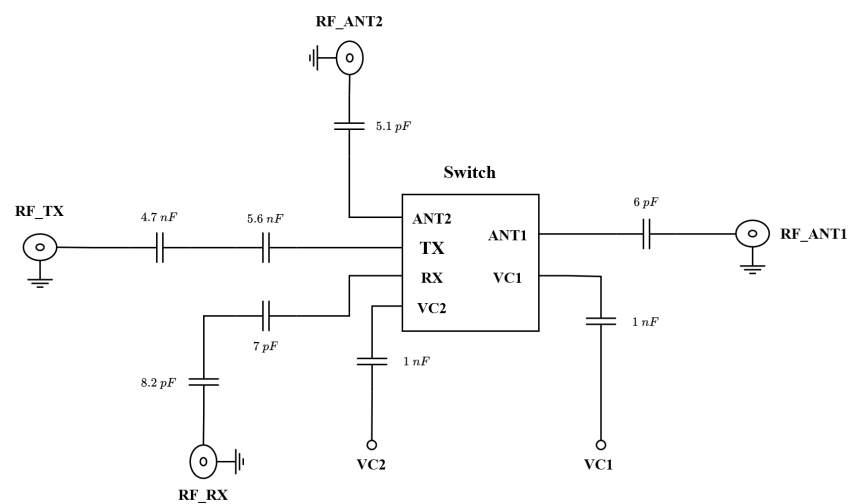


Figure 4.20: Switch test board schematic

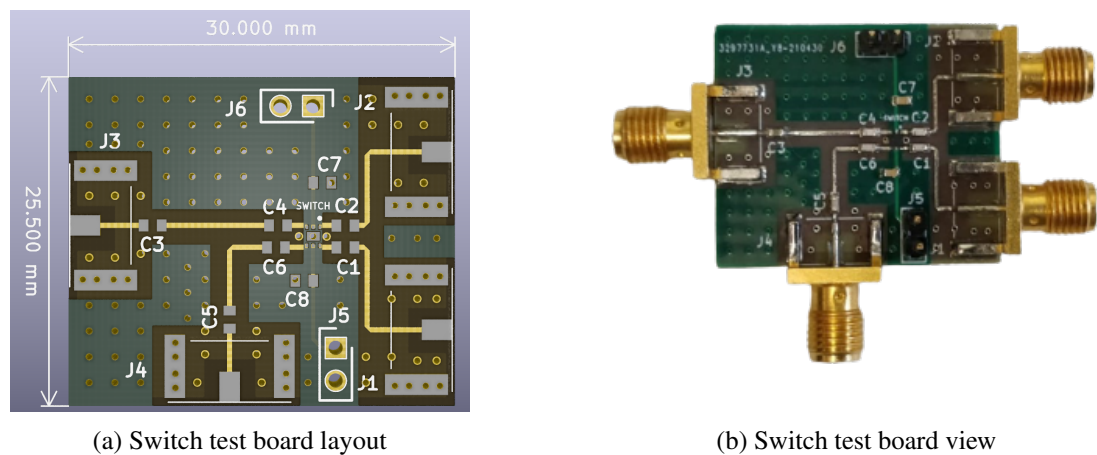


Figure 4.21: Switch test board design

4.6 Control Loop

As stated before, the resulting front-end of this thesis work, besides being able to receive and transmit C-band signals, is also expected to switch between TX and RX contexts based on RMS power detection. In this case, this task is achieved by the Control Loop circuitry represented in Figure 4.22. This circuit corresponds to the blocks: Voltage Reference, PWR Detector, Comparator, MS, XOR, and NOT of the detailed block diagram previously showed in Figure 3.3.

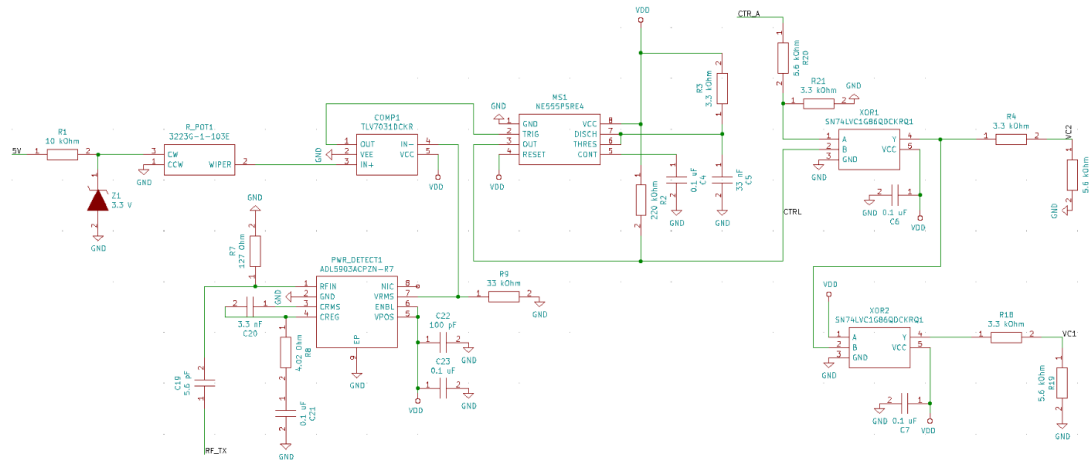


Figure 4.22: Control loop circuit

Regarding the circuit of Figure 4.22, from left to right, it starts with a voltage reference created by a Zener Diode and Potentiometer. By using these components, it is possible to adjust this reference so that the corresponding voltage level is not higher than the output voltage of the power detector. At the same time, it shall be higher than the noise coming out of the power detector voltage so that there is no confusion when RF power is entering the TX chain. In this circuit, the frontier between RF signals and control signals happens at the Power Detector component. This device produces an output voltage in relation to the corresponding input RF power. From the information present in the datasheet of the ADL5903 (present in Table 3.2) the surrounding passive components were carefully selected with considerations for RF adaptation, output rise time, output noise filtering, and decoupling aspects.

The voltages created at the Reference Voltage and PWR Detector blocks converge to the Comparator block, which in turn, will be in a high voltage state until there is a significant power detection voltage coming in. At that point, the Power Detector will drive the output of the Comparator to a low voltage state. This output connects to a NE555 timing circuit that is in a Monostable (MS) configuration. By having a low voltage state in the TRIG input, this component can create an approximately 100 μ s pulse. The duration of this pulse can be adjusted, but it was configured in a way that creates the expected window for full transmission of the front-end module.

The XOR module present after the MS block is intended to carry out the switching function earlier described in section 3.2 and represented in Figures 3.4 and 3.5. Depending on the CTR_A GPIO input, it will either leave the MS output as is or it will invert it. The output of this XOR will then serve as one of the control inputs to the considered RF switch.

The other XOR module present in Figure 4.22 is there to invert the output of the first XOR block. Effectively it is carrying out the NOT function of the detailed block description of Figure 3.3. The inverter block is used merely because of the nature of the RF switch. In the case of the CG2164X3 component (refer to Table 3.2 for the respective datasheet), the control inputs must always be in inverted logic states.

In this way, this sub-circuit is expected to carry out the control loop functionality considering the components used, namely the RF switch and the Power Detector.

4.7 Final Front-end Design

Up to this point, the design of all RF components of the TX, RX chains and RF switching, as well as the control loop circuit that allows the half-duplex operation of the front-end module, were described. This section will discuss the integration of all these parts to form the final front-end design and disclose some relevant details of the overall system.

Considering all the intermediate designs of the considered RF components, the design of the RF part of the final front-end module was also aided by the ADS software. Figures 4.23a and 4.23b show the considered ADS design for electromagnetic simulation and co-simulation with library components. This design mainly consists of a junction of all the previous designs of each component. At this point, adjustments in the considered passive components were made whenever it produced better results.

Once the ADS design was completed, the PCB layout was implemented considering the layout of Figure 4.24. For this board, the same four-layer board stackup was considered. At the same time, for the RF paths, it was requested the removal of the solder mask of the PCB to avoid influencing the RF results.

Apart from including the TX and RX paths, the RF switching, and the control loop circuit, this board also includes a power interface circuit that is relevant to distribute the appropriate operating voltage to every component in the board. From the SARIA Student Experiment Documentation (SED), it can be seen that this board will receive 12 V for its operation. [4] But in fact, most components of this module function at 5 V operating voltage. These aspects are both accommodated using a DC-DC converter - the TSR1-2450 by Traco Power². This component receives the

²<https://pt.mouser.com/datasheet/2/687/tsr1-537631.pdf>

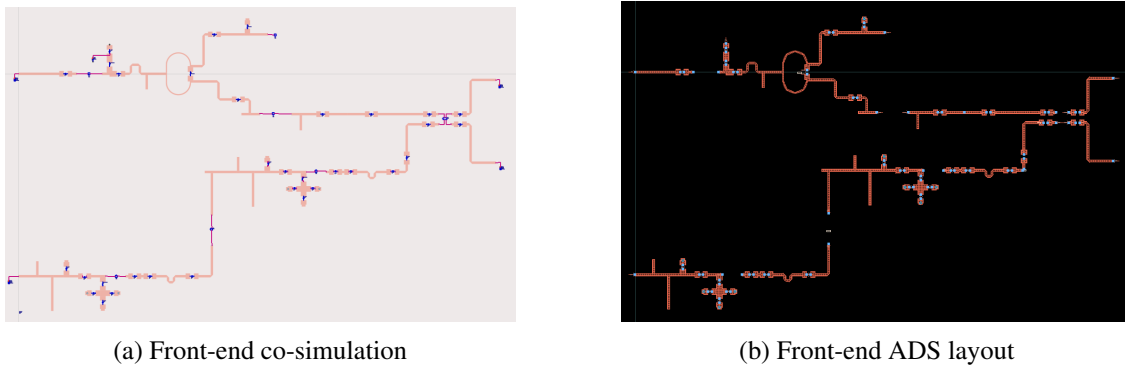


Figure 4.23: Front-end ADS design

incoming 12 V and converts it to 5 V with an efficiency of 92 %. At the output of this converter, the 5 V voltage is distributed to all the components on the board except for both LNAs.

In the case of the LNAs, a lower noise voltage source is recommended to avoid oscillations at the RF output port that can ruin the device expected performance. Using a low-dropout voltage regulator (LDO), it is possible to meet this specification. In this case the LDLN025 by ST³ was selected. The LDO regulator is based on linear voltage regulation and, when placed after the switching DC-DC converter like the TSR1-2450, has the advantage of reducing the noise at its output. These types of components are useful to mitigate high-frequency noise but have little effect on low-frequency noise. To prepare against this issue, a π -network was added to serve as a LPF at the output of the Traco Power converter.

With these considerations in mind, the final front-end board was designed and fabricated. Figure 4.25 shows the Front-end module after being properly manufactured and soldered.

³<https://pt.mouser.com/datasheet/2/389/dm00312779-1798765.pdf>

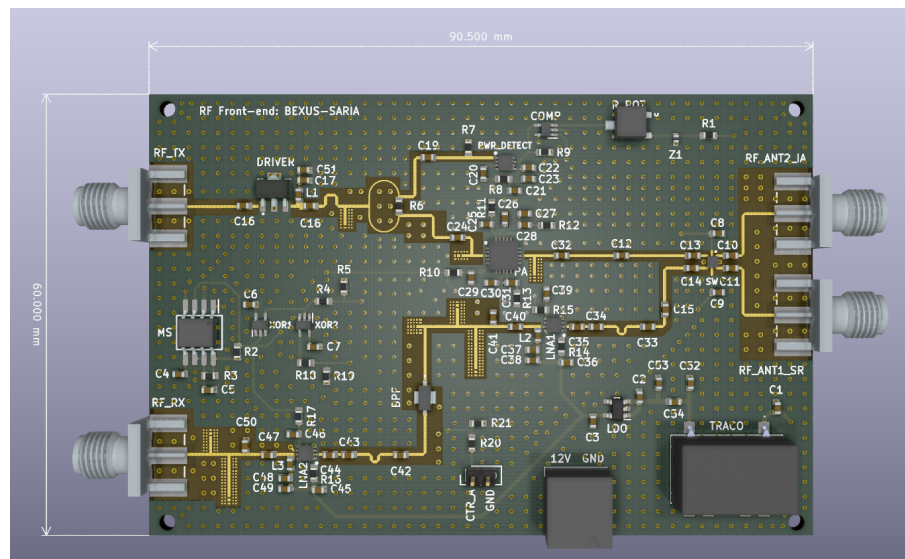


Figure 4.24: Front-end board expected board view

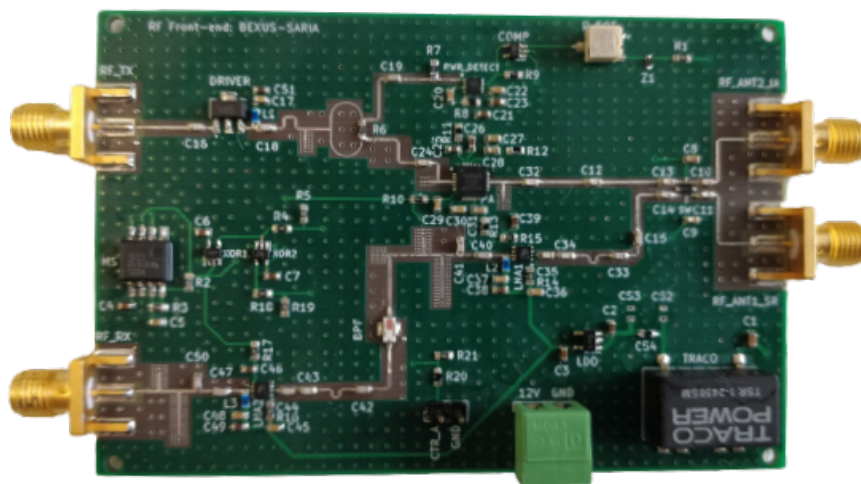


Figure 4.25: Front-end board real board view

Chapter 5

Testing and Validation

This next chapter will focus on the testing phase of both the intermediate and final boards designed and manufactured. As stated all through this document, the final front-end board is expected to properly receive and transmit C-band signals around the frequency of 5.8 GHz, while successfully commute between TX and RX modes in a half-duplex manner and also be able to switch the antenna of operation. With that in mind, this chapter will start by describing the testing setup used in all measurements. It will then move on to present the results that will eventually allow an evaluation if the board is working with the expected functionality and how well it is doing so in terms of RF and timing performances. In this way, the chapter will present and discuss the obtained results and conclude if they are acceptable considering the system specification established in Section 3.3.

5.1 Testing Setup

Regarding the setup used for testing, it was necessary to have adequate equipment to make S-parameter characterization of all intermediate boards as well as the final module. In the case of the final board, it was also required to use tools for the respective timing measurements.

Regarding the S-parameter characterization, the E8363B Vector Network Analyzer (VNA) by Agilent Technologies was used. For calibration purposes the Agilent 85052D calibration kit presented in Figure 5.1a was utilized. In order to connect the VNA ports to the ports of each one of the tested boards, RG142 SMA cables were used. Figure 5.2 shows the setup for the testing of the BPF as an example.

The S-parameter testing was made in the same way for all boards, except for the PA and the final front-end board, when testing the respective TX chain. In these cases, the output power was in the order of 30 dBm. The maximum power rating for the ports of the considered VNA was also



(a) Agilent 85052D calibration kit



(b) Midwest Microwave 10 dB attenuator

Figure 5.1: Supports tools for VNA testing

30 dBm so, as a precaution, a 10 dB attenuator was included at the output port of these boards. Figure 5.1b shows the incorporated attenuator.

In every performed test, the calibration considered the SMA cables connecting the VNA ports to the ports of the board under test. In this way, the calibration process placed the reference planes at the SMA connectors of the PCBs. No port extension was performed.

At this stage, the testing procedure consisted of measuring the S-parameters of a given device and comparing them with the ADS simulation results. Whenever the circuit was found to need some adjustment of the passive components, the ADS tool was utilized to understand how the S-parameters results would react to the variation of each passive element.

For the timing tests regarding the half-duplex operation of the final front-end module, the DSO3202A oscilloscope by Agilent Technologies was used to obtain the desired waveforms.

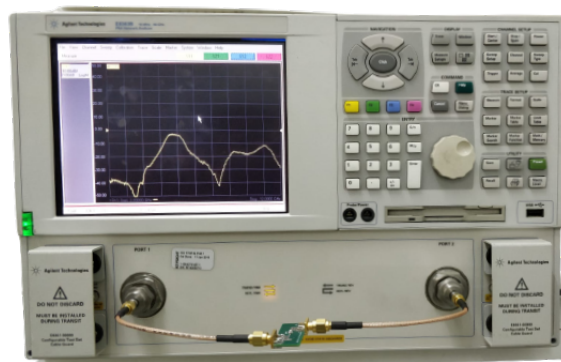
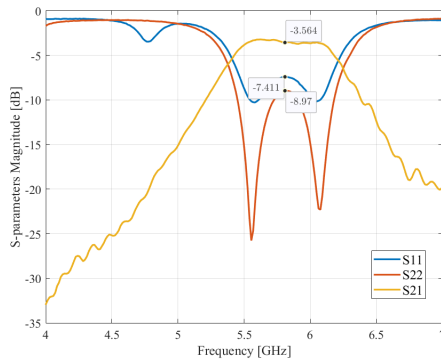
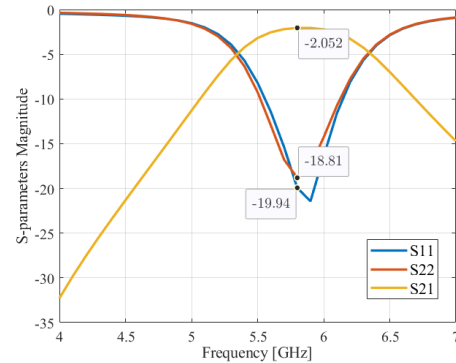


Figure 5.2: Example of the setup used for S-parameter characterization



(a) BPF results from the test board measurements



(b) BPF simulation results from the ADS software

Figure 5.3: BPF obtained results from test board measurements versus ADS simulation

5.2 Band Pass Filter (BPF)

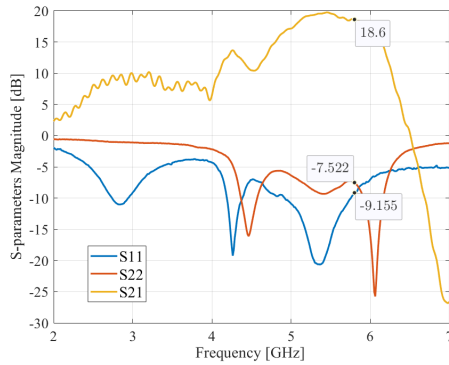
Starting with the testing of BPF, the design of this component described in section 4.3.2 was relatively straightforward. This device is a fully passive component and is expected to filter the signal coming after the first LNA around the frequency of 5.8 GHz without reducing too much the respective insertion loss. Figure 5.3a shows a full S-parameter characterization of this filter. From that picture, a value of around -3.3 dB was obtained for the frequency of 5.8 GHz. This value was expected to be around -2.1 dB from the ADS simulation present in Figure 5.3b. Since the component is in between two LNAs, this difference is not considered to be very significant. All in all, the BPF was rated as acceptable, keeping in mind its role in this front-end system.

5.3 Low Noise Amplifier (LNA)

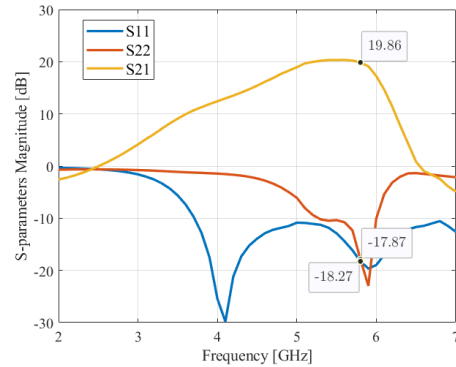
Following the testing of the BPF, the LNA was the next component to evaluate. From section 3.2, the RX chain of the front-end module is composed by 2 LNAs with a BPF in between them. The design for each LNA was the same, so a validation of the LNA board in conjunction with the already tested BPF, completes the testing phase of the RX chain.

Considering the design of the LNA described in section 4.3.1, a full S-parameter characterization was simulated using the ADS software and then tested using a VNA. Figures 5.4a and 5.4b show the obtained results.

Looking into the obtained values for the S-parameters of the LNA, it can be seen a decent value of S21 of 18.6 dB against an expected value of 19.8 dB from the ADS simulation. Unlike the S21 results, the return loss results are not exactly as expected. From Figure 5.4a, one can see -9.1 dB and -7.5 dB for the S11 and S22 parameters respectively.



(a) LNA results from the test board measurements



(b) LNA simulation results from the ADS software

Figure 5.4: LNA obtained results from test board measurements versus ADS simulation

Regarding the noise figure results of the LNA component, this data was simulated using ADS and the obtained result was 1.4 dB. This result was not tested in the respective intermediate board.

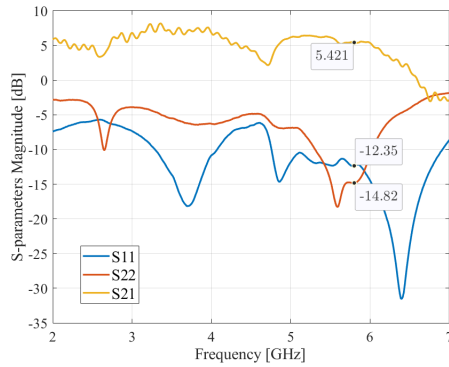
Even though these results reveal a noticeable vulnerability to the return loss performance of the LNA component, the overall functionality of this block is verified. In the end, this step was crucial to adjust the S-parameters of this block to the best performance possible considering the initial design. At the same time, it also exposed performance issues that can justify future issues in the RX chain of the front-end module.

5.4 Driver

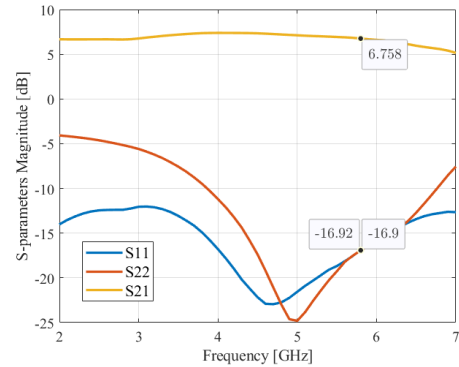
With the RX chain components completed tested, the validation of the key blocks of the TX chain was next. One of these components is the Driver block that interfaces the TX port of the front-end board with the Wilkinson Power Divider (WPD). Considering the role of this component to the overall system described in section 3.2, besides looking for an appropriate adaptation at the 5.8 GHz frequency, it was especially important to obtain interesting levels of input return loss (S11) with sufficient gain (S21) to overcome the inevitable attenuation that takes place with the power divider located after the Driver block.

Figures 5.5a and 5.5b compare the S-parameter results of the ADS simulation with the obtained parameters from the VNA. In this case, the S11 was at -12.4 dB and the S22 at -14.8 dB. These values were considered satisfactory for the purposes of this component. The measured S21 was 5.4 dB with an isolation (S12) of -22.8 dB.

The fact that this amplifier had an operating current of 0.09 A provoked noticeable heating of the device, but it was not considered a problem at all, to the point where no action was taken to mitigate it. In conclusion, after some adjustments to obtain the results shown in Figure 5.5a, the



(a) Driver results from the test board measurements



(b) Driver simulation results from the ADS software

Figure 5.5: Driver obtained results from test board measurements versus ADS simulation

component was evaluated as appropriate to fulfill its role.

5.5 Power Amplifier (PA)

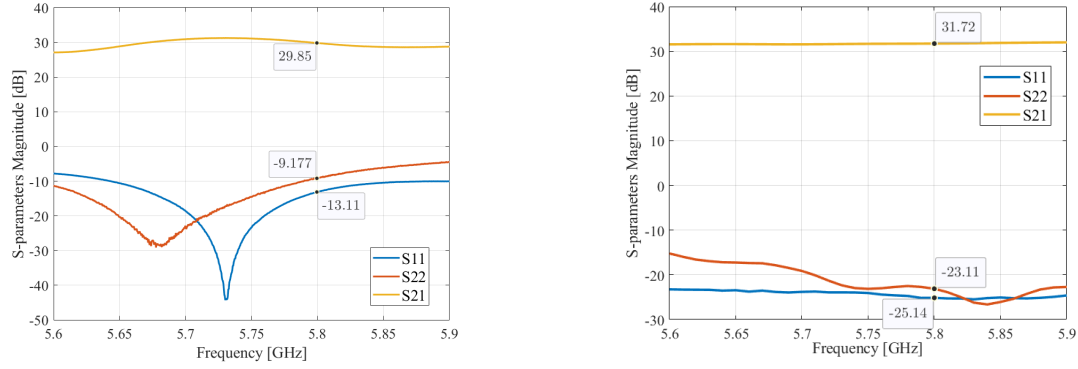
The other component to be tested in the TX chain is the PA. Unlike the Driver block, the PA was a much more sensitive component since it requires some caution in the testing with the VNA and it ultimately plays a fundamental role of the TX chain of delivering the expected output power to the following RF switch.

Regarding its power consumption, it was concluded that, when the PA is disabled, there is no power consumption from this component, and there is no RF amplification. On the other hand, if the PA is enabled, the current consumption is dependant on the Input Power applied to the amplifier starting at 0.3 A and going up to 0.8 A.

Since the VNA had a maximum power rating of 30 dBm, a 10 dB attenuator was used at the output port of the PA test board. The calibration of the VNA took this aspect into account to ensure reliable measurements.

Once again, the ADS software that was used to assist the RF design of this test board produced simulation results that are visible in Figure 5.6b. After some adjustments in the passive components of the RF paths of this board, the set of S-parameters results shown in Figure 5.6a were attained. From said Figures, it is noticeable an adequate level of input return loss (S11) of -13.0 dB with an acceptable output return loss (S22) of -9.0 dB.

As far as the S21 measurement goes, a value of 29.8 dB was achieved. Individually, this result is actually reasonable when compared to the expected 31.7 dB from the ADS simulation. However, this measurement was done with an input power of -15 dBm from the VNA. Considering the requirements of this system, present in section 3.3, and to more thoroughly characterise this PA, it



(a) PA results from the test board measurements

(b) PA simulation results from ADS software

Figure 5.6: PA obtained results from test board measurements versus ADS simulation

was important to get more S21 measurements at different input power levels. With that in mind, Figure 5.7 shows the evolution of the S21 parameter with Input Power (P_{in}) and Figure 5.8 plots the Output Power (P_{out}) versus the Input Power (P_{in}).

From these plots, it is possible to see that there is an expected decay in gain and Output Power as the Input Power increases, signaling that the PA is approaching saturation. However, according to the datasheet of this device (available in Table 3.2), a value of 30-32dBm was expected for the corresponding P1dB. Yet, these plots estimate an output P1dB value of only around 24 dBm for the considered PA and the respective board design. This difference seems to expose the RF adaption of the test board that initially did not appear to cause much difference between the simulated and practical results. As it turns out, as the Input Power increases, this difference becomes more evident and reveals a performance issue with this PA design. Such difference is significant enough to condition the maximum output power possible to be transmitted by the overall front-end module. The conclusion at this stage was that the PA indeed appears to be functionally correct in its role of amplifying C-band signals around the frequency of 5.8 GHz, but it has performance issues in achieving a transmission power of up to 30 dBm that do not go unnoted.

5.6 RF Switch

Up to this point, the testing of the RX and TX chains were described, leaving the RF switching for last before beginning the testing of the final front-end module. For this component, it was relevant to develop a test board since this block interfaces the TX and RX paths with both the inflatable antenna and the static reflector. With that in mind, it is crucial to ensure proper RF performance in terms of verifying low insertion loss when there is supposed to be a low-loss connection and high isolation when there is not supposed to be a connection between two of the four ports of the RF switch.

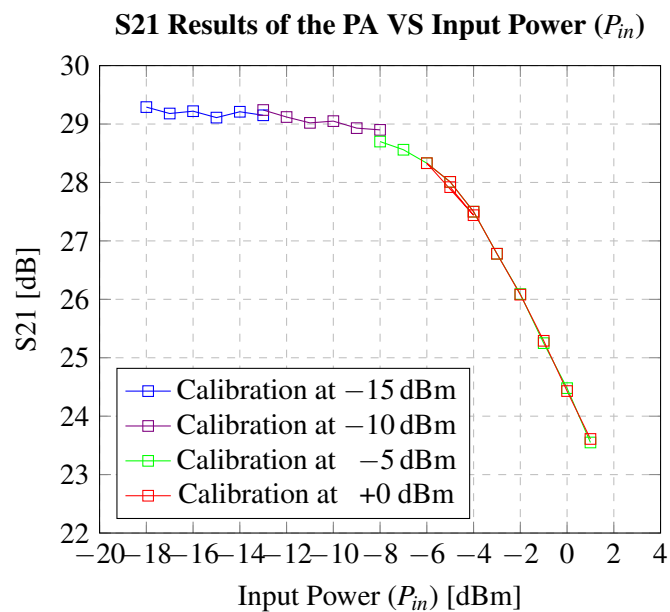


Figure 5.7: S21 results of the PA from the VNA measurements VS input power

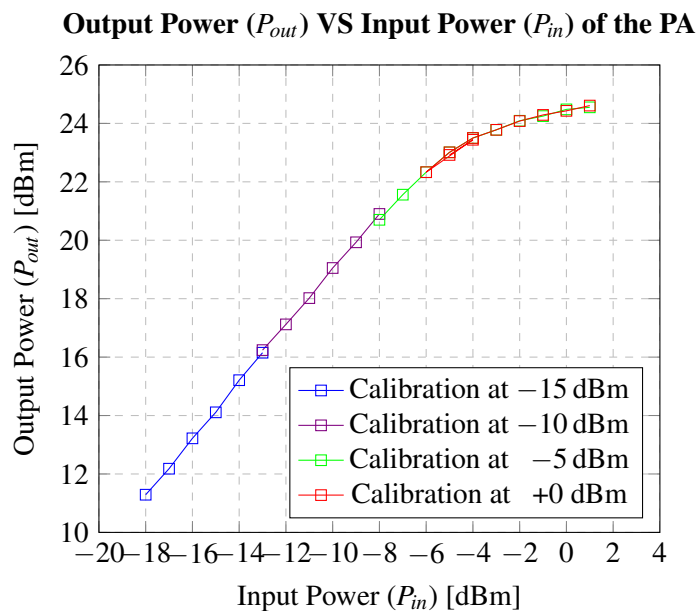


Figure 5.8: Calculated output power of the PA from the VNA measurements VS input power

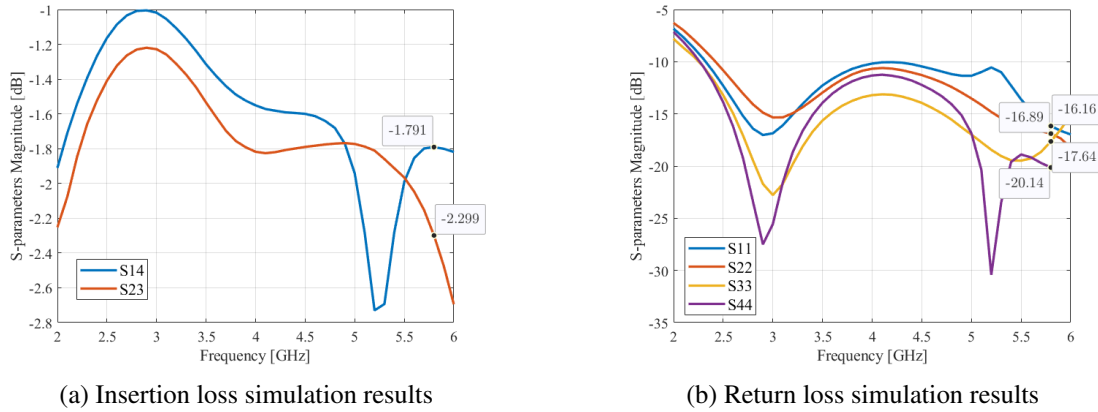


Figure 5.9: ADS simulation results of the RF Switch

That said, the upcoming sub-sections will specify the four different possible connections for each control configuration. Recalling that this switch has two control pins, in the results presented, it was considered the configuration where VC1 is at a low voltage state (kept at 0 V), and VC2 is at a high voltage state (this pin was controlled at 3 V). In similarity to the previous test boards, the RF design was aided by the ADS software. Figures 5.9a and 5.9b show the simulation results from this tool for the RF switch design.

RX-ANT1

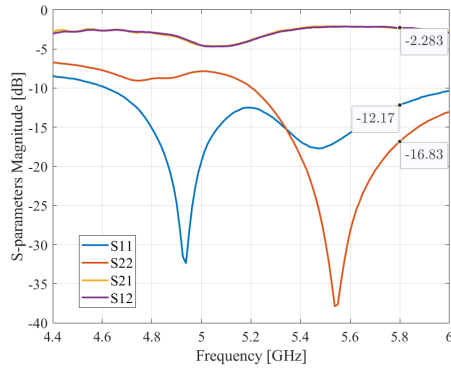
With the control setup considered for all these tests, it was expected a low insertion loss between the RX to ANT1 port of the switch, with appreciable levels of return loss. Figure 5.10a show the results for this scenario. The results for return loss at input and output are adequate for this application at -12.2 dB for S11 and -16.8 dB for S22. The insertion loss is expected to be at a low level, and indeed, there is an obtained value of -2.3 dB for the S21 measurement. This value is a bit higher than the expectation of -1.8 dB but was considered acceptable for this application.

RX-ANT2

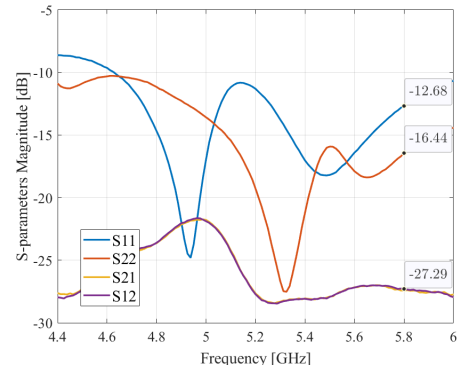
Using the same setup as before, this time the measurement was between the RX and the ANT2 ports. Because the antenna port changed, it would be expected to see relatively high isolation between these two ports at the S21 measurement. Figure 5.10b shows the results for this case, and the S21 is at an expectedly low value of -27.4 dB indicating that if one of the antennas is connected (in this case ANT1 is connected) to the RX port, there is isolation between RX and the other antenna port. On the other hand, Figure 5.10b also shows that the return loss results are also at a relatively low level around the frequency of 5.8 GHz.

TX-ANT1

Following the logic of the previously presented, the connection ANT1-RX should be verified. That means that the connection between the TX port and the same antenna port must have a relatively



(a) RX-ANT1 connection of the Switch



(b) RX-ANT2 connection of the Switch

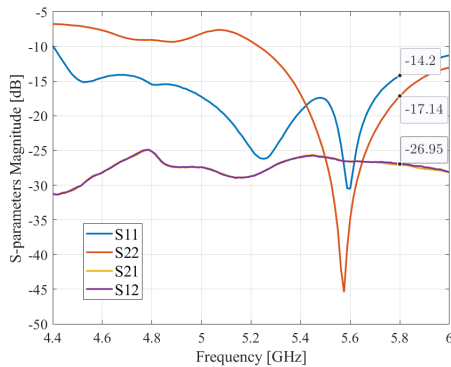
Figure 5.10: S-parameter characterization of the connection of the RX port with both antennas

low value at the S21 measurement. Figure 5.11a, shows exactly this. From that Figure, appropriate values of return loss were also verified.

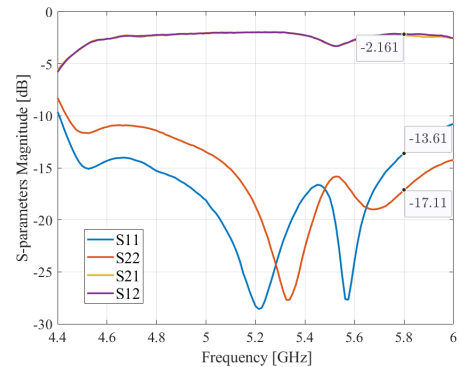
TX-ANT2

At last, in the connection TX-ANT2, following the same logic as before, one should expect to see a low insertion loss. From the ADS simulation results, it would be expected a value of around -2.3 dB. Figure 5.11b shows an insertion loss of around -2.3 dB as well. Once more, the return loss results were also satisfactory.

Overall, the results of the RF switch were considered adequate for this application, and the device was validated. In these conditions, the functionality of all the components from the RX, TX chains and the RF switching was successfully verified. The adjustments made at this intermediate testing step were taken into consideration in the final board integration and eventual testing.



(a) TX-ANT1 connection of the Switch



(b) TX-ANT2 connection of the Switch

Figure 5.11: S-parameter characterization of the connection of the TX port with both antennas

5.7 Final Front-end module

After assembling the final front-end board with the tuned components from the previous validation of each intermediate board, the final testing phase followed. That said, with the design presented in section 4.7, it was fundamental to verify not only proper RF behavior around the frequency of 5.8 GHz, but also correct switching between TX and RX and the two antennas of operation. In a first stage, it was important to test if the board was working as expected. After that, it was necessary to look into the respective performance.

Regarding the power interface of the module, a nominal consumption was verified when powering the board with 12 V. An important aspect to notice at this point was to look for possible noise in the supply of the main RF components. In particular, it was crucial to find out if there was any meaningful variation in the supply of the LNAs. If so, the LPF block (for which there was space reserved in the board with a π -network) should be implemented. Since this was not the case, the decision was to not include this filter.

The next step was to validate the half-duplex operation that allows for a switching of contexts from the default RX mode to TX mode. The approach to test this issue was to use a SDR to send the SARIA transmitting signal described in section 3.1.1 and measure the waveform coming out the Monostable (MS) that essentially triggers the PA, the LNAs and the switch for the TX state of either antenna. Figure 5.12 shows the results of this measurement. The Figure highlights the signal produced by the Power Detector block in yellow and the signal produced by the MS. In that Figure, it is noticeable that the dummy signal part of the transmitted signal triggers the MS to produce a square wave of around 115 μ s. This interval of transmission is well under the specification determined requirement P.2-1. That said, this aspect was considered verified.

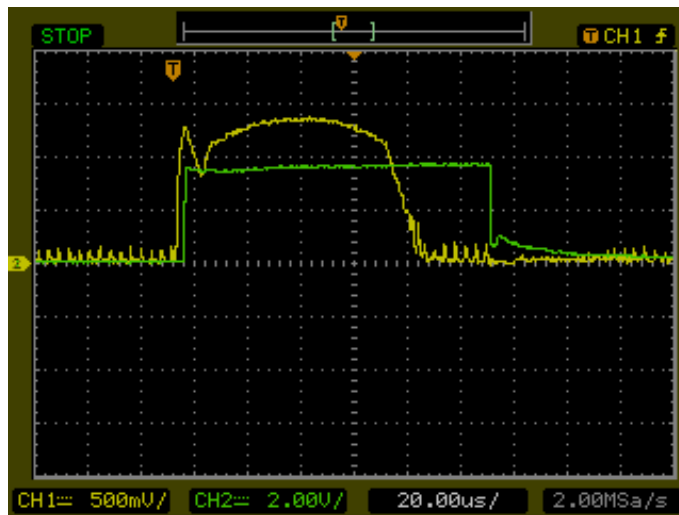
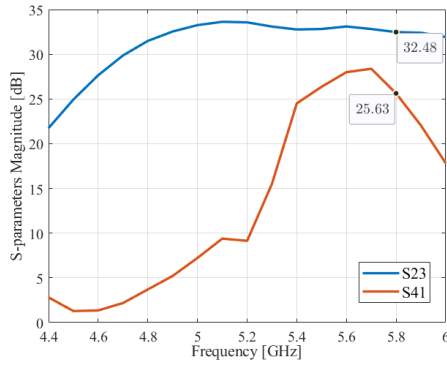
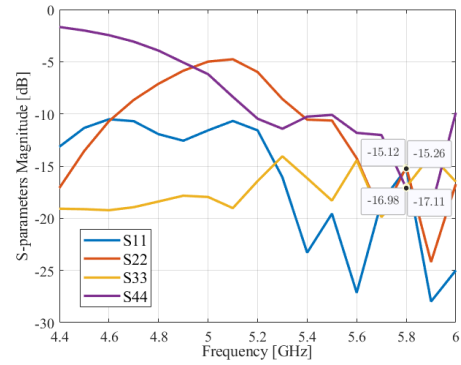


Figure 5.12: Half-duplex operation of the front-end module

The final stage of testing of this module had to do with the RF performance around the frequency



(a) Simulation results of the gain of the TX (S23) and RX (S41) chains



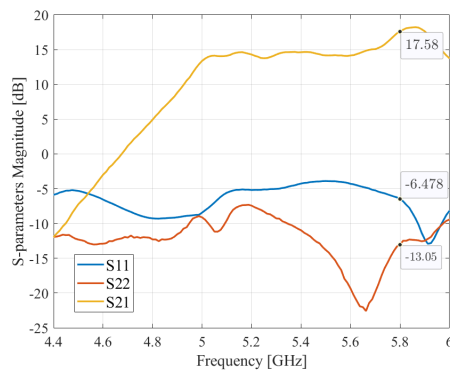
(b) Simulation results of the return loss of the final front-end board

Figure 5.13: ADS simulation results of the final front-end board

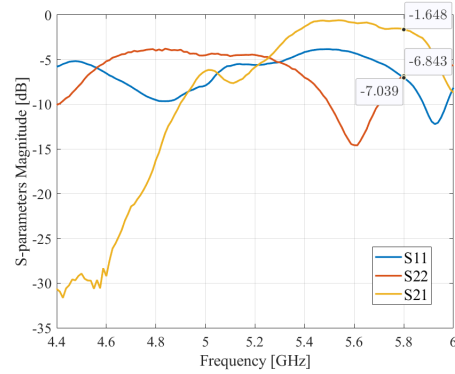
of 5.8 GHz. At this point, both the antenna switching and the quality of transmission and reception of C-band signals were tested, using the ADS simulation results present in Figures 5.13a and 5.13b as a baseline. Figure 5.14 show the obtained results for the ANT1-RX connection with both the CTR_A input connected to 5 V (ON) and to GND (OFF). Figure 5.15 does the same thing for the ANT2-RX connection. Looking into the transmission chain, Figures 5.16 and 5.17 show the S-parameter characterization again with CTR_A ON and OFF.

From the previous results, it can be noticed proper switching between antennas since the attenuation/gain measured at the S21 parameter matches the expected connections to be established by the RF switch in relation to the CTR_A input. That said, the switching between antennas was successfully verified based on the results of the S21 parameter with different combinations of input and output ports and the actuation or not of the CTR_A input.

On the other hand, the RF performance did present some problems in terms of the expected performance. Specifically, the S21 results of the connections of both antennas with the RX chain fell a bit short of the expected results. The consistency between different combinations of inputs and outputs with the CTR_A input suggests a problem with the adaption of the LNAs present in the RX path, which was expected to be very sensitive from the ADS simulations. In the end, the performance of both the Transmission and Reception chain was not exactly as anticipated. Yet, the initial functionality of transmitting and receiving C-band signals can still be verified since there is a noticeable adaptation of the front-end module towards this frequency band, even if not with the initially proposed performance.

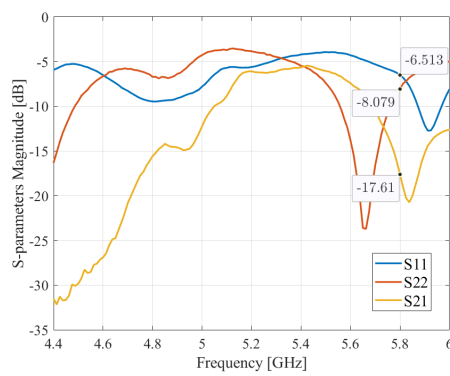


(a) With CTR_A ON

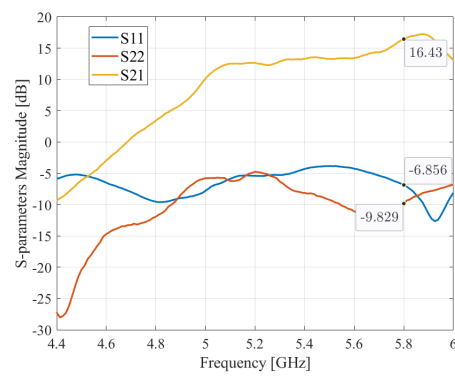


(b) With CTR_A OFF

Figure 5.14: ANT1-RX connection S-parameters results

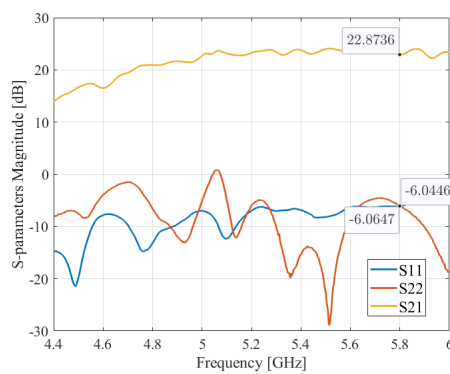


(a) With CTR_A ON

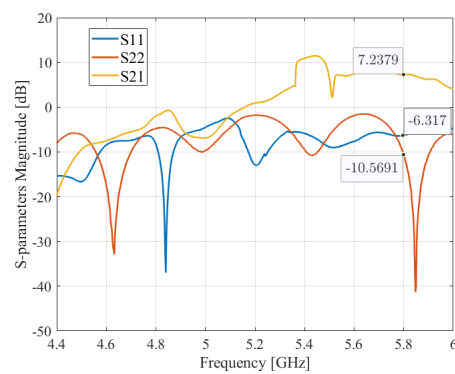


(b) With CTR_A OFF

Figure 5.15: ANT2-RX connection S-parameters results

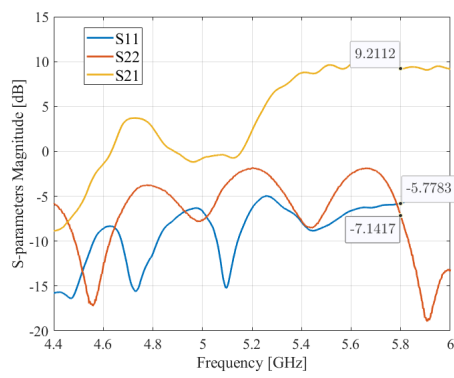


(a) With CTR_A ON

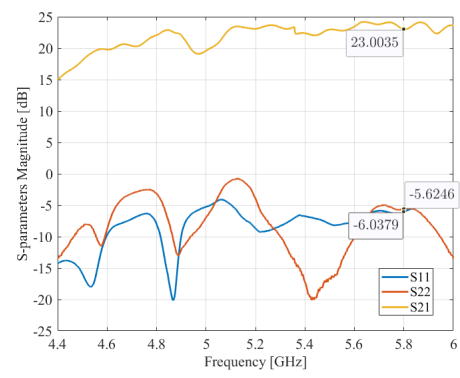


(b) With CTR_A OFF

Figure 5.16: ANT1-TX connection S-parameters results



(a) With CTR_A ON



(b) With CTR_A OFF

Figure 5.17: ANT2-TX connection S-parameters results

Chapter 6

Final Remarks

This chapter concludes the description of this dissertation work. Up to this point, the design, manufacturing, assembling, and testing of the front-end module were described. This part of the document will focus on some of the difficulties encountered in the development of the module that, to some extent, condition the execution time and obtained performance of the final board. It will then detail some of the conclusions and lessons learned realized after the development of this project. The chapter finishes with a reference for future work to be done related to this thesis.

6.1 Difficulties Encountered

The development of this thesis work did not happen without some setbacks worth to be mentioned at this point. The most notable complication had to do with delays in the arrival of material for the assembly of the intermediate and final boards. Understandably, this lag in the distribution of the selected components was motivated by the Covid-19 pandemic that affected the supply of electronic components worldwide. Even though some delay was expected, the actual waiting time ended up being longer than anticipated. The solution to mitigate this issue was to look for alternative suppliers that could deliver products quicker and reuse components left over from previous experiments. In the end, even if taking a bit longer, all the material arrived, and the final board was built with the correspondent material.

Another relevant issue encountered was the difficulty in obtaining a good similarity between the ADS simulation results and the measurements made with the VNA at the laboratory. Ideally, the ADS results would be very approximate to the VNA ones, indicating that the resulting boards were as close as possible to the initial designs. That was not the case. This disparity was assumed to be related to aspects such as the fact that the SMA connectors, used in the intermediate and final boards, were not simulated at the ADS level, and the manufacturing process of every board by JLCPCB included a surface finish called Lead-free Hot Air Solder Leveling (HASL). This type

of finishing is commonly used as a cheap finish for most general-purpose projects, but it leaves uneven surfaces that can be a problem with more sensitive designs. [19] This surface finish was selected to reduce costs in manufacturing, and there was a difficulty in simulating this layer that is applied on top of the copper RF traces, that possibly affected the correspondence between the ADS simulations and the VNA measurements.

6.2 Conclusions

From the exercise of designing and implementing a front-end module, there were some important conclusions and lessons learned gathered. The overall front-end system is a relatively complex one, for which there are a set of subsystems, that need to be carefully thought so that the final integration produces adequate results. That made the design phase represent around half the time devoted to the development of this thesis work. It was also important to dedicate some effort towards the RF simulations of the various subsystems and to the eventual final system, to get a baseline for how the design would look like to accommodate the system specifications.

Another aspect that became very relevant in the overall development of this module was the decision to develop intermediate boards for the more sensitive RF components. This suggestion came from the thesis Co-Supervisor, and it ended up becoming immensely helpful to better characterize each device before actually including them all together in the final board. This approach allowed to better isolate and solve the issues with each component and ultimately facilitate the debugging process of the resulting system.

Looking into the obtained intermediate and final results, and as stated over the course of chapter 5, the Driver, the Switch, and the BPF presented an appropriate and expected functionality. However, the same cannot be said for the PA and the LNA components. Their performance fell a bit short of the expected one, and eventually, the final front-end board accused similar issues. Section 6.1 highlights the difficulties encountered at the simulation stage of the connectors and the lower quality surface finish used in the manufacturing of each board. Besides these, other aspects such as soldering defects (considering that the soldering process was done at the faculty laboratory by the author) and the tolerance of passive components used can perhaps justify the disparity observed between the ADS simulations and the VNA results.

In the end, it is factual that, even though the performance of the final front-end module did not meet every standard set in the Performance Requirements of section 3.3, all the Functional capabilities initially planned were accomplished. The implemented board is expected to fulfill its role within the context of the SARIA experiment albeit with some performance imperfections.

6.3 Future Work

Chapter 5 displayed the S-parameter characterization for both the TX and RX chains of the final front-end module. From those results, it was clear that the respective RF performance was not particularly great in terms of return loss and gain of both paths. It would be interesting to have a different fabrication method and to adjust the ADS simulations to consider the implemented SMA connectors. This would allow to get a sense of the quality of the executed design. Furthermore, now that the design considered in this thesis work is functionally validated, it could be interesting to look for a RF design with the same system architecture and components but with a better consideration for the area since that was not the focus in the presented design.

Another aspect that is very relevant for the development of a RF front-end module, is the measurement of the noise figure value of the RX chain. This is something that is further detailed in section 2.3.3 but was not included in the scope of the integration, manufacturing, and testing phase of this document.

Appendix A

Detailed Information

A.1 RADAR Equation

Figure A.1 shows a basic representation of a parabolic antenna attached to a balloon, transmitting RF signals to the overflowed area and receiving the respective echoes from an arbitrary target. This figure will be used to aid the explanation of the *radar equation*.

From Figure A.1, one can identify the blue dotted circles representing the transmitted signals from the parabolic antenna and the red dotted circles representing the refracted energy from the target that will be received again by the transmitting antenna. In these kinds of setups, the same antenna is used for transmission and reception, and half-duplex operation is considered. The process starts with the parabolic antenna transmitting RF/microwave signals with a power density, S_t , that is dependent on the transmitted power, P_t , the antenna gain G_t , and the distance to the target, R , in the following way:

$$S_t = \frac{P_t \cdot G_t}{4\pi \cdot R^2} [\text{W/m}^2] \quad (\text{A.1})$$

Then, this power will reach the target, and some of it will be scattered back in a way that it can be received by the reflector. The amount of power scattered, P_s , depends on the Radar Cross Section (RCS), σ , of the target. This value is essentially dependant on the material, geometry, and size of the target and represents the power scattered in relation to the received power density:

$$\sigma = \frac{P_s}{S_t} [\text{m}^2] \quad (\text{A.2})$$

At this point, some of the transmitted power gets scattered and it reaches the parabolic antenna after traveling the same initial distance R (distance from antenna to target is the same as the one

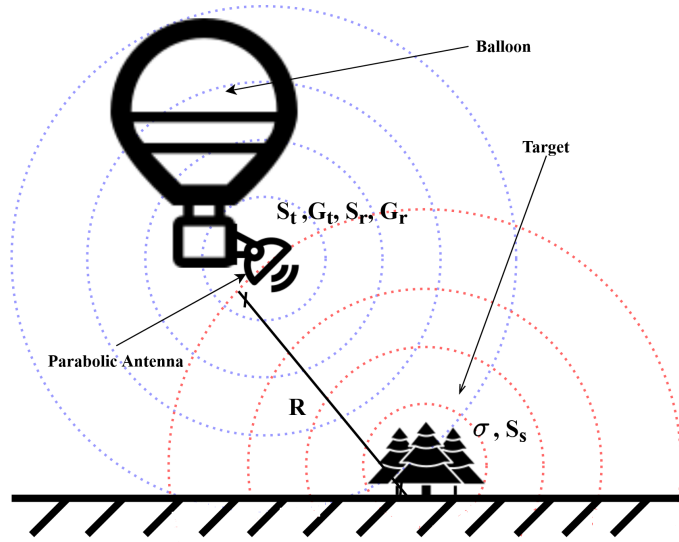


Figure A.1: Representation of a basic RADAR system

from target to antenna). Thus, the received power density, S_r , can be given as:

$$S_r = \frac{P_s}{4\pi \cdot R^2} = \frac{P_t \cdot G \cdot \sigma}{(4\pi \cdot R^2)^2} [\text{W/m}^2] \quad (\text{A.3})$$

To obtain the power received, P_r , one must now account for the effective area of the receiving antenna, A_{eff} . This value depends on the gain of the antenna G_r (since the transmitting antenna is the same as the receiving one, $G_r = G_t = G$) and the wavelength of the signal, λ .

$$A_{eff} = \frac{G \cdot \lambda^2}{4\pi} [\text{m}^2] \quad (\text{A.4})$$

When combining all these factors, the radar equation is finally unveiled as:

$$P_r = \frac{P_t \cdot G^2 \cdot \lambda^2 \cdot \sigma}{(4\pi)^3 \cdot R^4} [\text{W}] \quad (\text{A.5})$$

A.2 SAR Slant and Azimuth Resolutions

The slant range resolution, δ_R , corresponds to the minimum distance at which it is possible to distinguish two separated objects. If two targets are separated by a distance smaller than δ_R then, from the point of view of the SAR system, they will be indistinguishable. This value is dependant

on the speed of light, c , and bandwidth, β in the following way:

$$\delta_R = \frac{c}{2 \cdot \beta} \quad (\text{A.6})$$

Notice the division by 2 in equation A.6, indicating that an electromagnetic signal has to travel to the target and come back. Also from this equation, one can notice that it is possible to improve the slant range resolution by increasing the bandwidth.

In the slant range direction, SAR is not very different from Real Aperture Radar (RAR). It is in the orthogonal direction to the radar beam (also called the cross-range, azimuth or along-track range), that SAR is distinctive in its use of synthetic aperture to improve spatial resolution. In conventional RAR, the along-track resolution, δ_{AT} , is determined by the physical dimensions of the antenna aperture. This value is directly proportional to the RADAR wavelength, λ , and the target range, R , and it is inversely proportional to the antenna dimension, D : [5]

$$\delta_{AT} = \frac{\lambda \cdot R}{D} \quad (\text{A.7})$$

Equation A.7 highlights the problem with using Real Aperture Radar (RAR) technology for C-band imaging, previously mentioned in section 1.1. With RAR, the problem of poor slant range resolution can be solved (by increasing the bandwidth), but the problem with along-track resolution cannot. This is because, to obtain a good along-track resolution at C-band frequencies, the antenna dimension (D) would have to be in the order of kilometers or even tens of kilometers! To reduce δ_{AT} , the remaining options would be to increase frequency (moving out of C-band) or decreasing distance to the target. None of these options are valid if the goal is to perform RAR in the stratosphere. This means that it is not viable to use conventional RAR for this purpose. [5]

However, thanks to Carl Wiley's discovery, it is possible to achieve good along-track resolution by using SAR technology. In this case, the along-track resolution is determined by the Doppler bandwidth of the received signal instead of using the antenna real aperture in that direction. [5]

Figure A.2 refers to 3 stages that are relevant to describe the behavior of the Doppler frequency shift, as the balloon is passing by a fixed target. Notice that, for simplicity, it was represented a fixed balloon with a moving target, when in fact, in the SAR system to be developed here, the target is fixed and the balloon is moving. This is not a problem as it is just a matter of selecting one of the two as a reference for velocity. With that said, Figure A.3 plots the variation of Doppler shift frequency, f_d over time to obtain the so-called Doppler bandwidth, β_d . As the figure suggests β_d , can be defined as the difference between the Doppler frequency shift at stage 1 (initial stage), f_{d1} and the Doppler frequency at stage 3 (final stage), f_{d3} : [20]

$$\beta_d = f_{d1} - f_{d3} = \frac{2 \cdot v \cdot \theta}{\lambda} \quad (\text{A.8})$$

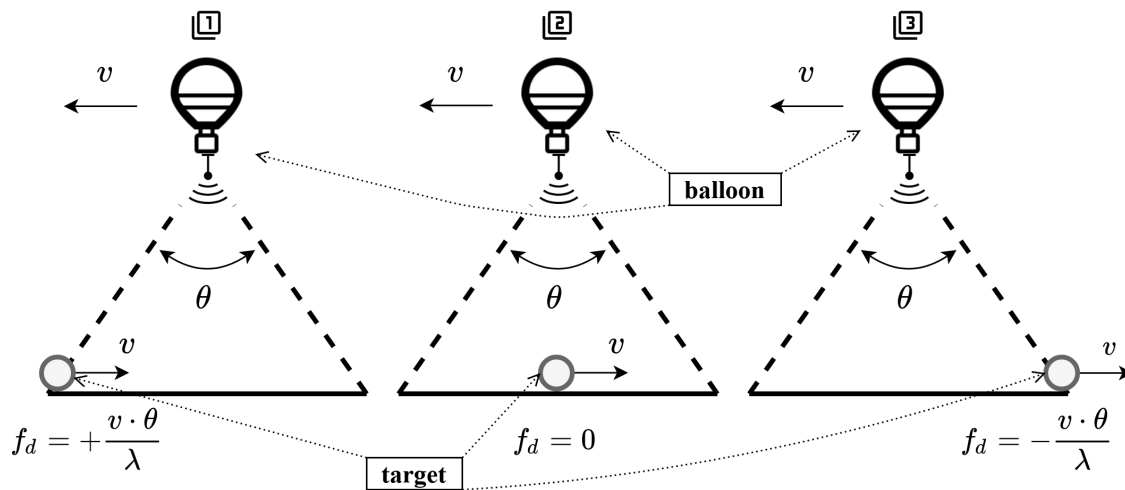


Figure A.2: Azimuth movement of a target relative to a balloon

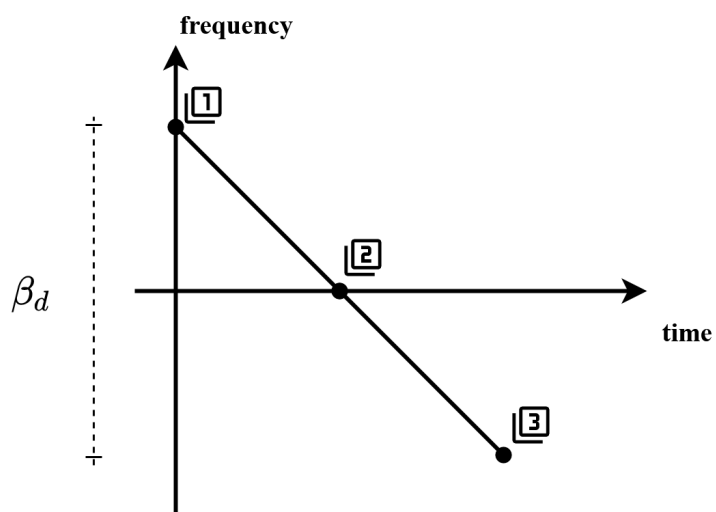


Figure A.3: Frequency vs time in a Chirp signal

In equation A.8, v represents the velocity of the balloon in the along-track direction, θ represents the antenna beam angle, and λ the wavelength of transmission. [20] The beam angle, θ can be related to the wavelength, λ , and the antenna length, D in the following way:

$$\theta = \frac{\lambda}{D} \quad (\text{A.9})$$

Furthermore, the Doppler bandwidth, β_d , can be converted to time, Δt to obtain the time difference between two distinct target points at the considered velocity of the balloon:

$$\Delta t = \frac{1}{\beta_d} = \frac{\lambda}{2 \cdot v \cdot \theta} \quad (\text{A.10})$$

Combining the equations A.8, A.9 and A.10, one can finally find a relation for the along-track resolution, δ_{AT} :

$$\delta_{AT} = v \cdot \Delta t = v \cdot \frac{\lambda}{2 \cdot v \cdot \frac{\lambda}{D}} = \frac{D}{2} \quad (\text{A.11})$$

Equation A.11 shows the ingenious idea that makes SAR technology viable! The fact is that, by using a moving RADAR system to obtain a synthetic aperture of a relatively small and feasible antenna, it is possible to obtain useful along-track resolutions. Even better, the resolution possible to achieve in the along-track direction depends only on the antenna length. [20]

In Figure A.4, the same results for resolutions are obtained with a different (but also valid) way of thinking. Here, the value for the antenna length, D_{AT} is directly replaced by $2L_{SA}$, to obtain the desired along-track resolution, δ_{AT} . Additionally, a relation for the Dwell or Integration Time is presented. The resolutions for slant range and along-track dimensions are summarized in table 2.1 for Synthetic Aperture RADAR and Real Aperture Radar.

A.3 SARIA Link Budget

The Link Budget from Table 3.1 starts by defining a TX power of 20 dBm. This value was carefully selected to be in agreement with the limitations of the ISM band of operation (5.725 GHz - 5.875 GHz). The TX antenna gain corresponds to the gain of a parabolic dish with diameter of 1 m. This antenna gain needs be accounted for at TX and at RX. For the considered frequency and considering an antenna efficiency of 50%, this value can be given by equation A.12.

$$G = \left(\frac{\pi D}{\lambda} \right)^2 e_A \approx 30 \text{ dB} \quad (\text{A.12})$$

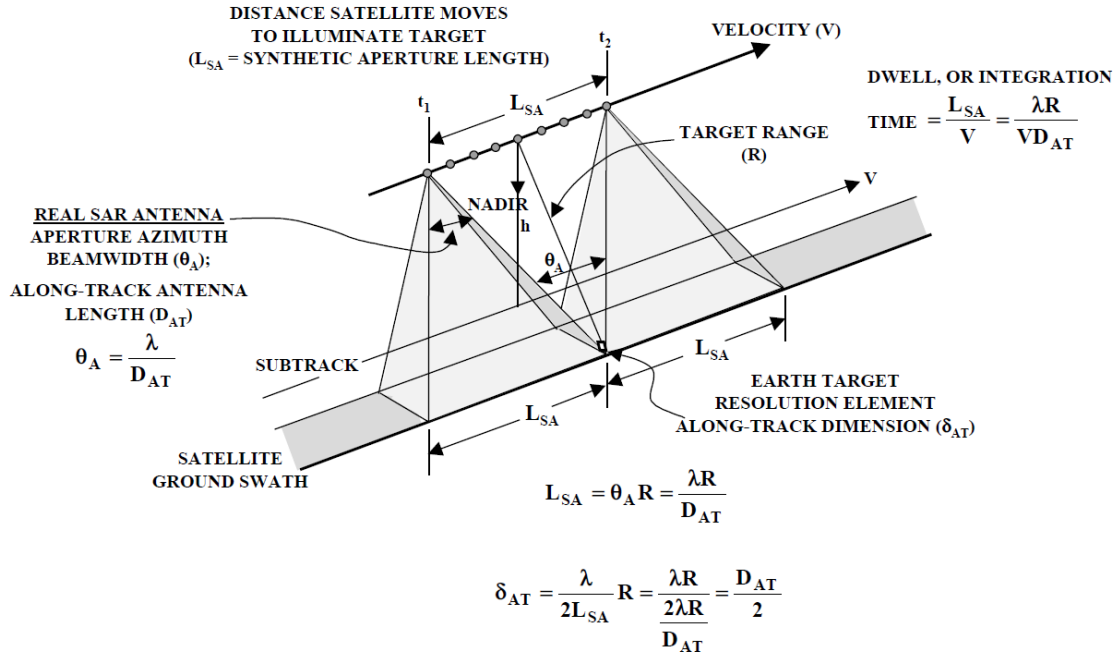


Figure A.4: Basic principles of SAR. Credit: SAR Marine User's Manual [5]

Regarding the Free Space Losses (FSPL), this value merely introduces the attenuation the RF signals suffer when traveling to the target. It is dependant on the signal wavelength, λ , and the distance to the target, R . For a frequency of 5.8 GHz and a distance to the target of 45 km. Naturally, the value obtained from A.13 needs to be considered twice for forward and backward motions of the RF signals and it has a negative impact on the link budget.

$$FSPL = \left(\frac{4 \pi R}{\lambda} \right)^2 = 141 \text{ dB} \quad (\text{A.13})$$

The Link Budget considered an effective signal bandwidth of 7.5 MHz and an antenna length of 1 m. Recovering the results for slant-range and along-track resolutions, summarized in table 2.1, it is possible to obtain, for these conditions, a slant-range resolution of 20 m and a along-track resolution of 0.5 m. This leads to an effective area of the resolution cell, A_{eff} , of 10 m^2 . From equation A.14, it is possible to obtain the value for the Projected Radar Cross Section.

$$\sigma_P = \frac{A_{eff} 4\pi}{\lambda^2} = 47 \text{ dB} \quad (\text{A.14})$$

At the same time, it is important to consider minimum target reflectivity. This value indicates what is the minimum gain associated with the target Projected Radar Cross Section (RCS) in order for it to be distinguishable from noise. This value was estimated as 25 dB. Combining this value with the initial Projected RCS, a value of 22 dB is achieved for the effective RCS.

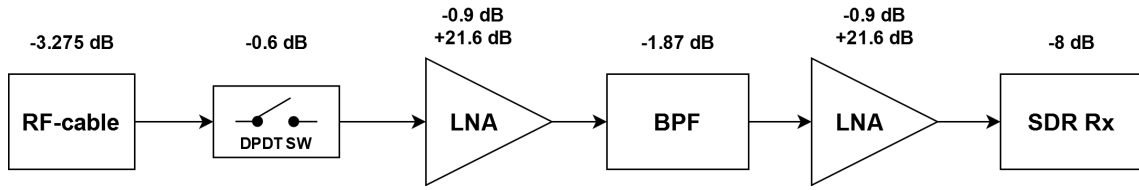


Figure A.5: Receiver system losses

The next value to analyze is the system losses. This factor relates directly to the RF front-end design since it amounts to the losses in the reception chain of the front-end module that affect the overall Link Margin. In the current implementation of the SARIA RF front-end, this value is affected by the RF cable (from the inflatable antenna to the front-end), switches, LNAs, BPF and the SDR receiving noise figure. Figure A.5 represents the receiver front-end architecture and the respective losses at each component. Following the Friis formula for noise factor, indicated in equation 2.5, a value of 6 dB was obtained.

In the eventual final design of the RF front-end module, it will be crucial to make sure the system losses associated with the receiving chain are low enough to comply with a positive Link Margin.

At this point, a SNR of -80 dB is obtained. Without any more adjustments, the experiment would not be viable but, when working with SAR technology, there are still signal processing techniques that are used to obtain RF feasibility. With that in mind, a compressing of the signal is made at the slant-range direction, to add a gain given by the product of the signal length and its bandwidth. Considering a signal length of $70\mu\text{s}$ and an effective bandwidth of 7.5 MHz , a gain of 27 dB is achieved. Following the slant-range compression, a similar operation is performed in the along-track direction. This time, the gain obtained depends on the Pulse Repetition Frequency (PRF) and the time interval at which a target is illuminated. Considering a velocity of 200 km (55 m s^{-1}), the distance between 2 points will be 0.055 m (with $\text{PRF} = 1\text{ kHz}$). In turn, the synthetic aperture length, L_{SA} , for $R = 45\text{ km}$, $\lambda = 0.05\text{ m}$ and $D = 1\text{ m}$ will be 2250 m . From the relations presented in Figure A.4, with this velocity and synthetic aperture length, the integration time will be around 41 s . Finally, the correspondent azimuth compression gain will be 46 dB .

As a final signal processing technique, a LPF is passed to the along-track image, in order to worsen the target resolution from $0.5 \times 20\text{m}^2$ to $20 \times 20\text{m}^2$. Even though it deteriorates the along-track resolution, it provides an additional SNR gain of 16 dB (factor of $40\times$).

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