The Compensated Balun

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Abstract—The compensated balun, first described by Marchand and later reinvented by Roberts, is found to have much broader bandwidths than realized by either author. This balun is analyzed here and the parameters which determine its bandwidths are discussed. Curves are presented which allow the design of a balun having any bandwidth. The practical considerations which preclude realization of infinite bandwidths are pointed out. Since the balun is one-quarter wavelength long at center frequency, the balun can be less than one-twentieth wavelength at its lowest operating frequency. Because impedance transformation is generally required with balanced-to-unbalanced line transformation, two techniques including impedance transformation are discused. The reactance slope of the compensated balun is controllable, which allows the balun to directly compensate for the reactance slope of some loads. An example of the compensation of a dipole antenna over a 17-percent bandwidth is discussed. Finally, experimental verification of the balun theory is presented, along with an application to an S-band impedance-transforming compensated balun.

List of Symbols

λ = Wavelength
ω = Radian frequency
f = Frequency
f_0 = Band-center frequency $= (f_{x1}+f_{x2})/2$
f_{x1} = Lower band-edge frequency
f_{x2} = Upper band-edge frequency
r = Resistive component of balun impedance
R = Load resistance (balanced line)
S = Optimum generator resistance
x = Reactive component of balun impedance
Z = Balun impedance
Z_a = Characteristic impedance of unbalanced input line ($Z_a = S$)
Z_{ab} = Characteristic impedance of unbalanced two-wire line formed by outer conductors Z_a and Z_b
Z_b = Characteristic impedance of compensating line
Z_c = Characteristic impedance of balanced output line ($Z_c = R$)
$\alpha = Z_b/R \cdot Z_{ab}/R = \beta\gamma$
$\beta = Z_b/R$
$\gamma = Z_{ab}/R$.

I. Introduction

THE FIRST transmission line balun (balanced-to-unbalanced line transformer) was described in the literature by Lindenblad[1] in 1939 and variations on his original scheme soon followed. Among these was that of Marchand,[2] who introduced a series open-circuited line (Z_b in Fig. 1) to compensate for the short-circuited line reactance of the two wires (Z_{ab}). Three other variations were cataloged and described by the Harvard Radio Research Laboratory staff.[3] Harvard's Type II balun is similar to Marchand's variation but without compensation. Because of the compensating feature of Marchand's variation the author suggests the name "compensated balun" to distinguish it from the others.

In 1957, Roberts[4] apparently reinvented the compensated balun and the author used Roberts' paper as a starting point of his initial analysis.[5] In 1958, McLaughlin, Dunn, and Grow,[6] using Marchand's paper, designed a 13-to-1 bandwidth compensated balun which is the broadest thus far published. Bawer and Wolfe[7] subsequently applied the compensated balun to a broadband spiral antenna.

The analysis presented here is an expansion of the author's initial analysis. The latter redetermines the optimization criterion of Marchand and includes two methods of combining impedance level transformation with balun action. This paper points out that Marchand's optimization criterion is not optimum after all. It continues with presentation of curves for rapid optimum design of a compensated balun. The reactance slope of the balun can be made either positive or negative, as will be discussed. Hence, the balun can be used to create a complementary reactance to some balanced loads and provide an improved match over a band. Oscilloscope curves of the inverse balun impedance[8] are presented, which have been computed and displayed by the TRW Systems Group On-Line Computer. These are useful in designing a complementary matching balun.

Experimental verification of the theory is presented using a 4-to-1 bandwidth compensated balun without

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[1] N. E. Lindenblad, "Television transmitting antenna for Empire State Building," *RCA Rev.*, vol. 3, pp. 387–408, April 1939.

[2] N. Marchand, "Transmission line conversion transformers," *Electronics*, vol. 17, pp. 142–145, December 1944.

[3] Radio Research Laboratory Staff, *Very High-Frequency Techniques*. New York: McGraw-Hill, 1947, vol. 1, p. 88.

[4] W. K. Roberts, "A new wide-band balun," *Proc. IRE*, vol. 45, pp. 1628–1631, December 1957.

[5] H. G. Oltman, Jr., "Analysis of the compensated balun," Rantec Corp., Calif., Tech. Rept., May 1961.

[6] J. W. McLaughlin, D. A. Dunn, and R. W. Grow, "A wide-band balun," *IRE Trans. on Microwave Theory and Techniques*, vol. MTT-6, pp. 314–316, July 1958.

[7] R. Bawer and J. J. Wolfe, "A printed circuit balun for use with a spiral antenna," *IRE Trans. on Microwave Theory and Techniques*, vol. MTT-8, pp. 319–325, May 1960.

[8] The impedance looking into the balun from the balanced terminal with a matched load on the unbalanced terminal. It is obtained by inverting the impedance matrix.

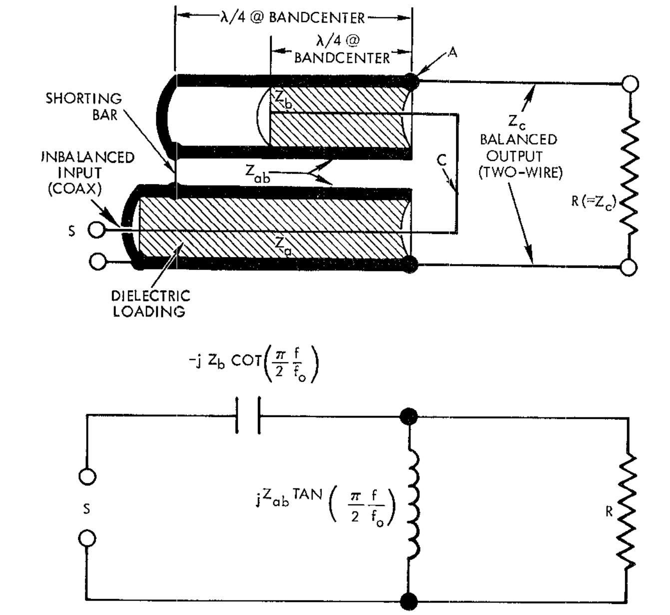

Fig. 1. Schematic and equivalent circuit of the compensated balun.

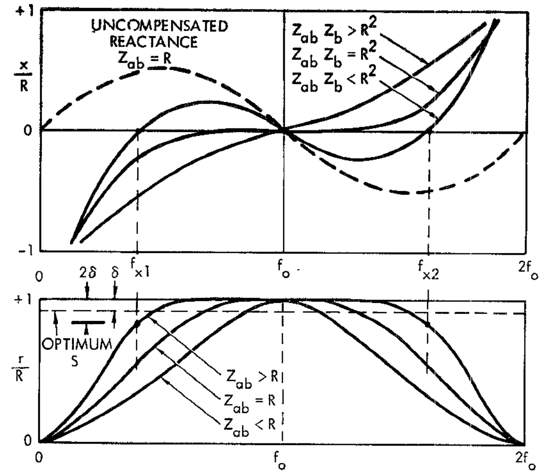

Fig. 2. Variation of the compensated balun impedance with the line and load parameters Z_{ab}, Z_b, and R. (a) Reactance. (b) Resistance.

impedance transformation. As an example of a balun with impedance transformation, a 4-to-3 bandwidth balun with a 50- to 380-ohm impedance transformer on the balance end is described.

II. Principle of Operation

Figure 1 is a schematic of the compensated balun using Robert's notation. It differs from the RRL Type III balun[3] in that, in the latter, the center conductor of Z_a terminates at point A rather than passing into Z_b. Lines Z_{ab} and Z_b are one-quarter wavelength long at center frequency. Therefore, at center frequency the load resistance R is seen unchanged at the right-hand terminals of Z_a, i.e., $Z=R+j0$. This is true for all baluns of this general type. Off band center, however, the shunting reactances of Z_{ab} alters the impedance seen at the end of Z_a. As seen in Fig. 2(a), the compensating line Z_b can reduce the balun reactance about center frequency, or reverse its sign, depending on the magnitudes of Z_b, Z_{ab}, and R. The compensating line, in essence, provides control of the reactance about center frequency. The band over which control is useful depends only on the relative magnitudes of Z_{ab} and R, as is indicated in Fig. 2(b). For broadest bandwidth, it is desirable to make Z_{ab} as large as possible relative to R. This fact is more readily appreciated if Z_{ab} is allowed to increase towards infinity. Then, the current entering Z_{ab} approaches zero and becomes negligible compared to the current entering the load resistance R. Consequently, the current entering the unbalanced line terminals passes out the balanced line terminals unimpeded.

Marchand's optimization criterion $Z_b=R^2/Z_{ab}$ determines the value of Z_b to cause the slope of the balun reactance to be zero at band center. This is close, but not quite the optimum condition for broadest band with a maximum mismatch less than a specified amount. The optimum value of Z_b and generator impedance is developed in Section III. There it is shown that the mismatch VSWR obtained using Marchand's criterion can be reduced by approximately a factor of two.

III. Analysis

The balun input impedance derived from the equivalent circuit of Fig. 1(b) is

$$Z = r + jx = \frac{jRZ_{ab}\tan\left(\frac{\pi}{2}\frac{f}{f_0}\right)}{R + jZ_{ab}\tan\left(\frac{\pi}{2}\frac{f}{f_0}\right)} - jZ_b\cot\left(\frac{\pi}{2}\frac{f}{f_0}\right), \tag{1}$$

where

$$f_0 = \frac{f_{x1}+f_{x2}}{2}.$$

The resistive and reactive components (normalized by the load resistance R) are:

$$\frac{r}{R} = \frac{\frac{Z_{ab}^2}{R^2}\tan^2\left(\frac{\pi}{2}\frac{f}{f_0}\right)}{1+\frac{Z_{ab}^2}{R^2}\tan^2\left(\frac{\pi}{2}\frac{f}{f_0}\right)} \tag{2}$$

$$\frac{x}{R} = \frac{\left(\frac{Z_{ab}}{R}\right)\left(1-\frac{Z_{ab}Z_b}{R^2}\right)\tan\left(\frac{\pi}{2}\frac{f}{f_0}\right) - \frac{Z_b}{R}\cot\left(\frac{\pi}{2}\frac{f}{f_0}\right)}{1+\left(\frac{Z_{ab}}{R}\right)^2\tan^2\left(\frac{\pi}{2}\frac{f}{f_0}\right)}. \tag{3}$$

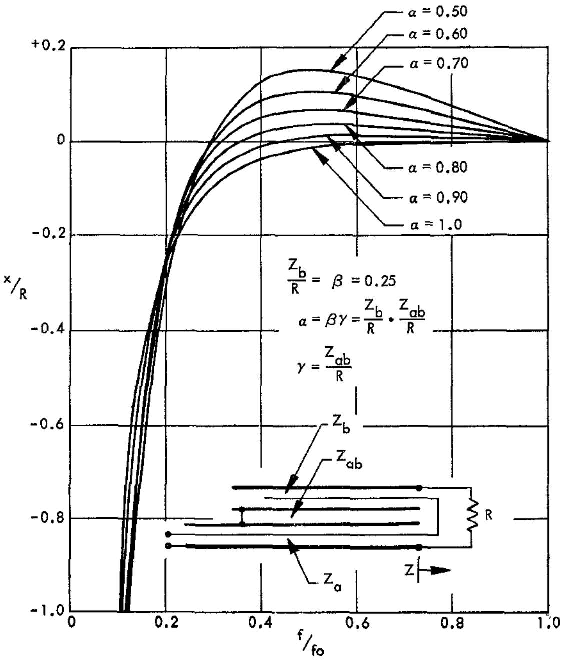

Fig. 3. Compensated balun reactance as a function of α for $\beta = 0.25$.

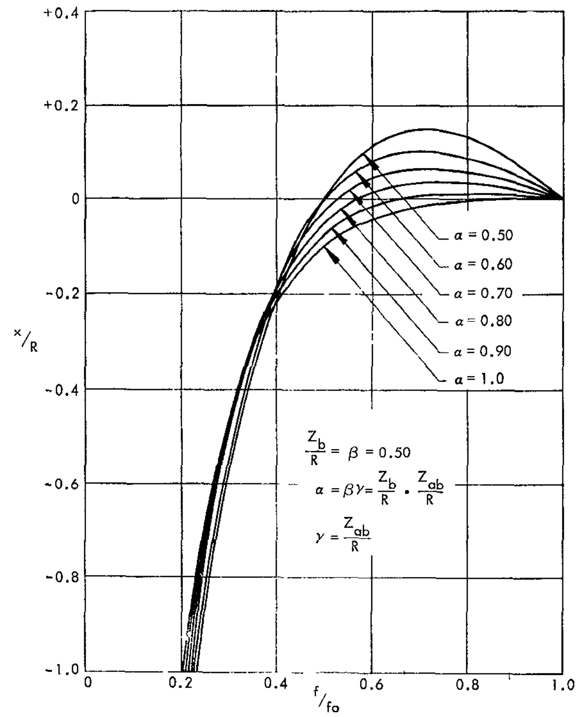

Fig. 4. Compensated balun reactance as a function of α for $\beta = 0.50$.

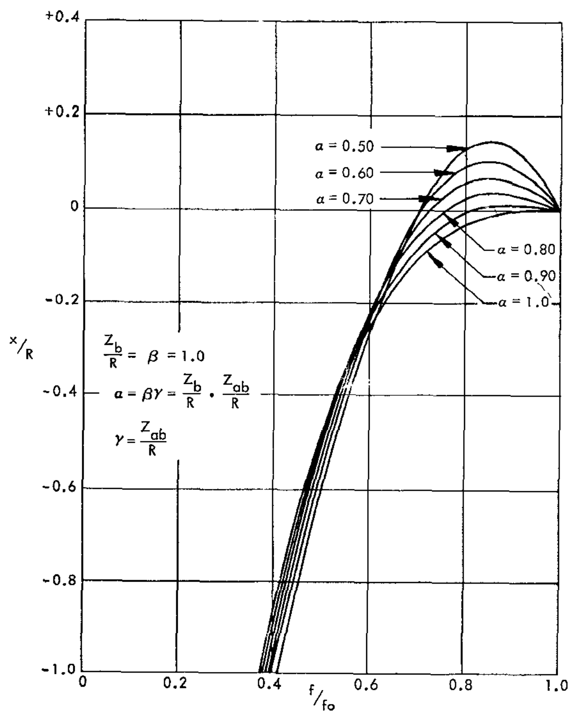

Fig. 5. Compensated balun reactance as a function of α for $\beta = 1.0$.

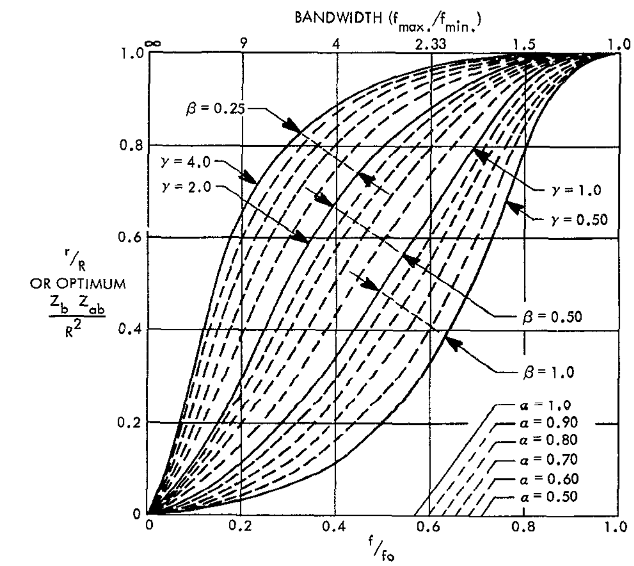

Fig. 6. Compensated balun resistance or optimum value of α. Parameters marked for comparison with Figs. 3, 4, and 5.

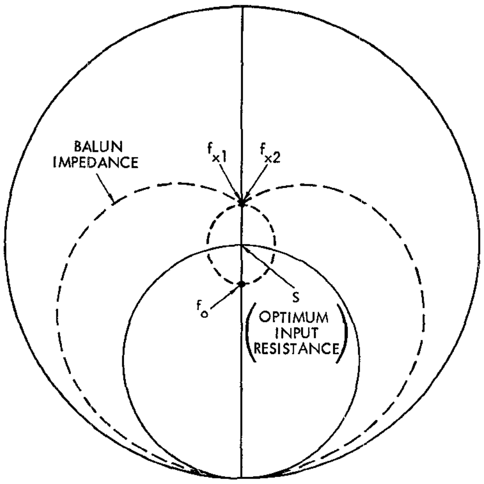

Fig. 7. Typical impedance plot of an optimum design compensated balun.

It is seen that the resistive term is independent of the compensating reactance Z_b. At center frequency, $r=R$ and $x=0$ independent of Z_{ab} or Z_b. Thus, at center frequency the output impedance is exactly transformed into the input. The reactive term (3) is plotted in Figs. 3, 4, and 5 for three values of Z_b/R. On each graph curves are drawn for six values of $\alpha=Z_{ab}Z_b/R^2$. These graphs show the control exercised by the compensating reactance Z_b and the quantity α. Figure 6 is a plot of the resistive term and has been labeled so that it may be easily compared with Figs. 3, 4, and 5. Figure 6 is, actually, a plot of r/R for values of $\gamma=Z_{ab}/R$, varying from 0.5 to 1.0 in 0.1 steps, 1.0 to 2.0 in 0.2 steps, 2.0 to 4.0 in 0.4 steps. This graph will also be useful in determining the reactance zero crossings and band edges, as will be discussed later in this section.

The factors which determine the slope of the balun impedance at center frequency are determined by differentiation of Z with respect to frequency. The results are:

$$\frac{dZ}{df} = -j\frac{\pi}{2}\frac{R^2 - Z_{ab}Z_b}{Z_{ab}} \quad \text{at } f \approx f_0. \tag{4}$$

If $R^2=Z_{ab}Z_b$, $dZ/df=0$. The latter is the optimization criterion originally derived by Marchand. Because of the lack of a real term, the resistance slope is zero independent of the parameters at center frequency. Thus, Marchand's criterion can be considered to be the criterion for zero reactance slope at center frequency. If $R^2>Z_{ab}Z_b$, the slope is negative, and if $R^2<Z_{ab}Z_b$, it is positive. It is convenient to define the parameter $\alpha=Z_{ab}Z_b/R^2$ as an indicator of the reactance slope. Marchand's criterion is indicated by $\alpha=1$.

These results along with Fig. 2 show several interesting features. First, the band over which the balun input resistance r is nearly constant depends only on the ratio Z_{ab}/R. The larger the characteristic impedance of the two-wire line segment Z_{ab}, the larger the band over which the balun input resistance is constant. A similar situation exists for the balun input reactance. Because of the term Z_{ab} in the denominator of the reactance slope equation (4), the larger the value of Z_{ab}, the smaller the slope. Thus, the deviation of the reactance from zero (which holds throughout the band) is reduced. In general, therefore, it is desirable to choose Z_{ab} as large as possible in order to achieve either the largest bandwidth the best match, or both.

Second, Z_b can be considered to be the control on the balun input reactance slope and/or the control of the band-edge reactance zero crossing. Since Z_{ab} is generally chosen as large as possible, this leaves Z_b as a variable to control the reactance. As mentioned before, the balun input resistance is independent of Z_b.

A. Criterion for Minimum Mismatch Throughout Band

Marchand's criterion is not the criterion for minimum mismatch throughout the band. It can be rigorously shown, by minimizing the variation of the vectorial quantity $\boldsymbol{Z}-\boldsymbol{S}$ with respect to variations in the compensating reactance Z_b, that the minimum mismatch at a specific off-center frequency occurs when that frequency is a reactance zero crossing. The minimizing process (not reproduced here) yields the requirement that,

$$Z_b = \frac{Z_{ab}\tan^2\left(\dfrac{\pi}{2}\dfrac{f_x}{f_0}\right)}{1+\left(\dfrac{Z_{ab}}{R}\right)^2\tan^2\left(\dfrac{\pi}{2}\dfrac{f_x}{f_0}\right)}, \tag{5}$$

where f_x is either the upper or lower zero-crossing frequency. Equation (5) reduces to Marchand's criterion for $f_x \approx f_0$. It is worth while noting that if (5) is multiplied by Z_{ab}/R^2, and if f_x is replaced by f, then the right-hand equality of (5) becomes identical to (2), the normalized balun input resistance. The latter is plotted in Fig. 6. Thus, Fig. 6 is useful in determining the optimum value of Z_b after Z_{ab} and the band edges f_x have been chosen.

With the restriction of (5), the balun impedance curve, when plotted on a Smith Chart, has a loop in it. It then remains to adjust the position of the loop so that it symmetrically encircles the generator impedance S, as shown in Fig. 7. This insures that the mismatch throughout the band is less than, or equal to, the zero-crossing mismatch. It can be shown by equating the values of $|\boldsymbol{Z}-\boldsymbol{S}|$ at band center and either band edge that the optimum generator impedance S is given by:

$$S = \frac{R}{2}\left(1+\frac{Z_bZ_{ab}}{R^2}\right). \tag{6}$$

This yields the best match over the band. Using it, the balun has a response somewhat similar to that of equal-ripple filters and transformers. If the actual generator impedance is not equal to the optimum generator impedance, which is most likely, an impedance transformer can be used with some degradation in match.

IV. Balun with Impedance Transformation

Step transformers lend themselves well to incorporation into the existing balun structure. The unspecified line length Z_a can be replaced by a step transformer with no change in balun dimenions. Using ordinary dielectrics, almost two quarter-wave transformer sections can be incorporated into Z_a. Extensive tables of quarter-wave transformers are in the literature.[9] If the balanced line impedance level is not too large, more than eight sections of a short-step[10] transformer can be incorporated into Z_a.[11]

In addition to the restrictions on maximum impedance level practical in coaxial lines, operating the balun terminals at a high level narrows the potentially available bandwidth. Practical balanced line and load impedances are normally always higher than their coaxial counterparts. But, because maximum balun bandwidth varies in proportion to the ratio $\gamma = Z_{ab}/R$, it is desirable to make the impedance looking toward the balanced load small. Thus, it is preferable to transform the high-balanced load impedance towards the lower coaxial line impedance. Unfortunately, this increases the physical length according to the number of transformer sections required. However, use of short-step transformers can appreciably limit the necessary length. Trade-offs must be chosen according to the requirements of each application. It also appears practical to split the transformers and incorporate one part into the coaxial line Z_a and the other into the balanced line Z_c.

V. Balun as a Complementary Matching Network

The reactance control exhibited by the compensating line Z_b has a valuable application as a complementary matching network. An example of this application is matching a dipole. A dipole has an increasing resistance and reactance with frequency. The dipole would be perfectly matched if it were attached to a source having its complementary impedance. The resistance of the dipole complementary impedance should also increase with frequency, but the reactance should decrease. This can be accomplished over a limited range as can be seen in the plots of the inverse balun impedance in Fig. 8(a) and (b).

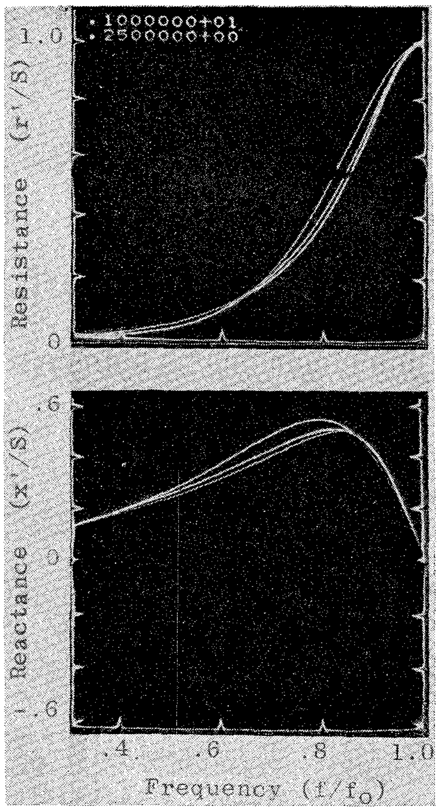

Fig. 8. Computed inverse compensated balun impedance for $\gamma = 0.25$. β takes on values of 1, .5, and .25.

Figure 8(a) is a group of successive curves of the inverse normalized balun resistance,[8] r'/S. Figure 8(b) is the corresponding group of reactance curves x'/S. The varying parameter is $\beta' = Z_b/S$. Just below center frequency, some of these curves show an increasing resistance and a decreasing reactance as required by a generator that is complementary to a dipole impedance. These curves were calculated and displayed on a memory oscilloscope by the author using the TRW On-Line Computer manufactured by Bunker-Ramo.

Figures 9 and 10 show the results of using the compensated balun to improve the match of a dipole antenna. The antenna impedance is that of a half-wave dipole with a length-to-diameter ratio of 20-to-1.[12] It has a resistance of 72 ohms at quarter-wave resonance. In Figs. 9 and 10, the optimum generator impedance S has been set equal to this value. Thus, an impedance transformer will be required if the actual generator impedance is not 72 ohms.

The inverse compensated balun impedance plotted is that for $\gamma' = 0.25$ and $\beta' = 0.25$, the lowest amplitude curves of Fig. 8(a) and (b). Below center frequency, the resistive components track each other well, and the band

[9] L. Young, "Tables for cascaded homogeneous quarter-wave transformers," *IRE Trans. on Microwave Theory and Techniques*, vol. MTT-7, pp. 233–237, April 1959; also, *IRE Trans. on Microwave Theory and Techniques (Corrections)*, vol. MTT-8, pp. 243–244, March 1960.

[10] G. L. Matthaei, E. G. Cristal, and D. B. Weller, "Novel microwave filter design techniques," Stanford Research Institute, Menlo Park, Calif., Rept. 6, Contract DA-36-039-AMC-00084E, U. S. Army Electronics Labs., Fort Monmouth, N. J., July 1964.

[11] Coaxial lines or strip lines are limited to a maximum of about 200 ohms and at least one section of a short-step impedance transformer is at a higher impedance level than the terminal impedance.

[12] Radio Research Laboratory Staff, *op. cit.*, p. 7, Figs. 1–7.

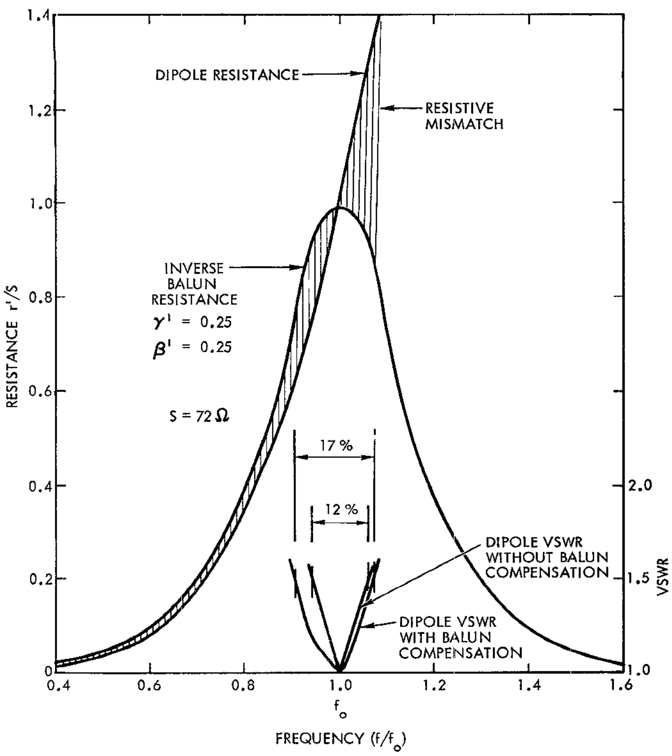

Fig. 9 Results of using the compensated balun as a complementary matching network on a dipole antenna (resistance).

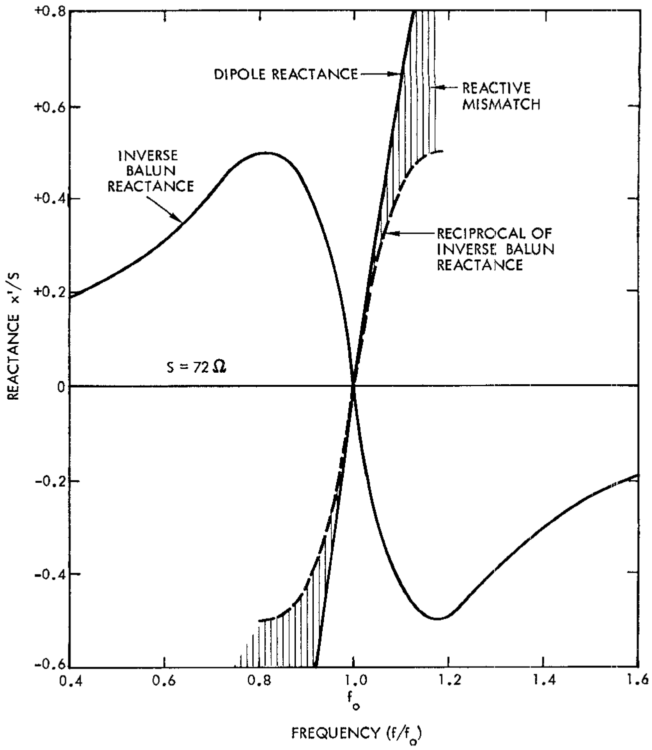

Fig. 10. Results of using the compensated balun as a complementary matching network on a dipole antenna (reactance).

over which a good match would be achieved would be quite large if the reactance also tracked. However, such is not the case. The reciprocal of the inverse balun reactance is drawn to more easily determine the magnitude of this mismatch. The shaded areas show the mismatch between the dipole and its effective source.

Included in Fig. 9 is an approximate curve showing the VSWR that will be seen by a 72-ohm generator. Two curves are shown—the mismatch without the balun improvement and the mismatch with the balun improvement. There is an increase in bandwidth of nearly 50 percent at a VSWR level of 1.5-to-1. The increase is 75 percent at a VSWR level of 1.2-to-1. Better results could have been achieved using a balun with γ' less than the value chosen here. A lower value would narrow balun bandwidth and, thus, create an off-center crossover for both resistance and reactance, and lower the mismatch.

A value of β' and γ' that will yield an off-center resistance crossover can be obtained from (7) and (8)

$$\gamma < \frac{1}{T}\left(\frac{\rho}{1-\rho}\right)^{1/2} \tag{7}$$

$$\beta = T\left[\gamma T \pm \left(\frac{\gamma^2 T^2}{\rho} - 1\right)^{1/2}\right] \tag{8}$$

where $T=\tan(\pi f_p/2f_0)$. Equation (8) is a solution of (2) using values of $r'/S=\rho$ and f_p/f_0 chosen from the resistance curve to be matched. Equation (7) is a restriction on the value of γ' derived from (8). It is the smaller value of β' obtained using the negative sign in (8) that creates the negative reactance slope required to match the dipole antenna. The larger value of β' creates curves with a positive reactance slope and also a humped resistance response near f_0.

Using (7) on the dipole impedance of Fig. 9, it is found that $\gamma' < 0.198$ for a crossover at $f_p/f_0 = 0.9$ and $r'/S = 0.61$. This confirms the observation from Fig. 9 that a closer complementary match could have been obtained using a smaller value of γ' and thus a faster falling resistance and reactance.

VI. Experiment

A. Test of Two Balun Arranged Back-to-Back

Figure 11 is a photograph of a pair of 4-to-1 bandwidth baluns arranged back-to-back so that the input and load are both in 50-ohm coaxial line. This arrangement was used because a well-matched load in two-wire line was not available. The pictured structure is completely symmetrical. On the left is a 50-ohm connector attached to a small diameter 50-ohm coaxial line Z_a. This line passes through the fingered support block and forms one conductor of the balun. The small diameter was chosen to make the two-wire characteristic impedance Z_{ab} as large as practical in accordance with the discussion in Sections II and III. The center conductor of the coax passes across the gap and into the other small diameter coax line Z_b, where it terminates in an open circuit.

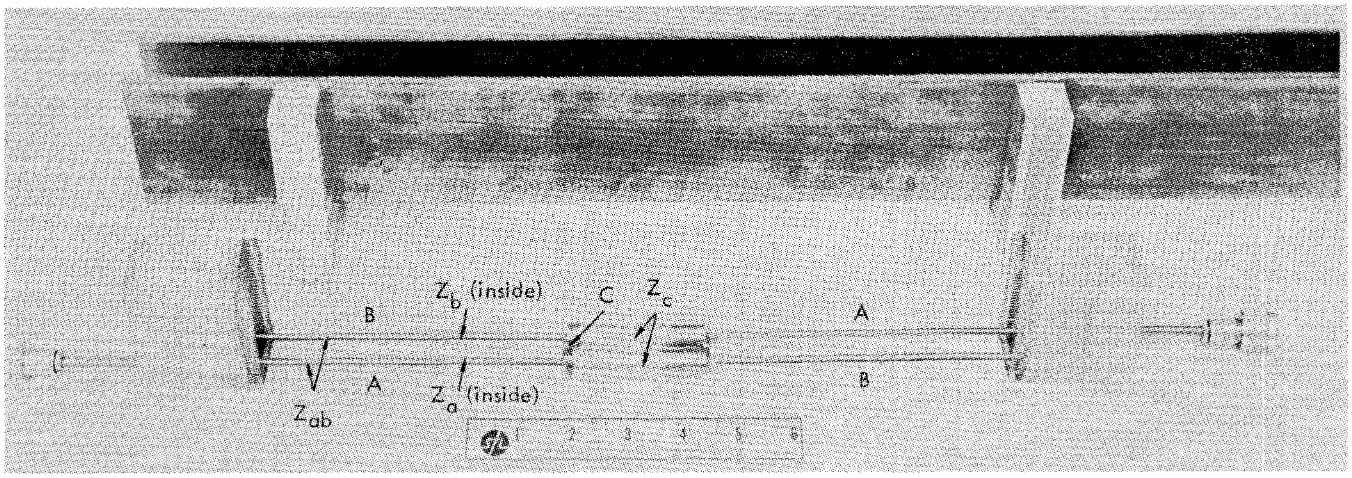

Fig. 11. Experimental arrangement of back-to-back compensated baluns.

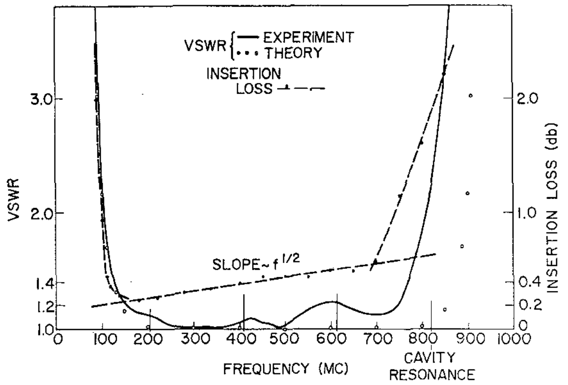

Fig. 12. VSWR and insertion loss of two back-to-back compensated baluns.

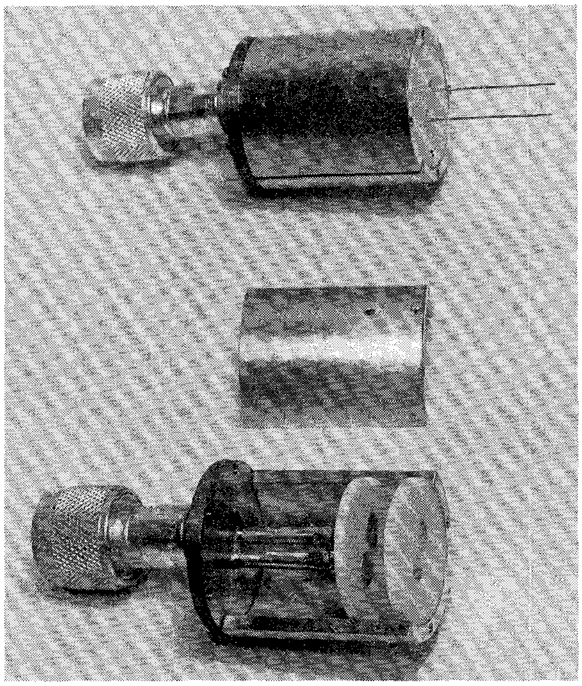

Fig. 13. Two S-band compensated baluns with a four-section short-step transformer on the balanced end.

The large diameter two-wire balanced output line Z_c is attached to the outer conductor of each coax line. Z_c, in turn, feeds the second balun which is terminated at its coax output with a well-matched 50-ohm load. Tightly wrapped aluminum foil was used to join the two baluns. The length of the line Z_c was made variable by use of additional lengths of tubing which screwed into place. Z_c could be made as long as two wavelengths. The entire structure was placed into the metal trough formed from S-band waveguide.

It was desired to operate the balun with the same balanced and unbalanced line impedances ($S=R=50\ \Omega$), so that the balun operation could be measured independent of any impedance transformers. This presented tolerance difficulties because a 50-ohm two-wire line requires large diameter, very closely spaced conductors. The close spacing requirement was somewhat alleviated by constructing the line symmetrically in the trough made from S-band waveguide. This increased the line capability and allowed an increase in line spacing.

The balun component line impedances were $Z_a=50$ ohm, $Z_b=12.5$ ohm, $Z_{ab}=200$ ohm, and $Z_c=50$ ohm. These values yield a minimum VSWR at center frequency which increases monotonically away from center frequency. These baluns were not designed for optimum equal ripple match (discussed in Section III) throughout the band because that technique had not been developed at the time of the test.

The trough introduced an unforeseen problem which shows in the data. The trough provides another conductor, which allows not only the desired balanced mode to propagate, but also an unbalanced mode; that is, a mode with the equal voltages on the two wires operating against the trough "ground." Thus, the structure with the two support blocks short-circuiting the lines forms an unbalanced mode "cavity." The "cavity" is lightly coupled to the input circuit through tolerance asymmetrics introduced in the fabrication.

This shows up in the data of Fig. 12 as an increase in attenuation and VSWR near the frequency at which the "cavity" is one wavelength long. Minor VSWR effects are also observed at $3/4\lambda$, $1/2\lambda$, and possibly $1/5\lambda$. The frequency of the VSWR peaks moves orderly as the length of the cavity is varied. It was observed that the peaks of the VSWR could be reduced by inserting resistance card segments between the two-wire conductors. This card was perpendicular to the electric field lines of the balanced mode and thus left that field unaffected. However, it was parallel to some lines of the unbalanced field and thereby attenuated those fields. This reduced the unbalanced mode Q and coupling, thus reducing the undesired perturbation. Also, it definitely proved the existence of the unbalanced mode.

In spite of troublesome unbalanced mode, at the low-frequency end the observed VSWR closely follows that predicted for two identical 4-to-1 bandwidth baluns operated back-to-back. It should be pointed out that the VSWR of one of the baluns taken separately would rise faster than that shown in Fig. 12, and the bandwidth would extend from 200 to 800 Mc/s as designed. The difference in Fig. 12 from the design is attributable to the back-to-back arrangement of two baluns. However, the data points closely match the theoretical back-to-back data at the low end and thus confirm the theory and the design in the low UHF band. These test baluns were fabricated from brass, yielding fairly high insertion loss, but closely following the expected $f^{1/2}$ frequency characteristic of ohmic losses as shown.

B. S-Band Balun with a Short-Step Transformer

Figure 13 is a photograph of a compensated balun with impedance transformation designed for a 1000-Mc/s band centered at 3500 Mc/s. The input impedance is 50 ohms and the output impedance is 380 ohms. This impedance transformation is accomplished by a short-step transformer[10] with four $\lambda/16$ sections having a total length of one-quarter wavelength. The design maximum passband attenuation was chosen as 0.01 dB.

Incorporating the transformer into the balanced output line leaves the balun looking into a load resistance $R=50$ ohms. The balun line impedance Z_{ab} was chosen to be 100 ohms, which yields a value of two for the parameter γ. This is sufficiently large for the intended bandwidth as it produces a maximum VSWR due to balun transformation of only 1.03-to-1. This value of impedance permits close spacing of the lines forming Z_{ab} and maintains a low value of uncompensated inductance across the gap.

Because of the large balanced line impedance, it is definitely preferable to provide the impedance transformation in the balanced line. If R were 380 ohms, a line spacing diameter ratio of approximately 90-to-1 would be required to yield a value of two for γ $(=Z_{ab}/R)$. At 3500 Mc/s such a ratio would require a large spacing with an intolerably large gap inductance.

An impedance transformer in balanced line presents two problems in design, however, which are not present in the coaxial counterpart. These problems are associated with the calculations of line diameters, spacings, and lengths. While accurate formulas and curves of balanced line spacing-diameter ratios are available for unshielded lines,[13,14] and shielded lines with a shield-spacing ratio of two,[13] the information is not sufficient to cover many design requirements. Such was the case here, and the available data had to be extrapolated.

The second problem which is encountered in any balanced line transformer is that of discontinuity capacities at the junctions of the transformer sections. To the author's knowledge, no tables or curves exist for determining discontinuity capacities for balanced lines. These capacities must be taken into consideration if the impedance steps are large. This is a larger problem with short-step transformers, since the impedance steps are unusually large compared to equivalent transformations using quarter-wave transformers. It is necessary to make extrapolations from the curves of coaxial discontinuity capacitances[15] to obtain the necessary information.

Both these problems were quite important in the design of the 3.5-Gc/s transformer. The wire diameters and discontinuity capacities were extrapolated as described above; the length changes were calculated to compensate for the capacities. The center high-impedance section was lengthened to allow use of a larger diameter wire. The calculated diameter for this section was 0.006 inches. The new diameter was equal to the output line diameter, 0.026 inches, which allowed use of a continuous length of wire for both parts. For support purposes and to decrease the diameters of the two low-impedance lines, these lines were dielectrically loaded with Teflon. This required a length change for those sections.

The assembled results, unhappily but not unexpectedly, were not optimum, and it was necessary to increase the shunt capacity at the high-impedance section nearest to the balun. The best results of the matching adjustments yielded a maxium VSWR of slightly over 1.3-to-1. The design value was 1.11-to-1.

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[13] H. A. Wheeler, "Transmission-line impedance curves," *Proc. IRE*, vol. 38, pp. 1400–1403, December 1950.

[14] International Telephone and Telegraph Corp., *Reference Data for Radio Engineers*, 4th ed. New York: American Book-Stratford Press, 1957, p. 588.

[15] J. R. Whinnery, H. W. Jamison, and T. E. Robbins, "Coaxial line discontinuities," *Proc. IRE*, vol. 32, pp. 695–709, November 1944.