

Metamaterial Ridged Waveguides with Wavelength Control for Array Antenna Applications

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1. Introduction

Till now, phased array antennas have only been used in high cost applications such as passenger airplanes or satellites. To break through this barrier we proposed a new type of phased array antenna for 76.5GHz consumer automotive radar [1][2]. The proposed phased array antenna has mechanical phase shifters that use magnetic walls with no metal contacts enabling the realization of a low cost and stable phased array antennas that can operate over a wide temperature range. The use of waffle-iron ridged waveguides makes it easy to build multi-layered structures. However, the proposed antenna has grating lobes at +60deg in the elevation plane since the wavelength in the waveguide is longer than that in free space. Employing a dielectric rod was not sufficient to suppress the grating lobes, which remained at the -20dB level to the main lobe. In this paper, we propose a method to control the wavelength in the waveguide and to reduce it to less than it is in free space in order to suppress the grating lobes.

2. Configuration of the Metamaterial Ridged Waveguide

Fig. 1 shows the configuration of the proposed waveguide which is based upon the waffle-iron ridged waveguide described in our previous papers [1][2]. In this paper, the "waffle-iron ridged waveguide" is renamed the "metamaterial ridged waveguide (MRW)" since the side walls are made of a metamaterial. It is known that the wavelength in the waveguide increases when TEM waves are reflected at both side walls [3]. Here we assume that the same phenomenon occurs in a MRW whose side walls are the equivalent magnetic boundary (EMB) formed by the metamaterial. This means the only difference between an ordinary ridged waveguide and the MRW is the characteristics of the side walls, one having an electric boundary and the other an EMB. The EMB, however, vanishes from the stop band, and similar dispersion curves for the waveguides were described in our previous paper [1]. In other words, the MRW can be modelled as a distributed transmission line, the same as other waveguides.

With the assumption that the MRW can be modelled as a distributed transmission line, we propose a method to shorten the wavelength in the MRW by adding a periodic step onto the ridge as shown in Fig. 1. Periodic steps with depth d give additional distributed elements to shorten the wavelength. Fig. 2 shows an equivalent circuit of the MRW, where ΔL and ΔC are the additional inductance and capacitance respectively. The wavelength without periodic steps λ_{g0} and the one with periodic steps λ_g are related by

$$\frac{\lambda_g}{\lambda_{g0}} = \sqrt{\frac{LC}{(L + \Delta L)(C + \Delta C)}} \quad (1)$$

Equation 1 means the wavelength in the MRW can be shortened by changing either ΔL , ΔC or both.

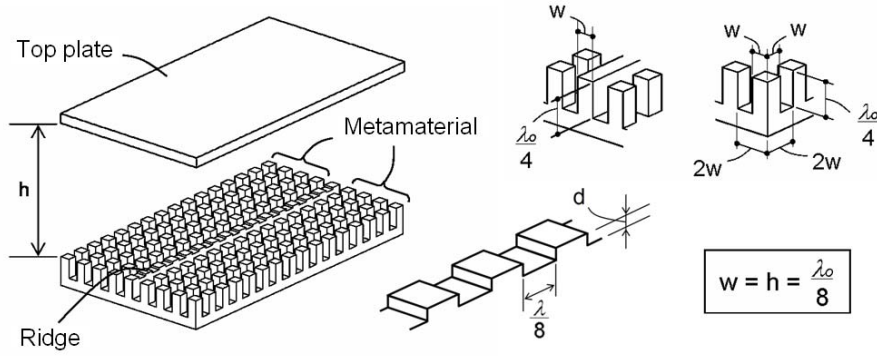


Fig. 1 Newly proposed MRW

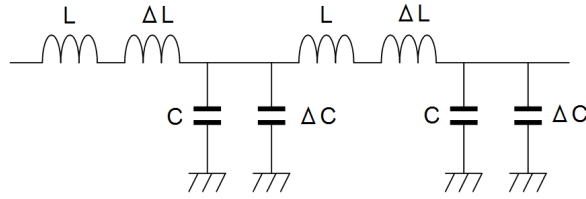


Fig. 2 Equivalent circuit of the MRW

3. Analysis of the MRW

Prior to calculating the wavelength, we provide confirmation of above assumption that the MRW can be modelled as a distributed transmission line. As shown in Fig. 3, the transverse resonating method can be applied to demonstrate that the same system operates in the MRW as in an ordinary waveguide. In Fig. 3 λ , λ_t and λ_g are the wavelength of the TEM wave in free space, the wavelength in the transverse direction and the wavelength in the waveguide direction, respectively. Additionally in Fig. 3, θ is the angle of direction of the TEM wave in the waveguide and ϕ is the phase of the round trip if the TEM wave is input in the transverse direction from the centre of the ridge.

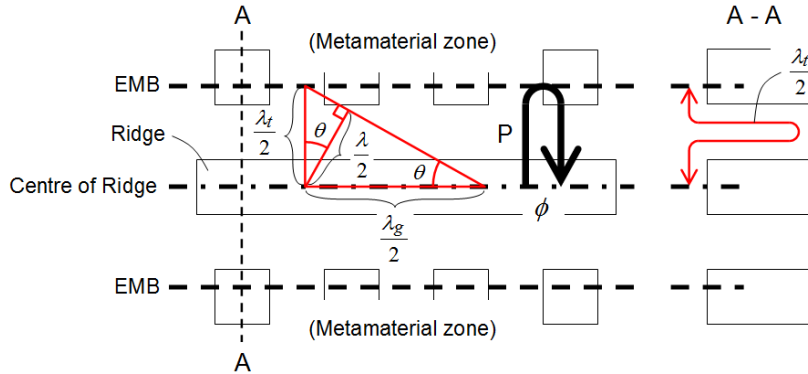


Fig. 3 Geometric relationships in the wavelength in the transverse resonating method

From Fig. 3, the relationship between λ , λ_t and λ_g is

$$\frac{1}{\lambda^2} = \frac{1}{\lambda_t^2} + \frac{1}{\lambda_g^2} \quad (2)$$

Using the propagation coefficient β , ϕ can be described as

$$\phi = -2\beta \frac{\lambda_t}{2} = -2 \frac{2\pi}{\lambda} \frac{\lambda_t}{2} = -2\pi \frac{\lambda_t}{\lambda} \quad (3)$$

Substituting λ_t from equation 3 into equation 2, λ_g is obtained.

$$\lambda_g = \lambda \sqrt{\frac{1}{1 - (2\pi/\phi)^2}} \quad (4)$$

Calculating ϕ analytically is beyond the scope of this paper. Instead, we used 3D EM simulation to find $\phi = 53\text{deg}$. Since $360\text{deg} < \phi < 720\text{deg}$ ($\lambda/2 < \lambda_t < \lambda$) from Fig. 3, we set $\phi = 667\text{deg}$, and using equation 4, $\lambda_g = 1.19\lambda$. This differs by only 3.5% from $\lambda_g = 1.15\lambda$ given in our previous paper for the wavelength in two waveguides of different lengths calculated by 3D EM simulation [2]. This result confirms that the above assumption is reasonable.

After confirming the assumption, the wavelength in the MRW was calculated. Fig. 4 shows the relationship between the depth of the steps d and the wavelength calculated by 3D EM simulation, where the wavelength is determined from the phase difference between the S21 parameter in two different length waveguides. As shown in Fig. 4, the wavelength in the MRW decreases as the depth of the steps increases, which indicates that the steps generate additional inductance and capacitance. In other words those additional elements can be understood as follows. The additional inductance is generated because of the higher characteristic impedance of the line in the stepped area, while the additional capacitance is generated with the electric field normal to the walls of the steps. This is demonstrated by the relationship between the normalised wavelength, λ_g/λ , and the depth of the steps, d , shown in Fig. 4, in which λ_g/λ decreases with d . As shown in Fig. 4, it is possible to shorten the wavelength to less than that in free space, which means the grating lobe of the resonating feed line of the array antenna would be suppressed.

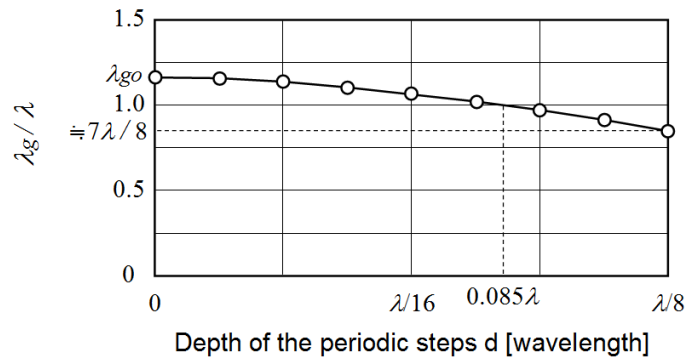


Fig. 4 Relationship between the wavelength in the MRW and the depth of the periodic steps

4. Application to Array Antennas

A problem with array antennas is that the longer the wavelength on the feed line, the higher the level of the grating lobes. In particular, for resonating feed lines reported in the literature [1][2], the distance between the radiation elements is equal to the wavelength of the feed lines, thereby grating lobes are generated at the same level as the main lobe. Although employing dielectric rods on the elements is a good way of suppressing the grating lobes, these are still too high to be employed in actual radar systems. Because of this difficulty we used our proposed structure to reduce the grating lobes.

Fig. 5 shows the arrangements of the radiation elements on the resonating feed line for the different depths of steps, d . As shown in Fig. 5, the distance between the radiation elements decreases with the depth of the steps.

Fig. 6 shows an example of the directivity calculated for the array antenna in Fig. 5 in which the array number is 10 and the element factor is the half wave dipole characteristic assuming the effect of dielectric rods. As shown in Fig. 6, the dielectric rods are insufficient for suppressing the level of the grating lobes, but the suppression increases with the depth of the steps and vanishes at depth $d = \lambda/8$.

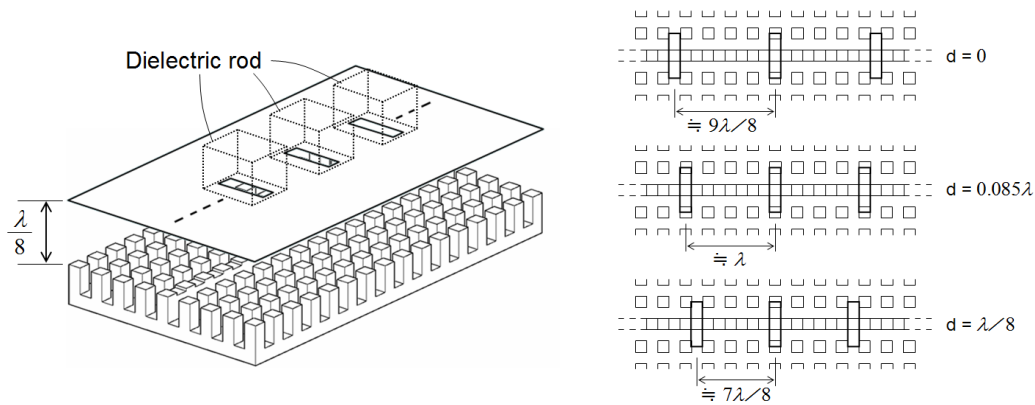


Fig. 5 Arrangement of the radiation elements on the resonating feed line for different depth d

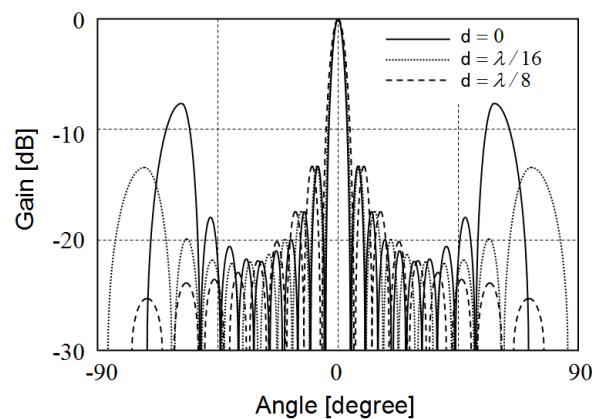


Fig. 6 Directivity of the array antenna for various depth d (calculated)

5. Summary

A new structure for suppressing the grating lobes of MRW array antennas is proposed. After confirming that the MRW could be modelled as a distributed transmission line, it was shown that periodic steps on the ridge of the MRW decrease the wavelength. The distance between the radiation elements in the resonating feed line, which is equal to the wavelength, also decreases, contributing to suppression of the grating lobes. The MRW with wavelength control is expected to be valuable not only in decreasing the size of array antennas but also that of many other microwave and millimetre-wave circuits.

References

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