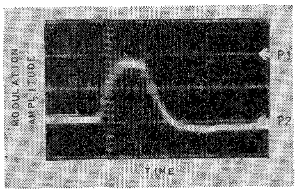

Fig. 3—Microwave absorption modulation pattern. P_1 and P_2 are the initial and absorption power levels respectively. Time scale=0.25 μsec/division (E=4000 v/cm, 40 pps).

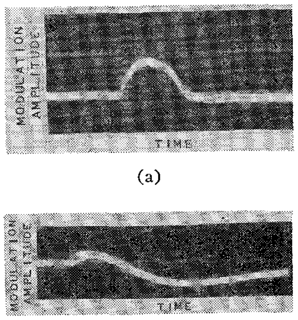

Fig. 4—Modulator operation dependence on junction polarity. Time scale=0.5 μsecond/division (E=2000 v/cm, 40 pps). (a) N-N^+ junction connected to positive-voltage lead. (b) N-N^+ junction connected to negative-voltage lead.

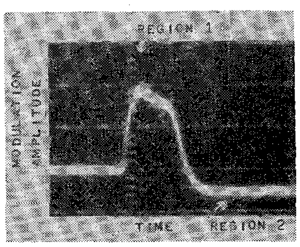

Fig. 5—Modulation changes due to minority carriers. Time scale=0.5 μsec/division (E=4000 v/cm, 40 pps).

both during and following the 0.5-μsec voltage-pulse application. This may be attributed to hole injection or to an ionization effect in these crystals. The region following the high-voltage pulse was of very long duration.

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Coaxial to Strip Transmission Line Adapter*

Over the past few years, strip transmission line has shown itself capable of being utilized in a large number of microwave configurations which were previously constructed in coaxial line or waveguide. In many cases the components fabricated in strip transmission line are simpler to design and produce particularly where, as in the sandwich type of line, advantage may be taken of photo-etching techniques. This type of construction using copper-foil-clad dielectric material enables the foil to be used for both the center conductor and the ground planes.

However, the use of such thin conducting material with relatively poor adhesion between the metal and dielectric often leads to difficulties where the coaxial line is attached to the strip. The action of soldering the center pin of the coaxial line to the strip tends to destroy the adhesion between the foil and the dielectric. The first time the connection is made the results may be quite satisfactory, but in development work it is often necessary to assemble and dismantle a filter or other device many times to make adjustments and alterations. If a soldered connection is used in such a situation, the end of the strip is soon distorted and loosened to such an extent that measurements made through the junctions are meaningless. To overcome this difficulty, a solderless transition has been devised for use in our laboratory work. An exploded view of this transition showing the important dimensions is given in Fig. 1.

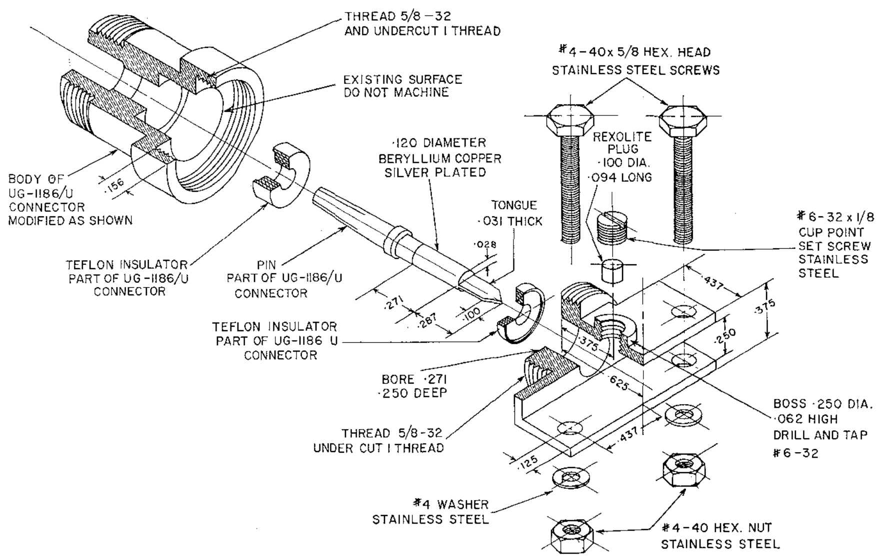

Fig. 1.

The adapter could have been of either the right-angle or in-line type, but the latter was chosen for mechanical strength and convenience, in addition to the fact that a symmetrical transition is less likely to introduce spurious modes. The channelled transition block is fastened to the stripline by two machine screws. This provides the mechanical attachment and the electrical connection to the ground planes. The flattened pin is accommodated in a slot milled in the lower surface of the upper dielectric sheet. The pin is pressed against the etched center conductor by a button of the same material as the dielectric of the line which is a loose fit in a hole in the upper dielectric sheet. A set screw in the threaded boss in the transition block holds the button tightly against the flattened pin, insuring good electrical contact between the pin and the center conductor.

The parts used for the coaxial end of the adapter were originally taken from a standard UG-1186/U connector with the body shortened and threaded as shown. However, after the initial connectors proved successful, a large quantity was made to our design by a manufacturer.

These connectors have been used over the range of frequencies from 300–6000 Mc and have given very satisfactory results. In order to confirm our opinion, measurements were made on a number of these connectors using the method of measuring a junction described by Wentworth and Barthel.[1]

Six adapters, taken at random from stock, were each measured with six different lengths of strip transmission line. This involved 36 assemblies and disassemblies of the transitions, but the uniformity of the results indicated that this had no adverse effect. At any frequency and with one particular length of line, the variation of the position of the voltage minima on the slotted line was less than ±0.2 mm with reference to the mean for the six transitions. This is of the same order as the expected experimental error and represents a spread of less than ±0.005 wavelength at the highest frequency. This close agreement between the measurements on the various adapters enabled the mean VSWR of the junctions to be computed from the average readings (Fig. 2). The true VSWR of a particular junction would not differ from this mean value by more than ±0.03.

* Received by the PGMTT, December 30, 1960.

[1] F. L. Wentworth and D. R. Barthel, "A simplified calibration of two port transmission line devices," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-4, pp. 173–175; July, 1956.

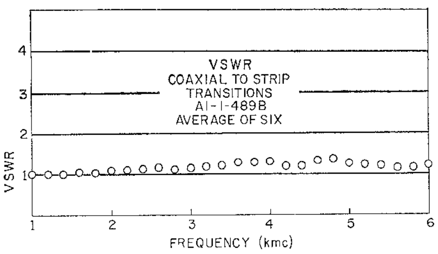

Fig. 2.

Our experience has shown that these connectors have a low enough VSWR to be adequate for all but the most critical situations and may be assembled and dismantled numerous times without damage to either the connector itself or to the strip to which it has been connected.

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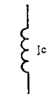
A Note on Loaded Line Synthesis*

The purpose of this note is to present an alternate derivation of a formula for the synthesis of a loaded line. The problem is to determine the values of the normalized susceptances a and c mounted an exact quarter-wavelength apart along a uniform line. These are arranged in the order a, c, $c, \cdots c$, a to achieve a loaded line with a given phase shift and perfect match. In order to analyze a particular loaded line design for standing-wave ratio and phase shift over a band of frequencies on a digital computer, it is worthwhile to know the values of the susceptances to many more decimal places than one would achieve from a simple graph.

In branch transmission line couplers, where the length of branches and the spacing between them are each exactly a quarter-wavelength long, the calculated response (standing-wave ratio and fraction of power out each arm) is symmetrical on a normalized frequency basis F where $F=f/f_0$ or $F=\lambda g_0/\lambda g$ when waveguide is used. For example, the response is the same at $F=0.80$ as it would be at $F=1.20$.

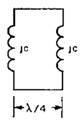
The response is not symmetrical in the case of a half-wave or quarter-wave rotating plate consisting of round waveguide loaded with a cascade of irises which are alternately all capacitive or all inductive as the guide is rotated. However, if the values of a and c are assumed to have a frequency variation which is reasonable (inductive varying as $1/F$ and capacitive varying as F), the difference in phase between the two positions of the plate does not change with frequency near the design center. That is, the phase shift of the plate, when it is used as a capacitive iris array minus that when it is an inductive iris array, has the desired value, and the rate of change of this value with respect to frequency is zero at the design center. The assumption is made that the magnitude of the susceptances are the same when the irises are inductive as when the irises are capacitive.

Consider a single shunt susceptance of normalized value, jc. Its matrix is



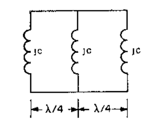
$$M = \begin{bmatrix} 1 & 0 \\ jc & 1 \end{bmatrix}. \quad (1)$$

For two such susceptances separated by a quarter wavelength of unity impedance line



$$M = \begin{bmatrix} 1 & 0 \\ jc & 1 \end{bmatrix}\begin{bmatrix} 0 & j \\ j & 0 \end{bmatrix}\begin{bmatrix} 1 & 0 \\ jc & 1 \end{bmatrix} = \begin{bmatrix} -c & j1 \\ -j(c^2-1) & -c \end{bmatrix}. \quad (2)$$

For three susceptances



$$M = \begin{bmatrix} c^2-1 & j(-c) \\ -j(-c^3+2c) & c^2-1 \end{bmatrix}. \quad (3)$$

For n susceptances separated by $n-1$ quarter wavelength of line

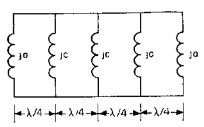
$$M = \begin{bmatrix} S_{n-1}(-c) & jS_{n-2}(-c) \\ -jS_n(-c) & S_{n-1}(-c) \end{bmatrix} \quad (4)$$

where

$$\begin{aligned} S_{-1}(-c) &= 0 \\ S_0(-c) &= 1 \\ S_1(-c) &= -c \quad (5) \\ S_2(-c) &= c^2-1 \\ S_3(-c) &= -c^3+2c \\ S_{n+1}(-c) &= -cS_n(-c) - S_{n-1}(-c). \quad (6) \end{aligned}$$

The polynomials $S_n(-c)$ are Chebyshev [2] polynomials of the second kind.

If a shunt susceptance ja is put a quarter-wavelength away from each end to achieve a matched condition, the matrix becomes



$$M = \begin{bmatrix} -aS_n(-c) - S_{n-1}(-c) & jS_n(-c) \\ -j(a^2S_n(-c) + 2aS_{n-1}(-c) + S_{n-2}(-c)) & -aS_n(-c) - S_{n-1}(-c) \end{bmatrix}. \quad (7)$$

This matrix represents a line of length θ and a normalized impedance of unity if each term in it is equal to those of a matrix of this length of line:

$$M = \begin{bmatrix} \cos\theta & j\sin\theta \\ j\sin\theta & \cos\theta \end{bmatrix}. \quad (8)$$

On equating matrix (7) with (8), the value of c can be found by setting $S_n(-c) = \sin\theta$ by inverse interpolation in the tables [2]. Then the value of a to give a matched line is

$$a = \left| \frac{|\cos\theta| - |S_{n-1}(-c)|}{\sin\theta} \right|. \quad (9)$$

There are two values of a which will match out the loaded line; the one which should usually be used is that with the smaller magnitude. The algebraic sign of a is the same as c and may be either positive or negative, depending on whether the resulting loaded line is to be effectively longer or shorter than the unloaded line.

An earlier contribution [3] has been based on using the value of a as exactly one-half that of c. The element spacing was then computed so that the differential phase shift was as desired. This scheme results in spacings which are not an exact quarter-wavelength and therefore may not give a symmetrical frequency response especially for small numbers of elements.

Appendix

The general circuit matrix of a dissipationless reciprocal symmetrical two-port network can be represented as

$$\begin{bmatrix} A & B \\ C & A \end{bmatrix}. \quad (10)$$

In this $A^2-BC=1$ and A is a purely real quantity and B and C are purely imaginary. The matrix of a cascade of n of these networks may be found by raising (1) to the nth power.

For $n=2$

$$\begin{bmatrix} A & B \\ C & A \end{bmatrix}^2 = \begin{bmatrix} 2A^2-1 & B(2A) \\ C(2A) & 2A^2-1 \end{bmatrix}. \quad (11)$$

For $n=3$

$$\begin{bmatrix} A & B \\ C & A \end{bmatrix}^3 = \begin{bmatrix} 4A^3-3A & B(4A^2-1) \\ C(4A^2-1) & 4A^3-3A \end{bmatrix}. \quad (12)$$

For the general case of n identical networks the matrix product may be expressed as [5]

$$M^n = \begin{bmatrix} T_n(A) & BU_{n-1}(A) \\ CU_{n-1}(A) & T_n(A) \end{bmatrix}. \quad (13)$$

$T_n(A)$ and $U_{n-1}(A)$ are Chebyshev polynomials of the first and second kind in A [2].

* Received by the PGMTT, October 29, 1959; revised manuscript received, January 3, 1961.