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NEW STRUCTURES FOR IMPEDANCE TRANSFORMATION IN FIN-LINES

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Summary

Cascaded transitions between unilateral and bilateral fin-line show unique features for impedance transformation. This is illustrated with a new broadband switch circuit, which has been realized by c.a.d..

Introduction

The basic building blocks of fin-line circuits are various discontinuities in the slot width. We have recently shown[1] how impedance transformation can be performed with either one or two steps in the slot width. With two cascaded steps one can generate either a notch or a strip. Almost all known components are realized in this way. These structures can be analyzed by combining an eigenmode with a modal analysis[2]. The procedure has been carried through in Reference 1. We will apply this method here to new configurations.

The structures for impedance transformation to be described show both electrical and practical advantages over the known ones. Their slot patterns are sketched in Fig.1. These structures consist of two cascaded transitions between unilateral and bilateral fin-lines of equal slot widths. The slot

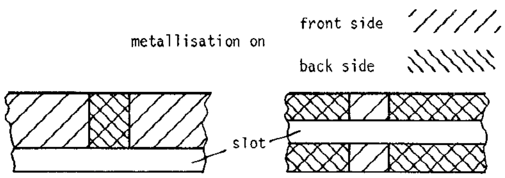

Fig.1 Slot patterns of transitions between unilateral and bilateral fin-lines

may be located either symmetrically or unsymmetrically with respect to the waveguide axis. A large range of impedances can be generated by varying two geometrical parameters: The common slot width 2s and the length 2l of the middle section. Such a line section can therefore be used in either of two ways: as an impedance transformer or as a semiconductor device mount. In the latter application, the circuit patterns show a practical advantage over conventional ones: While the circuit at the front side which contains the semiconductor devices is protected against damage, one can carelessly alter the transforming section on the back side of the substrate, in order to optimize performance. In addition, there are, however, even electrical advantages over the usual notch and strip patterns, which will be derived in the following.

Theory

In order to analyse the structures shown in Fig.1, one must know both the propagation constants and the field distributions of the hybrid eigenmodes of unilateral and bilateral fin-lines. The problem has been solved by a number of authors. We have adopted the spectral domain technique presented e.g. in Reference 3 and modified it to comprehend in a unified form the eigenmodes of both unilateral and bilateral fin-line. For the latter case, one must also take the odd modes into account (here 'odd' is defined with respect to that axis which is perpendicular to the substrate plane), because these modes are excited at a junction between unilateral and bilateral fin-line.

The modal analysis for computing the characteristics of an abrupt transition between a bilateral and a unilateral fin-line shall be briefly described. As shown in the slot patterns of Fig.1, both the slot widths and the location of the slots above the broad wall of the waveguide are assumed to stay constant at both sides of the junction. The electric and magnetic fields of the i-th eigenmode are written as

$$e_i(x,y,z) = a_i\, \bar{e}_i(x,y)\, e^{\pm\gamma_i z} ,\qquad h_i(x,y,z) = a_i\, \bar{h}_i(x,y)\, e^{\pm\gamma_i z} \tag{1}$$

with $\bar{e}_i$ and $\bar{h}_i$ the transverse vector functions of the electric and magnetic field, respectively. Denoting the transverse field distribution of the left waveguide 'a' at the junction $(z=0)$ by $\bar{E}_a$, $\bar{H}_a$ and expanding it in terms of the eigenmodes of that waveguide reads

$$\bar{E}_a = (1+\rho)\, a_1\bar{e}_{a1} + \sum_{i=2}^{\infty} a_i\bar{e}_{ai} ,\qquad \bar{H}_a = (1-\rho)\, a_1\bar{h}_{a1} - \sum_{i=2}^{\infty} a_i\bar{h}_{ai} . \tag{2}$$

Similarly, one writes for waveguide 'b'

$$\bar{E}_b = \sum_{j=1}^{\infty} b_j(\bar{e}_{bj} + \sum_{k=1}^{\infty} s_{jk}\bar{e}_{bk}) ,\qquad \bar{H}_b = \sum_{j=1}^{\infty} b_j(\bar{h}_{bj} - \sum_{k=1}^{\infty} s_{jk}\bar{h}_{bk}) . \tag{3}$$

ρ means reflection coefficient of the incident mode $(i=1)$, s_{jk} are the scattering coefficients of the next discontinuity located at $z>0$ in waveguide 'b'.

In the following we will regard the so-called 'boundary reduction problem' (terminology taken from Reference 2), i.e. the cross section of waveguide 'a' is larger than that of waveguide 'b'. In our case, the former is a unilateral and the latter a bilateral fin-line. The boundary enlargement problem can be treated in the same way as described below, if the subscripts 'a' and 'b' in (2) and (3) are interchanged.

Eqs. (2) and (3) are used to formulate the boundary conditions at the junction, which are manipulated in the following way: The cross product of the electric field in (2) with $\bar{h}_{am}$ is taken and integrated over the cross section of waveguide 'a'. For the unknown field on the left-hand side, $\bar{E}_b$ is inserted from (3). Similarly, the cross product of the magnetic field in (2) with $\bar{e}_{bn}$ is taken and integrated over the cross section of waveguide 'b'. For the unknown field on the left-hand side, H_b is inserted from (3). One obtains

$$(1+\rho)a_1 \int_{(a)} \bar{e}_{a1}\times\bar{h}_{am}\cdot u_z dxdy + \sum_{i=2}^{\infty} a_i \int_{(a)} \bar{e}_{ai}\times\bar{h}_{am}\cdot u_z dxdy = \sum_{j=1}^{\infty}\left[b_j\left(\int_{(b)} \bar{e}_{bj}\times\bar{h}_{am}\cdot u_z dxdy + \sum_{k=1}^{\infty} s_{jk}\int_{(b)}\bar{e}_{bk}\times\bar{h}_{am}\cdot u_z dxdy\right)\right], \quad (4a)$$

$$(1-\rho)a_1 \int_{(b)} \bar{e}_{bn}\times\bar{h}_{a1}\cdot u_z dxdy - \sum_{i=2}^{\infty} a_i \int_{(b)} \bar{e}_{bn}\times\bar{h}_{ai}\cdot u_z dxdy = \sum_{j=1}^{\infty}\left[b_j\left(\int_{(b)} \bar{e}_{bn}\times\bar{h}_{bj}\cdot u_z dxdy - \sum_{k=1}^{\infty} s_{jk}\int_{(b)}\bar{e}_{bn}\times\bar{h}_{bk}\cdot u_z dxdy\right)\right]. \quad (4b)$$

u_z means unit vector in z-direction. Because $\bar{E}_a$ exists on the common aperture only and vanishes elsewhere, the integrals on the right-hand side of (4a) must only be taken over the cross section of waveguide 'b'. These integrals can be solved once the eigenmodes have been formulated.

In order to solve (4) for the amplitudes of the excited modes, one must first determine the scattering coefficients s_{jk}. For a single transition waveguide 'b' may be assumed to be terminated in a matched load: $s_{jk}=0$. For two transitions in cascade, we can take advantage of the symmetry of the structure with respect to the plane $z=1$. The equivalent circuit can then be found from both an even and an odd excitation of both ports, so that $s_{jk}=0$ for $j \neq k$ and

$$s_{jj} = (1-y_{bj}) \,/\, (1+y_{bj}) = \pm\exp(-2j\beta_{bj}l). \quad (5)$$

y_{bj} is the normalized input admittance of the j-th mode in waveguide 'b' measured at $z=0$, β_{bj} means propagation constant of this mode.

The single transition is characterized by the equivalent circuit of Fig.2 whose elements are computed from the input admittances at both

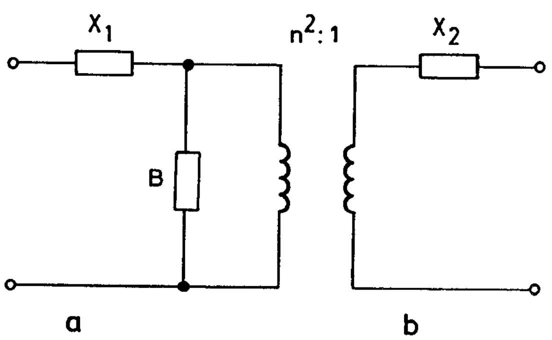

Fig.2 Equivalent circuit of a single transition (port 1 = bilateral, port 2 = unilateral fin-line)

ports. Two cascaded transitions are represented by a T-equivalent circuit with series reactances X_{sc} and shunt reactance $(X_{oc}-X_{sc})/2$. (Subscripts 'sc' denote a short circuit, subscripts 'oc' an open circuit at the symmetry plane $z=1$.)

Results

The modal analysis has been applied to a single transition between unilateral and bilateral fin-line and to two cascaded transitions of the types sketched in Fig.1. Some results for the elements of the equivalent circuit of a single transition (Fig.2) are shown in Fig.3. The turns ratio of the ideal impedance transformer has been omitted because it is very close to unity. The general shape of the 3 remaining elements (note that subscript '1' refers to the bilateral fin-line port) can be explained

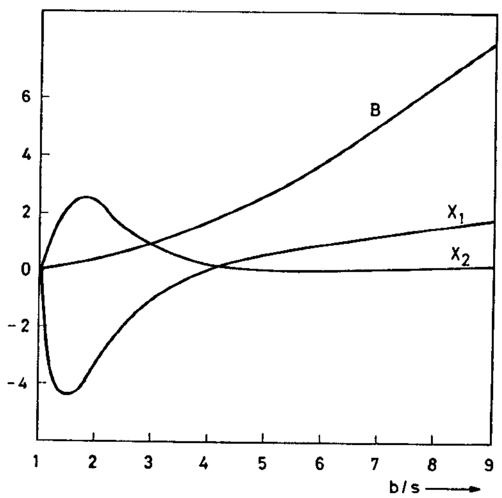

Fig.3 Elements of the equivalent circuit of a single transition (Fig.2) versus normalized slot width The elements are normalized against the wave impedance of the adjacent port. (frequency 30 GHz, WR-28 waveguide with 2a=7.112 mm, 2b=3.556 mm, symmetrically located slot of width 2s, RT-duroid 5880 substrate of thickness 2d=0.254 mm and relative dielectric constant ε_r= 2.22)

by assuming an electric field perpendicularly directed through the dielectric substrate from front to back fin of the bilateral fin-line. This field can of course only exist in the immediate vicinity of the transition between the two fin-lines. For larger slot widths, d is much smaller than s with 2d the substrate thickness. Hence, a capacitive series reactance X_1 appears in the equivalent circuit. The shunt susceptance B is small because $b \approx s$. - The electric field concentration in the slot increases with decreasing s. Hence, B must increase monotonically and X_1 must increase, too. Finally, X_1 becomes inductive because of an increasing magnetic field concentration which is due to a surface current filament on the transversely oriented fin edge of the bilateral fin-line.

The equivalent circuit elements of two cascaded transitions (T-circuit, X_{sc} = series reactance, X_p = parallel reactance, subscript 'sc' = short circuit, subscript 'oc'

= open circuit) are shown in Fig.4 versus the length of the middle section. The slot width is relatively small: s = 0.2 mm. The shape of the curves can be explained by cascading the equivalent circuits of two single transitions. (They have, however, been computed by taking higher-order mode coupling into account.) A significant feature is the parallel

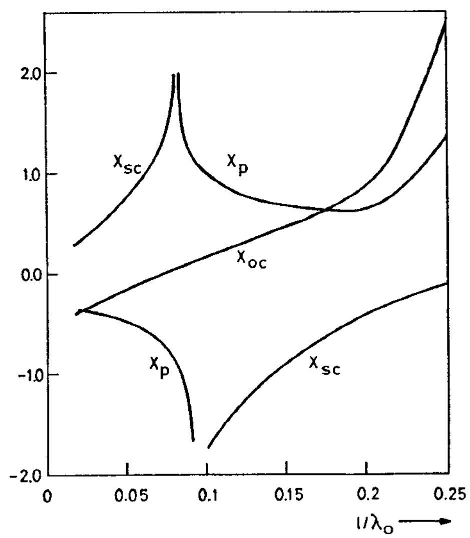

Fig.4 Elements of the equivalent T-circuit of a section of bilateral fin-line (length 2l) embedded into a unilateral fin-line
The elements are normalized against the wave impedance of the unilateral fin-line. λ_o means free space wavelength.
(parameters as in Fig.3 with 2s = 0.4 mm)

resonant behaviour of the series reactance for small lengths l, which can be explained if the two cascaded transitions are separately regarded (i.e. if higher-order mode coupling between them is neglected). The series reactance is calculated for odd excitation so that one port of the equivalent circuit of Fig.2 is terminated in a short-circuited transmission line of length l. Hence, the parallel resonant behaviour of X_{sc} is due to B resonating with X_1 and the input reactance of the transmission line.

The series reactance is capacitive for larger lengths l. This is of great importance for circuit design because there is no other discontinuity known showing a capacitive series reactance. (Both notch and strip show an inductive series reactance [1].) This unique feature of a double transition between unilateral and bilateral fin-line offers an interesting alternative in the design of pin-diode switches as will be illustrated below.

Application

For an illustration a broadband microwave switch with two pin-diodes will be designed. The slot pattern of the switch is shown in Fig.5. It consists of two sections of bilateral fin-line within a unilateral fin-line, which are properly spaced. The diodes are directly soldered across the slot. The equivalent circuit is shown in Fig.6. Here the diodes have been modelled by an inductive reactance (on-state) and by a capacitive reactance (off-state). The equivalent T-circuit of the bilateral fin-line section consists of capacitive series elements and an inductive shunt element. The design of the equivalent circuit elements of such a switch is treated in Reference 4. What we are concerned with here, is the realization of this equivalent circuit in fin-line. The transmission state of the switch corresponds to the off-state of the diodes. The corresponding equivalent circuit is then designed to represent a bandpass filter with e.g. a maximally flat response. This is achieved if X = 2B. X and B are defined in Fig.6. The isolation state of the switch corresponds to the on-state of the diodes. Then B must be series-tuned by the capacitive series reactances of the bilateral fin-line section. This determines the length of the section. The same task could not be fulfilled by any other discontinuity because they altogether produce purely inductive series reactances.

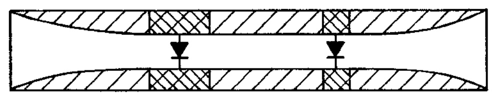

Fig.5 Slot pattern of a switch with two pin-diodes

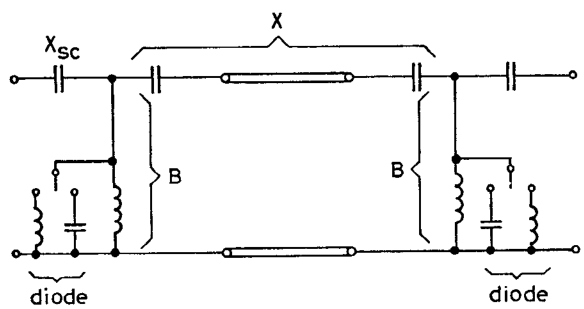

Fig. 6 Equivalent circuit of the switch of Fig.5

The design of the switch proceeds in the following way: The slot width of the unilateral fin-line is chosen so that it corresponds to the physical size of the diode. Then the diodes can directly be soldered across the slot without generating further parasitic elements. The two steady-state impedances of the pin-diode are then measured in a bilateral fin-line of equal slot width. The length of the bilateral fin-line sections in the switch circuit can now be calculated from the condition of isolation. The series-resonant frequencies of the two sections with the diodes in on-state should differ from another in order to enhance the isolation. These frequencies are related to the midband frequency in Reference 4. Hence, the two sections show different lengths. - Only one parameter has still to be fixed: The distance between the sections. It is found from the condition of transmission X = 2B.

A switch with two pin-diodes has been designed and realized in Ka-band. Its insertion loss is about 1 dB, the isolation amounts to about 20 dB without readjusting the slot pattern.

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