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Design and Development of Timmi - An Interferometric Radar

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DESIGN AND DEVELOPMENT OF TIMMI - AN INTERFEROMETRIC RADAR

A Thesis Presented

by

KARTHIK SRINIVASAN VENKATASUBRAMANIAN

Submitted to the Graduate School of the University of Massachusetts Amherst in partial fulfillment of the requirements for the degree of

MASTER OF SCIENCE IN ELECTRICAL AND COMPUTER ENGINEERING

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Electrical and Computer Engineering
To my Father, Mother and Sisters
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ABSTRACT

Interferometry has gained importance as a remote sensing technique to study topography, topographic change and volume and surface scattering properties of various natural targets. Interferometric radars rely on the ability to accurately measure amplitude and phase between signals received on two spatially separated antennas. The accuracy required for interferometric measurements place tight constraints on the performance of the radar hardware. This thesis details the development, construction and testing of a two-stage, two-channel Ku band downconverter (also referred to as Dual Channel Downconverter or DDC)- which forms the core of the interferometer - to meet the requirements to make highly accurate interferometric measurements.
CHAPTER 1
INTRODUCTION

Microwave Radar Interferometry is a technique that has been successfully used to produce topographic maps of the Earth’s surface at frequencies ranging from P-band to X-band, with the choice of frequency being application specific [8]. A conventional Synthetic Aperture Radar (SAR) can only measure the location of a target in two dimensions. An Interferometric SAR, on the other hand, adds a third dimension - altitude - providing a full three dimensional map of the target. In addition to topographic features, interferometry has been used to study changes in land and ocean surface over time [6].

Jet Propulsion Laboratory’s Ocean Surface Topography Mission (OSTM) is a joint NASA, the French Space Agency (CNES), NOAA and European Organization for the Exploitation of Meteorological Satellites (EUMETSAT) project to study sea surface topography. This effort has numerous applications, such as, global environment monitoring, including predicting hurricane intensification, tide models, measuring rate of global sea level rise and mapping ocean currents. As part of the OSTM project, the Wide Swath Ocean Altimeter(WSOA), a Ku-band Interferometer was proposed as an optional instrument for measuring mesoscale oceanographic topography from space. A complete link budget and preliminary system configuration was made out for the WSOA project. However, the WSOA did not form part of the final OSTM mission, and the link budget and proposed hardware were not constructed or tested for accuracy. Nevertheless, the need for precise interferometric measurements remains and hence the need to characterize individual components of the interferometer.
Interferometric accuracies required for applications, such as those proposed as part of WSOA, typically are in the centimeter scale. Achieving such high-precision interferometric measurements places significant and challenging restrictions on the hardware design. Interferometry relies on the highly accurate measurement of relative phase difference between antennas separated by a baseline. It is the ability to accurately measure this phase difference that drives the overall system requirements of the interferometer.

The RF receiver chain is one of the critical components of the interferometer and is highly sensitive to factors such as operating temperature which can affect the phase performance, hence, the accuracy of the interferometer. The push to move to higher frequency bands, in an attempt to achieve a more compact system with a much shorter baseline, increases the importance of a highly stable, accurate RF receiver. At high frequencies, achieving RF receiver relative phase stability of the order of magnitude required to make highly accurate interferometric measurements is very challenging. A literature search for previous work in characterizing the phase of a two channel Ku-band downconverter, however, did not yield any results.

1.1 Summary of Chapters

The work carried out as part of this thesis is the design of a High Performance Dual Channel Downconverter at Ku-band (13.28 GHz). It is envisioned that this downconverter will be used in TIMMi - Topographic Ice Mapping Mission - a proposed High Performance Interferometer which will be used to study the properties of snow, sea ice and also to make topographic measurements. The following chapters detail the work carried out to develop, test and characterize the downconverter and discuss some results from the initial deployment of the interferometer.

Chapter 2 describes the interferometric principle, outlining a mathematical analysis of the interferometric setup, leading to an error budget analysis for different deployment sce-
narios. The results of the error budget analysis are used to derive system parameters for an interferometric deployment.

Chapter 3 describes the design process followed in the development of the high performance Dual Downconverter. Two versions of the downconverter were built and tested for amplitude and phase performance. The first version was a breadboard version built from connectorized components. This section details the choice of components used for the breadboard version. The second version of the downconverter built was fabricated on a Printed Circuit Board (PCB). A detailed description of the design of various stages of the PCB version is laid out in the section, along with simulation results of the individual components which form part of the downconverter.

Chapter 4 of the thesis presents the results of the amplitude and phase measurements of the two versions of the downconverter. This is followed by a discussion on the measurements.

The Dual Downconverter was designed and built with the intention that it would eventually form the core of a airborne or space-borne interferometer. As a first step in the progress towards implementing an airborne system, a ground-based interferometer was built using the downconverter. An antenna was required to launch and receive signals from the interferometer. Hence, an Aperture Coupled Patch Antenna array was built and tested. Results from the antenna development process and initial field deployments are presented in Chapter 5.
2.1 Description of Interferometric Radar

An interferometer is a phase-interference device consisting of two antennas separated by a baseline. The signals from the two antennas are down-converted, and for a cross-track interferometric configuration, the phase difference between the two signals can be converted to the corresponding height of the target being illuminated. Figure 2.1 illustrates the setup using a simple geometrical model.

The two antennas, labeled A1 and A2, are separated by a distance B, referred to as the baseline. The angle of the target direction with respect to the nadir is termed the look-angle, $\theta$, and the orientation of the interferometric baseline relative to the horizontal is the baseline tilt angle, $\tau$. The antennas are mounted on a platform at a height $H$ from a reference point. The slant range to the illuminated target is $r$ and $r + \delta r$ for the two antennas respectively. The height of the target, $h$, can be determined from the differential path length $\delta r$, which is determined from the phase difference $(\delta \phi)$ of the electric fields observed by the two antennas.

The interferometer being proposed, operating in a cross-track configuration, will be used to study properties of snow, such as volume scattering, and snow depth.

2.1.1 Mathematical Background

Interferometric techniques are considered prime candidates to produce systematic, high quality, accurate digital elevation models (DEM) for mapping topography within an accuracy of meters and determining topographic change to an accuracy of centimeters. This is
achieved by utilizing the phase difference between the two antennas, A1 and A2, which can be translated to a corresponding height with respect to a reference height $H$. From the geometry of Figure 2.1,

$$\cos(\theta) = \frac{H - h}{r}$$

(2.1)

Figure 2.1. Geometric Model of an Interferometric Setup
Figure 2.2 magnifies the antenna positions. This geometry is used to determine $\theta$ used in (2.1). From Figure 2.2,

$$
\theta = \arcsin \left( \frac{\lambda \cdot \delta \phi}{2 \cdot \pi \cdot B} \right) + \tau \quad (2.2)
$$

where $\delta \phi$ is,

$$
\delta \phi = \frac{2\pi \cdot \delta r}{\lambda} \quad (2.3)
$$

The height of the target can be written as,

$$
h = H - r \cdot \cos(\theta) \quad (2.4)
$$
It can thus be seen that an accurate estimate of topographic height depends on the
accurate measurement of $\theta$, which in turn, depends on the relative phase difference $\delta \phi$.

A parameter often used in the characterization of interferometric radars is the ambiguity
height ($h_a$). This is defined as the amount of height change that leads to a $2\pi$ change in
interferometric phase. This parameter is given by,

$$h_a = 2\pi \frac{\partial h}{\partial \phi} = \frac{\lambda \cdot r \sin(\theta)}{B \cdot \cos(\theta - \tau)}.$$  \hspace{1cm} (2.5)

The ambiguity height is related to the baseline length. Small ambiguity heights result
from long baselines. Hence, the baseline should be as small as possible. At the same time,
small baselines require very precise phase measurements, as small phase changes would
map to significant changes in height estimate.

The measured topographic height is therefore a function of the measured phase difference
between the antennas, baseline length and the look angle. In order to make accurate
topographic height measurements, appropriate choice of baseline length and look angle are
required. The following section presents an error analysis for the interferometric system,
which is used to derive optimal system parameters such that errors in height measurement
are minimized.

2.2 Error Budget Analysis

In order to define system parameters, an error budget analysis was carried out on a
hypothetical interferometric radar system. A propagation of errors method analysis was
carried out as outlined in [3] and [5], and the results are presented below.

From the geometry of Figure 2.1,

$$v = \arccos \left( \frac{B^2 + r^2 - (r + \delta r)^2}{2 \cdot B \cdot r} \right)$$ \hspace{1cm} (2.6)

and

$$h = H - r \cos(\gamma + v)$$ \hspace{1cm} (2.7)
where,
\[ \delta r = \frac{\lambda \cdot \delta \phi}{2\pi} \]  
(2.8)

The absolute height error \( (\sigma_h) \) is given by the sum of the error variances, such as,

\[ \sigma_h^2 = A_r \sigma_r^2 + A_B \sigma_B^2 + A_H \sigma_H^2 + A_\gamma \sigma_\gamma^2 + A_\phi \sigma_\phi^2 + A_\lambda \sigma_\lambda^2 \]  
(2.9)

where individual terms are broken out into contributions to height error of the range error \( (\sigma_r^2) \), baseline length error \( (\sigma_B^2) \), platform height error \( (\sigma_H^2) \), baseline tilt angle error \( (\sigma_\gamma^2) \), phase error \( (\sigma_\phi^2) \) and wavelength error \( (\sigma_\lambda^2) \). The coefficients which convert these errors to a topographic height error are given through the sensivity of (2.4) to the various error sources, as in,

\[
A_r = \left( \frac{\partial h}{\partial r} + \frac{\partial h}{\partial \theta} \cdot \frac{\partial \theta}{\partial r} \right)^2
\]
\[
= \left( -\sin(\gamma + v) \cdot \left( \frac{r^2}{2B^2} - \frac{B}{2r} \right) \right)^2
\]
\[ \cdot K \]
(2.10)

\[
A_B = \left( \frac{\partial h}{\partial v} \cdot \frac{\partial v}{\partial B} \right)^2
\]
\[
= \left( \sin(\gamma + v) \cdot \left( \frac{1}{2} + \frac{(r + \delta r)^2 - r^2}{2B^2} \right) \right)^2
\]
\[ \cdot K \]
(2.11)

\[
A_H = \left( \frac{\partial h}{\partial H} \right)^2 = 1
\]
(2.12)

\[
A_\gamma = \left( \frac{\partial h}{\partial \gamma} \right)^2 = (-r \cdot \sin(\gamma + v))^2
\]
(2.13)
\[ A_{\phi} = \left( \frac{\partial h}{\partial v} \cdot \frac{\partial}{\partial \delta r} \cdot \frac{\partial }{\partial \phi} \right)^2 \]  
(2.14)

\[ A_{\lambda} = \left( \frac{\partial h}{\partial v} \cdot \frac{\partial}{\partial \delta r} \cdot \frac{\partial }{\partial \lambda} \right)^2 \]  
(2.15)

\[ K = \sqrt{1 - \frac{B^2 + r^2 - (r + \delta r)^2}{2 \cdot B \cdot r}}. \]  
(2.16)

with \( K \) is given by

System parameters for an interferometer used to study the properties of snow are derived based on the above theoretical formulation. The frequency of operation was chosen to be compatible with the frequency allocation for NASA's Wide Swath Ocean Altimeter (WSOA) experiment - which operates at Ku-band (13.285 GHz) with a system bandwidth of 20 MHz and a baseline of 6.4 m. For the work of this thesis, the error analysis was carried out for a similar system which would be deployed on a rooftop. Parameters such as required baseline length for this new system will be derived from the error budget analysis. The standard deviations assumed for the error calculations are listed in Table 2.1.

Some initial parameters, such as transmit power (+20dBm), antenna gain (10 dBi) and pattern were assumed. The antenna considered was a horn antenna with a rectangular aperture. Based on this assumption, the beamwidth was chosen as 20° x 162° (azimuth x
elevation). The antenna size was 0.05 x 0.0068 m. Figure 2.3 shows the total error as a function of Baseline Length for a look angle of 50°. The standard deviation of the phase error was computed using the SNR. The relationship between SNR and phase error is shown in Table 2.1. For this case, the standard deviation was 0.07°. The second trace in Figure 2.3 shows the total error as a function of baseline for a standard deviation of phase of 0.05° (which is assumed to be the phase error introduced by the downconverter hardware). If the SNR is increased more than 30 dB, the error in estimating height due to SNR becomes negligible compared to the error introduced due to phase errors of the downconverter. A high phase accuracy of the interferometer, therefore, depends on a highly phase accurate downconverter.

Based on these initial assumptions, a baseline of 0.2 m was chosen. A longer baseline could also lead to problems with height ambiguity due to phase wrapping. Figure 2.4 shows the different error components plotted as a function of look angle for a phase standard deviation of 0.07°. The height error at look angles less than 60° is less than 10 cm. The major source of error for small look angles is due to baseline length. This can be reduced by maintaining a very stable baseline over the short period of observation. Long term changes in baseline length and orientation can be removed through analysis of the trend in interferometric phase as a function of time. These results are for single look cases, and the phase error can be reduced further by utilizing multiple looks, as presented later.
Figure 2.3. Total Height Error as a Function of Baseline Length

Figure 2.4. Contribution of Error Components to Total Height Error-SNR dependent
Figure 2.5. Contribution of Error Components to Total Height Error-Hardware Dependent

In order to ensure that the phase errors induced due to the downconverter hardware of the interferometer is the limiting case, a high SNR is required. Figure 2.5 shows the results when the antenna gain is increased from 10 dBi to 20 dBi. The antenna dimensions for this case is 0.55 x 0.06 m. The new antenna has a beamwidth of 1.5° x 38°. This ensures that the phase standard deviation due to SNR is negligible for the bore-sight case. A higher gain, and hence narrower antenna beamwidth also increases the azimuth resolution, which is good in the sense that the topographic variation within a resolution cell will be less with the increase of resolution. As before, the error due to uncertainty due to baseline length is the dominant source of error.

The contribution to the total height error due to phase uncertainty can be minimized by increasing the number of looks. In Figure 2.6, the error due to phase uncertainty has been set equal to the error due to uncertainty in baseline tilt angle. This requires at least 21 looks. If the baseline is doubled, then the number of looks to achieve this condition becomes 6. For an interferometer mounted on a rooftop, the error due to the baseline tilt angle should,
ideally, be non-existent since the antennas are static and do not move. However, there could be movement of the baseline due to wind and hence, a potential error source.

![Figure 2.6](image)

**Figure 2.6.** Contribution of Error Components to Total Height Error - Multiple Looks

During the actual deployment of the interferometer (presented in Chapter 5), some changes had to be made in terms of the transmit power. The transmit power had to be calculated in terms of maximum input power to the downconverter that was developed as part of this thesis. This constraint, for a rooftop deployment at a height of 16 m forced the transmit power to be reduced to $-5$ dBm and the antenna that was built had a gain of 19.8 dB. The following plots show the error budget with these changes. In this case, the baseline was maintained at 0.2 m. The results in Figure 2.7 are for the single-look case only. These results show that the error due to the baseline length is dominant in the far range and the error due to baseline orientation dominates in the near range.

A second rooftop deployment scenario which was considered was a platform height of 50 m height. The error analysis, similar to the 15 m case, is presented in Figure 2.8.
Figure 2.7. Contribution of Error Components to Total Height Error with New System Parameters - $H = 16$ m $P_t = -5$ dBm

Figure 2.8. Contribution of Error Components to Total Height Error with New System Parameters - $H = 50$ m $P_t = -5$ dBm
Based on this analysis, system parameters to achieve a height accuracy of better than 10 cm have been summarized in Table 2.2.

**Table 2.2. System Parameters for the Interferometer**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Center Frequency</td>
<td>13.285 GHz</td>
</tr>
<tr>
<td>Bandwidth</td>
<td>20 MHz</td>
</tr>
<tr>
<td>Transmitter Power</td>
<td>-5 dBm</td>
</tr>
<tr>
<td>Antenna Gain</td>
<td>20 dBi</td>
</tr>
<tr>
<td>Beamwidth</td>
<td>1.5° x 38°</td>
</tr>
<tr>
<td>Baseline Length</td>
<td>0.2 m</td>
</tr>
<tr>
<td>Number of Looks</td>
<td>21</td>
</tr>
<tr>
<td>Range Resolution</td>
<td>7.5 m</td>
</tr>
<tr>
<td>Azimuth Resolution</td>
<td>1-3 m</td>
</tr>
</tbody>
</table>
CHAPTER 3
DUAL DOWNCONVERTER DESIGN

3.1 Introduction

The error budget analysis presented in Chapter 2 shows that the error due to phase uncertainty has a significant bearing on the performance of the interferometer. Previously published literature in this area of research has concentrated on system design aspects of the interferometer, along with significant work on processing and analysis of interferometric data. However, there is very little published about the challenges involved in the design of highly accurate receivers for interferometric measurements. This is, in part, due to the fact that, based on the application, a lot of interferometric hardware development work has been carried out at lower frequency bands. At higher frequencies, the wavelength of operation becomes very sensitive to very small changes and makes the hardware design more challenging. As an example, the wavelength at P-band is 1 meter. Changes of the order of 1 millimeter constitute a 0.1% change with respect to the wavelength of operation. However, at Ku-band (wavelength 2.25cm), a 1mm change translates to a 4.5% change. Such changes, which could possibly be ignored at P-band, will become significant error sources at Ku-band.

This simple example shows that increasing the frequency of operation presents a lot of restrictions on the design and fabrication of circuits. The advantages of moving to higher frequencies - such as smaller and more compact systems, along with opening up a whole range of science data that can be obtained - are significant enough to justify the effort required to design a high performance system. For the interferometric deployment scenarios
outlined in Chapter 2, a relative phase accuracy of close to 0.05° is required to make topographic measurements with an error of less than 10 cm over a wide range of look angles.

The first step towards making such interferometric measurements is to design a down-converter which is capable of ensuring a high degree of phase accuracy. The down-converter required is a common heterodyne receiver with two independent channels, which down-converts the signal from Ku-band to a frequency which is low enough to be digitized and analyzed. Although such downconverters are commonly used in altimeters and polarimeters, an interferometer requires a higher degree of phase and amplitude stability. A polarimeter, for example, typically has phase stability of the order of 1°. This is insufficient for an interferometer, which needs phase stability that is at least an order magnitude better, to make highly accurate interferometric measurements. There is, however, very little (if any) previously published work which looks at the phase stability of such downconverters. Conventional methods employed to measure phase of a downconverter have residual errors which can be greater than those being measured with the downconverter. Hence, a novel method of phase measurement was developed in [1]. This test bed, which had a residual phase error of close to 5 millidegrees, was be used to test the performance of this down-converter.

The following sections present the design of two versions of the downconverter that were developed.

3.2 System Design

The Interferometer and hence the dual channel downconverter (also known as DDC) was chosen to operate at Ku-band (13.285 GHz), with a bandwidth of 20 MHz. The operating frequency was chosen to be compatible with NASA’s Wide Swath Ocean Altimeter (WSOA) mission. The important parts of a DDC are the RF section, IF section and the
control and telemetry section. The RF signals enter the DDC at Ku-band (13.275 - 13.295 GHz). This is downconverted to an intermediate frequency at L-band (1.275 - 1.295 GHz). Noise bandwidth filtering is performed at this frequency. This frequency has been chosen to reduced the constraints on the performance of the noise bandwidth filter. This stage also includes most of the gain of the downconverter. This signal is again downconverted to baseband (5 - 25MHz) in the IF section. The baseband frequency is chosen so that the Analog to Digital conversion stage which follows the downconverter can sample the signals at 60 Msamples/sec. A higher baseband frequency would require a higher sampling rate, making the digital circuitry more expensive and difficult to build. The control and telemetry section provides test and monitoring ports at various points on the downconverter. The downconverter was designed to be highly symmetric to minimize imbalances between the channels in order to meet the stringent phase and amplitude requirements of the downconverter.

Due to the lack of adequate knowledge of the factors that influence phase performance, two prototypes of the dual downconverter were constructed. The first prototype being a fully functional unit, built using commercial, off-the-shelf connectorized components, with minimal overhead circuitry. This version (referred to as the breadboard version) served as a testbed for the second version, which was fully integrated onto a printed circuit board. The objective of this two-step development process was to use the breadboard version to identify potential problem areas, enabling a better and more robust design for the final version to meet the tight constraints on amplitude and phase stability for the downconverter. Table 3.1 lists the proposed system specifications for the UMass DDC. These parameters are for a critically sampled system, with no averaging.
Table 3.1. Dual Downconverter Design Specifications

<table>
<thead>
<tr>
<th>Design Constraint</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency Range</td>
<td>13.275 – 13.295 GHz</td>
</tr>
<tr>
<td>Output Frequency Range</td>
<td>5 – 25 MHz</td>
</tr>
<tr>
<td>Effective Noise Bandwidth</td>
<td>&lt; 30 MHz</td>
</tr>
<tr>
<td>Operating Temperature</td>
<td>−10 to 50°C</td>
</tr>
<tr>
<td>Noise Figure</td>
<td>&lt; 4.5 dB</td>
</tr>
<tr>
<td>Channel to Channel Isolation</td>
<td>&gt; 80 dB</td>
</tr>
<tr>
<td>Input/Output VSWR</td>
<td>&lt; 1.5 : 1</td>
</tr>
<tr>
<td>Relative Channel to Channel Phase Stability</td>
<td>0.050° RMS over BW</td>
</tr>
<tr>
<td>Receiver Phase Variation over Best Quadratic Fit</td>
<td>3° RMS over BW</td>
</tr>
<tr>
<td>Receiver Amplitude Variation</td>
<td>2 dB over BW</td>
</tr>
<tr>
<td>Receiver Amplitude Variation over Best Linear Fit</td>
<td>0.3 dB RMS over BW</td>
</tr>
<tr>
<td>Input Signal Range</td>
<td>−100 to −65 dBm</td>
</tr>
<tr>
<td>DDC End to End Gain</td>
<td>60 to 65 dB</td>
</tr>
<tr>
<td>Image Rejection</td>
<td>&gt; 30 dB</td>
</tr>
</tbody>
</table>

3.3 Dual Downconverter Version 1

3.3.1 Hardware Description

Figures 3.1 and 3.2 show the block diagram and the actual hardware of the breadboard two-stage, two-channel downconverter. The Ku-band section of the downconverter consists of a low-noise amplifier, band-limiting filter and a mixer which downconverts the Ku-band signal to L-band. This is followed by an identical second stage which mixes the signal down to an intermediate frequency of 15 MHz.
Figure 3.1. Block Diagram of the Breadboard Dual Downconverter
Figure 3.2. Photograph of the Breadboard Dual Downconverter
3.3.2 Ku-Band Section

The Ku-band section of this version of the downconverter consisted of a Low noise Amplifier (Hittite Microwave Corporation Part Number : HMC516), a Band-pass, pre-select filter with a 100 MHz bandwidth centered at 13.285 GHz (Manufacturer : Lorch Microwave) and a mixer from Marki Microwave. The Amplifier had a gain of 19 dB in the 13.2 – 13.3 GHz range and the conversion loss of the mixer of 7 dB with a local oscillator power of 7 – 10 dBm.

3.3.3 L-Band to IF Conversion Section

The mixer and amplifiers used in this section were purchased from Cougar components. The gain of the amplifiers in this section was 20 dB and the mixer had a conversion loss of 7 dB for a local oscillator power of 7 – 10 dBm. The filter at L-band was the noise bandwidth filter. It was a cavity resonator filter with a bandwidth of 25 MHz.

Table 3.2 lists the system-level specification of the breadboard dual downconverter.

<table>
<thead>
<tr>
<th>Design Constraint</th>
<th>DDC v1.0</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency Range</td>
<td>13.275 – 13.295 GHz</td>
</tr>
<tr>
<td>Output Frequency Range</td>
<td>5 – 25 MHz</td>
</tr>
<tr>
<td>Effective Noise Bandwidth</td>
<td>25 MHz</td>
</tr>
<tr>
<td>Operating Temperature</td>
<td>–10 to 50 °C</td>
</tr>
<tr>
<td>Noise Figure</td>
<td>2.12 dB</td>
</tr>
<tr>
<td>Channel to Channel Isolation</td>
<td>78 dB</td>
</tr>
<tr>
<td>Channel to Channel Phase Stability</td>
<td>0.050° RMS over BW</td>
</tr>
<tr>
<td>Receiver Amplitude Variation</td>
<td>2dB over BW</td>
</tr>
<tr>
<td>DDC End to End Gain</td>
<td>62 dB</td>
</tr>
</tbody>
</table>
3.4 Dual Downconverter Version 2

Version 2 of the Dual Downconverter was designed, fully integrated onto a printed circuit board (PCB). The boards were made highly symmetric to ensure thermal and electrical stability to the two channels of the downconverter. In an attempt to simplify the design of the complex PCB, it was decided to split the downconverter into two sections - a Ku-band section and an L-band section, each on a separate PCB. This design approach was favored, as, in addition to simplifying the design process, it provided an easier solution to isolate any potential problems that might be encountered.

The downconverter was fabricated on two 4-layer PCBs. The Ku-band section consisted of one layer of low loss, Rogers 4350B PCB ($\epsilon_r = 3.45$) and two layers of FR4, while all layers on the L-band board were FR4. The RF signals on the Ku-band board were routed on the low-loss Rogers material, while the FR4 layers were used for power and telemetry lines. Rogers 4350B was chosen for the high frequency portion of the circuit since it had a relatively low dielectric constant and a low loss tangent. This ensured the insertion loss of a coplanar waveguide etched on this board was low. A low dielectric constant resulted in an easier integration of some of the microstrip elements, such as filters, couplers and power dividers on the board. Cadence Allegro and PCB Layout tools were used to create the PCB. Figures 3.3 and 3.4 are pictures of the Ku-band and L-band boards respectively. The schematics for the two boards have been included in the Appendix.
Figure 3.3. Photograph of the Ku-Band Section

Figure 3.4. Photograph of the L Band Section
3.4.1 Ku-Band to L-Band Downconversion

Figure 3.5 (Ku-band RF section) shows the RF section of each channel of the Ku-to-L band downconversion section. Figures A1 and A2 in the Appendix show a more detailed schematic. This section consists of one stage of low noise amplification followed by a pre-select filter and a mixer stage. The two channels of the downconverter are exactly the same, and have similar components. The Ku-band to L-band downconversion section has an end-to-end gain of 10 dB.

![Figure 3.5. Ku-Band RF Chain](image)

3.4.2 Ku-Band LNA

The first stage of any RF system sets the noise figure of the system. An ideal first stage device is one with a very low noise figure. A component with a gain at this stage ensures that the noise figure of the system is driven primarily by the noise figure of this first stage. The low noise amplifier chosen for the Dual Downconverter is a Hittite Microwave Corporation amplifier, Part Number HMC516LC5. This device has a small signal gain of 20 dB with a noise figure of only 2 dB. The amplifier is a GaAs PHEMT MMIC LNA which is available conveniently housed in a SMT package with a small footprint (5 mm x 5 mm). The gain of HMC516LC5 is stable with temperature, showing $< 2$ dB variation between $-40^\circ$ C and $+85^\circ$ C. Figure 3.6 shows the S-parameters obtained from the manufacturer.
The rectangular marker in the plot is used to highlight the gain of the amplifier in the frequency band of interest.

### 3.4.3 Pre-select Filter

The pre-select filter limits the band of frequencies that pass through the downconverter. This filter is important especially in a noisy RF environment when there are other devices which operate at frequencies near the band of interest for this downconverter - 13.275 to 13.295 GHz. The pre-select filter used in this downconverter was designed in-house with a bandwidth of 300 MHz. The insertion loss in the frequency band of interest for the interferometer is 2.5 dB with a return loss better than $-15$ dB. The filter was designed as a 4-stage coupled microstrip-line filter with a Butterworth response in the passband [4]. The dielectric used for the filter was the same as the substrate used for the rest of the
PCB. This was a Rogers 4350B substrate with a relative dielectric constant of 3.48 and a substrate thickness of 10 mils. In order to reduce spurious radiation, the pre-select filters were enclosed in a metallic RF shield. The filter was designed using ADS and Ansoft Designer.

Figure 3.7 shows the schematic of the filter. Figure 3.8 shows the simulated S-parameters of the filter. The markers indicate the performance of the filter in both the passband and the stopband.

**Figure 3.7.** Layout of the Pre-Select Filter

**Figure 3.8.** Simulated S-Parameters of the Preselect Filter
3.4.4 Ku-band to L-band Mixer

The mixer used in the dual downconverter is a double balanced GaAs MMIC mixer purchased from Hittite Microwave Corporation, Part Number HMC553LC3B. This device has a conversion loss of between 7 and 9 dB and provides a high LO-RF isolation of 50 dB. HMC553LC3B requires a Local Oscillator drive level of +9 dBm or higher. In order to maintain symmetry of the PCB layout, the RF and LO ports on one channel were interchanged, to enable easier PCB routing.

The output of the mixer, at L-band (1.275 GHz to 1.295 GHz) goes to an SMA connector (Johnson components Part Number 142 – 0711). This signal is connected to the L-band to IF conversion board using a short length of SMA (Male) to SMA (Male) cable in order to allow direct access to the signal path for characterization and testing purposes.

3.4.5 Local Oscillator Distribution Network

The Local Oscillator for the Ku-band section of the downconverter operates at 12 GHz. The oscillator was a Dielectric Resonator Oscillator purchased from Luff Research Corp. It uses a 10 MHz reference signal and has a very low phase noise of $-110$ dBc at 100 KHz. The oscillator has an output power of $+12$ dBm. The 10 MHz signal is provided by a Wenzel Associates Oscillator. Figure 6.3 shows the schematic Local Oscillator distribution network.
Figure 3.9. Local Oscillator Distribution Network
3.4.6 Local Oscillator Amplifier

The Mixers used for downconversion from Ku-band to L-band require a minimum Local Oscillator power of +9 dBm with and ideal power of +13 dBm. As mentioned above, the Luff Research Oscillator provides +12 dBm power at its output. Hence, in order to provide sufficient LO drive at the mixers, it was necessary to amplify the local oscillator signal. The amplification stage also compensates for the power lost in the RF splitter and filters which follow this stage. The amplifier chosen for this application is a Hittite Microwave Corporation HMC490LP5 GaAs PHEMT device, which is available in a leadless QFN SMT package. This device has a gain of +23 dB and a noise figure of 2.5 dB. It has an output Third Order Intercept point of +34 dBm. In order to ensure the amplifier operates in the linear region, the output of the oscillator is reduced by a 6 dB pad before the signal reaches the amplifier. Figure 3.10 shows the S-parameters of the amplifier.

![S Parameters of the HMC490 Amplifier](image)

**Figure 3.10.** S-Parameters of the HMC490 Amplifier
3.4.7 RF Power Splitter

In order to distribute the Local Oscillator power equally to the two channels of the downconverter, a Wilkinson Power Splitter was designed. This was implemented in a microstrip line form and a Vishay Electronics 100 Ω resistor was used as the isolation resistor. The design was carried out using Ansoft Designer. The substrate used to realize the splitter was Rogers RO4350B, 10 mils thick. The simulated insertion loss was 3.5 dB and isolation between the two arms of the splitter was $<-20$ dB.

The splitter’s performance in the laboratory showed a dependence on the position of the 100 Ω resistor. It was found that the resistor, which had a circular cross-section, was not making proper contact with the solder pads on the board. A future version of the downconverter will dispense with the Wilkinson splitter and employ a Hybrid Splitter instead. The hybrid not only provides better isolation between the two arms of the splitter, but also eliminates the dependence of performance on the isolation resistor.

3.4.8 Local Oscillator Filters

Similar to the design of the Ku-band pre-select filters, the Local Oscillator filters were also designed as coupled microstrip filters, on the same Rogers RO4350B substrate. Figures 3.11 and 3.12 show the filter structure and the simulated S-parameters. The filter has a bandwidth of 600 MHz with an insertion loss of 3.83 dB at 12 GHz. The center frequency of the filter was set to a slightly higher frequency to account for a discrepancy in the specified dielectric constant of the substrate. The substrate manufacturer (Rogers Corp) indicated that the dielectric constant used in the CAD package might be slightly lower than what they had found from feedback from PCB manufacturers. As with the pre-select filters, the local oscillator filter was enclosed in a metallic shield as well. This was done to minimize potential radiation from the filter which could interfere with signals from the RF sections, reducing the channel-to-channel isolation.
The local oscillator network can provide a path for the RF signal from one channel to couple into the second channel. In order to reduce coupling between the two channels through this path, a high level of isolation between the arms of the Local Oscillator network is required. Figure 3.13 is the setup used to simulate and verify the isolation specification and Figure 3.14 shows the simulated results. The first curve corresponds to the S-parameters between the Wilkinson Divider and output of the filter, while the second plot is the S-parameter measured between the outputs of the two filters. This is the isolation
between the two RF channels. It can be seen that the isolation between the two channels is better than 70 dB at 13.285 GHz. In this simulation, the worst case was assumed. It was assumed that the filters were terminated in unmatched loads. If the filters were terminated in matched loads, the isolation will improve because of the suppression of reflected waves.

**Figure 3.13.** Layout Used to Simulate Local Oscillator Channel Isolation

![Layout Used to Simulate Local Oscillator Channel Isolation](image)

**Figure 3.14.** Simulated LO Channel Isolation

![Simulated LO Channel Isolation](image)

<table>
<thead>
<tr>
<th>Frequency (GHz)</th>
<th>Isolation (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>12.3</td>
<td>34</td>
</tr>
<tr>
<td>13.21</td>
<td>56</td>
</tr>
<tr>
<td>11.01</td>
<td>68</td>
</tr>
<tr>
<td>13.21</td>
<td>79</td>
</tr>
</tbody>
</table>

3.4.9 Local Oscillator Test Point

The Local Oscillator power needs to be sufficient to drive the mixers at the correct level to ensure good performance from the mixers. Hence, it is important to track the power being supplied to the mixer, and for this purpose, a Local Oscillator test point was added to
the circuit, shown in Figure 3.15. This consists of a microstrip coupler which taps a small portion of power from the Local Oscillator chain right after the Local Oscillator amplifier. Figures 3.16 and 3.17 show the microstrip layout and S-parameters of the Coupler used to tap off power from the Local Oscillator network to the Power Detection Circuit. Power detection is performed using an Analog Devices Log Detector, Part Number AD8317. The detected power level is then read by one of the channels of an Analog-to-Digital Converter in the Telemetry section which can then be read from a computer. In addition, there are two LEDs which indicate if there is sufficient Local Oscillator power. The LEDs serve as a quick visual indication of the presence of sufficient Local Oscillator drive power.

![Figure 3.15. Local Oscillator Test Point Schematic](image)

![Figure 3.16. Layout of the 12 GHz Hybrid Coupler](image)
3.4.10 Telemetry and Power Conditioning

The importance of monitoring temperature of the downconverter was obvious from the phase measurements made on the breadboard version of the Dual Downconverter. Hence, the PCB version of the downconverter was equipped with on-board temperature sensors which monitor temperature at various points on the board. The Ku-band to L-band conversion stage has five precision integrated circuit temperature sensors. These devices (National Semiconductors Part Number: LM61) can sense a $-30^\circ$ C to $+100^\circ$ C temperature range. Figure 3.18 shows the sensor along with its biasing circuit. Telemetry data obtained from the power detector and the temperature sensors is digitized by two Maxim-IC Analog-to-Digital Converters (ADC), Part Number MAX1168. Each ADC can read up to 8 single ended analog channels, or 4 differential channels. This ADC is a 16 bit Successive Approximation ADC and can sample at up to 200,000 samples per second. In addition, the

![Figure 3.17. Simulated S-parameters of the Hybrid Coupler](image-url)
ADC offers multiple options to interface to the computer. For this application, a program was written in the C programming language to communicate with the ADC over the parallel port of the computer. Providing the correct clock signal and synchronizing data with this clock proved tricky using this simple interface. Due to these timing issues, the ADC was clocked well below the maximum rated clock frequency to ensure correct timing. In order to operate the ADCs at or near the maximum rated clock frequency, a microprocessor would be required, which will act as an interface between the remote computer and the ADCs. One of the two ADCs on this board measure 5 channels of temperature data. The second ADC monitors the current drawn by the amplifiers on this board. Table 3.3 shows the channel allocation for the two ADCs.

<table>
<thead>
<tr>
<th>Channel Number</th>
<th>ADC 1</th>
<th>ADC2</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>+Z Mixer Temperature</td>
<td>+Z Amplifier Current Positive</td>
</tr>
<tr>
<td>2</td>
<td>+Z Amplifier Temperature</td>
<td>+Z Amplifier Current Negative</td>
</tr>
<tr>
<td>3</td>
<td>LO Amplifier Temperature</td>
<td>-Z Amplifier Current Positive</td>
</tr>
<tr>
<td>4</td>
<td>-Z Mixer Temperature</td>
<td>-Z Amplifier Current Negative</td>
</tr>
<tr>
<td>5</td>
<td>-Z Amplifier Temperature</td>
<td>LO Amplifier Current Positive</td>
</tr>
<tr>
<td>6</td>
<td>LO Power</td>
<td>LO Amplifier Current Negative</td>
</tr>
<tr>
<td>7</td>
<td>5.5V DC Positive</td>
<td>No Connection</td>
</tr>
<tr>
<td>8</td>
<td>5.5V DC Negative</td>
<td>No Connection</td>
</tr>
</tbody>
</table>

Power conditioning is provided by linear voltage regulators and associated power conditioning filters which ensure a clean bias voltage on all the amplifiers. The Ku-band to L-band board receives a 5.5 V DC supply from the L-band to IF conversion board. This
5.5 V is regulated by a linear voltage regulator which converts the 5.5V DC level to a 5 V level. This is then further regulated to the 3V DC level required by the amplifiers. Each amplifier has its own 5V to 3V regulator and associated filter (shown in Figure 3.19). The 5V signal is used to power the ADCs, power detector and the temperature sensors.

![Figure 3.19. Linear Voltage Regulator Circuit](image)

### 3.4.11 L-Band to IF Conversion

### 3.4.12 RF Chain

The RF subsystem of the L-band board consists of two stages of amplification, two stages of filtering and a mixer stage. In addition, this section has an L-band RF Test Point. A small fraction of power is tapped off the RF chain and this goes to an SMA connector where the signal can be monitored.

### 3.4.13 L-Band Filters

The two filters used in the L-band section are packaged, ceramic resonator filters, sourced from Lorch Microwave Inc. The first filter is a 6 section filter with a 100 MHz bandwidth, centered at 1.28 GHz. This filter provides additional attenuation to spurious signals outside the 1.23 to 1.32 GHz bandwidth.

The second filter is a 6 section filter with a bandwidth of 30 MHz. This is the filter which sets the system noise bandwidth. Figures 3.20 and 3.21 show the S-parameters of these filters measured by the manufacturer.
Figure 3.20. Measured S-Parameters of the L-Band Filter

Figure 3.21. Measured S-Parameters of the Noise Bandwidth Filter
3.4.14 L-Band and IF Amplifiers

The first stage of amplification directly follows the 100 MHz filter. This is a simple gain block purchased from WJ Associates. This device typically provides 19 dB gain at 1 GHz, with a noise figure of 3.3 dB. The second amplifier directly follows the hybrid coupler used to tap off power for the RF test point.

At the intermediate frequency (5 - 25 MHz), a Sirenza Microwave Gain block, Part Number SGA-6589 was used. This provides an additional 25 dB of gain.

3.4.15 L-Band Mixer

The mixer chosen for the L-band to IF frequency conversion was a Minicircuits Part Number: ADE-11X mixer. This provides a high RF-LO isolation of 36 dB (typical) and a conversion loss of 7 dB (typ). As with the Ku-band mixer, the LO and RF ports on one channel were interchanged.

3.4.16 Low Pass Filter

The Low pass filter used at IF (5-25 MHz) was purchased from Lark Engineering. This filter has a bandwidth of 30 MHz.

3.4.17 Local Oscillator Distribution Network

The Local Oscillator that was used at 1.3 GHz was a Crystal Oscillator, also purchased from Luff Research. This oscillator also uses a 10 MHz reference signal. The output power of the 1.3 GHz oscillator is +15 dBm. The Local Oscillator distribution network consists of an amplifier (Part Number: ECG005B), a directional coupler (Part Number: Minicircuits TCD-9-1W) to tap off power to the local oscillator test point and a power splitter (Part Number: Minicircuits QCN-19). Local Oscillator filters, with a center frequency of 1.3 GHz follow the power splitter. These filters were purchased from Lorch Microwave.
3.4.18 Telemetry and Power Conditioning

The Dual Downconverter takes a 15 V DC external power supply from two pins of a Type D, high density, sub-miniature connector, which is connected to an external power supply. The other pins of the D-Sub connector are used by the computer to remotely communicate with the ADCs on board. In addition to the D-Sub connector, there is an additional DC connector which allows connecting a general purpose DC supply to the board. This is an additional feature, which allows one to isolate potential power supply problems from any other problem. The DC supply is monitored by one of the on-board ADCs along with LEDs which provide a visual check on the presence of the supply voltage.

The L-band boards have a switching, step-down voltage regulator, which provides an efficient 15 V external supply conversion to an intermediate 7 V supply. The switching regulator has an internal fixed oscillator frequency of 260 KHz. A filtering circuit follows the switching regulator to filter out any leakage of the oscillator into the DC supply, thus ensuring a clean supply voltage to all components. The clean 7V supply is then fed to multiple linear voltage regulators which step the voltage down to 6 V and 5 V that is used by the active and telemetry components. The L-band board uses the same temperature sensors as the Ku-band board. The L-band board as a total of eight temperature sensors on board. The sensors are monitored using three Maxim-IC ADCs. In addition to monitoring

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**Figure 3.22.** Switching Regulator Circuit

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the temperature sensors, the ADCs also monitor current and Local Oscillator power. Table 3.4 shows the channel allocation for each of the three ADCs.

**Table 3.4. ADC Channel Allocation -2**

<table>
<thead>
<tr>
<th>Channel Number</th>
<th>ADC 1</th>
<th>ADC 2</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>+Z L-Band Amplifier Temperature</td>
<td>-Z L-Band Amplifier Temperature</td>
</tr>
<tr>
<td>2</td>
<td>+Z Video Amplifier Temperature</td>
<td>-Z Video Amplifier Temperature</td>
</tr>
<tr>
<td>3</td>
<td>+Z Mixer Temperature</td>
<td>-Z Mixer Temperature</td>
</tr>
<tr>
<td>4</td>
<td>+Z L-Band Amplifier Voltage</td>
<td>-Z L-Band Amplifier Voltage</td>
</tr>
<tr>
<td>5</td>
<td>+Z 3V Supply Voltage</td>
<td>-Z 3V Supply Voltage</td>
</tr>
<tr>
<td>6</td>
<td>+Z L-Band Amplifier Voltage</td>
<td>-Z L-Band Amplifier Voltage</td>
</tr>
<tr>
<td>7</td>
<td>L-Band RF Power</td>
<td>L-Band RF Power</td>
</tr>
<tr>
<td>8</td>
<td>RF Power</td>
<td>RF Power</td>
</tr>
</tbody>
</table>

**Table 3.5. ADC Channel Allocation -3**

<table>
<thead>
<tr>
<th>Channel Number</th>
<th>ADC3</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>LO Amplifier Temperature</td>
</tr>
<tr>
<td>2</td>
<td>LO Power</td>
</tr>
<tr>
<td>3</td>
<td>LO Filter Temperature</td>
</tr>
<tr>
<td>4</td>
<td>No Connection</td>
</tr>
<tr>
<td>5</td>
<td>No Connection</td>
</tr>
<tr>
<td>6</td>
<td>DC Supply</td>
</tr>
<tr>
<td>7</td>
<td>3.5V Positive</td>
</tr>
<tr>
<td>8</td>
<td>3.5V Negative</td>
</tr>
</tbody>
</table>
Figure 3.23. Analog to Digital Converter Circuit
CHAPTER 4
RESULTS AND DISCUSSION

4.1 Performance of the Dual Downconverter Version 1

The amplitude and phase performance of the downconverter was tested in the laboratory. This section presents the amplitude and phase performance results of the breadboard version of the downconverter.

4.1.1 Amplitude Performance

Figure 4.1 shows the gain of each channel of the downconverter over the 20 MHz bandwidth at the intermediate frequency. The variation of gain over the bandwidth for each channel is $< 2$ dB. Figure 4.2 shows the channel-to-channel variation in gain. Version 1.0 of the downconverter has a gain of 62 dB (mid-band) with a variation of $< 2$ dB within the bandwidth. It exhibits $< 1$ dB variation in gain between the two channels. The two channels exhibit a high isolation of 78 dB. The isolation between the two channels was high since almost all components connectorized and were well packaged in their own metallic enclosures. Hence, spurious radiation was almost non-existent. The active components were initially powered using a standard laboratory power supply. This was later changed to a stand-alone power supply unit which was built specifically to power the dual downconverter.
Figure 4.1. Gain of the Dual Downconverter

Figure 4.2. Channel-to-Channel Gain Variation
4.1.2 Phase Performance

The importance of the two-channel relative phase difference has been emphasized throughout this thesis. Conventional methods of measuring phase difference using a network analyzer were insufficient for the purposes of this work. The phase stability of the network analyzer was greater than the phase difference of the downconverter. Hence, the technique discussed in [1] and [7], which uses a Maximum Likelihood Estimator, was used to measure the phase difference between the two channels to millidegree accuracy. The test setup used for phase measurements is shown in Figure 4.3.

![Test Setup used for Phase Measurements](image)

**Figure 4.3.** Test Setup used for Phase Measurements

Figure 4.4 shows the results of initial phase measurements performed on the two-channel downconverter. The first plot is the phase of each channel, represented by circles and dots. An ideal response, for a system with zero phase difference, would have each dot perfectly centered within the circles. Additionally, this plot shows the absolute phase stability of the system, as there is no strong trend in one single channel phase over time. The center plot shows the phase difference between the two channels as a function of time. The phase difference between the two channels show a clear periodic variation with time. The period of each cycle is of the order of 20 minutes. During this test, the downconverter
was left open to ambient surroundings in the laboratory. The first suspicion for the cause of this periodic variation was that there was a periodic variation in the ambient temperature in the room. In order to gain better insight on the reasons for the periodic variation in phase over time, temperature sensors were used to monitor temperature at a number of points on the downconverter - this constitutes the third plot in Figure 4.4. From these plots, it can be seen that the phase difference of the downconverter, was indeed following the ambient temperature cycle. At this point, it was noticed that the downconverter had been placed below an air conditioning vent in the room, and the temperature cycle corresponded to the cycle of the air conditioning thermostat. At this point, it was becoming clear that in order to meet any phase specification for the downconverter, a more stable thermal environment was required.

![Figure 4.4. Phase estimates of DDC](image)

Figure 4.4. Phase estimates of DDC
In order to thermally isolate the downconverter from ambient temperature fluctuations, it was placed in a rudimentary, thermally isolated environment - a commonly available cooler. Figure 4.5 shows the setup after partial thermal isolation. Figure 4.6 shows the phase performance of the downconverter in this test setup. The short-term periodic fluctuations of phase were nearly eliminated. However, there was a longer-term trend which corresponded to the increasing temperature inside the cooler due to the components of the downconverter heating up with time. It can be seen that though the swing in phase difference is close to $0.5^\circ$ over the 8 hour period over which the test was run, the phase difference over smaller intervals of time, of the order of a few minutes is $< 0.05^\circ$. Hence, the downconverter meets the required phase specifications under a stable thermal environment.

Figure 4.5. Dual Downconverter in a Stable Thermal Environment

Once it was established that, given a stable ambient temperature the downconverter performed as expected, the next step was to try to change the temperature inside the cooler to see how the downconverter would perform under different temperature conditions. Further moderately controlled temperature tests were performed. These tests, however, provided only a small degree of control over the ambient temperature inside the cooler. Figure 4.7
Figure 4.6. Phase estimates of DDC - Slow Increase in Temperature

shows the result of artificially increasing and decreasing the ambient temperature inside the cooler. This was accomplished by adding ice bags inside the cooler to lower temperature and using electrically controlled heating pads on the base plate of the downconverter to artificially introduce thermal gradients across the downconverter. The response shows a clear and strong sensitivity to thermal shocks in the system, with recovery time constants on the order of minutes. These long time lags are likely due to the natural thermal interia of the connectorized components themselves and the thermal path provided by the aluminium plate that they are mounted on.

The amplitude and phase measurements made on the breadboard version show that, given stable room temperature conditions, it is possible to achieve a highly phase stable
receiver which would provide the required accuracy when used as part of a high precision interferometer. However, the connectorized version occupies a lot of space which is not desirable in a space-borne system, where volume occupied and weight of the instrument is as important as the electrical performance. Hence, the PCB version is more attractive as a space-borne version for the downconverter. Challenges such as the effect of channel cross-talk were not encountered in the breadboard version as each component was well shielded in its own enclosure, minimizing any spurious radiation.

4.2 Performance of the Dual Downconverter Version 2

The breadboard version provided interesting and important insight into the performance of a two channel downconverter. It was clear that any attempt to miniaturize the system should be done with much importance given to thermal symmetry and stability. With this in mind, the PCB version of the downconverter was designed to be highly symmetric.
Metallic shields were placed around filters and switching circuitry in the power conditioning network. This was done to minimize spurious radiation between channels which could potentially affect the phase performance of the system.

4.2.1 Amplitude Performance

Figures 4.8 and 4.9 show the measured channel gains and channel-to-channel gain difference for the downconverter. The PCB has a mid band gain of 61dB. It shows between 0.5dB and 1.5dB variation in gain between the two channels and less than 2dB gain variation over the bandwidth for each channel. The RF and LO ports of the Ku-band mixer on one channel were flipped to enable easier RF trace routing. A mixer evaluation board was tested to see the effect of flipping the two ports. It was found that the mixer suffered an additional 0.5 dB conversion loss due to this arrangement, an acceptable loss considering the advantage of easy signal routing on the PCB.

![Gain of the Dual Downconverter v2.0](image)

**Figure 4.8.** Gain of the Dual Downconverter v2.0
4.2.2 Phase Performance

Figure 4.10 shows the phase performance of the PCB version of the DDC. Over a period of two hours, the standard deviation of phase difference between the two channels of the DDC was of the order of 65 millidegrees. Phase measurements as a function of temperature for the breadboard version were performed using a rudimentary setup which did not provide sufficient control over the temperature through which the breadboard was cycled. In order to make more controlled temperature measurements, Temptronic Corporation, based in Sharon, MA, graciously allowed the use of their temperature control equipment in their facility. Temptronic specializes in designing and manufacturing localized thermal inducing systems. They design and manufacture precise thermal control equipment primarily for the semi-conductor industry, for characterization, qualifying and performing fault isolation on a component, PCB, module or wafer to meet military specifications, design or production testing. For the purpose of this experiment, the Dual Downconverter was tested
using a ThermoSystem\textsuperscript{TM} TPO4300 Series environmental test system. This test system can provide a stable thermal environment from $-85^\circ$ to $225^\circ$C and can temperature can be changed at the rate of $1^\circ$C/minute. Figures 4.11 and 4.12 show the setup used to make the temperature measurements at Temptronic and Figure 4.13 shows the Dual Downconverter at Temptronic’s facility at Sharon, MA.
Figure 4.11. Phase Measurement Setup at Tempronic Corporation

Figure 4.12. Tempronic’s ThermoSystem\textsuperscript{TM} TPO4300 Series
Figure 4.13. Dual Downconverter Inside the Thermal Chamber

The Downconverter was cycled through a temperature range of $-10^\circ$ C to $40^\circ$ C. The downconverter and phase measurement system was left at Temptronic’s facility for a week and the system was controlled over the ethernet, with a program written using MATLAB. After each measurement cycle, the Temptronic system was turned off and on a few occasions, the system was turned off over-night. Since the downconverter was placed in an air-tight enclosure, there was some concern about the downconverter over-heating over a period of time, when the system was not in use. This concern was addressed by a feature on the Temptronic system which allowed the system to periodically cool the enclosure, even after the system was placed on stand-by mode.

Figure 4.14 shows one of the temperature cycles which the downconverter was subjected. The first plot shows the absolute phase of the two channels of the downconverter. The second plot shows the relative phase difference between the two channels, and the third plot shows the standard deviation of phase difference as a function of ambient temperature. This temperature cycle was run over a period of over three hours. It can be seen that there is a clear temperature dependence, as was seen on the breadboard version. The lowest phase difference occurs at room temperature and is of the order of 70 millidegrees.
In the first measurement, the system was cycled through temperature with a dwell time of 15 minutes at each temperature point. In the next set of measurements, the dwell time was increased to 20 minutes. The thermal chamber was allowed to stabilize at each set point for 5 minutes and phase measurements were made for the next 15 minutes at each temperature setpoint. Allowing the ambient temperature settle at a particular set point ensured a more stable ambient temperature while making phase measurements - a lesson learned from the performance of the breadboard version. Figure 4.15 shows the phase measurement and temperature plots for this test case. As was observed in the previous measurements, the downconverter continued to show the same temperature dependence. When any system...
is put onto a spacecraft, it is encapsulated in a thermal blanket to protect it from extreme temperatures the spacecraft is subject to during its lifetime. However, as the spacecraft moves into and away from sunlight, the temperature inside the spacecraft and inside the thermal blanket can fluctuate at the rate of a 1° C/min. From an operational stand-point, it is important that when the system experiences such a thermal shock, the system should not exhibit any hysteresis. The system should return to its earlier state once the thermal shock is removed.

An interesting test case for the downconverter would, therefore, be one in which the downconverter encounters a sudden thermal shock, similar to what it might experience...
when onboard a spacecraft. This case was tested in another set of tests. Figure 4.16 shows the case where the downconverter was subject to a similar temperature profile as was done in the previous case. But, once the temperature reached 40° C, it was suddenly lowered to 15° C and held constant for a while. There is a big jump in phase difference, however, when the temperature returns to 15° C, the phase difference returns to the same difference as was seen when the downconverter was cycled through the slow temperature profile. In the initial temperature buildup, the phase difference was 0.568° when the temperature was held at 15° C. After the abrupt temperature change, the phase difference was 0.575°. Hence, from these tests, it appears that the phase difference as a function of temperature does not exhibit hysteresis and the two channel phase difference returns to the same level before and after a thermal shock.

The performance of the dual downconverter shows that achieving relative phase stability of 50 millidegrees is a challenging task. At room temperature, the PCB version shows a relative phase difference of 65 millidegrees. This can be improved by a few millidegrees with better isolation between the channels. In order to achieve 50 milligree phase stability over extended periods of time, it appears that the operating temperature of the downconverter will have to be controlled to be close to 20 – 25° C, but this characterization component of the work remains ongoing.
Figure 4.16. Downconverter Subject to a Thermal Shock
CHAPTER 5
INTERFEROMETRIC MEASUREMENTS

The development of a high performance downconverter was the first, and important, step towards the construction of an interferometric radar. The interferometer was named TIMMi - Topographic Ice Mapping Mission, as one of its intended missions is to study glaciers, sea ice and the volume scattering properties of snow. Deploying the interferometer required building a transmitter and antennas to launch the signals out at the intended target. The transmitter uses a frequency conversion scheme which was similar to the downconverter, only in reverse, that is, an intermediate frequency signal at 15 MHz was up-converted first to 1.285 GHz and then to 13.285 GHz. The advantages of this scheme are two-fold.

1. Utilizing the same frequency scheme enabled using the same local oscillators that were being used for the downconverter, and,

2. As the transmitter utilizes the same structure as the downconverter, spare components that were procured for the breadboard version of the downconverter could be used to build the transmitter. This saved time and money involved in buying additional components.

In addition to the upconverter, a 2 W power amplifier was bought to increase the output power of the transmitter.

The interferometer is currently configured to operate with a 20 MHz bandwidth. This bandwidth results in a range resolution of 7.5 m. Though this resolution might be sufficient for an air-borne or space-borne application, it is not ideal for a ground-based interferometer.
The next version of the interferometer will operate with a bandwidth of 100 MHz, resulting in a range resolution of 1.5 m. The downconverter for this version is currently being built.

A network analyzer (HP8753C) was used as the data acquisition system for TIMMi. Figure 5.1 shows the TIMMi measurement setup using the Network Analyzer. A similar technique, outlined in [9], has been used to make scatterometric and polarimetric measurements. The RF out port of the network analyzer is used to generate the IF signal which is then upconverted and transmitted. The signal reflected of the target being observed is received by the two antennas, downconverted by the DDC and is fed into the receive ports of the network analyzer. The network analyzer has a built-in Frequency-to-Time-Domain conversion utility which reduces the need for any significant additional post processing and displays data directly in time domain. The interferometer requires four sets of data - Amplitude and Phase from each channel. The network analyzer, however, was only capable of simultaneously making only two of the four required measurements, while also performing a time-domain transformation. To overcome this problem, the measurement scheme was broken down to make two measurements at a time. The first set of data acquired were the amplitudes for both channels simultaneously, followed by a second measurement of the phase of both channels. There was a 15 second time lag between both sets of measurements. As long as the system (including the antennas) and the target being illuminated are static in that 15 second time period, the time lag between measurements should not have any impact.
Figure 5.1. TIMMi Measurement Setup

5.1 Antenna Design

In order to make interferometric measurements, antennas were required to launch and receive signals from the intended target. The azimuth beamwidth of the antenna determines the azimuth resolution of the interferometer. Ideally, this should be as small as possible. A 1 m azimuth resolution requires a 3 dB beamwidth of close to 2°. The elevation beamwidth of the antenna should be as wide as possible so that the interferometer sees as long a range as possible. The resolution in this direction is set by the system bandwidth and not the antenna pattern.

Two design approaches were initially considered. The first option was to fabricate a 600 mm long slotted waveguide antenna to meet the azimuth beamwidth requirement. However, this option worked out to be very expensive and the manufacturer quoted an unusually long leadtime. The second option was to build a patch antenna array [2]. An aperture coupled patch array was chosen at it is the most versatile of the different patch antenna configurations. Figure 5.2 shows a section of the aperture coupled patch antenna array. Figure 5.3 shows the antenna which was tested in a near-field anechoic chamber at

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the Center for Advanced Sensor and Communication Antennas (CASCA) in the University of Massachusetts.

The antenna structure consisted of two PCBs - the corporate feed network for the antenna was designed on a low-loss Rogers 5880 material. This was chosen to minimize the loss of the feed network. The patch radiators were etched on an FR4 PCB. This PCB also served as the radome for the antenna. The two PCBs were spaced 62 mils apart. Initially, nylon spacers were used between the boards. However, it was found that the Rogers 5880 material was warping. Because of this effect, it was difficult to keep both boards at a constant distance from each other. In order to overcome this difficulty, a 62 mil Rohacell HF32 board was added in between the FR4 and Rogers 5880 board. The Rohacell board has a relative dielectric constant $\epsilon_r$ of 1.03. The low dielectric constant of this spacer does not affect the antenna performance, but serves to improve the mechanical stability of the antenna. The Rohacell board was fairly fragile and was prone to break easily. Hence, much care had to be taken to ensure the spacer board did not break while handling the antenna. The antenna parameters are tabulated in Table 5.1. The simulated and measured patterns are shown in Figure 5.4.

The antenna had a $-15$ dB return loss bandwidth of 150 MHz. A comparison of the simulated and measured pattern shows close agreement for the most part. However, in the azimuth cut, the measured results show a higher first side-lobe level on one side. This could potentially be due to the PCB warping. The higher than expected side-lobe in effect decreases the azimuth resolution. Given that the sidelobe is 10 dB lower than the main lobe, the bias in interferometric height that targets illuminated by the high side-lobe will introduce, is expected to be fairly small.
Table 5.1. Antenna Specifications

<table>
<thead>
<tr>
<th>Antenna Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency Range</td>
<td>13.275 – 13.295 GHz</td>
</tr>
<tr>
<td>Polarization</td>
<td>Horizontal</td>
</tr>
<tr>
<td>Gain</td>
<td>20 dBi</td>
</tr>
<tr>
<td>VSWR</td>
<td>&lt; 1.5 : 1</td>
</tr>
<tr>
<td>Azimuth Beamwidth</td>
<td>2°</td>
</tr>
<tr>
<td>Elevation Beamwidth</td>
<td>38°</td>
</tr>
<tr>
<td>Physical Size</td>
<td>55cm x 6cm</td>
</tr>
</tbody>
</table>

Figure 5.2. Cross-Section of the Patch Array Antenna

Figure 5.3. Patch Array Antenna - Patch Radiators (left-above), Corporate Feed network (left-below), Antenna under test at CASCA (right)
Figure 5.4. Antenna Pattern - Simulated (top-left) and Measured (azimuth - top right and elevation - bottom)
5.2 Field Deployments

TIMMi was deployed on three separate experiments, twice at the Center for Atmospheric Research Experiments (CARE), Ontario, Canada, and once on top of the Lederle Graduate Research Center Tower at the University of Massachusetts, Amherst. With each experimental deployment, improvements were made and some of the results are presented below.

5.2.1 CARE Expedition

Test Site Description

The interferometer was deployed in January, 2007 at the Center for Atmospheric Research Experiments (CARE), Ontario, Canada for a period of four days, followed by a second deployment in February, for a period of 5 days. The interferometer was installed on the roof of the main building of the CARE facility, 16 meters from ground level. The target was a piece of snow covered land, with a slow gradient away from the building. Figure 5.5 shows the interferometer deployed on the roof of the CARE facility.

Figure 5.5. TIMMi Deployed at the CARE Site
The antennas were mounted on a pole such that there was a 1.5 m clearance between the lowest antenna and the wall on the roof of the building. The antenna pole was guyed down to the roof with a rope. During this deployment, the antennas did not have a positioner. Azimuth positioning was done by rotating the antenna pole by hand.

**Interferometric Measurement Results**

The interferometric correlation is a measure of the similarity of the signals obtained by the two antennas. This is given by,

\[
\gamma = \frac{\langle E_1 \cdot E_2^\ast \rangle}{\sqrt{\langle |E_1|^2 \rangle \cdot \langle |E_2|^2 \rangle}}
\]  

(5.1)

where, \( E_1 \) and \( E_2 \) are the magnitudes of the signals received at Antenna 1 and 2 respectively. The asterix indicates the complex conjugate, and the brackets indicate averaging over independent samples (or looks). The phase of this complex correlation coefficient is the interferometric phase which gives rise to the topography. The magnitude indicates the consistency of the interferometric correlation over the number of looks. Note that nominally, these looks are taken spatially, as a measure of topographic variation. These looks however may also be taken as a function of time, to gauge the consistency of measurement.

Figures 5.6 and 5.7 show the interferometric correlation (amplitude and phase averaged over time) between the signals received at the two antennas and the estimated topography. These are plotted as a function of ground range, from the base of the building, for 13 azimuth positions. These azimuth angles, though approximately adjacent, were positioned by hand and are not well known. It can be seen that the signal decorrelates at ground ranges close to the building and then improves. At large distances from the building, the signal decorrelates again. This behaviour is consistent with a signal being reflected and dissipating as it travels further away from the antennas. The distinct line of correlation that appears at 35 meters from the building was likely due to low SNR, or, as would be shown later, potentially due to phase errors in the network analyzer.
Figure 5.6. Interferometric Correlation as a function of ground range for 13 azimuth angles - January 25, 2007

Figure 5.7. Measured Topography from ground at the CARE site - January 25, 2007
The correlation plot shown above was averaged over time. Signal decorrelation can occur if

1. the SNR is low

2. the target or the observing platform moves between measurements, or,

3. there is any change in the measurement system itself.

From the data collected during this expedition, it appeared that the signal was decorrelating in every dataset. One of the major causes for concern was that the antennas were not stable enough and were moving in windy conditions during the measurement runs.

The interferometer was deployed for a second time in February, over a period of five days at the same test site at CARE. On this occasion, the antenna pole was fitted with an antenna positioner. Fixing the antenna pole to the positioner, while providing improved the mechanical stability of the antennas, it gave better control over the azimuth positions.

During this test run, the ground target was covered with close to 6 to 7 inches of snow. Figures 5.8 and 5.9 show correlation and topography measured on one of the days during this deployment. When the individual samples are not averaged, the measured correlation is close to unity. However, time averaging the amplitude and phase leads to decorrelation. This again shows that there was something in the system which was changing randomly with time. Though the antennas were more stable in this measurement setup than in the previous setup, it was still suspected that the antennas were the major source of error.
Figure 5.8. Interferometric Correlation as a function of ground range for 13 azimuth angles - February 27, 2007

Figure 5.9. Measured Topography from ground at the CARE site - February 27, 2007
At this time, it was suggested that one source of error that was being overlooked was that the interferometer was installed too close to the ground. Hence, the variation of phase within each resolution element was more than $2\pi$ radians. Due to such a large phase variation within a cell, the phase signature measured would lead to erroneous results.

5.2.2 Lederle Graduate Research Center Deployment

In June 2007, we obtained permission to deploy TIMMi on the 17th floor of the Lederle Graduate Research Center Tower (LGRT) at the University of Massachusetts, Amherst. This deployment provided the opportunity to test the interferometer mounted at a good height away from the ground. The height of the tower was estimated to be 50m. A further improvement was made to the antenna mounting structure. This time, a wooden frame was made for the antennas. This ensured that the antennas were more stable in windy conditions than they were during the deployments in Canada. Azimuth positioning was achieved by moving the antenna structure along the wall.

Figures 5.10 and 5.11 are the correlation and measured topography plots for this deployment. Once again, the signal appears to decorrelate very quickly. The horizontal axis in Figure 5.10 has been calibrated in terms of number of points of the Network Analyzer frequency sweep. This has been done in order to compare these results with the measurements made of the stability of the network analyzer presented later. In order to relate the correlation plot and the topography plot, ground range between 20 meters and 100 meters on the topography plot correspond to points 110 to 201 on the correlation plot.
Figure 5.10. Interferometric Correlation as a function of ground range for 6 azimuth angles at the Lederle Graduate Research Center, University of Massachusetts, Amherst - June 26, 2007

Figure 5.11. Measured Topography from the Lederle Graduate Research Center, University of Massachusetts, Amherst - June 26, 2007
All the measurements made in Canada as well at the University of Massachusetts showed significant decorrelation in the signals led us to take a closer look at the stability of the data acquisition system. As a test, the output from the network analyzer which was being fed to the transmitter, was connected back into the network analyzer through a power splitter using very short cables. The two channel data was processed similar to the earlier interferometric measurement datasets. Figures 5.12 and 5.13 show the correlation and the measured phase difference between the two channels of the network analyzer. This data was acquired over a period of 30 minutes and has been averaged over time. If the network analyzer was stable, then the correlation should be exactly unity, because the signal, as a function of time, would be unchanging. However, the correlation is less than one over the entire time period and it drops to very low values at certain instances. The phase difference plot clearly shows that the measured phase of the network analyzer is not as stable as previously thought and fluctuates in a random manner. This random phase fluctuation causes the correlation to drop below unity. It can be seen that the low correlation regions correspond to the spikes in the phase measurements. These results show that the earlier interferometric measurements might have been corrupted because of the phase instability in the data acquisition system. These results show that the network analyzer might have been the major source of error in all the interferometric datasets.
Figure 5.12. Correlation between the Two Network Analyzer Channels

Figure 5.13. Phase Difference Between the Two Network Analyzer Channels
CHAPTER 6
CONCLUSIONS AND FUTURE WORK

6.1 Summary of Results

Radar interferometry has been receiving much attention in the remote sensing community and there are many new applications that are being proposed for interferometric measurements. Unlike polarimeters and other radar techniques, an interferometer relies on receiving signals on multiple antennas separated by a baseline, which needs to be accurately known. The need for a baseline makes the deployment of interferometers at lower frequencies more difficult and expensive, especially for space-borne applications. As the frequency of operation increases, the length of the baseline reduces, making it more attractive for space applications. However, designing radio frequency equipment at higher frequencies is more challenging. As the frequency increases, the performance requirement of the entire system becomes more stringent.

A two-channel downconverter forms the core of the interferometric measurement setup. Knowledge of the performance (and limitations) of the downconverter provides the system engineer with the information needed to design a highly accurate interferometer. The emphasis of this thesis has been to design, implement and characterize a high performance, two-channel downconverter which operates at Ku-band. It has been an effort to determine the performance of such a downconverter and try to identify potential issues which could affect especially the phase performance of the downconverter. The lack of widespread knowledge of such performance metrics for downconverters operating at Ku-band emphasizes the importance of the work done in this thesis.
A system-level error budget analysis was performed to determine the contribution of major error sources in topographic measurements with an interferometer. The analysis shed light on the performance requirements of the interferometric hardware, specifically, the receiver chain of the interferometer. A breadboard version of the dual downconverter was built using connectorized components. This version served to verify the performance of such a downconverter. Measurement results showed that the breadboard version performed very well under stable thermal conditions. The downconverter showed a relative phase difference of less than $0.05^\circ$ in a stable thermal environment. This value of phase difference was then set as the benchmark that the second version of the downconverter was expected to meet. This new version was integrated onto a PCB, thus making the system very compact. The integration had problems of its own. Components being packed so close to each other resulted in less than satisfactory isolation between the two channels. The presence of RF shields did not provide adequate isolation between the channels. Nonetheless, the measured phase difference between the two channels of this new downconverter was of the order of $0.065^\circ$ under stable thermal conditions. The PCB version was subject to controlled thermal cycles to simulate potential environments it might be subject to in various configurations. Improvement in isolation between the two channels is expected to result in marginally better phase difference performance. A custom metallic enclosure is currently being fabricated for the downconverter. This enclosure is expected to improve the isolation performance of the downconverter.

An interferometer - TIMMi - was built, with the downconverter developed in this work, as the core. The interferometer was deployed at the Center for Atmospheric Research Experiments (CARE), in Canada on two separate occasions, and once at the University of Massachusetts, Amherst. The results of these deployments showed the importance of a very stable measurement system in order to make accurate interferometric measurements. Any small movement in the antennas could result in significant errors in the measurement. It was also found that the data acquisition system being used - an old HP8753C network analyzer.
- was introducing significant error in the measurement. A study of the network analyzer showed that the phase measured by the network analyzer was varying with time and this was potentially masking any interferometric measurement. Though the deployments did not yield exciting results due to these issues, it gave a picture of the importance of a highly stable measurement system. These results enforce the significance of the need to build and characterize the receiver chain of the interferometer (especially the downconverter), as was done in this thesis.

6.2 Future Work

A new version of the downconverter, operating with a 100 MHz bandwidth, has been designed and is currently being fabricated. The wider bandwidth provided by this system will improve the range resolution of the interferometer from the present 7.5 m to 1.5 m. In addition, a new metallic enclosure has been designed for the downconverter. Careful design of the enclosure to isolate each section of the downconverter in individual cavities is expected to improve channel-to-channel isolation from the present 70 dB to 80 dB. This improvement is also expected to marginally improve the phase stability performance of the downconverter. The enclosure is currently being fabricated and is expected to be delivered soon.

The success of the interferometer will also depend on an updated data acquisition system. Different options for this system are currently being explored. It is expected that addressing concerns of the data system stability would alleviate many errors currently being observed in the interferometric data.
Figure 6.1. Ku-Band RF Channel + Z
Figure 6.2. Ku-Band RF Channel -Z
Figure 6.3. Ku-Band LO Channel
Figure 6.4. L-Band RF Channel + Z
Figure 6.5. L-Band RF Channel -Z
Figure 6.7. Ku-Band LO Test Point
Figure 6.8. L-Band LO Test Point

Figure 6.9. L-Band RF Channel Test Point
Figure 6.10. Power Conditioning Network

Figure 6.11. Analog-to-Digital Converters


