Correspondence

Coaxial Isolator Utilizing Yttrium-Iron Garnet*

The inherently narrow ferromagnetic resonance line width attainable with yttrium-iron garnet (YIG) material permits building isolators with greatly improved characteristics at the lower microwave frequencies. An isolator utilizing YIG designed for operation in the UHF range has been reported.[1] This letter describes a coaxial isolator designed to operate at 2250 ± 50 mc employing yttrium-iron garnet.

A $\frac{5}{8}$-inch diameter, 50-ohm coaxial line was half filled with a high dielectric constant material (Stycast Hi-K-15) with $2\frac{1}{2}$-inch tapers on the ends and the YIG rods placed on the interface; the magnetic biasing field was parallel to the interface in the manner reported by Duncan, *et al.*[2] Two YIG rods of square cross section (0.070 = 0.070 × 2.450 inches long with $\frac{1}{4}$-inch tapers on the ends) were located on both sides of the inner conductor. With a permanent magnet biasing field adjusted for resonance at 2250 mc, isolator characteristics as shown in Fig. 1 were obtained.

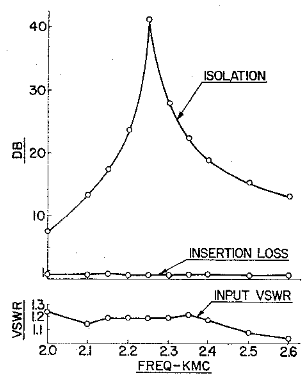

Fig. 1—Characteristics of yttrium-iron garnet coaxial isolator.

By replacing the permanent magnet with an electromagnet and adjusting the biasing field for maximum reverse attenuation at each frequency, the maximum reverse attenuation as a function of frequency was obtained as shown in Fig. 2. All measurements were made in the milliwatt power range. Since this material evidenced such reasonable isolation-to-insertion loss ratios in the 1-kmc region, additional work was initiated and is presently in progress to develop an isolator for this frequency range.

This material was prepared by the Magnetics Research Group of the Solid State Electronics Department of Lockheed Missile Systems Division, and had a measured

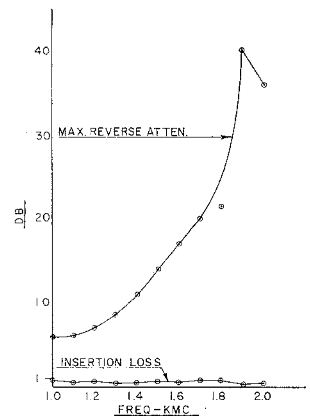

Fig. 2—Maximum reverse attenuation of coaxial isolator vs frequency by varying biasing field.

density of 98 per cent of theoretical maximum, a Curie temperature of 318°C, and a measured X-band line width of 60 oersteds.

L. FREIBERG
Lockheed Missile Systems Div.
Sunnyvale, Calif.

* Received by the PGMTT, June 27, 1958.

[1] F. R. Morgenthaler and D. L. Fye, "Yttrium garnet UHF isolator," PROC. IRE, vol. 45, pp. 1551–1552; November, 1957.

[2] B. J. Duncan, L. Swern, K. Tomiyasu, and J. Hannwacker, "Design considerations for broad-band ferrite coaxial line isolators," PROC. IRE, vol. 45, pp. 483–490; April, 1957.

A Broad-Band Microwave Coaxial Connector with Capacitive RF Coupling and Isolated DC Returns*

INTRODUCTION

A modified type N connector for passing RF while grounding the center conductor of the coaxial line has been reported by McLaughlin and Dunn.[1] This device is useful for many applications but in microwave measurements it is sometimes necessary to have a high pass filter with a different dc voltage connected to each side of the filter. One such situation occurs in the connection of the input terminal of a microwave tube at one dc potential to a signal generator at essentially ground potential. Such a filter has been made with a type N connector having a series capacitance in the center conductor and a high resistance wire connected to one side of the split center conductor and passing through a small hole in the outer conductor. Since the resistance wire has very little effect on the RF properties, very little difference would be observed by having a dc return to both sides of the split center conductor. The capacitance consists of parallel plates separated by an insulator with a large dielectric constant. From 1.5 to 10 kmc the connector has less than 3 db insertion loss and a VSWR less than 1.5. The dc current rating of the resistance wire is 100 ma and the dc voltage breakdown between two faces of capacitance is greater than 1500 volts.

* Received by the PGMTT, July 18, 1958.

[1] J. W. McLaughlin and D. A. Dunn, "Wide band dc return connector," IRE TRANS. ON ELECTRON DEVICES, vol. ED-4, pp. 310–311; October, 1957.

PHYSICAL PROPERTIES

The center conductor of a type N(UG29 B/U) connector is severed and insulated with dielectric material to form a blocking condenser which will pass RF but not direct current. The capacitance should be made as large as possible for low impedance at the operating frequency. Various materials were tested and the best results were obtained by inserting a piece of ceramic obtained from an ordinary Erie 0.001 μf disk ceramic condenser. It was estimated that the dielectric constant of this type ceramic was in the order of 100, which greatly exceeds the dielectric constant of the usual construction materials. With a piece of ceramic having a diameter of 0.168 inch and a thickness of 0.040 inch the inserted capacitance was 7 to 12 $\mu\mu$f at 1 mc and the dc working voltage was greater than 1500 volts. The dc path was formed by placing a fine Karma alloy resistance wire inside a small teflon tubing and feeding this unit through a $\frac{1}{16}$-inch diameter hole in the outer conductor in such a way that the Karma wire overlapped the teflon tubing and was pressed against the center conductor by screwing on a bakelite cap. With this arrangement the insulation between the Karma wire and the outside conductor was also greater than 1500 volts. The bakelite cap also formed the terminal for the dc connection. A drawing of the modified connector is shown in Fig. 1.

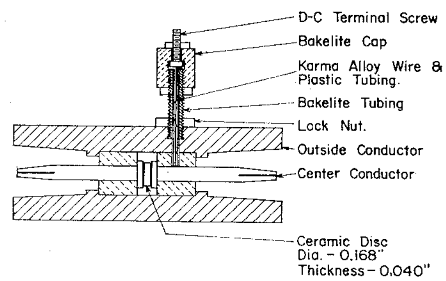

Fig. 1—Microwave coaxial capacitance using an N-type connector fitted with one dc connection and a series capacitance in the center conductor.

The Karma alloy wire was used because it was readily available in small sizes and has excellent mechanical properties. Its physical properties were as follows:

Karma alloy wire	0.0012 inch diameter
Resistivity of wire	521.5 ohms per foot
Resistance of dc path	30–40 ohms
Current carrying capacity	100 ma

The specifications of the unit were selected to satisfy our own requirements but could be suitably modified to allow for a different breakdown voltage or frequency bandwidth for other applications.

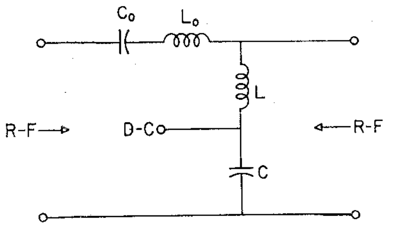

Fig. 2—Schematic diagram of a microwave coaxial capacitance of the type employed. Only one dc voltage is applied in this device but another terminal is possible.

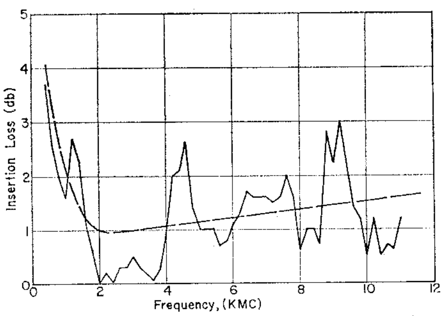

Fig. 3—Experimental measurement of insertion loss of the microwave N-type coaxial capacitance with one dc terminal, showing also the probable value of the connector loss. The connector has a sharp low frequency cutoff and a very gradual high frequency cutoff.[2]

RF Properties

The connector is really a band-pass filter[2] with a sharp low frequency cutoff and a gradual upper frequency cutoff so that it makes an excellent high-pass filter in the range from 1.5 to 10 kmc. A schematic diagram of the equivalent circuit is shown in Fig. 2. In this case C_0 represents the capacitance in the center conductor, L represents the inductance of the Karma alloy resistance wire, C represents the capacitance between this wire and the external shield, and L_0 is the natural series inductance between the capacitance C_0 and the dc resistance path. As L_0 approaches zero this becomes a high-pass filter.

The connector loss was measured by the insertion loss method. The power delivered to a matched load from a matched generator was detected with and without the "connector" inserted in the line. The ratio of these powers thus gave the insertion loss directly. As a result of this method the readings were sensitive to the VSWR at the terminals of the connector. The insertion loss varied from 0.01 to 3 db for the frequency range 1.5 to 10 kmc, as shown in Fig. 3.

The dashed average line on Fig. 3 indicates the probable value of the insertion loss across the band in the absence of reflections. Even though the device is a pass band filter the attenuation of the pass band apparently rises very slowly on the high frequency end making the device very broadband.

The VSWR of the connector, which may be of more interest than the insertion loss for some considerations, is shown in Fig. 4.

[2] W. P. Mason, "Electro-Mechanical Transducers and Wave Filters," D. Van Nostrand Co., Inc., New York, N. Y., p. 52; 1948.

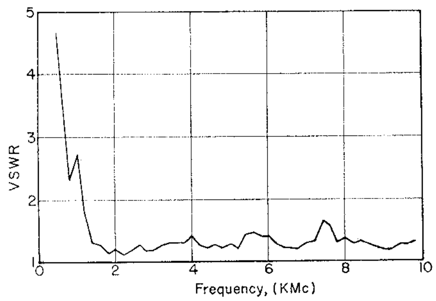

Fig. 4—Experimental measurement of VSWR of the microwave N-type coaxial capacitance with one dc terminal

The VSWR is less than 1.50 between 1.5 and 10 kmc.

No direct measurement was made to determine the radiation loss through the Karma wire terminal, but it was noted that placing a grounded copper shield ½-inch long around the terminal caused less than a 2 per cent difference in the VSWR reading when the dc terminal was grounded or open circuited.

C. M. Lin
R. W. Grow
Electronics Labs.
Stanford University
Stanford, Calif.

The Cutoff Wavelength of Trough Waveguide*

Introduction

The trough waveguide,[1] although in use for several years, is not widely known and has received only little attention in the literature. This waveguide was suggested by E. G. Fubini, and its fundamental mode may be compared to that of a TE mode in symmetrical strip transmission line. The configuration is shown in Fig. 1. This type of

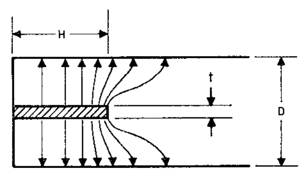

Fig. 1—Trough waveguide.

waveguide has several advantages, which include:

1) broad frequency range, because the cutoff frequency of the second mode is approximately three times that of the dominant mode;
2) low-reflection, broad-band transitions to TEM lines, easily made by an end-on connection of the center conductor of a coaxial line to a point on the center vane;[2]
3) line measurements, made with a minimum of disturbance because of the open side;
4) simple control of the propagation characteristics, possible by changing the center vane;
5) economical fabrication.

The application of this waveguide to line-source radiators has been investigated extensively by Rotman and Karas and excellent results have been obtained.[3]

A derivation of the cutoff wavelength for a TE mode in symmetrical strip transmission line in the case of a zero-thickness center strip was obtained independently by Jasik[4] and Oliner.[5] In both instances the result was based on the analogous E-plane bifurcation in rectangular waveguide, and this result applies also to the trough waveguide. In any actual waveguide, however, the center vane must have a finite thickness. It is of interest, therefore, to know how the cutoff wavelength depends on this parameter.

Derivation of the Approximate Cutoff Wavelength

By the transverse resonance procedure, the cutoff wavelength is given by

$$\lambda_c = 4(H + d) \tag{1}$$

where d is the distance from the physical edge of the center vane to the effective open circuit point. In the case of $t=0$, we have[4,5]

$$d = \frac{D}{\pi}\ln 2 + \frac{\lambda_c}{2\pi}\left[S_1\left(\frac{2D}{\lambda_c}\right) - 2S_1\left(\frac{D}{\lambda_c}\right)\right], \tag{2}$$

where

$$S_1(x) = \sum_{n=1}^{\infty}\left(\arcsin\frac{x}{n} - \frac{x}{n}\right).$$

We may relate d to an equivalent fringing capacitance which, because d is frequency dependent, will also be frequency dependent.

In the case of thick center vane let us consider the equivalent fringing capacitance from one edge of the vane. This is given by

$$C_f = 0.0885\epsilon\frac{d}{\frac{1}{2}(D - t)}, \; PF/\text{cm}, \tag{3}$$

so that

$$\frac{d}{D} = \frac{1}{2}\cdot\frac{C_f}{0.0885\epsilon}(1 - t/D), \tag{4}$$

and from (1), we have

$$\frac{\lambda_c}{D} = 4\frac{H}{D} + \frac{2C_f(1 - t/D)}{0.0885\epsilon}. \tag{5}$$

The second term on the right side of (5) is

* Received by the PGMTT, July 21, 1958.

[1] Airborne Instruments Lab., Mineola, N. Y., Advertisement, Proc. IRE, vol. 44, p. 2A; August, 1956.

[2] H. S. Keen, "Scientific Report on Study of Strip Transmission Lines," Airborne Instruments Lab., Mineola, N. Y., Rep. No. 2830-2; December 1, 1955.

[3] W. Rotman and N. Karas, "Some new microwave antenna designs based on the trough waveguide," 1956 IRE Convention Record, pt. 1, pp. 230–235.

[4] H. Jasik, private communication to E. G. Fubini; July 30, 1953.

[5] A. A. Oliner, "Theoretical developments in symmetrical strip transmission line," *Proc. Symp. Modern Advances in Microwave Techniques*, Polytechnic Inst. of Brooklyn, Brooklyn, N. Y., pp. 379–402; November, 1954.