A 32-element frequency-steered array antenna for reflectometry in W-band

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1 Introduction

For Doppler reflectometry [1, 2], a steerable and focussed beam is necessary. For this, current systems employ tiltable antennas or fixed antennas and tiltable mirrors [3, 4, 5]. Within the vacuum vessel of a fusion experiment, however, space is very limited and movable parts with articulated joints that need greasing are problematic. Also, mechanical tilting of mirrors or antennas is comparatively slow and might need a feedback system to be exact. Drive and feedback systems also must cope with the high magnetic field strengths and the radiation close to the plasma [5].

In view of this, a frequency-scanned phased array antenna is proposed as a Doppler reflectometry front-end. Frequency-scanning does not need any movable parts or electronics in the front-end which, therefore, can be a passive all-metal block with nothing but waveguide channels in it. This makes the system much sturdier and more reliable. With a series-fed system, one waveguide connection to the back-end per antenna suffices and the feed structure can be made very compact. By using an array of 32 close-set H-plane sector horns, a well focussed beam can be achieved while retaining the compact front-end size.

2 Phased array

An array antenna consists of multiple radiating elements. The antenna pattern is determined by the spacing and the amplitudes of the individual elements. The phase relations between the elements determine where in space the contributions from the individual elements interfere constructively and, thus, at which angle the beam is irradiated [6, 7].

![Diagram](image1)

Figure 1: Two 4 element antenna arrays. Left: parallel fed with variable attenuators (rectangles) and phase shifters (circles). Right: series feed with variable attenuators.
The left side of figure 1 shows an array with 4 identical radiating elements that are linearly arranged and equally spaced. In the feeding lines to the individual elements, variable attenuators and phase shifters determine the amplitude and phase distribution. At an angle $\theta$, the phase difference between the contributions of two neighboring elements is [7]

$$\psi = \beta d \sin \theta + \alpha,$$

(1)

where $\beta = 2\pi/\lambda$ is the free-space propagation constant, $d$ is the element spacing, and $\alpha$ is the phase difference between two neighboring radiating elements. The main beam will be irradiated at the angle that meets the condition $\psi = 0$,

$$\theta_0 = \arcsin \left( -\frac{\alpha}{\beta d} \right).$$

(2)

Without active phase shifting and with a fixed path difference $\Delta l$, as illustrated on the right in figure 1, the phase difference $\alpha = 2\pi \Delta l/\lambda_k$ is determined by the waveguide wavelength alone. The scan angle can, thus, be swept by sweeping the frequency. The beam shape is determined by the amplitude distribution. In a series feed, as illustrated on the right of figure 1, the element amplitudes can be determined by the coupling strengths towards the array elements.

3 Design

The principle is demonstrated for a W-band (75 – 110 GHz) Doppler reflectometry front-end. In the design process, CST Microwave Studio® (CST) is used to perform simulations in order to accurately define dimensions and determine parameters.

For beam shaping, a linear array of 32 identical H-plane sector horns is used. The scanning range is $-20^\circ \leq \theta_0 \leq 20^\circ$ around broadside. With the element spacing of $d = 2$ mm, grating lobes are suppressed for the entire scan range even at 110 GHz ($\lambda = 2.73$ mm) since it meets the requirement [8]

$$d < \frac{\lambda}{1 + |\sin \theta_{\text{max}}|} = 2.03 \text{ mm}.$$

(3)

The Taylor line source method [9, 6, 7, 10] is used for the array synthesis. The design goal is to achieve $\text{BW} \approx 4^\circ$ with $\text{SLL} = -35$ dB. The parameter $n$ is set to 5, to minimise the amplitude at the aperture edge and ensure a monotonic aperture distribution. The parameters directly related to the array pattern and the resulting half power beam width (BW) for the W-band centre frequency, $f_{\text{ctr}} = 92.5$ GHz, are [9]

$$A = \frac{1}{\pi} \arccosh(-\text{SLL}) = 1.503,$$

$$\sigma = \frac{n}{\sqrt{A^2 + \left(\frac{n-1}{2}\right)^2}} = 1.054,$$

(4)

$$\text{BW} = 2 \left| \arcsin \left( \frac{\sigma \lambda}{\lambda L} \sqrt{(\arccosh(-\text{SLL}))^2 - \left(\arccosh \left(\frac{-\text{SLL}}{\sqrt{2}}\right)\right)^2} \right) \right| \approx 3.6^\circ,$$

(5)

where $\lambda = 3.24$ mm is the free-space wavelength for 92.5 GHz and $L$ is the aperture length.

A helical WR10 waveguide serves as a phase delay line. By making the individual windings long enough, the phase difference between neighboring elements gets large enough to allow several complete scans in W-band, effectively dividing it into sub-bands, each roughly corresponding to a single radial measurement position. For the first prototype, the outer helix diameter is 15 mm resulting in a frequency agility, i.e. $\Delta \theta/\Delta f$, around broadside of circa $16^\circ$/GHz for $f \approx 90^\circ$. In total, one partial and five complete scans are possible in W-band with this helix diameter.

The 32 coupling elements are placed at the same azimuthal position, one on each winding. The CST model of the waveguide helix is shown on the left in figure 2. At the end of the feed line, a load is placed to absorb the amount of excess power that is needed since the last element cannot easily couple out all
the remaining power. The excess factor is chosen to be $\rho = 0.05$. To speed up the CST simulations, the model is simplified by introducing an equivalent straight structure with 4 coupling elements that was found by comparison with a helix with four windings. This equivalent structure can much more easily be scaled up to 32 elements for the necessary simulations.

![CST model of the helical series feed line with 32 coupling holes towards the radiating elements.](image)

Figure 2: *Left:* CST model of the helical series feed line with 32 coupling holes towards the radiating elements. *Right:* coupling strengths for the 32 elements. *Insertion:* CST model of a coupling hole.

Long holes are used as coupling elements. They are sized according to the aperture distribution, taking into account the losses along the delay line, the power coupled out or reflected at the previous coupling elements, and the excess power to be deposited in the load. The resulting coupling strengths are shown on the right in figure 2. The dimensions for the individual coupling holes are determined with CST models, like the one shown in the insertion in figure 2, to achieve the desired coupling strength distribution [10].

From a simulation of the complete equivalent straight structure, the S-parameters of all ports are determined and used for calculating the array factors for various frequencies, c.f. figure 3. There is an additional side lobe close to the main lobe that exceeds the $-35$ dB level. Also, the side lobes exactly opposite of the main beam are slightly more pronounced. Apparently, the coupling holes’ design does not perfectly reproduce the Taylor-$\pi$ aperture distribution. However, the deviations are very small.

![Array factors calculated from the coupling strengths of the simulated structure.](image)

Figure 3: Array factors calculated from the coupling strengths of the simulated structure.

Due to the regular spacing of the coupling elements on the series feed line, there are frequency ranges for which the reflection is drastically increased, the so-called stop-bands. For phase differences of $2\pi$ or $\pi$ between neighbouring elements, the reflections of the 32 coupling elements interfere constructively. Two approaches for reducing the stop-band effect are currently being considered. One uses a gradual change in the waveguide cross section breaking the periodicity of the array and reducing the resonant reflections. The other has additional reflecting elements in the waveguide, one close to each coupling element [11].
The main disadvantage of both approaches is that they increase the reflection level outside the resonances. Moreover, with a changing waveguide cross section, the array becomes focussing and needs a phase-correcting mirror which makes the front-end bulkier and more complex. On the one hand, the compensation elements do not increase the overall reflection level as much as the changing cross section does. On the other hand, they are not inherently broadband but designed for a specific frequency, achieving the desired effect only for a limited frequency range. The reflection coefficients for the three designs are compared in figure 4. Every second reflection maximum corresponds to broadside irradiation ($\Delta \varphi = 2\pi$).

![Figure 4: Simulated reflection coefficients $|S_{11}|$ of the three designs (uncompensated/continuous cross section change/discrete compensation elements). Note that the frequency differences between maxima for the feed structure with discrete compensation elements are smaller because of the larger helix diameter of 20 mm.](image)

4 Manufacturing

The first prototype was realised as an uncompensated feed with an outer helix diameter of 15 mm. First, the wave guide channels were milled into a brass rod. To reduce losses, the waveguide walls were covered with a thin copper layer by electroplating. Then it was filled with wax and, again by electroforming, closed with a thick layer of copper. The electroplating was done by a private company. The steps are shown in figure 5.

![Figure 5: The helical waveguide’s manufacturing process. Top left: milled into a brass rod. Top right: electroplated with a thin copper layer and filled with wax, photography courtesy of Galvano-T GmbH. Bottom: closed with a thick copper layer.](image)
The closed feed structure was then milled into a cuboid and the flanges and coupling holes were milled into its walls at the appropriate positions. The result is shown in figure 6. The dimensions are 110 mm × 35 mm × 19 mm. After removing the wax, the prototype was ready for testing.

Two further prototypes are being produced in order to test approaches for compensating the stopband effect. At the time of writing, a second prototype with a linearly changing cross section is ready for testing while a third one is in the last production stage. This third feed, featuring discrete compensation elements, has a larger outer helix diameter of 20 mm and a core of CuCrZr which has better mechanical properties than copper and better conductivity than brass.

![Prototype of the feed structure](image)

Figure 6: Prototype of the feed structure. Top: connection to the antenna array. Bottom: connections to WR10 waveguide.

5 Testing

Before starting the production of the actual feeds, the design and mechanical feasibility of the coupling holes and compensation elements were checked. A straight V-band waveguide test bed with exchangeable parts was used for the tests, since the results can easily be applied to W-band. The left of figure 7 shows the comparison of simulation and measurement for a long hole.

![Diagram comparing simulation and measurement](image)

Figure 7: Left: comparison of simulated and measured coupling strength for a long hole with length \(a\) = 2.4 mm, width \(b\) = 1.8 mm, and height \(h\) = 1.5 mm. Right: array amplitude and phase distribution measured at 90 GHz, the solid lines depict the desired values.

The test measurements for the compensation elements show a sharp reduction of \(|S_{11}|\) almost completely removing any reflection close to the compensation element’s design frequency but not over a large bandwidth.

The amplitude and phase distribution of the array elements were measured with a receiver horn close to the element under test while all other elements were terminated with absorbers. The results can be seen on the right in figure 7. The amplitudes of elements 16 to 32 are below the design values, especially for element numbers higher than 20. This might be due to additional reflections caused by mechanical imperfections in the waveguide. Elements 22 to 32 also show a clear phase deviations indicating that the helix diameter or the waveguide cross section deviate from the design there.

Figure 8 shows measured antenna patterns for scan angles of +7°, −10°, and −22°, roughly corresponding to the simulations in figure 3. The frequencies are shifted slightly, which is most likely due to
mechanical deviations from the design waveguide cross section or helix diameter. The conspicuous side lobes from figure 3 are much enhanced in the measurements. For $-10^\circ$, the side lobe level approaches $-10 \text{ dB}$. The source of the enhanced side lobes is not yet established beyond doubt. However, the deviations in the amplitude and phase distribution that were measured clearly point towards mechanical causes.

![Figure 8: Antenna patterns measured at different frequencies.](image)

### 6 Conclusion

The presented results show that using a frequency-steered phased array antenna for Doppler reflectometry is a promising approach. However, the frequency agility needs to be increased to allow enough radial measurement positions for obtaining conclusive radial profiles. This can be done by choosing larger radii.

A practicable manufacturing process was demonstrated by producing first prototypes of the feed. The tests show deviations from the desired aperture distributions and antenna patterns. These are attributed to mechanical errors in the production and, to a much lesser extent, to details in the coupling design that do not perfectly fit the desired aperture distribution. It remains to be proven that, by improving the fabrication process, the enhanced side lobes can be removed and the desired beam qualities can be fully achieved.

The horn array needed to complete the front-end also needs to be developed. Until now, preliminary straight H-sector horns were used for the testing purposes.

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References


