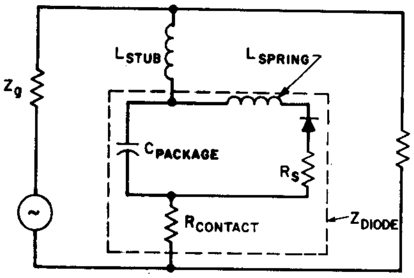

Fig. 1. Lumped element equivalent circuit.

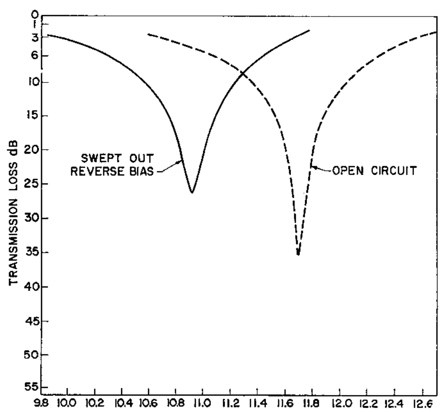

Fig. 2. "Off" transmission characteristic under high reverse bias (solid line) with an "open circuit" reference (dashed line) corresponding to "lossless" diode performance.

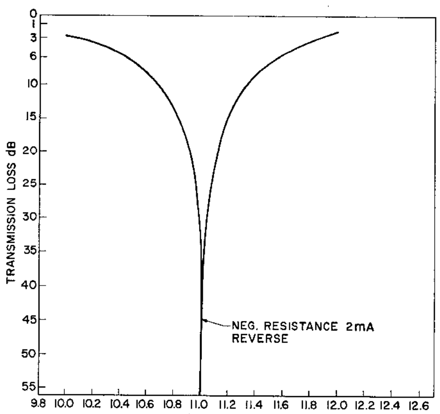

Fig. 3. "Off" transmission characteristic when reverse breakdown voltage is exceeded to the extent that 2 mA of current are flowing.

which can only result from cancellation of the various resistance losses.

An equally dramatic demonstration of the introduction of negative resistance to the circuit under IMPATT operation is to observe the reflection coefficient. At normal 30 to 35 dB transmission loss the return loss is very close to zero but as the diode avalanche current is increased from 2 to 5 mA the return loss passes through zero and reverses sign since more power is contained in the reflected wave than in the incident wave. In both transmission and reflection operations the filter characteristic is very sensitive to the amount of reverse current.

The increase of circuit Q is accompanied by an additional noise over normal passive operation and a reduction in effective bandwidth. All results described were obtained with CW measurements. Equivalent power relationships were also secured under bias pulse operation; however, pulse width reproducibility was reduced. Regular passive operation faithfully reproduced a triangular input bias pulse of 1×10^{-9} s at the base with approximately equal rise and fall times. When the diode was pulsed into IMPATT operation with the same shaped triangular input pulse the reproduced reflected pulse is stretched out to approximately 5×10^{-9} s. The RF input power level was approximately 0 dBm for all described tests. An appreciable increase of input power prevented successful operation.

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Microstrip Transmission-Line Transitions to Dielectric-Filled Waveguide

The highly asymmetrical electric fields of microstrip transmission lines have been found to make an ideal TEM (coaxial) transition to reduced height dielectric-filled waveguide. Such a transition from an OSM connector has been designed and optimized to provide a better than 1.5 to 1 VSWR over a 2.0 to 4.0 GHz frequency band into a slab of metalized dielectric of $K=15$. The transition takes the form of a microstrip transformer cut into the metalized surface of the dielectric-filled guide. Figure 1 shows an experimental transition. In use, the top surface of the transition would be shielded by a shallow cavity.

The problems associated with matching into waveguide can be appreciated by examining the calculated guide impedance as plotted in Fig. 2. Shown is the impedance as based on the ratios at power-voltage (Z_{WV}), power-current (Z_{WI}), and voltage-current (Z_{VI}) as a function of frequency.[1] It can be noted that the impedance depends on the method of calculation and on frequency, which is not the case with a TEM transmission line, where the

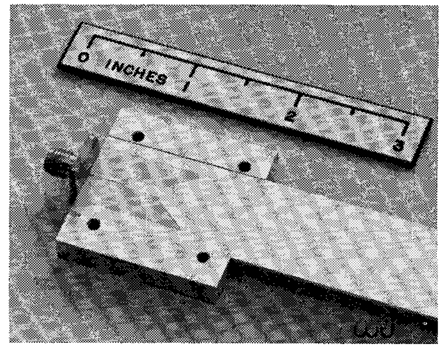

Fig. 1.

impedance would be the same and independent of frequency. Particularly distressing is the rapid change in impedance near the guide cutoff frequency. This rapid change was found to restrict the useful low-frequency limit of the transition to about 20 percent above the guide cutoff frequency. Normally waveguide can be used to about 13 percent above cutoff.

The final design of the transition was reached by empirical modifications using swept frequency techniques. The initial configuration was a two-step quarter-wavelength microstrip transformer designed to transform 50 ohms to the average guide impedance, the power-voltage (Z_{WV}) impedance of the guide being used.[2] The transformer was followed by a quarter-wavelength taper from a width corresponding to a microstrip impedance equal to the average guide impedance[3] (see Fig. 1). The net length of the transition was three quarter-wavelengths long.

Figure 3 is a cutaway drawing of the assembled transition. Adjustment of the shorting bar serves to equalize the VSWR over the frequency band. In particular, movement in the direction of the connector decreases the VSWR at the lower band edge with a corresponding increase at the higher frequencies. Swept data for the optimized transition is shown in Fig. 4. This data is for a 0.060 $\times 0.930 \times 10$ inch guide of $K=15$ dielectric, with a transition at each end.[3] The fine structure of the VSWR is due to the many wavelengths between transitions. The peak VSWR of the whole assembly is slightly greater than 1.5 to 1. The performance of the transition remains good above the TE_{20} cutoff frequency which is at 3.26 GHz for the case shown. Similar transitions were made for guide thickness as small as 0.038 inch. The same technique should be useful in making TEM transitions to guides of higher dielectric constants such as titanium dioxide of $k=96$ or for multioctave frequency bands using the dielectric-filled guide equivalent to ridge guide.

Manuscript received September 30, 1966.

[1] E. L. Ginzton, *Microwave Measurements*. New York: McGraw-Hill, 1957, p. 205, equations 4.7, 4.8, and 4.9

[2] Design curves for linewidths of microstrip transmission line on Al_2O_3 substrates are available. (B. T. Vincent, Jr., "Ceramic microstrip for microwave hybrid integrated circuits," *G-MTT Symp. Digest*, 1966.) In first approximation these curves can be used for other dielectric constants by multiplying the required impedance and dividing the wavelength by the square root of the ratio of the two dielectric constants.

[3] Subsequent Boonton bridge measurements on this dielectric material indicate that the actual dielectric constant was 11.9 and not its nominal value of 15. Thus the actual guide cutoff was 1.83 GHz and not the calculated value of 1.63 GHz based on a dielectric constant of 15. This would indicate that the transition is usable to within 6 percent of cutoff.

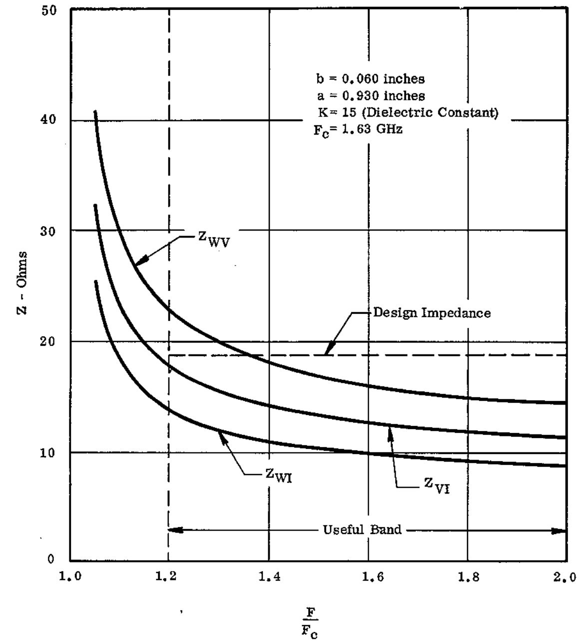

Fig. 2. Calculated waveguide impedance.

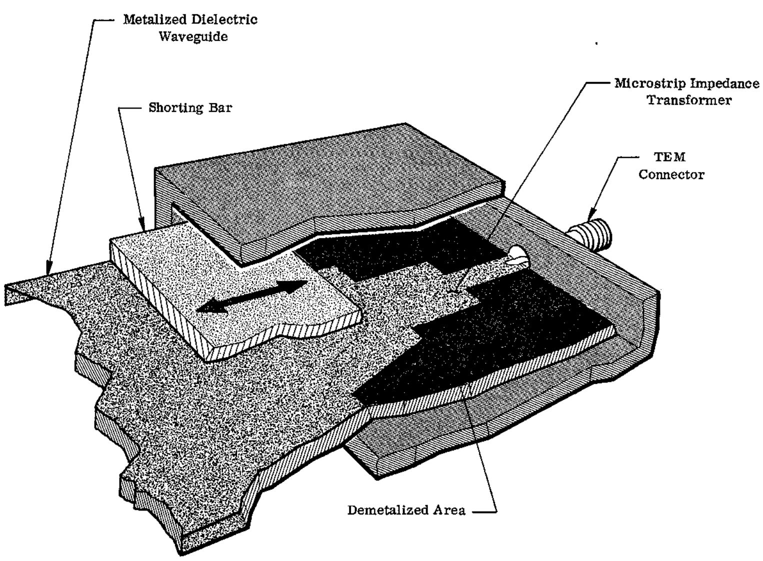

Fig. 3. TEM adapter cross section.

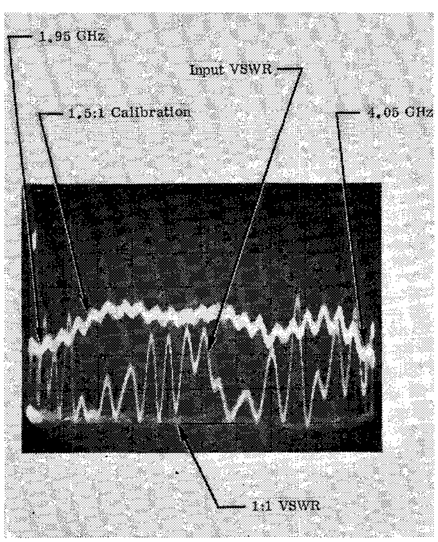

Fig. 4. Swept data input match adjusted adapter.

ACKNOWLEDGMENT

The efforts of G. F. Johnson who did the empirical optimizing of the transition are gratefully acknowledged.

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Guided Waves in Moving Media

In a recent paper,[1] the problem of electromagnetic wave propagation in a waveguide containing a moving dielectric medium was considered. It is shown in this correspondence that the approach used is unnecessarily complicated and that the mode behavior can be simply derived from the stationary problem by a Lorentz transformation to the frame of a moving observer.

The problem considered in Collier and Tai[1] was solved in the following manner: starting from Maxwell's equation and the constitutive equations in the axially-moving medium, the authors were able to derive vector and scalar potentials for both E- and H-type modes, satisfying a modified gauge condition. They were then able to write down a wave-type equation for the vector potentials which could be solved together with the boundary conditions at the waveguide walls. From this, the field components could be written down together with expressions for the wave impedance and axial propagation constant. These latter two were found to differ from their stationary values by a term independent of the waveguide dimensions.

In the present correspondence, it is shown that the problem may be solved in a much simpler way. The important point to note is that the movement of the medium relative to the waveguide is irrelevant to the solution of the problem. Since this movement is parallel to the axial velocity vector, the boundary conditions on the waveguide walls are exactly as in the stationary problem and the harmonic fields remain zero inside the waveguide walls. Physically, then, the problem is identical to that where the dielectric medium and waveguide move together along the z-axis at velocity $\bar{v}$ with respect to an observer, and the solution of this problem is related quite simply to the stationary problem (waveguide, medium, and observer all stationary) through a Lorentz transformation. In the stationary rectangular waveguide, a typical field component (unprimed system) is proportional to

$$\begin{matrix}\sin\\ \cos\end{matrix}\left(\frac{m\pi x}{a}\right)\begin{matrix}\sin\\ \cos\end{matrix}\left(\frac{n\pi y}{b}\right)\cdot \exp\left[i\Gamma_{mn}z-i\omega t\right] \tag{1}$$

where

$$\Gamma_{mn}^2 = \omega^2\mu\epsilon_{\text{eff}} - k_{mn}^2 = k^2 - k_{nm}^2$$

$$k_{mn}^2 = \left(\frac{m\pi}{a}\right)^2 + \left(\frac{n\pi}{b}\right)^2$$

$$\epsilon_{\text{eff}} = \epsilon\left(1 + \frac{i\sigma}{\omega\epsilon}\right).$$

Manuscript received October 20, 1966.

[1] J. R. Collier, and C. T. Tai, *IEEE Trans. on Microwave Theory and Techniques*, vol. MTT-13, pp. 441–445, July 1965.