3D PRINTED DOUBLE-RIDGED HORN ANTENNA WITH DIELECTRIC LENS FOR FREE-SPACE MATERIAL CHARACTERIZATION METHOD

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Abstract—This article deals with the design of a 3D printed wideband horn antenna with a dielectric lens for a material characterization workplace exploiting the free-space measurement method. In this work, various ridge profiles placed in the horn antenna are studied for the optimization of the impedance matching in a wide frequency range of 2–18 GHz. The workplace consists of two horn antennas with dielectric lenses and a material sample to be characterized.

Keywords—Material characterization, double-ridged horn antenna, dielectric lens, 3D printing, metal plating, measurement, permittivity

1. INTRODUCTION
The material characterization plays an important role in microwave circuit and antenna design. Since the area of material development in additive manufacturing technologies is rapidly advancing, the need of exact materials’ constitutive parameters retrieval, such as permittivity and permeability (dielectric and magnetic properties) is essential for the design process. The free-space characterization method is suitable for this purpose due to its wideband nature as the materials’ parameters are dispersive. In this work, the workplace as well as the antennas are 3D printed using the fused-decomposition modelling (FDM) technology. In the FDM, an object is created by printing individual layers, mostly from a thermoplastic filament.

2. DESIGN OF THE WIDEBAND HORN ANTENNA
The requirements for antenna design are as follows: reach the minimum value of the reflection coefficient S11, which is less than -10 dB in the frequency band of 2–18 GHz, and achieve a higher directivity in the E and H planes by an additional dielectric lens.

The core of the antenna is a rectangular waveguide of the width a calculated by (1), which assumes the higher TE30 mode (m = 3) and its cut-off frequency f_m = 18.44 GHz.

\[ a = \frac{m \cdot c}{2 \cdot f_m} \]  

For achieving a wider frequency band, a ridge is inserted in the waveguide, which is supposed to lower the frequency of the fundamental TE10 mode and to suppress the higher-order TE20 mode. The TE10 mode starts to propagate from 6.15 GHz in a ridge less waveguide. The TE20 mode has a cut-off frequency of 10.4 GHz. With the additional ridge, the TE10 cut-off frequency is moved to 1.72 GHz, the TE20 mode is suppressed and the TE30 mode has a cut-off at 19.28 GHz.

The waveguide width was firstly calculated to be 24 mm and further optimized to 24.34 mm. The height b was selected to be half of the waveguide width and optimized value reaches 18.88 mm. The ridge width was calculated to be 9 mm and the ridge gap was determined by simulations as g_r = 1.04 mm. The simulations were performed in CST Studio Suite.

For antenna excitation, a coaxial connector with an elongated dielectric, marked RF2-156-T-00-50-G was used. The conductor must be inserted in the waveguide at the place of the highest intensity of the excited mode. Further, the purpose of the resonance chamber is to reflect an electromagnetic wave from
the enclosed wall with a phase shift of 180° and guide it back to the feeding position, where it adds to the wave leading to the antenna aperture.

The horn transforms the waveguide impedance of 50 Ω to the impedance of free-space, which is 377 Ω. This is realized by a gradual opening of the waveguide. The horn dimensions were calculated by the corresponding equations listed in the literature [1]. To achieve compactness of the antenna, a gain of G = 18.4 dBi was chosen for the center frequency f = 10 GHz. The horn width was calculated to be W = 122 mm and optimized to 104 mm, the horn height was calculated to be H = 82.8 mm and optimized to 79 mm, and the length was calculated as L₀ = 114 mm. The simulation model is shown in the Fig. 1.

![Figure 1](image)

**Figure 1:** The simulation model of designed double-ridged horn antenna.

The ridge in the horn is an elongation of the ridge from the waveguide, however its shape can be defined by different functions which have a major impact on the final qualities of the antenna. In this work, 4 profiles were studied: quadratic, sinusoidal, exponential and Gaussian (shown in Fig. 2). Quadratic and sinusoidal profiles are similar to each other by the shape and qualities. They both have a worse reflection coefficient S₁₁ near frequency of 2 GHz, but they exhibit a higher gain. Further details of this analysis can be found in [2]. The exponential profile led to the best antenna performance. It is calculated by (2), where Z₀ is g₀/2 and Z₀ is H/2:

\[
f(z) = \frac{g_r}{z} e^{\frac{1}{z_h^2 - (W-a)^2}} \ln\left(\frac{Z_L}{Z_0}\right) z.
\]  \hspace{1cm} (2)

![Figure 2](image)

**Figure 2:** Ridges cutouts and corresponding reflection coefficients.

3. LENS DESIGN

For achieving a higher gain and a better focusing onto the characterized sample, two types of lenses were considered. A quasi-optical lens and a hyperbolical lens. Based on the power flow simulation, the quasi-optical lens was selected for final design because of its ability to focus the power better in the area of the sample placement. Based on this analysis, an appropriate position for the placement of the measured sample is in the distance of 115 mm from the lens aperture. In contrary, the hyperbolic lens
transforms the electromagnetic waves more into a planar wave, instead of one focal point. The lens profile is calculated by (3) [3] given that \( \lambda_0 = 0.03 \) mm at 10 GHz, tapering edge \( TE = 20 \) dB and the relative permittivity of the lens material equal to 2.6 for PLA at 10 GHz [2]. The resulting quasi-optical lens diameter and thickness are 123.2 mm and 28.6 mm, respectively. The lenses profiles and power flow simulation are shown in Fig. 3.

The quasi-optical lens is expressed in XZY Cartesian coordinate system as:

\[
x^2 + y^2 = (\varepsilon_r - 1)z^2 + 2f(\sqrt{\varepsilon_r} - 1)z.
\]  

(3)

The hyperbolic lens equation (4) is given as follows:

\[
y = -\left(\frac{\sqrt{\varepsilon_r(R^2+f^2)}-f}{\varepsilon_r-1}\right)^2 - \left(\frac{\sqrt{\varepsilon_r(R^2+f^2)}-f}{\varepsilon_r-1}\right)^2,
\]  

where \( R \) is the lens radius, and \( f \) represents focal distance.

![Figure 3: Lens cutout from the power flow simulation: (a) quasi-optical lens (b) hyperbolic lens.](image)

4. LENS ANTENNA FABRICATION AND MEASUREMENT

The double-ridged horn antenna was 3D printed using X material and X FDM 3D printer, the lens was 3D printed using the PLA material and the Prusa i3 MK3S+ FDM printer. The antenna radiation patterns, and gain were measured in an anechoic chamber. The measured gain of the antenna #1 was close to the simulated values. The maximum gain achieved by antenna #1 without the lens was 15.28 dBi at 11.6 GHz and 22.8 dBi with the lens at 13.8 GHz, respectively. The addition of the lens worsens the reflection coefficient due to reflection from the lens surface by 3 dB. However, the gain was improved by 6 dB at the center frequency of 10 GHz.

![Figure 4: (a) Simulated and measured reflection coefficient of two fabricated antennas, (b) normalized simulated and measured gain with and without lens in the E and H planes at 10 GHz.](image)
CONCLUSION

Based on a theoretical design, the double-ridged horn antenna with the quasi-optical lens was modelled and optimized. The antenna model was successfully designed in accordance with the requirements, achieving reflection coefficient close to -15 dB. The greatest impact on the change of antenna parameters have the position of the coaxial feed, the gap between ridges and the ridge profile. The manufactured antenna was plated with a copper foil, however, this method has a drawback in the insufficient copper foil bearing. A solution to this problem can be in metallization by a vacuum plating. The used FDM 3D printing method shown to be an inaccurate by means of manufacturing tolerance, e.g. the inner width of the waveguide should be 24.23 mm but after printing it was 23.8 mm, including the plated layer. The 3D printed antenna was adequately modified to comply with the requirements, especially the critical spacing between ridges, where the initial gap was 0.2 mm. By the influence of the printing imperfections and the insufficiency of the copper foil, the measured deviation of the reflection coefficient was significant between the frequency of 12 GHz and 18 GHz. Overall, the differences between the simulation and the measurement of manufactured antennas are caused mainly by the imperfection of the manufacturing process. The inaccuracies are more pronounced at higher frequencies. Subsequently, the functionality of the workplace for the verification of its function on a sample with known dielectric properties will be presented.

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REFERENCES

