EISCAT_3D
EISCAT_3D Radar Receiver/Antenna Subsystem Report
EISLAB, LTU

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1 Overview

1.1 Introduction

This document describes the result from work package WP4, phased array receivers, in the Eiscat 3D project. The presentation is intended to provide enough information to allow future development and evolution of the systems. Design requirements, limitations, trade-offs and solutions are presented for a number of separate system components.

The work in WP4 has been performed by staff at Luleå University of Technology (LTU) and EISCAT scientific organization. Staff from LTU has included senior researchers and Ph.D. candidates. A number of M.Sc. students have performed their final thesis work as part of the project, and thereby made valuable contributions.

1.2 Work package disposition

In the original Eiscat 3D proposal, the responsibility of the work package was defined to include research and development of three separate component parts in the receiver chain. These were:

- Antenna element
- Receiver front-end
- Time synchronization of antenna elements

The work in WP4 has since progressed on these three tracks, with a number of sub-activities in each track. The sections below give an overview of this work, as well as comments on changes to the initial project specification document.

1.2.1 Antenna element

At the onset, WP4 was outlined to perform work on the antenna element as well as on the antenna array design. It was however early on identified that the array design had large overlap with work packages 3 and 6. It was therefore recognized that the results from the active array design work performed in WP3 and WP6 would be applicable also to the design of the receive-only arrays. It was also recognized that the performance of a fully developed receiving array would not be needed to validate the concepts and hardware solutions being worked on in WP4 and WP5; a simpler array, steerable in one dimension only, would suffice.

Thus, the array-design part of WP4 was de-scoped into the construction of a simple, inexpensive and flexible antenna array for the EISCAT VHF frequency, 224 MHz, using commercially available, high-gain but narrowband off-the-shelf X Yagi antennas as array elements. This array would accommodate at least ten receivers to allow for digital beam-steering in a vertical plane and be sufficiently large to deliver a reasonably good signal-to-noise ratio (> 5%) when receiving incoherent-scatter signals from the ionosphere. It would be used during the 3D project as a test-bed on which to verify both receiver performance as well as a number of other mission-critical concepts (e.g. the digital beam-steering and beam-forming).
The work on the antenna element has been performed largely according to the original plan. The coupling effects between antenna elements in the array influences both receiver array and transmitter array with regard to performance and specifications in the beam forming. The interdependence can also be used actively as a means to support and add information to whichever system is used as the main array calibration system. The work on the Element antenna has specifically investigated and simulated the cross coupling effects between two antennas in order to evaluate the possibility to use this in a calibration purpose.

The effect of snow cover on antenna performance is crucial for the placement north of the Arctic circle. The work on the Element antenna has investigated the performance of the antenna elements, crossed yagi antennas through measurements and simulations. The results shows that during snowfall the performance of the antenna is degraded, and under severe conditions the antenna becomes non-operational. To guarantee operability of the system, the effect of snow cover should be taken into account when designing the final antenna.

Precipitation in the form of snow or rain can also severely degrade the performance of large antenna arrays, in particular if knowledge about the beam shape and pointing direction in absolute numbers is necessary. A method of estimating the far-field of each individual antenna element using the equivalent electric current approach has therefore been developed. Simulation results show that the far-field can be estimated even if the test signals used are noisy.

The work performed on the antenna element is covered in chapter 2 in this report.

1.2.2 Receiver front end

The performance of the receiver front end is crucial to achieve the desired capability in the beam steered array. Specifically, demands on inter-element timing set the beam widths and angle precision, and noise levels set the possibility to identify targets.

The initial plan had emphasis on the development of an ASIC solution for the receiver front end. Following an initial study, the decision was taken to abandon the ASIC track and focus on a solution where commercially available components are used. The main reasons for the decision were performance requirements and the relatively low frequency involved. On chip RF designs are commonly targeting higher frequencies than the Eiscat 3D system. Thus, the implementation of on chip filters and low noise amplifiers in this frequency region would require a substantial effort, with ensuing high risk. Also the requirements on the ADC are reaching available state-of the art, which renders an ASIC implementation insecure. A final consideration in the decision was the desire to keep a flexibility in the design, which by necessity will be under evolution throughout the development and into a possible implementation phase for the Eiscat 3D system.

With this decision taken, the work on the receiver front end has developed a low noise, under-sampling antenna front end with calibration possibilities in both time and amplitude. The design is centered around two main building blocks: a high performance Low Noise Amplifier (LNA) board with integrated couplers and switches for calibration signal injection, and a 16-bit, 80 Ms ADC board. The system involves active temperature control of the LNA, as well as feedback control of power supply voltages. An extensive control system has been designed to allow remote control and supervision.

An extensive description of the developed hardware is found in chapter 3 in this report.
1.2.3 Time synchronization

A crucial constraint in a direct sampled large-area array like the Eiscat 3D is the inter-element timing, which is dictated by requirements on beam pointing accuracy. The beam steering will be performed by Fractional Sample Delay (FSD), which must be able to delay a 30 MHz wide signal band 1/1024th of a sample without introducing phase shifts, and it must all be done in base-band. Also the calibration in regard of amplitude is critical for the performance of the array.

Two viable solutions to solve the timing distribution have been investigated. The first system is a cable-based system that continuously evaluates the delay between different parts of the timing system, thereby keeping track of any shifts in delay through the system. The second one is based on Global Navigation Satellite Systems (GNSS), which tracks the difference in phase of the signals from satellites to the timing units placed in the center of each sub-array. Both solutions have been evaluated in simulations as well as in measurements. It is shown that they have the capability to match the requirements, but both systems have their strong as well as weak points. Both systems have the capability to perform also calibration of amplitude from each antenna element receiver chain. The cable based system has been evaluated in field trials. The work performed on the timing and calibration system is described in chapter 4 in this report.

Directly coupled to the timing requirements is also the beam forming algorithms and filters used to achieve these. To evaluate the beam forming performance, and its connection to performance degrading aspects such as noise, jitter, bandwidth, an extensive Matlab simulation system has been designed. The simulation tool models each part of the physical system separately so that parts can be added, updated and removed without affecting the rest of the system. Noise and non-ideal timing performance is included in the simulation. The Matlab simulation tool has allowed the verification and evaluation of the proposed algorithms and timing requirements, as further described in chapter 5 in this report.

1.3 Dissemination of knowledge

The performed work has rendered a number of publications in international conferences and journals, as well as resulted in academic publications for M.Sc., licentiate and Ph.D. degree. Also Eiscat workshops have been attended. Below is a complete list of published, submitted, and upcoming work resulting from work package WP4.

1.3.1 Journal publications

- “A Pico-Second Level Cable-Based Calibration System for Large Aperture Array Radars”, G. Stenberg, J. Borg, and J. Johansson, To be submitted.


- “A Measurement System for the Complex Far-Field of Physically Large Antenna Arrays

1.3.2 Conference publications
- “Picosecond Level Error Detection using PCA in the Hardware Timing Systems for the EISCAT 3D LAAR”, G. Johansson, F. Hägglund, and J. Carlsson, To be published
- “Temperature stabilization of electronics module”, A. Gabert, J. Borg, J. Johansson, IMAPS Nordic Annual Conference 2006
- “Low-complex maximum likelihood estimator for the linear model” J. Ståbis and M. Lundberg-Nordenvaad, To be published

1.3.3 Doctoral thesis

1.3.4 Licentiate thesis
- “Advancement of atmospheric research tools”, G. Stenberg, 2006

1.3.5 M.Sc. thesis
- “Low-Noise Amplifier Design and Optimization”, M. Edvall, 2009
- “Temperature stabilization of electronics module”, A. Gabert, 2006
- “RF front end for atmospheric radar”, K. Söderström, 2007 (in writing)

1.3.6 EISCAT workshops and user meetings
- Kiruna 2006
- Åland 2007
- Uppsala 2009
2 Antennas

Antennas operating in an arctic environment may have their properties degraded significantly due to snow falling on the antennas. This could cause severe problems in applications where knowledge about the gain and pointing direction of the main beam of the antenna array is needed in absolute numbers. In Section 2.1 the performance of the antenna elements during snowfall is assessed both individually and in an array configuration. Some of the material presented in this section have previously been published in [1]. In Section 2.2 a measurement system for the EISCAT_3D antenna array is proposed, this section is a reformatted version of [2].

2.1 Environmental Effects on the ESCAT_3D Antenna Array

The performance of antennas during different weather conditions, particularly snowfall, have received significant attention. The effect a snow cover have on the input impedance of an antenna element in an array was studied experimentally in [3] where an array of dipoles was cover with snow layers of different depths. It was here found that the performance of the antenna array was degraded successively with increasing depth of the snow layer to the point of non-operational when the antenna array was completely covered. The effect of snow accretion on reflector antennas have long been of interest in the nordic countries. This was studied in [4] where it was found that the beam shape and pointing direction can be severely distorted under such conditions.

The investigation presented in this section is limited to the effect of a snow cover on the reflection coefficient of one antenna element and the mutual coupling between antenna elements. The reflection coefficient was measured during a snowfall and compared to meteorological data from the same period. The measured results are also compared with simulations. The mutual coupling was measured without any snow and the effect of a potential snowcover was assessed by simulations.

2.1.1 Dielectric Properties of Snow

Most work on determining the dielectric properties of snow has been done by the remote sensing community, where the main interest is in the permittivity, which is important when determining the refractive index and reflection coefficient. In [5] a theoretical model of the permittivity of snow is presented with the snow being modeled as a mixture of ice grains, water, and air. For dry snow the single most important factor affecting the permittivity is the density [6]. The permittivity of snow is complex, although the imaginary part is often neglected [7] since it is considerably smaller than the real part. It is, however, included here for completeness. A more comprehensive model including both the real and imaginary parts and their dependence on factors such as frequency, density, water content, temperature, and pollution effects is presented in [8]. So far only the permittivity of snow has been considered although limited information on the conductivity can be found in the literature [9]. The permeability can be assumed to be very close to unity.

In this paper, the interest is in the performance of snow covered antennas and only the permittivity is considered. The real part of the permittivity depends mainly on the density and
water content, although both temperature and pollution have a certain effect. The latter factors are, as mentioned above, included in the model presented in [8] but will be neglected here for simplicity. The real part of the relative permittivity, $\varepsilon_s$, of snow is in [8] given by

$$\text{Re}\{\varepsilon_s\} = 1 + 1.7\rho_d + 0.7\rho_d^2 + 8.7W + 70W^2,$$

(2.1)

where $\rho_d$ is the relative density of snow compared to water, and $W$ is the water content by volume. For the imaginary part, only the frequency and water content is of importance, and this is given by

$$\text{Im}\{\varepsilon_s\} = f_{10^9} (0.9W + 7.5W^2),$$

(2.2)

where $f$ is the frequency. The real and imaginary part of the relative permittivity of snow at 224 MHz is shown in Fig. 2.1 as a function of density and water content. It can be seen that the real part has a maximum of about 3.8 for the highest density and water content while the imaginary part is significantly lower with a maximum of 0.38. It should be noted the most heavy and wet snow is most likely to stick to surfaces and objects why this is also of most interest here.

2.1.2 The effect of Snow on a Single Antenna Element

In this Section the antenna’s performance when influenced by snow is studied in more detail. The antenna is a crossed yagi-antenna consisting of a feed element, which is a folded dipole, a reflector element, and three parasitic elements. The polarizations have an offset of about 0.33 m (corresponding to a phase-shift of approximately 90° at 224 MHz. Here, only one of the polarizations have been studied. It is assumed that the supporting bar have a negligible effect on the properties of the antenna.

2.1.2.1 Simulations

In the simulation the snow has been modeled as a complex dielectric medium as described above assuming a density of 600 kg/m$^3$ and a water content of 10 % (by volume). The real
and imaginary parts of the relative permittivity is then 3.7 and 0.037, respectively. This snow is heavier and wetter than the snow that typically falls during winter but it should have the most severe effect on the antennas as it have a large permittivity. It is also expected that this is the type of snow that would stick most efficiently to the antennas.

In Fig. 2.2 the simulated return loss is shown for bare antenna and the antenna covered with 0.5 mm and 1 mm of snow uniformly distributed on all conducting parts. The operating band of the antenna is shifted significantly down in frequency even for thin layers of snow. This is expected behavior for an antenna covered with a dielectric material with large real part of the permittivity [10]. In reality the snow will be distributed on one wine side of the wires only why the simulations gives a somewhat pessimistic estimation.

2.1.2.2 Measurements

The return loss of the antenna was measured on January 18-21, 2008, in Luleā, Sweden. During this period, a considerable amount of snow fell and the temperature oscillated around 0°C which resulted in a very heavy and wet snow/rain mix. In Fig. 2.3 the return loss of the antenna is shown for a frequency of 224 MHz (top right). It is also shown as a function of frequency (left) for two selected occasions marked in the plots to the right. In addition, the center frequency and bandwidth of the antenna is shown as well as the measured precipitation at a meteorological station in Luleå (this data was provided by the Swedish Meteorological and Hydrological Institute). Before any snow fell on the antenna the bandwidth was 6 MHz and the band was centered around 224 MHz. At the onset of the snow (close to midnight on January 19) the return loss is increased dramatically and the antenna is non-operational during this period. The reason for the increased loss is that the whole band of the antenna is shifted downward in frequency until it is entirely outside of the desired band. This behavior can be seen in the plots of the center frequency and bandwidth. In the plot of precipitation there seem to be a discrepancy between the start of the snow as indicated by the antenna measurements and the measured precipitation. This is most likely due to the distance between the meteorological station and the location of the antenna measurements.

Possible solutions to the problem are to place the antenna in a radome. This is, however, not practical in the EISCAT radar due to the physical size of the antenna. For the same reason
heating the antenna elements is not an option. To be able to have continuous operation of the antenna it should therefore be designed with a larger bandwidth than needed in order to be able to handle both the shift in frequency and the narrowing of the operational band.

2.1.3 Changes in the Mutual Coupling due to Snow Cover

Mutual coupling between antenna elements in an array can cause the amplitude of the far-field pattern of the whole array to be reduced in certain directions (referred to as scan blindness). Normally the antenna is designed to minimize these effects but during snowfall the properties of the antenna changes. It is therefore necessary to assess these effects before the design process of the full antenna array.

Measurement performed on the test-array outside Kiruna, Sweden, showed a mutual coupling of about -40 dB between adjacent antenna elements. During these measurements the antennas were free from snow. The mutual coupling have also been simulated for a four-by-four antenna array with the same dimensions as the test array. The simulations were done without snow on the antennas, with 0.25 mm of snow, and 0.5 mm of snow. The snow was
assumed to be distributed uniformly over the whole array. In all cases the coupling coefficient of less than -35 dB. Thus, the risk of scan blindness does not seem to increase due to snowfall. It should however be noted that these results are valid for the antenna elements used in the test array only. It is likely that the antenna elements used in the full-size array will be different. The coupling effects under different conditions (i.e. snowfall) should therefore be considered during the design process of both the antenna elements and the antenna array.

A particular yagi-antenna in an antenna array may also excite passive elements on another yagi-antenna. This will affect the far-field of the antenna elements. During snowfall it is likely that both the phase and the amplitude of these excitations will change which could alter the amplitude and phase of the antenna elements. This could potentially result in an error in the pointing direction of the beam formed by the whole array. The measurement method proposed in the next section will be capable of detecting these changes and therefore the availability of the antenna array will not be reduced.

2.2 Proposed Measurement System for the EISCAT_3D Antenna Array

In this section a measurement system using the equivalent electric current method for the EISCAT_3D antenna array is described. To reduce the size of the matrices in the system of equations to be solved the near-field of each antenna element is measured separately, with the probes used as test transmitters (the term probe will still be used throughout the text for simplicity). This is possible since all elements will have a separate front-end. The method presented has been simulated in order to illustrate the performance of the approach.

Since the radar will operate continuously the far-field of the antenna elements will need to be measured regularly. In particular, the measurement system must be able to detect any changes in the phase characteristics. The following limiting factors of the measurement system have been identified due to the specific application of ionospheric radar and the physical and electrical size of the system:

- The signals of interest to the users of the system are very weak. The probes used for the measurements must therefore be located outside the field-of-view of the antenna array in order to minimize interference and diffraction effects that could reduce the performance of the radar.

- The beamforming of the antenna array will be digital with each antenna element sampled individually. With up to 16000 antenna elements it is essential to keep the calculations to a minimum. Thus, the number of probes used for the measurements must be minimized.

- It is not practical to place the probes in the far-field of the antenna array due to the physical size of the antenna. Hence, the calibration system must be able to accurately measure the phase of the far-field with probes located at varying distances from the antenna elements.

As a result of the limiting factors mentioned above some of the conventional measurement methods may not be suitable for this type of antenna system. The far-field pattern of large aperture antennas is traditionally measured using radio sources such as quasars or distant galaxies [11]. With this approach it is possible to get accurate gain and phase characteristics of the antenna. The typical signal strength is, however, too low to be detected by the individual antenna
elements in an array. A similar technique, described in [12], uses spacecrafts as receivers for signals transmitted by the Antenna Under Test (AUT). To apply this to the EISCAT_3D radar, the spacecraft would need to act as the transmitter since not all sites will have transmitting capabilities. A major drawback of such a calibration system is that it is unlikely that a spacecraft can be dedicated to the radar and the calibration of the radar would therefore rely on signals transmitted for other purposes (signals-of-opportunity). Although this is not acceptable as the only calibration system such signals could be a valuable complement to signals transmitted by probes located in the near-field of the antenna array.

The far-field of an antenna can also be estimated using measurements in the near-field. This typically requires more calculations than the far-field techniques. A comprehensive introduction to near-field measurement techniques can be found in [13]. The methods described assume that all measured points lie on the same surface (planar, cylindrical, or spherical) within a fraction of a wavelength. This may not be possible to achieve in the EISCAT_3D radar due to the size of the system. The method proposed is instead based on the equivalent current approach where the antenna is replaced by an equivalent electric and/or magnetic current as described in [14]. A clear advantage is that the equivalent current approach is less sensitive to non-ideal probe positions although it is computationally heavier. It also suffers from ill-conditioned matrices. This problem have recently been assessed and regularization techniques was proposed as a solution in [15] and [16]. It was shown in [17] that the current measurement technique gives accurate results when estimating the far-field of large antenna arrays.

2.2.1 The Measured Electric Field

2.2.1.1 The Electric Field due to the Current Distribution on One Antenna

The electric field \( E \) at a point \( r \) in the near-field of an antenna with an electric current distribution can be calculated [18] using

\[
E(r) = C_k \eta \int_{S'} \left[ J C_{N1} - (J \cdot \hat{R}) \hat{R} C_{N2} \right] e^{-jkR/R} dS'
\]  

(2.3)

with

\[
C_k = -\frac{jk}{4\pi} \quad C_{N1} = 1 + \frac{1}{jkR} - \frac{1}{(kR)^2} \quad C_{N2} = 1 + \frac{3}{jkR} - \frac{3}{(kR)^2}
\]  

(2.4)

where \( S' \) is the surface of the antenna, \( J = J(r') \) is the electric current at the point \( r' \) on the antenna, \( R = |r - r'|, \hat{R} = (r - r')/|r - r'|, \) and \( \eta = \sqrt{\mu_0/\varepsilon_0} \approx 377 \) ohms is the wave impedance in free space.

The current distribution can be expanded using basis functions according to

\[
J(r') = \sum_{n=1}^{N} a_n f_n(r')
\]  

(2.5)

where \( a_n \) are constants and \( f_n(r') \) are suitable basis functions. The near-field integral (2.3) can then be written as

\[
E(r) = \sum_{n=1}^{N} a_n G_n(r)
\]  

(2.6)
with
\[ G_n = C_k \eta \int_S \left[ f_n C_{N1} - (f_n \cdot \hat{R}) \hat{R} C_{N2} \right] \frac{e^{-jkR}}{R} dS' \] (2.7)

where \( f_n = f_n(r') \) and \( G_n = G_n(r) \).

In a practical system, the field measured at the point \( r \) is affected by the probes’ polarization properties. We therefore have

\[ E(r) = E(r) \cdot \hat{p}^* = \sum_{n=1}^{N} a_n G_n \cdot \hat{p}^* \] (2.8)

where \( \hat{p}^* \) is the complex conjugate of the polarization vector \( \hat{p} \) which describes the polarization of the antenna.

2.2.1.2 The Difference Between Two Antennas in an Antenna Array

The expressions above relates the current distribution, described using basis functions, to the electric near- or far-field. Of interest when implementing the beamforming algorithms is the difference in the amplitude and phase of the far-field between two antenna elements, as this will affect the direction of the beam. Also, calculating this difference enables the use of signals-of-opportunity in the estimation process. One antenna element could then be temperature controlled (i.e. free from snow) and used as a reference element. To calculate the difference in the measured electric field between two antenna elements, \( p \) and \( q \), (2.8) is rewritten as

\[ E^{(q)}(r) - E^{(p)}(r) = \sum_{n=1}^{N} \left( a_n^{(q)} G_n^{(q)} - a_n^{(p)} G_n^{(p)} \right) \cdot \hat{p}^* \] (2.9)

Assuming that \( p \) is the AUT and \( q \) is the reference element, the constants \( a_n^{(q)} \) can be written as

\[ a_n^{(q)} = a_n^{(p)} + \delta_n \] (2.10)

where \( \delta_n \) is the difference between \( a_n^{(p)} \) and \( a_n^{(q)} \). This makes it possible to write (2.9) as

\[ \Delta E = \sum_{n=1}^{N} \left( a_n^{(p)} G_n^{(q)} - a_n^{(p)} G_n^{(p)} + \delta_n G_n^{(q)} \right) \cdot \hat{p}^* \] (2.11)

where \( \Delta E = E^{(q)}(r) - E^{(p)}(r) \). If the the test element is assumed to be unaffected by the environment the term \( a_n^{(p)} (G_n^{(q)} - G_n^{(p)}) \) will be known, with \( G_n^{(p)} \) and \( G_n^{(q)} \) given by (2.7). The only remaining unknowns are then the constants \( \delta_n \). Using matrix notation (2.11) is rewritten as

\[ G\delta = \Delta E - (G - G^{(p)})a^{(p)} \] (2.12)

where

\[ G = G^{(q)} \cdot \hat{p}^* \] (2.13)
2.2.2 Estimating the Current Distribution on the Antennas

2.2.2.1 Least Squares

In general, the system of equations presented in the previous section will be overdetermined. In this case it can be solved using the Least Squares method, which is optimal when the measurement errors have a Gaussian distribution, see chapter 6 in [19]. The system of equations that is to be solved can be written as

\[ G\delta = b + \varepsilon \]

where \( G \) is given in (2.7) and (2.13), \( \delta \) is the vector of unknowns, \( b \) is the observation vector given on the right hand side of (2.12), and \( \varepsilon \) is the observation noise. The solution to this system can be written as

\[ \delta = (G^H G)^{-1} G^H b \]

where \( G^H \) is the hermitian transpose of \( G \).

2.2.2.2 Kalman Filter

Since it is necessary to measure the performance of each antenna element continuously, a Kalman filter is used in order to reduce the errors induced by noisy measurements, see chapter 13 in [19] for a comprehensive introduction to Kalman filters. The performance of the AUT relative to the reference antenna at a given time, \( k \), is described by the state

\[ x_k = \begin{bmatrix} \delta_k & \dot{\delta}_k \end{bmatrix} \]

where \( \delta_k \) are the coefficients describing the difference in the current distribution on the AUT and the reference antenna element at time \( k \), from (2.12), and \( \dot{\delta}_k \) are the time derivatives of these coefficients. Assuming that there is a linear change of the coefficients \( \delta \) between time \( k - 1 \) and \( k \), the state \( x_k \) is related to the previous state \( x_{k-1} \) with

\[ x_k = F x_{k-1} \]

where

\[ F = \begin{bmatrix} 1 & \Delta t \cdot 1 \\ 0 & 1 \end{bmatrix} \]

where \( 1 \) is the identity matrix and \( \Delta t \) is the time between \( k - 1 \) and \( k \). Further, the observations, \( z_k \) at time \( k \) can then be related to the state using

\[ z_k = H x_k + \varepsilon_k \]

where \( \varepsilon_k \) is the measurement errors and

\[ H = [G \ 0] \]

where \( G \) is given by (2.7). The observations \( z_k \) are given by the right hand side of (2.12). The Kalman filter uses the previous state \( x_{k-1} \) and the observations \( z_k \) to estimate the current state \( x_k \). This is given by

\[ x_k = (1 - K_k H) F x_{k-1} + K_k z_k \]

where

\[ K_k = \frac{G}{G^H G} \]

and

\[ \varepsilon_k = z_k - H x_k \]
where the Kalman gain $K_k$ is given by

$$K_k = P_{k|k-1}H^TS_k^{-1}$$

Kalman gain

$$S_k = HP_{k|k-1}H^T + R_k$$

residual covariance

$$P_{k|k-1} = F_kP_{k-1|k-1}F_k^T + Q_{k-1}$$

predicted estimate covariance

$$P_{k|k} = (1 - K_kH)P_{k|k-1}$$

updated estimate covariance

where $R_k$ and $Q_{k-1}$ are the measurement and system covariance matrices. If these matrices are known or are possible estimate they are used in the Kalman filter to weight the signals. The system covariance matrix, $Q_k$, contains the information about the uncertainties affecting the estimated state, $x_k$. In an implementation of the measurement system this would, in addition to errors induced in the updating process, also include error sources related to the AUT. These could be both electrical, such as crosstalk and limited accuracy in the timing between front-ends, and mechanical, such as variations in the position of the AUT (e.g. due to wind).

The observation covariance matrix, $R_k$, includes information about uncertainties related to the observations. These are mainly caused by error sources affecting the probes. Apart from electrical and mechanical errors similar to the ones affecting the AUT, the main error is the thermal noise on the transmitted signal. There may also be a polarization error due to non-ideal probes.

All the uncertainties discussed here will have the effect of a amplitude and phase-shift of the far-field pattern of the AUT. The estimated far-field pattern of a given antenna element will thus include all other error sources in addition to the effect of snowfall, which is the desired information.

### 2.2.3 Numerical Results

In this section, the performance of an implementation of the method described above is assessed using numerical data from simulations using the Numerical Electromagnetics Code. In the simulations, all effects due to mutual coupling between nearby antenna elements (e.g. scan blindness) have been neglected. These effects can be taken into account when performing the measurements by using additional basis functions and careful selection of the probes used at a given time.

#### 2.2.3.1 System Setup

The antenna array considered here is similar to the EISCAT 3D test array outside Kiruna, Sweden. The antenna elements are the same crossed Yagi-antennas as the ones analyzed in [1], these are also shown in Fig. 2.4. The measurement system here is designed with the final radar system in mind, where the size of the antenna array could be on the order of 125-by-125 m. The antenna elements are mounted at an elevation angle of 55° and also rotated 45° with respect to its own axis. Again, it should be noted that this system setup is only considered to define realistic locations of the probes. It is thus not designed with any array parameters in mind.

The probes are assumed to be mounted on 150 m high towers which are spread out around the antenna array. Although there may be some interference due to reflections from the towers this effect should be small since there are no towers in the main beam direction of the antenna array. There are three probes on each tower at 50, 100, and 150 m. This setup is shown in Fig. 2.4. Note that in this figure, there are no probes in the main beam of the array, which was one the main requirements of the measurement system.
Figure 2.4: The proposed measurement system setup (b) and antenna element (a). The number of antenna elements plotted is in the figure for clarity lower than in the actual system. The probes are in (b) denoted by ‘◦’.
Figure 2.5: The far-field amplitude (a) and phase (b) for the antenna element in middle of the array. The probes are assumed to be noise free.

Fig. 2.5 shows the simulated far-field (solid line) of one antenna element and the far-field estimated using the equivalent electric current method (dotted line) for the element in the middle of the array in Fig. 2.4. This element is assumed to be used as the reference element and it is therefore temperature controlled and free from snow. The reference is the results from a full NEC-2 simulation while, for the estimation, only the field calculated at the probe locations was used. It can be seen that the results for both the amplitude and phase is accurate for the considered element. Only the results for one polarization is shown (the lower of the ones shown in Fig. 2.4).

Since the position of the probes relative to the AUT will be different for different antenna elements in the array, the error introduced in Fig. 2.5 could vary between the antenna elements. This is shown in Fig. 2.6 for the main beam direction of the antenna element (along the y-axis with an elevation angle of 55°). The error is significantly larger in the middle, front part of the array. This is because there are no probes in the main beam of these antenna elements.
Figure 2.6: The estimated far-field amplitude (a) and phase (b) error in the direction of the main beam of the antenna elements for different positions in the array.
2.2.3.2 Performance During Snowfall under Noise-free Conditions

Simulations were done for a test case to compare the performance of the Least Squares method with the Kalman filter approach. These simulations were done for one antenna element left, back corner of the antenna array with the antenna element located in the middle of the array used as the reference element. In this simulation the signal is assumed to be noise-free. The Kalman filter will in this case give the same result as the Least-squares solution since there is no need to use the previous state in the estimation process. Hence, only the results for the Least-squares solution is shown here.

The snow is modeled in the same way as in [1] where a model proposed in [8] was used. Since the permittivity of snow depends on a large number of parameters and can vary quickly with changing conditions it is not feasible to take every possible situation into account. Instead a worst case scenario was adopted. The water content of the snow was assumed to be 10 % by volume and the density 600 kg/m$^3$. This corresponds to wet and heavy snow, which have the highest effect on the antennas. The real part of the permittivity is in this case 3.7 while the imaginary part is 0.037.

The test scenario that is used both here, in the noise-free, and in the noisy conditions considered later is shown in Fig. 2.7. At the time $T = 10$, there is the onset of the snowfall. The thickness of the snow layer increases gradually until there is a 0.5 mm thick layer of snow covering the antennas. In the simulation process, the snow covers all wires of the antennas completely. This will most often not be the case in reality where the snow will cover only the top part of the wires. The simulation is divided into four different periods in order to evaluate the performance of the measurement system under different dynamic situations.

In Fig. 2.8 the amplitude of the difference between reference antenna’s and the AUT’s far-field is shown. The estimated far-field difference here gives accurate results. There is a small bias during the period when there is a constant snow layer on the antennas. This bias is, however, well within the acceptable limits of the system.

2.2.3.3 Performance During Snowfall under Noisy Conditions

In the simulations considered here, the signal transmitted by the probes is assumed to be noisy. The term noise is here includes all possible error sources on the observations. The observed signal from a probe is then modeled according to the right hand side of (2.14), where $\varepsilon$
is a complex number used to denote all errors. It is assumed to have a gaussian distribution with zero mean.

Since the amplitude of the signal received by the antenna elements will depend on both the distance to the probes and the antenna gain in the direction of the probes the signal-to-noise ratio (SNR) will vary significantly between different probes. The SNR for the AUT considered in this section varies from less than -7 dB to over 32 dB for different probes.

The amplitude as a function of time is shown in Fig. 2.9. It can be seen that the Kalman filter is clearly smoother than the least squares solution. There is also an amplitude offset which is clearly visible between time 20 and 30 (period (b) in Fig. 2.7).

To better evaluate the performance the complex error in the estimated far-field was calculated. This can be seen in Fig. 2.10. Four different periods were analyzed, (a), (b), (c), and (d) in the Figure and denoted by the same letters in Fig. 2.7. A total of 100 simulations were used to create these plots. The numerical values are also shown in Table 2.1, where the amplitude of the mean value is used as a measurement of the accuracy of the estimation methods. The standard deviation can be used as a measurement of the precision. During all periods, the Kalman filter
Figure 2.10: The complex error induced due to the noise on the test signal ($\varepsilon$ denotes the error). The four subplots (a) to (d) refers to the corresponding periods shown in Fig. 2.7. The units are V/m.

The return loss of an antenna when covered by snow has been simulated using models of the permittivity of snow. It was also measured during a period with significant snowfall. The results shows clearly that snow covering antennas could alter the characteristics of an antenna significantly which could be serious problem in cases when it is important that the antenna is operational continuously. If it is not possible to heat the antenna or placing it in a radome the bandwidth will need to be increased in order to be able to afford some movement of the
Table 2.1: Accuracy of the estimated electric field difference.

<table>
<thead>
<tr>
<th>Period</th>
<th>Least squares</th>
<th>Kalman Filter</th>
</tr>
</thead>
<tbody>
<tr>
<td>(a)</td>
<td>0.047 V/m</td>
<td>0.030 V/m</td>
</tr>
<tr>
<td>(b)</td>
<td>0.047 V/m</td>
<td>0.036 V/m</td>
</tr>
<tr>
<td>(c)</td>
<td>0.047 V/m</td>
<td>0.024 V/m</td>
</tr>
<tr>
<td>(d)</td>
<td>0.047 V/m</td>
<td>0.040 V/m</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Period</th>
<th>Least squares</th>
<th>Kalman Filter</th>
</tr>
</thead>
<tbody>
<tr>
<td>(a)</td>
<td>0.016 V/m</td>
<td>0.016 V/m</td>
</tr>
<tr>
<td>(b)</td>
<td>0.024 V/m</td>
<td>0.018 V/m</td>
</tr>
<tr>
<td>(c)</td>
<td>0.032 V/m</td>
<td>0.038 V/m</td>
</tr>
<tr>
<td>(d)</td>
<td>0.023 V/m</td>
<td>0.031 V/m</td>
</tr>
</tbody>
</table>

operational band.

Also, the change in coupling between antenna elements due to snowfall have been assessed. There does not seem to be any significant change in the coupling coefficients due to snowfall. The coupling under different conditions should, however, be considered during the design of the final antenna array.

The measurement system presented in Section 2.2 uses the equivalent electric current method to estimate the far-field of an antenna element in a physically large antenna array. The method enables continuous measurements of the performance of all antenna elements in array without the need of having any probes in the main beam of the antenna array. Also, using the difference of the far-field between two elements it is possible to incorporate far-field sources, which may be at an unknown distance from the antenna element, into the system of equations and retain the phase information. While it may be sufficient to use the least squares solution under low noise conditions, the Kalman filter have clear advantages as the noise level increases. This will be particularly important when far-field sources are added due to the low signal levels typically associated with these sources.

Although only one test scenario has been considered the results from the simulations shows that by using this type of measurement system, it is possible to improve the knowledge about the far-field characteristics of an antenna. This means that the availability of the radar system and the quality of the scientific measurements are increased.
3 Receiver Hardware

3.1 Receiver system overview

A block schematic of the antenna front end is shown in figure 3.1. The primary active parts in the receiver chain are the Low noise amplifier (LNA) and the A/D converter board.

The **LNA** is built around two amplifier stages and a band-pass filter. The input gate is protected by anti-parallel schottky diodes. The first stage is followed by a band-pass filter and then by a monolithic gain block. To facilitate system calibration the LNA includes a directional coupler and four analog switches. This allows the calibration of phase and amplitude variations between channels. Signal injection is performed locally at one LNA at a time, this signal is also distributed over coaxial cable to all other channels in the array. Doing this at all LNAs sequentially makes the calibration independent on the network of coaxial cables.

![Figure 3.1: Block schematic of one A/D card](image)

The **A/D converter board** contains additional filters and two additional gain stages. After amplification and filtering, the signal is digitized by a 16 bit ADC running at 80 MHz. The clock signal is generated by an on-board VCXO which is phase locked to a 10 MHz external clock reference. The last gain stages are fully capable of producing signal levels in excess of the tolerance of the A/D converter. Thus, a protection circuit consisting of a schottky diode limiter and a steerable attenuator is preceding the ADC. The A/D converter board delivers a 16 bit + clock LVDS output data stream.

The signal chain described above is for one single channel. Each A/D converter board does hold two identical units as described. This can be used to perform the A/D conversion of the signals from both polarisations on one antenna using the same A/D converter board. The following sections give an in depth description of the LNA, the A/D converter boards, and of the front end hardware implementation of the calibration system.
3.2 The LNA

3.2.1 Introduction

As a part of this pre-study a LNA has been designed to verify the feasibility building cheap, room temperature LNAs of sufficient performance. The design is aimed at fulfilling the specifications listed in Deliverable D2.1 [20]:

- a bandwidth of 30 MHz,
- a system noise figure of $50 \, \text{K}$ and,
- a spurious free dynamic range of 70 dB.

In addition, it was decided that the LNA should have a return loss of no less than 17 dB, as a high return loss is desirable when building antenna arrays, as this will help mitigate some of the problems with coupling between the antennas in the array. Furthermore it was deemed desirable to filter out unwanted signals as early as possible in the system, as these signals may otherwise inter-modulate and produce in-band signals which, once created, are impossible to get rid of. In case the final system use a smaller bandwidth, both noise figure and return loss can be improved by minor redesign. An additional requirement on the LNA was the possibility of signal injection for calibration purposes, as further discussed in section 3.5.

A block schematic of the LNA including signal injection logic is shown in Figure 3.2. The manufactured LNA boards (Figure 3.3) include input protection, two amplification stages, a passive filter, calibration signal injection circuitry and various voltage regulation and bias regulation circuits.

![Figure 3.2: Block schematic of the LNA including signal injection design](image-url)
3.2.2 RF Component characterization

As the applications for very low noise figure amplifiers operating at VHF frequencies are limited, so is the number of available semiconductor devices. The device with the best specified performance at a frequency not too different from the intended frequency range was the ATF-541M4 E-HEMT from Avago, with a specified noise figure at optimal noise match of 0.1 dB (7 K) at 500MHz. This device was thereafter characterized for the frequency of interest.

At RF and microwave frequencies the noise figure is calculated from the measured output noise when two or more different but well known noise levels are present at the input of the device. In order to fully model the noise in an active device (both voltage and current noise, and the correlation between these), it is necessary to measure the noise figure when several different input impedances are present at the input of the device. This is implemented by using a stub tuner or similar device for transforming the impedance the noise generating device(s) into the desired impedance, while (theoretically) preserving the different input noise levels.

In order to speed up the (manual) setup of the stub tuner, and also in order to use the additional degree of freedom available when using a triple stub tuner for minimizing the losses in the stub tuner a parametric model was fitted to a set of 125 measurements (all combinations of 5 different positions on each stub). Based on the initial suggestion and derivatives from this model it was usually possible to set the requested match to better than 0.05 (reflection coefficient difference) in two iterations. The characterization was performed by measuring difference in output noise level when using a 50 ohm termination at room temperature and when using a termination submerged in liquid nitrogen. This measurement was repeated at at least 5 different input matches, selected around an initial guess (or the output from an earlier experiment) of the optimum noise match of the device. A standard noisy two-port model was then fitted to the measured data, using the scattering parameters of the stub tuner as measured by the network analyzer and taking losses in cables etc into account. The measurement was complicated by the fact that the noise of the transistor was considerably smaller than the losses
in the measurement setup, making it necessary to accurately characterize all elements of the circuit. Nevertheless, a preliminary model was derived for the transistor operating at 224 MHz. This model was used in the optimization of the LNA.

Simpler (one or two-port network analyzer) measurements were used for characterizing various passive components. In particular, a number of wire wound surface mount inductors from different manufacturers was measured and evaluated, as it early became apparent that the performance (i.e., losses/Q-factor) of currently available inductors suitable for mass-produced electronics would limit the noise figure of the LNA. The electrical models derived from these measurements was used when designing the LNA.

3.2.3 Temperature Stabilization

Due to the strict timing requirements for the final system a study on temperature characterization was performed [21]. It was early concluded that it was not feasible to fully temperature stabilize the whole system. On the other hand, only parts critical to the timing are certain components in the calibration system. Because these components have been placed on the LNA boards, we temperature stabilize the whole LNAs in the current system. Further study will be required to determine if even this stabilization is necessary.

3.2.4 Directional coupler design

Each LNA board includes one directional coupler and a number of semiconductor switches for routing calibration signals into the receiver path.

Due to PCB size and insertion loss constraints a coupler much shorter than the normal 1/4 \( \lambda \) (\( \sim20 \) cm at 224 MHz) was used. The main implications are that the coupling will change significantly over the frequency band, and that the coupling per unit length will have to be fairly large. In order to minimize the insertion loss due to losses in the input signal path a coupler with transmission lines on opposite sides of a ground plane (with a coupling slot) was designed with a wide (for low loss and thus low noise figure) conductor for the incoming signal at one side of the board, a ground plane close to the other side of the board and a suitably narrower conductor for the calibration signal was designed (Figure 3.4).

Generic methods of calculating the scattering parameters of asymmetric coupled lines [22] and the FDTD simulation program atlc [23] was used to automatically optimize the geometry of a coupler given constraints on coupling and front to back ratio.

3.2.5 LNA design

A Matlab tool for simulating interconnected noisy two-ports was implemented based on the method described in [24, 25]. This tool was then used with a particle swarm optimizer [26] to find an “optimal” LNA given a number of constraints on noise figure, return loss, bandwidth, frequency selectivity and stability margin.

The coupler, input protection diodes, the filter following the transistor and a model of the second gain stage (an RFMD RF3376 monolithic gain block) were all included in the simulations used by the optimizer. In order to improve the yield of manufactured amplifiers it was required that the solution should remain stable when components deviated a certain amount from their optimal values.

It is worth noting that the requirement of a high return loss (17 dB) over a large bandwidth (30 MHz) leads to a significant trade off in noise figure from the sub 10 k noise figure (at optimal
Figure 3.4: The microstripline coupler implemented on the LNA boards

The noise match) of the transistor to a noise figure in the range of 40-50k, due to both suboptimal (from a noise perspective) input matching, and losses in said matching network.

After exploring the different possible trade offs and some different topologies the design shown in Figure 3.5 was selected for further testing.

Due to time constraints several interesting possibilities were however omitted, for example using two or more transistors in parallel or applying some source degeneration to the transistor [27].
3.2.6 Results

Overall, the design is working as intended. 24 units are currently undergoing a long-term test at the EISCAT site in Kiruna. Some variations in the frequency characteristics and gain of the LNAs has been observed, attributed mainly to the loose tolerances on available high quality factor surface mount inductors.

In preliminary tests (a single LNA board) 16kV human body model (20 pulses of each polarity) applied to input of the unpowered LNA did not induce measurable performance degradation.

The SFDR of the LNAs has been measured only when connected to the ADCs in the test receiver, but as this combination shows SFDR of 70dB or better, further tests to determine the linearity of only the LNAs has been postponed.

These results are highly encouraging and the performance is sufficient for the final system. However, the potential exists for future improvements. For example, it appears worthwhile to test if a small amount of source degeneration can be used without compromising stability, as this could result in an improved noise figure [27].

3.3 A/D converter board

3.3.1 Introduction

The advantage of direct undersampling is that no intermediate frequency is used, which in turn means no additional local oscillator or mixer. Instead the sampling frequency of the analog to digital converter (ADC) is chosen so that the frequency band of interest falls between the harmonics of the Nyquist frequency, and thus is sampled without loss of information. As with any sampling process, it is vital that suitable anti-aliasing filters are used, or signal and noise in other bands will fold into the signal.

In order to test the feasibility of using direct undersampling and to be able to test beamforming and calibration on array level an A/D converter board was designed. The A/D converter board as pictured in Figure 3.6 holds two channels which allow the use of one board per antenna with two polarisations. As shown in the block schematic in Figure 3.7, both channels are clocked by one on board VCXO which is phase locked to an array wide distributed reference clock. Main components and features of the A/D converter board are further discussed in the sections below.

3.3.2 Sample clock

Because the same sampling clock must be used by all receivers in the array, it must either be distributed from one oscillator to all ADCs, or if generated locally, must be phase locked to some other reference signal that is distributed to all receivers. The main argument for using a centrally generated clock directly is that a higher quality oscillator can be used, either a crystal based oscillator with additional filtering for wide-band noise, or an LC oscillator operating with a large stored energy and that is locked to a crystal oscillator for frequency stability. The drawbacks include the problems of distributing the signal without introducing more noise (or spurious signals) than a locally generated signal would contain, and that the noise in the sample clock would be correlated between channels, meaning that for some input signals and pointing directions a significant degradation in array noise figure would occur. For our design we have chosen to use local sample clock generation with one VCXO (Crystek CVHD-950)/PLL
(Analog Devices ADF4002) for each pair of ADCs. This means that the noise due to sampling clock noise will only be correlated between the two polarizations of each antenna, not between antennas. These PLLs are locked to a 10MHz reference signal that is distributed to all receivers. Even though this is the best commercially available VCXO we have been able to find it limits the maximum SNR of the ADCs to little more than 65dB, as explained below.

3.3.3 Anti-aliasing filters

The anti alias filtering is split between three filters. The first filter on the ADC board is a monolithic filter with a sharp characteristic but fairly high insertion loss, which implements the bulk of the out of band signal rejection, especially at frequencies close to the 200-240 MHz band. The two later filters are built from discrete surface mount components, have a much less sharp pass-band, and are used to attenuate out of band noise generated by the the amplifiers after the first filter and strong signals at frequencies far from the band of interest (e.g 100MHz, 450 MHz, 900MHz). These filters have much a significantly lower insertion loss, lowering required output level from the preceding amplifier stage, and thus the distortion generated in that amplifier.

3.3.4 Amplifier stages

One problem with directly sampling the RF signal is that of amplifying high frequency signals to the high levels required by an ADC (1-3Vpp), without introducing significant amounts of distortion. This doesn’t affect the first amplifier stage on the ADC board, where a generic high-linearity monolithic amplifier (Sirenza SBB-5089z) is used, but for the final stage a fairly powerful amplifier (HELA-10D) as the last stage was the only available solution. In order to protect the ADCs from being destroyed by the high output power this amplifier is capable of producing, e.g in case of abnormally high input signal to the system, two types of input pro-
tection was implemented: diodes attempt to limit the voltage from the amplifier to safe levels, and when significant current is flowing through these diodes an attenuator at the input of the ADC board is activated, attenuating the input signal by about 40dB, thus keeping the diodes from overheating. Another drawback of this amplifier is the rather high power consumption of about 3W, to be compared to the ADC which consumes about 1.3W. If more distortion can be tolerated a lower powered buffer could be used, with significantly lower power consumption. This would also make the input protection circuit obsolete.

3.3.5 ADC

The main figures of merit of any ADC operating at RF frequencies are the signal to noise ratio (SNR) and spurious free dynamic range (SFDR), typically measured at -1dB full scale (dBFS). Because both of these decrease with increasing signal frequency it becomes necessary to use a more expensive ADC when using direct undersampling than would have been the case if the signal had been converted to an intermediate frequency first. Furthermore, noise in the sampling clock will be scaled by $20\log_{10}(\text{Fsig}/\text{Fclk})$, which even for the best available VCXOs will result in a larger noise contribution than from the ADC itself, when sampling a 224MHz signal at 80MHz. In practice we have found that this limits the maximum achievable SNR over a 40 MHz bandwidth at 224 MHz to be little better than 65 dB, which is 8 dB or more lower than the theoretically achievable SNR when using a good ADC with a noise-free sampling clock. In our implementation we use LTC2208 ADCs from linear technology, selected mainly based on low distortion when undersampling a 200-240 MHz signal combined with a moderate power consumption.
3.3.6 Results and conclusions

A total of 13 ADC boards have been manufactured. Due to the issue with amplification of noise in the sampling clock when undersampling the SNR of all ADCs is limited to about 65 dB. The SFDR of all boards and channels is better than 70 dB.

As the quantization noise of the ADC will add directly to the noise temperature of the system, the noise contribution from the ADC must be balanced against the maximum SNR by placing an appropriate amount of gain before the ADC. In the current system the gain is somewhat too high, which places the noise from the ADC 20 dB below the other noise sources, with a contribution to the total system noise temperature of 2.9 K. This is of course very low, but comes at the cost of reducing the maximum system SNR to only about 45 dB. In our current tests this is of no consequence, but for a final system it would probably be desirable to reduce the gain preceding the ADC by about 6 dB, thus increasing the achievable system SNR to 51 dB at the cost of increasing the noise contribution to 12 K.

Generally speaking the use of undersampling as a means of simplifying the system has worked well. The higher cost of the ADC will have to be compared to the cost of the otherwise required additional local oscillator and mixer. The use of an intermediate frequency would also reduce the complexity of the anti-aliasing filters.

3.4 System control

The control system controls things such as the power supply outputs, the coupler modes for signal injection, signal injection frequency, LNA:s, A/D cards and temperature stabilization of the LNA boxes. The control system uses an 8-bit RISC microcontroller from Atmel and is interfaced through the USB port. The system can be controlled remotely through a server software which communicates with the different control boards over optic fiber. A bootloader software has also been implemented, which enables the possibility to upgrade firmware remotely. The control board is installed in the power supply box.

The current system is built mainly for the demonstrator array and for a larger system some changes would need to be made, such as increasing the speed by using for example an FPGA instead of the 8-bit microcontroller, but also implementing a bus system on which all boards can communicate instead of the current USB interface.

3.5 Calibration system implementation

The LNA PCBs include the electronics required for implementing a calibration system that will theoretically be capable of mitigating any timing/phase and amplitude errors in a multi-channel receiver system, without being dependant on any external precision components (such as long cables of accurately known length). This system is based on using a separate passive signal distribution network through which test signals can be sent.

The key is to have a passive signal distribution network connecting all LNAs to each other and/or to a reference station that is reciprocal, which means that the transfer function (both phase and magnitude) is exactly equal in both directions between any two ports of the network. As it turns out, almost any passive component (including cables and power splitters) are reciprocal, and so is any network built from only passive components.

The signal is injected at the originating LNA after some attenuation (in addition to the nominal -20 dB coupling of the directional couplers), and is distributed to the other LNAs where it is injected without additional attenuation. It is possible to select at which of the ports of the
directional coupler the signal is applied, thus injecting the signal either towards the LNA or towards the antenna, and not to inject the signal at all. These options allow us to measure and compensate for the coupling between the antennas, and to measure the reflection coefficient of each antenna (for diagnostic purposes).

It is important that the network is fully passive, and that the LNAs present the exact same impedance to their ports to the network regardless of the mode the LNA is in. Otherwise the reciprocity of the passive network will be useless with reduced calibration accuracy as a result.

The switchable attenuator allows for shorter measurement times by reducing the ratio of the signals seen at the injecting (local) and receiving (remote) LNAs to be more equal, thus allowing more signal to be used without saturating the local LNA, thereby improving the SNR. However, the impedance variations seen by the calibration network may be a major source of systematic error in the system. This aspect should be studied further, and it could also be interesting to build the network with the possibility of calibrating without switching the attenuator, using longer integration times to compensate for the lower SNR at the remote LNAs.

The topology of the calibration signal network, see Figure 3.8 remains an open question, in a large system it could be beneficial to use an asymmetric network with one or more reference systems where the signal is injected without coupler directly into an LNA, thus gaining about 20dB signal level. While not necessary for small arrays, this could be important when building larger systems. For our tests we have used a fairly symmetrical network to avoid the need for specialized hardware as shown in Figure 3.8.

This system is currently undergoing long-term testing in a small array at the EISCAT site in Kiruna. For more information on the capabilities and results of the system, see Section 4.2.
3.6 Demostration array photos

This section presents photographs of the hardware implementation performed in the Kiruna based demonstration array.

Figure 3.9: Photograph of four LNA channels in a temperature regulated box.
Figure 3.10: Photograph of the power supply rack unit.

Figure 3.11: Photograph of the whole RF frontend without copper shields
Figure 3.12: Photograph of the calibration signal injection box, the fans are for cooling the A/D cards

Figure 3.13: Photograph of the whole frontend
4 Timing

4.1 Introduction

The necessary accuracy of the timing system in the EISCAT_3D radar has been thoroughly investigated during the course of the project [28, 29]. The results show that for the receiver arrays to function within the defined parameters, the total error in the timing between any antenna elements in the array may not have a standard deviation of more than 160 ps. This error consists of three main contributors: the receiver hardware, the timing system, and antenna phase-center movement due to external factors such as weather. Thus, it is reasonable to have a target of a third of this error, \( \sim 50 \) ps, for the timing system by itself.

There are two different approaches to solving the timing distribution system for EISCAT_3D. The first system is a cable-based system that continuously evaluates the delay between different parts of the timing system, thereby keeping track of any shifts in delay through the system. The second one is based on Global Navigation Satellite Systems (GNSS), which tracks the difference in phase of the signals from satellites to the timing units placed in the center of each sub-array [30].

There are benefits and drawbacks of both systems: The cable-based system is less prone to disturbances but needs a lot of cables in the array and can only run when the radar is off. The GNSS-based system could be sensitive to disturbances since it is based on radio signals, but if disturbances are significant to the GNSS signal, chances are that the radar itself also is unusable. The GNSS system has a lot less hardware requirements, i.e. almost no cables, and thus could be a lot cheaper in a large array.

Both systems have been evaluated and are described in the following sections but at different levels. The cable calibration timing system has been simulated, built and tested in the test array in Kiruna, whereas the GNSS based system has been simulated and tested on an even smaller scale (single base-line test) at Luleå University of Technology.

4.2 Cable Calibration Timing System

One way of achieving phase synchronization was first suggested by Grover[31]. The idea is to send a pulse down a line and let it reflect at the end point, measuring the average time between the outgoing and reflected pulses. By doing so, one actually measures the time of the reflection, regardless of where on the line you are positioned when doing the measurement. Thus, using only one line for the whole array, each antenna element measures the same time with great accuracy. By synchronizing another distributed clock with this measurement, one would achieve phase-synchronized clocks in the whole array.

This method would prove difficult to achieve in the EISCAT_3D project, mainly because of difficulties in keeping the signal levels large enough through the array and to create a sensor with good enough accuracy to measure the time of each pulse. However, a version of the system has been developed where each LNA of the system is connected to every other LNA, see Figure 4.1. By injecting a sinusoid signal in the LNAs in different order and with different frequency, the actual relationship between the length of the cables used for the calibration can be decided. The direction of the injected signal can also be controlled, so that the length of signal path can be deduced not only between the LNAs, but also from each LNA to the phase center of the
connected antenna and also through the normal signal chain to the ADC. The implementation of the calibration system is further described in Section 3.5 in this report.
Figure 4.1: Diagram of the EISCAT_3D LAAR racks containing the cable calibration system. Each LNA is connected to every other LNA in the array via the CALIBRATION LINK to allow calibration of the different units to each other.
4.2.1 Simulation Results

The proposed calibration system has been simulated in Matlab by computing the resulting n-port when all the components of the system are interconnected by cables of random length. The directional coupler is modeled based on data extracted from four coupler boards, both in terms of capacitance/inductance matrix and the board to board variations of all the parameters. The coupler board switches are modeled by a 3 Ω on-resistance and 1.25 pF terminal capacitances, with deviations 0.1 common and 0.02 difference between switches. The resistors are configured for a 20 dB local signal injection attenuation. If we set the antenna and front-end reflections to -(10..13) dB and -(15..18) dB (chosen randomly), respectively, we get delay and amplitude standard variations of 32 ps and 6.0%, which can still be considered acceptable. Increasing LNA reflections to -(10..13) dB increases the error to 45 ps and 7.7% respectively. On the other hand, eliminating LNA reflections completely only improves the performance to 20 ps and 5%. Eliminating both LNA and antenna reflections reduces the errors to 14 ps and 4.5%, respectively. Thus, the -(15..18) dB range of return loss seem like a reasonable target for the LNA design.

The results above is of course specific for a specific set of variations, and should by no means be seen as the ultimate achievable accuracy, as better coupler and component tolerances, combined with a good method for compensating for known offsets (based on measurements during manufacturing and field calibrations), could be expected to yield significant improvements.

4.2.1.1 Notes

The current calculations of front-end doesn’t attempt to make any corrections for known errors, e.g. if we make production measurements on the coupler/LNA boards, nor does it compensate for the limited directivity of the coupler which results in antenna impedance measurements being way off.

4.2.2 Measurement Setup

To evaluate the actual performance of the cable-based calibration system, a C program has been developed which can be started through the server software installed in the test array in Kiruna. The calibration program is connected to the channel boards of the original EISCAT system, capable of capturing 256 samples of I- and Q-data for each row and polarization of the test array at 1.25 MHz. This limits the bandwidth of the system to 625 kHz, and with a sampling frequency of 80 MHz and a further digital down-mixing by 16.3 MHz, this places the signal band for the calibration system in the test array from 223.075 MHz to 224.325 MHz. In a future system, this limitation is not necessarily true as it is currently limited by the maximum frequency of the channel boards.

For each calibration measurement, the channel boards capture 50 µs of data every 50 ms for a period of 5 seconds. At every second 50 µs data block, the couplers and VCO outputs in each LNA are set by the calibration program. This is necessary to map up all the different signal paths between each ADC of the whole array. When the length of the signal paths are known, the time difference between each ADC can be calculated. At the start of each 5 second data block, the EISCAT radar controller sends a synchronization pulse to the calibration program so that each data block contains well defined settings of the couplers and VCOs.

The calibration program saves 3 different data blocks which each uses a different VCO frequency to allow resolving phase ambiguities since the signal paths between some rows in
the test array are greater than the wavelength of the injected frequency.

After analysis of the captured data in Matlab, a matrix of the correct time- and amplitude differences between each ADC in the array is stored and later used for correct determination of the beamforming filter settings.

4.2.3 Measurement Results

At the time of writing, no measurement results from the test array in Kiruna were available. While large sets of data has been collected, the analysis is not yet complete.

4.3 GNSS Timing System

In the system described here, low-end L1-only GNSS receivers are used in a highly application specific environment, to provide the picosecond accuracy timing. This is done by dividing the LAAR into small sub-arrays of e.g. 16 elements each. The maximum length of the cables distributing the clock is then reduced to 6 m which is short enough to be calibrated by length approximation only, assuming that the clock distributed to each sub-array is known. By inserting a Global Navigation Satellite System (GNSS) receiver at each of these sub-arrays, to provide a clock reference that is unaffected by changing conditions over the array, the antennas are now timed to the specified accuracy.

Other benefits of building a GNSS timing system include lower cost due to reduced amount of coaxial cable throughout the array, and that a continuous cable length calibration system that ensures the timing accuracy of the distributed clock system is no longer necessary.

The following sections will describe in detail the concept of designing an L1-only GNSS receiver based timing system which is highly application specific to meet the requirements of the EISCAT_3D LAAR. The simplifications that are possible to make to the GNSS receivers are stated and then a test setup made as proof of concept of the timing system is described.

4.3.1 GNSS Timing System Concept

Each of the sub-arrays will contain a Phase Locked Loop (PLL) in which the distributed frequency reference is reproduced and distributed to the local GNSS receiver, the radar ADCs, and a signal injection system located as close to the radar antenna elements as possible to calibrate the analog signal path of the system, as shown in Figure 4.2.

The main purpose of the PLL is to adjust the phase of the reference clock to be equal throughout the array. This is achieved by creating a closed loop feedback from the GNSS receiver to the PLL and adjusting the phase according to the phase differences in the received satellite signals in respect to a reference GNSS receiver. The reference receiver is a high-end receiver which is used in conjunction with application specific software to produce the information sent to each of the sub-array receivers that is necessary to calculate the phase difference of the local clock compared to the reference clock. The PLL can now be used to make the needed adjustments so that all local clocks have the same phase throughout the array.

The information sent from the reference GNSS receiver to each of the GNSS timing units is; which satellites to use, Doppler-shift, tracking chip, and expected phase and time. This information will allow the sub-array receivers to only be capable of tracking a low number of satellites, no more than six, and using the tracked phase differences to calculate the expected phase of the local PLL. Thus, full capability receivers are not needed, but instead a Field-Programmable Gate Array (FPGA) will be used with a GNSS RF-frontend to control the
Figure 4.2: Diagram of the EISCAT_3D LAAR receiver front-end and GNSS timing unit.

PLL.

4.3.2 GNSS Simplified Receiver

In general, an off-the-shelf GNSS L1-receiver is rated to produce a clock with an error of less than 50 ns, which is about 1000 times higher than the necessary 50 ps accuracy. However, specific conditions apply to this GNSS timing system that improves the accuracy and will relax the requirements on the receiver, such as:

- A short base-line system, i.e. the maximum distance between two GNSS antennas is 300 m which infer all significant external errors, such as atmospheric, ionospheric, and ephemeris errors, in this application to be common over the array.

- A common high accuracy reference clock is available throughout the array to all receivers, which removes a significant part of the clock drift errors between the receivers.

- Externally based selection of which satellites are to be used for the position & time solution to exclude any timing errors arising from the use of different geometry matrices in the position & time calculations.
- All receivers are stationary and the time constant of the change of cable length in the reference clock distribution can be expected to be in the order of 30 min. This enables the use of a very long integration time, up to 30 min, of the timing solution which will reduce thermal noise significantly.

- Phase measurements from one satellite only is sufficient to calculate the timing error between the sub-arrays since the relative position of each receiver is known with good accuracy. However, more satellites will increase accuracy and also reduce any positional error of the GNSS antenna.

- No integer ambiguity solution is necessary, since the relative position of the receivers is known with good accuracy and the absolute time difference between the receivers is insignificant, only the phase of the distributed clock is important.

These conditions all relax the requirements on each of the GNSS timing units to the point where each timing unit only need to be capable of tracking a low number of satellites, measuring the phase of each satellite and then calculate the timing and position error of its own location relative to the reference antenna. With the use of an FPGA, the correlators and tracking can be built in hardware and the calculation part can be built in software with the use of existing processing cores that are available for many FPGAs, all within a single chip.

### 4.3.3 Test Setup for Concept Evaluation

Test measurements have been performed in an outdoor environment during windy winter conditions, clear weather at -10°C and wind speeds up to 20 m/s in gusts, with two antennas (Novatel GPS-702-GG) placed randomly, but precisely surveyed, at approximately 5 m distance from each other placed on a rooftop to simulate the conditions in the EISCAT 3D LAAR. Intermediate Frequency (IF) data from the antennas were collected during a one hour measurement with a NordNav Multi-FrontEnd receiver and then post-processed in a Matlab script to calculate time and position difference between the antennas. As clock reference a rubidium frequency standard (PRS10, Stanford Research Systems, Inc.) was used.

The software used for post-processing is in-house developed and includes all the necessary functions that a future FPGA-based GNSS timing unit would need, such as tracking, phase calculation and position & time estimation. The position & time estimator is based on the Least-Squares estimator to provide better accuracy when the equation system is overdetermined. The actual equation used is based on equation (7.14) in [32]. Since the integer ambiguity \((N)\) is known in this case, the equation will be reduced to

\[
(\phi_{ar} + N_{ar})\lambda = G \begin{bmatrix} x_{ar} \\ b_{ar} \end{bmatrix} + \varepsilon_{\phi,ar}
\]  

where \(\phi_{ar}\) is the phase difference between the test- and reference antennas, \(N_{ar}\) is the integer ambiguity, \(\lambda\) is the wavelength of the carrier signal, \(G\) is the geometry matrix, \(x_{ar}\) is the position difference \(b_{ar}\) is the time difference, and \(\varepsilon_{\phi,ar}\) is the remaining errors. Subscripts \(a\) indicates test antenna and \(r\) the reference antenna.

This equation is solvable for the unknowns \(x_{ar}\) and \(b_{ar}\) giving the Least-Square estimation of the position and time difference between the test antenna and the reference antenna.

When running the post-processing a time resolution of 0.5 s was chosen for each position & time solution. This resolution should as a minimum be achievable in a future hardware implementation of the system.
4.3.4 Results

After running the post-processing on the collected data, the expected and measured phase differences between the test- and reference antennas were plotted for each tracked satellite, see Figure 4.3. The expected and measured values are almost identical except for a bias that differs from satellite to satellite. The bias is an indication of a timing difference between the two antennas, or rather the receivers connected to the antennas. The sign of the bias in this plot differs simply because the plot ignores the integer ambiguity for clarity.

![Figure 4.3: Measured and expected phase difference between test- and reference antennas. Dashed lines indicate the expected phase differences calculated from the antenna position difference and current satellite geometry, and the solid lines indicate the actually measured phase differences. Integer ambiguity is not resolved in this plot and the data has been unwrapped to make the plot more perspicuous. No averaging has been applied to the data.](image)

After solving Equation 4.1 for all collected data points, $d\mathbf{x}_{ar}$ was plotted to give an insight in the accuracy of the system, see Figure 4.4. The baseline error seen in the figure has a standard deviation of $\sigma_x = 14.77$ mm, $\sigma_y = 4.97$ mm, and $\sigma_z = 13.93$ mm respectively in ECEF-coordinates.

An averaging filter of 5 min of data was also applied to the solution to achieve a less noisy measurement which is indicated in the figure with a dashed line. When using the filter, the
Figure 4.4: Baseline error between test- and reference antennas. The dashed lines indicates that a 5 min averaging filter has been applied.

standard deviation is reduced to $\sigma_x = 4.50 \text{ mm}$, $\sigma_y = 1.88 \text{ mm}$, and $\sigma_z = 6.63 \text{ mm}$ respectively. The $b_{ar}$ values were also plotted, see Figure 4.5, with and without the 5 min averaging filter. It is obvious from the figure that a clock bias exists between the two receivers, which can be detected with this setup and calculated with relatively simple means, see Equation 4.1.

Without averaging, the standard deviation of the clock difference is $\sigma = 50.11 \text{ ps}$, which is within the usability range for the timing of the EISCAT_3D LAAR. With the 5 min averaging filter applied, this is reduced to $\sigma = 20.52 \text{ ps}$. To ensure that the oscillations in Figure 4.5 are actually a normally distributed timing error, a histogram of the time difference is shown in Figure 4.6.

4.3.5 Conclusions

The use of a GNSS based picosecond accuracy timing system has been shown to be feasible to achieve using relatively simple calculations. The need for a timing error of less than 50 ps for the EISCAT_3D LAAR has been met with margin using only 5 min of integration time, yielding a timing error of approximately 21 ps.

Due to the simplifications of the GNSS position & time calculations possible for the highly
Figure 4.5: Time difference between test- and reference antennas. The dashed lines indicates that a 5 min averaging filter has been applied.

application specific conditions that apply for the EISCAT 3D LAAR, a GNSS timing unit can be created in an FPGA to a relatively low cost compared to other timing distribution solutions.
Figure 4.6: Histogram of the time difference between test- and reference antennas using 50 bins. The standard deviation of the data is 50.11 ps. No averaging has been applied to this data.
5 Beamforming

5.1 Introduction

Since the EISCAT_3D radar will be a Large Aperture Array Radar (LAAR), it is necessary to use time-delay beamforming instead of the more commonly used phase-delay beamforming. This is necessary since the length of the shortest radar pulse that the radar needs to receive is smaller than the size of the array aperture. Therefore, if phase-delay beamforming were to be used, signal energy would be lost whenever the illuminated part of the array is less than the array itself. This is because the phase-delay method only cares about the phase of the signal, i.e., it can at most delay one whole wavelength of the incoming signal. At 224 MHz, the wavelength is roughly 75 cm which is a lot smaller than the maximum length inside the array aperture which can be up to 300 m. By using time-delay beamforming, this problem is removed since any delay can be compensated for by the system.

To achieve a versatile array radar, pointing angles, number of simultaneous beams and rapid redirection of beams are essential. In a time-delay beamforming array, this translates into a high number of different delays for each antenna element in the array that have a short switching time. This can be done in the analog domain by introducing different delay-lines in the signal path for each antenna element, or it can be done in the digital domain by delaying the signal through filters. The benefits of the digital solution easily outrank the analog counterpart; it is a lot faster in switching, it requires less hardware per beam and per antenna element, and it provides an easy way to increase the number of beams after the system has been built. The main drawback of the digital system is the difficulty to achieve the necessary timing accuracy, see Chapter 4, to achieve the beamforming. For the EISCAT_3D project, digital time-delay beamforming was chosen to be the most appropriate solution.

To have an expandable and reliable simulation tool, each part of the physical system has been designed separately so that parts can be added, updated and removed without affecting the rest of the system. The incoming signal to the array is created as highly over-sampled data and then fed through each step of the receiver system, i.e., anti-aliasing filters, S&H, ADC and beam-former. White noise is added to simulate background and temperature noise, and timing jitter is added to provide means to evaluate the effect of non-ideal timing.

5.1.1 Beamforming Layout

The positioning of the EISCAT_3D system receive sites are shown in Figure 5.1 and 5.2, yielding two different sets of sweep variables and mounting directions for the Yagi antennas used in the array. The values in Figure 5.2 are corrected for the curvature of Earth, which is 0.009°/km. This also affects the azimuth width of the receiver arrays, since the sending array have a 30° wide beam pointing straight up. As seen in Figure 5.2, the maximum beam-steering limits necessary to see the whole beam in Tromsø are larger than the -3 dB beam-width of the Yagi antennas, see Figure 5.3. Thus, the Yagi beam-width is a limit in the setup of the test array, and this limitation is used in the simulations done below. The antennas have been permanently mounted as the beam-steering limits listed in Table 5.1 suggest, but with a maximum steering limit of ±30°.
Figure 5.1: Geographical locations of the send and receive sites of the existing EISCAT UHF system. The EISCAT_3D system will be located at similar locations.

Table 5.1: Beam-steering limits for the different receive sites.

<table>
<thead>
<tr>
<th>Limit</th>
<th>Site 1</th>
<th>Site 2</th>
</tr>
</thead>
<tbody>
<tr>
<td>Elevation</td>
<td>46.09°±30.61°</td>
<td>65.22°±42.45°</td>
</tr>
<tr>
<td>Azimuth</td>
<td>0°±38.03°</td>
<td>0°±68.55°</td>
</tr>
</tbody>
</table>
Figure 5.2: Extents of the needed beam steering capabilities of the different receive sites of the EISCAT 3D system to be able to see the whole $\pm 30^\circ$ wide sending beam from Tromsö. $\alpha$ and $\beta$ are the two receive sites, $\sigma$ notates azimuth direction, and $l$ and $h$ is low and high steering limits in elevation. Earth’s curvature has been taken into account.
Figure 5.3: Gain pattern of the Yagi antennas used in the EISCAT_3D test array, as applied in the simulations. The pattern is a sinc curve fitted to the actual data points from the Yagi antenna data sheet.
5.2 Digital Time-Delay Beamforming

As stated in Section 4.1, the overall demand on the timing error of the EISCAT_3D radar is to have a standard deviation of less than 160 ps. To reduce the timing error introduced by the hardware, the Analog-to-Digital Converter (ADC) is placed as close to each antenna element as possible as described in Chapter 3. After the incoming signal has been digitized through under-sampling at 80 MHz in the ADCs, each signal needs to be digitally delayed so that the incoming signal from each antenna element is matched to every other antenna element in the array. After the antenna elements are lined up correctly, the sole remaining step is to sum the signals to create a beam.

The delaying of the signal can be divided into two parts, whole sample delay which delays by multiples of the sampling frequency, and Fractional Sample Delay (FSD) which delays the signal parts of a sample. This first part is easily achieved in the digital domain by dropping samples, and the second part can be implemented by digital filters. For the EISCAT_3D project, Finite Impulse Response (FIR) filters were chosen.

5.2.1 Finite Impulse Response Filters

To achieve the requested beam-steering granularity for EISCAT_3D, the minimum step size of the delay for each element must be below 13.1 ps [33]. This granularity is not affected by the standard deviation of the timing error, but instead contributes to the error.

There are five main design criteria to consider when designing a time-delay FIR filter; group delay error, phase error, amplitude error, filter length and coefficient resolution.

- The group delay error is important since creating a group delay is what we are trying to achieve. Thus any error in the group delay reduces the delay accuracy from each antenna element, causing less than optimal beam-forming. It also causes our broad band signal to be distorted since different frequencies in the band will have different delays. The design criteria for the beam-forming granularity sets the minimum fractional delay; 1/1024th of 12.5 ns yields delay granularity of \( \sim 12.2 \) ps, and thus a quantization error of \( \pm 6.1 \) ps. This leaves only \( \pm 0.55 \) ps of error to the filter design since the total error is set to be \( \leq 13.1 \) ps, which have proved very difficult to achieve. By increasing the number of filters to 8192, (for a memory cost of 81 kB for a set of 1024 filters), the beam granularity is reduced to \( \sim 1.5 \) ps, leaving \( \pm 5.8 \) ps for the filter design.

- Since group delay is the derivative of phase with respect to frequency, it should not be necessary to optimize on phase delay as well. However, this is only true for a single filter by itself. The phase shift of any one filter is not important, but the difference between filters is, i.e., if the phase is shifted 30° by some filters and -150° by others, they would cancel out and thereby degrade the beam. To prevent this from happening, the phase error of each filter must be small, i.e. within the previously stated \( \pm 5.8 \) ps, thus the difference between the filters will be equally small. Since both the group delay error and the phase error manifest themselves as time errors they can be added together. Thus, it is the sum of the two errors that must meet the \( \pm 5.8 \) ps accuracy demand.

- The amplitude error is important since it will affect the beamwidth, side-lobes and nulls. This is because beam-forming works by constructive and destructive interference between antenna elements through summing of signals, so if some signals are larger than others they can affect the summing adversely. The main effect is beam-widening, but
even a few percent of amplitude error will cause only limited beam-widening. Awaiting more specific simulation results, we have set a conservative limit of \( \leq 1\% \) for the amplitude error.

- The length of the filters, or the number of taps, directly affects the amount of hardware necessary to realize the filters. Assuming an FPGA with the same calculation clock speed as the sampling rate, a 36 tap long filter will require 36 multipliers to be realized.

- Similarly, the coefficient resolution also affects the necessary hardware as the resolution of the multipliers must match the filters.

The demands on the FIR-filter design are summarized in Table 5.2.

<table>
<thead>
<tr>
<th>Criteria</th>
<th>Limit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Group Delay Error</td>
<td>( \leq 5.8 \text{ ps} )</td>
</tr>
<tr>
<td>Phase Error</td>
<td>( \leq 1% )</td>
</tr>
<tr>
<td>Amplitude Error</td>
<td>( \leq 1% )</td>
</tr>
<tr>
<td>Filter length</td>
<td>36 taps</td>
</tr>
<tr>
<td>Filter resolution</td>
<td>18 bits</td>
</tr>
</tbody>
</table>

### 5.2.2 Filter Design Results

After optimizing the design of the FIR-filters created for the ESICAT_3D beamformers and studying data sheets of different hardware, it was concluded that FIR-filters with 18-bit coefficient resolution and a length of 36-taps would fulfill the set demands, see Figures 5.4 - 5.6. As can be seen, the criteria of less than 5.8 ps timing error is met for all delays over all frequencies.
Figure 5.4: Amplitude errors for each filter (fractional delay) over the EISCAT 3D frequency band for a 36-tap long filter set with a coefficient resolution of 18 bits.
Figure 5.5: Group delay errors for each filter (fractional delay) over the EISCAT_3D frequency band for a 36-tap long filter set with a coefficient resolution of 18 bits.
Figure 5.6: Phase errors for each filter (fractional delay) over the EISCAT_3D frequency band for a 36-tap long filter set with a coefficient resolution of 18 bits.
5.3 Simulation Results

While to the date of writing this document, there has been no testing of the beamforming filters in hardware, extensive simulations have been performed to evaluate the accuracy of the filters [28]. However, hardware testing is under way and should be completed before the end of the project.

All simulations in this section have been made with a 12-by-4 element array, see Table 5.3. This will be a sparsely populated array, but since no interference signals are generated in the simulations, no grating lobes will disturb the simulations. By using a sparsely populated array in simulation, calculation times are reduced dramatically.

Table 5.3: Simulation settings for the 12-by-4 test array that was used to generate the test plots.

<table>
<thead>
<tr>
<th>Setting</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency</td>
<td>205 MHz</td>
</tr>
<tr>
<td>Pulse length</td>
<td>200 ns</td>
</tr>
<tr>
<td>PSK # pulses</td>
<td>8</td>
</tr>
<tr>
<td>PSK Code</td>
<td>10101010</td>
</tr>
<tr>
<td>Time jitter</td>
<td>$\sigma = 160$ ps</td>
</tr>
<tr>
<td>Noise level</td>
<td>-60 dB vs ADC range</td>
</tr>
<tr>
<td>Signal level</td>
<td>-84 dB vs ADC range</td>
</tr>
<tr>
<td>ADC resolution</td>
<td>14 bit</td>
</tr>
<tr>
<td>Beam-form filter resolution</td>
<td>18 bit</td>
</tr>
<tr>
<td>S&amp;H bandwidth</td>
<td>640 MHz</td>
</tr>
<tr>
<td>Inter-element distance</td>
<td>220 m</td>
</tr>
<tr>
<td>Azimuth angle of array</td>
<td>$0^\circ$</td>
</tr>
<tr>
<td>Elevation angle of array</td>
<td>$15.48^\circ$</td>
</tr>
</tbody>
</table>

As can be seen in Figure 5.7, the pointing accuracy is well within the stated $0.06^\circ$ limit for all levels of introduced timing jitter. In Figure 5.8 the amplitude accuracy is plotted and it can be seen that for an introduced timing jitter of 160 ps, the -0.2 dB limit is maintained.
Figure 5.7: Steering accuracy plot for 36-taps optimized filter set. 1000 runs at each jitter setting on the 12-by-4 test array.
Figure 5.8: Amplitude accuracy plot for 36-taps optimized filter set. 1000 runs, array 12-by-4 at jitter 160 ps, amplitude jitter $\approx 0.187$ dB. The red bar marks the ideal amplitude that would be achieved if no errors were introduced.
Bibliography


