A New Model for Broadband Waveguide-to-Microstrip Transition Design

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A new model that permits the prediction of the resonance frequencies created by antipodal finline waveguide-to-microstrip transitions is presented. The transition is modeled as a tapered transmission line in series with an infinite set of coupled resonance circuits. The resonant circuits are modeled as simple microwave resonant cavities of which the resonance frequencies are easily determined. The model is developed and the resonance frequencies determined for several different transitions. Experimental results are given to confirm the models.

Introduction

The use of frequencies above 18 GHz for government and commercial communications applications is expected to increase rapidly in the next few years. Low cost systems in this frequency range will utilize solid-state devices in the form of monolithic microwave integrated circuits (MMICs). The current trend in solid-state microwave integrated circuits is to use microstrip transmission lines for device interconnections. However, all test equipment in the mm-wave frequency range utilizes rectangular waveguide. In addition, rectangular waveguide will be used in phased array antennas and other systems in which the signal must travel an appreciable distance in a medium other than free space. Therefore, a low cost, easily fabricated, low loss and broad bandwidth waveguide-to-microstrip transition is essential in the test and implementation of any system utilizing MMICs.

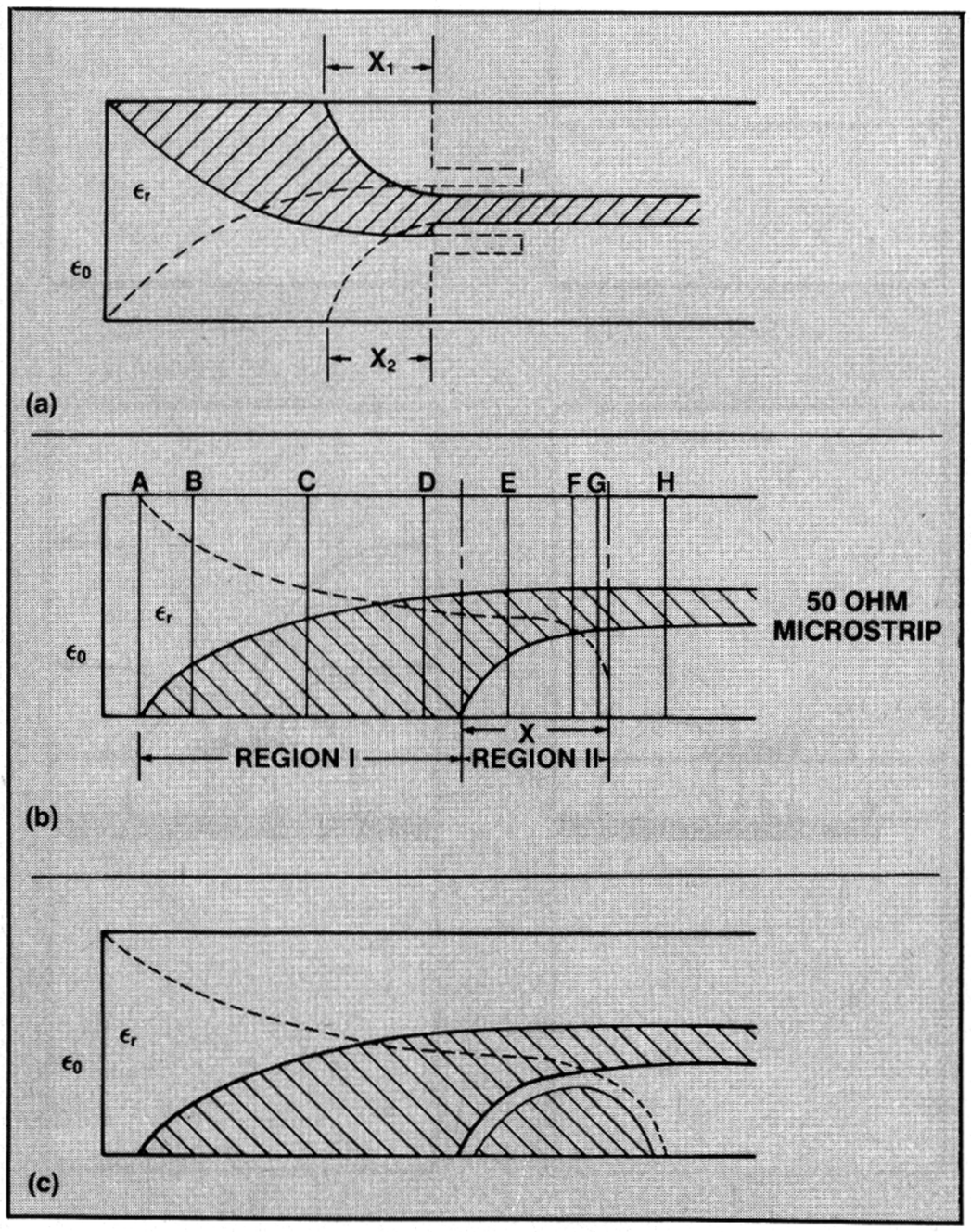

Fig. 1 (a) Van Heuven transition; (b) simplified van Heuven transition; and (c) simplified Van Heuven transition with semicircular fin.

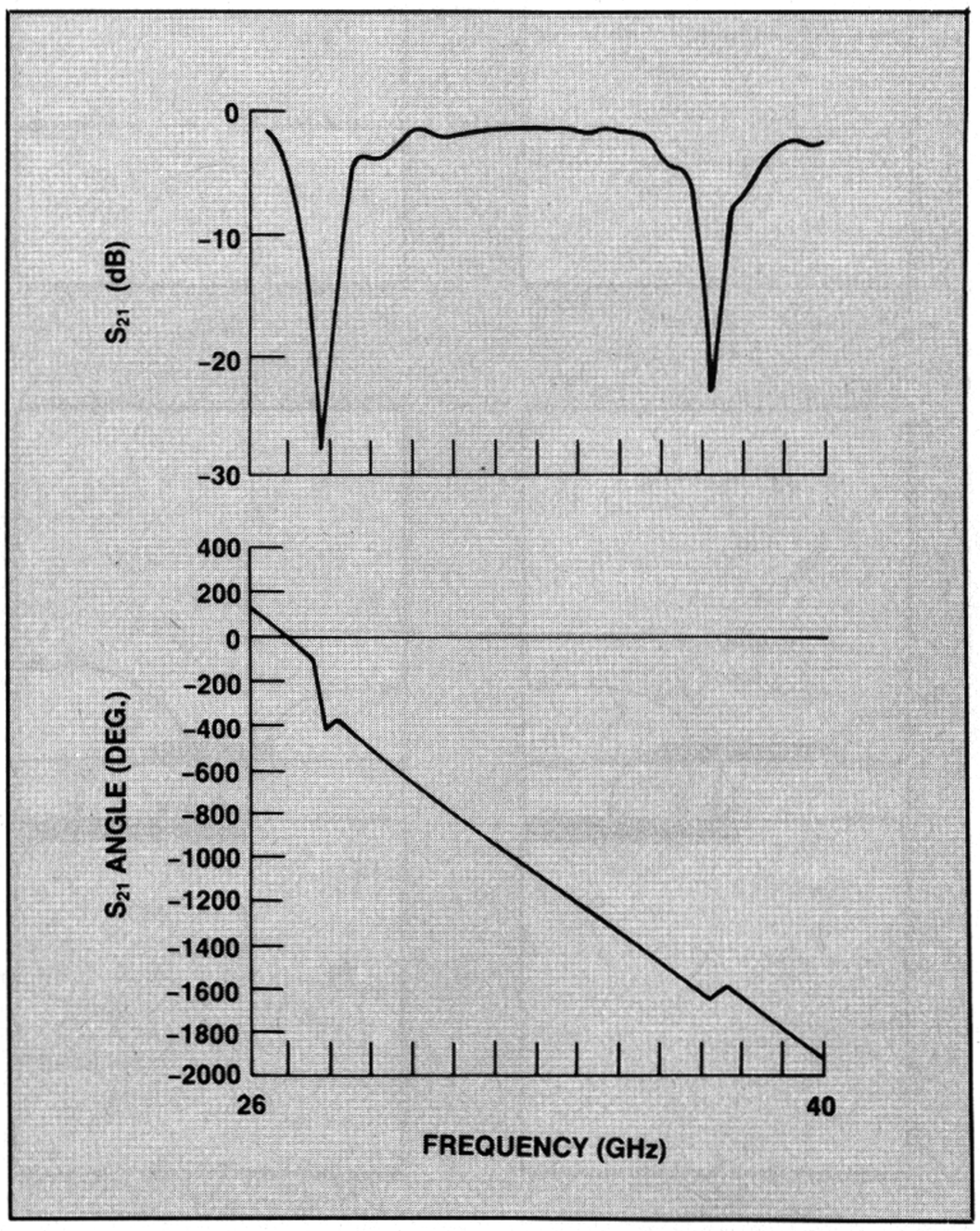

Fig. 2 S_{21} measurements of antipodal finline waveguide-to-microstrip transition.

The antipodal finline waveguide-to-microstrip transition was first demonstrated by J.H.C. van Heuven.[1] This and the many derivations of it seen in the literature[2-4] (Figure 1) are attractive from the system integration point of view since they are in line with the waveguide and can be manufactured easily on inexpensive, soft substrates using standard printed circuit board techniques. The problem with using antipodal finline transitions has been obtaining broadband response from a simple set of design rules. The design is complicated since the transition creates a set of resonant modes that limit the useful bandwidth of the transition. A typical S_{21} measurement of a transition demonstrating this phenomenon is shown in Figure 2.

The resonances are a major problem in transition design and may exclude this type of transition from being used under certain circumstances. In a standard configuration, the component under test is inserted between two transitions with a short section of microstrip separating the device from the transition to minimize loss. The resonances of the transition may couple to the device under test if the bandwidth of the device overlaps the resonant frequencies.

This may be observed by placing two different transitions back to back and measuring the resonance frequency. It will be seen that the resonance frequency is between the resonance frequencies of each individual transition. Therefore, the use of antipodal finline transitions of the type shown in Figure 1 may be limited for standard applications such as microstrip transmission line resonator studies and filters where the inserted device has a resonance frequency of its own. Applying transitions in phased array antenna systems may be limited by the associated phase shift at each resonance. In the passband portion of the transition, the phase shift is linear, but near a resonance a rapid phase shift that is difficult to predict occurs.

Existing design rules[5] do not address the problem of placement of the resonance frequencies. There-

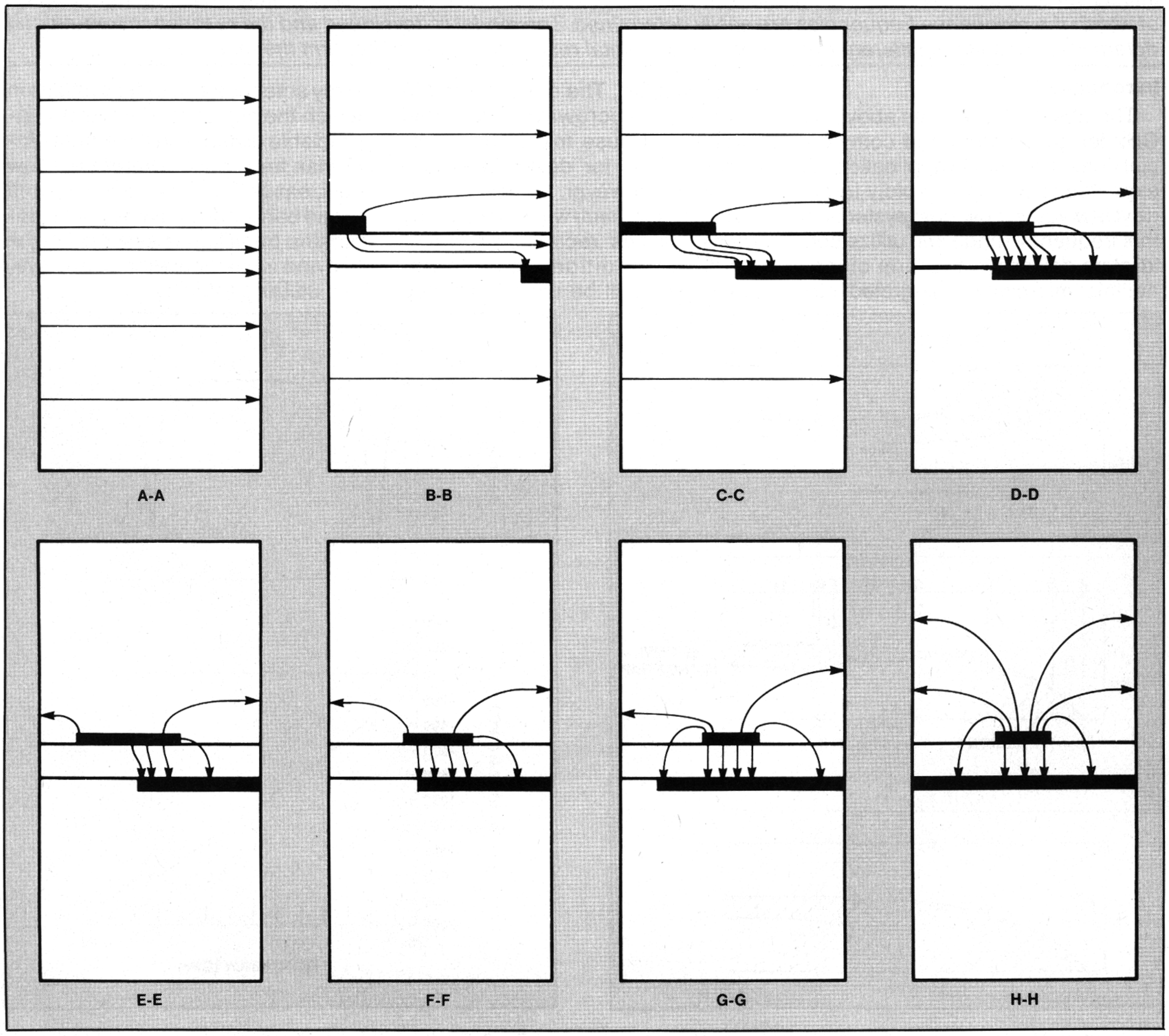

Fig. 3 Electric fields at various cross-sections along the transition from conventional rectangular waveguide to a 50 ohm microstrip, as shown in Figure 1b.

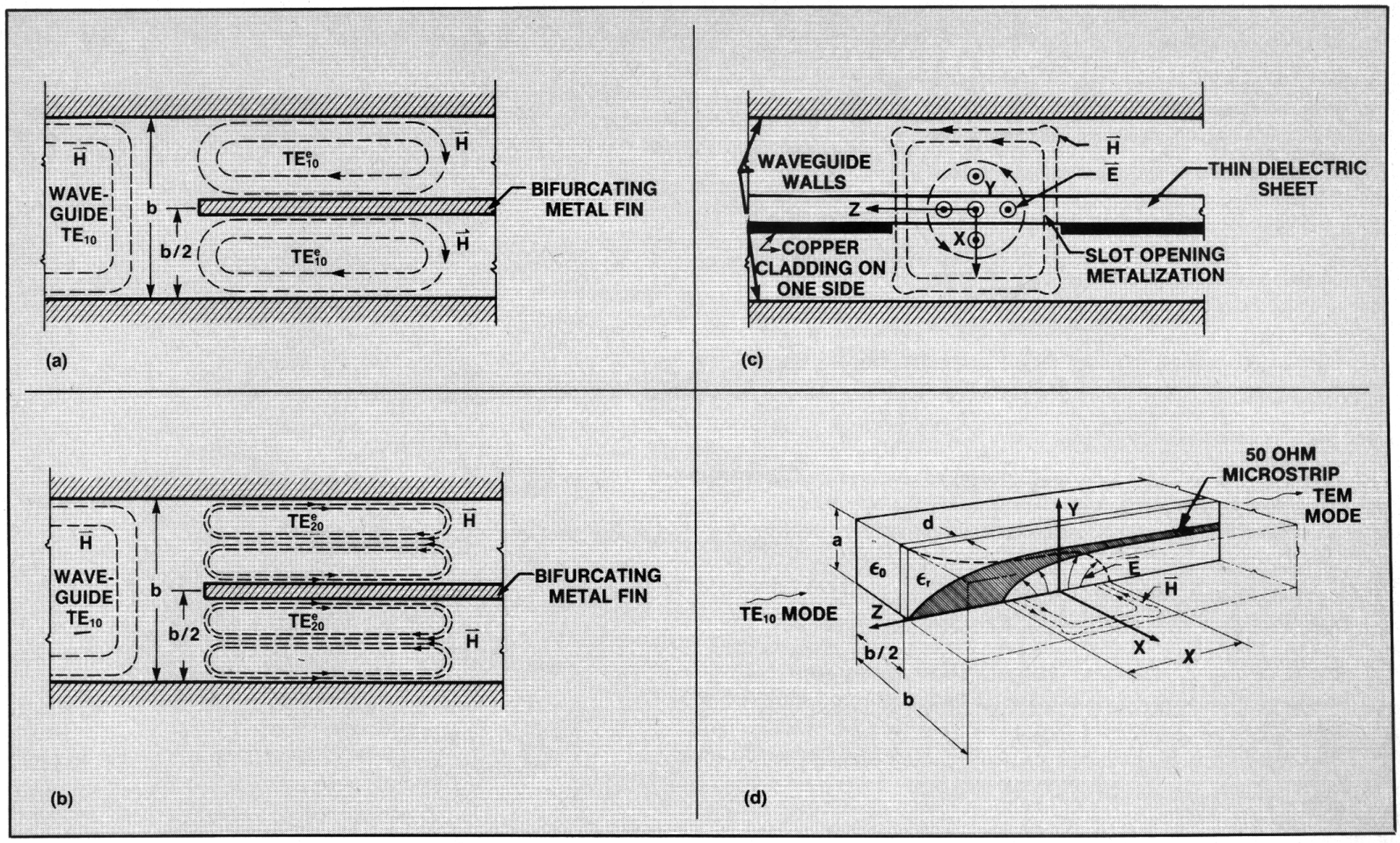

Fig. 4 $\vec{E}$ and $\vec{H}$ fields of region II: (a) TE_{10} evanescent mode; (b) TE_{20} evanescent mode; (c) resonant EM fields in the vicinity of a printed slot on a bifurcating metal wall; and (d) decay of electric fields to zero at waveguide walls.

fore, trial and error design still is used. We present a new model for predicting the resonances; it permits placement of the resonance frequencies outside the desired band during the design process. This eliminates the need for much of the iterative design process and therefore reduces the time and cost involved in implementing waveguide-to-microstrip transitions.

Physical Description of the Resonance

The transition shown in Figure 1b is the easiest to describe and is therefore used to develop a physical description of the resonances. The transition has been separated into two regions in order to describe

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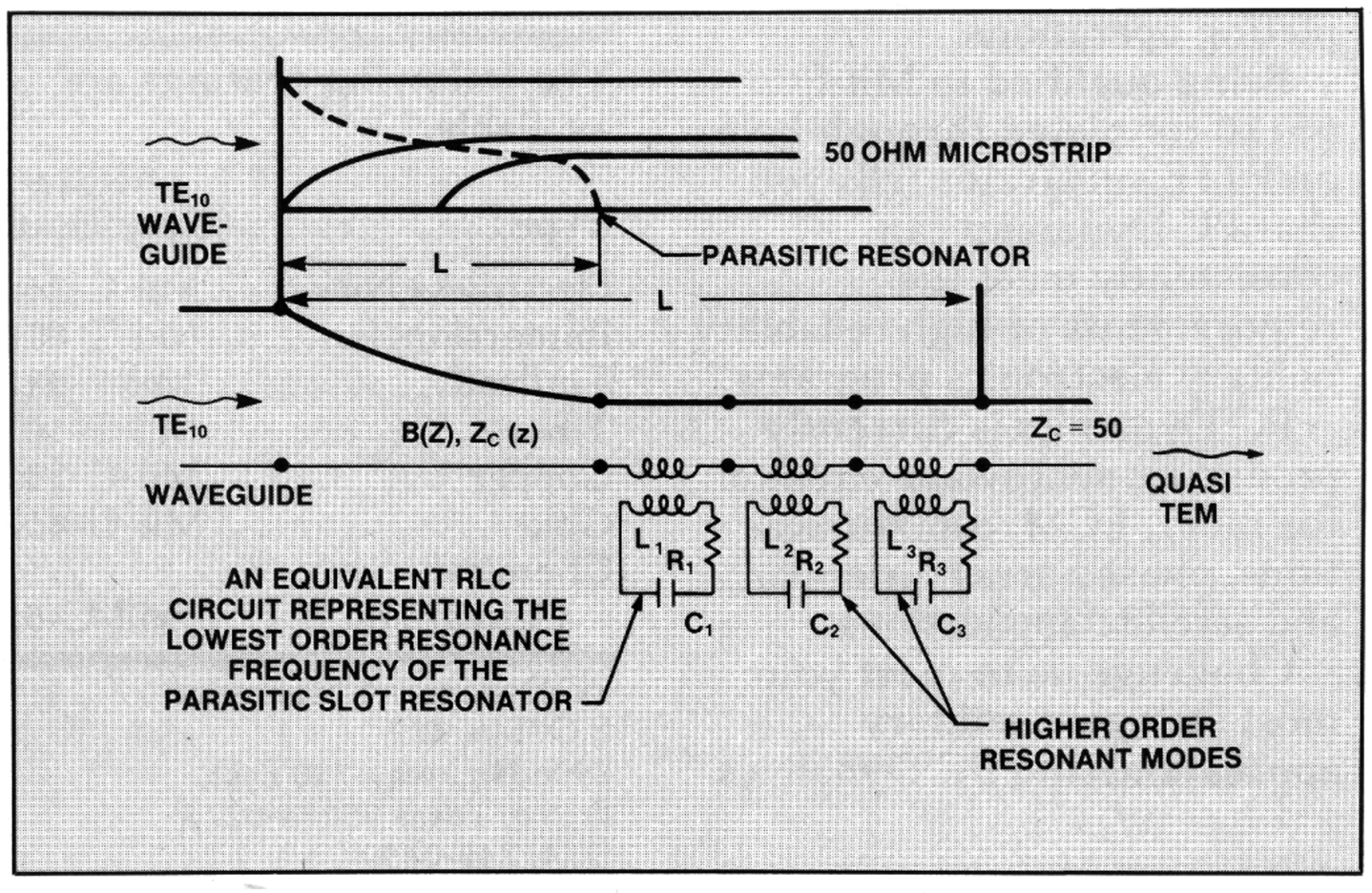
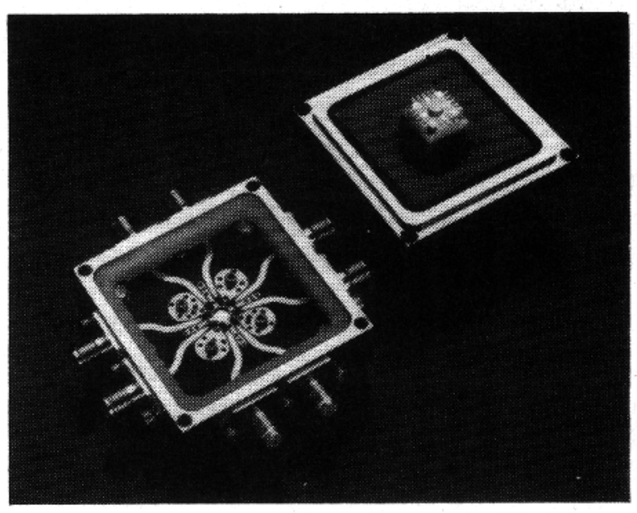


Fig. 5 An equivalent lumped-element model for a waveguide-to-microstrip transition.



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how the transition works. Region I is a tapered antipodal finline which concentrates and rotates the electric fields of the incident TE_{10} waveguide mode 90° into the quasi-microstrip mode that propagates in antipodal finlines with overlapping fins (Figure 3, sections AA-DD). In addition, region I transforms the high impedance of the TE_{10} waveguide mode to a lower impedance close to the 50 ohm impedance usually used for microstrip transmission lines. Region II makes the transition from antipodal finline to microstrip (Figure 3, sections EE-HH). The resonances are created in region II.

To develop the physical picture of the electromagnetic fields associated with the resonances that will be used to derive the model, consider the transition to look like an E-plane filter to a propagating TE_{10} mode signal in rectangular waveguide. If a solid metal sheet is inserted into a waveguide in the center of the H-plane and parallel to the E-plane of the waveguide, the TE_{10} waveguide mode is cut off and evanescent modes are created.

Figures 4a and 4b show the two lowest-order evanescent modes in the bisected waveguide region, labeled the TE^e_{10} and TE^e_{20} modes.[6] Inductive energy is stored in the evanescent modes. If a slot is placed in the bisecting metal sheet, electric fields will be created in the slot parallel to the E-plane of the waveguide. When the capacitive energy created by the slot equals the inductive energy stored in the evanescent modes, a resonance will occur. Figure 4c shows the electric and magnetic fields that are created when a slot is added to the bisecting metal sheet. The waveguide-to-finline interface and the finline discontinuities create a similar set of evanescent modes. The inductive energy stored in the evanescent modes is one component of the loss in prop-

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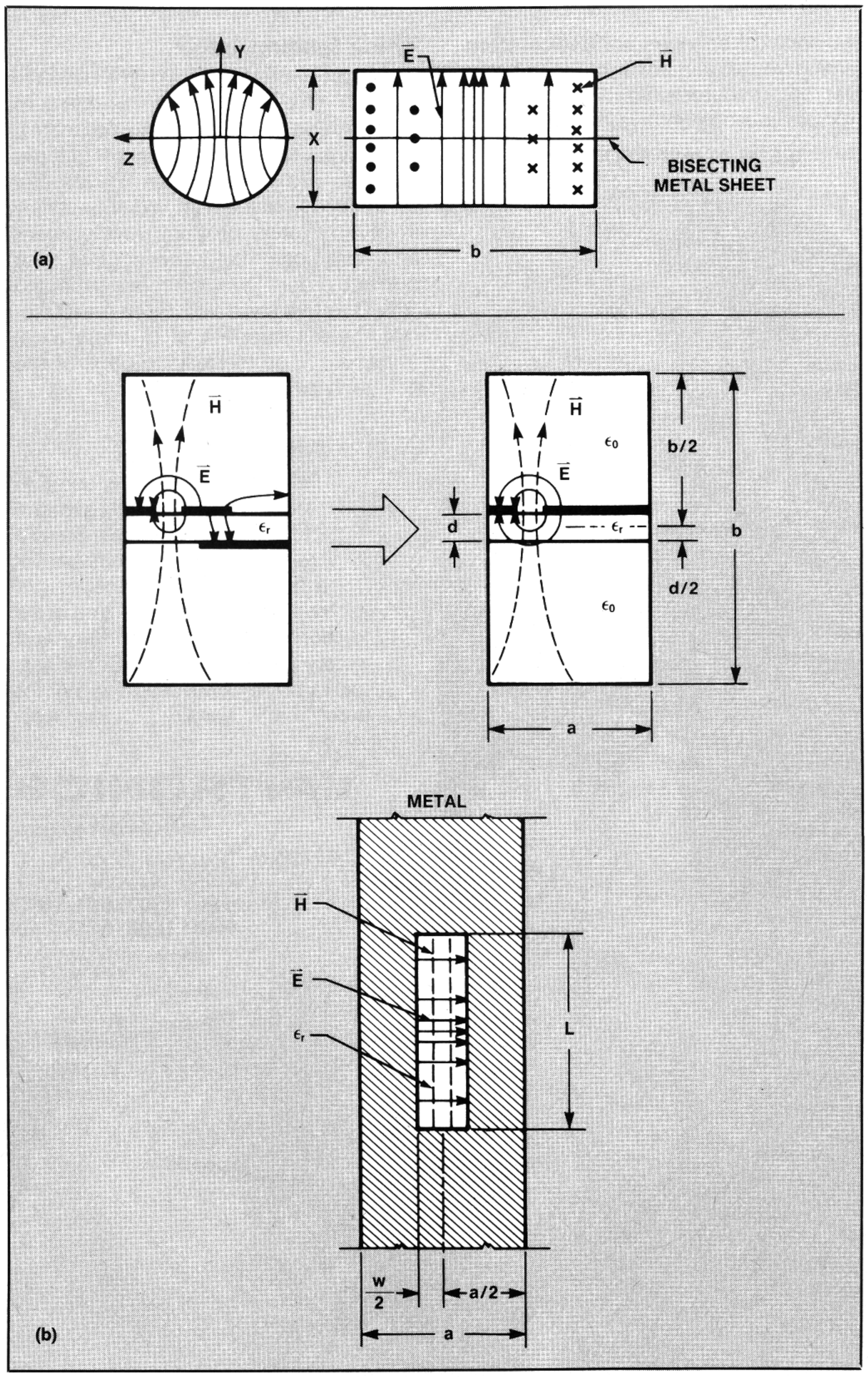
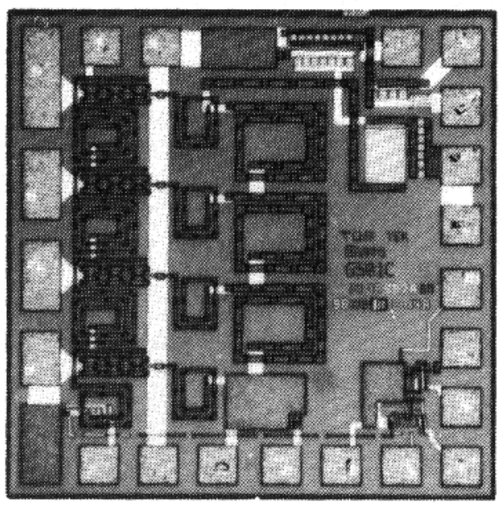


Fig. 6 Equivalent resonant cavities for transitions shown in Figures 1b and 1c: (a) TE_{111} mode of an equivalent cylindrical waveguide cavity, homogeneously filled with dielectric material of ϵ_{eff}; and (b) first order, resonant mode of an equivalent slot resonator.



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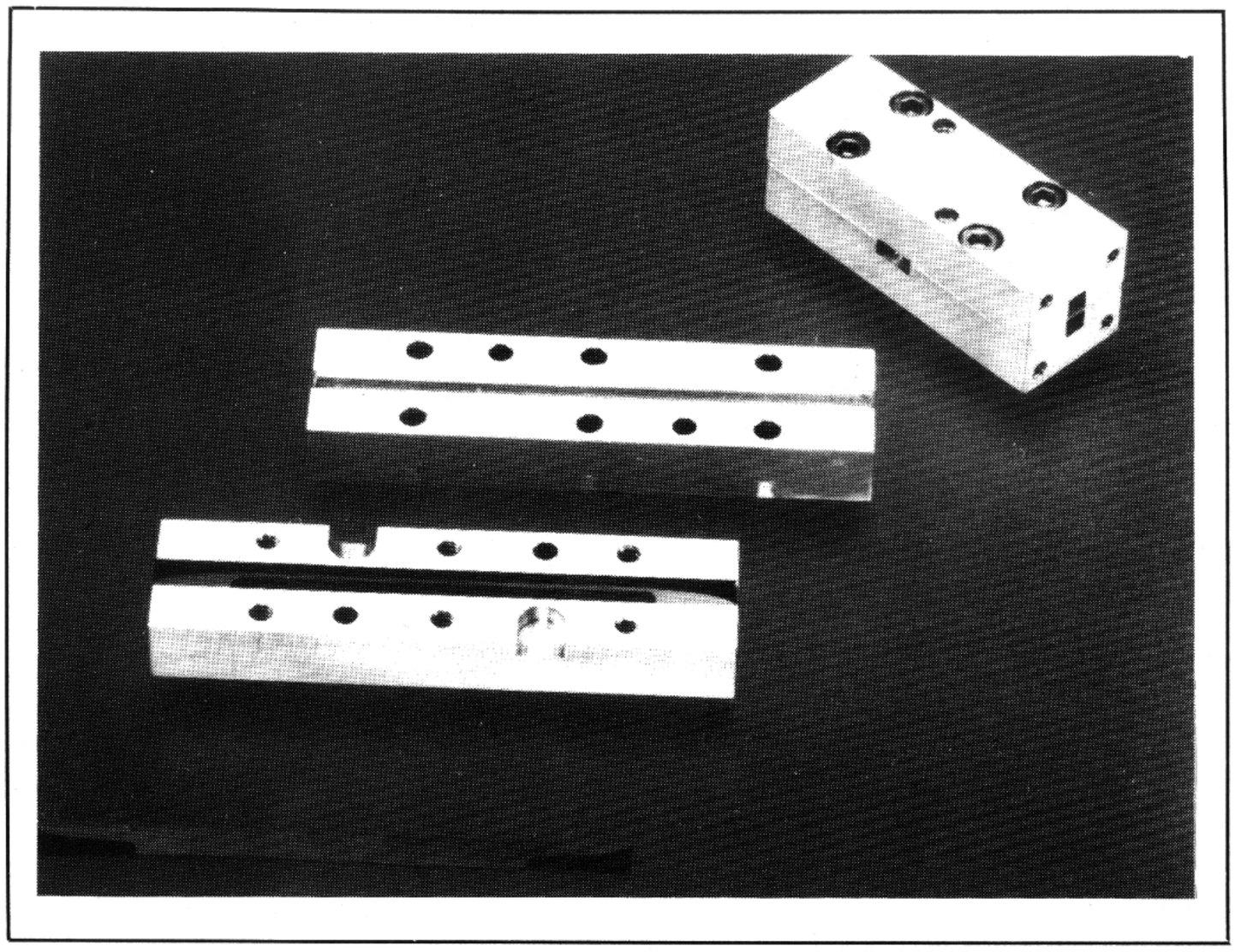

Fig. 7 26.5 to 40 GHz waveguide-to-microstrip transition in test fixture.

agating power associated with the transitions. Region II appears to the evanescent modes as a slot in a bisecting metal sheet and a source of capacitive energy. Therefore, region II is the source of the resonances that plague the transition.

Model Development

Analytical methods to obtain the resonance frequencies using the above description would be difficult to implement for a complicated structure such as the transition. Therefore, a model based on the electromagnetic field diagrams derived in figure 4 and experimentally derived parameters is developed. In the tapered antipodal finline region, a quasi-TE_{10} mode propagates. In region II and the microstrip, a quasi-TEM mode propagates. The evanescent modes decay rapidly and therefore do not propagate. Therefore, the quasi-TEM and the TE_{10} modes must propagate independently of the evanescent modes for all frequencies except those near the resonance frequency, F_r, of region II.

Near F_r the propagating wave couples energy to the resonator in the same manner as any transmission line would couple energy to a reactive resonant cavity placed within the electromagnetic fields of the propagating wave. Therefore, if the resonance frequency could be modeled by a cavity resonator of a form related to the shape of region II, an equivalent circuit model could be used such as the one shown in Figure 5. Each RLC circuit represents a mode of an equivalent resonant cavity. If we assume that the Q of each resonator is large and the resonant modes are not close together, then each resonant mode can be considered to be decoupled, and placement of the resonance frequencies may be made by a simple alteration to region II.

The choice of the resonant cavity to model region II must be dependent on the electric and magnetic fields of region II. The primary mode in region II is an LSE mode. The electric fields are concentrated in the slot, or region II, in the E-plane of the waveguide. Outside of the center of the H-plane, the electric fields decay to zero at the waveguide walls, $x = \pm b/2$ of Figure 4d. The magnetic field lines must encircle the electric field lines and are constrained by the waveguide walls and the metal sheet that forms region II.

The shape of region II and the LSE-mode fields in the region suggest a cylindrical waveguide cavity with a bisecting metal sheet in the E-plane. Resonant modes of the form TE_{n11}, $n \geq 1$, have fields which are similar to those of region II. In addition, unlike an elliptical waveguide resonant cavity, which would more closely describe region II, the resonance frequencies of cylindrical cavities are obtained easily. Therefore, a cylindrical cavity filled with a material of some ϵ_{eff} and having a radius of X/2 and a length b is proposed to model region II (see Figure 6a).

The resonance frequencies are dependent on the available cavity volume, so an effective permittivity is introduced to make the resonance frequency of the cylindrical cavity coincide with the experimental values. The proposed model is simple and its usefulness depends on whether or not a single value of ϵ_{eff} can be found such that various resonance frequencies can be predicted accurately. It will be shown that this is the case for many transitions that have been studied experimentally. We found that ϵ_{eff} is dependent on the length parameter X but was essentially independent of frequency. Thus, for a given transition, the various resonance frequencies can be predicted accurately from the model.

The resonance frequencies are derived by setting the propagation constant for cylindrical waveguide

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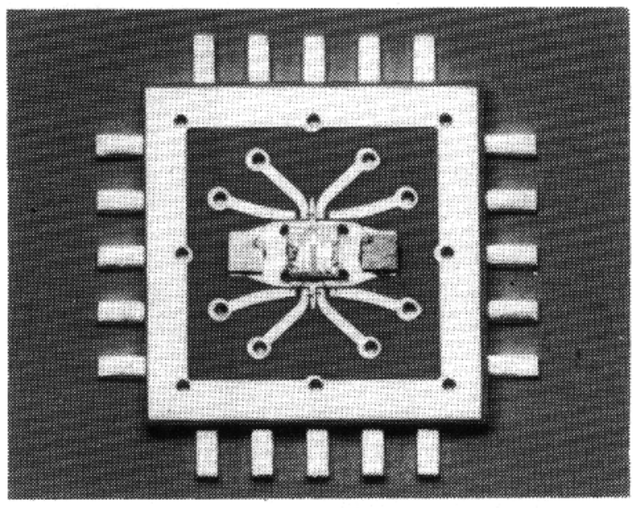

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equal to π/b, where b is the length of the waveguide cavity. The resulting equation for the resonance frequencies is:

$$F_{n11} = \frac{C}{2\pi\sqrt{\epsilon_{eff}}} \times \left[\left(\frac{P_{n1}}{X/2}\right)^2 + \left(\frac{\pi}{b}\right)^2\right]^{1/2} \quad (1)$$

where X/2 is the radius of the cylinder, P_{n1} are roots of the equation dJn(kcr)/dr = 0 (r =X/2 and kc = cutoff wave number of the waveguide), and ϵ_{eff} is some frequency-dependent effective dielectric constant.[7] The solution for ϵ_{eff} would permit the determination of the resonance frequency. However, a theoretical determination of ϵ_{eff} is very difficult. Therefore, an experimentally derived ϵ_{eff} was found.

Experimental Results

Transitions were fabricated for various lengths (X) of region II. In general, the total length of the transition (L = length of region I plus X) was kept constant. The transition length, L, was equal to λ_g (F = 32 GHz) of the TE_{10} rectangular waveguide mode for Ka-band transitions. The transitions were fabricated on 10 mil, ϵ_r = 2.22 RT/duroid* and 10 mil, ϵ_r = 2.17 Cu Clad** substrates. All testing was done for two back-to-back transitions interconnected by 3.80 cm of 50 ohm microstripline. The transitions were mounted for testing in a test fixture (Figure 7) that when clamped together supplied sufficient contact between the metal finlines and the waveguide walls. Measurements were taken on a modified Hewlett-Packard 8409 automatic network analyzer.

The measured resonance frequencies, F_r, are plotted versus X for two differently shaped region IIs in Figure 8. In addition, transitions were made with straight finline tapers and/or a triangular shaped region II. We found that the finline taper made only small differences in the resonance frequency. Also, as

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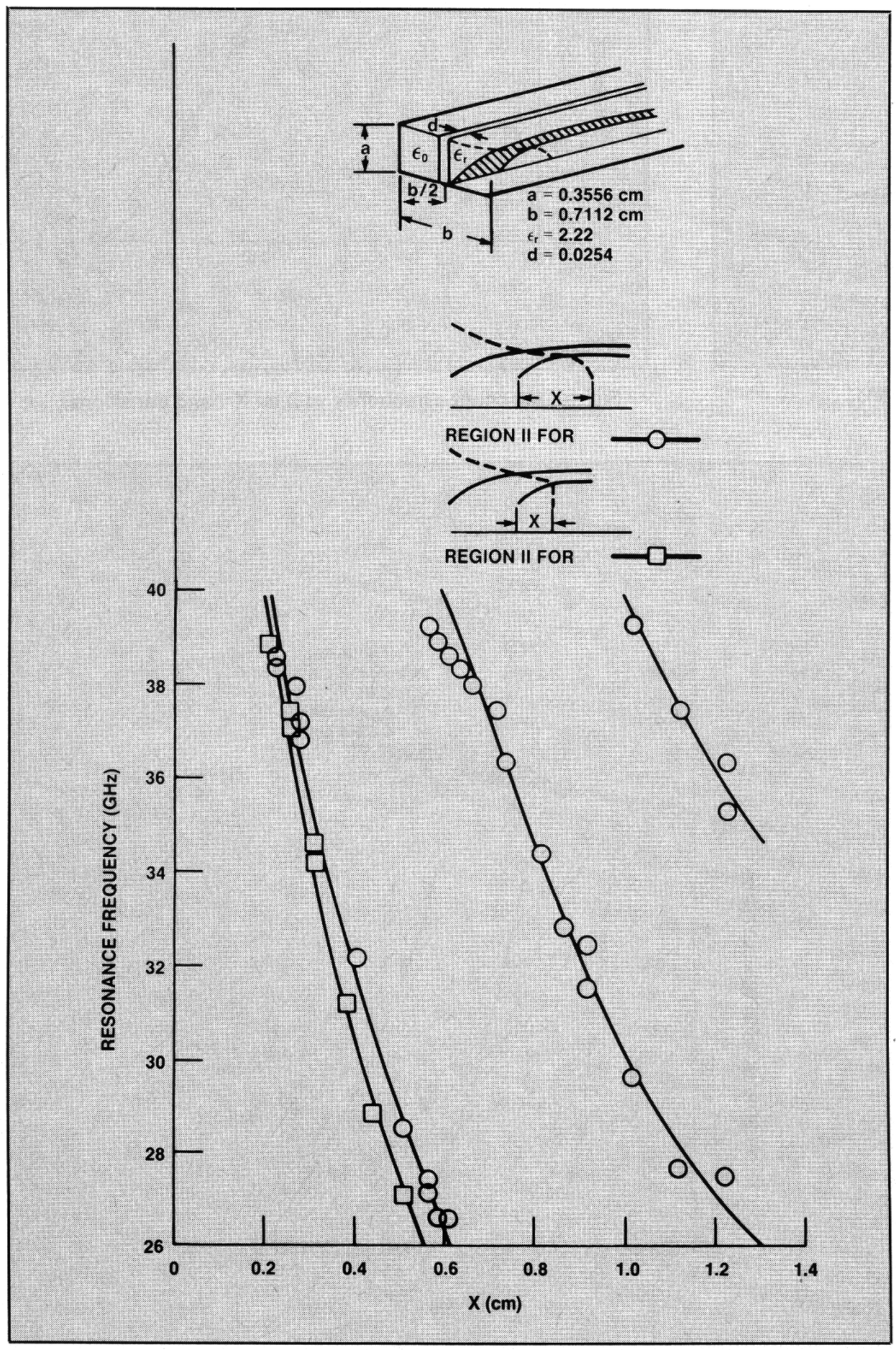
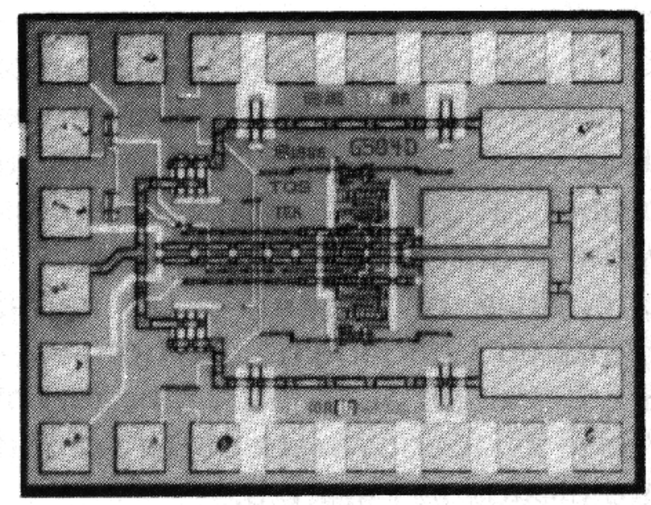


Fig. 8 Resonance frequency as a function of X for Ka-band transitions.



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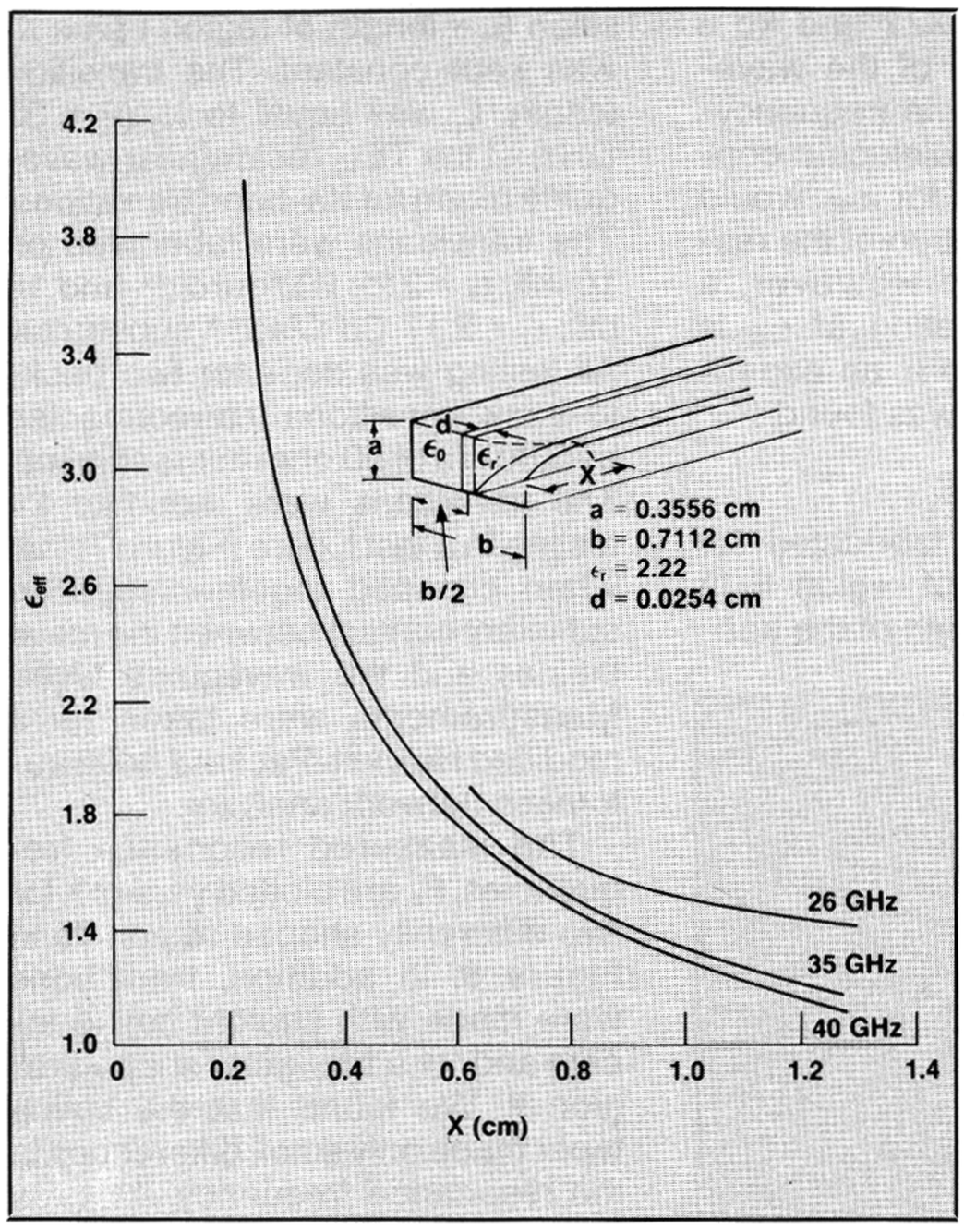

Fig. 9 ϵ_{eff} vs X for Ka-band transitions.

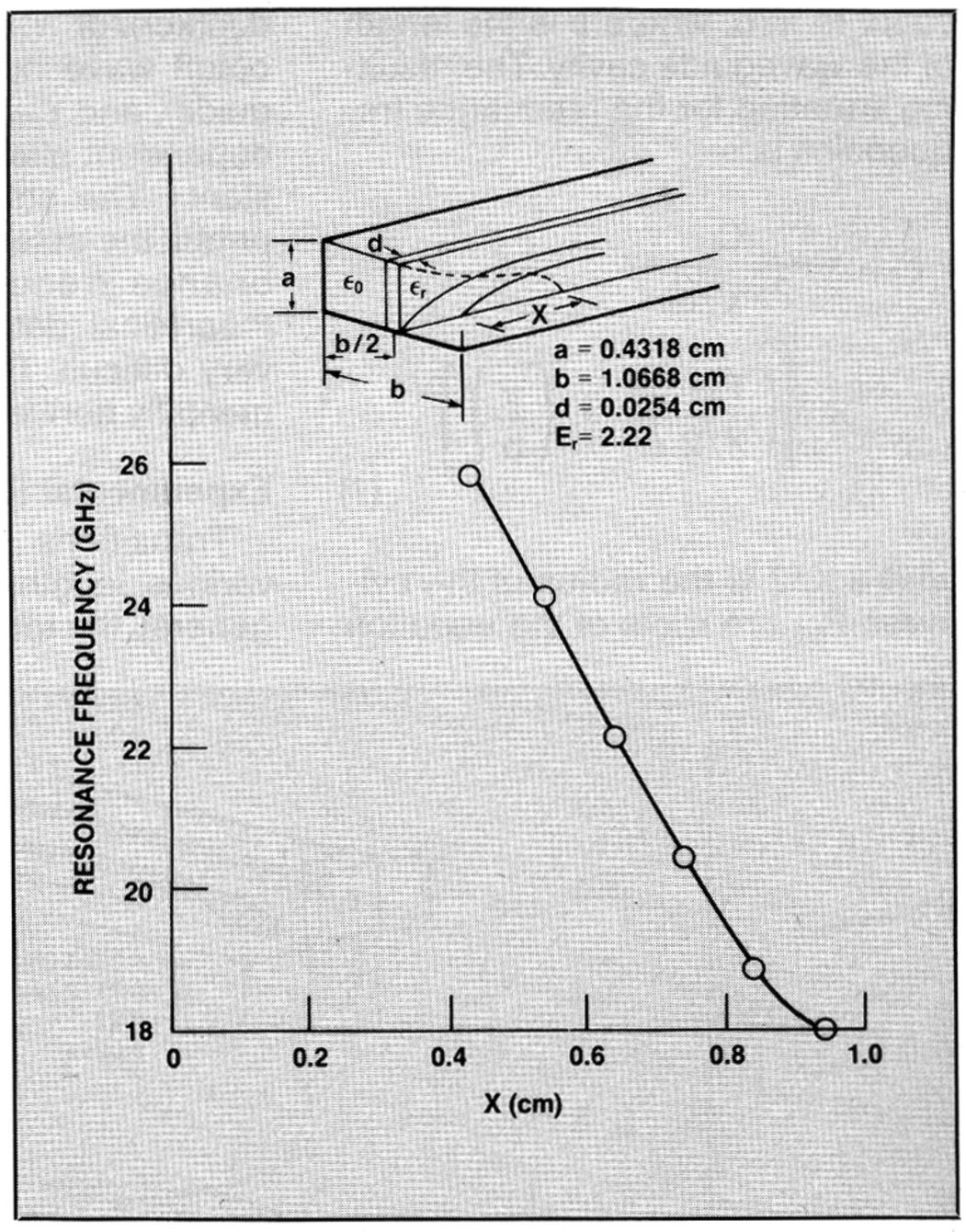

Fig. 10 Resonance frequency vs X for K-band transitions.

can be seen in Figure 8, the exact curvature of region II did not significantly alter the resonance frequency. The dominating determination of F_r is the length X. This permits the developed curves to be used for a broad range of antipodal finlines without an exact knowledge of the curvature of region II. By using Equation 1 and the results from Figure 8, values of ϵ_{eff} as a function of X and frequency were obtained and are plotted in Figure 9.

Model Derivation for Other Transitions

The curves drawn in Figure 8 may be obtained using only a few well placed data points. This allows a quick determination of the resonance frequency for a region II of any length X. This approach was followed for a determination of the resonance frequency for K-band transitions (Figure 10). All dimensions were scaled directly from Ka-band transitions.

The van Heuven transition may be described by the equivalent circuit model in Figure 5. A major complication is that the RLC circuits must describe two resonant cavities and all of their resonant modes. In addition, the resonant modes of the

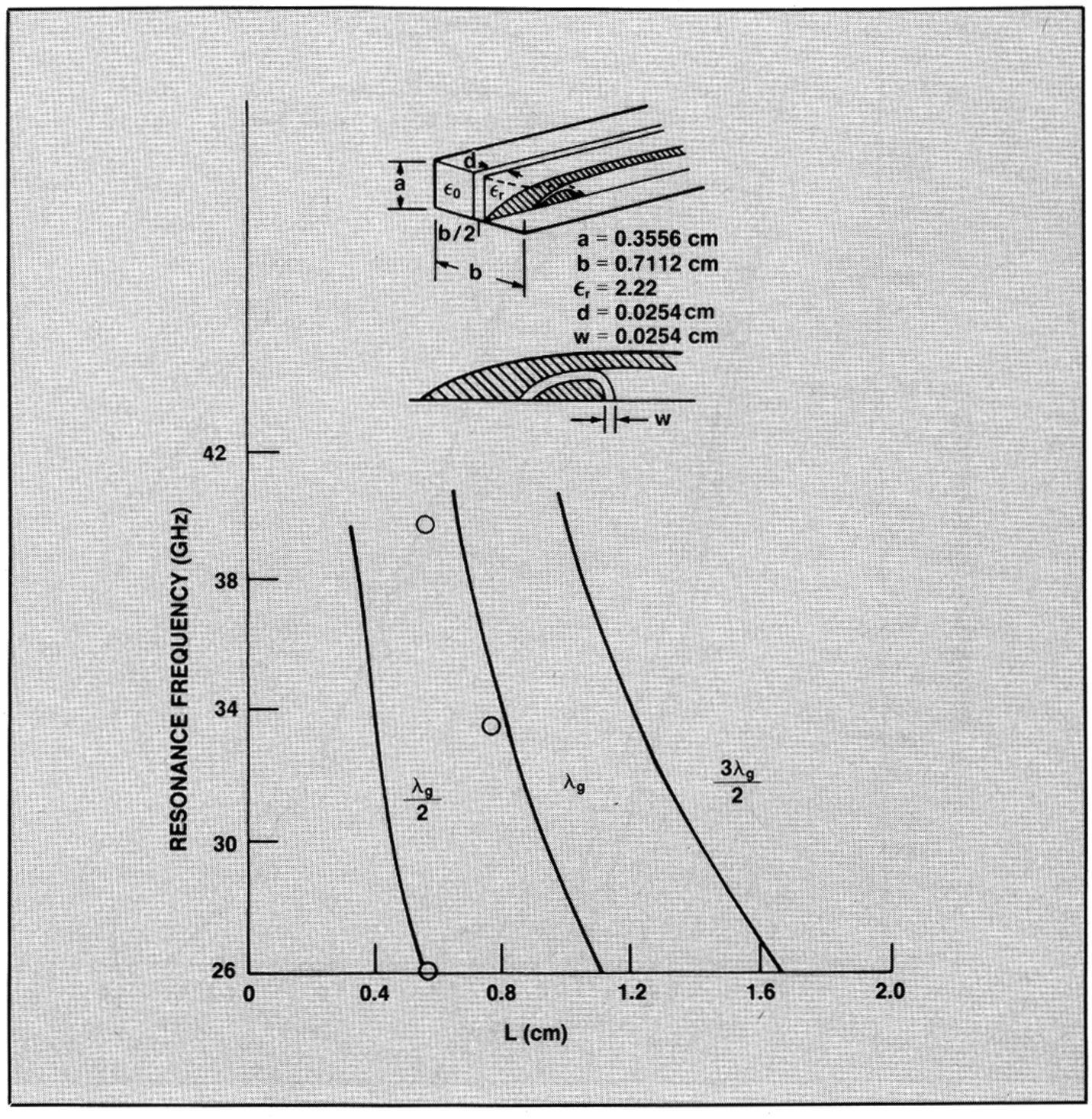

Fig. 11 Resonance frequency vs slot length L.

two cavities now are coupled closely. Therefore, the positioning of the resonance frequencies by an independent change of either X_1 or X_2 becomes impossible without an accurate model of the coupling coefficients. Empirically derived curves such as those in Figures 8 and 10 may be developed. When this was done, we noticed that the bandwidth of the transition was less than that for the transition in Figures 1b and c.

Transitions with a semicircular metal fin added to region II on the microstrip side of the transition (Figure 1c) may be modeled as a slotline resonator. The electric and magnetic fields are shown in Figure 6b. An advantage of this type of transition is the accuracy to which the resonance frequencies may be predicted. The guide wavelength of a straight slotline may be determined accurately by methods such as the spectral domain method[8] and Cohn's method.[9] It has been shown by Kawano[10] that the curvature of a slot line does not alter the resonance frequencies of a slotline resonator. Simons[11] has shown that the placement of a slot-line in rectangular waveguide has negligible effect on λ_g/λ_0. Therefore, the resonance frequencies may be modeled by a straight slotline resonator of length and width equal to those of the slot in region II located in the center of a rectangular waveguide.

The guide wavelength of slots fabricated on $\epsilon_r = 2.22$, 10 mil substrates has been determined by Simons[11] using Cohn's technique for various slot widths and frequencies in Ka-band, TE_{10}-mode rectangular waveguide. This data was used to determine the resonance frequencies (slot length $L = n\lambda_g/2$). The resonance frequencies are plotted in Figure 11. Transitions were made with slot widths of 0.0254 cm and tested. There was good agreement with the predicted values. Kawano[10] has shown that the resonance frequency increases as the slot width increases. Therefore, there are two parameters which may be varied to change F_r.

Conclusions

We have presented a physical description of the resonance frequency modes of antipodal finline waveguide-to-microstrip transitions. We developed a new model that permits the prediction of the resonance frequencies. Measurements verified the models. The models are scalable to other frequency bands. The model for the transition shown in Figure 1c is especially attractive since theoretical values for all the necessary parameters are available in the literature. The use of these models should reduce the time and cost involved in designing waveguide-to-microstrip transitions of a desired characteristic.

Acknowledgments

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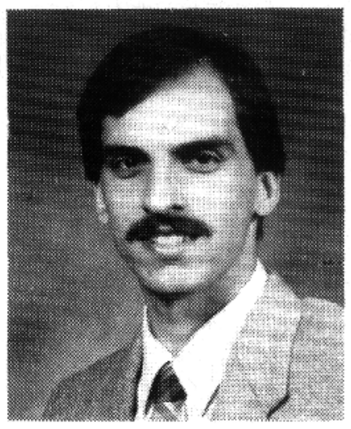

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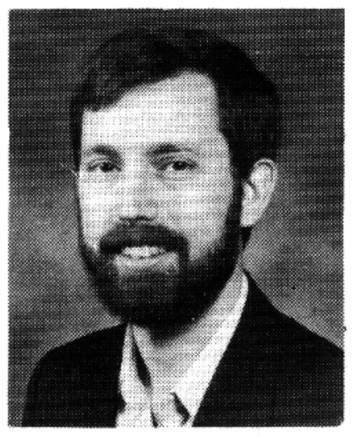
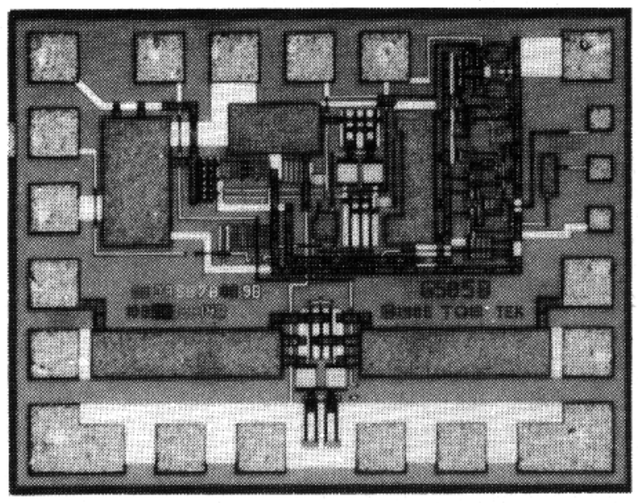


Alan N. Downey *received a BEE from Cleveland State University in 1979 and his MSEE from the University of Toledo in 1983. He joined NASA Lewis Research Center in 1977 as a co-op student and joined the Space Communications Division in 1979. From 1979 to 1985 he was engaged in microwave measurement and solid-state technology research. Since 1986, he has been involved in experiments management for the Applications Technology Satellites Program and beacon development for the Search and Rescue Satellite (SARSAT) Program.*



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