An Experimental S-Band Digital Beamforming Antenna

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Abstract — This paper describes a 12 channel S-band digital beamforming antenna aimed at radar applications. It has a 2.8-3.3 GHz agile band and the signal is sampled at IF. The design of antenna elements, calibration network, receiver modules and signal processing ASIC are discussed. Calibration procedures and experimental results, showing very good performance, are also presented. The purpose of the work has been to evaluate design parameters, requirements and limitations.

INTRODUCTION

With digital beamforming DBF [1] very accurate beam control can be obtained and various array signal processing methods can be applied. This includes multiple beams with very low sidelobes and adaptive pattern control, direction of arrival estimation, space-time adaptive processing, etc.

The different parts of a receiving DBF antenna are illustrated in Fig. 1. It consists of an array of antenna elements, a calibration network, receiver modules, AD-converters, digital down converters, channel equalizers, a digital beamformer and a control processor.

![DBF Antenna System](image1)

Fig. 1. A DBF antenna system.

The experimental antenna discussed here is shown in Figs. 2 and 3 and is designed with radar applications in mind. It includes the analog parts and AD-converters together with their control circuits plus relatively large digital memories in each channel. The digital signal processing in Fig. 1 is, due to cost and flexibility, performed off-line in a general purpose computer.

The antenna has 12 digital channels arranged as a linear horizontal array. It has an agile frequency band of 2.8-3.3 GHz and an intermediate frequency (IF) bandwidth of about 5 MHz. The AD-conversion is made at 25.8 MHz sampling rate with the IF at 3/4 of this frequency. The final down conversion and filtering is done digitally in the computer.

![Experimental Antenna Front View](image2)

Fig. 2. The experimental antenna front view.

![Experimental Antenna Rear View](image3)

Fig. 3. The experimental antenna rear view.

ANTENNA DESIGN

Array elements

The antenna array elements are stripline dipoles, arranged in a linear array of orthogonal linear subarrays with 4 dipoles each, as shown in Fig. 4. The dipoles are fed by a stripline to slotline transition forming a balun.

![Dipole Subarray](image4)

Fig. 4. Dipole subarray

The dimensions of the dipole, the open stub and shorted slotline forming the balun, and the transmission line transformer, are chosen to optimize the impedance match over frequency and scanning angle. The active impedance has been calculated using the method of moments on a some-
what simplified geometry and this result has been confirmed experimentally for single elements and for a subarray in a waveguide simulator. Advantages of dipole elements are that they are broadband and approximately single-mode elements which is an advantage when applying some corrections to obtain precise pattern control.

**Calibration network**

The calibration network by which the antenna, mainly the receiver modules, can be calibrated is a microstrip circuit behind the subarray substrates. It is of bidirectional type [1], where the thermal variations are minimized by using the geometric mean value of the signals when fed from two opposite directions. This direction is controlled by a diode switch within the network. The signal is coupled to each channel by a capacitive non-directional coupler. Together with a directional coupler on the antenna subarray substrate in Fig. 4 there is a total coupling of ~60 dB. This passive network can, like the passive antenna array which is not part of the internal calibration signal path, itself be calibrated by an external calibration procedure. This can be done with time intervals much longer than the calibration of the active, and hence more sensitive, receiver modules.

**Receiver Module**

The analog receiver module [2] (partly developed and manufactured by Ericsson Microwave Systems AB) that follows in each channel is a complete superheterodyne receiver with high dynamic range and good out of band signal suppression. It includes, as shown in Figs. 5 and 6, a protection limiter, a step attenuator, amplifiers, filters and mixers arranged in two down converting stages, in which the received radar signal is transformed into the final IF.

![Fig. 5. Schematic description of the receiver module.](image)

The antenna elements have a broad pattern so that any jammer will have almost full strength in the channels. Each receiver, including AD-conversion, must therefore have good image frequency rejection and high dynamic range.

The rejection also includes higher order image frequencies which requires a good front-end filtering also far above the passband. In order to obtain this, still using uniplanar techniques, a high performance microstrip edge coupled filter is used, which is compensated for the differences in phase velocity of the odd and even modes [3]. The first IF filtering (at 690 MHz) is done by a high-Q uniplanar compact hairpin filter [4]. The area of this filter is only \( \lambda / 8 \times \lambda / 4 \).

Variations within the final IF bandwidth, mainly due to filters, can be corrected for by digital equalizing filters but these variations should still be kept at a minimum.

Fig. 7 shows the budget for the receiver chain based on the component data given in TABLE I.

The individual gain of the twelve receiver modules were manually tuned to 50 ± 1 dB at 3 GHz. The output average thermal noise RMS voltage will then be 1.5 times the AD-converter quantization step. Typical measured data of the receiver modules are summarized in TABLE II.

**TABLE I**

<table>
<thead>
<tr>
<th>No</th>
<th>Comp. Type</th>
<th>G (dB)</th>
<th>NF (dB)</th>
<th>IP3i (dBm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>Limiter</td>
<td>-0.7</td>
<td>0.7</td>
<td>100</td>
</tr>
<tr>
<td>2</td>
<td>LNA</td>
<td>14.9</td>
<td>1.9</td>
<td>15.1</td>
</tr>
<tr>
<td>3</td>
<td>Step Att.</td>
<td>-0.6</td>
<td>0.6</td>
<td>100</td>
</tr>
<tr>
<td>4</td>
<td>RF BP-filter</td>
<td>-2.1</td>
<td>2.1</td>
<td>100</td>
</tr>
<tr>
<td>5</td>
<td>Mixer 1</td>
<td>-7.5</td>
<td>8</td>
<td>15</td>
</tr>
<tr>
<td>6</td>
<td>LP-filter</td>
<td>-0.5</td>
<td>0.5</td>
<td>100</td>
</tr>
<tr>
<td>7</td>
<td>Amplifier IF1</td>
<td>11.1</td>
<td>3</td>
<td>24</td>
</tr>
</tbody>
</table>

**Fig. 7. Receiver budget. Gain (G), noise figure (NF) and input third order intercept point (IP3i).**

**TABLE II**

<table>
<thead>
<tr>
<th>Radar Frequency (2.8-3.3 GHz)</th>
<th>Low band</th>
<th>Mid band</th>
<th>High band</th>
</tr>
</thead>
<tbody>
<tr>
<td>Gain</td>
<td>47.5 dB</td>
<td>50 dB</td>
<td>49 dB</td>
</tr>
<tr>
<td>Noise Figure</td>
<td>5.8 dB</td>
<td>4.6 dB</td>
<td>4.8 dB</td>
</tr>
<tr>
<td>Input 3rd order intercept point</td>
<td>-7 dBm</td>
<td>&gt;5 dBm</td>
<td>&gt;5 dBm</td>
</tr>
<tr>
<td>Image Rejection Mixer 1</td>
<td>&gt;50 dBc</td>
<td>&gt;65 dBc</td>
<td>&gt;65 dBc</td>
</tr>
<tr>
<td>Image Rejection Mixer 2</td>
<td>40 dBc</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Aliasing Rejection ADC</td>
<td>&gt;35 dBc</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Spurious Free Dynamic Range</td>
<td>&gt;108 dBHz²</td>
<td>&lt;65 dBHz²</td>
<td></td>
</tr>
</tbody>
</table>

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The AD-converter is a key component in any digital receiver since it implicitly limits the final IF frequency. A high IF permits the use of a receiver module with fewer mixer stages and filters with reduced complexity. An AD-converter with high dynamic range and high sampling frequency is thus required.

In the experimental antenna an 12 bit converter from Analog Devices (AD9026) is used. This makes it possible to use a sampling frequency of 25.8 MHz (equal to 4/3 of the IF frequency, and 1/26 of LO2 in Fig. 5). The sampled signals are temporarily stored in 12 digital memories of maximum 1 Mwords and later read by the controlling computer [5].

The digital down converter (DDC) in Fig. 1 transforms the digital signal to analytic form in an IQ filter and then feeds it through one or two decimate-by-two stages to a well filtered digital signal at 1/4 or 1/8 of the original sampling frequency rate. By this filtering the quantization noise from the AD converters is also reduced.

The equalizer (EQU) after the DDC is a finite impulse response (FIR) filter with programmable complex valued coefficients which should make the total transfer function in the channels frequency independent, or at least equal in all channels. The filter coefficients are obtained from a calibration procedure.

The beamformer calculates the output signals in a number of ordinary antenna beams by multiplying the signal vector (with one vector element per channel) with a number of weight vectors. In a more general situation the simple weight multiplication can be replaced by filters, for example to get true time delays or for space-time adaptive beamforming.

**Digital filter chip**

A digital CMOS ASIC containing digital down converter and channel equalizer stages has been developed in collaboration with the University of Linköping [6]. This circuit is intended to follow immediately after the AD-converter and to implement the digital filter chain of Fig. 8.

![Digital filter chip](image)

**Fig. 8.** Block diagram of digital filter chain.

The signal bandwidth is limited to 3 MHz in this case. A bandpass decimation in two separate stages is used to reduce the output sampling rate to 3.225 MHz. This chip performs nearly 700×10^6, 16 bit, fixed point arithmetic operations/second at an internal bit-rate of 51.6 MHz. The implementation is based on a multiple-path time-multiplexed architecture with distributed arithmetic processors. A similar architecture is also well suited for implementation of the digital beamformer in Fig. 1.

The chip is currently under test and the ultimate goal is to integrate it with the experimental antenna to provide real time processing in the channels.

**ANTENNA CALIBRATION AND MEASUREMENTS**

Antenna measurements have been performed in an anechoic chamber with a measurement distance of 6 meter. This is too short to be considered as a far field distance, in particular when trying to find the limitations of the antenna. However the main part of this can be compensated for as the other types of errors since the measurement geometry is known and the signal is measured in each channel.

Most corrections can be applied directly to the received signals or, more efficiently, to the weight coefficients. When using some adaptive and spatial signal processing algorithms where various manipulations are done, these two methods will not be identical (but still essentially equivalent).

Measurements have been done both with and without two passive dummy subarrays at each end of the active array. Fig. 2 shows the array with these included but the results reported below are for an array without these subarrays. This gives somewhat inferior results when simple corrections are applied but better results when the more complex corrections are applied.

Since all individual antenna element signals are measured separately, a very precise control of the conventional antenna patterns can be obtained.

By illuminating the antenna with a known incident wave, the amplitude and phase errors in the channels, as they appear in that particular calibration configuration, can be corrected for. Fig. 10 shows measured element patterns normalized to unity at θ = 0° and Fig. 11 total antenna pat-
terns when steering the main beam at various directions, applying −50 dB Chebyshev weights.

In these and all other plots the RF-frequency is 3 GHz.

Fig. 10. Antenna element patterns.

Fig. 11. Antenna patterns using simple channel corrections for 50 dB Chebyshev weights, steered at 0°, ±30° and ±60°.

Due essentially to mutual coupling the element patterns differ from each other, in particular for the edge elements, and the array sidelobes are higher than the nominal value.

The mutual coupling can, to a large extent, be compensated for, [7], [8]. If the antenna elements can be considered to be of the minimum scattering type, which is a good approximation for the dipole elements used in this array, the measured output signals \( \mathbf{v}_{\text{out}} \) and the isolated, open circuit, antenna voltage \( \mathbf{v}_{\text{oc}} \) are related by

\[
\mathbf{C}_{\text{cmp}}^T \mathbf{v}_{\text{out}} = \mathbf{v}_{\text{oc}}
\]

where \( \mathbf{v}_{\text{oc}} \) and \( \mathbf{v}_{\text{out}} \) are vectors of dimension \( N \) (the number of elements) and \( \mathbf{C}_{\text{cmp}} \) is an \( N \times N \) compensation (decoupling) matrix. This matrix is a function of the array scattering matrix and the scattering matrix of the receiver modules, both as seen from the element reference planes in the feeding gap. It also relates the dipole currents to the feeding wave amplitude in a hypothetical transmit mode via

\[
\mathbf{a}_{\text{trans}} = \mathbf{C}_{\text{cmp}} \mathbf{i}_{\text{dipole}}
\]

provided the scattering matrices of the receive and transmit modules are related by

\[
\mathbf{S}_{\text{trans}} = \mathbf{S}_{\text{receive}}^T
\]

The matrix \( \mathbf{C}_{\text{cmp}} \) is now estimated by a slightly different procedure as compared to [7]. By illuminating the array from a large number, \( M \), of directions, \( \theta_m \), \( N \) linear equations for each direction is obtained. The open circuit effective antenna length \( h(\theta) \) of the elements can be assumed to be unknown too. It will, however, be assumed that all elements have equal antenna lengths, which will be true for minimum scattering elements and large ground planes. Since the measurement distance is short, first order parallax corrections to the local angle, \( \theta \), are also applied.

Thus there are \( M \times N \) equations in \( N^2 + M \) unknowns,

\[
\mathbf{v}_{\text{out}}(\theta_m) \mathbf{C}_{\text{cmp}} = h(\theta_m) \mathbf{e}_{\text{inc}}(\theta_m), \quad m = 1, \ldots, M
\]

where \( \mathbf{e}_{\text{inc}}(\theta) \) is the incident electric field. From these equations the compensation matrix \( \mathbf{C}_{\text{cmp}} \), and the normalized antenna length \( h(\theta) \) can be found by iteratively solving for them in a least squares sense.

From the above calibration measurements a simple correction table with \( M \) entries are also obtained. It can be used in direction of arrival estimations where corrections can be applied for each test direction separately. It is also used as reference when later updating the calibrations.

Using this method corrected element patterns, shown in Fig. 12, are obtained by plotting the antenna signal

\[
\mathbf{v}_{\text{antenna}} = \mathbf{w}^T \mathbf{C}_{\text{cmp}}^T \mathbf{v}_{\text{out}}
\]

including a nearfield compensation, with only one element in the weight vector \( \mathbf{w} \) being nonzero.

Fig. 12. Antenna element patterns using channel decoupling corrections, and normalized effective area at the receiver outputs.

This figure also shows normalized effective antenna area in the main beam direction, referred to the receiver outputs, as function of scan angle for uniform nominal weights. The norm of the corrected weight vectors \( \mathbf{w}_{\text{cmp}} = \mathbf{C}_{\text{cmp}} \mathbf{w} \), is then kept constant. If the receiver modules has equal and uncorrelated output noise the plotted curve will also be proportional to the antenna gain over noise ratio, \( G/T \). Since the noise in the decoupled signals \( \mathbf{C}_{\text{cmp}}^T \mathbf{v}_{\text{out}} \) are cor-
related, this parameter will not of the same shape as the element patterns in Fig. 12.

As seen, the element patterns are now very similar to each other. There is some ripple of approximate magnitude ±0.05 dB close to broadside due to reflections, of -45 dB, between the flat antenna ground plane and the source wall of the anechoic chamber when they are parallel. (These reflections will of course mean that some performance will be lost if the calibration data should be applied in a free space environment.)

Fig. 13 shows nominal 50 dB patterns with the above type of correction applied which are much closer to the ideal patterns than those shown in Fig. 11.

![Antenna patterns using channel decoupling corrections for 50 dB Chebyshev weights, steered at 0°, ±30° and ±60°.](image)

The low level of the remaining differences between the channels after this calibration is further illustrated in Fig. 14. This shows the antenna output signal as function of the sines of the angle of incidence and beam scanning angle respectively. Very low sidelobe (-80 dB) Chebyshev weights has been applied so that any sidelobes above the plotting floor of -60 dB are essentially due to imperfections.

![3D antenna patterns using a compensation matrix](image)

The peaks at the corners are the edge of the main beam, folding into the visible space when scanning to 90°. There are some visible sidelobes when the beam is scanned close to 90°, which was also seen in the ±60° scan patterns in Fig. 13. This is due to edge scattering from the, now strongly illuminated, rolled edge of the ground plane. There is also a small sidelobe close to the main beam for small angle scanning, related to the broadside ripple of Fig. 12 discussed earlier. This has been demonstrated by replacing the rolled edges with large serrations giving decreased 90° scan sidelobes and increased small scan sidelobes.

Fig. 15 illustrates the stability of the antenna and the measurement system. Here a two week old $C_{cmp}$ was used, updated with a simple calibration at $\theta = 0°$ to correct for drift in the receiver modules. As seen the result is still very good.

![3D antenna patterns using an old compensation matrix](image)

Since each receiver module has relatively narrow band anti-aliasing and image rejection filters, there will be differences in the frequency dependance of the channels. The patterns above was obtained for a single frequency, where it was also calibrated. Fig. 16 shows the normalized transfer functions for all channels within a 6.45 MHz (= $f_{samp}/4$) bandwidth.

![Normalized gain in all channels versus frequency.](image)

The above frequency variations can be corrected for by an equalizing filter in each channel, in which the coeffi-
coefficients are based on calibration data. Such a filter is also included in the digital filter chip discussed earlier. Fig. 17 shows the transfer functions of the channels, relative to one reference channel, after the signals have passed through an optimized 15 tap equalizing FIR-filter. The sampled signals has here been decimated by a factor of 4 to 6.45 MHz and the equalization bandwidth is 5 MHz. The resulting RMS inter-channel deviation, or Cancellation Ratio, was then improved from −29 dB to −75 dB which corresponds to the noise limit of these calibrations. (For 12 taps, a 3 MHz band and a decimation by 8, as in the filter chip in Fig. 9, the Cancellation Ratio becomes −56 dB.)

Since the origin of most of this frequency variation is within the IF circuits it will be relatively insensitive to carrier frequency with only a slowly varying gain slope. These frequency variations are in these measurements also affected by the reflections in the chamber.

![Frequency variations after the channel equalization filters, normalized to one reference channel.](image)

Fig. 17. Frequency variations after the channel equalization filters, normalized to one reference channel.

Most of the variations have its origin in the receiver modules but a small part is also due to frequency variations of the mutual coupling and of the input match of the receiver modules. Thus the channel equalization is strictly valid for one angle of incidence and the compensation matrix used, $C_{emp}$, at one frequency only. Both corrections are however very useful for all angles and frequencies within the IF bandwidth of the receivers. A low sidelobe pattern will only get slightly increased sidelobes. The −50 dB sidelobe peaks for 0° scan in Fig. 13 will increase to a peak of −43 dB for a 5 MHz bandwidth.

Calibrating the array at different IF frequencies will give slightly different compensation matrices. By using a few such matrices, valid for different IF frequencies, a short FIR-filter with matrix coefficients, or $N$ scalar filters for each conventional beam, can be designed which will improve the results even further. Fig. 18 shows patterns for 13 frequencies within a 5 MHz bandwidth, and 6.45 MHz sampling rate, using such a filter with 3 taps, after the regular equalizers.

![Antenna patterns using 50 dB Chebyshev weights for 13 frequencies within a 3 MHz bandwidth.](image)

Fig. 18. Antenna patterns using 50 dB Chebyshev weights for 13 frequencies within a 3 MHz bandwidth.

**CONCLUSION**

A digital beamforming antenna has been designed, built and tested. It has 12 receiver channels including AD-converters. The receiver modules have high dynamic range and good out-of-band rejection, convenient for radar applications. The signal processing is performed off-line, but the feasibility of high performance ASIC implementation of the front end processing has also been demonstrated. Measurements show that, with appropriate corrections, very precise pattern control can be obtained.

**ACKNOWLEDGMENT**

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**REFERENCES**


