A SIMPLE METHOD FOR ANALYZING FIN-LINE STRUCTURES

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ABSTRACT

A simple description of fin-line circuits is introduced by establishing a correspondence between rectangular waveguides and a fin-line. The usefulness of the method is demonstrated by designing a bandpass filter.

Introduction

Integrated fin-line is an advantageous alternative to microstrip in the design of microwave integrated circuits at frequencies above 10 to 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, what makes the method cumbersome if being applied to an analysis of fin-line structures with varying slot pattern, the second paper does not show any way how to calculate the field distributions in the cross section of the waveguide. There is hence a lack for a first-order design theory as it already exists for microstrip circuits, which allows analyzing even complex fin-line structures with 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_{mo}-modes) is sufficient to analyze a wide class of fin-line circuits with varying slot pattern. Concerning these modes the fin-line can be described by equivalent rectangular waveguides.

The investigations to be presented are fourfold: In the first part 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 part 2. By matching the various tangential field components at the interface, the TE_{mo}-modes of the fin-line (which are of course different from the TE_{mo}-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 part 3, where directions will be given of how to replace a fin-line by an equivalent set of rectangular waveguides. Analysis of the same transition as in part 2 using this equivalent description then yields identical results. Finally part 4 is devoted to a discussion of the validity of the method.

1.Fin-line Eigenmodes

The propagating waves are in fin-lines neither TE nor TM but a combination of both. This is due to the dielectric substrate of integrated fin-lines (see cross section in fig.1!). It is impossible to simultaneously match all 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, as can be concluded from [4].

Since the hybrid eigenmodes of the fin-line contain TE and TM terms, they will be classified in the following way:

A HE-mode means an 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 means an hybrid eigenmode with a dominating TM part. Now the TE part vanishes at cutoff.

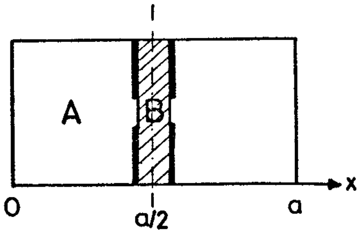

Fig.1 Fin-line cross section

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 [5]. Our calculations will be restricted to the case that ε_r is moderate and $c/a \ll 1$.

The eigenvalues have been calculated for the cross section of fig.1. At cutoff this configuration can be regarded as a discontinuity between two parallel plate waveguides which are shot-circuited at x=0 and open- or short-circuited, respectively, at x=a/2. Such step discontinuities have approximately been analyzed for the case of two infinitely long transmission lines in [6] by neglecting any frequency dependence of the phase constants. We have solved the problem accurately by expanding the fields in the two regions A and B and by matching the tangential field components at the interface. 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.

The characteristic equation has been solved twice: a first time for $\varepsilon_r=1$ and a second time for $\varepsilon_r=2.22$ (RT-Duroid as substrate material). The effective dielectric constant k_{emn} is then given by

$$k_{emn}=k_{cmn}^2(\varepsilon_r=1)/k_{cmn}^2(\varepsilon_r=2.22)$$

with k_{cmn} the wave number at cutoff [1,3].

The effective dielectric constant depends on frequency but very weakly, as has been shown in [2]. Hence the propagation constant can be written as $k_{zmn}^2=k_{emn}\cdot k^2-k_{cmn}^2(\varepsilon_r=1)$ where $k^2=\omega^2\mu\varepsilon$ and k_{emn} is taken from the relation above.

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

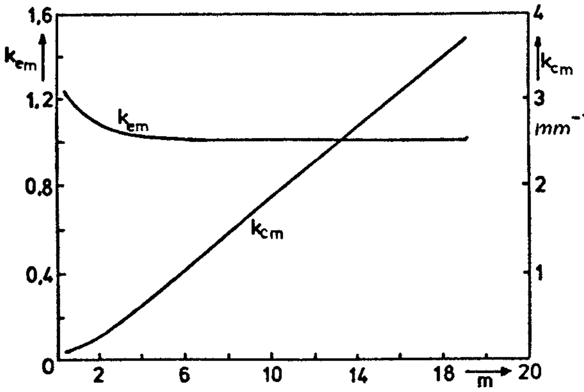

Fig.2 k_{em} and k_{cm} for TE_{mo} versus m

2. Metallic Strip In 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. It can, however, be shown that it suffices matching the total E_y- and H_x-components across the boundary planes z=0 and z=w. The other total field components are then matched, likewise [8]. From the matching conditions, the transmission and reflection coefficients turn out to depend only on the TE_{mo}-mode of the fin-line. Other modes are averaged out.

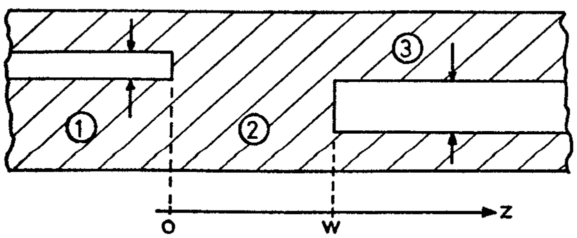

Fig. 3 Metallic strip in fin-line

Theoretical and experimental results for two different slot patterns are shown in fig. 4.

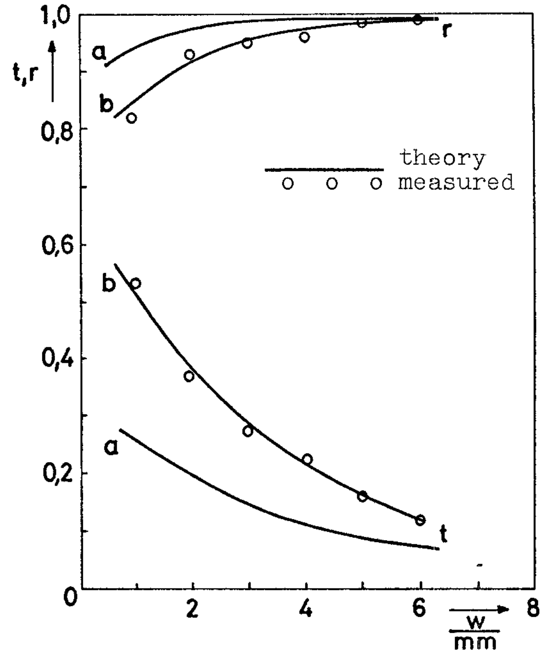

Fig. 4 Transmission t and reflection r of a metallic strip in fin-line
a: no symmetry with respect to y=b/2
b: symmetrical with respect to y=b/2

3. 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)

$$\lambda_{gm} = \lambda_o/(k_{em}-(\lambda_o/\lambda_{cm})^2)^{0.5}$$

with λ_o wavelength in free space, $\lambda_{cm}=2\pi/k_{cm}$ and by comparing it to the one for the guided wavelength of the TE_{mo}-mode in a rectangular waveguide, one sees that the TE_{mo}-modes of a fin-line can be thought to be supported by a singly infinite set of rectangular waveguides having the broad dimension

$a_m = m\pi/k_{cm}$ and being homogeneously filled with a dielectric of permittivity k_{em}. This equivalence is sketched in fig. 5.

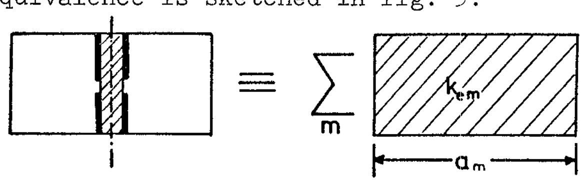

Fig. 5 Fin-line equivalent

Analyzing the structure of fig.3 now means matching the tangential field components

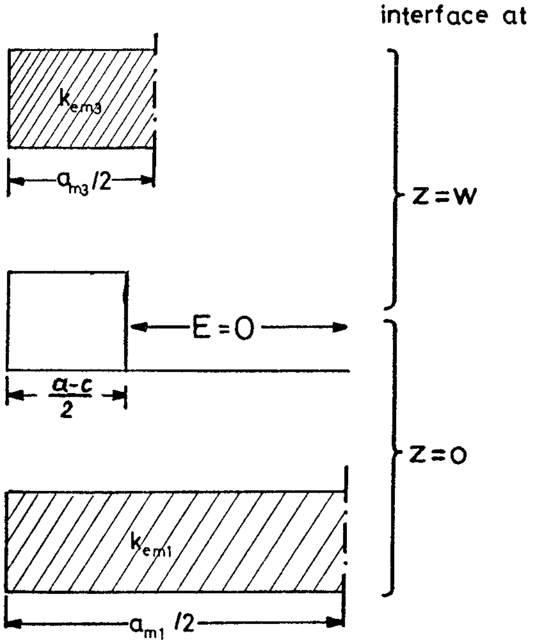

Fig. 6 Fin-line equivalent applied to the problem of fig. 3

between a set of rectangular waveguides, as has been shown for a fixed m in fig. 6. The analytical results obtained from this procedure are identical to those obtained from the mode matching method in section 2.

The dependence of a_m on m is shown for typical geometric parameters in fig. 7. From this diagram a rapid convergence of a_m against a can be seen.

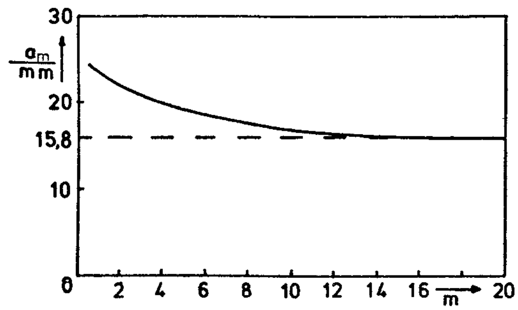

Fig. 7 Equivalent waveguide width a_m versus m

4. 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_{mo}-modes other modes have an influence on the transmission and reflection coefficients, too. In most of the cases, which have been investigated, the fin-line equivalent yielded results within the measurement accuracy. Work is in progress now, to derive analytical expressions, which enable one to estimate 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. Calculated and measured frequency response coincided within a 5 per cent error band with one exception: The loaded Q-factor of the transmission resonator showed deviations of up to 40 per cent between theory and measurements due to a neglect of ohmic losses in the calculations.

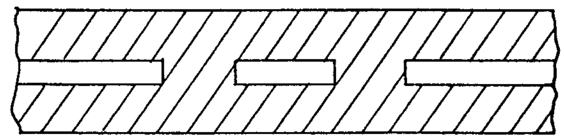

Fig. 8 Slot pattern of a bandpass filter

This difficulty can be overcome, however, if the losses are taken into account according to the guidelines given in [3].

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_{mo}-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 bandpass filters.

Acknowledgement

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