Design of Square Waveguide Corner

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***Abstract*—The paper presents a full-wave design for matching a square mitered waveguide corner. We compare the full-wave numerical simulations with experimental results. The corner can operate with both polarizations, i.e. TE_{10} for magnetic plane bend and TE_{01} for electric plane bend. That is very useful for intended meteorological radar. That could use dual polarization transmit and receive both horizontal and vertical polarizations to assess additional properties of the weather targets.**

Keywords—corner design; design optimization; dual-mode waveguide; full-wave design; meteorological radar; square waveguide; waveguide corner

I. INTRODUCTION

The electric field vector defines the direction of polarization of electromagnetic waves. Generally, the polarization is elliptical with special cases of circular and linear polarizations. If the radar is perfectly circularly polarized and the clutter is caused by spherical raindrops, the clutter cancellation will be perfect, enabling to detect the presence of an otherwise-hidden composite target, which reflects both right-hand and left-hand circular polarizations. When the radar antenna is elliptically polarized, the cancellation of an echo from symmetrical targets will not be complete; also, if the antenna is circularly polarized and the target is not symmetrical, cancellation will not be complete [1]-[6]. However, we could use slightly elliptical polarization to substantially reduce echoes from raindrops.

Various produced radars have been described [7], where variable elliptical polarization is created by rotating dielectric or metal parallel plates in front of horn, by the rotation of dielectric quarter wave plate in the circular waveguide or by the rotation of rectangular waveguide feeding the circular and subsequent square waveguide. Certain radars contain optional separate weather channel. The square waveguide elements were manufactured using casting forms [8].

In the very motivating recent paper [9], the writers suggested broadband dual-mode waveguide bends for linear and circular polarizations. It is very inspiring to obtain the certainly broadband corners or bends using totally new ideas for design.

Moreover, the effort to obtain the element without any phase difference in broad band is very hopeful for foundation of different elliptical polarizations, which can be used for numerous microwave systems. However, bearing in mind application point of view, various problems could be considered.

The polarization characteristics are regulated by the system requests [1], [2]. These could be verified using nomographs [3], which gives both the constraints for good circularity, and the sensitivity of the different parameters. The nomograph has been utilized widely in the measurement of electromagnetic-wave propagation in free space and in nonstandard waveguides together with the radiation characteristics of several antennas. It may be easily utilized to explain abundant problems including those connected with microwave antennas, waveguide components, and tolerance issues. Indisputably, it is not enough to design one exceptional component (such as waveguide bend or corner) as the polarization characteristic are created by the complete sum of phase differences of all components comprising antenna feed, which form the transmission line used for transmission and/or reception of orthogonal modes. Bearing in mind the waveguide with the length of l and necessary tolerances, i.e. possible sizes a_1 and a_2, the variation of phase difference δ between orthogonal polarizations is

$$\delta = l\left[\sqrt{\left(\frac{2\pi f}{c}\right)^2 - \left(\frac{\pi}{a_2}\right)^2} - \sqrt{\left(\frac{2\pi f}{c}\right)^2 - \left(\frac{\pi}{a_1}\right)^2}\right] \tag{1}$$

where c is free space velocity of light and f is frequency. That clearly validates that the importance of a special component could be diminished but the system requirements could be bigger.

Clearly, there is still a demand for waveguides in numerous circumstances, comprising high-power systems, millimeter wave systems or satellites. Consequently the design of septum with sharp wedges cannot be usually used due to likely discharges made by high electric fields.

Meteorological radars differ from other kinds of radars in the character of weather targets. The resulting features of the radar signal, and signal processing, are generations of important weather information. Important meteorological targets contain an extensive span of scattering echo intensities. They are dispensed from short to long ranges, close to ground to higher troposphere, where weather is substantial, and typically contain millions of three-dimensional cells detected by the radar. Additionally, it is essential to measure quantitatively the

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received signal properties of the cells for parameter estimations such as precipitation rate, precipitation type, air movement, turbulence, and wind characteristics. As most of radar resolution cells comprise useful information, meteorological radars ask for fast digital signal processors. That means the meteorological radars should extract a huge amount of data for various users. Meteorological radars could use dual polarization transmit and receive both horizontal and vertical polarizations to assess additional properties of the weather targets. Transmitting the two orthogonal polarizations allows estimation of differential quantities between echoes from two polarizations. We could obtain enhanced rainfall rates, as well as other physical information, as functions of these measurements that associate the differences in the horizontally and vertically polarized signals.

The paper deals with full-wave design of square mitered waveguide corner, which can operate with both polarizations. Fig. 1 shows the waveguide corner. We compare the full-wave design with experimental results. The Institute of Atmospheric Physics, Prague, is developing a 35 GHz Frequency-Modulated Continuous-Wave (FMCW) cloud radar [10], where the square corner will be implemented. The simulations in CST Microwave Studio were done.

II. DESIGN OF SQUARE WAVEGUIDE CORNER

Usually, a waveguide corner is done by using a mitered corner, which provides a diagonally oriented reflector surface for the propagation direction change [11]. We can use for waveguide discontinuities and their equivalent circuits the classic reference [12].

If we utilize a square waveguide, a mitered corner gives a good match for one mode but the same corner would provide poor performance for the other mode as Fig. 2 demonstrates. However, we could create the other mitered corner with different dimensions that gives a good match [13]. That means we should create two various effective reflecting surfaces simultaneously. The mitered corner has one effective reflecting surface for the E-plane mode and the different effective reflecting surface for the H-plane mode.

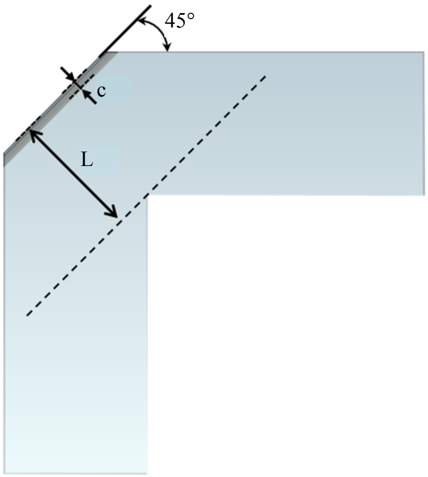


Fig. 1. Waveguide mitered corner with ridge - reflecting surface with ridge.

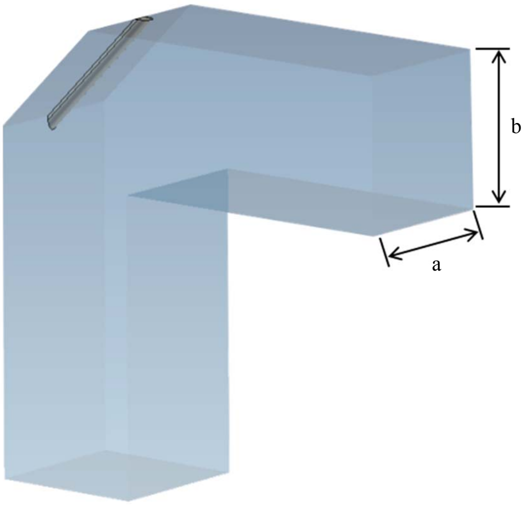
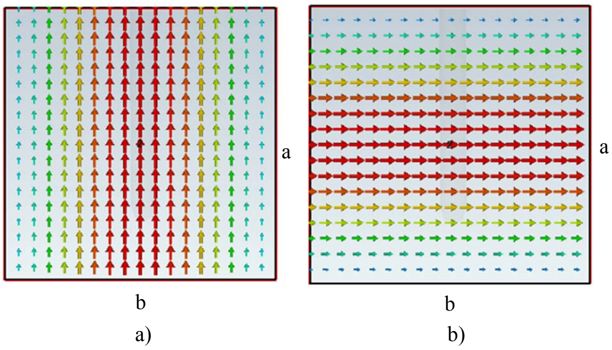


Fig. 2. Square waveguide mitered corner with ridge *a*=*b,* c.

The reflecting surface has a ridge placed in the middle (i.e. *a/*2), which is parallel to the electric field vectors for E-plane bend, so that the ridge top behaves approximately as the effective reflector for that mode.

We could roughly consider that operational frequencies are below the cut-off frequency in the region between the effective reflector and the back reflector. We utilized the “cut and try” method in 1970. That was very time-consuming. In the beginning, we used the proper wire and finally we employed the ridge. However, today approach is much faster and easier using full wave design. It could be noted that the same idea for matching of dual mode square waveguide corner has been proposed using a reflector with at least one, and rather several, ridges or a several conductive wires parallel to one another [14], [15]. To increase a bandwidth, a modification with successful experimental results has been published [16].

We present the full-wave design method for matching a square waveguide corner for an operation with both polarizations, i.e. TE_{10} for H-plane (magnetic plane) bend and TE_{01} for E-plane (electric plane) bend as is shown in Fig. 3 (TE mean transversely electric wave with mode 01). To validate the simulations, we used a model designated for frequencies about 9.1 GHz.



a) b)

Fig. 3. a) Input TE_{10} for H-plane bend, and b) Input TE_{01} for E-plane.

The chosen frequencies diminish the tolerances problems, which could be expected for substantially higher frequencies of the intended radar.

Fig. 4 shows S_{11} parameters. According to numerical simulations with $a=b=20.6$ mm and $l=18.9$ mm we could optimize the design to obtain the minimal S_{11} for TE_{01} with ridge height of $c=1.49$ mm. The solid line depicts the TE_{10} (H-plane bend) and dotted line shows the TE_{01} modes (E-plane bend). Measurements of TE_{10} are shown by crosses and TE_{01} by circles. Fig. 4a shows results of the corner without ridge and Fig. 4b with the optimized ridge. Obviously, the ridge effect is substantial for E-plane bend (TE_{01}). Clearly, the ridge affects TE_{10} for H-plane bend but this effect is much lower. That demonstrates above statements, which were gained experimentally.

We could compare numerical simulations with measurements of the model. However, that is rather complicated [17] as total reflection coefficient arises from several small reflecting obstacles. If the electric field hit the obstacles, the total reflected wave creates the complete total reflection coefficient. When the incident wave hits the first connection, a partial reflected wave is produced. A transmitted wave strikes on the second junction and a portion of this is reflected to give a wave incident from the right on the first junction. Similarly, the next junctions create reflections. A portion is transmitted, and a portion is reflected back.

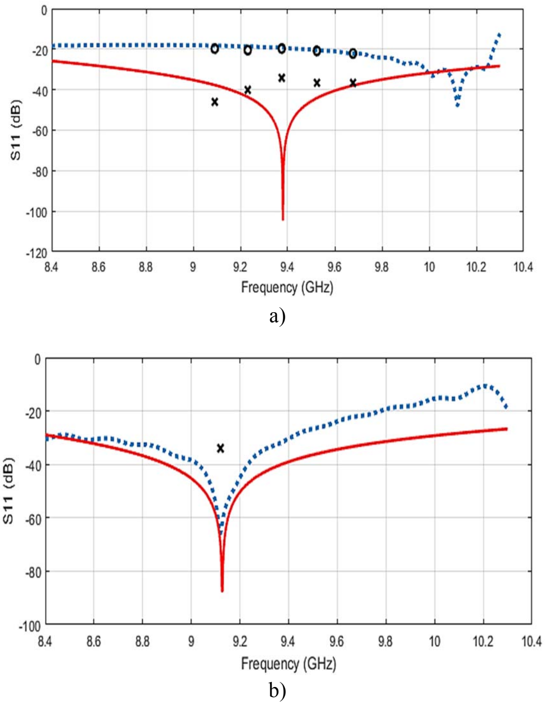


Fig. 4. Numerical simulations of S_{11} parameters for input TE_{10} for H-plane (solid line) and TE_{01} for E-plane bend (dotted line). Measurements of TE_{10} are shown by crosses and TE_{01} by circles. a) without a ridge and b) with ridge.

For small reflections, the resultant reflection coefficient could be obtained by taking only first-order reflections into account. In fact, we have to analyze the effect of several transitions such as rectangular-to-square waveguide and/or coax-to-rectangular transitions, which we use for the measurements of waveguide corner. Even if the reflection coefficients of individual components could be very low, we cannot expect that the accuracy of resulting coefficient S_{11} will be better than –40 dB.

Fig. 5 shows the phase differences between linear components. The solid line depicts numerical simulations and the dotted line shows measurement. It could be noted the calculated phase differences are changing insignificantly, if the dimensions are slightly distorted.

To show the tolerance issues, we present the numerical simulations of S_{11} parameters for input TE_{10} (for H-plane - solid line) and TE_{01} (for E-plane bend-dotted line), when the three dimensions are smaller. We chose $a=b=20.5$ mm and $c=1.39$ mm. Comparing Fig. 4b with Fig. 6 we can see that minima of S_{11} parameters are shifted. Clearly, the effect of tolerances is much bigger for E-plane than for H-plane bend. However, we can see that the reflection coefficients for input H-plane (solid line) are substantially smaller than –20 dB for the frequency band from 8.4 to 10.3 GHz and for input E-plane bend (dotted line) are smaller than –20 dB for the frequency band from 8.4 to 9.7 GHz.

Generally, the elliptical polarization of any antenna will be changed, if the phase differences between orthogonal linear components are changed. During the development of components and the whole system, both the amplitude and phase of the linear components are required. An amplitude-only measurement setup can measure the axial ratio and tilt angle of an elliptically polarized wave. The equipment measures the amplitudes directly. The nomograph [3] shows the relationships between the variables and gives both the requirements for good circularity, and the sensitivity of the various parameters. It has been extensively employed in the measurement of electromagnetic-wave transmission in free space and in nonstandard waveguides, and the radiation properties of various antennas. It may be used to solve numerous problems, including those associated with microwave antennas, waveguide devices, and tolerances.

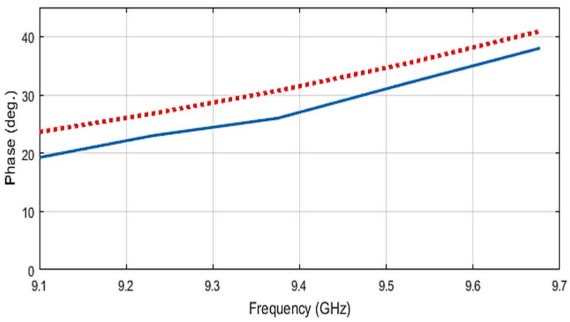


Fig. 5. Phase difference between linear components. Numerical simulations (solid line) and measurement (dotted line).

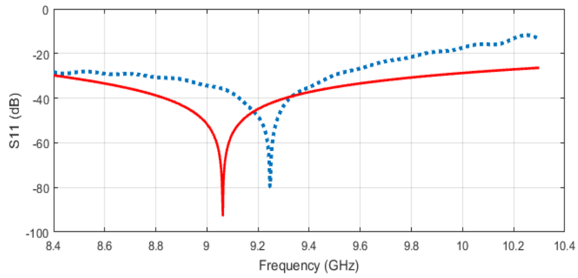


Fig. 6. Tolerance demonstrations of numerical simulations of S_{11} parameters for input TE_{10} (for H-plane - solid line) and TE_{01} (for E-plane bend - dotted line).

To clarify behavior of above stated features, Fig. 7 and Fig. 8 depicts the numerical simulations of E-fields for simulation of 9.126 GHz for E-plane bend (TE_{01}) and for H-plane bend (TE_{10}).

To consider practical design, it is obvious that pure aluminum certainly forms a narrow surface layer of aluminum oxide, which makes a physical barrier to corrosion or further oxidation in extraordinary environments. But some aluminum alloys do not create the oxide well, and consequently are not secured against corrosion. To reduce ageing, it is frequently compulsory passivate the element surfaces. Nevertheless, the passivation generates insulating layers, which change propagations of waveguide modes, and the prohibited leakage would be generated. Thus, the conducting connection of all waveguide corner parts should be preserved.

III. CONCLUSION

The paper deals with full-wave design of square waveguide mitered corner, which can be used for both polarizations, as is shown in Fig. 3. We compare numerical simulations for several cases (such as corner without and with the ridge, and changes of corner dimensions) with experimental results.

Obviously, the designed waveguide corner does not meet the requirement that the phase differences between linear components should be zero. Actually, the differences are changing for the other components as well. Therefore, we should apply the nonstandard waveguide with two transitions between various waveguide cross sections to establish the zero phase differences.

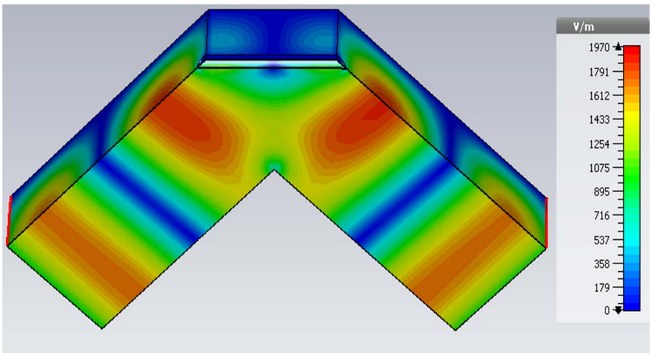


Fig. 7. E-field simulation for 9.126 GHz for TE_{01}.

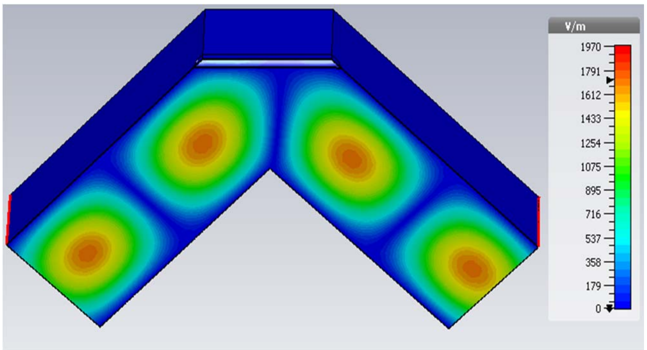


Fig. 8. E-field simulation for 9.126 GHz for TE_{10}.

Fig. 7 demonstrates that the assumption about the operational frequencies, which are below the cut-off frequency in the region between reflector and the back reflector, is roughly fulfilled (i.e. electric field is extremely attenuated). However, the ridge creates the discontinuity, and therefore the S_{11} parameter minima of input TE_{10} for H-plane (solid lines) are shifted as can be seen in Fig. 4. We can see the phase differences are created, if we compare Fig. 7 and 8, which depict the same time.

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