

# BROADBAND TRANSFORMERS AND POWER COMBINING TECHNIQUES FOR RF

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## INTRODUCTION

The following discussion focuses on broadband transformers for RF power applications with practical examples of various types given with performance data. Detailed design formula are available in the Reference section. Power combining techniques useful in designing high power amplifiers are discussed in detail.

## BROADBAND TRANSFORMERS

The input and output transformers are among the most critical components in the design of a multi-octave amplifier. The total performance of the amplifier (linearity, efficiency, VSWR, gain flatness) will depend on their quality. Transformers with high impedance ratios and for low impedances are more difficult to design in general. In the transmission line transformers very low line impedances are required, which makes them impractical for higher than 16:1 impedance ratios in a 50-Ohm system. Other type transformers require tight coupling coefficients between the primary and secondary, or excessive leakage inductances will reduce the effective bandwidth. Twisted line transformers (Figure 1C, D, F, G) are described in References 1, 2, and

4. Experiments have shown that the dielectric losses in certain types of magnet wire, employed for the twisted lines, can limit the power handling capability of such transformers. This appears as heat generated within the transformer at higher frequencies, although part of this may be caused by the losses in the magnetic core employed to improve the low frequency response. At low frequencies, magnetic coupling between the primary and secondary is predominant. At higher frequencies the leakage inductance increases and the permeability of the magnetic material decreases, limiting the bandwidth unless tight capacitive coupling is provided. In a transmission line transformer this coupling can be clearly defined in the form of a line impedance.

The required minimum inductance on the low impedance side is:

$$L = \frac{4 R}{2\pi f}$$

where

- L = Inductance in  $\mu\text{H}$
- R = Impedance in Ohms
- f = Frequency in MHz

This applies to all transformers described here.

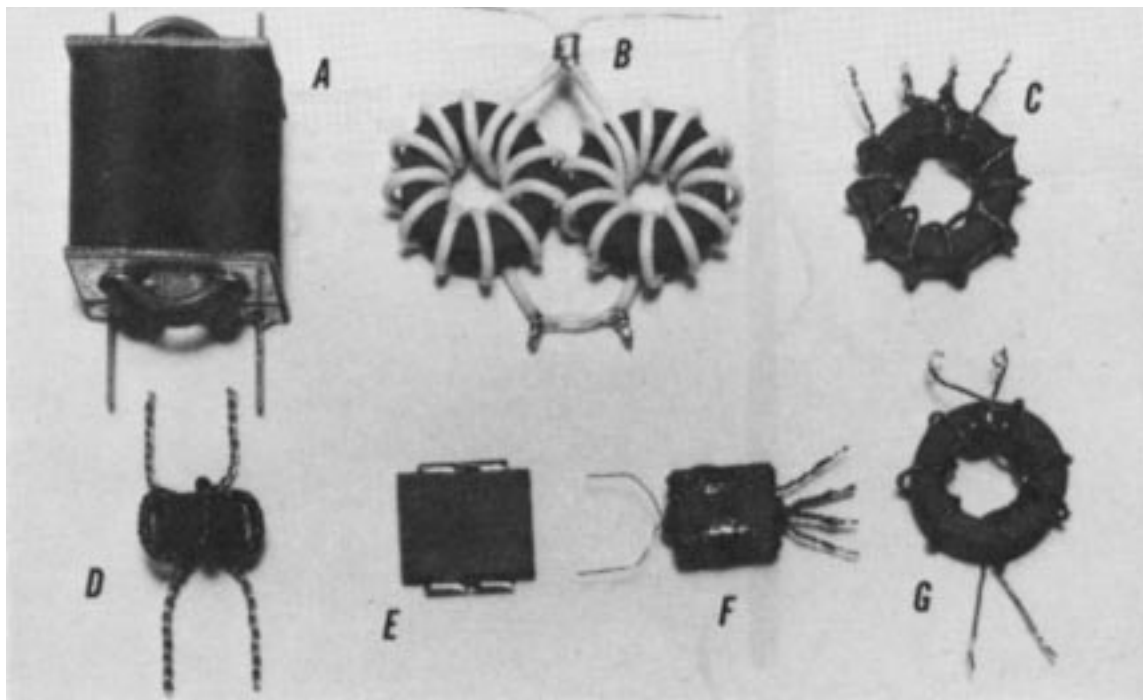


Figure 1. HF Broadband Transformers

Some transformers, which exhibit good broad band performance and are easy to duplicate are shown in Figure 1.

Transformers E and F are intended for input applications, although A in a smaller physical form is also suitable. In E, the windings are photo etched on double sided copper-Kapton\* (or copper-fiberglass) laminate. The dielectric thickness is 3 mils, and the winding area is 0.25 in<sup>2</sup>. Ferrite plates ( $\mu_r = 2000$  to 3000) are cemented on each side to improve the low frequency response. This type transformer in the size shown, can handle power levels to 10 W. Figure 2 shows curves for laminate thickness versus winding area for various impedance ratios.

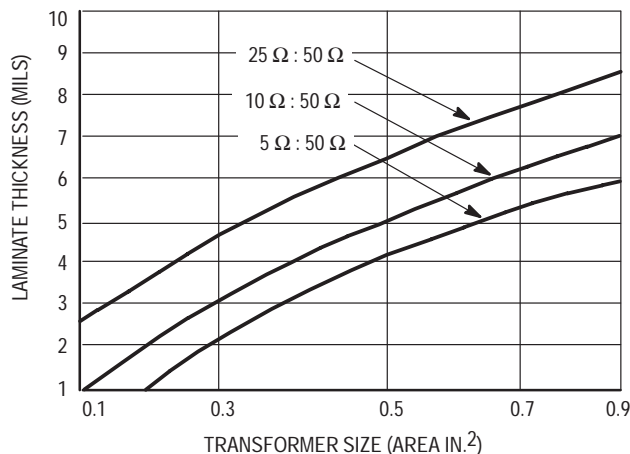
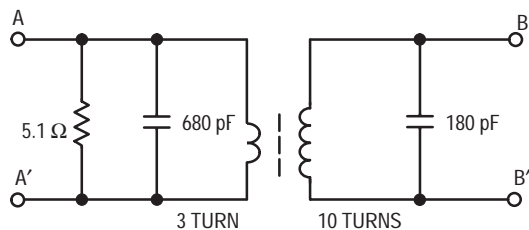


Figure 2. Laminate Thickness versus Winding Area

Impedance ratios of this transformer are not limited to integers as 1:1, 4:1 — N:L, and the dc isolated primary and secondary have an advantage in certain circuit configurations. This design will find its applications in high volume production or where the small physical size is of main concern. Table 1 shows the winding configuration and measured data of the transformer shown in Figure 3.

Table 1. Impedance at Terminals BB' Transformer Terminated as Shown



f (MHz)	R <sub>p</sub> (Ohms)	X <sub>p</sub> (Ohms)
1.0	50.7	+j 81
2.0	53.0	+j 185
4.0	53.1	+j 1518
8.0	53.5	-j 214
16.0	50.5	-j 79
32.0	52.9	-j 30

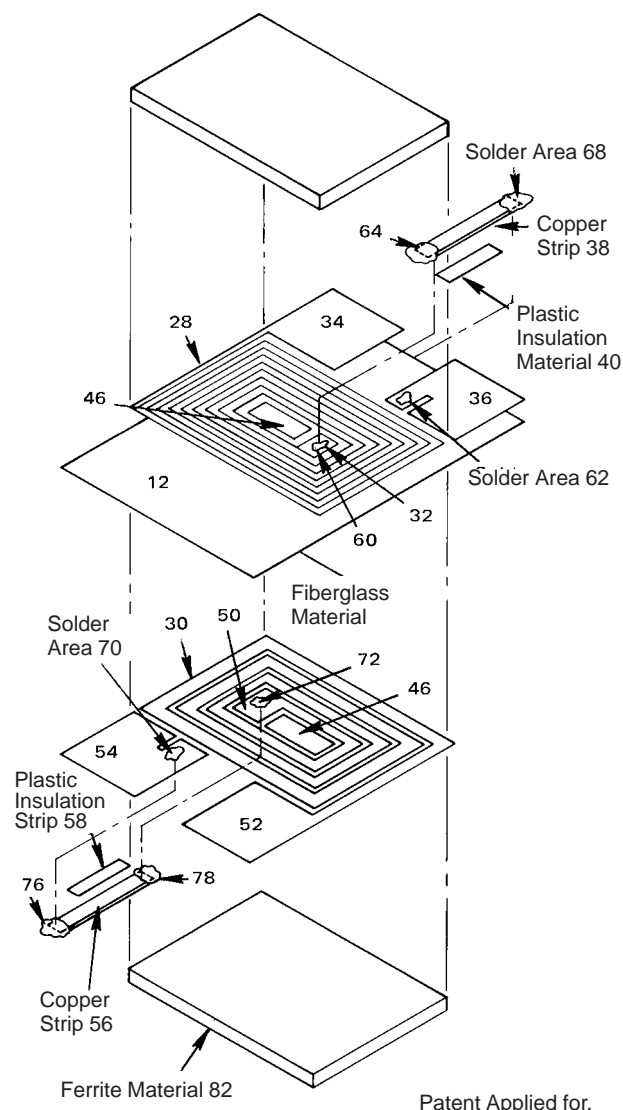


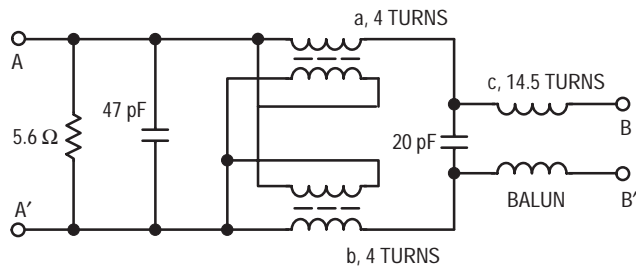
Figure 3. Detailed Structure of Transformer Shown in Figure 1E

In the transformer shown in Figure 1F and Table 2, a regular antenna balun core is employed (Indiana General F684-1 or equivalent). Lines A and B each consist of two twisted pairs of AWG #30 enameled wire. The line impedances are measured as 32 Ohms, which is sufficiently close to the optimum 25 Ohms calculated for 4:1 impedance ratio. ( $Z_0 = \sqrt{R_{in} R_L}$ ).

Windings a and b are wound one on top of the other, around the center section of the balun core. Line c should have an optimum  $Z_0$  of 50 Ohms. It consists of one pair of AWG #32 twisted enameled wire with the  $Z_0$  measured as 62 Ohms. The balun core has two magnetically isolated toroids on which c is wound, divided equally between each. The inductance of c should approach the combined inductance of Lines a and b (Reference 4, 6).

\*Trademark of E. I. DuPont, De Nemours and Co., Inc.

**Table 2. Impedance at Terminals BB' Transformer Terminated as Shown**



f (MHz)	R <sub>p</sub> (Ohms)	X <sub>p</sub> (Ohms)
1.0	53.0	+ j 185
2.0	52.6	+ j 330
4.0	52.9	+ j 430
8.0	53.1	+ j 600
16.0	53.2	+ j 750
32.0	53.5	+ j 3060

The reactance in the 50 Ohm port (BB') should measure a minimum of + j 200. To achieve this for a 4:1 transformer, a and b should each have three turns, and for a 9:1 transformer, four turns. When the windings are connected as a 9:1 configuration, the optimum Z<sub>0</sub> is 16.6 Ohms, and a larger amount of high frequency compensation will be necessary. Lower impedance lines can be realized with heavier wires or by twisting more than two pairs together. (e.g., four pairs of AWG #36 enameled wire would result in the Z<sub>0</sub> of approximately 18 Ohms.) Detailed information on the manufacture of twisted wire transmission lines can be found in References 2, 4, and 8.

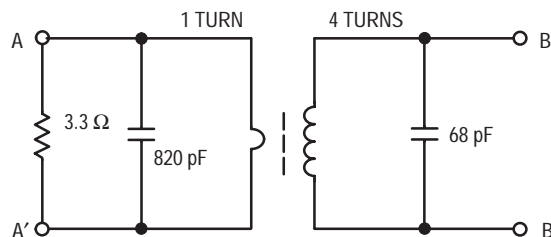
Figure 1A shows one of the most practical designs for higher impedance ratios (16 and up). The low impedance winding always consists of one turn, which limits the available ratios to integers 1, 4, 9 — N. Data taken of this type of a 16:1 transformer is shown in Table 3, while Figure 4 illustrates the physical construction. Two tubes, 1.4" long and 1/4" in diameter — copper or brass — form the primary winding. The tubes are electrically shorted on one end by a piece of copper-clad laminate with holes for the tubes and the tube ends are soldered to the copper foil. The hole spacing should be larger than the outside diameter of the ferrite sleeves.

A similar piece of laminate is soldered to the opposite ends of the tubes, and the copper foil is divided into two sections, thus isolating the ends where the primary connections are made. The secondary winding is formed by threading wire with good RF insulating properties through the tubes for the required number of turns.

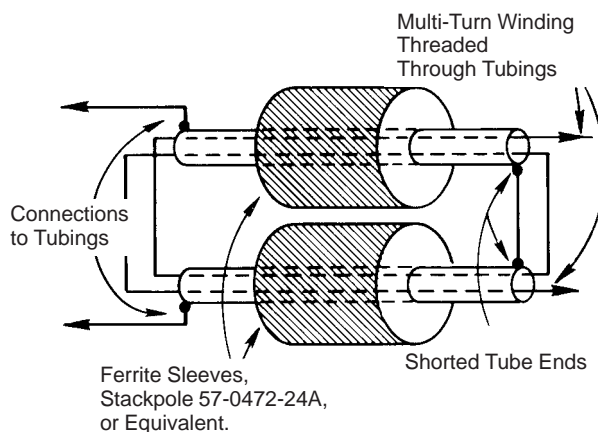
Although the measurements indicate negligible differences in performance for various wire sizes and types (stranded or solid), the largest possible diameter should be chosen for lower resistive losses. The initial permeability of the ferrite sleeves is determined by the minimum inductance required for the lowest frequency of operation according to the previous formula. Typical μ<sub>r</sub>'s can vary from 800 to 3000 depending upon the cross sectional area and lowest operating frequency. Instead of the ferrite sleeves, a number

of toroids which may be more readily available, can be stacked.

**Table 3. Impedance at Terminals BB' Transformer Terminated as Shown**



f (MHz)	R <sub>p</sub> (Ohms)	X <sub>p</sub> (Ohms)
1.0	54.0	+ j 1030
2.0	54.0	+ j 3090
4.0	54.0	+ j 5800
8.0	53.9	- j 300
16.0	53.1	- j 760
32.0	53.2	- j 600



**Figure 4. Physical Construction of a 16:1 Transformer (Actual Number of Turns Not Shown)**

The coupling coefficient between the primary and secondary is almost a logarithmic function of the tube diameter and length. This factor becomes more important with very high impedance ratios such as 36:1 and up, where higher coupling coefficients are required. The losses in the ferrite are determined by the frequency, permeability and flux density. The approximate power handling capability can be calculated as in Reference 4 and 6, but the ferrite loss factor should be taken into consideration. The μ<sub>r</sub> in all magnetic materials is inversely proportional to the frequency, although very few manufacturers give this data.

Two other variations of this transformer are shown in Figure 5. The smaller version is suitable for input matching, and can handle power levels to 20 W. It employs a stackpole dual balun ferrite core 57-1845-24B. The low impedance winding is made of 1/8" copper braid. The portions of braid going through the ferrite are rounded, and openings are made in the ends with a pointed tool. The high impedance winding is threaded through the rounded portions of the braid, which was uncovered in each end of the ferrite core. (See Figures 4 and 5.)

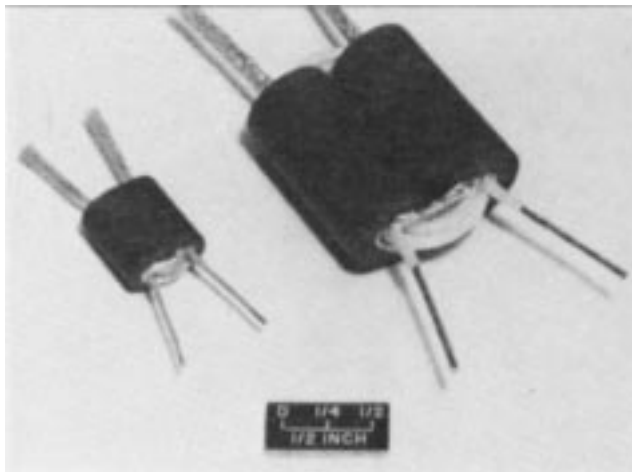
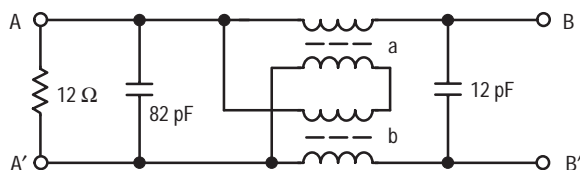


Figure 5. Variations of Transformers in Figure 1A

The construction technique of the larger version transformer is similar, except two separate ferrite sleeves are employed. They can be cemented together for easier handling. This transformer is intended for output applications, with a power handling capability of 200–250 W employing Stackpole 57-0472-27A ferrites. For more detail, see Reference 7.

The transformer shown in Figure 1B is superior in bandwidth and power handling capability. Table 4 shows data taken on a 4:1 transformer of this type. The transmission lines (a and b) are made of 25-Ohm miniature co-axial cable, Microdot 260-4118-000 or equivalent. Two 50 Ohm cables can also be connected in parallel.

Table 4. Impedance at Terminals BB' Transformer Terminated as Shown



f (MHz)	R <sub>p</sub> (Ohms)	X <sub>p</sub> (Ohms)
1.0	48.3	+ j 460
2.0	48.1	+ j 680
4.0	48.0	+ j 920
8.0	48.0	+ j 1300
16.0	48.1	+ j 900
32.0	48.1	+ j 690

The balun, normally required to provide the balanced to unbalanced function is not necessary when the two transmission lines are wound on separate magnetic cores, and the physical length of the lines is sufficient to provide the necessary isolation between AA' and BB'. The minimum line length required at 2.0 MHz employing Indiana General F627-19-Q1 or equivalent ferrite toroids is 4.2 inches, and the maximum permissible length at 30 MHz would be approximately 20 inches, according to formulas 9 and 10 presented in Reference 2. The 4.2 inches would amount to four turns on the toroid, and measures 1.0 μH. This complies with the results obtained with the formula given earlier for minimum inductance calculations.

Increasing the minimum required line length by a factor of 4 will provide the isolation, and the total length is still within the calculated limits. The power loss in this PTFE insulated co-axial cable is 0.03 dB/ft at 30 MHz in contrast to 0.12 dB/ft for a twisted wire line. The total line loss in the transformer will be about 0.1 dB.

The number of turns on the toroids has been increased beyond the point where the flux density of the magnetic core is the power limiting factor. The combined line and core losses limit the power handling capability to approximately 300 W, which can be slightly increased by employing lower loss magnetic material.

Note the connection arrangement (Figure 6), where the braid of the cable forms the high current path of the primary.

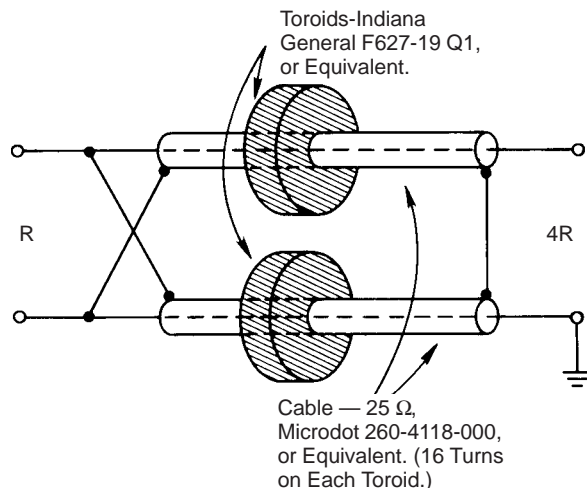


Figure 6. Transformer Construction (Figure 1B)

### HIGH-FREQUENCY POWER COMBINING TECHNIQUES EMPLOYING HYBRID COUPLERS

The zero degree hybrids described here are intended for adding the powers of a multiple of solid-state amplifiers, or to combine the outputs of groups of amplifiers, usually referred to as modules. With this technique, powers to the kW level at the high-frequency bands can be realized.

When reversed, the hybrids can be used for splitting signals into two or more equal phase and amplitude ports. In addition, they provide the necessary isolation between the sources. The purpose of the isolation is to keep the system operative, even at a reduced power level during a possible failure in one amplifier or module. The isolation is especially important in output combining of linear amplifiers, where a constant load impedance must be maintained. Sometimes the inputs can be simply paralleled, and a partial system failure would not have catastrophic effects, but will merely result in increased input VSWR.

For very high frequencies and narrow bandwidths, the hybrid couplers may consist of only lengths of transmission line, such as co-axial cable. The physical lengths of the lines should be negligible compared to the highest operating frequency to minimize the resistive losses, and to avoid possible resonances. To increase the bandwidth and improve the isolation characteristics of the line, it is necessary to increase the impedance for non-transmission line currents (parallel currents) without effecting its physical length. This can be done by loading the line with magnetic material. Ideally, this material should have a linear BH curve, high

permeability and low losses over a wide frequency range. For high-frequency applications, some ferrites offer satisfactory characteristics, making bandwidths of four or more octaves possible.

Depending upon the balance and phase differences between the sources, the currents should be mostly cancelled in the balun lines. In a balanced condition, very little power is dissipated in the ferrite cores, and most occurring losses will be resistive. Thus, a straight piece of transmission line loaded with a high permeability ferrite sleeve, will give better results than a multiturn toroid arrangement with its inherent higher distributed winding capacitance.

It is customary to design the individual amplifiers for 50 Ohm input and output impedances for testing purposes and standardization. 50- and 25-Ohm co-axial cable can then be employed for the transmission lines. Twisted wire lines should not be used at power levels higher than 100 Watts average, due to their higher dielectric losses.

Variations of the basic hybrid are shown in Figure 7A and B where both are suitable for power dividing or combining.

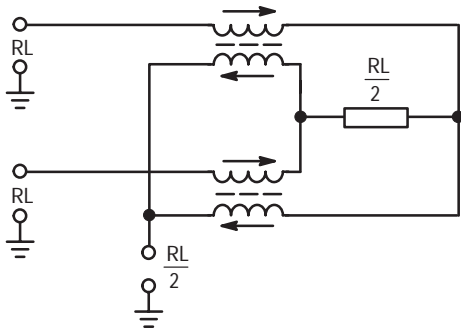


Figure 7A

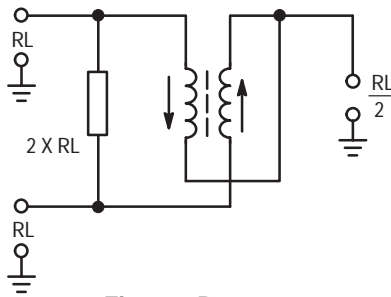


Figure 7B

Figure 7. Variations of Basic Hybrid

The balancing resistors are necessary to maintain a low VSWR in case one of the 50-Ohm points reaches a high

impedance as a result of a transistor failure. As an input power splitter, neither 50-Ohm port will ever be subjected to a short due to the base compensation networks, should a base-emitter junction short occur. An open junction will result in half of the input power being dissipated by the balancing resistor, the other half still being delivered to the amplifier in operation. The operation is reversed when the hybrid is used as an output combiner. A transistor failure will practically always cause an increase in the amplifier output impedance. Compared to the 50-Ohm load impedance it can be regarded as an open circuit. When only one amplifier is operative, half of its output power will be dissipated by R, the other half being delivered to the load. The remaining active source will still see the correct load impedance, which is a basic requirement in combining linear amplifiers. The resistors (R) should be of noninductive type, and rated for 25% of the total power, unless some type of automatic shutoff system is incorporated. The degree of isolation obtainable depends upon the frequency, and the overall design of the hybrid. Typical figures for 2 to 30 MHz operation are 30–40 dB. Figures 8A and B show 4 port “totem pole” structures derived from Figures 7A and 7B. Both can be used with even number of sources only, e.g. 4, 8, 16, etc. For type 8B, it is more practical to employ toroidal multi-turn lines, rather than the straight line alternatives, discussed earlier. The power output with various numbers of inoperative sources can be calculated as follows, if the phase differences are negligible: (Reference 2)

$$P_{\text{out}} = \left(\frac{P}{N}\right) N_1$$

where:

P = Total power of operative sources

N = Total number of sources

$N_1$  = Number of operative sources

Assuming the most common situation where one out of four amplifiers will fail, 75% of the total power of the remaining active sources will be delivered to the load.

Another type of multiport hybrid derived from Figure 7A is shown in Figure 9. It has the advantage of being capable of interfacing with an odd number of sources or loads.

In fact, this hybrid can be designed for any number of ports. The optimum values of the balancing resistors will vary according to this and also with the number of ports assumed to be disabled at one time. Two other power combining arrangements are shown in Figures 10 and 11. The isolation characteristics of the four-port output combiner were measured, the data being shown in Table 5. The ferrite sleeves are Stackpole 57-0572-27A, and the transmission lines are made of RG-142/U co-axial cable. The input power dividers described here, employ Stackpole 57-1511-24B ferrites, and the co-axial cable is Microdot 250-4012-0000.

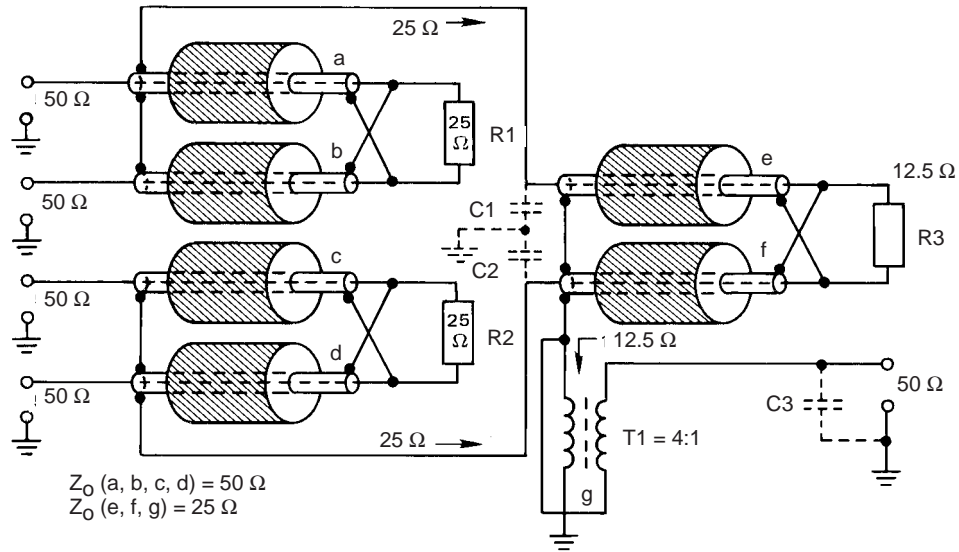


Figure 8A

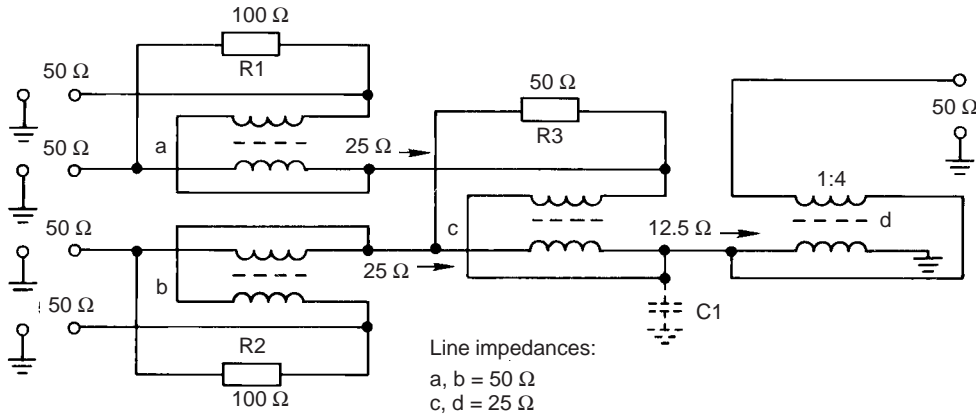
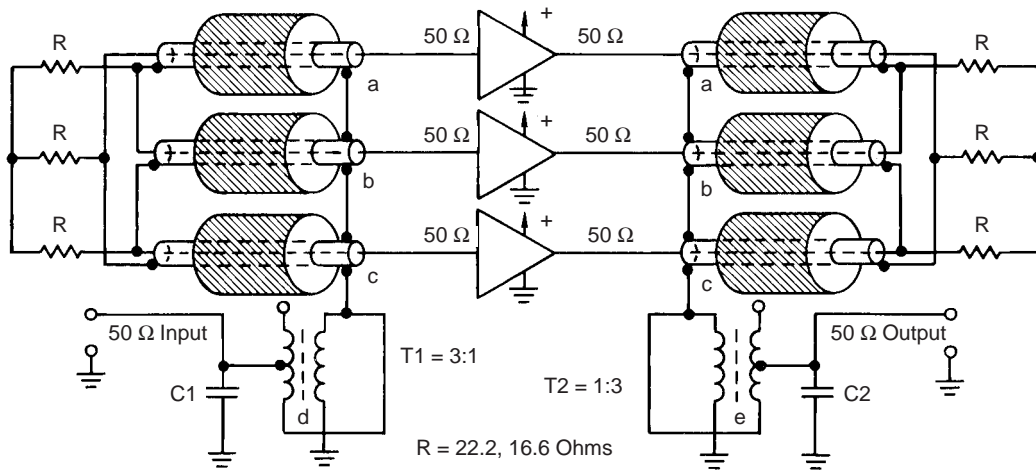


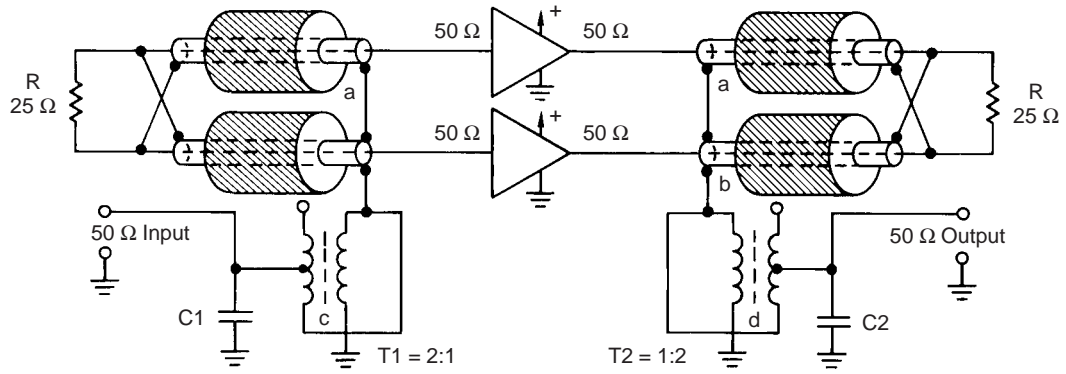
Figure 8B

Figure 8. Four Port "Totem Pole" Structure



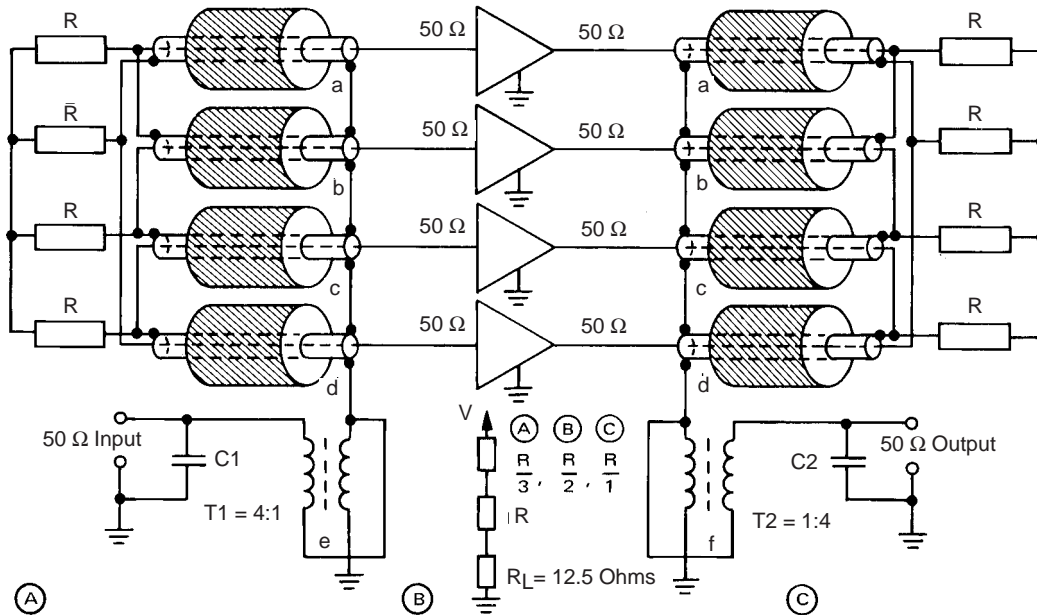
$Z_0(a, b, c) = 50 \Omega$   
 $Z_0(d, e) = 25 \Omega$  (optimum 28.9Ω)

Figure 9. Three-Port Hybrid Arrangement



$Z_0(a, b) = 50 \Omega$   
 $Z_0(c, d) = 25 \Omega$  (optimum  $35.4 \Omega$ )

Figure 10. Two-Port Hybrid System



- (A) 1 PORT INOPERATIVE.  
 Optimum  $R = 28.3 \text{ Ohms}$   
 $P_{out} = (P1+P2+P3) - (P_R + P \frac{R}{3})$
- (B) 2 PORTS INOPERATIVE.  
 Optimum  $R = 25 \text{ Ohms}$   
 $P_{out} = (P1+P2) - (P_R + P \frac{R}{2})$
- (C) 3 PORTS INOPERATIVE.  
 Optimum  $R = 18.75 \text{ Ohms}$   
 $P_{out} = P1 - (P_R + P \frac{R}{1})$

$P1, P2, P3 =$  Power at any operative port,  $P_R =$  Power dissipated in  $R$ , excluding  $R_L$ .  
 $V =$  RMS voltage at any 50 Ohm point.  
 (The phase differences are assumed negligible.)

$Z_0(a,b,c,d) = 50 \Omega$   
 $Z_0(e,f) = 25 \Omega$  (optimum)

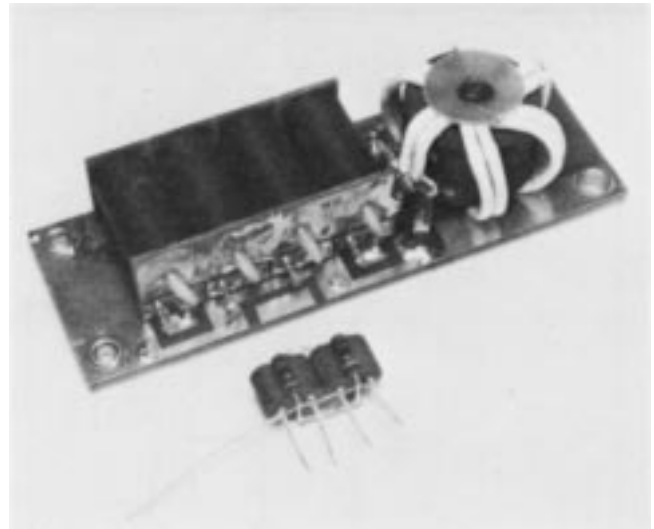
Figure 11. Four-Port Hybrid System

**Table 5. Isolation Characteristics of Four Port Output Combiner**

f (MHz)	Isolation, Port-to-Port (dB)
2.0	27.0 – 29.4
4.0	34.8 – 38.2
7.5	39.0 – 41.2
15	32.1 – 33.5
20	31.2 – 33.0
30	31.0 – 33.4

The input and output matching transformers (T1 – T2) will be somewhat difficult to implement for such impedance ratios as 2:1 and 3:1. One solution is a multi-turn toroid wound with co-axial cable, such as Microdot 260-4118-000. A tap can be made to the braid at any point, but since this is 25-Ohm cable, the  $Z_0$  is optimum for a 4:1 impedance ratio only. Lower impedance ratios will normally require increased values for the leakage inductance compensation capacitances (C1 – C2). For power levels above 500–600 W, larger diameter co-axial cable is desirable, and it may be necessary to parallel two higher impedance cables. The required cross sectional area of the toroid can be calculated according to the  $B_{max}$  formulas presented in References 4 and 6.

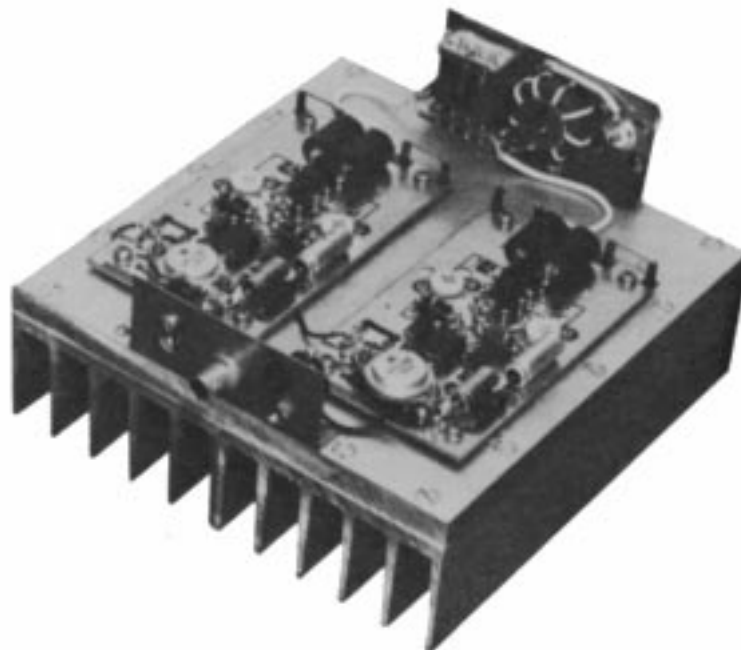
The 2 to 30 MHz linear amplifier (shown in Figure 13) consists of two 300 W modules (8). This combined amplifier can deliver 600 W peak envelope power. The CW power output is limited to approximately 400 W by the heatsink and the output transformer design.



**Figure 12. Two-Four Port Hybrids**

The one at the lower left is intended for power divider applications with levels to 20–30 W. The larger one was designed for amplifier output power combining, and can handle levels to 1 – 1.5 kW. The balancing resistors are not shown with this unit.)

The power combiner (Figure 13A) and the 2:1 step-up transformer (Figure 13B) can be seen in the upper right corner. The input splitter is located behind the bracket (Figure 13C). The electrical configuration of the hybrids is shown in Figures 7A and 10. Note the loops equalizing the lengths of the co-axial cables in the input and output to assure a minimum phase difference between the two modules.




**Figure 13. 2 to 30 MHz Linear Amplifier Layout**



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