Gain enhancement of dielectric resonator loaded waveguide antennas with dielectric overlays

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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 discs, 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 is excited by a coaxial probe so that the input impedance is nearest to 50 Ω. 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 \( \lambda_0/4 \) choke is introduced to reduce both sidelobe and cross-polarisation 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 \( \lambda_0 \) were found to be the smallest suitable to yield the best compromise for gain increase and sidelobe level decrease. Therefore in this study this parameter was fixed and attention was focused on the effect of distance above the aperture. Moreover, the thickness of the parasitic disc was chosen to be \( \lambda_0/2 \) 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 \( \lambda_0 \) is shown in Fig. 4. The first sidelobe has a level of about \( \lambda_0 \) 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 < 1.0) for a wider bandwidth, however.

**Conclusion:** An experimental study for gain enhancement of dielectric directors has been presented. Results indicate, however, 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.

**References**


**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 ∂/∂x and ∂/∂y are zero. The electric field is taken to have only an x component E(x, t) and, henceforth, the subscript is dropped.

Let us now say, at z = z₀, the desired signal is the Gaussian pulse

\[ E(z₀, t) = \varepsilon₀ \exp\left(-\beta t^2\right) \]

being defined for all time, -∞ < t < ∞, where ε₀ and β are constants. The width of the pulse is 1/β seconds. Noting that z₀ > 0, the objective is to determine the source signal E(t) = E(0, t)/ε₀ at z = 0 which will produce the desired Gaussian pulse at z = z₀.

To proceed we take Fourier transforms as follows:

\[ E(z₀, \omega) = \int_{-\infty}^{\infty} E(0, t) \exp(-j\omega t) \, dt \]

\[ E(0, \omega) = \int_{-\infty}^{\infty} E(z₀, t) \exp(-j\omega t) \, dt \]

Now clearly

\[ E(z₀, \omega) = E(0, \omega) \exp(-j\omega z₀) \]

where

\[ \gamma(\omega) = [J_{\omega}(\alpha + j\omega)]^{1/2} \]

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

The desired solution is the inverse Fourier transform

\[ E(0, t) = \frac{\varepsilon₀}{2\pi} \int_{-\infty}^{\infty} E(z₀, \omega) \times \exp\left(j\omega(t + z₀)/\omega\right) \, d\omega \]

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

\[ E(z₀, \omega) = [R(\omega) + j\omega](-\omega^2/4\beta^2) \]

Now clearly

\[ E(0, t) = \frac{\varepsilon₀}{\beta} \exp\left(-\beta t^2 + \frac{z₀}{\beta}\right) \]

which is also a Gaussian pulse with a maximum at t = z₀/β seconds.

When the conductivity is not zero, the integral given by eqn. 8 needs to be evaluated numerically. Results for E(t) = E(0, t)/ε₀ are shown in Fig. 1 as a function of βt where β = 10^7 s⁻¹, z₀ = 1 m, ε₀μ₀ = ε₀μ₀ = 10, and σ = 0, 1, 5, 10, and 50 mS/m. The zero conductivity case replicates the desired signal Gaussian pulse E(t₀) in accordance with the above characteristics.