Correspondence

Wide-Band Rectangular to Circular Waveguide Mode and Impedance Transformer

Transformers connecting rectangular waveguides to circular waveguides are widely used in transmission systems, aerial components, ferrite devices, and measuring instruments. Conventional transitions such as linear tapers [4] or metal [2] and dielectric transformers [3] provide satisfactory matching within a frequency band of a maximum of 10 or 15 per cent only.

Recent developments in measuring instruments—rotary vane attenuator, rotary vane load, rotary phase shifter—designed to cover the entire operating range of the rectangular waveguide, i.e., $1.25f_c$ to $1.90f_c$, require mode and impedance transformers to be introduced between the rectangular and the circular waveguide, which show low reflection over a far wider bandwidth.

This correspondence presents the theory and design of wideband mode and impedance transformer for a reflectionless connection of a standard rectangular waveguide transmitting the fundamental H_{10} mode into a circular waveguide transmitting the fundamental H_{11} mode. The construction uses quarter-wave truncated circular waveguide sections, the characteristic impedances of which are proportional to the Chebyshev coefficients.

The unit is designed to connect a standard rectangular waveguide having the dimension ratio of about 2-to-1 and transmitting the fundamental H_{10} mode to a circular waveguide transmitting the fundamental H_{11} mode. The ratios of the waveguide characteristic impedances are usually within the range 2 to 4 and, in general, vary with frequency. Wide-band characteristics of the described transformer are principally dependent on the circular waveguide dimensions which determine the cutoff wavelength, characteristic impedance, and its variations within the operating range. To obtain optimum operating conditions over the entire band, as recommended for the rectangular waveguide, circular waveguide dimensions have been chosen so that the cutoff wavelengths for both fundamental modes in the waveguides would be equal, thus providing the ratio of characteristic impedances of both waveguides to amount numerically to 2 and to be frequency independent:

$$\lambda_c^{H_{11}} = \lambda_c^{H_{10}}; \qquad \frac{Z_0^{(c)}}{Z_0^{(r)}} = \text{const.} = 2$$

thus

$$A = \frac{2}{1{,}706}\,a \tag{1}$$

where

a = broad wall of the rectangular waveguide.
A = circular waveguide diameter.

It is seen that for the connection of a rectangular waveguide to a circular one, both transmitting fundamental modes, an impedance transformer should be applied, having the characteristic impedance ratio amounting numerically to 2 while being simultaneoulsy a wave mode transformer.

The problem has been solved by a multistep, quarterwave Chebyshev transformer consisting of truncated circular waveguide sections. Figure 1 shows the schematic diagram of the unit. The proposed solution is natural and enables the introduction of a higher mode attenuator (flat absorption vane).

For the proper dimensioning of a multistep quarterwave transformer it is necessary to know how the characteristic impedance and cutoff wavelength of truncated circular waveguide depend on the waveguide dimensions, here on the ratio of B/A, Fig. 1. The problem is dealt with as follows.

The waveguide characteristic impedance and the cutoff wavelength for a truncated circular waveguide, as shown in Fig. 1, were obtained by Valenzuela [5], [6], but during the design of the described transformer these papers were not available to the authors. The approximate method has been applied for the determination of the truncated circular waveguide cutoff frequency. It is based on the assumption that the truncated waveguide cutoff wavelength is equal to that of a rectangular waveguide having the same cross section area, and the height equal to the height of the truncated circular waveguide. Such an assumption leads to the following expression for the truncated circular waveguide cutoff wavelength:

$$\frac{\lambda_c}{A} = \sqrt{1-\eta^2} + \frac{1}{\eta}\arcsin\eta, \tag{2}$$

where

$$\eta = \frac{B}{A} \qquad \text{(see Fig. 1).}$$

Results of the calculations obtained by this method are shown in Fig. 2 and Table I.

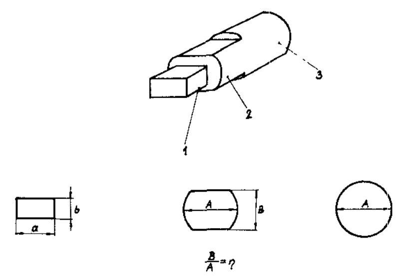

Fig. 1. Mode-and-impedance transformer connecting the rectangular waveguide 1 operating in H_{10} mode into the circular waveguide 3 transmitting the H_{11} mode; transformation takes place in the truncated circular waveguide quarter-wave, section 2.

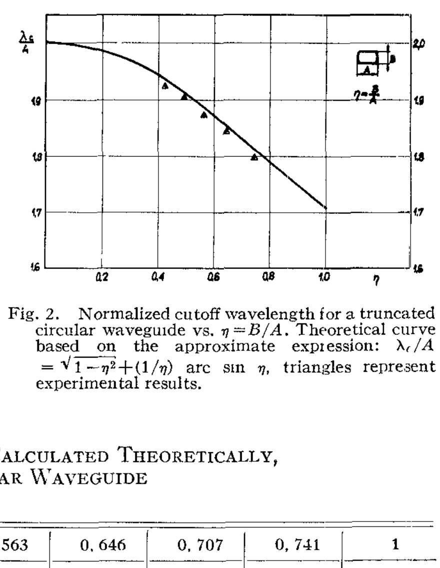

Fig. 2. Normalized cutoff wavelength for a truncated circular waveguide vs. $\eta = B/A$. Theoretical curve based on the approximate expression: $\lambda_c/A = \sqrt{1-\eta^2} + (1/\eta)\arcsin\eta$, triangles represent experimental results.

TABLE I

NORMALIZED CUTOFF WAVELENGTH OF A TRUNCATED CIRCULAR WAVEGUIDE CALCULATED THEORETICALLY, AS COMPARED WITH EXPERIMENTAL RESULTS. $\phi = 26.8$ MM CIRCULAR WAVEGUIDE AT THE OUTPUT, $f = 10.1$ Gc/s

$\eta = B/A$	0	0,125	0,250	0,379	0,425	0,493	0,50	0,563	0,646	0,707	0,741	1
(λ_c/A) theorem	2,00	1,994	1,978	1,951	1,937	1,914	1,912	1,888	1,850	1,817	1,813	1,7065
(λ_c/A) exper.	—	—	—	—	1,925	1,905	—	1,872	1,848	—	1,798	—
λ_g theorem	35,645	35,692	35,821	36,051	36,177	36,361	36,405	36,648	37,053	37,436	37,188	39,021
λ_g exper.	—	—	—	—	36,30	36,50	—	36,80	37,10	—	37,70	—
$\frac{(\lambda_c/A)\text{ theorem}}{(\lambda_c/A)\text{ exper.}}$ %	—	—	—	—	0,6	0,47	—	0,85	0,11	—	0,84	—

The triangles in Fig. 2 represent measured values. The approximate method for the calculation of the truncated circular waveguide cutoff wavelength has proven satisfactory for values of $\eta = B/A$ which are not greater than 0.707 and, as seen in Table I, the difference between theoretical and experimental values is well within 1 per cent. The

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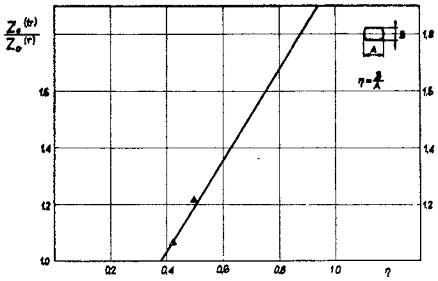

Fig. 3. Normalized characteristic impedance of a truncated circular waveguide vs. $\eta = B/A$. Theoretical curve shown continuously, triangles represent experimental results.

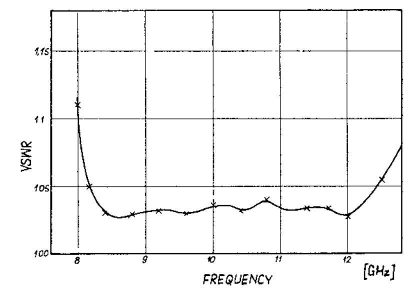

Fig. 4. Input VSWR of a four step Chebyshev's quarter-wave transformer connecting a standard rectangular waveguide WR-90 into circular waveguide with $\phi = 26.8$ mm.

systematic error seen in Fig. 2 is probably due to inadequate measurement accuracy of the truncated circular waveguide cutoff wavelength.

The determination of the truncated circular waveguide characteristic impedance is far more difficult. Also for this case an approximate method was applied by assuming that the truncated circular waveguide characteristic impedance, as normalized to the impedance of a corresponding rectangular waveguide, is a linear function of the ratio of $\eta = B/A$. The function is shown in Fig. 3.

The same assumption as previously used for wavelength calculation has been taken for the determination of the point $Z_0^{(tr)}/Z_0^{(r)} = 1$ while the point for $\eta = 1$ can be calculated theoretically. In the case in question [see (1)]:

for

$$\eta = 1$$

$$\frac{Z_0^{(tr)}}{Z_0^{(r)}} = 2$$

because

$$\lambda_c^{H_{11}} = \lambda_c^{H_{10}}.$$

An experimental verification of the above assumption is rather difficult due to deformations caused by the susceptance of the discontinuity at the junction of two sections, each having different η. However, the two measured points represented in Fig. 3 by triangles, which correspond to low values of η, seem to confirm the assumption at least for values of η close to 0.5.

In the transformer design, tables prepared by Young [7] were used and corrections were applied to the lengths of individual transformer steps as given by Cohn [1]. The transformer input was measured in rectangular waveguide, while the circular waveguide has been terminated by a matched sliding load. The transformer input VSWR as a function of frequency is given in Fig. 4.

In conclusion, an approximate theory was used for the determination of the truncated circular waveguide cutoff wavelength and its characteristic impedance. The obtained results were verified by experiment.

Basing on verified theoretical results a wideband mode-and-impedance transformer has been designed which provides reflectionless connection of a rectangular waveguide with H_{10} mode into a circular waveguide with H_{11} fundamental mode. The transformer consists of four quarter-wave truncated circular sections, each with its characteristic impedance being proportional to the Chebyshev coefficients. The unit connects a standard WR-90 waveguide to a ϕ-26.8 mm circular waveguide. The described transformer shows an input VSWR less than 1.05 over the frequency range of 8.2 to 12.4 Gc/s. The total transformer length is ca 35 mm ($1\frac{1}{2}$ inch).

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A Magnetic Loop, Diode Detector Compatible with Microwave and Beam Waveguides

Introduction

Improved detector performance is a common objective in any communication system. This is particularly true of the millimeter and submillimeter region [1] where the available signals are weak.

Point contact crystal detectors have almost a unique position in the microwave region. They also have been used at wavelengths as short as 0.5 mm. However, the conventional packaging configuration in a waveguide is difficult to realize and losses become excessive. Also, the typical electric probe arrangement cannot be adapted to the low loss, beam waveguide system [2], [3], more suitable for the submillimeter wave region.

In this letter a magnetic probe [4] crystal diode arrangement shown in Fig. 1 will be described. This detector configuration has the advantages of compatibility with both microwave and beam waveguides and, in measurements made to date, sensitivity somewhat better than a 1N53 cartridge in a conventional mount.[1]

As shown in Fig. 1, the semiconductor crystal chip is mounted on the end of the inner tungsten conductor of a coaxial line. The loop whisker wire is fastened to the outer conductor of the cartridge with the sharpened point making contact with the crystal in the normal fashion.

When used with a conventional waveguide, the loop detector is mounted at the end of the guide with the loop oriented parallel to the electric field. In this arrangement the maximum magnetic field will tend to remain at the loop and the coupling is nearly constant as the frequency changes.

In the case of a reflecting beam waveguide, the loop detector is placed in the center of one of the guide reflectors. Since the ratio of the loop to beam spot size is unfavorable, the coupling in general will be small. However, it can be greatly improved by making one section of the reflecting beam waveguide into a resonator as illustrated in Fig. 2. The resonator can be formed by placing a plane reflector and a grating coupler [5] (no. 1) in the positions shown in Fig. 2. In frequency mixing experiments, local oscillator power can be coupled into the system by forming a second resonator between the guide reflector and a second grating coupler (no. 2) placed as shown in Fig. 2. The resonator system is inherently narrow band (1–3 Mc) but is easily tunable by varying the spacing between the elements of the system.

Construction Procedure

The construction procedure for the prototype models of the cartridge shown in Fig. 1 is as follows:

A crystal chip is soldered to the end of a tungsten rod 0.030-inch in diameter and 0.750-inch long. Copper is then electroformed around the tungsten rod and crystal and the electroformed copper is then machined to a uniform diameter of 0.40-inch. The insulating layer can be made in either of two ways depending on the thickness of insulation desired. For operation at very short wavelengths (1–2 mm) a thin layer is desirable to prevent RF leakage into the coaxial system and is made by covering the post with a layer of glyptal. The outer conductor of the cartridge is then made by covering the glyptal with a layer of conducting silver paint and electroforming over the silver paint. It is desirable to have a hole in the outer conductor to house a 0.004-inch tungsten wire which serves as the coupling loop and whisker combined. This can be accomplished by electroforming to a diameter of approximately 0.080-inch, machining a

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[1] The mount used was a DE MORNAY-BONARDI type DBA-313.