

Radio Astronomy Instrumentation for the AVN



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Abstract

The African VLBI network (AVN) is a proposed network of VLBI capable radio telescopes to be situated in the African countries that will later host antenna elements of the Square Kilometre Array (SKA) radio telescope. This network was proposed to help develop human capital in the African partner countries in preparation for the SKA. Once completed the network will consist of telescopes implemented both by converting decommissioned telecommunication ground stations, and new purpose built radio telescopes. This thesis presents the design and development of a potential cryogenic radio receiver for one of the new purpose built AVN elements. Since the telescope is still under development, I will present my contribution towards this design effort.

The front end of the receiver features a corrugated wide-flare angle feed-horn design compatible with the SKA antenna optics for operation over 4.6 to 8.5 GHz bandwidth and a scaled version to cover 8.3 to 15.3 GHz. The design process is presented along with how the simulated performance compares with potential SKA feeds. Measurements of the prototype's beam pattern are also presented.

A novel 3.3:1 bandwidth, 3-probe triple-ridge transducer (TRT) was designed and prototyped. The simulated and measurement results agree and confirm the broad-band spurious-free characteristics of the TRT. The derived analytic expressions for extracting linear orthogonal polarisations is presented, along with the experimental results.

For receiver gain stabilisation, a new technique based on cross-correlation whilst only using a single receiver chain was explored. Experimental results confirm gain stabilisation using this technique is possible without noticeable degradation in the receiver noise temperature.

A brief description of the cryostat housing the LNAs and TRT is presented. The radio frequency (RF) Analogue signal processing architecture is based on quadrature sampling. This required the design and prototyping of the RF bandpass filter, anti-aliasing lowpass filter and mixer chip circuit board. For the receiver backend, an RF system on chip (RFSoc) FPGA based digital signal processor is proposed. This will provide the reconfigurability needed to switch between single dish and VLBI observations. Finally, the thesis closes with a presentation on the expected radio telescope system temperature and its implications on the sensitivity and mapping speed.

Statement of Originality

I declare that this thesis was written by myself and that the work contained within this thesis is my own except where explicitly stated otherwise. The work in this thesis has not been submitted for any other degree or professional qualification.

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Daliso Banda

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Chapter 1

Introduction

In the early years of radio astronomy observations were limited by poor angular resolution and the sensitivity of the instruments (Ryle 1975). Given that the resolution is proportional to λ/D (where λ is the wavelength at which the observation is being carried out and D is the diameter of the aperture), higher resolution was achieved by increasing the size or aperture of the collecting area to enable the sampling of a larger fraction of the wave front. Constructing a single aperture with resolution comparable to an optical telescope becomes prohibitive for radio wavelengths as a much larger aperture is required to compensate for the longer wavelength. However, consider a situation in which multiple smaller radio telescopes covering the equivalent area that a larger telescope would have covered, are simultaneously used to observe a source. These telescopes probe the wavefront in a similar way in which the single collector would have, albeit more sparsely. With appropriate phasing, each pair in the array can be treated as an interferometer. This is best explained with a diagram such as the one shown in Figure 1.1. If we consider the case of a large single dish telescope, the rays incident on the aperture are considered parallel given the large distance to the source. Each of the incident rays arrive in phase and are all brought to a focus where they interfere constructively. In the case of a two element array telescope also shown in the figure, when observing the same object, the wavefront arrives at antenna 2 before arriving at antenna 1. There is therefore a geometric delay (τ_{geo}) which must be compensated for. This is achieved by introducing an instrument delay

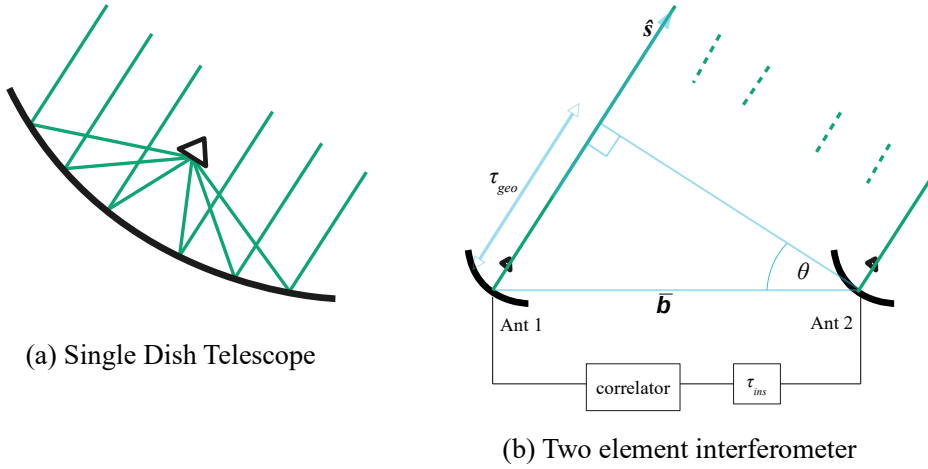


Figure 1.1: Simplified comparison of single-dish telescope with 2 element interferometric telescope. Where $\hat{s}$ is the unit vector in the direction of the source, τ_{geo} is the geometric delay, τ_{ins} is the instrument delay and $\bar{b}$ is the baseline vector of between the two elements.

(τ_{ins}) in the signal path of antenna 2 to insure constructive interference when the two signals are combined electronically. This combining of the signals is analogous to the focusing of the rays by the reflector of the single dish and can be thought of as a virtual focus. In this case the instrument's resolution is proportional to $\lambda/|\bar{b}|$. Provided the magnitude of $\bar{b}$ is comparable to the diameter of the single dish D , then their resolutions will be comparable when observing at the same wavelength. The sensitivity on the other hand will be lower for the array, as can be seen from the figure not all the flux is intercepted by the elements in the telescope array. In fact in order to reconstruct images more effectively a number of elements with various baselines are typically used to improve the sampling of the wavefront. Observations from each pair can be combined in such a way that the final results of the observation would achieve similar resolution as would have been achieved by a single dish (the wavelength of observation and the distance between the two furthest elements in the telescope array would determine the final resolution of the array). This technique, known as aperture synthesis was proposed by Sir Martin Ryle for which he shared the Nobel Prize for Physics in 1974 (Ryle 1975). Many radio telescopes have been constructed over the past decades using this technique, eg the VLA (and its upgrade

i.e the JVLA), ALMA, MeerKAT, ASKAP etc. The resolution limit of a conventional interferometer is limited by the need to bring all the signals over long distances into the correlator while preserving phase coherence.

Very-long-baseline interferometry (VLBI) extends the capabilities of astronomical interferometry by enabling the distance between telescopes to be much greater. Signals are recorded separately at each telescope, timed with an accurate atomic clock. The measured voltages and time data are then transported on storage devices or via high speed network to the correlator. At the correlator, the phase coherence is recovered by searching delay space to recover a detectable signal. Although this is computationally expensive, it allows baseline lengths up to the size of the earth. With VLBI a network of telescopes across the earth can be used for interferometric observations resulting in much greater resolution. For example this has enabled for the first time the direct imaging of the shadow of a black hole by the Event Horizon Telescope (EHT) ([Akiyama et al. 2019](#)). For this observation the EHT consisted of a global VLBI array of eight stations with baseline lengths ranging from 160 m to 10,700 km operating at 230 GHz (1.3mm). The advancement in global connectivity, fibre optic networks now make electronic Very Long Baseline Interferometry (e-VLBI) possible ([Szomoru 2009](#)). Rather than playing back the data and time recordings after the observation at the correlator, the radio telescopes taking part in the e-VLBI session connect to a central data processor for real time data processing. The results are available almost immediately rather than weeks as was the case when disks had to be shipped to the correlator. We are likely to see more of this as more stations come on-line and internet bandwidth of already existing infrastructure continues to improve.

In a bid to achieve greater sensitivity and resolution at centimetre wavelengths the Square Kilometre Array (SKA) was proposed. It is a large radio interferometer currently under development. Once completed it will be the world's largest and most powerful radio telescope ([Pellegrini et al. 2020](#)). The SKA radio telescope is to be co-located in Australia and South Africa. Due to the large scale of the project, it will be constructed in phases. In the first phase of construction (SKA1), SKA1-Low (50

- 350 MHz) will be located in Australia whilst SKA1-Mid (350 MHz - 13.8 GHz) will be located in South Africa. Phase II will bring a significant increase in capabilities and other sites will be constructed in the 8 African partner countries that supported South Africa's bid to host the SKA.

1.1 The African VLBI Network

1.1.1 Introduction

South Africa's Department of Science and Technology along with the South African SKA office (SKA-SA) proposed the deployment of radio telescopes with VLBI capabilities in African partner countries to help prepare them for the arrival of the SKA in Phase II ([Gaylard 2013](#)). This collection of telescopes is called the African VLBI network. One of the ways in which this network could be implemented is by repurposing decommissioned 30 m class dishes which are located in some of the African partner countries. Cost effective conversion of a dish to a radio telescope has been demonstrated ([Yonekura et al. 2016](#)) and is also currently being carried out in one of the partner countries (Ghana). However, not all partner countries have these facilities available and in some instances the condition and location of these dishes would require an investment comparable to building a modern antenna. In such cases a new build is the best option. Therefore, once completed the network will consist of telescopes implemented both by converting decommissioned telecommunication ground stations, and new purpose built radio telescopes ([Gaylard et al. 2014](#)). Rather than committing a full design effort from the ground up, the AVN could take advantage of the mass production of the antenna that has already been prototyped for the SKA. Furthermore, rather than using the SKA receiver package, the AVN could purpose build the receivers at a fraction of the cost thus considerably reduce the total cost of deployment. It is for this reason that this thesis focuses on the design and development of a potential AVN receiver. Figure 1.2 shows the geographical locations of the proposed sites of both the SKA and AVN in South Africa and the partner countries

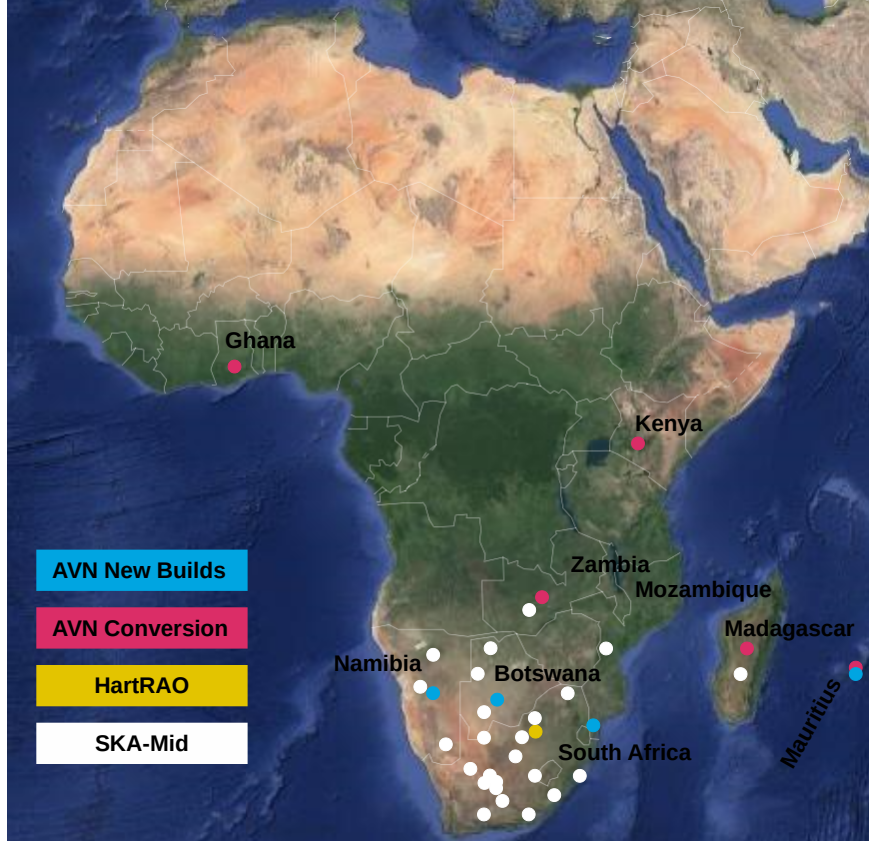


Figure 1.2: Map of Africa showing geographical locations of SKA and AVN sites. The blue dots are potential AVN new build sites. The AVN sites with decommissioned satellite dishes for conversion are the magenta dots. HartRAO's location is the yellow dot and the white dots refer to the location of SKA-Mid sites. Satellite image obtained from Google Maps.

(listed in Appendix A).

The AVN is greatly motivated by both science and potential social and economic benefits, these will be discussed in the next section.

1.1.2 Science case

Currently the AVN consists of two elements, the 26 m radio telescope at Hartbeesthoek observatory (HartRAO) in South Africa and the 32 m Kuntunse dish currently under refurbishment (Asabere et al. 2015). Until the latter and other telescopes come on line, HartRAO will continue to be the only VLBI-capable instrument on the

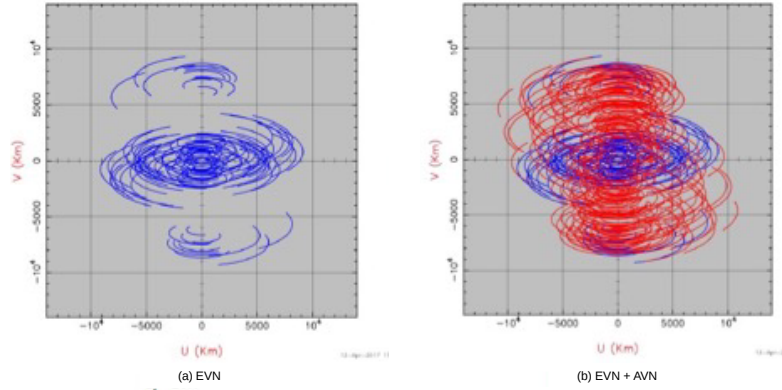


Figure 1.3: Example of EVN and AVN uv -coverage of a source from rise to set at declination $+20^\circ$. (a) shows the uv tracks for the EVN alone. (b) shows the tracks for the EVN alone in blue and those generated as a result of the additional AVN baselines are given in red. Figure taken from [Gurvits et al. \(2020\)](#). Permission to reproduce this diagram has been granted by Leonid Gurvits.

African continent. Whilst HartRAO has long been involved in VLBI observations with the European VLBI Network (EVN) the quality of the images has been limited by the lack of similar telescopes between Europe and South Africa. Thus the addition of the AVN sites will increase the science capabilities of the international VLBI community. This is best illustrated by looking at the uv -coverage of the EVN alone and the uv coverage of the EVN together with the AVN, shown in Figure 1.3. The uv track plots show the sampling of the spatial frequencies that are used to generate an image of the sky brightness. Higher image quality is achieved when there are fewer gaps between tracks. Figure 1.3 (a) shows the uv tracks for the EVN alone and there is a gap between the tracks concentrated in the centre (the European array) and those further away (the European-South Africa baselines). Observing with EVN and AVN results in the uv tracks shown in part (b) of the figure. The red tracks are those generated by the baselines using AVN stations and these baselines fill in the gap, hence improving the image quality as more information is available for image generation.

The specific science cases for the AVN are drawn from the usage of currently existing VLBI networks and by the proposed science case for the SKA ([Gaylard](#)

2013). This includes the study of objects typically observed using VLBI, such as radio-bright sources with small angular size. This criteria is also met by larger objects which are further away. The evolution of active galactic nuclei (AGN) and radio bright supernovae in external galaxies are examples of what is studied using VLBI. Methanol masers at 6.668 GHz and 12.178 GHz can also be observed. As new AVN sites come on-line they can immediately be used with the EVN. Once 4 or more sites of the AVN are completed, the AVN will also be able to carry out VLBI observations independent of the EVN.

Furthermore the sites will also have single dish capabilities enabling students and researchers to carry out research on selected projects in radiometry, spectroscopy and possibly pulsar timing. It is therefore important that the AVN antennas are properly equipped for both stand-alone single dish observations and VLBI observations.

1.1.3 Social and economic case

The second motivation and arguably the most important aspect, is the potential social and economic benefit of AVN. The project hopes to help develop skills, regulations and institutional capacity that will be vital for the partner countries' participation in the SKA and its pathfinder activities. It is envisaged that the programme will result in the transfer of knowledge and skills related to the building, maintenance, operation and use of radio telescopes. This has the potential to foster economic development as the majority of these skills are transferable to other industries that more readily contribute to economic activity (Govender 2009).

1.2 AVN receiver new build

In this thesis a potential receiver system for one of the AVN's new builds is proposed. The scope of the work presented in this thesis includes the developing of the feedhorn and transducer combination, as well as the radio frequency (RF) electronics architecture. In addition a receiver gain stabilisation technique is presented, which would allow the use of the receiver for precision single-dish observations.

1.2.1 Receiver system requirements

To enable the telescope to function both as an element in a VLBI network and as a single dish it needs to meet a number of specifications. The band of operation needs to match that of the target network. In this case the telescope will be part of the AVN, and SKA compatibility would be beneficial. The telescope in Ghana operates in C band at 5 GHz with a bandwidth of 124 MHz and at 6.7 GHz with a bandwidth of 370 MHz. These bands are both covered in SKA band 5A which runs from 4.6 GHz to 8.5 GHz. Therefore by building the telescope with operational bandwidth equivalent to the SKA band 5A will enable the telescope to operate in frequency bands which are compatible with both the AVN and the SKA1 Mid Telescopes. Consequently the telescope for this project will operate between 4.6 GHz to 8.5 GHz with the possibility of extending into SKA Band 5B which runs from 8.3 GHz to 15.3 GHz. To achieve shorter observation time for faint sources, the telescope needs to be as sensitive as possible; this translates into the requirement of a receiver system with a low noise temperature. To achieve this the first gain stages of the RF electronics should be cooled and temperature stabilised. It is vital that both intensity and polarisation are measured to a required accuracy. Therefore a full polarisation receiver front-end design is required with good isolation between polarisations. For radio continuum measurements a wideband multi-channel spectrometer is employed, where as for emission-line spectroscopy a narrowband multi-channel spectrometer is desirable. This versatility is easily achieved using Digital Signal Processing (DSP). Therefore a reconfigurable digital signal processor backend is necessary. Additional requirements and design specifications are listed in Table 1.1.

1.2.2 Receiver overview

Figure 1.4 shows an overview of the receiver system that can meet the specifications; this provides a starting point for the design work. To achieve the desired sensitivity the transducer and Low Noise Amplifier (LNA) are cooled. The remainder of the radio frequency (RF) processing is carried out at room temperature. Final processing is

Table 1.1: A table showing a list of the main telescope components discussed in this thesis. The second column shows the desired specifications of each of these components. The last column highlights the overall telescope requirements that each specification contributes to. The receiver front end in this case refers to the RF electronics from the Horn to the LNA and the other RF components are the filters and splitters.

Device	Specification	Telescope Requirement
Receiver front end	Bandwidth	4.6 to 8.5 GHz
	Noise temp.	< 12 K Sensitivity parameter, $A_{eff}/T_{sys} < 6.1 \text{ m}^2/\text{K}$ across the band.
Feedhorn	Bandwidth Ratio	1.85:1
	Return loss	< -15 dB Gain ripple $\leq 1.5 \text{ dB}_{p-p}$
	Cross-polar	< -15 dB Intrinsic cross-polarisation ratio, (IXR) < -15 dB within 3 dB beamwidth of the main beam of the telescope. To insure -40 dB cross-polarisation is attainable after calibration.
	-15 dB Beamwidth	120° Aperture efficiency $> 80\%$. Making it possible to attain the $6.1 \text{ m}^2/\text{K}$ sensitivity parameter.
Transducer	Bandwidth	4.6 to 8.5 GHz
	Return loss	< -15 dB Gain ripple $\leq 1.5 \text{ dB}_{p-p}$
Other RF components	Bandwidth Ratio	1.85:1
	Return loss	< -15 dB Gain ripple $\leq 1.5 \text{ dB}_{p-p}$

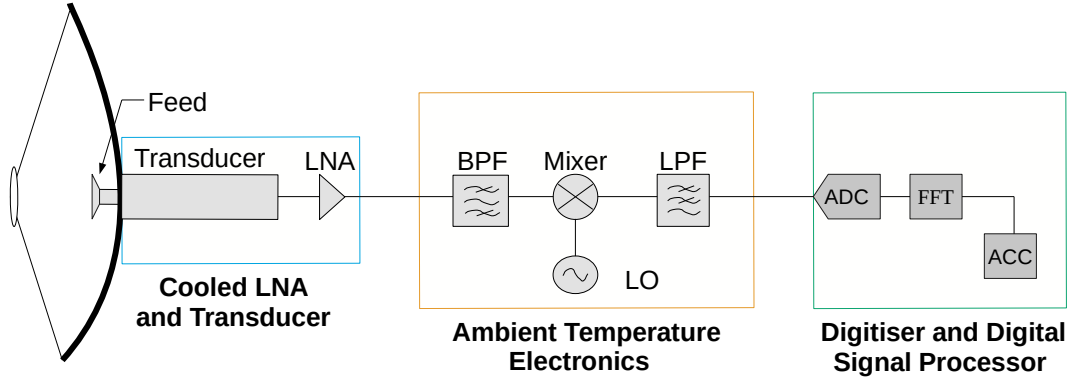


Figure 1.4: Block diagram showing main components of the Receiver to meet the system specifications. Transducer and Low Noise Amplifier (LNA) are kept cool for better sensitivity. RF electronics processing further conditions the received signal to enable digitisation. The filtering is done by the bandpass filter (BPF) and the low-pass filter (LPF). The local oscillator (LO) determines which section of the available bandwidth is being observed. The final section is the digital backend where final processing is carried out using Digital signal processing (DSP). This involves computing a Fast Fourier transform (FFT) followed by accumulation (ACC).

carried out using DSP as this provides the highest performance and reconfigurability.

1.3 Thesis layout

My contributions to the development of the receiver system are outlined in the remainder of this thesis. In Chapter 2 other systems with similar goals are presented and factors that informed the design decisions for the receiver are highlighted. In Chapter 3 the detailed design of the prototype feedhorn is presented. The simulated performance for the feed alone and with the full optical system is also presented. In addition to this the measured return loss and beam patterns for one of the prototype horns is given. The transducer design work is detailed in Chapter 4. A new 3-probe design that enables a wide bandwidth with very low waveguide spurious mode operation is presented. The simulated frequency response is shown followed by the measured performance. The chapter closes with a demonstration of how orthogonal linear polarisations can be extracted from the 3-probe system. In Chapter 5 we

explore the effects of receiver gain instability on system sensitivity. Stabilisation techniques that are currently available are also presented. A new stabilisation technique is proposed and tested via experimental work in the laboratory and the results are given. In Chapter 6 a more detailed system overview is given followed by a more detailed presentation on the signal processing blocks. The chapter closes by presenting an estimate of the overall system performance.

Chapter 2

Radio Telescope Receiver Systems

In this chapter we will discuss the main components of the radio telescope receiver. The aim of the chapter is to give a general overview of the telescope starting from the front end of the telescope, consisting of the antenna system to the backend digital signal processor. Telescope optics are discussed in Section 2.1. The feed design selection criteria are given in section 2.2. The analogue signal processing is then covered in Section 2.3. Lastly, section 2.4 discusses the digital signal processing backend.

2.1 Reflector systems and optics

2.1.1 Reflector systems

A reflector and feed antenna system is one of the most widely employed configurations in radio telescopes. Reflector antennas provide a large collecting area and can be used over a wide range of frequencies, compared to ‘wire’ antennas that are typically tuned to particular frequencies and have effective areas proportional to λ^2 ; where λ is the operating wavelength. The reflector collects the radio signals from the source and focuses them. The feed then couples the received signal into a transmission line. If the received signal is unpolarised then the power delivered to a single-polarised

transmission line P_{TL} by the feed is given by,

$$P_{TL} = \frac{S_v A_{\text{eff}} \Delta v}{2}, \quad (2.1)$$

where S_v is the flux density from the astronomical source incident on the primary reflector, A_{eff} is the effective area of the dish and Δv is the bandwidth over which the observation is being carried out. The effective area is related to the physical area of the primary reflector by,

$$A_{\text{eff}} = \pi \left(\frac{D}{2} \right)^2 \eta_a \quad (2.2)$$

$$= A_{\text{phy}} \eta_a \quad (2.3)$$

Where A_{phy} is the physical area of the primary mirror and η_a is the aperture efficiency. The aperture efficiency is the measure of how effective the dish is at collecting the incident power. It accounts for the feed illumination of the reflector, the reflection losses, blockage (caused by the feed and its support structure) and the accuracy of the reflectors surface profile.

Apart from enabling the detection of faint sources by increasing the area over which the flux of a source maybe collected, dishes also provide directivity. This refers to the ability to provide higher sensitivity in a specific direction depending on where the dish is pointing. This allows for differentiation of flux from sources at different positions in the sky. Mathematically, directivity, $D(\theta, \phi)$ is derived from radiation intensity, $U(\theta, \phi)$ (defined as the radiated power from an antenna per unit solid angle). Given the reciprocity of antennas this applies to receiving as well. Directivity is given by,

$$D(\theta, \phi) = \frac{U(\theta, \phi)}{\frac{1}{4\pi} \int_0^{2\pi} \int_0^\pi U(\theta, \phi) \sin \theta d\theta d\phi} \quad (2.4)$$

$$(2.5)$$

$$= \frac{U(\theta, \phi)}{U_0}, \quad (2.6)$$

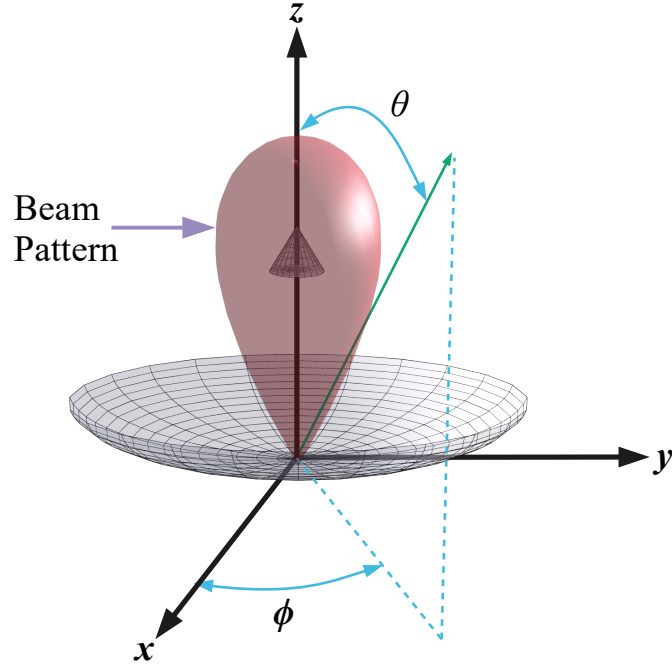


Figure 2.1: Coordinate system for prime focus analysis.

where θ and ϕ are the usual spherical polar co-ordinates, defined in Figure 2.1 and U_0 is the average power per unit solid angle. The directivity is also referred to as the beam pattern of the antenna. In Figure 2.1 an idealised beam pattern is also shown. This pattern determines the resolution of the telescope, which is defined as twice the angle at which the directivity drops by half. This is known as the full width at half maximum (FWHM) angle of the telescope's main beam. "Directivity" can also be used to describe the peak value of $D(\theta, \phi)$ in the forward direction. One other important antenna parameter is the antenna's power gain; it is an indication of how effective the antenna system is at converting incident radio waves to electrical power. The boresight forward gain G is given by

$$G = \frac{4\pi A_{eff}}{\lambda^2}. \quad (2.7)$$

The gain is related to the directivity through

$$G(\theta, \phi) = \eta_e D(\theta, \phi), \quad (2.8)$$

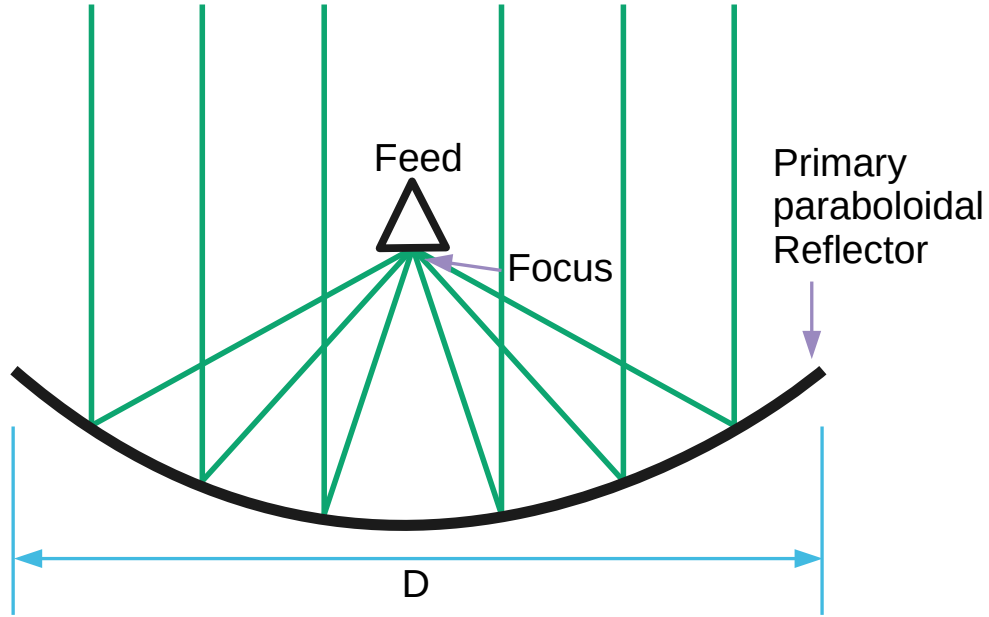


Figure 2.2: Geometry of a symmetric (centre-fed) prime focus dish.

where the electrical efficiency η_e accounts for electrical losses such as ohmic losses (Balanis 1997).

2.1.2 Reflector antenna types

If the feed is placed at the focal plane of the primary reflector (also known as a mirror) it is referred to as a prime focus dish as shown in Figure 2.2. This configuration is the most mechanically simple to manufacture. The dish usually consists of a number of identical panels and hence less moulds are needed. One of the disadvantages of the prime focus is that for large reflectors the feed is not easily accessible when the dish is pointing at zenith. By using a dual reflector system the feed can be relocated to a convenient position. Other benefits of dual reflector systems is the reduction of spillover and side-lobes (see e.g. Balanis 1997) and also includes being able to place the feed at a more mechanically optimal position, enabling the use of high gain feeds. By shaping the subreflector a more uniform illumination of the main reflector can be achieved (see e.g. Olver et al. 1994) thereby increasing aperture efficiency. Figure 2.3 shows the most common dual reflector configurations, Cassegrain and Gregorian. The

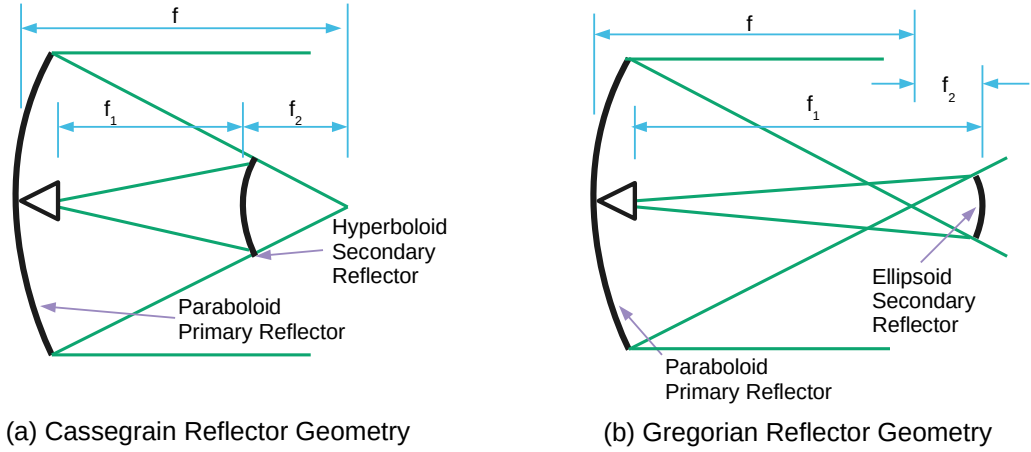


Figure 2.3: Cassegrain and Gregorian Dual-reflector System Geometry. (a) shows the geometry for the Cassegrain system which employs a paraboloidal primary reflector and a Hyperbolic reflector for the secondary reflector. (b) shows the geometry for the Gregorian system which utilises an ellipse for the secondary reflector (Olver et al. 1994).

Cassegrain employs a paraboloidal secondary reflector, whereas for the Gregorian it is an ellipsoid. In this case one of the foci of the ellipse is placed to coincide with that of the primary reflector whilst the phase centre of the feed is placed at the other focus.

One drawback that is common to both the prime focus and the dual reflector systems discussed so far is the blockage of the aperture by the feed or secondary reflector as well as the support structure that holds these in position. This translates to lower aperture efficiency. The supports also contribute to scattering away from the beam axis, which tends to increase the far-out sidelobes, resulting in higher system temperature. In situations where the receiver noise temperature contributes significantly more to system temperature than the blockage and scattering effects, these effects can be ignored. On the other hand, for higher performance systems with much lower receiver noise temperatures, the effects of blockage and scattering have a more significant effect on the overall system temperature and hence can not be ignored. To overcome these drawbacks dual-reflector offset-fed optics can be used, such as the ones shown in Figure 2.4. The main reflector is only a portion of a paraboloid. The feed is pointed at the centre of the subreflector to reduce spillover and achieve uniform

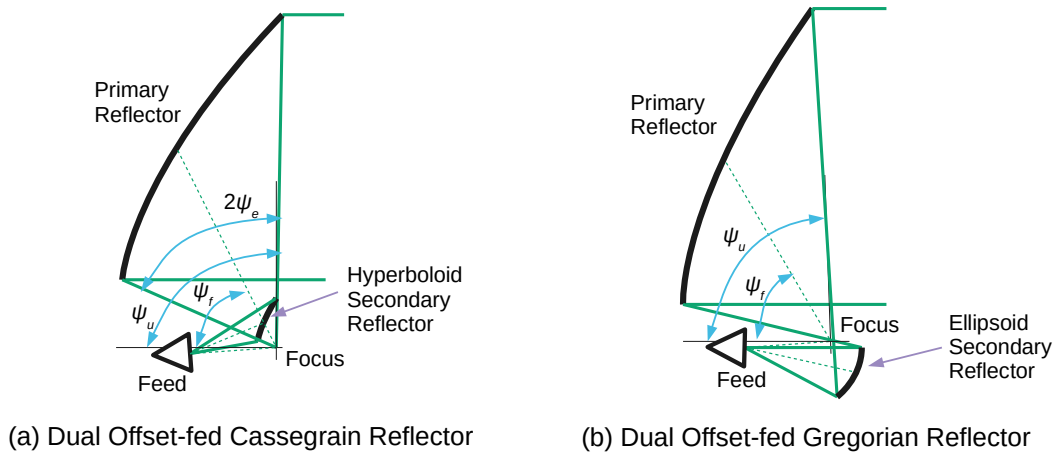


Figure 2.4: Offset-fed Cassegrain and Gregorian Dual-reflector System Geometry. The main and sub reflectors are sections of a paraboloid and hyperboloid/ellipsoids respectively. (a) shows the cross-section geometry for the Cassegrain system, cross section of the primary reflector is a portion of a paraboloid whilst the secondary is a portion of a Hyperboloid. (b) shows the geometry for the Gregorian system which uses a portion of an ellipsoid for the secondary reflector ([Milligan 2005](#)).

illumination of the aperture.

Another important antenna parameter worth mentioning is the focal-length to diameter (f/D) ratio. This is the ratio of the effective focal length of the reflector to the diameter of the reflector. Higher f/D ratios require feeds with narrower beamwidth (high-gain feeds) whilst lower f/D ratios require low-gain feeds, which have wider beamwidth.

Antenna types previously used in radio astronomy have mostly been prime focus or Cassegrain. In order to achieve low-cost, low-power and reduce the man-power required to build the antennas, the Australian Square Kilometre Array Pathfinder (ASKAP) uses prime focus antenna elements with a 0.5 f/D ([Chippendale & Schinckel 2011](#)). For the Very Large Array (VLA) a Cassegrain variant is employed. To maximise sensitivity, instead of the traditional paraboloid and hyperboloid surfaces, the surfaces were mathematically perturbed to achieve nearly uniform illumination over the entire aperture (apart from areas of the aperture affected by the blockage arising from the secondary mirror and its support structure). In addition the feeds are placed off-axis to allow for multiple feeds to be used ([Hunter & Napier 2016](#)). The latter

comes at the cost of poorer polarisation performance whilst the shaping of the surface for uniform illumination results in increased side-lobe levels. For the Allen Telescope Array (ATA) offset Gregorian antenna elements were selected in order to eliminate blockage (Welch et al. 2009). It is for this reason also that the SKA specifies an offset Gregorian dual-reflector dish system for the SKA-mid, more details of which are provided in the next section.

2.1.3 SKA dish description

Given that considerable effort has been put into developing the SKA-mid dishes, the most likely outcome is that the AVN new build receptors will be based on the SKA dish. This is cost effective and will make it easier for these stations to be integrated into the SKA. It is for this reason that we have designed the feed for the AVN receiver based on the optics of the SKA dish. We will spend the rest of this section reviewing the SKA dish whilst the feed design will be covered in greater detail in Chapter 3.

The receptor consists of a primary and secondary reflector in an offset Gregorian configuration as shown in Figure 2.5. This image is a picture of the first SKA dish prototype. A summary of the specifications are given in Table 2.1. As mentioned in the previous section this configuration offers an aperture area free of blockage. This translates into higher aperture efficiency which is a design requirement stated in the baseline SKA design documentation (Dewdney et al. 2015). Furthermore the system is less susceptible to ground-based and satellite radio frequency interference (RFI) as strut scattering which results in increased far-out sidelobe levels is not present (Jonas et al. 2018). Feed-down tipping along with extending the bottom edge of the secondary reflector further reduces the noise contribution from the ground and improves illumination efficiency. To further improve the sensitivity the reflector system was shaped and optimisation was carried out so that the best possible illumination efficiency is achieved over the SKA1-Mid frequency bands (Theron et al. 2013). Finally it is desirable to have a compact antenna system. To achieve this the horn and secondary are placed closer to the primary reflector resulting in low focal length to dish diameter (f/D) ratio. In order to effectively illuminate the secondary reflector



Figure 2.5: A picture of the first assembled SKA prototype Dish. Picture taken from the SKA organisation.

the feed needs to have a broad beamwidth, this necessitates the use of a low-gain feed (wide flare angle horn). The fact that the feed is not in the path of the aperture also allows for a number of horns to be located on the receiver indexer and swung into focus when needed.

Table 2.1: Summary of SKA-mid dish specifications.

Optics	offset-Gregorian
Frequency	0.35 to 20 GHz
Effective f/D	0.5
Elevation range	15° to 90°
Primary size	15 m by 18 m
Projected diameter	15 m
Secondary size	5 m by 5 m
Secondary illumination half-angle	58°

2.2 Feeds and transducers

With dual mirror systems the secondary reflector focuses the received electromagnetic wave to a focal plane. The feed's phase centre is placed at this focus. The feed provides the necessary conditions to transition the signal from free space to propagation in a waveguide structure. The wave is then converted into an electrical current by a transducer. It is vital that this is done as efficiently as possible across the entire band of operation. This is achieved by selecting the appropriate feed for the optics. We will now provide an overview of the two main types of feeds which have been widely used in radio telescopes, starting with the high-gain feed.

2.2.1 High-gain feeds

Even though the optics under consideration are not compatible with this kind of feed, this discussion is included for completeness. For Cassegrain optics high-gain feeds are commonly used. These have a narrow flare angle as depicted in Figure 2.6 (a). The semi-flare angle, θ_{semi} , of these feeds tend to be less than 60 degrees. Corrugated conical high-gain feed horns are almost universal due to their good circular beam pattern, lower sidelobes and good cross-polarisation performance. These have replaced the conical and pyramidal horns that were once extensively in use. They tend to provide good performance over a 2:1 bandwidth, but can also be designed to operate up to 3:1 especially for narrow flare angles.

2.2.2 Low-gain Feeds

Low-gain feeds are best suited for prime focus and Gregorian optics. The first version of this type of feed, which has also seen considerable use in ground based satellite communication terminals, consists of an open waveguide with a choked flange as shown in Figure 2.6 (b). The chokes suppress the sidelobes by reducing the surface currents on the outer edges of the feed. This type of feed is used on the SKA-mid precursor MeerKAT (Tan et al. 2016). The second type is also a corrugated conical horn like the high-gain feed but the flare angle is much broader; with θ_{semi} in excess

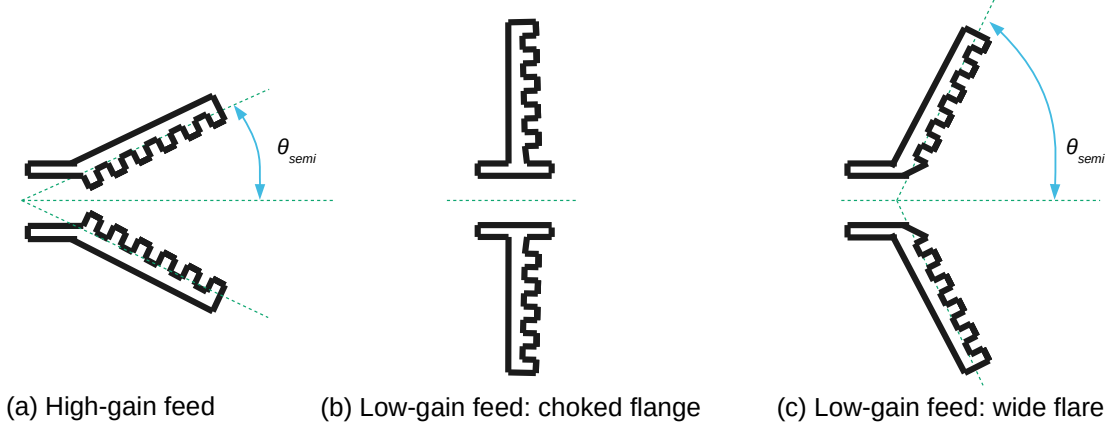


Figure 2.6: Cross-sections of basic feed Horn structures.

of 60 degrees. The simplest version of this is shown in Figure 2.6 (c). A third version of this type of feed we will be considering is the Quad-ridge flared horn (QRFH). These horns are made by flaring out one end of a quad-ridge waveguide structure to provide better transition from guided-wave to free-space propagation. This feed is usually chosen for its high bandwidth capabilities. Whilst in the past telescopes tended to be fairly narrow band, the trend has been to increase the bandwidth over which a given instrument can operate. Higher bandwidth has a number of benefits. The sensitivity to continuum radiation is proportional to $\sqrt{\Delta\nu}$ and mapping speed is proportional to $\Delta\nu$, therefore a larger bandwidth will translate into being able to detect fainter sources and greater mapping speed. For line emission, more bandwidth means more lines can be measured. In general for full frequency coverage over a large range (for example the SKA covers 0.35 to 15 GHz) the greater the bandwidth of each individual feed, the fewer the number of feeds required and hence the cheaper and simpler the entire system may be to implement. This is especially important for telescope arrays composed of a large number of elements. The reflector system can be designed and optimised for broadband operation, for example the SKA-mid optics are designed to cover the full 43:1 bandwidth. On the other hand designing a single feed and transducer that can operate with consistently good performance across this entire band is challenging if not impossible. Quad-ridge flared horns (QRFH) provide

relatively broadband operation which is why they were initially the first choice for a number of the SKA sub-bands. A 6:1 bandwidth has been reported for a QRFH ([Akgiray et al. 2012](#)). This implementation achieved a return loss of -10 dB and average peak cross-polarisation ratio of -10 dB. It is evident from these results that performance across such a large band is poorer than feeds of bandwidth 2:1 or less; the return loss, cross-polar, illumination efficiency and spillover tend to suffer. One other advantage of the QRFH is that the ridges in the feed can be extended into the OMT structure where the separation between opposing ridges can be reduced to match a 50 Ohm launch pin, thus creating a semi-continuous structure consisting of the Orthomode transducer (OMT) and feedhorn. Given that the whole SKA-mid band can not be covered using a single feed and transducer it is divided into 6 bands: 350 to 1050 MHz (Band 1), 950 to 1760 MHz (Band 2), 1650 to 3050 MHz (Band 3), 2.8 to 5.18 GHz (Band 4), 4.6 to 8.5 GHz (Band 5a) and 8.3 to 15.3 GHz (Band 5b). This entails separate feeds and transducer for each band. Band 1 will be covered using a 3:1 bandwidth QRFH ([Tan et al. 2016](#)), whilst Band 2 will adopt the axially corrugated/choked flange horn and corresponding compact OMT used on the MeerKAT ([Tan et al. 2016](#)). Initially Band 5 was to run from 4.6 to 13.8 GHz using a single QRFH. After comparing expected sensitivity and cross-polarisation for a scaled version of the Band 2 feed and two other potential QRFH feeds it was proposed that splitting the band into 2 sub-bands and using choked flange feeds for each sub-band would yield the best performance. The simulations revealed that neither QRFH feed could meet the sensitivity specifications whilst the choked flange horns do with reasonable headroom. Furthermore, even though the QRFH feeds meet the cross-polarisation specification, this performance is 20 dB below that of the choked flange feed and worse than the typical current-generation radio telescope ([Leech et al. 2016](#)). This is due to the inherent nature of the QRFH structure. It is for these reasons that we also chose to forego the QRFH and explore the possibility of using one of the other low-gain feed structures for the receiver described in this thesis. Details on the feed design process are covered in the Chapter [3](#).

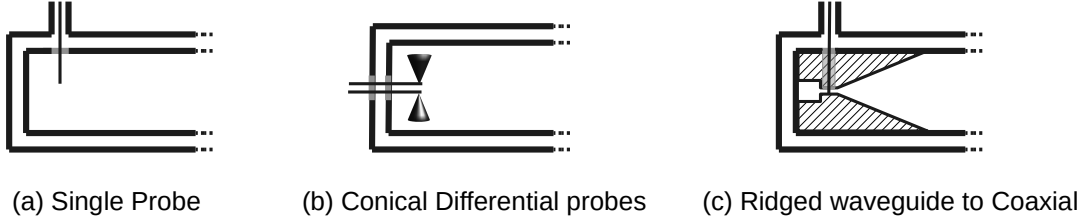


Figure 2.7: Basic single polarisation transducer cross-sections.

2.2.3 Probe type transducer

The simplest transducer is constructed by inserting a probe into a circular or rectangular waveguide at an appropriate location as shown in Figure 2.7 (a). These tend to provide a narrow bandwidth, although flaring the probe can increase the bandwidth. A good example of this is the orthomode transducer (OMT) that will be used for SKA-mid Band 2. It is based on conical orthogonal dipoles, and with this setup a 2:1 bandwidth is achieved (Lehmensiek & Theron 2012). The cross-section for a single polarisation is illustrated in Figure 2.7 (b).

2.2.4 Waveguide transducers

For the case of the probe transducers, separation of orthogonal polarisations is achieved by using two sets of orthogonal probes, whereas for a waveguide transducer it is accomplished using waveguide components. There are various types, a good example being the turnstile OMT. In this OMT a turnstile junction (illustrated in Figure 2.8) is used to carry out the polarisation filtering. The horn feeds into one of the five ports of the device, which then separates the two polarisations. The total power in each polarisation is split between two of the remaining four ports. A Y-junction can then be used to combine the signal in the appropriate ports to enable the extraction of the two polarisations independently. This type of OMT has been shown to offer good return loss (less than -19 dB), with low insertion loss (≈ 0.15 dB) and great cross-polarisation performance (less than -48 dB) (Navarrini & Plambeck 2006).

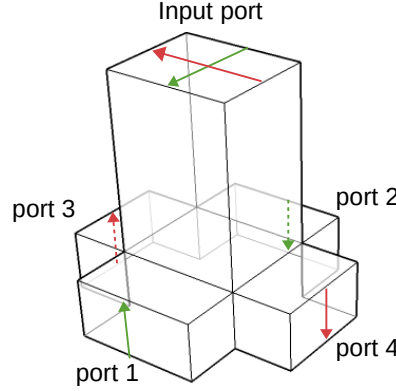


Figure 2.8: Illustration of a turnstile junction. The green and red arrows depict the orientation of the excited orthogonal electric fields. The orthogonal electromagnetic waves incident on the input port are separated and extracted using the other 4 ports of the device.

2.2.5 Ridged-waveguide transducers

The most basic form of this type of transducer, for a single polarisation, consists of a circular or rectangular waveguide with opposing ridges protruding through the waveguide as illustrated in Figure 2.7 (c). A coaxial probe is inserted through one of the ridges, with the centre conductor of the probe terminating on the opposite ridge. This structure can be designed to achieve broadband matching hence providing a wide bandwidth of operation. By using a quad-ridge structure; with two sets of ridges placed in quadrature, two orthogonal polarisations can be received. This structure is known as a quad-ridge orthomode transducer. It has been used in various radio telescopes. For example, in the L-band receiver of the eVLA (Coutts 2011), isolation and return loss of better than -30 dB and -15 dB respectively were achieved across an octave bandwidth in this implementation.

Even though the other transducers discussed in the preceding sections would satisfy our design requirements, we explored the use of ridged-waveguide based transducers as these have been demonstrated to provide higher bandwidths. Broader bandwidth capabilities of the transducer would enable the frequency range of operation to be extended by simply swapping out the horn rather than the entire transducer and RF electronics. Details on the design of the transducer are provided in Chapter 4.

2.3 Analogue signal processing

Traditionally the analogue RF signal processing in a radio telescope has included five main operations. The flow chart in Figure 2.9 shows these operations. The first is to amplify the received signal. This is necessary because the signal at the output of the transducer is still too low to be processed by most electronic detectors. Depending on the sensitivity of the detector, a total gain in excess of 80 dB is not uncommon. Even though this gain is typically distributed along the receiver front end, it is vital that the first stage of amplification contributes the least amount of noise to the incident signal. The noise performance of a receiver is typically modelled using noise temperature. To calculate the expected receiver noise temperature T_{RX} , Friis' formula for noise temperature, (Equation 2.9) is used. T_n and G_n correspond to the noise temperature and gain contributed by each individual component respectively. Where the subscript n indicates the position of the component in the receiver chain. The components that follow the first stage have less effect on the overall noise temperature of the receiver. From this we see the significance of minimising the noise contributed by the first stages of the receiver.

$$T_{RX} = T_1 + \frac{T_2}{G_1} + \frac{T_3}{G_1 G_2} + \frac{T_4}{G_1 G_2 G_3} + \dots + \frac{T_n}{G_1 G_2 G_3 \dots G_{n-1}} \quad (2.9)$$

To achieve this low noise amplifiers are used. These amplifiers are designed to minimise the noise contribution of the device whilst providing adequate gain such that the noise added by components that follow have minimal effect on the signal-to-noise ratio. After the initial stage of amplification, the other stages of amplification ensure that the signal is raised to the required level for final processing. The second operation is selecting the desired band for the observation; this is achieved by filtering. This bandpass filtering usually takes place at radio frequencies. The filters can be implemented using waveguides, discrete components or lumped elements depending on the frequency of operation. For our frequencies of interest, greater than 4 GHz, suitable RF filters can be implemented using PCB microstrip techniques. Lower frequency filters, less than 2 GHz, can be implemented using discrete components. More

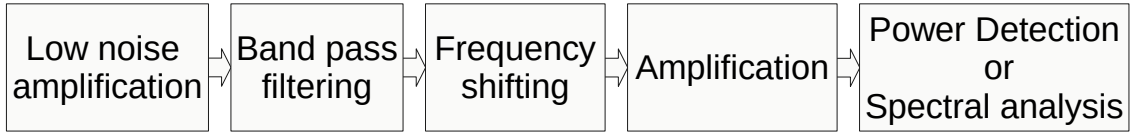


Figure 2.9: Block diagram of general operations performed by radio telescope receiver systems.

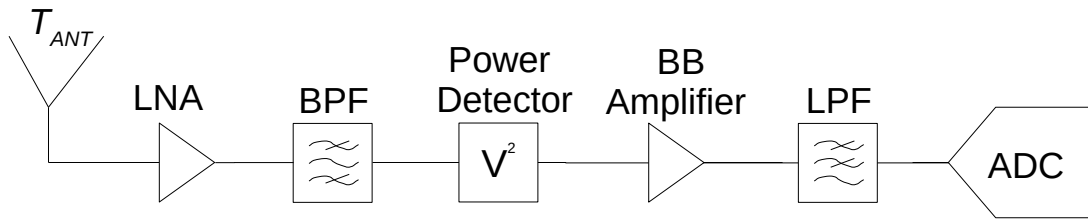


Figure 2.10: Block diagram of a total power radiometer. A local oscillator is not needed as the power detector squares the RF signal, the base band (BB) amplifier's output DC voltage is proportional to the incident power. After low-pass filtering to avoid aliasing during sampling, the signal is digitised using an ADC to allow for electronic storage.

details on both of these are provided in Chapter 6.

The third operation is to shift the selected RF passband to baseband by multiplying the RF signal with a continuous wave (CW) signal generated by the local oscillator or LO. The multiplication is accomplished using a mixer.

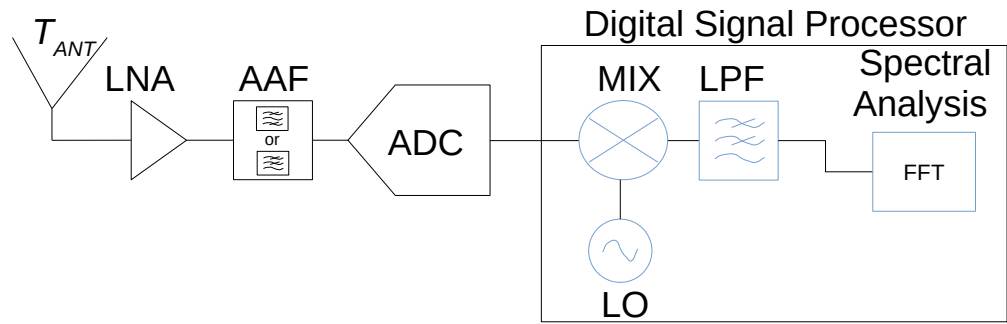
After further signal amplification, the last operation is detection. This involves extracting information from the signal's voltage that can be used to determine the state of the object under observation. This can simply be computing the total power received from the object using a diode detector or can involve the implementation of a spectrometer.

How these steps are realised in hardware is generally a function of what the science goals are, and the availability of technology and funding. The simplest receiver that can be implemented is a total power radiometer shown in Figure 2.10. With a broadband filter this receiver allows for continuum observations to be carried out. The filter selects the bandwidth of interest and the diode detector squares the input voltage signal over the entire bandwidth, giving a voltage output that is proportional

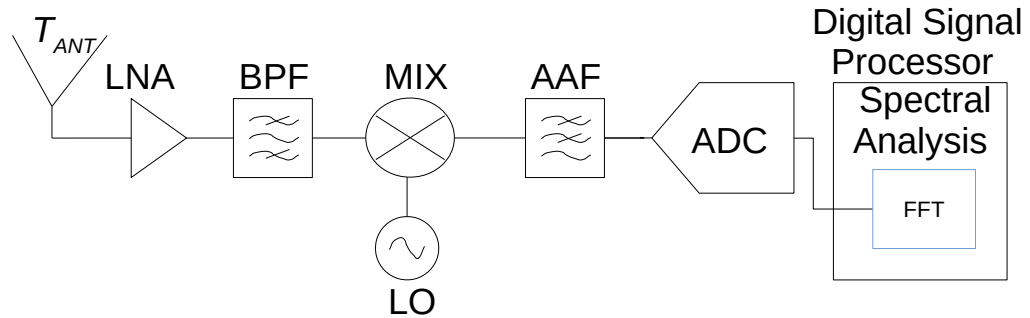
to the total power in that band. By calibrating and using the Raleigh-Jeans approximation, $P \approx kT$ we can determine the temperature of the source as a black body emitter. This architecture can be extended to implement a spectrometer by using multiple narrowband filters, each tuned to cover a sub-band of the total system bandwidth. This is how spectral line observations were originally carried out.

In modern telescopes the diode detector has been replaced with an Analogue to Digital Converter (ADC). The trend has been to push the ADCs as close to the antenna as possible in order to replace as many analogue operations with their Digital Signal Processing (DSP) counterparts. Figure 2.11 (a) illustrates the latter with a block diagram. The ADC is placed immediately after the anti-aliasing filter (AAF) which can be a bandpass or lowpass filter depending on the Nyquist zone in which the RF signal is located. The remainder of the processing such as the mixing, filtering and detection are then carried out using DSP. Digital techniques enable more cost-effective implementation of sophisticated signal processing functions. Spectrometers can be implemented with higher spectral resolution, with considerably more channels than was possible in analogue. Some signal processing techniques are only possible with digital techniques and last but not least digital systems provide more reliability and are less susceptible to component ageing resulting from environmental conditions (see e.g. [Manolakis & Ingle 2011](#)). However this is only possible in situations where the ADC has sufficiently high sampling rate and input bandwidth to directly sample the RF signal determined by the AAF. In situations where it is not possible to use the appropriate ADC because of power consumption constraints or non-availability, a mixing stage precedes the ADC. In this case mixing is implemented using analogue components, this shifts the band of interest to a suitable frequency for the available ADC as shown in Figure 2.11 (b).

The SKA will use a 30 GSps ADC implemented by interleaving two 15 GSps ADCs. This would have allowed for direct digitisation of the desired band for our receiver but it is currently still under development. Therefore the receiver architecture selected will include a mixing stage before a 4 GSps ADC as depicted in Figure 2.11 (b). More details on the architecture are provided in Chapter 6. In the next section



(a) Predominantly Digital Signal Processing Receiver



(b) Predominantly Analogue Signal Processing Receiver

Figure 2.11: Block diagrams of two modern radio telescope receiver architectures. (a) The ADC is placed as close to the antenna as possible, after the AAF (anti-aliasing filter). It has sufficient sampling rate to digitise the desired band. Mixing, filtering and detection are carried out in the digital signal processor. (b) The final detection is carried out using DSP, whilst frequency shifting is carried out in the analogue domain.

we will review digital signal processor hardware.

2.4 Digital signal processing

Initially digital signal processing hardware along with the analogue processing section were both implemented using custom discrete components mounted on printed circuit board (PCB). The discrete DSP components were later replaced with an application specific integrated circuit (ASIC). ASICs were used for DSP up until the early 2000s (e.g. Atacama Large Millimeter Array (ALMA) correlator ([Escoffier et al. 2007](#))), more recently FPGAs have become the better technology due to reconfigurability and lower cost compared to ASICs. Typical FPGA DSP hardware takes the form illustrated in Figure 2.12 (a). FPGAs enable real-time processing capabilities possible with ASICs (with less throughput but at greatly reduced development cost) whilst also providing reconfigurability offered by PC based processing. This is made possible by the integrated circuit architecture of FPGAs. These integrated circuits (ICs) consist of configurable logic blocks (CLB) and random access memory (RAM) blocks embedded in a "sea" of programmable interconnects as shown in Figure 2.13. Depending on the manufacture, the CLB is implemented using look up tables (LUT) or multiplexers (MUX). These can be programmed to implement any Boolean operation with the interconnects providing flexible routing. For the FPGAs intended for DSP implementation, DSP blocks are also included in the FPGA fabric. These blocks include hardware multipliers, adders and accumulators (operations that are common in DSP algorithms) allowing for higher throughput than would otherwise be possible using a purely CLB implementation.

Newer instruments can benefit more readily as software can be ported from older instruments with greater ease. This has especially been made easier by the the Collaboration for Astronomy Signal Processing and Electronics Research (CASPER) ([Werthimer 2011](#)). The collaboration has developed open source, scalable, modular, upgradeable DSP hardware and non-platform specific software libraries that enable an instrument's digital backend to be developed quickly and cost effectively.

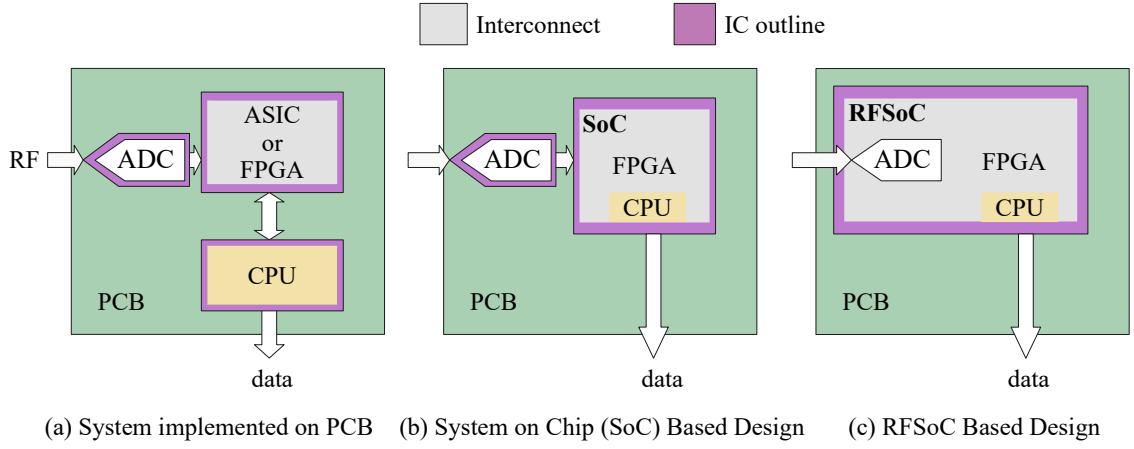


Figure 2.12: Block diagram illustrating the progression of DSP hardware implementation. (a) Early systems were implemented with each individual component housed in a separate IC package. The system was thus assembled on a printed circuit board (PCB). (b) System-on-chip based DSP backends utilise system-on-chip (SoC) ICs. These integrated circuits have other function blocks implemented on the same silicon as the FPGA. The integrated blocks include devices such as transceivers and CPU cores. (c) The RFSoc takes the SoC a step further by integrating ADCs capable of sampling the RF signal.

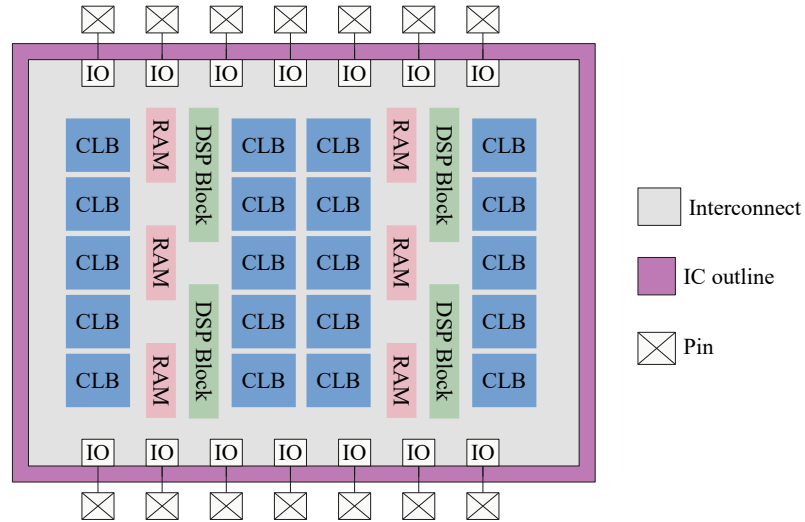


Figure 2.13: Block diagram of general FPGA architecture. The configurable logic blocks (CLB) allow for different logic operations to be programmed by the user. The interconnects connect the different logic blocks and DSP blocks and RAM according to the implemented firmware.

The first software based instruments were implemented using personal computers ([Hotan 2008](#)). These relied on transferring the digitised signals into the memory of the PC where processing was then carried out using high level programming. For higher, real-time bandwidth processing this approach was not practical. However with the advancements in general-purpose computing hardware, software based telescopes are making a come-back as it is now possible to implement more capable software-based instruments. A good example of this is the LOw Frequency ARay (LOFAR) which is frequently described as one of the first in the new generation of software telescopes ([Broekema et al. 2018](#)). LOFAR’s correlator and beamformer are based on high performance multi-core Graphics Processing Units (GPUs) which allow for parallel computing capabilities needed in DSP for astronomy.

Some systems employ hybrid solutions, for example FPGA-GPU hybrid systems have been implemented, as well as FPGA-ASIC hybrid systems in the case of ALMA ([Escoffier et al. 2007](#)).

To improve performance and reduce costs, the IC industry has introduced system-on-chip (SoC) FPGAs. In these devices, CPUs, transceivers and other commonly used peripherals are implemented on the same silicon as a traditional FPGA, as shown in Figure 2.12 (b) (this has enabled data to be read out from the digital backend over standard gigabyte Ethernet). A number of instruments have capitalised on the benefits of FPGA based digital processors (see e.g. [Hickish et al. 2016](#)). For example the AVN conversion in Ghana uses a ROACH (Reconfigurable Open Architecture Computing Hardware) board ([Copley et al. 2016](#)) developed by CASPER ([Hickish et al. 2016](#)). The ROACH is used to implement a 0.39 MHz resolution spectrometer with 400 MHz bandwidth. For the MeerKAT array the Square Kilometre Array Reconfigurable Application Board (SKARAB) is used. In this case the SKARABs are used in the processing nodes to carry out filtering, correlation and beamforming ([Jonas et al. 2018](#)).

Until recently the high speed ADCs have been discrete components mounted on the board and interfaced with the FPGA using a commensurate high-speed transceiver or the general purpose IO pins on the FPGA. This complicates the PCB layout and the

transceivers result in higher power consumption ([Collins 2017](#)). With the telecommunication and instrumentation industry in mind, to overcome these challenges, Xilinx extended the capabilities of SoC FPGAs by integrating RF sampling ADCs and DACs on the same silicon as the FPGA creating an RFSoc ([Xilinx 2019](#)). The resulting architecture is illustrated in Figure 2.12 (c).

For the receiver implementation in this thesis we chose to go with an RFSoc based digital backend whose firmware is currently under development in our group. More details on this hardware are provided in Chapter 6.

2.5 Summary

In this chapter the context of the design work was provided. The first section gave the background of the antenna system and why it was selected for this project. The second section reviewed the main two feed horn categories and the transducer types. The reasons for selecting a low-gain (wide flare angle corrugated) feed horn and ridged-waveguide transducer were highlighted. This was then followed by a discussion on the necessary analogue signal processing. In this case we will be adopting a homodyne receiver system followed by analogue-to-digital conversion. In the final section of this chapter the reasons for using DSP based backend were provided followed by an introduction to the RFSoc-based backend that will be employed in this project.

Chapter 3

Corrugated Wide-Flare Feedhorn

Electromagnetic waves incident on the primary reflector are directed to the secondary reflector which in turn redirects the waves into the feed horn. The horn couples the electromagnetic waves from free space to the waveguide and feeds them through to the transducer. In order for the instrument to provide the best possible performance the feed horn needs to be matched to the optics of the radio telescope's antenna system. In our case it is the SKA-Mid antenna system. As already stated, one of the characteristics of these optics is the low f/D ratio. This feature is a very important factor in determining what type of horn we can use. A low f/D ratio, as a result of the wideband requirement means that to effectively illuminate the secondary reflector the feed needs to have a wide flare angle.

Using a smooth-walled, wide flare-angled horn would allow for effective illumination of the secondary reflector, but the system sensitivity would suffer due to higher side lobes resulting in increased ground pick up, especially at lower elevations. Full polarisation measurements would also perform inadequately as this type of horn has poor cross-polarisation performance. To address this issue the profile of the horn could be designed in such a way that beam symmetry and hence sidelobe performance is improved as is shown by [Potter \(1961\)](#). Whilst these techniques have been shown to provide very good performance for narrow flare angle horns, it is not clear if this performance is transferable to wide flare horns, but even if it is possible the performance is not inherently broadband, which is not suitable for our system.

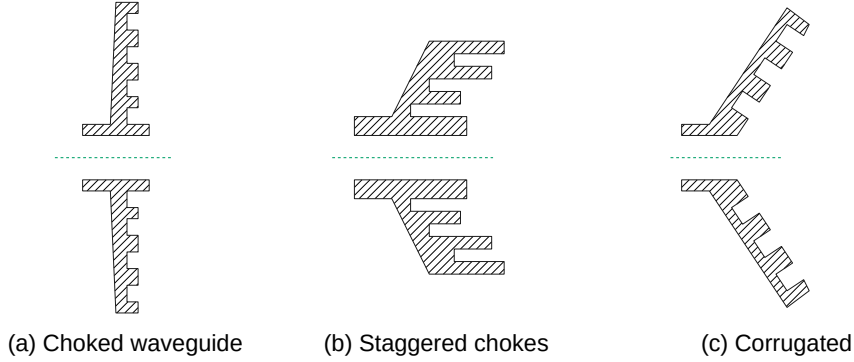


Figure 3.1: Wide-beamwidth Feed structure cross-sections (a) Standard choked waveguide. (b) Staggered chokes. (c) Corrugated wide-flare angle.

To overcome these challenges whilst providing the required illumination pattern, there are two possible options, the choked waveguide (Figure 3.1 (a) and (b)) and the wide flare angle corrugated horn (Figure 3.1 (c)). Like the narrow flare-angle corrugated horn, the wide flare angle version can be designed to provide a circular beam pattern with low cross-polarisation (Clarricoats & Olver 1984, Olver et al. 1994). These horns rely on generating a transverse magnetic (TM) field propagation mode such that the combination of this mode and the dominant mode (transverse electric, TE) results in a symmetric and smooth field distribution at the aperture of the feed, resulting in the desired beam pattern. In the narrow flare-angle corrugated feeds, the conversion of a fraction of the TE waves to TM waves is achieved using corrugations which are $\lambda/2$ deep. The depth of these corrugations reduces to $\lambda/4$ away from the horns throat, towards the horn's aperture. However, in the case of the wide-flare angle horn, the conversion is done over the smooth section at the throat. In both designs the $\lambda/4$ corrugations provide the appropriate surface impedance for the hybrid mode to then propagate to the aperture.

Even though the wide-flare-angle feed provides excellent performance in terms of low cross-polarisation and symmetrical E- and H-plane patterns, the geometry was considered to be complex and difficult to manufacture. This motivated the search for other feed designs with simpler structures. This led to the development of the choked

waveguide feed (Olver et al. 1994). The MeerKAT array makes use of this type of feed; it has a structure with staggered chokes similar to the illustration in Figure 3.1 (b) (Lehmensiek & Theron 2012). The operation of these feeds is fundamentally different from the corrugated feed horns. The waveguide radiates (for convenience considering it as a transmitter) the electromagnetic wave whilst the choke or chokes are added to control the beam symmetry, backradiation, and by extension the spillover efficiency. This is achieved by controlling the currents on the rim of the waveguide. Olver et al. (1994) shows that with an appropriately sized choke-less waveguide it is possible to generate low cross-polar performance but, by adding a choke, much better cross-polar performance can be achieved with a smaller waveguide diameter (which is advantageous to avoid higher mode propagation). A smaller waveguide aperture may result in a broader beamwidth than desired, but this can be controlled to achieve the desired beamwidth by optimising the geometry of the chokes and waveguide.

For this project the frequency of operation is low enough such that we can take advantage of conventional machining techniques to manufacture a wide flare-angle feed without difficulty. As a result we decided to explore how such a feed would perform in comparison to current potential SKA feeds, none of which is a corrugated wide flare-angle feed.

The next section outlines the requirements of the feedhorn. This will be followed by a description on how the feed was designed. After this section, the expected performance of the feed when coupled with the SKA optical system will be presented. The last discussion in this chapter will be on the measurement results of the prototype feed horn.

3.1 Requirements

For wide-flare-angle horns (semiflare-angle greater than 30°) the beamwidth is mainly determined by the semiflare angle (Olver et al. 1994). The -15 dB beamwidth is approximately equal to the semiflare-angle. That is, the edge taper is determined by the flare angle of the feed. To maximise aperture efficiency whilst restricting spillover,

an edge taper illumination of about -15 dB is required for the secondary reflector. For the SKA dish optics this corresponds to a flare angle of about 120° . Therefore the desired semiflare angle is 60° .

The peak cross-polarisation to co-polarisation ratio should be better than -15 dB. This specification is based on the SKA specification that requires this to be -15 dB without calibration (Kusel et al. 2017). Return loss needs to be better than -20 dB across the band so that sensitivity and gain flatness are not compromised. A summary of these specifications is given in Table 1.1.

The above performance should be sustained across the entire band of operation. Return loss and lowest frequency of operation are what determine the feed horns waveguide diameter. Provided the waveguide is sufficiently larger than the cut-off diameter, any diameter could be used, but just how large is restricted by the need to avoid the excitation of higher order modes of propagation in the waveguide that would have a negative impact on the feed's beam pattern. To allow for ease of testing, the input waveguide diameter was set to be equal to the already existing SKA horn. Since the frequency performance scales with the physical size of the feed, for the initial design process the proposed feed was designed to operate in SKA band 4 (3 to 5.5 GHz) to allow for comparison with already existing SKA feed designs available at the start of this project (the SKA band 2/MeerKAT L-band feed, designed by EMSS and the initially proposed SKA band 4 feed, designed by the Joint Laboratory for Radio Astronomy Technology (JLRAT)) (Leech et al. 2016). The input waveguide diameter was designed to be the same as that of the existing feed also to ease comparison. After the design was completed it was then scaled and fine-tuned to operate between 4.6 and 8.5 GHz (SKA band 5a). A second version was also produced by scaling this design to operate between 8.3 to 15.3 GHz (SKA band 5b).

3.2 Design and simulation

Having discussed the specifications of the feed horn in Section 3.1, we will now move on to how the feed horn was designed. The initial design presented in Section 3.2.1 was developed by following guidelines provided by [Olver et al. \(1994\)](#). This design was then refined and the final design is given in Section 3.2.3.

3.2.1 Initial design

The beam pattern is the Fourier transform of the field at the aperture. Therefore, by controlling the field pattern at the aperture the desired field pattern can be obtained. If the aperture field consists of 85% TE_{11} and 15% TM_{11} modes, this results in a symmetrical beam pattern (i.e. E and H field patterns coincide), which translates into low cross-polarisation ([Potter 1961](#)). This can be achieved in various ways, but one of the most popular techniques is by corrugating the horn's interior walls. This has the effect of altering the surface currents, resulting in the desired HE_{11} mode. For narrow flare angles (semiflare-angle < 30 degrees) the corrugations are usually perpendicular to the axis for ease of manufacturing. For wide flare angle horns, ideally the corrugations should be perpendicular to the flare to achieve the desired wall impedance. Given the frequency band of operation (translating to reasonably larger dimensions of the horn) it is possible to manufacture the horn using 5-axis milling without difficulty.

[Olver et al. \(1994\)](#) recommends using a smooth-wall conical horn section at the throat, as this provides the best compromise between achieving a good impedance match to free space whilst still restricting the generation of higher modes which degrade the beam pattern.

Based on the previous discussions, the initial design consisted of a waveguide input section, followed by smooth-wall section, feeding into the corrugated section as shown in Figure 3.2. Given the reciprocal nature of antennas we can describe the functions of the various geometric features as follows. The smooth walled section right after the waveguide serves two functions. Firstly, it provides impedance matching as the

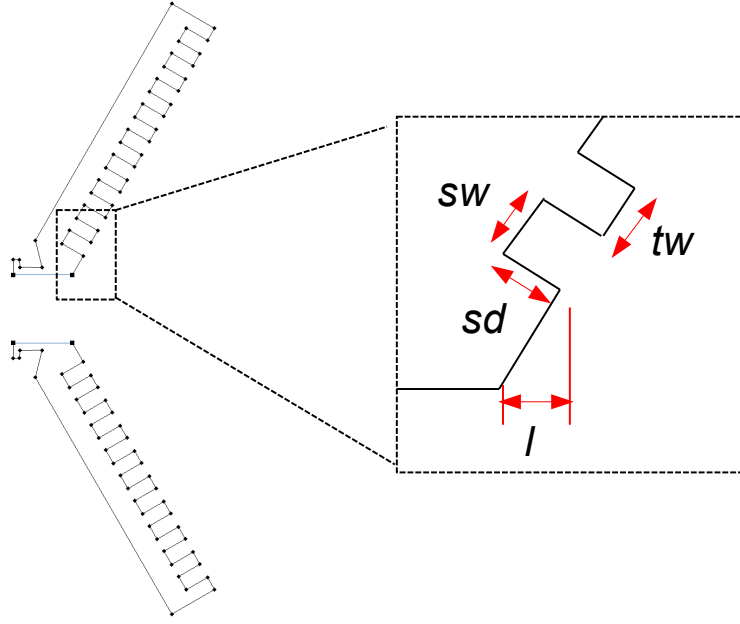


Figure 3.2: Initial Design Cross Section

waves transition from being bound in the waveguide to free space. Secondly it is the means by which part of the dominant TE_{11} waves are converted into the TM_{11} mode. The combination of these two give rise to the HE_{11} mode. The field distribution of this mode is very close to being Gaussian. Provided it is allowed to propagate to the feed aperture, the feed will exhibit a circular beam pattern and with low side lobes, as the Fourier transform of a Gaussian is Gaussian. The corrugations that follow after the smooth wall section provide the necessary surface impedance to allow the HE_{11} mode to propagate to the aperture and hence achieve the desired field distribution.

The semiflare angle was 60 degrees, this actually corresponds to the -15 dB beamwidth of the horn (selected to match that of the SKA band 4 horn). The aim was to have about 3.5 slots per wavelength with a total of 9 slots along the flare of the horn. The slot depth (sd), slot width (sw) and tooth width (tw) were $\lambda/4$ at the centre of the band, where the wavelength was considered to be the free-space wavelength. The feed was modelled and analysed in Corrugated Horn Analysis by Modal Processing (CHAMP)¹ software package. CHAMP works by sub-dividing the interior

¹CHAMP-TICRA, <https://www.ticra.com/>

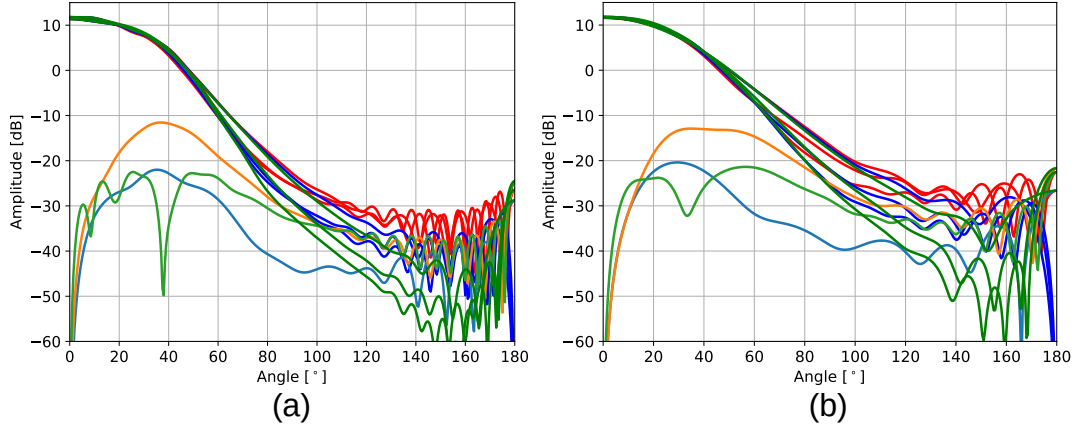


Figure 3.3: Beam Patterns for (a) Initial Design and (b) Second Design; where Red, Green and Blue traces are for Co-Polar plots for 2.8 GHz, 4 GHz and 5.18 GHz respectively. Light-blue, Orange and Light-Green are the corresponding Cx-Polar Patterns.

of cylindrically symmetric horns into a series of cylindrical waveguides and matching the reflection/transmission at each boundary. For non-cylindrical structures, it uses a method of moments technique. The resulting beam pattern of the initial design described above had relatively high cross-polarisation levels at higher frequencies and poor return loss at the lower frequencies. The slot depth was increased resulting in improved cross polarisation levels in the entire band of interest (not surprising, since the guide wavelength is longer than the free space wavelength; the initial depth was thus shorter than required). Further adjustments were made to all four parameters to improve the insertion loss and reduce the cross-polarisation levels. The beam pattern for frequencies 2.8 GHz, 4 GHz and 5.18 GHz are shown in Figure 3.3 (a). This design exhibited a circular beam pattern with low side lobes as can be seen in the figure, however the return loss at the lower end of the band was limited by the input waveguide diameter (shown in Figure 3.4).

3.2.2 Design refinement

The initial design had an aperture with a radius of 284.4 mm. This was considered a little larger than desired, especially since the cross-polarisation performance has

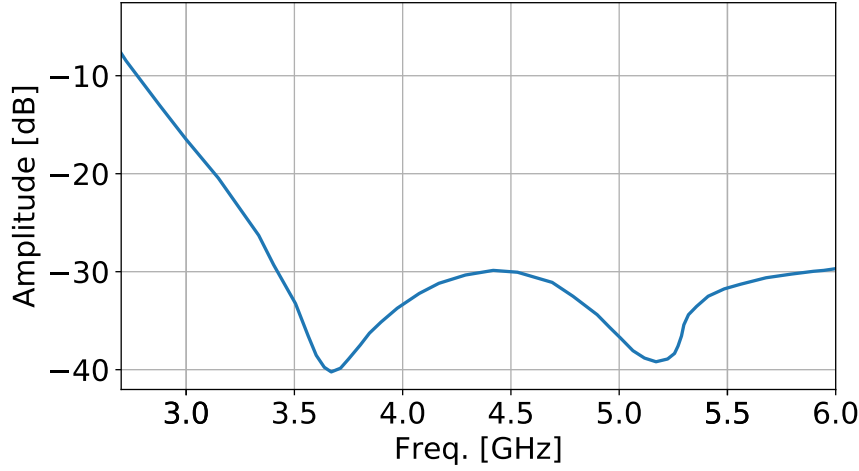


Figure 3.4: Reflection loss of the initial feed horn design.

little dependence on the number of corrugations for this type of horn. We also wished to make a fair comparison between the choked flange design already done for the SKA, which meant making the overall diameter the same (if possible). A larger feed will be able to give better sidelobe performance but may have additional costs, particularly if the feed is cooled as in the SKA band 5. Therefore, to reduce the size of the horn the flared corrugated section was shortened, resulting in the geometry shown in Figure 3.5 (a), where the outermost perimeter of the horn now has a radius of 155.9 mm. Unfortunately, although the cross-polarisation performance was not noticeably affected, the side lobe levels were approximately 10 to 15 dB higher than that of the initial design. To lower the side lobes or back radiation, a choke was added to the outer rim of the horn as this has been reported to lower backradiation (Olver et al. 1994). Thus, the geometry was as shown in Figure 3.5 (b). This had the effect of lowering the side lobes but it did not achieve the same performance as that of the initial design. This can be seen by comparing the patterns in Figure 3.3 (a) and (b). Nonetheless it was determined that the current sidelobe performance though not better than the initial design is superior to that achieved by the potential SKA horn designs of similar size (see Figure 3.8). Given that this feed satisfied the

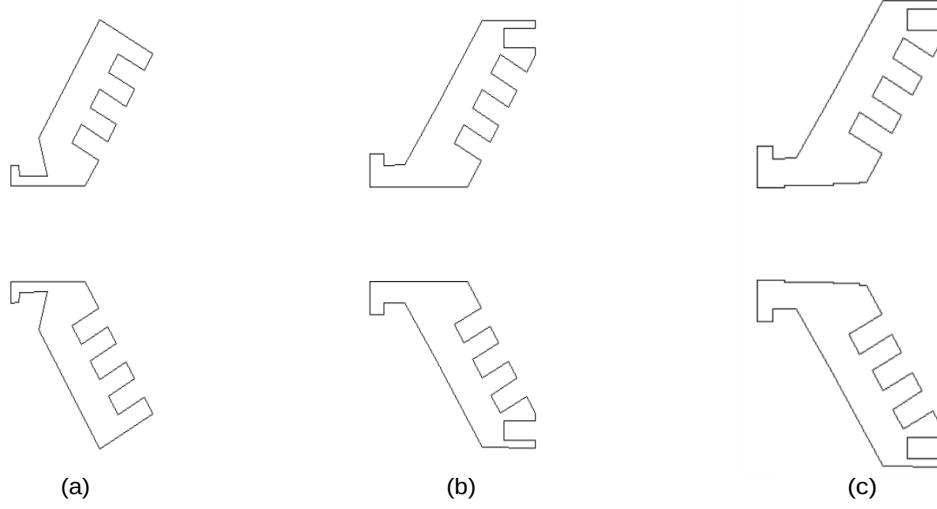


Figure 3.5: Feed Horn cross sections: (a) 2nd design corrugated slots perpendicular to surface. (b) In 3rd design added choke ring at rim of feed. (c) Final Design added stepped waveguide.

specifications it follows that the proposed design's sidelobe performance will be more than adequate. The return loss was unaffected by the shortening of the flare section. The design process is summarised in Table [3.1](#)

3.2.3 Final design

With the initial design of the feed complete, the feed dimensions were scaled to the desired band of operation 4.6 to 8.5 GHz. The input waveguide diameter was such that the lowest frequency of operation was relatively close to the cut-off frequency of the waveguide. The motivation for this was to try and maximise the bandwidth over which unwanted modes are not able to propagate in the waveguide, hence increasing the bandwidth of operation of the feed. Unfortunately this translates into poor return loss performance at the lower end of the band, as the feed operates close to the cut-off frequency of the dominant mode. To alleviate this whilst maintaining the same input waveguide diameter, steps were added to the waveguide section. The resulting waveguide structure is a waveguide whose diameter increases in steps from the input port towards the feed's throat. This improves the return loss by enhancing the match

Table 3.1: Summary of the feedhorn design process.

Design parameter	Design iteration			
	1	2	3	4
Aperture Size	568.8 mm diameter, 100° to 180° sidelobes are about -40 dBi (Figure 3.3 (a)).	Reduced to 311.7 mm, 100° to 180° sidelobe levels have been increased (Figure 3.3 (b)).	No change	No change
Semi-flare angle	120° flare angle achieves -15 dB edge taper.	No change	No change	No change
Corrugation	$sw = sd = tw = \lambda/4$, where λ is greater than the free-space wavelength at the centre of the band, improves cross-polar at high frequencies.	No change	No change	No change
Waveguide	63.3 mm diameter, achieves poor return loss at low frequencies (Figure 3.4).	No change	No change	Added steps, improved return loss at low frequency (Figure 3.6).
Outer Choke	None	None	Added to the outer rim, 100° to 180° side-lobes about -30 dBi (Figure 3.7).	No change

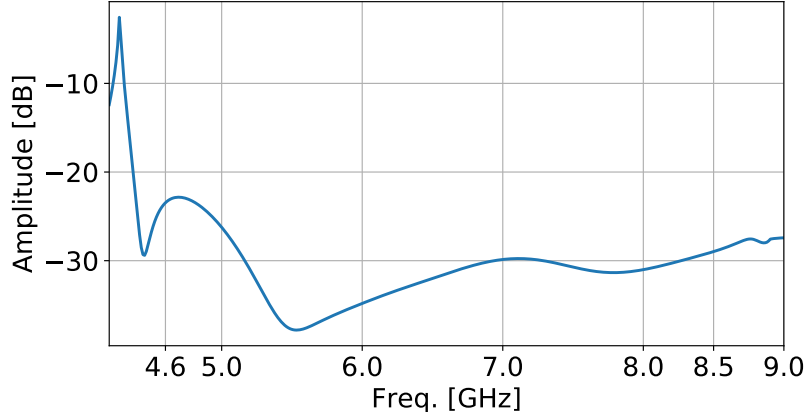


Figure 3.6: Final band 5a feedhorn design return loss.

to free space. Secondly, the length of the waveguide with the lower cut-off frequency is shortened thus we can expect reduced ohmic losses. Both of these result in better system temperature. The resulting return loss is given in Figure 3.6. The return loss across the desired band 4.6 to 8.5 GHz is well below 20 dB.

By changing the relative lengths and heights of the steps in the waveguide it was found that cross-polarisation performance at higher frequencies could also be improved. This was used to achieved good cross-polar performance across a broader portion of the desired band. The final design is given in Figure 3.5 (c).

The beam patterns of the final design are shown in Figure 3.7. The sidelobes and backlobes are below -20 dBi across the band; lower than those of the scaled MeerKAT L band horn (also one of the potential feeds for the SKA). We also see that the variation of performance with frequency is minimal. Figure 3.8 shows the simulated beam patterns for three feed horns scaled to operate between 3 to 5.5 GHz. Part (a) is the structure and beam pattern of one of the feeds initially proposed for SKA band 5 (feed A). This feed was developed by JLRAT. Part (b) is the proposed AVN feed (feed B) with its corresponding beam patterns. This horn is a wide-flare-angle design with corrugations perpendicular to the axis. Part (c) is the MeerKAT L-band feed, a choked flange design (feed C), along with beam patterns (this also a potential feed for the SKA band 5).

The sidelobe and backradiation performance of the proposed AVN feed is better

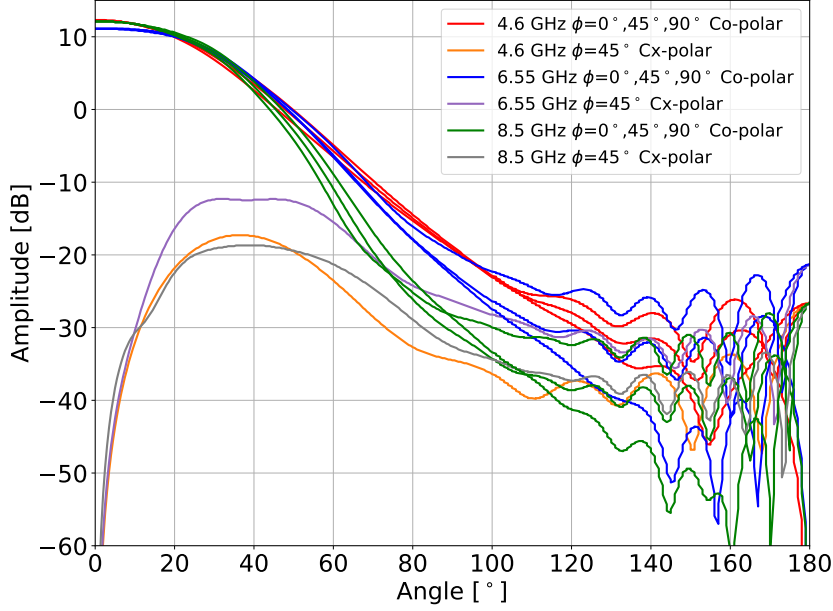


Figure 3.7: CHAMP generated beam patterns for the final band 5a feedhorn design.

than the other two. The co-polar pattern level drops to around -30 dB between 90° and 180° , whilst for the other two the level is around -20 dB (except for the MeerKAT feed which at the top edge of the band also drops to -30 dB). In principle we should expect better spillover efficiency from the AVN feed across the entire band as a result of this. Unavoidably the AVN feed ended up being slightly larger than the other two, with aperture diameters of 280.5 mm, 311.7 mm and 241.9 mm, for A, B and C respectively. Initially we thought that may be the reason why the AVN feed exhibits better sidelobe performance than the other two but, if that were the case then we would expect feed A to have achieved better side lobe performance than C across the whole band. Therefore the difference in diameter is not large enough to be the only reason for the almost 10 dB improvement in sidelobe levels.

The simulated cross-polar performance of all three feeds is better than the target specification of -15 dB at the lower end of the bandwidth. On the other hand, at the top end of the band at 5.5 GHz, feed A loses beam circularity and a diminished cross-polar performance is predicted. This is due to the structure of the feed. Generally

for narrow flare angles it is acceptable for the corrugations to remain perpendicular to the axis of the feed but not the wall of the flared wall section, whereas for wide flare angle horns, the corrugations are more effective when they remain perpendicular to the wall to ensure the appropriate surface impedance is achieved over much of the desired band as possible. The MeerKAT and AVN feeds generally exhibit the best cross-polar performance of the three and so we took a closer look at these at the actual desired design bandwidth of 4.6 to 8.5 GHz. The ratio of maximum cross-polar-to-peak gain over a broader band were plotted, these results are shown in Figure 3.9. Generally we observe similar cross-polar performance across the desired band, with the MeerKAT feed achieving better performance in the middle of the band. The AVN feed has better performance at the lower and top edges of the band. This is also true for frequencies higher than the top edge of the desired band. This indicates that this design may have broader bandwidth capabilities with respect to cross-polar performance. Therefore, the AVN feed could potentially be used over a 2:1 bandwidth, whilst the MeerKAT feed is limited to 1.85:1. Fundamentally, this stems from the cut-off frequency of the TM_{11} mode in circular waveguides, this is evident from the MeerKAT feed's diminished beam circularity at 6 GHz due to the propagation of the TM_{11} mode (Figure 3.8 (c)). The SKA's 1.85:1 bandwidth sub-bands are as a result of this consideration.

Lastly the ohmic losses of the MeerKAT and AVN feeds were compared at band 5b (8.3 to 15.3 GHz), since we expect higher losses at higher frequencies. These results were generated using CHAMP and are shown in Figure 3.10. There is not a significant difference between the two in terms of ohmic loss performance. It is still worth noting that the proposed AVN feed does exhibit slightly lower ohmic losses across the entire band, with the difference being greatest at the lower end. The reason for this disparity, despite both feeds having the same input waveguide diameter, is the variation in the geometry of the steps in the waveguide of each feed. The diameter is such that the feed operates relatively close to the cut-off frequency of the dominant mode (to maximise the spurious mode free bandwidth). As a result the attenuation constant of the waveguide is still reducing with increase in frequency

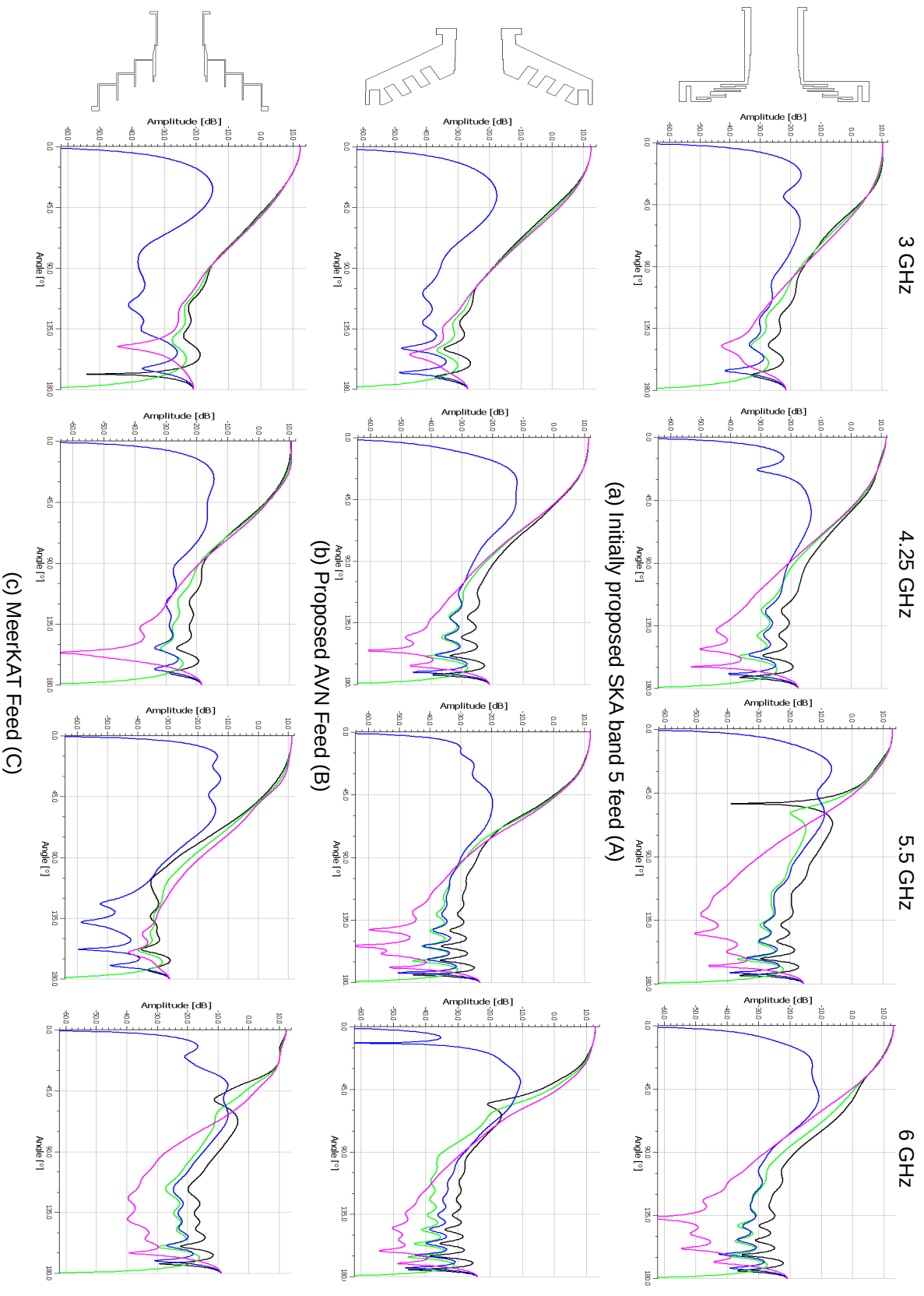


Figure 3.8: Comparison of simulated beam patterns for horns scaled to band 4 (3 to 5.5 GHz): (a) Initially proposed SKA band 5 feed (b) Proposed AVN feed (c) MeerKAT L-band feed. For all plots the y -axis (amplitude) scale runs from 10 db to -60 dB in 10 dB increments and the x -axis (angle) from 0° to 180° in 45° increments.

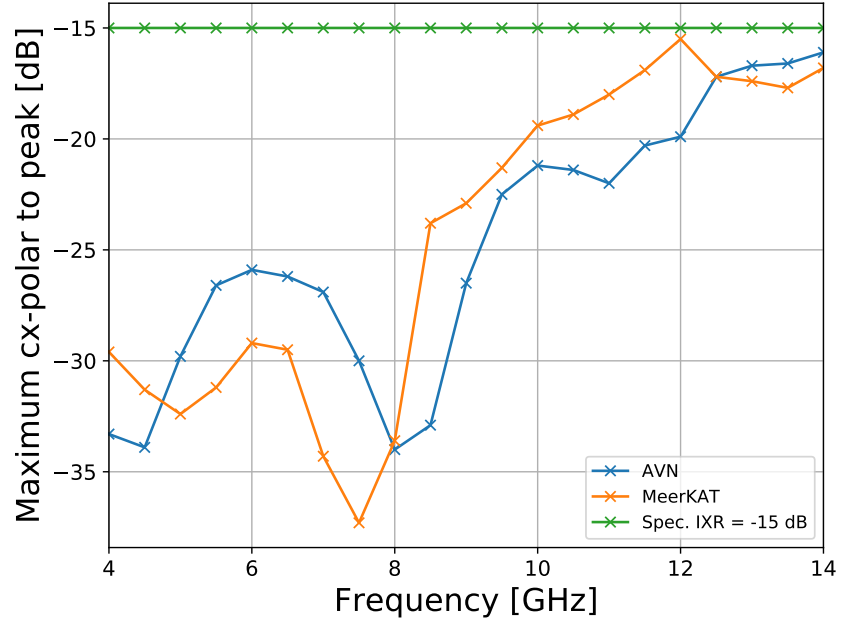


Figure 3.9: Comparison of the ratio of maximum cross-polar to peak gain of the MeerKAT and proposed AVN feeds scaled to band 5a (4.6 to 8.5 GHz).

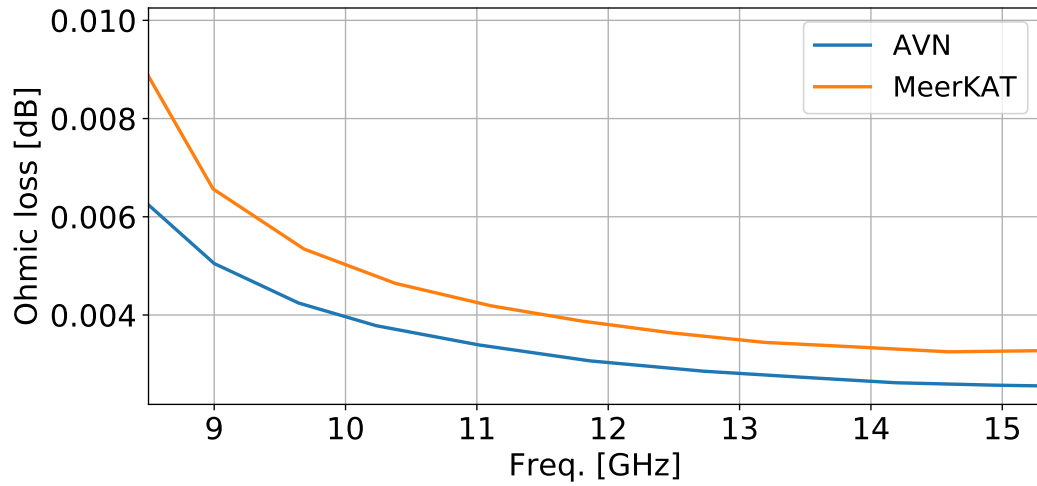


Figure 3.10: MeerKAT and proposed AVN band 5b feed horn Ohmic loss for aluminium material.

hence the negative slope we observe in both traces in Figure 3.10. For the AVN feed, the input waveguide diameter is maintained for a shorter distance before the first step occurs resulting in a broadening of the waveguide hence shortening the length over which the higher attenuation occurs. Therefore in addition to improving the return loss and cx-polar, the steps also reduce the ohmic losses of the AVN feed. In terms of system temperature, if cooled to 20 K the MeerKAT feed contributes 0.04 K whilst our feed would add 0.03 K at 8.5 GHz. At room temperature each feed would contribute 0.6 K and 0.4 K respectively at 8.5 GHz. For the AVN receiver the feed will not be cooled therefore the slight gain in system temperature will be beneficial.

The horn design was further verified using ANSYS HFSS². CHAMP simulates the feeds by applying the Mode Matching technique to the section defined as the horn interior, which in the case of our horns is the input waveguide up to the last step towards the aperture. The remainder of the structure, which are the chokes or corrugations, are defined as the horn exterior and are simulated using the Method of Moments. HFSS simulations are done differently, solving a full-wave electromagnetic simulation over the entire structure of the feed using a finite-element analysis algorithm. This allows for any arbitrarily shaped structure to be simulated whilst with CHAMP rotational symmetry is assumed. Therefore simulating the horn with HFSS allows us to further verify the simulation results. The results generated using HFSS are shown in Figure 3.11. The image shows how a TE_{11} mode excited at the waveguide port propagates through the feed and arrives at the aperture. The electric field intensity is plotted on cross sections at different positions along the feed. From these plots we see the distribution becoming progressively circular as the field propagates along the length of the horn. The results of the simulations for the beam patterns obtained using HFSS agreed very well with the results from CHAMP.

What differentiates this proposed design from previous designs presented by Miligan (2005), Abbas-Azimi et al. (2009) and Xie et al. (2016) are the use of a stepped input waveguide and the addition of a peripheral choke on a wide-flare feed horn. The transducer, which will be presented in the next section, covers the entire band from

²Ansyz High Frequency Structure Simulator (HFSS), <https://www.ansys.com>

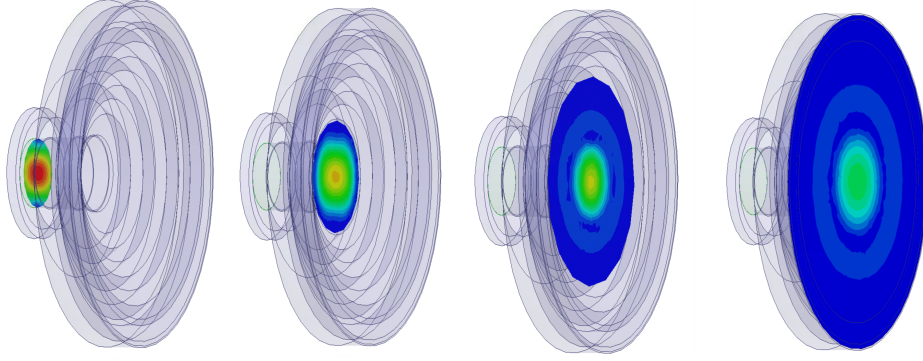


Figure 3.11: ANSYS HFSS electric field magnitude simulation at 6.9 GHz for proposed AVN band 5a feed.

4.6 to 15.3 GHz. Therefore in order to take full advantage and increase the science that can be done using this instrument it would be beneficial to have a feed that can also cover the top part of this band. Due to time constraints the most practical solution was to produce two feeds, one that could cover 4.6 to 8.5 GHz (band 5a) whilst the other would cover 8.3 to 15.3 GHz (band 5b). The higher frequency feed was produced by scaling the geometric dimensions of the designed prototype. The RF electronics were also designed to cover this wider bandwidth so that a future upgrade would only require the design of a single feed that can provide the required 3.3 to 1 bandwidth. One possible technique of achieving such a design is by using dielectric loading (Dunning et al. 2015). Another feature that can be explored is the use of ring-loaded slots as was done by Xie et al. (2016).

3.3 Physical optics simulations

The final stage of numerical validation involved simulating the coupling of the feed horns with the SKA dish optics. This was done using GRASP³, General Reflector Antenna Software (TICRA 2020), which uses the Physical Optics method. This is achieved by importing the spherical wave expansion of the feedhorn’s beam pattern (generated by CHAMP) into the GRASP model of the SKA antenna. The model used

³GRASP-TICRA, <https://www.ticra.com/>

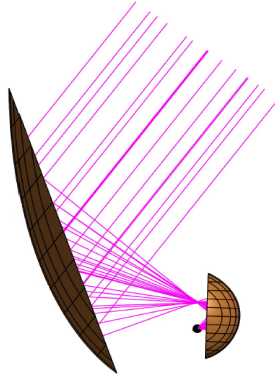


Figure 3.12: Physical optics GRASP simulation model of the SKA dish.

for the simulation is shown in Figure 3.12 (the GRASP simulation model assumes an elliptical, continuous surface primary rather than the segmented polygonal primary of the actual SKA design. It also ignores the surrounding structure, deformation due to wind loading, effects of gravity or temperature and also ignores the surrounding structure and any surface errors). In this section we present the key performance characteristics predicted by the simulations. Two main plots will be presented, the first being the main beam which lies between $-1 \leq \theta \leq 1$ degrees. This is the region of the beam that determines how the telescope will respond to the source being observed. The second set of plots will give the full beam pattern between $-180 \leq \theta \leq 180$ degrees. In these plots the focus is looking at the region of the beam outside the main beam: it is this region that determines the spillover performance of the telescope.

3.3.1 Beam patterns

We start by looking at the main beam patterns for four frequencies across the desired bands. The patterns for band 5a are shown in Figure 3.13 whilst band 5b patterns are in Figure 3.14. Looking at Figure 3.13 we see that the beam is circular across the whole band. This is also true for the scaled version in Figure 3.14. This is confirmed by the low peak co- to cross-polarisation ratio. In Figure 3.15 we see that the cross-polarisation performance is better than the allowable minimum of 15 dB. The first

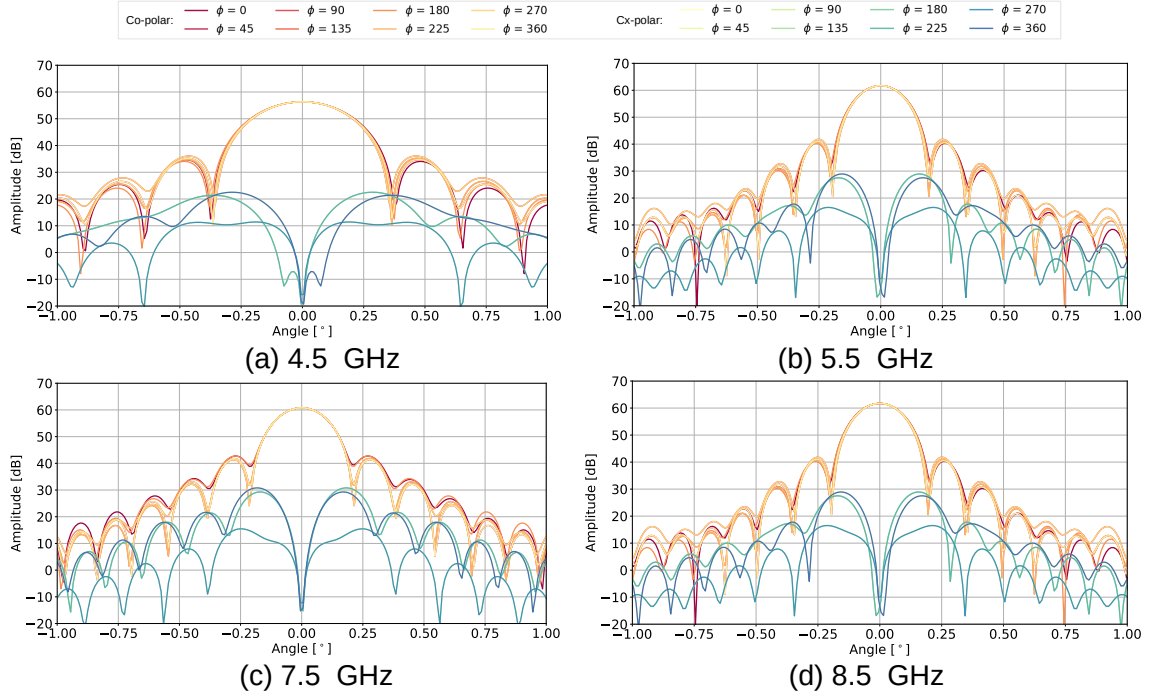


Figure 3.13: Far-field radiation patterns of the SKA-mid antenna illuminated with the AVN band 5a feed. Co- and cross-polar beam patterns for 45° Φ cuts, between $-1 \leq \theta \leq 1$ degrees.

sidelobes are about -20 dB for both feeds. The gain ranges from 56 to 62 dB and 62 to 65 dB in the two bands. Aperture efficiency is better than 80% across both bands, as shown in Figure 3.15.

Figures 3.16 and 3.17 show the beams for band 5a and 5b feeds respectively in their entirety. Here we focus on the area outside the main beam. The performance in this region determines how sensitive the telescope is to spillover (pick up from the ground and the area of the sky that the feed sees outside the edges of the secondary reflector). This has an effect on the system temperature of the receiver, hence has a bearing on the maximum attainable sensitivity. The closest sidelobes to the main beam appear between 25 to about 90 degrees. During observations these will most likely be pointing at the cold sky therefore should not have significant effect on the system temperature. The far out sidelobe (sidelobes beyond the eighth null) peaks are below the specified level, 6 dBi. These play a crucial role in determining the spill over efficiency as they affect the system temperature via ground power pick up.

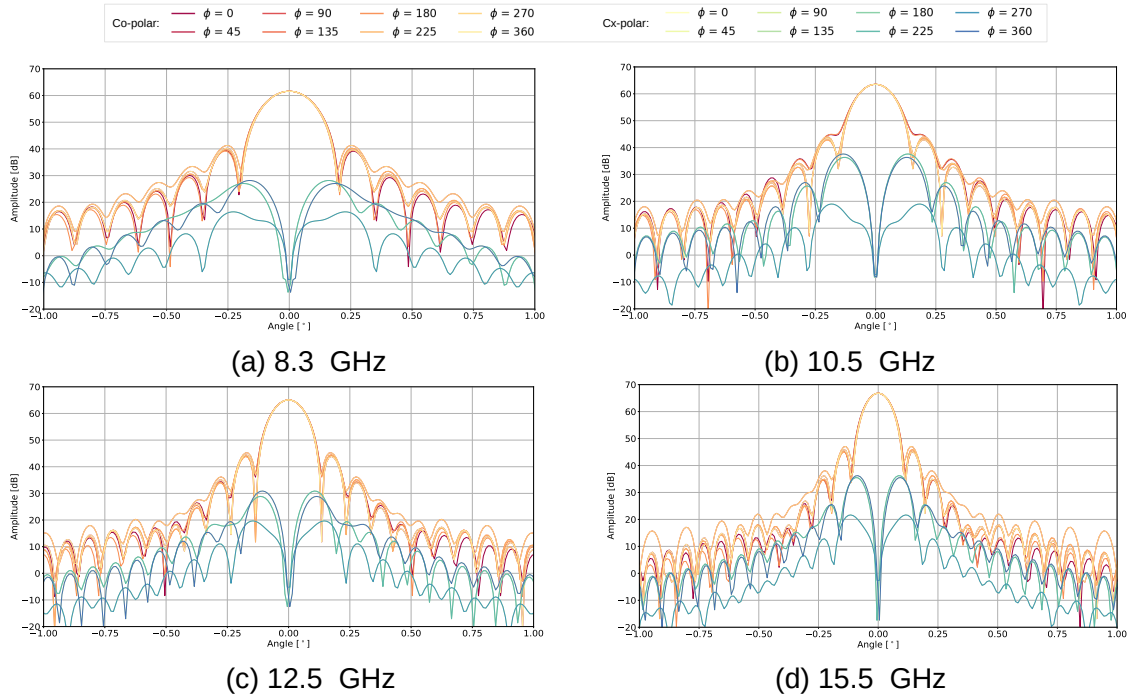


Figure 3.14: Far-field radiation patterns of the SKA-mid antenna illuminated with the AVN band 5b feed. Co- and cross-polar beam patterns for 45° Φ cuts, between $-1 \leq \theta \leq 1$ degrees.

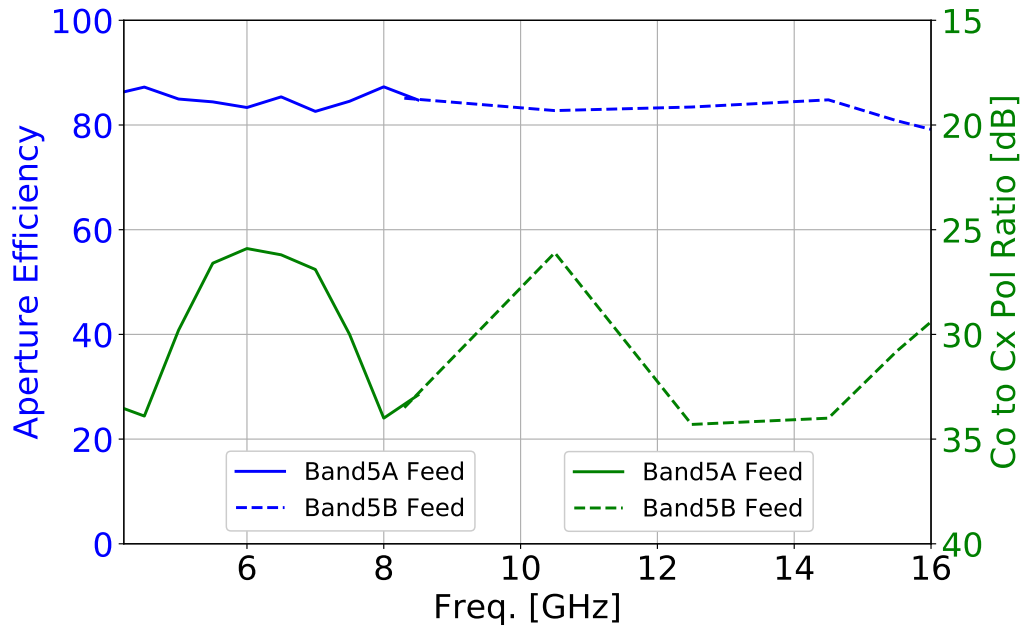


Figure 3.15: physical optics simulation aperture efficiency and cross-polar performance.

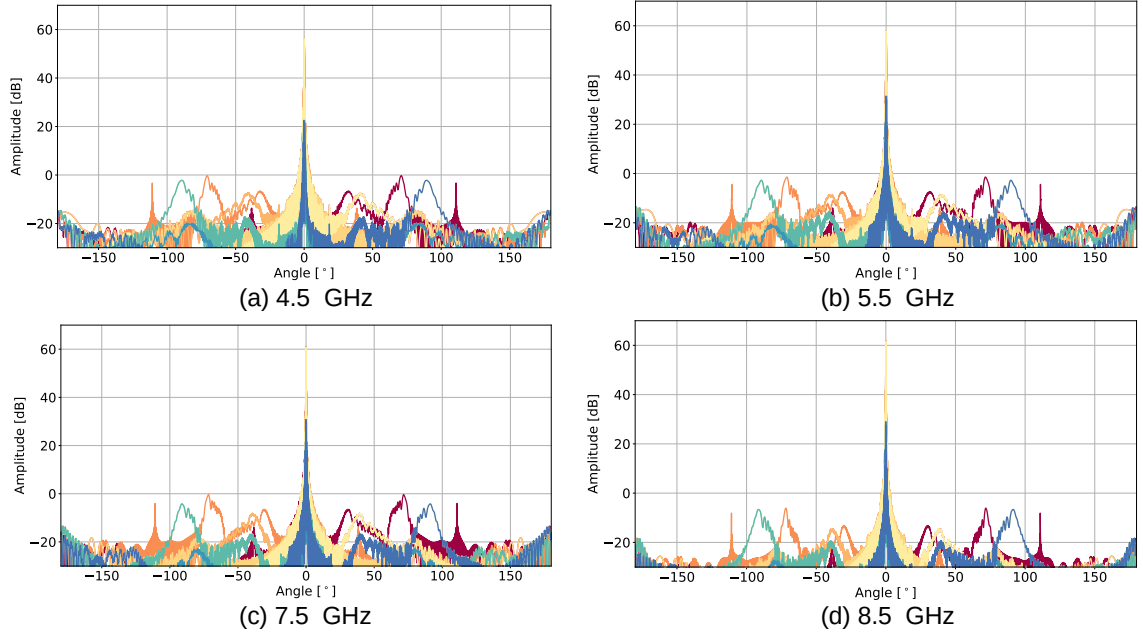


Figure 3.16: Far-field radiation patterns of the SKA-mid antenna illuminated with the AVN band 5a feed. Co- and cross-polar beam patterns for 45° Φ cuts, between $-180 \leq \theta \leq 180$ degrees.

How much they contribute to the overall system temperature will only be established after integrating over the whole sky. More details on this will be covered when the predicted system temperature is presented in Chapter 6.

From these results we conclude that the proposed feeds show good compatibility with the target telescope optics. Prototypes of both feeds were then manufactured for testing and verification.

3.4 Prototype feed measurement results

The measurement of the feedhorn's return loss was conducted by Andre Hector. The SKA band 5a orthomode mode transducer (OMT) (Du et al. 2019) was used to measure the return loss as shown in Figure 3.18. The network analyser was calibrated up to the coax ports of the OMT using a through, open, short and match (TOSM). The results of this measurement are given in Figure 3.19. Despite the ripple, at the lower end of the band, the measured return loss agrees with the simulated values.

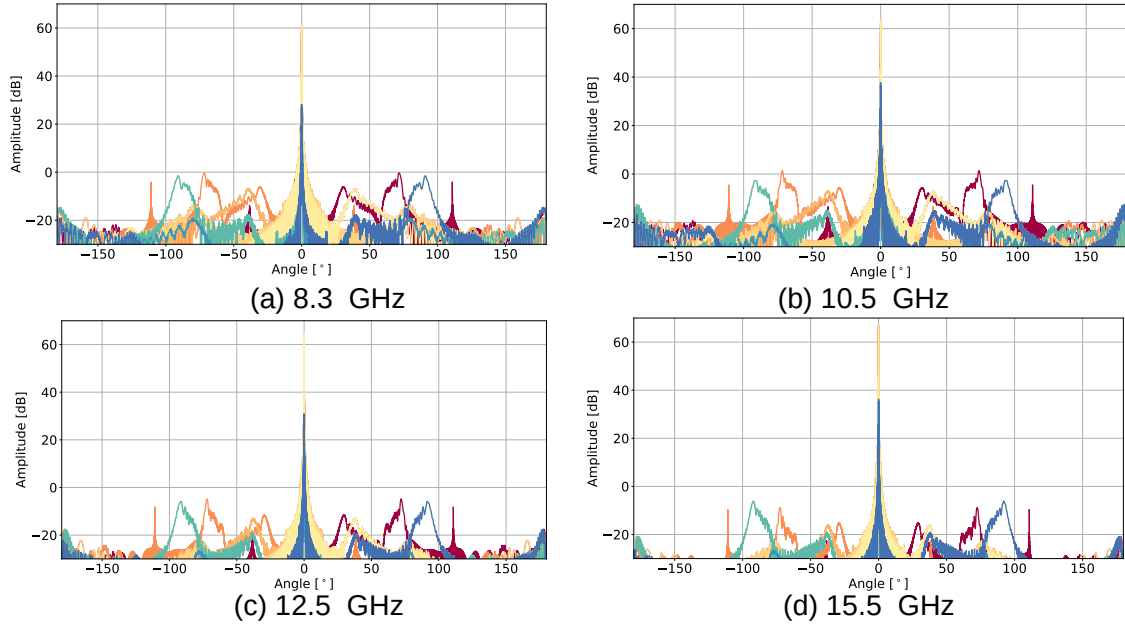
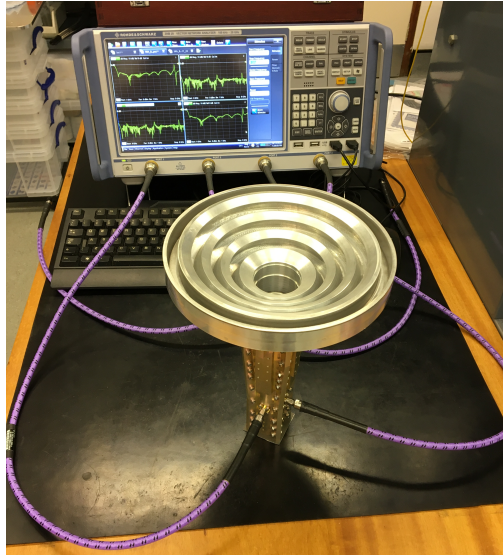


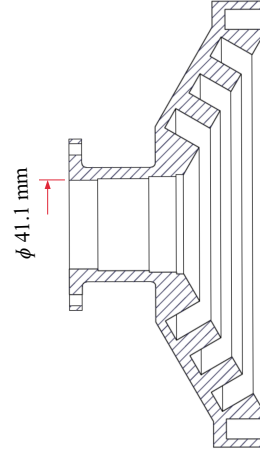
Figure 3.17: Far-field radiation patterns of the SKA-mid antenna illuminated with the AVN band 5b feed. Co- and cross-polar beam patterns for 45° Φ cuts, between $-180 \leq \theta \leq 180$ degrees.

At the top end of the band the measured is higher than the predicted values. This is as a result of the higher return loss of the OMT which limits our measurements. This is apparent from studying the green trace which is a measurement of the OMT's return loss (when its waveguide port faces a cone lined with a RF absorber). This measurement provides a good indication that the feedhorn does meet the better than 20 dB return loss specification across the desired band.

Using a waveguide to coaxial transition that covers 8.2 to 12.5 GHz the beam pattern for the AVN band 5b version was measured. These measurements were done at the technical University of Munich, courtesy of Dr Christian Holler. The results are presented in Figure 3.20. The measurement was carried out from -90° to 90° , the results only cover the positive angle. The measured and simulated co-polar patterns agree between 0 and 60 degrees. Unfortunately the peak cross-polar measurement was not carried out. None the less the fact that the co-polar plots agree well enough over the main beam is encouraging.



(a) AVN feed and SKA band 5a OMT



(b) AVN feedhorn Cross-section

Figure 3.18: AVN band 5a prototype feed and SKA OMT.

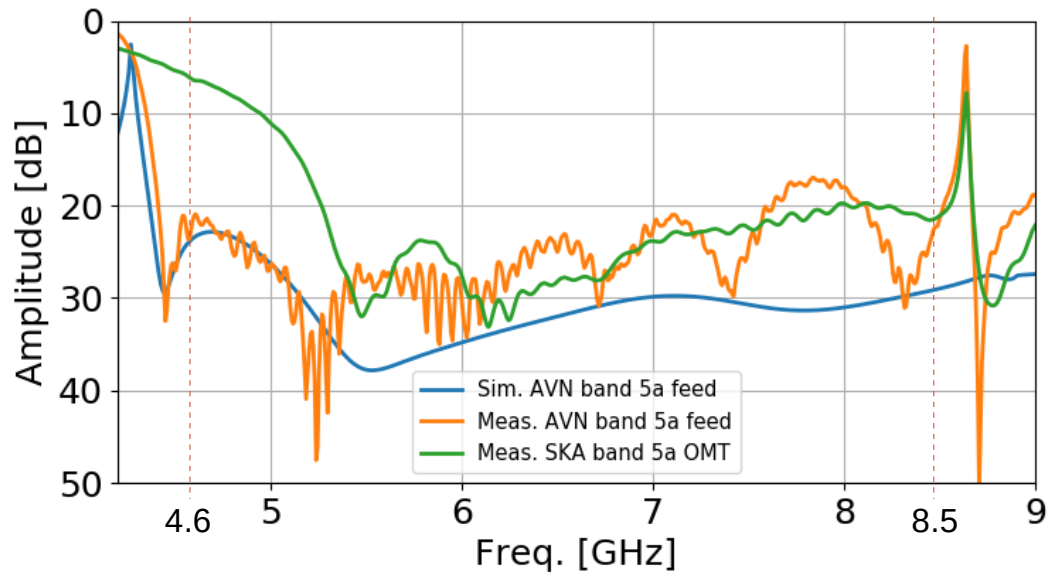


Figure 3.19: Return Loss: Feed and Flann Transition

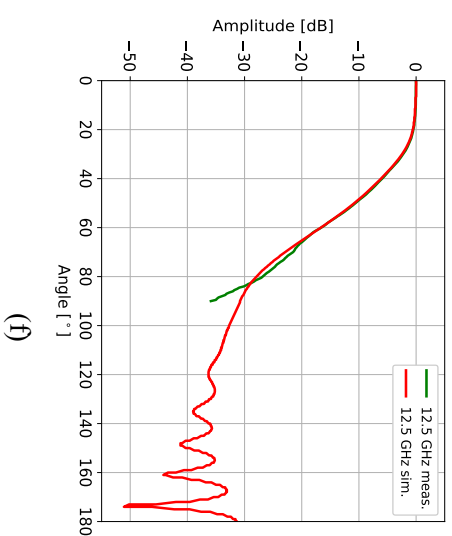
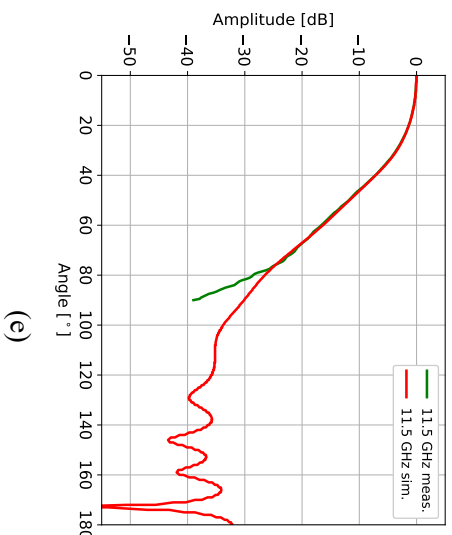
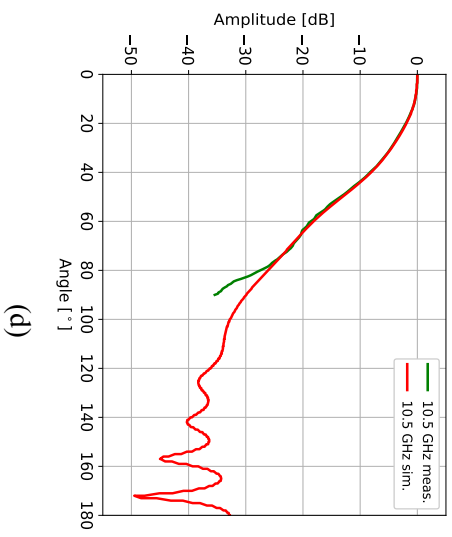
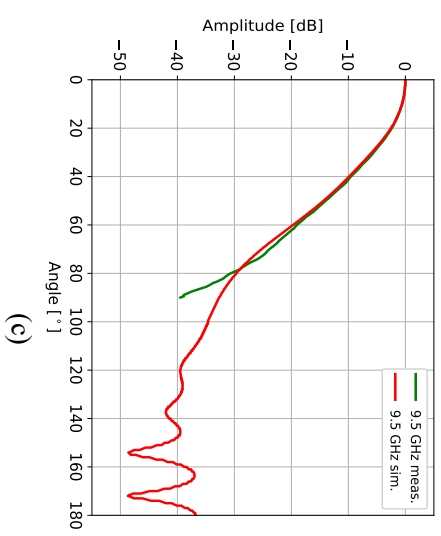
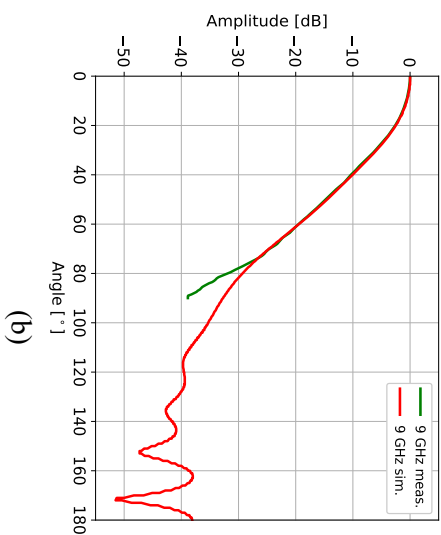
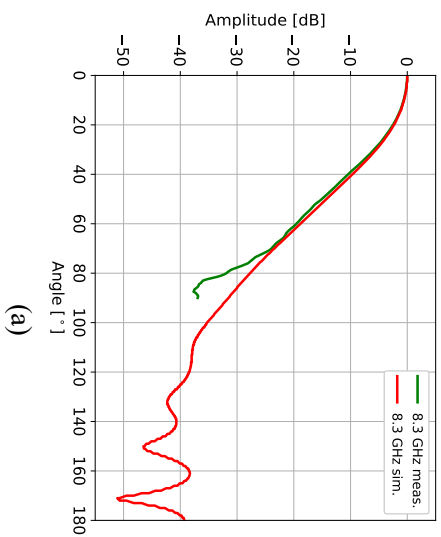


Figure 3.20: AVN band 5b feed measured and simulated E-plane beam pattern.

3.4.1 Summary

In this chapter the design of a wide-flare-angle corrugated feedhorn was presented. The design procedure of the feed was outlined. The predicted performance is a return loss less better than 20 dB and co-polar peak to cross-polar peak ratio better than -22 dB, over a 2.14 : 1 bandwidth. The feed was then scaled to produce a second version to allow for the instrument to cover 8.3 to 15.3 GHz. We then presented the expected far-field radiation pattern of the telescope when using the designed feeds. The performance was shown to satisfy the design requirements. The performance generally matches that of potential feeds for the SKA while for some parameters such as ohmic loss it performs better. The chapter was concluded by presenting the initial results of the measurement of one of the feed horns. The measurement results over the mean beam are shown to agree very well with the simulation results, therefore we expect this design to perform as desired.

Chapter 4

Triple Ridge Transducer

With the discussion on the feedhorn complete we now turn to the transducer. Although the feedhorn and transducer have been considered separately, these two devices are closely linked to each other. In some cases they are designed as a single unit. In this case for ease of design it was more beneficial to develop each one independently, but with a common waveguide interface port which is assumed to propagate the energy received by the feedhorn via the dominant degenerate modes only (Transverse Electromagnetic 11, TE_{11}). This is a reasonable assumption given that care was taken to design the feedhorn so that no higher modes are generated. The role of the transducer is to convert the electromagnetic radiation captured by the horn into electrical signals that can further be processed. In this chapter a novel transducer design will be presented. Section 4.1 will highlight the requirements. In Section 4.2 conventional transducers will be considered thereby motivating the need for the proposed transducer design. Section 4.3 will give a background on waveguide structures with three-fold symmetry. The design and simulation of the novel transducer will be covered in Section 4.4. The chapter will close with Section 4.5, where the measurement results of the transducer prototypes will be presented.

4.1 Requirements

The goal is to cover as much of the spectrum as possible between 4.6 GHz and 15.3 GHz. This corresponds to Band 5 of the SKA specification, which in the SKA implementation is split into two bands of frequency range $< 2:1$ each. This ensures compatibility of the AVN design with the SKA, while reducing the complexity of the system by having a single band. The return loss should be better than -15 dB across the band. For a full polarisation receiver the ideal transducer allows for the signals of different polarisation to be converted without any coupling between them over the entire band of interest. As a result the transducer should not introduce cross-polarisation coupling that can not be corrected via calibration in the digital signal processing. The input waveguide port should be the same size as the band 5a feedhorn discussed in the previous chapter. To use the transducer with the smaller, higher frequency feed an adapter waveguide can be attached.

4.2 Conventional transducers

The simplest transducer can be implemented by extending the centre conductor of a coaxial transmission line into a circular waveguide. The shielding or ground conductor is terminated on the waveguide. For optimum coupling the position of the probe relative to the backshort of the waveguide is critical; ideally it should be a quarter of the wavelength from the backshort and perpendicular to the circular waveguide. The length is also tuned to provide best performance. For full polarisation implementation a second probe can be inserted orthogonal to the first. Very good performance can be achieved this way, but over a narrow bandwidth, for example the Flann transducers used in the measurement of the feed in Chapter 3 provided a 1:1.5 bandwidth (-24 dB return loss).

For broadband operation, the Quad ridge waveguide (QRWG) orthomode transducer (OMT) is usually the first choice for transducer implementation e.g. (Dunning et al. 2013). The structure consists of two sets of protruding ridges from opposite

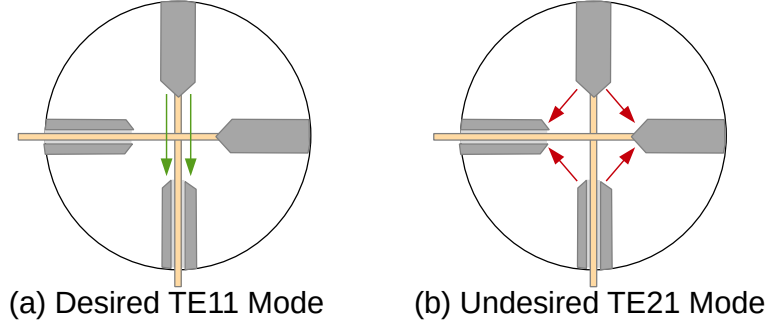


Figure 4.1: QRWG OMT cross-section illustrating the, (a) desired TE₁₁ mode electric field lines for the vertical polarisation and (b) the undesired TE₂₁ mode field lines.

faces of the waveguide where separation is progressively reduced. The orthogonal linear polarisations are then extracted using launch pins between two opposing ridges. Despite its popularity it is not without flaws. The most notorious of these is unwanted resonances also known as suck-out, which appear as a sharp and isolated notch in the insertion loss in the operating bandwidth. They result in loss of sensitivity at the frequency which the resonance occurs and maybe mistaken for a real signal. These suck-outs are generated when there is a trapped mode as a result of the quad-ridge structure being able to support two desired orthogonal TE₁₁ modes, along with the undesired TE₂₁ mode in the regions of the structure where the ridge separation is at its smallest. This property also results in poor polarisation isolation as some of the electromagnetic radiation contained in one of the desired degenerate modes maybe be converted to the TE₂₁ mode and couple into the probe extracting the other desired degenerate mode. This is easier to see by looking at the field lines as illustrated in Figure 4.1. From this Figure once can appreciate how the propagation of the TE₂₁ mode can result in compromised polarisation purity.

One of the ways in which this can be overcome is by using an Offset QRWG OMT (de Villiers et al. 2009, Dunning et al. 2013). The probes are geometrically separated such that only two opposing ridges are closely spaced in regions of the structure with a coax-ridge transition. In these regions the structure approximates a Dual ridged waveguide (DRWG) which are known for having a broad-bandwidth of single mode operation (Hopfer 1955). For some systems it is possible to use this technique to push

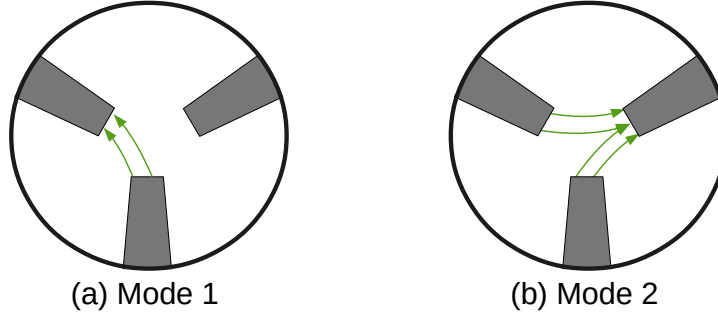


Figure 4.2: TRCWG structure cross-section illustrating mode 1 and mode 2, which are the two degenerate dominant modes of propagation.

the TE_{21} mode cut-off out of the desired band. Even though single mode bandwidths of 3:1 are possible with a DRWGs the offset QRWG OMT at best only approximates a DRWG therefore it will not achieve the same broadband characteristics as the DRWG. Consequently for broadband systems the spurious TE_{21} mode cut off still lies within the desired band and may result in some amount of spurious mode generation as shown by [Dunning et al. \(2013\)](#). Another disadvantage of this approach is that the OMT will tend to be longer than the design with co-located probes. For cryogenic receivers, it is desirable to have a more compact OMT to reduce the cooling load and cut down on cost, thus co-located probes are preferable in this regard.

4.3 Transducers with three-fold symmetry

Waveguide structures with three-fold symmetry such as triangular waveguides and triple-ridged waveguides have been reported to provide a wide bandwidth over which only two degenerate modes can propagate. [Dillon & Gibson \(1995\)](#) present the modal hierarchy and field patterns for a triply-ridged structure (TRCWG) generated using numerical techniques. It is further reported that an order of 30% improvement in single-mode bandwidth over QRWG structures can be achieved. This is confirmed experimentally using a triple-ridged waveguide resonator. A simplified illustration of the two orthogonal dominant modes in the TRCWG is shown in [Figure 4.2](#).

In addition to comparing the QRWG to the TRCWG numerically, [Morgan \(2011\)](#)

takes it further by comparing the single-mode bandwidth characteristics of triangular waveguides with that of the square and circular waveguides, as well as coaxial triangular and coaxial circular waveguides. It is also shown that a waveguide with a triangular cross-section has a broader single mode bandwidth than circular or square waveguides. The simulations predict that the triangular waveguide provides a $1 : \sqrt{3}$ spurious free bandwidth whilst the square waveguide achieves $1 : \sqrt{2}$. Of all the structures the TRCWG is shown to provide the greatest bandwidth. The simulation results for the QRWG and TRCWG indicate that (for relatively similar geometries) as the ratio of the ridge-gap to waveguide diameter reduces, the bandwidth free of spurious modes of the TRCWG increases exponentially whilst for the QRWG it reduces and approaches zero. For example, given a ridge-gap to waveguide diameter ratio of 0.2, the TRCWG achieves a $1 : 2.85$ ratio whilst the QRWG only achieves a $1 : 1.1$ bandwidth.

An intuitive explanation for why three-fold symmetry especially in the case of the TRCWG offers broader spurious-free bandwidth is given by considering that as the ridge-gap reduces the electric fields are increasingly concentrated in the opposing parallel surfaces of the ridges (the gap between the ridges). In the limit the remainder of the ridge serves no electrical purpose (for the dominant modes) and can thus be ignored. Each ridge gap can thus be thought of as one of the wires in an N -wire transmission line, with $N = 3$ in this case. Given that the total number of TEM modes the resulting structure can support is $N - 1$ it follows that the TRCWG with an appropriate ridge-gap will tend to support two modes only over a broad bandwidth (whilst the QRWG supports three). These two modes are the desired degenerate dominant modes.

[Morgan \(2011\)](#) also presents a prototype OMT based on a triangular- and tri-coaxial waveguide structure, but single-mode bandwidth possible with this structure is much less than what is achievable with a TRCWG structure. Despite the merits of using the latter structure, a transducer design that fully exploits these benefits has not been reported. For this thesis a novel transducer design that takes advantage of the TRCWG structure was implemented and is the subject of discussion for the

remainder of this chapter. From this point forward we will refer to the TRCWG as the triple-ridge transducer (TRT).

4.4 Design and simulation

The transducer was split into two sections. The first section provides a transition from a purely circular waveguide structure (where the horn is attached) to that of a circular waveguide with ridges protruding from the waveguide wall to the centre. The second stage is the transition from a triple-ridged waveguide to a coaxial transmission line. With no available literature on the latter, with regards to a 3-probe coaxial transition, it was tackled first.

4.4.1 Ridged-waveguide to coax transition

A feature of this transducer is that it has two degenerate lowest order modes. The main question that had to be addressed was how to extract the electrical signals from the structure. With dual or quad-ridge OMTs a probe is inserted through one of the ridges and terminates on the opposing ridge as shown in Figure 4.3 (a). With the TRT it was not immediately clear as to how the probe would be inserted given that when the probes are placed in the same plane it would not be possible to have mounting threads on the opposing ridge without mechanical interference. Rather than offsetting the probes the implemented solution was to limit the length of the mounting holes of the probes in order to avoid mechanical interference. The probes could then be fixed in place using conductive glue or simply rely on applied pressure. The initial structure is shown in Figure 4.3 (b).

A 3D model of the coaxial to ridged-waveguide transition was developed and simulated using ANSYS HFSS. The final HFSS model is shown in Figure 4.4. These images depict what would be seen by carving out the volume of air enclosed by the waveguide structure where all the dark purple surfaces correspond to the conductive waveguide walls. The image on the left shows an isometric view through a transparent backshort, while the image on the right is the direct view through a transparent

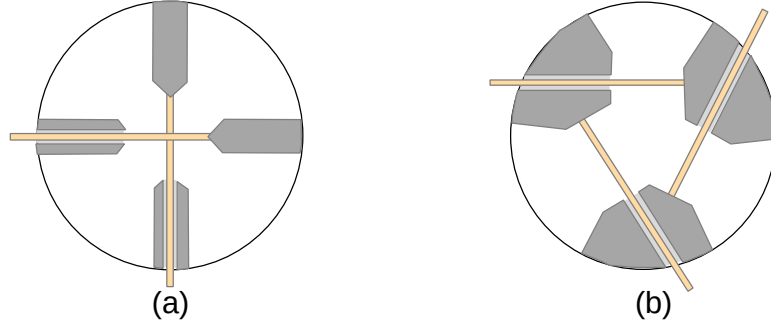


Figure 4.3: Cross section of Coaxial to ridge waveguide transitions, (a) QRWG OMT (b) Proposed TRT

backshort.

Initial design of the geometry involved making sure that the structure could support propagation of the desired modes over the target bandwidth, 4.6 to 15.3 GHz. The lowest cut-off frequency of a waveguide structure determines the lowest frequency that can be supported. The modes of propagation corresponding to the lowest cut-off are called the dominant modes and these are the desired modes of propagation for the TRT. The other modes which have higher cut-off frequencies are undesirable, for the reasons stated in Section 4.2.

Each mode has a frequency dependent propagation constant, which is the complex sum of two variables, the phase constant and the attenuation constant. These correspond to the imaginary and real parts of the propagation constant respectively. For modes that are in cut-off the phase constant is less than zero and the attenuation constant is greater than zero. For modes that are able to propagate at a given frequency if excited the phase constant will be positive whilst the attenuation constant will be small. By simulating the propagation constant for the various modes of interest, which modes can or cannot be supported for the particular frequencies can be determined. By adjusting the geometry the cut-off frequency of the dominant mode was set below 4.6 GHz whilst the cut-off frequency of the first higher order mode was set above 15.3 GHz. The variation of the propagation constant with frequency is given in Figure 4.5. The green traces correspond to the phase constants whilst the attenuation constants are the red traces. The first two modes which are the desired

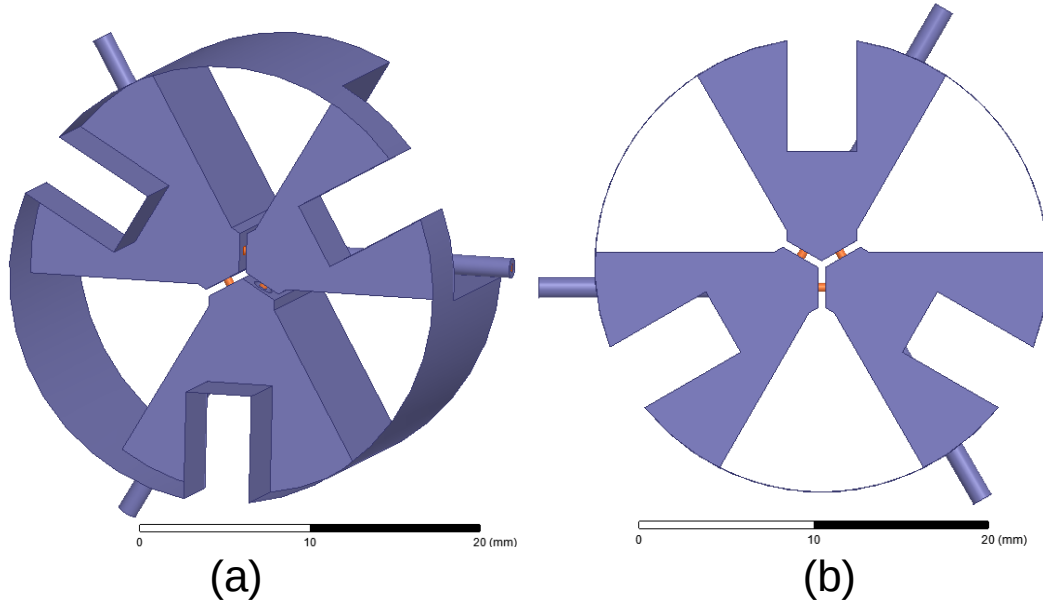


Figure 4.4: TRT coaxial to Ridged waveguide transition HFSS model, (a) Isometric View (b) Back view

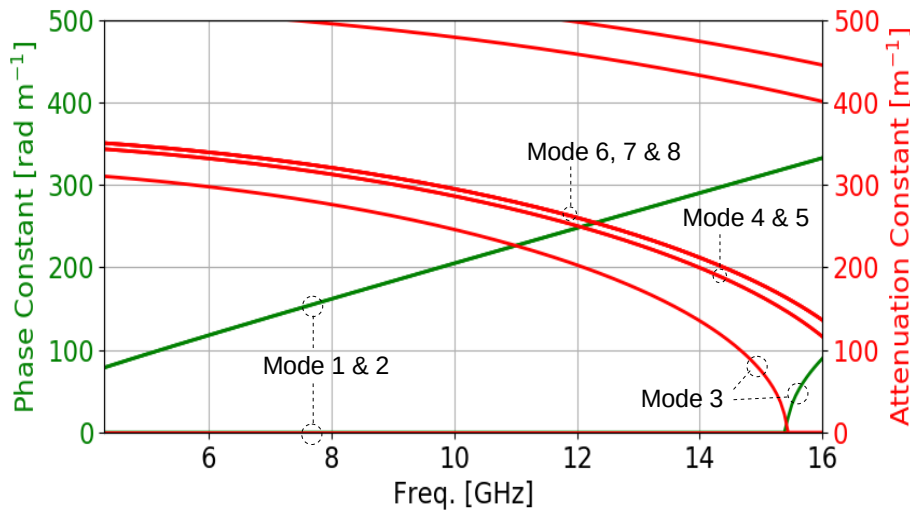


Figure 4.5: Simulation results showing the propagation constants for the coax-ridged waveguide transition. The attenuation constant for modes 1 and 2 are zero hence lie on the x-axis of the graph.

modes start to propagate well below 4.6GHz, as can be seen by the greater than zero phase constant and zero attenuation constant combination. In fact the dominant modes cut in at around 2.75 GHz whilst the next mode cuts in at around 15.7 GHz potentially giving a spurious-free bandwidth of 1 : 5.7. The electric field distribution for the first eight modes are shown in Figure 4.6. The electric field distribution for the two degenerate dominant modes are shown in Mode 1 and 2 images. The fields are concentrated in the gaps between the parallel sections of the ridges. Mode 3 is the first spurious mode. For this mode the electric fields extend from one ridge to another, starting on one and terminating on another. In order to increase the cut-on frequency of this mode, thereby increasing the spurious-mode-free bandwidth, the space between the ridges had to be made as small as possible. This was achieved in two steps. The first was to reduce the waveguide diameter; this had to be done moderately as reducing the waveguide diameter requires a smaller gap between the ridges to ensure the cut-on frequency of the dominant modes is below 4.6 GHz. The second step was to broaden the ridges where they attach to the waveguide wall. This also allowed for the coaxial probes to be inserted in the ridges. Given that all the other modes are also concentrated between the ridges, the latter design choices inevitably pushed the cut-on frequency of these modes further out of the desired band.

Along with ensuring that no spurious mode can propagate over the desired band, the structure had to be designed so that the coaxial probe-to-ridge waveguide transition provided sufficiently good match so that the majority of the electromagnetic radiation that is incident at the waveguide port is fed into the three coaxial ports with minimal reflection. In other words the return loss looking into the structure had to be better than -15 dB across the band for the dominant modes, and the sum of the power in the probes should equal that excited at the input. This is necessary if the instrument is to achieve high sensitivity. The electrical power in each of the coaxial ports is a linear combination of the orthogonal linear polarisations excited at the input. The desired linear polarisation or circular polarisations can be obtained using analogue signal processing or digital signal processing (DSP): this will be covered in greater depth in Section 4.5.2.

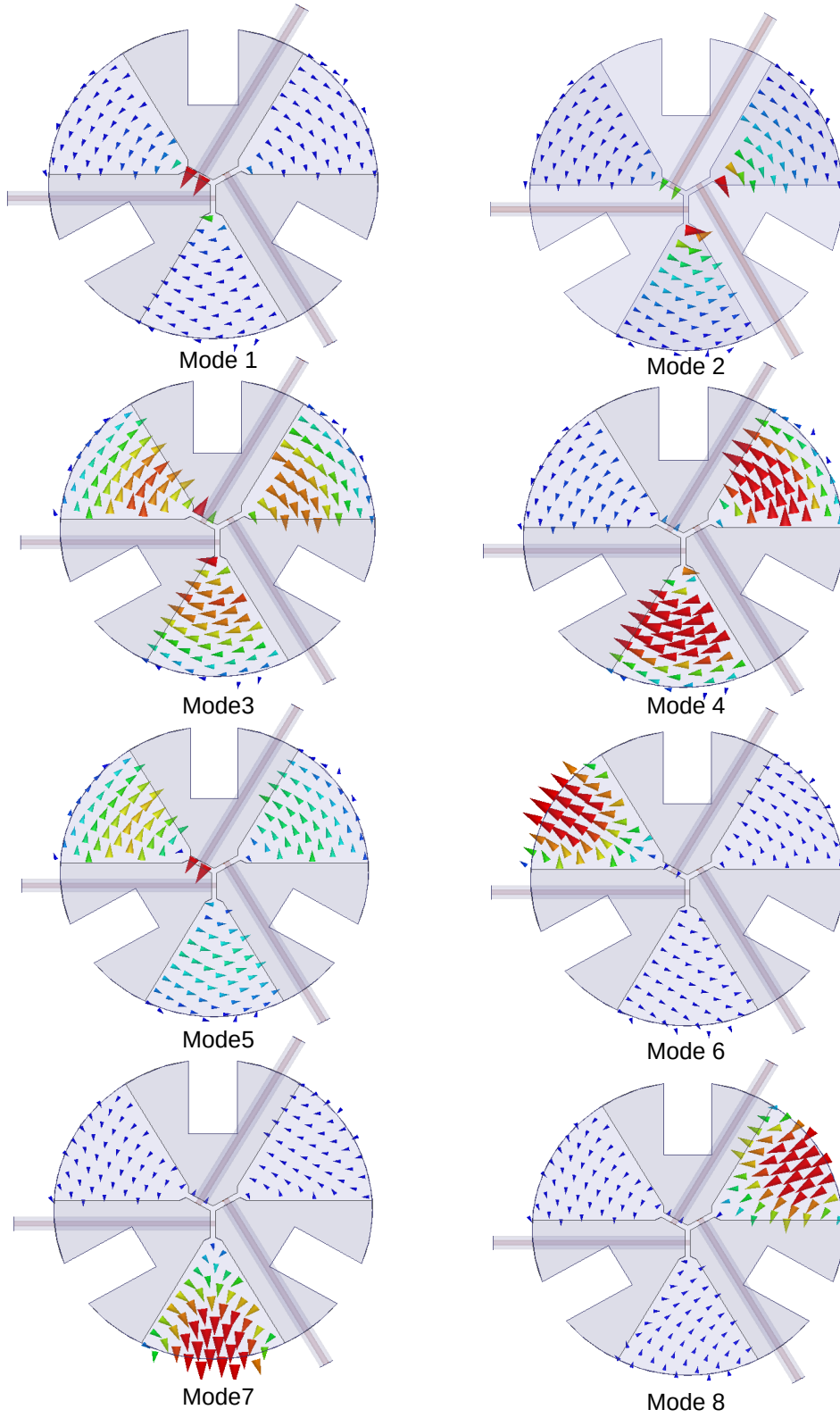


Figure 4.6: Coax-Ridged Waveguide Transition waveguide port propagation modes. Arrow colour is proportional to E field amplitude; ranging from blue to red with red being highest amplitude for that plot.

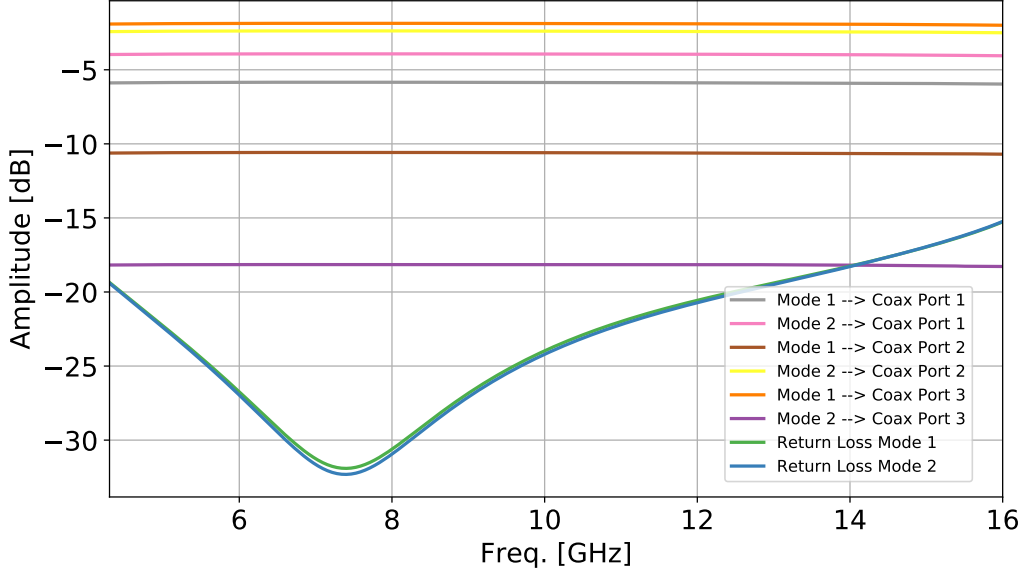


Figure 4.7: Coaxial to ridge-waveguide transition simulated S-parameters.

Figure 4.7 shows the S-parameters. The blue and green traces are the return loss for the two desired degenerate modes. We can see that the target return loss has been met over the desired band. The other traces correspond to the coupling of the excited TE_{11} at the waveguide port to each of the coaxial probes, a flat and broadband performance is predicted by these simulations. It can also be noted that for this section of the TRT the main limiting factor of the bandwidth is the impedance matching as indicated by the return loss. To achieve even broader bandwidth one could consider optimizing the coaxial probe geometry as was done by [Dunning et al. \(2013\)](#).

4.4.2 Ridged-waveguide to circular waveguide transition

The first section of the transducer transitions the circular waveguide to the ridged waveguide. For a purely circular waveguide the characteristic impedance is greater than the 50Ω at the coaxial probes. The ridge transition therefore has to provide adequate impedance matching across the frequency of operation. To achieve this the ridge profile was selected to be of the form $C + A * \sin(B)^2$ and the parameters A,

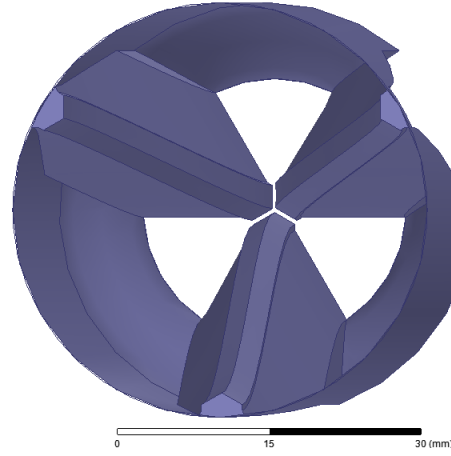


Figure 4.8: Ridged-waveguide to circular-waveguide transition.

B, C were varied to optimize the match. The diameter of the circular waveguide is determined by the low-frequency feedhorn, whilst the output cross-section is determined by the coaxial transition covered in the previous section. The final length of the ridges was determined by the need to achieve the desired return loss at the lower end of the band and by the need to minimise the amount of the higher order Mode 3 being generated. Both of these were satisfied by sufficiently lengthening the ridge without altering the output cross-section. The simulation model of the final transition is shown in Figure 4.8. Once the ridges are sufficiently close to each other the characteristic impedance becomes only weakly dependent on the waveguide diameter. This allows for the waveguide diameter to be reduced in a relatively shorter section to match that of the coaxial to ridge waveguide transition without affecting the return loss.

Figure 4.9 shows the final simulated performance of this section. The simulations assumed that the circular waveguide was excited with orthogonal linear polarisations X and Y. The graph shows return loss for the excited modes X and Y as well as their coupling to each of the simulated 7 modes at the cross-section just before the coaxial probes. Across the band of operation very little of the excited signals are converted to the higher spurious modes. By studying the blue traces we can confirm that very little of the excited signal is reflected from the device. Grey and pink traces further

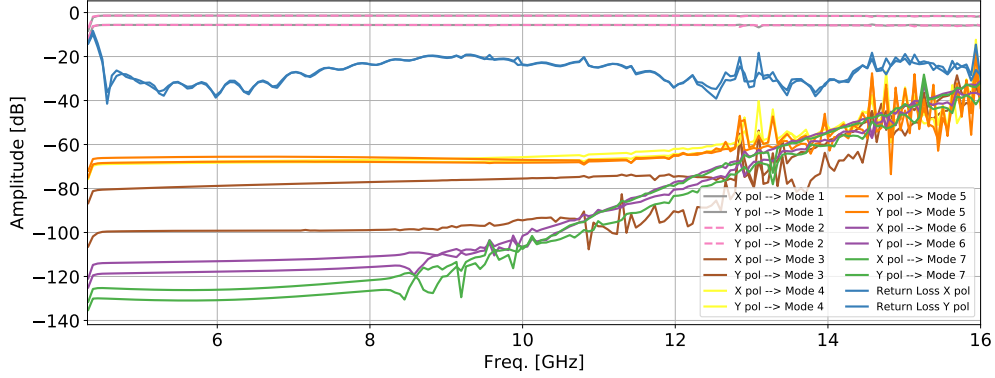


Figure 4.9: Ridged-waveguide to circular-waveguide transition mode coupling.

confirm that the incident signal ends up in the desired Modes 1 and 2 where they will be extracted by the coaxial probes.

4.4.3 Full TRT simulation

Combining the two sections gives the results shown in Figure 4.10. It is evident from the results that the return loss is better than 15 dB across the target bandwidth. Furthermore, we see that each of the modes couple into each of the 3 probes as expected, the degree of coupling depending on the orientation of the mode relative to the probe. By summing all the power from each mode received at each probe we can confirm how much of the incident power has been received. This is shown in Figure 4.11; we see that we can expect greater than 95 percent transmission across the band. For a polarization axes orientation of 45 degrees with respect to the geometry in Figure 4.12, the coupling of the X-polarisation into ports 1, 2 and 3 are 31.6%, 63.1% and 4% respectively. For the Y-polarisation it is 31.6%, 4% and 63.1% respectively (the relationship between the excitation's orientation and the resulting coupling is presented in greater detail in Section 4.4.4). The power transmitted starts to decrease from around 10 GHz. This is not as a result of Ohmic losses as the simulation was conducted under perfect electrical conductor conditions. The loss is therefore a combination of the worsening return loss of the coax to ridge transition section and the conversion of the desired modes to the higher order modes in the waveguide to triple-

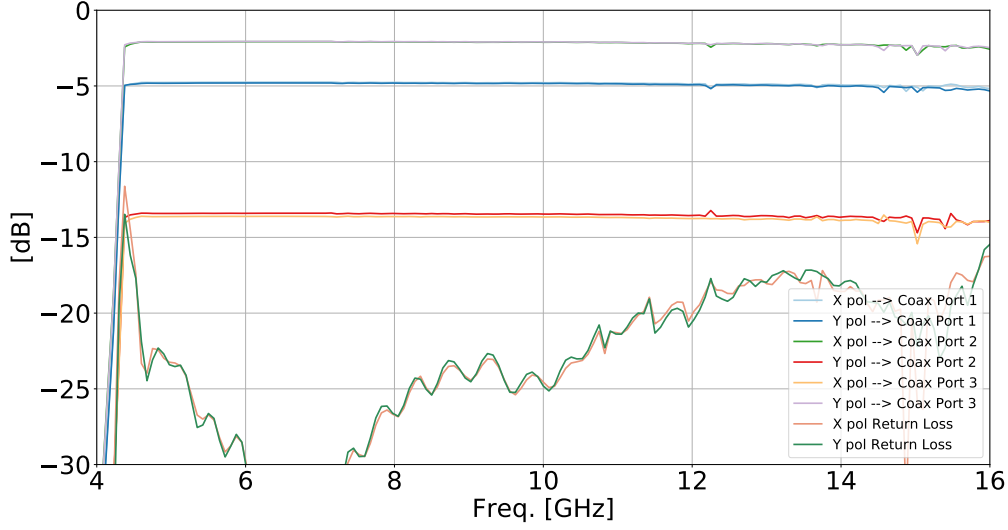


Figure 4.10: TRT waveguide port excitation to coaxial port coupling.

ridge transition. This will not be a problem for our cryogenic receiver. The sharp dips observed at the higher end of the band are artefacts from the simulation, possibly a consequence of not including enough of the higher order modes in the simulation; this is confirmed by the fact that they are not observed in the measurement results.

4.4.4 Waveguide to coaxial port coupling coefficients

With the design now complete we can now look at how excitation in the TRT's waveguide port couples into each of the coaxial ports. As already stated, the output of each coaxial port is a linear combination of each of the TE_{11} modes that are excited at the input waveguide port. Regardless of the orientation of the excited mode, it can be decomposed into two orthogonal modes X and Y. Therefore if the coupling of each of these modes into each of the ports is known, it is possible to determine the orientation or polarisation of the received signal. Assuming that one of the ports is parallel to the x -axis, then the y -axis will lie symmetrically between the other two ports. In this case we assume that port 1 is the port parallel to the x -axis then the y -axis lies between port 2 and 3 as shown in Figure 4.12. Let us define E_{inc} as the

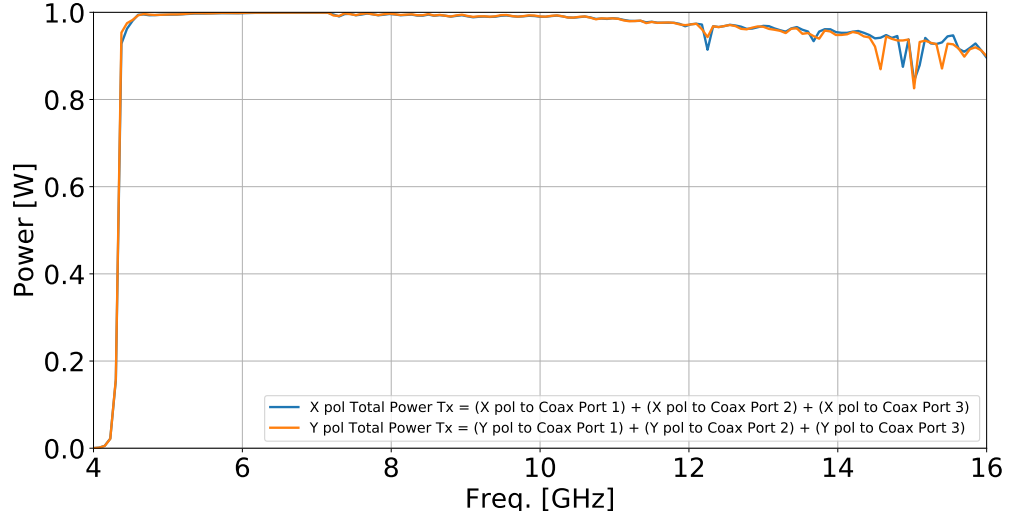


Figure 4.11: TRT simulated insertion loss per polarisation.

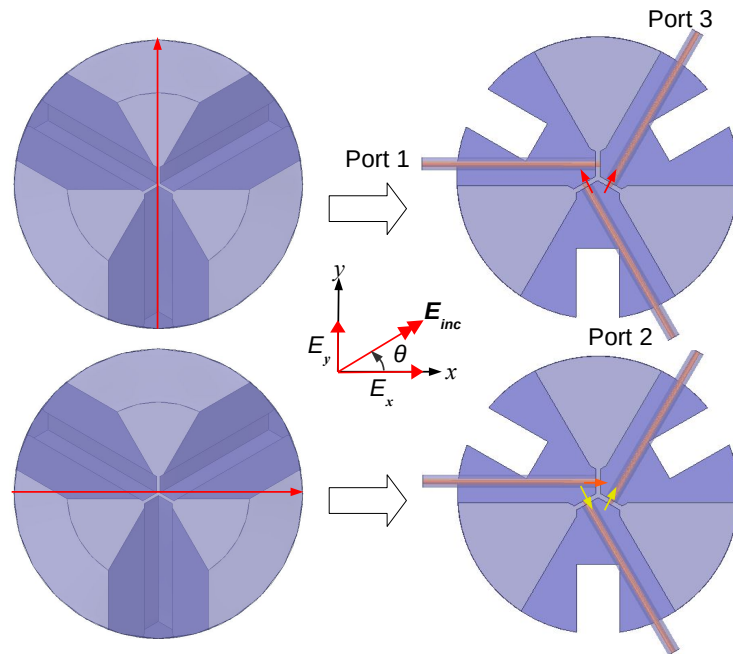


Figure 4.12: Linear waveguide polarisation excitation to coax port coupling.

electric field amplitude incident on the circular waveguide port where

$$E_x = E_{inc} \cos \theta \quad (4.1)$$

$$E_y = E_{inc} \sin \theta \quad (4.2)$$

as depicted in Figure 4.12. Using this Figure expressions can be obtained that predict the coupling of the input signal to each of the coax ports, hence determine the polarisation of the incident signal. Taking port 1 as the reference port then the electric field amplitude incident on the circular waveguide port E_{inc} can be related to the electric fields arriving between the 3 ridges by the following expressions

$$E_1 = E_{inc} \cos \theta \quad (4.3)$$

$$E_2 = E_{inc} \cos(\theta - 120^\circ) \quad (4.4)$$

$$E_3 = E_{inc} \cos(\theta + 120^\circ), \quad (4.5)$$

where E_1 , E_2 and E_3 correspond to the amplitude of the electric fields arriving at the gaps between the ridges for port 1, 2 and 3 respectively. The voltage developed at each port is directly proportional to the electric field amplitude, in the ideal case the constant of proportionality k_{EV} (determined during calibration) is the same for each port therefore the induced voltages are related to the incident Electric field by

$$v_1 = k_{EV} E_{inc} \cos \theta \quad (4.6)$$

$$v_2 = k_{EV} E_{inc} \cos(\theta - 120^\circ) \quad (4.7)$$

$$v_3 = k_{EV} E_{inc} \cos(\theta + 120^\circ), \quad (4.8)$$

Using the trigonometric identity

$$\cos(\alpha \pm \beta) = \cos \alpha \cos \beta \mp \sin \alpha \sin \beta \quad (4.9)$$

to expand v_2 and v_3 we get

$$v_2 = k_{EV} E_{inc} \left(-\frac{1}{2} \cos \theta + \frac{\sqrt{3}}{2} \sin \theta \right) \quad (4.10)$$

$$v_3 = k_{EV} E_{inc} \left(-\frac{1}{2} \cos \theta - \frac{\sqrt{3}}{2} \sin \theta \right). \quad (4.11)$$

We can now write v_1 , v_2 and v_3 in terms of E_x and E_y as follows

$$v_1 = k_{EV} E_x \quad (4.12)$$

$$v_2 = k_{EV} \left(-\frac{1}{2} E_x + \frac{\sqrt{3}}{2} E_y \right) \quad (4.13)$$

$$v_3 = k_{EV} \left(-\frac{1}{2} E_x - \frac{\sqrt{3}}{2} E_y \right). \quad (4.14)$$

In Matrix form we have

$$\begin{bmatrix} v_1 \\ v_2 \\ v_3 \end{bmatrix} = k_{EV} \begin{bmatrix} 1 & 0 \\ -\frac{1}{2} & \frac{\sqrt{3}}{2} \\ -\frac{1}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix} \begin{bmatrix} E_x \\ E_y \end{bmatrix}. \quad (4.15)$$

Therefore it follows that the power coupling from each polarisation to each port when normalised to 1 is given by

$$\begin{bmatrix} p_1 \\ p_2 \\ p_3 \end{bmatrix} = \begin{bmatrix} \frac{2}{3} & 0 \\ \frac{1}{6} & \frac{1}{2} \\ \frac{1}{6} & \frac{1}{2} \end{bmatrix} \begin{bmatrix} P_x \\ P_y \end{bmatrix}. \quad (4.16)$$

Based on the coupling coefficients given in Equation 4.16 it is expected that for an incident signal that is parallel to the designated x -axis we expect to see -1.76 dB of the total power in port 1 and -7.78 dB in each of the other ports, whereas for a signal purely aligned with the y -axis we expect to see no power in port 1 and -3 dB in the other two ports.

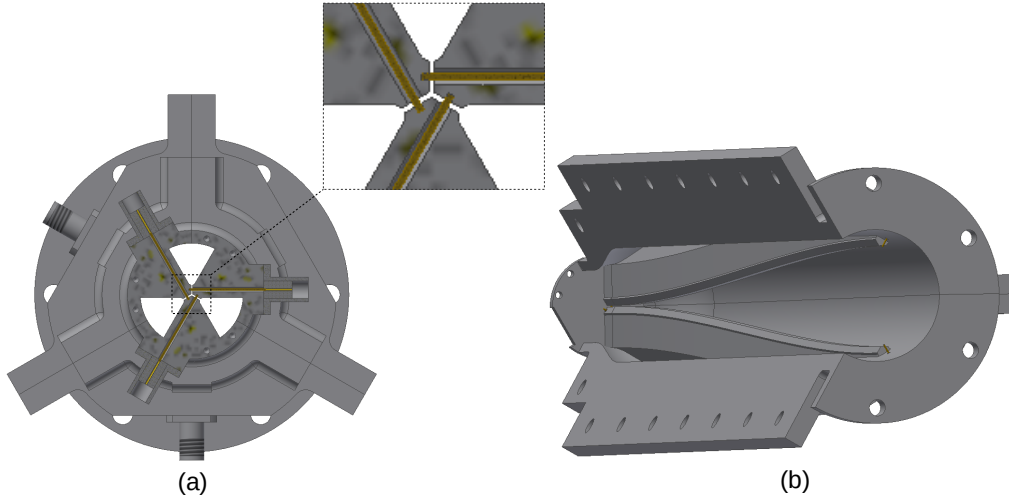


Figure 4.13: Solid Model of the TRT which consists of 3 individual sections: (a) cross-section view at coaxial to ridge transition of the fully assembled TRT (b) isometric view with one of the three sections hidden to show only two of the ridges.

4.5 Prototyping and measurement

To prototype the transducer a solid model was developed as shown in Figure 4.13. Two versions were developed: one with noise injection probes (shown in Figure 4.13) and one without. Initially these were to be produced using 5-axis machining and Electrical Discharge machining (EDM). The 5-axis would do the majority of the work whilst the EDM would be used to cut out the gap between the ridges which required the greater accuracy offered by EDM. Unfortunately the TRT was too long for the EDM machines available in the departments workshop and so only 5-axis machining was used. The TRT is fabricated out of three separate essentially identical sections (the only difference being the noise injection port, it is not included on one of the sections). Each of the sections was machined separately; the gap was machined last by widening it in 100 μm increments and measuring using a microscope between each pass. More details on this are given in Appendix B.

The prototypes are shown in Figure 4.14. The image on the left shows one of the prototypes looking from the rear with the backshort removed, whilst (b) shows the two prototypes side by side.

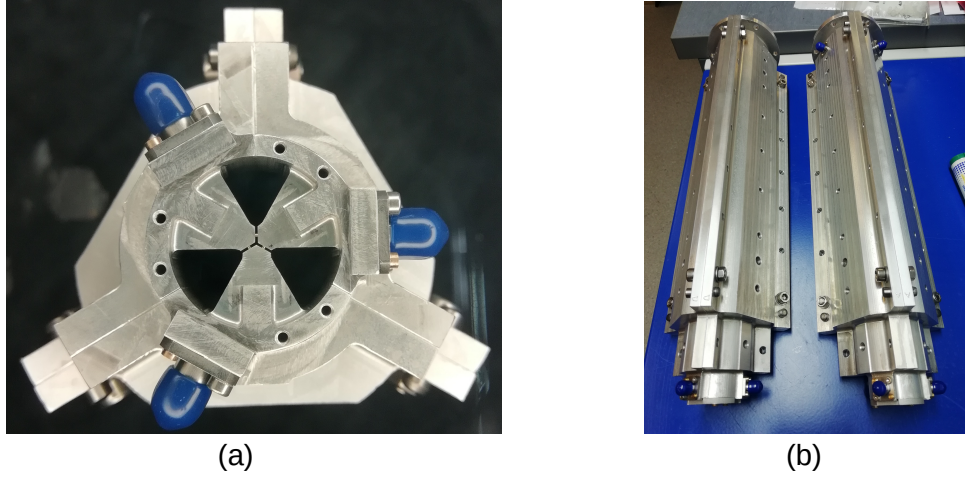


Figure 4.14: Prototype TRTs: (a) rear view with backshort removed (b) TRT side view.

4.5.1 Frequency response

The first measurement involved using a Flann coaxial to waveguide transition and a Vector Network Analyser (VNA) to measure the performance over 4.64 to 7.05 GHz (the model numbers of the Flann transitions are 13094-SF40 and 13644, their specifications are detailed in the Flann catalogue). The results are shown in Figure 4.15. The excitation was a single linear TE_{11} mode hence we only see 3 traces showing the coupling of the signal to each of the 3 coaxial probes. The coupling takes place in the expected proportions just as predicted in the simulation, the main difference being slightly higher loss in the measured data (the simulations assumed perfect electrical conductors). The return loss measurement is limited by the Flann transition. Even so, this serves to illustrate that it lies below the targeted -15 dB over the measurement bandwidth.

In QRWG OMTs the return loss looking into the coaxial port is equal to that looking into the waveguide port, therefore in that case the return loss looking into the waveguide can be obtained by measuring that at the coaxial ports. In the case of the TRT this is not possible, as the return loss is not reciprocal. Nonetheless, provided the measured coaxial port return loss closely matches that of the simulations,

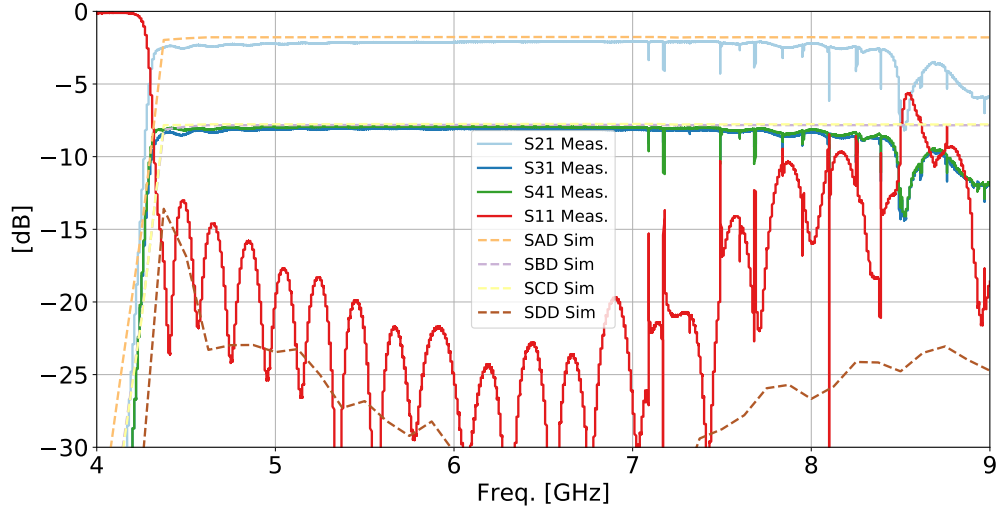


Figure 4.15: TRT S-parameter measurement and simulation results.

then the return loss looking into the waveguide port should also match the design specifications. For this measurement the low-frequency horn was bolted onto the TRT to improve the match to free space and a pyramidal absorber was inserted into the feed to minimise reflections back into the TRT. The results are given in Figure 4.16. The green traces correspond to the cross-coupling between the probes, whilst the red trace is the return loss looking into the coaxial probes. In both cases there is good agreement between the simulated and the measured values. One of the main differences is the sharp peaks between 10 and 12 GHz occurring in the return loss. It is possible due to the narrowness of these features that they were missed in the simulation. The other is the drop in both traces at the top of the band, but this is mostly beyond the 15.3 GHz upper limit of the design bandwidth. The spikes that are observed in the simulation data do not appear in the measured data (which has higher spectral resolution), hence they are artefacts from the simulation, probably as a result of restricting the number of simulated higher-order modes at the waveguide port. Given that the variation between the measured and simulated values across the band are less than 2 dB, the return loss should meet the design specification.

The measurement setup used to obtain Figure 4.15 was limited in that only a

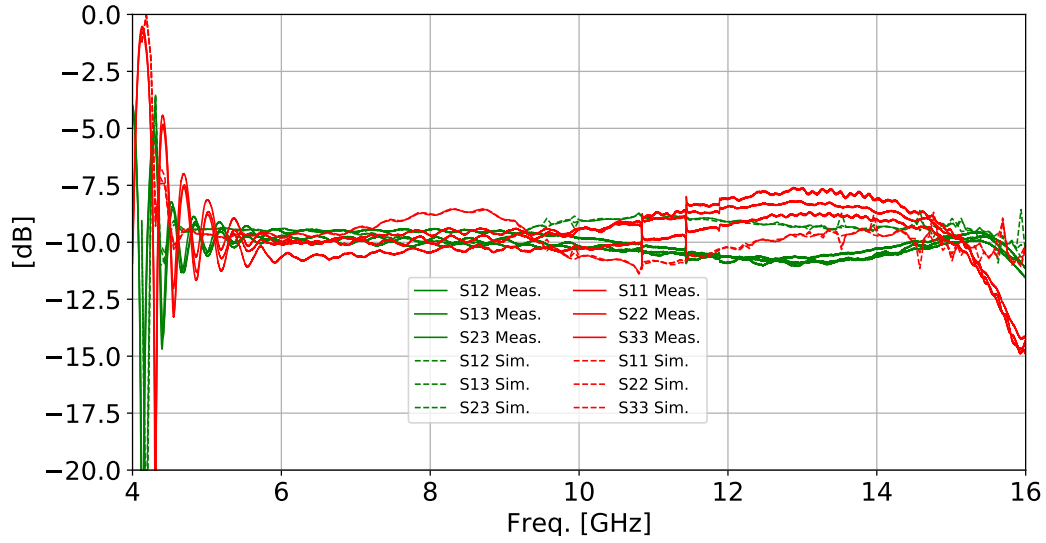


Figure 4.16: TRT Simulated and measured coaxial port cross coupling and return loss

fraction of the band could be measured to determine the insertion loss or the coupling of the input to each of the coaxial probes. Due to the limited bandwidth of the Flann transition, to cover the whole band the best available option was to bolt the two prototypes together and use one of them for the excitation whilst measuring the coupling into each of the ports on the other using the VNA. The problem with this method is that over such a broad bandwidth we are bound to have cavity resonances, and these appear as sharp dips in the insertion loss as shown in Figure 4.17 (a). One way of addressing this is to insert pieces of waveguide with varying lengths between the two TRTs. This has the effect of changing the length of the cavity, hence shifting the frequency of resonance. This allows for data from each measurement to be used to obtain the overall response without the resonance spikes. Results obtained using this technique are given in Figure 4.17 (b). Three different waveguide spacers were available; different combinations of these were used to obtain a total of 8 separate measurements. Given that the majority of resonances appear as destructive interference, resulting in increased insertion loss, the algorithm to generate the graph compared each of the measurement points and plotted the points for each frequency

that had the lowest insertion loss. This assumption is not valid above 15 GHz and at some points at lower frequencies. A more elaborate algorithm could be used to clean this up, nonetheless the current implementation does serve to show that the majority of the narrow features are as a result of the cavity rather than the structure of the TRT and therefore under normal use will not be a problem. The only exceptions are the dips at around 10.8 and 11.4 GHz which appear in both Figure 4.16 and 4.17. These were not observed during the simulation phase due to resolution/duration trade off for such a broadband. Future work can explore how these resonances can be resolved.

The other method of dealing with cavity resonance is to insert a piece of RF absorber in one of the TRTs (the absorber should have the same diameter as the waveguide; ideally shaped like two cones sharing with a shared base for impedance matching). This has the effect of attenuating the reflected energy to the extent that resonance is completely dampened. Although this has the effect of increasing the measured insertion loss as shown in Figure 4.17 (c), this can be corrected for by using the measurement results obtained without the absorber. The other drawback is that unless the absorber has the appropriate shape for matching purposes the return loss measured at the coaxial ports will be inaccurate. This is not a concern given that this measurement is used solely for the purpose of confirming if there is any suck-out resulting from higher-order mode excitation due to the geometry of the TRT. As can be seen, the cavity resonances are dampened, but because the absorber used was not properly shaped (in this case it was simply a 1 cm thick disk) for impedance matching gain ripple is present. At least these results serve as a sanity check for those presented in (b).

The last set of measurements under frequency response involved confirming the coupling of the noise probes to each of the coaxial ports. The protrusion depth of the noise probe into the waveguide was used to set the amount of coupling. The goal was to have it around -30 dB across the band. Both the simulation and measurement results are shown in Figure 4.18. To avoid clutter, traces for only one of the noise probes are shown. The measurement results agree very well with the simulation. This was achieved by trimming the noise probe's protrusion progressively. The final

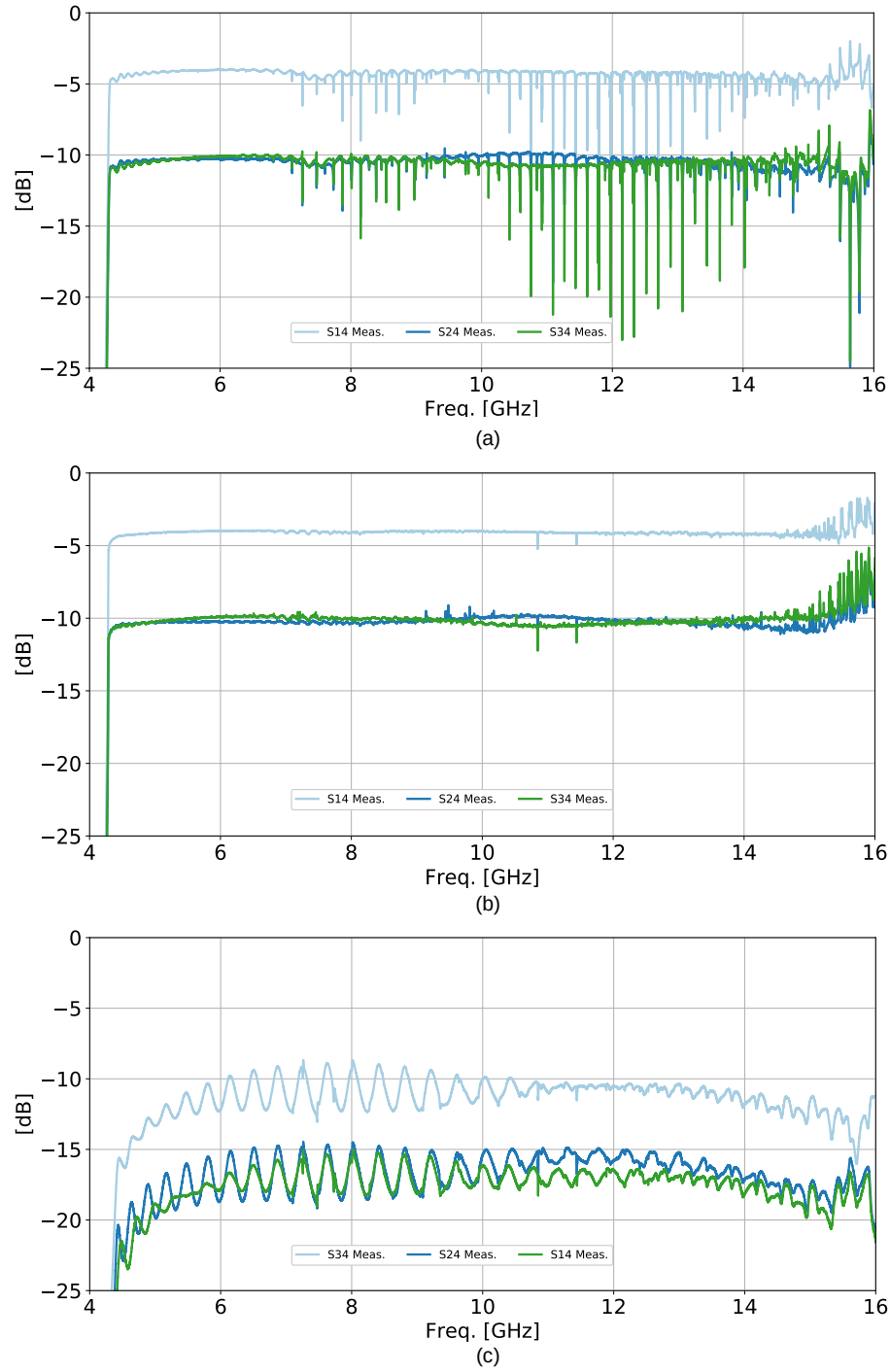


Figure 4.17: Measured Waveguide to coaxial port coupling using TRT for excitation, (a) raw measurement (b) combined results using different spacers between TRTs (c) With Eccosorb Insert.

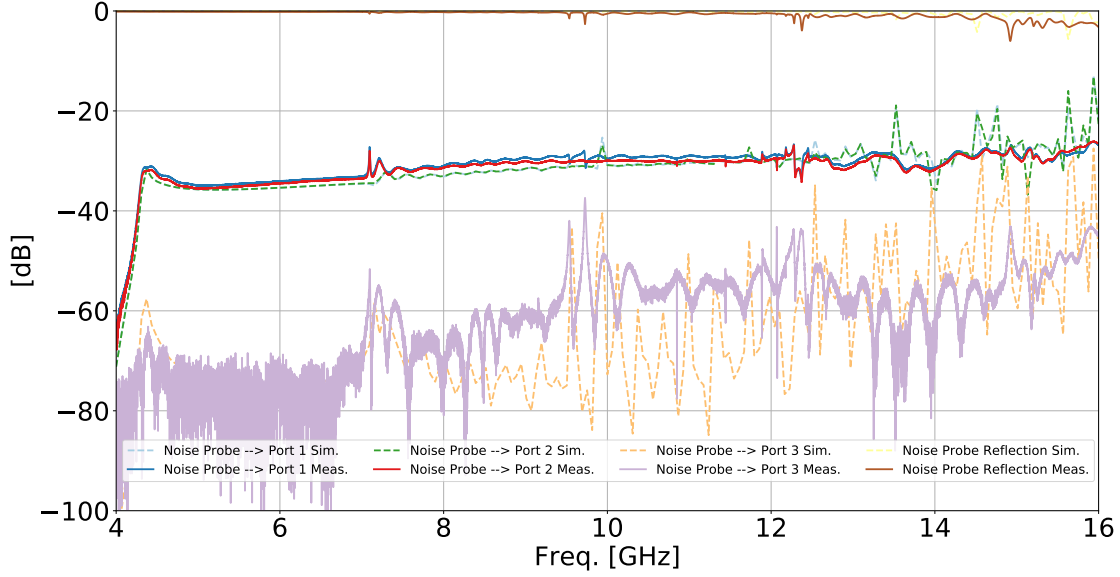


Figure 4.18: TRT noise probe to coax port coupling: measurement and simulation results.

protrusion into the waveguide is 3.3 mm.

4.5.2 Extracting linear polarisations

We will now look at how to determine the polarisation state of E_{inc} using the voltages measured at the three ports. E_x couples into all three ports therefore we have to combine these. To isolate E_x components in v_2 and v_3 we add Equations 4.13 and 4.14. This gives

$$v_2 + v_3 = -k_{EV} E_x \quad (4.17)$$

Subtracting Equation 4.17 from 4.12 yields

$$v_1 - (v_2 + v_3) = 2k_{EV} E_x. \quad (4.18)$$

For E_y all the power is in ports 2 and 3. To isolate it we subtract Equation 4.14 from

4.13.

$$v_2 - v_3 = \sqrt{3}k_{EV}E_y \quad (4.19)$$

Rearranging for E_x and E_y in Equations 4.18 and 4.19 respectively gives

$$E_x = \frac{1}{2k_{EV}}(v_1 - (v_2 + v_3)) \quad (4.20)$$

$$E_y = \frac{1}{\sqrt{3}k_{EV}}(v_2 - v_3). \quad (4.21)$$

To determine the polarisation state we can simply use V_x and V_y , which are given by

$$V_x = \frac{1}{2}(v_1 - (v_2 + v_3)) \quad (4.22)$$

$$V_y = \frac{1}{\sqrt{3}}(v_2 - v_3), \quad (4.23)$$

or in matrix form

$$\begin{bmatrix} V_x \\ V_y \end{bmatrix} = \begin{bmatrix} \frac{1}{2} & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{1}{\sqrt{3}} & -\frac{1}{\sqrt{3}} \end{bmatrix} \begin{bmatrix} v_1 \\ v_2 \\ v_3 \end{bmatrix} \quad (4.24)$$

These expressions assume perfect manufacturing such that the return loss is the same for each probe and the angle between each probe is equally 120 degrees. With these prototypes as would be the case in most situations this is not so. Figure 4.19 shows both the magnitude and phase response obtained using a Network analyser when the TRT is excited with a linear polarised signal that is parallel to port 1 (x -axis).

In Figure 4.19 (a) at 6.5 GHz the insertion loss is 0.39 dB. The coupling into port 1, 2 and 3 is worse than the ideal by 0.24, 0.32 and 1.12 dB respectively. These are acceptable variations and can be corrected for by calibration. For the phase, assuming port 1 is the reference then port 2 has a phase of 173.4 degrees and port 3, 158.7 degrees instead of 180 degrees. Correcting for these in the frequency domain

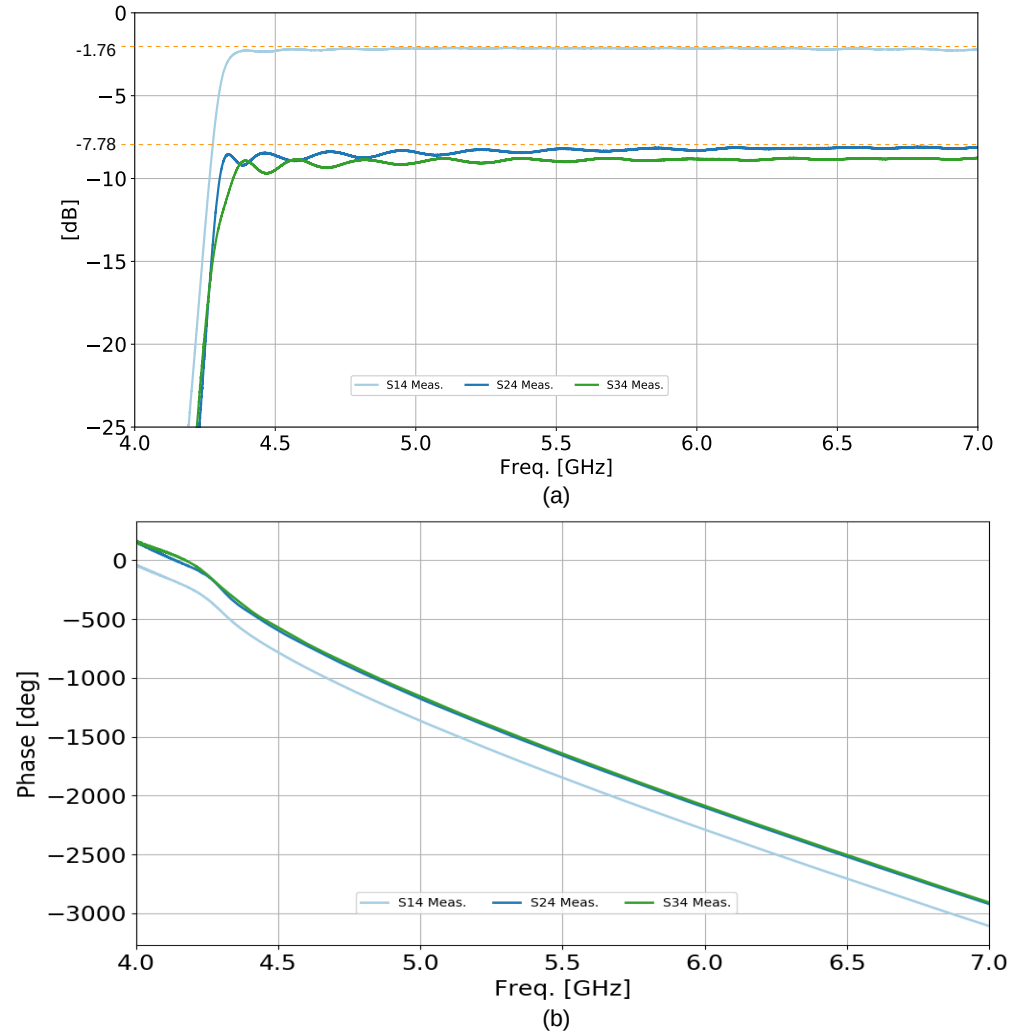


Figure 4.19: Measured TRT frequency response for x -polarised excitation: (a) amplitude response (b) phase response, with the reference plane being the SMA connectors at the end of the cables running from the VNA.

where amplitude and phase are separate is possible.

The excitation was rotated so that it was now perpendicular to port 1, the results obtained are given in Figure 4.20. Again considering values at 6.5 GHz, as expected most of the power is in ports 2 and 3, with insertion loss of 0.55 dB. The difference between ports 2 and 3 is now only 0.06 dB. Given that the coupling into port 1 is less than -20 dB phase consideration for this port is not necessary, what matters is the phase between port 2 and 3. Assuming port 2 is the reference then port 3 has a phase of 158 instead of 180 degrees. Again this can be corrected in the frequency domain.

With the VNA it was not possible to access the actual port voltages as will be the case with the final receiver. Therefore to verify that x and y components of the incident waveform could be recovered from the port voltages, a scaled-down version of the receiver was assembled. The block diagram is given in Figure 4.21. The receiver included some of the RF blocks that will be used in the final system, the only exception being the Reconfigurable Open Architecture (ROACH) board which in this case was used to digitize and read out the IF signal. All the components apart from the mixers were purchased. I developed the mixer chip's circuit board (this is covered in greater detail in Chapter 6) as well as the firmware running on the ROACH. The experiment involved capturing the IF signal for various values of θ , the angle of inclination between the excited TE_{11} mode and port 1 probe of the TRT (as shown in Figure 4.12). The number of sample points captured for each port of the TRT was 4096. These were then read out to a server where they are saved to allow for off-line processing. This was done for three frequencies across the measurable bandwidth.

The following expressions are used to obtain the x and y components of the excitation:

$$V_x(f) = \frac{1}{2} \left(V_1(f) - (\Lambda V_2(f) + \Gamma V_3(f)) \right) \quad (4.25)$$

$$V_y(f) = \frac{1}{\sqrt{3}} \left(\Lambda V_2(f) - \Gamma V_3(f) \right), \quad (4.26)$$

where $V_1(f)$, $V_2(f)$ and $V_3(f)$ are the complex numbers representing the Fourier

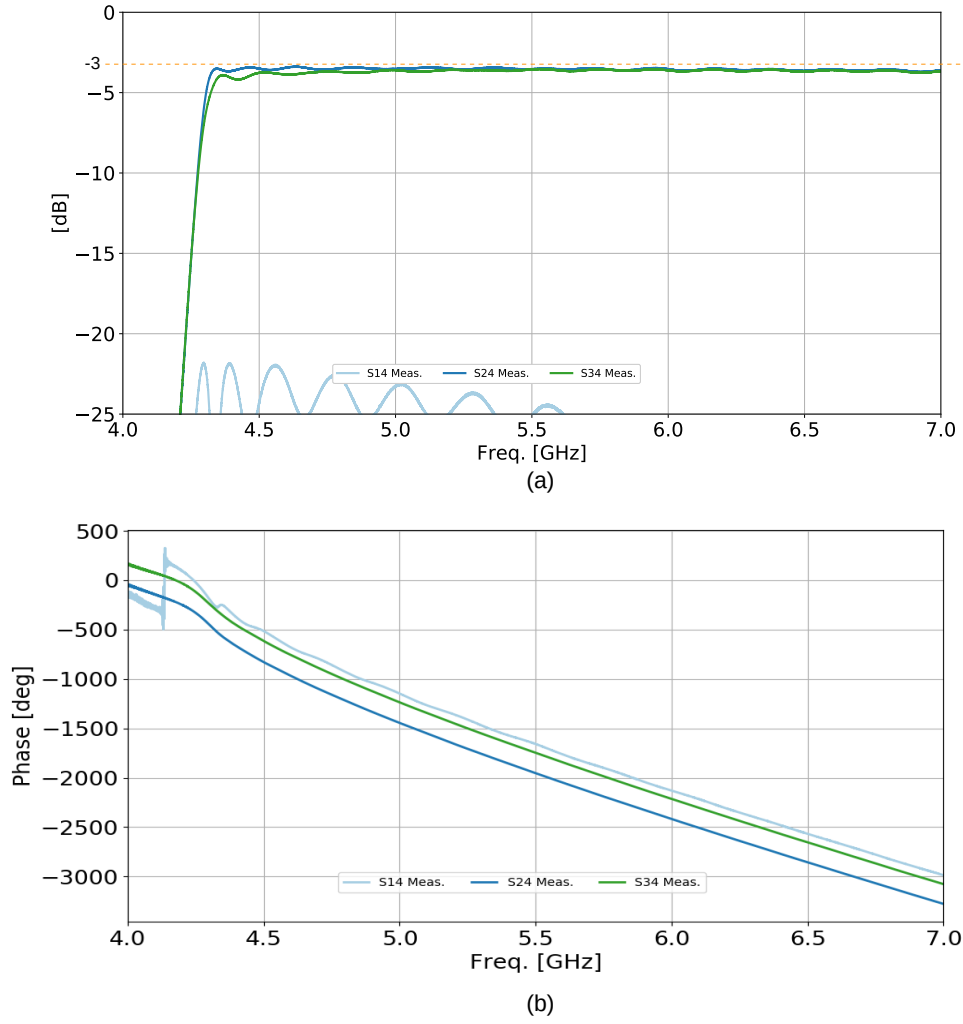


Figure 4.20: Measured TRT frequency response for y -polarised excitation: (a) amplitude response (b) Phase response, with the reference plane being the SMA connectors at the end of the cables running from the VNA.

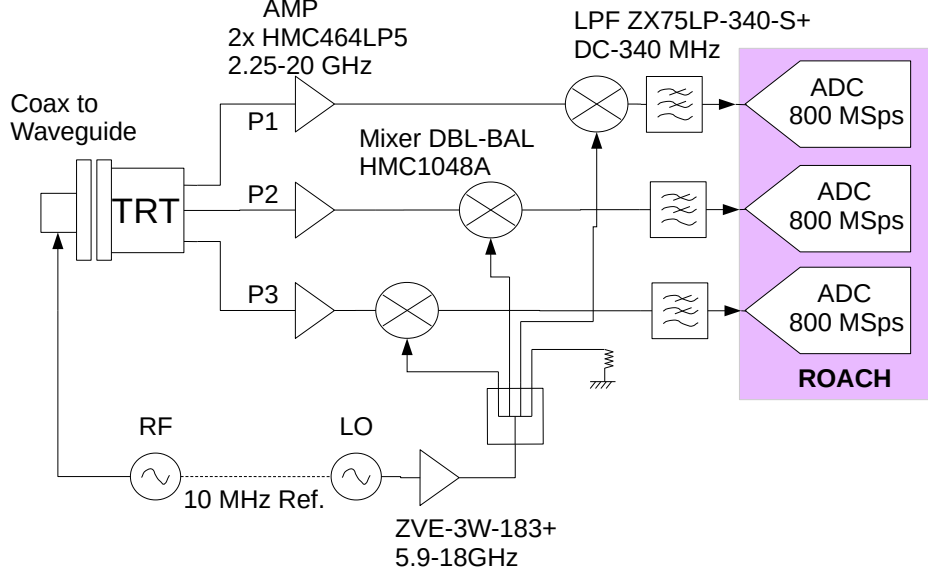


Figure 4.21: Polarisation extraction experiment receiver block diagram.

transform of each of the voltages measured at the TRT ports. The complex phase calibration factors Λ and Γ are applied to port 2 and 3 respectively to correct for the phase difference with respect to port 1. This is obtained by comparing the expected phase difference to that which is measured when $\theta = 0$. $V_x(f)$ and $V_y(f)$ are the synthesised x and y components. θ is then calculated using,

$$\theta = \arctan\left(\frac{\text{Mag}(V_y(f))}{\text{Mag}(V_x(f))}\right). \quad (4.27)$$

Tables 4.1, 4.2 and 4.3 show the results obtained for three different frequencies. Across the 1 GHz bandwidth the results are fairly consistent. These demonstrate that it is possible to extract the x and y components of the received signal from the three probes of the TRT. In the final receiver, the synthesised x and y linear polarisation signals can then be used to compute the Stokes parameters.

Table 4.1: Polarisation measurement RF = 6.02 GHz, LO = 6.00 GHz.

θ [deg]	Measured Value [deg]	error [%]
0	0.06	-
30	30.67	2.23
60	60.77	1.28
90	88.11	2.10

Table 4.2: Polarisation measurement RF = 6.52 GHz, LO = 6.50 GHz.

θ [deg]	Measured Value [deg]	error [%]
0	0.33	-
30	30.24	0.81
60	60.79	1.32
90	89.10	1.00

Table 4.3: Polarisation measurement RF = 7.02 GHz, LO = 7.00 GHz.

θ [deg]	Measured Value [deg]	error [%]
0	0.60	-
30	30.11	0.37
60	60.83	1.38
90	88.70	1.44

4.6 Summary

In this chapter the design and measurement of a triple-ridged waveguide transducer was presented. The transducer's novel structure takes advantage of the benefits of using three-fold symmetry to achieve broadband operation. The measured return loss at the coaxial probes agrees well with the simulations, hence it should meet the requirement of better than -15 dB return loss at the waveguide port across the desired 3.3:1 bandwidth.

The noise injection probes were tuned to produce the appropriate coupling of approximately -30 dB across the whole bandwidth.

Finally the extraction of the two linear orthogonal components of the excitation was demonstrated. The worst error being 2.23% which could be improved by employing a more comprehensive calibration routine and integrating over a number of measurements.

Chapter 5

Gain Calibration

Gain instability of receiver systems tends to limit the maximum achievable sensitivity. As a result, various techniques of stabilising the receiver have been developed. In this chapter a new method of stabilising a receiver will be proposed. The entire RF chain for the AVN receiver was not yet available, so the method was tested on a scaled-down version of the receiver in order to validate it and these results are presented here. The method can easily be integrated into the final receiver, as the firmware needed has already been developed and tested here.

The outline of the chapter is as follows: Section 5.1 will discuss receiver sensitivity and the effects of gain stability. This will be followed by a presentation of the various stabilisation techniques that are currently used in Section 5.2. In Section 5.3 the new gain stabilisation technique will be discussed. The chapter will be concluded with a summary in Section 5.5.

5.1 Receiver sensitivity and stability

5.1.1 Radio receivers and sensitivity limits

There are two main categories of receivers that can be used in the implementation of a single-dish radio telescope. The first is an incoherent receiver, also known as a bolometer. In this type of receiver the phase information of the signal received from

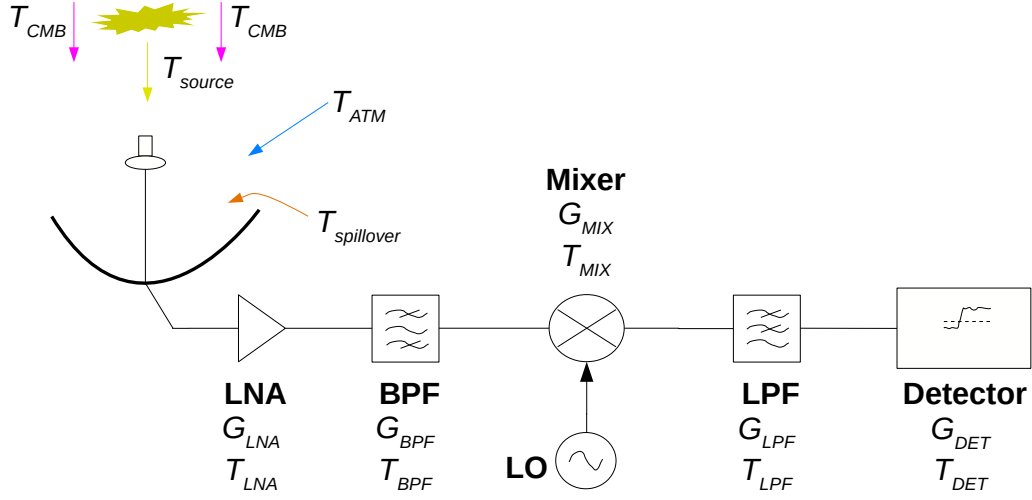


Figure 5.1: Homodyne receiver block diagram.

the astronomical source is not preserved. They tend to be used for the detection of continuum emission and not spectral lines as they are inherently broadband devices (see e.g. [Marr et al. 2015](#)). The second type of receiver is the coherent receiver. For this receiver the phase information is preserved. The receiver's operating bandwidth is shifted from RF to a lower range of frequencies to allow for further processing. This shifting is often accomplished by multiplying or mixing the received signal with a local oscillator (LO) having a frequency within the receiver's bandwidth of operation. The mixing can be accomplished in multiple stages, in which case this is called a heterodyne receiver, but when only one stage of mixing is employed it is known as a homodyne receiver. Figure 5.1 shows a block diagram of a basic homodyne receiver. This corresponds to the receiver architecture used in this project.

The signal detected by the receiver is a sum of all the signals incident at the antenna (antenna temperature T_{ANT}) and the noise contributed by each block in the receiver chain (T_{RX}) as illustrated in Figure 5.1. Each contribution is given in terms of equivalent noise temperature defined as

$$T_{eq} = \frac{P}{k \Delta\nu}, \quad (5.1)$$

where P is the actual power contributed by the signal, k is Boltzmann's constant and

$\Delta\nu$ is the bandwidth. This convention of representing the power of a noise signal in terms of an equivalent temperature is often used in electronics and has also been adopted by astronomers. The system temperature (T_{SYS}) determines the overall sensitivity of the instrument. If we assume atmospheric loss is negligible, $G_{ATM} = 1$, then T_{SYS} is approximately given by

$$T_{SYS} = T_{ANT} + T_{RX} \quad (5.2)$$

$$T_{SYS} = T_{CMB} + T_{source} + T_{ATM} + T_{spillover} + T_{RX}. \quad (5.3)$$

The first term is the cosmic microwave background, which is $T_{CMB} \approx 2.73 \text{ K}$. T_{source} represents radiation from the source and is usually much smaller than the other terms. T_{ATM} is atmospheric emission picked up by the telescope's beam. $T_{spillover}$ accounts for radiation beyond the edge of the reflector. T_{RX} represents the noise generated by the receiver, taking into account the contributions from each component in the receiver chain. It is determined as follows using the noise temperature equivalent of Friis' formula ([Pozar 2012](#)):

$$\begin{aligned} T_{RX} = & T_{pLNA} + T_{LNA}L_{pLNA} + \frac{T_{BPF}L_{pLNA}}{G_{LNA}} + \frac{T_{MIX}L_{pLNA}}{G_{LNA}G_{BPF}} \\ & + \frac{T_{LPF}L_{pLNA}}{G_{LNA}G_{BPF}G_{MIX}} + \frac{T_{DET}L_{pLNA}}{G_{LNA}G_{BPF}G_{MIX}G_{LPF}}. \end{aligned} \quad (5.4)$$

This shows the total noise contributed when each noise contribution is referenced to the input of the receiver. T_{pLNA} and L_{pLNA} account for the effective noise and loss contributions of the feed horn, transducer and other components preceding the low noise amplifier (LNA). An important point to note is that the first two terms dominate the receiver's noise temperature. It is for this reason that in some telescopes the feed horn, transducers and LNAs are cooled to achieve higher sensitivity. Furthermore, LNA gain is in excess of 20 dB to ensure that the noise contributions of the stages that follow have minimal effect on the signal-to-noise ratio (SNR).

In most cases $T_{ANT} \ll T_{RX}$ making it difficult to measure T_{ANT} directly with a

total power measurement. The simplest way of overcoming this is to use switched power measurements (Dicke 1946). Two measurements are carried out, the first being the 'on-source' observation in which $T_{ANT} + T_{RX}$ are measured, the second being the 'off-source' observation; the telescope is pointed at least a beam width away from the source. In this observation T_{RX} and only some of the other terms in T_{ANT} are measured. Subtracting the two sets of results should result in a measurement of T_{source} . In general a more complex scanning strategy that varies the telescopes pointing with respect to the source can be used; the complex large-scale motion is a function of the science goals. This alone is not enough to obtain a sufficiently accurate measurement of T_{ANT} as the result is still dominated by fluctuations in the noise power. It is these fluctuations that limit receiver sensitivity as they dominate the uncertainty in the antenna temperature. Just as with other measurements involving random errors, this uncertainty can be reduced by integrating or averaging a number of measurements. The factor of reduction is $\sqrt{N}$, where N is the number of samples or independent measurements. Assuming Nyquist sampling and an integration time of τ , then $N = 2 \Delta\nu\tau$. The RMS error in each measurement of noise power T_{SYS} is an indication of the fluctuations. Given the statistics of the noise power the RMS error in each sample is given by $\sqrt{2}T_{SYS}$. After integration the RMS fluctuation of the antenna temperature $\sigma(T_{ANT})$ is given by,

$$\sigma(T_{ANT}) = \frac{T_{SYS}}{\sqrt{\Delta\nu\tau}} \quad (5.5)$$

This important result is known as the radiometer equation. It is a measure of the achievable sensitivity for a given bandwidth and integration time for an ideal radiometer. For a detection to occur, the RMS fluctuations after integration should be less than T_{source} . Typically, source power greater than 5 times the noise fluctuation is desired; also known as a 5σ detection.

5.1.2 Radio receiver stability and sensitivity

Equation 5.5 does not take into account all the phenomena that may bring about fluctuations in the noise power and hence limit receiver sensitivity. For example erratic fluctuations in atmospheric emission limits the sensitivity of single-dish continuum observations. Another major source of sensitivity degradation is the stability of the receiver gain. As already mentioned, the power received from astronomical sources is usually small, and consequently a large amount of amplification is needed to enable further processing with the electronics. Receiver gains (G) greater than 80 dB are typically required. Factoring this into Equation 5.1 we get

$$P_{SYS} = Gk\Delta vT_{SYS}. \quad (5.6)$$

As a result of the gain being high, small variations in it appear as fluctuations in the measurement as shown in Equation 5.7 below.

$$P_{SYS} + \Delta P_{SYS} = (G + \Delta G)k\Delta vT_{SYS} \quad (5.7)$$

These variations are indistinguishable from the variations introduced in the system temperature by the astronomical source and thus limit the overall sensitivity:

$$\Delta T_{stability} = \frac{\Delta G}{G}T_{SYS} \quad (5.8)$$

Using Equations 5.8 and 5.5, coupled with the fact that the noise and gain variations are random independent processes, therefore their variances add up, the overall RMS error in the measurement of the system temperature for a practical receiver is given by,

$$\sigma(T_{ANT}) = T_{SYS} \sqrt{\frac{1}{\Delta v\tau} + \left(\frac{\Delta G}{G}\right)^2}. \quad (5.9)$$

The system temperature has a flat power spectrum, where the power spectral density $PSD \propto f^0$, and is referred to as ‘white noise’, with equal power levels per unit loga-

rithmic bandwidth. The spectrum of the gain variation on the other hand exhibits a power law, $PSD = f^\alpha$, (Press 1978). The power increases from high to low frequency. It is more generally known as pink or $1/f$ noise (although this does not imply $\alpha = 1$).

White noise is caused by the thermal agitation of charge carriers, whilst gain variation is a result of various physical processes and factors. One of these factors is atmospheric emissivity; which directly affects T_{ANT} . Atmospheric emissivity is a function of a number of meteorological variables such as ambient temperature and relative humidity. Variations in these properties on similar time-scales as the measurement appear as $1/f$ noise. Power supply and temperature variations can also bring about pink noise. The gain of the amplifier is a function of it's DC biasing and temperature, and changes in either of these results in a corresponding change in gain.

Another source of pink noise in amplifiers stems from the subatomic processes in the semiconductor material used in implementing the amplifiers. When an electric field E is applied across the semiconductor, the charge carriers experience a force, resulting in an electric current with total current density J given by

$$\begin{aligned} J &= \sigma E \\ &= J_e + J_h \\ &= (-N_e \mu_e e + N_h \mu_h e) E, \end{aligned} \tag{5.10}$$

where σ is the effective conductivity, N_e and N_h are the number of free electrons and holes respectively per unit volume, μ_e and μ_h are the electron and hole mobility and e is the absolute charge of a single hole or electron. Assuming E is constant then it follows that for the noise current to exhibit $1/f$ characteristics it must stem from a variation in conductivity of the semiconductor material in the amplifiers. Thus the pink noise is a result of $1/f^\alpha$ variations either in the number of charge carriers $N_{e,h}$, or their mobility $\mu_{e,h}$.

Two theories that explain pink noise in semiconductors have emerged from this, the first considers the variation to be in the number of charge carriers (McWhorter et al. 1955). The pink noise is proposed to arise from the trapping and subsequent release

of charge carriers by surface states. They are named as such because they only exist at the material interface and hence are different from the electronic band structure of the bulk material. The time that the charge carrier spends in the trap before being released is characterised by a time constant τ_{trap} , which is assumed to vary with a distribution of $1/\tau_{trap}$. It is this distribution that is postulated to give rise to noise with a $1/f^\alpha$ PSD. These results were arrived at by considering germanium filaments but can, and have, been applied to metal-oxide field effect transistors (MOSFETs) (Christensson et al. 1968). Pink noise is considered to be generated by the capture and release of the MOSFET's channel charge carriers by traps in the silicon-silicon dioxide interface and the oxide layer. The time constants of the traps are assumed to increase exponentially from the interface; the further into the oxide layer the higher the τ_{trap} . This results in the appropriate distribution to generate a $1/f$ spectrum. This can also be extended to high electron mobility (HEMT) devices that are used in the LNA and second stage amplifiers selected for this project.

The second theory considers that pink noise is generated by variation in the carrier mobility $\mu_{e,h}$ brought about by lattice scattering (Hooge 1994). Atoms of the semiconductor's crystal lattice with temperatures above zero will vibrate creating an acoustic (pressure) wave. This discrete unit of mechanical energy is called a phonon and can be treated as a particle in quantum mechanical terms. Lattice scattering is the interaction of the charge carrier with the phonon; this interaction can be thought of as a collision of the two particles resulting in the scattering of the carrier hence reducing its mobility. It is suggested that the number of phonons fluctuates with $1/f$ spectrum (Hooge & Vandamme 1978) thus giving rise to noise with $1/f^\alpha$ PSD. It is worth noting that it is also through this mechanism that changes in temperature can give rise to pink noise, since an increase or decrease in temperature results in a corresponding change in the number of phonons and consequently a variation in carrier mobility.

The effects of atmospheric emissivity variation can be reduced by using an appropriate scanning strategy; by modulating the desired sky signal to shift it up and away from the low frequency time variation of the atmosphere. Pink noise generated

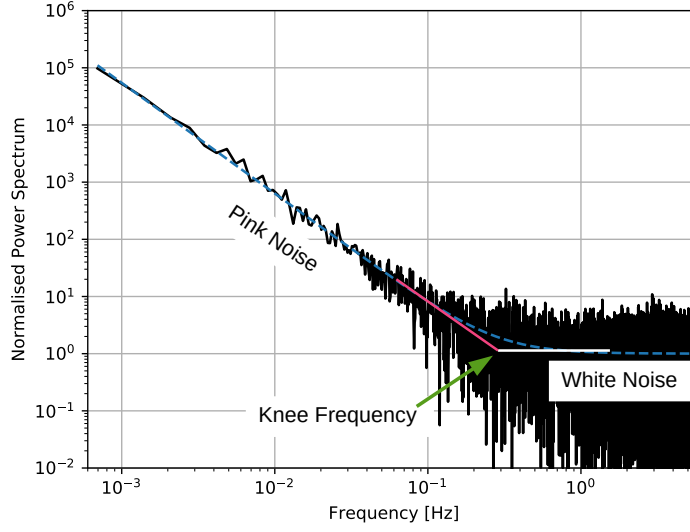


Figure 5.2: Example spectrum showing $1/f$ and white noise.

by power supplies can also be minimised by using a properly regulated power, supply whilst the effects of temperature variations can be reduced by temperature stabilising the RF electronics. However, pink noise will still be present as a result of the imperfections of the latter measures and the other mechanisms of pink noise generation. To address these, other methods of receiver gain stabilisation have to be employed.

Plotting the spectrum of a practical total power receiver will result in a spectrum very similar to that given in Figure 5.2. The pink noise dominates at lower frequencies and decreases with frequency until the white noise component, T_{sys} dominates. The frequency at which the white noise is equal to the pink noise is the knee frequency, f_{KNEE} . The knee frequency is significant as it indicates the maximum integration time possible for a given receiver setup before the pink noise starts to dominate and limit system sensitivity. This is given by $\tau = 1/f_{KNEE}$. For the example given in Figure 5.2, $f_{KNEE} = 0.2$ Hz and hence $\tau = 5$ s. Although f_{KNEE} is often used to characterise $1/f$ plus white noise spectra, it is an unsatisfactory parameter to fit for since it depends on both noise components. A model where the noise power spectrum

is given by Equation 5.11 is more useful (Jew 2017),

$$P_{noise} = \sigma_w^2 + \sigma_r^2 \left(\frac{f}{f_0} \right)^{-\alpha}, \quad (5.11)$$

where σ_w is the white noise level, and σ_r is correlated or $1/f$ noise at the reference frequency f_0 .

To implement a sensitive receiver it is therefore necessary to keep T_{SYS} and ΔG as small as possible. In the case of ΔG this maybe difficult to achieve, therefore a method of correcting for these variations may be implemented. This is discussed in the next section.

5.2 Gain calibration techniques

5.2.1 Dicke switching

In order to measure a source that has $T_{ANT} = 10^{-4}T_{SYS}$, using Equation 5.8, would require $\Delta G/G \ll 10^{-4}$. Hence G would have to be kept constant to within 1 part in 10^4 , which is difficult to achieve. Consequently, this limits direct measurement of received noise power. To overcome this, a differential measurement can be used. One of the earliest implementations of this was proposed by Dicke (1946). Figure 5.3 shows a receiver that makes use of Dicke switching. This refers to the act of switching the receiver input between the sky and a reference load and then subtracting the detected reference power from the sky power. In this way the effects of gain fluctuations can be minimised directly without tracking the gain. The switching frequency has to be sufficiently high to ensure Nyquist sampling of gain variation as well as ensure that the reference and sky measurements experience the same gain. The gain fluctuation resulting from this implementation is given by,

$$\Delta T = \frac{\Delta G}{G} (T_{ANT} - T_{REF}). \quad (5.12)$$

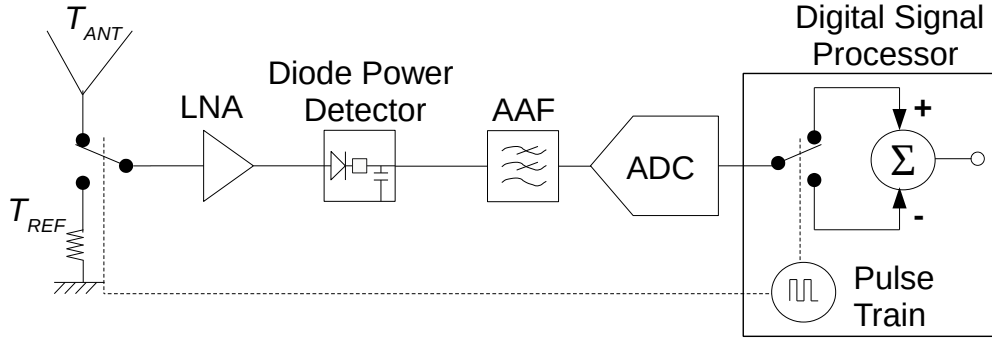


Figure 5.3: Block diagram of a Dicke receiver. The dotted line shows the path of the pulse train generated in the signal processor for switching the receiver input between T_{ANT} and T_{REF} and simultaneously the ADC output between positive and negative for T_{ANT} and T_{REF} respectively. This results in the subtraction of T_{REF} from T_{ANT} in the integrator.

For a balanced receiver where $T_{ANT} = T_{REF}$, the measurement is completely free of the effects of gain variation but in practical receivers, T_{ANT} is not exactly equal to T_{REF} . As a result there is residual variation in the gain, ΔG_{res} . In this case since the gain is not being tracked, ΔG_{res} is defined as the residual variation in the gain after the reference is subtracted and is given by

$$\Delta G_{res} = \Delta G T_{ANT}(1 - \iota), \quad (5.13)$$

where ι is the ratio of T_{REF} to T_{ANT} . Even if the residual gain variation is reduced to zero, half of the observation time is spent on measuring the reference load, resulting in the overall sensitivity being reduced by a factor of $\sqrt{2}$. Furthermore, subtracting two uncorrelated, random noise-like signals results in additional loss of sensitivity by another factor of $\sqrt{2}$. The practical radiometer equation for this receiver is thus given by,

$$\sigma(T_{ANT}) = T_{SYS} \sqrt{\frac{4}{\Delta v \tau} + \left(\frac{\Delta G_{res}}{G}\right)^2}. \quad (5.14)$$

5.2.2 Pseudo correlation

The pseudo-correlation receiver such as the one shown in Figure 5.4 makes use of 3 dB hybrids to combine and subsequently separate a stabilised reference signal with the sky signal (see e.g. [King et al. 2013](#), [Mennella et al. 2003](#)). The output at each of the LNAs is $\frac{1}{\sqrt{2}}G_{LNA1}(v_{ANT} + v_{REF})$ and $\frac{1}{\sqrt{2}}G_{LNA2}(v_{ANT} - v_{REF})$. The second hybrid separates the two signals into v_{H1} and v_{H2} shown below:

$$v_{H1} = \sqrt{2}v_{ANT}(G_{LNA1} + G_{LNA2}) \quad (5.15)$$

$$v_{H2} = \sqrt{2}v_{REF}(G_{LNA1} + G_{LNA2}). \quad (5.16)$$

Given that the reference signal is stabilised, then any change in measured power for this signal is as a result of variation in receiver gain, which is exactly the same for the sky signal. Therefore subtracting this from the antenna temperature should eliminate ΔG or minimise ΔG_{res} . The definition of ΔG_{res} is similar to that of the Dicke switch in that it is proportional to the difference between the antenna and reference noise temperatures. Unlike the Dicke receiver, observation occurs continuously. However, because two uncorrelated random noise like signals are being subtracted the sensitivity is still reduced by a factor of $\sqrt{2}$. The sensitivity for this receiver is given by,

$$\sigma(T_{ANT}) = T_{SYS} \sqrt{\frac{2}{\Delta\nu\tau} + \left(\frac{\Delta G_{res}}{G}\right)^2} \quad (5.17)$$

The disadvantages of this system are that twice the number of components are needed, and the insertion of the lossy hybrid before the low noise amplifiers results in raised system temperature. The Planck low frequency instrument (LFI) ([Davis et al. 2009](#)) and the C-Band All-Sky Survey (C-BASS) instruments ([King et al. 2013](#), [Copley 2014](#)) both employ this architecture for each polarisation. Unlike the Planck LFI and C-BASS north, where both entangling and disentangling of the reference and sky voltages is done using hybrids, C-BASS south uses digital signal processing to separate the sky and references signals.

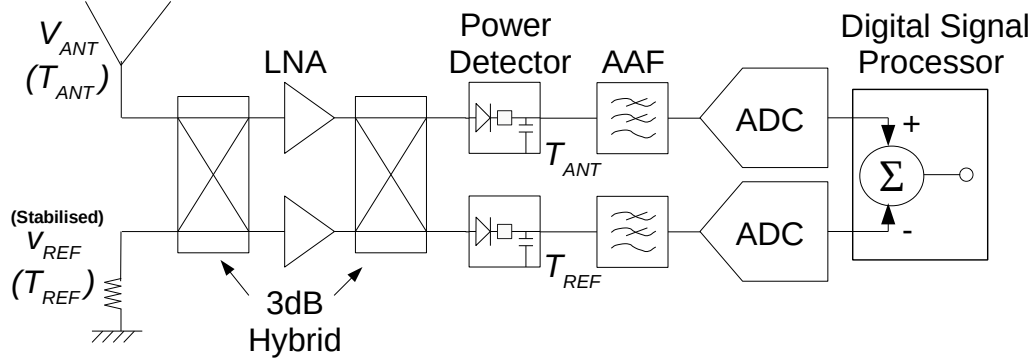


Figure 5.4: Block diagram of a pseudo-correlation receiver. The first 3 dB hybrid is used to generate the sum and difference of the two input signals to allow for each of the input signals to experience the same gain variations. Once separated by the 2nd hybrid each of the signals can be digitised by the ADC and the stabilised reference can then be used to correct for the gain variation by the signal processor.

5.2.3 White-noise stabilisation

The white-noise stabilised receiver is extensively described in (Pollak 2018). A variation of this is shown in Figure 5.5. Unlike the previous methods, this technique relies on tracking the gain. A stabilised white-noise signal is coupled into the RF chain. The duty cycle of the noise injection is controlled by the back end. This allows for two measurements to be made. By comparing these two measurements and making use of the known value of v_{REF} , variation in G is determined. At worst the duty cycle is 50%, giving similar performance to that of the Dicke receiver. If the dynamic range of the receiver is sufficiently high then a high value of T_{REF} can be used. Therefore the receiver sensitivity is given by

$$\sigma(T_{ANT}) = T_{SYS} \sqrt{\frac{K}{\Delta v \tau} + \left(\frac{\Delta G_{res}}{G} \right)^2}, \quad (5.18)$$

where the sensitivity reduction factor K is $1 < K \leq 4$ depending on the duty cycle, with a 50% duty cycle yielding $K = 4$, but less for smaller duty cycle (Pollak et al. 2019). Here ΔG_{res} is the uncorrected gain fluctuations due to errors in the gain

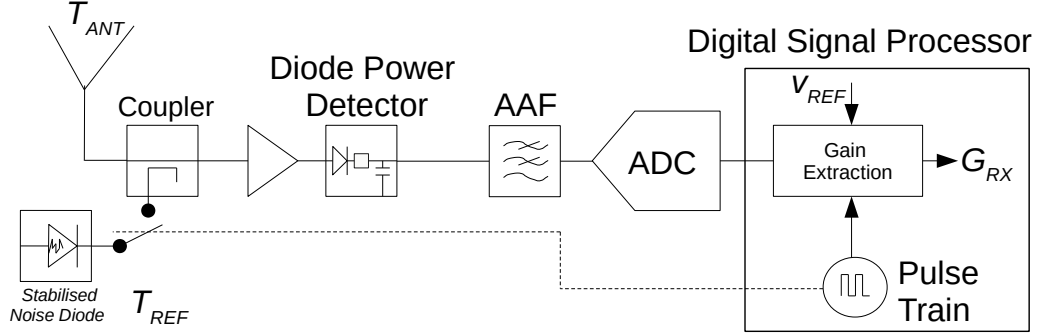


Figure 5.5: Block diagram of a white noise stabilised receiver. The noise source is switched in and out of circuit allowing for the backend to make two separate measurements of system temperature, with or without the noise source coupled into the RF chain. Given that v_{REF} is known, the receiver gain can be tracked and variations can be corrected for.

estimation. A relatively high dynamic range is required especially for lower duty cycles as more reference signal power is required to compensate for the shorter integration time available. To relax this requirement higher duty cycle can be used.

5.2.4 Continuous-wave stabilisation

For this stabilisation technique a continuous-wave (CW) signal is coupled into the RF chain as early as possible, as shown in Figure 5.6. The most recent work on this was carried out by (Pollak 2018). It was demonstrated that by injecting a CW signal that occupies only one of the frequency channels of the spectrometer, one could monitor the power of the CW in that single channel to measure the gain instabilities. The advantage of this technique is that the reference signal only occupies one channel, thereby allowing observation to take place continuously over the rest of band. The sensitivity is given by

$$\sigma(T_{ANT}) = T_{SYS} \sqrt{\frac{1 + \epsilon}{\Delta v \tau} + \left(\frac{\Delta G_{res}}{G} \right)^2}, \quad (5.19)$$

where ϵ is the fraction of the total bandwidth dedicated to monitoring the CW signal and the other terms are as previously stated. Since this also relies on gain tracking,

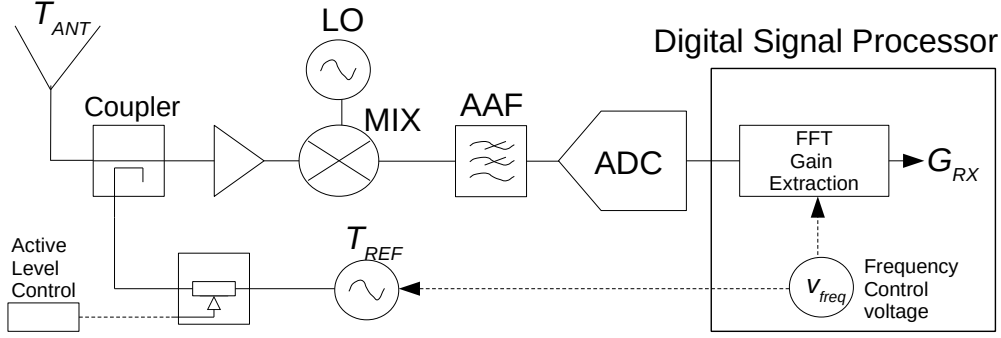


Figure 5.6: Block diagram of a continuous-wave stabilised receiver. A stabilised CW signal is continuously coupled into the RF chain. The digital backend performs a Fourier Transform on the received signal to allow for the CW signal’s amplitude to be monitored by observing the corresponding bin. Variations in this amplitude correspond to gain variation, which can then be used to corrected the other bins.

ΔG_{res} in this case is also due to the uncorrected gain fluctuations as a result of errors in the gain estimate. To achieve good signal-to-noise ratio (SNR) the reference signal power level does not have to be as high since it occupies only one channel. As a result it does not require higher ADC dynamic range. However, high channel isolation is required in the spectrometer implementation. This is achievable by using a polyphase filter bank as part of the signal processing; isolation better than 60 dB between adjacent channels and 80 dB for more widely spaced channels is possible (Pollak et al. 2019). In practice it was found that the thermal noise in one channel did not limit the accuracy of the gain measurement. Instead, the inherent instability of the CW signal source was the limiting factor. To minimise this, active level control can be employed as shown in the Figure 5.6.

5.2.5 Summary of stabilisation techniques

The pseudo-correlation receiver stabilisation technique given in Section 5.2 improves on the Dicke switch receiver’s sensitivity but at the expense of hardware complexity and cost. The white-noise stabilised receiver improves on the Dicke switch’s sensitivity with less hardware complexity but, requires high dynamic range to maximise the sensitivity improvement. The CW-stabilised receiver overcomes this hurdle and

provides even better sensitivity performance than the Dicke switch, at the expense of a more complicated backend processing system. In the following section we discuss an alternative receiver stabilisation technique which improves upon those described here.

5.3 Noise correlation gain calibration technique

5.3.1 Principle of operation

In the rest of this chapter an alternative receiver stabilisation technique is proposed, the noise-correlation gain stabilisation method. It can provide the same receiver sensitivity as the CW technique, but without the loss of any frequency channels and also with no high dynamic range ADC requirement. Furthermore, this method enables the measurement of complex gain, hence both amplitude and phase variations can be corrected. The basic principle is that a white-noise signal is injected as a reference into the receiver chain. The output of the receiver is then correlated with the original reference signal using an XF or FX correlator (for an XF correlator the correlation is carried out in the time domain after which the spectrum is calculated, whereas for the FX variant, the spectrum is computed first). Any variation observed in the cross-correlation is proportional to the gain instability, therefore one can either determine the actual gain or simply track its variation by monitoring the result of the cross-correlation. For both XF (with sufficient number of lags) and FX implementations it is possible to extract the phase variations across broad bandwidths and apply corrections for each frequency channel independently.

Two variants of this system are shown in Figure 5.7 and Figure 5.8. They both rely on injecting a broadband noise signal into the receiver frontend just as with the CW method. The first version uses a noise diode to provide the white-noise reference signal. The signal is split into two. One arm is attenuated (to control the power level relative to the system temperature) and is injected into the signal path in front of the LNA. The other arm is down converted using the same LO as the main RF chain,

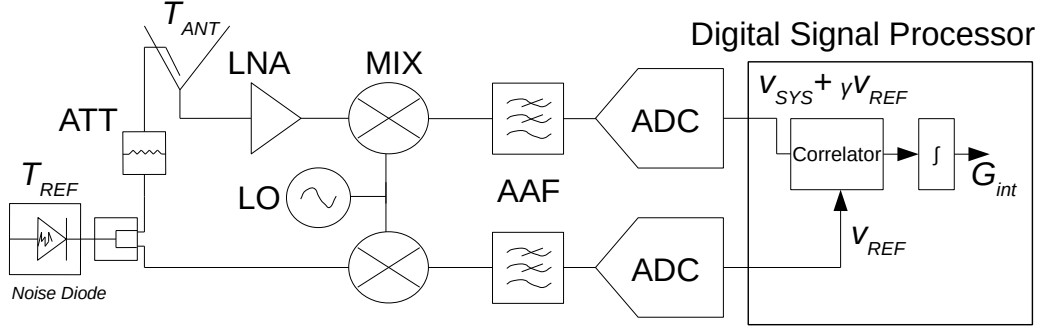


Figure 5.7: Block diagram of noise-correlation stabilised receiver using noise diode. Signal from a stabilised noise diode is simultaneously digitised and weakly coupled into the RF chain such that $T_{REF} \ll T_{SYS}$. The digital backend tracks the gain by correlating these two signals.

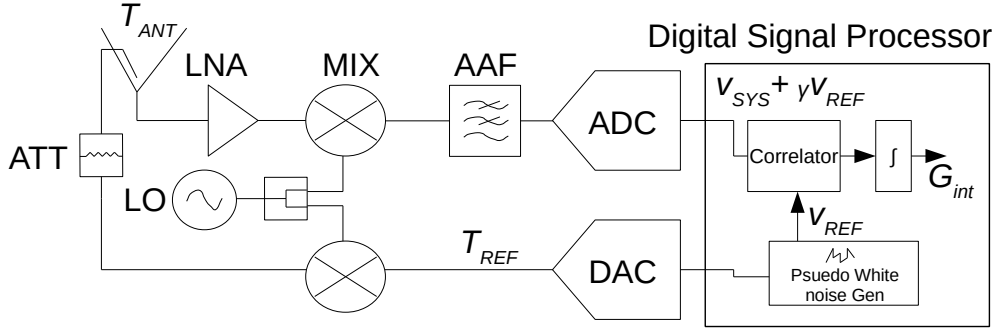


Figure 5.8: Block diagram of noise-correlation stabilised receiver using pseudo Noise Generation. Signal from a psuedo-noise generator in the digital backend is fed through the DAC and weakly coupled into the RF chain such that $T_{REF} \ll T_{SYS}$. The digital backend tracks the gain by correlating the digitised receiver voltage and the internally available pseudo-noise signal.

so that the same band of the noise diode's output is selected, and digitised by the ADC to allow for correlation to proceed in the DSP. The second receiver of this kind, shown in Figure 5.8, makes use of a digital-to-analogue converter (DAC) to generate the noise-like signal. The DAC generates the noise from a predetermined sample sequence or from a noise-generating algorithm. In this way the reference signal is available in the DSP and can thus be used in the correlation. The hardware for this version is simpler than that required for the noise diode version, especially for the AVN receiver whose DSP will be based on an RFSoc which has on-chip DACs.

As mentioned above, before the reference is coupled into the RF chain it is attenuated. This enables the control of the reference signal's power-level relative to the system temperature. Assuming no gain instability in the system, consider the case where T_{REF} is equal to T_{SYS} . The sensitivity would be given by,

$$\sigma(T_{ANT}) = \frac{2T_{SYS}}{\sqrt{\Delta\nu\tau}}. \quad (5.20)$$

Which is the same as the Dicke switch receiver with perfect instability correction. A very large correlation gain, allows for $T_{REF} \ll T_{SYS}$, thus enabling even better system sensitivity.

5.3.2 Noise correlation gain measurement SNR

A question still remains, how accurate is the gain measurement given that the reference power level is much lower than the system noise? The answer is that correlation followed by integration allows for very good signal-to-noise ratio for the gain measurement. The correlation takes the reference power in the entire system bandwidth and concentrates it in one of the lags of the correlator. Accumulating over N_c ($\tau_c\Delta\nu/2$) samples further beats down the noise by a factor of $1/\sqrt{N_c}$. This correlator integration time, τ_c , not to be confused with the observation signal integration time τ , also determines the sampling rate of the gain variation. Therefore, in order to satisfy Nyquist's criterion, the maximum integration time should not exceed $1/2f_{KNEE}$.

Assuming the peak of the cross-correlation is located in the zeroth lag, let v_o be the output of the zeroth lag every clock cycle, then Equation 5.22 expresses the output of the integrator for the zeroth lag after a single integration cycle.

$$v_o = \sum_{i=0}^{N_c} G_{int} (\gamma v_{REFi} + v_{SYSi}) v_{REFi} \quad (5.21)$$

$$= G_{int}\gamma \sum_{i=0}^{N_c} v_{REFi}^2 + G_{int} \sum_{i=0}^{N_c} v_{SYSi} v_{REFi} \quad (5.22)$$

$$= v_a + v_b \quad (5.23)$$

where G_{int} is the receiver gain over one integration cycle, which varies over much longer time scales than the integration time τ_c and, hence can be considered constant over a particular integration cycle. v_{REF} and v_{SYS} are the voltages corresponding to T_{SYS} and T_{REF} . γ accounts for the attenuation and any other stationary effects the receiver chain may have on v_{REF} . N_c is the number of integrated samples. The two variables v_a and v_b have been introduced for convenience to allow the two components of v_o to be considered separately. Starting with v_a we have

$$v_a = G_{int}\gamma \sum_{i=0}^{N_c} v_{REFi}^2. \quad (5.24)$$

The mean power of v_{REF} over an integration cycle is given by the variance as follows,

$$\sigma_{REF}^2 = \sum_{i=0}^{N_c} \frac{(v_{REFi} - \bar{v}_{REF})^2}{N_c - 1}, \quad (5.25)$$

where σ_{REF} is the standard deviation and root mean square (RMS) voltage of the reference. Rearranging and noting that the mean is zero gives,

$$(N_c - 1)\sigma_{REF}^2 = \sum_{i=0}^{N_c} v_{REFi}^2. \quad (5.26)$$

Using Equation 5.26 to substitute in Equation 5.24 yields,

$$v_a = (N_c - 1)\gamma G_{int}\sigma_{REF}^2. \quad (5.27)$$

Looking at Equation 5.27 it can be appreciated that this process is actually computing an estimate of the variance of the reference signal which is a Gaussian random variable with zero mean. The error on this estimate or measurement is the standard deviation divided by the square root of the number of independent samples. The complete expression is thus given by,

$$v_a = (N_c - 1)\gamma G_{int} \left(\sigma_{REF}^2 \pm \sigma_{REF}^2 \sqrt{\frac{2}{N_c}} \right). \quad (5.28)$$

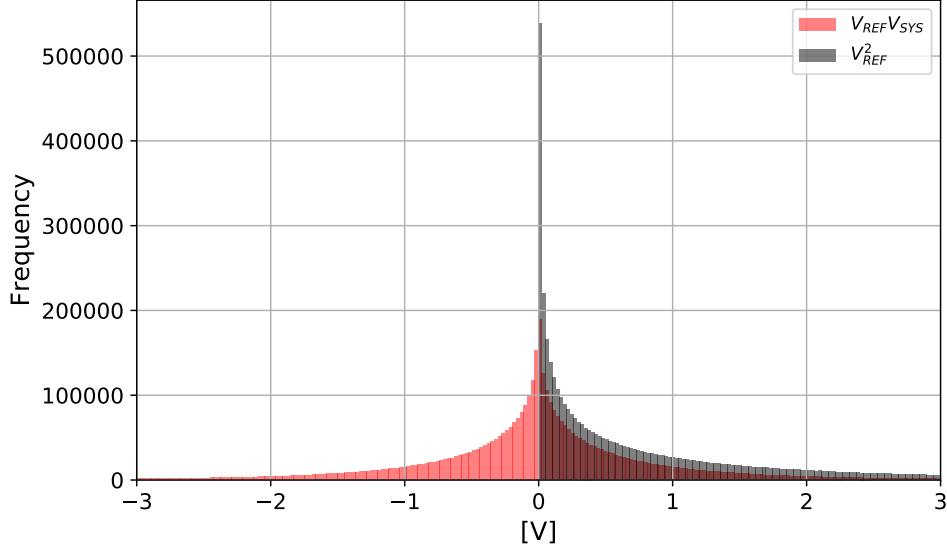


Figure 5.9: Histogram for 2^{22} points of V_{REF}^2 and $V_{SYS}V_{REF}$ in 1000 bins. The RMS voltages in this case are, $\sigma_{REF} = \sigma_{SYS} = 1$ V.

Moving on to the second term of v_o , we start with

$$v_b = G_{int} \sum_{i=0}^{N_c} v_{REFi} v_{SYSi}. \quad (5.29)$$

Equation 5.29 can be re-written in terms of the mean μ_{SR} , and standard deviation of the product of V_{SYS} and V_{REF} as given by Equation 5.30.

$$v_b = G_{int}(N_c - 1) \left(\mu_{SR} \pm \sigma_{SYS} \sigma_{REF} \sqrt{\frac{1}{N_c}} \right), \quad (5.30)$$

where σ_{SYS} is the RMS voltage of the system noise.

If we plot the histograms for V_{REF}^2 and $V_{SYS}V_{REF}$ for a sufficiently large N_c and observe their distributions, we can draw some useful conclusions to simplify Equations 5.28 and 5.30 under certain conditions. Figure 5.9 is an example of such a plot. For this plot the RMS voltages were assumed to be 1 V for both the reference and system noise. As expected, all values for V_{REF}^2 are positive, summing them gives the variance of the reference signal scaled by N_c . On the other hand $V_{SYS}V_{REF}$ appears to have a

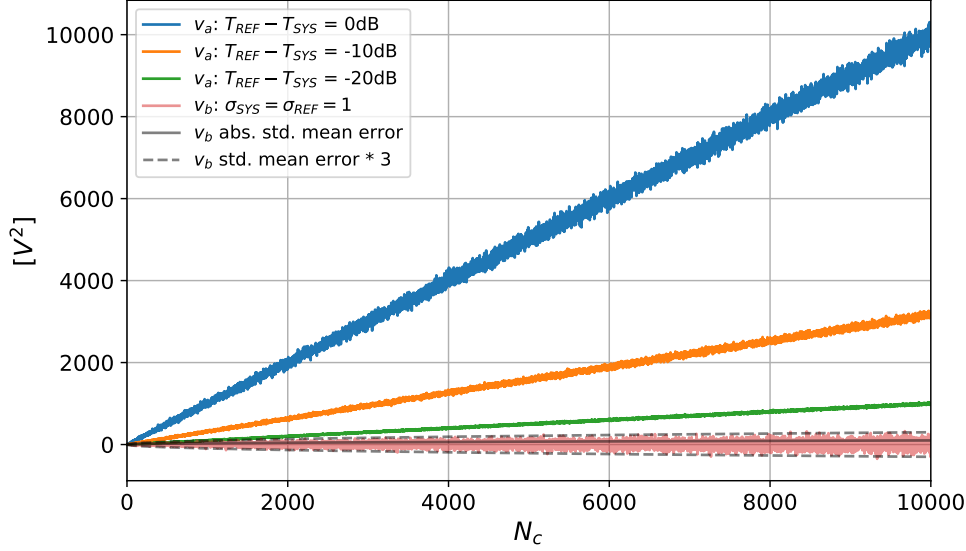


Figure 5.10: Plots of v_a and v_b for $0 < N_c \leq 10^4$. Gain is constant; $G_{int} = 1$. σ_{REF} and σ_{SYS} are equal to 1 V. $T_{REF} = K\gamma^2 v_{REF}^2$ and $T_{SYS} = K v_{SYS}^2$, where K is the conductance and is equal to 0.02Ω for a 50Ω system. γ is varied to achieve $T_{REF}/T_{SYS} = 1, 0.1$ and 0.01

symmetrical distribution, therefore for an infinite number of points the sum would be zero. For a finite number of points this is not the case, but provided N_c is sufficiently large, it follows that

$$\mu_{SR} \ll \frac{\sigma_{SYS}\sigma_{REF}}{\sqrt{N_c}}. \quad (5.31)$$

This was verified to be true for the numerical simulation used for Figure 5.9 where $N_c = 2^{22}$; in this case μ_{SR} was found to be less by approximately 3 orders of magnitude.

A second numerical simulation was carried out. In the second simulation Equations 5.24 and 5.29 were computed for different values of N_c and γ whilst the RMS voltages were kept constant at 1 V and $G_{int} = 1$. No seeding was done for the random number generator and so different sequences of σ_{SYS} and σ_{REF} were generated for each value of N_c . Figure 5.10 shows the graphs that were obtained. The results are consistent with Equations 5.28 and 5.30. The best fit line for the traces of v_a

are straight lines with a gradient scaled by γ . The error terms for both v_a and v_b plots grow much more slowly than the other terms as a result of the $1/\sqrt{N_c}$ factor. The average μ_{SR} from v_b is close to zero just as predicted in the previous simulation. Therefore, given that in practice large values of N_c will be used, this term can be neglected without significant error. This results in a simplification of Equation 5.30 as follows

$$v_b \approx \pm G_{int}(N_c - 1)\sigma_{SYS}\sigma_{REF}\sqrt{\frac{1}{N_c}}. \quad (5.32)$$

The final expression for v_o is obtained by substituting Equations 5.28 and 5.32 into 5.23 and noting that since the error terms are uncorrelated random variables they add up in quadrature. The output is thus given by

$$v_o \approx (N_c - 1)G_{int} \left(\gamma\sigma_{REF}^2 \pm \gamma\sigma_{REF}^2 \sqrt{\frac{2}{N_c} + \frac{\sigma_{SYS}^2}{\gamma^2\sigma_{REF}^2} \frac{1}{N_c}} \right) \quad (5.33)$$

$$= (N_c - 1)G_{int} \left(\gamma\sigma_{REF}^2 \pm \gamma\sigma_{REF}^2 \sqrt{\frac{2}{N_c} + \frac{\zeta^2}{\gamma^2} \frac{1}{N_c}} \right) \quad (5.34)$$

where $\sigma_{SYS}/\sigma_{REF} = \zeta$. The variables N_c , γ , σ_{SYS} and σ_{REF} are constant over the observation integration time τ , therefore the only variable that changes over this time is G_{int} and the error term. By using a sufficiently large N_c the error term can be minimised allowing for the accurate measurement of variations in G_{int} .

Finally the signal to noise ratio is given by,

$$SNR_{G_{int}} = \sqrt{\frac{N_c}{2 + \zeta^2/\gamma^2}}. \quad (5.35)$$

5.4 Experimental verification of gain calibration by correlation

5.4.1 Hardware experimental setup

To experimentally verify that gain tracking could be achieved and hence corrected for using the proposed correlation technique a receiver was implemented. Given that the target backend with digital-to-analogue capabilities was not available we used a noise diode as the reference source. This would allow us to use an existing Reconfigurable Open Architecture Computing Hardware (ROACH) board along with the Collaboration for Astronomy Signal Processing and Electronics Research (CASPER) (see e.g. [Hickish et al. 2016](#)) toolflow to implement a receiver similar to that given in Figure 5.7. The receiver RF and mixer modules were also not available and so a baseband version of the receiver was implemented as shown in Figure 5.11. Not having the RF stages of the receiver chain means that there is less gain than that which will be available in the real receiver. Consequently the amplitude of gain variation will be less for this setup than the complete receiver. Therefore if we are able to detect and correct for stability for this section of the receiver it should follow that this technique will work for the full receiver whose gain variation would be easier to detect, due to the fact that it would produce greater gain variation. The reference signal is split in half using a 3 dB power divider; one arm is digitised after being filtered and amplified whilst the other arm is attenuated before combining it with T_{SYS} (the goal is to further attenuate T_{REF} such that $T_{REF} \ll T_{SYS}$). With T_{REF} switched off and T_{SYS} switched on, the power detected by the ROACH was -20.54 dBm. With T_{SYS} switched off and zero attenuation in the path of T_{REF} the power detected was -23.48 dBm, thus confirming the 3 dB ratio introduced between the two signals by the splitting of T_{REF} . The arm with T_{SYS} and T_{REF} (i.e the signal path in a real receiver) was designed with higher gain so that the gain variation of this path dominates over any variation that would be experienced in the other leg with just T_{REF} , as would be the case in the complete receiver. The output of the two ADCs

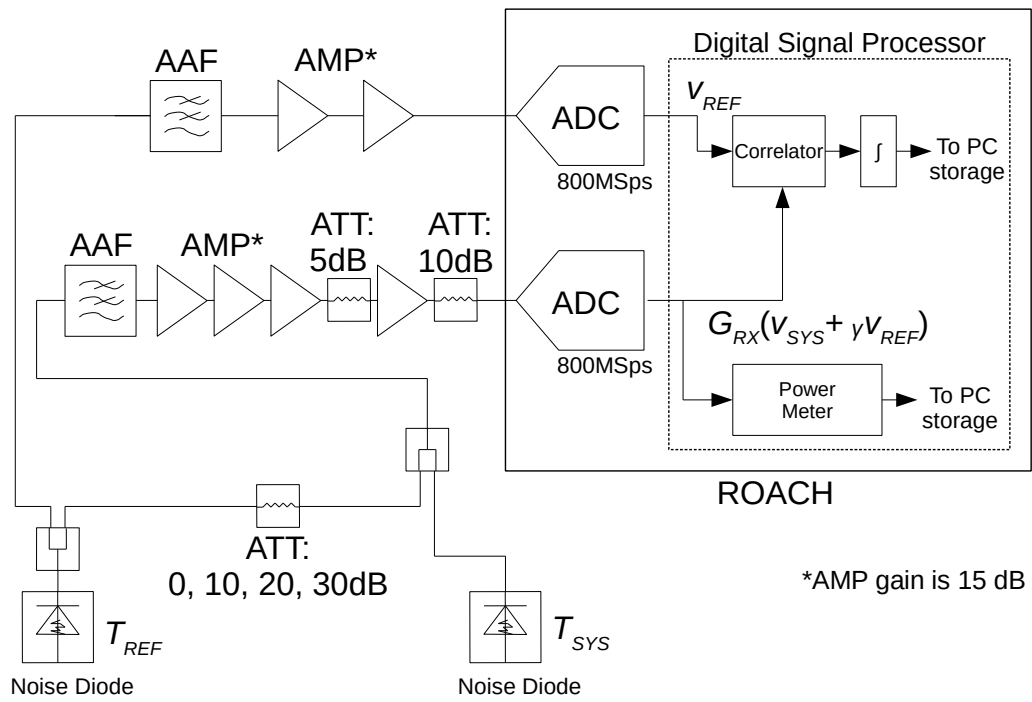


Figure 5.11: Block diagram of the experimental setup to verify the noise-correlation technique. T_{REF} and T_{SYS} are modelled using noise diodes. The reference signal is attenuated before adding it to system temperature. The FPGA correlator in the ROACH is used to carry out correlation and implement a power meter.

are the digitised versions of V_{REF} and $G(V_{SYS} + \gamma V_{REF})$, which are further processed by the digital signal processor.

5.4.2 Firmware experimental setup

As shown in Figure 5.11, the firmware implements two main functions. The first is a total power meter whilst the second is a cross-correlator with accumulation.

The total power meter is a digital implementation of a square-law detector and integrator. This sub-system implements a simple total-power radiometer over a 400 MHz bandwidth. The leg with $T_{SYS} + T_{REF}$ feeds into this subsystem, and the output generated from this is the actual desired measurement whose gain variation needs to be reduced or completely corrected for. This firmware was adopted from Pollak (2018), and the block diagram is given in Figure 5.12. The iADCs (CASPER. 2012) attached to the ROACH are clocked at 800 MHz whilst the FPGA on the ROACH is clocked at 200MHz. Therefore on each clock cycle of the FPGA, the ADC block delivers 4 sequential 8-bit samples to the FPGA per channel. The first section of the power meter implemented on the FPGA takes each of these samples and squares them. This is followed by summing and accumulation. The accumulation occurs over 2^{13} clock cycles, after which the value is fed into one of the buffer registers shown in yellow on the far right of Figure 5.12. The power meter therefore samples the input noise power at $200 \text{ MHz} / 2^{13} = 24.414 \text{ kHz}$. This is acceptable as it is much faster than the knee frequency and hence Nyquist samples the gain variations.

The second function implemented on the ROACH is the correlator . It correlates the two digitized data streams of the V_{REF} and $G(V_{SYS} + \gamma V_{REF})$, which in Figure 5.12 are designated as $y = \{y_1, y_2, y_3, y_4\}$ and $x = \{x_1, x_2, x_3, x_4\}$ respectively, tapped directly from the ADC's de-multiplexer. With 4 sequential samples available at each clock cycle, a 7-lag correlator can easily be implemented. Even though more lags can be generated it is not necessary for this application and would only add complexity to the design. However, more lags would be useful in an application that needed to apply a frequency-dependent gain correction. The outputs of the ADC, x and y , can

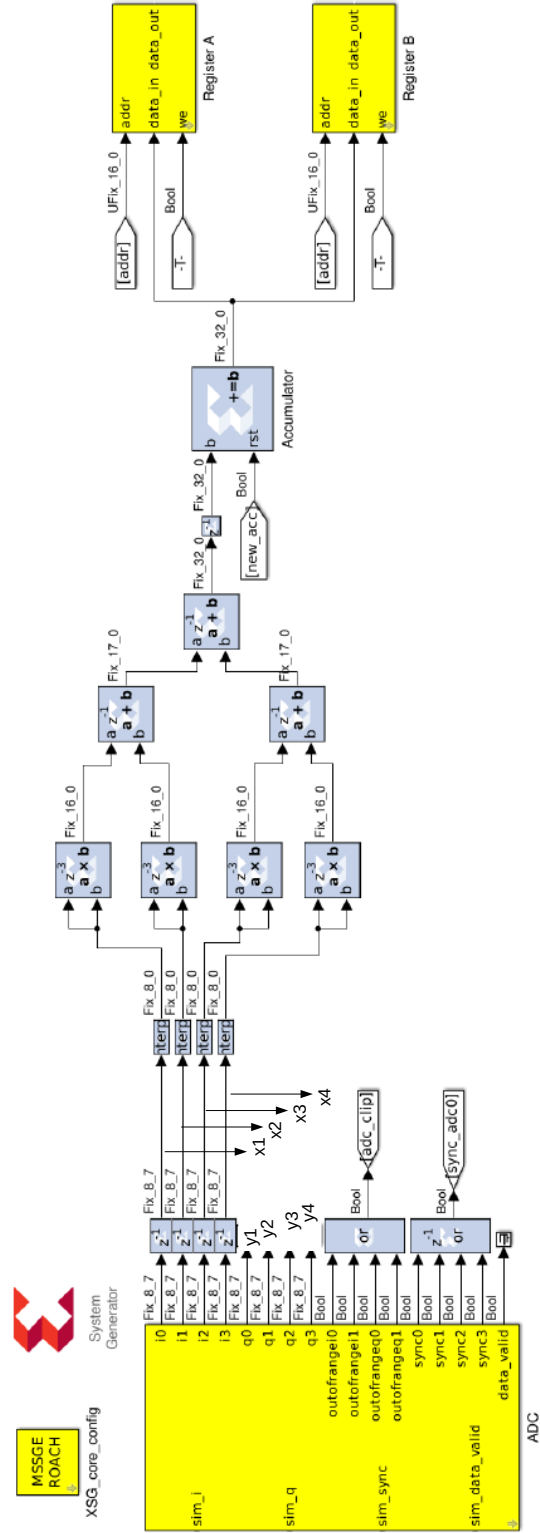


Figure 5.12: Block diagram of the power meter subsystem developed by Alexander Pollak (Pollak 2018).

also be written as

$$x = \{x[k], x[k-1], x[k-2], x[k-3]\} \quad (5.36)$$

$$y = \{y[k], y[k-1], y[k-2], y[k-3]\}. \quad (5.37)$$

Therefore the firmware carries out the following computation every clock cycle ([Hawkins 2002](#)),

$$r_{xy}^+[n] = \sum_{k=0}^3 x[k]y[k+n] \quad (5.38)$$

where k is the sample number and n the lag index, $n \in [-3, -2, -1, 0, 1, 2, 3]$. The result for each lag is then accumulated for N_c clock cycles. The Simulink FPGA correlator firmware schematics for each of the 7 lags are shown in Figures [5.13](#), [5.14](#) and [5.15](#).

The output of each lag feeds into a multiplexer (MUX) as shown in Figure [5.16](#). After the accumulation is complete, the multiplexer is activated and the values for each lag are fed into one of the registers (yellow blocks on the right) of the dual-buffer readout subsection.

5.4.3 Experimental results

For the first experiment no additional attenuation was added and so T_{REF} was about 3dB less than T_{SYS} . The cable length was adjusted so that the correlation peak was in the zeroth lag. The correlator lag outputs were integrated over 2^{21} clock cycles. This corresponds to about 10.49 ms, which means gain variation with frequency less than 47.68 Hz can be corrected for. The experiment ran over approximately 15 minutes. The expected ideal signal-to-noise for the gain measurement (using Equation [5.35](#)) is 28.6 dB. The initial stage of processing the results involved looking at the time-series data of the zeroth lag values to see if there is any variation over time. This is then compared to the waveform captured by the power meter. Given that the power meter values were obtained by integrating over 2^{13} samples, it follows that for each zeroth

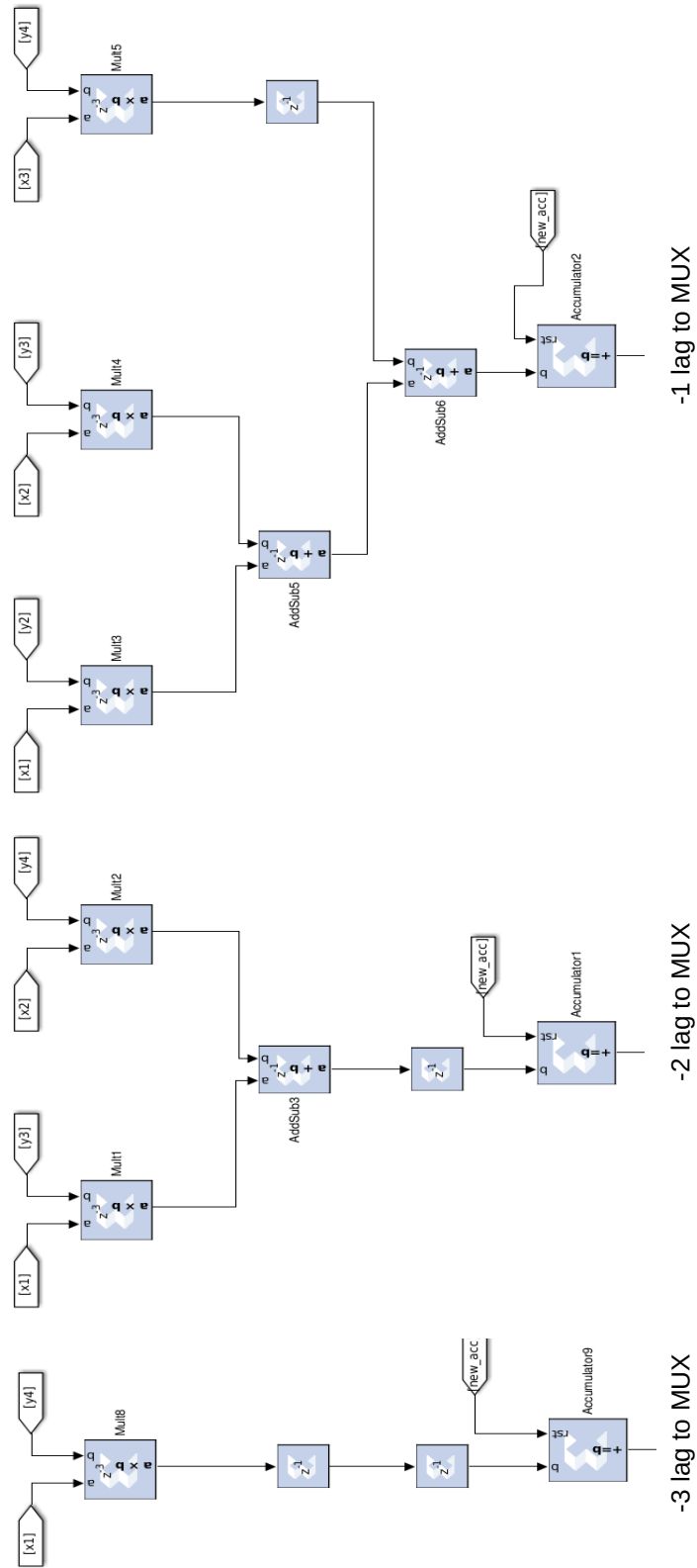


Figure 5.13: FPGA correlator firmware schematic for lags $n = -3, -2$ and -1 .

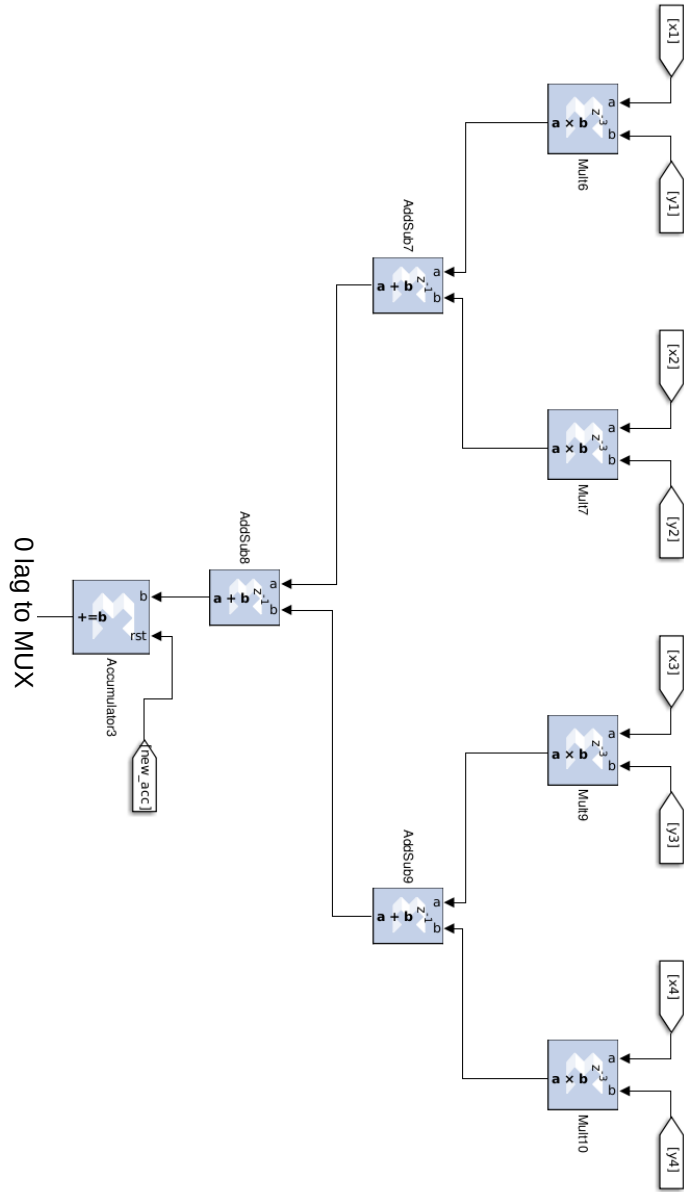


Figure 5.14: FPGA correlator firmware schematic for lag $n = 0$.

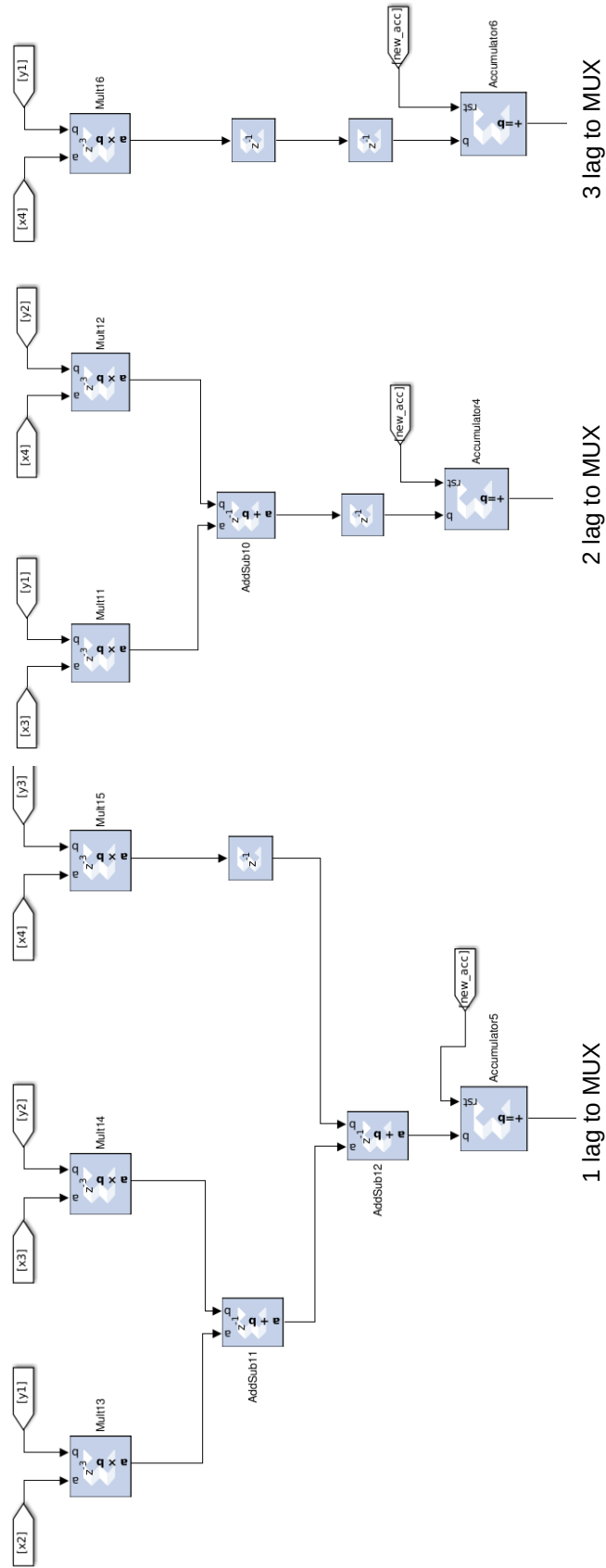


Figure 5.15: FPGA correlator firmware Schematic for lags $n = 1, 2$ and 3

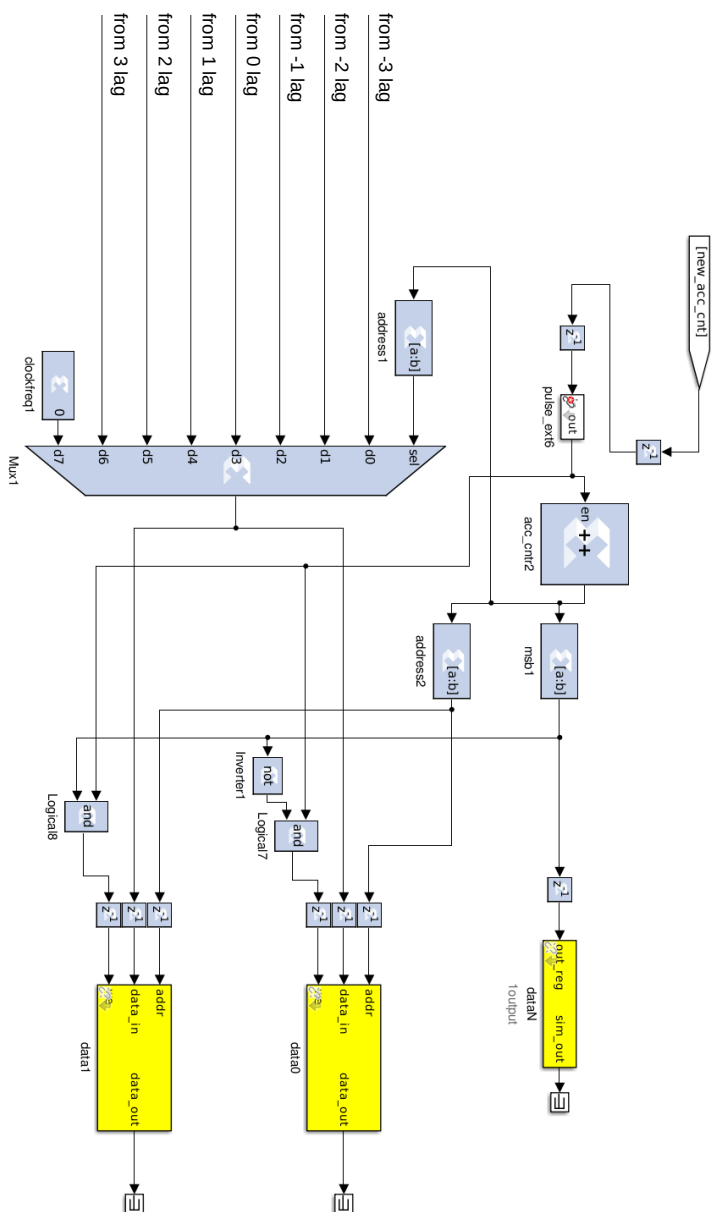


Figure 5.16: FPGA correlator firmware schematic MUX and readout section.

lag sample we have 256 samples in the power meter. Therefore the two waveforms can be compared by either interpolating the zeroth/correlation data or by carrying out a secondary integration on the power meter data. The latter was chosen as it results in a smaller file size which is easier to work with, i.e. it improves the signal-to-noise of the power meter data. It is equivalent to applying a low pass filter which allows us to see the variation in the time domain, which would otherwise be hidden by the white noise. Figure 5.17 shows the time series plot. These plots were obtained by dividing each data set by its sample mean. Over long time scales we can see that there is relatively good agreement between the measurements and so it is worth proceeding to applying the correction to the actual raw power meter data.

As can be seen in Figure 5.17, data from the correlator is noisy, therefore using it as is would result in adding more noise to our measurement data. This can be avoided by applying a filter that cuts out frequencies higher than f_{KNEE} since these will only contribute noise when the correction is applied. Therefore the data passes through 2 stages of preprocessing before it is applied in the correction. The data is first filtered using a low-pass raised-cosine filter (Alagha & Kabal 1999). A relatively broad-band filter is used in order to reduce the number of samples that are discarded at the start and end of the data sequence due to filter latency. As a consequence there is still some residue noise as is shown by the blue trace in Figure 5.17. To enable further rejection of the noise without loss of data points we then fit a 5th order polynomial to this filtered data. The resulting curve is shown by the green trace in Figure 5.17. This data sequence is then interpolated to match the number of samples available in the raw power-meter data sequence. The gain correction coefficient is calculated using:

$$\boldsymbol{\eta} = \frac{mean\{\mathbf{G}_{INT}\}}{\kappa \mathbf{G}_{INT}}, \quad (5.39)$$

where $\boldsymbol{\eta} = \{\eta_1, \eta_2, \dots, \eta_m\}$ is the gain stabilisation coefficient matrix, the set of gain estimates is $\mathbf{G}_{INT} = \{G_{int_1}, G_{int_2}, \dots, G_{int_m}\}$ and κ is a scaling factor. The scaling factor is adjusted so that the filtered estimated gain coincides with the centre of the

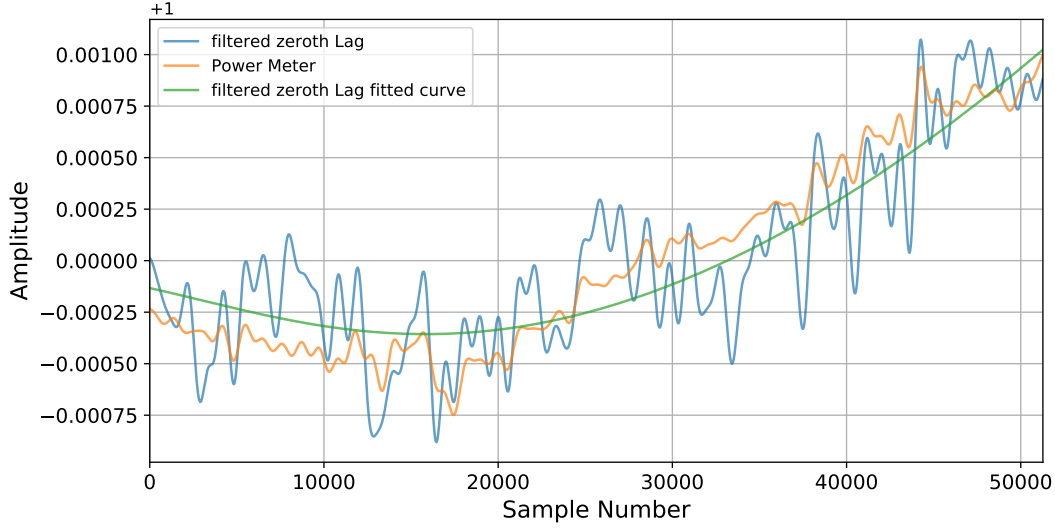


Figure 5.17: Time series data plots for the power meter, filtered-zeroth lag and curve fitting. T_{REF} attenuation was 3 dB and the correlation output was integrated for 2^{21} cycles.

filtered raw power measurement. As a result of this rescaling, the mean of the raw power data is adjusted so that it is equal to that of the correction factor. This is accomplished as follows:

$$PDO = 1 - \kappa \quad (5.40)$$

$$P_{M_n} = \frac{P_M}{\text{mean}\{P_M\}} - PDO \quad (5.41)$$

where PDO is the raw power meter data offset, P_{M_n} is the normalised power meter data and P_M is the raw power meter data. The normalised power meter data is then multiplied with the gain correction coefficient to obtain the gain stabilised power measurement P_{M_c} , as given in Equation 5.42

$$P_{M_c} = \eta P_{M_n} \quad (5.42)$$

The Fast Fourier transform is then used to compute the amplitude spectrum of both

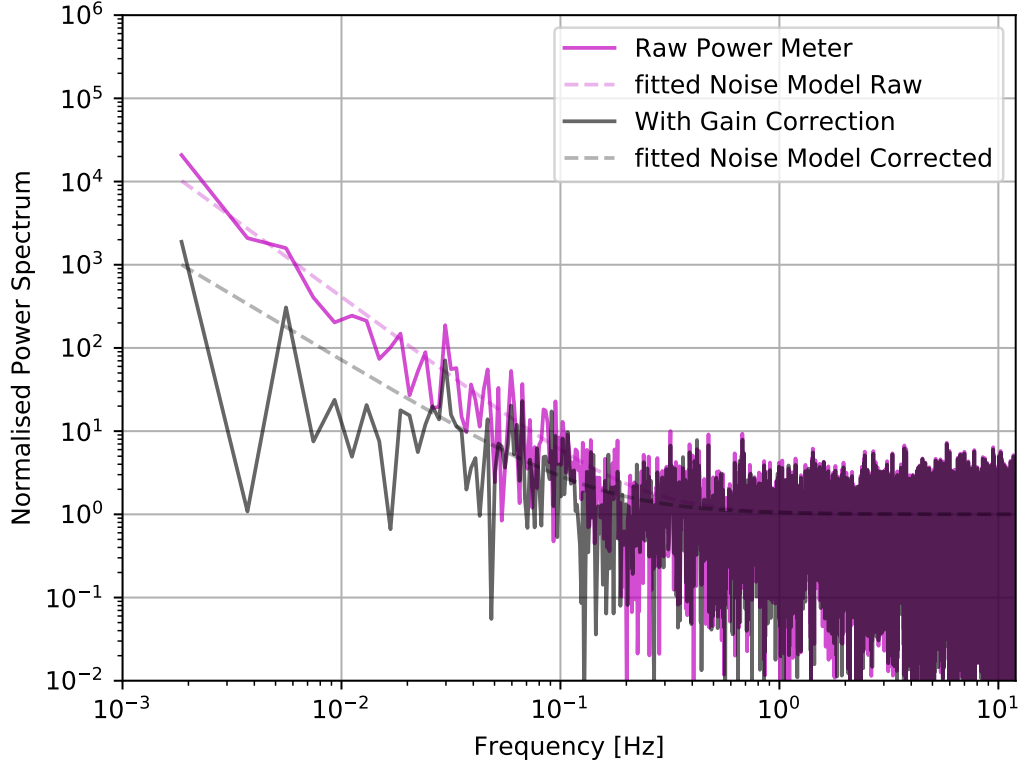


Figure 5.18: Normalised power spectrum of raw and corrected power meter data along with fitted curves: 3 dB T_{REF} attenuation and 2^{21} correlator integration.

the raw power meter data and the gain corrected power meter data. The spectra are then run through the noise-fitting algorithm discussed in Section 5.1.2. The parameters of the fit are α , σ_w and σ_r whilst f_{KNEE} is derived from this fit. The main goal of the stabilisation is to reduce σ_r without making σ_w worse. The reference frequency f_0 was set to 0.004 Hz. This provides another dimension to compare the raw and corrected spectrum in addition to f_{KNEE} . The results for the first experiment with no extra reference attenuation and with $N_c = 2^{21}$ are given in Figure 5.18. The spectra are each normalised by their white noise component σ_w . The stabilised trace shows that some correction has taken place but is relatively noisy, it could be that a longer integration time was needed. The noise model parameters are given in Table 5.1. From there we see that unstabilised trace exhibits pink noise with a knee frequency of 0.24 Hz. The stabilisation reduced it to 0.16 Hz. At the reference

frequency 0.004 Hz the pink noise was reduced by 8.9 dB.

Table 5.1: Unstabilised and stabilised noise fitting parameters for different combinations of T_{REF} level and integration time. The last experiment with 10 dB attenuation and 2^{22} integration cycles was conducted using different amplifier setup to the other measurements.

Parameter		3dB 2^{21}	3dB 2^{22}	13dB 2^{21}	13dB 2^{22}	10dB 2^{22}
α	Unstab.	1.91	1.93	1.42	1.54	2.01
	Stab.	1.58	1.23	1.03	0.98	1.24
σ_r	Unstab.	48.56	118.46	27.26	43.63	90.05
	Stab.	17.30	11.13	11.02	10.01	3.75
σ_w	Unstab.	1.00	1.00	1.00	1.00	1.00
	Stab.	1.00	1.00	1.00	1.00	1.00
f_{KNEE} [Hz]	Unstab.	0.24	0.57	0.43	0.55	0.37
	Stab.	0.16	0.21	0.43	0.45	0.04

A second experiment was carried out in which the correlator output integration time was doubled to $N_c = 2^{22}$. The ideal SNR for this gain measurement is 30.1 dB. The same procedure was followed to process the results, and the resulting normalised power spectra are given in Figure 5.19. In this experiment the correction is much better, as indicated by the noise model parameters shown in Table 5.1. The knee frequency for the raw spectrum is calculated to be 0.57 Hz, whilst the corrected knee frequency is 0.21 Hz. This more than doubles the possible integration time. Furthermore, the pink noise level at the reference was reduced by 20.5 dB. This is much better than the previous experiment.

So far the T_{REF} power level is only 3 dB below T_{SYS} , which results in a system sensitivity of $1.5T_{SYS}/\sqrt{\Delta\nu\tau}$. Although this is better than the Dicke switch is it is far below the potential of this stabilisation technique. To test if it is indeed possible to achieve stabilisation when $T_{REF} \ll T_{SYS}$, T_{REF} is further attenuated by 10 dB, bringing the total attenuation to 13 dB. For this setup the ideal system sensitivity is $1.05T_{SYS}/\sqrt{\Delta\nu\tau}$. Two measurements were then carried out, one with correlator output integration over 2^{21} points and the other 2^{22} points; with expected SNRs for

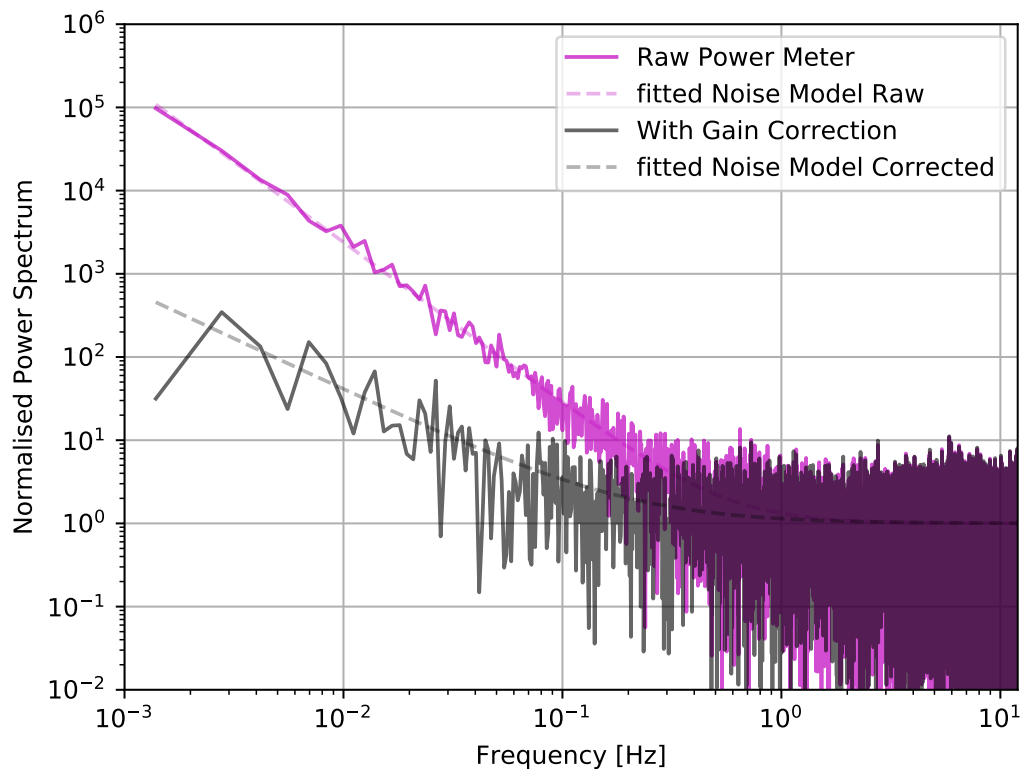


Figure 5.19: Normalised power spectrum of raw and corrected power meter data along with fitted curves: 3 dB T_{REF} attenuation and 2^{22} correlator integration.

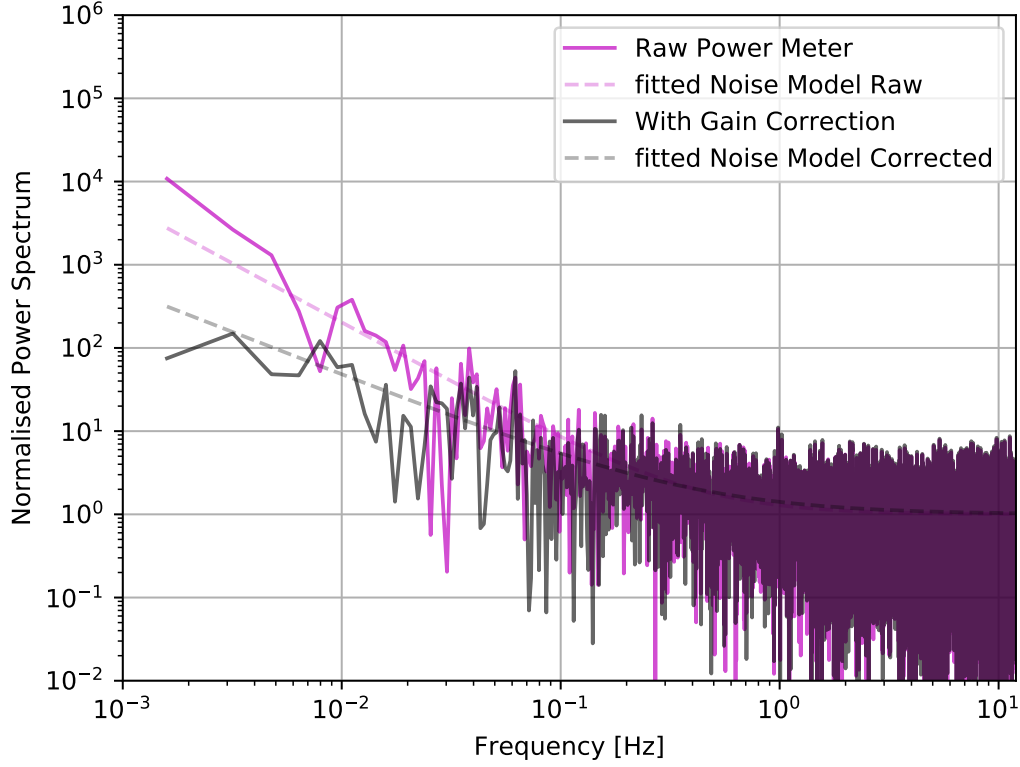


Figure 5.20: Normalised power spectrum of raw and corrected power meter data along with fitted curves: 13 dB T_{REF} attenuation and 2^{21} correlator integration.

gain estimation of 24.9 and 26.4 respectively. The spectra for the former is given in Figure 5.20. Despite the poorer SNR the stabilised data does show some improvement, in fact at the lower end it appears to have almost flattened. As expected due to lower SNR, with reference to Table 5.1, for this experiment σ_r shows the lowest improvement after stabilisation. There is also no change in the knee frequency.

The spectra for the 2^{22} point integration is given in Figure 5.21. An improvement can be seen: just as in the previous experiment the stabilised trace appears to have been flattened for the frequency components below f_{KNEE} . The noise fitting parameters are given in the second last column of Table 5.1. The noise level at the reference frequency has been reduced by 12.8 dB. There is a corresponding slight reduction in knee frequency from 0.55 Hz to 0.45 Hz.

Generally the resulting effect of this gain stabilisation on the power data is to

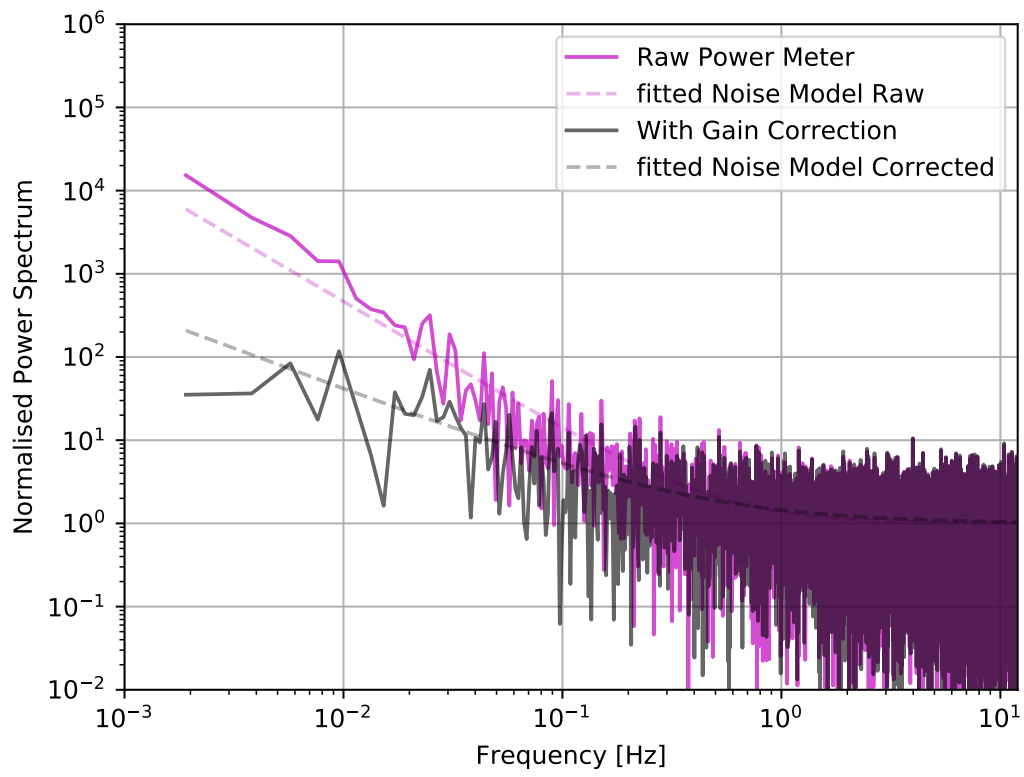


Figure 5.21: Normalised power spectrum of raw and corrected power meter data along with fitted curves: 13 dB T_{REF} attenuation and 2^{22} correlator integration.

reduce the magnitude and slope of the pink noise below f_{KNEE} with minimal effect on f_{KNEE} and σ_w . These results are positive as they show that to some extent gain stabilisation can be achieved even when T_{REF} is much smaller than T_{SYS} .

However, for the last four experiments we expected a signal-to-noise ratio in excess of 20 dB, therefore the gain correction is not being limited by noise. The ADCs have 8-bit resolution with a full-scale deflection of 500 mV peak-to-peak, -22 dBm will result in at least 2 to 3 bit sampling. Sampling at this resolution should have a quantisation efficiency of 0.86 to 0.96 (Thompson et al. 2017). With the additional 10 dB attenuation we have $T_{REF} = -33.48$ dBm at the ADC's input. This results in the signal occupying fewer quantisation levels, which at 1-bit resolution offers a quantisation efficiency of 0.637 (Thompson et al. 2017). Although this is the case it is not a problem because of dithering made possible by the white noise of the system temperature. On the other hand, for the other leg of our setup which allows for the acquisition of T_{REF} the power at the ADC is fixed (at around -36.85 dBm) there is no additional noise input to enable effective dithering hence resulting in a lower quantisation efficiency. This is probably the reason why even after filtering the gain estimation using the raised-cosine filter the resulting data was still significantly noisy. Apart from reducing the SNR the reduction in the number of bits also results in poor gain tracking as the ability to resolve variations in the gain is greatly reduced. This limits how well the gain estimate can resolve variations in the gain even in the absence of noise. The other possible source of error for this particular system are v_{REF} 's amplifiers; even if the total gain is lower than the other leg, the instability of these amplifiers may be enough to compromise the correlation (this is less true for the noise diodes which are temperature stabilised). To overcome some of these limitations, the experimental hardware setup was revised. The updated block diagram is given in Figure 5.22. The amplifiers for v_{REF} were moved to precede the splitter, thereby totally removing the effect of their instability on the correlation. Secondly, an additional amplifier was included to raise the power level at the ADC for T_{REF} acquisition. For the receiver portion of the setup the power levels were adjusted by the 56 dB attenuation so that even with T_{REF} 10 dB lower than T_{SYS} , it was still being

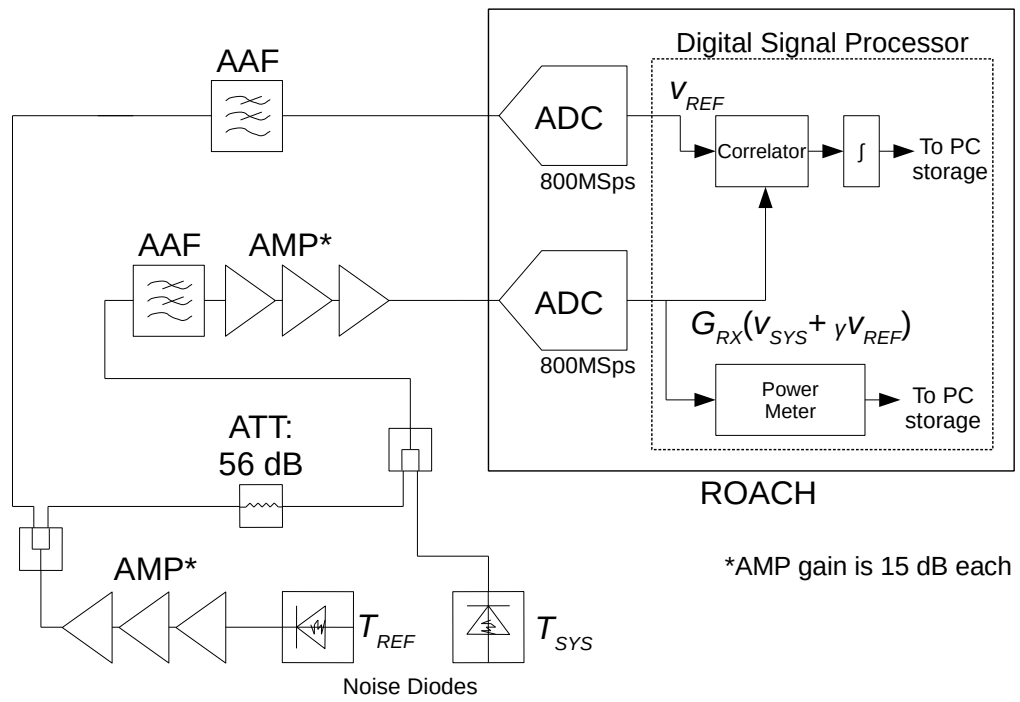


Figure 5.22: Revised experimental setup to verify the noise correlation technique. T_{REF} and T_{SYS} are modelled using noise diodes. The reference signal amplifiers have been relocated to negate their instability. Power levels at both ADCs now allow for better than 3 to 4 bit sampling resolution even when $T_{SYS} = 0.1T_{REF}$.

digitised with 2 to 3 bits resolution. This reduces the reliance on dithering. With these changes the gain variation in the T_{REF} amplifiers is no longer a problem as it is common to both legs of our setup, so better correlation coefficients are achievable. In addition, T_{REF} is acquired with greater resolution (3 to 4 bits), hence we expect higher SNR for the gain measurement.

The integration time was maintained at 2^{22} , as it provided both sufficient SNR (27.7 dB) and good time resolution. The gain variation's sampling frequency was 47.7 Hz, so variation in gain of up to 23.8 Hz could be tracked. Processing of the results was similar to the previous measurements, but in this case it was not necessary to fit a curve to the gain estimate before applying the gain correction, as the noise could be filtered to the required bandwidth without needing a very narrow low-pass filter. The best measurement result is shown in Figure 5.23. The noise fitting parameters are given in the last column of Table 5.1. The pink noise level at the reference has been reduced by 27.6 dB and the knee frequency has been reduced from 0.37 Hz to 0.04 Hz, thereby increasing the possible integration time from 2.7s to 25s. In practical terms, C-BASS required an integration time of 20s to achieve the desired sensitivity (Copley 2014). These results demonstrate that this method of stabilisation would have met that specification. This clearly demonstrates the potential that this technique has to offer.

The target system's ADCs on the other hand will provide 4 GSps with 12 bit depth (Collins 2017). This is 4 times more sample points to average and a greater dynamic range to allow for further reduction of T_{REF} relative to T_{SYS} . Therefore even better performance in the target system can be expected.

The correlator operates on the full system bandwidth, the same as the power meter RF bandwidth, therefore the gain estimate generated by the 7 lags of the correlator is an indication of the gain variation across the entire RF bandwidth of the system. This means that, for instance, when $N_c = 2^{22}$ we can measure gain variations which vary slower than 23.8 Hz across the entire RF bandwidth. In the case of a total-power radiometer such as the system implemented in the experiment, only the zeroth lag was used for gain estimation. In the event that a spectrometer is implemented would

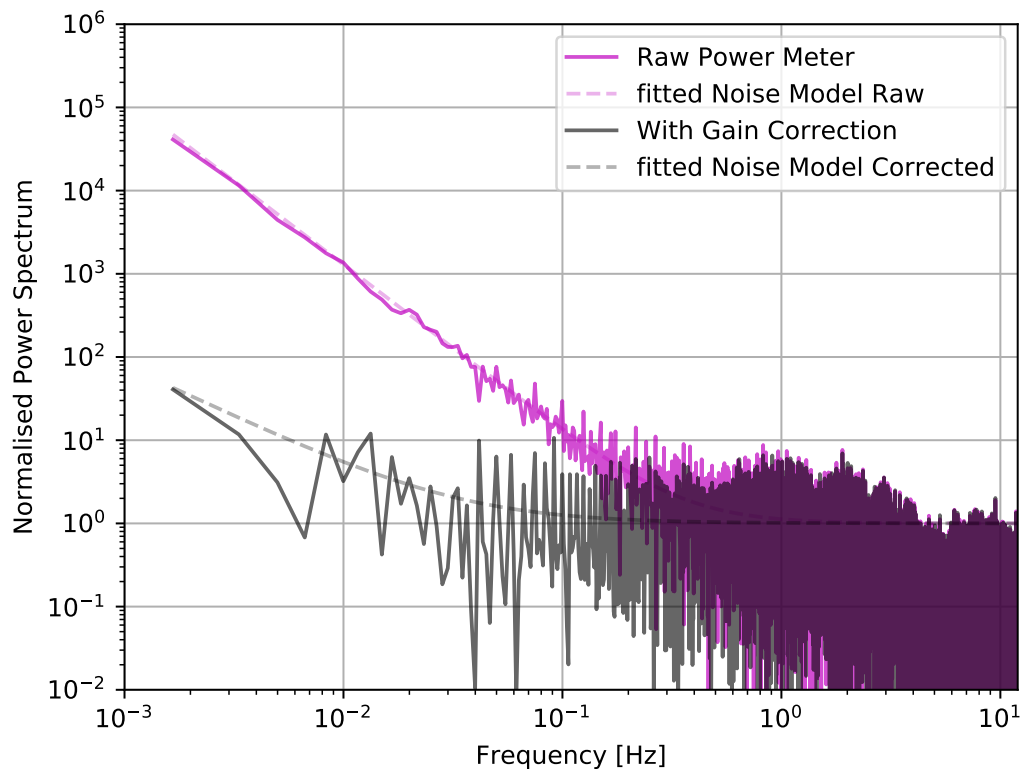


Figure 5.23: Normalised power spectrum of raw and corrected power meter data along with fitted curves. 10 dB T_{REF} attenuation and $N_c = 2^{22}$.

it still be possible to use the zeroth lag data only for each of the frequency channels, or would each channel require a different correction factor determined using all 7 lags of the correlator? A similar question had to be answered for the CW stabilisation technique presented in (Pollak et al. 2019). By computing the Pearson Correlation Coefficient (PCC) between each of the frequency channels it was shown that gain drifts of the different frequency channels were correlated. In other words, the gain drift did not show any frequency dependence, therefore one frequency channel could be used to correct for gain variation in all the other channels. In the remainder of this chapter we will investigate whether the zeroth lag data would be sufficient to estimate the gain variation in each frequency channel had a spectrometer been implemented instead of a total power radiometer.

For the experiments carried out in this thesis a spectrometer was not implemented. As a result there was no way to directly access individual frequency bins. To overcome this hurdle convolution theory and the Wiener-Khinchin theorem can be used (Wilson et al. 2009). Let $v_c[n]$ be the output of the correlator each clock cycle, where n is the lag index. The Fourier transform of v_c can be written as,

$$\mathcal{F}\{v_c[n]\} = H_{SYS}(f) \left(\gamma V_{REF}(f) + V_{SYS}(f) \right) V_{REF}^*(f) H_{SYS}^*(f) \quad (5.43)$$

$$= |H_{SYS}(f)|^2 \left(\gamma |V_{REF}(f)|^2 + V_{SYS}(f) V_{REF}^*(f) \right), \quad (5.44)$$

where the gain of the system has been represented by a more general term $H_{SYS}(f)$, the frequency response of the system. The second term in the brackets corresponds to the correlation of v_{SYS} with v_{REF} , which is zero as previously stated. Thus we have

$$\mathcal{F}\{v_c[n]\} = \gamma |H_{SYS}(f)|^2 |V_{REF}(f)|^2. \quad (5.45)$$

Given that $|V_{REF}(f)|^2$ is the power spectral density (PSD) of the reference signal which has a flat frequency response, the Fourier transform of the correlator lags will give an estimate of the system response after each integration. With 7 lags, 4 spectral

bins will be available, with one of them being the DC component. By calculating the PCC between each of these we can determine if gain tracking using this method has any frequency dependence and hence determine whether zeroth lag data only can be applied to the entire bandwidth. If the correlation between channels is poor, then more lags would be needed to achieve the desired resolution in the frequency domain and each channel's gain estimate would then be obtained from the Fourier transform of the lags.

Using the results from the last experiment (10 dB | 2^{22}), a real FFT was carried out on the 7 lags for each correlation integration cycle over the entire measurement time. The PCC was then calculated for each combination of the frequency bins. The results are shown in Table 5.2. The diagonal corresponds to the autocorrelation for

Table 5.2: PCC: For each bin across the band for 10 dB attenuation and $N_c = 2^{22}$.

	0	133 MHz	267 MHz	400 MHz
0	1.000	0.992	0.905	0.957
133 MHz	0.992	1.000	0.882	0.963
267 MHz	0.905	0.882	1.000	0.844
400 MHz	0.957	0.963	0.844	1.000

each frequency bin, hence the coefficient of 1. The channels exhibit good correlation with an average PCC of 0.94 ± 0.06 . Simulating a seven lag correlator in Python under the same conditions of N_c , ζ and γ yielded a PCC of $1.00 \pm 5 \times 10^{-6}$. A lower PCC in the measurement data maybe due to other physical characteristics such as quantisation and variations in the signal path not common to both legs of the system (for example the ADCs not being perfectly identical) that are not accounted for in the simulation. This is an indication that potentially better results could be achieved using the implementation depicted in Figure 5.8. The correlation coefficients do not exhibit any pattern, but appear random as is expected in a system where uncorrelated noise is present.

These results show that gain tracking achieved using the zeroth lag data can be used to correct for receiver instabilities even in multichannel spectrometers for this hardware setup. For a system where the gain variation is not correlated across the

band, all lags of the correlator would have to be used to achieve reasonable spectral resolution to apply independent correction coefficients; this could also include phase variation calibration.

5.5 Summary

In this chapter receiver sensitivity was discussed. The effects of gain instability on receiver sensitivity were also presented. This was then followed by a discussion on the various techniques that are currently available to stabilise the receiver, improving the sensitivity. Advantages and disadvantages of each technique were presented. For some of the methods, better stability comes at the cost of increased hardware complexity, such as the pseudo correlation receiver. The CW stabilisation receiver offers the best sensitivity without complicating the hardware; this comes at the expense of losing one frequency bin of the spectrometer. A new stabilisation technique was then presented. It offers the same sensitivity performance as the CW stabilised receiver without the loss of any frequency channels. It requires similar hardware to the CW with the main difference being the requirement of an additional ADC or DAC depending on the specific implementation. This requirement is not significant as most DSPs can easily be equipped with an additional ADC; this is especially true for the target system whose digital signal processor will be based on the RFSoc which has a plethora of ADC channels and DACs.

The mathematical working principle of the proposed technique was presented along with the derivation of the expected signal-to-noise ratio of the gain measurement. The last section in the chapter presented results obtained from experimental testing of the proposed technique. The results demonstrate the effectiveness of this technique. The final set of results confirm that this technique can be used to stabilise receiver gain variations for spectrometer receivers such as the target system.

Chapter 6

Receiver System

At the time of writing the system has not yet been completed, but the design is complete and all key components have been tested. In this chapter we present an overview of the entire system and then focus on the design of the receiver system, as well as the components that have been prototyped. In Section 6.2 the design choices are presented. This is followed by a discussion on the RF section of the receiver in Section 6.3. In Section 6.4 the hardware implementation of the digital backend is presented, along with the functional block diagram of the main signal processing steps. In the final section all the results from the other chapters are combined to estimate the expected system performance.

6.1 System overview

The AVN receiver is composed of three main systems as illustrated in Figure 6.1. The figure highlights what is covered within the scope of this thesis. The antenna system consists of the antenna optics and the cryostat subsystems. The antenna optics were covered in greater detail in Chapter 3. The cryostat provides the appropriate conditions to cool down the transducer and LNA. More detail will follow in Section 6.2.

Signal conditioning has two main subsystems to account for the two domains in which the signal is processed. Analogue processing takes place first; the RF signal

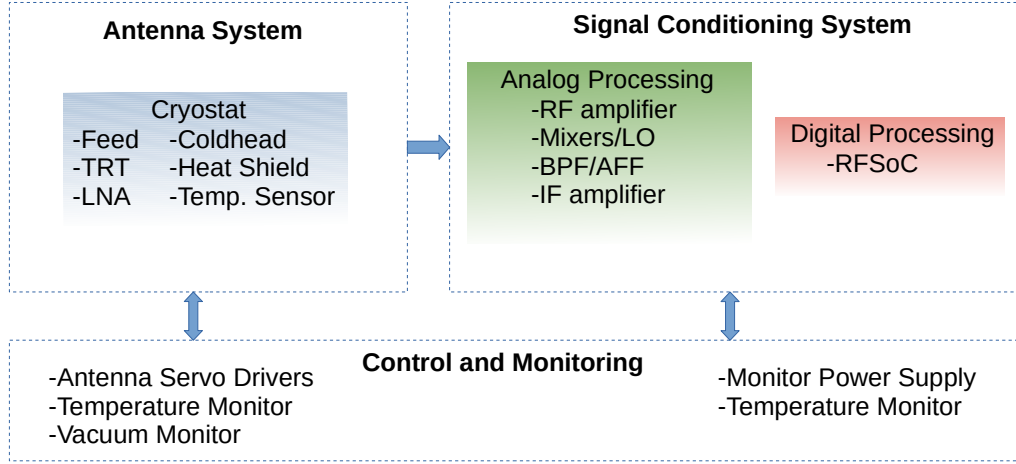


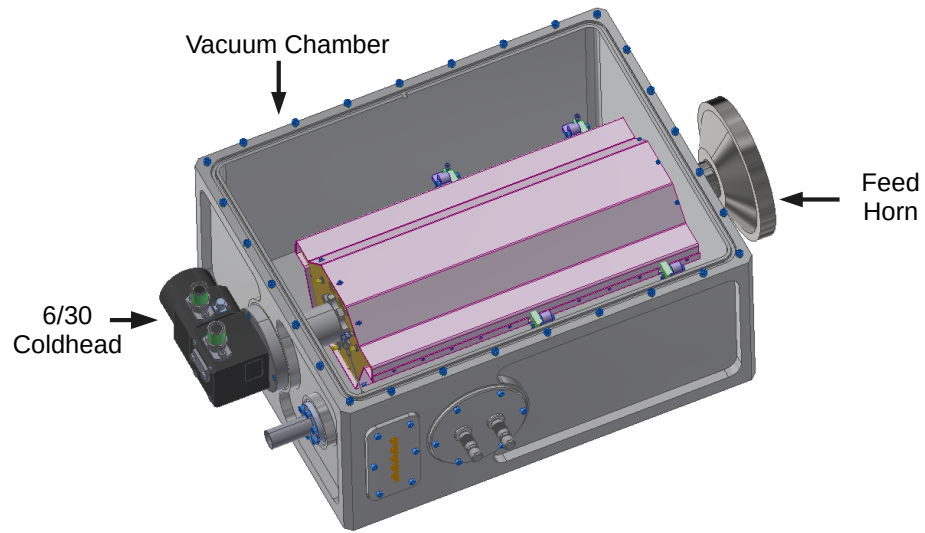
Figure 6.1: Block diagram of system overview.

from the LNAs is filtered, amplified and down-converted to baseband. The final stage in the signal processing is the digital processing which starts with the digitisation of the baseband signal. This allows the remainder of the processing to be carried out in the digital domain.

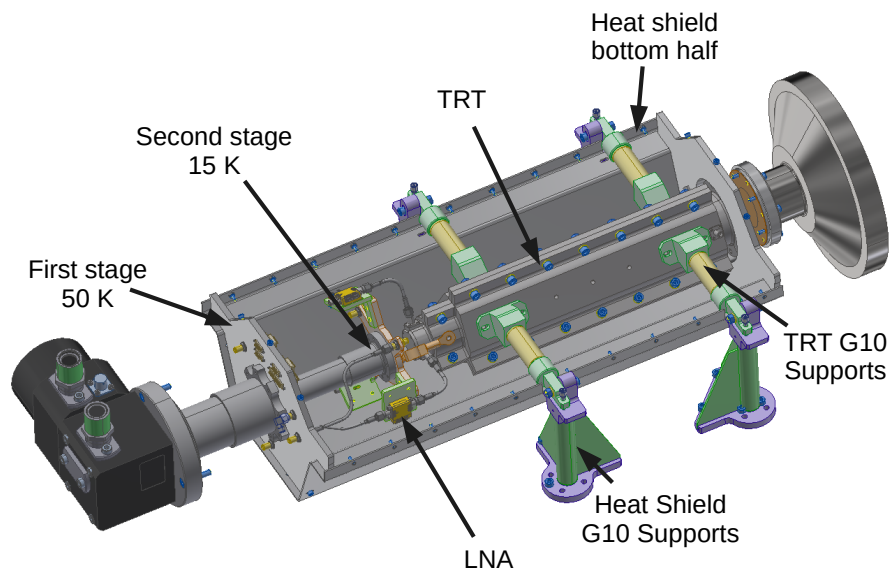
The third system is the control and monitoring system, which manages the overall operation of the system. It is responsible for the pointing of the antenna and monitoring of the vacuum in the cryostat. It also monitors the temperature of the components which are temperature sensitive. The power supplies of both the cryostat and signal conditioning subsections are also monitored and controlled by this system.

6.2 Cryostat

Each component in the receiver RF chain contributes noise to the overall system temperature as discussed in Chapter 5. To achieve the highest possible sensitivity it is important to minimise the amount of noise each component contributes. This is especially critical for components in the front end of the receiver preceding the LNAs. For passive devices such as the TRT, the noise power contributed is proportional to



(a) Cryostat without Top Lid



(b) Heat shield Sub-assembly Without the Top Half

Figure 6.2: Cryostat 3D CAD model.

its equivalent noise temperature (e.g. [Pozar 2012](#)) given by,

$$T_e = \frac{1 - G}{G} T_{phy}, \quad (6.1)$$

where G is the gain of the device and T_{phy} is the physical temperature of the device. Assuming the TRT has a gain of 0.9, equivalent to an insertion loss of -0.45 dB, and its physical temperature is 300 K, this would result in an equivalent noise temperature of 33.3 K. On the other hand if it is held at a temperature of 15 K the equivalent noise temperature is only 1.7 K. Therefore cooling the TRT along with the LNAs yields much better system sensitivity.

In order to keep the LNAs and TRT cold the heat flow towards these devices has to be controlled. The mechanisms by which heat transfer occur are conduction, convection, phase-change and radiation ([Bejan & Kraus 2003](#)). The cryostat is designed to minimise the amount of heat transferred via any number of these mechanisms. If a temperature gradient exists in air, the warmer air's density will be lower resulting in it rising whilst the cooler denser air falls to replace it. This results in a convective current that would facilitate heat transfer. This is avoided by forming a vacuum in the cryostat. Consequently this also inhibits phase-change heat transfer; as ideally the water vapour or any other substances are also evacuated. Conduction is unavoidable as the TRT and LNA's need to be mounted and supported. Fourier's law, Equation 6.2, quantifies the conductive heat current q in Watts for an object with a temperature gradient,

$$q = -kA \frac{dT}{dx}, \quad (6.2)$$

where k is the thermal conductivity, A is the cross-section area, whilst dT/dx represents the temperature gradient. From this we see that conduction can be minimised by selecting the appropriate material with low thermal conductivity and by restricting the cross sectional area. These are both employed in the cryostat design. A CAD model of the cryostat, designed by Paul Brentnall is shown in Figure 6.2. Figure 6.2 (b) shows more clearly the G10 struts that support the TRT off the heat shield.

This provides a thermal buffer between the 15 K and 50 K stages as G10 has a low thermal conductivity at the temperatures of interest. The same material is also used for the heat shield supports. These provide conductive thermal isolation between the chamber sitting at 300 K and the heat shield. Stainless steel instead of aluminium coaxial cables are used between the LNA and the bulkhead SMA connector as these have a lower thermal conductivity. The heat flow is also directly proportional to the temperature gradient and so by minimising the gradient heat flow can also be reduced. This is achieved by thermally bonding the RF and bias cables from the 15 K stage on the 50 K stage using bulkhead feed-through connectors rather than directly terminating them on the 300K chamber bulkhead. Similarly the supports of the TRT are attached to the heat shield rather than supporting them directly off the vacuum chamber.

The other means by which heat is transferred is radiative heat transfer and this ends up being by far the greatest contributor to heat flow. The radiative heat transfer current between two surfaces in watts is given by the following expression,

$$q = \frac{\sigma A_2 (T_1^4 - T_2^4)}{\frac{1-\epsilon_2}{\epsilon_2} + \frac{1}{F_{21}} + \frac{A_2}{A_1} \left(\frac{1-\epsilon_1}{\epsilon_1} \right)} \quad (6.3)$$

where σ is the Stefan–Boltzmann constant, ϵ_1 and ϵ_2 are the emissivities of the surfaces, A_1 and A_2 are the surface areas and F_{21} is the configuration factor defined as the fraction of heat flow leaving surface 2 that strikes 1 directly; it is solely dependent on the geometry of the two surfaces. For simplicity if we assume that surface 1 and 2 are concentric it follows that every ray leaving surface 2 is intercepted by 1, therefore $F_{21} = 1$. If we also assume that surface 1 behaves like a perfect blackbody ($\epsilon_1 = 1$) then Equation 6.3 simplifies to

$$q = \sigma \epsilon_2 A_2 (T_1^4 - T_2^4). \quad (6.4)$$

Even if the cryostat walls are not black bodies, they behave as such if the cryostat is large compared to the cold part inside, because any ray leaving the cold part will

reflect many times and eventually be absorbed. Furthermore, the surfaces involved are not perfectly concentric therefore using Equation 6.4 will predict a higher heat load, but this will provide an extra margin of safety. It is worth noting here that temperature difference has a significant influence on heat flow. The main role of the heat shield is to act as a thermal intercept as it stops the objects at 15 K from being directly illuminated by the 300 K walls of the chamber (similar to how the bulkhead on the second stage is used in the case of conductive heat transfer). This way the heat load on the second stage is significantly reduced. There is still radiative heat transfer between the heat shield and the chamber, given the temperature difference but, there is also more power available for cooling on the first stage than there is on the second stage. Efforts were also made to minimise the surface area of the heat shield, as well as lowering its emissivity by constructing it out of polished aluminium. Provision has also been made for mounting temperature sensors on the LNA, TRT, cold plate and heat shield. There is also capability to mount heaters on the LNA bracket should there be a need to temperature stabilise them.

The estimated conductive heat load for each stage is given in Table 6.1. TRT G10

Table 6.1: Conductive heat transfer breakdown.

Object	A [m ²]	L [m]	k [W/mK]	T_1 [K]	T_2 [K]	Power [W]
TRT G10 Strut	1.21×10^{-3}	0.082	0.2	50	15	1.03×10^{-1}
HS G10 Strut	1.01×10^{-3}	0.116	0.4	300	50	8.68×10^{-1}
Coax cables	-	-	-	50	15	2.29×10^{-3}
Coax cables	-	-	-	300	50	1.59×10^{-1}

Strut refer to the TRT's support structure whilst HS G10 Strut refers to the heat shield's support structure. The values in the Coax cable rows refer to the total heat load contributed by all the RF coaxial cables between each stage. As can be seen in the table using G10 supports and limiting the area of conduction has minimised the conductive heat load. For the coaxial cable calculations it was assumed that the cable is composed of a stainless steel outer-conductor with a beryllium copper inner-conductor separated by a PTFE dielectric. These materials provide lower thermal conductivity and enough flexibility to be bent into the appropriate shapes. Even

though they have lower electrical conductivity than copper it is not a problem as these cables are used after the signal has been amplified by the LNA. Given that the digitiser has sufficient dynamic range any slope introduced in the RF passband will not pose any problems.

Table 6.2: Radiative heat transfer breakdown.

Object	A [m ²]	ϵ	σ [W/m ² K ⁴]	T_1 [K]	T_2 [K]	Power [W]
TRT Sides	7.14×10^{-2}	0.4	5.67×10^{-8}	15	50	1.00×10^{-2}
TRT Flange	3.51×10^{-3}	0.4	5.67×10^{-8}	15	300	6.45×10^{-1}
Heat shield	3.80×10^{-1}	0.1	5.67×10^{-8}	50	300	1.74×10^1

The breakdown of the radiative heat load is presented in Table 6.2. If the TRT were to be polished its radiative heat load could be reduced further, but even without polishing the load is not significant. The greatest load on the second stage is as a result of the TRT flange exposure to the cryostat’s 300 K internal walls. The total

Table 6.3: Combined heat load and estimated available cooling capacity.

	15K stage [W]	50K stage [W]
Heat load	0.76	18.5
Cooling capacity	6	25

heat load on each stage is shown in Table 6.3. The total heat load on the first stage is under 1 W whilst on the second stage we expect about 19 W. The bulk of this comes from the contribution of the radiative heat transfer from the walls of the vacuum chamber to the heat shield. In principle we could reduce this load by using multi-layer insulation but, it was considered safer to use a cold head with sufficient cooling power. The proposed 6/30 coldhead has enough power for the predicted loads.

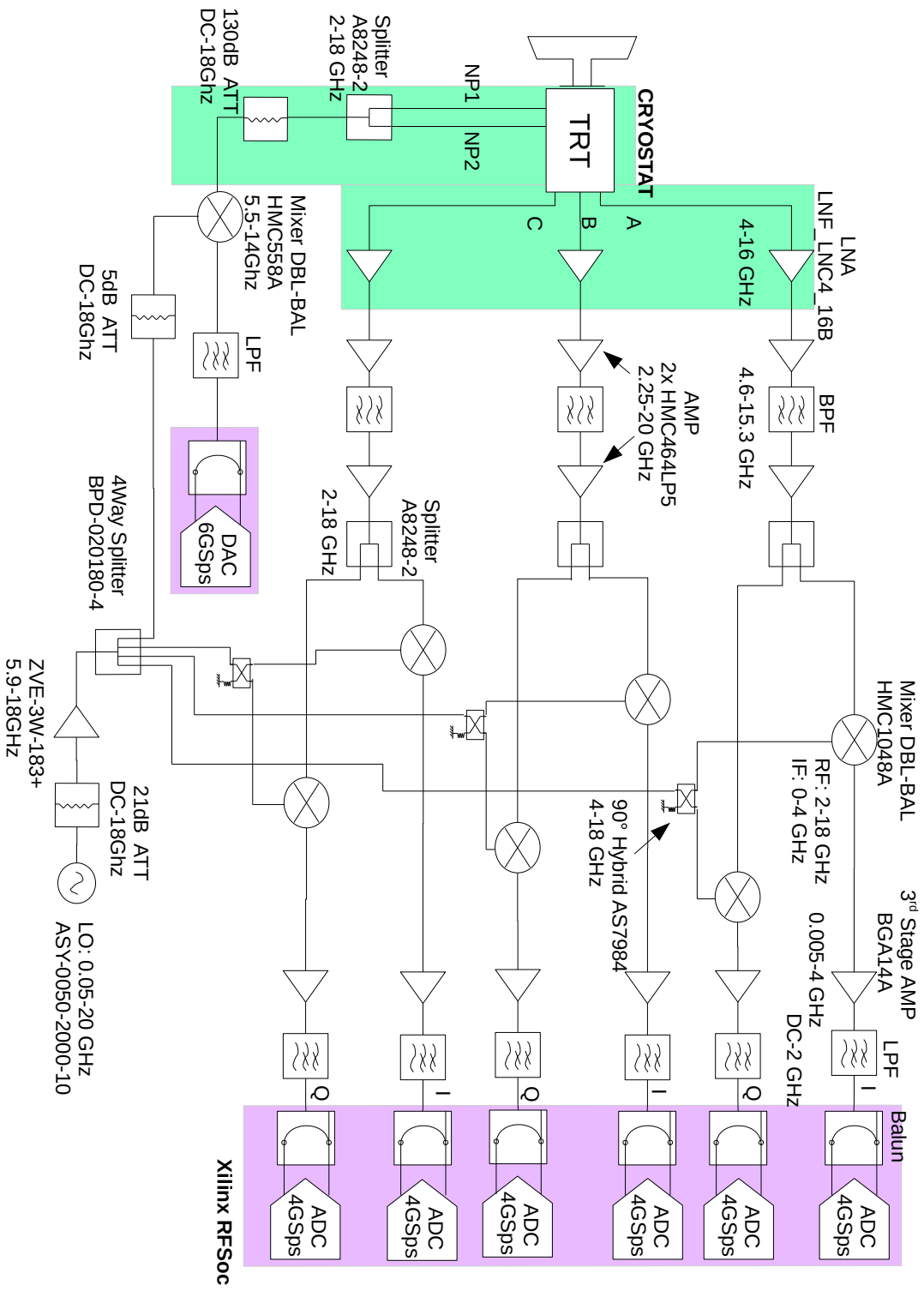


Figure 6.3: AVN receiver block diagram.

6.3 Analogue signal processing

6.3.1 Receiver architecture

A block diagram of the receiver is given in Figure 6.3. Even though the primary band of interest is 4.6 to 8.5 GHz, the receiver chain has been designed to cover the full bandwidth of the transducer making future upgrade easier. The main function of the RF electronics is to condition the analogue signal without degrading it so that it can be digitised by the backend. The signal received by the feed horn is fed into the transducer which converts the received electromagnetic waves into electrical signals which are then amplified by the LNAs. The cryogenic amplifiers amplify the signal without significantly degrading SNR. Band pass filtering is then carried out to select the relevant frequency band. The signal is then split in two to allow for I/Q mixing. Mixing then shifts each leg to baseband; in this case 0 to 2 GHz. After the final stage of amplification the signals are filtered by low pass filters to avoid aliasing. Finally digitisation occurs which allows for the remainder of the signal processing to go ahead in the digital signal processor.

I/Q mixing is utilised to allow for complex sampling of the received signal. This coupled with the use of the complex Fourier transform allows for the desired band to be recovered in the digital signal processor whilst only using one LO and ADC's running at half the frequency that would otherwise be required. Figure 6.4 helps illustrate how this is achieved. Figure 6.4 (a) shows an illustration of the band pass filter's output signal. It is a real signal which is then split in half by the RF splitters. One of these is multiplied with the LO directly resulting in the in-phase component, $i(t)$. The other half is multiplied with a 90° shifted version of the LO signal resulting in the out-of-phase version of our signal, $q(t)$. These are illustrated in (b). The results of low pass filtering and amplification are shown in (c). These signals are then digitised by the ADCs. Carrying out a complex Fast Fourier Transform on the data will allow for both the real and negative frequency components to be available. Therefore the lower part of the band that was moved to the negative frequencies by the mixing can be recovered. Given that $I(f)$ and $Q(f)$ are the spectrum of the

in-phase and out-of-phase signals the desired signal spectrum can be recovered by computing $I(f) - jQ(f)$; this simply reverses what was done in the time domain. The results of this operation are shown in part d) of the figure.

6.3.2 Component design

Some of the components required for the implementation were not available in the laboratory and were also not readily available off the shelf. These had to be developed in-house. The first of these is the bandpass filter. For this system a 4.6 to 15.3 GHz bandpass filter is required. Microstrip filter implementation was chosen as good performance can be achieved cost effectively. The goal was to use a shorted-stub type bandpass filter; the reason for this is that the shorted-stubs provide more effective capacitive and inductive shunt paths to ground, which can be tuned to provide high impedance over the passband and low impedance to ground in the stop-bands of the filter. There are various ways in which such filters can be designed. [Hill-Valler \(2019\)](#) designed a 7 to 15 GHz bandpass filter using the insertion loss method ([Pozar 2012](#)). This technique allows for greater control over the passband, stop-band amplitude and phase characteristics. This is made possible because the filter response is defined by the insertion loss. Depending on the desired filter characteristics an appropriate polynomial is selected to define the insertion loss, and component values are then determined from this polynomial. However, for substrates with convenient ϵ_r such as Rogers R04350 and Rogers 6010-2LM, it was shown that the impedance of the transmission lines (which is a function of the width of the lines) was either too high or too low to be manufactured. Therefore for our filter design another design technique was explored. Optimum distributed highpass filters ([Hong & Lancaster 2011](#)) have been known to provide very broadband lowpass filters. As with most filter implementations based on the shorted stub structure (periodic structures) it has harmonic passbands that occur periodically. In this case rather than being undesirable it can be taken advantage of. By setting the highpass cut-on to 4.6 GHz and the upper frequency cut-off (due to periodicity) to 15.3 GHz, this highpass filter can be used as a bandpass. In order to ensure the higher harmonic passbands are suppressed, a

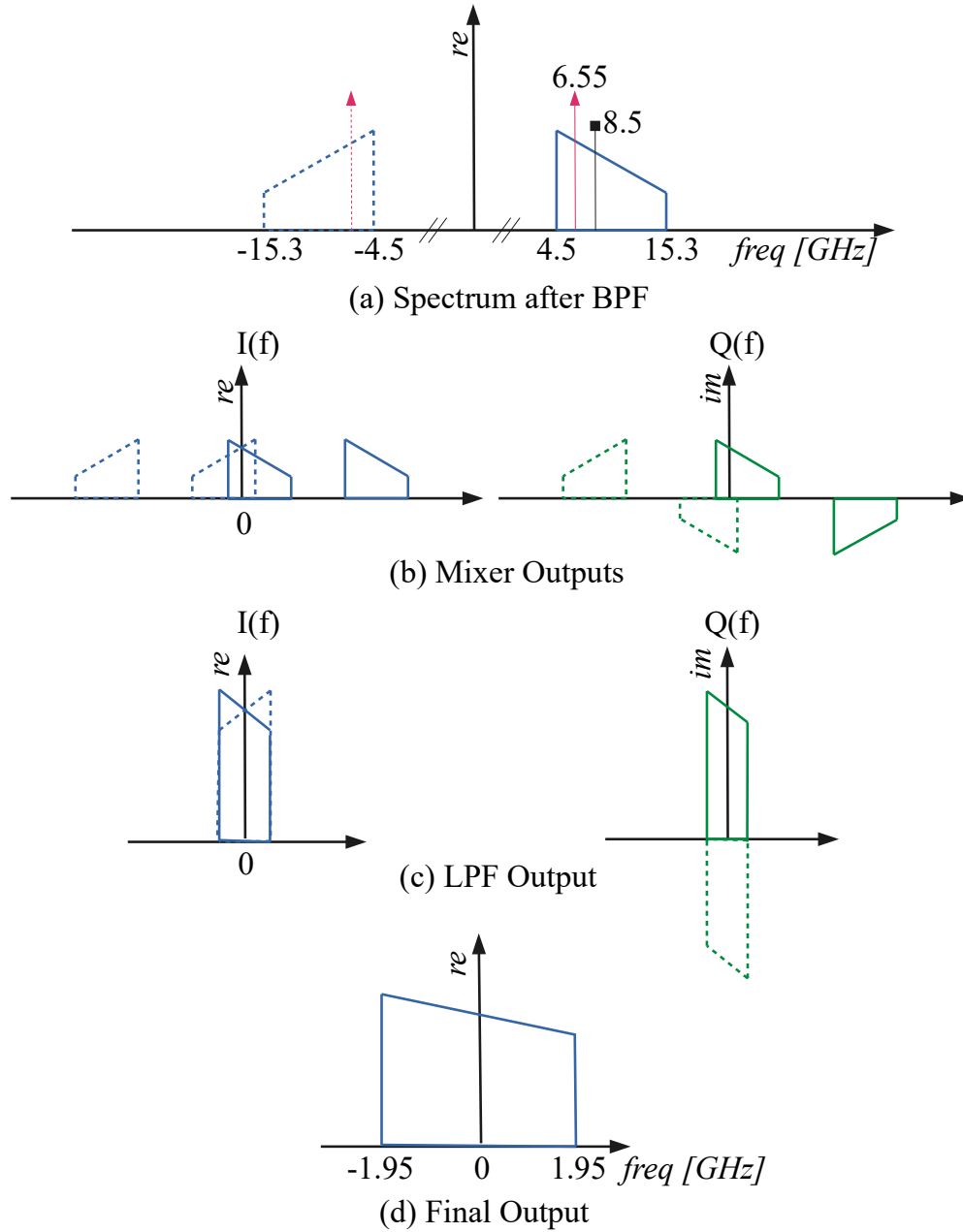
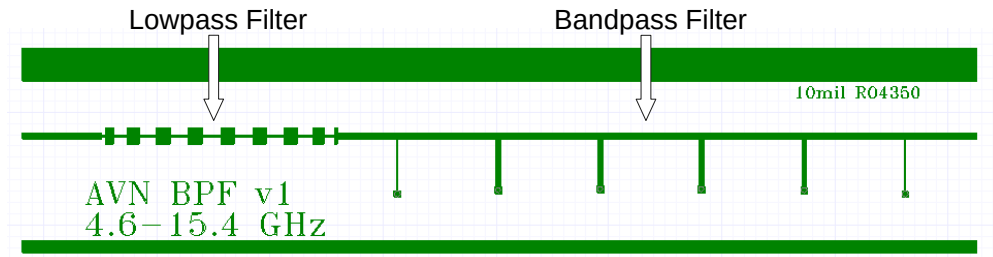
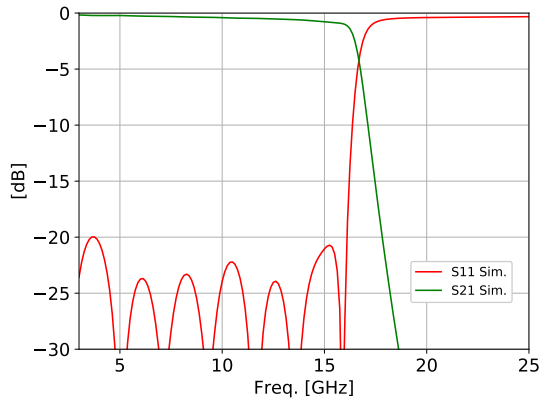


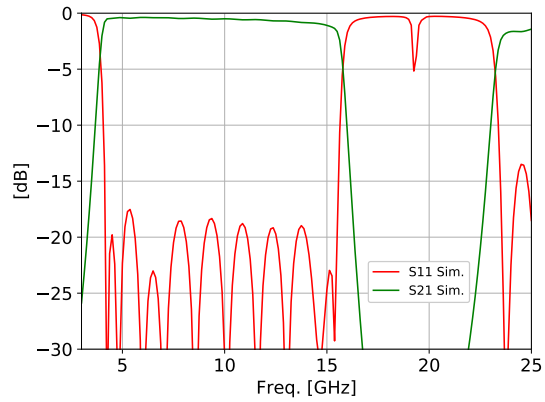
Figure 6.4: Illustration of the downconversion signal processing for each channel of the transducer. Starting from the bandpass filter to the re-combination of I and Q channels by DSP to give the desired spectrum.



(a) Microstrip implementation of the filters.



(b) Lowpass Filter Section Sim.



(c) Bandpass Filter Section Sim.

Figure 6.5: (a) Shows the simulation model of the complete bandpass filter implemented using microstrip technology (the ground plane is hidden). (b) Shows the simulation results of the lowpass section and (c) shows the simulation results of the bandpass filter section.

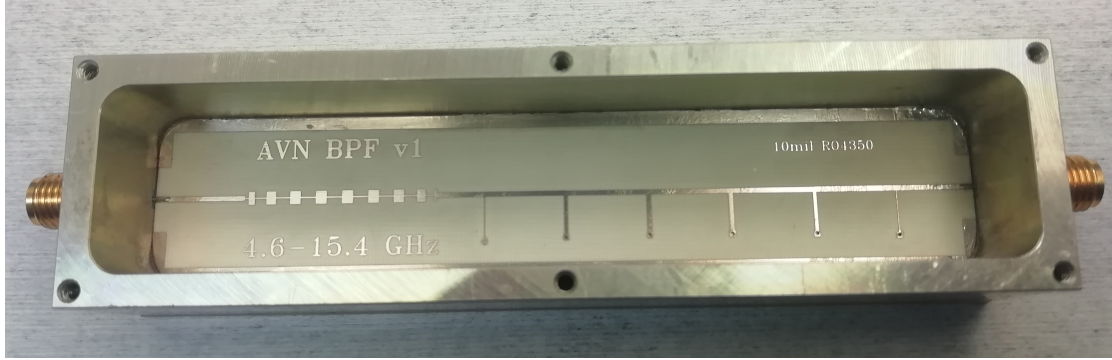


Figure 6.6: Boxed version of a complete 4.6 to 15.3 GHz prototype bandpass filter.

lowpass filter with a cut-off slightly above 15.3 GHz can be applied in series with the bandpass filter. A stepped-impedance structure is used for the low pass filter due to their ease of design. ANSYS Designer tools¹ were used for the design work. For the bandpass filter section, the procedure outlined by Hong & Lancaster (2011) was followed for the initial design. The lower cut-off frequency is used to calculate the electrical length of the stubs. Depending on the desired filter order, passband ripple and the calculated electrical length, the corresponding prototype element values are selected from a set of tabulated values. The impedance of the transmission line sections and stubs is then calculated based on the target characteristic impedance. The PCB layout and optimisation was then done in ANSYS Designer. Figure 6.5 (a) shows the final simulation model of the complete bandpass filter. Figure 6.5 (c) shows the bandpass filter section frequency response; we have a passband in the desired location, but then the first higher harmonic passband emerges just after 20 GHz. In order to suppress this and any other harmonics, the lowpass filter is designed to have a cut-off at round 16 GHz as can be seen in part (b) of the figure. Prototypes of the filter were produced on Rogers R04350 substrate. Figure 6.6 shows an image of one of the prototypes. Production of the mask was outsourced whilst the remainder of the fabrication and assembly was done in the department. The measured frequency response of the prototype filter is given along with the simulation results of the complete filter in Figure 6.7. The response for the prototype and the simulation agree

¹Ansyes Designer 8.0, <https://www.ansys.com>

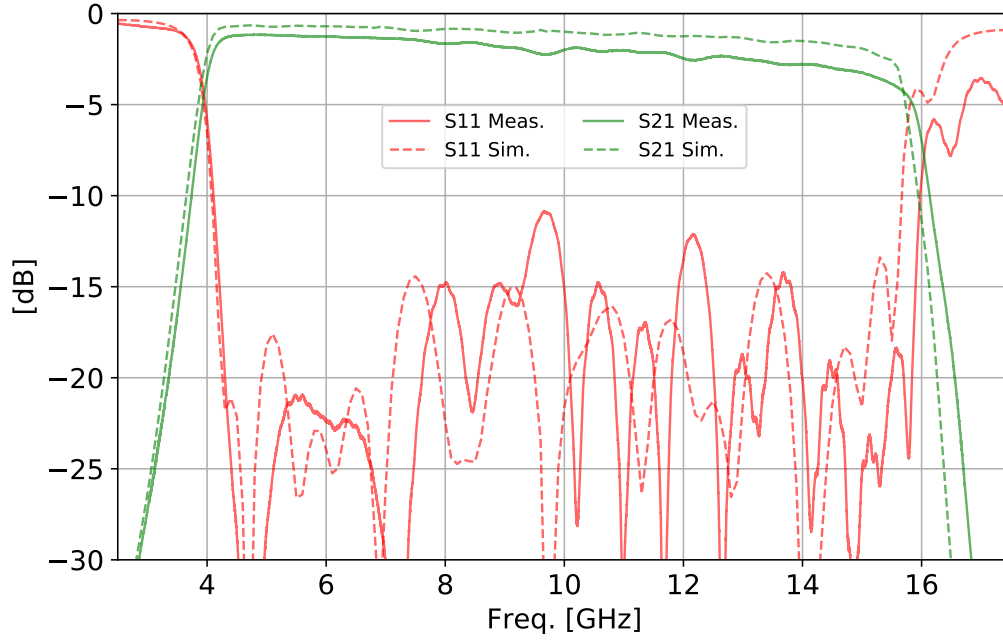
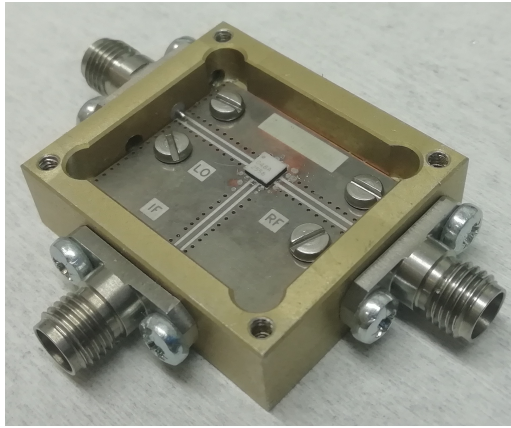


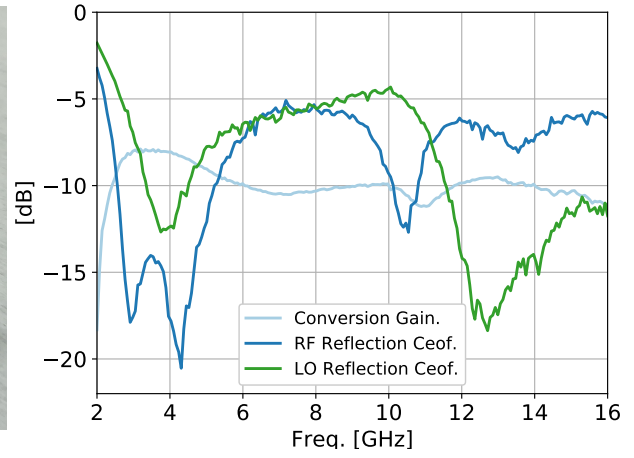
Figure 6.7: 4.6 to 15.3 GHz bandpass filter simulated and measured frequency performance.

to a large extent. There is a slight difference in the high frequency cut-off, as well as the insertion loss. The measured return loss is slightly higher but still well within the acceptable range. The correct material properties were included in the simulation therefore the difference maybe arising from the coaxial connector (or the connection between the coaxial and microstrip), as these were not simulated.

I/Q mixers have been implemented on single chips and are readily available. These devices have the splitter and 90° hybrid on the chip, and as a result only a single RF and LO input would be required. Such a device for our desired bandwidth was not available off the shelf and so the alternative was to implement the I/Q mixer using individual building blocks as shown in Figure 6.3. The appropriate 2-way splitter and hybrid were obtained off the shelf whilst the mixer package had to be fabricated and assembled in-house. The HMC1048A mixer chip from Analog Devices was selected. It is a double-balanced downconverter, which enables better port isolation (Yang et al. 2012). It provides an RF bandwidth of 2.25 to 18 GHz with IF bandwidth



(a) Prototype Mixer

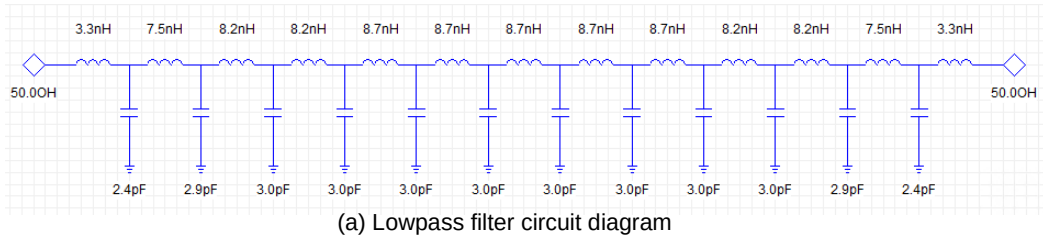


(b) Measured Conversion Gain and Reflection Coef.

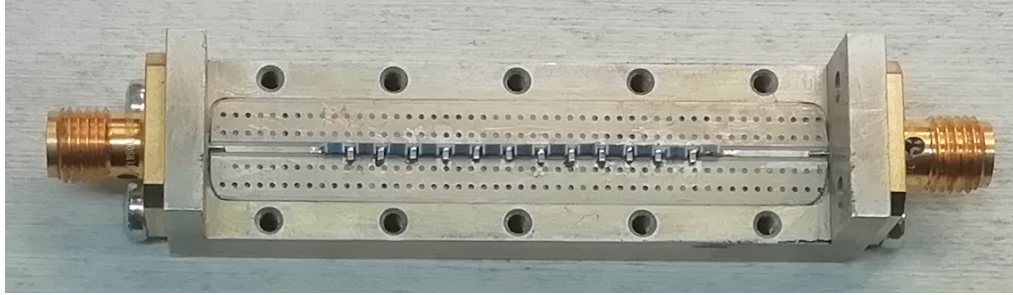
Figure 6.8: (a) shows the picture of the prototype packaged mixer and (b) shows the measured conversion gain, RF and LO reflection coefficient amplitudes.

of dc to 4 GHz at a conversion gain of about -8 dB. A PCB was designed using grounded coplanar waveguide (GCPW). Fabrication and assembly proceeded as for the bandpass filters. A picture of a prototype and the measured conversion gain, RF and IF reflection loss are given in Figure 6.8. The performance is similar to what is given in the device datasheet.

The final component to have been designed is the anti-aliasing lowpass filter, from dc to 2 GHz. The filter was designed using the tools in ANSYS Designer 8.0. The suggested component values by the software had to be tweaked by hand to achieve the desired performance, based on commercially available capacitor and inductor values. The circuit diagram is given in Figure 6.9 (a). The upper half of the filter lies in the microwave region for which the use of typical lumped components can become problematic as a result of parasitic effects. This was overcome by careful selection of the components as suggested by Copley (2014). The components need to have low parasitic resistance, inductance and capacitance over the design bandwidth, with well-characterised published component values. They should also be able to withstand the high temperatures present during soldering. For these reasons the LQW18 wire wound inductors from Murata were used. These are specified for high frequency use. For the capacitors the GJM15 variants from Murata were selected. These are capable



(a) Lowpass filter circuit diagram



(b) Lowpass filter Prototype

Figure 6.9: DC to 2 GHz Anti-aliasing Lowpass filter circuit diagram and assembled prototype.

of operating up to 10 GHz. These are all surface-mount devices and had the same package size as components in filters which were previously developed in the group. The filter was designed with an order of 25 so that an already existing PCB and enclosure designs could be used for this implementation. A picture of the prototype is given in Figure 6.9 (b). The lid has been removed to expose the populated PCB. The measurement and simulation result of the filter are shown in Figure 6.10. The measured response shows slightly elevated return loss. The roll off also starts sooner than the simulated results. With this filter we have a usable bandwidth of 1.75 GHz, using I/Q mixing this gives an aggregate bandwidth of 3.5 GHz. We suspect this discrepancy arises from component tolerance and parasitics. To maximise the available bandwidth the design could be further tweaked by taking into account the tolerance of the components and their parasitics, as well the PCB parasitics in the simulation. Nonetheless the current filter does serve the purpose for demonstrating that the system works.

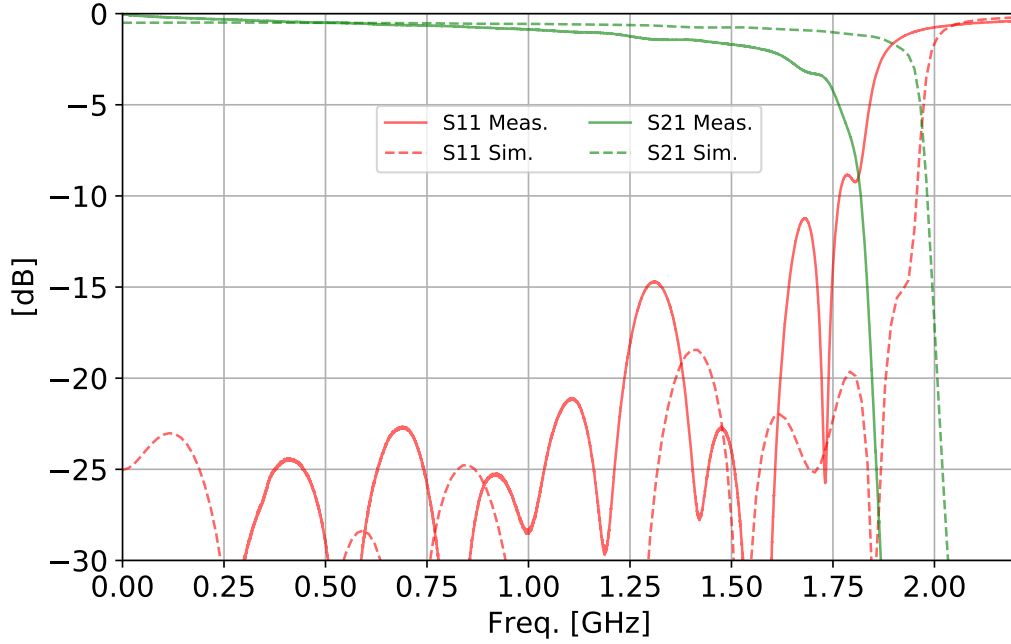


Figure 6.10: DC to 2 GHz Anti-aliasing Lowpass filter simulated and measured frequency performance.

6.3.3 receiver noise temperature

To calculate the expected receiver noise temperature T_{RX} Friis' formula for noise temperature (Equation 2.9) is used. From this we see that the later stages in the receiver chain contribute less to the overall noise temperature of the receiver and therefore generally need not be considered. Therefore in practise it is sufficient to calculate the noise temperature based on the output of the LNA. For this reason the receiver noise model given in Table 6.4 shows the contribution of each component ΔT_{RX} and the cumulative noise temperature only up to the LNA. The effects of the Ohmic losses of the primary and secondary mirrors have also been included. They are assumed to have an insertion loss of 0.25% each. This value was taken from the Detailed Design Review documentation for the SKA (Leech et al. 2017). Similarly the values of the gains for the feed, cryostat window and the cable were also obtained from there. The TRT's gain value was obtained from simulation results, whilst the LNA's gain and noise temperature were as specified by the manufacturer in the datasheet.

Table 6.4: Receiver front end noise model at 4.5 GHz. Phy/Gain Temp refers to the physical temperature for lossy devices and equivalent noise temperature for the LNA.

Component	Gain	Phy/Gain Temp	ΔT_{RX}	T_{RX}
Primary Mirror	-0.011 dB	330K	0.84K	0.84K
Secondary Mirror	-0.011 dB	330K	0.84K	1.68K
Cryostat Window	-0.020 dB	330K	1.53K	3.21K
Horn	-0.045 dB	330K	3.47K	6.68K
TRT	-0.094 dB	15K	0.33K	7.01K
Cable	-0.095 dB	15K	0.34K	7.35K
LNA	35.00 dB	2.8K	2.92K	10.27K

The LNA (LNF-LNC4_16B) has an average gain and noise temperature of 35 dB and 2.8 K respectively across a 4 to 16 GHz bandwidth. Looking at the column showing the contribution per stage we see that the greatest contributor is the feed horn. Given that the feed is at 330 K better performance can be achieved if the feed horn is cooled, as is being proposed for the SKA Mid. Some of the components have

Table 6.5: Receiver noise temperature: 4.5 to 8.5 GHz.

Freq. [GHz]	4.5	5.5	7.5	8.5
T_{RX} [K]	10.27	9.66	9.77	10.36

a noise contribution with stronger frequency dependence. As a result, the receiver noise temperature was calculated for four points in the frequency bands 4.5 to 8.5 GHz (the primary band) and 8.3 to 15.5 GHz. The components whose gain was changed depending on frequency were the LNA, Cable and TRT as this information was readily available. The other components' properties were held constant for all calculations. The results are summarised in Tables 6.5 and 6.6. For 4.5 to 8.5 GHz

Table 6.6: Receiver noise temperature: 8.3 to 15.5 GHz.

Freq. [GHz]	8.3	10.5	12.5	15.5
T_{RX} [K]	10.14	9.96	10.92	13.85

less than 1 K variation was observed whilst for 8.3 to 15.5 GHz a variation of 3.71 K is noted. The later can be attributed to the general increase in contribution with

frequency of those components but, is predominantly as a result of the LNA’s noise temperature. It hovers around 2.5 K between 4 to 8 GHz but starts to rise after 9 GHz and reaches 4.8 K at 15.5 GHz. The receiver noise temperature for the primary band does satisfy the requirement as it is less than 12 K, therefore the desired sensitivity parameter should be attainable. For the auxiliary band, band 5b on the other hand the noise temperature is higher than the specification at the top end of the band, this will most likely result in the failure to attain the desired sensitivity at the top end of the band. To verify these predictions, the receiver’s equivalent noise temperatures presented in this section are combined with the expected antenna noise temperature to determine T_{sys} and subsequently the sensitivity parameter in Section 6.5.2.

6.4 Digital signal processing

6.4.1 Digital backend hardware

The Digital Signal processing backend is based on the HTG-ZRF8 board developed by Hitechglobal ([Hitechglobal 2020](#)). Our variant is populated with a Xilinx Zynq UltraScale+ RFSoc, ZU28DR. The block diagram of the board in Figure 6.11 shows some of the main features available on the board. Eight channels of 12-bit 4.096 GSPS analogue-to-digital conversion are available via the SSMC connectors. Also 8 channels of the 14-bit 6.4 GSPS DAC are provided. PCI Express and 100G Ethernet are also supported for control and data readout. The plan is to have the board work in standalone mode so it can be mounted in a rack rather than tethered to a PC via the PCI slot. For that reason the 100G Ethernet port will be used for the communications. Figure 6.12 shows an image of one of the prototype DSP backend modules developed by a member of the group. The board has been mounted and a copper heat sink has been added. It makes thermal contact with the FPGA and base plate. The RF ports have been patched to the module’s bulkhead SMA connectors. This has also been done for the 100G Ethernet. The blue PCB allows the board to draw power from the rack’s back plane and controls the status LEDs on the front panel of the module.

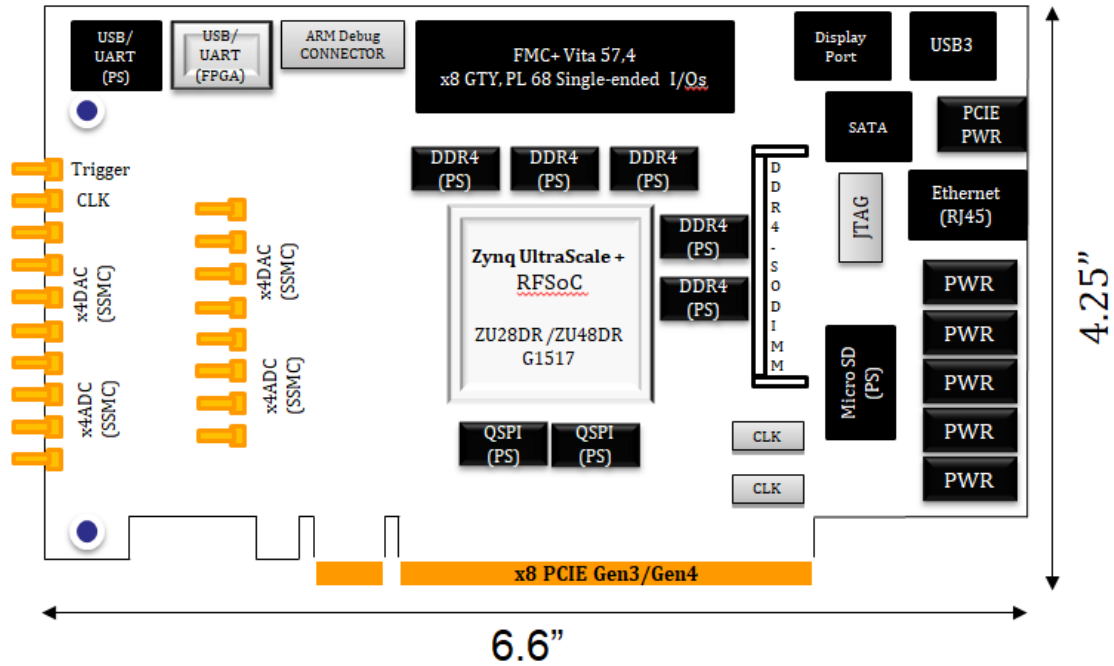


Figure 6.11: HTG-ZRF8 DSP board block diagram obtained from Hitechglobal (Hitechglobal 2020).

Some work has been carried out to verify that the ADCs provide the performance stated in the datasheet. Tests have been conducted to test spurious free dynamic range (SFDR) of the ADC. The achieved results agree with the values provided in the datasheet (Liu 2019b). A basic spectrometer has also been implemented (Liu 2019a). More capable firmware for the backend is currently under development.

6.4.2 Digital signal processing overview

Another member of the group will develop the firmware that will run on the DSP. My contribution to this was to propose a block diagram of the key function blocks that should be implemented for the AVN receiver, this is the subject of this section. For single-dish operation the final DSP firmware will implement a spectro-polarimeter. In this section we will highlight the key functional blocks. A block diagram of the major operations is given in Figure 6.13. Part (a) shows the overall signal flow. The input signals are the digitised I and Q of each of the TRT channels A, B and C. The first step

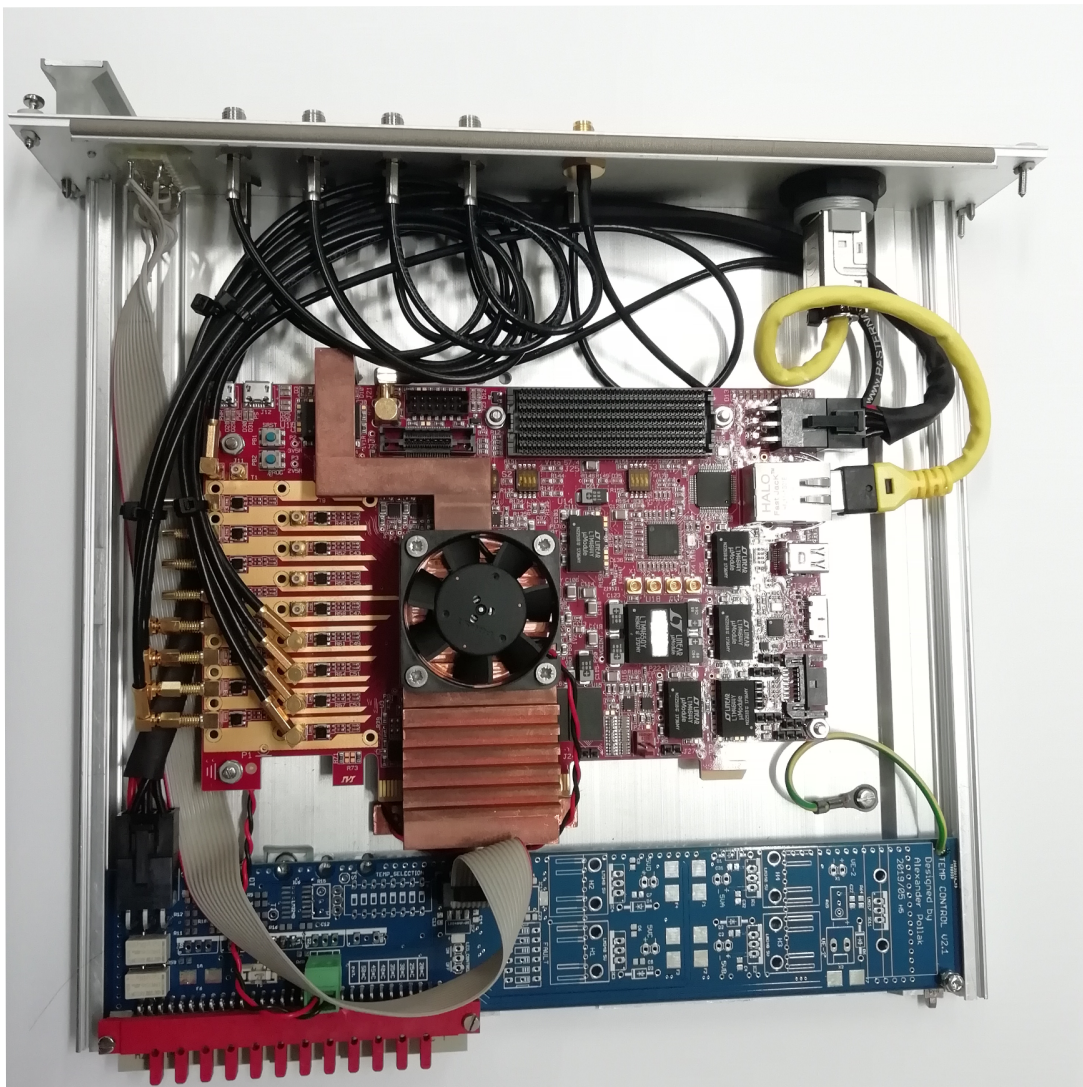
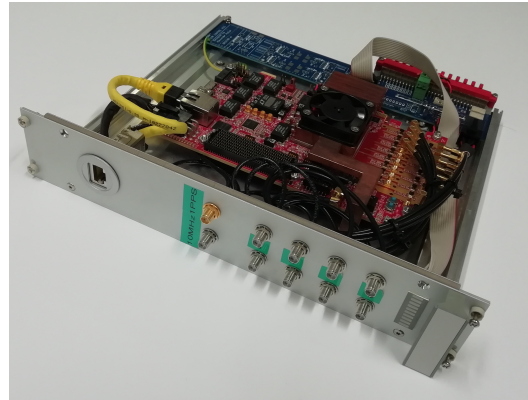


Figure 6.12: DSP Module.

is to calculate the spectrum for each of the 6 signals using polyphase filter banks (PFB) and a complex FFT. The PFB is used for its ability to offer lower spectral leakage than a simple FFT without significantly increasing computation requirements ([Price 2016](#)). The complex voltage amplitude spectra of the 6 channels are then calibrated. The block on the left in part (b) of the figure provides more detail on this computation. Each channel is calibrated individually per frequency bin by multiplying it with its specific complex calibration factor. This corrects for any gain variations introduced by the receiver chain. In the next step, I and Q channels are combined to generate the complete complex amplitude spectrum for each channel of the TRT. Extraction of the orthogonal linear polarisations is then carried out. The details of the process are given in the centre block of part (b) of Figure [6.13](#). Given that calibration also includes compensating for the imperfections in the TRT, here we can apply the ideal equations to generate V_X and V_Y from V_A , V_B and V_C (details in Chapter [4](#)). The following equations describe the operation,

$$V_X = \frac{1}{2}(V_A(f) - (V_B(f) + V_C(f))) \quad (6.5)$$

$$V_Y = \frac{1}{\sqrt{3}}(V_B(f) - V_C(f)). \quad (6.6)$$

The polarimetry function is achieved by calculating the Stokes parameters. These parameters fully describe the polarisation state of a given electromagnetic wave. They are defined as follows,

$$I(f) = |V_X(f)|^2 + |V_Y(f)|^2, \quad (6.7)$$

$$Q(f) = |V_X(f)|^2 - |V_Y(f)|^2, \quad (6.8)$$

$$U(f) = 2\text{Re}\{V_X(f)V_Y^*(f)\}, \quad (6.9)$$

$$V(f) = -2\text{Im}\{V_X(f)V_Y^*(f)\}, \quad (6.10)$$

Where $I(f)$ in this case is the total power of the incident wave, $Q(f)$ and $U(f)$ are the power in the linear polarisations and $V(f)$ is the power in circular polarisation. In the final step the Stokes parameters are accumulated before being read out.

For VLBI applications circular polarization is preferred as it provides the simplest implementation of the correlator. In this case there are two possible options in order to use this system in an already existing VLBI setup. The first option is to implement the conversion from A, B, C space to the X,Y space using RF components right after the TRT. This could then be followed by a 90° hybrid to convert linear to circular polarisation. Any imperfections could in principle be corrected for by calibrating in the DSP. The second option is to implement a circular polariser in the FPGA as presented by [Das \(2013\)](#). This polariser could be inserted after the V_X and V_Y extraction block, with the advantage of needing no additional RF components. In interferometry the Stokes calculation has to be computed after the baselines have been correlated. Therefore this computation in our DSP, along with the accumulation shown in Figure 6.13, would be replaced with the polariser and a packetiser. The 100G Ethernet will provide the necessary data bandwidth to transport the measured voltages to the correlator at full time resolution for immediate correlation, or for storage in the case of VLBI observations.

6.5 Estimation of system performance

Now that the estimated receiver noise temperature has been determined and the full optical beam patterns are available two key indicators of performance can be computed. These are the sensitivity parameter and the point-source mapping speed figure-of-merit. These are arrived at by considering the system noise temperature power in terms of the flux density this would correspond to i.e the system equivalent flux density ([Sutinjo et al. 2015](#)). This is given by

$$SEFD = 2k \frac{T_{SYS}}{A_{eff}}, \quad (6.11)$$

where A_{eff} is the effective collecting area introduced in Chapter 3, T_{SYS} is the system temperature discussed in Chapter 5, and k is the Boltzmann constant; when k is equal to 1380 the SEFD is given in units of Jansky (Jy) ($10^{-26} \text{ Wm}^{-2}\text{Hz}^{-1}$). Using

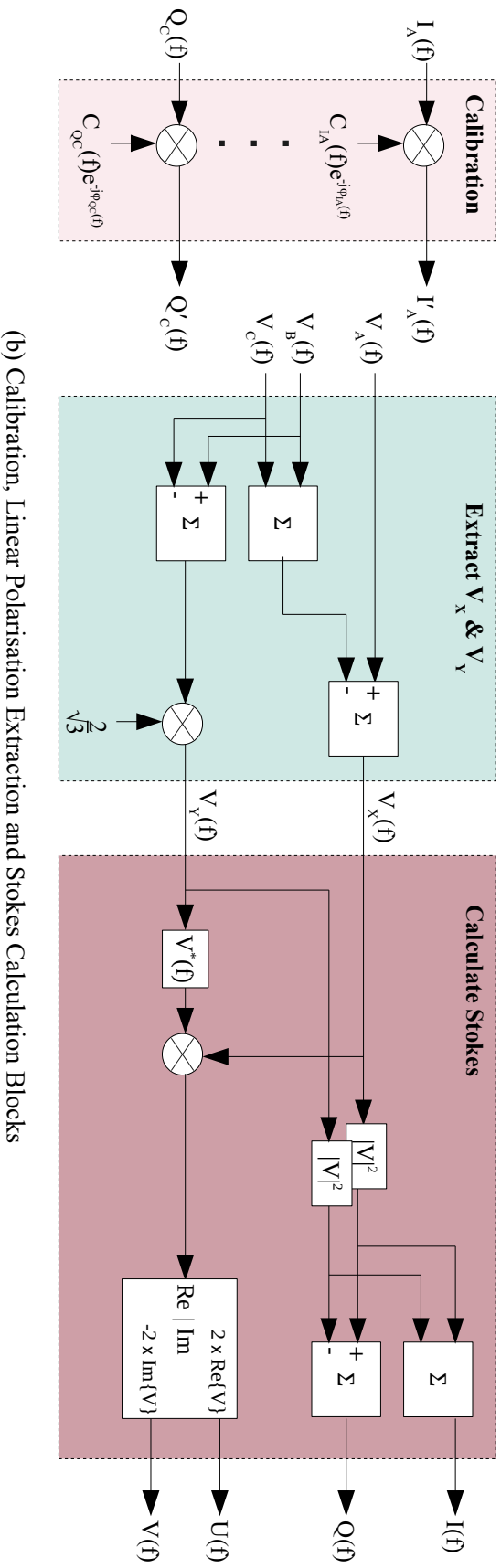
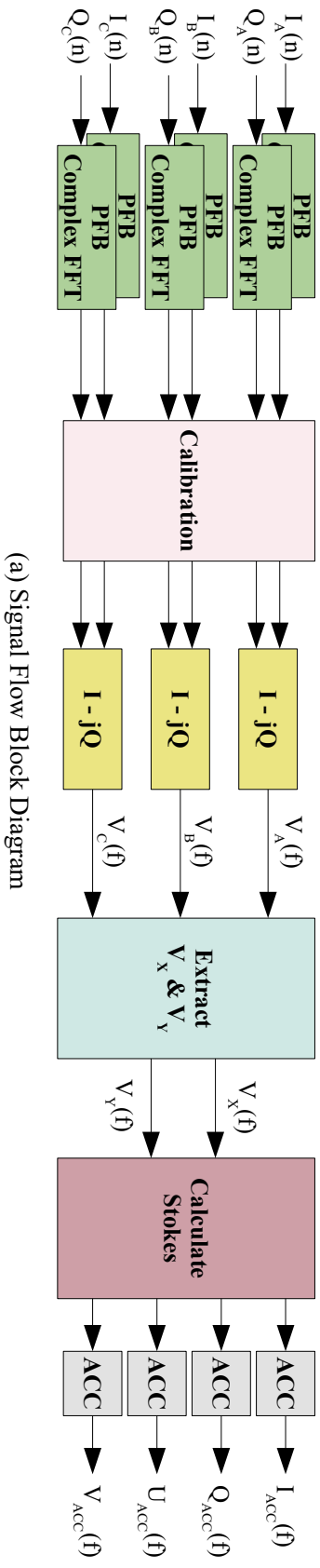


Figure 6.13: (a) Shows the overall signal flow in the digital signal processing firmware to be implemented. (b) zooms in on key function blocks, these include the calibration, linear polariser and Stokes calculation blocks.

Equation 6.11 and the previous discussion on the ideal radiometer equation the flux density equivalent of the radiometer equation is given as,

$$S_{RMS} = \frac{SEFD}{\sqrt{\tau \Delta v}}, \quad (6.12)$$

$$= \frac{2k}{\sqrt{\tau \Delta v}} \frac{T_{SYS}}{A_{eff}}, \quad (6.13)$$

where S_{RMS} is the rms variation in flux density and hence the minimum flux density that can be detected. Note that

$$S_{RMS} \propto \left(\frac{A_{eff}}{T_{SYS}} \right)^{-1}. \quad (6.14)$$

This ratio is a measure of the combined performance of the antenna and the receiver. The higher the ratio the better, as it indicates the source signal will receive greater amplification with less noise added, hence shorter integration times are needed, achieving better sensitivity. Since it is independent of the source power level and the path loss it is used as a means to compare different systems. The second figure of merit, the point-source mapping speed, is related to this. Rearranging Equation 6.13 to make τ the subject of the equation yields,

$$\tau = \frac{4k^2}{\Delta v S_{RMS}^2} \left(\frac{T_{SYS}}{A_{eff}} \right)^2. \quad (6.15)$$

Equation 6.15 gives the integration time required to observe a point source with a given sensitivity S_{RMS} . The integration time is the inverse of the mapping speed hence the point source mapping speed (M),

$$M \propto \left(\frac{A_{eff}}{T_{SYS}} \right)^2, \quad (6.16)$$

As previously discussed the system temperature is given by,

$$T_{SYS} = T_{ANT} + T_{RX}. \quad (6.17)$$

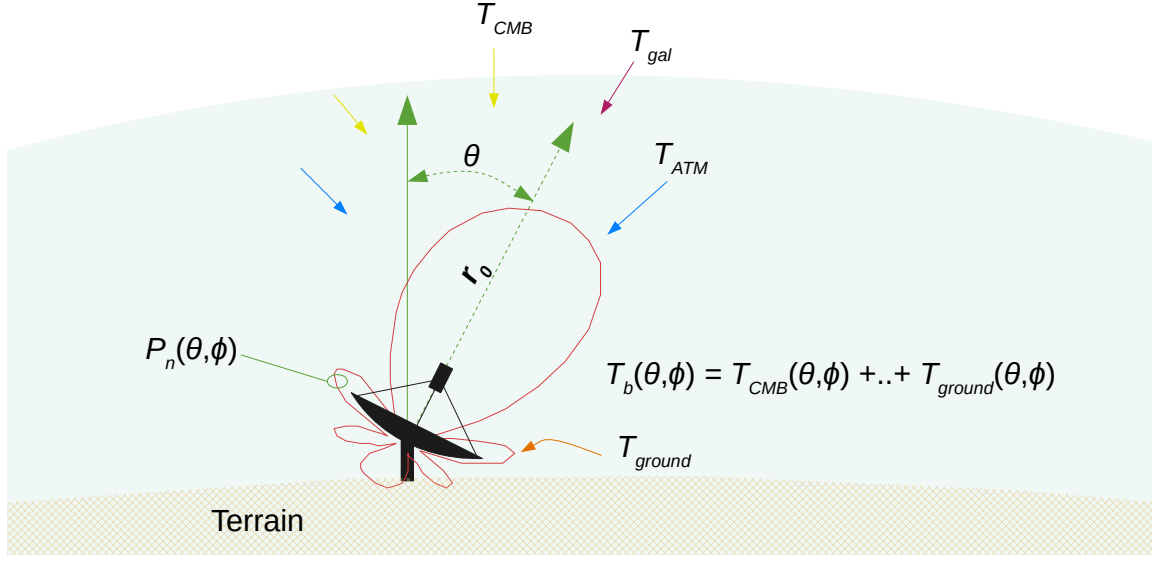


Figure 6.14: A model for antenna noise temperature calculation given the antenna's beam pattern $P_n(\theta, \phi)$ and the brightness temperature of $T_b(\theta, \phi)$ for the site. θ is the angle from Zenith and r_0 is the unit vector describing the pointing direction of the telescope.

T_{RX} was presented in Section 6.3.3. To calculate the system temperature T_{ANT} has to be determined. This is covered in the next section.

6.5.1 Computation of T_{ANT}

Other signals apart from the source also couple with the antenna's beam pattern as shown in Figure 6.14. For any given direction of pointing r_0 , and particular frequency ν , these add up to give an effective brightness temperature. This is referred to as the brightness temperature distribution $T_b(\nu, \theta, \Phi)$ for that particular site. $T_{ANT}(\nu|\hat{r}_0)$ can therefore be calculated by integrating the antenna beam pattern $P_n(\nu, \theta, \phi|\hat{r}_0)$, over the brightness distribution as shown in Equation 6.18 (Medellin 2007).

$$T_{ANT}(\nu|\hat{r}_0) = \frac{\iint_{4\pi} T_b(\nu, \theta, \phi) P_n(\nu, \theta, \phi|\hat{r}_0) \sin \theta d\theta d\phi}{\iint_{4\pi} P_n(\nu, \theta, \phi) \sin \theta d\theta d\phi} \quad (6.18)$$

Given that the target site is not yet known, the SKA-Mid site's $T_b(\nu, \theta, \Phi)$ is used to determine $T_{ANT}(\nu|\hat{r}_0)$. The brightness distribution model is described by Medellin

(2007). The model includes cosmic microwave background and galactic emissions. Also, Realistic opacity levels are included, as well as wet and dry continuum components, water, oxygen lines and the reflection of the sky's emission from the ground modelled as a dielectric. While this might vary from the final site it still gives an indication of the kind of performance that can reasonably be expected.

$T_{ANT}(v|\hat{r}_0)$ was calculated in two steps. First the sky/ground brightness for each frequency is generated using a MatLab script which implements $T_b(v, \theta)$. This generates a text file of brightness temperatures for a given elevation, in this case from 0 to 180° in steps of 0.1°. In the second step, brightness temperatures are integrated with the simulated physical optics beam patterns. The C++ program used to perform the integration was provided by Jamie Leech. The output is the antenna noise temperature for a specific angle θ from zenith (as depicted in Figure 6.14) from 0 to 90° in 10° steps. The results are given in Figure 6.15 (a). This plot also includes the contribution from T_{RX} . As expected, T_{ANT} increases with increase in θ as the antenna sidelobes start to pick up more of the ground. Part (b) of the Figure shows T_{SYS} which at zenith is around 15 K in the middle of band 5a. This degrades to around 22 K at the top of band 5b. Whilst the band 5a feed was specifically designed to operate at this frequency band 5b feed is simply a scaled version of band 5a hence no optimisation was carried out for actual operation at band 5b. In addition to the aperture efficiency not being the same at the top of band 5a as the bottom of 5b, this could also contribute to the step increase in T_{SYS} from band 5a to 5b. With regards to the general upward trend observed in the system temperature, this is caused by RF component losses and atmospheric emission increasing with frequency. Furthermore, there is a water vapour emission line at 22 GHz which starts to become more important.

6.5.2 Sensitivity and mapping speed figures of merit

With T_{SYS} having been determined, the sensitivity parameter can be calculated for each angle of elevation. The results are given in Figure 6.16. To put it into perspective, consider that for the SKA band 5, the specifications require a sensitivity greater than

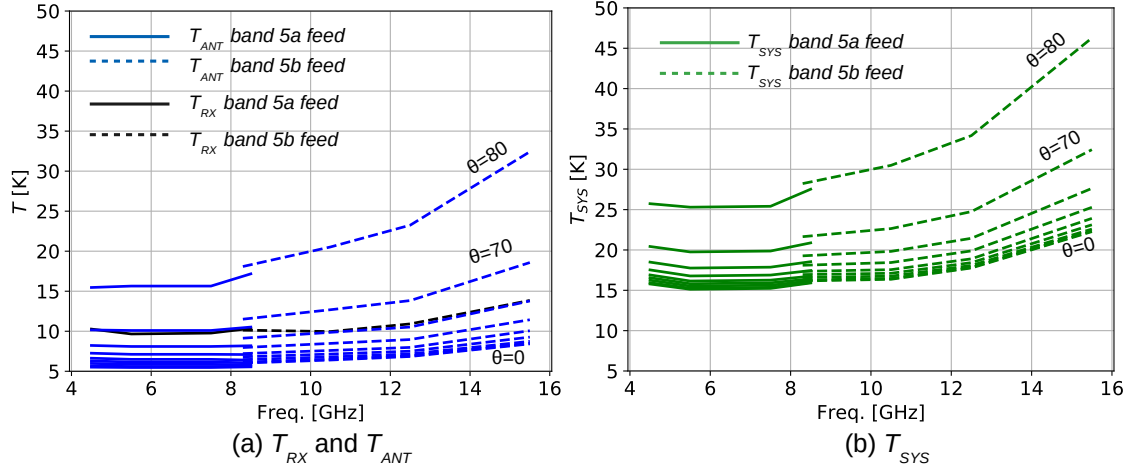


Figure 6.15: (a) Shows T_{RX} and T_{ANT} versus frequency. Each T_{ANT} curve is for a specific angle α from Zenith. They range from 0 to 80° in steps of 10° (b) Shows T_{SYS} for each angle θ from Zenith, again ranging from 0 to 80° in steps of 10°.

6.1 m² K⁻¹ for all zenith angles between 0 and 60 degrees and that it should not fall below 80% of this value (4.88 m² K⁻¹) across the band. Looking at the graph one notes that this specification has been met with reasonable margin for the primary band (5a). It has also been satisfied with the un-optimised feed for band 5b but with less of a margin; it would probably fail for the actual implementation, therefore for future upgrades it would be better to optimise the feed design. The mapping speed figure-of-merit is given in Figure 6.17. The specification for this is simply the square of A/T, which comes to 37.2 m⁴ K⁻²: the same conclusions drawn from the previous plots apply here.

6.6 Summary

In this chapter we started by presenting the three main systems that make up the entire telescope. These are the antenna, signal conditioning and control and monitoring systems. Under the antenna system we have the antenna optics and cryostat subsystems. More details were presented on the latter subsystem. The need to cool the LNA and the TRT for higher sensitivity motivated the relevance of the cryostat.

Under the signal conditioning subsystem we presented two subsystems namely,

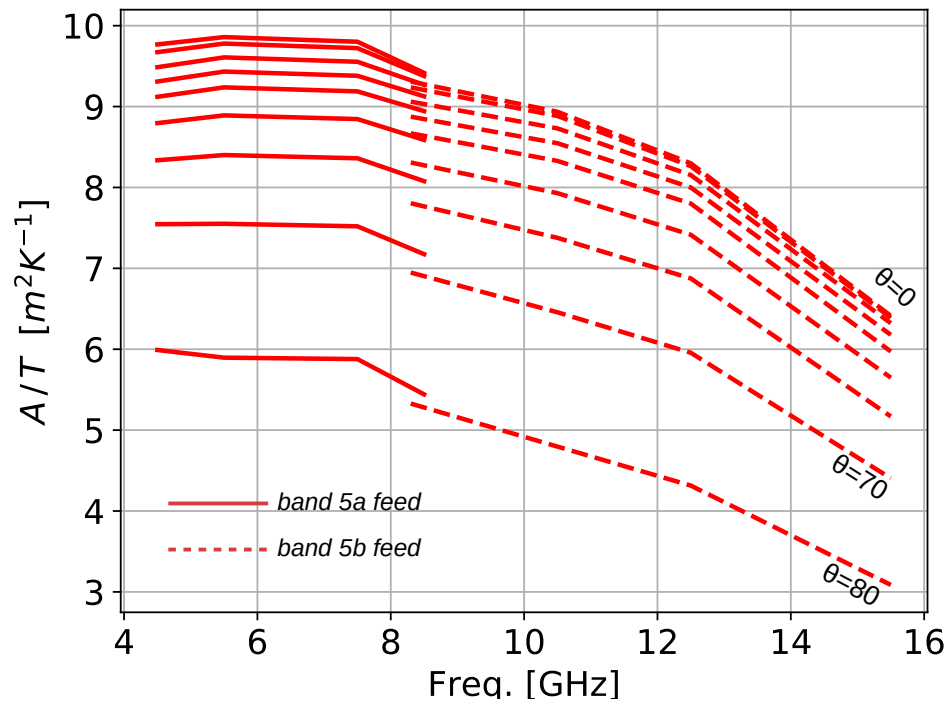


Figure 6.16: Shows $A_{\text{eff}}/T_{SY S}$ for each angle θ from Zenith. The top plot corresponds to 0° and the bottom plot corresponds to 80° . The angle increases in increments of 10° .

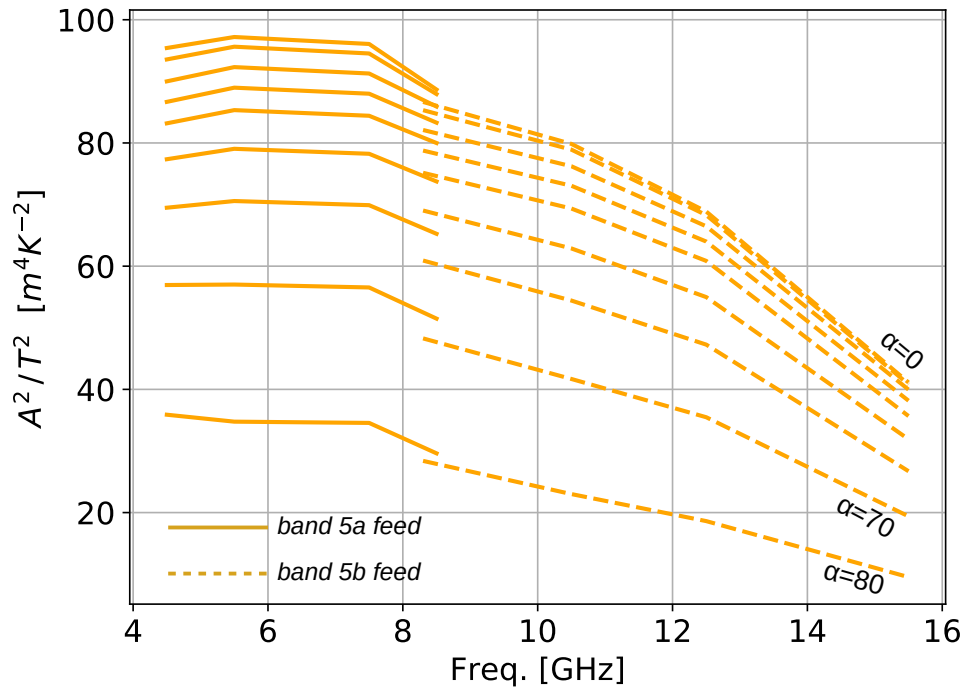


Figure 6.17: Shows the mapping speed $(A_{\text{eff}}/T_{\text{SYS}})^2$ for each angle θ from Zenith. The top plot corresponds to 0° and the bottom plot corresponds to 80° . The angle increases in increments of 10° .

the analogue signal processing and the digital signal processing subsystems. For the analogue signal processing the receiver architecture was presented followed by the design of some of the RF components. These included the bandpass filter, mixer circuit boards and anti-aliasing low pass filters. Finally under this section the receiver noise temperature was also tabulated. The DSP hardware was also presented. Given that the digital signal processing firmware is currently under development only the functional block diagram of the main operations were presented.

In the final section of this chapter we combined the work done on the feed horn and its expected coupling with the antenna optics, together with the proposed new transducer design and the expected noise performance of the receiver system to calculate an estimate of the expected overall system sensitivity and mapping speed.

Chapter 7

Conclusion

To conclude this thesis, in this chapter a summary of the work carried out thus far towards the development of a receiver system for an AVN new build is presented. Lastly, a discussion on the future work to be carried out to complete the receiver implementation is highlighted.

7.1 Summary

In this thesis I have presented the design work carried out towards developing a cryogenic 4.6 to 15.3 GHz radio telescope receiver for one of the AVN's telescopes. The scope of my contribution has included developing and prototyping a feed horn compatible with the SKA-Mid dish optics. A novel wide-band triple ridged transducer (TRT) has also been designed and prototyped. To enable observation across the entire bandwidth of the TRT a wideband homodyne receiver architecture has been proposed. In addition a novel receiver gain stabilisation technique has also been proposed and verified using experimental testing.

The presented feed was designed to give better than 80% aperture efficiency over 4.6 to 8.5 GHz with the SKA-Mid optics. The feed structure consists of corrugations which are perpendicular to the horn's flared walls to enable a circular beam illumination of the primary, thereby enabling low cross-polarisation levels. A choke was added to the outer rim of the feed to lower the sidelobes without increasing the

physical size of the feed. Two versions of the feed were prototyped; one to cover the original 4.6 to 8.5 GHz (band 5a) and another scaled down in size to enable operation between 8.3 to 15.3 GHz (band 5b). The measured return loss, though limited by the measurement setup, gives an indication that a good match to free space has been achieved. Furthermore, initial beam pattern measurements show that the simulated and actual performance agree over the main beam of the feed.

To implement the transducer a triple-ridge waveguide structure was employed. Owing to the inherent nature of this type of structure to support two degenerate modes of propagation over broader bandwidths, a bandwidth much broader than the equivalent quad ridge structure is achieved. In theory, a transducer based on this structure should provide better polarisation purity and suffer less from the "suck-out" observed in quad-ridged OMTs. The TRT prototype does offer broadband performance. In this prototype better than 3:1 bandwidth has been achieved, covering both band 5a and 5b. Expressions for generating orthogonal linear polarisations from the signals generated by the three probes have also been derived. The experimental work verified the effectiveness of the expressions. One of the two prototypes of the TRT included a built-in probe for noise injection calibration. This was also verified by measurement. This has the advantage of enabling the insertion of a calibration signal without degrading the system noise temperature caused by inserting a lossy coupler before the LNAs.

Given that the TRT performance far exceeded the initial targeted bandwidth, the receiver's electronics was designed to enable operation across both band 5a and 5b. We have employed a homodyne architecture with quadrature mixing and sampling. To cover such a broad band effectively some custom components had to be designed. These included an RF bandpass filter, a PCB for a wideband double balanced mixer chip and an IF anti-aliasing filter. With this setup, along with the selected RF-SoC DSP backend, the observation band of the telescope can be changed by simply swapping out the feed and tuning the LO to an appropriate frequency.

One other factor that contributes to system sensitivity, is the gain stability of the receiver chain. The possibility of injecting a known-white noise calibration signal

and then using cross correlation to track the changes in the receiver gain were also explored. Experimental work was used to verify this novel technique. The results show that it is possible to track the gain with SNR greater than 20 dB (and subsequently apply correction) by injecting a calibration signal much smaller than the system temperature, hence allowing for receiver sensitivity levels much closer to the ideal radiometer than most of the techniques currently in use.

In conclusion, the goals of the project have been met. The AVN receiver design work has been completed successfully. All the components under my scope of work have been designed, prototyped and tested. Including the initial investigation on the feasibility of a cross-correlation based gain stabilisation technique.

7.2 Future work

The key components of the AVN receiver have now been designed and verified. However, it still remains to complete the receiver and have it tested on a telescope. Although the AVN project has not yet acquired any SKA-mid antennas, we expect that this will happen within the next couple of years. The SKA antennas themselves will also soon be available - the first prototype of the SKA-Mid antennas is currently undergoing testing in South Africa. In the remainder of this section we list the steps that will be needed to provide an initial test receiver for the AVN, and also consider some of the longer-term developments which could follow.

7.2.1 Cryostat

The design of the cryostat has been completed, although the interruption of the corona virus pandemic prevented it being manufactured during the course of this thesis. The cryostat is based on similar design concepts to the SKA Band 5 cryostat, but is significantly smaller and simpler (as it only needs to accommodate one transducer and LNAs for a single band), so we do not anticipate any problems with the cryogenic engineering. Completion of the cryostat will allow testing of the cold TRT and feed-horn combination, and verification of the system temperature modelling presented in

Chapter 6. This will be done using hot-cold load (Y-factor) measurements, using the sky as the cold load.

7.2.2 Backend hardware

All the hardware components for the implementation of the complete backend down-conversion system exist. The RF components need to be mounted in rack-mountable modules such as the one housing the DSP hardware discussed in Chapter 6. The complete system including power supplies and monitoring assembled will also be mounted in similar modules. The final system will be housed in a rugged 19-inch card-frame, allowing it to be installed in any standard 19-inch rack assembly. This will allow it to be moved from the lab to various testing sites with minimal interfacing issues. The development of the RFSoc-based digitization and processing system is proceeding in parallel in the Oxford group, with the development of the firmware and software for the data converters (ADCs and DACs), and IP blocks for channelization, evaluation of Stokes parameters, data packetization, and high-speed communication to external networks. When this is complete, it will be relatively straightforward to implement the calibration and gain stabilization algorithms that have been developed and presented in Chapter 5 of this thesis across the full bandwidth of the receiver.

7.2.3 Antenna testing

Once the cryostat and backend are completed, it will be possible to test the functionality and reliability of the complete receiver system, including noise performance, with on-sky tests using the feeds alone. However, the final test will be to mount the receiver on an antenna. Although it has been designed specifically for the SKA antenna optics, the receiver would be compatible with other antennas with similar effective focal ratios, for example the MeerKAT antennas, or even a prime focus antenna such as the Lovell 76 m. The tests of the horn beam patterns presented in Chapter 3 of this thesis give excellent confidence that the receiver would achieve its design performance when coupled to full telescope optics.

7.2.4 Future developments

At present, radio telescope feeds are limited to around 2:1 in bandwidth, or pay a heavy penalty in feed efficiency if broadband design such as quad-ridged feed horns (QRFH) are used. The development of the ultra-high bandwidth TRT, along with a backend architecture that uses commercially-available digitization and processing hardware to provide many GHz of processed bandwidth, provides the prospect of very wide bandwidths with good performance. However, the remaining limiting factor is the feedhorn. Future developments should look to providing a feedhorn with bandwidth matched to the TRT. This can be accomplished in several ways, depending on the requirements of the telescope optics. For low focal ratio designs such as the SKA, a triple-ridged analogue to the QRFH could be investigated. For higher f/D ratio optics that require higher gain feeds, variations on the corrugated feed design could be investigated, using features known to improve their bandwidth such as ring-loaded corrugations or dielectric inserts in the feedhorn.

The use of coherent calibration techniques, as presented in Chapter 5 of this thesis, is also capable of much further development. The availability of DACs of equivalent bandwidth to the ADCs, tightly coupled to the signal processing system, means that arbitrary calibration signals can be generated and processed coherently. This offers the possibility of coherent calibration across large arrays independently of observing astronomical sources, as well as accurate bandpass and phase calibration on individual antennas.

Appendix A

AVN and SKA-SA Partner Countries

A list of the 8 partner countries that supported South Africa's bid to host the SKA:

1. Ghana
2. Mauritius
3. Kenya
4. Zambia
5. Madagascar
6. Mozambique
7. Botswana
8. Namibia

Appendix B

TRT Gap Machining

Two prototypes of the TRT were manufactured. The first prototype has the noise coupling probes built into it whilst the second prototype does not. Each of them were machined in 3 parts and bolted together. Each section was labelled with two letters (A to F) on each bolting flange to ensure they were bolted correctly each time they were being assembled. Since the EDM machine could not be used to cut out the desired ridge gap, 5 axis machining was also used to successively cut away $100\ \mu m$ until the gap was as close as possible to the design width of $433.01\ \mu m$. Between each pass the ridge gap was measured using a microscope to monitor the spacing. Figure B.1 shows the final measurements of the gaps for the first prototype. The

Table B.1: TRT first prototype final gap measurement.

Meas. No.	Ridge Pair [μm]		
	AA	BB	CC
1	245.77	216.61	249.39
2	286.88	323.65	343.75
3	442.95	406.20	456.16

measurements carried out between each pass are shown in Table B.1. The last row corresponds to the final measurement of the first prototype's ridge gap. The image of the final measurements for the second prototype's ridge gap are given in Figure B.2 and the successive measurements are shown in Table B.2.

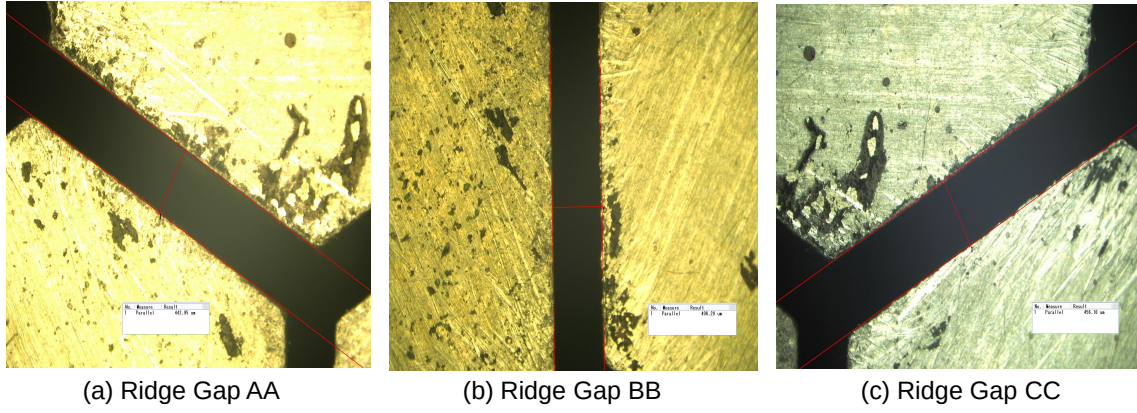


Figure B.1: TRT first prototype final gap measurement.

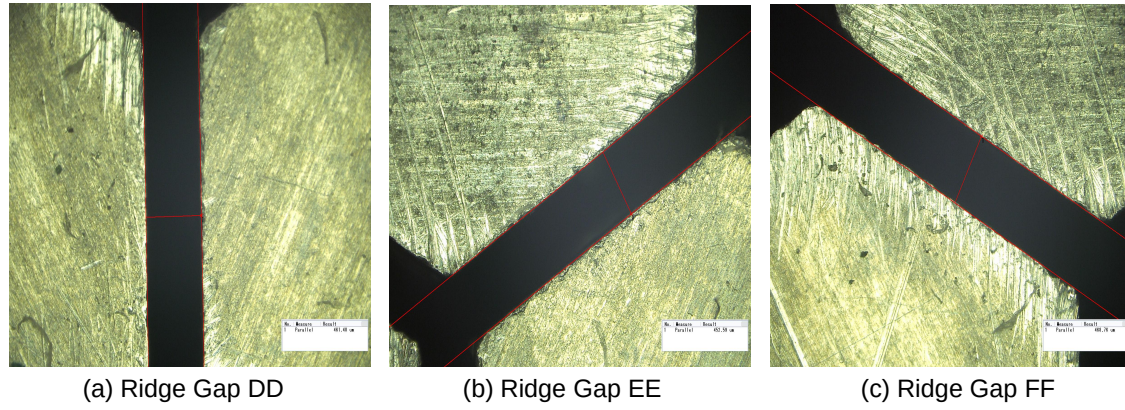


Figure B.2: TRT second prototype final gap measurement.

Table B.2: Measurements of ridge gaps for prototype 2.

Meas. No.	Ridge Pair [μm]		
	DD	EE	FF
1	288.05	246.60	243.53
2	327.60	305.76	363.35
3	461.40	452.59	468.76

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