Broad-Band Stepped Transformers from Rectangular to Double-Ridged Waveguide*

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Summary—**The design of a series of broad-band Tchebycheff-type stepped waveguide transformers from various sizes of standard rectangular waveguides to a double ridged waveguide covering the frequency range of 4750 to 11,000 mc is described. Four separate transformers employing RG-67/U (WR-90), RG-68/U (WR-112), RG-106/U (WR-137), and WR-159 to Airtron ARA-133 double-ridged waveguide have been designed using this technique and cast in aluminum. The complete frequency range is covered by several pairs depending on which sizes of mating rectangular waveguides are desired. The RG-106/U design covers a frequency range of 53 per cent with a maximum VSWR of 1.08, while the other three designs each cover a slightly smaller frequency band with a VSWR not exceeding 1.05. Along with the experimental results obtained, an outline of the design method is given which can be used to design similar transformers between any compatible rectangular and double ridged waveguides.**

Introduction

CONSIDERABLE information on the design of stepped single-ridged transformers to rectangular waveguide is available in the literature. However, despite the wide usage of double-ridged waveguides in present day systems, information on the design of well-matched broad-band double-ridged transformers to rectangular waveguide has been noticeably lacking. In this article, the work of Cohn[1] and Hopfer[2] dealing with single-ridged transformers is combined, and suitably modified to yield an accurate design method for double-ridged stepped transformers made between waveguides of arbitrarily selected aspect ratios and cross-sectional dimensions.

General Considerations

The use of stepped impedance sections proportioned to Tchebycheff polynominals will always result in better broad-band results for a given length and bandwidth than the use of tapers or other step functions. The amount of calculation involved in designing Tchebycheff transformers may be somewhat more laborious, but the extra effort is more than justified by the improvement in the end results.

The form of the transformer is shown in Fig. 1. The structure consists of a symmetrical series of steps on the top and bottom of the waveguide with approximately quarter-wavelength spacing between steps. The sections

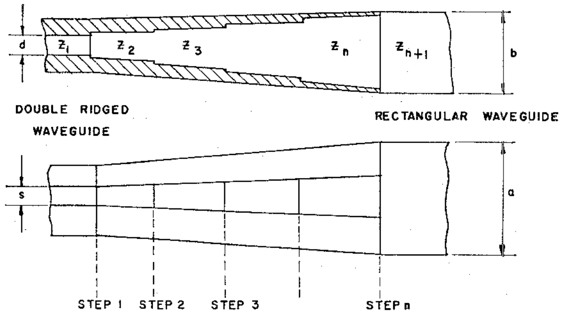

Fig. 1—The double-ridged stepped transformer.

between steps are designated by Z_2, Z_3, etc. Each of these sections is considered to maintain an approximately constant value of impedance throughout its length. For example, if section Z_3 is designed for 215 ohms, the impedance is very close to 215 ohms from one end of section Z_3 to the other end. The symbols d, b, s, and a given in Fig. 1 are intended to cover the gap distance, height of the waveguide, width of the ridges, and width of the waveguide, respectively, at any point throughout the length of the transformer.

As a rule, the rectangular waveguide will be larger in cross section than the ridged waveguide to which it is to be mated. This is due to the large reduction in TE_{10} cutoff wavelength accomplished by the ridge loading. Therefore, for compatible frequency characteristics, the unloaded guide will be considerably larger. The dimensions of the ARA-133 waveguide used in these particular transformer designs are shown in Table IV in a paper by Anderson.[3]

The problem of mating double-ridged waveguide to a rectangular waveguide which is larger in cross section, and still maintaining intermediate sections with constant impedance, can be solved in two ways. The outside dimensions of the waveguides can be blended together by either smooth tapers in both the E and H planes, or by abruptly stepping the waveguide along with the steps in the center ridges. The latter solution is rather impractical from a fabrication standpoint, and more important, the electrical design would be unduly complicated by the large discontinuity susceptances which would have to be compensated. Therefore, the transition is made with smooth tapers in the outside dimensions and in the width of the ridges, thereby main-

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[1] S. B. Cohn, "Optimum design of stepped transmission-line transformers," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-3, pp. 16–21; April, 1955.

[2] S. Hopfer, "The design of ridged waveguide," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-3, pp. 20–29; October, 1955.

[3] T. N. Anderson, "Rectangular and ridge waveguide," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-4, pp. 201–209; October, 1956.

taining a constant ratio of ridge width to guide width (s/a ratio) throughout the length of the transformer. To complete the necessary requirements for sections with constant impedance, the ratio of the gap between the ridges to the total height of the waveguide (d/b ratio) is calculated for each section by the method given below.

Design Method

The number of steps required for a given maximum VSWR can be roughly calculated by a trial and error solution of (5) in Cohn's article.[4] This equation is given below. Z_1 and Z_{n+1} are the terminating characteristic impedances as defined in Fig. 1. $T_m(x)$ is the Tchebycheff polynomial of mth degree, and

$$S_{\max} = 1 + \frac{\ln\left[\dfrac{Z_{n+1}}{Z_1}\right]}{T_{n-1}\left[\dfrac{1}{\cos\phi_1}\right]}.$$

ϕ_1 is the electrical spacing of the steps at the low-frequency end of the band. The equation is derived on the basis of small steps; however, Cohn has shown that the equation holds very well even with fairly large steps. For this type of design where over-all VSWR's of 1.10 or less are desired, it is recommended that at least four or five steps be used to minimize the discontinuity susceptances presented by the steps and the small inherent errors in the design method.

Once the number of steps has been chosen, the Tchebycheff coefficients and the resulting characteristic impedances required for each section are calculated by the methods outlined by either Cohn or Hopfer. As Hopfer points out, use of the impedances at infinite frequency ($Z_{0\infty}$) yields better results than the use of the actual impedances at midband, and this fact has been further substantiated by the results given in this article.

Calculation of the dimensions of the stepped ridge sections according to the prescribed impedances is then undertaken. Since the a and b dimensions of the transition section are tapered, the problem arises as to what specific values of a and b should be used. It has been found from experiment that very good results are obtained by dividing into $(n-1)$ equal parts the differences between the a and b dimensions of the two waveguide sizes used, and then calculating from these values the mean values of a and b in each impedance section. Use of these values for determination of the d/b ratio will yield the mean value of impedance in the section being considered. A slight variation in impedance from one end of the section to the other usually occurs; however, this variation is only about ± 1 ohm if reasonable tapers have been employed, and therefore, the assumption of constant impedance sections is considered to be valid. If the tapers are quite pronounced, the use of the mean values of a and b in each section will introduce objectionable deviations from the mean values of impedance. This can be easily overcome by determining the d/b ratio at each end of the section using the a's and b's at these points instead of the mean values. This method actually yields a much closer approximation to a constant impedance section, and is to be preferred. However, the amount of calculation is doubled, and is only necessary for severe tapers. The actual values of d/b ratio required at each point are determined with the aid of Fig. 2. This is a graph of relative admittance vs d/b ratio taken from information given on Fig. 16 of Hopfer. Hopfer's graph was worked out for single-ridged guide, but since there is no similar graph available for double-ridged guide, it was decided to try these data. Fortunately, actual values of impedance are secondary to relative values in waveguide transformer considerations and excellent results have been achieved using this graph, even though the data apply primarily to single-ridged guide.

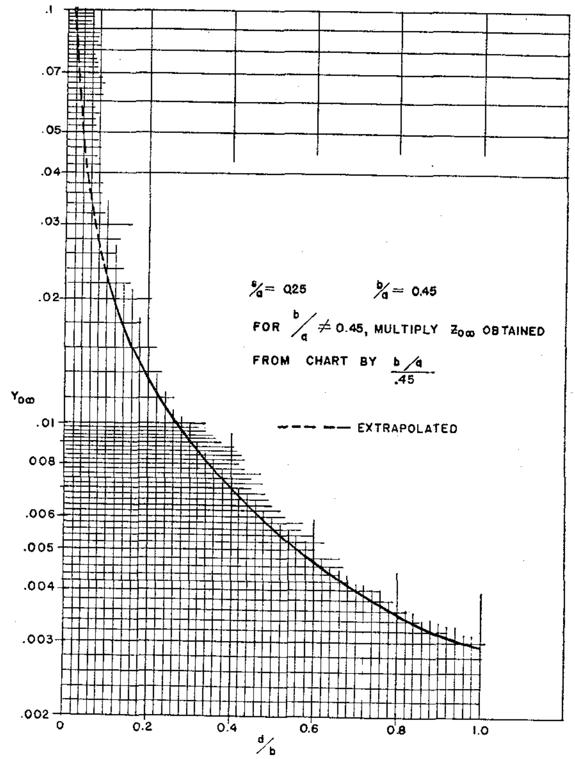

Fig. 2—Relative admittance chart for double-ridged waveguide transformer sections. Plotted from Fig. 16 of Hopfer[2] for constant value of $s/a=0.25$.

If Fig. 2 is plotted on linear coordinates, it can be seen that the resultant curve is very closely a straight line up to approximately a value of $d/b=0.8$. After this, the function drops off more and more rapidly from the straight line. This fact is significant because heretofore both double- and single-ridged transformers were designed using a straight line function to determine the relative impedance values between Z_1 and Z_{n+1}.

Fig. 2 is plotted for a constant s/a ratio of 0.25. This value of s/a has been chosen because most double-ridged waveguides are designed with this ratio which

[4] *Op. cit.*, p. 17.

represents the optimum in attenuation and high power handling characteristics. If other s/a ratios are desired, Hopfer's original graph should be used.

The length required for each of the constant impedance sections is then calculated. Since the a and b dimensions vary according to the outside tapers, the cutoff characteristics vary throughout the length of each section, thereby introducing varying λ_g values. To overcome this problem, the TE_{01} mode cutoff frequencies are determined from Hopfer's Fig. 2 for both ends of each section (TE_{20} mode cutoff wavelength should also be checked to be sure that this mode cannot be supported). The mean λ_g at one end of the section under consideration is then calculated from

$$\lambda_{gm} = \frac{2(\lambda_{g1}\lambda_{g2})}{\lambda_{g1} + \lambda_{g2}}.$$

λ_{g1} and λ_{g2} correspond to the guide wavelengths at one end of the section at the low- and high-frequency ends of the design band, respectively. The mean λ_g at the opposite end of the same section is then calculated in the same manner after its TE_{01} mode cutoff frequency has been determined. The actual length of the section (S) is found by combining the λ_{gm} values in the next equation. This procedure is employed for all of the constant impedance sections in the transformer. In general,

$$S = \frac{\lambda_{gm1}\lambda_{gm2}}{2(\lambda_{gm1} + \lambda_{gm2})},$$

all of the section lengths will be different.

Since all of the lengths are different, the original assumption of $(n-1)$ equal length sections for calculation purposes is now slightly in error for constant tapers. However, considerable variation in length can be tolerated since the angles of the tapers are so small that the change in a and b is negligible in most cases. For the transformers designed, except for the WR-159 case, the maximum discrepancy in a and b did not exceed 0.003 inch. If the discrepancies exceed this figure, as in the case of the WR-159 transformer design, the calculations may be easily corrected by using the lengths arrived at in the first approximation to recalculate a new series of d/b ratios and cutoff frequencies with the adjusted a and b dimensions.

Cohn has included in his article a method for compensating the step discontinuity susceptances by decreasing the length of the sections so affected. If four or less steps are used for the transformer, it is suggested that Cohn's procedure for compensation be employed to modify the section lengths. However, if more than four steps are used, the steps are generally small enough for the compensation to become negligible.

Experimental Results

Four transformers from different sizes of standard rectangular waveguide to ARA-133 double-ridged waveguide were designed according to the above principles to cover the frequency range of 4750 to 11,000 mc. Fig. 3 is a tabulation of the pertinent data on these transformer designs. Fig. 4 gives the final VSWR results on a typical RG-68/U aluminum production unit. It can be seen that the VSWR does not exceed 1.05 throughout the frequency range of this unit. The RG-67/U and WR-159 designs yielded comparable results over their respective frequency ranges; however, the RG-106/U design exhibited a maximum VSWR of 1.08. This higher value of VSWR can be attributed to the fact that this design not only covered the entire recommended operating range of RG-106/U, but it extended another 1100 mc below this range as well. A typical production sample of this design is shown in Fig. 5.

No preliminary fabricated design models were made; the designs were calculated and the resulting dimensions were transferred directly through the tools to one-piece castings. None of the designs required further design effort except the WR-159 transformer. The first castings on this unit exhibited the typical Tchebycheff

Waveguide Size	Transformer Frequency Range (mc)	Per cent Bandwidth	Number of Steps	*Length (inches)
RG-67/U	8200–11000	29	5	1.48
RG-68/U	7050–10800	44	6	1.90
RG-106/U	4750– 8200	53	6	3.16
WR-159	4750– 7350	43	7	4.09

* Transformer section only.

Fig. 3—Tabulation of transformers designed to mate with ARA-133 double-ridged waveguide.

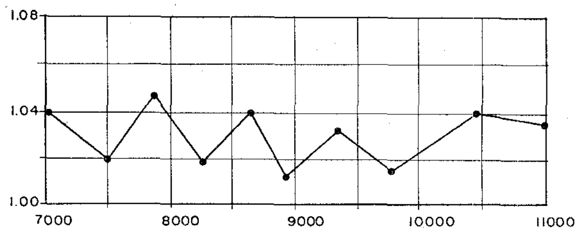

Fig. 4—VSWR vs frequency for RG-68/U to ARA-133 six-step transformer.

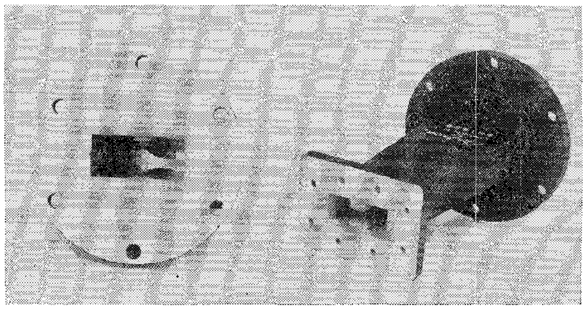

Fig. 5

ripple response, but the ripple varied between VSWR's of approximately 1.09 to 1.05. This difficulty was not surprising due to the extreme tapers required in the *a* and *b* dimensions, but a maximum VSWR of 1.05 was arrived at fairly easily by trial and error experimentation with the step dimensions.

It is interesting to note that further improvements probably could have been made on these transformer designs, since one of the RG-68/U castings tested did not exceed a maximum VSWR of 1.02 over the 7050 to 10,800-mc range. This outstanding performance can be attributed to the small dimensional variations which occur among many castings of the same design. Obviously, this particular casting possessed exactly the right combination of dimensions. However, for production purposes, a VSWR limit of 1.05 is much more practical since the tolerances which would have to be held to maintain the 1.02 limit are much too tight to be attainable.

Mechanical tolerances on these units presented some problems. In general, it was found that the distance between steps was relatively uncritical, but the height of the steps proved to be quite critical. Tolerances in the order of ± 0.004 inch were found to be adequate for the section lengths, but changes as small as 0.0015 inch in the step heights introduced measurable differences in the VSWR patterns.

Despite the simplifying assumptions and small errors in the design method, the results obtained are very satisfactory, and it is felt that the design method has been proven to be reliable and accurate for this type of waveguide transformer.

A Wide-Band Balun*

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Summary—**Experimental results are given for a transformer from an unbalanced 50-ohm coaxial line to a balanced pair of 50-ohm coaxial lines. The design is one proposed by Marchand. The balance, standing wave ratio, and insertion loss are nearly constant over a 13 to 1 frequency range from 650 mc to 8500 mc. The standing wave ratio is less than 2.1 to one and the insertion loss is about 0.5 db over this band of frequencies.**

Introduction

SEVERAL authors[1–4] have presented theoretical analyses of devices suitable for transforming from a balanced to an unbalanced transmission line over a wide frequency range. It is the purpose of this paper to present some experimental results on a balun of the type described by Marchand[1] which provides satisfactory performance over a greater than ten to one band of frequencies extending from 650 to 8500 mc. One use for a balun with this sort of frequency coverage is in connection with microwave oscillators that can tune over frequency ranges of this order of magnitude and that can often be most effectively designed with balanced two-conductor interaction circuits.[5–7] The output of such tubes is in many cases most conveniently brought out of the vacuum envelope by means of two separate coaxial lines having a balanced signal between their center conductors. The balun to be described has been constructed to operate between such a two-coax system and a single coax, but only a slight change of the construction of the balanced output would be required to convert to a balanced shielded pair instead of two separate coaxial lines. Another possible use for a balun of this type would be as a microwave power splitter to obtain two equivalent outputs which differ in time phase by 180°.

Fig. 1 is a photograph of the balun. It transforms an unbalanced input signal supplied to the 50-ohm Hewlett-Packard G-76A receptacle input mounted on the brass cylinder into a balanced output signal appearing between the center conductors of the two 50-ohm RG5/U cables.

Fig. 2 is a schematic diagram of the device as proposed by Marchand.[8] Z_{OC} is the impedance of the unbalanced line C, Z_{OT} is the impedance of the large outer

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† Electronics Res. Lab., Stanford University, Stanford, Calif.

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[8] N. Marchand, *op. cit.*, Fig. 10.