Leaky-Wave Antennas Realized by Using Artificial Surfaces

S. Maci, R. Magliacani, A. Cucini

Dept. of Information engineering, University of Siena, Via Roma 56,53100, Siena, Italy
macis@ing.unisi.it, maglia@dii.unisi.it cucini@dii.unisi.it,

1.INTRODUCTION

Within the class of electromagnetic bandgap structures, periodic surfaces printed in stratified dielectric media occupy an important role, as testified by the recent literature [1], [2]. Solution have been devised for antenna applications, often oriented to realization of compact antennas, or to suppression of surface- and space-wave propagation for reducing diffraction lobes. From the various works in literature also emerge a different application which uses the same printed periodic surface technology, to the gain enhancement [3], [4]. The basic mechanism to use for the gain enhancement is the excitation of weakly attenuated leaky waves, at the same type to that described in the past for unbounded layered media with high dielectric contrast [5].

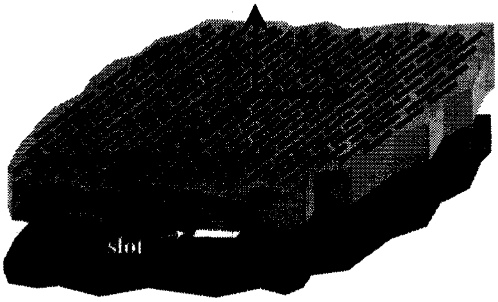

Fig. 1 *Geometry of the antenna. A gangbuster dipole-FSS is printed on a grounded slab and fed by a slots on a ground plane*

The excitation and leaky-wave propagation can be obtained by printing over a grounded slab a frequency selective surface (FSS). Within the bandwidth where the FSS resonates, the structure behaves as a weakly transparent parallel-plate waveguide. A point-source (dipole or slot) can thus excite a cylindrical leaky wave, thus implying a large phased aperture-field distribution on the FSS. The radiation of this large aperture produces a directive conical beam. For an appropriate design of the slab thickness, the leaky-wave phase velocity approaches infinity and the conical radiation collapses into a broadside beam. In this paper, a systematic procedure for the antenna analysis is presented, which uses a synthesis of the FSS via equivalent network, based on the data obtained by full-wave analysis.

2. GEOMETRY OF THE ANTENNA AND FSS HOMOGENEIZATION

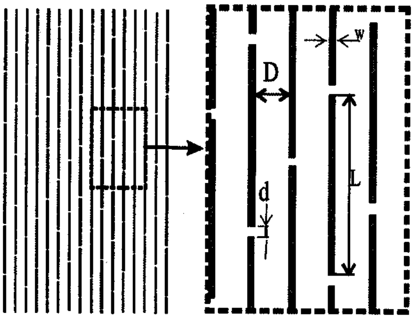

Fig. 2 *Geometry of the gangbuster surface. (L= 10 mm, D= 2.51 mm, d= 0.571 mm, w= 0.25 mm.)*

The antenna we are investigating is constituted by a "homogenizable" frequency selective surface printed on a grounded slab. For the sake of simplicity, we will consider the a dipole surface often denoted as "gangbuster" (Fig. 2), [6][7] but similar structures may be obtained by the use of other types of FSS. The term "homogenizable" is referred to the fact that the FSS can be viewed as a continuous surface with an equivalent reflection coefficient. This property arises from the fact that the basic elements (slot or dipole), although resonant at the operating frequency, are compactly assembled, thus resulting in a certain number of elements per square wavelength. This justify a description in terms of an equivalent

0-7803-7846-6/03/$17.00 ©2003 IEEE

concentrated LC network. When the structure is composed by slots, the inductance dominates and the equivalent circuit is more appropriately seen as a parallel L-C circuit. For printed dipoles the dominant term is an L-C series circuit. The equivalent circuit can be derived by a pole-zero synthesis from full wave data.

3. EQUIVALENT NETWORK SYNTHESIS OF THE FSS

For the sake of simplicity, consider a type-2 gangbuster surface of dipoles [7], printed on a semi-infinite medium. The geometry is that presented in Fig. 2. The equivalent network for the TE and TM (w.r.t. the surface normal) polarization is represented in the same figure. As shown in Fig. 3, the scan in the E- and H-plane are relevant to TE and TM polarization, respectively, in order to ensure the tangent E-field component parallel to (i.e., interacting with) the dipoles. For the opposite polarization (TE on E-plane and TM on H-plane) the dipoles, are almost invisible, because very thin

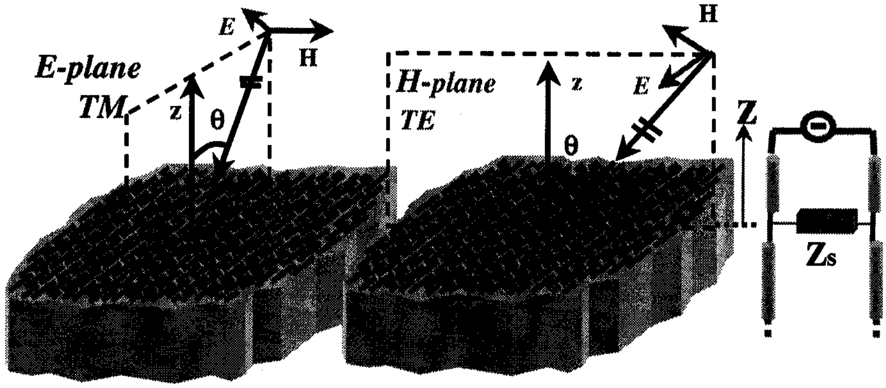

Fig. 3 *TM and TE polarization for plane wave incident on a FSS printed on an semiinfinite slab and relevant scan planes which realizes an interaction with the printed dipoles. On the right, representative z-transmission line is depicted, whose parameters (characteristic impedances and propagation constants) depend on the TE and TM polarization*

Fig. 4 shows the equivalent impedance versus the frequency for the equivalent impedance $Z_s= Z_s(\theta, f)$ The direction of incidence is perpendicular to the surface ($\theta=0$). For the sake of simplicity, the lower dielectric is assumed as a free-space. The results have been obtained by a full-wave method after identification of the impedance via the equivalent circuit in Fig. 3. The full-wave results have been obtained by a Method of Moments code in the spectral domain, and next successfully validated by a commercial full-wave FSS software [8]. As apparent from Fig. 4, by a simple inspection of the poles and zeros of $Z_s(0,f)$, the equivalent circuit in the inset of Fig. 4 has been assumed to synthesize the broad-band behavior of the full-wave data. In this circuit, the L-C series block describes the behavior at lower frequencies, and reproduces the first zero at 17.7 GHz. The parallel L-C blocks accounts for the pole at 34.1 GHz. The remaining zero at 36.7 GHz is caused by the interaction between the series-type and the parallel-type blocks. In order to evaluate the values of the various *L* and *C* parameters, a least-square procedure with conjugate gradient optimization has been implemented, thus leading to the values indicated in the caption of Fig. 4. The

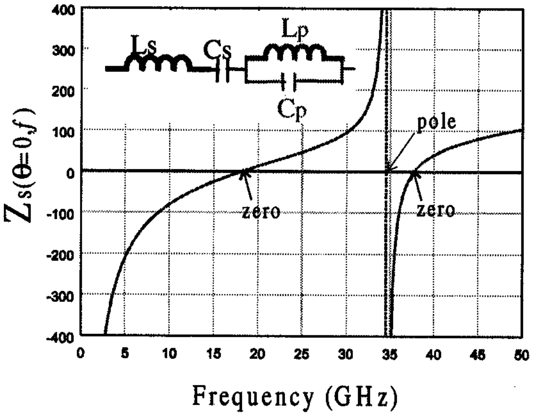

Fig. 4 *Equivalent impedance versus frequency and L-C network interpretation. The values which matches the full wave data are: Lp= 0.1372* nH, *Ls=0.8914* nH *Cp=.0.1384* pF, *Cs=0.07559* pF.

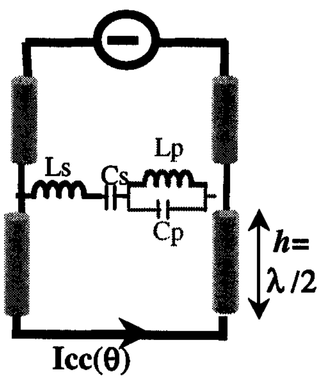

Fig.5 *Equivalent circuit for the leaky-wave antenna. The distance of the ground plane is a half-wavelength from the FSS*

circuit in the inset matches the full-wave data without appreciable difference in the range of drawing.

Now, let us consider the original structure in Fig. 1, and its equivalent z-transmission line shown in Fig. 5. The ground plane is located at a half-wavelength from the FSS surface (λ associated with 17.7 GHz, i.e., the frequency corresponding to the first series resonance of the FSS). In general, the presence of the ground plane may modify the equivalent circuit deduced from the geometry in Fig.3; this occurs when higher order (evanescent) Floquet modes are not significantly attenuated at the ground plane. In our case, due to the relatively large distance between the FSS and the ground plane, we successfully verified the applicability of the network synthesized without ground plane. The parallel LC resonance is almost short-circuited at the operating frequency, but its presence is required due to the non-negligible perturbation it exerts on the first resonance.

Fig. 6 shows the current in the short circuit as a function of the frequency. This current corresponds to the tangential magnetic field at the ground plane for a plane wave incident from broadside. It is evident that around 18.40 GHz the current exhibits an extremely selective peak. Other peaks are found for the other resonant frequencies of Zs. The high-Q of thr resonance at 18.4 GHz is the result of a two selective mechanism, *i.e.* the resonance of the FSS and the resonance of the half-wavelength slab.

Let us now consider the short-circuit current as a function of the incident angle θ at the resonant frequency of 18.40 GHz. The equivalent impedance Z_s is set as that obtained for the perpendicular incidence.

The selective behavior in frequency also implies a selective behavior with respect to the scan angle. We observe that, being the short circuit current associated to the tangent magnetic field at the ground plane, the relevant pattern is that received by a small magnetic dipole (slot). Thus, the TE and TM responses represent the co-polar E- an H-plane far field components for the antenna shown in Fig.1. It is apparent that the beam-widths in the two planes are almost identical.

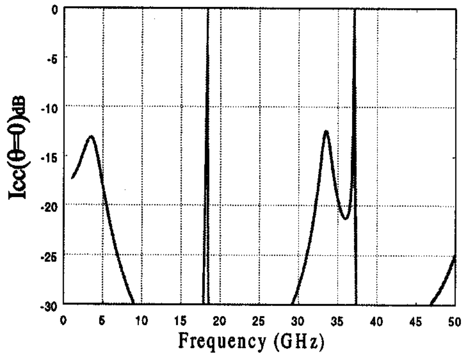

Fig. 6 *Large bandwidth short circuit current of the equivalent circuit in Fig. 5 (θ=0)*

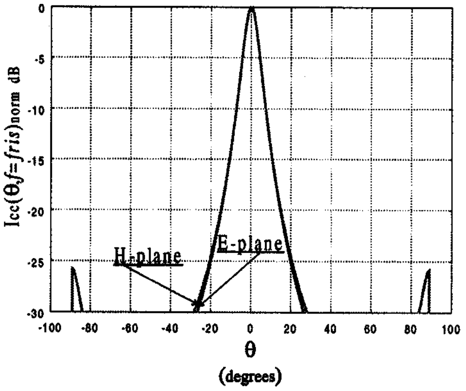

Fig 7. *Ground plane current as a function of the incidence angle in the H and E plane at f=18.40 GHz.*

4. INTERPRETATION IN TERMS OF LEAKY WAVE EXCITATION

If the antenna is seen as a transmitting one, the phenomenon of high-gain can be associated to the excitation of a leaky wave. To this end, the equivalent circuit in Fig. 5, in which a matched load is take the place of the generator is used to find the dispersion (transverse-resonance) equation. From this analysis it is seen that a solution of the transverse resonance has been found for complex value of k_x (TM-case) or k_y (TE-case) The complex wavenumber exhibits a very low imaginary part and an almost vanishing real part. The details for this analysis are not presented here.

5. NUMERICAL RESULT FOR FINITE FSS AND CONCLUSIONS

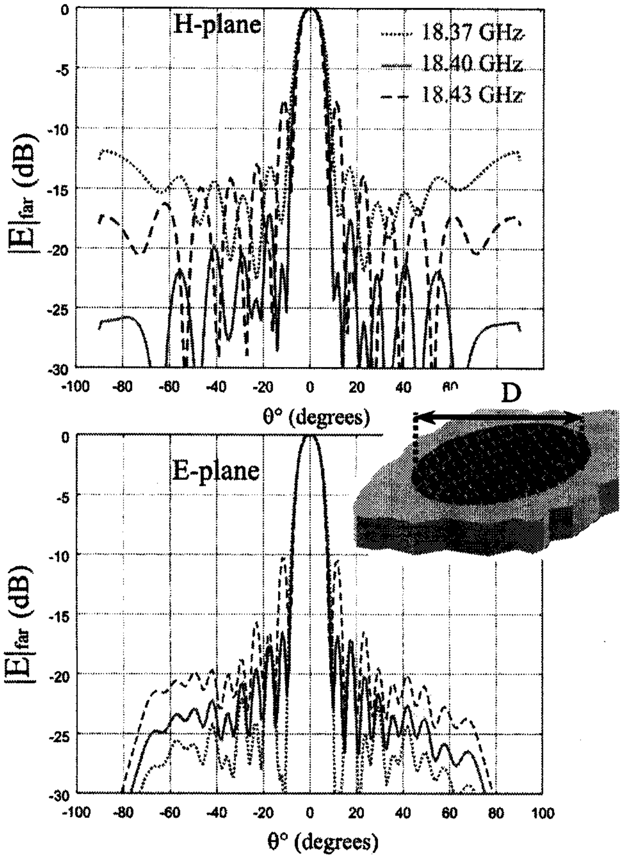

Fig. 8. *Finite circular FSS gangbuster printed on an infinite slab and relevant radiation pattern in the principal planes. The surface is fed by a small slot; the diameter D is equal to 16,2 cm (10λ). Results obtained by ENSEMBLE™, using a relative*

In order to test the practical applicability of the concept, the antenna has been studied with a finite FSS printed on an infinite substrate. The analysis has been performed by using the software ENSEMBLE™. The same elements used in the previous case have been considered, but distributed over a circular surface with a diameter of 10 wavelength (Fig. 8). The radiation pattern is shown for various frequencies around the resonant frequency (18.40 GHz). The radiation pattern exhibits the expected directive property (gain of 26 dB), with a good rotational symmetry. The diffraction lobes, obviously assent for the ideal infinite configuration, increases dramatically in the *H*-plane when the frequency deviates from the central frequency. The relative bandwidth of the antenna is then limited to 0.2%.

In conclusions, the major limitation of the antenna is the narrow bandwidth. Larger bandwidth may be obtained by using a less tied gangbuster surface ("type 1", [7]). In this case, the structure becomes similar to that presented in [4]. The price which is paid is however a drastic decrease of gain.

REFERENCES

[1] D. Sievenpiper *et al.*vol.47, IEEE Trans. Microwave Theory Tech, no 11, Nov. 97, pp. 2059-2074,

[2] Fei-Ran Yang, *et al.* vol. 47, no 11, Nov. 97, pp. 2092-2098

[3] H. D. Yang and D. R. Jackson, IEEE Trans. Antennas Propagat., Vol. 48, N. 4, pp. 556-564, Apr. 2000

[4] T. Zhao et al., AP-S Symp. Digests, , July 8-13 2001, Boston, Massachusetts, Vol. 3, pag. 248-252

[5] D.R. Jackson *et al.*, IEEE Trans. Antennas Propagat, vol. AP-41,pp. 344-348, March 1993

[6] N. Engheta, URSI meeting, Boston, Massachusetts, July 8-13 2001, pag. 389

[7] B. A. Munk , *Frequency Selective Surfaces:Theory and Design* , Wiley, 2000

[8] R. Mittra, V. V. S Prakash, S. Chakravarty, N. Ti Huang , JiFu Ma , *Frequency Selective Surfaces Simulator*, July 2002