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A New Impedance-Matched Wide-Band Balun and Magic Tee

GORDON J. LAUGHLIN

***Abstract*—A new wide-band microwave balun particularly attractive for microstrip circuitry is described in which the normally balanced line is in the form of a pair of equal-amplitude and antiphase unbalanced lines. This novel method of input–output coupling allows a coplanar arrangement of input and output microstrip lines.**

Often the balanced and unbalanced line impedances in a balun are unequal, necessitating an impedance-matching network. A first-order reflection coefficient theory that mutually considers the impedance effects of the balun cavity, a compensating stub, and a quarter-wave transformer is used to design wide-band impedance-matched baluns. Curves of VSWR versus bandwidth are presented for several balanced-to-unbalanced line-impedance ratios. Experimental results are given for an octave-band impedance-matched balun with a balanced-to-unbalanced impedance ratio of 2:1.

The new wide-band balun is adaptable to a microstrip magic tee. A proposed magic tee that relies on circuit symmetry for operation has multioctave bandwidth potential.

INTRODUCTION

MICROWAVE baluns are devices used for converting balanced transmission lines to unbalanced lines. A variety of descriptions for these devices have appeared in the literature [1]–[6]. In general, microwave baluns have inherent in their operation a cavity which appears as a resonant line shunting a balun junction. An example of such a balun is given in Fig. 1(a). The balun may be modified as shown in Fig. 1(b) to drive a pair of unbalanced lines in antiphase with equal amplitude. In this

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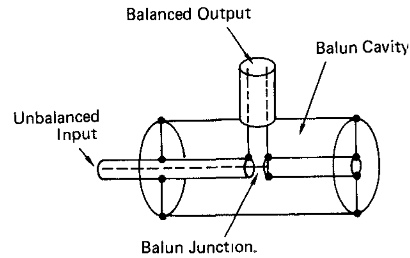
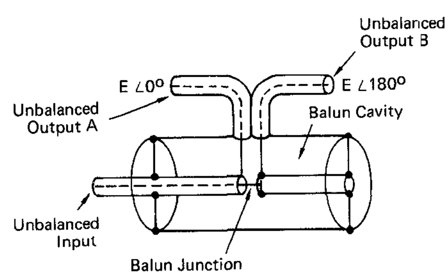


(a)



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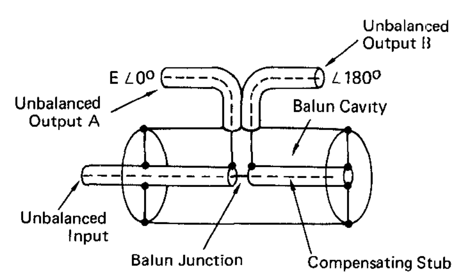

(c)

Fig. 1. Typical configuration of microwave balun. (a) Uncompensated balun. (b) Modified uncompensated balun; antiphase and equal amplitude outputs. (c) Modified compensated balun; antiphase and equal amplitude outputs.

case the normally balanced arm is split into two ports each of which is typically an unbalanced line. To achieve maximum bandwidth with these baluns the shunt impedance of the resonant line is often designed to have negligible effect, and Phelan [5] has presented curves of VSWR and bandwidth as a function of cavity impedance for baluns that are naturally impedance matched at band center. An open-circuited stub that partially cancels the cavity effects can be introduced [Fig. 1(c)], resulting in a compensated balun. Oltman [6] has given curves of real and reactive parts of the input impedance for compensated baluns.

Wide-band impedance matching of baluns that are not naturally matched at midband is complicated by the presence of the shunting balun cavity and, in compensated baluns, by the open-circuited stub. Fubini and Sutro [7] have shown that improved VSWR's can be achieved in an uncompensated balun (i.e., no open-circuit stub) by inserting a properly designed pair of quarter-wave transformers in the balance line. Their analysis considered the balun impedance to approximate the ideal load for a pair of transformer sections. Transformer pairs were designed by an iterative method to make the approximate ideal load and transformer pair correspond. Balun designs for balanced-to-unbalanced impedance ratios of 3:1 and 2.56:1 resulted in VSWR's below 1.25:1 for 4:1 bandwidth ratios in both examples. Oltman [6] has given an example of impedance matching a compensated balun with an impedance ratio of 7.6:1 to a predicted VSWR under 1.11:1 over a 1.33:1 bandwidth. A short-step transformer with four $\lambda/16$ sections was used, but an analysis of the interaction between transformer and balun was not given.

In this paper a new compensated balun and procedure for impedance matching are described. The particular configuration considered has the normally balanced line in the form of a pair of equal-amplitude and antiphase unbalanced lines. The form of the balun differs from earlier devices in the method by which input–output coupling is accomplished resulting in an easily fabricated balun applicable to microstrip circuitry and coaxial lines. In this balun the balanced-to-unbalanced impedance ratio is matched by mutually considering the effects of the cavity, open-circuited stub, and a quarter-wave transformer, using a first-order reflection coefficient theory to design an impedance-matched device. The impedance analysis is applicable to other compensated baluns.

BALUN DESCRIPTION

The balun investigated here is shown in a coaxial configuration in Fig. 2 [9]. Output coaxial lines C and D are formed from a single coaxial line with a gap in the outer shield at balun center. The center conductors of the output lines are common across the gap. Similarly, the input line, A, and compensating stub, B, are formed with common center conductors. The outer shields of A and C are electrically joined as are B and D. Both gaps are aligned.

A cavity, nominally a half-wavelength long, encloses the balun junction and is shorted to the outer shields of lines A, B, C, and D at the cavity ends. The cavity has the

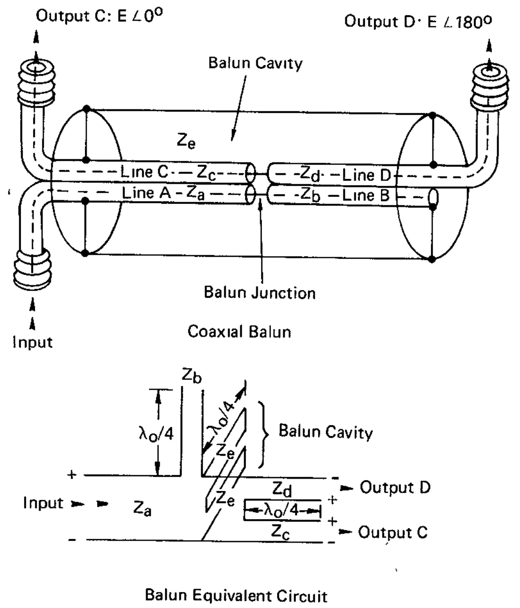

Fig. 2. Schematic and equivalent circuit of balun.

effect of shunting the junction with a pair of $\lambda/4$ series-connected short-circuit stubs. Each of the short-circuit stubs is a coaxial line where the outer conductor is the cavity wall and the shields of the enclosed coaxial lines form the center conductor. The reactance of the compensating line, an open-circuit stub in the figure, is connected directly to the input center conductor and appears in the equivalent circuit as a reactance in series at the input.

MICROSTRIP BALUN

A microstrip version of the aforementioned balun is depicted in Fig. 3 as a two-level microstrip circuit with three conductor planes. The uppermost conductor plane contains the input line (balun unbalanced line), the output lines (balun balanced line), and series-compensating stub. These several lines are separated by nominally one linewidth, except in the vicinity of the balun junction, to essentially eliminate line coupling. Normally the compensating stub will have an impedance different from the input line resulting in different linewidths. Since it is desirable that, in particular, both output lines have identical performance, the spacings between output lines and input line and compensating stub are made the same. Immediately at the balun junction the input line is brought close to the output lines so as to make the junction small and to avoid impedance transformation by line length between output and input lines. Each of the two output lines includes a quarter-wave transformer and the two branches sum in series at the balun junction. The quarter-wave transformer could be in the input arm, but it will become apparent in the section on an experimental balun that impedance values in the balun would be less attractive.

The middle conductor plane carries the ground plane for the lines in the uppermost plane. A single planar ground plane replaces the normal coaxial shields of input line A and output line C of Fig. 3. Similarly, a second ground

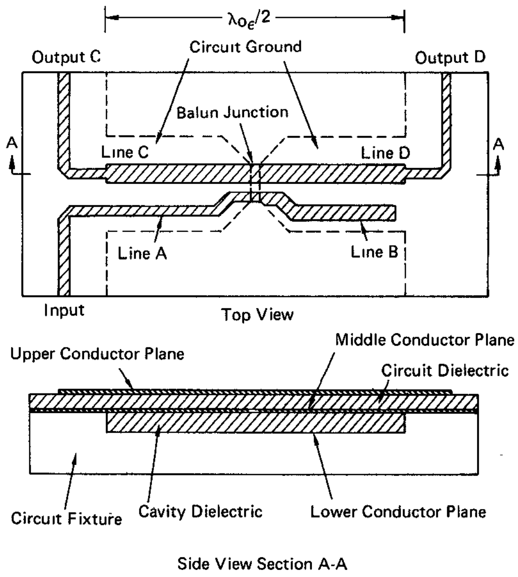

Fig. 3. Microstrip balun.

plane replaces the shields of compensating line B and output line D. Since all of the current in a microstrip-line ground plane is, for the most part, confined to three linewidths, the two planar ground planes extend one linewidth to either side of the upper-plane circuit lines without radiation loss. The two ground planes should be the same width to maintain symmetry and, due to upper-plane line edges not being collinear, the criterion for ground-plane width must be compromised somewhat. The balun junction is formed at the center of the circuit by a small gap between ground-plane ends. At the balun junction the ground planes are tapered to lessen end effects and reduce to a width nominally equal to the distance from outer edge to outer edge of the circuit lines at the junction.

The balun resonant cavity is formed by the region between the middle and lowest conductor planes. A channel for the cavity is cut in the circuit fixture, filled with dielectric, and the middle conductor plane placed on top. The resonant cavity is nominally a half-wavelength end-to-end at midband where the cavity ends are formed by shorting the ground planes in the middle conductor plane to the circuit fixture. The cavity depth is determined by the characteristic impedance of each cavity arm, having a value one-half the total cavity characteristic impedance. Line coupling from cavity to upper lines (by other than junction coupling) is minimal due to the three-linewidth criterion between upper circuit lines and the ground plane (if the radiation loss from the upper lines is negligible then the lines will not accept radiation from an outside source).

ANALYSIS

Fig. 4 is an equivalent circuit for a compensated balun that has a quarter-wavelength impedance transformer in the balanced line. Here the balanced line has an impedance twice that of each output microstrip line in the balun described above. The balun cavity, represented by the

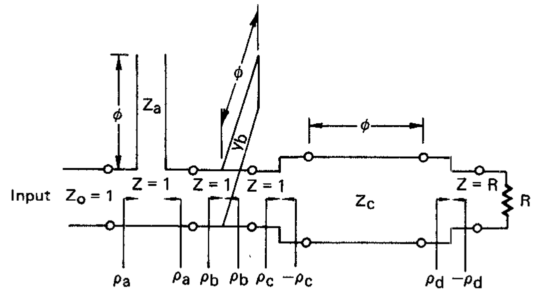

Fig. 4. Equivalent circuit of compensated balun with impedance transformer in balanced line. $\phi = (\pi/2)(f/f_0)$, $\rho_a = (-jZ_a \cot\phi)/(2 - jZ_a \cot\phi)$, $\rho_b = (jY_b \cot\phi)/(2 - jY_b \cot\phi)$, $\rho_c = (Z_c - 1)/(Z_c + 1)$, $\rho_d = (R - Z_c)/(R + Z_c)$.

short-circuited shunt stub, and the open-circuited series-compensating stub are resonant at band center. In this analysis small reflections at the discontinuities are assumed, and second-order reflection terms can be considered negligible compared to unity. The balun input reflection coefficient for small junction reflections can be written in terms of s parameters as

$$\rho \approx S_{11}^a + (S_{12}^a)^2 S_{11}^b + (S_{12}^a)^2 (S_{12}^b)^2 S_{11}^c + (S_{12}^a)^2 (S_{12}^b)^2 (S_{12}^c)^2 S_{11}^d e^{-j2\phi} \quad (1)$$

where

$$S_{11}^a = \rho_a \quad (S_{12}^a)^2 = -(1 - |\rho_a|^2)\frac{\rho_a}{\rho_a^*}$$

$$S_{11}^b = \rho_b \quad (S_{12}^b)^2 = -(1 - |\rho_b|^2)\frac{\rho_b}{\rho_b^*}$$

$$S_{11}^c = \rho_c \quad (S_{12}^c)^2 = 1 - \rho_c^2$$

$$S_{11}^d = \rho_d$$

and the reflection coefficients, the ρ_i, are given in the figure. It is of interest to note that ρ_a and ρ_b are complex quantities whereas ρ_c and ρ_d are real. For convenience let

$$\rho' = \rho \frac{\rho_a^*}{\rho_a}$$

where ρ_a^* is the conjugate of ρ_a and the magnitudes of ρ' and ρ are identical. Substituting into (1) and dropping appropriate second-order terms result in

$$\rho' \approx \rho_a^* - \rho_b + \frac{\rho_b}{\rho_b^*}(\rho_c + \rho_d e^{-j2\phi}). \quad (2)$$

Equation (2) can be reduced to a more convenient form by first observing that

$$\frac{\rho_b}{\rho_b^*} = -\frac{2 + jY_b \cot\phi}{2 - jY_b \cot\phi}$$

which can be rewritten as

$$\frac{\rho_b}{\rho_b^*} = -e^{j2\tan^{-1}}\left(\frac{Y_b}{2}\tan\left(\tfrac{1}{2}\pi - \phi\right)\right).$$

Using the approximations

$$\tan\theta \approx \theta \qquad \text{for small } \theta$$

and

$$e^{jx} \approx 1 + jx \quad \text{for small } x$$

it follows that

$$\frac{\rho_b}{\rho_b^*} \approx -\left(1 - \frac{Y_b}{2} - \frac{Y_b}{2} e^{-j2\phi}\right)$$

$$\text{for small values of } \tfrac{1}{2}\pi - \phi \quad \text{and} \quad \frac{Y_b}{2}\left(\tfrac{1}{2}\pi - \phi\right).$$

The reflection coefficients due to the balun-compensating stub, ρ_a, and cavity, ρ_b, are approximated near resonance as

$$\rho_a \approx \frac{Z_a}{4}\cos\phi(Z_a\cos\phi - j2\sin\phi) \qquad \frac{Z_a^2 - 4}{4}\cos^2\phi \ll 1$$

and

$$\rho_b \approx -\frac{Y_b}{4}\cos\phi(Y_b\cos\phi - j2\sin\phi) \qquad \frac{Y_b^2 - 4}{4}\cos^2\phi \ll 1.$$

The small reflections ρ_c and ρ_d at the quarter-wave transformer steps can be approximated by

$$\rho_c \approx \tfrac{1}{2}\ln Z_c$$

and

$$\rho_d \approx \tfrac{1}{2}(\ln R - \ln Z_c).$$

Substituting the foregoing several approximations in (2) and expanding the trigonometric functions in exponential form results in

$$\begin{aligned} \rho' \approx & \left[\frac{Z_a^2}{16} + \frac{Y_b^2}{16} + \frac{Z_a}{8} - \frac{Y_b}{8}\right] e^{j2\phi} \\ & + \left[\frac{Z_a^2}{8} + \frac{Y_b^2}{8} - \frac{1}{2}\left(1 - \frac{Y_b}{2}\right)\ln Z_c\right] \\ & + \left[\frac{Z_a^2}{16} + \frac{Y_b^2}{16} - \frac{Z_a}{8} + \frac{Y_b}{4}\ln Z_c \right. \\ & \left. - \frac{1}{2}\left(1 - \frac{Y_b}{2}\right)(\ln R - \ln Z_c)\right] e^{-j2\phi} \\ & + \left[\frac{Y_b}{4}(\ln R - \ln Z_c)\right] e^{-j4\phi}. \end{aligned} \tag{3}$$

Equation (3), with the exception of a phase term that is of no consequence, has the form of a three-step quarter-wavelength transformer written as

$$\Gamma_t = \Gamma_1 + \Gamma_2 e^{-j2\phi} + \Gamma_3 e^{-j4\phi} + \Gamma_4 e^{-j6\phi} \tag{4}$$

where the Γ_i are the reflection coefficients at the impedance steps. For small Γ_i

$$\Gamma_1 + \Gamma_2 + \Gamma_3 + \Gamma_4 \approx \tfrac{1}{2}\ln R_L$$

where R_L is the transformer load.

A design procedure for the Chebyshev-type quarter-wave transformer is to first let the junction discontinuities be symmetrical, i.e., $\Gamma_1 = \Gamma_4, \Gamma_2 = \Gamma_3$ [8]. Equation (4) can then be rewritten as

$$\Gamma_t e^{j3\phi} = 2\Gamma_1\cos 3\phi + 2\Gamma_2\cos\phi \tag{5a}$$

or

$$\Gamma_t e^{j3\phi} = 8\Gamma_1\cos^3\phi + (-6\Gamma_1 + 2\Gamma_2)\cos\phi. \tag{5b}$$

For a fractional bandwidth $\omega_q = (f_2 - f_1)/f_0$, let $x = \cos\phi/\cos\phi_1$ where ϕ_1 is the electrical length of each transformer step at f_1 and $|x| \le 1$ over the bandwidth ω_q. With proper assignment of impedances, the transformer reflection coefficient can be made to have a Chebyshev response given by

$$\Gamma_t e^{j3\phi} = \epsilon T_3(x) \tag{6}$$

where

$T_3(x) = 4x^3 - 3x$ the Chebyshev polynomial of degree 3

and

ϵ = maximum reflection for $|x| \le 1$.

Equating the coefficient in (6) and using the identity $\cos\phi_1 = \sin\frac{1}{4}\pi\omega_q$, the Γ_i for a Chebyshev-type transformer are given as

$$\Gamma_1 = \frac{\epsilon}{2\sin^3\frac{1}{4}\pi\omega_q} \tag{7a}$$

$$\Gamma_2 = \frac{3\epsilon}{2\sin^3\frac{1}{4}\pi\omega_q}\left(1 - \sin^2\tfrac{1}{4}\pi\omega_q\right). \tag{7b}$$

The maximum reflection coefficient is related to the fractional bandwidth and transformer load by

$$\epsilon = \frac{\ln R_L \sin^3\frac{1}{4}\pi\omega_q}{2(4 - 3\sin^2\frac{1}{4}\pi\omega_q)}. \tag{7c}$$

The variables in (3) can be solved in terms of the Γ_i resulting in

$$Z_a = (1 + \ln R)8\Gamma_1/\ln R - \ln R + 4\Gamma_1 - 4\Gamma_2 \tag{8a}$$

$$Z_b = 8\Gamma_1/\ln R \tag{8b}$$

$$Z_c = R^{1/2} \tag{8c}$$

where ϵ and ω_q are found to be related by

$$\begin{aligned} & \epsilon^2\left[\frac{32}{(\ln R)^2\sin^6\frac{1}{4}\pi\omega_q} + \frac{48}{\ln R\sin^4\frac{1}{4}\pi\omega_q} + \frac{36}{\sin^2\frac{1}{4}\pi\omega_q}\right] \\ & + \epsilon\left[-\frac{16}{\sin^3\frac{1}{4}\pi\omega_q} + \frac{12(1 - \ln R)}{\sin\frac{1}{4}\pi\omega_q}\right] \\ & + (\ln R)^2 - 2\ln R = 0. \end{aligned} \tag{8d}$$

Curves of maximum VSWR versus bandwidth ratio, f_2/f_1, for several values of balanced-line impedance are given in Fig. 5.

It should be noted that the quarter-wave transformer

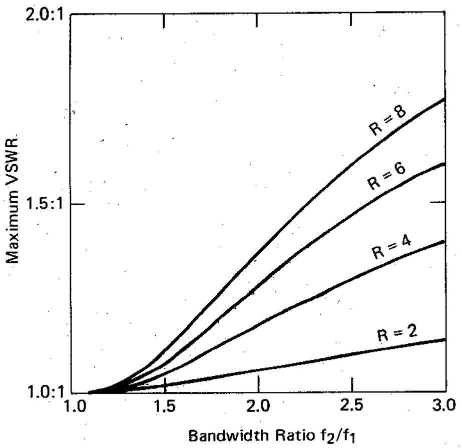

Fig. 5. Balun maximum VSWR versus bandwidth ratio for Chebyshev-type design (small reflection theory). R is the ratio of balanced-line impedance to input-line impedance.

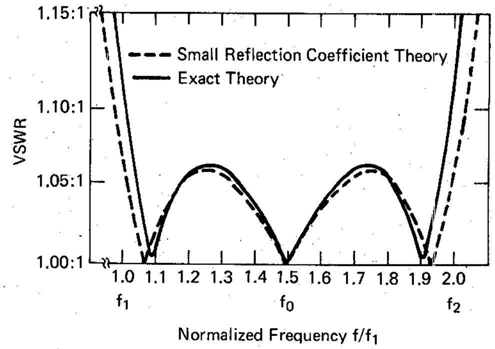

Fig. 6. VSWR versus frequency for Chebyshev-type design with $\omega_q = 0.67$ and $R = 2$.

could be placed in the input arm with equal ease.[1] The results for that configuration can be obtained directly from the preceding equations

$Z_a' = Y_b R$ compensating-stub impedance

$Y_b' = Z_a / R$ balun-cavity admittance

$Z_c' = R^{1/2}$ quarter-wave-transformer impedance

where R is the balanced-line impedance.

EXPERIMENTAL BALUN

An octave-band Chebyshev-matched balun was designed for a balanced-line impedance of 100 Ω and an unbalanced line of 50 Ω. Using the small-reflection-coefficient theory for $\omega_q = 0.67$, the expected maximum VSWR is 1.06:1 over the octave band when the balun has the impedances

$$Z_a = 46.2\ \Omega$$

$$Z_b = 39.0\ \Omega$$

$$Z_c = 70.7\ \Omega.$$

Curves of VSWR versus frequency from the small-reflection-coefficient theory and from exact theory are given for comparison in Fig. 6. Exact theory predicts a maximum VSWR of 1.11:1 at the band edges and 1.06:1 over the remaining portion. The small-reflection-coefficient theory is seen to be in reasonably small error for this design.

Fig. 7 shows a microstrip balun with this design at S band (2–4 GHz) in which the balanced line contains the quarter-wave transformer and is in the form of series-connected unbalanced lines. The input line, output line, and compensating stub are etched from one side of a doubly copper-clad 0.010-in-thick Teflon-fiberglass printed circuit board ($k = 2.55$). The input and output impedances are 50 Ω and the compensating stub is 46 Ω. Each of the output arms contains a 35-Ω quarter-wavelength

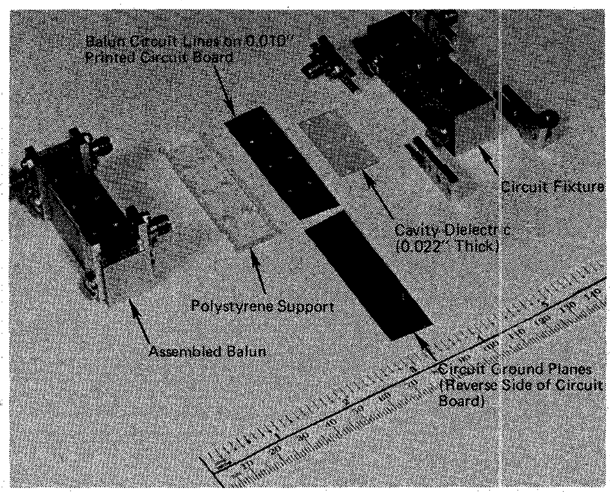

Fig. 7. Photograph of assembled balun and exploded view of balun circuitry.

transformer. Separation between adjacent lines is nominally one linewidth except at the balun junction where lines are brought to a 0.020-in separation. The circuit ground planes are etched from the reverse side of the printed circuit board and form a 0.020-in gap for the balun junction.[2] The ground plane is 0.200 in wide and is tapered by 45° angles to a 0.089-in width at the balun junction. From the foregoing design parameters each arm of the balun cavity, formed by cutting a channel in the circuit fixture, is 20 Ω (substantially lower than most conventional baluns) dictating a cavity thickness of 0.022 in when filled with polystyrene ($k = 2.56$). The cavity length, end to end, is nominally 1.230 in. The circuit fixture width is 0.600 in in keeping with being three linewidths, but could probably be made narrower since the field of these low-impedance lines is tightly confined under the line. The entire circuit assembly is held down and kept flat by a 0.125-in polystyrene block secured with nylon screws.[3]

Measured performance of the microstrip balun is given in Fig. 8. The maximum VSWR of 1.23:1 over the octave-band and the midband VSWR of 1.22:1 are higher than predicted (1.11:1 and 1.0:1, respectively), probably due

[1] In this configuration Fig. 4 is modified by placing the transformer, Z_c', between the input line and the open-circuit-compensating stub, Z_a'. A balun with matching in the input arm is the example given in [6].

[2] The balun gap is a compromise between keeping the gap less than 0.01 λ at the high frequency (0.018 in) and easily attained gap width (0.02 in).

[3] Circuit linewidths were selected assuming a medium of $k = 2.55$ both above and below the circuit lines.

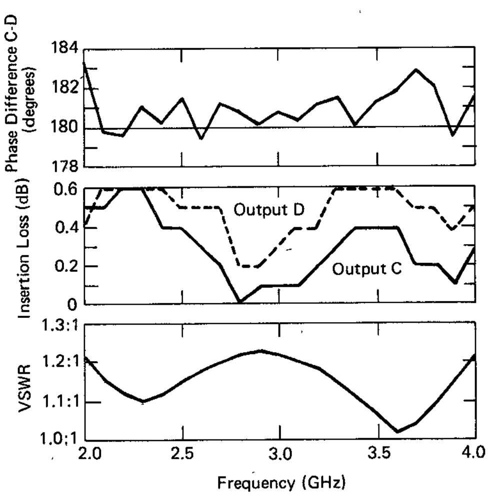

Fig. 8. Performance of Chebyshev-type design balun.

to effects at the balun junction. The balun cavity was shortened slightly in an unsuccessful effort to cancel a suspected capacitance at the junction (data now shown). This leads to the belief that the high midband VSWR is caused by impedance transformation through line length between input and output at the junction as well as by a capacitive effect. The unbalance between output ports is 0.3 dB maximum in amplitude, and 3.6° in phase, and the maximum insertion loss is 0.6 dB.

It was mentioned in an earlier section that if the quarter-wave transformer were in the unbalanced line the impedance values would be less attractive than previously indicated. For the octave-band Chebyshev-matched balun discussed the impedance would be

$Z_a' = 128.2$ compensating stub

$Z_b' = 108.2$ cavity impedance

$Z_c' = 70.7$ quarter-wave transformer.

The compensating impedance, Z_a', is nearing the practical limit of microstrip; the higher cavity impedance requires a thicker cavity; and the transformer impedance, being in a single line rather than split between the two output lines, is higher and consequently more lossy. Clearly the form of the experimental balun is more effective.

MICROSTRIP MAGIC TEE

It has occurred to the author that some microwave baluns have potential for adaptation to wide-band magic tees (180° hybrids). The microstrip balun which has been described can be configured into a magic tee as shown in Fig. 9 where the balun cavity arms are folded so as to be parallel, and a fourth conductor is added to drive the pair of output lines. The new input is in the form of quasi coplanar-microstrip line in that the ground plane is split into two halves while the circuit line is in a plane above the ground plane. A secondary ground plane, the balun cavity ground, lies in a level below the primary ground plane and will affect the new line impedance since some coupling through the primary ground plane slot is inevitable.

The operation of the magic tee relies on symmetry of the device as does the described balun. When ports C and D are driven in phase with equal amplitudes no field is excited across the junction gap and consequently power is not coupled into the difference port (the original balun input port). However, lines C and D are connected in parallel to the sum port, and under proper impedance matching, power flow is out the sum port. Driving lines C and D in antiphase with equal amplitudes excites a field across the junction gap thus coupling power into the difference port. The asymmetrical excitation in line E cannot couple into the sum port. Equivalent circuits for the sum mode, difference mode, and the total circuit are given in Fig. 10. The cavity impedance, Z_c, in this configuration has the same value as in a balun design, but due to the proximity of the two cavity arms the odd-mode impedance must be considered.

In designing a magic tee of this form, the objective is to

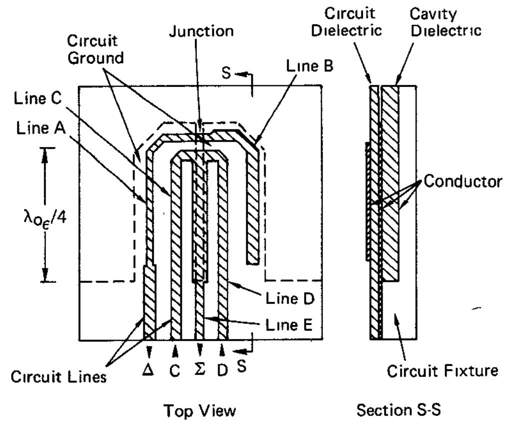

Fig. 9. Microstrip magic tee.

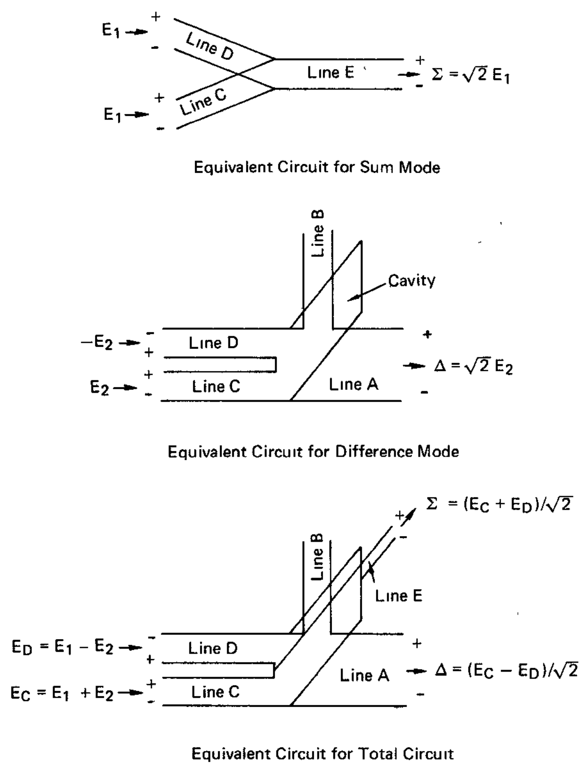

Fig. 10. Microstrip magic tee equivalent circuits.

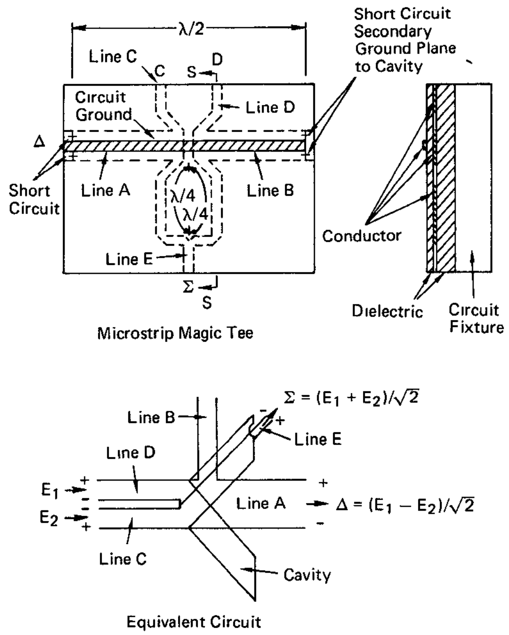

Fig. 11. Magic tee adapted from Bawer and Wolfe [4] and equivalent circuit.

design for minimum mismatch at the sum and difference ports (if these ports are matched and total isolation between them exists, then it immediately follows that for a lossless device the output ports are matched with total isolation). Since hybrid designs similar to the aforementioned experimental balun would transform 50-Ω output levels to series-connected 25-Ω loads at the balun junction, the sum port would see 12.5 Ω requiring a 4:1 impedance transformation. To achieve maximum bandwidth it would probably be more advantageous to include quarter-wave transformers in the input lines rather than the output, even though the impedance values would be less attractive. The result is a required 2:1 impedance transformation in the sum line.

Another form of a wide-band magic tee may be realized by modifying the balun developed by Bawer and Wolfe [4]. Their balun differs from the balun described in this paper in that the output is taken directly off the circuit ground planes at the balun junction and is in the form of a balanced-line output, and the resonant cavity is in the form of a short-circuited balanced line which is not in the form of microstrip. If a secondary ground plane is included in their balun, the cavity and balanced-line output can be made in the form of pairs of microstrip lines. Additional microstrip lines can be added as shown in Fig. 11 resulting in a magic tee. The added lines, which comprise the sum port, are a pair of lines connected to the circuit ground planes at the balun junction and are shorted together a quarter-wavelength from the junction. As in the previously discussed magic tee, the sum and difference ports are isolated. Impedance matching of this magic tee is somewhat different from that of the other because the pair of new lines appears in the equivalent circuit as additional susceptance across the junction when the difference port (balun input) is driven. Similarly, when the sum port is driven, the balun cavity appears as a susceptance shunting the load. Naturally these effects must be considered to impedance match the magic tee.

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