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A WIDEBAND SLOTTED WAVEGUIDE ANTENNA ARRAY FOR SAR SYSTEMS

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Abstract—Possibilities for the extension of the operational frequency band of slotted waveguide antennas are studied. It is shown that by using both conventional longitudinal slots and subarraying techniques it is possible to reach the relative bandwidth of about 15%. This result is illustrated by the development of a novel slotted waveguide antenna for high-resolution SAR applications. The antenna operates in the X-band and forms the beam of $4^\circ \times 6^\circ$ with the gain of about 30 dB.

1. INTRODUCTION

The extension of the operational bandwidth of antennas is stimulated by the development and applications of synthetic aperture radars (SAR) and other radar systems. Potential radar applications call for the bandwidth of about 10% and higher, to achieve, for example, the spatial range resolution of about 20 cm with X-band radars.

The indicated value of the relative bandwidth is easily achieved by using reflector or horn antennas. However, both large volume and weight of these antennas essentially complicate their application in airborne and spaceborne systems. Slotted waveguide antennas due to their planar form, strength of the construction, and high efficiency are more attractive candidates for applications in such systems. But the typical bandwidth of practical slotted waveguide antennas is usually of about few percents.

Recent studies [1–5] have, however, demonstrated that there are promising approaches to the extension of the bandwidth of conventional resonant slotted waveguide arrays. In this paper, we study these and some other approaches to the extension of the

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bandwidth of such type of antennas. We apply these results to the development of a novel X-band slotted waveguide antenna with the relative bandwidth of about 15%. The proposed antenna shows also attractive radiation and impedance matching properties, which are discussed as well.

2. ANTENNA BANDWIDTH ENHANCEMENT

Let us consider the frequency limitations of a resonant planar slotted array of longitudinal slots made in the broad wall of rectangular waveguides. It is well known that the bandwidth of such arrays is determined by both the resonant properties of individual slots and the bandwidth of the waveguide structure. Also, it is well known that the frequency bandwidth of slots can be extended by reducing the waveguide wall thickness and/or increasing the slot width. The bandwidth can be additionally enhanced by using waveguides with reduced cross-section and/or cutting slots at a maximum possible distance from the waveguide centerline. The application of slots of special forms, e.g., dumbbell slots [2], is another way to enhance the slotted waveguide antenna bandwidth. But such slots are hard-to-manufacture and they provide a high level of cross-polarization components.

It should be also reminded that conventional longitudinal slots themselves demonstrate a rather large bandwidth under a proper choice of their parameters. Our simulations have shown that the bandwidth of about 20% can be easily achieved. For example, such bandwidth is realized by using longitudinal radiating slots with the width of 2.5 mm made in the broad wall of the X-band rectangular waveguide with the reduced cross-section of 20.5×7 mm and the wall thickness of 0.5 mm.

In order to increase the bandwidth of a slotted waveguide antenna, wideband waveguides are used and/or the antenna array is divided into short sections (subarrays) with individual feeding. The usage of broad rectangular waveguides is the simplest way to extend the frequency bandwidth. However, such waveguides are only useful in the case of linear arrays. For planar arrays, this solution results in the appearance of intensive grating lobes due to a large interelement spacing. Evidently, the interelement spacing can be reduced by using dielectrically filled waveguides, but this approach is not too practical. Actually, the same factor limits the application of wideband ridged waveguides [3, 5]. The subarraying technique is considered now as the most effective and preferable approach to the extension of the antenna bandwidth.

As for the selection of the number of slots in a subarray, several factors should be accounted. Generally, the less the number of slots the wider the array bandwidth can be achieved. But if rather short subarrays are selected, their number is raised and a complex feeding system is required. Moreover, the efficiency of short arrays of one or two slots is typically too low. So the optimal number of slots in subarrays should be determined for each particular case. This problem does not have a universal solution, and various approaches have been proposed. For example, the maximum number of slots can be estimated from the analysis of the VSWR versus the number of slots. An approximate expression for such dependence has been proposed in [6]:

$$\begin{aligned} VSWR &= 1+\frac{2}{\alpha^2}+\frac{2}{\alpha}\left(1+\frac{1}{\alpha^2}\right)^{\frac{1}{2}} \\ \alpha &= \frac{1+\frac{(\pi NB)^2}{3\times10^4}}{\frac{\pi NB}{300}\left(1+\frac{(\pi NB)^2}{2\times10^4}\right)} \end{aligned} \tag{1}$$

where N is the number of slots in a subarray, and B is the frequency bandwidth in percents. Thus specifying an acceptable VSWR and a required B-value, it is possible to estimate the maximum number of slots.

The above approach does not account for the waveguide dispersion and the aperture distribution degradation. To take into consideration these effects, we propose to use the following idea. At the resonant frequency, slots in a resonant array are placed in the maxima of the standing wave. If the frequency is varying, the maxima of the standing wave are shifted. It is possible to determine analytically the frequency band Δf, in which the amplitude and the phase variations of the wave on the slots are within some limits. Provided that the relative amplitude variation is limited to the value less than $1/\sqrt{2}$ with respect to the maximum of the standing wave, and the phase error does not exceed 45°, one can find:

$$\Delta f=\frac{c}{2a}\cdot\left[\left(1+\left(\frac{N-\frac{1}{4}}{N-\frac{1}{2}}\cdot\frac{a}{d}\right)\right)^{\frac{1}{2}}-\left(1+\left(\frac{N-\frac{3}{4}}{N-\frac{1}{2}}\cdot\frac{a}{d}\right)\right)^{\frac{1}{2}}\right] \tag{2}$$

where a is the waveguide width, $d=\frac{\lambda_g}{2}$ is the slot spacing (λ_g is the guide wavelength at the central frequency), and N is the number of the slots.

An example of the dependence of Δf versus the number of slots and the waveguide width as determined by Eq. (2) is shown in Fig. 1.

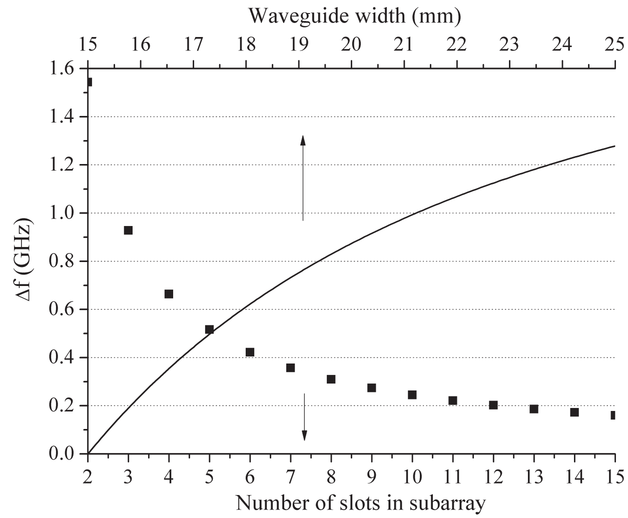

Figure 1. Bandwidth of subarray versus the number of slots (scatters; $a = 20.5$) and the waveguide width (solid line; $N = 3$).

The bandwidth of subarray is increased with decreasing the number of slots and with increasing the waveguide width. If the waveguide width is close to the cut-off value, then the waveguide dispersion rises and the subarray bandwidth Δf goes to zero.

If the antenna operational frequency is outside the indicated bandwidth (2), the aperture distribution is essentially modified. Evidently, in this case the field amplitude on the slots distant from the short-circuit is notably less than that on the nearest slots because of the shift of the standing wave maxima. Such distortion in the aperture distribution results in reducing the subarray efficiency, beam broadening, and partial null-filling.

Both, end- and center-fed subarrays can be used. The formula of (2) is valid for the end-feeding. The center-fed subarrays can be approximately considered as two coupled end-fed subarrays. In this case, the maximum number of slots is increased by factor two without reducing the bandwidth.

Our analysis of the expressions (1) and (2) shows that the maximal bandwidth of conventional end-fed slot arrays with three and more radiating slots does not typically exceed 15%. It means that the bandwidth of subarray is a critical limiting factor for the overall antenna bandwidth rather than the frequency performance of individual slot radiators.

3. ANTENNA SYNTHESIS

Our design goal was to develop an X-band slotted waveguide antenna with a $4^\circ \times 6^\circ$ antenna beam and the relative operational bandwidth of about 15%. A rectangular array of radiating slots has been selected. The rectangular aperture is separable one [7], so it is possible to use linear aperture distributions to form independently the radiation pattern with different values of the beam width in the antenna principal planes. The 20 dB Chebyshev amplitude distribution for linear arrays has been used for this purpose. In order to realize such distribution in the antenna aperture, the antenna is designed in the manner shown in Fig. 2, where two building blocks: Left and Right, from total 16 are shown. There are two radiating subarrays in each block.

The wideband slots and the waveguide described in Section 2 have been used in the antenna design. Assuming the 15% bandwidth, we find from (2) that the maximum value of the slots number in the subarrays is six and three for center-fed and end-fed subarrays, respectively. In this case, the maximum VSWR, as predicted by (1), is 2 for the end-fed arrays.

The antenna shown in Fig. 2 contains two main layers: Radiating and feeding ones. The center-fed design is used for radiating layer subarrays and end-fed design is used for feeding layer subarrays. The radiating layer consists of 32 slotted subarrays with longitudinal slots (16 in each half), and the number of slots was selected to be equal six per subarray. The feeding of the radiating subarrays is organized by means of crossed feeding subarrays, which are coupled with radiating ones via inclined slots. The number of slots was selected to be equal two per feeding subarray. Due to the subarraing used, the feeding layer has 16 individual inputs (8 in each half), and a feeding network is assumed.

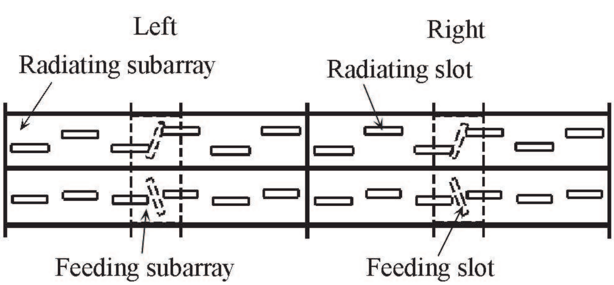

Figure 2. Two building blocks (Left and Right) of 16 of the slotted waveguide antenna.

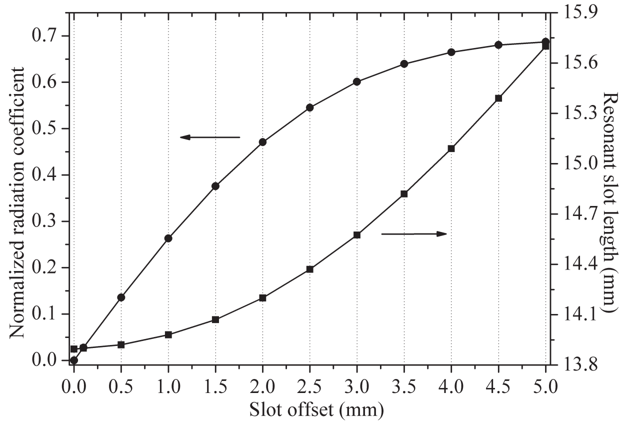

Figure 3. Normalized radiation coefficient and the resonant slot length versus the slot offset from the waveguide centerline. The waveguide cross section is 20.5×7 mm, the slot width is 3 mm, and the waveguide upper wall thickness is 0.5 mm.

The method of moments [7, 8] has been applied to evaluate characteristics of the individual slot in the radiating waveguide used. In particular, we have determined the dependences of the normalized radiation coefficient and the resonant slot length versus the slot offset from the waveguide centerline, which are shown in Fig. 3.

These dependences have been used for the antenna synthesis by using the energy method [9] in the way described in [10–12]. In accordance with this method, the mutual coupling of the radiating slots through the internal and external spaces is ignored. It is a valid assumption for resonant slotted arrays [9]. The phase shift between neighboring slots is determined by the electrical distance $2\pi d/\lambda_g$. The phase distribution along the antenna is linear. The conductance of i-th slot in j-th subarray is [9]:

$$g_{ij} = g_j^{in} \frac{A_{ij}^2}{\sum\limits_{i=0}^{N_j} A_{ij}^2} \tag{3}$$

A_{ij} is the desired amplitude at this slot, N_j is the number of slots in the subarray, and g_j^{in} is the input admitance of the j-subarray. Using these conductances, the radiation coefficients read:

$$S_{ij} = \frac{g_{ij}^{\frac{1}{2}}}{1 + \frac{g_{ij}}{2}} \tag{4}$$

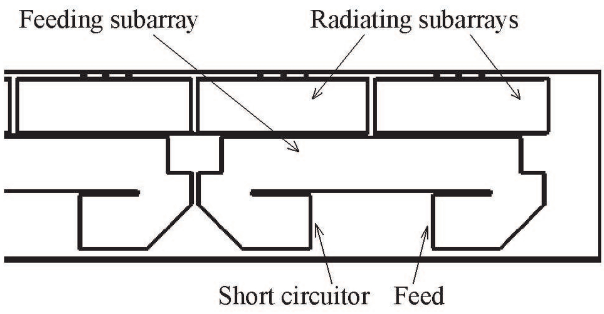

Figure 4. Using of bended feeding waveguides in the antenna building blocks.

Obtained radiation coefficients are used to calculate the offsets and the lengths of the radiating slots.

The radiating waveguides are divided into groups with common feeding as shown in Fig. 4. Each of these groups is fed by a feeding waveguide via inclined feeding slots in their common broad wall. The feeding waveguides are resonant ones so their width a_f should be selected in accordance with the required interelement spacing of the feeding slots. It is clear that this spacing should be equal to the distance between the centers of the neighboring radiating waveguides. Obvious calculations give the following expression for a_f

$$a_f = \frac{\lambda_0}{2 \cdot \left[1 - \left(\frac{\lambda_0}{2 \cdot (a_r + t)}\right)\right]^{\frac{1}{2}}} \tag{5}$$

where t is the wall thickness between the radiating waveguides, a_r is the broad wall width for the radiating waveguides. In accordance with this formula, the crossection of each feeding waveguide is $20.1 \times 7\,\text{mm}$.

The method of moments described in [7, 8] has been used in our computer simulations to determine the coupling coefficient of the feeding slot cut in the common broad wall of the radiating and feeding waveguides. It should be noted that in contrast to the radiation slots, the resonant length of the feeding slots depends slightly on the inclination angle only. This feature simplifies the synthesis and makes possible the usage of feeding slots with identical lengths.

The synthesis of the feeding subarrays is performed by using the energy method as well. But here we take into account that the power distributed into the radiating subarrays depends not only on the slot inclination angle but also on the radiation efficiency of the connected radiating subarrays.

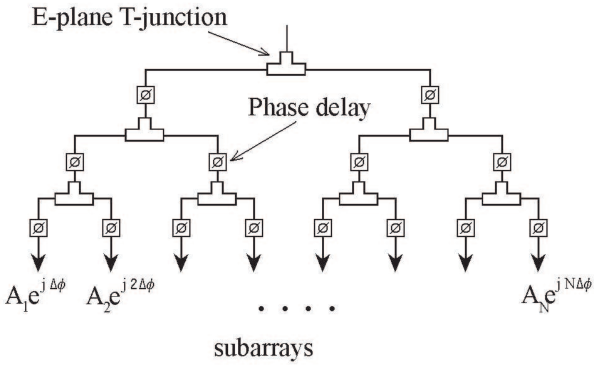

Figure 5. The structure of the feeding network for the antenna half.

It is known that the array of inclined slots is similar to a two-wire line with series obstacles. Therefore, the position of the short circuit measured from the end slot equals $\lambda_f/2$ (λ_f is the guide wavelength of the feeding waveguide). So the overlapping of neighboring feeding waveguides occurs. In order to prevent such overlapping, the feeding waveguides are bended down as indicated in Fig. 4.

The feeding network for the 16 inputs of the antenna feeding layer is of a parallel type based on matched E-plane T-junctions. It is organized identically for each 8 inputs of both antenna halves in the manner shown in Fig. 5. The feeding networks of the antenna halves are joined by an additional E-plane T-junction. The distribution of the subarray feed amplitudes A_i ($i = 1, 2, \dots 8$) approximates the required aperture distribution, and the $\Delta\phi$ is a deliberately introduced phase shift. This phase shift is the sum of phase shifts in each branch of the feeding tree (see Fig. 5). The phase distribution inside each building block is constant. So, the phase is changed discontinuously from one block to another.

The reason for the introduction of the phase shift $\Delta\phi$ is as following. In the case of in-phase antenna feeding ($\Delta\phi = 0$), reflections from the all inputs are summarized and the overall VSWR can be high. In order to prevent this effect, we introduce this phase distortion in the antenna E-plane phase distribution. To suppress completely the reflection and to achieve the ideal case of VSWR $= 1$, the phase shift should be equals 45°, which is determined as 360° divided on the number of the building blocks.

However, this phase shift results in some undesirable distortions of the antenna radiation pattern, like inclination of the antenna beam and growth of the side lobes. To reduce these distortions, the phase shift should be minimized. So, some compromise should be found for this shift. In our case, this compromise value is 30°, when VSWR is less than 1.4 in the operating frequency band, and the above distortions have acceptable values.

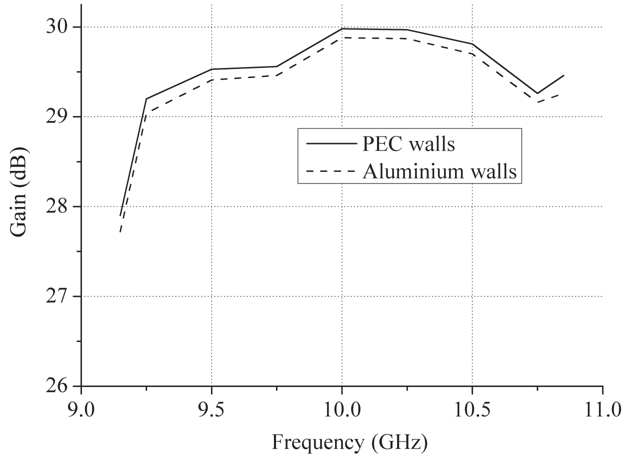

Figure 6. Gain of the designed antenna vs. the frequency.

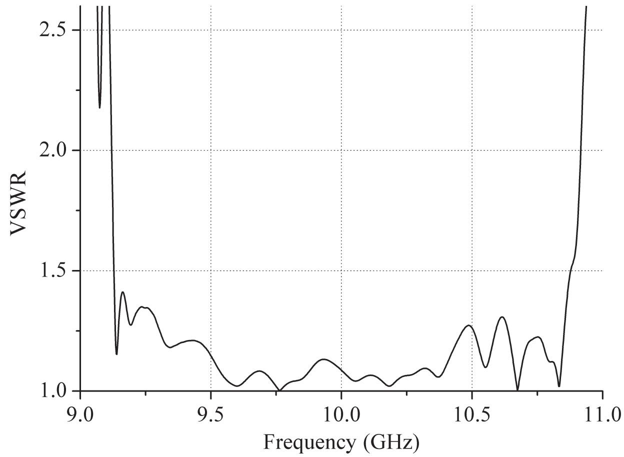

Figure 7. VSWR of the antenna with the matched feeding network versus the frequency.

4. SIMULATION RESULTS

To validate the antenna design, an FDTD simulation of the complete antenna has been performed. The simulation has confirmed that the beam of $4° \times 6°$ is formed. The antenna gain, calculated by taking into account the feeding network, in the operating frequency band is shown in Fig. 6. It is seen that the gain is around 30 dB, and it has a small

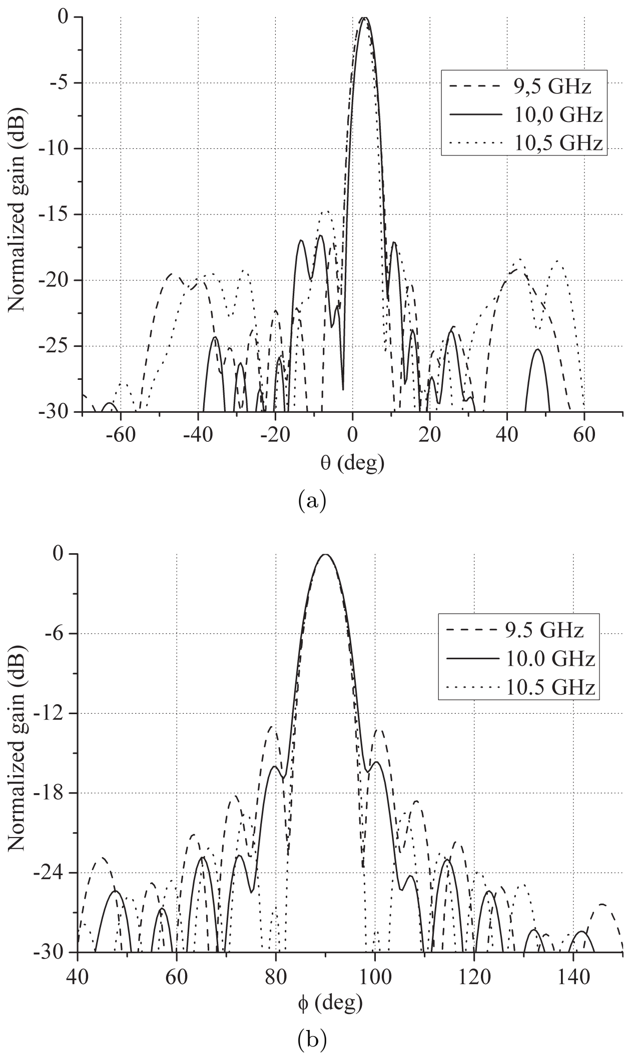

Figure 8. Simulated radiation pattern of the designed antenna.

variations (± 0.5 dB) within the bandwidth of about 1.7 GHz.

We have also found out that the total losses in the feeding system are low. This is also evident from Fig. 6, where the gain calculation is presented for the case of PEC and aluminum walls. There is only a minor decrease of the antenna gain due to the losses in the walls.

Simulation results on VSWR are illustrated in Fig. 7. It is seen that this value is less than 1.4 in the frequency band from 9.15 GHz to 10.85 GHz, which is larger as compared to the required frequency band. It should be stressed that this result is due to the usage of the introduced phase shifts between the feeding blocks as described in the previous section.

The phase shifts, as it was noted above, may cause some antenna pattern distortions in the E-plane. The corresponding pattern is shown in Fig. 8(a) for various values of the frequency. It is seen that there is a beam inclination of about 3°, which slightly depends on the operating frequency. The beam width in this plane is 4°.

The radiation pattern in the H-plane is illustrated in Fig. 8(b). There is no beam inclination because of the homogeneous phase distribution in this plane. However, the frequency value has a rather pronounced effect on the level of sidelobes.

5. CONCLUSION

In this paper, we have investigated possibilities of the development of wideband slotted waveguide arrays. The bandwidth limitations imposed by both individual slots and slot arrays have been analyzed. It is confirmed that the bandwidth of the conventional longitudinal slots made in a low profile waveguide can be as high as 20%. The application of subarraying technique allows reaching the bandwidth of such antennas as high as 15%.

These findings have been used in order to develop a novel X-band antenna with the beam of $4° \times 6°$ and the gain of about 30 dB. The dimensions of the antenna are $365 \times 260 \times 34$ mm which make it attractive for applications in SAR and other radar systems.

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