Tapered Slotline Antennas at 802 GHz

Pranay R. Acharya, *Student Member, IEEE,* Hans Ekström, Steven S. Gearhart, Stellan Jacobsson, *Member, IEEE*
Joakim F. Johansson, Erik L. Kollberg, *Fellow, IEEE,* and Gabriel M. Rebeiz, *Member, IEEE*

Abstract— **Tapered endfire slotline antennas, of the BLTSA type, have been fabricated on 1.7 μm thin SiO_2/Si_3N_4 dielectric membranes. Antenna patterns of the E-, H-, and D-planes have been measured at 802 GHz. The –10 dB beamwidths were found to be approximately 40° in all planes, with side lobe levels below –11 dB (–19 dB in the E-plane). The cross-polarized peaks in the D-plane are 8 dB below the co-polarized peak. A theoretical model for calculating the E- and H- plane patterns of tapered slotline antenna has been extended to include the co- and cross-polarized D-planes. Measured and calculated patterns show good agreement.**

I. Introduction

The growing interest in submillimeter wave technology, e.g. for radio astronomy, has led to the development of a variety of planar antennas, such as the integrated horn antenna [1], the corner reflector antenna [2], the log-periodic antenna [3], the double dipole antenna [4], the double slot antenna [5], the dielectric waveguide antenna [6], and various types of Tapered Slotline Antennas (TSA) [7]–[10]. The TSA family belongs to the group of endfire travelling wave antennas. The planar geometry of the TSA allows it to be easily integrated with other planar devices such as filters, SIS and Schottky-diode mixers, or bolometers. The TSA can be designed to produce symmetrical radiation patterns, with a half-power beam width between 20° and 40°, despite the planar structure. The performance of the antenna is determined both by its geometry and by the thickness and permittivity of the supporting dielectric substrate. There is a trade-off in substrate thickness; a too "thick" substrate launches surface modes, which degrade the antenna performance, whereas a too "thin" substrate gives higher cross-polarization and less symmetrical beam patterns. The optimum substrate thickness, t, has experimentally been found to be $t \approx \gamma\lambda_o\left(\sqrt{\varepsilon_r}-1\right)^{-1}$ [8], where ε_r is the permittivity of the supporting substrate and $0.005 \leq \gamma \leq 0.03$. Hence, in the submillimeter wave region, the optimum substrate thickness is a few micrometers, which makes the antenna fabrication complicated.

There are several different designs of the TSA; the Linearly Tapered Slotline Antenna (LTSA) [7,8], the Constant Width Slotline Antenna (CWSA) [8], the exponentially tapered slotline antenna ("Vivaldi") [9], and the Broken Linearly Tapered Slotline Antenna (BLTSA) [10], which is an antenna with a tapered slot consisting of three linear sections (Fig. 1). The TSAs generally exhibit a high cross-polarization level in the diagonal plane (D-plane), which is a common disadvantage of this antenna type. In this paper, we report measurements on a BLTSA at 802 GHz (the frequency of a convenient laser line). This antenna exhibits a lower (around 2 dB) cross-polarization ratio in the diagonal plane than the LTSA and the CWSA, due to a shorter distance between the phase centers of the E- and H-planes. Furthermore, the BLTSA design can provide smaller dimensions than the Vivaldi, and therefore the fragile supporting membrane can be made smaller.

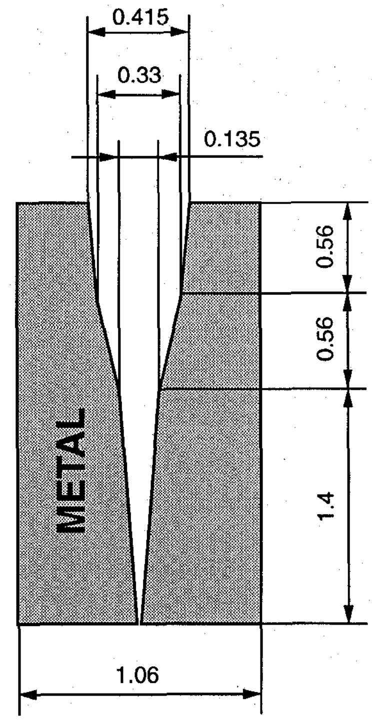

Fig. 1. Dimensions (in millimeters) of the broken linearly tapered slotline antenna (BLTSA) at 802 GHz.

II. Theory

Previously, the TSA radiation patterns have been theoretically predicted in the E- and H-planes [11]. However, for a better understanding of the slotline antenna, it is important to have knowledge of the D-plane pattern as well. Therefore, in this paper, the theoretical model has been extended to predict the co- and the cross-polarization patterns in the D-plane.

In this analysis the TSA is approximated by a number of slot sections (usually five sections per free space wavelength)

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P.R. Acharya, H. Ekström, S. Jacobsson and E.L. Kollberg are with Chalmers University of Technology, Dept. of Radio and Space Science, S-412 96 Göteborg, Sweden. J.F. Johansson is with Chalmers University, Dept. of Radio and Space Science.
S.S. Gearhart and G.M. Rebeiz are with NASA/Center for Space Terahertz Technology, Dept. of Electrical Engineering and Computer Science, University of Michigan, Ann Arbor, MI 48109–2122, USA.
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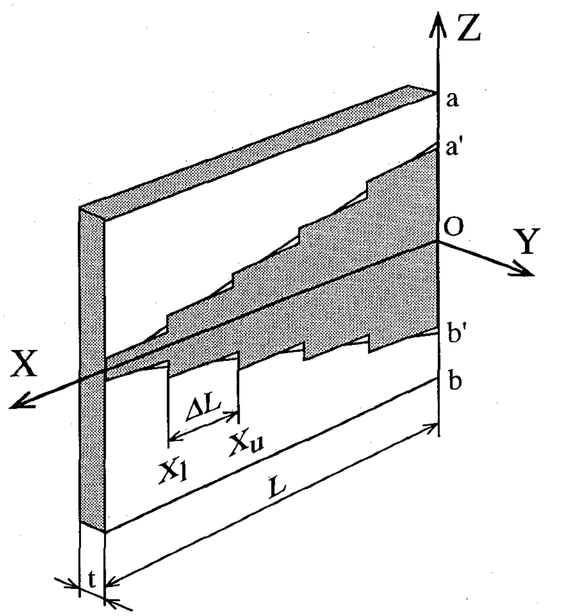

Fig. 2. Linearly tapered slotline antenna (LTSA) with step approximation.

of different widths and lengths, as shown in Fig. 2. The impedance and the effective wavelength of the slot changes from section to section, and are determined by the slot width and the properties of the supporting dielectric substrate. It is assumed that the lateral edges of the antenna (points a and b in Fig. 2) are at infinity, and that the small discontinuities between the slot sections do not generate higher order modes. The fields in the different slot sections are related through power conservation and the effects of the discontinuities, due to the step approximation, are taken into account by adding a reflected wave, calculated by using the theory of small reflections. The far field electric field (1) is obtained by integrating the aperture distribution, $E_z(x,z)$ [11] over each slot with a far field pattern Green's function, $f(x,z,\psi)$.

$$E(\psi) = \iint_S f(x,z,\psi) E_z(x,z)\, dx\, dz \tag{1}$$

The far field pattern function is calculated using the Green's function of a radiating slot situated near the edge of a semi-infinite metal sheet [12]. The aperture distribution is determined by taking the dielectric substrate into account, which is ignored in the far-field Green's function. A previously developed computer program [13] was used to calculate the effective wavelength and characteristic impedance of the slotline.

However, for permittivities less than 10 and thin substrates, it is also possible to use approximate expressions for the effective wavelength and impedance of the slotline [14]. The closed form far-field expressions of the electric field in the E-, H-, and D-planes for the ith section of the slot (of length ΔL^i) are given by:

$$\begin{aligned} E_E^i(\psi) = {} & C\sqrt{\frac{Z_s^i}{\cos\psi}}\, J_o\left(\frac{k_o w^i}{2}\sin\psi\right) \\ & \cdot\left[e^{-j\phi^i}\frac{\left\{F(p_2 k_o x_u^i) - F(p_2 k_o x_l^i)\right\}^*}{\sqrt{p_2}} \right. \\ & \left. + \Gamma^i e^{j\phi^i}\frac{\left\{F(p_3 k_o x_u^i) - F(p_3 k_o x_l^i)\right\}}{\sqrt{p_3}} \right] \end{aligned} \tag{2}$$

$$\begin{aligned} E_H^i(\psi) = {} & C\sqrt{Z_s^i}\cos\frac{\psi}{2} \\ & \cdot\left[\frac{e^{-j\phi^i}}{p_2}\left(\sqrt{2}\sin\frac{\psi}{2}\left\{F(p_1 k_o x_u^i)e^{jp_2 k_o x_u^i} \right.\right.\right. \\ & \left. - F(p_1 k_o x_l^i)e^{jp_2 k_o x_l^i}\right\} \\ & \left. + \sqrt{p_4}\left\{F(p_4 k_o x_u^i) - F(p_4 k_o x_l^i)\right\}^*\right) \\ & + \Gamma^i\frac{e^{j\phi^i}}{p_3}\left(-\sqrt{2}\sin\frac{\psi}{2}\left\{F(p_1 k_o x_u^i)e^{-jp_3 k_o x_u^i}\right.\right. \\ & \left. - F(p_1 k_o x_l^i)e^{-jp_3 k_o x_l^i}\right\} \\ & + \sqrt{p_5}\left\{F(p_5 k_o x_u^i)\right. \\ & \left.\left.\left. - F(p_5 k_o x_l^i)\right\}\right)\right] \end{aligned} \tag{3}$$

$$\begin{aligned} E_{D_{cross}^{co}}^i(\psi) = {} & \pm C\sqrt{Z_s^i}\cos\frac{\psi}{2}\, J_o\left(\frac{k_o w^i}{2}\frac{\sin\psi}{\sqrt{2}}\right) \\ & \cdot\left[\frac{e^{-j\phi^i}}{p_2}\left(\sin\frac{\psi}{2}\left\{F(p_1 k_o x_u^i)e^{jp_2 k_o x_u^i}\right.\right.\right. \\ & \left. - F(p_1 k_o x_l^i)e^{jp_2 k_o x_l^i}\right\} \\ & + C_F\sqrt{p_4}\left\{F(p_4 k_o x_u^i)\right. \\ & \left.\left. - F(p_4 k_o x_l^i)\right\}^*\right) \\ & + \Gamma^i\frac{e^{j\phi^i}}{p_3}\left(-\sin\frac{\psi}{2}\right. \\ & \cdot\left\{F(p_1 k_o x_u^i)e^{-jp_3 k_o x_u^i}\right. \\ & \left. - F(p_1 k_o x_l^i)e^{-jp_3 k_o x_l^i}\right\} \\ & + C_R\sqrt{p_5}\left\{F(p_5 k_o x_u^i)\right. \\ & \left.\left.\left. - F(p_5 k_o x_l^i)\right\}\right)\right] \end{aligned} \tag{4}$$

where $F(x)$ is the Fresnel integral of the form

$$F(x) = \int_0^x \frac{e^{-jt}}{\sqrt{2\pi t}}\, dt \tag{5}$$

and the other parameters are given as:

$$p_1 = (\sin\theta - \cos\psi)$$
$$p_2 = (\lambda_o/\lambda_s^i - \cos\psi)$$
$$p_3 = (\lambda_o/\lambda_s^i + \cos\psi)$$
$$p_4 = (\lambda_o/\lambda_s^i - \sin\theta)$$
$$p_5 = (\lambda_o/\lambda_s^i + \sin\theta)$$
$$\sin\theta = \begin{cases} 1 & \text{for the H-plane} \\ \sqrt{\frac{1+\cos^2\psi}{2}} & \text{for the D-plane} \end{cases}$$
$$\phi^i = k_o \Delta L^i \cos\psi + k_s^i (L - (N-i)\Delta L^i) + \Delta L^i \sum_{n=1}^{i-1} k_s^n$$
$$C_{F_{Cross}}^{Co} = \frac{\sqrt{\frac{\sin\theta+\cos\psi}{1+\cos\psi}}}{2p_4 \sin\theta} \cdot \left[\frac{\lambda_o}{\lambda_s^i}(2\sin\theta - \cos\psi \pm 1) \mp \cos\psi - 1\right]$$
$$C_{R_{Cross}}^{Co} = \frac{\sqrt{\frac{\sin\theta+\cos\psi}{1+\cos\psi}}}{2p_5 \sin\theta} \cdot \left[\frac{\lambda_o}{\lambda_s^i}(2\sin\theta - \cos\psi \pm 1) \pm \cos\psi + 1\right]$$
$$k_0 = 2\pi/\lambda_o \qquad k_s^i = 2\pi/\lambda_s^i$$

Notation:

- C a constant
- ψ the angle from boresight, $0 \le \psi < \pi (0 \le \psi < \pi/2$ for the E-plane)
- $*$ complex conjugate
- i the slot section index
- x_u^i, x_l^i the upper and lower x limits of the i'th radiating slot, see Fig. 2
- λ_0, λ_s^i the free space wavelength and effective wavelength in the i'th slot section
- $\pm/\mp$ the upper and lower signs correspond to the co- and cross-polarization, respectively
- Γ^i the total reflection coefficient seen from the sides of the i'th narrow slot
- $J_0(x)$ the Bessel function of the first kind of order zero
- N the total number of slot sections
- ΔL^i the length of the ith slot section

III. Fabrication

The membrane is made of a 1.7 μm thick dielectric layer of SiO_2/Si_3N_4, which is deposited on a double-side polished silicon substrate [15]. The membrane, with dimensions slightly larger than the antenna, is formed by anisotropic EDP-etching of the silicon substrate. The membrane is supported by the thick silicon along three sides, leaving the endfire direction of the antenna undisturbed, see Fig. 3. The membrane is therefore fragile. The antenna, made of chrome-gold, and bismuth micro-bolometer detector are patterned on the etched membrane by lift-off. Fig. 1 shows the dimensions of the BLTSA antenna, and Fig. 3 the 5 μm wide bismuth micro bolometer in the 10 μm wide antenna slot. The total length of the antenna is 6.75 λ_0, the width is 2.83 λ_0, and the opening angle is 9 degrees. Since larger membranes are more fragile, antenna designs are limited to membrane areas smaller than approximately 3 × 7 mm^2 (>300 GHz for the BLTSA).

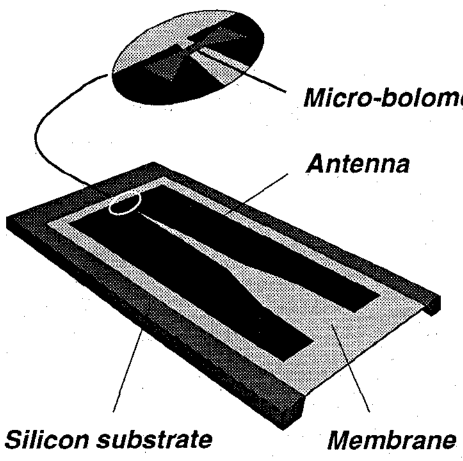

Fig. 3.Broken linearly tapered slotline antenna (BLTSA) on a $SiO2/Si_3N_4$ membrane supported on three sides by a silicon substrate.

IV. Measurements

The E-, H-, and D-planes of the 802 GHz antenna have been measured using a far-infrared laser as signal source. The antenna was placed in the far-field of the output Gaussian beam to ensure a plane wave incidence. The bolometer impedance was $150 \pm 20\Omega$ with a responsivity of 8 V/W at 100 mV bias, and the S/N ratio was better than 25 dB.

The -10 dB beamwidths of the measured patterns at 802 GHz were $\approx 40°$ for all three planes, and the side lobe levels in the E-, H-, and D-plane were -19 dB, -11 dB, and -10 dB, respectively. The cross-polarized level in the D-plane, which is known to be high for this type of antenna, was 8 dB below the co-polarized peak (Fig. 4). By taking into account the measured patterns in the E-, H-, and D-planes, the directivity and the feed efficiency of the antenna were calculated to be around 13 dB and 50%, respectively.

The feed efficiency is calculated for a plane wave incident on a parabolic reflector and peaks in this case for a subtended half-angle of $\psi = 20°$, which corresponds approximately to the -10 dB level in the far field radiation pattern. Similar calculations for a conical horn measured at 40 GHz yielded a feed efficiency of 66%.

The 802 GHz antenna had a narrower and more circular beam than a scaled antenna measured at 348 GHz [15]. Furthermore, the D-plane cross-polarized level was 2 dB lower in the 802 GHz design than for the 348 GHz antenna. The relatively thicker membrane at 802 GHz is the reason for the improved radiation pattern at this frequency. According to numerical simulations [16], a 1.7 μm membrane thickness should give better antenna patterns around 3 THz, with a

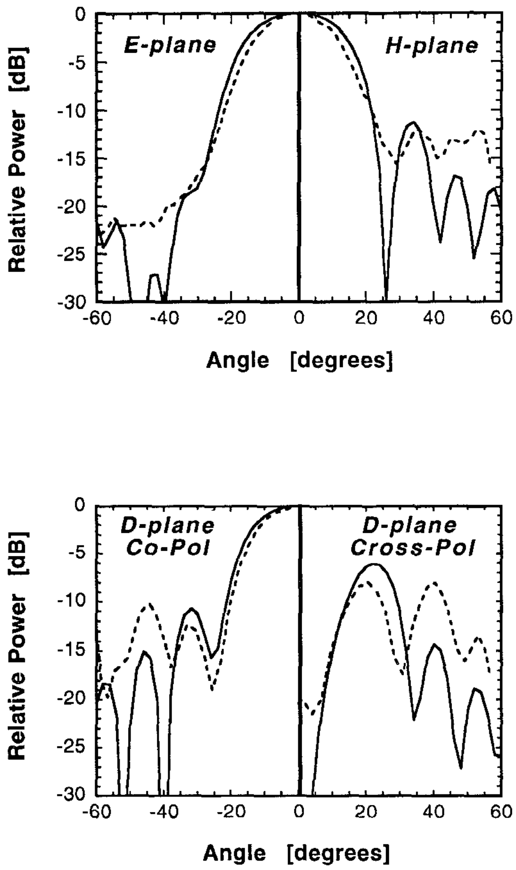

Fig. 4. Radiation patterns of a BLTSA at 802 GHz. Calculated (solid line) and measured (dashed line).

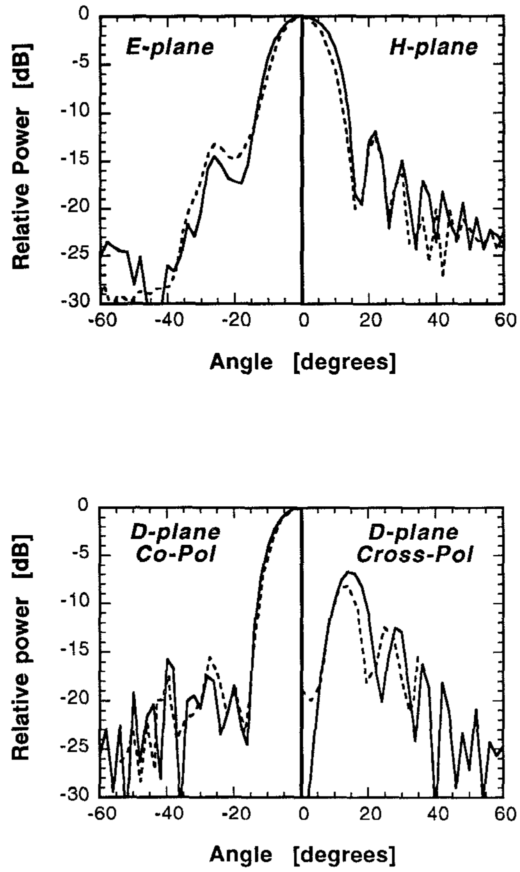

Fig. 5. Radiation patterns at 802 GHz for a BLTSA designed for 348 GHz (see text). Calculated (solid line) and measured (dashed line).

directivity and D-plane cross-polarization level around 15 dB and −13 dB, respectively. The predicted feed efficiency is then 60% for a subtended half-angle of 14°.

The antenna scaled to 348 GHz and built on a 1.7 μm thick membrane [15] was also measured at 802 GHz. The 348 GHz antenna is identical in shape to Fig. 3 but with a scale factor of 2.305. This antenna is of course very long in terms of the wavelength at 802 GHz, but, as seen in Fig. 5, still provides good patterns at this higher frequency. The 802 GHz antenna patterns, as well as the patterns of the 348 GHz antenna measured at 802 GHz, are compared with calculated patterns (Figs. 4 and 5), and show good agreement. The beamwidths and the side lobe levels are predicted to within 4° and 3 dB, respectively. The predicted D-plane cross-polarized level is generally 2 dB higher than the measured value.

V. Conclusions

The BLTSA on a thin (1.7 μm) dielectric membrane shows very good performance at 802 GHz, with a symmetric beam and low sidelobes. Measured E-, H-, D-plane antenna patterns are accurately theoretically predicted. These antennas can be fabricated on 0.5–1.5 μm thick membranes for frequencies up to 10 THz.

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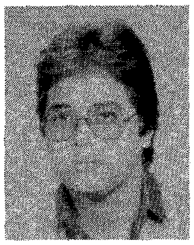

Pranay Raj Acharya (S'83) was born in Nepal on November 25, 1962. He received his M. Eng. in Radio Electronics from the Higher Institute of Mechanical and Electrical Engineering, Varna, Bulgaria, in 1987. In 1992 he received the licentiate of Engineering degree from Chalmers University of Technology, Gothenburg, Sweden. His research interests include design and manufacture of slot-line antennas on thin substrates and in sandwich structures at millimeter and submillimeter wavelengths.

Hans Ekström received his M.Sc. and Licentiate degrees from Chalmers University of Technology, Sweden, in 1987 and 1992, respectively. His research interests include design and fabrication of integrated superconducting mixers and planar antennas for microwaves and millimeter waves.

Steven S. Gearhart, photograph and biography not available at the time of publication.

Stellan Jacobsson received his M.Sc. and Ph.D. degrees from the School of Electrical Engineering at Chalmers University of Technology in 1979 and 1988, respectively. His present research activities are in the fields of millimeter and submillimeter wave technology.

Joakim F. Johansson (S'83–M'88) was born in Töllsjö, Sweden, on November 3, 1959. He received the degrees of Civilingenjör (M.Sc.), Tekn. Lic., and Tekn. Dr. (Ph.D.) in Electrical Engineering from Chalmers University of Technology, Gothenburg, Sweden, in 1983, 1986, and 1988, respectively.

He is currently a Research Fellow at the Department of Radio & Space Science with Onsala Space Observatory, Chalmers Univ., doing research on millimeter and sub-millimeter wave radio astronomy instrumentation and quasi-optics.

Dr. Johansson is a member of Eta.

Erik L. Kollberg (M'83–SM'83–F'91) received his Teknologie Doktor degree in 1970 from Chalmers University of Technology, Gothenburg, Sweden, where he was employed as associate professor 1974. He because professor at the Electrical Engineering Department of Chalmers 1979. He was on six months sabbatical leave to CalTech, Pasadena, until March 1991.

Most of his work has been focused on low noise receiver technology for applications in radio astronomy at Onsala Space Observatory, and he has published more than 150 papers. From 1963 to 1976 he made research on low-noise maser amplifiers. Various types of masers were developed for the frequency range 1 to 35 GHz. In 1972 research was initiated on low-noise Schottky diode mixers, and in 1981 also on millimeter and submillimeter wave superconducting quasiparticle mixers (SIS). Recently he has broadened his interest in high Tc superconducting circuits and semiconductor devices, including various types of diodes and three terminal devices.

Dr. Kollberg was the winner of the 1982 Microwave Prize given at the 12th European Microwave Conference in Helsinki, Finland, and was awarded the Gustaf Dalén gold medal in 1986.

Gabriel M. Rebeiz (S'86–M'88) was born in December 1964 in Beirut, Lebanon. He graduated in 1982 from the American University in Beirut with a B.E. (Honors) in electrical engineering. In September 1982, he joined the California Institute of Technology, and earned the Ph.D. in electrical engineering in June 1988.

He joined the faculty of the University of Michigan in September 1988 where he is now an Assistant Professor in the Electrical Engineering and Computer Science Department.

Dr. Rebeiz has been awarded a NASA-Certificate of Recognition Award for his contribution to the millimeter-wave space program (March 1990) and the Best Paper Award at the 1990 International Conference on Antennas, Nice, France. He received an NSF Presidential Young Investigator Award in 1991. His research interests lie in planar millimeter-wave antennas, receivers and transmitters, and fabrication and measurements of novel millimeter-wave transmission-lines and devices.