Correspondence

The Cutoff Characteristics of the Channel Waveguide

In order to increase the power-handling capacity of a TE_{10} mode rectangular waveguide, the channel waveguide[1] as shown in Fig. 1 was proposed. To the authors' knowledge, the application of channel waveguides for this purpose is not yet known in the United States, although in a paper by Iashikin[2] received on September 26, 1956 by the A. S. Popov Technical Society of Radio Engineering and Telecommunications U.S.S.R., the author states that channel waveguides "are widely used at present."

The waveguide was thought to have a greater power-handling capacity than the regular rectangular waveguide. Increase of the power-handling capacity is dependent on the channel waveguide configuration. There would be many things to compromise for a practical channel waveguide design. The impedance, the cutoff wavelength, the mode stability and the attenuation and phase characteristics are to be considered together with the power-handling capacity to determine the waveguide configuration. For simplicity, only the power-handling capacity is considered here and it will be compared with a regular rectangular waveguide of the same height b_1, (see Fig. 1). The power-handling capacity of a waveguide is usually limited by arcing. Arcing field strength can be assumed to be, for purposes of approximation, proportional to the distance between the extreme top wall and bottom wall of the waveguide at the maximum field intensity position. Therefore, the increase of the power-handling capacity can be estimated approximately by 20 log b_2/b_1 in db measurements. Therefore, if $b_2/b_1=2$, then the power-handling capacity would increase approximately 6 db or 4 times the regular waveguide power-handling capacity. This means that if the arcing starts at 100 kw for a regular waveguide, then the channel waveguide can carry power up to approximately 400 kw. The only method commonly used today to increase the power-handling capacity of a waveguide is pressurizing the waveguide. But even this technique has it's arcing limits and the waveguide configuration must be modified to attain very high power-handling capacity. The proposed channel waveguide is one example of many possible solutions. Other methods are conceivable such as increasing the height of a rectangular waveguide. In any method, the impedance, attenuation and phase characteristics, frequency characteristics, mode stability and fabrication techniques need to be considered and compromised. Therefore, it would be difficult to say at this stage of investigation, whether any one method is superior. The criteria depends on the characteristics desired.

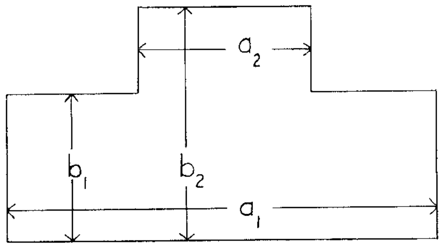
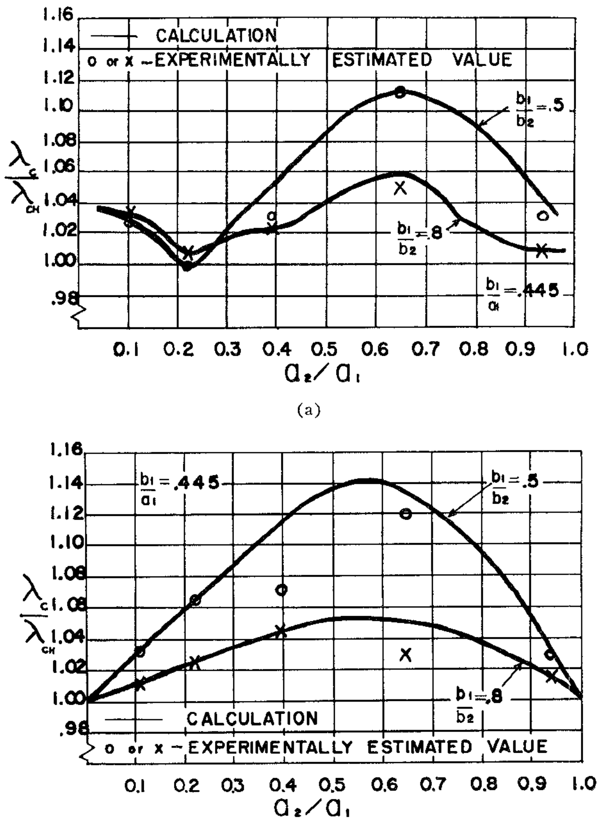


Fig. 1—Cross-sectional view of the channel waveguide.



(a)

(b)

Fig. 2(a) —$\lambda_c/\lambda cH$ ratio predicted by using Cohn's method. (b) $\lambda c/\lambda cH$ ratio predicted by using Iashikin's method. λ_c is the cutoff wavelength of the regular rectangular waveguide and λcH is the cutoff wavelength of the channel waveguide.

The main purpose of this note is neither to demonstrate the increased power-handling capacity nor to show the superiority of the channel waveguide over regular waveguides, but to report a finding of peculiar cutoff characteristics of the channel waveguide. Vilmur and Ishii theoretically analyzed the cutoff wavelength of the channel waveguide based on Iashikin's method[2,3] and Cohn's method.[4] Theoretical investigation based on the two different methods produced two different conclusions[1] as shown by the solid lines in Fig. 2 (a)–(b). The purpose of this investigation was to solve this discrepancy by experiments, and it was found that there was a "mode" of propagation which followed Iashikin's method [Fig. 2(b)] while there was another "mode" of propagation which followed Cohn's method [Fig. 2(a)].

Ten different channel waveguides were made in order to completely correlate the experimental data with the theoretical data. Five of these had a depth ratio, (b_1/b_2) of 0.8, and five had a ratio (b_1/b_2) of 0.5. Five different widths were used for each depth, and a base section with a RG 52/U waveguide was used. A 50-cm long channel waveguide with tapered sections (6-cm long) at

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[1] R. J. Vilmur and K. Ishii, "The channel waveguide," IRE Trans. on Microwave Theory and Techniques, vol. MTT-10, pp. 220–221; May, 1962.

[2] A. I. Iashikin, "The calculation of the fundamental critical wavelength for a rectangular waveguide with longitudinal rectangular channels and ridges," *Radio Engrg.*, vol. 13, pp 8–14; March, 1958.

[3] A. I. Iashikin, "A method of approximate calculation for waveguides of triangular and trapezoidal cross-sections," *Radio Engrg.*, vol. 13, pp. 1–9; October, 1958.

[4] S. B. Cohn, "Properties of ridge waveguides," *Proc. IRE*, vol. 35, pp. 783–788, August, 1947.

both ends to match the standard RG 52/U waveguide was mounted in the experimental circuit as shown in Fig. 3. The cutoff wavelength was *estimated* from a rising characteristic of the input VSWR when the operating wavelength approached the cutoff. In order to operate near the cutoff, a special long-slotted section (28.5 cm) was used for the measurement of VSWR. It is true that the method of finding the cutoff wavelengths in the channel waveguide mentioned above is not the only one that could be used, and may not be the most convenient or accurate. In fact, the "insertion" method was also tried to make sure that the power was cut off. The standing wave detector method had, however, the advantage of detecting both "cutoff modes." "Insertion" methods could give only the lowest frequencies cutoff mode. If the "insertion" method alone were used, both "cutoff modes" could never have been found.

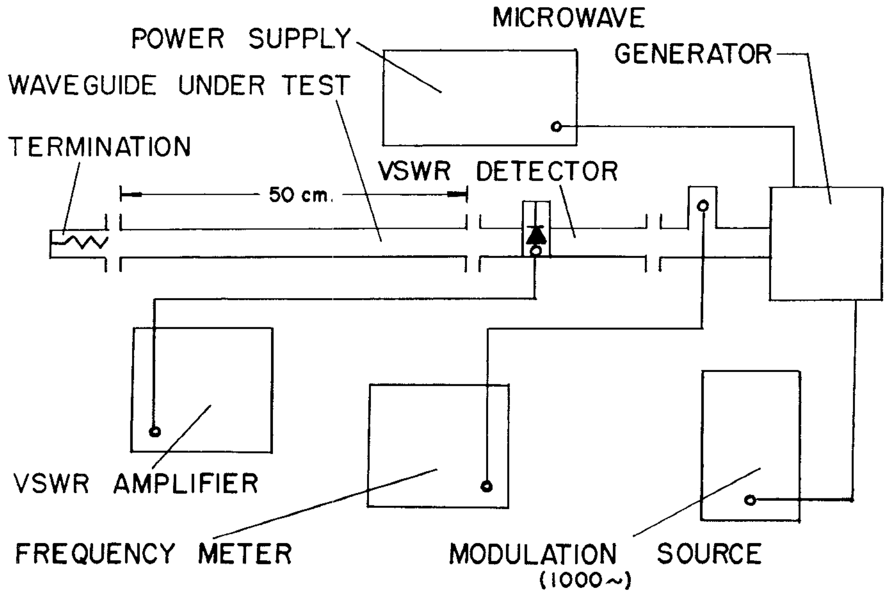

Fig. 3—A schematic diagram of the experimental set up for VSWR measurement of channel waveguide.

A typical example of the experimental results of VSWR and frequency characteristics is shown in Fig. 4. In this particular case of $b_1/b_2=0.5a_2/a_1=0.214$, as seen from the experimental results, there was a transmission of microwaves far below the cutoff frequency predicted by Iashikin's method. It was also certain that a sharp rise of VSWR existed near the cutoff frequency predicted by Iashikin's method. As seen from Fig. 4, another rise of VSWR was observed near the cutoff frequency predicted by Cohn's method. Similar phenomena were observed for channels of different dimensions. When the cutoff wavelengths of various channel waveguides were estimated from the experimental VSWR curves by observing the rises of VSWR at wavelengths near the theoretical cutoff, the results shown by plots in Fig. 2 were obtained. The experimental results seemed to follow the theoretically predicted values based on both Iashikin's and Cohn's methods. The experimental results based on the VSWR measurements of the channel waveguide's cutoff characteristics indicated there were two different "modes" of waves existing in the channel waveguide. One "mode" follows Iashikin's method and the other "mode" follows Cohn's method. The reason why the two modes exist and how the transition between two modes takes place are considered as follows: Iashikin[2] started his analysis with the wave equation

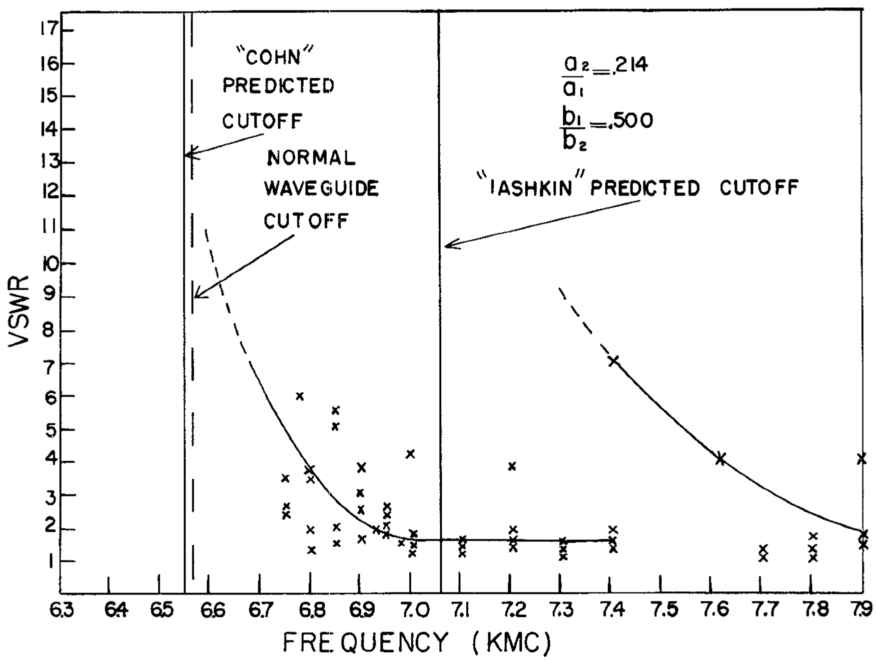

Fig. 4—An example of experimental VSWR vs. frequency characteristics of channel waveguide.

$$\frac{\partial^2 H_z}{\partial x^2}+\frac{\partial^2 H_z}{\partial y^2}+k^2H_z=0$$

with H_z is the longitudinal magnetic field intensity. The boundary condition used was $\partial H_z/\partial n=0$ at the waveguide wall and the cross section of the waveguide was divided vertically into subsections and at every section boundary conditions $H_i=H_{i+1}$ and

$$\frac{\partial H_i}{\partial x}=\frac{\partial H_{i+1}}{\partial x}$$

were applied. Thus Iashikin's solution was an infinite series and contained all possible space harmonics which may exist in the channel waveguide. The cutoff wavelength was determined by $\lambda_c=2\pi/k$.

On the other hand, Cohn started with an equivalent circuit of a ridge waveguide operated at cutoff using the concept of Ramo and Whinnery.[5] Cohn's equivalent circuit consists of two transmission-line sections joined together at a discontinuity admittance calculated by Whinnery and Jamieson.[6] The cutoff wavelength was calculated as a resonance wavelength of transversely resonating jointed transmission lines. Thus, according to Cohn's approach, in this particular case of channel waveguide, the method predicts only the longest dominant mode cutoff wavelength. In other words, Iashikin's approach assumes that all possible modes exist near the cutoff and Cohn's approach assumes only a single dominant mode at the cutoff.

According to the results shown in Fig. 4, when the operating frequency decreased gradually starting from 7.8 kMc, the channel waveguide seemed to follow Iashikin's mode, and at 7.06 kMc, the operating frequency reaches cutoff of Iashikin's mode. If the operating frequency is reduced further, high modes cancel each other and the waveguide is cut off to them, then only the dominent mode can propagate. The channel waveguide is switched to Cohn's mode, since Iashikin's mode is in cutoff, and the channel waveguide will still transmit microwaves. Insertion loss of the channel wave-

[5] S. Ramo and J. R. Whinnery, "Fields and Waves in Modern Radio," John Wiley and Sons, Inc., New York, N. Y.; 1944.

[6] J. R. Whinnery and H. W. Jamieson, "Transmission line discontinuities," *Proc. IRE*, vol. 32, pp. 98–116, February, 1944.

guide section was only 3.5 db at 6.8 kMc. When the operating frequency is reduced further, the cutoff of Cohn's mode appears. In the case of Fig. 4, Cohn's cutoff is close to the actual cutoff of the channel waveguide.

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A Coaxial Adjustable Sliding Termination

INTRODUCTION

The accuracy of impedance measurements using modified reflectometer techniques depends mainly upon the tuning of the reflectometer. This tuning is accomplished by sliding first a low-reflection termination and, then, a high-reflection termination (sliding short circuit) in the output waveguide of the reflectometer. The actual error that can occur due to imperfect tuning can be computed[1] and depends to a large extent upon the size of the reflection coefficient of the low-reflection sliding termination. The lower this reflection coefficient is, the smaller the error will be. The adjustable sliding termination described in this paper was developed to reduce this reflectometer tuning error; hence, the main emphasis was on obtaining a stable, very low-reflection coefficient.

DESCRIPTION OF INSTRUMENT

A drawing showing the principle of the instrument is shown in Fig. 1. The principle of operation is similar to the one described by Ellenwood and Ryan.[2] The main difference is that, instead of using a double slug tuner in front of a terminating element, the reflection coefficient of the actual terminating element is variable. It is varied by moving a lossy cylinder inside of a lossy taper. When the face of the cylinder is positioned immediately in front of the edge of the taper, a maximum reflection is obtained; when it is completely withdrawn inside of the taper, a minimum reflection is obtained.

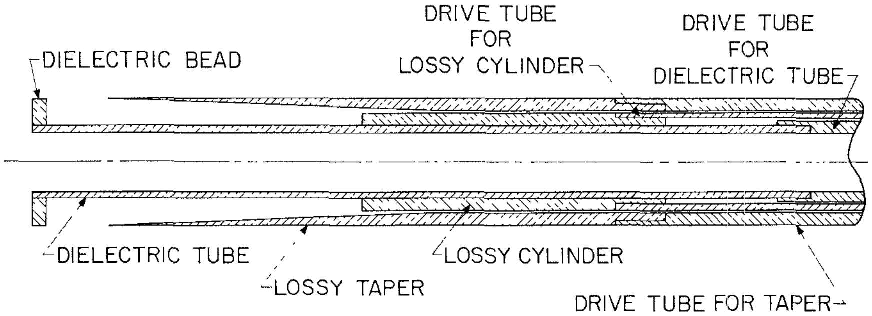

Fig. 1—Cross section of the adjustable sliding termination.

The total reflection coefficient of the sliding termination is a combination of the reflection from the terminating elements and the reflection from the bead in front of the terminating element. The bead is mounted on a very thin dielectric tube that extends through the lossy terminating elements. A maximum reflection occurs when the two reflections are in phase, and a minimum reflection occurs when they are out of phase.

THEORY OF OPERATION

The equation[3] for the total reflection coefficient of the termination in terms of the reflection coefficients of the bead and the terminating elements is

$$\Gamma = \Gamma_1 + \frac{\Gamma_L(1 - \Gamma_1^2)e^{j\phi}}{1 + \Gamma_1\Gamma_L e^{j\phi}}, \qquad (1)$$

where Γ_1 is the reflection coefficient of the bead, Γ_L is the reflection coefficient of the terminating element and ϕ is the angle of combination of the two reflection coefficients. The assumption that the bead and transmission line are lossless has been made in order to use this equation.

The minimum value of Γ is

$$\Gamma_{min} = \frac{|\Gamma_1| - |\Gamma_L|}{1 - |\Gamma_1\Gamma_L|}. \qquad (2)$$

For Γ_{min} to equal zero, $|\Gamma_1|$ must equal $|\Gamma_L|$. In practice, the range of $|\Gamma_L|$ that can be obtained is measured and then the bead is designed to give a reflection coefficient of the desired value. If a small reflection coefficient is of prime interest, the bead is made to have a reflection just slightly larger than the smallest obtainable value of $|\Gamma_L|$. If a wide range of reflection coefficient is desired, the bead is made to have a reflection coefficient just slightly smaller than the largest obtainable value of $|\Gamma_L|$.

The reflection coefficient of a single bead can be computed from (3).[4]

$$\Gamma_1 = \frac{-j(\epsilon - 1)\tan\dfrac{2\pi\sqrt{\epsilon}\,l}{\lambda}}{2\sqrt{\epsilon} + j(\epsilon - 1)\tan\dfrac{2\pi\sqrt{\epsilon}\,l}{\lambda}}, \qquad (3)$$

where Γ_1 is the reflection coefficient of the bead, l is the length of the bead, ϵ is the dielectric constant of the bead and λ is the wavelength in free space. If the value of

$$\tan\frac{2\pi\sqrt{\epsilon}\,l}{\lambda}$$

is small, which is usually the case, the magnitude of the reflection coefficient can be closely approximated by (4).

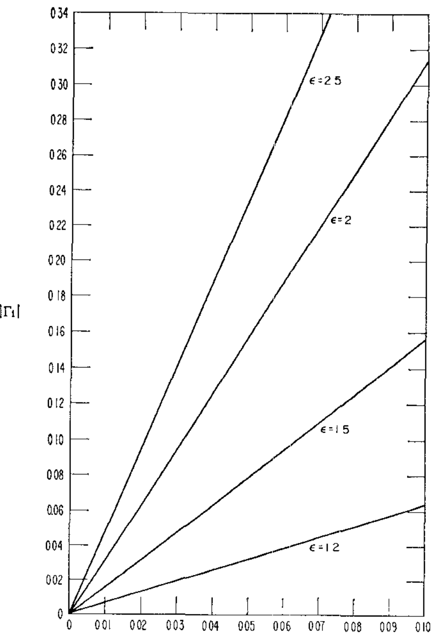

Fig. 2—Graph of (3).

$$\Gamma_1 \cong \frac{\pi l}{\lambda}(\epsilon - 1). \qquad (4)$$

Eq. (4) is graphed in Fig. 2 for values of ϵ from 1.2 to 2.5.

RESULTS

Figs. 3 and 4 show instruments that have been made using the above principles. The instrument in Fig. 3 was designed for a $\frac{3}{4}$-inch transmission line and the one in Fig. 4 was designed for a $\frac{9}{16}$-inch line. In each case, stable reflection coefficients of less than 0.005 were readily attained. The range of reflection coefficient of the terminating element of the $\frac{9}{16}$-inch instrument was from 0.02 to 0.15 at a frequency of 4 Gc. The bead was designed to give a reflection coefficient of approximately 0.05. This was done by making the outer diameter of the bead considerably less than the inner diameter of the outer conductor to give the bead an equivalent dielectric constant[5] of approxi-

Manuscript received August 14, 1963.

[1] W. J. Anson, "A guide to the use of the modified reflectometer technique of VSWR measurement," *J. Research Natl. Bur. Standards*, vol. 65C, pp. 217–223; October–December, 1961.

[2] R. C. Ellenwood and W. E. Ryan, "A UHF and MW matching termination," PROC. IRE, vol. 41, pp. 104–107; January, 1953.

[3] T. Moreno, "Microwave Transmission Design Data," Dover Publications, Inc., New York, N. Y., p. 31; 1958.

[4] *Ibid.*, see p. 84.

[5] J. W. E. Griesmann, "Handbook of Design and Performance of Cable Connectors for Microwave Use," Report No. R-520-56; PIB-450 for Bureau of Ships Contract No. NObsr-52078 Index No. NE-110718; 1956.