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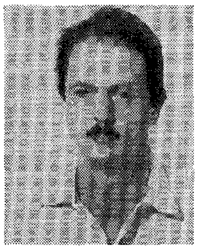

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Synthesis of Optimum Finline Tapers Using Dispersion Formulas for Arbitrary Slot Widths and Locations

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Abstract —**The theory of TEM matching sections has been modified so that it can be applied to finline tapers. A step-by-step procedure is given to calculate the taper contour for a given maximum VSWR. The taper is optimum in the sense that its length is the shortest possible for the required VSWR. To achieve fast convergence, a transversal resonance method was developed to calculate finline dispersion, which is valid for arbitrary slot widths and slot locations. The finline can be unilateral as well as bilateral, and the slot may be off-centered. The dispersion data are compared with values found in the literature, and the calculated taper performance with the authors' own measurements, both showing good agreement.**

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I. Introduction

FINLINE COMPONENTS have attracted much attention due to their favorable properties, such as broad single-mode bandwidth, moderate attenuation, sim-

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ple fabrication, and integration feasibility. For the design of such components, even if they are planned for a later integration, low return loss transitions to other waveguides are essential. Transitions in the final device should be as short as possible to achieve compactness and low insertion loss. This contribution deals with tapers between different slot widths with the above-mentioned features, and particularly transitions between finline and rectangular hollow waveguide. The slot may be located arbitrarily on the substrate. The computation is very fast, because an easily evaluated transversal resonance method is sufficient to obtain the slot profile.

II. TEM Theory

The taper design for TEM structures is well-known from the literature. The input reflection coefficient is [1]

$$R(\beta) = -\int_0^l \kappa^{-+}(z') e^{-2j\beta z'} dz' \qquad (1)$$

with $\kappa^{-+}(z)$ the z-dependent coupling coefficient between the backward and the forward traveling fundamental wave, l the length of the taper, and $\beta = 2\pi f\sqrt{\varepsilon\mu}$ the non-z-dependent phase constant. From (1), $\kappa^{-+}(z)$ can be deduced by Fourier transformation, and solutions are known which hold the input reflection coefficient R below a certain value $R_{\max}$ for phase constants $\beta > \beta_0$, thus achieving a high-pass performance.

III. Taper Synthesis for Non-TEM Structures

According to [2], an expression equivalent to (1) can be found for non-TEM waves if we approximate the coupling coefficient

$$\kappa^{-+}(f, z) \simeq \kappa^{-+}(f_0, z) \qquad (2)$$

by its value at a fixed frequency f_0. Furthermore, we approximate the integral

$$\int_0^z 2\beta(f, z')\, dz' \simeq \eta(f) \cdot \xi(z) \qquad (3)$$

as a product of a purely frequency-dependent and a purely z-dependent factor. $\eta(f)$ is normalized so that $\eta(f_0) = 1$. This results in

$$R(\eta) = \int_{-\theta}^{\theta} C \cdot K(\xi) e^{-j\eta\xi} d\xi \qquad (4)$$

with $\theta = \xi(l) = -\xi(0)$ and

$$C \cdot K(\xi) = -\frac{\kappa^{-+}(f_0, \xi)}{2\beta(f_0, \xi)}. \qquad (5)$$

The integral in (4) is of the same type as (1), and the coupling distributions for $C \cdot K(\xi)$ known from TEM theory can be applied.

IV. Non-TEM Theory for the Coupling Coefficient in Finline

If we have a relation between $\kappa^{-+}(f_0, \xi)$ and $\beta(f_0, \xi)$, the function $\beta(f_0, \xi)$ can be evaluated from $CK(\xi)$. In fact, on certain assumptions, both values can be expressed with the local cutoff frequency $f_c(z)$ as a parameter.

According to [3], the coupling coefficient κ^{-+} in an empty waveguide with arbitrary and varying cross section is

$$\kappa^{-+} = C_{11} - \frac{1}{\sqrt{Z_1}} \frac{d\sqrt{Z_1}}{dz} \qquad (6)$$

where Z_1 is the field-wave impedance.

The finline taper is modeled by a double ridge waveguide with varying gap s. With $\tan\phi_0 = ds/dz, C_{11}$ for the fundamental mode (H mode) can be written as an integral along both foreheads of the ridges

$$C_{11} = -\frac{1}{Z_1} \tan\phi_0 \int_{L_1, L_2} E_{tn}^2\, dl_s. \qquad (7)$$

E_{tn} is the electric-field distribution normalized to the power 1 of the fundamental mode if $\phi_0 = 0$.

According to [2], (7) simplifies to

$$C_{11} = -\frac{1}{f_c} \frac{df_c}{dz}. \qquad (8)$$

With

$$Z_1 = \sqrt{\mu_0/\varepsilon_0} \Big/ \sqrt{1 - (f_c/f)^2}$$

(6) and (8) result in

$$\kappa^{-+} = -\frac{1 - (f_c/f)^2/2}{1 - (f_c/f)^2} \frac{1}{f_c} \frac{df_c}{dz}. \qquad (9)$$

Equation (9) has been derived on the following assumptions:

- $|\tan\phi_0| \ll 1$, i.e., smoothly varying slot contour,
- negligible longitudinal magnetic-field components on and negligible transversal current density across the forehead of the ridge (i.e., along the thickness of the finline metallization) [2],
- TE character of the fundamental mode.

The latter assumption follows from the ridge-waveguide model. Although the field distribution is not the same, the dielectric substrate in real finlines can be considered by an effective dielectric constant k_e, which lowers the cutoff frequency f_c. With this cutoff frequency inserted in (9) and the relation

$$\beta = 2\pi f \sqrt{\varepsilon_0 \mu_0} \sqrt{k_e} \sqrt{1 - (f_c/f)^2} \qquad (10)$$

(5) can be evaluated.

V. Distribution of the Coupling Coefficient

In (4), the integrand is split into a ξ-dependent part and the normalizing constant C, so that

$$\int_{-\theta}^{\theta} K(\xi)\, d\xi = 1. \qquad (11)$$

Integrating (5) from $\xi = -\theta$ to θ with (9), (2), and $d\xi = 2\beta(f_0, z)\, dz$ from (3) yields

$$C = \frac{1}{4} \ln\left[\left(\frac{f_c(0)}{f_c(l)}\right)^4 \cdot \frac{1 - (f_c(l)/f_0)^2}{1 - (f_c(0)/f_0)^2}\right]. \qquad (12)$$

Integrating (5) from $-\theta$ to ξ results in the ξ-dependent cutoff frequency

$$f_c(\xi) = f_c(0)\cdot\left[F/2+\sqrt{F^2/4+(1-F)\cdot\exp(4CI(\xi))}\right]^{-1/2} \tag{13}$$

with $F=(f_c(0)/f_0)^2$ and

$$I(\xi)=\int_{-\theta}^{\xi} K(\xi')\,d\xi'.$$

The coupling distribution $K(\xi)$ along the taper must be chosen so that the reflection coefficient is below a certain value $R_{\max}$ for frequencies $f > f_0$. In [3], a procedure is given to evaluate the coefficients a_i in the Fourier series

$$K(\xi)=\sum_{i=1,3,\cdots}^{2n+1} a_i \cos\left(i\frac{\pi}{2}\frac{\xi}{\theta}\right) \tag{14}$$

so that the reflection coefficient $R(\eta)$ from (4) reaches the value $R_{\max}$ n times for $\eta>1$ (i.e., n equal ripples for $f>f_0$). With an infinite number of terms, (14) becomes the Dolph–Chebyshev distribution.

It is characterized by

$$K(\xi)=\frac{D}{2}\left\{\frac{I_1\left(\theta\sqrt{1-(\xi/\theta)^2}\right)}{\sqrt{1-(\xi/\theta)^2}}+\delta(\xi-\theta)+\delta(\xi+\theta)\right\} \tag{15}$$

for $-\theta \leqslant \xi \leqslant \theta$ with $D=R_{\max}/C, \theta=\operatorname{arcosh}(1/D)$, $I_1(x)$ the modified Bessel function of the first order, and $\delta(t)$ Dirac's delta function.

The shape of the function $K(\xi)$ in (14) with $n=4$ compared with that of (15) is shown in Fig. 1 with $D=0.01$ in both cases. The functions are even in ξ, so only the branch for $\xi>0$ is shown. The shapes of $K(\xi)$ are similar where both are nonzero. The Dirac's function at $\xi=\theta$ makes the Dolph–Chebyshev distribution the shortest possible taper for a given reflection loss.

Due to the normalization (11) and the symmetry $K(\xi)=K(-\xi)$ we can evaluate the integral in (13)

$$I(\xi)=\begin{cases}\dfrac{1}{2}+\dfrac{D\theta^2}{2}\displaystyle\int_0^{\xi/\theta}\frac{I_1\left(\theta\sqrt{1-y^2}\right)}{\theta\sqrt{1-y^2}}\,dy, & \text{for } |\xi|<\theta\\ 0, & \text{for } \xi=-\theta\\ 1, & \text{for } \xi=\theta\end{cases} \tag{16}$$

The Bessel function in (16) can be expanded in a power series [4], and the integral in (16) can be evaluated by integrating it term by term.

This expansion was previously published in [5], but because of some misprints there, we give the result once more

$$\int_0^x \frac{I_1\left(\theta\sqrt{1-y^2}\right)}{\theta\sqrt{1-y^2}}\,dy=\sum_{k=0}^{\infty}a_k b_k$$

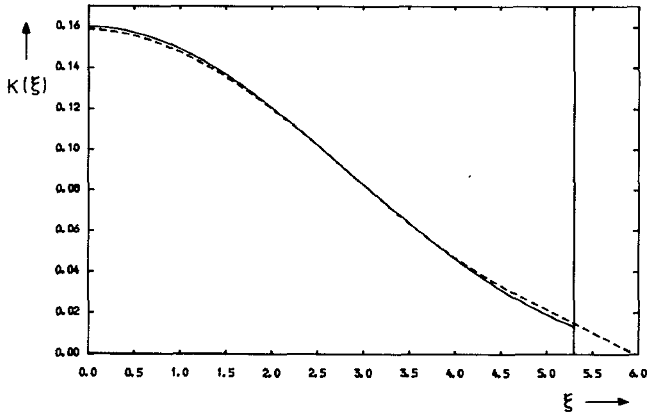

Fig 1. Normalized coupling distribution for $D = 0.01$. ——Dolph–Chebyshev (15). ----cosine-series $n=4$ (14).

with

$$a_0=1 \qquad a_k=\frac{\theta^2}{4k(k+1)}a_{k-1}$$

$$b_0=\frac{x}{2} \qquad b_k=\frac{\frac{x}{2}(1-x^2)^k+2kb_{k-1}}{2k+1}. \tag{17}$$

The series converges rapidly, and taking only 12 terms makes the relative error smaller than 10^{-5}.

Dirac's function in (15) produces discontinuities at the taper ends, a fact which might cause problems due to the excitation of higher order modes. The step in cutoff frequency can be calculated from (12), (13), and (16)

$$\frac{f_c(+0)}{f_c(0)}=1-R_{\max}\cdot\frac{f_0^2-f_c^2(0)}{2f_0^2-f_c^2(0)}. \tag{18}$$

The discontinuity at the taper end with the narrow slot is even smaller. The lower the $R_{\max}$ chosen, the smaller the step. For $R_{\max}=0.01$ ($\triangleq -40$ dB) and $f_0=1.4\ f_c(0)$, the ratio in (18) is 0.997. The Fourier series distribution (14) does not have this discontinuity. Comparative measurements with both taper types showed that the influence of the step is negligible.

VI. Synthesizing the Slot Contour

Up to now, we have been able to synthesize the function of the cutoff frequency f_c along the taper. It should be noted that no wave impedances are required for this synthesis, thus avoiding the problem of finding the appropriate impedance definition.

To obtain the slot width, we need a relation between slot width and cutoff frequency. This relation should be easy to evaluate, because it is called at every knot on the z axis. It is inconvenient to apply spectral-domain techniques [6]–[9], which are very precise, but time-consuming. The simple formulas of [10] are limited to special values for the permittivity of the substrate and are not valid for large slot widths, [11] gives no data for small slot widths, and the accuracy of the results in [12] seems too poor.

VII. Transversal Resonance Condition

A new attempt was therefore made to evaluate cutoff frequency and effective permittivity by a transversal resonance method. The method is valid for arbitrary slot widths and slot locations. The computed results are compared with data from [8], [10], [11], and [13].

The analysis is based on the following simplifications:

- isotropic, homogeneous, and lossless dielectric layer,
- zero-thickness metallization with infinite conductivity,
- thickness of substrate small compared with waveguide width,
- symmetrically located substrate.

In the dispersion relation (10) for the phase constant β, which underlies our taper synthesis, the effective dielectric constant k_e may be approximated as frequency-independent [14]

$$k_e \simeq (f_{c0}/f_c)^2 \tag{19}$$

with f_{c0} the cutoff frequency of a finline of the same dimensions and a substrate's permittivity $\varepsilon_r = 1$. Equation (10) is then the dispersion of a homogeneously filled waveguide. In the following, our task will be to find the cutoff frequencies f_c and f_{c0}.

A. Bilateral Finline

The cross section of a bilateral finline and its equivalent transverse network at cutoff are shown in Fig. 2. The finline is symmetrical with respect to the $x = a/2$ plane, where the admittance in the x-direction at cutoff is zero. The equivalent transverse network for the dominant mode consists of a capacitive susceptance shunting the TEM transmission line with short-circuit termination. The cutoff wavenumber k_x is determined by the resonance condition at reference plane T

$$-\cot(k_x \cdot l) + \frac{B_b}{Y_b} = 0 \tag{20}$$

with

$$\frac{B_b}{Y_b} = \frac{b}{\pi} \cdot k_x \cdot (P_w + \varepsilon_r \cdot P_d) \tag{21}$$

where

$$P_w = \ln(\csc(\alpha_w) \cdot \csc(\beta_w)) \tag{22}$$

$$P_d = r_d \cdot \arctan\left(\frac{1}{r_d}\right) + \ln\sqrt{1 + r_d^2} \tag{23}$$

and

$$\alpha_w = \frac{\pi}{2} \cdot \frac{s}{b} \qquad r_d = \frac{d}{s}.$$

$$\beta_w = \frac{\pi}{2}\left(1 - 2\frac{e}{b}\right).$$

The field distortions by the metal fins have been modeled by the susceptance jB_b which is composed of two parts. The first part models the field distortions left of reference plane T. It is taken from Marcuvitz's *Waveguide Handbook* [15, p. 218], where the susceptance of a window of zero thickness in a rectangular waveguide has been derived. In fact, it is sufficient to take just the first term in P_w into account (22) because the other terms influence the results by less than 0.2 percent.

The second part models the field distortion between plane T and the symmetry plane $x = a/2$. It has been taken from the equivalent circuit of an open E-plane T junction (see [15, p. 337]). Due to the symmetry with respect to $x = a/2$, this part of the susceptance is characterized only by the susceptance jB_a of the equivalent circuit in [15].

The first zero of k_x in (20) is the cutoff wavenumber for the fundamental mode. The cutoff frequencies f_{c0} and f_c are obtained from k_x with $\varepsilon_r = 1$ and $\varepsilon_r \neq 1$, respectively. Numerical results have been compared to those taken from [10] and a disagreement of less than 1 percent has been found. (Setting $\varepsilon_r = 2.22$, $\left(f_c\sqrt{\varepsilon_0\mu_0} \cdot b\right)$ varies between 0.1 and 0.2 for d/b between 1/32 and 3/4. In all cases, the deviations are smaller than 0.001.) Fig. 3 shows dispersion in a bilateral finline of relatively small slot width. The agreement with results taken from [13] is very good.

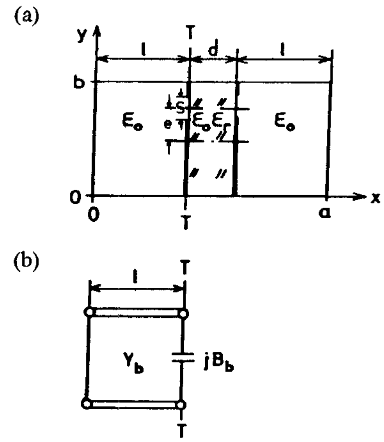

Fig. 2. The cross section of (a) a bilateral finline and (b) its equivalent transverse network.

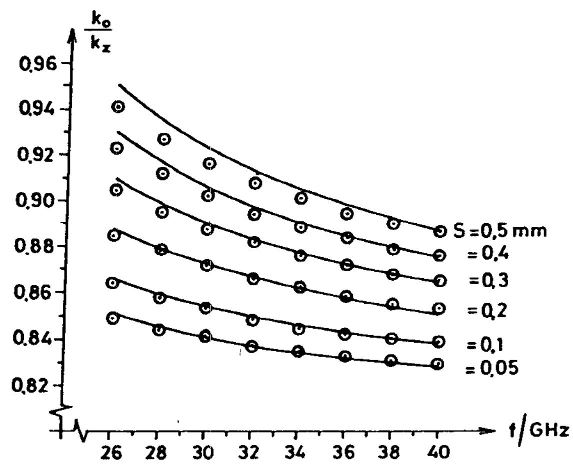

Fig. 3. Dispersion of a bilateral finline with slot centered in waveguide. ——— Simons and Tech [13]. ⊙ Our approximation.

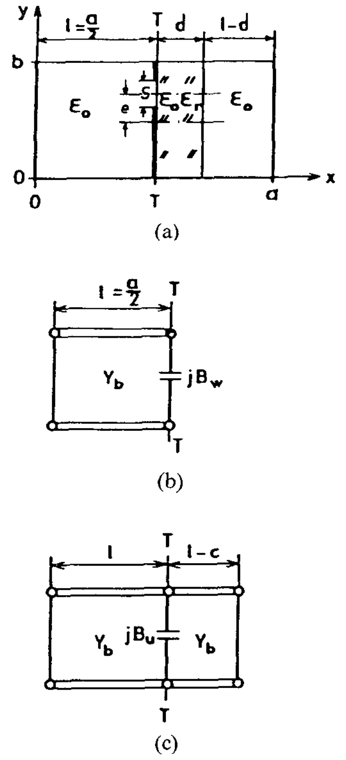

Fig. 4. The cross section of (a) a unilateral finline, (b) its equivalent transverse network for $\varepsilon_r = 1$, and (c) for $\varepsilon_r \neq 1$.

B. Unilateral Finline

Unilateral finline is often preferred because it is easy to fabricate and semiconductor devices are simple to mount. The cross section of this finline is shown in Fig. 4. The thin metal layer is placed at the $x = a/2$ plane, so that the structure is symmetrical for $\varepsilon_r = 1$. It is useful to construct two equivalent transverse networks at cutoff for the dominant mode: one for $\varepsilon_r = 1$ and another for $\varepsilon_r \neq 1$ (Fig. 4). The equivalent network for $\varepsilon_r = 1$ consists simply of a transmission line of length $a/2$ which is short-circuited at one end and shunted by jB_w at the other. The shunt susceptance can be taken from (21) and (22).

The cutoff wavenumber k_{x0} in the air-filled ridged waveguide is determined by the resonance condition at reference plane T

$$-\cot(k_{x0} \cdot l) + \frac{B_w}{Y_b} = 0 \tag{24}$$

with

$$\frac{B_w}{X_b} = \frac{b}{\pi} \cdot k_{x0} \cdot P_w. \tag{25}$$

P_w is given by (22). The equivalent network in Fig. 4(c) is formed by a susceptance jB_u shunted by two TEM transmission lines with short-circuit terminations. The relation governing cutoff of the H_{10} mode is given by

$$-\cot(k_x \cdot l) - \cot[k_x \cdot (l-d)] + \frac{B_u}{Y_b} = 0 \tag{26}$$

with

$$\frac{B_u}{Y_b} = \frac{b}{\pi} \cdot k_x \cdot [2P_w + \varepsilon_r(P_d + P_b)] \tag{27}$$

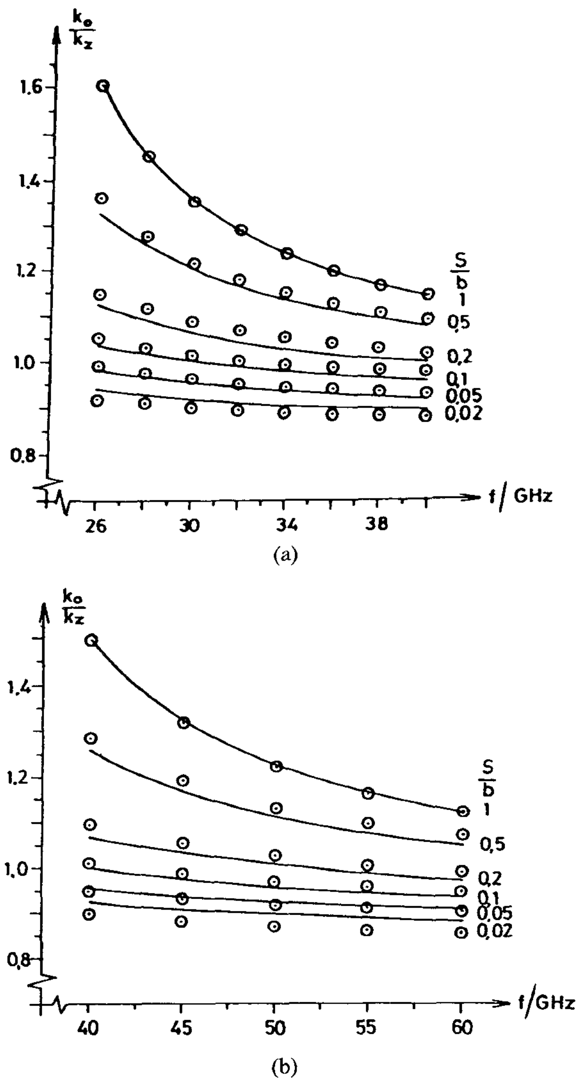

Fig. 5. Dispersion of unilateral finline with (a) WR(28) and (b) WR(19) shield. Fins are centered. ——Spectral-domain technique; Knorr and Shayda [8]. ⊙ Our approximation.

where

$$P_b = r_b \cdot \arctan\left(\frac{1}{r_b}\right) + \ln\sqrt{1 + r_b^2} \tag{28}$$

and

$$r_b = \frac{d}{b}.$$

The susceptance jB_u has been constructed by superposing the susceptance of the window (the term proportional to $2P_w$) and another susceptance (the term proportional to $\varepsilon_r(P_d + P_b)$) representing the influence of the dielectric and the transformation through the layer of length d. While P_w (22) and P_d (23) have already been given, P_b in (28) has been taken from [15, p. 337].

The unilateral finline with metal fins centered in the waveguide (Fig. 4) has been analyzed in [8], [10], and [13]. Other authors have considered a unilateral finline with the dielectric layer symmetrically located in the waveguide [6], [9]. The modifications of (24) and (26) for this case are easy to carry out.

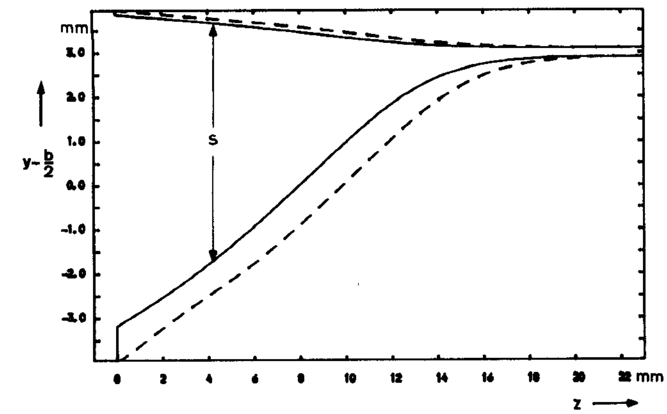

Fig. 6. Taper contour. Design data: $f_0 = 12.3$ GHz, $R_{max} = -30$ dB. ——Dolph–Chebyshev. ----cosine-series $n = 4$.

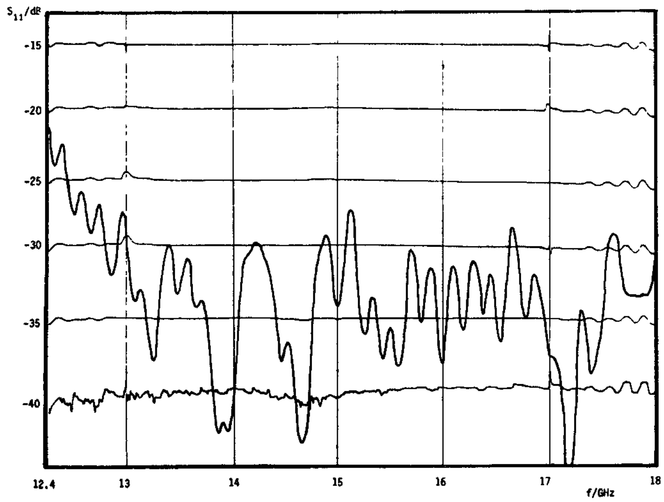

Fig. 7. Return loss of double taper according to Fig. 6, Dolph–Chebyshev.

Comparing the cutoff wavenumber of the unilateral finline with results published in [10] yields deviations of less than 1.5 percent if $d/a < 1/8$. This restriction is, however, usually fulfilled for practical finlines. Some dispersion curves are presented in Fig. 5 showing good agreement (better than 3 percent) with published results. The authors' results deviate from those calculated with Hoefer's diagram for the slot capacitance [11] less than 0.5 percent. This diagram, however, gives only data for a normalized slot width $s/b \geqslant 0.1$.

VIII. Synthesis Procedure

The synthesis procedure can now be summarized as follows.

1) Choose the cutoff frequency f_0 of the taper and the maximum input reflection coefficient R_{max} for $f > f_0$. Choose an appropriate step width Δz for the longitudinal coordinate. If the taper is eccentric, select a relation between slot width s and eccentricity e (Figs. 2(a) and 4(a)) to achieve a unique function $s(f_c)$ from the characteristic equation. A linear function

$$e(s(z)) = e(0) + \frac{e(l) - e(0)}{s(l) - s(0)} \cdot (s(z) - s(0)) \quad (29)$$

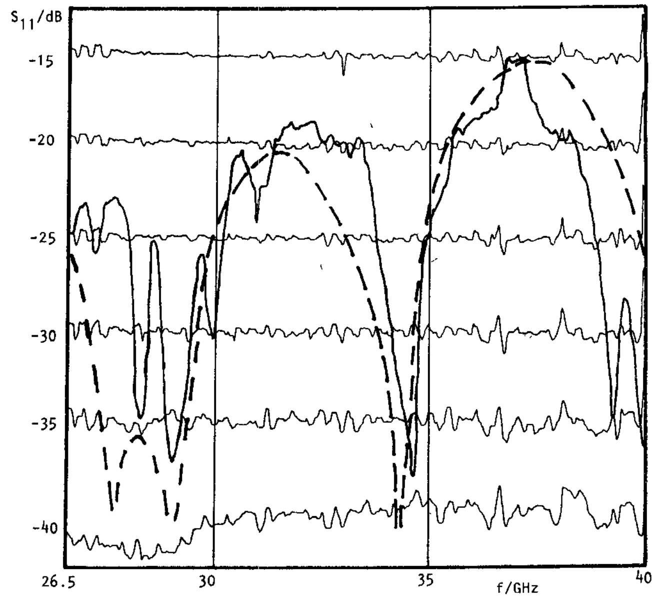

Fig. 8. Return loss of double taper $R_{max} = -20$ dB, Dolph–Chebyshev. ——Measured, ----calculated.

guarantees that the slot edges never exceed the waveguide height.

2) Determine $f_c(0)$ and $f_c(l)$ from $s(0)$ and $s(l)$ and the transverse resonance condition.

3) Determine the normalization factor C from (12).

4) Set initial values $z = 0$, $\xi = -\theta$, and $s = s(0)$.

5) Next take $z := z + \Delta z$. Evaluate the related $\xi(z)$ from (3) as $\xi := \xi + 2\beta(f_0, z) \cdot \Delta z$, f_c from (13), and s from the transversal resonance condition with $k_x = 2\pi f_c \sqrt{k_e} \sqrt{\mu_0 \varepsilon_0}$.

6) Repeat 5) until the final slot width is reached.

IX. Experimental Results

Several double-tapers were fabricated on RT/Duroid 5880 in WR-62 (and WR-28) housing. They lead from the empty waveguide to a slot width of 0.2 mm and back again to the empty waveguide. The slot was located either in the center or 3 mm (1.5 mm) below, R_{max} was chosen to be -20, -30, -35, -40, and -60 dB. To estimate the influence of the steps in the Dolph–Chebyshev profile, a set of tapers with the cosine-series coupling distribution (14) was also made.

The dynamic range of the measurement equipment was -40 dB (-35 dB) for the reflection coefficient, restricted by the waveguide termination behind the tapers. No double-taper (except the 20-dB type) turned out to be worse than -25 dB in the return loss from 12.4 to 18 GHz (26.5 to 40 GHz), and most of them were better than -30 dB. The taper length for a taper designed for -30 dB is only 18 mm (8.2 mm), which is 0.75 free-space wavelengths at the lower frequency limit.

The results of the measurements can be summarized as follows.

1) There is no significant difference in the performance of cosine-series and Dolph–Chebyshev profile tapers. The steps in the profile apparently have no influence.

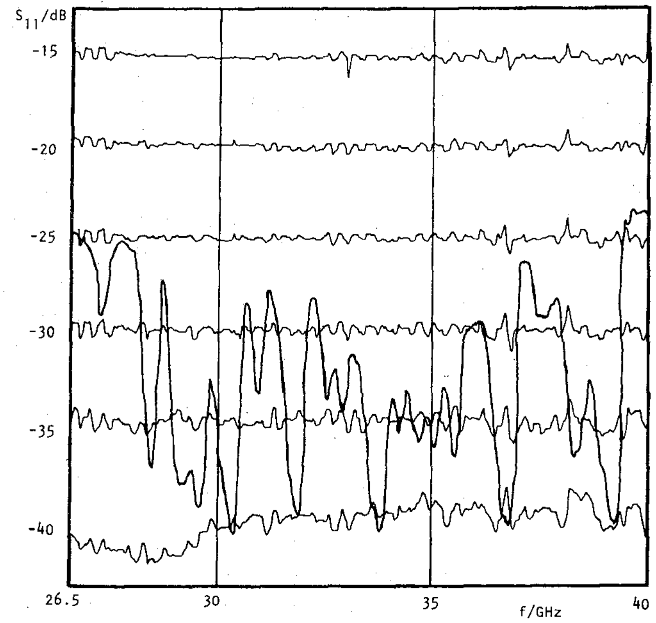

Fig. 9. Return loss of double taper. Design data: $f_0 = 26$ GHz, $R_{max} = -30$ dB, length 8.2 mm, centered slot.

2) Designing the taper for reflection loss better than -35 dB does not improve the performance. The taper may even become worse. This probably means that the accuracy limit of the transversal resonance method has been reached.

3) Tapers with a return loss of -30 dB or better should be fabricated within ± 5 percent in slot width.

For lower requirements, however, tolerances are not so critical: For a maximum reflection coefficient of -20 dB, a profile shift of ± 20 percent is admissible.

Fig. 6 shows the slot profile of a -30-dB *Ku*-band eccentric taper with cosine-series and Dolph–Chebyshev profile. Fig. 7 shows the input reflection of the corresponding double taper. Fig. 8 compares the measured input reflection of a 20-dB double taper compared with values calculated by direct integration of the coupled wave equations [3] with phase constant $\beta(f, z)$ and coupling coefficient $\kappa^{-+}(f, z)$. Finally, Fig. 9 shows the performance of a -30-dB taper in *Ka*-band. Insertion loss was about 0.3 dB for any double taper.

X. Conclusion

A fast and precise method has been presented for the design of optimum finline tapers with arbitrary slot widths and locations. The method can be applied for tapers down to -35-dB return loss. An even better performance may be obtained with a more exact relation between slot width and cutoff frequency, involving more computer time. The practical results would be doubtful, however, because requirements for mechanical tolerances become very stringent. The design procedure can also be applied to antipodal finlines, provided that a relation between slot width and cutoff frequency is known. The tapers have already been successfully applied to finline circulators, phase modulators, mixers, and oscillators.

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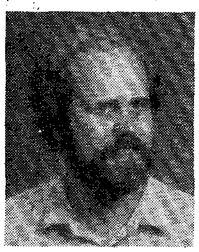

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Short Papers

Quarter-Wave Matching of Waveguide-to-Finline Transitions

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Abstract —**This paper presents closed-form expressions for the design of a quarter-wave transition-matching transformer. This structure takes the form of a notch or protrusion cut in the finline substrate at the waveguide-to-finline interface. The dimensions of the transformer are calculated using a homogeneous waveguide model for the partially loaded sections. The characteristics of this model are found with perturbation theory. Several transformers were designed and measured. A 5-dB improvement in return loss over a full waveguide band is typical.**

I. Introduction

E-PLANE CIRCUIT technology is now well established as a viable approach to millimeter-wave circuit realization and integration. Indeed, almost all important circuit functions have been successfully realized in this technology using integrated finline as the principal transmission medium. E-plane circuits consist of metallic fin pattern deposited on a thin substrate which is suspended in the E-plane of a standard waveguide enclosure.

For obvious reasons, such circuits must be made compatible with standard waveguide components and test equipment. In most cases, this is accomplished through a printed taper, a finline section in which the gap between the fins is gradually narrowed from the waveguide height b, to its final width d, as shown in Fig. 1.

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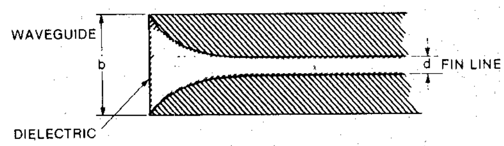

Fig. 1. Tapered waveguide to finline transition.

A critical look at Fig. 1 reveals that even for an optimal taper profile the transition could not be reflectionless because of the dielectric discontinuity at the taper front. For typical finline substrates and geometries, the return loss due to this discontinuity is approximately 27 dB.

In order to minimize the effect of the "dielectric step," various researchers [1]–[5] have introduced a quarter-wave transformer section in the form of either a notch or a protrusion, as shown in Fig. 2.

One can only surmise from the literature that the dimensions of these transformers have been determined by trial and error. In this paper, therefore, design expressions will be developed to determine the dimensions of quarter-wave notches and protrusions. Measured data will be presented to demonstrate the validity of the new formulas as well as the improvement in reflection loss achieved by such a structure.

It may be noted here that in spite of the inherently narrow bandwidth of quarter-wave transformers, the improvement in return loss is significant over a complete waveguide band, about 5 dB. Only for very broad-band applications will it become necessary to use multistep transformers. However, in such a case, the waveguide en-

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