GAIN ENHANCEMENT OF DIELECTRIC RESONATOR LOADED WAVE-GUIDE ANTENNAS WITH DIELECTRIC OVERLAYS

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model correctly describes the electromagnetic nature of the media throughout the considered frequency bandwidth.

22nd January 1992

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GAIN ENHANCEMENT OF DIELECTRIC RESONATOR LOADED WAVEGUIDE ANTENNAS WITH DIELECTRIC OVERLAYS

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Indexing terms: Antennas, Dielectric and dielectric devices, Resonators, Antenna radiation patterns

The gain characteristics of a dielectric resonator loaded coaxial probe fed circular waveguide antenna (DRLWA) with overlaying parasitic discs have been investigated experimentally. Results indicate that, when properly spaced, the overlays can enhance the gain by more than 6dB.

Introduction: Dielectric resonator antennas (DRAs) have recently been proposed [1] as simple efficient nonmetallic radiators, especially useful for microwave and millimetre-wave bands. Recently, it was also shown [2] that when the dielectric resonator is embedded inside a circular waveguide the radiation pattern of the waveguide aperture may be further improved by optimising the relative dielectric and waveguide dimensions. In particular, the waveguide dimension adjustment helps in tuning the resonance frequency to a desired one. In this Letter, experimental results on the effect of overlays on the gain of the latter radiator are reported. It may be mentioned that in a study in this area, Lee and Lee [4] and Afzalzadeh [5] noticed a significant improvement in the gain of bare microstrip antennas with parasitic overlays, a matter that prompted the present investigation.

Experimental results: A typical DRLWA is shown in Fig. 1. In the present study the dielectric disc was chosen to have $\varepsilon_r = 8$ and dimensions for resonance frequency of 11 GHz. The disc

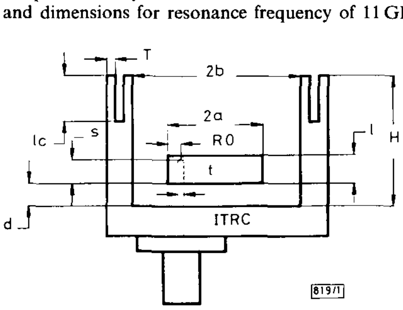

Fig. 1 *Geometry*

is excited by a coaxial probe so that the input impedance is nearest to 50Ω. When the disc is encapsulated in a waveguide of internal diameter 22·3 mm the resonance frequency shifted to 11·35 GHz. As mentioned in Reference 3, the input impedance for the probe can be predicted by solving for the excited modes after determination of the proper Green function of this particular arrangement. The $\lambda/4$ choke is introduced to reduce both sidelobe and crosspolarisation levels of the waveguide radiation pattern. Fig. 2 shows the E-plane pattern with a beamwidth of approximately 52°.

When parasitic elements are placed above the waveguide aperture, an appreciable increase in the forward gain is observed. To study the effect the various models shown in Fig. 3 were examined. In all these models, discs of $\varepsilon_r = 4$ and diameter 6·6 cm corresponding to 3·9 λ_g were found to be the smallest suitable to yield the best compromise for gain increase and sidelobe level decrease. Therefore in this study this

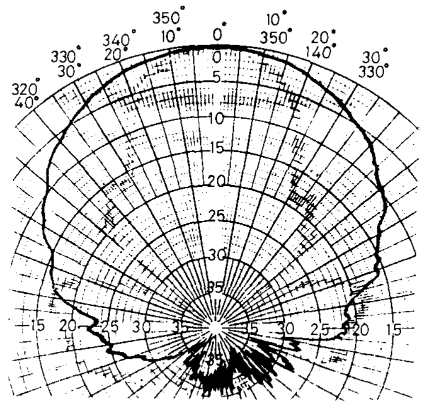

Fig. 2 *E-plane radiation pattern of DRLWA*

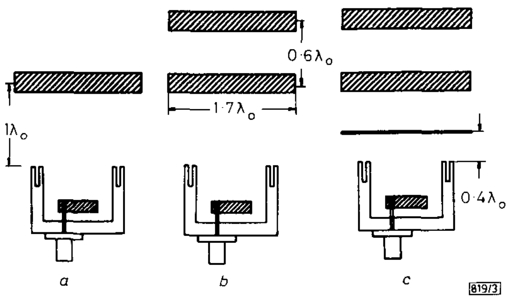

Fig. 3 *DRLWA with overlays*

a Single overlay $\varepsilon_r = 4$
b Double overlay $\varepsilon_r = 4$
c Double overlay $\varepsilon_r = 4$ with additional $\varepsilon_r = 2{\cdot}33$ parasitic

parameter was fixed and attention was focused on the effect of distance above the aperture. Moreover, the thickness of the parasitic discs was chosen to be $\lambda_g/2$ to avoid strong multiple reflection effect within the dielectric and thus allow maximum transmission through it.

The best pattern obtained from model *a* in Fig. 3 in which the lowest sidelobe occurred when the disc was at a distance λ_0 is shown in Fig. 4. The first sidelobe has a level of about 17 dB and also a large cluster sidelobe of the same level near the horizon. When another parasitic element was added (model *b* of Fig. 3), the first sidelobe was smoothed and a slight increase in gain observed. However, the horizontal lobe still persisted. To reduce the level of this sidelobe, a thin PTFE disc of the same diameter was introduced in the vicinity of the aperture (model *c*). In this way we could reduce all sidelobe levels below about 22 dB. The gain of this model turns out to be about 12 dB.

The above characteristics were observed to remain almost constant for at least 10% bandwidth. The return loss was better than 14dB ($VSWR < 1{\cdot}6$) for a wider bandwidth, however.

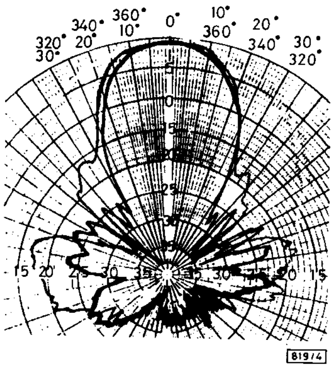

Fig. 4 *Radiation pattern of DRLWA with dielectric overlays*

Conclusion: An experimental study for gain enhancement of dielectric resonator loaded waveguide antennas using parasitic dielectric directors has been presented. Results indicate that, with two directors a gain improvement of about 7dB or more may be obtained over a bandwidth of at least 10%. The whole unit may be packaged inside lower dielectric capsule with elements placed properly using foam spacers.

14th January 1992

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PREFORMING AN ELECTROMAGNETIC PULSE IN LOSSY MEDIUM

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Indexing terms: Electromagnetic waves, Wave propagation

Plane wave propagation of a transient signal in a lossy homogeneous medium is considered. It is shown that it is possible to prescribe a source signal waveform which will produce a Gaussian signal waveform after propagating a specified distance.

As is well known, a transient signal propagating in lossy media suffers both attenuation and waveform distortion [1]. We pose the question: can the source signal waveform be tailored to produce an impulse after propagating a given distance in the medium? A solution for an ideal case is presented here. In some sense, the problem is analogous to the design of matched filters in radar signal processing [2].

We consider an unbounded homogeneous region of permittivity ε, conductivity σ and magnetic permeability μ. We deal with one dimensional plane wave propagation in the positive z direction. Thus the transverse derivatives $\partial/\partial x$ and $\partial/\partial y$ are zero. The electric field is taken to have only an x component $e_x(z, t)$ and, henceforth, the subscript is dropped.

Let us now say, at $z = z_0$, the desired signal is the Gaussian pulse

$$e(z_0, t) = e_0 \exp -\beta^2 t^2) \tag{1}$$

being defined for all time, $-\infty < t < \infty$, where e_0 and β are constants. The width of the pulse is $1/\beta$ seconds. Noting that $z_0 > 0$, the objective is to determine the source signal $e_s(t) = e(0, t)/e_0$ at $z = 0$ which will produce the desired Gaussian pulse at $z = z_0$.

To proceed we take Fourier transforms as follows:

$$E(z_0, j\omega) = \int_{-\infty}^{+\infty} e(z_0, t) e^{-j\omega t}\, dt \tag{2}$$

$$E(0, j\omega) = \int_{-\infty}^{+\infty} e(0, t) e^{-j\omega t}\, dt \tag{3}$$

Now clearly

$$E(z_0, j\omega) = E(0, j\omega) \exp\left[-\gamma(j\omega) z_0\right] \tag{4}$$

where

$$\gamma(j\omega) = [j\mu\omega(\sigma + j\varepsilon\omega)]^{1/2} \tag{5}$$

is the propagation factor for plane wave transmission at an angular frequency ω.

The desired solution is the inverse Fourier transform

$$e(0, t) = \frac{e_0}{2\pi} \int_{-\infty}^{+\infty} E(z_0, j\omega) \times \exp\left[\gamma(j\omega) z_0 + j\omega t\right] d\omega \tag{6}$$

where, on employing eqns. 1 and 2, we deduce that

$$E(z_0, j\omega) = [e_0\sqrt{(\pi)}/\beta] \exp(-\omega^2/4\beta^2) \tag{7}$$

Eqn. 6 may then be written in the form

$$e_s(t) = \frac{1}{\beta\sqrt{(\pi)}} \int_0^{\infty} \exp\left[R(\omega) z_0 - (\omega^2/4\beta^2)\right] \times \cos\left[I(\omega) z_0 + \omega t\right] d\omega \tag{8}$$

where $R(\omega) + jI(\omega) = \gamma(j\omega)$. Here we have also used $R(-\omega) = R(\omega)$ and $I(-\omega) = -I(\omega)$.

In the relatively trivial case, where $\sigma = 0$ and both ε and μ are independent of frequency, $\gamma(j\omega) = j\omega/v$ where $v = (\varepsilon\mu)^{-1/2}$. On using eqn. 6, we obtain

$$e(0, t) = e_0 \exp\left[-\beta^2\left(t + \frac{z_0}{v}\right)^2\right] \tag{9}$$

which is also a Gaussian pulse with a maximum at $t = -z_0/v$ seconds.

When the conductivity is not zero, the integral given by eqn. 8 needs to be evaluated numerically. Results for $e_s(t) = e(0, t)/e_0$ are shown in Fig. 1 as a function of βt where $\beta = 10^7\,\mathrm{s}^{-1}$, $z_0 = 1\,\mathrm{m}$, $\varepsilon/\varepsilon_0 = \varepsilon_r = 10$, $\mu_r = \mu/\mu_0 = 1$, and $\sigma = 0, 1, 5, 10$, and $50\,\mathrm{mS/m}$. The zero conductivity case replicates the desired signal Gaussian pulse $e(z_0, t)$ in accordance