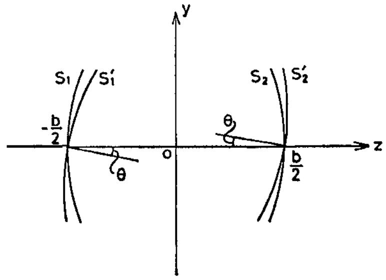

Fig. 2. Approximated optimum mirrors (S_1', S_2') obtained by rotating usual confocal mirros (S_1, S_2).

$$R_0 = \frac{k}{\omega\epsilon_{33}}$$

$$k^2 = \omega^2\epsilon_{22}\mu_0\left(1 - \frac{\epsilon_{23}^2}{\epsilon_{22}\epsilon_{33}}\right)$$

$$p = \frac{\epsilon_{22}}{\epsilon_{33}}$$

$$q = \frac{\epsilon_{23}}{\epsilon_{33}}$$

$$c_0 = ka^2\left(\frac{1}{pb} - \frac{1}{2p_0}\right)$$

$$c_1 = \frac{ka^2}{pb}$$

$$c_2 = ka\left(A - \frac{q}{p}\right)$$

$$c_2' = ka\left(B - \frac{q}{p}\right)$$

$$b_a = b\left(1 + \frac{q^2}{2p_0}\right)$$

$$\epsilon_{s,y}(S_i) = \frac{2E_{s,y}(S_i)}{R_0}\cdot \quad (6)$$

In order that this FPR is equivalent to the confocal FPR filled with isotropic medium, the following condition must be satisfied:

$$c_0 = c_2 = c_2' = 0. \quad (7)$$

Consequently, from (6) and (3), we obtain the optimum mirror equations

$$g_1(y) = \frac{\epsilon_{33}}{2\epsilon_{22}b}y^2 + \frac{\epsilon_{23}}{\epsilon_{22}}y - \frac{b}{2}$$

$$g_2(y) = -\frac{\epsilon_{33}}{2\epsilon_{22}b}y^2 + \frac{\epsilon_{23}}{\epsilon_{22}}y + \frac{b}{2}\cdot \quad (8)$$

When the isotropy of the medium is small ($\epsilon_{23}/\epsilon_{22} \ll 1$, $\epsilon_{33}/\epsilon_{22} \cong 1$), the optimum mirrors can be obtained approximately by rotating the usual confocal mirrors by the angle

$$\tan\theta = -\frac{\epsilon_{23}}{\epsilon_{22}} \quad (9)$$

as shown in Fig. 2. In this case, θ agrees closely with the angle between the propagation vector and the Poynting vector of the extraordinary wave.

For an example, we consider the second harmonic generator using an FPR filled with KDP. If we choose 1.1526-μ laser light as the fundamental, the index matching occurs when the angle between the optic axis of the KDP and the z axis is 42.7° [4], and we obtain $\epsilon_{11} = 2.32\epsilon_0$, $\epsilon_{22} = 2.26\epsilon_0$, $\epsilon_{33} = 2.25\epsilon_0$, $\epsilon_{23} = \epsilon_{32} = -0.0658\epsilon_0$. Therefore, we must rotate the mirrors by $\theta \cong 1.67°$ to obtain effective resonance of the second harmonic extraordinary light.

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Use of the TEM Mode in Microwave Heating Applicators

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***Abstract*—A combination of hot air and microwave heating is generally achieved by placing air knives alongside the waveguide applicators when drying web materials. If the air knife is placed inside the applicator, using it as the center conductor of a TEM structure, better or additional moisture removal may be achieved. The TEM mode also permits more control of the wave attenuation. Perturbation formulas are developed to compare conventional slotted waveguide applicators with two TEM systems. Experimental results are given. Improved uniformity of heating and lower leakage are indicated.**

The use of microwave power in the 915±13-MHz and 2450±50-MHz ISM bands for heating web materials is increasing. The material is generally fed through a slot in the center of the broadside wall of a rectangular waveguide operating in the dominant TE_{10} mode. Several configurations are possible [1]. Other waveguide structures, including fringe field applicators [2], [3] have been used for this purpose. A traveling wave in a slotted waveguide applicator of simple geometry has many mechanical advantages but uniformity of heating has been difficult to achieve because the available power for heating decays exponentially and reflections occur. The problem has been that the TE_{10} structure provides poor control of the coupling of power to the load.

Currents in a waveguide wall have a transverse component except along a line down the center of a broadside wall, as shown in Fig. 1. This is the only location for a slot that will be nonradiating. Any structure bearing a transverse electromagnetic (TEM) mode will be nonradiating if it is slotted longitudinally at any location. The TEM mode will propagate at all frequencies down to zero, whereas TM and TE modes are cut off unless the largest dimension is of the order of half of a free-space wavelength, or greater. In rectangular waveguides, air knives must be placed outside the waveguide, whereas with TEM structures the center conductor forms an obvious air conduit, although the resulting air stream is not an impinging jet. The slot may be placed anywhere and the cross section may be changed to control the attenuation. Thus it is possible to combine the hot air and microwave applicator sections and to reduce, for example, the size of 915-MHz applicators, depending upon the power-handling capacity required.

Altman [4] derives the following formula, from perturbation theory, for the attenuation of a lossy dielectric sheet in a guided

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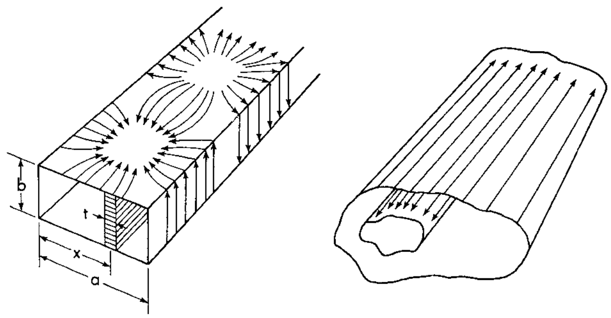

Fig. 1. Current in rectangular and TEM guides.

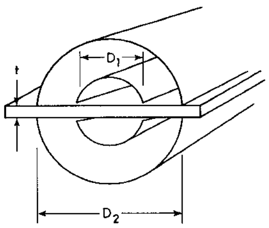

Fig. 2. Circular coaxial TEM guide.

wave system:

$$\alpha = \epsilon_0\omega\epsilon_r''(j\omega)\,\frac{\int_{\Delta S}(\bar{E}_0{}^*\cdot\bar{E})dS}{2\int_S(\bar{E}_0{}^*\times\bar{H})\cdot\bar{n}dS} \tag{1}$$

where S is the cross-sectional area of the guide and ΔS is that of the perturbed region (the web in our case). $\bar{E}_0$ is the unperturbed electric field, $\bar{E}$ is the perturbed field quantity, and the asterisk indicates a complex conjugate. ϵ_r'' is the imaginary part of the complex relative dielectric constant, $\bar{n}$ is a normal unit vector, and α is the attenuation of the perturbed guide, in nepers per meter.

For rectangular TE_{10} waveguides, (1) becomes

$$\alpha_g = \epsilon_0\omega\epsilon_r''F_g\eta_0\left(\frac{\lambda_g}{\lambda_0}\right)\sin^2\left(\frac{\pi x}{a}\right) \tag{2}$$

where $F_g=(t/a)$ and a and t are as in Fig. 1. The guide wavelength λ_g is greater than the free-space wavelength λ_0. The accuracy is good up to $(t/a)\simeq 0.1$ for common practical applications [5]. For a circular coaxial line we obtain:

$$\alpha_l = \tfrac{1}{6}\epsilon_0\omega\epsilon_r''F_l\eta_0\left(1+\frac{D_1}{D_2}+\frac{D_2}{D_1}\right) \tag{3}$$

where the filling factor F_l is given by $F_l=4t/\pi(D_1+D_2)$ and the dimensional quantities are as shown in Fig. 2. This structure has been built in standard $1\frac{5}{8}$-in coaxial line and the above attenuation formula confirmed for light webs (about 2 mm thick, 50-percent water, dry basis).

Comparing the TE_{10} rectangular guides to a 50-Ω coaxial line, we obtain

$$\frac{\alpha_l}{\alpha_g} = 0.623\left(\frac{F_l}{F_g}\right)\left(\frac{\lambda_0}{\lambda_g}\right). \tag{4}$$

In Fig. 3, computed results for attenuation in a 0.24-kg/m² web at about 25-percent water content are shown. Since it has no cutoff frequency, the TEM structure can give higher attenuation, i.e., F_l can be made large by making the structure small. Cutoff effects prevent this in the rectangular guide, since if the wide dimension a and web thickness t are fixed, F_g is fixed.

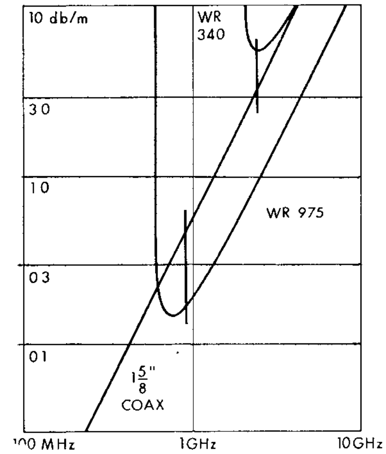

Fig. 3. Results of perturbation theory.

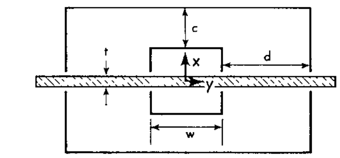

Fig. 4. Rectangular TEM structure.

In Fig. 4, another TEM system is shown, chosen mainly to show that low attenuation factors can also be achieved. If we assume that $w\gg c$, we can assume that the voltages $V=V_c=V_d=(E_d)d$. If the perturbation is small and the center conductor square, we can expand (1) as follows:

$$\alpha = \frac{\eta_0\omega\epsilon_0\epsilon_r''\int_0^d\left(\frac{V}{d}\right)^2 t\,dy}{2\int_0^c\left(\frac{V}{c}\right)^2 w\,dx + 2\int_0^d\left(\frac{V}{d}\right)^2 w\,dy + \iint_{\text{corners}}(E_c)^2dS}. \tag{5}$$

If $d\gg c$ the first of the lower three integrals dominates and the power is carried mainly above and below the center conductor. The attenuation is then proportional to c/d. This result is for the asymptotic case: low characteristic impedance and c/d very small. Using a 50-Ω rectangular structure it has been demonstrated that this technique does indeed give the desired control. For $w/c=1.6$ a change in c/d of from 1 to 0.25 gave 2 to 1 attenuation control.

Similar results have been demonstrated by Jull *et al.* [6]. The main problem in either case will be arcing, but with coaxial lines this should be easier to control since the TEM structure permits wider regions for high power transport. The corners must obviously be rounded.

There are several advantages in reducing the attenuation of the web in the applicator. If the input side of the web absorbs a large part of the power the other side will remain wet until the input side is nearly dry. Also high attenuation occurs in conjunction with a large reflection coefficient for the edge of the web, since both increase with high electric field concentration in the web region of the applicator. This gives hot spots, which dry first, rather than leveling [7]. The power deposition is given by

$$W = \omega\epsilon_0\epsilon_r''(j\omega)\iint_{\Delta S}\int_{z_1}^{z_2}(\bar{E}^*\cdot\bar{E})dz \tag{6}$$

where $\Delta S(z_2-z_1)$ is the volume of the web under consideration. Normally, as a web dries it absorbs less power, but it has been shown [8] that drying of a wet cellulose mixture does not guarantee that ϵ_r'' will be reduced, and it can even increase due to heating. If the heating is

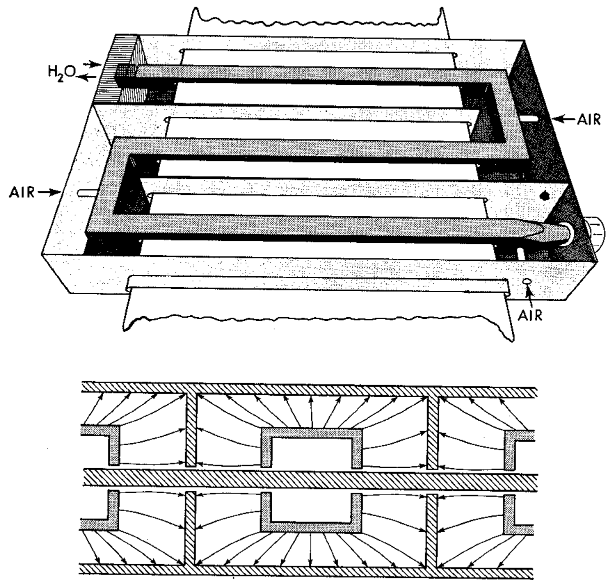

Fig. 5. Schematic diagram of a complete applicator.

uneven and the temperature quite high, thermal runaway and scorching can result. As the fields near the web are reduced to control attenuation, leakage from the slots will also be reduced. Experiments with a Narda 8110 Radiation Monitor [9] showed that leakage power is proportional to attenuation and otherwise unchanged by going from TE_{10} rectangular waveguide to a TEM structure with the same wall thickness and web load. All these problems are compounded by thick or very wet webs and the TEM structure thus permits heating heavier loads.

Fig. 5 shows a sketch of an implementation of this system. Air can be introduced through chokes or dielectric ducts, flowing from the center conductor directly over the web. The inner conductor can be formed from channel aluminum, rounded at the corners to prevent arcing. The front face, or the dividers, can be set at an angle to control uneven heating. If the center conductor is fairly wide, the spacing to the upper and lower ground plane will dominate the characteristic impedance, permitting reduced sensitivity to manufacturing irregularities.

Conclusions

It has been shown that by going to the TEM mode it is possible to build a simple web applicator that is well designed for both microwave power and air. The applicator size can be reduced for 915 MHz and there is, in principle, no lower frequency limit. Perturbation theory has been found to handle the analysis adequately as is expected for light dielectric loads. Better methods are available to control electric fields in the vicinity of the web and thus the structure is adaptable to a much wider variety of loads.

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Feedthrough for Digital Latching Ferrite Phasers

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***Abstract*—Feedthroughs are used in latching ferrite phasers having control wires entering the microwave region. The feedthrough damps out the troublesome resonances that occur as a result of air gaps at the ferrite–waveguide wall interface and at the same time chokes out the radiation from the holes through which the latching wires pass. The equations and curves necessary to design the feedthrough are presented, as well as some experimental results obtained at *C* band on a practical production configuration.**

Introduction

In digital latching ferrite phasers, the presence of air gaps between the surface of the ferrite and the walls of the waveguide cause higher order modes to be excited which couple onto the latching wires and can cause absorption resonances in the dominant-mode output and excessive radiation leakage out of the waveguide. The amount of higher order mode excitation, and consequently the coupling, depends upon the size of the air gap; small air gaps of the order of 0.0003 wavelength can cause relatively large coupling. Although special techniques have been applied to reduce these air gaps, such as finish grinding the ferrite surface flat and using some sort of prestressed cover for the waveguide, air gaps that can significantly degrade performance still occur. Instead of going to extremes with these expensive and time-consuming measures to prevent the air gaps from occurring, the effects of the higher mode excitation can be eliminated cheaply and efficiently by terminating the latching wires in a feedthrough that damps out the resonances and chokes out the radiation leakage.

In the past, tubes of lossy dielectric were placed on the wires or in the waveguide walls through which the wires pass. These tubes produced some damping of resonances, but exhibited high RF leakage and increased the overall loss of the device. Recently, resonances have been damped by a resistive sheet across the guide [1]. The concept of this feedthrough is to introduce a very low impedance at the wire exit hole, giving low RF leakage (good isolation), and to effect a damping of resonances by adding loss outside the propagating waveguide. Design equations and curves are presented for choosing both the dimensions of the device and the type of material. Also described are a theory of operation of the feedthrough and experimental results at *C* band obtained using different types of lossy dielectric as the feedthrough material in a production design.

Theory of Feedthrough Operation

The dimensions of the feedthrough are shown in Fig. 1. The shank of the feedthrough fits into a hole in the side of the waveguide and the latching wire is brought out through the center of the feedthrough and is soldered to the hat. The outer diameter is chosen such that a very low impedance appears at the wire exit hole. The outside wall of the waveguide acts as one conductor of the radial line while the hat of the feedthrough acts as the other. The feedthrough should be made as thin as possible so that the characteristic impedance of the radial line at $r=a$ is small, making the normalized impedance very high at that point. The size of the mismatch as transformed to the input of the radial line determines the isolation achieved. The characteristic impedance of the radial line at $r=a$ is given by

$$Z_0(a) = \frac{b}{2\pi a}\sqrt{\frac{\mu}{\epsilon}} \quad (1)$$

where

μ permeability of radial line material,
ϵ permittivity of radial line material,

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