Letters

A Waveguide to Suspended Stripline Transition

B. GLANCE AND R. TRAMBARULO

Abstract—**A launcher for coupling a 50-Ω suspended stripline to a waveguide by means of a probe has been built and tested at 30 GHz. Matching has been obtained between 28.8 and 32.6 GHz with a return loss better than 35 dB. The launcher can be fabricated as an integral part of the stripline components.**

Millimeter stripline sources, circulators, multipliers, and downconverters are being studied for future millimeter integrated radio systems. At present these stripline components must be tested with waveguide equipment, and a well-matched transition from waveguide to stripline is a necessity.

Two main types of transition from waveguide to standard microstrip line have been reported [1], [2]. One uses a launcher which couples the stripline to the waveguide by means of a probe. A transition of this kind, designed for *C* band, employs an alumina substrate to support the microstrip line [1]. The other transition uses a stepped ridgeline transformer to couple the waveguide to a standard line supported on a silica substrate. An electrical contact is required in this transition between the ridgeline transformer and the microstrip line. A transition of this latter kind has been operated at millimeter wavelengths [2].

This letter reports a 30-GHz probe-type launcher which couples a WR-28 waveguide to a 50-Ω suspended stripline on a silica substrate. A probe was selected rather than a stepped ridgeline transition because the former can be fabricated as an integral part of the stripline and the difficulty of making reliable electrical contacts is avoided.

The line configuration and its dimensions are shown in Fig. 1. The transition circuit which has been used to optimize the 30-GHz launcher consists of a 1.050-in long 50-Ω suspended stripline coupled with similar probes at each end to two WR-28 waveguides. The return loss was measured at the input waveguide with the output waveguide terminated by a matched load. The transition circuit has been built with very small tolerances in order to define precisely the launcher dimensions. The silica substrate is cut to match exactly the distance between the opposite walls of the two waveguides. The length of the probe extending into the waveguide was measured to ±0.0001 in with reference to the end of the substrate. The tolerance of the other circuit dimensions is ±0.0005 in.

The simplest transition consists of a section of 50-Ω stripline which extends into the waveguide. The optimum length of probe is 0.0583 in, which gives 22-dB return loss. The matching is improved if the line is loaded capacitively at discrete positions. The widest bandwidth is obtained by loading the line at the position nearest to the launcher. The configuration for the transition is shown in Fig. 2, and the return loss measured for two transitions in tandem is shown in Fig. 3. It is difficult to assign a lower limit for the return loss of each individual transition because of the possibility of destructive interference between reflections on the pair of transitions. However, when attenuation was inserted in the section between the transitions, the return loss was affected very little over the region where the match was good (>30-dB return loss). From these results we believe that the return loss for a single transition is better than 35 dB between 28.8 and 32.6 GHz.

Transmission losses measured from the input waveguide to the output waveguide are about 0.4 dB. They include: 1) loss in the 2 in of brass waveguide, 2) loss in the two transition regions, 3) loss in the waveguide shorts, and 4) loss in the 50-Ω suspended stripline 1.050 in long. The waveguide loss calculated at 30 GHz is 0.07 dB, and the loss calculated for the stripline connecting the transitions is 0.14 dB. Thus the loss for each transition is about 0.1 dB. The loss for each waveguide short was found to be 0.05 dB greater than for a good fixed short, and therefore transitions with losses less than 0.1 dB may be possible.

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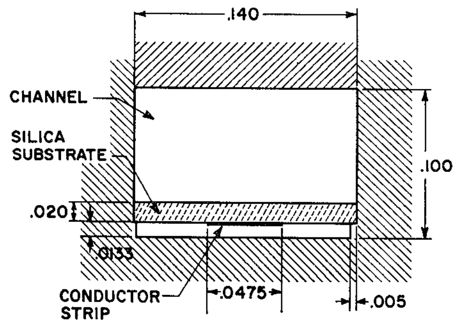

Fig. 1. Cross section and dimensions of a 50-Ω suspended stripline for use at 30 GHz. (Dimensions are in inches.)

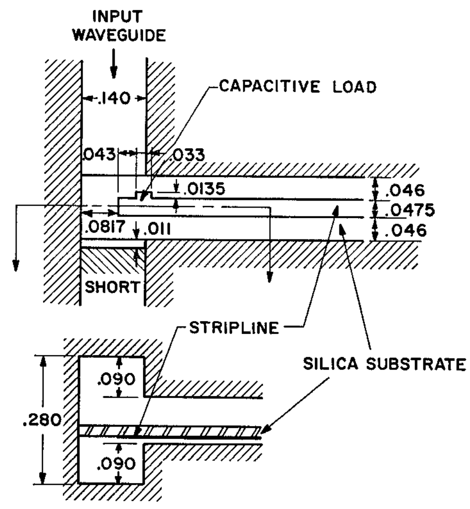

Fig. 2. Dimensions (in inches) of the transition.

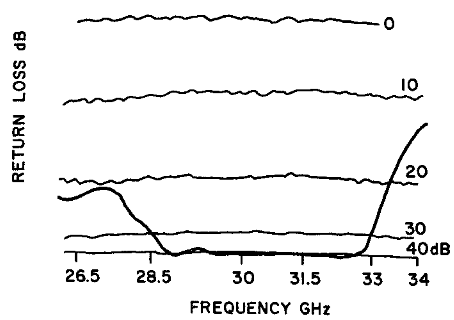

Fig. 3. Return loss versus frequency for two transitions in tandem.

The transitions previously described are made with a silica substrate in contact with the opposite walls of the two waveguides. This arrangement insures the high precision required for the two symmetrical launchers. In some applications, it is necessary to provide a gap between the substrate and the waveguide wall. The return loss has been measured for increasing gaps from zero to 0.0054 in with the waveguide short position adjusted for optimum return loss, which is essentially the same as in Fig. 3. The optimum return loss is maintained if the distance from the short to the substrate shown in Fig. 2 is increased by the size of the gap. All other dimensions in Fig. 2 are unchanged.

A transition that couples a 50-Ω suspended stripline to a WR-28 waveguide by means of a probe has been designed for the 30-GHz band. The launcher is simple and can be an integral part of stripline circuits. The return loss is larger than 30 dB over the frequency interval from 28.5 to 33 GHz. The insertion loss of a transition over this band is 0.1 dB.

References

[1] R. H. Knerr, "A new type of waveguide-to-stripline transition," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-16, pp. 192–194, Mar. 1968.

[2] M. V. Schneider, B. Glance, and W. F. Bodtmann, "Microwave and millimeter wave hybrid integrated circuits for radio systems," *Bell. Syst. Tech. J.*, vol. 48, July–Aug. 1969.

On the Definitions of Parameters in Ferrite–Electromagnetic Wave Interactions

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***Abstract*—Conflicting definitions of the parameters characterizing the material in ferrite–electromagnetic wave interactions occur in the current literature, as well as do actual sign errors that are difficult to detect because of the variety of signs appended to the relevant quantities by different authors. A harmonious set of definitions are set out and their consistency illustrated using the case of a uniform circularly polarized plane wave.**

This letter is prompted by the fact that in the literature dealing with problems in the interaction of electromagnetic waves with ferrite media, conflicting definitions for the parameters describing the material are current. Indeed, actual sign errors, which are difficult to detect, occur in the literature.[1] Especially when one author borrows a result from another, great care in checking for consistency of sign is needed. It would be highly desirable to have a self-consistent set of equations on record in which negative signs are removed from quantities, such as frequencies, which are normally taken as positive.

Consider the system of an electron with its spin axis displaced from the direction of a superimposed dc magnetizing field $\boldsymbol{H}_0$ (see Fig. 1). The equation of motion for a conglomerate of such elementary systems becomes, in this MKS system of units

$$\frac{d\boldsymbol{M}}{dt} = -\gamma'\boldsymbol{M} \times \boldsymbol{B} = -\gamma\boldsymbol{M} \times \boldsymbol{H} \tag{1}$$

where $\boldsymbol{M}$ is the total magnetic moment per unit volume, $\boldsymbol{H}$ the total magnetic field intensity, and the gyromagnetic ratio is given by

$$\gamma = \frac{g\,|e|\mu_0}{2m} \tag{2}$$

(when nuclear spin or a positron is considered, then the sign of the right-hand side of the equation of motion (1) will change accordingly. As here defined γ is an essentially positive quantity).

Taking

$$\boldsymbol{M} = \boldsymbol{M}_0 + \boldsymbol{m}e^{j\omega t}, \qquad \boldsymbol{M}_0 = M_0\boldsymbol{a}_z \tag{3}$$

$$\boldsymbol{H} = \boldsymbol{H}_0 + \boldsymbol{h}e^{j\omega t}, \qquad \boldsymbol{H}_0 = H_0\boldsymbol{a}_z \tag{4}$$

where $\boldsymbol{M}_0$ and $\boldsymbol{H}_0$ are positive quantities, and impressing a circularly polarized electromagnetic wave propagating in the direction of the dc magnetizing field of the form

$$\boldsymbol{e}^{\pm} = (\boldsymbol{a}_x \mp j\boldsymbol{a}_y)e^{j(\omega t - \beta z)} \tag{5}$$

$$\boldsymbol{h}^{\pm} = \frac{\omega\epsilon}{\beta}(\pm j\boldsymbol{a}_x + \boldsymbol{a}_y)e^{j(\omega t - \beta z)} \tag{6}$$

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[1] As an example, the classic work *Microwave Ferrites and Ferrimagnetics* by Lax and Button contains a sign error in their equation (4.5) (seemingly based on a confusion of sign of the gyromagnetic ratio which is positive in their use throughout). As a consequence all their subsequent equations in chapter 4, though self-consistent, are in error in the same way, i.e., resonance occurs for *left*-handed circular polarization, the same as the precession. But in chapter 7, equation (7.9) gives $h_x = +jh_y$ for the right-handed circular polarization, but the plus sign does *not* go with the plus in $\Gamma_{\pm} = (-\omega^2/c^2)K_e(\mu \pm \kappa)$ of eq. (7.10) if the quantities of chapter 4 are used; though subsequent usage, e.g., fig. 7.2 and accompanying equations, rather implies that the correspondence was intended. These errors are difficult to disentangle because, for the most part, they refer merely to one direction of circular polarization or its opposite, rather than to right- and left-handed. Normally, one uses $\pm$ in this latter sense.

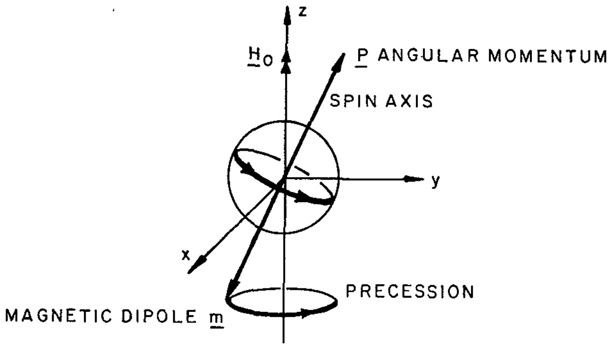

Fig. 1. Note that the letters with underbars in the figure appear bold face in the text.

the following result is obtained

$$(\beta^{\pm})^2 = \omega^2\mu_0\epsilon\left(1 + \frac{\omega_m}{\omega_0 \mp \omega}\right) \tag{7}$$

upon employing (1) and the Maxwellian equation

$$\nabla \times \boldsymbol{e} = -j\omega\mu_0(\boldsymbol{h} + \boldsymbol{m}). \tag{8}$$

The superscripts refer to the sense of the circular polarization with the plus sign indicating right-hand polarization. The latter is here defined as being characterized by a clockwise rotation of the field vectors when viewed in the *direction of propagation*. The radian frequencies are defined by

$$\omega_0 = \gamma H_0 \qquad \omega_m = \gamma M_0 \tag{9}$$

and are positive quantities.

The susceptibility matrix for the material takes the form, which is the commonest, though not a universal choice,

$$\bar{\boldsymbol{\mu}} = \mu_0\begin{bmatrix} 1+\chi & -j\kappa & 0 \\ +j\kappa & 1+\chi & 0 \\ 0 & 0 & 1 \end{bmatrix} \tag{10}$$

provided that χ and κ are determined by

$$\chi = \frac{\omega_0\omega_m}{\omega_0^2 - \omega^2}$$

$$\kappa = -\frac{\omega\omega_m}{\omega_0^2 - \omega^2}. \tag{11}$$

The propagation factors for the circularly polarized waves are then given by

$$\begin{aligned}(\beta^{\pm})^2 &= \omega^2\mu_0\epsilon(1 + \chi \mp \kappa) \\ &= \omega^2\mu_0\epsilon\left(1 + \frac{\omega_m}{\omega_0 \mp \omega}\right).\end{aligned} \tag{12}$$

Note that the $\mp\kappa$ and $\mp\omega$ occur together.

Thus the above definitions lead to consistent results with the strong interaction occurring for the circularly polarized wave that has the same rotational sense as the precession of the magnetic dipole, as it must. In addition, confusion is minimized in that the signs associated with the effective permeability when expressed in χ and κ carry over when ω_0 and ω_m are used instead.

It should also be noted that when the direction of the dc magnetizing field is reversed the positive nature of ω_0 and ω_m is retained, but the signs of ω_0 and ω_m in (11) and (12) are reversed. Hence the strong interaction will now occur, as it should, for the left-hand polarized wave.

It is perhaps worth emphasizing here that positive or right-handed polarization is defined in relation to the direction of propagation, not the applied field. This is in accordance with conventional use for plane waves, for which an applied field is irrelevant. Defining the sense of circular polarization with respect to the applied field, as is sometimes done, introduces yet a further source of sign confusion to this subject.

The above analysis is based on a time reference $e^{+j\omega t}$. Some users prefer $e^{-j\omega t}$, in which case the sign of j in the preceding results needs to be changed throughout.