SEQUENTIAL QUADRATURE MEASUREMENTS FOR PLASMA DIAGNOSTICS

by

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Abstract

Sequential Quadrature Measurements for Plasma Diagnostics

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Utah State University, 2014

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The study of the ionosphere has been very important due to its effects on terrestrial and satellite communications. This thesis presents an introduction of the ionosphere effects, its modeling and measurement methods that have been used along the history.

The Sweeping Impedance Probe (SIP) has proven over the years to be a reliable method based on the radio frequency (RF) behavior of the plasma. A new SIP architecture is presented based on the latest techniques available, using a Vector Network Analyzer (VNA) detection and employing dynamic correction of errors with Correlated Double Sampling (CDS) and a reference channel. The design will be detailed showing the component selection based on their performance parameters. In this sense, several analyses have been made to ensure that the sweep rate and accuracy requirements can be met. The testing and calibration methodology is developed to further increase the final accuracy of the instrument.

Lastly, the main conclusions of the project are summarized and new and exciting lines of work are presented for what is expected to be the next generation of SIP instruments.
Public Abstract

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The ionosphere is the atmosphere layer characterized by its high concentration of ionized plasma. It has a great impact on radio communications with satellites, causing disturbances and disruptions. Therefore, it is important to understand and predict the ionosphere characteristics.

The Sweeping Impedance Probe (SIP) is an instrument for characterizing the ionosphere used for many decades with great success. In this thesis, a new SIP architecture design is presented using the latest techniques and components available. The design is detailed and analyses have been performed to ensure the required performances. The new SIP will be flown in the Auroral Spatial Structures Probe (ASSP) sounding rocket mission in early 2015, and it is expected it will make the most accurate measurements to date.

Lastly, the conclusions of this project are presented and future work is outlined for what will become the next generation of SIP instruments.
To my family...
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Julio Martin-Hidalgo
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<th>Definition</th>
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<td>Alternating Current</td>
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<tr>
<td>ADC</td>
<td>Analog-to-Digital Converter</td>
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<td>AM</td>
<td>Amplitude Modulation</td>
</tr>
<tr>
<td>ASSP</td>
<td>Auroral Spatial Structures Probe</td>
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<td>AWG</td>
<td>American Wire Gauge</td>
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<tr>
<td>CDS</td>
<td>Correlated Double Sampling</td>
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<tr>
<td>CSV</td>
<td>Comma-Separated Value</td>
</tr>
<tr>
<td>CT</td>
<td>Current Transformer</td>
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<tr>
<td>DAC</td>
<td>Digital-to-Analog Converter</td>
</tr>
<tr>
<td>DAS</td>
<td>Data Acquisition System</td>
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<tr>
<td>DC</td>
<td>Direct Current</td>
</tr>
<tr>
<td>DCR</td>
<td>Direct Conversion Receiver</td>
</tr>
<tr>
<td>DDS</td>
<td>Direct Digital Synthesizer</td>
</tr>
<tr>
<td>DUT</td>
<td>Device Under Test</td>
</tr>
<tr>
<td>EMI</td>
<td>Electromagnetic Interference</td>
</tr>
<tr>
<td>FDTD</td>
<td>Finite Difference Time Domain</td>
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<tr>
<td>FPGA</td>
<td>Field-Programmable Gate Array</td>
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<tr>
<td>GBW</td>
<td>Gain-Bandwidth Product</td>
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<tr>
<td>GSE</td>
<td>Ground Support Equipment</td>
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<tr>
<td>IGRF</td>
<td>International Geomagnetic Reference Field</td>
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<tr>
<td>INL</td>
<td>Integral Nonlinearity</td>
</tr>
<tr>
<td>LDO</td>
<td>Low-Dropout Regulator</td>
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<td>LNA</td>
<td>Low-Noise Amplifier</td>
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<td>LO</td>
<td>Local Oscillator</td>
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<td>LPF</td>
<td>Low-Pass Filter</td>
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<td>Acronym</td>
<td>Full Form</td>
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<tr>
<td>LSB</td>
<td>Less Significant Bit</td>
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<tr>
<td>MF</td>
<td>Major Frame</td>
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<td>PFP</td>
<td>Plasma Frequency Probe</td>
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<td>Plasma Impedance Probe</td>
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<td>PLL</td>
<td>Phase Locked Loop</td>
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<td>RAM</td>
<td>Random-Access Memory</td>
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<td>RF</td>
<td>Radio Frequency</td>
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<td>RMS</td>
<td>Root Mean Squared</td>
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<tr>
<td>RSS</td>
<td>Root Sum Squared</td>
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<tr>
<td>RTL</td>
<td>Register-Transfer Level</td>
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<td>SDL</td>
<td>Space Dynamics Laboratory</td>
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<tr>
<td>SF</td>
<td>Sub Frame</td>
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<td>SFDR</td>
<td>Spurious-Free Dynamic Range</td>
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<td>SIP</td>
<td>Sweeping Impedance Probe</td>
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<td>SNA</td>
<td>Scalar Network Analyzer</td>
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<td>SS</td>
<td>Stainless Steel</td>
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<td>SSB</td>
<td>Single-Sideband</td>
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<tr>
<td>TCR</td>
<td>Temperature Coefficient of Resistance</td>
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<td>TEC</td>
<td>Total Electron Content</td>
</tr>
<tr>
<td>TM</td>
<td>Telemetry</td>
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<tr>
<td>UART</td>
<td>Universal Asynchronous Receiver/Transmitter</td>
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<td>USU</td>
<td>Utah State University</td>
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<td>VNA</td>
<td>Vector Network Analyzer</td>
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Chapter 1

Introduction

1.1 History

When Marconi transmitted the first transatlantic radio waves in 1901, the scientific community was astonished. It was well known that electromagnetic waves propagate in straight lines, and the Earth’s curvature should it have made impossible this kind of long distance transmission. Heaviside and Kennelly speculated that a layer of ionized plasma high in the atmosphere could have been reflecting radio waves. Later in 1932, the term ionosphere proposed by Watson-Watt was adopted.

The ionosphere is composed of ions and free electrons forming a plasma. The high-energy ultraviolet and X-rays solar radiation are absorbed by the neutral atoms and molecules causing them to release electrons. Due to this ionization mechanism, the plasma is quasi-neutral, that is, the number of ions and electrons are practically the same. As will be explained, it is often more convenient to determine the number of electrons instead of the different ions. The fundamental parameter which defines the level of ionization is the electron density or Total Electron Content (TEC).

The ionosphere has a great temporal and spatial variability. In altitude, the different species and concentrations found in the atmosphere absorb different wavelength and amounts of energy, creating layers of higher concentration as shown in Fig. 1.1 [1]. The ionosphere extends from 70 km in altitude to thousands of km, with the electron density being zero at the surface (total recombination). During the night, two layers named E and F are present. During the day, the D layer appears and layer F is divided into two sublayers. In addition to the daily cycle, differences can be also be observed with the seasons and with the solar cycle. The spatial variability is mainly influenced by the Earth’s magnetic field (Fig. 1.2 [2]). At low latitudes, plasma is concentrated on either side of the magnetic equator forming what
is known as the equatorial anomaly. Of even more importance are the solar storms which cause disturbances in the ionosphere, or plasma bubbles which form at the bottom of the ionosphere and float to the top leaving a turbulent wake that is impenetrable to radio waves. Both are unpredictable with the current models.

The ionosphere has numerous effects on radio waves depending on their frequency. Low frequency radio waves, like the one transmitted by Marconi, are bounced from the ionosphere, while the higher frequency waves are capable of passing through. However, they also suffer from different effects like absorption, Faraday rotation, or phase change due to an increased electrical path. The highest frequency of radio waves that will reflect
back to Earth when transmitted vertically is called the critical frequency of the ionosphere. Radio waves are reflected from the ionosphere when their frequency matches the resonance of the ionospheric plasma. This resonance is related to the local electron density so that radio waves are reflected at different altitudes depending on their frequency, with the lower frequencies reflected at the E layer and the higher frequencies at the F layer. Changes in the electron density can allow radio waves to be transmitted further away (like AM radios during the night) or be completely disrupted. In the same way, satellite communications get affected with distortions and disruptions. The Global Positioning System (GPS) signals must be corrected for the total electron content line-of-sight from the satellite to the receiver to achieve high accuracy. This is done by sending signals at two or more distinct frequencies and comparing the relative delay between them to determine the total electron content. The GPS constellation includes a TEC instrument to account for the increased delay of the radio waves, improving the positioning accuracy, as well as providing high redundancy to mitigate disruptions.

Because of its practical implications on radio wave transmission and its high variability, the ionosphere has been studied for decades with many different methods. Nowadays TEC real-time maps are available online, but there is still a high demand for better models and predictability. For example, the United States government spends $500 million annually to predict and mitigate the disturbances produced by the ionosphere. One of the best models available is GAIM-GM by Utah State University (USU) Space Weather Center, which provides real-time information on the structure of the ionosphere based on 357 TEC stations.

1.2 Ionospheric Measurements

After its discovery, the ionosphere was studied with the same mechanism Marconi used to communicate over the Earth’s horizon that is, by reflecting radio waves from it. Ionosondes operate by emitting a signal of time varying frequency to the ionosphere and determining the propagation time of the reflected signal. With this information the virtual height and electron density can be calculated at different altitudes. Another approach is the
incoherent scatter radar, measuring the reflected power scattered by the cloud of electrons around individual ions. These techniques were later implemented in the mid 50s on satellites making the first topside sounders (i.e. above the F layer).

While most of the current TEC data and models are developed with ground instruments, in-situ measurements provide better accuracy and are more appropriate to measure localized plasma anomalies, turbulences, or bubbles. Because the ionosphere minimum height is about 70 km, the use of high altitude atmospheric balloons is impossible, reserving in-situ measurements for sounding rockets and satellites. In fact, the first in-situ measurement was made by error. After World War II, Americans and Soviets acquired V2 rockets and scientists from Germany. Post war testing of the V2 rockets implemented a radio for telemetry purposes. On some the first flights the radio signal was lost when the rockets exceed a certain altitude and recovered again once it fell below the same altitude. It was understood that the local ionospheric plasma was interacting with the antenna and detuning it from the transmitter. This and other phenomena were used to study the density and structure of the ionosphere. After those first experiments, space weather measurements with sounding rockets became an important scientific goal. They are still in use for low altitudes where satellites can not survive due to high atmospheric drag or during short lived phenomena when it is difficult to get a satellite at the correct location.

Two groups of instruments have been widely used over the years for in-situ measurements of the electron density: Langmuir probes and impedance probes. The first uses probes that collect the free electrons and measure the Direct Current (DC) current collected by the probe as a function of voltage. The latter are based on probes immersed in the plasma and studying the varying electrical properties of its impedance. USU and the Space Dynamics Laboratory (SDL) have an extensive history of over 50 years studying the ionosphere on-board sounding rockets and small satellites [3], and are one of the pioneers of the impedance probes.
1.3 Plasma Physics

1.3.1 Plasma as a Dielectric Medium

In order to understand the impedance probe methods, it is convenient to first understand the plasma properties of the ionosphere. A cold plasma with a magnetic field aligned with the Z-axis, can be approximated as a dielectric medium with the following form for the tensor permitivity:

$$
\epsilon_r = \begin{bmatrix}
\varepsilon_1 & -i\varepsilon_2 & 0 \\
 i\varepsilon_2 & \varepsilon_1 & 0 \\
 0 & 0 & \varepsilon_3 \\
\end{bmatrix},
$$

(1.1)

where

$$
\varepsilon_1 = 1 + \frac{1}{\omega} \sum_k \frac{\omega_k^* \omega_{pk}^2}{\Omega_k^2 - \omega_k^2},
$$

(1.2)

$$
\varepsilon_2 = \frac{1}{\omega} \sum_k \frac{\Omega_k}{\Omega_k^2 - \omega_k^2},
$$

(1.3)

$$
\varepsilon_3 = 1 - \frac{1}{\omega} \sum_k \frac{\omega_{pk}^2}{\omega_k^*},
$$

(1.4)

$$
\omega^* = \omega - j\nu_k,
$$

(1.5)

$$
\omega_{pk}^2 = \frac{n_k q_k^2}{\epsilon_0 m_k},
$$

(1.6)

$$
\Omega_k = \frac{q_k B_0}{m_k},
$$

(1.7)

$n_k, m_k, q_k,$ and $\nu_k$ are the density, mass, charge, and ion or electron neutral collision frequency of the $k$th species, with the first species being electrons followed by various ion species.

For frequencies between 100 kHz and 20 MHz, the ions are unmoving due to their higher mass. Focusing on the electrons, the plasma presents several interesting natural resonant frequencies.
• Plasma frequency \( (\omega_p) \): the natural resonant frequency of the electrons on a plasma in the absence of a magnetic field. It is dependent on the electron density \( n_e \),

\[
\omega_p = \frac{\epsilon_0 m_e}{e^2 n_e}.
\]  

(1.8)

• Electron gyrofrequency \( (\Omega_e) \): the natural resonant frequency of the electrons under a constant magnetic field,

\[
\Omega_e = \frac{eB_o}{m_e}.
\]  

(1.9)

1.3.2 Capacitor Model

The most common geometry used for impedance probes used short dipoles and monopoles. These probes are electrically short, and therefore its free space impedance is purely capacitive. As a first approach, we can consider the probe as a capacitor filled with dielectric plasma (see Fig. 1.3). It can be shown that the impedance under this condition follows the equation

\[
C(\theta) = C_0 \frac{1}{2} [\epsilon_1 (1 + \cos^2(\theta)) + \epsilon_3 \sin^2(\theta)],
\]  

(1.10)

where \( \theta \) is the angle between the probe and the magnetic field and \( C_0 \) is the free space capacitance for a cylindrical capacitor with inner radius \( a \) and outer radius \( b \) given by

\[
C_0 = \frac{2\pi \epsilon_0 L}{\ln\left(\frac{b}{a}\right)}.
\]  

(1.11)

When the probe is excited at low frequencies between 100 kHz and 20 MHz, only the electrons are capable of moving, while the ions that are heavier than the electrons remain static. The electron-neutron collision frequency in the altitudes of interest (E and F regions) is only a few kHz, much lower than the cyclotron frequency and can be eliminated from the
expression. With these considerations the impedance is

\[
Z = \frac{2}{j\omega C_0 \left[ \left( 1 + \frac{\omega_p^2}{\eta^2 - \omega^2} \right) \cos^2(\theta) + \left( 1 - \frac{\omega_p^2}{\omega^2} \right) \sin^2(\theta) \right]}.
\]  \hspace{1cm} (1.12)

The impedance is shown in magnitude and phase in Fig. 1.4. Two resonant frequencies are observed.

- A low impedance resonance (similar to a R-L-C series resonant circuit) at the electron gyrofrequency.

- A high impedance resonance (similar to a R-L-C parallel resonant circuit) at the upper-hybrid frequency. This frequency is the combination of the plasma frequency and gyrofrequency

\[
\omega_{uh}^2 = \omega_p^2 + \Omega_e^2.
\]  \hspace{1cm} (1.13)

The electron density can be obtained by measuring the upper-hybrid frequency as follows

\[
n_e = \frac{4\pi \epsilon_0 m_e}{e^2} (f_{uh}^2 - f_c^2),
\]  \hspace{1cm} (1.14)

where the only unknown parameter is the electron gyrofrequency. Because it is only dependent
of the magnetic field, it can be calculated from magnetic field models (IGRF), or with the use of local measurements of the magnetic field.

1.3.3 Balmain Model

The capacitor approach fails to account for the effects of the plasma on the charge distribution on the probe surface. It is also limited to cold plasmas, but the general impedance shape and resonant frequencies are the same as other models, and it is very intuitive. More complex models have been developed over the years, with Balmain’s model [4] being the most accepted and used. It treats the probe as an antenna with an assumed triangular current distribution. The modeled impedance can be seen in Fig. 1.4, and has been validated several times in plasma chambers as well as sounding rocket flights. The impedance expression is as follows

\[
Z = \frac{a}{j\omega 2\pi \varepsilon_0 \varepsilon_1 L F^{1/2}} \left[ \ln \frac{L}{\rho} - 1 - \ln \frac{a + F^{1/2}}{2F} \right],
\]

(1.15)
in which \( F = \sin^2 \theta + a^2 \cos^2 \theta \) and \( a^2 = \varepsilon_1 / \varepsilon_3 \).

![Fig. 1.4: Impedance of a small antenna in plasma.](image-url)
Another approach is to numerically solve for the impedance of an antenna in a plasma. This has been done using a Finite Difference Time Domain (FDTD) model and plasma fluid equations [5]. One of the key observations of this model is that when near the upper hybrid resonance the current distribution on the antenna deviates from the triangular shape as was assumed by Balmain. The result is that the magnitude of the impedance is an order of magnitude less than what is predicted by the analytical theories.

The impedance or radio frequency (RF) methods have several advantages over Langmuir probes. They are loosely dependent on the rocket or spacecraft surface charge (providing an absolute measurement), the magnetic field orientation, or the plasma temperature. Although the free space capacitance changes with the geometry of the probe, the resonant frequencies remain unchanged, making this kind of instruments ideal and accurate for plasma diagnostics. The upper-hybrid frequency usually has a high quality factor, being very sharp, and the phase crosses zero, making it very convenient and simple for instruments to track this frequency.

One source of error of this kind of probes is the shunt capacitance that is in parallel with the probe and sense electronics input. It is the total capacitance of the sensor geometry that is not filled with plasma and is electrically parallel with the actual probe input. The physical cause of the shunt capacitance are the stray capacitances of traces or wires inside the instrument before its connection to the probe input. The result is a reduced sensitivity to the plasma effects to be observed. It can be effectively mitigated with electrical guard designs as will be explained in the following chapters.

1.4 RF Impedance Probe Methods

Several approaches have been used over the years for measuring the parallel resonance of an impedance probe which is associated with the upper hybrid frequency of a plasma. They can mainly be classified into two groups, those that sweep and measure the magnitude of the impedance curve, and those that observe the phase change associated with the parallel resonance.
1.4.1 Plasma Frequency Probe

The Plasma Frequency Probe (PFP) measures the upper-hybrid frequency by tracking when the phase crosses zero as the resonant frequency is changed. The usual method is some kind of analog Phase Locked Loop (PLL), although there are several variations. They perform a wide sweep to locate the resonant frequency and then they lock in. If for any reason the lock is lost, the search procedure is started again. It has also been implemented a Sweeping PFP, where a continuous sweep is performed until the resonant frequency is found. The number of steps taken in the sweep is recorded and the procedure begins again.

1.4.2 Sweeping Impedance Probe

The Sweeping Impedance Probe (SIP) excites the probe with a constant voltage, and measures the current flowing. The method for measuring the current has been changed over the years as new electronic components became available, and with new innovative designs. They normally measure it in quadrature to obtain the complex impedance. Because the measure is proportional to the admittance, they are sometimes referred as admittance probes (and it is usually more convenient for expressing errors or saturation ranges). When the measurement is finished, the frequency is swept over the full range, obtaining a very detailed impedance curve. The main disadvantage is the lower sweep rate compared with the PFP and the huge amount of telemetry required for a single sweep.

Only the variations of the impedance with respect the free space capacitance $C_0$ are of interest. Several techniques have been used trying to subtract $C_0$ from the probe impedance, obtaining a narrower range and increased sensibility. However, determining $C_0$ is not easy because it requires a rocket mock-up hanging in free space in the absence of any element for many meters.

The latest SIP instruments flown are the STORMS mission [6], and the Japanese S-520-26 [7], in 2012.

1.5 ASSP Mission

The Auroral Spatial Structures Probe (ASSP) [8] is a NASA mission intending to
study the energy flow around the aurora. It will consist of a sounding rocket, which ejects six subpayloads during the flight. This will be the first mission directed by NASA with a constellation deployed from a sounding rocket. This formation will take measurements simultaneously providing unique spatial and temporal variations of the different magnitudes. ASSP is scheduled to be launched in the early 2015 from Poker Flats Flight, and will measure the ionosphere under aurora conditions from 110 km to 606 km.

The ASSP instruments are being developed and built by SDL-USU. All seven payloads include several instruments such as electric field, magnetometer and a Langmuir probe. Additionally, the main payload in the rocket includes a SIP and a Multi Fix-Bias Langmuir Probe. The position of the different instruments can be seen in Fig. 1.5.

The ASSP SIP is implemented using a new architecture, intending to be more accurate
and fast than the previous generations by using the latest electronic components and techniques, and a very careful design and calibration methodology. The ASSP mission is ideal for a SIP due to the high altitude, which will cover regions E and F.

1.6 Thesis Outline

The objective of this thesis is to present the work done and the lessons learned during the development of the new SIP architecture.

Chapter 1 has covered the ionosphere history, past measurements, in particular the RF methods, where an introduction of the plasma physics behind these instruments is given for clarity. Lastly, the ASSP mission has been introduced where the SIP instrument will be used. This chapter should give the reader a general idea of the motivation behind this project and thesis.

The ASSP SIP overview is covered in Chapter 2, including the requirements driving the design and the theory of operation of the instrument.

Chapter 3 details the different components and units comprising the ASSP SIP. The different analyses during the design are presented here showing the performances achieved compared with the requirements.

SIP testing, calibration, and performances results are presented in Chapter 4. Calibration is a critical step during testing, based on the error model analyzed, well known calibration loads, and a careful methodology.

In Chapter 5, the conclusions of this project are described, giving the reader a summary of the achievements and tasks that could be improved in the future.

As part of any engineering job, multiple design files have been generated. These are included in the different Appendices for reference. Appendix A contains the current transformer analyses and testing. The design schematics of the different boards can be found in Appendix B. The digital design implemented in the Field Programmable Gate Array (FPGA) is shown in Appendix C. Appendix D includes the characterization of the calibration loads. Software were developed to operate the test equipment and the instrument, and they are detailed in Appendix E. All the information related to this thesis is included in
electronic format in the DVD attached. Its contents are described in Appendix F.
Chapter 2

Instrument Overview

2.1 Vector Network Analyzer Concept

The Sweeping Impedance Probe developed for the Auroral Spatial Structures Probe sounding rocket is based on concept and designs developed by the amateur radio community for vector network analyzers. This community has developed a variety of low-cost vector network analyzers (VNA) and released the designs and build experiences on the Internet. Assembled or unassembled kits for these VNAs can be purchased. These VNAs are able to make transmission and reflection measurements from 0.05 to 60 MHz, with about 0.035 Hz frequency resolution and over 110 dB of dynamic range. This capability spans the needs for an impedance probe. The ASSP SIP is greatly influenced by the VNA developed by the radio amateur with call sign N2PK and is referred to as the N2PK VNA.

The block diagram for the N2PK VNA [9] is presented in Fig. 2.1 where a Device Under Test (DUT) is shown undergoing a simple transmission or Wheatstone type reflection bridge measurement. The RF Direct Digital Synthesizer (DDS) block generates an RF voltage at a $0^\circ$ reference phase which is applied to the input of the DUT. The output from the DUT to the RF Detector input is a signal to be measured with a given amplitude and phase. In addition, the RF signal at the Detector input is measured with a short length of transmission line in place of the DUT that is assumed to have unity gain and zero phase, and with an open at the Detector input. From these three vector measurements at a single frequency, all DUT transmission characteristics, such as gain and phase, can be calculated [9]. In the same way, impedance and reflection characteristics of the DUT can be measured with the use of a reflection bridge. In this case, three accurate terminations, open, short, and load, are used as references.

The N2PK VNA uses a narrowband direct-conversion architecture to convert the
Fig. 2.1: Block diagram of the N2PK network analyzer.

detected signal to DC base band through mixing. The DC voltage is dependent not only on the magnitude of the RF voltage at its input, but also its phase relative to the RF signal at the local oscillator (LO) input. Highly accurate measurements of this amplitude and phase dependent DC voltage are obtained using a precision linear analog detector, a 24-bit analog-digital converter, and precise phase control of the LO DDS. The phase information is obtained by making two sequential DC measurements for each frequency and test condition (reference loads and DUT). In each case, the first measurement is made with the LO at the reference phase of 0°; the second measurement is made with the LO phase shifted by 90°. This process results in the quadrature or vector components of each signal at the Detector RF input.

2.2 Sweeping Impedance Probe Requirements

A Sweeping Impedance Probe operating in the Earth’s ionosphere has a number of requirements that are significantly different than the N2PK VNA, which is primarily used for measuring the performance of HF antennas and communication systems. These requirements labeled R1 through R6 are presented below with discussion.
R1 The frequency of operation shall be between 1 MHz and 20 MHz and shall be programmable.

When the ionosphere is driven near the local electron cyclotron frequency significant energy can be transferred to the plasma heating the electrons and disturbing the medium to be probed. On previous SIP instruments the sweep range had started as low as 0.1 MHz and strong interference was generated for all diagnostic instruments on the sounding rocket through a process known as sheath rectification when the SIP was at the low end of its sweep. The low end of the sweep needs to be adjusted based on the expected cyclotron frequency of plasma to be measured.

R2 The number of sample points in a frequency sweep shall be at least 128 points.

Previous experience with impedance probes on the STORMS mission has shown that this is a sufficient number of samples to locate the various resonances of the probe-plasma system [6]. In general, the points are distributed linearly in two band with higher resolution at low frequencies where the upper hybrid frequency is expected and few points at the highest frequency ranges.

R3 The sweep rate shall be at least 10 Hz.

The sweep rate determines the along track resolution of the measurements. Given that the plasma environment is dynamic, it is desirable to accomplish the measurement before the underlying media changes significantly.

R4 The admittance measurement shall have an accuracy of greater that 1% in magnitude and 1° in phase across the measurement range.

It is not essential that the SIP be extremely accurate because of the morphology of the parallel resonance associated with the upper hybrid frequency. A small error in magnitude or phase does not change the estimation of the frequency of this resonant condition significantly. However, it is often desirable to obtain as noiseless and accurate values as possible. These requirements were made base on the design capabilities and their relative impact on other important parameters like power, size, and mass.
Because the SIP is actually an admittance probe, where the magnitude sensed is the current proportional to the admittance, it is more convenient to express the accuracy requirement or the error model in terms of the admittance instead of impedance. However, because traditionally impedance has been used more and it is a more intuitive magnitude, both terms will be used in the thesis.

**R5** The magnitude range of the impedance measurements shall be between 100$\,\Omega$ and 100$\,k\Omega$.

This is one of the driving requirements for the SIP because most network analyzers operate relatively near a reference frequency of 50$\,\Omega$. The need to observe the high impedance of the plasma-antenna parallel resonance drives this extended operating range.

**R6** The maximum voltage presented at the probe input shall be less than 0.5 V peak.

The ambient plasma can be driven to nonlinear states such that the simple linear theory for the impedance of an antenna in plasma is not applicable. The desire to drive the plasma harder to improve the signal to noise ratio for the instrument need to be balanced by the need to not unduly perturb the plasma. This level is based on best practices from years of impedance probe work [10].

### 2.3 Block Diagram

The ASSP SIP instrument is composed of the functional blocks indicated in Fig. 2.2. The electronics are divided into two boards. The Main Board is located inside the rocket and contains the major functionality. It includes the VNA, which produces the constant voltage sine wave for exciting the probe, and the detector to acquire and digitize the measured current in quadrature, coming from the RF Head. All the VNA sequencing and frequency sweep is managed by the Controller, which is also in charge of the digital processing and packing the values into the Telemetry (TM) format. The different electronic components are powered from a single high-voltage bus. The Power Conditioning and Filtering stage generates all the voltages required.
The second board is inside the RF Head. It is located inside the antenna boom, and closely placed to the feed. It connects the VNA excitation voltage to the probe, while measures the current through it. It is then amplified by a Low-Noise Amplifier (LNA), and transmitted to the VNA for its acquisition.

2.4 Theory of Operation

The Sweeping Impedance Probe is based on the N2PK VNA, which has proven over 10 years to be very accurate and similar to commercial equipments in performance. By basing the SIP off an existing design, it is ensured it has the required simplicity and compaction to be implemented as part of a flight instrument. The RF Head is based on the N2PK design, but more heavily on the experience of previous SIP instruments developed at Utah State University. Both of them have been adapted for the SIP admittance ranges, sweep rate, and accuracy requirements. The error model has been extensively analyzed compared with the N2PK VNA, which relied more on prototype measurements.

The way the SIP VNA operates is indicated in Fig. 2.3. A DDS will output from one of its channel a sine wave of the desired frequency. After removing the harmonics produced by the DDS with the antialias filter, the signal will be ideally a pure sine wave

\[
V_{RF}(t) = A \sin(2\pi f_0 t),
\]

where \(A\) is the amplitude of the sine wave, and the initial phase can be considered zero.

![ASSP SIP block diagram](image-url)
without loss of generality.

This voltage will be directly applied to the probe. In the RF Head a current sensing circuit is implemented consisting of a Current Transformer (CT) and a LNA. Considering a CT with a turn ratio of $n$, a burden resistor $R_b$, and an amplifier of gain $G_{\text{amp}}$, the incoming signal to the VNA has the following expression:

$$V_{\text{sens}}(t) = \frac{I(t)}{n} R_b G_{\text{amp}} = \frac{A|Y(f_0)| R_b G_{\text{amp}}}{n} \sin(2\pi f_0 t + \arg(Y(f_0))). \quad (2.2)$$

In the VNA, a second DDS channel will generate a sine wave with the same frequency and phase as the first channel. It is multiplied with the incoming signal from the RF Head producing the following output:

$$V_{\text{det}}(t) = G_{\text{det}} V_{\text{sens}}(t) V_{\text{LO}}(t) = G_{\text{chain}} |Y(f_0)| [\cos(\arg(Y(f_0))) - \cos(4\pi f_0 t + \arg(Y(f_0)))], \quad (2.3)$$

where $G_{\text{det}}$ and $G_{\text{chain}}$ are the detector and the measurement chain gain, respectively,

$$G_{\text{chain}} = \frac{A R_b G_{\text{amp}} G_{\text{det}}}{n}. \quad (2.4)$$

Fig. 2.3: SIP VNA block diagram and operation.
The doubled frequency signal can be easily removed with a Low-Pass Filter (LPF). The filtered DC signal is digitized with a regular Analog-to-Digital Converter (ADC),

\[ V_{\text{ADC},0} = G_{\text{chain}} |Y(f_0)| \cos(\arg(Y(f_0))). \]  \hspace{1cm} (2.5)

This value corresponds with the in-phase or real component of the admittance.

After that measure has been performed, the DDS second channel will shift its phase by \(90^\circ\),

\[ V_{\text{LO},0}(t) = A_{\text{LO}} \sin(2\pi f_0 t) \rightarrow V_{\text{LO},90}(t) = A_{\text{LO}} \sin(2\pi f_0 t + \frac{\pi}{2}) = A_{\text{LO}} \cos(2\pi f_0 t), \]  \hspace{1cm} (2.6)

that, when multiplied by \( V_{\text{sens}}(t) \) and filtered, gives the following output,

\[ V_{\text{ADC},90} = G_{\text{chain}} |Y(f_0)| \sin(\arg(Y(f_0))), \]  \hspace{1cm} (2.7)

which is the quadrature component or imaginary part of the admittance.

Given both components the admittance can be calculated directly as

\[ |Y(f_0)| = G_{\text{chain}} \sqrt{V_{\text{ADC},0}^2 + V_{\text{ADC},90}^2}, \]  \hspace{1cm} (2.8)

\[ \arg(Y(f_0)) = \tan^{-1}\left(\frac{V_{\text{ADC},90}}{V_{\text{ADC},0}}\right). \]  \hspace{1cm} (2.9)

This operation is similar to the calculation of the components of a vector with respect to base vectors. A graphical representation of the admittance plane can be seen in Fig. 2.4, where the base is formed by the LO different phases.

Compared with other SIP architectures, this method relies on two things: the use of a digital source generator and sequential measurements. Both DDS channels must have the exact same frequency, otherwise the ADC input will be a slowly modulated signal. Moreover, even if we could get two signals with the same frequency, being able to shift the local oscillator by exactly \(90^\circ\) at any frequency is impossible without the use of digital methods.
With respect to the sequential measurements, they provide an excellent accuracy because both values are affected by the same gains and delays. In the STORMS SIP an error of $1^\circ$ or $2^\circ$ was assumed in phase because the in-phase and quadrature components had different tracks and lengths [6]. The importance of the sequential measurements will be even more noticeable when the use of Correlated Double Sampling (CDS) and the reference channel are explained.

From Equation (2.8) we can see that for obtaining an accurate magnitude value the measurement chain gain must be known with very good accuracy. Even more, any drift of this gain will ruin the instrument accuracy. As for the phase, obtained in Equation (2.9), a simplification has been made. Any element in the measurement chain (amplifiers, transformers, mixer) will introduce phase delays that must be corrected. Hence, a very good calibration procedure, including temperature drifts, must be made to ensure final performances. In the next subsections, two methods will be explained to improve the inherent accuracy and alleviate the calibration.

### 2.4.1 Correlated Double Sampling

The first technique described is the Correlated Double Sampling (CDS). In general, the measurements $V_{ADC,0}$ and $V_{ADC,90}$ will have a certain offset error from different components
such as the mixer, amplifiers, and the ADC. The digitized signals can be expressed more generally as

\[ V_{\text{ADC},0} = G_{\text{chain}}|Y(f_0)| \cos(\arg(Y(f_0)) + V_{\text{off}}), \quad (2.10) \]

\[ V_{\text{ADC},90} = G_{\text{chain}}|Y(f_0)| \sin(\arg(Y(f_0)) + V_{\text{off}}), \quad (2.11) \]

which results in a measurement error of the calculated admittance,

\[ |Y(f_0)| = G_{\text{chain}}\sqrt{(V_{\text{ADC},0} - V_{\text{off}})^2 + (V_{\text{ADC},90} - V_{\text{off}})^2}, \quad (2.12) \]

\[ \arg(Y(f_0)) = \tan^{-1}\left( \frac{V_{\text{ADC},90} + V_{\text{off}}}{V_{\text{ADC},0} + V_{\text{off}}} \right). \quad (2.13) \]

By taking two additional measurements using LO phases of 180° and 270°,

\[ V_{\text{LO},180}(t) = A_{\text{LO}} \sin(2\pi f_0 t - \pi) \rightarrow V_{\text{ADC},180} = -G_{\text{chain}}|Y(f_0)| \cos(\arg(Y(f_0))) + V_{\text{off}}, \quad (2.14) \]

\[ V_{\text{LO},270}(t) = A_{\text{LO}} \sin(2\pi f_0 t - \frac{3\pi}{2}) \rightarrow V_{\text{ADC},270} = -G_{\text{chain}}|Y(f_0)| \sin(\arg(Y(f_0))) + V_{\text{off}}, \quad (2.15) \]

two new values are obtained with a similar offset, but with the opposite value for the signal of interest. Subtracting these new measurements from the original ones results in a value free of offset error (see Fig. 2.5). In reality, the offset can change slightly between the measurements, but after the CDS correction it will be much lower. The residual offset error will be corrected by calibration. It is important to remark that the CDS technique is only possible by acquiring sequential measurements using the same acquisition chain.

For sake of simplicity, the corrected values with CDS are defined as

\[ V_I = \frac{V_{\text{ADC},0} - V_{\text{ADC},180}}{2}, \quad (2.16) \]

\[ V_Q = \frac{V_{\text{ADC},90} - V_{\text{ADC},270}}{2}. \quad (2.17) \]
2.4.2 Reference Channel

The use of a reference channel has been used in many of the previous SIP instruments. Measuring the reference channel duplicates the number of measurements, but it also provides several advantages, correcting gain errors and phase delays. The impedance is calculated as follows:

\[
|Y(f_0)| = \sqrt{\frac{V^2_{I,\text{ant}} + V^2_{Q,\text{ant}}}{V^2_{I,\text{ref}} + V^2_{Q,\text{ref}}}} |Y_{\text{ref}}|, \quad (2.18)
\]

\[
\arg(Y(f_0)) = \tan^{-1}(\frac{V_{Q,\text{ant}}}{V_{I,\text{ant}}}) - \tan^{-1}(\frac{V_{Q,\text{ref}}}{V_{I,\text{ref}}}). \quad (2.19)
\]

In the graphical plane, shown in Fig. 2.6 for a resistive reference, it can be seen as both measurements are affected by the same delay that can be corrected. If gain errors affect both channels in the same way, the triangles formed by the real admittances, and the ones affected by the gain error are similar.

While the gain and phase delay of the measurement chain can be effectively corrected, this is only true if the reference impedance is very stable and very well known.
2.4.3 Summary of Instrument Operation

The use of sequential measurements provide a very good inherent accuracy for the instrument. Most of the systematic errors (offset, gain, and phase delay) and their drifts are corrected. The residual errors can be corrected by calibration as explained in Chapter 4. However, the main disadvantage is the sequential use of the ADC, requiring a device with a high sampling rate. To allow a high sampling rate, the input signal cannot be filtered without impacting the settling time. Thus, there is a trade-off between the sweeping rate and the maximum admissible noise in the system. The noise analysis has been one of the main concerns for the selection of components. Appropriate power distribution and grounding is essential to obtain a reduction of the Electromagnetic Interference (EMI) noise.

There are several sources of nonlinearity errors in the system as will be explained. Although they may be systematic, its calibration is often difficult and tedious. Therefore, the main objective has been minimizing its influence at the source, in order to do a simpler linear calibration.
Chapter 3
Design and Analyses

This chapter explains the different design decisions taken during the component selection and their main parameters. It also presents the analyses that were performed to ensure the final performance of the instrument.

3.1 Detailed Design

The detailed design section should provide enough insight to the reader to perform the next SIP generation instruments based on this architecture. The order of the subsections tries to replicate the flow of the signal from the source (DDS) up to the final destination (ADC).

3.1.1 Direct Digital Synthesizer

The DDS must output two signals synchronized in frequency. For that matter two DDS chips can be used with the same clock oscillator, or like in this case, a single 2-channel chip. The DDS selected presents several advantages against others.

- It uses a serial port for configuring the DDS, that while slow, occupies less pins in the Field-Programmable Gate Array (FPGA).
- It supports different kinds of modulations by direct action on some special pins. In this case for the LO channel, it is very convenient for changing its phase.
- It presents a low phase noise and Spurious-Free Dynamic Range (SFDR).

It also has disadvantages like the unique voltage used, 1.8 V, or having more functionality than needed in this application.
The DDS output, shown in Fig. 3.1, consists of a complementary differential current sink, which must be converted with a balun for driving the unbalanced antenna impedance. R147 sets the DC voltage of the outputs around 1.55 V, and R148 helps maintain the maximum output voltage inside their compliance range (there can be problems driving a filter with inductive input impedance).

While the use of a DDS is very convenient for frequency and phase accuracy, they have larger phase noise and harmonics than other signal generator methods. In order to reduce the harmonics and the digital aliases, a reconstruction filter is implemented at the output with a frequency cut-off of 25 MHz. When the generated signal has a low frequency, not every harmonic can be filtered. The filter is implemented with passive components to avoid any additional noise or error. Additionally it is known that DDS produce higher harmonics when the output frequency is a sub-multiple of the reference clock frequency, 64 MHz. In those cases, the frequency has been slightly shifted (1 LSB introduces a shift of less than 1 Hz, negligible for the instrument operation).

Because the output of the DDS is the result of a reconstructed digital sine wave, it follows a $\text{sinc}(f_0/f_{\text{REF}})$ response. There is a small decay of 15% when the output is at 20 MHz that will be considered as part of the gain error.

![Fig. 3.1: DDS output configuration.](image-url)
3.1.2 RF Head

The RF Head block diagram can be seen in Fig. 3.2. The excitation voltage received from the Main Board drives the antenna through a Current Transformer (CT) that senses the current. It is converted to voltage in the secondary with a burden resistor and an amplifier closely placed augments the signal to appropriate levels for its transmission to the Main Board.

The reference channel is configured in the same way replacing the antenna by a fixed and very accurate resistor. In this sense two changes have been made with respect to previous SIP instruments.

- First, the reference impedance selected is mechanically less representative, but on the other hand is much more accurate and stable having an initial tolerance of 0.01% and a Temperature Coefficient of Resistance (TCR) of 20 ppm/°C.

- With the exception of the CT, burden resistor, and amplifier, all the other elements of the measurement chain are the same to the antenna channel. And all the mentioned components have been implemented as symmetrically as possible to the antenna channel.

- The CT for each channel are positioned perpendicular to each other, avoiding magnetic coupling between them.

The CT analysis and testing took a fair amount of time, and all the know-how has been compiled in Appendix A. The final configuration is a 1:5 turns ratio transformer with a burden resistor of 50Ω. To avoid phase differences between channels, an amplifier with very high Gain-Bandwidth Product (GBW) has been selected. Additionally, only one signal is sent back to the Main Board with the help of a multiplexer made from RF switches. In this way, the uncertainty in the phase delay produced by the cable is eliminated. The amplifier is configured in a non-inverting configuration to increase the input impedance compared with the burden resistor. Small resistors values have been used to reduce their thermal noise.
There always exist stray capacitances between the antenna feed and ground. There are also border effects of the antenna. Figure 3.3(a) shows the three possibilities. Stray capacitances before the CT load the driver, but it does not produce any output on the CT. However any current after the CT and returning through the enclosure or other ground track, will be sensed by the CT and produce an undesirable output. All these capacitances combined are the shunt capacitance referred in the introduction, which reduces the instrument sensitivity. To mitigate this problem the use of a guard is implemented, where the enclosure is grounded through the primary wire shield, which is also coupled to the CT. As seen in Fig. 3.3(b), any current through these stray capacitances or from the antenna to the guard return through the shield canceling its effect on the output.

The guard design has been improved over previous instruments obtaining a very enclosed guard. At the moment of writing this thesis, a new design based on an insulated material (green in the figure) coated with a high conductivity spray (orange in the figure) is being studied by the mechanical team. It will provide a very good mechanical fixation for the guard.
and the antenna, a continuous and conductive guard from the connector to the enclosure and will ease its manufacturing. This new guard design together with a very close transformer to the antenna connector (just a few centimeters) should reduce the shunt capacitance effect to a minimum.

3.1.3 Detector and ADC

The detector is a Direct-Conversion Receiver (DCR) (or zero-IF receiver), consisting of a frequency shifter by combining the current sensed with the LO signal from the DDS, both of the same frequency. For the detector there are two possible designs.

- A frequency multiplier or linear mixer. It generates the sum and difference frequencies with much less harmonics. The main disadvantages are that LO signal noise is present in the output (together with its mixing with RF noise [11]) and they have less conversion gain.

- A frequency mixer or modulator. In these kind of devices the LO input is overloaded.

![Fig. 3.3: Mitigation of shunt capacitance with a guard.](image)

(a) Stray capacitances effect without guard.

(b) Stray capacitances effect with guard.
into a square signal and acts like a switch control for the RF signal and its opposite. The main advantage is that it is much less sensible to the LO noise and has higher conversion gains, but it generates multiple harmonics and intermodulation products.

For the ASSP SIP a conservative approach has been taken. The AD831 active mixer has been selected having a high conversion gain with low harmonics. The only disadvantage is its elevated power consumption.

The differential output of the mixer is filtered and buffered for driving the ADC. The frequency cut-off of this filter is critical for the instrument timing. The fundamental requirement is that it must be much lower than 2 MHz (twice the minimum frequency), because the mixer will produce an output with the same level as the DC signal we want to digitize. Furthermore, to reduce aliasing and limit the noise, the filter must reject frequencies higher than 200 kHz, which is half the sampling frequency when oversampling (see Section 3.1.4). However, reducing the bandwidth increases the settling time of the filter after a change of phase or frequency, being the major contributor to the instrument measurement time. The final bandwidth selected is around 100 kHz, obtaining a good rejection at 2 MHz of 25 dB. With the settling time configured to 9 ms, the value has been settled to 99.92 % of the final value (in average of the four coadded samples).

The ADC is also a critical element of the system for determining the dynamic range. A 24-bit Delta-Sigma ADC would have been preferred but they have a very limited sampling rate, insufficient for this application. A very accurate 16-bit ADC with a higher sampling rate has been used.

### 3.1.4 Controller

The Controller is implemented in an Igloo FPGA (AGL1000V5). The main block diagram of the digital design can be seen in Fig. 3.4.

- The Data Acquisition System (DAS) Controller acquires all the different information to be transmitted by Telemetry (TM), it controls the SIP instrument, changing the frequency and phase of the DDS and acquiring the ADC, and acquires houskeeping
information.

- The Ground Support Equipment (GSE) Interface implements a Universal Asynchronous Receiver/Transmitter (UART) interface for on-ground debugging and testing.

- The Payload Controller modifies the acquisition modes of the DAS Controller according to the commands received from the GSE.

- The Telemetry Matrix Former groups the information from the DAS Controller and other information relevant to the status of the SIP instrument, and transmits it to the NASA TM&Power unit. From this unit, some synchronization signals are received which are used for triggering the acquisition of the telemetry.

- The Flash interface implements a counter that is incremented after each reset.

Many of the blocks have been reused from another programs, or from other ASSP units some modifications. The DAS Controller is specific of the SIP instrument. The complete Register-Transfer Level (RTL) diagrams can be found in Appendix C.

The general timing of the SIP acquisition can be seen in Fig. 3.5. The SIP instrument has nine 16-bit channels allocated in each Sub Frame (SF), where the raw data of a single frequency is transmitted. A Major Frame (MF) consists of 32 SF, and the SIP instrument requires four complete MF to transmit a single sweep data (128 frequency points). Being the SF period 140μs, and the MF period 5.88 ms, this results in a maximum sweep rate of about 42.5 Hz. Because of the huge amount of allocated data rate, it was decided to push the design, increasing the sweep rate at the expense of reduced accuracy. The NSROC Payload from NASA generates the timing signals indicating to the SIP Payload the beginning of SFs and MFs. This signals are used for synchronizing and timing the SIP operations. After the DDS initial configuration, the FPGA remains idle until the next MF. Then, it starts acquiring each frequency values at the beginning of each SF, and stores them into a memory. These values are transmitted in the next MF, i.e. data is delayed by one MF period but time-tagged for further post-processing.
Fig. 3.4: FPGA design block diagram.
Fig. 3.5: FPGA and SIP overall timing diagram.
It should be noted that there is an overlap between SFs. While the ADC has sampled the data and initiates the conversion phase, the next frequency and phase is loaded. It should also be noted that because frequency is loaded with a serial line it requires some time. To avoid any additional delay to the settling time, frequencies values are preloaded in the previous SF and loaded immediately into the DDS registers at the beginning of the SF.

With the limited processing made in the FPGA, the amount of logic needed is very limited too. In Table 3.1, it is shown a comparative of different resources between the specific SIP modules, the complete FPGA design and the total amount provided by the FPGA. The Random-Access Memory (RAM) blocks utilization could be optimized by delaying the values acquisition by only two SF instead of a MF. It can be seen that the SIP digital design does not require a high demand of the FPGA and it could be integrated together with other instruments.

For debugging purposes, different test modes have been implemented in the FPGA design which are only accessible through the UART. These test modes allow to establish the VNA in a particular frequency (which in VNA is called zero-span), with/without CDS or the amount of coadded samples.

### 3.1.5 Power Conditioning and Filtering

All the components mentioned previously are powered from a single high-voltage and high-power unregulated bus. Different voltage rails are generated and conditioned from the primary bus, as shown in Fig. 3.6, by using DC/DC converters, Low-Dropout Regulators (LDOs), and references. The power consumption associated with each converter is summarized in Table 3.2. The VNA dissipates around 1.3 W with 1 W coming from the mixer. The RF Head dissipates 230 mW due to the amplifiers. The Controller and

<table>
<thead>
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<th>Module</th>
<th>SIP Controller</th>
<th>Complete Design</th>
<th>Total FPGA</th>
</tr>
</thead>
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<td>3131 (13%)</td>
<td>24576</td>
</tr>
<tr>
<td>I/O pins</td>
<td>19 (11%)</td>
<td>51 (29%)</td>
<td>177</td>
</tr>
<tr>
<td>RAM blocks</td>
<td>19 (59%)</td>
<td>28 (87%)</td>
<td>32</td>
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</table>

Table 3.1: SIP modules occupation on the FPGA.
housekeeping circuits take only 220 mW and all the remaining power is lost in the converters, which are working at a very low load with low efficiency.

### 3.2 Analyses

Different analyses were made for the selection of components and assuring performances. In particular, the driving requirements for the design were the impedance range, and measurement uncertainty in magnitude and phase. This measurement error will be determined by two different kind of errors, systematic and random errors. Systematic errors are comprised by:

- **Offset error**: mitigated by the CDS technique. The offset voltage limits the dynamic range of the ADC, but with enough margin it is not a critical parameter. The residual error is difficult to predict and it will be considered only during testing and calibration.

- **Gain error**: mitigated by the reference channel. Like the offset error, it will be considered later during testing and calibration.

Fig. 3.6: Power generation and conditioning from the primary bus.
Table 3.2: Power consumption of the SIP instrument.

<table>
<thead>
<tr>
<th>Component</th>
<th>+1.5VD</th>
<th>+3.3VD</th>
<th>+5VD</th>
<th>+1.25VA</th>
<th>+2.5VD</th>
<th>+1.8VA</th>
<th>+3.3VA</th>
<th>+5VA</th>
<th>-5VA</th>
<th>+15VA</th>
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<td>3.4 W</td>
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</tbody>
</table>

Note: all values in mA unless otherwise noted.
• Nonlinearities: there are several sources of nonlinearities and each one requires a different approach that will be explained. In particular, the loading effect will affect the gain of the instrument for different admittance ranges.

These errors are repetitive under the same operating conditions and will be ultimately calibrated. On the other hand random errors will be present.

- Noise: it has been properly limited by design in order to assure it does not degrade the measurement when acquiring low signals.

- Resolution: although limited by the device used, some techniques can be used to improve the effective resolution.

- Nonlinearities: some of the nonlinearities are not only dependent on the value being measured but on other characteristics of the DUT or the test equipment. An example would be the harmonic distortion.

Some of the random errors presented, like resolution or the ADC Integral Nonlinearity (INL), are not really random but dependent on the specific component. However, they are considered random because of its difficulty to be calibrated.

3.2.1 Noise

Because of the high sampling rate, noise was one of the main concerns, and as such, a detailed staged analysis was performed in order to select the components. During this analysis the noise values are provided in dBc/Hz of the Single-Sideband (SSB) noise with respect to the fundamental (i.e. graphs represent the offset frequency with respect to the fundamental frequency and with the relative magnitude value to the fundamental magnitude integrated over 1 Hz bandwidth). This representation is convenient for this application where the frequency is swept.

The phase noise of the DDS can be expressed as [12]

\[
L_{DDS}(f, r) = \frac{1}{2} r^2 L_{CLK}(f) + \left(\frac{r}{r_R}\right)^2 L_{1/f}(f, r_R) + \kappa(r) L_{floor},
\]

(3.1)
where \( r = \frac{f_0}{f_{CLK}} \) is the ratio of the output frequency and the reference clock, and \( \kappa(r) \) is a weak function of \( r \). It has three components:

- The phase noise from the reference clock \( L_{CLK}(f) \);
- The \( 1/f \) or flicker noise from DDS \( L_{1/f}(f) \), which is given at a particular reference ratio \( r_R \);
- Floor white noise of the output DAC of the DDS \( L_{floor} \).

Shown in Fig. 3.7 is the phase noise at the output of the DDS for the worst case, i.e. the maximum output frequency. Because it is an accessible point, during testing it can be observed the expected performance of the DDS.

The signal will be sensed by the CT and amplified by the LNA. Because of the low signals in this stage, the noise of the components affect in great measure the total amount of noise. Shown in Fig. 3.8 are the different contributors, including the burden resistor thermal noise, the amplifier, and the resistors used for the amplification. As it can be observed, the LNA used has very low floor noise (its flicker noise at low frequencies is irrelevant because it will be rejected by the balun and mixer), and being the major contributors the thermal noise of the amplifier resistors.

The mixer will introduce its intrinsic noise while rejecting all noise from the LO input. Unfortunately, the datasheet does not provide the flicker noise information, and it is something that should be checked during testing. After that, the noise from the ADC buffer

![Fig. 3.7: Phase noise spectrum at the DDS output.](image-url)
amplifiers (including their flicker noise) is added taking into account an increase of 3dB for the differential path. The ADC noise is negligible according to the datasheet (0.137 LSB), and it has not been considered. The different contributors can be seen in Fig. 3.9.

The use of CDS improves not only the offset rejection but also the flicker noise [13]. It can be considered as a digital filter with a sampling rate equal to the time difference between phases 17µs (see Fig. 3.10). The frequency response is a high pass-filter with a cut-off frequency of $f_s/4$, which in this case is 14.7 kHz. This results in a very good rejection of the low frequency noise as seen in Fig. 3.11.

Although many sources of noise have not been considered like EMI or components noise that is not given in the datasheet, it is expected that their contribution will be negligible. This is the result of a very low bandwidth digitized (and further reduced by coadding) and the use of CDS.
Fig. 3.10: CDS filter response in time and frequency domains.

Fig. 3.11: Phase noise spectrum reduction with the use of CDS.
3.2.2 Nonlinear Errors

3.2.2.1 Loading Effect

The DDS output gets loaded with the different elements shown in Fig. 3.12. Usually, the antenna impedance is much higher than $Z_0$ and $R_b/n^2$, there is no loading effect and the voltage $V_{RF}$ is constant and about 250 mVp. However, as its impedance lowers, the voltage in the antenna will decrease. With 100 Ω, the voltage decreases to 200 mVp, and with 50 Ω to only 110 mVp.

The loading effect is systematic and can be easily corrected. The main concern for the loading effect is that only affects low antenna impedances where the most accurate calibration loads are found. For future designs it would be interesting analyzing the convenience of implementing a low output impedance buffer to drive the antenna and reference taking into account the degradation of noise and harmonics.

3.2.2.2 Harmonic Analysis

The main source of harmonics is the DDS. It has a high SFDR of 65 dBC (see Fig. 3.13(a)). However, if the fundamental is located in the resonant frequency (where the impedance is maximum) and its harmonics are located in the low impedance, they will get amplified to higher levels. Figure 3.13(b) shows the output of the LNA superimposing the

![Diagram](image_url)

Fig. 3.12: Loading effect on the DDS output.
impedance curve. Supposing a resonant impedance of 100 kΩ and an impedance of 1 kΩ for the harmonics, the 65 dBc margin gets reduced to only 25 dBc.

The mixer will produce a DC output of these harmonics when mixed with the LO harmonics. The conversion gains for the different harmonics are shown in Fig. 3.13(c). With the harmonic level at 25dBc below the fundamental and with the mixer rejection of 40dB, it produces an output of -65dBc. Although this value is negligible, it is only achieved by the use of the high-power mixer. With other lower power DDS and mixers, this value can get as worse as -35dBc (1.7% in linear units), and it would require special measures to mitigate its influence.

3.2.2.3 ADC Resolution

The ADC resolution is 16 bits, which may seem not enough for the lowest admittances. However due to noise dithering at the ADC input, and the coadding implemented in the Controller, the effective resolution is enhanced. Each value transmitted quadrature component is the result of eight ADC samples, four coadded values in one phase, and another four for the complementary phase. This results in an effective resolution of 17.5 bits. If datarate is not the limiting factor, future designs should implement higher resolution ADC.

3.2.2.4 ADC Linearity

The main problem of the ADC that cannot be mitigated is its lack of linearity. This ADC has a typical INL of 0.4 LSB (≈ 6 ppm). While the INL of the ADC can be potentially calibrated it requires a very specific test set-up, and it is not practical for this kind of instrument. Oversampling and coadding does not reduce the INL because the oversampled values have the same INL error. However, CDS can reduce the INL by a factor of $\sqrt{2}$, because for this particular ADC the INL has a zero mean value and looks like white noise (is the same consideration as it is normally made for considering the quantization error a white noise). The ADC nonlinearity influence was discovered too late in the design process, when the boards were already manufactured.
Fig. 3.13: Nonlinear error produced by harmonic mixing of the local oscillator with DDS harmonics.
3.2.2.5 Other Sources of Nonlinearity

In general, all components can generate nonlinear behavior near saturation. All the amplifiers are used with low-level signals compared to the supply voltage, and therefore should not present any appreciable saturation effect. This however, should be taken into account if low-power devices are used.

3.2.3 Saturation: Maximum Admittance

The different gains and voltages for the minimum impedance are shown in Fig. 3.14. The voltage at the ADC input is ±0.7 V (for 0° and 180°, respectively), which has been properly adjusted to the ADC reference voltage of ±1.25 V. The LNA gain has been limited to avoid saturation of the mixer RF input. A short-circuit cannot be used as an antenna impedance because it will enter the nonlinear region generating compression on the output signal and excessive harmonics.

3.2.4 Measurement Uncertainty

The real admittance $Y$ will be measured with some error giving the estimation $Y_M$. The admittance error is due to the in-phase and quadrature components errors $\epsilon_I$ and $\epsilon_Q$ as shown in Fig. 3.15. The expression of the magnitude error in percentage is

$$
\epsilon_{mag} = \left( \frac{|Y_M|}{|Y|} - 1 \right) \cdot 100 = \left( \frac{|Y + \epsilon_I + j\epsilon_Q|}{|Y|} - 1 \right) \cdot 100, \quad (3.2)
$$

![Fig. 3.14: Saturation and gain analysis of the measurement chain.](image)
and it can maximized for the minimum admittance and for the error aligned with the admittance vector as in Fig. 3.15,

$$\max(\epsilon_{\text{mag}}) = \pm \left( \frac{\min(|Y|) + \sqrt{\epsilon_I^2 + \epsilon_Q^2}}{\min(|Y|)} - 1 \right) \cdot 100 = \pm \frac{\sqrt{\epsilon_I^2 + \epsilon_Q^2}}{\min(|Y|)} \cdot 100. \quad (3.3)$$

The phase error is calculated as the angle between both vectors,

$$\epsilon_{\text{ph}} = \arg(Y_M) - \arg(Y), \quad (3.4)$$

and the worst case is found for the minimum admittance and the errors aligned 90° to the admittance vector (see Fig. 3.15),

$$\max(\epsilon_{\text{ph}}) = \pm \tan^{-1} \sqrt{\frac{\epsilon_I^2 + \epsilon_Q^2}{\min(|Y|)}} \approx \pm \sqrt{\frac{\epsilon_I^2 + \epsilon_Q^2}{\min(|Y|)}}. \quad (3.5)$$

---

Fig. 3.15: Admittance gain and phase error as a function of the quadrature components errors.
Assuming that the in-phase and quadrature errors will have the same standard deviation, the previous expressions can be simplified as

\[
\max(\epsilon_{\text{mag}}) = \pm \frac{\sqrt{2}\epsilon}{\min(|Y|)} \cdot 100, \tag{3.6}
\]

\[
\max(\epsilon_{\text{ph}}) = \pm \frac{\sqrt{2}\epsilon}{\min(|Y|)}. \tag{3.7}
\]

Assuming a perfect calibration of the systematic errors (offset, gain, loading effect) the measurement uncertainty will be limited by the random errors, noise, resolution, and nonlinearities (see Table 3.3). This assumption is not entirely true, because the calibration has some issues that will impact the performances. They are explained in the following chapter.

The measurement uncertainty has been calculated to predict the compliance with the accuracy requirements. Without taking into account the ADC, the requirements can be met with margin because of the new architecture with great selectivity and dynamic range, and with a very conservative selection of components. All the elements (DDS, amplifiers, mixer, etc.) have extremely good performances at the expense of an elevated power consumption. The ADC lack of resolution and linearity is the only design drawback, but it has been analyzed and it should be noted and corrected for future instruments. It is believed that all the analyses and considerations that have been collected here should prove a resourceful guide for the next SIP instruments based on this architecture.

<table>
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<tr>
<th>Error</th>
<th>Value</th>
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<tr>
<td>Resolution</td>
<td>±6.74 µV</td>
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<tr>
<td>INL</td>
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<td>Total (RSS)</td>
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<td>Min. Signal ((Y_{\text{min}}))</td>
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</tr>
<tr>
<td>Phase Accuracy</td>
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</table>
47

Chapter 4

Calibration and Testing

In this chapter, the calibration methodology is explained, including the systematic error model derived from the analyses and the different calibration loads available. The ASSP SIP boards have been recently manufactured and the initial testing showing some the main parameters are presented.

4.1 Calibration Methodology

4.1.1 Error Model

To perform the calibration the instrument error model must be defined, including all the sources of systematic errors present. The error model is developed from the admittance measurement expression derived in the previous chapter,

\[ Y_M = \frac{V_{I,\text{ref}} + jV_{Q,\text{ref}}}{V_{I,\text{ref}} + jV_{Q,\text{ref}} \left( R_{\text{ref}} + R_b/n^2 \right)^*}, \]  

where \( (R_{\text{ref}} + R_b/n^2)^* \) is the nominal resistance of the reference channel, including the impedance loading of the burden resistor. Using the complete expression along with the systematic errors of the raw values,

\[ V_{I,\text{ref}} = \frac{G_{\text{ref}}}{|1/Y_{\text{ref}} + Z_{s,\text{ref}}|} \cos(\arg(1/Y_{\text{ref}} + Z_{s,\text{ref}}) - \phi_{\text{ref}}) + V_{\text{off}}, \]  

\[ V_{Q,\text{ref}} = \frac{G_{\text{ref}}}{|1/Y_{\text{ref}} + Z_{s,\text{ref}}|} \sin(-\arg(1/Y_{\text{ref}} + Z_{s,\text{ref}}) - \phi_{\text{ref}}) + V_{\text{off}}, \]  

\[ V_{I,\text{ant}} = \frac{G_{\text{ant}}}{|1/Y_{\text{ant}} + Z_{s,\text{ant}}|} \cos(\arg(1/Y_{\text{ant}} + Z_{s,\text{ant}}) - \phi_{\text{ant}}) + V_{\text{off}}, \]  

\[ V_{Q,\text{ant}} = \frac{G_{\text{ant}}}{|1/Y_{\text{ant}} + Z_{s,\text{ant}}|} \sin(-\arg(1/Y_{\text{ant}} + Z_{s,\text{ant}}) - \phi_{\text{ant}}) + V_{\text{off}}, \]
where $G_{\text{ref}}$, $G_{\text{ant}}$ are the gains for the different channels, $\phi_{\text{ref}}$, $\phi_{\text{ant}}$ are the phase delays associated with each path, and $Z_{s,\text{ref}}$, $Z_{s,\text{ant}}$ are the impedance in series with the reference resistor and the antenna (including the burden resistor). The first approximations are then taken by assuming that the offset will be the same for all the measurements, and by having the same gain and phase delay for the in-phase and quadrature components. Introducing these expressions on the impedance measurement equation results in

$$Y_M = \frac{G_{\text{ant}} e^{-j\phi_{\text{ant}}}}{G_{\text{ref}} e^{-j\phi_{\text{ref}}}} \frac{1}{1/Y_{\text{ant}} + Z_{s,\text{ant}} + Y_{\text{off}}} \frac{1}{(R_{\text{ref}} + R_{b}/n^2)^2}.$$

(4.6)

where $Y_{\text{off}}$ represents the equivalent admittance of the offset voltage. For the reference channel the admittance selected is high enough so that the offset can be neglected. The connection to the reference resistor is also short so $Z_{s,\text{ref}} \approx R_{b}/n^2$. On the other hand, for the antenna channel the ground impedance could be a source of error to be determined during testing. Finally, the previous equation can be reduced to obtain the error model based on the minimum number of independent parameters,

$$Y_M = \epsilon_g e^{-j\phi_d} \frac{1}{1/Y_{\text{ant}} + Z_{s,\text{ant}} + Y_{\text{off}}},$$

(4.7)

where $\epsilon_g$ includes the mismatch between channel gains and the knowledge uncertainty of the reference resistor, and $\phi_d$ represents the phase mismatch between channels.

The calculated error model has three complex parameters requiring three independent measurements. A higher order model could have been obtained without making approximations. During the initial testing and characterization, it will be possible to observe if the assumptions can be validated. Once the error parameters are determined the inverse function can be used to correct the measured values,

$$Y_{\text{ant}} = \frac{1}{1/Y_M e^{-j\phi_d} - Y_{\text{off}} - Z_{s,\text{ant}}}.$$  

(4.8)
4.1.2 Calibration Loads

The mechanical configuration of the RF Test Head was designed in order to reuse the calibration loads built for the STORMS mission. These loads are canned shaped similar to the antenna allowing for easier connection to the instrument. They were also constructed with a great range of impedances comprised by:

- Open Load;
- Short Load;
- Resistors: 50 Ω, 100 Ω, 200 Ω, 500 Ω, 1 kΩ, 2 kΩ, 5 kΩ, 10 kΩ, 20 kΩ, 50 kΩ, 100 kΩ, 200 kΩ, 300 kΩ;
- Capacitors: 0.5 pF, 1 pF, 1.5 pF, 2 pF, 2.5 pF, 3 pF, 3.5 pF, 4 pF, 4.5 pF, 5 pF, 6 pF, 8 pF, 10 pF, 12 pF, 15 pF, 18 pF;
- Inductors: 1 µH, 2 µH, 5 µH, 10 µH, 20 µH, 50 µH, 100 µH, 200 µH, 500 µH, 1000 µH, 2000 µH;
- Resonant Circuits: 1.4 MHz, 2 MHz, 4 MHz, 6 MHz, 8 MHz, 10 MHz, 12 MHz.

The resonant circuits present a similar impedance to the plasma with different parallel resonant frequencies.

Ideally, the calibration loads must be characterized with higher accuracy than the instrument under calibration. Unfortunately, there is no equipment in the market that has the required accuracy over the impedance and frequency range. For example, commercial LCR meters have a great accuracy but only at a fixed frequency. For this project, a commercial VNA was used and the complete measurements can be found in Appendix D.

The problem of using a VNA is that they are only accurate for resistive impedances around the characteristic impedance $Z_0$ (Agilent recommends using them only for 3 Ω around $Z_0$ for better performances). In the results, the measurement uncertainty in magnitude and phase is shown along with the measured impedance. Because of this, only the 50 Ω load is known with enough accuracy, around ±1% in magnitude and ±0.7° in phase. Additionally,
it can be said that the OPEN load presents a high impedance (more than 3 kΩ) and the SHORT load presents a low impedance (lower than 0.6 Ω), but the exact value is uncertain as well as its phase. In Appendix D the problems associated with the OPEN load are explained. Furthermore, it is discussed the convenience of not connecting any load, instead of the OPEN load, for having a higher impedance.

4.1.3 Calibration Procedure

Given the unavailability of very well known calibration loads, the following approximations are made.

- First, assume $Z_{s,\text{ant}}$ is equal to $R_b/n^2$. In either case, the ground impedance cannot be calibrated with the calibration loads, but with the real antenna and rocket (or in more practical terms, with an analysis or simulation).

- It is foreseen that $Y_{\text{off}}$ will be very small because of a very low offset voltage after the CDS correction. In that case, it can be simply discarded. Because the OPEN load can not be properly characterized, the result will only be valid if the parallel is larger than 100 kΩ.

- Characterize the gain and phase mismatch with the 50 Ω load.

In principle, temperature drifts should be mostly corrected by the CDS and reference channel. However, to improve the final accuracy it is also recommended to do a calibration with a controlled temperature in 10°C steps.

4.2 Initial Tests

After solving some problems associated with layout errors, ADC common mode range, obsolete parts, etc., the ASSP SIP boards were ready to be tested functionally and obtain some of its performance parameters. The schematics of both boards are included in Appendix B and the GSE used during the testing is described in Appendix E. Some of the solutions adopted had a negative impact on the performances.
• Noise from the digital logic (specially the master clock oscillator) is coupled into the analog plane. This will increase the total amount of noise in all frequencies.

• The reconstruction filter cut-off frequency is lower than expected (around 18 MHz), due to the components tolerance and $Q$ factor of the inductances.

• To solve a problem with the ADC common mode range, the low pass filter after the mixer had to be simplified into a single pole. With this change, the image frequency when measuring the lower frequencies (below 6 MHz) cannot be rejected completely and this results in a higher amount of noise, and directly proportional to the signal being measured.

In Fig. 4.1, the standard deviation noise measurements has been represented. The CDS reduces the noise drastically because of the flicker noise filtering. On the other hand, the coadding reduces the high-frequency noise (including the image frequency output of the mixer). Gain at high frequencies has degraded and thus, the amount of noise increases. Combining the CDS and coadding the total amount of noise is very reduced but higher than expected for the reasons stated before.

In Fig. 4.2, the measurement offset has been represented. The CDS reduces in almost
an order of magnitude the total admittance offset, but it is still high enough to be considered
during the calibration. The offset in the lower frequencies is consistent with a shunt
capacitance of 0.6 pF which will need to be further investigated.

Finally to assess the capability of the instrument for measuring variations on the free
space capacitance, several capacitor loads were measured. In Fig. 4.3, three nominal loads of
1, 4, and 10 pF have been represented. As can be seen, the dynamic range of the instrument
is large enough to measure the 10 and 4 pF, but it is not enough for measuring the 1 pF
(only with averaging the capacitor curve can be distinguished).

Focusing on the 4 pF capacitor which is similar to the expected free space capacitance, a
curve fitting is performed to find the true capacitance being measured and express the error
as the residual with respect to this fitting. This has been represented in Fig. 4.4, dividing
the total error between the mean and the standard deviation. The mean deviation is due to
the lack of calibration and nonlinearity of the instrument, while the standard deviation is
purely due to the instrument noise. At low frequencies the error is primarily due to the noise
of the image frequency, and at high frequencies the loss of gain and dissimilarities between
the antenna and reference channel. It should be noted that the curves here represented
have not been calibrated in gain or phase, and therefore the mean error can potentially be

![Graph](image.png)

Fig. 4.2: Offset measurements with and without CDS.
Fig. 4.3: Dynamic range of the instrument measuring different capacitors.
reduced. Additionally, correcting the low-pass filter and reconstruction filter, the error curve should be approximately flat like it is in mid-frequencies with a total error of 3% and 4° in magnitude and phase.

Finally, a resonant circuit at 4 MHz has been measured and represented in Fig. 4.5 comparing the results with the curve obtained with a commercial VNA. Even with an error in general higher than expected, the resonant frequency can be easily determined.

Fig. 4.4: Residual error for a 4 pF capacitor measurement.
Fig. 4.5: Measurement comparison of a resonant frequency with a commercial VNA.
Chapter 5

Conclusions

In this thesis, a new SIP architecture has been outlined and analyzed. It is considered a mature design and it can achieve the required performance metrics. The main advantages of this architecture are the following.

- It has a high selectivity. The energy of the in-phase or quadrature component is integrated over many cycles providing good rejection of spurious signals and noise. The main limitation is usually found in the ADC dynamic range.

- The accuracy is enhanced by techniques like CDS or the reference channel. This provides an inherent accuracy that can be further improved by calibration.

- An error model has been developed and some guidelines are given for the calibration. In principle, many of the systematic errors will be corrected or limited.

- It is a compact design requiring very few components.

The design and manufacturing errors will degrade the accuracy, but they have been identified and will be corrected in future versions. Because enough margin was taken in the impedance range, it is anticipated that the actual antenna impedance will be accurately measured. The ASSP mission presented a wonderful opportunity to test this new architecture by providing very few restrictions in terms of size or power. This SIP architecture based on a VNA, along with techniques like CDS, is very compact and provides a very good accuracy. However, for future missions, the power consumption should be reduced, and the design optimized in general, by for example, reducing the amount of different voltages required. If the noise level is kept under appropriate levels, additional drivers and amplifiers are suggested to avoid loading effect and to better adjust the dynamic range to the ADC input.

The RF Head is another block that has undergone important improvements.
• The reduced frequency span with respect to previous instruments has made it possible to use a very small transformer, with an exceptional flat response and high repeatability.

• The symmetry between channels is increased, and the reference channel should be very accurate and stable.

• The guard design has a better electrical continuity, mechanical support, and ease of manufacturing.

Minor improvements are suggested for the next SIP. The burden resistor could be changed for an active load [14]. In this way, the impedance loading in the primary would be negligible and it would simplify the error model and calibration, while also improving the low frequency response. To improve the lower cut-off frequency even further, the primary cable should be changed to a material with lower resistivity. This will allow selecting a core magnetic material with lower permeability, which are more stable with temperature and frequency. Another idea to increase the symmetry between channels would be using a low gain amplifier for each one, and a larger gain after the multiplexer.

For the calibration it is strongly recommended to use an impedance analyzer with more calibration loads. Even if they are not capable of reaching the highest impedances, it would allow to characterize with 1%/1° accuracy impedances between 10 and 100 kΩ loads. Characterizing with precision capacitors and inductors would improve the knowledge of the instrument, by observing the possible differences between the in-phase and quadrature components. Moreover, having more calibration loads, even in the lower range, would enable the use of more terms in the error model. In particular, a SHORT load is easy to construct with high accuracy (ignoring the electrical length) and an effort should be made to make possible its use inside the VNA range.

More ambitious projects include the miniaturization of this instrument for its integration into small satellites like CubeSats. From an electrical perspective, this presents several challenges. Low-power components must be selected and the accuracy degradation must be assessed. However, even more challenging will be the mechanical deployment of the antenna while maintaining a good electrical guard.
Even if the SIP highest advantage is providing a very accurate impedance curve, it requires a high data rate that is not always available. There is work in progress for implementing a digital phase locked loop. Because the frequency update is limited by the serial line data rate, the tracking cannot be performed like that of an analog PLL. It is proposed to implement a small span sweep centered around the parallel frequency. The raw data must be converted to impedance values and the parallel resonance estimated in just a few milliseconds. Therefore, it seems reasonable to perform those operations with limited accuracy, even without calibration, while down-linking enough raw data to ground where the parallel frequency could be calculated with higher accuracy.

Going even further, it is the intention of SDL-USU to implement a CubeSat top-side sounder based on this same architecture. This will present many challenges and possible projects, including the RF design of a tunable impedance matching network, the signal post-processing for reducing the data rate or the on-ground post-processing. If successful, it will provide invaluable data being the first top-side sounder on-board a CubeSat.
References


Appendices
Appendix A

Current Transformer

The Current Transformer (CT) has a large influence on the overall instrument performances. It has to measure the current over a wide frequency range (1.3 decades), with very high sensitivity (currents in the order of microamperes in the upper-hybrid frequency) while introducing low noise. This appendix explains the different transformer cores, configurations and decisions that were made during the design and manufacturing process.

A.1 Theory

The CT is used in the typical configuration shown in Fig. A.1, with a small burden resistor that senses the current reflected from the primary, and with a Low Noise Amplifier (LNA) connected to it. This together with the CT ratio 1:n assures that the impedance insertion in the primary $R_b/n^2$ is minimum. The CT general model [15, 16] can be seen in Fig. A.2, where $R_1$ and $R_2$ are the resistive losses of the windings, $L_1$ and $L_2$ are the leakage inductance of the windings, $C_1$ and $C_2$ are the capacitance of the windings, $R_C$ represents the losses on the core, and $L_M$ is the magnetizing inductance of the core (where the hysteresis and saturation can be modeled). The capacitance between windings has been neglected because the braid of the primary wire grounded acts like a Faraday shield [17].

This model is approximated by the lumped elements model of Fig. A.3, where

\[ R_{1,2} = R_1 + R_2/n^2, \]  
\[ L_{1,2} = L_1 + L_2/n^2, \]  
\[ C_{1,2} = C_1 + n^2C_2. \]

For low frequencies, the model gets simplified as shown in Fig. A.4. In general, all inductances and capacitances present a low and high impedance, respectively, and can
Fig. A.1: Typical circuit using a current transformer.

Fig. A.2: General model of the current transformer.

Fig. A.3: Lumped elements model of the current transformer.
be ignored. However, the core inductance at low frequencies presents a low impedance comparable to the burden resistor load, thus decreasing the sensitivity. The transfer function is similar to a high-pass filter where the cut-off frequency is

\[ f_L = \frac{R_{1,2} + R_b/n^2}{2\pi L_C}. \] (A.4)

For increasing the bandwidth in the lower frequencies the core inductance must be increased. For example, toroidal cores have the following expression:

\[ L = \frac{\mu n^2 A}{2\pi r}, \] (A.5)

where \( A \) is the cross-sectional area and \( r \) is the radius. In general, manufacturers use the \( A_L \) parameter which indicate the inductance per turn. Taking this into account, it is usually selected a high permeability material, with big dimensions and a large number of turns. It is also convenient to reduce the burden resistor load in the primary, and using wires with low resistivity (mostly in the primary because the secondary impedance gets reduced by \( n^2 \)).

In high frequency the transformer model gets simplified as shown in Fig. A.5. The high cut-off frequency can be calculated as

\[ f_H = \frac{1}{2\pi \sqrt{(R_b/n^2C_{1,2})^2 + (L_{1,2}/R_C)^2}}. \] (A.6)

For increasing the bandwidth in the upper frequencies, it is recommended the least number of turns (reducing in this way all parasitic elements) and low permeability materials. These materials have a lower permeability at low frequencies, but they extend much further in
high frequencies without dropping. They also have less losses and are more stable with temperature.

Additional requirements are having enough sensitivity, which is given by

$$S_V = \frac{V_O}{I} = \frac{R_b}{n},$$

(A.7)

and low noise, which results in a low burden resistor. As can be seen, there exists a trade-off between the material, number of turns, and burden resistor.

A.2 Configurations Tested

Different configurations were tested varying the core geometry (toroidal, small binocular, or big binocular), the number of turns of the primary and secondary, the wire type, and the burden resistor. The configurations used are summarized in Table A.1. In order to compare the different transformers, the transfer functions have been normalized to the ideal gain (i.e. 0dB represents the gain of an ideal transformer with that configuration).

The transformers are characterized with the test set-up shown in Fig. A.6. The series resistance is used to excite the primary with a constant current independent of the impedance load. The input impedance of the network analyzer is used as the burden resistor.

The conclusions extracted from the tests are:

- Because the ASSP SIP transformer is situated very close to the antenna, the use of a big impedance controlled coaxial was considered not necessary. Instead of that, a thin shielded wire was used, allowing the employment of the binocular core. This core presents the advantage with respect toroidal cores of very reduced leakage inductance.

![Fig. A.5: High frequency model of the current transformer.](image)

$$L_{1,2}$$

$$R_C$$

$$C_{1,2}$$

$$R_b/n^2$$

$$1: n$$
<table>
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<th>$n_2$</th>
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<th>$S_V$</th>
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<td>61</td>
<td>1</td>
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Fig. A.6: Current transformer test set-up.

(in mundane words, very little wire is not contributing to the magnetic field), with higher $A_L$, and it is very easy to wire allowing a high repeatability. A comparative of toroidal and binocular cores of similar dimensions can be seen in Fig. A.7.

- The highest permeability core is preferred. Even if the permeability is reduced at high frequencies or has more temperature degradation, to achieve a good low frequency response is the best option. In Fig. A.8, three different materials were tested in the same configuration, 73 with $\mu_r = 2500$, 61 with $\mu_r = 125$, and 43 with $\mu_r = 800$.

- The burden resistor was selected to be 50Ω. It is an often convenient load to use [17], because it is the load used by the manufacturer to provide its data, an arbitrary
length of coaxial wire and common instrumentation can be used without changing its properties, it requires a low number of turns to have a low impedance insertion, and it imposes a low requirement on the amplifier input impedance. In this particular application, it also presents an adequate low thermal noise.

• The number of turns was limited to five turns in the secondary. By having this low number of turns, the high frequency response and the sensitivity are improved. On the other hand, the impedance insertion is high, but limited by the low burden resistor. In Fig. A.9, it can be seen how a big turns ratio (1:6) improves the low frequency response, but with 1:5 is enough. It was also tested with half turn in the primary (making it similar to a toroidal), but the response was not acceptable.

• The wire selected for the primary is a custom manufactured Stainless Steel (SS) shielded wire. Unifilar shielded wires are not very common and this wire was already in use in other ASSP instruments. As indicated some tests were performed with enamel coated wire with a small braid (configurations 15 to 17), but even if the electrical properties were good, the mechanical result was much better with the shielded wire.

• For the secondary, an enamel coated wire was used. This type of wire is often referred as magnetic wire, because it is used in transformers and motors. It presents the
advantages of having a copper core and a very thin insulation. This results in a high conductivity and density, needing less turns than other wires. In the tests, this advantage can be observed with respect to regular Teflon coated wire, or the custom SS wire used in the primary. The final gauge selected, 28 AWG, was chosen from the ones that were performing well electrically (24 and 28), and made easier the winding. For example, a 36 AWG gauge was available, but at these frequencies the performance was poor.

Configuration 19 is the final one. It consists of a big binocular core (2873000302), with the custom shielded SS wire in the primary (N12-50F+00007-5), and the 28 AWG enamel coated wire in the secondary (MW-MC5516-038). It presents a very flat frequency response and a fair sensitivity. The transformer selected has enough space in the holes in case more turns are considered necessary in the future. To fix the cables into a right position, the holes are potted without affecting the electrical performances.

A.3 Final Transformer Tests

In order to improve the performances of the instrument, it is interesting to have two similar transformers for the antenna and reference channels. Several transformers were wired, potted, and tested with the configuration selected. The results presented in Fig. A.10,
and zoomed in the bandwidth of interest in Fig. A.11, show a good repeatability, and they were used to select the more similar couple in terms magnitude and phase. In this case, transformers 8 and 11 are quite similar and have some of the higher gain.

The primary wire resistance was measured to be about $5.7\,\Omega$ and the secondary resistance

![Diagram showing frequency response with magnitude and phase plots.](image)

Fig. A.10: Final tests for transformer selection.
Fig. A.11: Final tests for transformers selection with detail in the operational frequency bandwidth.
of $0.1\Omega$. It would have been interesting to measure the capacitance and inductance (which are normally not provided in the wire datasheet), to check the the expected cut-off and resonant frequencies.

### A.4 Hysteresis and Saturation

Saturation and hysteresis affects the CT by producing harmonics on the secondary output. A specific test was performed to assure that the level was acceptable. The CT was driven with the maximum current at the minimum frequency and the secondary was measured with a spectrum analyzer. The result can be seen in Fig. A.12, where the third harmonic is $62\,\text{dB}$ below the fundamental. This harmonic is in fact produced by the signal generator (two different signal generator were used with similar results). Even if it were a harmonic produced by the CT, it is low enough to discard any saturation degradation effects, which was foreseen due to the low current levels, but it also discards any hysteresis problem.

### A.5 Shield Rejection

The guard is only effective if the coupling between the primary core and secondary, and the primary shield and secondary are very similar. To test this, the set-up of Fig. A.13 was used, where the same current is flown through primary and shield. In ideal conditions

![Fig. A.12: Harmonics testing of the current transformer.](image)
the secondary output would be zero. The result have been normalized to the output of the secondary when only current is flowing through the primary shown in Fig. A.14. As frequency increases, the different geometry of the core and shield are more pronounced. However, there is still a good rejection of at least 20 dB. This means that for having a similar output to the actual signal, the stray capacitances should be 20 dB higher than the antenna capacitance (10 times in linear units), when in fact they are much less.

Fig. A.13: Current transformer shield rejection test set-up.

Fig. A.14: Rejection ratio of currents through the shield of the current transformer.
Appendix B

PCB Schematics

(7 pages)
Solder wires for these connectors directly to board.

Manually surface wire the VDD and GND lines to these modules. Fill all available wiring side PCB space with ground planes and lock a wire in the ground plane specially adjacent to the switch modules to allow the addition of a metal shield over U1-U5 to further improve isolation.

Solder wires for these connectors directly to board.

Space Dynamics Lab - ASSP
186-0190 SIP RF Board
Made Cox
7/18/2013
Appendix C

FPGA RTL Diagrams

(14 pages)
All parameters are big-endian.
ASSP Digital Design
J. Martin-Hidalgo
Data Acquisition Functional Partition
3
ASSP
DESCRIPTION
FILE NAME
ITEM
SHEET
SHT REV
TITLE
ENGINEER
LAST MODIFIED
SPACE DYNAMICS LABORATORY
UTAH STATE UNIVERSITY RESEARCH FOUNDATION
North Logan, Utah 84341

PROGRAM
SPDASCTRL
VSDX

NEXT ASSY
ASSP Science Firmware

This drawing contains information that is proprietary to SPACE DYNAMICS LABORATORY (SDL).

REFERENCE USURF BP 409.1

Form No. QF 0423 Rev A

ADS 8343
HK
_HK
_DClk
_HK
_Shdn
_HK
_Busy
_HK
_DIn
_HK
_DOut

DAS Control Blocks

Sweeping Impedance Probe

Combinatorial Logic Decoder

Combinatorial Logic Decoder

SFID [4:0] HkAddr [3:0]
0x7 0x8 0x9 0xA 0xB 0xC 0xD 0xE 0xF 0x10 0x11 0x12 0x13
Note: * during sweep, c-add is disabled on DCUP_H, DCUP_L and the FPP. The sampled values are piped into a FIFO before being sent out in the Telemetry matrix. The sweeps occur every 11.8 major frames.
Data Acquisition Detailed Timing Diagram (A/D)

- DataValid_H
- LastMsrmnt_H
- InitCnv_sync
- tconv = 1.6 μs
- tPULSE = 300 ns
- tACQ = 0.66 μs
- tSMPL = 9.12 μs
Appendix D

Calibration Loads Characterization

To characterize the calibration loads a commercial VNA (FieldFox 44914A) was used. The calibrations loads were connected to the VNA by using a base adapter (see Fig. D.1), which provides an standard SMA connector on the VNA side, and an SMB and ground connection for the can ground on the load side.

Before measuring the calibration load cans, the VNA is calibrated by using an standard N-type connectors calibration kit with an SMB to type-N adapter (Pasternak PE9313) as shown in Fig. D.2. Ideally, the calibration kit should have SMB connectors but these are very rare. The introduction of the SMB to N-type adapter introduces a phase delay (the reference plane is located at the adapter output) and could limit the effective directivity measured [18]. Although the adapter datasheet does not provide electrical information, based on the results of the calibration kit LOAD, with a return loss of about 70 dB, the directivity is not limited by it. The electrical length introduced by the adapter was characterized to be 66.8 mm and it needs to be subtracted from all the calibration loads measurements. With this correction, the measurement at the SMB connector without any load is very similar to an OPEN.

With this set-up, all the calibration loads were measured. Apart from the instrument

Fig. D.1: Test set-up for the calibration loads characterization.
uncertainty, additional disturbances can be observed in the graphs. At low frequencies, the 
VNA shows an abnormal response and what it seems a scale change. There is also observed 
a shunt capacitance of about 3 pF that affects the measurement of high impedances, such as 
resistors, low value capacitors, and inductors. The shunt capacitance seems to be associated 
with the mechanical configuration. Between the measurements of the SMB without any load 
and the OPEN load there is about 0.7 pF. The OPEN load has the can configuration and a 
PCB without any trace. Even more, with a high resistor value 300 kΩ at high frequencies 
it can be observed about 0.7 pF. It can be concluded that the parasitic capacitance of the 
mechanical can and the PCB traces affect the measurements, especially for high impedances.

In the next set of figures, the measurement of the pertinent loads and the unloaded 
SMB connector are presented. It can be seen that due to the nonlinear relation between the 
reflection coefficient and the impedance, when the former is acquired with low uncertainty 
the latter has very large uncertainty and vice versa. In the case of the 50 Ω load, shown in 
Fig. D.3 and D.4, the impedance uncertainty is minimum. The SHORT load, shown in Fig. 
D.5 and D.6, is a very good reflector but the impedance uncertainty is large, especially its 
phase. The same happens with high impedances. Additionally, it can be observed that the 
OPEN load, shown in Fig. D.7 and D.8, presents a smaller impedance than the unloaded 
SMB connector, shown in Fig. D.9 and D.10.
Fig. D.3: $s_{1,1}$ magnitude and phase of the 50Ω load.

Fig. D.4: Impedance magnitude and phase of the 50Ω load.
Fig. D.5: $s_{1,1}$ magnitude and phase of the SHORT load.

Fig. D.6: Impedance magnitude and phase of the SHORT load.
Fig. D.7: $s_{1,1}$ magnitude and phase of the OPEN load.

Fig. D.8: Impedance magnitude and phase of the OPEN load.
Fig. D.9: $s_{1,1}$ magnitude and phase of the unloaded SMB connector.

Fig. D.10: Impedance magnitude and phase of the unloaded SMB connector.
Appendix E

Ground Support Equipment

Two GSE specific software were developed during the development of the project. The framework chosen was Labview due to its ease of programming and fast integration with instruments controlled by GPIB.

E.1 Scalar Network Analyzer

Prior to the acquisition of the commercial VNA, a Scalar Network Analyzer (SNA) was constructed with a signal generator and a spectrum analyzer controlled from a PC as can be seen in Fig. E.1. The process followed is the same as with any SNA: both instruments are initialized with the required parameters (power, IF bandwidth, span, etc.), the signal generator is set to the minimum frequency, an appropriate settling time is waited and the measured value by the SA is obtained. Then, the frequency is changed to the next value and the process starts again. Because of the multiple interaction with the equipments, and the settling time of the SA, the process requires some minutes to complete for a regular frequency span and resolution. As indicated in the figure, the signal generators used had a limited frequency range of operation, and for covering the complete range both of them were used requiring a manual disconnection and connection for each measurement.

The software interface, shown in Fig. E.2, allows the configuration of the frequency span (with linear or logarithmic sweep), averaging and exporting the results in Comma-Separated Value (CSV) format. It is intended to emulate a real SNA interface with the most common options.

E.2 SIP Ground Support Equipment

When using the SIP board the telemetry must be acquired by a specific GSE with
Fig. E.1: SNA GSE block diagram.

Fig. E.2: SNA GSE software interface.
a PCM telemetry module. Because at the time neither the GSE nor the FPGA module were finished, a specific GSE was developed through the SIP UART. It allows configuring the FPGA in all the test modes (fixed frequency, CDS disabled, and variable number of coadded samples) and acquiring the measured values in small chunks. Because of the limited buffering in the FPGA and UART data rate (115.2 kbps), the complete frequency range has to be acquired over multiple sweeps. This does not introduce any error as long as the load is static like in the case of the SIP. In the same manner than in the SNA, the interface emulates a real VNA, providing real-time impedance values (without calibration) and a table with raw values for post-processing in more specialized software.

Fig. E.3: SIP GSE software interface.
Appendix F

DVD Contents

The DVD attached to this thesis contains the following materials:

/  
  thesis .................................................. LaTeX files of this thesis  
  figures ........................................... Figures used in the thesis  
  tables .................................. Tables used for generating plots with PGFPPlots  
  matlab  
    analyses  
      noise ...................................... Noise analysis and figures  
      harmonics .................................. Harmonic analysis and figures  
      lpffilter .................................. Low pass filter analysis  
      aafilter ................................... Reconstruction filter analysis  
      loading .................................. Loading effect analysis  
    calloads .................................. Calibration Loads tests and analyses  
    fpga ........................................ Support files for the FPGA design  
    transformers ................................ Transformers tests and analyses  
    tests ........................................ SIP tests and analyses  
  labview  
    sna ........................................ SNA Labview software  
    gse ........................................ SIP GSE Labview software  
  fpga  
    Common ..................................... Common files of the ASSP FPGAs  
    SIP  
      ActelDesigner ................ Place-and-Route project and physical contraints  
      Documentation ............................ RTL diagrams  
      Simulation ................................ Simulation models and testbench  
      Synplify .................................. Synthesis project and constraints  
      VHDL ....................................... VHDL source code and cores  
  pcb ........................................ PCB schematics, layout and part list