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*Research Article*

# Slotted waveguide antenna with a near-field focused beam in one plane

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**Abstract:** A slotted waveguide providing a two-dimensional near-field focused beam is presented. This antenna focuses the beam in the *E*-plane and provides a wide beam in the *H*-plane in order to illuminate a linear array as reflect- or transmit-array antenna with a small width (100 mm) and a very large length (1530 mm) located in the near-field region. The simulated field distribution on the array is found to be in very good agreement with the measurement of a prototype at 9.41 GHz.

## 1 Introduction

The near-field focusing technique is used in several applications, such as imagery sensing, characterisation of materials or in biomedical devices. To focus a beam at a distance $d_f$, called the focal distance, from the antenna, the phase has to be adjusted along the radiating aperture to obtain a locally plane wave-front (Gaussian beam property) in the desired spatial location. Today, many antennas allow focusing a beam in the near-field region: planar array [1], horn with a dielectric lens [2, 3] or rectangular slotted waveguide [4]. But all these antennas focus a beam in three dimensions (spot). In some cases, an antenna that focuses along one plane and provides a broad beam in the orthogonal plane (two-dimensional focused beam) can be interesting, such as in optics for testing a long material with a small width and a long length or in microwave application for feeding a linear reflect- or transmit-array antenna.

In [5–7], the authors design two-dimensional focusing antennas using leaky-wave principle considering a tapered microstrip line [5], a rectilinear slotted waveguide [6] or substrate integrated waveguide technology [7]. However, these three geometries use partly or completely a printed circuit technology. In our case, the focused antenna will be used as a feed for a transmit-array antenna in navigation radar technology. In such applications, the antenna has to accept high power with a good radiation efficiency, so we will favour a complete waveguide technology. In [8], a horn antenna with a biconvex dielectric lens inserted in its aperture has been investigated to provide a two-dimensional focused beam. But the addition of the lens in the aperture of the *H*-plane sectoral horn reduces the beam width in the *E*-plane, and the maximum ratio length/width of the beam at the focal point was equal to 8.

This paper presents a slotted waveguide antenna dedicated to the illumination of an array with a length of 1530 mm and a width of 100 mm (ratio length/width equals to 15.3), located in the near-field region of the feed. This structure focuses the field along one plane in order to illuminate the width of the array and provides a wide beam along the orthogonal plane to illuminate the length of the array. Therefore the spot size specifications are determined by the dimensions of the illuminated array. As the whole system (array + feed) has been developed for a navigation radar embedded inside a compact integrated mast, the architectural restrictions enforce the focused beam characteristics. For example, applying the transmit-array technology to design the illuminated antenna, the standard beamwidth aperture of a navigation radar leads to a 96-element linear array in the horizontal plane, which corresponds to a length of 1530 mm. The restriction in the dimensions of the mast induces a maximum vertical size of the transmit-array antenna of 100 mm, with a maximum distance of 500 mm between the feed and the transmit-array antenna. To minimise the spill-over losses, the normalised field level at the edge of the array has to be close to −10 dB. So, the focused antenna has to be designed to reach these spot sizes restrictions, which are summed up in Table 1.

In this paper, we present a focused antenna that reaches these specifications with low losses and high power handling. This paper is organised as follows: In the first part, we carry out the theory of the near-field focusing technique with a slotted waveguide. In the second part, we present the results of the simulation with CST Microwave Studio. Finally, these theoretical results are compared with the measurement results.

## 2 Design process of the near-field focusing slotted waveguide antenna

### 2.1 Presentation of the near-field focusing technique

Let us consider a rectangular antenna aperture ($A$) of length $L$, as shown in Fig. 1, with a field distribution $E_{AP}(x, z)$ over this aperture as expressed in (1)

$$E_{AP}(x, z) = E_0(x, z) \cdot e^{-j\varphi(x,z)}\hat{z} \quad (1)$$

where $E_0(x, z)$ and $\varphi(x, z)$, respectively, are the magnitude and phase of the field distribution, and $(x, z)$ is the coordinates of the aperture.

As seen in [8] and considering the reference plane at $z = 0$, the phase distribution along the aperture needed to focus the beam at a distance $d_f$ must be equal to

$$\varphi_f(x, z) = \varphi_f(z) = k \cdot d_f\left(1 - \frac{1}{\cos(\beta)}\right) = k \cdot \left(d_f - \sqrt{z^2 + d_f^2}\right) \quad (2)$$

where $k = (2\pi/\lambda)$ is the free space constant.

Such a phase variation along the aperture can be obtained in different ways: the first solution consists in modifying the field distribution of horn antenna by inserting a dielectric lens inside its

**Table 1** Characteristics of the focused beam

| Characteristics | Value, mm |
|---|---|
| beamwidth at −10 dB in the horizontal plane | 1530 |
| beamwidth at −10 dB in the vertical plane | 100 |
| maximum distance of the focal point | 500 |

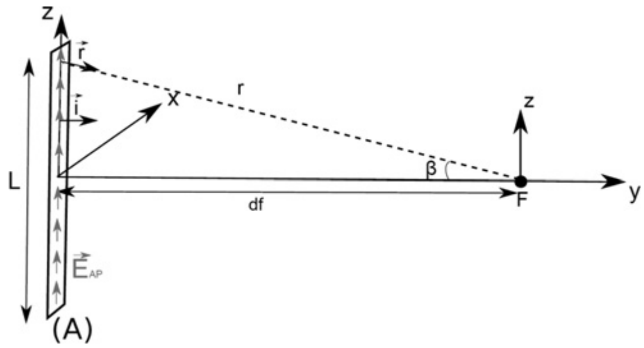


**Fig. 1** *Illustration of near-field focusing theory*

aperture, as shown in [8]; the second solution uses the modification of the complex propagation wavenumber to obtain the required phase variation; another solution is based on the synthesis of such a variation, thanks to discrete radiating sources. The second solution is proposed in this paper: a slotted waveguide antenna is designed, optimising the slots and their positions on the waveguide to obtain the desired field distribution.

It is important to notice that the previous formulation (2) uses the ray-tracing theory. Of course in the reactive near-field region, the electric-field behaviour is more complex and its computation needs much more efforts. But even if the theory presented in this paper is limited, it has the advantage to be very easy to implement and gives a good approximation of the waveguide design.

### 2.1 Design of the antenna

*2.2.1 Slot geometry:* The shape of the slots has been chosen using [9]. The I-slot design (Fig. 2*a*) provides a good coupling with the waveguide and radiates a relatively low cross-polarisation level compared to rotated linear slots. This slot geometry has been optimised on CST Microwave Studio at 9.41 GHz.

*2.2.2 Arrangement of the slots on the waveguide:* To focus the electrical field radiated by the slotted waveguide, the phase of the field radiated by each slot must be the same at the focal point, called *F* (Fig. 1).

Then, the location of each slot is enforced considering the phase distribution given in (2) and the propagation characteristics inside the waveguide; each slot location corresponds to a phase of the field inside the waveguide $\varphi_{wg}$, and must be equal to the desired phase $\varphi_f$.

The phase of the field, for a $TE_{10}$ mode, inside the waveguide is equal to

$$\varphi_{wg}(z) = \frac{2\pi z}{\lambda} \cdot \sqrt{1 - \left(\frac{f_c}{f}\right)^2} \quad (3)$$

where $f_c = (c/2a)$ is the cut-off frequency of the waveguide for a $TE_{10}$ mode and $a$ is the width of the waveguide.

The previous equation uses the propagation constant of the $TE_{10}$ mode of an unperturbed rectangular waveguide. However, the assignment of slots on the waveguide can obviously perturb the field inside the waveguide. This perturbation can lead to an error in the location of the slots and induces a shift in the position of the focal point. To highlight the impact of the slots assignment on the phase distribution, we compare in the following section the theoretical phase with the simulated one.

It is important to notice that the orientation of the slots allows us to use both the negative and the positive components of the field (Fig. 3). This capability will optimise the size of the antenna, considering the location where $\varphi_f(z) = \varphi_{wg}(z)$ and also the location where $\varphi_f(z) = \varphi_{wg}(z) + \pi$. Moreover, this capability to use both negative and positive components of the field also prevents grating lobes which appear when the distance is higher than $\lambda_0$. To take advantage of the alternation of the orientation, the slots have been done on the small side of the waveguide (called 'b').

Fig. 4 shows $\varphi_f(z)$, $\varphi_{wg}(z)$ and $\varphi_{wg}(z) + \pi$ as a function of $z$. The slot positions to focus the beam at a distance $d_f$ correspond with intersections indicated by '*'. In this example, $d_f = 500$ mm, $f = 9.41$ GHz and $a = 23$ mm. The position of each slot is presented in Table 2. We have chosen the maximum distance of 500 mm between the antenna and the focal point to reach a −10 dB beamwidth in the horizontal plane. The number of slots has been chosen to match with the specification of a −10 dB beamwidth of 100 mm in the vertical plane.

The final structure is a symmetrical centre-fed waveguide with 14 slots. The drawing of the antenna is shown in Fig. 5.

The slots layout is symmetrical according to the feed. The right-side slot locations are given in Table 2, considering the feed as the reference point.

## 3 Simulation and measurement results

### 3.1 Simulation results

The final slotted waveguide has been simulated using CST Microwave Studio (Fig. 5). Loads have been added at the ends of the waveguide to suppress reflected energy and quarter wave transformer inserted to obtain good reflection coefficient (Fig. 9). As expressed in the previous section, the insertion of slot on the waveguide can modify the field inside the waveguide and leads to

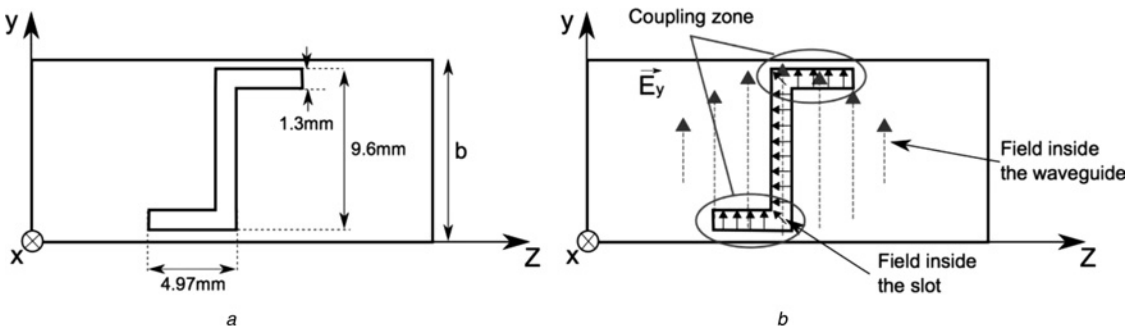


**Fig. 2** *Slot geometry*

*a* Design of the slot
*b* Field distribution inside the slot

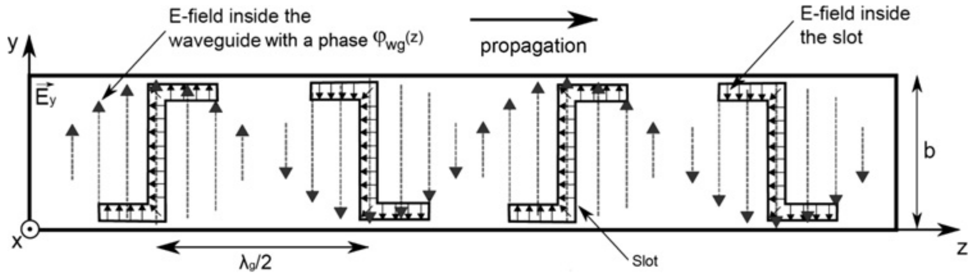


**Fig. 3** *Orientation of the slots to use both positive and negative components of the field*

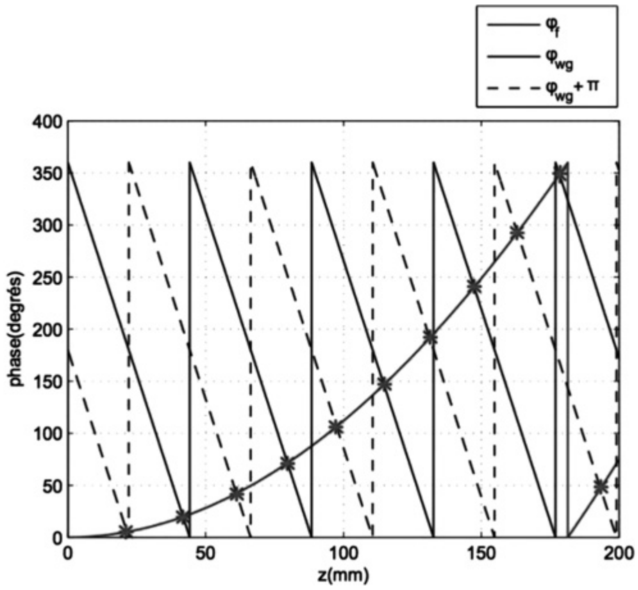


**Fig. 4** *Determination of the position of each slot on the waveguide*

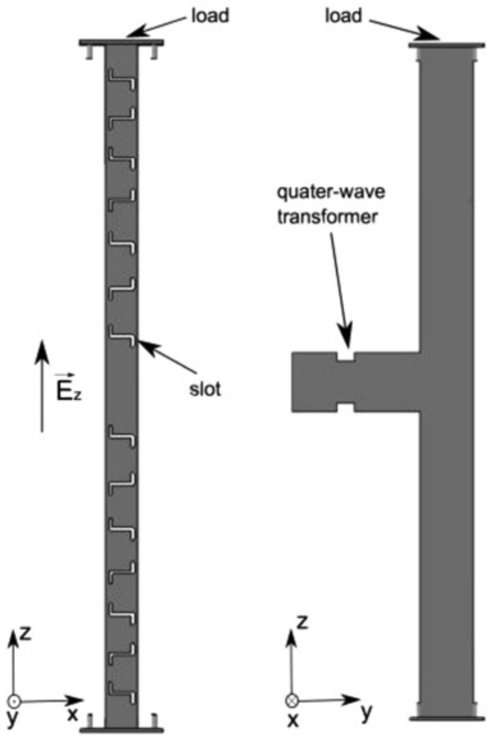


**Fig. 5** *Design of the slotted waveguide with CST Microwave Studio*

**Table 2** Slot positions

| Slot no. | 1 | 2 | 3 | 4 | 5 | 6 | 7 |
|---|---|---|---|---|---|---|---|
| *d*, mm | 21 | 42 | 61 | 79 | 97 | 114 | 131 |

a perturbation of the field and a wrong phase distribution. To verify the impact of the slot on the field inside the waveguide, we represent in Fig. 6, the desired phase variation along *z*-axis of the plane where the slots are located and the simulated variation. We can notice small differences in the phase distribution that could lead to a small shift in the plane-wave location.

Fig. 7 shows the $E_z$-field results in the *E*-plane in both amplitude and phase at 9.41 GHz. We can notice that the $E_z$-field is focused with a minimal width for the beam at 360 mm from the slotted waveguide. Moreover, at 360 mm from the slotted waveguide, we can notice a wavefront locally plane. The difference of focal distance between theory (500 mm) and simulation (360 mm) can be explained through two different reasons:

- First, as said above and shown in Fig. 6, the error in the phase distribution because of the wave perturbation inside the waveguide induced by slots insertion can induce a small difference between the desired and the actual plane-wave position.
- Second, the most probable reason of this difference between the theoretical location of the plane-wave and the simulated one is owing to the approximation used in the theoretical study. Indeed, the theoretical positions of the slots on the waveguide have been determined through (2). This equation is based on the well-known expression of the phase of the field as a function of the distance and the wavelength ($\varphi = 2\pi l/\lambda$). However, as said in the Section 2, in the near-field zone, the propagation of the field is more complex than the applied ray-tracing theory, which does not take into account coupling between I-slot. Moreover, in the theoretical part, we have considered that each slot radiates an isotropic pattern. All these approximations in the theoretical study can explain the

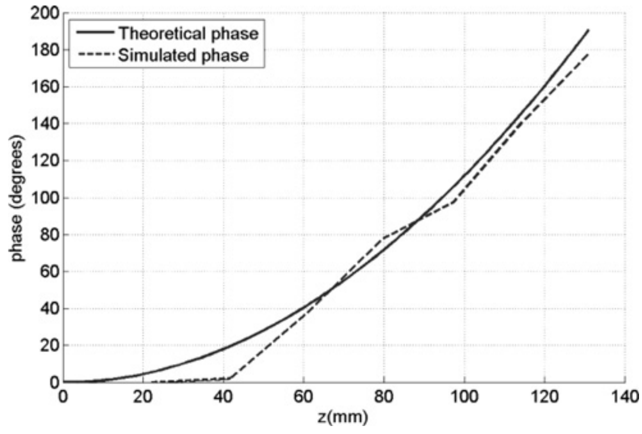


**Fig. 6** *Theoretical and simulated phase along z-axis in the slots cut-plane*


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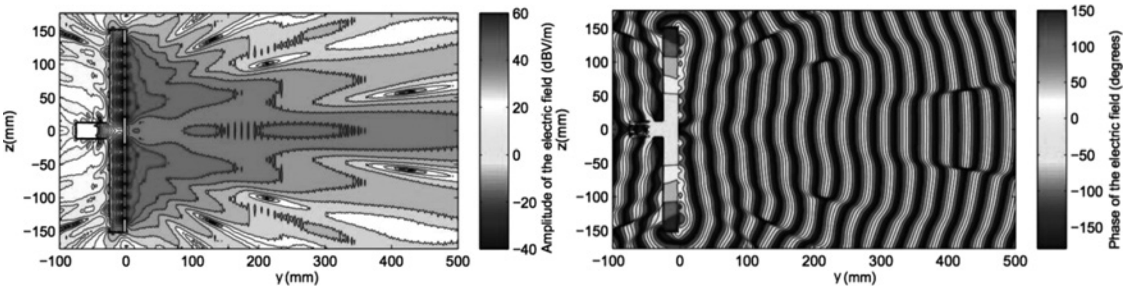


**Fig. 7** *Simulated electric field $E_z$ in amplitude and phase in the E-plane at 9.41 GHz*

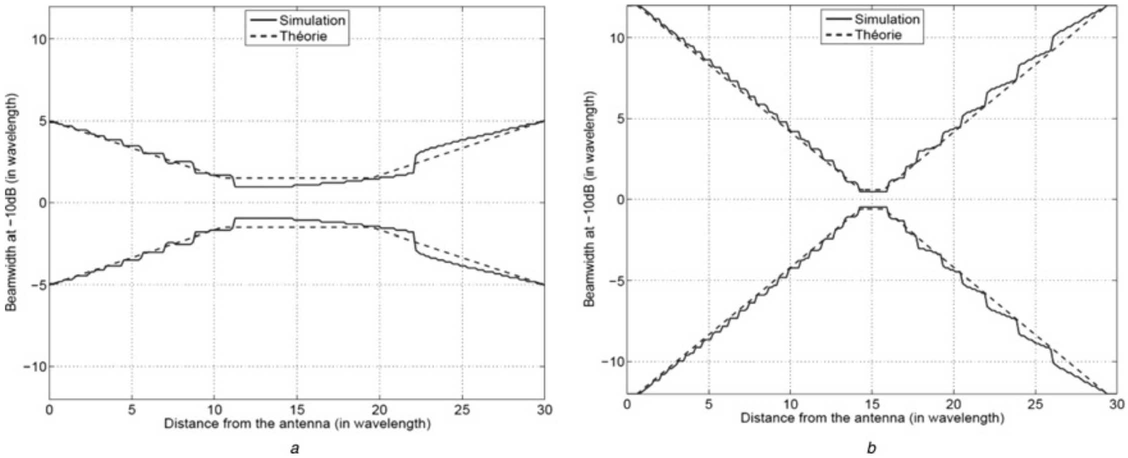


**Fig. 8** *Theoretical and simulated beamwidth at −10 dB for a radiating aperture equals to*

*a* 10λ
*b* 25λ

difference with the simulation results. However, the approach presented in this paper has the advantage to be easy to implement with rather good results. Indeed, as shown in Fig. 7, the beam is focused on a wide range of distance and the error in the position of the plane-wave will have no impact on the system performance.

Indeed, as expressed in [10, 11], the focus is not a single point, and we can consider that the field is still focused on a large area called the focal depth. Theoretically, the focal depth corresponds to the half power beam size in the axial direction. We can clearly notice in Fig. 7 that the field is still focused at 500 mm from the antenna.

As we can see in Fig. 8, this focal depth depends on the size of the antenna. When the radiating aperture is very large compared to the wavelength, the focus depth is very small and the focus is closed to a single point. Conversely, when the aperture becomes smaller (around 10λ in our case) the focal depth is larger with a maximum of intensity before the desired focal distance [10, 11].

As reported in Table 1, a 10 dB tapering is required in both vertical and horizontal plane for the focused beam at the focal distance. We specify that for the theoretical calculation of the −10 dB beamwidth presented in Fig. 8, we first calculate the field radiated by a discrete array of isotropic sources using the following equation (4) and then, we extract the beamwidth as a

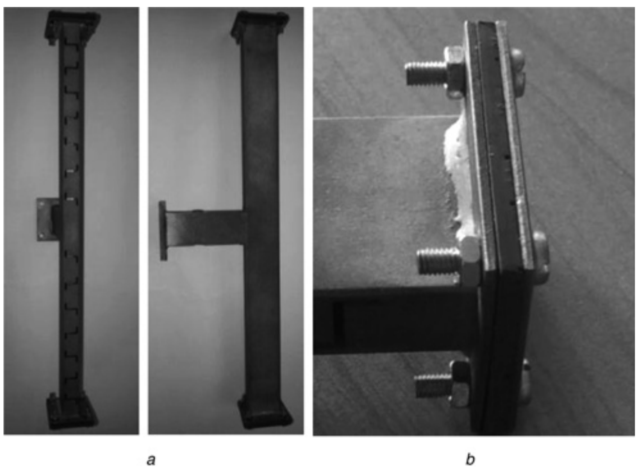


**Fig. 9** *Prototype of the proposed antenna*

*a* Proposed slotted waveguide antenna
*b* With a zoom on the load

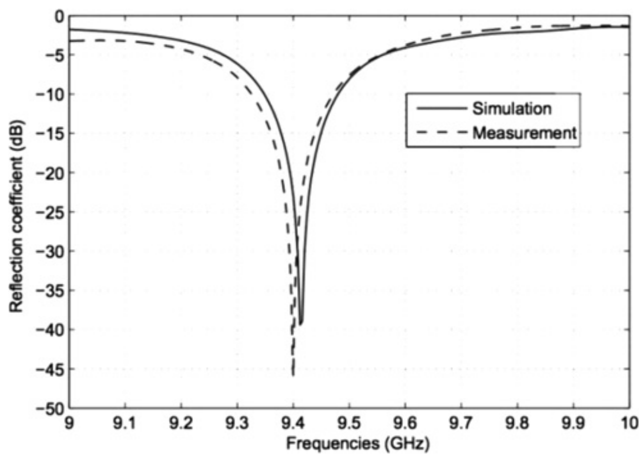


**Fig. 10** *Reflection coefficient of the slotted waveguide in simulation and in measurement*

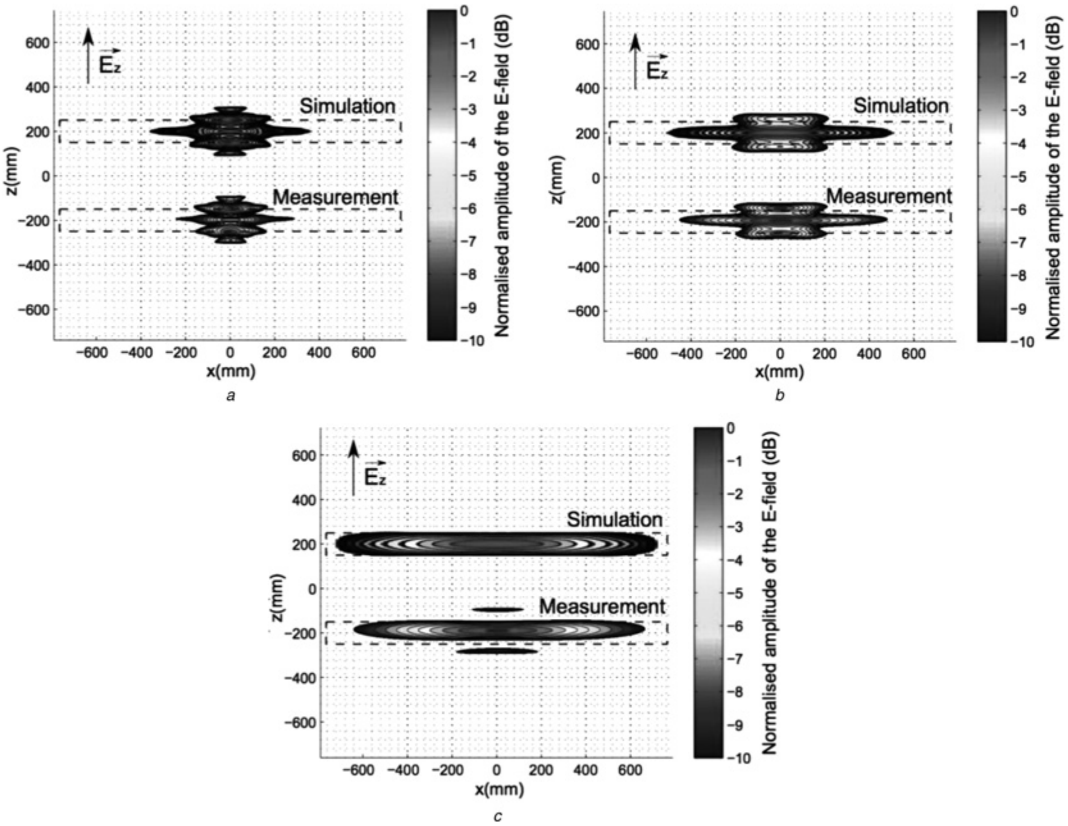


**Fig. 11** *Simulated and measured mapping of the normalised $E_z$-field*

*a* At 100 mm
*b* At 200 mm
*c* At 500 mm from the slotted waveguide at 9.41 GHz

function of the distance

$$\boldsymbol{E}_z(\boldsymbol{x}, z) = \sum_{i=1}^{N} \mathrm{e}^{-\mathrm{j}\left(k_0 \cdot \sqrt{((z-z_{\mathrm{elem}}(i))^2+x^2)}+\varphi_{\mathrm{elem}}(i)\right)}\hat{z} \qquad (4)$$

where $N$ is the number of discrete elements in the aperture, $k_0$ is the wavenumber in free space, $z_{\mathrm{elem}}(i)$ is the position of the $i$th discrete element in the aperture and $\varphi_{\mathrm{elem}}(i)$ is the phase of the $i$th discrete element in the aperture.

The previous equation considers a linear aperture along $z$-axis (for $x = 0$), with a uniform amplitude distribution. Each discrete source radiates an isotropic pattern. The variable $x$ represents the distance from the antenna aperture.

### 3.1 Measurement results

A prototype of this antenna has been manufactured, as depicted in Fig. 9. Fig. 9*b* shows the load inserted at the two ends of the waveguide. We have used a flexible absorbing material (reference ASI 10 from Siepel) held by a metallic square screwed to the waveguide. This absorbing material provides a reflection coefficient of −25 dB at 9.41 GHz.

The antenna has been measured at 9.41 GHz, and Fig. 10 shows the measured and simulated reflection coefficients of the proposed antenna. The simulation and the measurement results are in good agreement with a reflection coefficient less than −28 dB at 9.41 GHz.

Thanks to a back-propagation technique applied to far-field measurement results, the $E_z$-field of the antenna in the near-field region has been determined at 9.41 GHz for different distances. Fig. 11 shows the mapping of the $E_z$-field in simulation and in measurement at 100, 200 and 500 mm from the slotted waveguide. The dashed line represents the contour of the desired −10 dB area with a length of 1530 mm and a width of 100 mm. For clarity, the simulated and measured spots are drawn in the same graph and shifted along $z$-axis. Simulation and measurement are considered to be in good agreement, even if small discrepancies can be noticed.

First, a slight tilt of the beam can be observed in measurement results in both planes. In the $E$-plane, this tilt is equal to 2 mm at 100 mm from the slotted waveguide and 11 mm at 500 mm. In the $H$-plane, this tilt is equal to 2.4 mm at 100 mm from the slotted waveguide and 12 mm at 500 mm. This tilt can be explained by a both prototyping errors and slight misalignment of the antenna in the measurement setup.

Second, in the $H$-plane, we can notice that the −10 dB beamwidth obtained in measurement is smaller than simulation results. Indeed, at 500 mm from the slotted waveguide, the −10 dB beamwidth in the $H$-plane equals to 1307 mm in measurement and 1440 mm in simulation. This difference is partly due to the post-process which calculates the near-field mapping.

## 4 Conclusion

A near-field focused antenna providing a two-dimensional focus beam, with a focus beam in the $E$-plane and a wide beam in the $H$-plane, has been investigated in this paper. This slotted waveguide can be easily designed with the formulas given in this paper. The simulation and the measurement are in very good agreement. As shown, this near-field focused antenna can be used as a feed for a linear array with a ratio length/width equal to 15, with a focal point located at $15.7\lambda_0$ from the slotted waveguide.

## 5 Acknowledgments

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