A Simple Method for Analyzing Fin-Line Structures

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Abstract—**A first-order design theory for fin-line circuits is developed by establishing a correspondence between a fin line and a set of rectangular waveguides. The usefulness of the method is demonstrated by applying it to the analysis of a transition from a fin line to a below-cutoff waveguide. Finally, theoretical and experimental results are given for a bandpass filter, which show a fairly good agreement.**

I. Introduction

THE INTEGRATED fin line is an advantageous alternative to the microstrip in the design of microwave integrated circuits at frequencies above 20 GHz [1]. The only theoretical approaches being available until now are those of [2] and [3], which both deal with dispersion of the phase coefficient and of the wave impedance. While the first paper requires extensive and complicated mathematics, which makes the method cumbersome if it is applied to an analysis of fin-line structures with a varying slot pattern, the second paper does not show any way to calculate the field distributions in the cross section of the waveguide. There is, hence, a need for a first-order design theory as it already exists for microstrip circuits, which allows analyzing even complex fin-line structures with a limited effort. The present work shall fill this gap.

Our aim is to present an equivalent description of fin-line circuits, which establishes a one-to-one correspondence between the field expansion in a rectangular waveguide homogeneously filled with dielectric and in a fin line. It will be shown that the knowledge of only one group of eigenmodes (the TE_{m0} modes) is sufficient to analyze a wide class of fin-line circuits with a varying slot pattern. Concerning these modes the fin line can be described by equivalent rectangular waveguides.

The investigations to be presented are fourfold. In Section II, the complete eigenmodes of fin lines of arbitrary configurations will be derived. Based on this field expansion, the transition from a fin line to a rectangular waveguide, which is separated by the closed slot of a fin line in two parts being below cutoff, is analyzed in Section III. By matching the various tangential field components at the interface, the TE_{m0} modes of the fin line (which are, of course, different from the TE_{m0} modes of a rectangular waveguide) turn out to be the only components which determine the transmission and reflection coefficient of the transition. This forms the foundation for Section IV, where directions will be given on how to replace a fin line by an equivalent set of rectangular waveguides. Analysis of the same transition as in Section III using this equivalent description then yields identical results. Finally, Section V is devoted to a discussion of the validity of the method.

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II. Fin-Line Eigenmodes

The eigenvalues of a fin line will now be calculated for the cross section of Fig. 1. The configuration is regarded at the cutoff frequency where it forms a parallel-plate waveguide which is short-circuited at $x=0$ and $x=a$. The dielectric substrate with its metal fins can be regarded as a discontinuity between the parallel plates. Such step discontinuities have been analyzed approximately for the case of two infinitely long transmission lines in [4] by neglecting any frequency dependence of the phase constants. We will solve the problem accurately by expanding the fields in the two regions A and B and by matching the tangential field components at the interface.

The TE_{mn} modes with m odd will be regarded as an example. Such a mode is characterized at cutoff by an E_y and an H_z component. For region A one can write

$$E_{yA}=A_n\cos(k_{ynA}y)\sin(k_{xmA}x)+\sum_{p\neq n}A_p\cos(k_{ypA}y)\sin(k_{xpA}x) \quad (1)$$

$$H_{zA}=A_nY_A\cos(k_{ynA}y)\cos(k_{xmA}x)+\sum_{p\neq n}A_pY_{pA}\cos(k_{ypA}y)\cos(k_{xpA}x) \quad (2)$$

and in region B

$$E_{yB}=B_n\cos(k_{ynB}(y-b_1))\cos(k_{xmB}(a/2-x))+\sum_{s\neq n}B_s\cos(k_{ysB}(y-b_1))\cos(k_{xsB}(a/2-x)) \quad (3)$$

$$H_{zB}=B_nY_B\cos(k_{ynB}(y-b_1))\sin(k_{xmB}(a/2-x))+\sum_{s\neq n}B_SY_{sB}\cos(k_{ysB}(y-b_1))\sin(k_{xsB}(a/2-x)). \quad (4)$$

The various constants in (1)–(4) read

$$k_{ynA}=n\pi/b \quad k_{ypA}=p\pi/b$$

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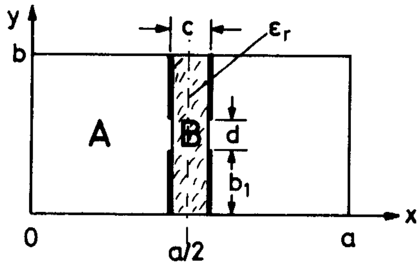

Fig. 1. Cross section of a fin line.

$$k_{xmA}^2 + k_{ynA}^2 = k_{cmn}^2 \quad k_{xpA}^2 + k_{ypA}^2 = k_{cmn}^2$$
$$k_{ynB} = n\pi/d \quad k_{ysB} = s\pi/d$$
$$k_{xmB}^2 + k_{ynB}^2 = \epsilon_r k_{cmn}^2 \quad k_{xsB}^2 + k_{ysB}^2 = \epsilon_r k_{cmn}^2$$
$$Y_A = -k_{cmn}^2/(j\omega\mu k_{xmA}) \quad Y_{pA} = -k_{cmn}^2/(j\omega\mu k_{xpA})$$
$$Y_B = -\epsilon_r k_{cmn}^2/(j\omega\mu k_{xmB}) \quad Y_{sB} = -\epsilon_r k_{cmn}^2/(j\omega\mu k_{xsB}). \tag{5}$$

$\lambda_{cmn} = 2\pi/k_{cmn}$ is the cutoff wavelength of the structure.

The boundary conditions at $x = l/2 = (a-c)/2$ read

$$E_{yA} = E_{yB} \quad H_{zA} = H_{zB} \text{ at } b_1 \leqslant y \leqslant b_1 + d$$
$$E_{yA} = 0 \text{ at } 0 \leqslant y \leqslant b_1 \quad b_1 + d \leqslant y \leqslant b. \tag{6}$$

Equation (6) yields the characteristic equation

$$Y_B/d \tan(k_{xmB}c/2) = Y_A/b \cot(k_{xmA}l)\phi(n,n) + \sum_{p\neq n} Y_{pA}/b \cot(k_{xpA}l)\phi(p,n) \tag{7}$$

where $\phi(n,n)$ and $\phi(p,n)$ must be calculated from the following system of linear equations:

$$\phi(t,n) = k_{tn} + F_1(t,n)\phi(n,n) + \sum_{p\neq n} F_2(t,n,p)\phi(p,n), \quad t = 1,2,\cdots. \tag{8}$$

The quantities k_{tn}, $F_1(t,n)$, and $F_2(t,n,p)$ depend on ϵ_r, k_{cmn}, and the dimensions of the fin-line cross section.

The eigenvalues k_{cmn} can now be calculated by solving the characteristic equation. A similar procedure must be undertaken for m even and for the TM_{mn} modes.

The foregoing analysis is based on a suggestion of Cohn for calculating the eigenvalues of ridge waveguides [5]. It can be modified to include the case of unilateral fin lines (with metal fins covering only one side of the substrate) according to the calculations in [6] for slot lines, because a unilateral fin line is similar to a boxed-in slot line.

The eigenvlaue approach shall now be used to construct the fin-line eigenmodes. To proceed in this, we must remember that the propagating waves in fin lines are neither TE nor TM but a combination of both. This is due to the dielectric substrate of integrated fin lines (see the cross section in Fig. 1). It is impossible to simultaneously match all the field components of either a TE or a TM mode in the slot interface between regions A and B except at cutoff frequency and for perfectly conducting walls [7].

Since the hybrid eigenmodes of the fin line contain TE and TM terms, they will be classified in the following way. An HE mode indicates a hybrid eigenmode, in which the TE part is much larger than the TM part. When the frequency approaches its cutoff value, the TM part vanishes and the HE mode becomes purely TE. An EH mode indicates a hybrid eigenmode with a dominating TM part. Now the TE part vanishes at cutoff.

From the foregoing analysis, we know both the TE and TM modes at cutoff. In order to extend this solution beyond this frequency, we will utilize the fact that the ratio between the TE and TM parts in an hybrid eigenmode primarily depends on the magnitude of the dielectric constant ϵ_r and on the substrate thickness c. For moderate ϵ_r and $c/a \ll 1$, the dielectric plays a minor role [1] so that the eigenmodes may be considered to be either TE or TM. This holds exactly for the fin line of Konishi, which equals a ridge waveguide with a thin ridge [8]. Our calculations will henceforth be restricted to the case that ϵ_r is moderate and $c/a \ll 1$. Equations (1)–(6) are then approximate expressions for the HE modes of a bilateral fin line. Slightly modified relations hold in the case of a unilateral fin line which can be deduced utilizing the results in [6].

A characteristic quantity of fin-line eigenmodes is the effective dielectric constant k_{emn}, which has been defined in [1] as the squared ratio of the cutoff wave numbers k_{cmn} in the case of an air-filled fin line to that of the original fin line with the dielectric substrate between the fins. In order to calculate the effective dielectric constant k_{emn}, the characteristic equation has been solved twice—a first time for $\epsilon_r = 1$ and a second time for $\epsilon_r = 2.22$ (RT-duroid as substrate material). k_{emn} is then given by [1], [3] as

$$k_{emn} = k_{cmn}^2(\epsilon_r = 1)/k_{cmn}^2(\epsilon_r = 2.22). \tag{9}$$

The effective dielectric constant depends only very weakly on frequency, as has been shown in [2]. Hence the propagation constant can be written as

$$k_{zmn}^2 = k_{emn}k^2 - k_{cmn}^2(\epsilon_r = 1) \tag{10}$$

where $k^2 = \omega^2\mu\epsilon$ and k_{emn} is taken from the relation above.

With (9) and (10) the solution for the fin-line eigenmodes has been completed. Its basic assumptions are the moderate ϵ_r, $c/a \ll 1$, and the k_{emn} constant versus the frequency. The first two are well fulfilled for usually used fin lines; the third holds according to the numerical results in [2]. The eigenmode approach presented here hence is believed to form a basis for a first-order design theory for fin-line circuits. A similar though approximate attempt to the solution of this problem has already been given in [7], where the tangential electric field at the interface has been matched while the magnetic field has not.

Numerical results are shown for the TE_{m0} fin-line mode in Fig. 2.

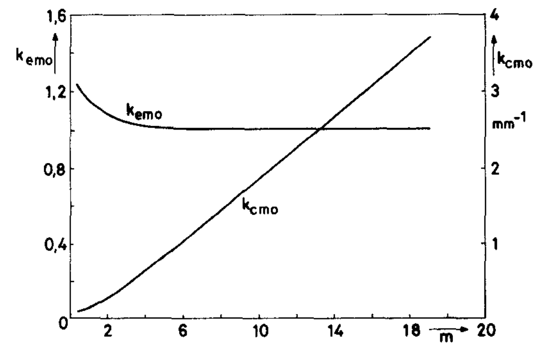

Fig. 2. Effective dielectric constant k_{em0} and wave number k_{cm0} for TE_{m0} modes versus m. The dimensions of the fin line are $a=15.8$ mm, $d=1.58$ mm, $b=7.9$ mm, $b_1=3.16$ mm, $c=0.3$ mm, and $\epsilon_r=2.22$.

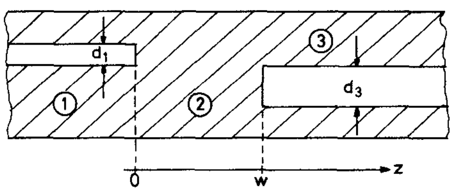

Fig. 3. Slot pattern of a metallic strip in a fin line.

III. Metallic Strip in a Fin Line

The eigenmodes of the fin line being known, the transition from a fin line to a below-cutoff waveguide (which is realized by short-circuiting the slot between the fins) and back to another fin line can be analyzed. The corresponding slot pattern on the fin-line substrate is shown in Fig. 3.

The discontinuity will, in general, excite all types of eigenmodes in the three regions. The incidental wave in region 1 is a TE_{10} mode.

$$E_{y1}^{TE}=\Big(A_0 \sin(k_{x11}x) + \sum A_p \cos(p\pi y/b)\sin(k_{xp1}x)\Big)\exp(-jk_{z10}z), \quad \text{for } 0\leqslant x\leqslant 1 \quad 0\leqslant y\leqslant b$$

$$E_{y1}^{TE}=\Big(B_0\cos(k_{x12}(a/2-x))+\sum B_s\cos(s\pi(y-b_1)/d_1)\cdot\cos(k_{xs2}(a/2-x))\Big)\exp(-jk_{z10}z), \quad \text{for } 1\leqslant x\leqslant a/2 \quad b_1\leqslant y\leqslant b_1+d_1 \quad (11)$$

$$H_{x1}^{TE}=Y_{10}E_{y1}^{TE} \quad (12)$$

with

$$k_{x11}=k_{c10} \quad k_{x12}=\sqrt{\epsilon_r}\,k_{c10} \quad k_{xp1}^2+(p\pi/b)^2=k_{c10}^2$$
$$k_{xs2}^2+(s\pi/d_1)^2=\epsilon_r k_{c10}^2 \quad Y_{10}=-k_{z10}/(\omega\mu). \quad (13)$$

The excited waves in regions 1 and 3 consist of both TE_{mn} and TM_{mn} terms. The E_y^{TE} component is given by the expressions in (1) and (3) times $\exp(-jk_{zmn}z)$, and $H_{x1}^{TE}=Y_{mn}E_{y1}^{TE}$ with $Y_{mn}=-k_{zmn}/(\omega\mu)$. The TM components of E_y and H_x have the same form as E_y in (1) and (3) but with different constants.

The electromagnetic field in region 2 can be described by

$$E_{y2}^{TE}=A_{rq}\sin(r\pi x/1)\cos(q\pi y/b)\exp(-jk_{zrq}z) \quad (14)$$
$$H_{x2}^{TE}=Y_{rq}E_{y2}^{TE} \quad (15)$$

and by a TM term, which is identical to the TE term with one exception. Y_{rq} has to be replaced by Y'_{rq}. The various constants read

$$Y_{rq}=-k_{zrq}/(\omega\mu) \quad Y'_{rq}=-\omega\epsilon/k_{zrq}$$
$$k_{zrq}^2+(r\pi/1)^2+(q\pi/b)^2=k^2=\omega^2\mu\epsilon. \quad (16)$$

Based on (11)–(16), the total fields are given by

$$\left.\begin{matrix}E_{y1}^t\\H_{x1}^t\end{matrix}\right\}=\sum_{m=1,n=0}^{\infty}A_{mn}\big(\epsilon_{m0}\pm R_{mn}\exp(2jk_{zmn}^{TE}z)\big)\left\{\begin{matrix}E_{y1}^{TE}\\H_{x1}^{TE}\end{matrix}\right\} + \sum_{m=1,n=1}^{\infty}A'_{mn}R'_{mn}\exp(2jk_{zmn}^{TM}z)\left\{\begin{matrix}E_{y1}^{TM}\\H_{x1}^{TM}\end{matrix}\right\} \quad (17)$$

in region 1 with $\epsilon_{m0}=1$ for $m=1$ and 0 otherwise, and

$$\left.\begin{matrix}E_{y2}^t\\H_{x2}^t\end{matrix}\right\}=\sum_{r=1,q=0}^{\infty}A_{rq}\big(1\pm R_{rq}\exp(2jk_{zrq}^{TE}z)\big)\left\{\begin{matrix}E_{y2}^{TE}\\H_{x2}^{TE}\end{matrix}\right\} + \sum_{r=1,q=1}^{\infty}A'_{rq}\big(1\pm R'_{rq}\exp(2jk_{zrq}^{TM}z)\big)\left\{\begin{matrix}E_{y2}^{TM}\\H_{x2}^{TM}\end{matrix}\right\} \quad (18)$$

in region 2, and

$$\left.\begin{matrix}E_{y3}^t\\H_{x3}^t\end{matrix}\right\}=\sum_{s=1,t=0}^{\infty}A_{st}\left\{\begin{matrix}E_{y3}^{TE}\\H_{x3}^{TE}\end{matrix}\right\}+\sum_{s=1,t=1}^{\infty}A'_{st}\left\{\begin{matrix}E_{y3}^{TM}\\H_{x3}^{TM}\end{matrix}\right\} \quad (19)$$

in region 3.

It can be shown that it suffices to match all of the E_y and H_x components across the boundary planes $z=0$ and $z=w$. The other total field components then are matched, likewise [5].

The boundary conditions read

$$E_{y1}^t=E_{y2}^t, \quad \text{for } 0\leqslant x\leqslant 1 \quad E_{y1}^t=0, \quad \text{for } 1\leqslant x\leqslant a/2$$
$$H_{x1}^t=H_{x2}^t, \quad \text{for } 0\leqslant x\leqslant 1 \text{ and } 1\leqslant x\leqslant a/2. \quad (20)$$

They must be applied at $z=0$ and $z=w$. Multiplying these relations by $\sin(r\pi x/1)$ and integrating over the cross section yields, e.g., at $z=0$,

$$\sum_{m=1,3,5,\cdots}(\epsilon_{m0}+R_{m0})A_{m0}A_0b\cdot\int_0^1\sin(r\pi x/1)\sin(k_{xm1}x)\,dx=(1+R_{r0})A_{r0}b1/2 \quad (21)$$

$$\sum_{m=1,3,5,\cdots}(\epsilon_{m0}-R_{m0})A_{m0}A_0Y_{10}b\cdot\int_0^1\sin(r\pi x/1)\sin(k_{xm1}x)\,dx=(1-R_{r0})A_{r0}Y_{r0}b1/2. \quad (22)$$

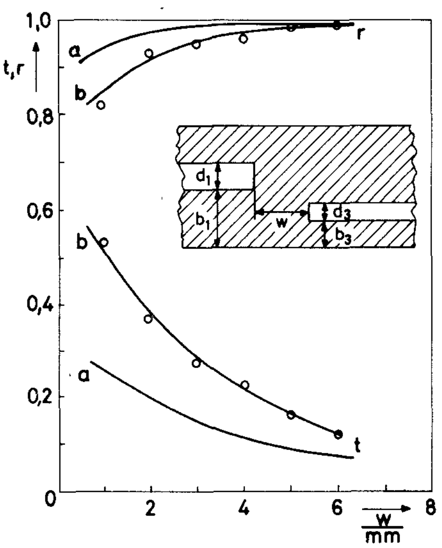

Fig. 4. Transmission coefficient t and reflection coefficient r of a metallic strip in a fin line versus the length w of the strip. Curves a: $d_1=d_3=3.1$ mm, $b_1=4.77$ mm, $b_3=0.03$ mm. Curves b: $d_1=d_3=b/2$, $b_1=b_3=b/4$. The common dimensions are $a=15.8$ mm, $b=7.9$ mm, $c=0.3$ mm, and $\epsilon_r=2.22$.

Two similar equations can be obtained from the boundary conditions at $z=w$. Thus the transmission and reflection coefficients turn out to depend only on the TE_{m0} mode of the fin line. Other modes are averaged out. Theoretical and experimental results for two different slot patterns are shown in Fig. 4.

IV. A Fin-Line Equivalent

We are in a position now to introduce an equivalent for a fin line. By regarding the expression for the guided wavelength ($n=0$)

$$\lambda_{gm0}=\lambda_0/\left(k_{em0}-(\lambda_0/\lambda_{cm0})^2\right)^{1/2} \tag{23}$$

with a λ_0 wavelength in free space, and by comparing it to the one for the guided wavelength of the TE_{m0} mode in a rectangular waveguide, one sees that the TE_{m0} modes of a fin line can be thought to be supported by a singularly infinite set of rectangular waveguides having the broad dimension $a_m=m\pi/k_{cm0}$ and being homogeneously filled with a dielectric of permittivity k_{em0}. This equivalence is sketched in Fig. 5.

A first-order design theory for fin-line structures can now be formulated based on this equivalence.

Assumptions: 1) The HE modes of a bilateral fin line might be replaced by their TE parts as in Section II (i.e., ϵ_r is moderate and $c/a \ll 1$). 2) The geometry of the boundary value problem allows neglecting all other modes except the TE_{m0} modes (as is, e.g., valid for the problem treated in Section III).

Conclusion: The bilateral fin line might then be replaced by a singularly infinite set of rectangular waveguides having a broad dimension a_m and equal height b and being homogeneously filled with a dielectric of permittivity k_{em0}. Each equivalent waveguide supports its dominant TE_{m0} mode. Its parameters a_m and k_{em0} are related to the fin-line modes by the analysis of Section II.

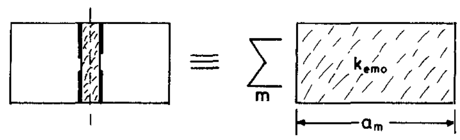

Fig. 5. A fin-line equivalent.

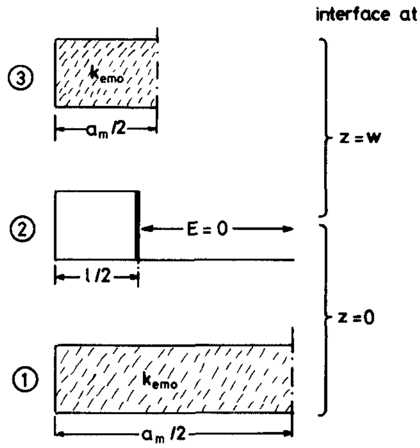

Fig. 6. A fin-line equivalent applied to the problem of Fig. 3.

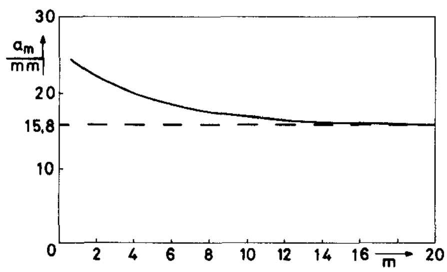

Fig. 7. Equivalent waveguide width a_m of a fin line versus m. The dimensions of the fin line are $a=15.8$ mm, $d=1.58$ mm, $b=7.9$ mm, $b_1=3.16$ mm, $c=0.3$ mm, and $\epsilon_r=2.22$.

This procedure leads to a considerable simplification in the mathematics. It can be further simplified by using closed form solutions for a_m and k_{em0}, which have been derived for $m=1$ in [3]. These formulas which can easily be extended to cover the case of $m\neq1$ represent approximations with an error of less than 5 percent.

As an example, we will apply the fin-line equivalent to the problem of a metallic strip in a fin line. Analyzing the structure of Fig. 3 now means matching the tangential field components between a set of rectangular waveguides, as has been shown for a fixed m in Fig. 6. The total field in region 1 at $z=0$ is given by

$$E_{y1}^{TE}=\sum_{m=1,3,5,\cdots}(\epsilon_{m0}+R_{m0})A_{m0}\sin(k_{xm0}x)$$

$$H_{x1}^{TE}=\sum_{m=1,3,5,\cdots}(\epsilon_{m0}-R_{m0})Y_{m0}A_{m0}\sin(k_{xm0}x). \tag{24}$$

The various constants have been defined in Section III. Similarly, the total field in region 2 reads

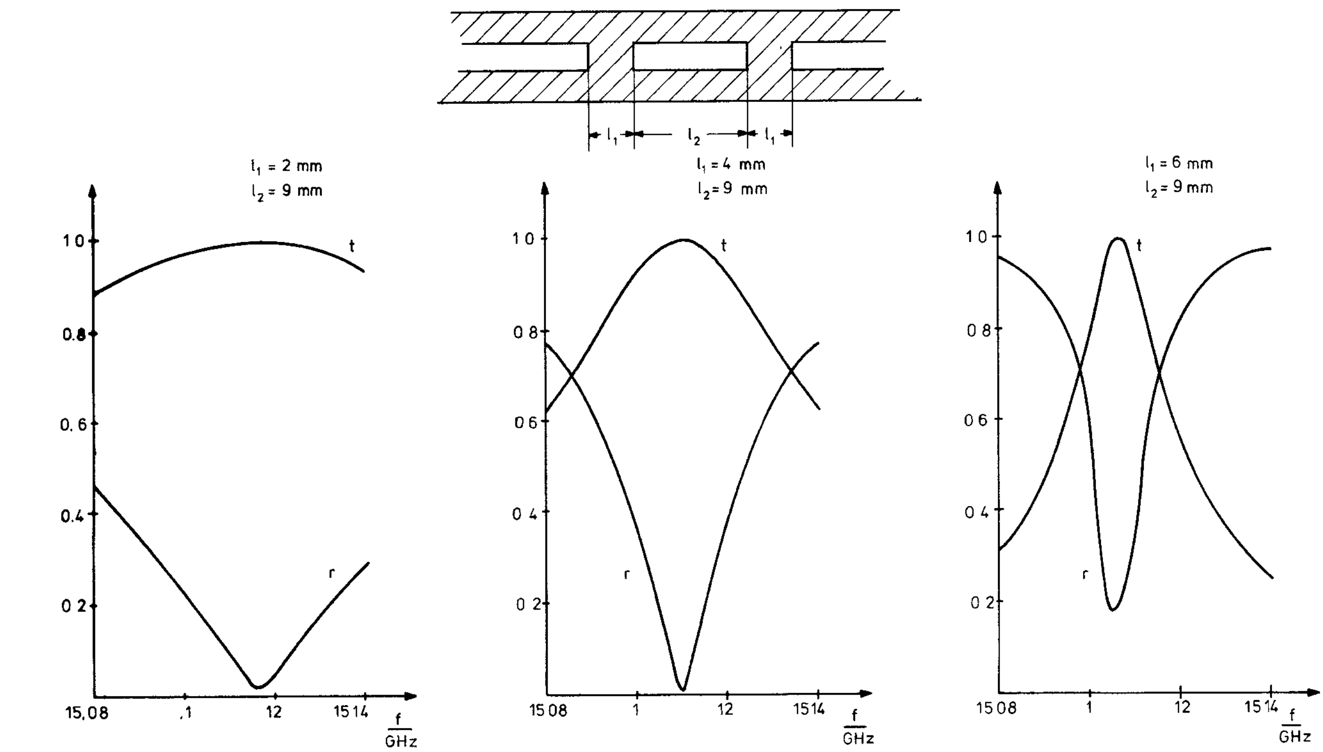

Fig. 8. Slot pattern of a bandpass filter and transmission coefficient t and reflection coefficient r versus frequency f for various strip widths. The dimensions of the fin line are $a=15.8$ mm, $d=1.58$ mm, $b=7.9$ mm, $b_1=3.16$ mm, $c=0.3$ mm, and $\epsilon_r=2.22$.

$$E^{\mathrm{TE}}_{y2} = \sum_{r=1,2,3,\cdots} (1+R'_{r0})A'_{r0}\sin(r\pi x/l)$$

$$H^{\mathrm{TE}}_{x2} = \sum_{r=1,2,3,\cdots} (1-R'_{r0})Y'_{r0}A'_{r0}\sin(r\pi x/l). \qquad (25)$$

Matching the tangential field components at the interface $z=0$, multiplying these equations by $\sin(k_{xm0}x)$, and integrating over the cross section yields

$$(\epsilon_{m0}+R_{m0})A_{m0}ba_m/4 = \sum_{r=1,2,3,\cdots}(1+R'_{r0})A'_{r0}b \cdot \int_0^{l/2}\sin(r\pi x/l)\sin(k_{xm0}x)\,dx$$

$$(\epsilon_{m0}-R_{m0})A_{m0}Y_{m0}ba_m/4 = \sum_{r=1,2,3,\cdots}(1-R'_{r0})A'_{r0}Y'_{r0}b \cdot \int_0^{l/2}\sin(r\pi x/l)\sin(k_{xm0}x)\,dx + \sum_{m=1,3,5,\cdots} A_{m0}(\epsilon_{m0}-R_{m0})Y_{m0} \cdot \int_{l/2}^{a_m/2}\sin^2(k_{xm0}x)\,dx. \qquad (26)$$

From these equations (and two similar ones at $z=w$) the reflection and transmission coefficients can be calculated. The analytical results from (26) on one side and from (21) and (22) on the other turn out to be identical, indicating that the problem of a metallic strip across a fin line can be completely described by the set of TE_{m0} modes.

The second sum in the last of the two equations of (26) represents the surface current density due to a discontinuity in the magnetic field component H_x. This term might be simplified considerably by assuming a constant surface current density, as is well known from similar problems (see, e.g., [10]). The dependence of a_m on m is shown for typical dimensions in Fig. 7. From this diagram a rapid convergence of a_m against a can be seen.

V. Validity of the Method

In order to check the validity of the method, it has been applied to different fin-line structures. In the case of a discontinuity in the slot width, the method is no longer exact, because in addition to the TE_{m0} modes other modes have an influence on the transmission and reflection coefficients. In most of the cases which have been investigated, the fin-line equivalent yielded results within the measurement accuracy. This is also true for circuits with unilateral fin lines. Work is in progress now to derive analytical expressions for the error which has to be expected when applying the method to various classes of problems.

Finally, a bandpass filter whose slot pattern is shown in Fig. 8 has been designed using the fin-line equivalent. The calculated and measured frequency response coincided within a 5-percent error band with one exception. The loaded Q factor of the transmission resonator showed deviations of up to 40 percent between theory and measurements. This is due to the neglect of ohmic losses in the calculations. The difficulty can be overcome, however, if the losses are taken into account according to the guidelines given in [3].

VI. Conclusions

A fin-line equivalent has been developed, which is thought to fill the gap for a first-order design theory. This method reduces boundary value problems in complex fin-line structures to the problem of matching the TE_{m0} modes between two sets of equivalent rectangular waveguides. Its usefulness has been checked by applying it to the analysis of fin-line discontinuities and of a bandpass filter.

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An *X*-Band Balanced Fin-Line Mixer

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Abstract—**The fin-line technique has been used in a balanced 9–11-GHz mixer with a 70-MHz intermediate frequency. The mixer without an IF amplifier has an available conversion loss of less than 5 dB with a 3.8-dB minimum and a SSB noise figure of less than 6.9 dB with a 5.3-dB minimum. The mixer is tunable by variable shorts. It is possible to scale the device to millimeter-wave frequencies.**

I. Introduction

THIS PAPER describes the design and performance of a microwave integrated-circuit (MIC) balanced mixer that covers the bandwidth of 2 GHz within the *X* band with available conversion losses of less than 5 dB and a noise figure of less than 6.9 dB. Not included is the noise contribution from the IF amplifier. The mixer operates with an IF of 70 MHz, but the device is able to handle higher IF's up to some gigahertz. For this purpose, the low-pass filter coupling out the intermediate frequency must have a suitable cutoff frequency.

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In the circuit considered here, a fin-line technique [1] has been used to realize a mixer which is capable to work well up to millimeter-wave frequencies. To this end the mixer is equipped with connections of rectangular waveguides both at the signal and the local oscillator input.

Because the fundamental mode of a fin-line (H_{10} mode) is the same as the one of a rectangular waveguide, transitions between these two guides are easy to handle and have a very small insertion loss and a VSWR over the entire waveguide bands. Parasitic radiation which often is a problem connected with planar waveguides especially at higher frequencies can be avoided. So the fin-line has very low losses. Moreover, it offers the same possibilities of integration as other planar circuits.

The most essential part of the mixer is a planar magic T completely integrated in a rectangular waveguide. The magic T proved itself as a rather broad-band and low-loss device. The purpose of the magic T is twofold. First, it distributes the signal and local oscillator voltages with their proper phase relationships to the two nonlinear elements, and, second, it blocks the local oscillator input from the signal frequency input and vice versa.

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