

École polytechnique de Louvain

CubeSat GNSS-R receiver design: Filtering, Synchronisation and Antenna design

GLITTER Project

Author: **Loïc TUMELAIRE**

Supervisor: **Christophe CRAEYE**

Readers: **Claude OESTGES, Luc VANDENDORPE, Laurent PAUCOT**

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Abstract

This thesis explores CubeSat satellites as an innovative solution to achieve an antenna array for satellite-based Earth observation, using GNSS reflectometry technology.

The study identifies a suitable circuit for amplifying and filtering the incoming Global Navigation Satellite System (GNSS) signals to the CubeSat using PCB technology. Three validation steps were performed: simulation linking equations to theoretical behaviour, fabrication with respect to PCB standards and testing in multiple phases.

Testing was first carried out using a home-made Global Navigation Satellite System (GNSS) signal simulated with only one incoming signal, to ensure the operation of the whole chain. Capturing a real direct signal was the next step with one CubeSat, and the following stage involved receiving the reflected signal in addition to the direct one. Then, the last part was to challenge the model with the synchronization between two CubeSats.

These synchronization challenges are caused by the inability to transmit the local oscillator through a cable and the analysed solution is to use the incoming Global Navigation Satellite System (GNSS) signal with u-blox module as a frequency reference.

In a parallel part, patch antenna design for the direct and reflected signal was investigated using mathematical design, simulation with CST and ADS software and measurement.

Keywords: Satellite-based Earth observation, CubeSats, GNSS reflectometry, Multi-feedback filter, PCB, Wireless synchronization, Antenna design.

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Acronyms

AC Alternating Current. 73

ADC Analog-to-digital Converter. xii, 12, 13, 15, 17–20

ADEV Allan Deviation. 42–44, 46–50, 65

AltBOC Alternative Binary Offset Carrier. 5

API Application Programming Interface. 42

AR Axial Ratio. x, 59, 60

AVAR Allan Variance. 42, 43

CBOC Composite Binary Offset Carrier. 5

CDMA Code Division Multiple Access. 5

DC Direct Current. 14, 17–21, 45

DUT Device Under Test. 44

EC European Commission. 2, 6

EMI Electromagnetic Interference. 30, 36

EUSPA European Union Agency for the Space Programme. 3

FDMA Frequency Division Multiple Access. 5

GLITTER Gnss-r sateLLITe earTh obsERvation. iii, 2, 6, 11, 12, 65, 66

GNSS Global Navigation Satellite System. ii, 1–3, 5–8, 11–14, 41, 42, 45–49, 51

GNSS-R GNSS reflectometry. ii, xii, 1, 2, 6, 14, 53, 65, 66, 95

GNSSDO GNSS Disciplined Oscillator. 8, 14, 15, 41, 42, 47, 48, 65, 66

GPS Global Positioning System. 1

HAS High Accuracy Service. 4

I/Q In-phase and Quadrature. 5, 13, 14, 17, 19, 20

IF Intermediate Frequency. 11, 13, 14

IIP2 Second-Order Intercept Point. 20

IIP3 Third-Order Intercept Point. 20

KCL Kirchhoff's Current Law. 76

LDO Low Dropout Regulator. 27, 28, 31, 33

LED Light Emitting Diode. 31, 33

LEO Low Earth Orbit. 7

LHCP Left-Hand Circularly Polarized. x, xi, 9, 12, 53, 55–62, 65, 66, 95–99

LNA Low Noise Amplifier. 19

LO Local Oscillator. 2, 11, 13, 14, 17, 20, 37, 41

MEO Medium Earth Orbit. 1, 3, 7

MTIE Maximum Time Interval Error. ix, x, 44, 47–51, 65

OADEV Overlapping Allan Deviation. x, 43, 46, 47, 49–51

OAVAR Overlapping Allan Variance. 43

OS Open Service. 4

OSNMA Open Service Navigation Message Authentication. 4

PCB Printed Circuit Board. ii, ix, xi, xii, 15, 17, 27, 30, 35–37, 77, 78, 100

PLL Phase-Locked Loop. 14, 41, 51

PP Peak-to-Peak. 17, 20, 39

PPS Pulse Per Second. 41, 42

PRS Public Regulated Service. 4

PSD Power Spectral Density. 5

RAAN Right Ascension of the Ascending Node. 3

RF Radio Frequency. 13, 14, 18, 20, 37

RHCP Right-Hand Circularly Polarized. x, xi, 5, 9, 12, 53, 55–62, 65, 66, 95–98

RMA Relative Mean Anomaly. 3

RMS Root Mean Square. 15, 17, 42, 44, 51

SAR Search and Rescue Service. 4

SAR Synthetic Aperture Radar. 1

SMA SubMiniature version A. 18, 30

SNR Signal-to-Noise Ratio. 14

TIE_{RMS} Time Interval Error Root Mean Square. ix, x, 44, 47–51, 65

UCLouvain Université catholique de Louvain. 11

UTC Coordinated Universal Time. 14

VCTCXO Voltage-Controlled Temperature-Compensated Crystal Oscillator. 42

VIA Vertical Interconnect Access. 30

VNA Vector Network Analyzer. 63

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1 Introduction

Remote sensing provides essential data for understanding and managing the Earth's environment, with applications ranging from climate monitoring to disaster response and natural resource management. Starting in the 1980s, Global Navigation Satellite System (GNSS) signals such as the well-known Global Positioning System (GPS) were proposed for remote sensing of the Earth's atmosphere, ionosphere and geodynamic processes by performing radio occultation measurements [5].

Observing from the ground the variations of the GNSS signals as they pass through the atmosphere, scientists can infer valuable information about the long term trends in global climate change, the consequences of a global greenhouse effect and the dynamics of the Earth's atmosphere and ionosphere.

In 1993, in parallel to the radio occultation technique, the concept of GNSS reflectometry (GNSS-R) emerged as a technique to sense the Earth's surface using reflected GNSS signals [6]. For example in 1998, airborne experiments demonstrated the potential of GNSS reflections to retrieve ocean surface roughness and related wind information [7].

Other missions such as the STS-59 Space Shuttle mission in 1994 [8] were using Synthetic Aperture Radar (SAR) technology onboard satellites to perform Earth observation. However, SAR systems require high power transmitters and large antennas, which limits their deployment to large and expensive satellites. If we suppose solar irradiance of around 1.3 kW/m^2 and photoelectric conversion efficiency of around 30%, the typical power delivered by solar panels for MEO satellites is around 3.7 kW to 2 kW at the end of their life [9]. On the other hand, SAR technology has typical radiated power of around 0.5 to 3 kW [10], which leaves little power to the other instruments on board the satellite.

In recent years, many companies and research groups have been working on the development of GNSS-R, but this time based on satellite platforms. The main advantage of GNSS-R with satellite receivers is that it can provide global coverage

of the Earth's surface without relying on high power transmitters, as the GNSS signals are already present in the environment and can be received by small and low-cost receivers. As the receiver of this bistatic radar configuration is very far from the surface, it requires a very directive antenna to have the ability to focus on very narrow regions of the Earth's surface. Recent projects, such as the HydroGNSS mission [11], rely on large aperture antennas which implies the launch of large satellites with high cost. The use of multiple small satellites such as CubeSats in formation to implement an irregular antenna array has emerged as a promising approach [12].

An antenna array of space-based movable antennas is an innovative concept that enables cost-effective and high-resolution Earth observation. This approach is studied by the G_{nss}-r sateLLITe earTh obsERvation (GLITTER) project founded by European Commission (EC), which investigates GNSS-R technology using CubeSats in formation in space for Earth observation. This project presents several challenges, and this thesis focuses on developing solutions for some of them.

The GLITTER project aims to demonstrate the feasibility of using CubeSats in formation for GNSS-R Earth observation. This project will unfold in several phases, starting with the design and development of the hardware and software components, followed by laboratory testing and validation. In the short term, a first demonstration will be conducted using drones, and in the long term, a swarm of CubeSats will be launched to achieve a full demonstration of the concept.

One of the challenges is the low signal strength of the reflected GNSS signals, which requires the design of a sensitive and low-noise receiving chain. The second challenge is the synchronization of the Local Oscillator (LO) of the different CubeSats, which is essential for the performance of the system. The third challenge addressed in this thesis is the design of the antennas, which must be optimized for the specific requirements of the GNSS-R application.

This thesis is structured as follows. First, the context is presented, with a focus on the GNSS systems and the GLITTER project and its objectives. Secondly, the proposed receiving chain implemented in the project is presented with the specifications, where the specific hardware contributions of the thesis are highlighted. These contributions, detailed in the following chapters, are the design of a low-noise baseband circuit, a synchronization method for the LO of the different CubeSats and the design of dedicated antennas for the reception of GNSS signals. Finally, a summary of the main contributions of each chapter is provided and some future perspectives of the project are discussed.

2 Context

2.1 GNSS systems

The European Union Agency for the Space Programme (EUSPA) defines the Global Navigation Satellite System (GNSS) as "Any satellite constellation that provides global positioning, navigation, and timing services." [13]. The term "GNSS" is often used to refer to the entire constellation of satellites, such as Galileo (Europe), GPS (United States), GLONASS (Russia) and the BeiDou (China) systems. EUSPA explains that these systems use signals from space to transmit ranging and their timing information to GNSS-enabled receivers, which then use this very accurate data to determine location.

For example, the Galileo system consists of 30 satellites [14] in Medium Earth Orbit (MEO) ¹ [1] at an altitude of approximately 23 222 km [16], with an orbital period² of about 14 hours [17]. The satellites are arranged in three orbital planes, as shown in Figure 2.1. Currently, 24 satellites are operational, while the remaining 6 are auxiliary satellites or are not in service. The 24 nominal satellites are distributed in 3 orbital planes, each containing 8 satellites all separated by an angular distance of 45° in the Relative Mean Anomaly (RMA) and with a Right Ascension of the Ascending Node (RAAN) of 0°, 120° and 240° respectively. The first satellite in the figure, A01, is the reference. The constellation is designed to ensure that at least 4 satellites are visible from any point on Earth at any time, providing global coverage. A GNSS receiver needs signals from at least 4 satellites to compute its three-dimensional position and time.

¹MEO satellites operate at altitudes of 19,000 to 23,000 km and at a velocity of under 4 km/s [15].

²The orbital period, T , of a satellite is the time it takes to complete one full orbit around its planet.

The Galileo system is designed to provide the following services [4] worldwide:

Service	Description	User segment
Open Service (OS)	Free access to positioning and timing information	General public
High Accuracy Service (HAS)	Centimeter-level accuracy for precision applications	Surveying, agriculture, and autonomous vehicles
Public Regulated Service (PRS)	Encrypted service for government-authorised users	Government and security agencies
Search and Rescue Service (SAR)	Detection and location of distress signals	Emergency responders
Open Service Navigation Message Authentication (OSNMA)	Complements the OS by delivering authentication data	General public and professional Open Service users

Table 2.1: Galileo main services and their target user segments [4].

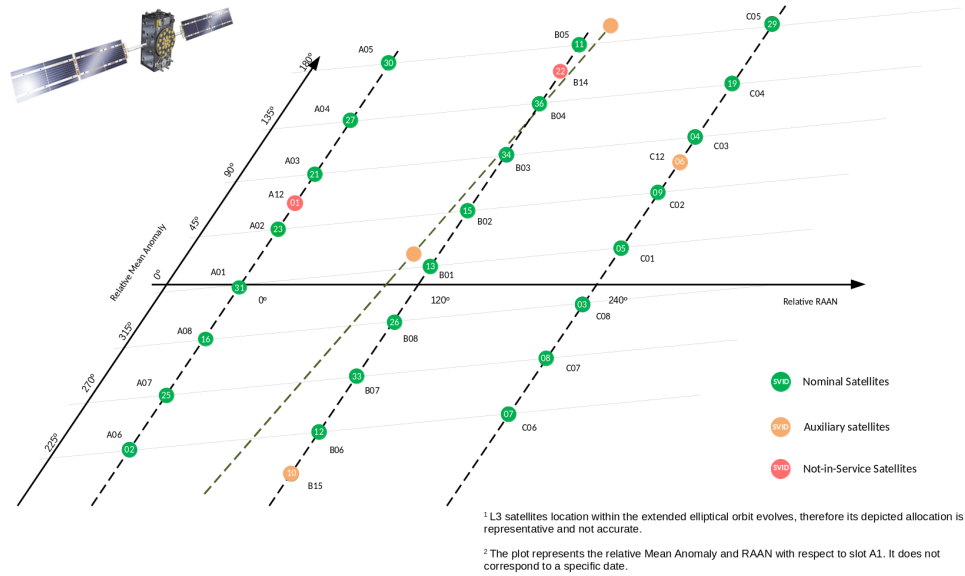


Figure 2.1: Orbital planes and slots of the Galileo constellation [1].

In GNSS systems, signals are transmitted on specific frequency bands, which are allocated for different purposes. Each band has a specific center frequency ("carrier wave") and a bandwidth that determines the range of available frequencies. These carrier frequencies are derived from a fundamental frequency of 10.23 MHz. For instance, the L1 carrier which is the most popular GPS carrier, is obtained as

$$f_{L1} = 154 \times 10.23 \text{ MHz} = 1.57542 \text{ GHz}. \quad (2.1)$$

Similarly, the other L bands are different multiples of the fundamental frequency.

Galileo bands have other names, such as E1, E6, E5, E5a, and E5b, each with specific characteristics and applications [2]. For example, the In-phase and Quadrature Power Spectral Density (PSD) of the Galileo E5 and GPS L5 bands are shown in Figure 2.2. The I and Q components can have different Power Spectral Density (PSD)'s because of the modulation scheme used. For example, the Galileo E5 band uses the Alternative Binary Offset Carrier (AltBOC) modulation [18], where the I component is used for the data stream and the Q component for the pilot stream used for tracking and coherent integration. On the other hand, the E1 band uses the Composite Binary Offset Carrier (CBOC) modulation [19]. The I channel transmits both the data and pilot streams for civil applications, while the Q channel is used for governmental and authorised military encrypted services.

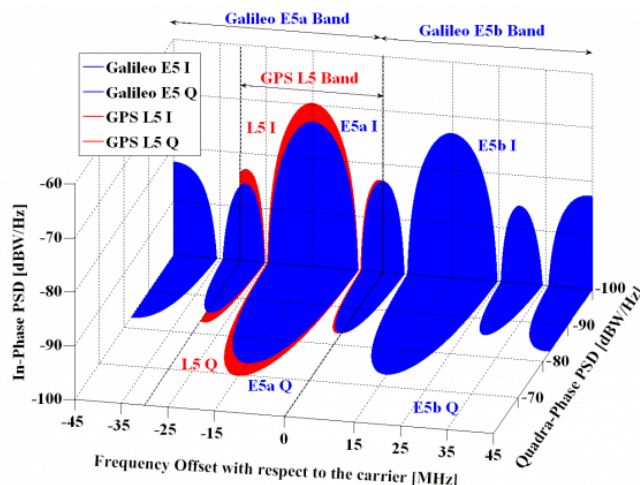


Figure 2.2: Galileo E5 and GPS L5 signals modulation spectrum [2].

For signal modulation, GNSS systems like GPS, Galileo and BeiDou use Code Division Multiple Access (CDMA), while GLONASS is the only one that still uses Frequency Division Multiple Access (FDMA) for the transmission of its navigation signals [20]. The GNSS signals sent from all of these satellites are Right-Hand

Circularly Polarized (RHCP). More details on polarisation are given in Chapter 6.

This is the context in which the European Commission (EC) GLITTER project takes place. Being part of it, we mainly focus on Galileo signals and the L-band microwave frequencies for which signals are only slightly affected by clouds, rain and weather conditions and have better vegetation and soil penetration than higher frequency radar [11].

2.2 GLITTER Project

The G_{NSS}-r sateLLite earTh obsERvation (GLITTER) project [21, 22, 23] is an EC project funded by the Marie Skłodowska-Curie Actions. The objective is to investigate GNSS reflectometry (GNSS-R) technology using small satellites such as CubeSats in formation in space for Earth observation. CubeSats are nanosatellites with a standard (1U) size of $10 \times 10 \times 10 \text{ cm}^3$ and GNSS-R is a remote sensing technique that exploits the reflection of GNSS signals off the Earth’s surface to retrieve geophysical information, such as those presented in Table 2.2.

Parameter	Purpose
Soil moisture (quantity of water present in the surface layer of the soil)	Agriculture, hydrology, and climate studies
Sea level and floods monitoring	Coastal monitoring and climate change studies
Geodesy (Earth’s shape, orientation, and gravity field)	Geophysics and Earth science research
Humidity	Weather forecasting and climate studies
Ice detection, density and classification	Climate change studies and polar research
Moving targets	Traffic monitoring and maritime surveillance

Table 2.2: Examples of geophysical parameters that can be retrieved with GNSS-R.

The proposed observation concept relies on a multi-static architecture. A multi-static architecture is a sensing (or communication) system that uses multiple, spatially distributed transmitters and receivers at different locations, enhance detection, improve coverage and ensure resilience. This concept involves several CubeSats arranged as a distributed antenna array that receives the reflected GNSS signals from the Earth’s surface. The main advantage of this architecture is that it allows an improvement of the effective resolution of the system by increasing the size of the synthetic aperture and does not rely on powerful transmitters onboard

the CubeSats. The CubeSats in formation will act as receivers which operate in Low Earth Orbit (LEO), approximately 400 km above the Earth, while GNSS transmitters remain in Medium Earth Orbit (MEO), approximately 20,000 km above the Earth. As explained in the work of L. Paucot [3], this geometry offers a favorable compromise between observation capabilities such as resolution and minimization of the sidelobes in the area where the reflected signal is received, called the glistening zone, and the strict constraints of mechanics and costs. One of the formation geometries proposed in this work is illustrated in Figure 2.3. A 'master' satellite is located at the center of the figure. To approximate a uniform density, radial density has been made equaled to the azimuthal density of satellite by increasing the spacing of them along a given branch, starting from the master satellite.

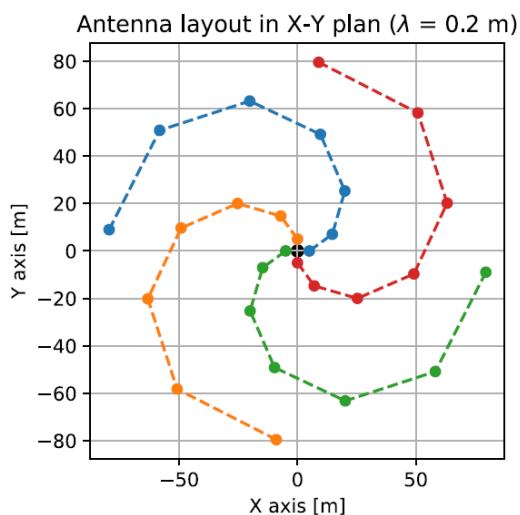


Figure 2.3: Example from the works of L. Paucot of a formation geometry with 29 antenna elements [3].

From a mission-performance perspective, the objective is to improve the effective ground resolution from approximately $10 \text{ km} \times 10 \text{ km}$ with a single satellite to about $1 \text{ km} \times 1 \text{ km}$ with a formation of 20 CubeSats. For example, Figure 2.3 illustrates a formation geometry with 29 antenna elements spaced by approximately 5 m with a radius around 80 m, is expected to achieve a resolution of approximately $0.8 \text{ km} \times 0.8 \text{ km}$. Other configurations have been proposed, keeping a minimal safe distance between the elements of around 1 m.

In this context, synchronization between the CubeSats remains a critical challenge because the relative geometry evolves continuously under orbital dynamics and

perturbations, causing time-varying delays and Doppler shifts that must be compensated for coherent operation. These physically admissible relative trajectories can be described by the Clohessy-Wiltshire equations [24].

A first solution is to synchronize the satellites through a wireless distribution of a common local oscillator, but this approach is non-trivial and variations in inter-satellite spacing induce additional phase shifts. A practical mitigation strategy studied in this work consists of using the incoming GNSS signal itself, through a u-blox-based receiver chain. This approach, known as GNSS Disciplined Oscillator (GNSSDO), is based on the inherent stability and accuracy of GNSS signals as a shared frequency reference. Each CubeSat in the formation would then synchronize its local oscillator to this common reference, thereby maintaining coherence across the entire CubeSat array.

2.3 Short-term GLITTER objectives

The first upcoming experimental objective linked to this thesis is to implement two processing chains that emulate two CubeSats (see Figure 2.4) receiving different Global Navigation Satellite System (GNSS) signals, one direct from the satellite and one reflected by the ground. In this experiment, the two CubeSats are replaced by two drones, flying in the air with all the necessary hardware running on-board. The goal of this experiment is to validate the hardware design and the synchronization method by demonstrating the reception of a GNSS signal during 100 ms with two independent processing chains and to evaluate the performance of the system in terms of signal quality, synchronization accuracy, and data transmission.

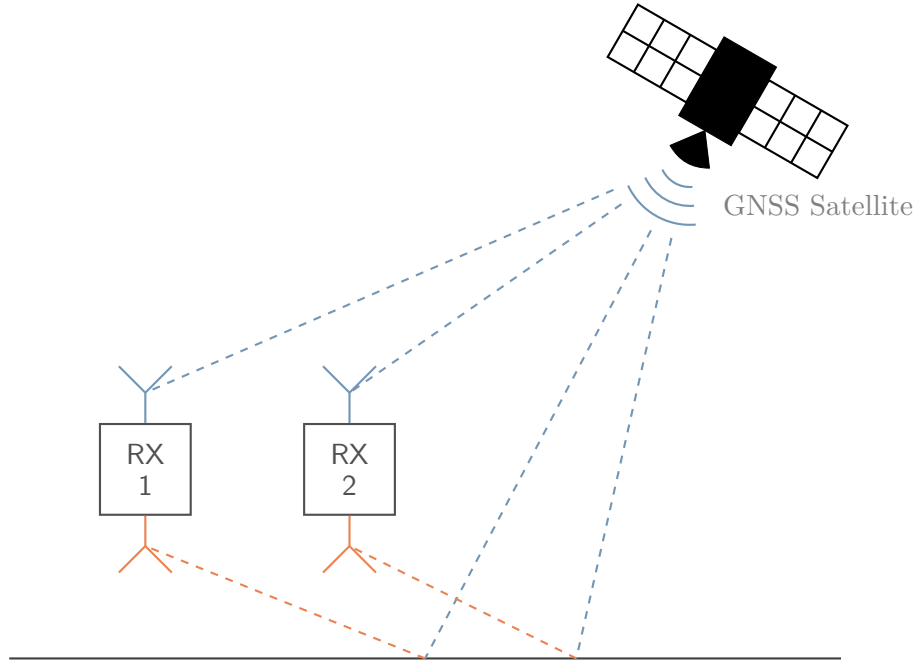


Figure 2.4: Global schematic of the experiment goal with two CubeSats.

The two drones are represented by RX~1 and RX~2 respectively in the schematic. Each of the two is equipped with two antennas: a top antenna (in **blue**) for receiving the direct signal from the satellite, and a bottom antenna (in **orange**) for receiving the ground-reflected signal. The signal sent by the GNSS satellite is a Right-Hand Circularly Polarized (RHCP) signal and the reflected signal is transformed into a Left-Hand Circularly Polarized (LHCP) signal due to the reflection from the ground [25]. The RHCP ray paths are in blue and the LHCP paths are in orange in the schematic.

3 System specifications

To meet the experiment objectives, each CubeSat processes both of the two received signals (direct and reflected) with the same heterodyne¹ architecture developed by the GLITTER UCLouvain team. This hardware processing chain takes place between the antenna and the digital processing, as illustrated in Figure 3.1.

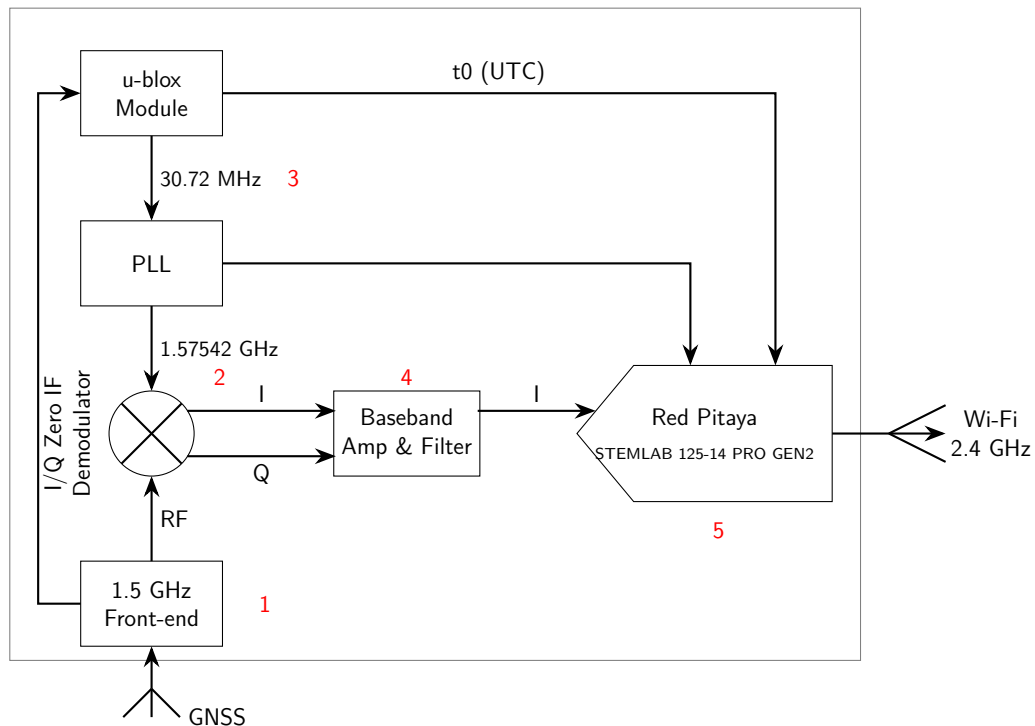


Figure 3.1: Global schematic of the processing chain.

As a first step, the GNSS signal is received by the antenna. Two antenna designs are required. The first design is intended for the ground-reflected signal and the

¹A heterodyne architecture is a signal processing technique that involves mixing a received signal with a locally generated signal (Local Oscillator) to produce an Intermediate Frequency signal, which is easier to digitize.

second one for the direct signal coming from the satellite. The design is based on the requirements of the signals of interest of the GLITTER mission, such as the operating frequency (around 1575 MHz for the E1 band), high possible gain, Left-Hand Circularly Polarized (LHCP) for downward and Right-Hand Circularly Polarized (RHCP) polarization for upward facing antennas, efficiency, directivity and physical constraints of the CubeSat platform.

As a first prototype, due to cost constraints and the need for rapid prototyping of the hardware chain, a corner trimmed patch design with an affordable substrate has been selected. The design process is detailed in Chapter 6 and involves electromagnetic simulation to optimise the antenna parameters, such as the dimensions of the patch, the size of the corners and the feed point location to achieve the desired performance.

After the GNSS signal has been captured by the antenna, it enters the Front-end circuit, numbered (1) in the schematic. This circuit and the linked budget of the hardware system are designed in the link budget chapter of the work of Ir. Ashish Mhaske, GLITTER–GNSS-R Satellite Earth Observation: Design of RF Components, Computing Platform and Array Configuration for Drone Experiment [26], where the design considerations are discussed in detail below.

The very weak GNSS direct signal entering the antenna has an estimated minimum input power of around -128.5 dBm. Supposing an antenna gain of around 8 dBi and a surface reflection loss of around 15 dB, the minimum power at the output of the antenna is around $-128.5 + 8 = -120.5$ dBm for the direct signal and around $-128.5 + 8 - 15 = -135.5$ dBm for the reflected signal.

Given that the ADC full-scale is taken as $+10$ dBm ($\pm 1 V_{\text{peak}}$ into 50Ω) and the thermal noise density is approximated by $k \times T$, where k is the Boltzmann constant and $T = 290$ K is the standard temperature for antenna beam fully illuminating the ground, the total gain of the processing chain must satisfy two constraints : (i) the minimum value is set to ensure that the analog noise is above the quantization noise of the ADC with some margin and (ii) the maximum value is chosen to have an output signal below the ADC full-scale to avoid saturation. The analog noise power at the input of the ADC over 20 MHz bandwidth ² is estimated in this work to be around -100 dBm and the quantization noise power is around -68 dBm over the same bandwidth.

²A bandwidth is the difference between the upper and lower frequencies at which the signal power drops by 3 dB from its maximum value.

The 20 MHz bandwidth has been chosen in order to fulfill the E1, E5, L1 and L5 bands requirements, with the Galileo E5 being the widest one. As the main-lobe bandwidth of the E5 band is 51.15 MHz and the signal will be demodulated at 0 Hz, the low-pass cut-off frequency will be around $51.15 / 2 = 25.575$ MHz. For practical reasons³ 20 MHz has been chosen.

The total gain should be between $G_{\min} = -68 - (-100) + 6 = 38$ dB and $G_{\max} = +10 - (-100) - 10 = 100$ dB. The 6 dB margin is added to ensure that the analog noise is sufficiently above the quantization noise and the 10 dB margin is subtracted to avoid saturation of the ADC. The total gain of the hardware chain is chosen to be around 70 dB, which is a good compromise between the two limits and allows the first low-noise stages to dominate the noise figure of the receiver, according to Friis' formula⁴. It will lead to signal power at the input of the ADC of around -51 dBm for the direct signal, -66 dBm for the reflected one and -30 dBm for the thermal noise. It is important to note that the GNSS signals remain below the thermal noise level even after amplification and can only be recovered through correlation processing.

This total gain is separated into two parts: the first part is the gain of the Front-end, which is around 32.06 dB, and the second part is the gain of the baseband circuit, which is around 40 dB. A significant portion of the gain is placed in the RF Front-end to minimize the impact of noise from subsequent stages, while the baseband stages provide additional gain without risking saturation due to RF interferers. As the carrier frequency is around 1575 MHz, the Front-end operates at Radio Frequency from 1530 MHz to 1620 MHz, while the baseband circuit operates at IF from 1 kHz to 20 MHz.

The RF signal delivered by the Front-end enters the In-phase and Quadrature (I/Q) demodulator in (2). The demodulator downconverts the RF signal to an Intermediate Frequency (IF) signal by separating the I/Q components with a 90° phase shift⁵. An example of an I/Q signal is the Galileo E5 signal shown in Figure 2.2. The frequencies of the IF signal are given by the following equation :

$$f_{IF} = f_{RF} - f_{LO} \quad (3.1)$$

where f_{RF} is the frequency of the incoming RF GNSS signal and f_{LO} is the frequency of the LO signal. The selected demodulator is the LT5575 from Analog Devices [28],

³According to the software team, they don't need more than 20 MHz.

⁴The Friis' formula can be formulated as follows: $NF_{total} = NF_1 + \frac{NF_2-1}{G_1} + \frac{NF_3-1}{G_1G_2} + \dots + \frac{NF_n-1}{G_1G_2\dots G_{n-1}}$, where NF is the noise figure and G is the gain of each stage in the signal chain.

⁵Explanations about In-phase and Quadrature (I/Q) components can be found in [27].

which operates up to 2.7 GHz and is well suited for weak-signal GNSS-R reception thanks to its high linearity and low noise figure⁶.

After mixing by the demodulator, the In-phase and Quadrature components are expressed as :

$$\begin{cases} x_I(t) = x(t) \cos(2\pi f_{LO}t) \\ x_Q(t) = x(t) \sin(2\pi f_{LO}t) \end{cases} \quad (3.2)$$

The I/Q demodulator needs a Local Oscillator (LO)⁷ signal to downconvert the received Radio Frequency (RF) signal to Intermediate Frequency (IF). In this setup, the LO signal is provided by the Phase-Locked Loop (PLL) (3), which is locked to a stable frequency reference generated by the u-blox module [29] that is synchronized with the incoming GNSS signal received through the Front-end. This u-blox module is a commercial GNSS receiver that provides a stable frequency reference derived from the received GNSS signal, but also Coordinated Universal Time (UTC) information which can be used to timestamp the down-converted sampled received signal. The adopted approach, which uses the incoming GNSS signal as a shared frequency reference via u-blox receivers, is known as GNSS Disciplined Oscillator (GNSSDO). It is a type of frequency reference that uses the timing signals from the GNSS signals to maintain a highly accurate and stable frequency output by combining the long-term frequency accuracy of satellites with the short-term stability of a local crystal oscillator. This method avoids the complexity of distributing a physical LO reference among all the CubeSats and relying on heavy and difficult inter-satellite synchronization links. The synchronization of the LO between multiple receivers is a critical aspect of the system design. This ensures that the down-converted signals from different receivers are coherent and can be properly combined in the post-processing step. This synchronization method will be explained and characterized in Chapter 5.

The demodulated signal is then filtered and amplified by the baseband processing chain (4). The chain is designed to provide around 40 dB of gain, together with a band-pass filter starting from 1 kHz to 20 MHz. The 1 kHz high-pass filter has been selected in order to remove the DC component in the baseband signal and the unnecessary information close to the center of the band for the processing team.

The filter is implemented using three stages: first low-pass filter with a cut-off frequency of 20 MHz and 10 dB of gain, followed by a high-pass filter with a cut-off

⁶The noise figure is a measure of the degradation of the Signal-to-Noise Ratio (SNR) caused by components in the signal chain.

⁷A Local Oscillator (LO) is a signal source that generates a stable frequency reference for the demodulation process.

frequency of 1 kHz and 15 dB of gain, and finally a second low-pass filter with a cut-off frequency of 20 MHz and 15 dB of gain.

The choice of the components such as amplifiers, resistors and capacitors, is based on the type of circuit (differential or not), the required gain, bandwidth, noise performance, as well as on the availability of components that can be easily integrated into a PCB design.

Particular attention is also paid to impedance matching⁸ and isolation in the PCB implementation in order to have good demodulation performance and limited interference. The validation plan follows three consecutive steps: simulation with LTspice software [30] to verify the theoretical behavior, fabrication according to PCB design rules, and experimental testing. More information about the baseband circuit is provided in Chapter 4.

Finally, the output signal is sampled by the Analog-to-digital Converter (ADC)⁹ of the Red Pitaya [31] (5) for digitization. However, the Red Pitaya has only two input channels, so only the I component of the two signals (direct and reflected) is sampled for the first hardware demonstration. The sampled signal is transmitted via Wi-Fi to a master computer for further processing. After the amplification of 32.06 dB in the Front-end and 40 dB in the baseband circuit, the signal power level at the input of the Red Pitaya for the reflected signal should be $-64 \text{ dBm} = 0.11 \text{ mV}_{\text{RMS}}$ and $-48.5 \text{ dBm} = 0.84 \text{ mV}_{\text{RMS}}$ for the direct signal.

The components introduced earlier and designed in this thesis are detailed in the following three sections: baseband circuit, characterization of the GNSSDO and patch antenna design.

⁸Impedance matching is the process of designing the input and output impedances of a circuit to maximize power transfer and minimize signal reflections.

⁹An Analog-to-digital Converter is a system that converts an analog (continuous) signal into a digital (discrete) signal.

4 Baseband circuit

This chapter explains the design and characterisation of the baseband circuit introduced in Chapter 3. The design process is structured into four main phases: the functional objectives of this circuit, the design and validation of it with simulations, the PCB implementation and finally the characterisation of the manufactured PCB. The baseband circuit will be produced four times, two for each drone as they have two antennas, and will have the identical paths: one for the I channel and one for the Q channel. The two I/Q circuits will be implemented on the same Printed Circuit Board (PCB) and designed identically to ensure symmetry and reduce phase imbalance.

4.1 Dimensioning and specifications

As explained in Chapter 3, the main objective of the baseband chain is to provide an overall low-noise gain of approximately 40 dB while implementing a band-pass filter between 1 kHz and 20 MHz. A first-order high-pass filter is used to remove the DC component and the LO leakage and low-frequency drift, and does not need to be very selective. A seventh-order low-pass filter is used to provide a fast out-of-band attenuation for anti-aliasing and to limit the integrated noise delivered to the ADC within the 20 MHz bandwidth. This bandwidth should have a gain as flat as possible in order to avoid distorting the signal and degrading the correlation performance. The Front-end circuit connected at the input of the baseband circuit provides the following output specifications:

Parameter	Specification
Output signal type	Differential pair (I+/I- or Q+/Q-)
Minimal output signal power	-103.44 dBm
Minimal output signal voltage (PP)	0,0042563 mV
Minimal output signal voltage (RMS)	0,0015048 mV
Output impedance	50 Ω (to be matched with baseband input)

Table 4.1: Output specifications of the Front-end circuit.

At the output stage, the chain is designed to interface with the Red Pitaya platform [31] which has the Radio Frequency input specifications listed in Table 4.2.

Parameter	Specification
Number of channels	2
Full scale voltage range	± 1 V
Absolute max. input voltage	± 6 V
Type of input	Single-ended
Sample rate	125 MS/s
ADC resolution	14 bits
Input coupling	DC
Bandwidth	DC - 50 MHz
input impedance	1 M Ω
Connector type	SMA
ADC	LTC2145-14

Table 4.2: Red Pitaya ADC input specifications.

To meet these requirements, the baseband chain is implemented as a three-stage cascade of active filters and amplifiers, as shown in Figure 4.1. This approach overcomes the gain-bandwidth limitations of a single amplifier while enabling optimised noise performance according to the Friis formula, as the noise contribution of each stage is divided by the cumulative gain of the preceding stages. The filtering strategy combines the following three stages:

Stage	Type	Cut-off frequency	Gain
1	Low-pass	20 MHz	10 dB
2	High-pass	1 kHz	15 dB
3	Low-pass	20 MHz	15 dB

Table 4.3: Filtering strategy for the baseband chain.

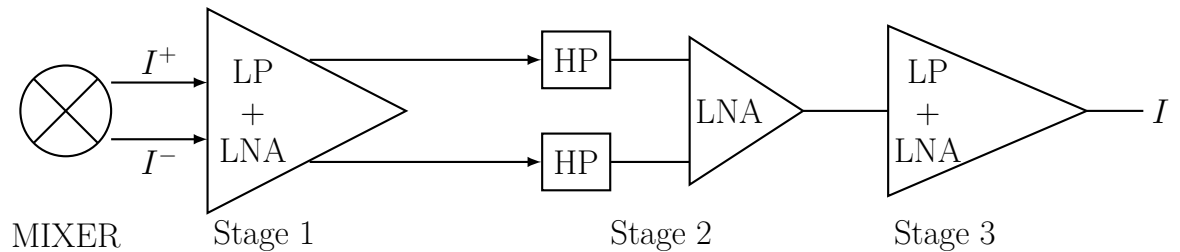


Figure 4.1: Block diagram of the baseband chain.

The final stage includes an ADC driver. This kind of ADC is designed to rapidly charge the sampling capacitor of the Red Pitaya with a high slew rate¹ and ensure good settling, thereby reducing sampling-induced distortion.

4.2 Component selection

From an implementation perspective, all selected components must be available on DigiKey [32] and all amplifiers should be Low Noise Amplifier (LNA)² in order to ensure a good noise figure for the overall chain. The active components selected for the three stages of the baseband chain are listed in Table 4.4 with their specifications.

Component	Stage	Type	Gain	Noise Spec	Supply	Bandwidth
LT5575	1	I/Q Demodulator	3.5 to 4.2 dB	12.8 dB (NF)	5 V	800 MHz – 2.7 GHz
ADA4940	1	Fully Diff. Amp	Configurable	3.9 nV/ $\sqrt{\text{Hz}}$	5 V	260 MHz
LT1818	2	Op-Amp	Configurable	6.0 nV/ $\sqrt{\text{Hz}}$	± 3.3 V	400 MHz
AD8065	3	Op-Amp	Configurable	7.0 nV/ $\sqrt{\text{Hz}}$	± 3.3 V	145 MHz

Table 4.4: Summary of active components selection and key specifications.

All components have a bandwidth much higher than the 20 MHz target, noise figure of 12.8 dB for the demodulator and a maximum input noise voltage density in the order of 7 nV/ $\sqrt{\text{Hz}}$ for the amplifiers. In addition to frequency-domain performances, power consumption, thermal dissipation, weight, size and cost are treated as explicit design constraints throughout the development.

The selected mixer, the LT5575, is described in Table 4.5 with its key specifications. It provides conversion to baseband with a gain of approximately 4 dB, which is beneficial for the overall noise performance of the chain. The I/Q components are well balanced with a gain mismatch of 0.04 dB and a phase mismatch of 0.4 degrees which is needed for accurate I/Q demodulation, and the output bandwidth, DC to 490 MHz, is sufficient for the 20 MHz bandwidth requirement.

¹The slew rate is the maximum rate of change of the output voltage per unit time, typically expressed in volts per microsecond ($V/\mu s$). A high slew rate allows the amplifier to quickly respond to rapid changes in the input signal such as high frequency components, ensuring that the signal settles within the required time frame for sampling.

²Low Noise Amplifier (LNA) are amplifiers designed to provide an high gain while adding as little noise as possible to the signal.

Parameter	Specification
Architecture	Direct conversion I/Q demodulator
RF and LO Inputs	Single-ended
IIP3 ³	28 dBm (at 900 MHz) / 22.6 dBm (at 1.9 GHz)
IIP2	54.1 dBm (at 900 MHz) / 60 dBm (at 1.9 GHz)
I/Q Gain Mismatch	0.04 dB (typical)
I/Q Phase Mismatch	0.4deg. (typical)
LO Input Power Range	−13 dBm to 5 dBm
Output Bandwidth	DC to 490 MHz
RF to LO Isolation	57.1 dBc (at 1900 MHz)

Table 4.5: Summary of the LT5575 mixer specifications.

4.3 Design and validation

The circuit design is developed iteratively, starting with the first stage and progressively adding subsequent stages while validating the overall response at each step. Simulations are performed in LTspice [30] with an input signal of $-135.5 + 32.06 = -103.44$ dBm = 0.0042563 mV_{PP} (on $50\ \Omega$) going from 1 Hz to 1 GHz in order to cover the full band of interest of the circuit (1 kHz – 20 MHz). The global schematic is illustrated in Figure 4.2.

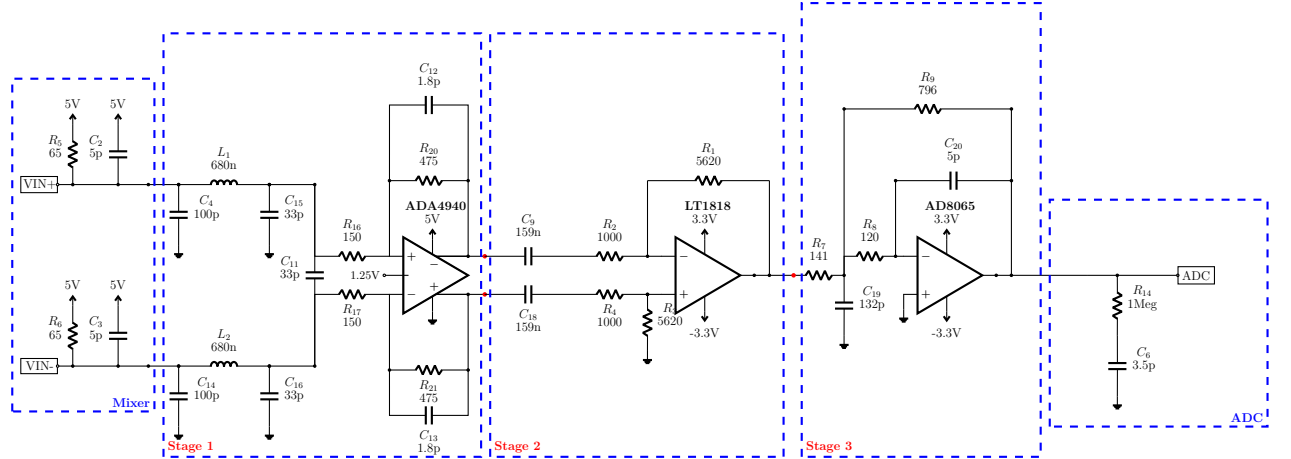


Figure 4.2: Final circuit design implemented in LTspice.

³Third-Order Intercept Point (IIP3) is a measure of the linearity of a mixer or amplifier, indicating the power level at which third-order intermodulation products would equal the fundamental output power.

4.3.1 First stage

The first stage is based on the LT5575 demodulator, following its corresponding design notes [33]. This stage is fully differential (input and output are differential signals) and provides approximately 10 dB of gain and a low-pass filter with a cut-off frequency of 20 MHz.

The difficulty of this differential stage is to deal with the DC level of 3.9 V provided by the LT5575 circuit. This is why a differential amplifier, here the ADA4940 [34], with a common-mode voltage of 1.25 V is used to shift the signal down to the input range for the subsequent stage.

The development of the transfer function of the first stage can be found in Section A.1 leading to the expression :

$$H_1(s) = \frac{V_{\text{out}}^+(s) - V_{\text{out}}^-(s)}{V_{\text{in}}^+(s) - V_{\text{in}}^-(s)} = -\frac{R_f}{R_{in}} \cdot \frac{1}{1 + s\frac{L}{R_{in}} + s^2LC_{eq}} \cdot \frac{1}{1 + sR_fC_f} \quad (4.1)$$

where s is the Laplace variable defined as $s = \sigma + j\omega$, σ being the exponential decay rate and $\omega = 2\pi f$ the angular frequency. The gain is set by the ratio R_f/R_{in} with $R_f = R_{20} = R_{21}$ and $R_{in} = R_{16} = R_{17}$, and the low-pass response is defined by the LC tank with $C_{eq} = 2C_{11} + C_{15}$, $L = L_1 = L_2$ and $C_f = C_{12} = C_{13}$, giving the cut-off frequency of $f_c = \frac{1}{2\pi\sqrt{LC_{eq}}} \approx 20$ MHz. It yields the following parameter values:

Component	Description	Value
L_1, L_2	Inductors	680 nH
C_4, C_{14}	Shunt capacitors (input)	100 pF
C_{11}, C_{15}, C_{16}	Shunt capacitors (LC filter)	33 pF
R_{16}, R_{17}	Input resistors	150 Ω
R_{20}, R_{21}	Feedback resistors	475 Ω
C_{12}, C_{13}	Feedback capacitors	1.8 pF
U2	Fully differential amplifier	ADA4940

And the simulated frequency response is shown in Figure 4.3.

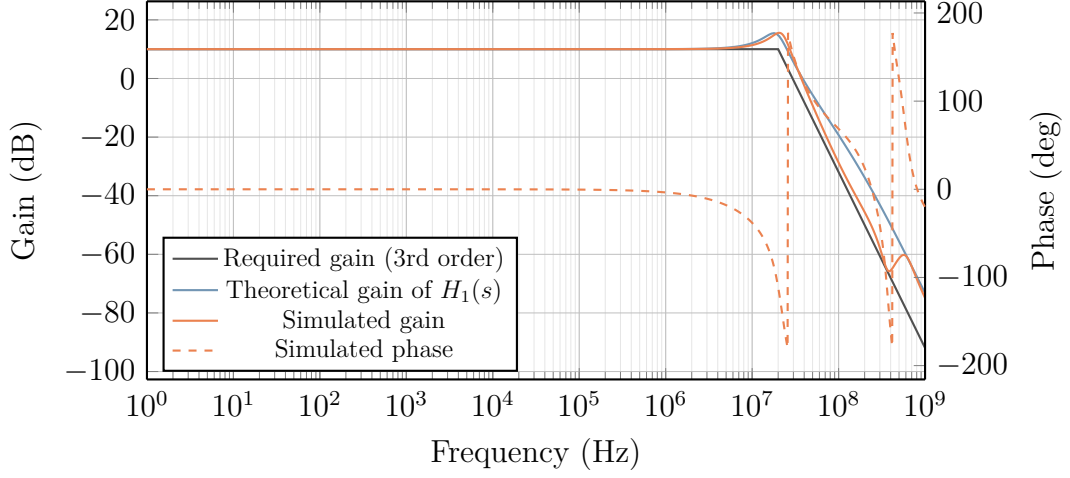


Figure 4.3: Bode diagram of the simulation result for the first stage.

We can remark that the simulated gain is around 10 dB at low frequency and starts to decrease after 20 MHz with a slope of approximately -60 dB/decade as a 3rd order system, which is consistent with the design specifications.

4.3.2 Second stage

The second stage uses the LT1818 amplifier to implement an active high-pass filter at 1 kHz while providing approximately 15 dB of gain. It also performs the differential-to-single-ended conversion required for the subsequent stage.

The development of the transfer function of the second stage can be found in Section A.2, leading to the following expression:

$$H_2(s) = \frac{V_{\text{out}}(s)}{V_{2,\text{in}+}(s) - V_{2,\text{in}-}(s)} = \frac{R_1}{R_2} \cdot \frac{sR_2C_9}{1 + sR_2C_9} \quad (4.2)$$

where the gain is set by the ratio R_1/R_2 and the cut-off frequency is defined by $f_c = \frac{1}{2\pi R_2 C_9}$. The obtained parameters values can be found in Table 4.6 and the simulated frequency response is shown in Figure 4.4.

Parameter	Value
Cut-off frequency (kHz)	1
Gain (linear)	5.623
Gain (dB)	15
$R_1 = R_3$	5620 Ω
$R_2 = R_4$	1000 Ω
$C_9 = C_{18}$	159 nF

Table 4.6: Parameter values for the second stage design.

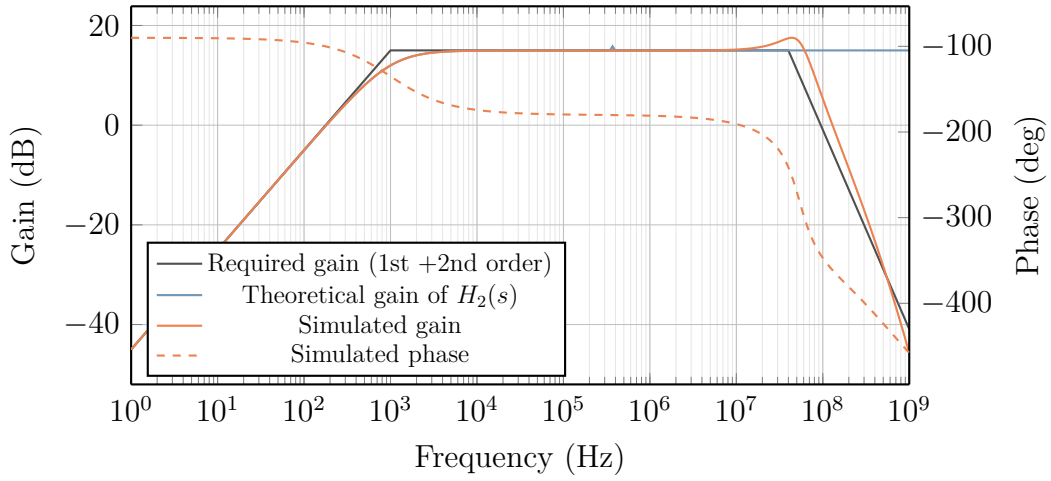


Figure 4.4: Bode diagram of the simulation result for the second stage (including amplifier saturation).

The simulated gain is around 15 dB at high frequency and decreases for frequencies below 1 kHz, which is consistent with the design specifications. We can also notice a small increase in the gain at about 20 MHz and then a decrease after around 40 MHz. The increase in gain is due to the non-linearity of the amplifier and will be compensated by a reduced cut-off frequency of the low-pass response of the third stage. The decrease after 40 MHz is due to the limited gain-bandwidth product of the amplifier. The limited gain-bandwidth product will add an additional second order low-pass filter at around 40 MHz, which will also contribute to the overall response of the chain by increasing the slope of total low-pass response of the gain.

4.3.3 Third stage

The third stage is built around the AD8065 circuit from Analog Devices in a multiple-feedback low-pass topology for an effective cut-off frequency of around

10 MHz. The natural pole frequency is set to 20 MHz, but the effective cut-off frequency is reduced to around 10 MHz by the non-idealities of the amplifier. This reduced cut-off frequency will minimize the non-idealities from the second stage, as explained in Section 4.3.2. This final single-ended stage adds about 15 dB of gain.

The development of the transfer function of the third stage can be found in Section A.3 leading to following expression :

$$\begin{aligned}
 H_3(s) = \frac{V_{out}}{V_{in}} &= \frac{\frac{-R_9}{R_7}}{1 + R_8 R_9 C_{20} \left(\frac{1}{R_7} + \frac{1}{R_8} + \frac{1}{R_9} \right) s + R_8 R_9 C_{19} C_{20} s^2} \\
 &= \frac{\frac{-1}{R_7 R_8 C_{19} C_{20}}}{s^2 + s \frac{1}{C_{19}} \left(\frac{1}{R_7} + \frac{1}{R_8} + \frac{1}{R_9} \right) + \frac{1}{R_8 R_9 C_{19} C_{20}}}
 \end{aligned} \tag{4.3}$$

where the parameters are defined with the following approach [35]⁴ :

$$\begin{cases} \text{Choose } H \text{ and } C_{20}, \\ \text{Choose } \alpha = \frac{1}{Q}, \\ \text{Define } k = 2\pi F_0 C_{20}, \end{cases} \tag{4.4}$$

and the other components can be calculated as follows :

$$\begin{cases} C_{19} = \frac{4}{\alpha^2} (H + 1) C_{20} \\ R_7 = \frac{\alpha}{2Hk} \\ R_8 = \frac{\alpha}{2(H + 1)k} \\ R_9 = \frac{\alpha}{2k} \end{cases} \tag{4.5}$$

It yields the following parameters values :

⁴There is a small mistake in the tutorial [35], the MULTIPLE FEEDBACK LOW-PASS DESIGN EQUATIONS transfert function has a variable H too much at the numerator.

Parameter	Value
Natural pole frequency = f_0 (MHz)	20
Effective cut-off frequency = $f_{c,-3dB}$ (MHz)	10
Quality factor = Q	1
Gain (linear) = $H = R_9/R_7$	5.623
Gain (dB) = $G = 20 \log_{10}(H)$	15
R_7	141.511 Ω
R_8	120.146 Ω
R_9	795.775 Ω
C_{19}	132.5 pF
C_{20}	5 pF

Table 4.7: Parameter values for the third stage design.

The quality factor should be set to $Q = 0.707$ for a Butterworth response, which provides a maximally flat frequency response in the band, but it is set to 1 in order to have a small ripple around the cut-off frequency. This peak minimizes the impact of the attenuation of the amplifier at high frequencies. The simulated frequency response of the third stage, highlighting this real-world attenuation against the ideal transfer function, is shown in Figure 4.5.

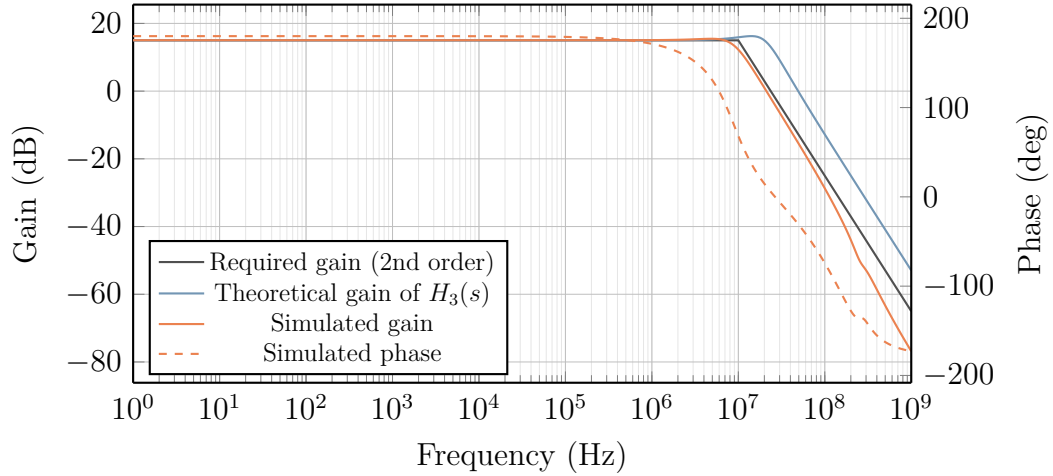


Figure 4.5: Bode diagram of the simulation result for the third stage.

The simulated gain is around 15 dB at low frequency and starts to decrease after 10 MHz with a slope of approximately -40 dB/decade as a 2nd order system, which is consistent with the design specifications.

4.3.4 Overall circuit design

With the hypothesis of non-interacting stages, i.e. that the output impedance of each stage is negligible compared to the input impedance of the subsequent stage, the overall transfer function can be obtained using the cascade rule as follows :

$$H_{total}(s) = H_1(s) \cdot H_2(s) \cdot H_3(s) \quad (4.6)$$

where $H_1(s)$, $H_2(s)$, and $H_3(s)$ represent respectively the transfer functions of the first, second, and third stages respectively. By replacing the expressions of each stage, we can obtain the following expression for the total transfer function :

$$H_{total}(s) = \left[\frac{R_f R_1 R_9}{R_{in} R_2 R_7} \right] \cdot \left[\frac{s R_2 C_9}{1 + s R_2 C_9} \right] \cdot \left[\frac{1}{1 + s \frac{L}{R_{in}} + s^2 L C_{eq}} \right] \cdot \left[\frac{1}{1 + s R_f C_f} \right] \cdot \left[\frac{1}{1 + s \frac{1}{C_{19}} \left(\frac{1}{R_7} + \frac{1}{R_8} + \frac{1}{R_9} \right) + s^2 R_8 R_9 C_{19} C_{20}} \right] \quad (4.7)$$

The total theoretical mid-band gain is $A_{v,tot} \approx 100 = 40 \text{ dB}$, confirming the design target. The total simulated frequency response is presented in Figure 4.6.

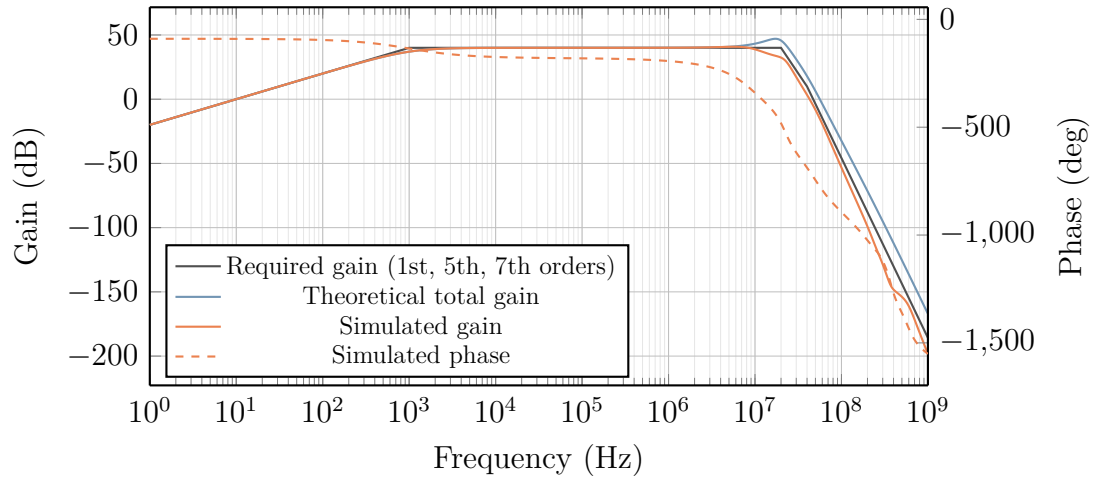


Figure 4.6: Final result of the simulation with LT-Spice of the circuit.

As required from the specifications, the final transfer function of the baseband circuit shows a passband of 1 kHz to 20 MHz with a low-pass filter of order 1, a

total low-pass of order 5 at around 20 MHz and a total low-pass of order 7 at around 40 MHz. The gain is close to 40 dB in the passband region, as expected from the design specifications.

4.4 PCB design

The PCB implementation is developed using the Kicad Software [36] from the validated LTspice schematic. Several practical additions, such as power supply regulation, debugging helpers or test points are integrated to make the board working properly and to ease the debugging process if needed.

4.4.1 Power supply

The power supply uses a symmetric rail configuration of +6 V, 0 V and −6 V but the circuit needs +5 V, +3.3 V, +1.25 V and −3.3 V for the different components. These voltage levels must be as stable as possible to ensure that a minimum amount of noise is added to the signal by the power supply itself.

Positive rails

The +6 V input generates the required +5 V and +3.3 V through Low Dropout Regulator (LDO)⁵. In this circuit, we used the ADP7118 as LDO. In total, three ADP7118 will be used, one for the +3.3 V and two for the +5 V rail in order to divide this rail in two: one rail for the demodulator and one rail for the first stage. It follows parts from design equations (in the adjustable mode of the ADP7118):

$$V_{OUT} = 1.2 \text{ V} \cdot \left(1 + \frac{R_1}{R_2}\right) \quad (4.8)$$

where R_1/R_2 is the ratio of the resistors in the feedback loop of the regulator. This ratio has the value of $R_1/R_2 = 1.75$ for +3.3 V and $R_1/R_2 = 3.167$ for +5 V. It is recommended that the R_2 value be less than 200 k Ω to minimize errors in the output voltage caused by the input current. The following values are selected :

⁵A Low Dropout Regulator (LDO) is a simple inexpensive way to regulate an output voltage that is powered from a higher voltage input.[37]

Parameter	Value
R_1 for +3.3 V	175 k Ω
R_2 for +3.3 V	100 k Ω
R_1 for +5 V	316.7 k Ω
R_2 for +5 V	100 k Ω

Table 4.8: Parameter values for the design of the positive rails.

The 1.25 V reference is produced with a voltage divider and a precision micropower shunt reference, the LM4040, in order to provide a stable voltage for the first stage of the circuit. The LM4040 has a reverse-breakdown voltage of 2.5 V, so it is powered from the +3.3 V rail and provides a stable voltage of +2.5 V. This voltage is then divided by the voltage divider, with the following equation:

$$V_{OUT} = V_{IN} \cdot \frac{R_2}{R_1 + R_2} \quad (4.9)$$

where V_{IN} is the imposed voltage of +2.5 V and R_1 and R_2 are the tuned resistors of the voltage divider. As we want a +1.25 V output, the ratio of the resistors must be $R_2/R_1 = 1$. We choose $R_1 = R_2 = 100$ k Ω .

Negative rail

The negative rail of -3.3 V is generated from the -6 V input with the ADP7182, which is a negative LDO. It follows the design equation (in the adjustable mode of the ADP7182):

$$-V_{OUT} = -1.22 \text{ V} \cdot \left(1 + \frac{R_{FB1}}{R_{FB2}}\right) \quad (4.10)$$

where R_{FB1}/R_{FB2} is the ratio of the resistors in the feedback loop of the regulator. This ratio has the value of $R_{FB1}/R_{FB2} = 1.705$ for -3.3 V. R_{FB2} must be less than 120 k Ω to minimize the output voltage errors due to leakage current. The following values are selected :

Parameter	Value
R_{FB1} for -3.3 V	170.5 k Ω
R_{FB2} for -3.3 V	100 k Ω

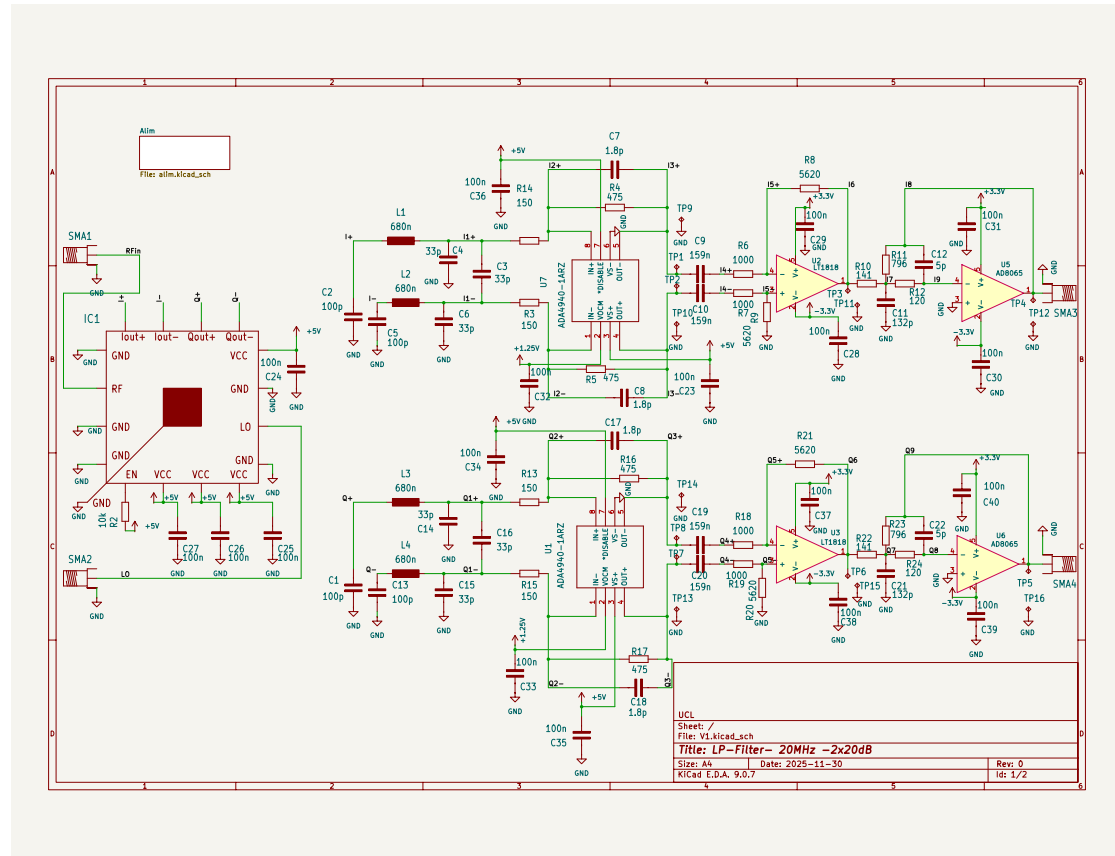
Table 4.9: Parameter values for the design of the negative rails.

4.4.2 Debugging and layout considerations

For debugging, a status LED with its series resistor is included after each power stage for basic power indication. In order to have testability, debugging and to

choose which component to supply or not, some jumpers are added to the design and dedicated test points are placed at each stage output. Local decoupling capacitors near active devices are added to the design to stabilize the supplies and to reduce the noise.

It leads to the global Kicad schematic based on the Ltspice design shown in Figure 4.7.



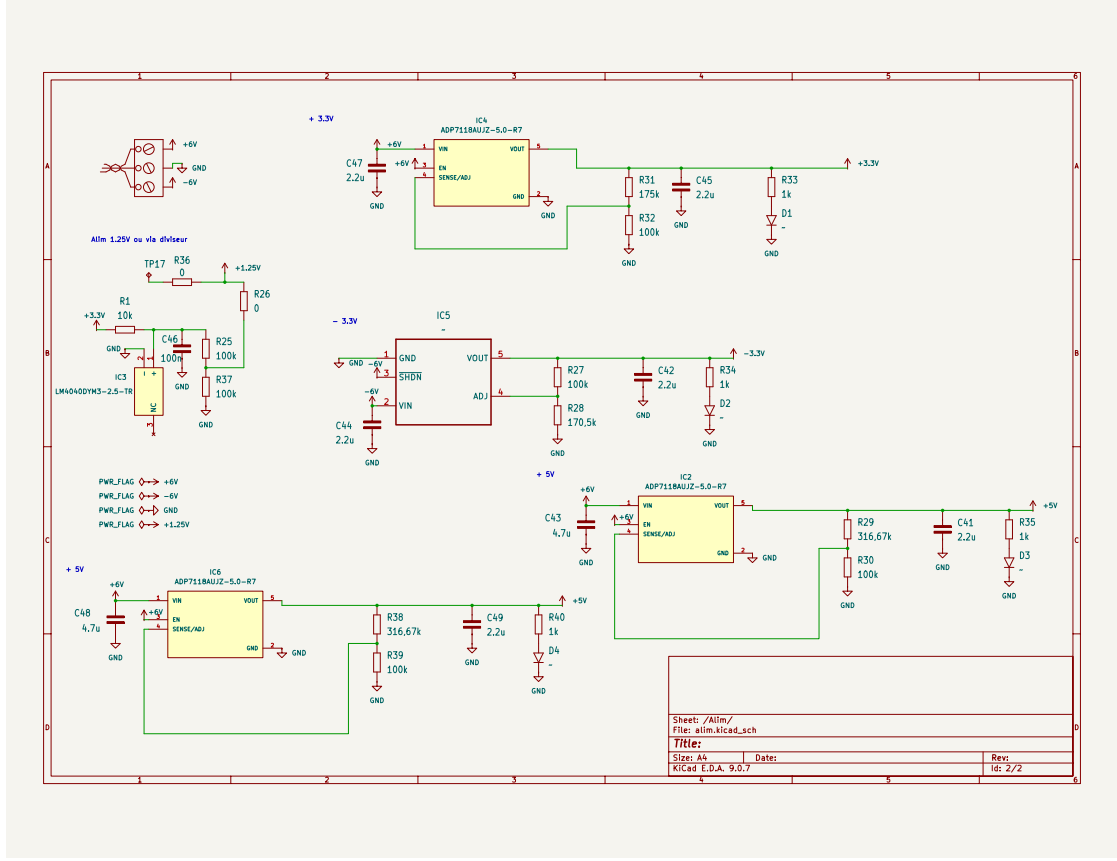


Figure 4.7: Global schematic of the final PCB design.

The layout based on this schematic includes SMA connectors used for RF/analog input and output interfaces to simplify measurements and compatibility with the Red Pitaya setup. Additional layout constraints include a minimum spacing of 0.625 mm between adjacent resistors, a trace width of 0.7 mm for 50 Ω interfacing near SMA transitions, and 0.3 mm power-line routing. Vertical Interconnect Access (VIA) stitching⁶ is applied around sensitive analog areas to improve return-current continuity and mitigate Electromagnetic Interference (EMI). Finally, the I and Q paths are kept as symmetric as possible to reduce any phase imbalance.

4.4.3 Power budget

The maximal total power consumption of the whole system can be estimated from the sum of the power consumed by the active components such as the demodulator

⁶Vertical Interconnect Access (VIA) are small copper-plated holes in a PCB that connect conductive traces across different layers. VIA stitching is the technique to add some VIA in a regular pattern to connect the ground planes together to avoid floating grounds values.

and the amplifiers. The other components such as LED's, jumpers, test points and passive components are not included in this estimation as they consume a negligible amount of power compared to the active components. The active components are supplied by the LDO's, as the power consumed by the active components is also the useful power provided by the regulators. The maximal useful power provided by the regulators must be higher than the power consumed by the active components and the neglected ones. This can be expressed with the following relations :

$$P_{max-useful} \geq P_{active} + P_{neglected} \quad (4.11)$$

where $P_{max-useful}$ is the maximal useful power that can be provided by the regulators, P_{active} is the power consumed by the active components and $P_{neglected}$ is the power consumed by the neglected components.

The useful power provided by the LDO's can be linked to the total power consumed, the thermal power due to dissipation and the power lost in the regulators due to leakage and quiescent currents⁷ :

$$P_{tot} = P_{lost} + P_{thermal} + P_{max-useful} \quad (4.12)$$

where P_{tot} is the total power provided by the regulators, P_{lost} is the power lost in the regulators due to leakage and quiescent current and $P_{thermal}$ is the power lost in the regulators due to thermal dissipation.

The total power consumed by the LDO can be estimated with the following equation:

$$P_{tot} = V_{in} \cdot I_{in} = V_{in} \cdot (I_{out} + I_{lost}) \quad (4.13)$$

where V_{in} is the input voltage of the regulator, I_{in} is the current provided by the power supply to the regulator, I_{out} is the current provided by the regulator to the load and I_{lost} is the current lost in the regulator due to leakage and quiescent current of the regulator.

This current lost can be neglected as it is much smaller than the current provided to the load, so we have $I_{in} \approx I_{out}$ and give the final equation :

$$P_{tot} \approx V_{in} \cdot I_{out} \quad (4.14)$$

The maximal total power consumed by the LDO can be estimated from the input voltage and the maximum current that can be provided by the regulator

⁷Quiescent currents are the currents consumed by the regulator itself to operate, even when there is no load connected to the output.

in normal conditions, which is 200 mA for the ADP7118 and ADP7182 circuits. The estimation of the total power consumed by the regulators is shown in Table 4.10.

The useful power delivered by the regulators is the maximal power available for the active components of the circuit, so it can be estimated knowing the output voltage of the regulators and the maximum current that can be provided to the load by the regulators, which is 200 mA for the ADP7118 and ADP7182 circuits.

$$P_{max-useful} = V_{out} \cdot I_{out-max} \quad (4.15)$$

The estimation for the maximum useful power provided by the regulators is also shown in Table 4.10.

LDO model	V _{in} (V)	V _{out} (V)	I _{out-max} (mA)	P _{tot} (W)	P _{useful} (W)
ADP7118	6	5	200	1.2	1
ADP7118	6	5	200	1.2	1
ADP7118	6	3.3	200	1.2	0.66
ADP7182	-6	-3.3	-200	1.2	0.66
Total	-	-	-	4.8	3.22

Table 4.10: Estimation of the maximal total and useful power of the regulators.

The maximum total power consumed by the regulators is estimated to be around 4.8 W, which is the estimate of the maximum total power consumed by the whole circuit. The total maximum useful power provided by the regulators is estimated to be around 3.22 W, which corresponds to the maximum power available for the active components.

The power consumed by the active components of the circuit can be estimated from the supply voltage and the maximum supply current of each component. It follows the relation:

$$P_{tot} = V_{supply} \cdot I_{supply} = (V_+ - V_-) \cdot I_{supply} \quad (4.16)$$

where P_{tot} is the total power consumed by the active component, V_{supply} is the supply voltage of it, so the output voltage of the regulator that supply it, V_+ and V_- are the positive and negative supply voltages respectively and I_{supply} is the supply current of the component. The power budget of the amplifiers is estimated in Table 4.11.

Component	V_{supply} (V)	I_{supply} (mA)	P_{tot} (mW)
LT5575	$5 - 0 = 5$	155	775
ADA4940 x 2	$5 - 0 = 5$	$46 \cdot 2$	460
LT1818 x 2	$3.3 - (-3.3) = 6.6$	$14 \cdot 2$	184.8
AD8065 x 2	$3.3 - (-3.3) = 6.6$	$6.4 \cdot 2$	84.48
Total	-	-	1504.28

Table 4.11: Estimation of the maximal power consumption for the active components.

Knowing the estimation of the power provided by the LDO and the power consumed by the active components, the power distribution for each voltage rail can be analysed. For active components which are supplied by a positive and a negative voltage, the power consumed has been equally divided between the two rails. It leads to the power distribution for each voltage rail shown in Table 4.12.

Rail voltage level	P_{useful} (W)	P_{tot} (W)
5V	1	0.775
5V	1	0.460
3.3V	0.66	$0.269/2 \approx 0.135$
-3.3V	0.66	$0.269/2 \approx 0.135$
Total	3.22	1.504

Table 4.12: Maximum power budget from regulators to active components.

We can see that the maximum power provided by the regulators is higher than the maximum power consumed by the active components, which leaves remaining useful power for the other components of the design such as LEDs, jumpers, test points and for the power losses in the regulators.

A current budget can also be estimated knowing the maximal current that can be provided by the regulators (200 mA for the ADP7118 and ADP7182) and the maximal current consumed by the active components. The current has been equally divided between the positive and negative rails for components supplied by two power rails. For each voltage rail, it is estimated as follow :

Voltage rail	Current provided (mA)	Current consumed (mA)
5V	200	155
5V	200	$46 \times 2 = 92$
3.3V	200	$(14 \times 2 + 6.4 \times 2)/2 = 20.4$
-3.3V	200	$(14 \times 2 + 6.4 \times 2)/2 = 20.4$

Table 4.13: Maximum current budget from regulators to active components.

For each voltage rail, the current provided by the regulators is higher than the current consumed by the active components, which is consistent with the power distribution results and confirms that the design has a margin for the current consumption of the other components of the design and for the current leakage in the circuit.

4.4.4 Thermal dissipation

For room temperature, thermal verification has been completed using the thermal models provided in the datasheets. For each regulator, the power dissipation in the die⁸ is given by:

$$P_D = [(V_{IN} - V_{OUT}) \cdot I_{OUT}] + (V_{IN} \cdot I_{GND}) \quad (4.17)$$

where V_{IN} is the input voltage, V_{OUT} is the output voltage, I_{OUT} is the load current and I_{GND} is the ground current. Power dissipation due to ground current such, as the leakage and quiescent current of the regulator, is quite small and can be ignored. The power dissipation is then estimated :

$$P_D = [(V_{IN} - V_{OUT}) \cdot I_{OUT}] \quad (4.18)$$

Then, the junction temperature⁹ can be estimated with the following equation:

$$T_J = T_A + (P_D \cdot \theta_{JA}) \quad (4.19)$$

where T_A is the ambient temperature and θ_{JA} is the thermal resistances between the junction and ambient air. It yields the final model:

$$T_J = T_A + \{[(V_{IN} - V_{OUT}) \cdot I_{OUT}] \cdot \theta_{JA}\} \quad (4.20)$$

Considering that the ambient temperature is $T_a = 25^\circ\text{C}$ and the thermal resistance in the worst case scenario for each regulator ($\theta_{JA} = 170^\circ\text{C/W}$ for all), it provides

⁸A die is a small piece of semiconductor material on which the circuit is fabricated.

⁹Junction temperature is the temperature of the semiconductor inside the regulator [38].

the following results in Table 4.14.

For the active components of the circuit, the power dissipation can be approximated by the total power consumed by the component, which is estimated in Table 4.11 since the power of the useful outgoing signal and the power due to leakage are negligible compared to the power consumed by the component. The thermal dissipation of the active components is shown in Table 4.15, using the specific thermal resistance θ_{JA} of each respective component's package.

Regulator	V_{IN} (V)	V_{OUT} (V)	I_{LOAD} (mA)	P_D (mW)	T_J ($^{\circ}C$)
ADP7118 (5V)	6	5	155	155	51.4
ADP7118 (5V)	6	5	76	76	37.9
ADP7118 (3.3V)	6	3.3	20.4	55	34.4
ADP7182 (-3.3V)	-6	-3.3	-20.4	55	34.4

Table 4.14: Thermal dissipation estimation of the regulators.

Component	$P_{tot} = P_D$ (mW)	θ_{JA} ($^{\circ}C/W$)	T_J ($^{\circ}C$)
LT5575	775	37	53.7
ADA4940	230	158	61.3
LT1818	92.4	250	48.1
AD8065	42.24	180	32.6

Table 4.15: Thermal dissipation estimation of the individual active components.

The maximum junction temperature recommended in the datasheets for reliable operation is $125^{\circ}C$. All the estimated junction temperatures are below this limit, which confirms that the design is thermally safe and that the regulators and the active components will operate reliably without overheating issues.

4.4.5 Final Design

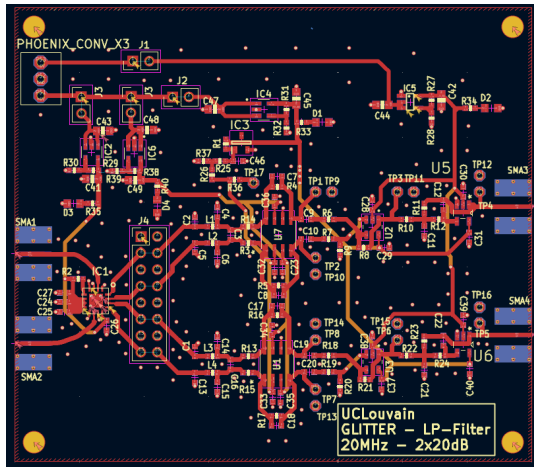
The schematic has been transformed into a PCB layout, as shown in Figure 4.8, Figure 4.9 and Figure F.1. This PCB uses a 4-layer copper stackup with the following configuration:

Layer	Function
Top	Signal and components
Layer 2	Ground plane
Layer 3	Power plane
Bottom	Signal and components

Table 4.16: Copper stackup of the PCB.

Layer 2 is a continuous ground plane to reduce noise and EMI, while layer 3 is a power plane providing a path for the supply currents and sometimes a path for the signals when the top layer is congested. Each of these layers are separated by an FR4 dielectric layer with characteristics of $\epsilon_r = 4.5$ and $\tan\delta = 0.02$.

The wires from the schematic are transformed into tracks and the components are placed on the board according to the design rules and constraints defined in the previous sections and from the manufacturing capabilities of the PCB manufacturer. The red tracks are on the top layer and the orange tracks are on the bottom layer. The red areas are the top surface ground planes and the green one is the second ground plane under the top layer.



4.5 Characterisation

The PCB was received from the manufacturer and the different components were soldered onto the board. The dimensions of the received PCB are $80\text{ mm} \times 70\text{ mm} \times 1.59\text{ mm}$, and it is pictured in Figure 4.10.

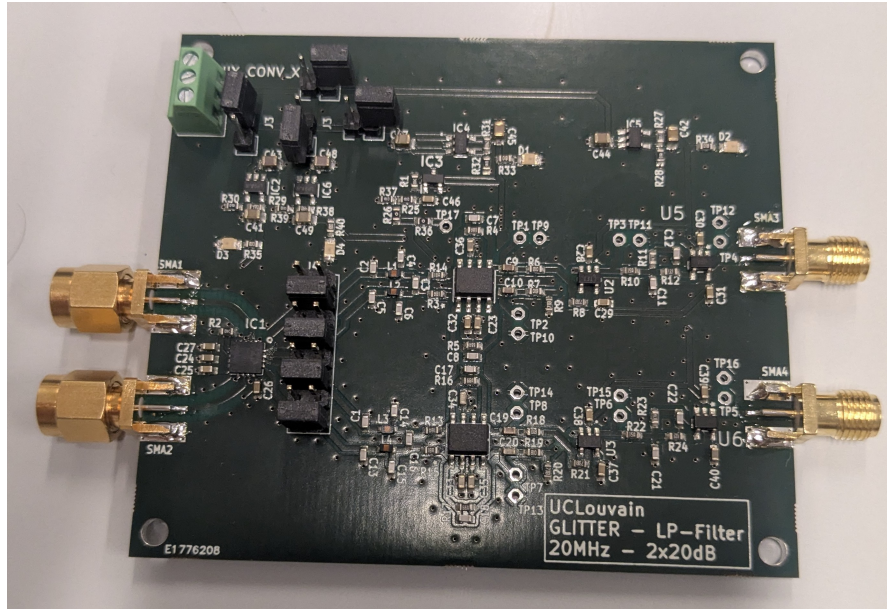


Figure 4.10: Photo of the received PCB.

It was characterized using two signal generators and a Picoscope 3405D [39] device, used as an oscilloscope.

The LO signal is generated by a signal generator which provides a sinusoidal signal at 1.57542 GHz into the LO input of the PCB.

The front-end RF signal has been generated by another signal generator, which is well synchronized with the LO signal. It provides a sinusoidal signal at frequencies from 40 Hz to 50 MHz added to the carrier frequency of 1.57542 GHz. The RF signal is swept in amplitude to verify the gain and the bandwidth of the circuit in different signal input circumstances. This signal is then injected into the RF input of the PCB.

The resulting measured signals at the output are plotted in a Bode diagram illustrated in Figure 4.11 and Figure 4.12 for a zoomed-in view of the passband region.

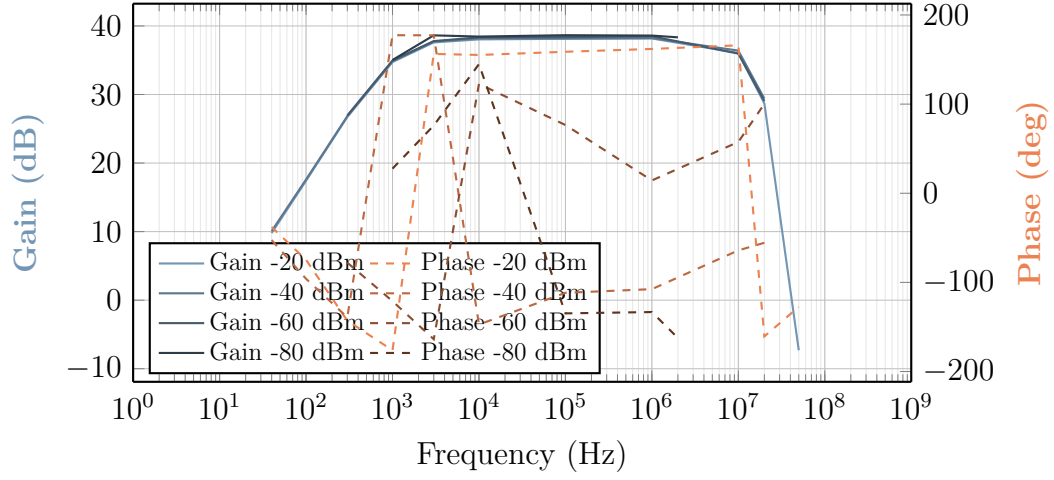


Figure 4.11: Measured Bode diagram of the complete circuit at different input power levels.

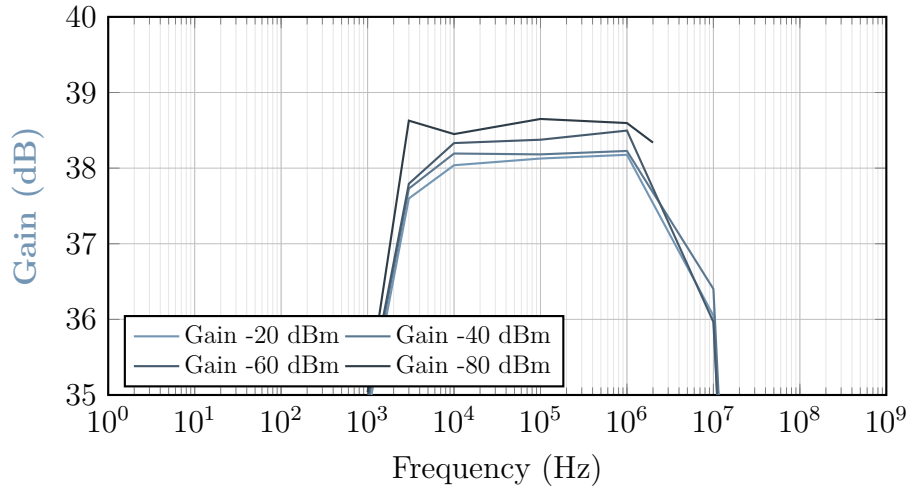


Figure 4.12: Measured Bode diagram of the complete circuit at different input power levels (Gain restricted to 35–40 dB).

The measured gain is around 38.5 dB in the passband region between the two cut-off frequencies that are 1 kHz and 20 MHz. This result is close to the design target of 40 dB, and the small difference can be explained by the tolerances of the components used in the design, the losses in the cables and the non-idealities of the amplifiers.

Moreover, we can see that the gain is higher for lower input power, which is consistent with the non-linearity of the amplifiers used in the design.

Saturation of the signal at high input power can be observed, as shown in Figure 4.13 where the input signal is around -17 dBm (which corresponds to 89.337 mV_{PP}) and the output signal is saturated at around 1800 mV_{PP} .

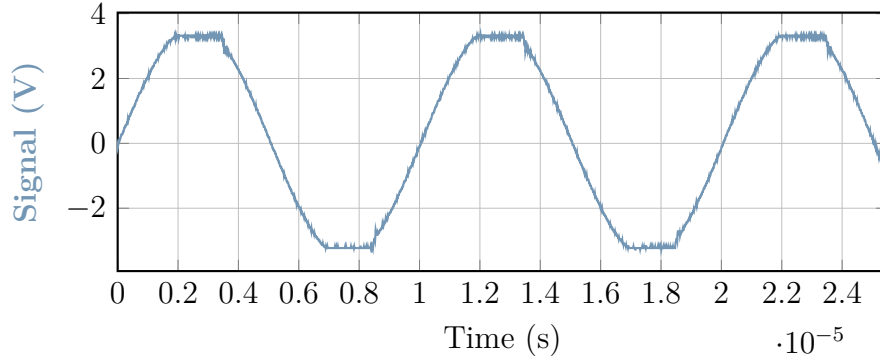


Figure 4.13: Simulated output signal at saturation for a 89.337 mV_{pp} input.

5 Characterisation of the GNSS disciplined oscillator

5.1 Objectives

The objective of this section is to evaluate the feasibility of frequency synchronization for the CubeSat formation using the u-blox EVK-M8F as a reference. This involves assessing the stability of the frequency signal provided by the u-blox module used as frequency reference for the satellite, thereby assessing whether synchronization can be maintained over time.

This GNSSDO provides a 30.72 MHz reference signal and a 1 PPS timing information output. As explained in Chapter 3, the 30.72 MHz reference is used as an input to the onboard PLL, then distributed across the CubeSat to provide the necessary clock signals, such as the LO signal for the demodulator in the baseband circuit. The analysis is conducted by characterizing short-term, mid-term, and long-term stability of the provided reference signal. The influence of the distance between the satellites' antennas is considered. The selection of the GNSS constellation was not evaluated as this functionality is limited to the generation of u-blox used, leading to the use of the GPS and Galileo constellation concurrently. The precise selection of the constellation is more fully supported with newer generations.

5.2 Methodology

The stability of the u-blox reference signal is evaluated by measuring with a Picoscope 3405D oscilloscope the 30.72 MHz output signal generated by the u-blox module from received GNSS signals. As explained in Chapter 2, this situation is representative of the intended use of the u-blox module as a GNSSDO in the CubeSat formation, where it will receive GNSS signals and provide a stable frequency reference for synchronization.

5.2.1 PicoScope 3405D

The time-domain measurements are performed with a PicoScope 3405D [39] used as an oscilloscope with the python psospa API [40]. The instrument provides 8-bit resolution at 1 GS/s with a 100 MHz analog bandwidth. Its input range spans from ± 20 mV to ± 20 V peak, and the generator output supports up to ± 5 V peak. The typical sample jitter¹ is 3 ps RMS, the typical noise level is about 160 μ V RMS on the ± 20 mV range and the specified timebase accuracy is ± 2 ppm. This timebase accuracy leads to a frequency deviation of $30.72 \times 10^6 \cdot 2 \times 10^{-6} = 61.44$ Hz at 30.72 MHz, which corresponds to a timing error of 2 μ s over one second.

5.2.2 U-blox

The signal reference is based on the u-blox EVK-M8F platform [29, 41]. The module is a GNSSDO using a Voltage-Controlled Temperature-Compensated Crystal Oscillator (VCTCXO) as a local oscillator, which is disciplined by the GNSS signal received from the antenna. The module provides a 1PPS timing output and a 30.72 MHz reference signal, corresponding to a period of approximately 32.55 ns. The theoretical frequency accuracy of the u-blox module is specified to be below 5 ppb = 5×10^{-9} in GNSS locked mode². This corresponds to a characteristic frequency deviation at 30.72 MHz of $30.72 \times 10^6 \cdot 5 \times 10^{-9} = 153.6$ mHz, which corresponds to a timing error of 5 ns over one second. This timing error is probably acceptable for the intended application, as it is much smaller (by a factor of 20000) than the 100 ms coherent acquisition window planned for the CubeSat formation. The u-blox module configuration is performed with the u-center software [42].

5.2.3 Stability metrics

A first metric to analyse the stability of the frequency reference is obtained by looking at the Allan Deviation (ADEV), i.e., the square root of the Allan Variance (AVAR), which quantifies frequency fluctuations over different averaging times [43, 44].

¹Jitter refers to the deviation from true periodicity of a signal, often in relation to a reference clock. It can be caused by various factors such as noise, power supply variations or temperature changes.

²GNSS locked mode means that the u-blox module is receiving and tracking signals from the GNSS satellites, allowing it to use the satellite's atomic clocks as a reference for its own frequency output.

The Allan Variance is defined as :

$$\sigma_{AVAR}^2(\tau) = \frac{1}{2} \langle (\bar{y}_{n+1} - \bar{y}_n)^2 \rangle \quad (5.1)$$

where $\bar{y}_n$ is the average fractional frequency over the n -th interval of duration τ :

$$\bar{y}_n = \frac{1}{\tau} \int_{t_n}^{t_n+\tau} y(t) dt \approx \frac{x(t_n + \tau) - x(t_n)}{\tau} \quad (5.2)$$

where $x(t)$ is the phase/time-error measurement, t_n is the start time of the n -th interval and $\langle \cdot \rangle$ denotes the expectation value.

For a finite record of M adjacent τ -averaged values, we estimate it with (fixed τ estimators) :

$$\sigma_{AVAR}^2(\tau) \approx \frac{1}{2(M-1)} \sum_{n=1}^{M-1} (\bar{y}_{n+1} - \bar{y}_n)^2 \quad (5.3)$$

Allan Deviation is given with the square root of it :

$$\sigma_{ADEV}(\tau) = \sqrt{\sigma_{AVAR}^2(\tau)} \quad (5.4)$$

A variation of the ADEV is the Overlapping Allan Deviation (OADEV), which provides a more efficient use of the data by allowing overlapping intervals. In this context, overlapping means that the intervals used to compute the average frequency can overlap with each other, allowing for more estimates to be computed from the same dataset, improving the statistical confidence of the results. The Overlapping Allan Variance (OAVAR) is defined at an averaging time of $\tau = m\tau_0$:

$$\sigma_{OAVAR}^2(\tau = m\tau_0) = \frac{1}{2m^2(M-2m+1)} \sum_{n=1}^{M-2m+1} \left\{ \sum_{i=n}^{n+m-1} [y_{i+m} - y_i] \right\}^2 \quad (5.5)$$

where y_i is the i -th fractional frequency measurement, τ_0 is the sampling interval, m is the averaging factor and M is the total number of frequency measurements. It can be estimated from a set of $N = M + 1$ time measurements as :

$$\sigma_{OAVAR}^2(m\tau_0) = \frac{1}{2(m\tau_0)^2(N-2m)} \sum_{n=1}^{N-2m} (x_{n+2m} - 2x_{n+m} + x_n)^2 \quad (5.6)$$

where x_n is the n -th phase/time-error measurement, N is the number of phase/time-error samples, and $N = M + 1$.

The OADEV is given with the square root of it :

$$\sigma_{OADEV}(m\tau_0) = \sqrt{\sigma_{OAVAR}^2(m\tau_0)} \quad (5.7)$$

The overlapping version of the ADEV provides superior performance over the non-overlapping estimators as m rises and the time-series is of moderate length.

The last metric used to evaluate the stability of the frequency reference is the Time Interval Error Root Mean Square (TIE_{RMS}) and the Maximum Time Interval Error (MTIE) [45, 46, 47].

$$TIE_{RMS}(\tau) = \sqrt{\frac{1}{N-m} \sum_{n=1}^{N-m} (x_{n+m} - x_n)^2} \quad (5.8)$$

$$MTIE(\tau) = \max_{1 \leq i \leq N-m} \left(\max_{i \leq j \leq i+m} x_j - \min_{i \leq j \leq i+m} x_j \right) \quad (5.9)$$

where x_n is the n -th phase measurement, $\tau = m\tau_0$ is the observation time, m is the number of samples in each observation interval, and N is the total number of samples. They provide a measure of the time error of the signal, with TIE_{RMS} quantifying the root mean square of the time-error variation over an interval τ and MTIE quantifying the maximum peak-to-peak time-error variation observed over any interval of duration τ .

5.3 Experimentation

As the Picoscope has a lower theoretical time stability than the u-blox reference, another time reference than the internal clock of the picoscope is needed to evaluate the stability of the u-blox signal. This is why the measurement campaign has been carried out using two u-blox units, one as a reference and the other as a Device Under Test (DUT). The ADEV and the TIE_{RMS} of one u-blox unit is then estimated by taking the measured ADEV divided by $\sqrt{2}$ while the MTIE is divided by 2. Each campaign has been made in a relatively open outdoor environment with globally clear-sky conditions and the Picoscope and the u-blox modules have been running for one hour before the measurement in order to ensure stable operating conditions.

The setup for this measurement is illustrated in Figure 5.1.

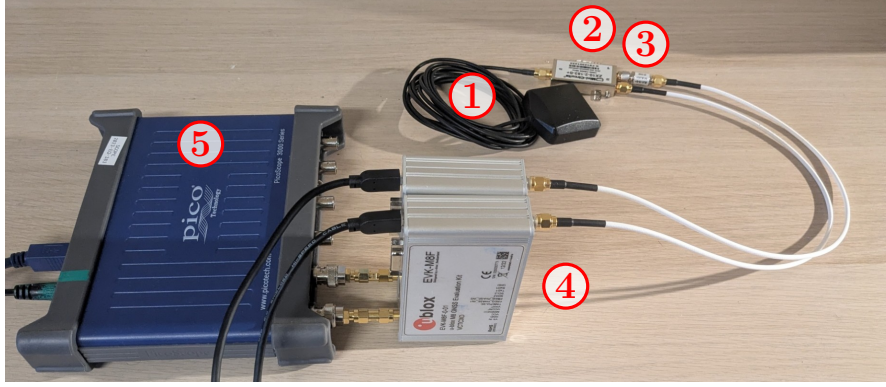


Figure 5.1: Measurement setup for the u-blox with common antenna configuration.

The GNSS signal is captured by the u-blox antenna (1), then, if the common antenna configuration is used, it is split with a ZX10-2-183-S+ power splitter (2) [48] which covers 1500 to 18000 MHz and supports DC pass-through. Supporting DC pass-through means that the splitter allows the DC component of the signal to pass through, which is important for supplying the antenna, but this antenna only requires one power source (here one u-blox module). This is why when the power splitter is employed, a DC block (3) is added to prevent the DC pass-through of the two u-blox modules from interfering with each other. This signal then enters the u-blox units (4), which processes it and outputs the 30.72 MHz reference signal sampled by the Picoscope (5) and then sent to the computer by data streaming.

The 30.72 MHz reference signal is sampled by the Picoscope at 125 MS/s more than three times the reference signal frequency. It ensures the respect of the Nyquist-Shannon sampling theorem³ and allows one to recover the measured signal frequency, as illustrated in Figure 5.2.

³The Nyquist-Shannon sampling theorem states that a band-limited baseband signal can be accurately reconstructed from its samples if the sampling rate is at least twice the bandwidth of the signal.

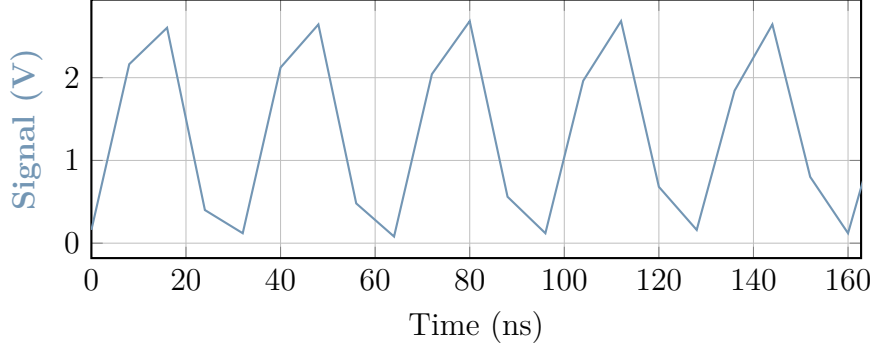


Figure 5.2: Sampled 30.72 MHz reference signal from the u-blox module.

The signal exhibits a clear sinusoidal waveform with a period of approximately 32.55 ns, corresponding to the expected frequency of 30.72 MHz. The number of acquired samples was limited to avoid saturation of the memory of the oscilloscope or of the computer. The code used to process the sampled signal is available in Appendix D.

5.3.1 U-blox with common antenna

A first measurement campaign has been performed using two u-blox units with a shared antenna path, as illustrated in Figure 5.1. This configuration allows to evaluating the stability of the u-blox reference signal while minimizing the influence of external factors such as the variability of the received GNSS signal, as both u-blox units are fed with the same signal from the antenna. The ADEV and the OADEV of this measurement campaign are presented in Figure 5.3.

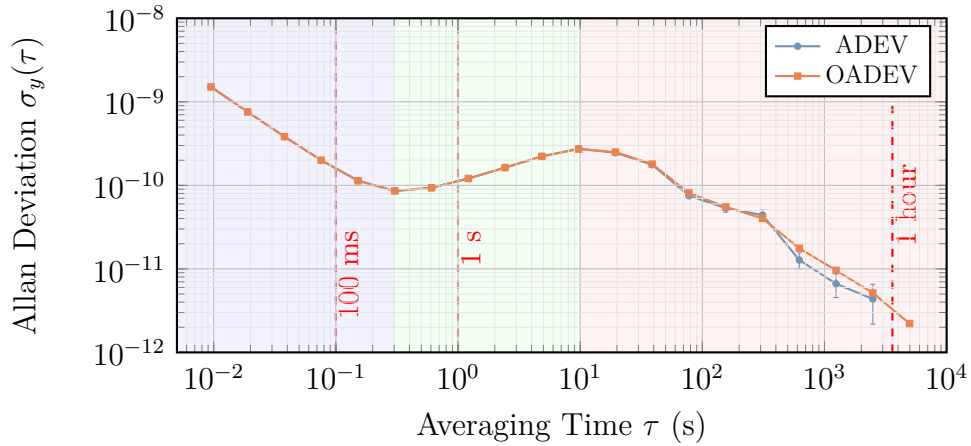


Figure 5.3: Allan deviation of the u-blox signal with a shared antenna path.

We can observe three different regimes in the Allan deviation curve :

- (i) For $\tau < 0.3$ s, the ADEV decreases with a slope of approximately -1 on the log-log plot, which is characteristic of white phase noise. This indicates that at short averaging times, the frequency stability is dominated by the local oscillator of the u-blox.
- (ii) For $0.3 \text{ s} < \tau < 10$ s, the ADEV increases with a slope of approximately $+0.2$ on the log-log plot. This is due to the locked system behaviour of the GNSSDO, where the system switches from relying on the short-term stability of the local oscillator in the u-blox module to the long-term stability of the external reference, the GNSS signal.
- (iii) For $\tau > 10$ s, the ADEV decreases again with a slope of approximately -1 on the log-log plot, which is characteristic of white phase noise. This indicates that at long averaging times, the frequency stability is dominated by the GNSS signal, which is expected to be more stable than the local oscillator, as GNSS satellites rely on a highly stable atomic clock⁴.

At $\tau = 1$ s, the ADEV is approximately 1×10^{-10} , which is consistent with the theoretical frequency accuracy of the u-blox module specified to be below $5 \text{ ppb} = 5 \times 10^{-9}$ in GNSS locked mode.

At $\tau = 100 \text{ ms} (= 10^{-1} \text{ s})$, the ADEV is approximately 2×10^{-10} . It means that the frequency stability of the 30.72 MHz u-blox signal at an averaging time of 100 ms can produce a characteristic frequency deviation of $30.72 \times 10^6 \cdot 2 \times 10^{-10} = 6.144 \text{ mHz}$.

The OADEV provides a more efficient use of the data by allowing overlapping intervals, resulting here in a poorer Allan Deviation value at longer averaging times compared to the non-overlapping version. Furthermore, the OADEV extends the valid range of analysis, as the overlapping method extracts more comparison samples for the estimation at large values of τ than the non-overlapping method.

The TIE_{RMS} and MTIE of this measurement campaign are shown in Figure 5.4.

⁴It has been shown that the frequency stability of the Galileo satellite clocks at 10^3 s and 10^4 s averaging intervals reaches the 10^{-14} level [49].

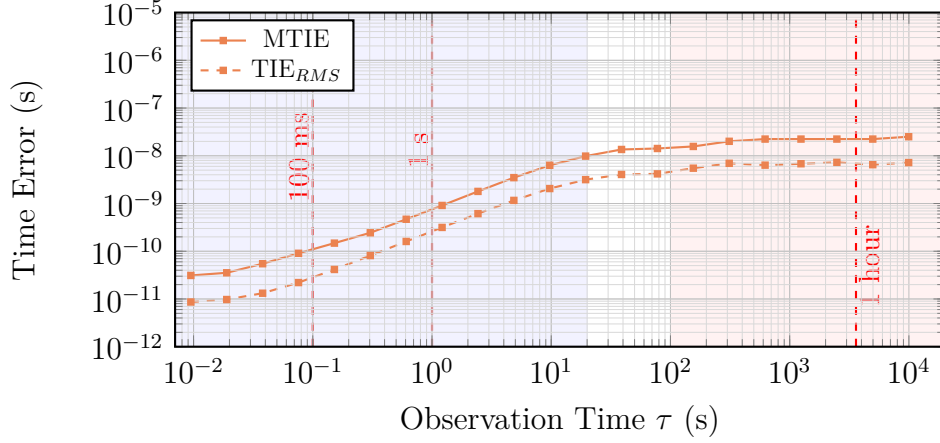


Figure 5.4: TIE_{RMS} and MTIE of the u-blox reference signal with a shared antenna path.

Two different regimes in the MTIE and TIE_{RMS} curves can be observed :

- (i) For $\tau < 20$ s, the TIE_{RMS} and MTIE increase with a slope of approximately +1 on the log-log plot, which is due to the drift of the local oscillator of the u-blox module.
- (ii) For $\tau > 100$ s, the TIE_{RMS} and MTIE stop increasing and start to stabilize at around 2×10^{-8} s, which explains that the u-blox is locked in the GNSS signal which relies on a highly stable atomic clock.

At $\tau = 1$ s, the MTIE is approximately 8×10^{-10} s, which means that the Maximum Time Interval Error at $\tau = 1$ s is around 0.8 ns. The TIE_{RMS} at $\tau = 1$ s is approximately 3×10^{-10} s, meaning that the Time Interval Error Root Mean Square at $\tau = 1$ s is around 0.3 ns. This confirms that the TIE_{RMS} and the MTIE of the u-blox reference signal at an observation time of 1 s in GNSS locked mode are below the specified theoretical limit of 5 ns.

At $\tau = 100$ ms, the MTIE is approximately 1×10^{-10} s = 100 ps and the TIE_{RMS} is approximately 3×10^{-11} s = 30 ps. It means that the 30.72 MHz u-blox reference signal at an observation time of 100 ms has a typical frequency variation error of $30.72 \times 10^6 \cdot \frac{3 \times 10^{-11}}{100 \times 10^{-3}} = 9.22$ mHz and a maximum frequency variation error of $30.72 \times 10^6 \cdot \frac{1 \times 10^{-10}}{100 \times 10^{-3}} = 30.72$ mHz.

The same order of magnitude and shape for the ADEV, TIE_{RMS} and MTIE curves have been observed with a different measurement setup and a different GNSSDO in the work of Bauer et al. [50], which confirms the consistency of the results obtained in this measurement campaign.

5.3.2 U-blox with different antennas

The second measurement campaign has been performed using two u-blox units with different antennas, which allows one to evaluate the stability of the u-blox reference signal while considering the influence of external factors such as the variability of the received GNSS signal. This configuration is more representative of the upcoming experimentation where two receiving chains are placed on two different drones, each one with its own antenna.

The ADEV and the OADEV of this measurement campaign are presented in Figure 5.5.

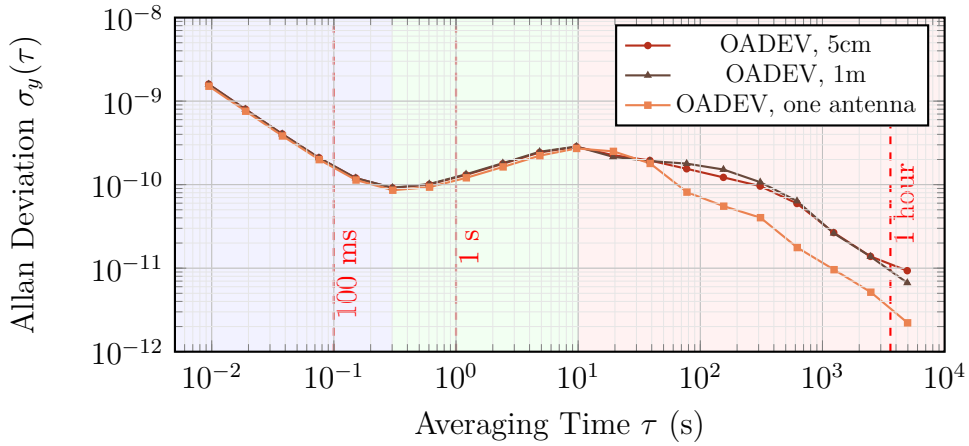


Figure 5.5: OADEV of the u-blox signal with different antennas spaced by 5cm and 1m compared to the single antenna case.

The same three different regimes in the OADEV curve can be observed as in the single antenna case, but with a higher OADEV level at averaging times greater than 40s. This is due to the fact that with different antennas, the variability of the received GNSS signal can influence the stability of the u-blox reference signal. The OADEV levels for the two-antennas configuration (1 m and 5 cm) are globally the same. Other distances of interest between the two antennas could be tested in order to evaluate the influence of the antenna spacing on the stability of the u-blox reference signal.

The TIE_{RMS} and MTIE of this measurement campaign are shown in Figure 5.6.

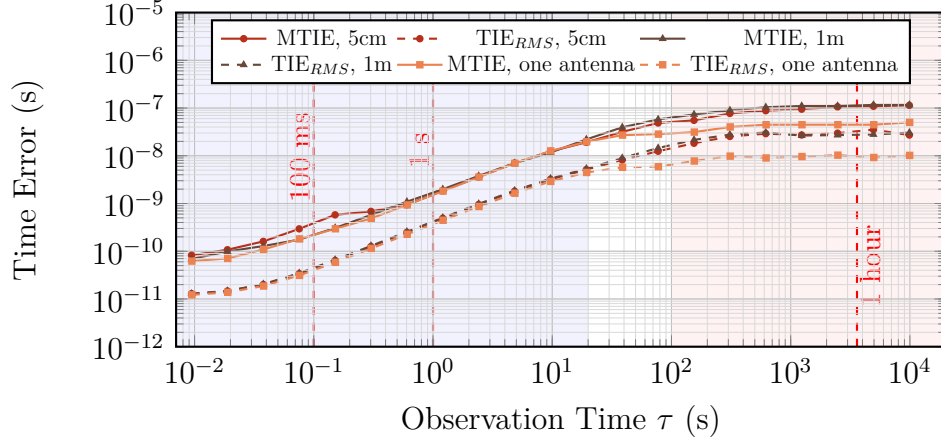


Figure 5.6: TIE_{RMS} and MTIE of the u-blox reference signal with different antennas spaced by 5cm and 1m.

As in the OADEV graph, the TIE_{RMS} and MTIE curves for the two antennas configuration (1 m and 5 cm) are globally at the same level, but with a higher level at observation times greater than 40 s compared to the single antenna case.

5.4 Results summary

In summary, the stability of the u-blox reference signal has been evaluated with two different measurement campaigns, one with a shared antenna between two u-blox receivers and one campaign with two different antennas for the two u-blox receivers.

The results show that the u-blox reference signal exhibits a good stability with an Allan Deviation of approximately 1×10^{-10} at an averaging time of 1 s, which is consistent with the ≤ 5 ppb frequency accuracy specified by the manufacturer. For an averaging time of 100 ms considered in the two drones experimentation, the ADEV is approximately 2×10^{-10} , which corresponds to a characteristic frequency deviation of 6.144 mHz at 30.72 MHz.

The TIE_{RMS} and MTIE of the u-blox reference signal are also below the specified theoretical limit of 5 ns at an observation time of 1 s. For the observation time of 100 ms, the TIE_{RMS} is approximately 30 ps and the MTIE is approximately 100 ps, which corresponds to a typical frequency variation error of 9.22 mHz and a maximum frequency variation error of 30.72 mHz at 30.72 MHz.

The stability of the u-blox reference signal is influenced by the variability of the received GNSS signal, as shown by the higher OADEV, TIE_{RMS} and MTIE levels observed in the two antennas configuration compared to the single antenna case. This highlights the importance of considering the influence of external factors such as the variability of the received GNSS signal.

5.5 PLL

The generated signal is used as a reference for the onboard Phase-Locked Loop (PLL), which is responsible for distributing the clock signal at different frequencies across the satellite. The PLL configuration must ensure that the output clocks maintain the required stability and phase noise characteristics for the synchronization needs. Good perspective about the stability of the PLL performance can be obtained by analyzing the phase jitter of the output clocks, which is a measure of the short-term variations in the phase of the clock signal. The manufacturer of the proposed PLL, the Si5344H-EB, specifies a maximum Root Mean Square (RMS) phase jitter of 0.140 ps for a frequency range of 12 kHz to 20 MHz. At 30.72 MHz input clock frequency, this short-term time fluctuation represents a relative variation of only $0.14 \times 10^{-12} \cdot 30.72 \times 10^6 = 4.3 \times 10^{-6}$ of the signal period (an RMS phase error of $2\pi(\frac{180}{\pi}) \cdot 4.3 \times 10^{-6} = 0.0015^\circ$) [51].

PLL configuration and clock-tree exploration should be carried out with the Skyworks ClockBuilder Pro environment [52] and linked to this Allan deviation analysis in a further work.

6 Patch antenna design

6.1 Objectives

The objective of this part of the thesis is to design and validate two patch antennas dedicated to GNSS-R reception in the target frequency band. A trimmed-patch topology (patch with opposite corner truncations) is selected to generate circular polarization (one LHCP and one RHCP) due to its manufacturing simplicity and low cost. The main design specifications are summarized in Table 6.1.

Antenna type	Trimmed patch (truncated corners)
Center frequency	1.57542 GHz
Target bandwidth	20 MHz
Polarization	Circular
Axial ratio	RHCP for direct signal, LHCP for reflected signal
Substrate	FR4 - IS400 ($\epsilon_r = 4.35$, $h = 1.6$ mm, $\tan\delta = 0.022$)

Table 6.1: Target specifications for the patch antenna design.

6.2 Design

The electromagnetic design is carried out in CST Studio Suite [53], following the methodology presented in the CST tutorial from A. Abazi [54]. The design workflow includes geometry definition, substrate parameterization, feed adjustment, polarization and matching tuning to meet the specification targets.

6.2.1 Dimensions calculation

Initial patch dimensions are derived from the analytical formulas for trimmed patch antennas provided by A. Balanis in ANTENNA THEORY [55, p. 820–821]. This yields the following equations for the patch dimensions :

$$W = \frac{c}{2f} \sqrt{\frac{2}{\varepsilon_r + 1}} \quad (6.1)$$

$$\varepsilon_{\text{eff}} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left(1 + 12 \frac{h}{W} \right)^{-\frac{1}{2}} \quad (6.2)$$

$$\Delta L = 0.412 h \frac{(\varepsilon_{\text{eff}} + 0.3) \left(\frac{W}{h} + 0.264 \right)}{(\varepsilon_{\text{eff}} - 0.258) \left(\frac{W}{h} + 0.8 \right)} \quad (6.3)$$

$$L_{\text{eff}} = \frac{c}{2f \sqrt{\varepsilon_{\text{eff}}}} \quad (6.4)$$

$$L = L_{\text{eff}} - 2\Delta L \quad (6.5)$$

where all symbols are defined in Table 6.2 :

Symbol	Definition	Value
c	Speed of light	$\approx 3 \times 10^8$ m/s
f	Center frequency	1.5 GHz
ε_r	Relative permittivity of the substrate	4.35
h	Substrate thickness	1.6 mm
W	Width of the patch	58.215 mm
L	Length of the patch	45.378 mm
ε_{eff}	Effective permittivity	4.128
ΔL	Length extension due to fringing fields	0.743
t	Corner truncation size	4.538 mm

Table 6.2: Symbols, definitions and values for the patch antenna design equations.

The corner truncation size is then estimated using the empirical relation :

$$t \approx 0.1 L \quad (6.6)$$

By truncating two opposite corners of a square patch, it creates a small perturbation that separates the degenerate orthogonal modes of the patch (TM₁₀ and TM₀₁), resulting in two resonant frequencies with nearly equal amplitudes and a phase difference of approximately 90°, which enables circular polarization. In order to obtain a circular polarisation, the patch should be near to a square patch, so we set :

$$W = L = 45.378 \text{ mm} \quad (6.7)$$

The website "pasternack.com" [56] has been used for dimensions validation and gives the following results which are consistent with the above calculations :

Parameter	Value
W	58.17 mm
L	45.35 mm

Table 6.3: Patch dimensions calculated using the Pasternack calculator.

These closed-form expressions provide the starting point for corner-trim sizing and matching optimization. It yields the final design after optimization (see Section E.2), illustrated in Figure 6.3.

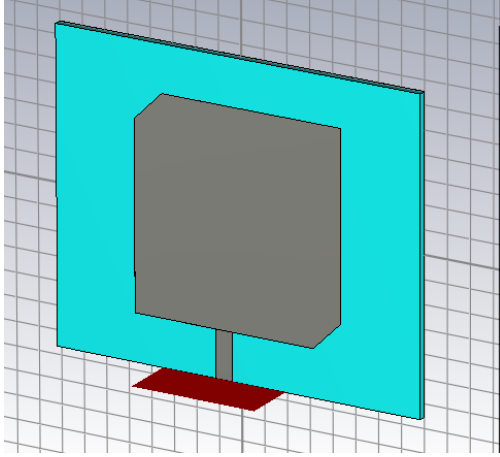


Figure 6.1: LHCP patch antenna.

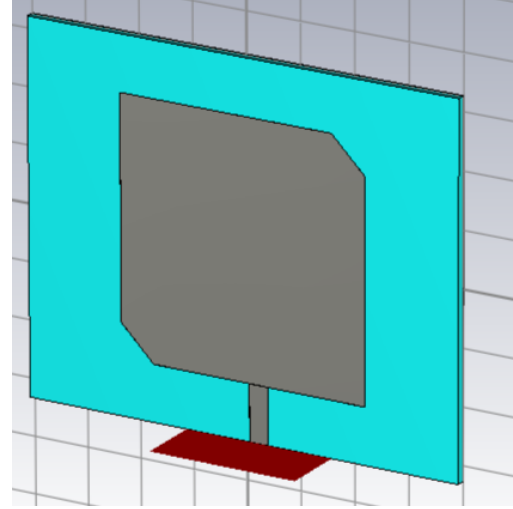


Figure 6.2: RHCP patch antenna.

Figure 6.3: Final geometries of the LHCP and RHCP corner trimmed patch antennas based on analytical calculations.

6.3 Simulation

Circuit-level validation and matching refinement are performed with CST [53] and compared to ADS [57]. The combined CST-ADS workflow is used to cross-validate that the antenna matching and polarization objectives are met. In the following subsections, the final results obtained from CST are presented and discussed, while the comparison with ADS is provided in section E.6.

The antenna's performances have been assessed through different simulated figures of merit such as matching (S-parameters), Impedance (Z-parameters), gain, axial ratio, directivity, efficiency and Left-Hand/Right-Hand Circular Polarization (LHCP/RHCP) gain across the frequency band of interest.

6.3.1 S-Parameters

The scattering parameters¹ are defined by S_{11} as there is only one port at the antenna. The S-parameters are normalized to a reference impedance of around $50\ \Omega$, which depends on the geometry of the feed structure, the permittivity of the substrate and on the frequency. This reference impedance is not a fixed ideal load, but rather the analytically calculated line impedance of the feed structure at the port and is shown in Figure 6.4.

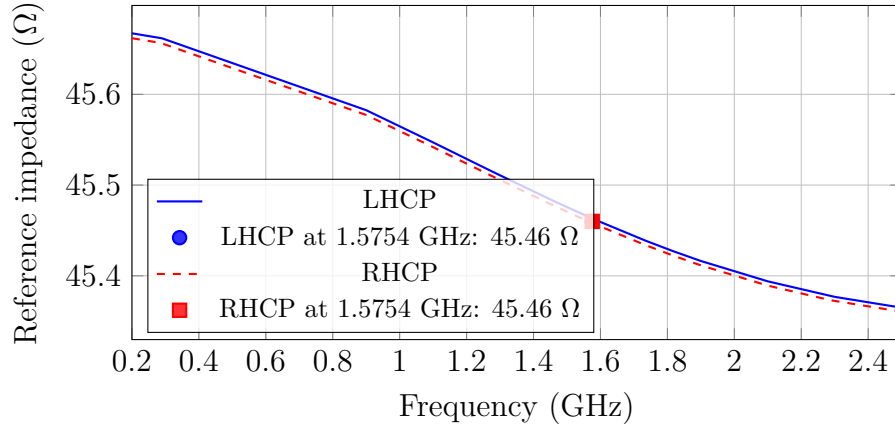


Figure 6.4: Reference impedance used for the LHCP and RHCP antennas.

A well-matched antenna will have a low S_{11} (e.g., below $-10\ \text{dB}$) across the target bandwidth, indicating that most of the power is being radiated by the antenna rather than reflected back to the source (generator). The variation of the S_{11} parameter with respect to the frequency is shown in Figure 6.5 for both the LHCP and RHCP antennas.

¹Scattering parameters describe how high-frequency electrical signals travel through and reflect off circuits. They are defined as $S_{ij} = b_i/a_j$ where a_j is the incident wave at port j and b_i is the reflected wave at port i [58].

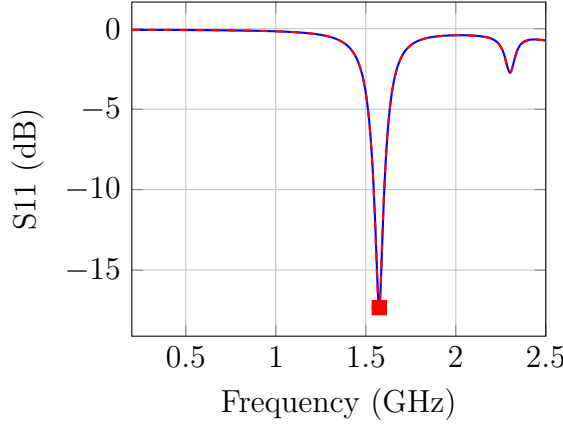


Figure 6.5: Reflection coefficient of the LHCP and RHCP antennas across 0.2 to 2.5 GHz.

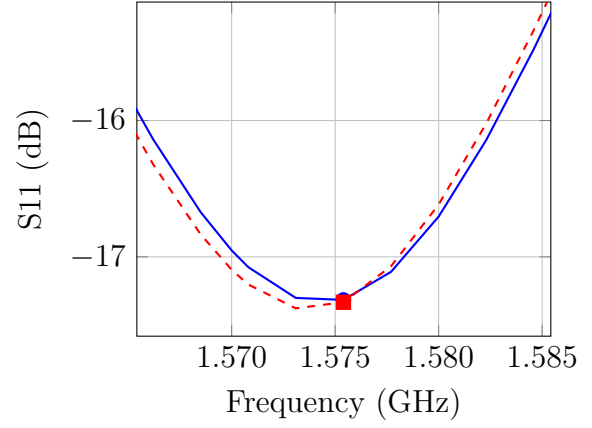


Figure 6.6: Zoomed-in view of S_{11} LHCP and RHCP across the frequency band of interest.

—●— LHCP -■- RHCP

For both antennas, the S_{11} parameter is around -17.33 dB at 1.575 GHz, which indicates that little energy is reflected back to the source, so we can say that the antenna is well matched at the center frequency.

Moreover, we can zoom in the frequency band of interest ($1.57542 \text{ GHz} \pm 10 \text{ MHz}$) to check that the S_{11} is below -10 dB across the whole bandwidth. The results are shown in Figure 6.6 and confirm that both antennas have a minimum margin of 5 dB below the -10 dB threshold across the target bandwidth. This is not enough regarding the uncertainty of the substrate permittivity (ϵ_r) and the manufacturing tolerances, but this is a good starting point and will be improved in a future version with a different substrate.

6.3.2 Z-Parameters

The input impedance of the antenna is evaluated through the Z-parameters. A good impedance match is typically indicated by a real part of Z_{11} close to the characteristic impedance of the feed line, here in CST 50 ohms, and a small (close to zero) imaginary part across the target bandwidth.

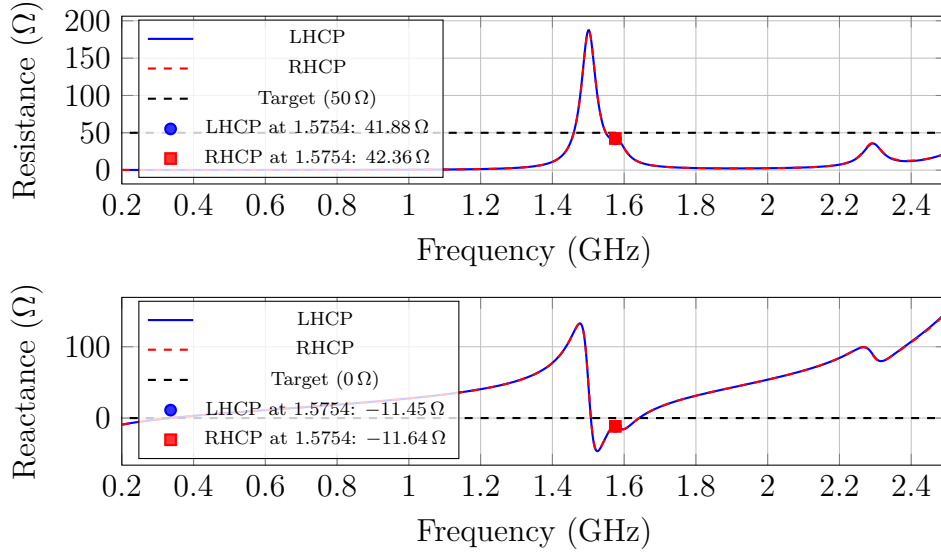


Figure 6.7: Input impedance of the LHCP and RHCP antennas across the frequency band of interest.

The real part of Z_{11} is around 50Ω and the imaginary part is close to zero at the center frequency for both antennas, which is consistent with the relative good matching observed in the S_{11} parameter.

6.3.3 Farfield Gain

The farfield absolute gain of the antenna is a measure of its ability to amplify the incoming radiated power in a specific direction. It is evaluated at the center frequency of 1.57542 GHz across different incident angles to confirm that the antenna provides a radiation pattern looking in the normal incidence direction and a sufficient gain. The farfield gain is shown in Figure 6.8 and the radiation pattern in polar coordinates is shown in Figure 6.9 for both the LHCP and RHCP antennas.

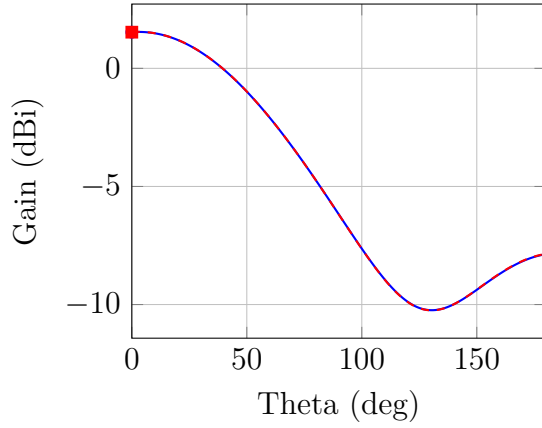


Figure 6.8: Farfield gain of the LHCP and RHCP antennas across different incident angles.

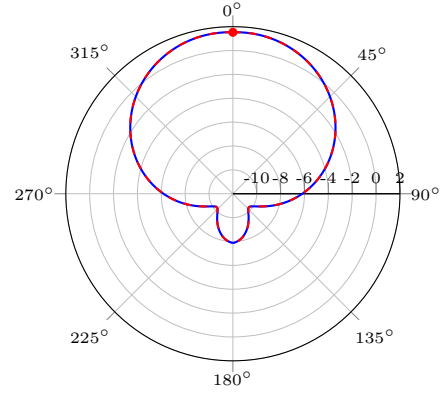


Figure 6.9: Polar radiation pattern of the antennas.



The farfield absolute gain in the normal incidence direction is around 1.5 dB for both antennas and starts to decrease as the incident angle increases, until reaching around -8 dB at 180 degrees. This indicates that the antenna will have a larger receiving cross-section in the normal incidence direction.

6.3.4 Axial Ratio

The Axial Ratio (AR) is defined as [59] :

$$AR = \frac{E_{\max}}{E_{\min}} \quad (6.8)$$

Where $E_{\max}$ and $E_{\min}$ are the maximum and minimum electric field strengths of the antenna in orthogonal polarizations, respectively.

An AR of 1 (0 dB) indicates perfect circular polarization, while higher values indicate increasing ellipticity. The AR evaluated across different angles of incidence is shown for both antennas in Figure 6.10 and Figure 6.11 with a zoomed-in view around the normal incidence angle.

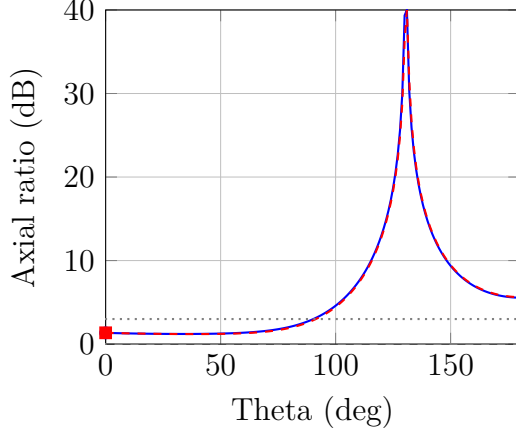


Figure 6.10: AR of the LHCP and RHCP antennas across different incident angles.

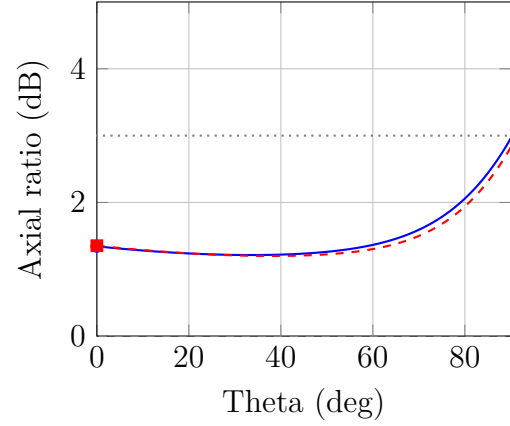


Figure 6.11: Zoomed-in view of the AR of the LHCP and RHCP antennas across different incident angles.

—●— LHCP -■- RHCP --- Ideal AR = 0 dB Limit AR = 3 dB

The AR is around 1.17 (1.36 dB) at the normal incidence angle for both antennas, which is close to the ideal value of 1 and far below the 3 dB threshold representing a degradation by a factor of 1.41 in the circular polarization quality. We could expect to observe a purely linear polarization (and thus an infinite AR) at exactly $\theta = 90^\circ$, relying on the assumption of an infinite ground plane. This mathematically forces the tangential electric field to zero at the horizon. However, in this practical design, the finite size of the substrate and ground plane introduces edge diffraction. Consequently, this transition point is pushed further around $\theta \approx 130^\circ$ rather than at 90° . However, since the antenna's gain is extremely low at such wide angles, the AR value becomes less meaningful in practice.

6.3.5 LHCP/RHCP gain

The LHCP and RHCP gains are evaluated to confirm that the antennas provide the desired circular polarization characteristics. A well-designed LHCP antenna should exhibit a significantly higher gain in the LHCP mode compared to the RHCP mode, and vice versa for the RHCP antenna.

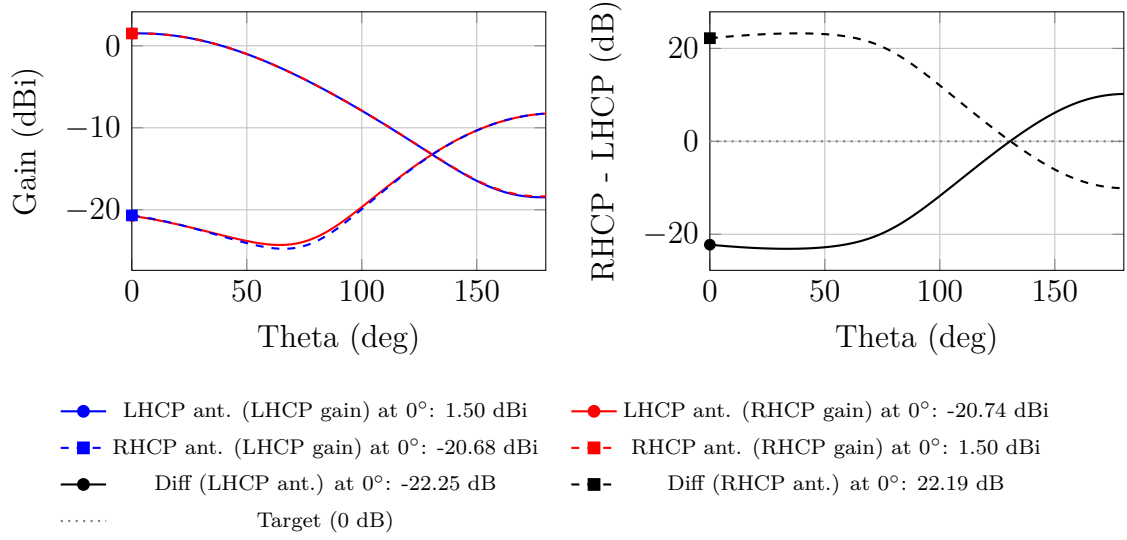


Figure 6.12: Comparison of LHCP and RHCP gain for the LHCP and RHCP antennas across angles of incidence.

The LHCP gain in the normal incidence angle of the LHCP antenna is around 1.5 dB, while its RHCP gain is around -21 dB. This corresponds to a polarization discrimination of about 20 dB between the desired and undesired polarizations. But with an incident angle of about 180 degrees, the LHCP gain drops to about -17 dB while the RHCP gain increases to about -7 dB. It indicates a change of polarization (left to right and right to left) at the horizon, but these back side gains are still low compared to the front side ones. The same observation can be made for the RHCP antenna but with the opposite polarization.

6.3.6 Efficiency

The efficiency of the antenna is a measure of how effectively it converts input power into radiated power. In CST, two types of efficiency are calculated, the radiation efficiency and the total efficiency. The total efficiency is described in Appendix E, leading to the same matching conclusions as covered in Section 6.3.1. The radiation efficiency accounts for losses due to the antenna's materials and design, and is defined as follows :

$$\eta_{\text{rad}} = \frac{P_{\text{rad}}}{P_{\text{accepted}}} \quad (6.9)$$

where P_{rad} is the power radiated by the antenna and P_{accepted} is the power accepted by the antenna (after accounting for mismatch losses). This efficiency metric is evaluated across different frequencies and compared between the LHCP and RHCP antennas as shown in Figure 6.13.

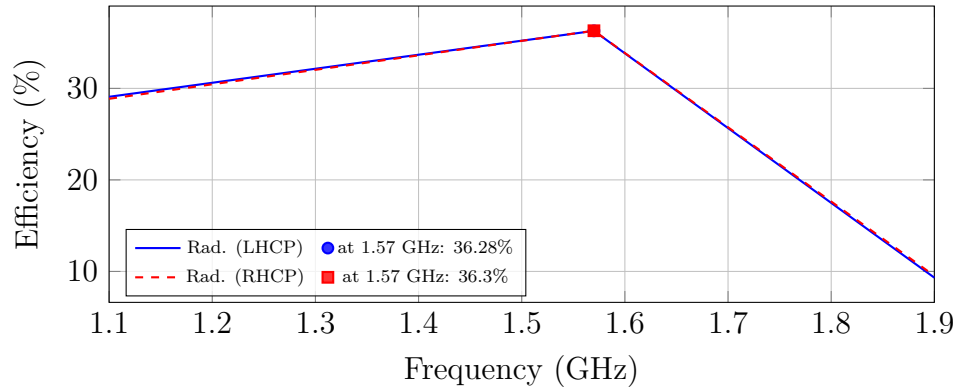


Figure 6.13: Efficiency of the LHCP and RHCP antennas across the frequency band of interest.

The radiation efficiency at the center frequency is around 36% for both antennas, which is relatively low and can be attributed to the use of a low-cost substrate with a relatively high loss tangent. This means that a significant portion of the input power is lost within the antenna materials rather than being radiated.

6.4 Characterisation

The manufactured antennas are illustrated in Figure 6.14 and Figure 6.15 and the S11 parameter is compared between simulation and measurement for the RHCP antenna in Figure 6.16.

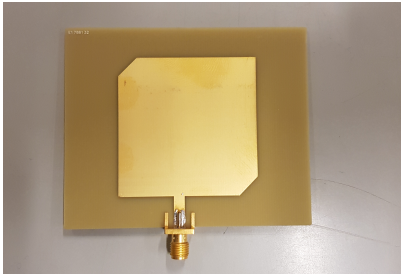


Figure 6.14: Manufactured LHCP antenna.

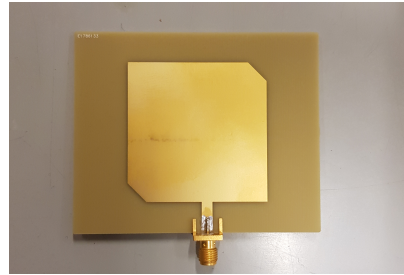


Figure 6.15: Manufactured RHCP antenna.

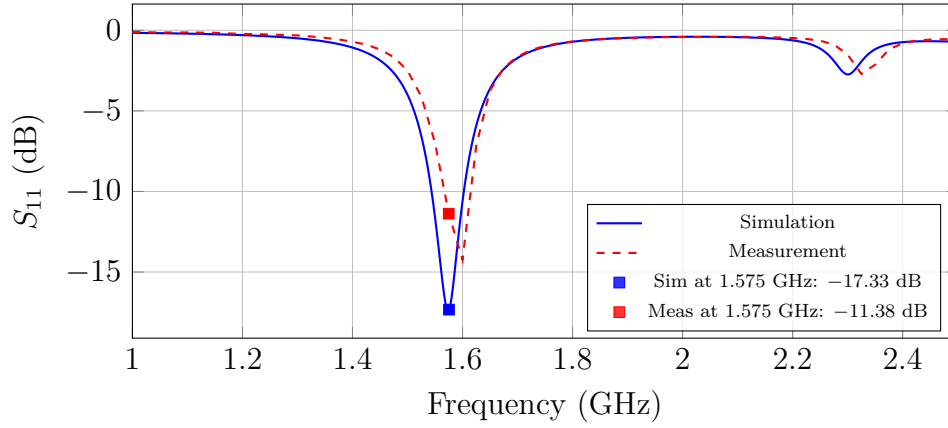


Figure 6.16: Comparison of simulated and measured S_{11} parameter for the RHCP antenna.

We can observe a shift of the measured S_{11} curve towards higher frequencies compared to the simulation, which can be attributed to fabrication tolerances, variations in the substrate properties, and measurement setup imperfections. It results in a measured S_{11} of around -11.38 dB at the center frequency of 1.575 GHz.

Other experimental characterizations should be conducted in an anechoic chamber using a Vector Network Analyzer (VNA) in order to measure the S-parameters, gain, axial ratio, and other relevant performance metrics of the two fabricated antennas. The measured results should be then compared with simulation results.

However, another antenna design with a better substrate and a feed centered on the antenna patch is in preparation, which will be fabricated and characterised in the next phase of the project, so the experimental validation of the current design will be performed later on.

7 Conclusion and perspectives

This thesis has contributed to the GLITTER project by addressing three key hardware challenges for GNSS-R Earth observation with CubeSats in formation: a low-noise baseband circuit, inter-receiver synchronization, and dedicated antenna design. The work has focused on enabling the reception and digitization of very weak direct and reflected GNSS signals, requiring a careful reception, amplification and stable timing/frequency synchronization.

First, a complete baseband amplification and filtering chain was designed to meet a 1 kHz to 20 MHz passband with approximately 40 dB of gain. The design was validated through a three-step methodology : analytical transfer-function development, LTspice simulations, and PCB fabrication followed by laboratory measurements. Experimental characterization confirmed a passband gain close to the target ($= 38.5$ dB) and the expected cutoff behavior, while highlighting the impact of component tolerance and amplifier non-idealities. Beyond frequency response, power-budget and thermal analysis were carried out to ensure integration on the intended platform, such as the drones.

Second, the feasibility of synchronizing independent receiver chains using a GNSS Disciplined Oscillator approach was investigated with u-blox EVK-M8F modules providing a 30.72 MHz reference. Stability was assessed using Allan Deviation as well as time-error metrics (TIE_{RMS} and MTIE) in two configurations: a shared-antenna setup to reduce external variability and a two-antenna setup representative of distributed receivers. The results indicate that GNSSDO-based referencing can provide the level of frequency stability required for the planned 100 ms coherent acquisition windows, while also showing that separate antennas introduce additional variability at longer observation times.

Finally, two corner-trimmed patch antennas (RHCP for the direct signal and LHCP for the reflected signal) were designed using analytical sizing and electromagnetic simulation. The simulated results demonstrate adequate matching around the GNSS-R L1/E1 band, good axial-ratio performance near boresight, and strong

polarization discrimination between LHCP and RHCP, which is essential to limit mixing between direct and reflected components.

In the short term, the next steps of the GLITTER project will focus on addressing the key questions raised in the introduction regarding the practical feasibility of GNSS-R observations using distributed low-cost platforms. This will be achieved through end-to-end validation during drone-based experiments, including simultaneous reception of direct and reflected signals, as well as the assessment of coherence after digitization and processing. In this context, particular attention will be given to the validation of inter-receiver synchronization under realistic conditions, the impact of the GNSSDO-PLL clock chain in terms of phase noise, and the experimental characterization of the two antennas. These experiments will provide the first demonstration of the system in a controlled but representative environment, linking laboratory validation to real-world deployment.

In the long term, the project aims to fully answer the initial objective of demonstrating the feasibility of GNSS-R Earth observation using a formation of CubeSats. Building upon the results obtained with drone experiments, the system will evolve towards a space based implementation. This will enable the deployment of a swarm of CubeSats capable of performing coordinated GNSS-R measurements at a larger scale, opening the way to new Earth observation capabilities based on distributed sensing.

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Microsoft Copilot with UCLouvain license was used to convert Python graphs into LaTeX format, both using the same dataset. This tool was also used for validation of the spelling and the grammar. All technical claims, references, calculations, and final wording were written by the author.

A Transfer function of the Baseband circuit development

The hypothesis of an ideal op-amp has been used, implying that the gain provided is infinite, the input impedance is infinite and the output impedance is zero, we can derive the transfer function of each stage of the baseband circuit.

A.1 First stage

Define the differential input and output voltages as :

$$V_{id} = V_{in+} - V_{in-} \quad (\text{A.1})$$

$$V_{od} = V_{out+} - V_{out-} \quad (\text{A.2})$$

To simplify the analysis, the circuit is split into two cascaded sections: the passive high-pass input filter $H_{1a}(s)$ and the active amplifier $H_{1b}(s)$. Since the fully differential circuit is horizontally symmetric, it is solved using the half-circuit equivalent model, where the horizontal axis of symmetry acts as an Alternating Current (AC) virtual ground.

For the half-circuit, the differential capacitor C_{11} is split into two equivalent grounded capacitors of value $2C_{11}$. The total equivalent shunt capacitance to ground is therefore :

$$C_{eq} = C_{15} + 2C_{11} \quad (\text{A.3})$$

$$L = L_1 = L_2 \quad (\text{A.4})$$

The impedances of the passive components are $Z_L = sL$ and $Z_C = \frac{1}{sC_{eq}}$. The passive LC network is loaded by the amplifier's input resistor $R_{in} = R_{16} = R_{17}$. As the operational amplifier maintains a virtual ground at its input terminals, R_{in} appears in parallel with Z_C . The transfer function of the passive filter defined as

$H_{1a}(s)$ is given by the voltage divider :

$$H_{1a}(s) = \frac{V_{\text{node}}}{V_{id}/2} = \frac{Z_C \parallel R_{in}}{Z_L + (Z_C \parallel R_{in})} \quad (\text{A.5})$$

Substituting the impedances and simplifying, we obtain the 2nd-order response :

$$H_{1a}(s) = \frac{\frac{R_{in}}{1 + sR_{in}C_{eq}}}{sL + \frac{R_{in}}{1 + sR_{in}C_{eq}}} = \frac{1}{1 + s\frac{L}{R_{in}} + s^2LC_{eq}} \quad (\text{A.6})$$

The gain of the fully differential amplifier stage $H_{1b}(s)$ is set by the feedback network. Taking into account the feedback capacitor defined as $C_f = C_{12} = C_{13}$ and the feedback impedance as $Z_f = \frac{R_f}{1 + sR_fC_f}$. The active half-circuit transfer function is :

$$H_{1b}(s) = \frac{V_{od}/2}{V_{\text{node}}} = -\frac{Z_f}{R_{in}} = -\frac{R_f}{R_{in}} \cdot \frac{1}{1 + sR_fC_f} \quad (\text{A.7})$$

Finally, combining both sections, the total differential transfer function of the first stage is :

$$\begin{aligned} H_1(s) &= \frac{V_{od}(s)}{V_{id}(s)} = H_{1a}(s) \cdot H_{1b}(s) \\ &= -\frac{R_f}{R_{in}} \cdot \frac{1}{1 + s\frac{L}{R_{in}} + s^2LC_{eq}} \cdot \frac{1}{1 + sR_fC_f} \end{aligned} \quad (\text{A.8})$$

A.2 Second stage

Let V_{in+} and V_{in-} be the incoming signals coming from the positive and negative inputs of this stage, respectively. The differential input voltage is $V_{in,diff} = V_{in+} - V_{in-}$.

To simplify the analysis, we group the components into generalized complex impedances. Given the symmetric design of the circuit, we define the input impedances Z_{in} and the feedback/ground impedances Z_f as follows :

$$\begin{cases} Z_{in} = R_2 + \frac{1}{sC_9} = R_4 + \frac{1}{sC_{18}} \\ Z_f = R_1 = R_3 \end{cases} \quad (\text{A.9})$$

Assuming an ideal operational amplifier, we can determine the output voltage V_{out} using the superposition principle. First, we turn off V_{in-} (shorted to ground). The circuit acts as a non-inverting amplifier, where the non-inverting input voltage V_+ is set by the voltage divider formed by Z_{in} and Z_f :

$$V_+ = V_{in+} \frac{Z_f}{Z_{in} + Z_f} \quad (\text{A.10})$$

The partial output voltage V_{out1} due to V_{in+} is amplified by the non-inverting gain :

$$V_{out1} = V_+ \left(1 + \frac{Z_f}{Z_{in}} \right) = \left(V_{in+} \frac{Z_f}{Z_{in} + Z_f} \right) \left(\frac{Z_{in} + Z_f}{Z_{in}} \right) = V_{in+} \frac{Z_f}{Z_{in}} \quad (\text{A.11})$$

Next, we turn off V_{in+} (shorted to ground). The circuit now acts as a standard inverting amplifier. The non-inverting input is at 0 V, creating a virtual ground at the inverting input. The partial output voltage V_{out2} due to V_{in-} is :

$$V_{out2} = -V_{in-} \frac{Z_f}{Z_{in}} \quad (\text{A.12})$$

According to the superposition theorem, the total output voltage is the sum of the partial outputs :

$$V_{out} = V_{out1} + V_{out2} = (V_{in+} - V_{in-}) \frac{Z_f}{Z_{in}} = V_{in,diff} \frac{Z_f}{Z_{in}} \quad (\text{A.13})$$

Substituting the actual component values back into the equation, we obtain the transfer function $H_2(s)$:

$$H_2(s) = \frac{V_{out}}{V_{in,diff}} = \frac{R_1}{R_2 + \frac{1}{sC_9}} = \frac{sR_1C_9}{1 + sR_2C_9} \quad (\text{A.14})$$

To highlight the physical parameters of the filter, the equation can be rewritten in the canonical form of a first-order high-pass filter :

$$H_2(s) = \frac{R_1}{R_2} \cdot \frac{sR_2C_9}{1 + sR_2C_9} \quad (\text{A.15})$$

A.3 Third stage

Let V_1 be the voltage at the intermediate node connecting R_7 , R_8 , R_9 , and C_{19} . With the assumption of ideal operational amplifier, we have $V_- = 0$ V.

Applying Kirchhoff's Current Law (KCL) at the inverting input node (V_-) :

$$\frac{V_1}{R_8} + \frac{V_{out}}{\frac{1}{sC_{20}}} = 0 \implies V_1 = -V_{out} \cdot sR_8C_{20} \quad (\text{A.16})$$

Applying KCL at the intermediate node V_1 :

$$\frac{V_1 - V_{in}}{R_7} + \frac{V_1}{R_8} + \frac{V_1 - V_{out}}{R_9} + \frac{V_1}{\frac{1}{sC_{19}}} = 0 \quad (\text{A.17})$$

Rearranging the terms to isolate V_1 :

$$V_1 \left(\frac{1}{R_7} + \frac{1}{R_8} + \frac{1}{R_9} + sC_{19} \right) = \frac{V_{in}}{R_7} + \frac{V_{out}}{R_9} \quad (\text{A.18})$$

Substituting Equation (A.16) into Equation (A.18) :

$$-V_{out} \cdot sR_8C_{20} \left(\frac{1}{R_7} + \frac{1}{R_8} + \frac{1}{R_9} + sC_{19} \right) = \frac{V_{in}}{R_7} + \frac{V_{out}}{R_9} \quad (\text{A.19})$$

Grouping the V_{out} terms on the right side of the equation :

$$-\frac{V_{in}}{R_7} = V_{out} \left[\frac{1}{R_9} + sR_8C_{20} \left(\frac{1}{R_7} + \frac{1}{R_8} + \frac{1}{R_9} \right) + s^2R_8C_{19}C_{20} \right] \quad (\text{A.20})$$

Dividing by V_{in} and multiplying the numerator and the denominator by R_9 yields the first form of the transfer function, highlighting the static gain $H = \frac{R_9}{R_7}$:

$$H_3(s) = \frac{V_{out}}{V_{in}} = \frac{\frac{-R_9}{R_7}}{1 + R_8R_9C_{20} \left(\frac{1}{R_7} + \frac{1}{R_8} + \frac{1}{R_9} \right) s + R_8R_9C_{19}C_{20}s^2} \quad (\text{A.21})$$

To obtain the standard canonical form of a second-order low-pass filter, both the numerator and the denominator are divided by the coefficient of s^2 , which is $R_8R_9C_{19}C_{20}$.

$$\begin{aligned} H_3(s) &= \frac{\frac{-R_9/R_7}{R_8R_9C_{19}C_{20}}}{s^2 + s \frac{R_8R_9C_{20}}{R_8R_9C_{19}C_{20}} \left(\frac{1}{R_7} + \frac{1}{R_8} + \frac{1}{R_9} \right) + \frac{1}{R_8R_9C_{19}C_{20}}} \\ &= \frac{\frac{-R_9R_8C_{19}C_{20}}{R_7R_8C_{19}C_{20}}}{s^2 + s \frac{1}{C_{19}} \left(\frac{1}{R_7} + \frac{1}{R_8} + \frac{1}{R_9} \right) + \frac{1}{R_8R_9C_{19}C_{20}}} \end{aligned} \quad (\text{A.22})$$

B Packages of the components of the PCB

This appendix provides a table listing the active components used in the PCB design along with their corresponding packages.

Component	Package
adp7118	TSOT-5
adp7182	5-Lead TSOT
LT5575	UF PACKAGE 16-LEAD (4mm x 4mm) PLASTIC QFN
ADA4940	8-Lead SOIC (Single)/4-Layer Board
LT1818	S5 PACKAGE 5-LEAD PLASTIC TSOT-23
AD8065	5-Lead Small Outline Transistor Package [SOT-23] (RJ-5)

C Bill of Materials

This appendix provides the bill of materials used for the PCB implementation provided by KiCad.

"Id";	"Designator";	"Footprint";	"Quantity";	"Designation";	"Supplier and ref";
1;	"C30,C40,C35,C46,C33,C31,C36,C28,C32,C34,C27,C23,C37,C39,C26,C25,C38,C24,C29";	"0603_V4";	19;	"100n";	;;;
2;	"TP3,TP2,TP11,TP14,TP1,TP7,TP4,TP8,TP10,TP5,TP6,TP12,TP13,TP17,TP15,TP16,TP9";	"TESTPAD_JUMPERLIKE";	17;	"~";	;;;
3;	"C9,C19,C10,C20";	"0603_V4";	4;	"159n";	;;;
4;	"R19,R18,R7,R6";	"0603_V4";	4;	"1000";	;;;
5;	"R13,R3,R14,R15";	"0603_V4";	4;	"150";	;;;
6;	"C42,C44,C47,C45,C41,C49";	"0805_V2";	6;	"2.2u";	;;;
7;	"SMA3,SMA1,SMA2,SMA4";	"SMA8401-0000TER";	4;	"~";	;;;
8;	"C1,C13,C5,C2";	"0603_V4";	4;	"100p";	;;;
9;	"R40,R34,R35,R33";	"0603_V4";	4;	"1k";	;;;
10;	"R21,R20,R8,R9";	"0603_V4";	4;	"5620";	;;;
11;	"D1,D4,D3,D2";	"0805_V2";	4;	"~";	;;;
12;	"R1,R2";	"0603_V4";	2;	"10k";	;;;
13;	"R29,R38";	"0603_V4";	2;	"316,67k";	;;;
14;	"IC3";	"SOT95P237X111-3N";	1;	"LM4040DYM3-2.5-TR";	;;;
15;	"R28";	"0603_V4";	1;	"170,5k";	;;;
16;	"R4,R17,R16,R5";	"0603_V4";	4;	"475";	;;;
17;	"PHOENIX_CONV_X3";	"TERMINAL_BLOC_3X_254";	1;	"~";	;;;
18;	"IC2,IC6,IC4";	"ADP7118AUJZ-5.0-R7";	3;	"ADP7118AUJZ-5.0-R7";	;;;
19;	"R24,R12";	"0603_V4";	2;	"120";	;;;
20;	"J1,J3,J3,J2";	"PinHeader_1x02_P2.54mm_Vertical";	4;	"PinHeader_1x02_P2.54mm_Vertical";	;;;
21;	"C17,C18,C8,C7";	"0603_V4";	4;	"1.8p";	;;;
22;	"R11,R23";	"0603_V4";	2;	"796";	;;;
23;	"IC5";	"SOT95P270X145-5N";	1;	"~";	;;;
24;	"L2,L4,L3,L1";	"0603_V4";	4;	"680n";	;;;
25;	"C43,C48";	"0805_V2";	2;	"4.7u";	;;;

```

26;"U7,U1";"R_8_ADI";2;"ADA4940-1ARZ";;;
27;"R32,R39,R27,R30,R25,R37";"0603_V4";6;"100k";;;
28;"R22,R10";"0603_V4";2;"141";;;
29;"C11,C21";"0603_V4";2;"132p";;;
30;"C12,C22";"0603_V4";2;"5p";;;
31;"R31";"0603_V4";1;"175k";;;
32;"C6,C15,C3,C14,C4,C16";"0603_V4";6;"33p";;;
33;"IC1";"QFN16-HITTITE";1;"~";;;
34;"R36,R26";"0603_V4";2;"0";;;
35;"U3,U2";"LT1818IS5#TRPBF";2;"LT1818";;;
36;"J4";"PinHeader_2x08_P2.54mm_Vertical";1;"
    PinHeader_2x08_P2.54mm_Vertical";;;
37;"U5,U6";"AD8065";2;"AD8065";;;

```

D Allan deviation Python code

```
#####
# Dependencies
#####

! python.exe -m pip install --upgrade pip
! pip install --upgrade pandas matplotlib numpy scipy pypicosdk
    picosdk allantools

# imports
import numpy as np
import matplotlib.pyplot as plt
import matplotlib.ticker as mticker
import time
import os
import pandas as pd
from ctypes import c_int16, c_int32, c_float, byref, POINTER
from picosdk.ps3000a import ps3000a as ps3k
from picosdk.functions import assert_pico_ok
from datetime import datetime
import allantools

#####
# Useful functions:
#####

def init_picoscope(channel='A', v_range=ps3k.PS3000A_RANGE['PS3000A_5V']):
    """ Initialize the PicoScope for data acquisition on a
        specified channel and voltage
        range.

    Args:
        channel (str, optional): The channel to initialize.
                                Defaults to 'A'.
        v_range (PS3000A_RANGE['...'], optional): The input
                                                    voltage range. Defaults
                                                    to ps3k.PS3000A_RANGE['
```

```

        PS3000A_5V'].

Returns:
    chandle (c_int16): Handle for the opened PicoScope unit
    ch (int): Channel identifier for the initialized channel
    max_adc (int): Maximum ADC value for the configured
                   voltage range, used for
                   converting raw ADC values
                   to millivolts
"""

# Possible channels and their corresponding identifiers in the
# PicoScope API
ch_map = {
    'A': ps3k.PS3000A_CHANNEL['PS3000A_CHANNEL_A'],
    'B': ps3k.PS3000A_CHANNEL['PS3000A_CHANNEL_B'],
    'C': ps3k.PS3000A_CHANNEL['PS3000A_CHANNEL_C'],
    'D': ps3k.PS3000A_CHANNEL['PS3000A_CHANNEL_D'],
}

chandle = c_int16() # Handle for the opened PicoScope unit
status = {}

# Open the PicoScope unit
status['openunit'] = ps3k.ps3000aOpenUnit(byref(chandle), None)
assert_pico_ok(status['openunit'])

# Configure the specified channel with the given voltage range
# and DC coupling
ch = ch_map[channel]
status['set_ch'] = ps3k.ps3000aSetChannel(
    chandle, ch, 1,
    ps3k.PS3000A_COUPLING['PS3000A_DC'],
    v_range, 0.0
)
assert_pico_ok(status['set_ch'])

# Get max ADC once
max_adc = c_int16() # Maximum ADC value for the configured
                    # voltage range
status['max_adc'] = ps3k.ps3000aMaximumValue(chandle, byref(
    max_adc))
assert_pico_ok(status['max_adc'])

return chandle, ch, max_adc.value

def acquire_block(chandle, ch, max_adc,
                  samples,

```

```

        timebase,
        trigger_mv=0,
        pre_trigger_samples=None,
        V_pico=ps3k.PS3000A_RANGE['PS3000A_5V']):
    """ Acquire a block of data from the PicoScope.

    Args:
        chandle (str): PicoScope handle
        ch (PS3000A_CHANNEL['...']): Channel identifier for data
                                   acquisition
        max_adc (int): Maximum ADC value for the configured
                       voltage range, used for
                       converting raw ADC values
                       to millivolts
        samples (int): Number of samples to acquire
        timebase (int): Timebase identifier for data acquisition
        trigger_mv (int, optional): Trigger voltage in millivolts.
                                   Defaults to 0.
        pre_trigger_samples (int, optional): Number of samples to
                                             acquire before the
                                             trigger event. Defaults
                                             to None.

    Returns:
        time_axis (np.array): Time points of the acquired block
        signal_mv (np.array): Signal values in mV corresponding to
                               time_axis

    """
    if V_pico == ps3k.PS3000A_RANGE['PS3000A_5V']:
        V_range = 5000
    elif V_pico == ps3k.PS3000A_RANGE['PS3000A_2V']:
        V_range = 2000
    elif V_pico == ps3k.PS3000A_RANGE['PS3000A_1V']:
        V_range = 1000
    elif V_pico == ps3k.PS3000A_RANGE['PS3000A_500mV']:
        V_range = 500
    elif V_pico == ps3k.PS3000A_RANGE['PS3000A_200mV']:
        V_range = 200
    else:
        raise ValueError("Unsupported voltage range")

    if pre_trigger_samples is None:
        pre_trigger_samples = samples // 2

    post_trigger_samples = samples - pre_trigger_samples

    # Trigger (can also move outside loop if constant)
    threshold_adc = int(trigger_mv * max_adc / 5000)

```

```

ps3k.ps3000aSetSimpleTrigger(
    chandle, 1, ch, threshold_adc,
    ps3k.PS3000A_THRESHOLD_DIRECTION['PS3000A_RISING'],
    0, 1000
)

# Timebase
time_interval_ns = c_float()
returned_max_samples = c_int32()

ps3k.ps3000aGetTimebase2(
    chandle, timebase, samples,
    byref(time_interval_ns), 1,
    byref(returned_max_samples), 0
)

# Run block
ps3k.ps3000aRunBlock(
    chandle,
    pre_trigger_samples,
    post_trigger_samples,
    timebase,
    1, None, 0, None, None
)

# Wait ready
ready = c_int16(0)
while not ready.value:
    ps3k.ps3000aIsReady(chandle, byref(ready))

# Buffer
buffer_adc = np.zeros(samples, dtype=np.int16)

ps3k.ps3000aSetDataBuffer(
    chandle, ch,
    buffer_adc.ctypes.data_as(POINTER(c_int16)),
    samples, 0, 0
)

c_samples = c_int32(samples)
overflow = c_int16()

# Get values
ps3k.ps3000aGetValues(
    chandle, 0,
    byref(c_samples),
    1, 0, 0,
    byref(overflow)
)

```

```

# Convert
signal_mv = (buffer_adc[:c_samples.value] / max_adc) * V_range

dt = time_interval_ns.value * 1e-9
time_axis = np.arange(c_samples.value) * dt

return time_axis, signal_mv

def acquire_dual_block(chandle, chA, chB, max_adc,
                       samples=50000,
                       timebase=3,
                       trigger_mv=0,
                       V_pico=ps3k.PS3000A_RANGE['PS3000A_5V'],
                       auto_trigger_ms=1,
                       enable_trigger=1):
    """ Acquire a dual-channel block of data from the PicoScope.

    Args:
        chandle (str): PicoScope handle
        chA (PS3000A_CHANNEL['...']): Channel identifier for the
            first channel
        chB (PS3000A_CHANNEL['...']): Channel identifier for the
            second channel
        max_adc (int): Maximum ADC value for the configured
            voltage range, used for
            converting raw ADC values
            to millivolts
        samples (int): Number of samples to acquire
        timebase (int): Timebase identifier for data acquisition
        trigger_mv (int, optional): Trigger voltage in millivolts.
            Defaults to 0.
        V_pico (PS3000A_RANGE['...'], optional): Voltage range
            setting for the PicoScope
            . Defaults to ps3k.
            PS3000A_RANGE['PS3000A_5V
            '].
        auto_trigger_ms (int, optional): Auto-trigger delay in
            milliseconds. Defaults to
            1.
        enable_trigger (int, optional): Flag to enable/disable
            trigger. Defaults to 1.

    Returns:
        time_axis (np.array): Time points of the acquired block
        signalA_mv (np.array): Signal values in mV corresponding
            to time_axis for channel
            A

```

```

        signalB_mv (np.array): Signal values in mV corresponding
                                to time_axis for channel
                                B
    """
    if V_pico == ps3k.PS3000A_RANGE['PS3000A_5V']:
        V_range = 5000
    elif V_pico == ps3k.PS3000A_RANGE['PS3000A_2V']:
        V_range = 2000
    elif V_pico == ps3k.PS3000A_RANGE['PS3000A_1V']:
        V_range = 1000
    elif V_pico == ps3k.PS3000A_RANGE['PS3000A_500mV']:
        V_range = 500
    elif V_pico == ps3k.PS3000A_RANGE['PS3000A_200mV']:
        V_range = 200
    else:
        raise ValueError("Unsupported voltage range")

    post = samples // 2      # Number of samples to acquire after
                             # the trigger event
    pre = samples - post     # Number of samples to acquire before
                             # the trigger event

    # Calculate the trigger threshold in ADC units based on the
    # specified trigger voltage in
    # millivolts and the maximum
    # ADC value for the configured
    # voltage range
    threshold = int(trigger_mv * max_adc / V_range)

    # Trigger on channel A only
    ps3k.ps3000aSetSimpleTrigger(
        chandle, enable_trigger, chA, threshold,
        ps3k.PS3000A_THRESHOLD_DIRECTION['PS3000A_RISING'],
        0, auto_trigger_ms
    )

    # Get the time interval and maximum samples for the specified
    # timebase
    dt_ns = c_float()
    max_samp = c_int32()

    ps3k.ps3000aGetTimebase2(
        chandle, timebase, samples,
        byref(dt_ns), 1,
        byref(max_samp), 0
    )

    # Run block acquisition
    ps3k.ps3000aRunBlock(

```

```

        chandle, pre, post,
        timebase, 1, None, 0, None, None
    )

    # Wait
    ready = c_int16(0)
    while not ready.value:
        ps3k.ps3000aIsReady(chandle, byref(ready))

    # Set up data buffers for both channels
    bufA = np.zeros(samples, dtype=np.int16)
    bufB = np.zeros(samples, dtype=np.int16)

    ps3k.ps3000aSetDataBuffer(chandle, chA, bufA.ctypes.data_as(
        POINTER(c_int16)), samples, 0
        , 0)
    ps3k.ps3000aSetDataBuffer(chandle, chB, bufB.ctypes.data_as(
        POINTER(c_int16)), samples, 0
        , 0)

    # Retrieve the acquired data for both channels
    c_samples = c_int32(samples)
    overflow = c_int16()

    ps3k.ps3000aGetValues(chandle, 0, byref(c_samples), 1, 0, 0,
        byref(overflow))

    # Convert raw ADC values to millivolts for both channels
    dt = dt_ns.value * 1e-9
    t = np.arange(c_samples.value) * dt

    scale = V_range / max_adc

    return t, bufA[:c_samples.value] * scale, bufB[:c_samples.
        value] * scale
def zero_crossings(time, signal):
    """ Detect rising zero crossings in a time-domain signal.

    Args:
        time (np.array): Time points of the signal
        signal (np.array): Signal values

    Returns:
        np.array: Time points of the zero crossings
    """
    # Remove DC offset
    signal = signal - np.mean(signal)

    # Detect rising zero crossings

```

```

idx = np.where((signal[:-1] < 0) & (signal[1:] >= 0))[0]

# Estimate crossing times using linear interpolation
t_cross = []
for i in idx:
    t0, t1 = time[i], time[i+1]
    y0, y1 = signal[i], signal[i+1]

    if y1 == y0:
        continue

    t = t0 - y0 * (t1 - t0) / (y1 - y0)
    t_cross.append(t)

return np.array(t_cross)

def estimate_frequency_from_block(time_axis, signal_mv):
    """ Estimate the frequency of a signal from a block of time-
        domain data by detecting
        rising zero crossings.

    Args:
        time_axis (np.array): Time points of the acquired block
        signal_mv (np.array): Signal values in mV corresponding to
            time_axis

    Returns:
        float: Estimated frequency in Hz, or None if estimation
            fails
    """

    # Detect zero crossings in the acquired block
    t_cross = zero_crossings(time_axis, signal_mv)

    # Need at least 2 crossings to estimate frequency
    if len(t_cross) < 2:
        return None

    # Calculate periods between crossings and estimate frequency
    # as 1/mean_period using linear fit

    n = np.arange(len(t_cross))

    # Fit  $t = nT + t_0$ 
    p = np.polyfit(n, t_cross, 1)
    T = p[0]

    return 1.0 / T

```

```

def fractional_frequency(freq):
    """ Convert a frequency time series to fractional frequency
        deviation from the mean.

    Args:
        freq (np.array): Time series of frequency estimates

    Returns:
        y (np.array): Fractional frequency deviation from the mean
        f0 (float): Mean frequency used as reference for
                    fractional deviation
    """

    f0 = np.mean(freq)
    y = (freq - f0) / f0

    return y, f0

def calculate_allan_deviation(y, sample_times):
    """ Calculate the Allan deviation of a fractional frequency
        time series.

    Args:
        y (np.array): Fractional frequency deviation time series
        sample_times (np.array): Time points corresponding to the
                                samples in y

    Returns:
        tau_out (np.array): Averaging times for which the Allan
                            deviation is calculated
        adev (np.array): Allan deviation values corresponding to
                        tau_out
        adev_err (np.array): Error estimates for the Allan
                            deviation values
        ns (np.array): Number of samples used for each Allan
                        deviation estimate
    """

    # Estimate sampling interval
    dt = np.mean(np.diff(sample_times))

    n = len(y)
    taus = 2 ** np.arange(0, int(np.log2(n//2))) # Possible
                                                averaging times (powers of 2)

    adev = []
    adev_err = []
    ns = []

```

```

tau_out = []

for tau in taus:
    m = n // tau # Number of non-overlapping segments of
                  length tau

    if m < 2:
        continue

    # Compute the average frequency for each segment of length
    # tau
    ybar = y[:m*tau].reshape(m, tau).mean(axis=1)

    # Compute the Allan variance (see formula before)
    avar = np.sum(np.diff(ybar)**2) / (2*(m-1))

    # Store the Allan deviation, corresponding tau, error
    # estimate, and number of
    # samples used
    adev.append(np.sqrt(avar))
    adev_err.append(np.sqrt(avar / (2 * (m - 1))))
    ns.append(m)
    tau_out.append(tau * dt)

return np.array(tau_out), np.array(adev), np.array(adev_err),
       np.array(ns)

#####
# Main Execution
#####

ps3k.ps3000aCloseUnit(chandle) # Just in case it was left open the
                                PicoScope from a previous run

### Plot few period of the signal (30.72 MHz) to verify
                                acquisition ###

f = 30.72e6
T = 1 / f # = 32.55 ns

# initialize PicoScope and acquire a block of data
chandle, ch, max_adc = init_picoscope()
time_axis, signal_mv = acquire_block(chandle, ch, max_adc, samples
                                     =3000, timebase=3)

ps3k.ps3000aCloseUnit(chandle)
plt.figure(figsize=(10, 4))
plt.plot(time_axis * 1e9, signal_mv)
plt.xlabel("Time (ns)")
plt.ylabel("Signal (mV)")
plt.xlim(0, 5*T*1e9) # limit x from 0 to 5 periods of the 30.72
                     MHz signal

```

```

plt.title("Acquired Signal Block")
plt.grid()
plt.savefig(f"Results/output_signal_{datetime.now().strftime('%Y-%m-%d_%H-%M-%S')}.pdf", format="pdf")

plt.show()

# Save data to txt file in Time Voltage format
data_to_save = np.column_stack((time_axis, signal_mv))
np.savetxt(f"Results/output_signal_{datetime.now().strftime('%Y-%m-%d_%H-%M-%S')}_signal_data.txt",
           data_to_save, header="Time(s)\nVoltage(mV)", comments='')

### TWO U-BLOX COMPARISON ###
duration_total_s = 60*60*4 # Total duration of acquisition (
                             seconds)
samples = 2000 # Can be changed to adjust the time resolution (eg
                5000 vs 40000)
f = 30.72e6 # Frequency of the signal (Hz)
period_s = 1.0 / f # Period of the signal (s)

print("Starting dual-channel acquisition...")

# Init the PicoScope
chandle, ch, max_adc = init_picoscope()

# Choose channels
chA = ps3k.PS3000A_CHANNEL['PS3000A_CHANNEL_A']
chB = ps3k.PS3000A_CHANNEL['PS3000A_CHANNEL_B']

t_start = time.perf_counter()

# Lists to store phase difference and corresponding time tags
# Phase difference is calculated as the time difference between
# zero crossings of the two
# channels, averaged over each
# block

phase_diff_series = []
time_tags = []

# Acquire data in blocks until the total duration is reached
try:
    while (time.perf_counter() - t_start) < duration_total_s:

        # Acquire dual-channel block of data
        t, sigA, sigB = acquire_dual_block(
            chandle, chA, chB, max_adc,
            samples=samples,

```

```

        timebase=3,
        auto_trigger_ms=1
    )

    # Extract zero crossings for both channels
    tA = zero_crossings(t, sigA)
    tB = zero_crossings(t, sigB)

    n = min(len(tA), len(tB))

    if n < 5:
        continue

    # Calculate phase difference (time difference between zero
                                crossings) for the first
                                n crossings

    phase = tB[:n] - tA[:n]

    # average per block
    phase_diff_series.append(np.mean(phase))
    time_tags.append(time.perf_counter() - t_start)

finally:
    ps3k.ps3000aCloseUnit(chandle)

phase_diff_series = np.array(phase_diff_series)
time_tags = np.array(time_tags)

print("Done.")

# Remove DC offset from phase difference series
y = phase_diff_series - np.mean(phase_diff_series)

# We have the phase differences in seconds (Phase Error, x)
x_phase = phase_diff_series

# Unroll the phase if the difference exceeds half a cycle (Cycle
                                Slip)
x_phase_unwrapped = np.unwrap(x_phase / period_s * 2 * np.pi) *
                    period_s / (2 * np.pi)

# Ensure dt_tags are positive and print the number of invalid
                                points
dt_tags = np.diff(time_tags)
valid = dt_tags > 0
print(f"Number of valid points: {np.sum(valid)}")
print(f"Number of invalid points: {np.sum(~valid)}")

# Convert the Phase (x) to Fractional Frequency (y)

```

```

# Formula :  $y_i = (x_{i+1} - x_i) / (t_{i+1} - t_i)$ 
y_fractional = np.diff(x_phase_unwrapped)[valid] / dt_tags[valid]

# Adjust the time tags to have the same size as the fractional
# frequency (y)
time_tags_y = np.array(time_tags)[1:][valid]

# Allan deviation of the fractional frequency time series using
# custom implementation
tau, adev, adev_err, ns = calculate_allan_deviation(y_fractional,
                                                    time_tags_y)

# Calculate Allan deviation using allantools
taus_allantools, adev_allantools, adev_err_at, ns_at = allantools.
    adev(
        y_fractional, data_type="freq", rate=1/np.mean(dt_tags[valid])
    )
# Calculate Overlapping Allan deviation using allantools
taus_ov, adev_ov, adev_err_ov, ns_ov = allantools.oadev(
    y_fractional, data_type="freq", rate=1/np.mean(dt_tags[valid])
)

# Supposing the two u-blox are independent, the Allan deviation of
# each individual u-blox
# should be the Allan deviation of their phase difference divided
# by sqrt(2).

# In the data-sheet of the u-blox EVK-M8F:
# Frequency accuracy:
#   GNSS locked:      <= 5 ppb
#   Hold-over, 24 hours <= 100 ppb max. (< 25 ppb typ. under
#                                       stable operating conditions)

# So we expect the Allan deviation of each individual u-blox to be
# around 5 ppb at short averaging
# times
# (when GNSS locked) and to degrade to around 25-100 ppb at longer
# averaging times (hold-over).

# Individual u-blox Allan deviation estimate
TWO_UBLOX = True
if TWO_UBLOX:
    adev = adev / np.sqrt(2)
    adev_allantools = adev_allantools / np.sqrt(2)
    adev_ov = adev_ov / np.sqrt(2)
    adev_err = adev_err / np.sqrt(2)
    adev_err_at = adev_err_at / np.sqrt(2)
    adev_err_ov = adev_err_ov / np.sqrt(2)

```

```

fig, ax = plt.subplots(figsize=(7,5))
line_custom = ax.loglog(tau, adev, 'o-', markersize=4, label="
                                Custom Non-overlapping ADEV")[0]
line_at = ax.loglog(taus_allantools, adev_allantools, 's-',
                    markersize=4, label="Allantools
                                Non-overlapping ADEV")[0]
line_ov = ax.loglog(taus_ov, adev_ov, 's-', markersize=4, label="
                                Overlapping ADEV")[0]
ax.errorbar(tau, adev, yerr=adev_err, fmt="none", ecolor=
            line_custom.get_color(), alpha=0.
            9, capsize=4, capthick=1.0,
            elinewidth=1.2, zorder=4)
ax.errorbar(taus_allantools, adev_allantools, yerr=adev_err_at,
            fmt="none", ecolor=line_at.
            get_color(), alpha=0.9, capsize=4
            , capthick=1.0, elinewidth=1.2,
            zorder=4)
ax.errorbar(taus_ov, adev_ov, yerr=adev_err_ov, fmt="none", ecolor
            =line_ov.get_color(), alpha=0.9,
            capsize=4, capthick=1.0,
            elinewidth=1.2, zorder=4)

# Vertical line for tau = 1 hour
ax.axvline(3600, color="red", linestyle="--", linewidth=1.0, label
            ="1 hour")

ax.set_xlabel("Tau (s)")
ax.set_ylabel("Allan deviation (phase difference)")
ax.set_title("Allan Deviation of Phase Difference (Multi-block) -
            u-blox A vs B")

ax.grid(True, which="both")
ax.legend()
plt.show()

# TIE_RMS(tau) and MTIE(tau) using allantools
time_tie = time_tags - time_tags[0]
tie = x_phase_unwrapped - np.mean(x_phase_unwrapped)
rate_tie = 1.0 / np.mean(np.diff(time_tie))
tierms_out = allantools.tierms(tie, rate=rate_tie)
mtie_out = allantools.mtie(tie, rate=rate_tie)
taus_tierms, tierms_vals = tierms_out[0], tierms_out[1]
taus_mtie, mtie_vals = mtie_out[0], mtie_out[1]

# Individual u-blox TIE/MTIE estimate
TWO_UBLOX = True
if TWO_UBLOX:
    tierms_vals = tierms_vals / np.sqrt(2)
    mtie_vals = mtie_vals / 2

fig, ax = plt.subplots(figsize=(7,5))

```

```

ax.loglog(taus_tierms, tierms_vals, 'o-', markersize=4, label="
                                TIE_RMS")
ax.loglog(taus_mtie, mtie_vals, 's-', markersize=4, label="MTIE")
ax.set_xlabel("Tau (s)")
ax.set_ylabel("Time error (s)")
ax.set_title("TIE_RMS and MTIE vs Tau - u-blox A vs B")
ax.grid(True, which="both")
ax.legend()
plt.show()

# Saving datas in Datas folder for later use in the report and the
                                date (day, hour, minute) of the
                                acquisition in the name

os.makedirs("Datas", exist_ok=True)
df = pd.DataFrame({
    "time_s": time_tags,
    "phase_difference_s": phase_diff_series,
    "fractional_phase_difference": y
})
df.to_csv(f"Datas/phase_difference_timeseries_{datetime.now().
                                strftime('%Y-%m-%d_%H-%M-%S')}.
                                csv", index=False)

```

E Other results for the patch antenna

E.1 CST Mesh

The mesh used in CST for the electromagnetic simulations of the patch antenna is illustrated in Figure E.3.

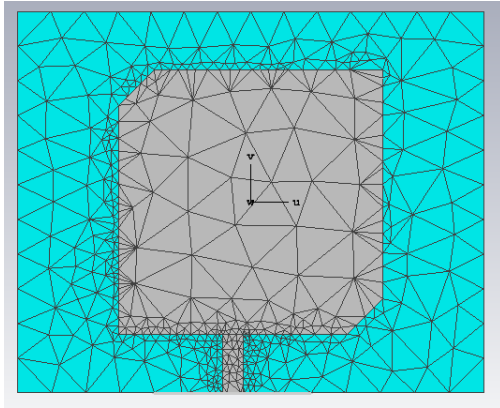


Figure E.1: LHCP patch antenna.

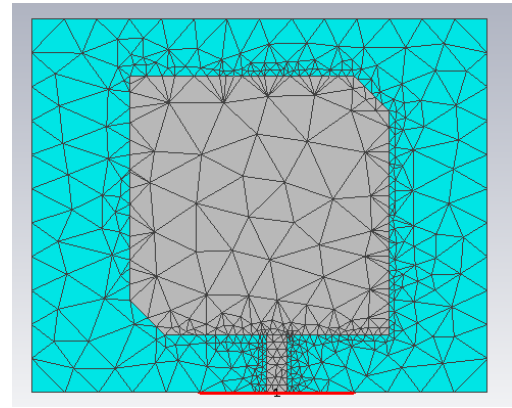


Figure E.2: RHCP patch antenna.

Figure E.3: Mesh used for the CST simulations of the corner-trimmed patch antennas.

E.2 Optimisation

All the metrics shown for the patch antennas were used to optimise the antenna design iteratively. A particular focus has been put on the axial ratio and the gain, as for GNSS-R applications we want to increase as much as possible the power of the received signal without mixing the direct (RHCP) and reflected (LHCP)

signals. This trade-off is achieved by tuning the corner truncation size t , which gives the following results:

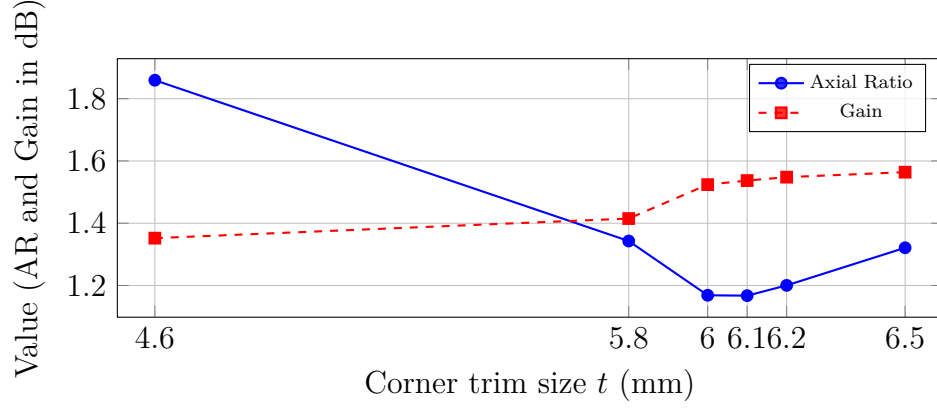


Figure E.4: Axial Ratio and Gain vs. Corner trim size.

The value of t that minimizes the axial ratio is around 6.1 mm, which corresponds to a gain of around 1.537 dB and is therefore the one used for all the results presented in the CST simulations.

E.3 CST surface current

The surface current distribution on the patch antennas at the 1.57542 GHz is shown in Figure E.7.

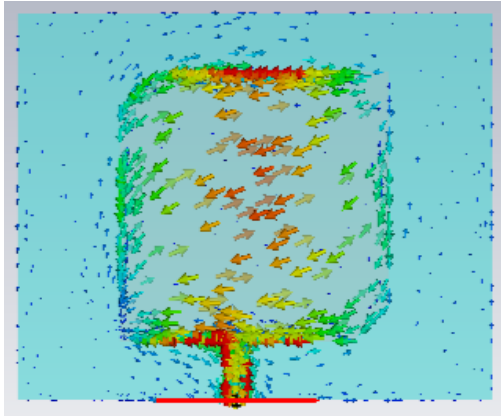


Figure E.5: LHCP surface currents.

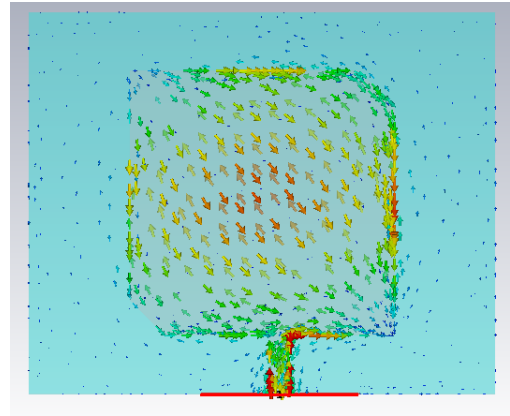


Figure E.6: RHCP surface currents.

Figure E.7: Surface current distribution at 1.57542 GHz.

The rotation of the surface currents on the patch antennas is consistent with the expected circular polarization (LHCP or RHCP).

E.4 Directivity

The directivity of the antenna is a measure of how focused the radiated power is in a particular direction compared to an isotropic radiator¹.

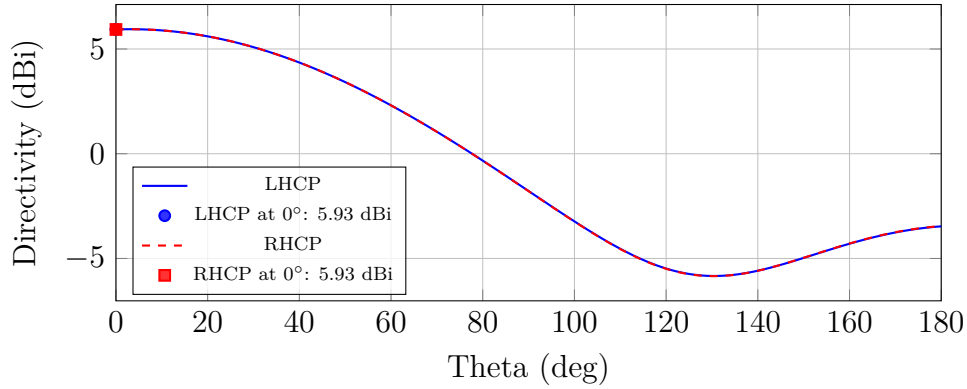


Figure E.8: Directivity of the LHCP and RHCP antennas across different incident angles.

The directivity in the normal incidence direction is around 6 dB for both antennas, which indicates that the antenna radiates four times the power density of an isotropic radiator in the normal incidence direction. The directivity also decreases as the incident angle increases, which is consistent with the gain pattern observed in the previous section.

E.5 Total efficiency

The total efficiency is defined as the radiation efficiency but also includes mismatch losses due to impedance mismatch between the antenna and the feed line. It is defined as :

$$\eta_{\text{total}} = \frac{P_{\text{rad}}}{P_{\text{available}}} = \eta_{\text{rad}} \times (1 - |S_{11}|^2) \quad (\text{E.1})$$

where P_{rad} is the power radiated by the antenna, P_{accepted} is the power accepted by the antenna (after accounting for mismatch losses), $P_{\text{available}}$ is the power available

¹It is a theoretical component which is a lossless point-source antenna that radiates energy with equal intensity in all directions (uniformly over a sphere).

from the source and S_{11} is the reflection coefficient of the antenna. This efficiency metric is evaluated across the frequency band of interest and compared between the LHCP and RHCP antennas as shown in Figure E.9.

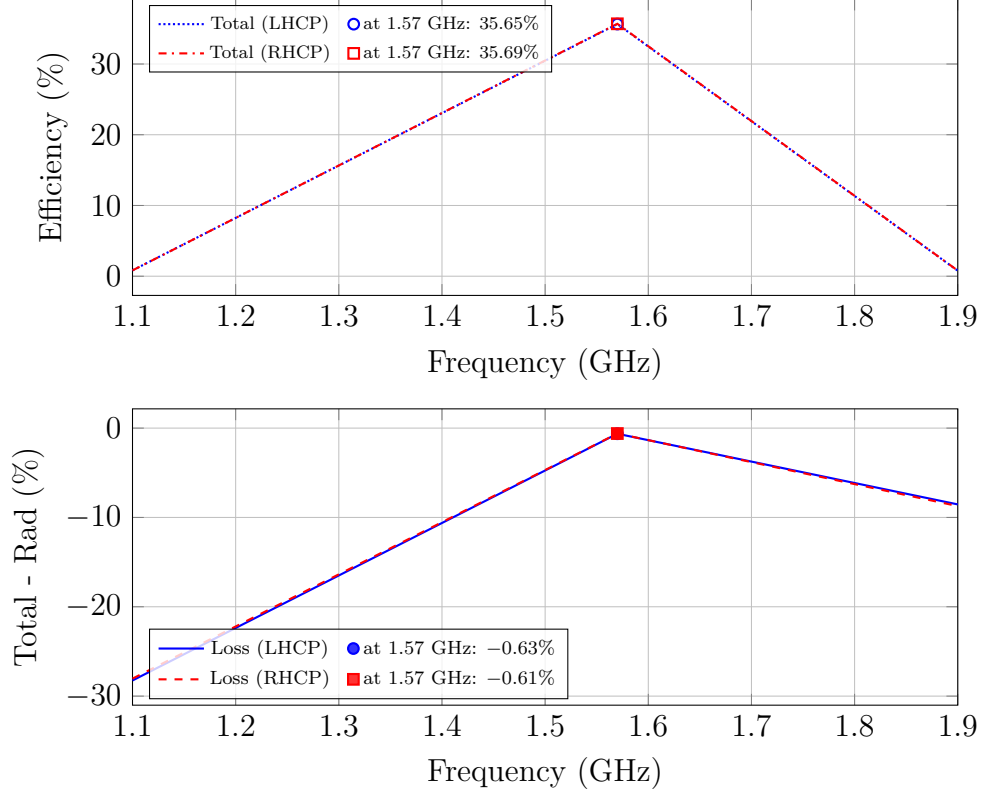


Figure E.9: Efficiency of the LHCP and RHCP antennas across the frequency band of interest.

The total efficiency at the center frequency is around 36% for both antennas, which indicates that the antenna delivers 36% of the available power from the source as radiated power and that the mismatch losses, the difference between total and radiation efficiency so $\eta_{\text{rad}} - \eta_{\text{total}} = \eta_{\text{rad}} \times |S_{11}|^2$, are around 1%.

E.6 ADS results

This section presents the results obtained with ADS software [57] for the LHCP patch antenna with the mesh illustrated in Figure E.10. Similar results are obtained for the RHCP one.

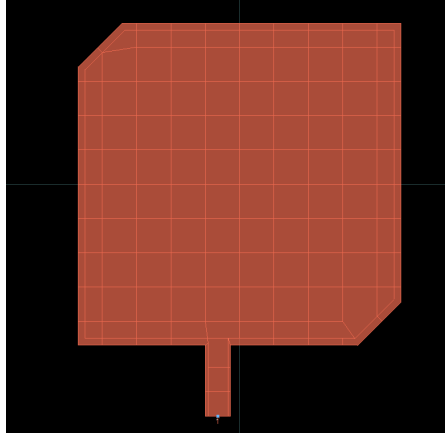


Figure E.10: Mesh used for the ADS electromagnetic simulations of the patch antenna.

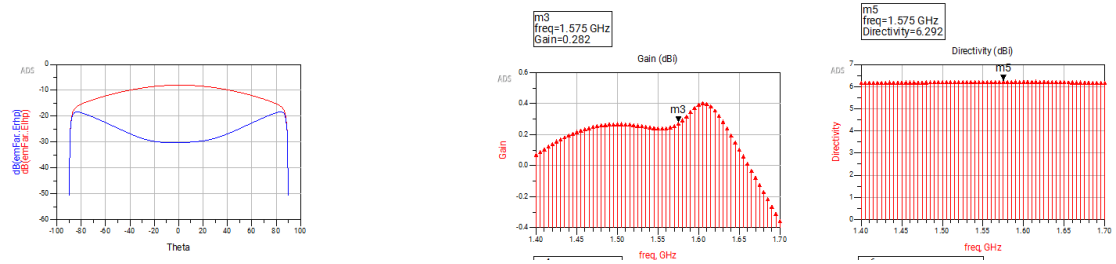
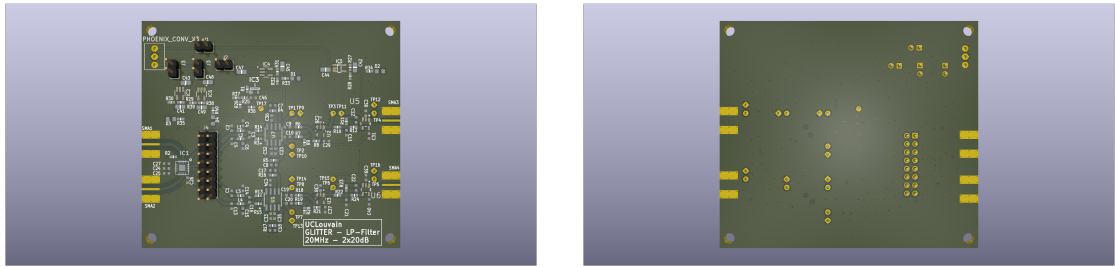


Figure E.11: ADS results for the LHCP patch antenna: (a) E-field components in the boresight direction, (b) gain across frequency and (c) directivity across frequency.

F Additional views of the PCB

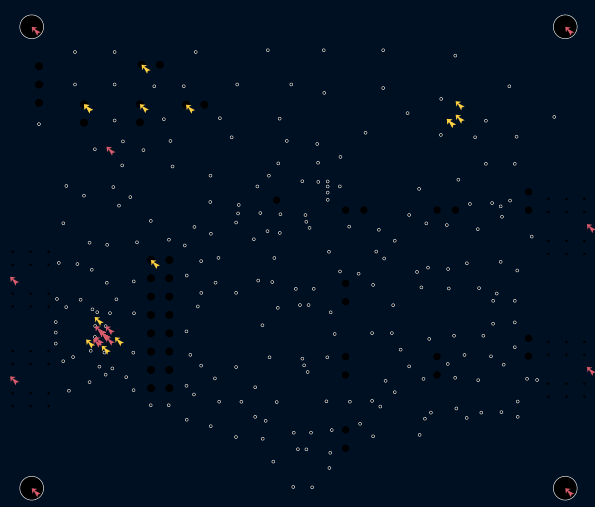
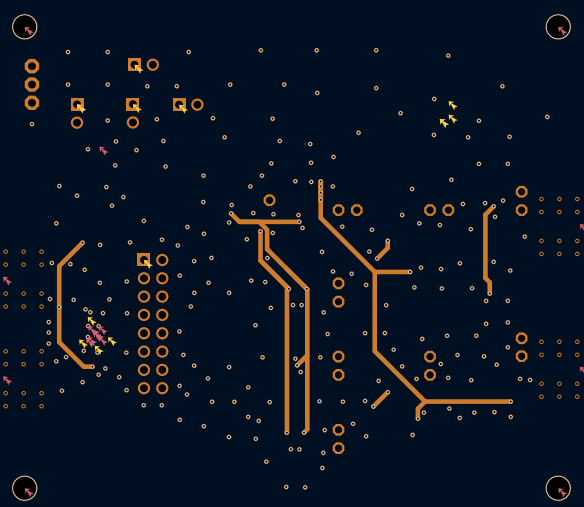
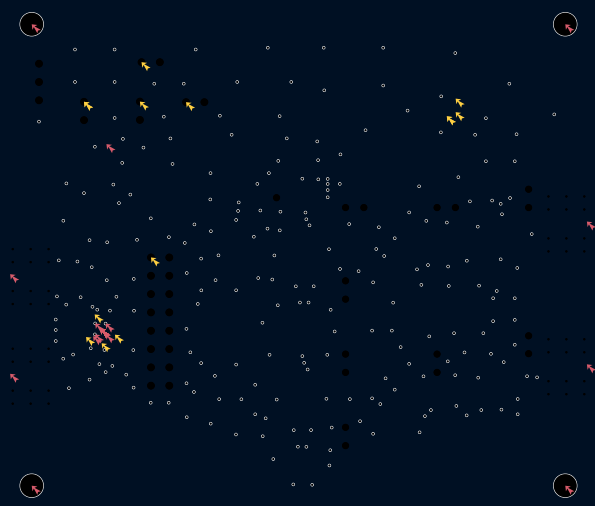
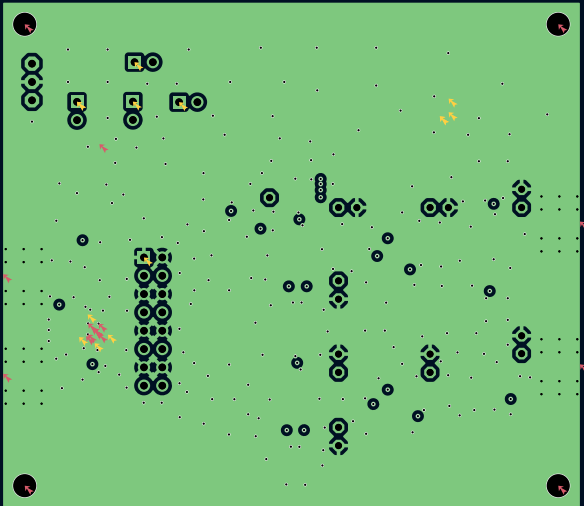
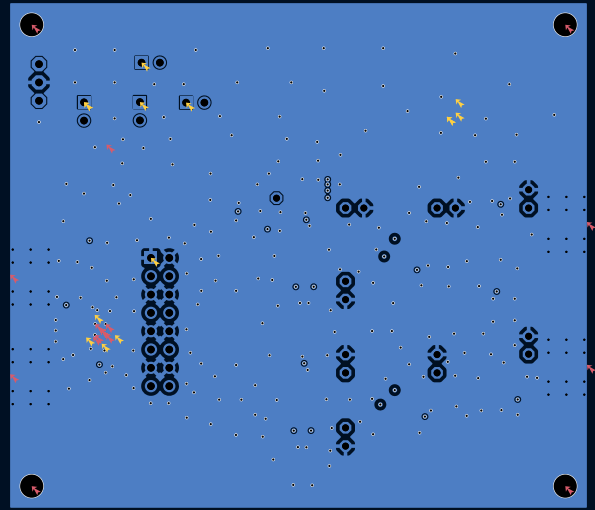
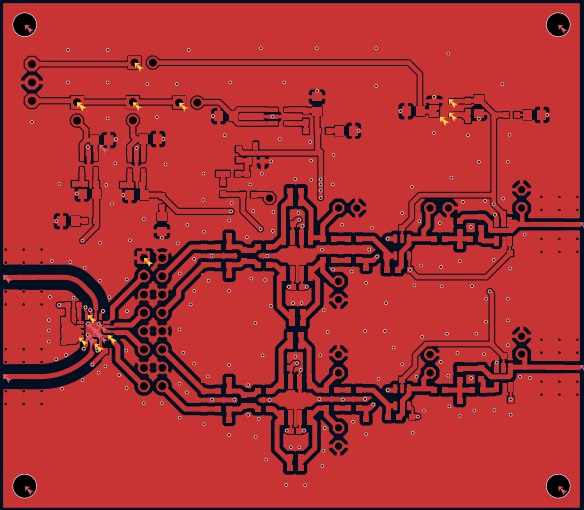
The 3D rendering of the PCB design is shown in Figure F.1.

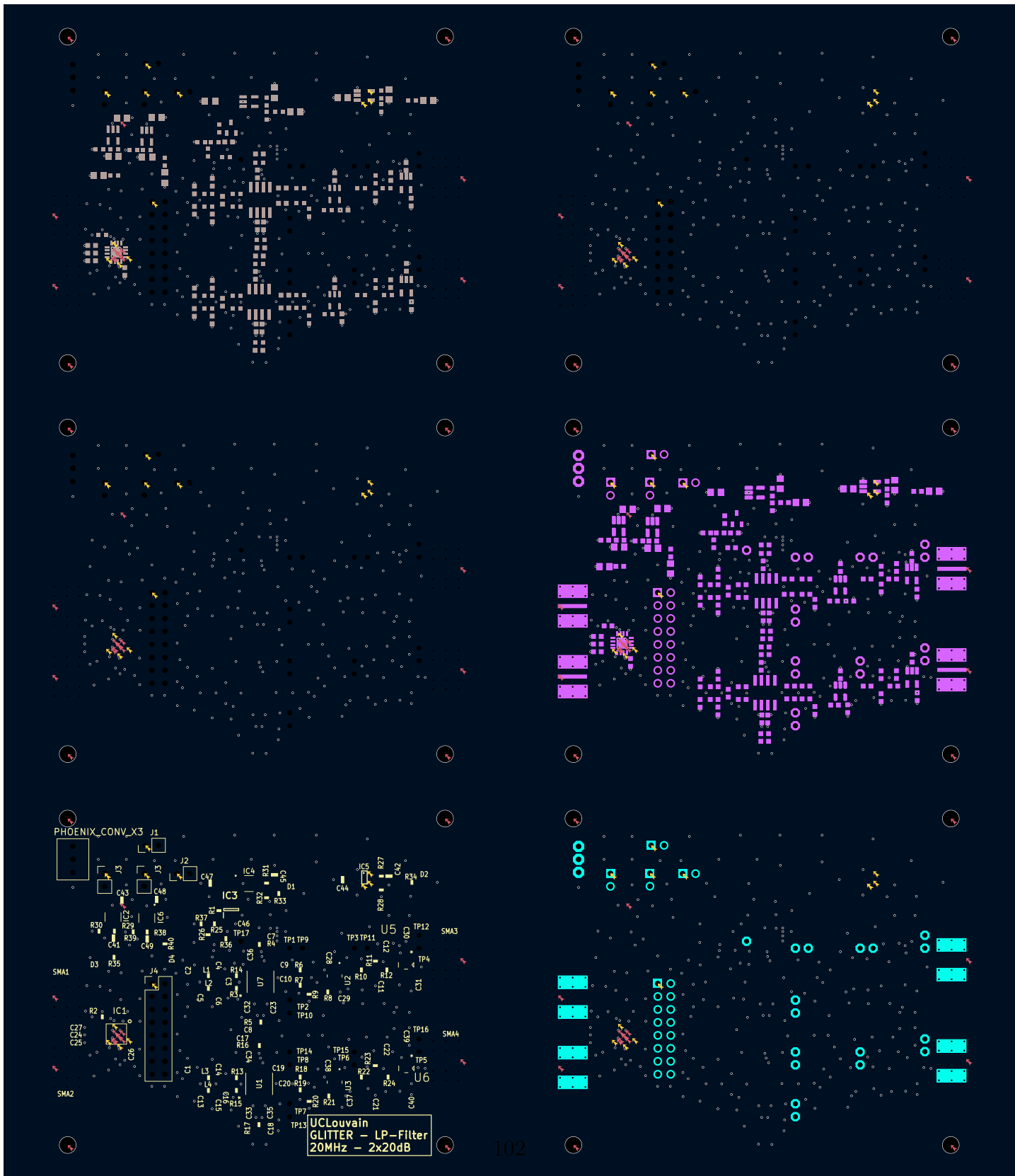


(a) 3D rendering of the PCB (top view) (b) 3D rendering of the PCB (bottom view)

Figure F.1: 3D rendering of the PCB (top and bottom views).

The complete PCB layout exported as a PDF file is illustrated in Figure F. Each page correspond to a specific mask layer, including V1-NPTH, V1-PTH, V1-B_Adhesive, V1-B_Cu, V1-B_Mask, V1-B_Paste, V1-B_Silkscreen, V1-Edge_Cuts, V1-F_Adhesive, V1-F_Courtyard, V1-F_Cu, V1-F_Fab, V1-F_Mask, V1-F_Paste, V1-F_Silkscreen, V1-In1_Cu, V1-In2_Cu, V1-job and V1-Margin. This view allows for detailed inspection of the PCB design and facilitates fabrication and assembly processes.





UNIVERSITÉ CATHOLIQUE DE LOUVAIN

École polytechnique de Louvain

Rue Archimède, 1 bte L6.11.01, 1348 Louvain-la-Neuve, Belgique | www.uclouvain.be/epl