Receiver Architecture of the THousand Element Array (THEA)

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ABSTRACT

As part of the development of a new international radio-telescope SKA (Square Kilometre Array), an outdoor phased-array prototype, the THousand Element Array (THEA), is being developed at NFRA. THEA is a phased array with 1024 active elements distributed on a regular grid over a surface of approximately 16 m². The array is organised into 16 units denoted as tiles. THEA operates in the frequency band from 750 to 1500 MHz.

On a tile the signals from 64 antenna elements are converted into two independent RF beams. Two times 16 beams can be made simultaneously with full sensitivity by the real-time digital beam former of the THEA system. At the output of each tile the analog RF signal from a beam is converted into a 2×12 -bit digital quadrature representation by a receiver system.

A double super-heterodyne architecture is used to mix the signal band of interest to an intermediate frequency of 210 MHz. The IF-signal is shifted to baseband by means of a partly digitally implemented I/Q mixer scheme. After a quadrature mixer stage, the I and Q signals are digitised by means of 12 bit A/D converters at 40 MS/s. Implementing a part of the mixing scheme digitally offers the flexibility to use different I/Q architectures, e.g. Hartley and Weaver mixer setups. This way the effect of RFI in different mixing architectures can be analyzed. After the digital processing, the samples are multiplexed, serialised and transported over fibres to the central adaptive digital beam former unit where the signals from all tiles are combined giving 32 beams.

This paper focuses on the design choices and the final implementation of the THEA system. In particular, the receiver architecture is addressed. A digital solution is presented, which enables switching between a Hartley and a Weaver based mixer scheme.

Keywords: Radio astronomy, phased array, RFI rejection, receiver concept, digital receiver, broad band

1. INTRODUCTION

The international radio astronomy community can hardly wait for the next generation radio telescope to come. A sensitivity, comparable to the Hubble Space Telescope at optical frequencies, should become available at radio frequencies by means of the Square Kilometre Array SKA.^{1,2}

An important boundary condition of (future) radio telescopes is that they should be able to operate despite the ever increasing pollution of the radio spectrum. An example of increased usage of the electromagnetic spectrum is the exponential growth of mobile phone systems all over the world. Also, for new systems, that are and will be developed by the telecommunication industry, more frequency space is needed. In the past, only small parts of the available radio spectrum were dedicated to radio astronomy, but today almost every frequency becomes attractive for radio science. A typical example of this is the development of the LOw Frequency ARray (LOFAR).^{3,4}

To build such a large radio telescope which should be robust to radio frequency interference (RFI), new and attractive technologies are investigated. A very promising and technology pushing concept is a solution based on phased array principles. Beam steering is realized with electronic circuits and no moving mechanical parts are required. Another advantage is the possibility to build a telescope with multiple independent beams. Several astronomers can use the instrument and make independent observations simultaneously. At NFRA, a research and development path has been outlined focused on the phased array concept. A road-map for SKA has been defined and it is based on the development of a series of prototypes with increasing complexity. By organising the development this way the necessary expertise to build SKA is developed along the time path. Already in 1996 the basic idea was

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implemented by means of the first prototype, the Adaptive Antenna Demonstrator (AAD).⁵ This was an indoor laboratory scale model and adaptive RFI nulling based on spatial filtering implemented by means of a phased array was demonstrated.

The second demonstrator in the row was the One Square Meter Array (OSMA).⁶ In Fig. 1 a picture of OSMA in the NFRA anechoic antenna measurement room is shown. With OSMA, a collecting antenna area of 1 m^2 was implemented with 64 active antenna elements and the system operates from 2 to 4 GHz. Several adaptive beam forming algorithms and calibration techniques are developed and tested with OSMA.^{7,8}

Figure 1. A picture of the One Square Meter Array (OSMA) in the NFRA antenna test range. The array consists of 8×8 active Bow-Tie antenna elements. Two rows with dummy elements are situated at the edges to decrease the finite array edge effects.

The third demonstrator, the THousand Element Array (THEA), is an outdoor system and is currently developed at NFRA. The array consists of 1024 antenna elements distributed over an area of 16 m^2 . The system operates in an optimised frequency range from 750 to 1500 MHz and it is designed to provide an instantaneous bandwidth of 20 MHz.

A general overview of the system is given in the next section. After this, the RF beam former is discussed and last the receiver architecture is described. In particular, attention is paid to the last stage of the receiver. A novel digital implementation makes it possible to switch online between two different mixer schemes.

2. THEA SYSTEM OVERVIEW

Besides the increased complexity, a new aspect of the third demonstrator (THEA) is that the antenna array will be operated outdoors. The 1024 antenna elements of the array are equally distributed over 16 units denoted as tiles. This array topology allows us to experiment with several array configurations by simply changing the tile positions.

2.1. Architecture outline

In Fig. 2 an overview of the THEA system is presented. In this diagram the tiles are configured on a regular grid. Each tile provides two digitised baseband I/Q output signals which correspond to the two independent RF beams. The signals from the tiles are transported over optical fibres to the THEA back-end. The back-end will be situated in the NFRA main building at a distance of approximately 150 m from the tiles. The second beam forming stage is implemented digitally in the THEA back-end. There, the 2×16 I/Q signals are combined in the real-time Adaptive Digital Beam Former (ADBF) providing 2×16 output beams. The Reduction Acquisition and Processing (RAP) unit

Figure 2. *THEA system overview*

performs the post processing on the beams. It provides a configurable 16 GFlops processing platform. Examples of its capabilities are: implementation of a real-time spectrometer operating on 2 full bandwidth beams; implementation of a real-time cross-correlator using 2 beams; a 1 second simultaneous record of 32 full bandwidth beams. The main control of the system is implemented by means of the Thea Experimental CHassis (TECH) subsystem, which provides also the main user interface. TECH also plays an important role during the calibration phase of the system. Furthermore, the settings for a set of RF beams can be calculated and the data required to set the pre-calculated beams can be stored into memory on the corresponding tiles. An actual beam is achieved by programming the right combination of phase shift and gain into each signal path of the array. During the measurement phase, when all the required beam data is available locally on each tile, fast beam switching can be performed. Equipment for power supply and required Local Oscillator (LO) signals is hosted in a small building in the neighbourhood of the tiles.

2.2. Tile architecture

The antenna system and the front-end electronics are integrated in a tile housing to protect it against fluctuating weather conditions and moisture. The housing of a tile is made of a low cost foam. For maintenance purposes a tile can be opened as shown in Fig. 3(a). The topside of a tile is protected by a radome, which covers the antenna system.

A cut out in Fig. 3(a) gives an impression of the column beam former boards with integrated antenna elements oriented vertically below the radome. The beam former consists of 16 column beam former boards. For reasons of manufacture-ability two column beam former boards are required to span the width of a tile. Each column beam former contains 4 antenna elements and the size of a tile is approximately $1 \times 1 \text{ m}^2$. The column beam formers are connected directly to a “row” beam former board placed horizontally below the column beam former boards. Here, the signals from the column beam formers are combined resulting in two independent RF beams. The beam former topology is discussed further in the next subsection. The receiver electronics, Front-End Controller (FEC) and tile DC supply system are placed in the bottom compartment of the tile housing.

Figure 3(b) shows a functional block-diagram of a tile. Since the RF beam former features two independent RF beams, two receivers are provided within a tile. An analysis of the sensitivity and dynamic range requirements is described in Ref. 9. A receiver is divided into three units: an IF unit, an IF to Digital Conversion (IDC) unit and an Optical Tile Transceiver (OTT) interface. Three LO signals are provided to a tile and a beam is tuned to a specific

Figure 3. (a) Mechanical representation of an opened tile. (b) Functional block-diagram of a tile.

RF frequency by altering the frequency of the first LO signal (LO_1). The other two LO frequencies can be fixed. By applying two different LO_1 frequencies the two independent RF beams are tuned to a different RF input frequency.

In the IDC unit, the I/Q baseband signal is digitised by means of two 12-bit ADC's with a sample frequency of 40 MHz. A central 40 MHz sampling clock is distributed optically to each receiver by means of the incoming fibre channel. The I/Q output data is multiplexed and transported to the back-end over the outgoing fibre.

The Front-End Controller takes care of setting the phase shifters, variable gains and time-delays in the beam former to switch to a specific antenna beam. The FEC is capable of storing data corresponding to pre-calculated RF beams. The data is provided to the FEC by means of RS422 serial communication links.

3. WIDE BAND BEAM FORMER

The beam formers applied in THEA are based on Vivaldi antenna elements. The elements are linear polarised and one polarisation is implemented with the beam former. First an array operating at a higher frequency was developed. This scale model was suitable for measurements in an anechoic chamber and results showed good agreement with the predicted results from electromagnetic simulations with the Method Of Moments (MOM).¹⁰

Based on these results, an array for the THEA frequency band is developed. In Fig. 4(a) the shape of a Vivaldi element is shown and in Fig. 4(b) the predicted Voltage Standing Wave Ratio (VSWR) is shown of an element in an infinite array radiating broadside.

(a)

(b)

Figure 4. (a) Vivaldi notch element for THEA. Overall length = 31.5 cm; overall width = 11.25 cm. Notch aperture is 9.25 cm wide. Triangles indicate grid for MOM analysis. (b) Predicted VSWR of a THEA antenna element in an infinite array configuration radiating broadside. It is optimized for a frequency band of 0.5 - 1.5 GHz.

Figure 5. Block-diagram of the beam former architecture.

In Fig. 5 a block-diagram of the complete beam former is shown. It shows the 16 column beam former boards on the left side and the row beam former board on the right side. Four antenna elements are integrated on a column beam former board. Each element is, via a balun, connected to the input of a Low Noise Amplifier (LNA). After the LNA the signal is split and this enables the two independent RF beams feature of the beam former. Hereto, all the

beam former electronics required for steering one beam is implemented twice. The first beam forming stage consists of phase shifters and gain control implemented with vector modulators. Each vector modulator has a digital interface by which a required phase shift and gain combination can be programmed. On the row beam former board, for each quadrant, the signals from four column beam formers are combined. Digital programmable time delay units* can be applied to the four signals corresponding to each quadrant before they are combined to give the resulting RF beam. With this beam former architecture a solution with a high integration level is achieved.

An important function of the RF beam former is to provide RFI suppression. The beam former is not designed to provide frequency selectivity, on the contrary, it has a broad band frequency response. However, the frequency characteristics of the antenna elements and the RF beam former will provide some frequency selective attenuation for out of band RFI signals. Prominent RFI signal components will be attenuated by means of deterministic nulling. Hereto an array antenna gain pattern with a null into the direction of the RFI is programmed into the beam former. Realizing an attenuation of an RFI signal of more than 30 dB is possible.

4. THE IF UNIT

The main task of the IF unit is to select a channel from the broad band signal coming from the beam former and translate this to an output Intermediate Frequency with a centre frequency of 210 MHz and an output bandwidth of 20 MHz.

Figure 6. Functional block-diagram of the IF unit.

A conventional super-heterodyne architecture is used for the IF unit as shown in Fig. 6. The input stage provides switchable gain control. The extra input gain is mainly meant for the case that the higher part ($f \geq 800$ MHz) of the input frequency band is selected. In this case, the high pass filter F_1 attenuates signals at the lower part of the spectrum resulting in less spurious frequency products caused by intermodulation. The IF unit is designed to perform sufficient without the extra input gain, but for experimental purposes the performance can be increased when this feature is used. With the first mixer stage (M_1) the input signal frequency band of interest is selected and mixed to an IF frequency of 2680 MHz. Here the first (spectral) channel selectivity is implemented with band pass filter F_2 . Notice that spatial selectivity is already realized with the RF beam former. The bandwidth of filter F_2 is relatively broad (200 MHz), however its main task is to decrease performance degradation caused by intermodulation. The second mixer stage shifts the signal to the output IF centre frequency of 210 MHz. The second band pass filter (F_3) in the output stage of the IF unit provides most of the IF channel selectivity. Last, the signal is amplified to a level which is suitable for the IDC unit described in the next section. For increased flexibility the output amplifier stage has a gain control range of more than 25 dB.

5. IF TO DIGITAL CONVERTER

The functionality of the IDC Subsystem is to transform a 200-220 MHz real IF input signal into a digital single-sided baseband signal with spectral contents ranging from 0 till 20 MHz. The input signal consists of white receiver noise, one or more strong narrow-band signals and a very weak white-noise signal originating from an astronomical source. The power of the narrow-band signal can be up to 45 dB above the noise-floor.

*The time delay units are optional. They can be integrated easily on the row beam former board.

The basic requirements for the frequency translation and digitisation are:

- The suppression of the image must be better than 35 dB.
- The sum of the power of all spurious products must be 10 dB below the total noise power.
- The quantisation step-size should be one fourth of the RMS receiver noise power. This way the receiver noise is quantised using 2 bits. Furthermore, the A/D converter must have sufficient dynamic range to cope with the strong narrow-band signals.

5.1. General mixer concepts

Two different mixer concepts will be presented for implementing the functionality described above: the Weaver based mixing scheme¹¹ and the Hartley based mixing scheme.¹² In Fig. 7 a block-diagram of the Weaver based implementation of the IDC unit is given.

Figure 7. Block-diagram of the Weaver based scheme

The first quadrature LO-signal is located in the middle of the band of interest which runs from 200 to 220 MHz. After the first mixer stage the band of interest runs from -10 MHz to +10 MHz when both mixer output signals are considered as one complex signal. Although, the bandwidths of the separate output signals corresponding to the band of interest are 10 MHz, the bandwidth of the complex signal is 20 MHz. For a real baseband signal, the bandwidth is defined as the distance between the highest frequency component and zero while for the complex signal we define the bandwidth as the distance between the highest and the lowest frequency component. To distinguish between these bandwidth definitions from here on the bandwidth of a complex signal will be referred to as complex bandwidth. The low pass filters limit the bandwidth of the (real) mixer output signals to 30 MHz. This signal is digitised and filtered digitally in order to limit the complex bandwidth to 20 MHz. Notice that this requires a bandwidth of 10 MHz for both digital low pass filters (LPF_B). In the output stage the combination of mixers and adders can be interpreted as a full complex mixer, i.e. a mixer which realizes the operation $z(t) = \exp(j\omega_{lo2}t) \times w(t)$ where both $z(t)$ and $w(t)$ are complex. From this it follows that with $\omega_{lo2} = 2\pi \times 10$ Mrad/s, the band of interest is positioned at 0 to 20 MHz.

The general characteristics of this architecture are:

- Because of crosstalk of the LO input from the first mixer to its RF input, a relatively strong DC component will be present after the mixer. After the second mixer stage this is translated into a “spike” at 10 MHz.

- The image frequencies are within the band of interest from 200-220 MHz so it is not possible to benefit from the suppression of the band pass filter of the IF unit.
- The requirements for the analog Low Pass Filter are relatively relaxed because the sampling is done at 40 MHz. This implies that the filters have a roll-off area from 10 to 30 MHz.
- The implementation of the digital mixers is straightforward. The sample frequency equals 40 MHz while the two LO signals, cosine and sine, have a frequency of 10 MHz. The multiplication with the LO can be interpreted as multiplying the subsequent samples with 1, 0, -1, 0, ... for the cosine and with 0, 1, 0, -1, ... for the sine.

Figure 8. Block-diagram of the Hartley based scheme. The $\mathcal{H}$ represents the Hilbert operation.

A block-diagram of the Hartley based scheme is depicted in Fig. 8. Instead of locating the LO at 210 MHz, it is located at one of the edges of the band being 220 MHz. After the mixer, the band of interest will be located at 0-20 MHz. The analog signals are converted into digital signals after the low pass filter. After the A/D converters the combination of Hilbert transforms and adders realize an additional complex filter stage to eliminate any residual neighbour channel components. It can be verified that a digital implementation resembles

$$Z(\omega) = 2\epsilon(\omega)W(\omega), \quad \text{for } -\omega_s/2 < \omega \leq \omega_s/2, \quad (1)$$

where ω_s is the angular sample frequency, $Z(\omega)$ and $W(\omega)$ are the resp. Fourier transform of $z(t)$ and $w(t)$ and

$$\epsilon(\omega) = \begin{cases} 0 & \text{for } \omega < 0, \\ 1 & \text{for } \omega > 0. \end{cases} \quad (2)$$

The characteristics of the Hartley based scheme are:

- The DC component after the mixer which is caused by the crosstalk from the LO onto the RF input is located at the edge of the band of interest.
- The image frequencies which are within the band from 220 to 240 MHz are subject to extra suppression by the band pass filter preceding the IDC subsystem.
- In order to loose as less bandwidth as possible and to prevent severe aliasing, the Low Pass Filter before the A/D converter must have a roll-off just before 20 MHz.
- The Hilbert transform is realised in the digital domain. In order to loose as less bandwidth as possible, a Hilbert filter using a relatively large number of taps is required.

5.2. Non-idealities in mixer structures

Practical implementations of the discussed mixer architectures result in a degradation of signals in the band of interest and a finite suppression of signals out the band of interest. The effects of three sources of non-idealities are discussed in this section for both mixer architectures:

- non-ideal LO signals;
- non-ideal filtering;
- sampling.

A gain or phase difference between both branches of a mixer structure results in an undesired mixing product, known as the image. The gain or phase differences between both branches of a complex mixer structure such as the Weaver and Hartley mixer can in fact be represented by non-ideal LO signals. In stead of a pure cosine and sine as LO signal, there will be an amplitude and a phase difference. Writing this as a complex LO signal yields

$$l_o(t) = \exp(j\omega_{lo}t) + \frac{1 - (1 + \epsilon) \exp(-j\Delta\theta)}{1 + (1 + \epsilon) \exp(j\Delta\theta)} \exp(-j\omega_{lo}t), \quad (3)$$

where $\Delta\theta$ is the phase difference and ϵ the amplitude difference. The spectrum of the output signal of a practical complex mixer can be written as

$$Z(\omega) = W(\omega - \omega_{lo}) + \delta \times W(\omega + \omega_{lo}), \quad (4)$$

where δ is complex and defined by

$$\delta = \frac{1 - (1 + \epsilon) \exp(-j\Delta\theta)}{1 + (1 + \epsilon) \exp(j\Delta\theta)}. \quad (5)$$

Taking the logarithm of the absolute value of δ results in the image rejection ratio (IRR). The relation between the phase error (θ) and the gain error ($1 + \epsilon$) with the IRR as parameter is shown in Fig. 9.

Figure 9. The image rejection ratio as function of the gain error and phase error.

From Fig. 9 it is seen that much effort must be paid in making both branches of the mixer equal. In Weaver the image band falls within the signal band. Assuming a source at $220 - \Delta$ MHz at the input of the IDC results in a spurious product at $20 - \Delta$ MHz at the output of the IDC. This spurious product is the power of the original source suppressed with IRR. An IRR of 35 dB is assumed to be feasible for the used mixer structures.

To increase the selectivity for the IDC unit a bandpass filter is inserted in the IF unit. This filter and the analog low pass filter together determine the degree of aliasing, which is inherent to sampling. Non ideal filtering together with sampling results in signals out the band of interest which are folded over the original signal band.

The last digital filter stage in the IDC unit (a complex filter for Hartley and two real filters for Weaver) is used for suppressing signals from -20 till 0 MHz in the output signal, because a complex output signal is required. To preserve the 20 MHz signal band, the cut off region of the digital filter must be small.

The specifications of the filters are set, assuming a worst case scenario. This worst case situation occurs when measuring close to two TV transmitters located at 655.25 MHz and 679.25 MHz. In fact, the receiver is tuned through this range. At each position the sum of the total spurious products generated in the IDC unit is determined. The main goal is to tune as close as possible to the TV transmitters, still fulfilling the total spurious products demand. This can be interpreted as a quality measure of the unit.

From this it was found that for Weaver an area of 34 MHz (from 650.25 till 684.25 MHz) cannot be measured, i.e. the spurious product requirement cannot be met, within this range. For Hartley it was simulated that every frequency component could be measured. The following assumptions were used during simulation:

- no spurious products are generated in front of the IDC unit, so only the two TV transmitters are assumed at the input;
- if a TV transmitter is in the band of interest, then its power is not accounted as a spurious product;
- the IRR is 35 dB;
- the used filter characteristics are approximated.

Theoretical motivations for using Hartley or Weaver results in the conclusion that the performance of both depends on the specific situation. The selectivity in front of the A/D converters gives different results and the position of the image is not equal in both situations. Because of these differences it is advantageous to use different architectures for different situations. For example for an RFI signal in the band of interest Hartley should be applied, while for an RFI at 20 MHz above the band of interest Weaver is more attractive.

5.3. Implementation of the IDC unit

For using both Hartley and Weaver the same analog part can be used. For Hartley the LO frequency must be set at 220 MHz, while for Weaver the LO frequency equals 210 MHz. The implementation is depicted in Fig. 10.

Figure 10. *The architecture of the IDC unit.*

A phase shift circuit delivers in phase and quadrature LO signals for both LO frequencies (210 and 220 MHz). Both mixers are implemented in an ASIC (ASTRON01) to reduce gain error and phase error between both branches as much as possible. The IF amplifier is inserted for driving the A/D converter full scale.

It is desirable to implement only one type of analog filter for both mixing schemes. To prevent aliasing in the Weaver architecture (the aliasing is already limited due to the bandpass filter) a certain suppression must be reached at least at 30 MHz. For the Hartley architecture this frequency is at 20 MHz, which means that the cut off region

is within the signal band. Therefore an analog low pass filter is used with a cut off frequency of 20 MHz. At 30 MHz a suppression of 30 dB is reached. As discussed the digital part of both mixer structures differs. For Hartley a complex digital filter must be used, while in the case of Weaver two real digital filters in combination with a complex mixer can be used.

When implementing both digital architectures, three filters can be implemented, which requires additional logic. In fact the complexity doubles, because the complexity of a complex filter is two times the complexity of one real filter, assuming a similar performance. To reduce the complexity, the Hartley architecture is build in the same way as the Weaver architecture with one additional complex mixer before the digital low pass filter. This complex mixer is added to shift the 0-20 MHz band to a 20 MHz band around zero. Then, the same digital filter as in Weaver can be used. Switching between both mixer architectures is reduced to just bypassing the additional complex mixer or not bypassing it. Schematically this is shown in Fig. 11. The signal mix_{sel} controls which mixer structure is used. Note that also the LO frequency has to be changed from 210 MHz for Weaver to 220 MHz for Hartley.

Figure 11. The digital architecture of the IDC unit.

The part which requires the most gates in the digital architecture is the digital filter. It is chosen to use a Finite Impulse Response (FIR) filter because of its inherent phase linearity and stability. Enlarging the number of taps, results in enlarging the complexity because the number of multipliers also enlarges. The coefficients are quantised in 14 bits and the output of the digital filter is quantised to 12 bit. This results in a frequency response given in Fig. 12. The quantisation of the output limits the frequency response. The complete digital circuit is implemented using programmable logic of Altera. In this way the functionality of the digital circuit can be changed in system, or functionality can be added in the near future.

Figure 12. Two digital filter responses are shown. One shows the result after re-quantization of the output to 12 bit and the other one is for the case no output re-quantization is applied.

6. CONCLUSION

The third step towards SKA is now being developed at NFRA. THEA will receive astronomical sources by a phased array concept and is able to operate outdoors in a RFI hostile environment. A significant part of the RFI sources are reduced due to a beam former. In order to process the information of astronomical sources, the high frequency band is down converted to an IF band in the IF unit. From there, the signal is converted to baseband using a Hartley or Weaver architecture in the IDC unit. Within this scheme the A/D conversion takes place. Because the architecture is partly designed digitally, it is easy to switch between a Hartley or Weaver architecture. In this paper it was shown that only bypassing a digital mixer and tuning the LO frequency results in switching from Hartley to Weaver.

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