time. The time from 50-percent logic to 10-percent detected RF signal is divided into 40-ns delay time and 15-ns fall time. The maximum power handling capability was not measured, but it can be assumed that 26–30-dBm power handling is possible due to the diodes total power dissipation of 24 dBm.

IV. Conclusion

It has been shown that the design of finline mixers and p-i-n diode attenuators is possible at frequencies around 140 GHz with acceptable results. Besides the electrical data, excellent mechanical stability is achieved with this approach. For example, a finline balanced mixer on a quartz substrate with even larger dimensions (E-band) than in this case has been successfully exposed to shock accelerations up to 30 000 g, a value unthinkable for whisker-contacted mixers.

Furthermore, production is made easier and cheaper by the planar circuit in conjunction with beam-lead diodes. Last, but not least, integration techniques have become possible, including E-plane circulators, leading to very compact front ends with losses lower than the sum of the discrete components as demonstrated at 94 GHz [4], [6].

References

[1] K. Solbach, "The status of printed millimeter-wave E-plane circuits," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-31, pp. 107–121, Feb. 1983.
[2] P. J. Meier, "Millimeter integrated circuits suspended in the E-plane of rectangular waveguide," *IEEE Trans Microwave Theory Tech.*, vol MTT-26, pp. 726–733, 1978.
[3] R. N. Bates, S. J. Nightingate, and P. M. Ballard, "Millimeter-wave E-plane components and subsystems," *Radio Electron. Eng.*, vol. 52, pp 506–512, 1982.
[4] W. Menzel and H. Callsen, "Integrated finline components and subsystems at 60 and 94 GHz," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-31, pp. 142–146, Feb. 1983.
[5] H. Meinel and H. Callsen, "Fin-line p-i-n diode attenuators and switches for the 94 GHz range," *Electron. Lett.*, vol. 18, pp. 541–542, 1982.
[6] W. Menzel, H Callsen, and K. Solbach, "An integrated receiver front-end for 94 GHz dual polarization radar," in *Proc. 13 Europ. Microw. Conf.* (Nürnberg), 1983, pp. 142–147.
[7] R. S. Tahim, G. M. Hayashibara, and K. Chang, "Design and performance of W-band broad-band integrated circuit mixers," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-31, pp. 277–283, Mar. 1983.
[8] H. Meinel and H. Callsen, "Fin-line detectors for frequencies up to 260 GHz," in *Proc. 7th Intern. Conf on Infrared and Millimeter Waves* (Marseille), 1983, p. 34.
[9] H Barth, "141 GHz generation by a GaAs gunn-oscillator upconverter chain," in *Proc. IEEE MTT-S Conf. 1982* (Dallas), pp. 364–366.
[10] H. Barth, "140 GHz impatt oscillators," private communications, 1983.
[11] K. Solbach, "E-plane circulators aid millimeter-wave design," *Microwave & RF*, pp. 73–79, Dec 1983
[12] W. Menzel, "A 140 GHz balanced mixer for fin-line integrated circuits," in *Proc. 13th Europ Microwave Conf* (Nürnberg), 1983, pp. 179–182.
[13] L.-P. Schmidt, T. Itoh, and H Hofmann, "Characteristics of unilateral fin-line structures with arbitrarily located slots," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-29, pp. 352–355, 1981

Radial-Line/Coaxial-Line Stepped Junction

A. G. WILLIAMSON, SENIOR MEMBER, IEEE

***Abstract*—A radial-line/coaxial-line junction having a step in the inner conductor in the coaxial aperture is considered. It is shown that the stepped junction may be modeled by an equivalent circuit obtained by a simple modification of that for the unstepped junction. Comparison of theoretical and experimental results has established this representation to be very accurate over a wide frequency range.**

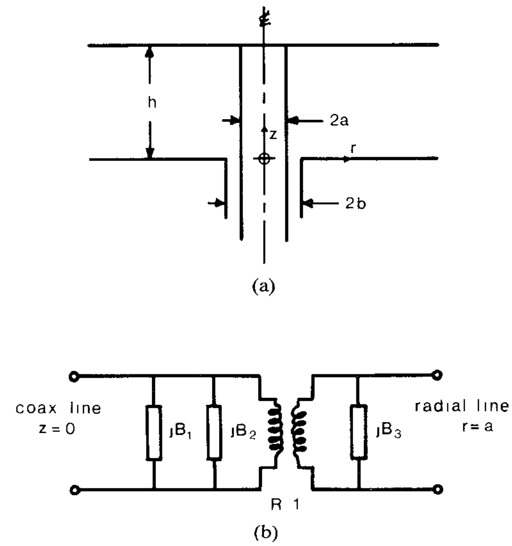

Fig. 1. Radial-line/coaxial-line junction: (a) cross-sectional view, (b) equivalent circuit.

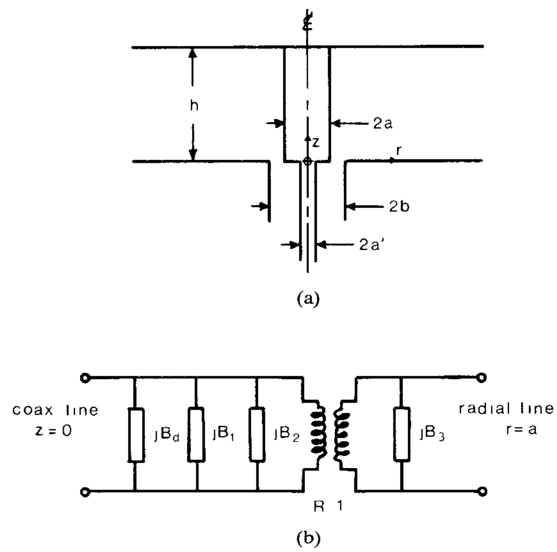

Fig. 2. Radial-line/coaxial-line stepped junction: (a) cross-sectional view, (b) equivalent circuit.

I. Introduction

There are many microwave devices which employ, in some form or other, a radial-line/coaxial-line junction (e.g., antenna feeds, power dividers, and combiners). In designing such devices, it is advantageous to have an equivalent circuit representation for the junction. However, in spite of the wide application of radial-line/coaxial-line junctions, there have been relatively few studies reported. Indeed, it is only recently that equivalent circuits have been developed for the junction [1]–[3].

Recently, Allison [2], [3] investigated the design of wide-band transitions. He concluded that for the coaxial line he was employing, transitions of the form shown in Fig. 1(a) were unsuitable for wide-band performance, and that it was necessary to enter the

Manuscript received March 12, 1984; revised July 16, 1984.
The author is with the Department of Electrical and Electronic Engineering, University of Auckland, Private Bag, Auckland, New Zealand, currently on leave at the Department of Electrical Engineering and Electronics, University of Liverpool, P.O. Box 147, Liverpool L69 3BX, England.

0018-9480/85/0100-0056$01.00 ©1984 IEEE

radial line at a radius greater than that of the inner conductor of the coaxial line. This led him to the development of a transition of the form shown in Fig. 2(a).

In this paper, an equivalent circuit is deduced for the stepped junction (Fig. 2(a)) by a simple modification of the equivalent circuit for the unstepped junction (Fig. 1(a)). The deduced equivalent circuit predicts results for the input admittance which are in excellent agreement with experimental results.

The equivalent circuits presented here relate to the axially symmetric situation (i.e., no circumferential variation of the fields) where only the dominant modes can propagate in either the coaxial or radial lines (i.e., $h < \lambda/2$). In addition, it is assumed that the transmission lines are lossless and have an air dielectric. Generalization of the results to other dielectrics is straightforward.

II. Equivalent Circuit for Radial-Line/Coaxial-Line Junction

The radial-line/coaxial-line junction shown in Fig. 1(a) can be represented by the equivalent circuit given in Fig. 1(b) where [1]

$$B_1 = -\frac{2\pi}{\eta_0 \ln(b/a)} \cot(kh) \tag{1}$$

$$B_2 = \frac{4\pi}{kh\eta_0 \ln^2(b/a)} \sum_{m=1}^{\infty} \frac{1}{q_m^2} \frac{K_0(q_m kb)}{K_0(q_m ka)} \cdot \left(I_0(q_m ka) K_0(q_m kb) - I_0(q_m kb) K_0(q_m ka)\right) \tag{2}$$

$$B_3 = \frac{2\pi ka}{\eta_0 kh} \cdot \frac{J_1(ka)Y_0(kb) - J_0(kb)Y_1(ka)}{J_0(ka)Y_0(kb) - J_0(kb)Y_0(ka)} \tag{3}$$

and

$$R = \frac{(2/\pi)\ln(b/a)}{J_0(ka)Y_0(kb) - J_0(kb)Y_0(ka)} \tag{4}$$

where

$$q_m = \sqrt{\left(\frac{m\pi}{kh}\right)^2 - 1} \tag{5}$$

and $k = 2\pi/\lambda$, η_0 is the intrinsic impedance of free space, and J_0, J_1, Y_0, Y_1, I_0, and K_0 are Bessel functions and modified Bessel functions of the first and second kinds.

The above parameters were deduced from analyses which used the TEM approximation for the electric field in the coaxial aperture. Recently, an analysis has been undertaken without approximation of the aperture field. It has shown [4], [5] that the form of the exact equivalent circuit is also given by Fig. 1(b), and that B_1 and B_3 are independent of the form of the coaxial aperture electric field. Only R and B_2 are dependent on the coaxial aperture electric field, and, in most problems of interest, R is not very sensitive to the precise form of the field.

Only the result for B_2, which relates to the susceptance contribution of the cutoff radial-line modes [4], [5], is affected to any extent by the use of the TEM approximation. Even so, the effect of the TEM approximation is not all that great, as will be seen shortly. Thus, the equivalent circuit shown in Fig. 1(b) with the parameters as above is a quite accurate model.

III. Junction with Step in Aperture Plane

Consider now the stepped junction shown in Fig. 2(a), and let us consider how the equivalent circuit shown in Fig. 1(b) would need to be modified in order to model this new problem.

Compare the two junctions shown in Figs. 1(a) and 2(a), considering them excited at the coaxial ports. Now in both cases the radial-line fields are excited by the fields in the coaxial aperture from $r = a$ to $r = b$. The essential difference between the two problems is the step on the inner conductor, and the resulting "fringing effect" this produces. In the unstepped case, the coaxial aperture electric field is singular at $r = b$, while in the stepped case, the field is singular at both $r = a$ and $r = b$. The dominant mode field in the radial line will be almost identical in the two cases (for the same voltage excitation) as it is relatively insensitive to the form of the coaxial aperture electric field [4], [5]. Significant differences only occur in the extent of the excitation of the higher order modes. Since these higher order modes are cutoff, their effect is very local to the junction. The differences between the higher order mode fields of the two problems is even more localized to the junction, and in particular to the region of greatest difference between the exciting coaxial aperture electric fields of the two cases, i.e., near the step.

Because the dominant mode fields excited in the radial line in both cases are almost the same, so will be the input conductance. The input admittance of the two junctions will thus significantly differ only in the susceptive component. If we denote the susceptive difference as B_d, we could therefore represent the stepped junction by the equivalent circuit shown in Fig. 2(b). Having deduced the form of the stepped-junction equivalent circuit, it now only remains to determine an appropriate value for B_d, which accounts for the disturbance caused by the step on the inner conductor.

If the step was located not in the aperture, but some distance further down the coaxial line, we could account for the step as a discontinuity in the coaxial line [6]. Moreover, the step would not have to be very far from the aperture for this approach to be valid. Even when the step is in the aperture plane the distortion of the fields caused by the step will still be quite localized to the step. The major effects of the abrupt ending of the coaxial line at $z = 0$ have already been accounted for in the other terms of the equivalent circuit. As such, it would seem that a reasonable value for B_d could be obtained from the discontinuity capacitance C_d of a step (from $r = a'$ to $r = a$) in a coaxial line having an undisturbed outer radius of b.

Let us therefore take

$$B_d = \omega C_d \tag{6}$$

and obtain C_d from [6, p. 49].

In order to test this representation consider a stepped junction having $a = 1.75$ cm, $b = 2.20$ cm, $h = 1.60$ cm with a short circuit located at a radius of 4.70 cm for various values of a', namely 0.60, 1.00, 1.27, and 1.75 cm, and over a frequency range from 1.2 for 2.6 GHz. Theoretical results for the input susceptance at the coaxial port obtained from the equivalent circuit given in Fig. 2(b) are shown in Fig. 3. Also shown are experimental results obtained by the author. These results were obtained by slotted-line techniques in the coaxial line where the VSWR is very high, and thus the input susceptance is very easily and accurately obtained because of the very low (theoretically zero) input conductance resulting from short-circuiting the radial line.

The case $a' = 1.75$ cm considered in Fig. 3 corresponds to the unstepped junction. Note the good agreement between the theoretical and experimental results. This is the basis for the statement earlier that the equivalent circuit given in Fig. 1(b), with the parameters B_1, B_2, B_3, and R as given, provides a quite accurate representation for the unstepped junction.

Now considering Fig. 3 further, it can be seen that the agreement between the theoretical and experimental results for the stepped-junction cases is also very good. Note that the dif-

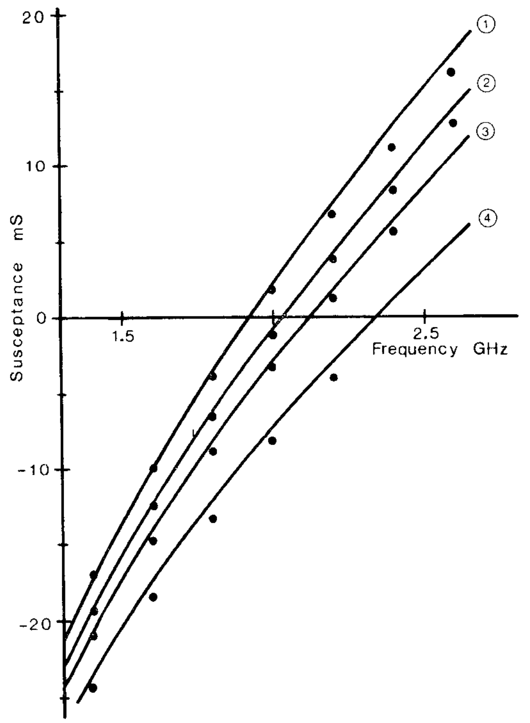

Fig. 3. Input susceptance at the coaxial port of a stepped junction having $a = 1.75$ cm, $b = 2.20$ cm, $h = 1.60$ cm with a short circuit at a radius of 4.70 cm for various values of a', namely ① $a' = 0.60$ cm, ② $a' = 1.00$ cm, ③ $a' = 1.27$ cm, ④ $a' = 1.75$ cm. —— theoretical results, ··· experimental results.

ferences between the theoretical results for different values of a' agree very well with the differences between the corresponding experimental results, that is, the representation of the additional discontinuity effects by (6) is quite satisfactory. Note also that the experimental results are, in general, slightly less than the theoretical results. This aspect will be further discussed shortly.

Let us now use the equivalent circuit (Fig. 2(b)) to consider the stepped junction investigated by Allison *et al.* [2], [3] which had $a = 0.283$ cm, $a' = 0.152$ cm, $b = 0.350$ cm, and $h = 0.254$ cm with the radial line matched. The admittance presented to the coaxial line is shown in Fig. 4 for the frequency range to 18 GHz. (*NB.* In terms of fractions of a wavelength, the dimensions of the cases considered in Figs. 3 and 4 are similar.) The admittance results predicted from the equivalent circuit are given by the solid conductance (G) curve and the dashed susceptance (B) curve. Also shown in Fig. 4 are experimental results from Allison's study [2], [3].

Considering Fig. 4, one sees excellent agreement for the conductance results and reasonable agreement for the susceptance results, although the difference between theory and experiment increases with increasing frequency. At the high frequency end of the study the junction is electrically quite large and the TEM approximation is not as satisfactory as it is at lower frequencies.

If one knew the actual aperture electric field one could correct for the use of the TEM approximation. Such corrections have been calculated for narrow aperture coaxial cases [7] but a wide range of data for other cases is not available. Fortunately, the case under consideration here has a ratio $b/a \simeq 1.2$, that is, the annular gap is quite narrow. Thus, the result in [7] may be used to slightly modify B_2 for the over estimate provided by the TEM approximation. Let us therefore apply a correction ΔB_2 to the value of B_2 given by (2) of

$$\Delta B_2 = -\frac{4ka}{\eta_0}\ln(4/\pi). \tag{7}$$

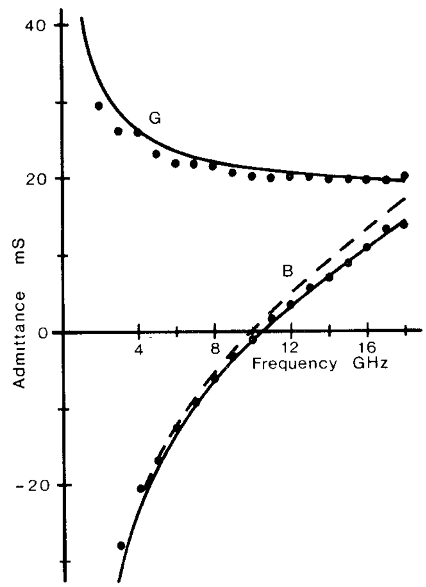

Fig. 4. Input admittance at the coaxial port of a stepped junction having $a = 0.283$ cm, $a' = 0.152$ cm, $b = 0.350$ cm, $h = 0.254$ cm with the radial line matched. ---- theoretical results (without B_2 correction), —— theoretical results (with B_2 correction), ··· experimental results.

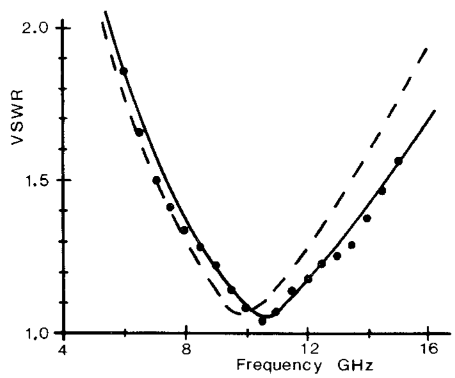

Fig. 5. VSWR of results shown in Fig. 4. ---- theoretical results (without B_2 correction), —— theorectical results (with B_2 correction), ··· experimental results.

Theoretical results for the susceptance of the junction using the modified value of B_2 are given by the solid curve in Fig. 4. (The conductance results are unchanged of course.) The improved agreement between theoretical and experimental results is easily seen.

For near matched situations (as here), the VSWR is a sensitive gauge to the accuracy of the equivalent circuit. The VSWR of the stepped junction is shown in Fig. 5. Computed results (both with and without the correction to B_2) are shown together with the measured results of Allison *et al.* [2], [3]. The agreement between the computed (especially the B_2 corrected) and measured results is excellent, thereby establishing the equivalent circuit to be very accurate.

IV. Conclusion

The radial-line/coaxial-line stepped junction has been considered, and it has been shown that it may be accurately modeled by the equivalent circuit for the unstepped junction with the addition of a discontinuity susceptance. Comparison of theoretical and experimental results has shown this representation to be very accurate over a wide frequency range.

References

[1] A. G. Williamson, "Equivalent circuit for radial-line/coaxial-line junction," *Electron. Lett.*, vol. 17, pp. 300–301, 1981.
[2] R. C. Allison, "A matched coaxial-radial transmission line junction," M.S. thesis, University of California, Los Angeles, 1977.
[3] R. C. Allison, R. L. Eisenhart, and P. T. Greiling, "A matched coaxial-radial transmission line junction," in *IEEE Int. Microwave Symp. Dig.*, 1978, pp. 44–46.
[4] A. G. Williamson, "Radial-line/coaxial-line junctions," University of Auckland, School of Engineering, Rep. no. 332, 1984.
[5] A. G. Williamson, "Radial-line/coaxial-line junctions—Analysis and equivalent circuits," to be published in the *Int. J. Electron.*
[6] P. I. Somlo, "The computation of coaxial line step capacitances," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-15, pp. 48–53, 1967.
[7] D. C. Chang and T. T. Wu, "A note on the theory of end-corrections for thick monopoles," *Radio Sci.*, vol. 3 (new series), pp. 639–641, 1968.

New Analysis of Semiconductor Isolators : The Modified Spectral Domain Analysis

S. TEDJINI AND E. PIC

Abstract—**This paper addresses semiconductor isolators of the field displacement effect type. The semiconductor is modelized by its surface impedance tensor. This description allows an extension of the well-known spectral domain method to the analysis of the semiconductor isolators. Different configurations are studied and numerical results are given. The finline isolator with InSb is shown to be the best choice—indeed, insertion loss is less than 3 dB/cm and isolation is greater than 18 dB/cm. Experimental results supporting these calculations will be published in a following paper.**

I. Introduction

In view of the increasing use of the millimeter-wave band, one of the urgent problems of today is the development of nonreciprocal devices. The purpose of this paper is to describe some millimeter-wave semiconductor isolators using the field displacement effect.

The field displacement effect in gyroelectric structures at room temperature was demonstrated by Hirota and Suzuki [1]–[3] during 1969–1971 and is now well known.

The basic structure is a rectangular waveguide loaded in its E-plane by a thin slab of high-mobility semiconductor (such as InSb) which is transversally magnetized. The field displacement effect result from the superposition of TE_{m0} modes excited by the displacement current and the $TM_{m,n}$ modes excited by the conduction current on the semiconductor. The resulting electromagnetic field shows an asymetric distribution along the y-axis (see Fig. 1) which reverses with the direction of the wave propagation (or with the direction of the magnetostatic field).

This basic is a reciprocal one. In order to obtain some nonreciprocity, it is necessary to introduce a geometrical dissymmetry. Two nonreciprocal structures will be specially considered: the air-gap semiconductor loaded waveguide isolator, and the finline isolator.

Manuscript received March 2, 1984; revised July 9, 1984.
S. Tedjini is with the Laboratoire d'Electromagnétisme, Unité Associée au CNRS no. 833, ENSERG, 23 rue des Martyrs, 38031 Grenoble Cedex, France.
E. Pic is with the Laboratoire de Génie Physique, BP 46, Domaine Universitaire, 38402 St. Martin D'heres, Cedex, France.

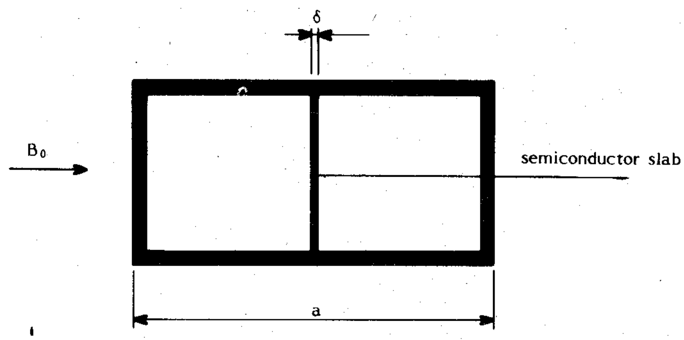

Fig. 1. Basic structure. In the E-plane of the waveguide we have a semiconductor slab submitted to magnetostatic field B_0. The semiconductor thickness is $\delta \ll a$.

A unified method will be presented to study the above structures. It is an extension of the well-known spectral domain technique (SDT) which takes into account the losses of the structure. It is then possible to compute the attenuation constant, and to derive the main parameters of the isolator: insertion losses and isolation.

In this paper (the first of two parts), the theoretical treatment is given with some details, together with computational results.

Experimental setups and results will be described in a second paper (to be published).

II. Historical Review

There are two main classes of nonreciprocal semiconductor devices. In the first class, the semiconductor sample is longitudinally magnetized (the static magnetic field is parallel to the propagation direction). Such structures have been considered in [12], [13]. For instance, in [12], an InSb semiconductor at 77 K (mobility = 48 $m^2/V \cdot s$, conductivity = 1.5 10^3 s/m) with a length l = 28.9 mm exhibits a 2-dB insertion loss and a 30-dB isolation (frequency was 35 GHz and magnetic field 0.2 T). These results are very interesting. Unfortunately, the realization of such a device is difficult. Two drawbacks are low temperature operation (77 K) and longitudinal magnetization.

In the second class, the semiconductor slab is transversally magnetized. Operation at room temperature may be considered. This requires a very thin semiconductor plate to overcome the losses due to the semiconductor conductivity.

Several methods have been proposed to compute the propagation constant of a waveguide partially filled by a semiconductor and submitted to a magnetostatic field.

For instance, Gabriel and Brodwin [4] have used a perturbation method, where the nondisturbed field is the TE_{10} mode of the waveguide. The semiconductor introduces a perturbation operator: the field perturbation is projected on the basis of the eigenmodes of the waveguide. However, this method is not valid for a strong perturbation, namely for a high semiconductor conductivity.

Hirota and Suzuki [2] used a variational method assuming a lossless semiconductor. This method determines the phase constant and exhibits the field displacement effect. But their calculation did not take into account the dielectric substrate.

Another pertinent treatment is the Schelkunoff method [6]. This method has been used by Arnold and Rosenbaum [7], and by J. L. Amalric [8] in the case of anisotropic inhomogeneous waveguide. This last author gave an interesting discussion of the advantages and drawbacks of this method.

0018-9480/85/0100-0059$01.00