BROADBAND TRANSFORMER DESIGN FOR RF TRANSISTOR POWER AMPLIFIERS

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Summary

Various forms of broadband transmission line transformers for use through VHF, and concepts basic to their operation are summarized and described. Broadband power adding and power dividing hybrids for push-pull amplifiers and parallel additive uses are given similar attention. The design of broadband input and output matching elements for a single ended stage using a 2N5071 transistor for 30 watts of output power from 30 MHz to 80 MHz is illustrated. Useful curves for guiding the design of 4:1 transformers with resistive loads are given. Curves showing typical input impedance versus frequency of a pair of cascaded 4:1 transformers loaded by the large signal R-L-C equivalent input circuit of the 2N5071 are included. The use of hybrids to operation is described.

Introduction

In the past few years RF power transistors have attained performance levels which have made them attractive for implementing many solid state transmitters intended for military and commercial communications. Many of these applications are for selective frequency operation over one or more frequency bands, a band being typically between 0.4 and 0.8 of an octave in bandwidth. With modern trends stressing minimized tuning operations, the use of untuned broadband interstage and output matching networks has become popularly sought. The most suitable broadband impedance matching networks in the realm of lumped components, applicable from LF into UHF, have been broadband transmission line transformers and baluns of the type originally proposed by Guanella[1] and more recently expanded upon by Ruthroff[2]. Short lengths of transmission line (parallel strip, coaxial or twisted pair) provide upper frequency capability free from normal transformer interwinding resonance limitations. High permeability materials, such as ferrites extend the lower bandwidth capabilities. Though the concept of these broadband transformers is not new, to date little analysis of their two port parameters as a function of frequency, or transmission line length, and characteristic line impedance has been available.

Survey of Some Popular Broadband Transmission Line Components

The 4:1 Impedance Transformer

The 4:1 transmission line transformer is a popular element for impedance matching a single ended source to a single ended load. A close analogue among conventional transformers is the autotransformer with a 2:1 turns ratio as depicted in Fig. 1a, showing a load R_L transformed to $\frac{R_L}{4}$ at the input.

The transmission line counterpart is shown schematically in Fig. 1b. The transmission line provided excellent coupling of the magnetic and electric fields. This ensures a minimum of leakage reactance over the useful frequency range of the line. With conventional windings, interwinding capacitance limits the performance at high frequencies. Between transmission line conductors, the uniformly distributed capacitance acts harmoniously with series distributed inductance to give reactance free operation over a broad frequency range.

Upper Frequency Cutoff Factors. The transmission line transformer has upper frequency limitations related to the length of the transmission line in wavelengths and to the ratio between the characteristic impedance Z_0 of the line and the load impedance. In most cases, the length of transmission line is chosen to be $\lambda/8$ or less at the uppermost frequency desired. The optimum match is $Z_0 = 1/2\ R_L$ for step down of a purely resistive load and $Z_0 = 2\ R_L$ when the transformer is being used to step up the load from R_L to $4\ R_L$. As expected with transmission line matching, the shorter the physical length of the line in wavelengths the less important exact matching becomes. The interconnection of the transmission line is important in keeping with the two general matching principles for a flat line: 1) The voltage between conductors must be maintained constant as a wave propagates from the input until it reaches the load. This is illustrated in Fig. 1b by the voltage E, between points 1 and 3 at the driven end and also at the load end between points 2 and 4. Differences in the signal path length to corresponding points along the line cause phase differences increasing with frequency. Such phase differences in the 4:1 connection, (not typical of normal, matched transmission line) determine the final bandwidth capability of the transformer. 2) For a current flowing in one conductor (i.e. 1/2 flowing from 1 to 2) an equal current in the opposite direction should exist in the other

conductor (i.e. I/2 flowing from 4 to 3).

Low Frequency Cutoff Factors. Transmission line transformers have a low frequency cutoff determined by the falloff of primary reactance as frequency is decreased. This reactance is determined by the series inductance of the transmission line conductors. Therefore, the longer the length of line, the greater the series inductance, extending performance to lower frequency. This is in direct conflict with extending the upper frequency performance which, as previously stated, is enhanced by shortening the physical length of the transmission line. Minimizing the overall length consistent with achieving the suitable reactance for low frequency requirements can be accomplished by using high permeability materials such as ferrites. Inductance of a conductor is directly proportional to the relative permeability of the surrounding medium. A high permeability material placed m close to the transmission line conductors acts on the external fringe field present and can magnify the inductance appreciably and thereby provide a lower cutoff frequency. There is no influence upon internal magnetic fields nor upon the characteristic impedance of the line. The power transferred from input to output is not coupled through the ferrite material, but rather through the dielectric medium separating the transmission line conductors. This is an important concept for design. Relatively small cross section ferrite material can perform unsaturated at impressively high power levels. In contrast to this, conventional transformers couple power from primary to secondary entirely through a high permeability core which must be chosen to suitably carry the total power without saturating. Fig. 1c illustrates a popular construction for a 4:1 transformer using twisted pair wound on a ferrite toroid. A basic advantage of this construction is the ability to minimize the length of the interconnection from point 1 to point 4. Fig. 1d is a photograph of several 4:1 transformers made in this manner. The larger ones in the upper row have windings from 3.5 to 4.5 inches long. The transformer at the upper left uses a loosely twisted pair of number 20 formvar coated transformer wire having a $Z_o \simeq 50\Omega$. The middle unit is twisted of two bifilar windings of number 26 formvar wire for a $Z_o = 20\Omega$. The unit at the upper right is constructed of homemade strip transmission line. Flat formar transformer wire about 0.10 inches wide paired together with contact bond cement gives a $Z_o \simeq 10\Omega$. These transformers have each carried 100 watts of RF power without saturating the ferrite cores. With properly matched resistive loads they have less than 0.1 dB insertion loss at an upper frequency of 100 MHz and a lower frequency of 2.0 MHz. The lower two transformers are constructed with smaller toroids and have 2.5 inch lengths of transmission line wound about them. The transformer at the lower right uses RG-196 (50Ω coaxial line) with the outer insulation removed. This improves its flexibility and permits winding the

the shield closer to the ferrite which makes more effective gains from the high permeability material. In the lower left of Fig. 1d is a small unit wound with the strip line of $Z_o \simeq 10\Omega$ described earlier. The small toroid demonstrates the relative flexibility of this ribbon type line. The 4:1 transformer can be used singly or in cascade for stepping up or stepping down a load impedance.

Higher Ratio Impedance Transformers

A brief mention of some other possible extensions of the basic 4:1 transformer maybe informative. Fig. 2a shows a 9:1 impedance transformer in schematic form. Two equal lengths of transmission line are used. Construction forms usually follow the style of the 4:1 type described above. Each length of transmission line ordinarily uses a separate ferrite core. The cores may be stacked to keep interconnections as short as possible. The transformer operation should be clear with the voltages and currents as indicated. For optimally matched conditions and negligible phase shift:

$$R_{IN} = \frac{E}{I} \tag{1}$$

$$R_L = \frac{E/3}{3I} = \frac{E}{9I} = \frac{R_{IN}}{9} \tag{2}$$

$$Z_o = \frac{E/3}{I} = \frac{R_{IN}}{3} = 3 \cdot R_L \tag{3}$$

Figure 2b shows the next logical extension of the 9:1. One additional length of transmission line is employed to achieve a 16:1 impedance transformer. The general winding approach may be observed from Fig. 2a and 2b and extended to higher ratios by induction.

The available ratios will progress 1:1, 4:1, 9:1, 16:1, 25:1 ... n:1 (n always an integer). Also, the optimum Z_o choice will always be given by:

$$Z_o = \sqrt{R_L R_{IN}} \tag{4}$$

Broadband Power Adding and Power Dividing Networks

It is frequently necessary to use two or more transistors to attain some output power beyond that of any available single device. Broadband power adding and dividing can be nicely accomplished with hybrids constructed simply as transmission line structures. Hybrids permit summing the power from two or more sources with insignificant interaction between the individual sources. The hybrid provides excellent isolation between the input ports when separate sources of power are to be added (or alternately isolation between two output ports acting as separate sources when power splitting). The input impedance (or output impedance in the alternate case) at one port remains essentially invariant regardless of

conditions of an open circuit or a short circuit at the second port. Obviously, such characteristics will effectively eliminate chain type failures common in multiple transistor output stages with transistors directly paralleled at their outputs.

The Zero Degree Hybrid. A hybrid can be constructed as a zero degree phase type for adding inputs which are approximately in phase. Similarly, it will split a single input into two identical in-phase sources.

Ruthroff described several zero degree phase hybrids. Fig. 3 illustrates schematically in transmission line form one such hybrid together with voltage, current and impedance relationships. The component R_T is a physical resistor which dissipates non-addable input power. This waste power is related to phase and/or amplitude differences between the signals at ports 1 and 2. For the extreme case where one port is open circuited or short circuited, one half of the total input power from the other port is dissipated in R_T and half is delivered to the load R_L.

The 180 Degree Phase Hybrid. The push-pull operation of two power transistors is often desirable in output stages. Push-pull usually reduces the second harmonic output power by about 8 dB over single ended stages, thereby easing the demands for second harmonic rejection in the low pass output filter.

The 180 degree phase hybrid affords all the virtues of isolation between adjacent ports, described previously, plus 180 degree phase splitting (or an adding for two signals 180 degrees out of phase).

Fig. 4a illustrates the interconnections and operating relationships for a 180 degree phase hybrid being used for power adding. Fig. 4b shows the voltage and current relations when one port is either open or shorted. As can be noted, under this condition, one half of the output at the working port is dissipated in R_T. In general for this hybrid:

$$P_{out} = \frac{P_1 + P_2}{2} - \sqrt{P_1 P_2} \cos \theta \qquad (5)$$

where:

P_{out} = power into the load R_L
P_1 = power into port 1
P_2 = power into port 2
θ = phase angle between signals at ports 1 and 2.

When used for power splitting, this hybrid will contribute equal out of phase power to each transistor load. If one load is open circuited, short circuited or in general dissimilar from the alternate load, the input drive to the alternate load will remain unchanged. This is controlled by a change in impedance at the input port.

With $Z_o = R_L$, the length of the series output line section is unimportant, since it performs as a simple length of transmission line terminated in its characteristic impedance. With very low values of load impedance it is impracti-

cal to exactly match Z_o to R_L. The length of this section must then be maintained quite short relative to $\lambda/8$ if reactive effects due to mismatch are to be avoided.

The Application of Broadband Transmission Line Structures to the Design of Untuned Broadband RF Transistor Amplifiers

Requirements for broadband untuned VHF transistor amplifiers often demand operation over two or more bands representing over an octave of frequency range. For single-ended stages, the sole purpose of the broadband elements is for impedance matching. The 4:1 impedance transformer is a popular broadband component for this. For power additive stages, signal phasing and isolation between transistors must also be considered. The design of broadband components for both a single-ended output stage and a push-pull output stage will be demonstrated.

Extensive characterization of the 4:1 transformer will be described as part of this design.

Single-Ended RF Power Stage Design

Matching for power transistor stages consists of 1) Converting some fixed real antenna load to the appropriate load line for the final power stage such that some specified power output can be realized with good efficiency at all operating frequencies, and 2) Interstage matching, which requires transforming the reactive low input impedance of a transistor into a suitable load line for the driving stage at all operating frequencies.

To demonstrate the design of broadband elements to accomplish matching, a typical power output stage will be investigated. The transistor is a 2N5071 operating from a 28 volt d.c. supply. Operation is specified to be 30 watts of RMS output power from 30 MHz to 80 MHz. The antenna load is 50Ω.

Output Matching

The optimum load line for producing 30 watts of output power from this transistor operating from 28 volts d.c. is 13Ω to 15Ω. One 4:1 transformer will step down the 50Ω antenna load to 12.5Ω. Equations for terminal impedance of the 4:1 transformer can be derived from two Kirchoff loop equations for the circuit of Fig. 1b solved simultaneously with the two general defining equations for voltage and current at one end of a lossless transmission line. At the low impedance port ($R_{IN} = R_L/4$) i.e. the load is stepped down:

$$Z_{IN} = Z_o \left[\frac{Z_L \cos\beta\ell + j Z_o \sin\beta\ell}{2Z_o (1 + \cos\beta\ell) + j Z_L \sin\beta\ell} \right] \qquad (6)$$

At the high impedance port ($R_{IN} = 4R_L$) i.e. the load is stepped up:

$$Z_{IN} = Z_o \left[\frac{2Z_L (1 + \cos\beta\ell) + j Z_o \sin\beta\ell}{Z_o \cos\beta\ell + j Z_L \sin\beta\ell} \right] \qquad (7)$$

Where Z_L is the load impedance, β the phase constant of the transmission line in radians per unit length, and ℓ is the length of the line section. For the broad number of applications where the load impedance is purely resistive, normalized design curves can describe Z_{IN} as a function of frequency and length of line (in terms of $\beta\ell$) and as a function of Z_o.

Figs. 5, 6, and 7 display R_{IN}, X_{IN}, and phase respectively, at the low impedance port (i.e. the transformer used for load step down) as a function of length and Z_o of the transmission line. The R_{IN} and X_{IN} values can be unnormalized by multiplying by $R_L/4$. The various Z_o values are simply related to R_L as shown. The curves apply only to high frequency cutoff effects and assume nonexistence of low frequency effects. With Z_o equal to Z_o optimum, R_{IN} drops off gradually, from a value of $R_L/4$, with increasing length of line (i.e. the equivalent of increasing frequency for a fixed length of line). An inductive X_{IN} becomes apparent beyond 0.10λ. With Z_o values greater than Z_o optimum the inductive character of the input impedance is accented. The behavior closely approaches a fixed R-L series circuit when Z_o is $2Z_o$ optimum or greater. With Z_o values less than Z_o optimum, the input impedance behaves as a parallel R-C circuit.

Fig. 8 shows curves of insertion loss as a function of line length for various values of Z_o. Insertion loss is defined as the ratio of power into the load to power available from a source E_g, with internal resistance R_g, matched to the ideal R_{IN} value of the transformer. It should be clearly understood that the transformer is as internally lossless as the transmission line comprising it. As typical of lossless filters, power transfer is determined by the variation of terminal impedance with frequency. There is insignificant accuracy loss in assuming that real power into the transformer P_{IN} equals real power into the load P_{out}. An expression for insertion loss with any generalized impedance ($Z_{IN} = R_{IN} + j X_{IN}$) represented at a particular value of is:

$$\text{Insertion Loss} = \frac{P_{out}}{P_{available}} =$$

$$\frac{4 R_g R_{IN}}{(R_g + R_{IN})^2 + X_{IN}^2} \qquad (8)$$

Where $R_g = R_L/4$ the value of R_{IN} at $\ell = 0$ for load step down or

$R_g = 4R_L$ the value of R_{IN} at $\ell = 0$ for load step up.

Note that in Fig. 8 pairs of Z_o values, related by $K Z_o$ optimum and Z_o optimum/K, where K can be any given constant, yield identical insertion loss.

For each of the curves described above the abscissa is given is wavelengths λ, which allows relating units of length and frequency for any design case. Knowing the highest operating

frequency f_u in MHz and next judging from the chart the largest fractional part of a wavelength n, which affords R_{IN} and X_{IN} within acceptable limits, the physical length of line is given by:

$$\ell \approx \frac{7200n}{f_u} \text{ inches} \qquad (9)$$

This expression holds for lines constructed with conventional dielectrics of relative dielectric constants ranging from 2.0 to 2.5.

The required minimum length for the lower frequency cutoff requirements should next be examined. A conservative empirical relation which holds quite closely is:

$$\ell \geq \frac{20 R_L}{\left(1 + \frac{\mu}{\mu_o}\right) f_L} \text{ inches} \qquad (10)$$

Where R_L is the load resistance in ohms, f_L the lowest operating frequency in MHz, and μ/μ_o the relative permeability of the ferrite at f_L. This expression relates a length corresponding to the input impedance having a 10 to 15 degree inductive phase angle at f_L when the transformer is terminated in a real load R_L. A typical ferrite for use at 30 MHz (e.g. Q-2 Ferramic) will have $\mu/\mu_o = 15$. With $R_L = 50\Omega$, and $f_L = 30$ MHz, equation (9) gives $\ell = 2.08$ inches. Reference to the curves for the low impedance port (Fig. 7) with $Z_o = Z_o$ optimum = 25Ω at $.08\lambda$; R_{IN} is about 6% down from the ideal 12.5Ω value and X_{IN} is essentially zero.

Using equation (8) with $f_u = 80$ MHz and $h = 0.08$ it follows that $\ell = 7.5$ inches. Therefore any length between 2.15 inches and 7.2 inches wound on a ferrite toroid, will offer excellent performance. Arbitrarily let the choice be a length of 4 inches. It is apparent that for this frequency range of 30 MHz to 80 MHz the design of the transformer is quire flexible. If the frequency range were expanded (e.g. 2 MHz to 80 MHz) the length determined from equation (8) would converge upon that of equation (9).

The curves are very useful in other ways. For example, assume it to be more convenient to use transmission line of $Z_o = 50\Omega$ than the optimum value of 25Ω. At $n = 0.08$ and observing the $Z_o = 2 Z_o$ optimum curves, R_{IN} is almost ideally equal to $R_L/4$ or 12.5Ω, X_{IN} is about 5Ω inductive. The effect of this value of reactance may be difficult to place in perspective when attempting to estimate performance. The insertion loss curves of Fig. 8 will be useful for this. Note that at 0.08λ on the $Z_o = 2 Z_o$ optimum curve the insertion loss is less than 0.2 dB. This given more specific meaning to the effect of the 5Ω inductive reactance. Fig. 9 illustrates the output matching circuit in block diagram.

Measurements. Measurements to verify calculated results were made over a frequency range of 1 MHz through 108 MHz using an HP 4815A Vector Impedance Meter. A 9-inch length of RG-58 cable ($Z_o = 50\Omega$ and $\ell = 0.25\lambda$ at 100 MHz) with outer

insulation removed was interconnected as a 4:1 impedance step down transformer. Ferrite toroids, CF 102 size of type Q-1 material (μ/μ_o = 100 up to 10 MHz), covered the entire transmission line section. With the fixed Z_o = 50Ω line terminated in R_L of 50Ω, 100Ω and 200Ω, the 0.5 Z_o optimum, Z_o optimum and 2.0 Z_o optimum conditions were simulated. Measured and calculated insertion loss with frequency are compared in Fig. 9. Differences between measured and calculated insertion loss with frequency are compared in Fig. 9. Differences between measured and calculated was less than 0.2 dB at 100 MHz for the 0.5 and 2.0 Z_o optimum conditions. The Z_o optimum simulation gave performance hardly distinguishable from the theoretical from 10 MHz to 108 MHz.

The measured curves evidence low frequency effects. As apparent from equation (9) low frequency cutoff is a function of the load impedance, the length of the ferrite covered section, and the permeability of the ferrite at the lower frequencies. The low 1 dB frequencies were 1 MHz or less.

Input Matching

The broadband input matching of the transistor is considerably more difficult than the matching at the output. The transistor base input is a low impedance which makes it impractical to select the value of characteristic impedance considered optimum for the transformer. The input impedance appears to be an R-L-C series circuit from 30 MHz to 80 MHz. Large signal base input impedance measurements of this transistor at 30 MHz, 40 MHz, 60 MHz, and 80 MHz indicate the resistance R_B = 2Ω, the inductance L_B = 3nH and the capacitance C_B = 2500 pF. The L_B is due to parasitic lead inductance in the transistor package plus any contributed by circuit interconnections. Extreme care in circuit construction is required to minimize this series base input inductance. This input circuit is resonant at about 58 MHz and has a Q = 0.55. Minimized input Q is desirable for transistors intended for broadbanding. Fig. 9 depicts the transistor input circuit.

Universal design curves for the 4:1 transformer, with various reactive terminations, are not practical, but design attitudes evident from this example are applicable to other transistors and other frequency ranges. The equations (6) and (7) can always be applied to any particular design situation.

Some general 4:1 transformer characteristics apparent from Fig. 6 and 7 with real loads may be used to guide design attitudes. The transistor input impedance is inductive above 58 MHz. The design curves for reactance at the high impedance port suggest that the use of low Z_o (i.e. Z_o optimum or less) would lend an increasingly capacitive phase with frequency. This would offset the increasingly inductive tendency of the transistor input impedance with frequency. Unfortunately, this specifies Z_o values from 1Ω to 4Ω. Such unusually low values have been achieved with homemade coaxial and strip transmission line, but have not been useful because of the unavoid-

ably bulky construction. The minimum interconnection lengths contribute parasitic inductance that more than offsets the improvements anticipated over higher Z_o lines with compact interconnections.

The lowest practical values of Z_o are 10Ω to 12Ω. Assuming the use of 12Ω for design, the previous curves indicate that inductive reactance contributed by the transformer is minimized by minimum length. The smallest practical construction length of transmission line with Z_o = 12Ω is about 2 inches. Fig. 10 and 11 show the variation of R_{IN1} and X_{IN1} with Z_o = 12Ω transmission line of lengths 2.0, 2.5 and 3.0 inches. The differences in performance are seen to be relatively minor. These curves were developed from computer calculations using equations (7) with Z_L being the transistor input impedance.

The transistor input was carefully simulated with a 2Ω carbon resistor, three leadless capacitor chips totaling 2500 pF. With interconnecting inductance of 3 to 4 nH. The dotted curve shows measured performance for a 4:1 transformer constructed with a 2.5 inch length of Z_o = 12Ω line. The transformer was identical in structure to the strip line unit at the lower right of Fig. 1d. Measurements for these curves were made with a HP 4815A Vector Impedance Meter. Corrections must be made in low impedance measurements to account for the parasitics in the probe. These values are typically a resistance of 0.6Ω and an inductance of 12 to 15 nH in the probe. The exceptionally close correlation of measured and calculated curves emphasizes that the calculated values for these transformers are reliably accurate in predicting performance.

Curves of calculated R_{IN1} and X_{IN1} for various values of Z_o with line length fixed at 2.5 inches are shown in Fig. 12 and 13. These calculations do not include the effects of parasitic series inductances. It is likely that larger parasitics would accompany the lower Z_o values thereby nullifying much or all of the slight performance edge indicated by the curves.

It is always desirable to maintain an operating load line as purely resistive as possible. Where this is not possible best stability often favors capacitive rather than inductive reactance. Assuming this to be true, the addition of a parallel capacitor added as shown at point 1 in Fig. 15 can maintain effectively the phase of Z_{IN1} capacitive over the entire 30 MHz to 80 MHz range rather than growing increasingly inductive above 50 MHz. For zero phase at 80 MHz:

$$C = \frac{X_{IN1}}{\omega \left(R_{IN1}^2 + X_{IN1}^2 \right)} \tag{11}$$

where ω, X_{IN1}, and R_{IN1} are defined at 80 MHz. This given C close to 100 pF. The use of a variable capacitor would allow trimming to the fixed value for the best overall operation. The cascading of an additional 4:1 transformer to step up Z_{IN1} paralleled by the capacitor gives an input Z_{IN2} at point 2 as shown in Fig. 15.

The large signal dynamic swing of a transistor stage is limited by available voltage. For

this reason load lines are conventionally represented by the equivalent parallel resistance for partly reactive loads. Real output power can then be defined easily by $V_{pK}^2/2R_L$ where R_L is the equivalent parallel load resistance.

R_{IN2} represents the equivalent parallel resistance looking in at point 2. Fig. 15 gives a clear representation of this. Fig. 14 shows the behavior of R_{IN2} and the phase of Z_{IN2} with frequency for two values of Z_0. The transmission line length is 2.5 inches. A Z_0 of 8Ω gave results hardly distinguishable from the 16Ω value plotted. From Fig. 14 the better choice of Z_0 is the 16Ω value which produces the smaller overall variation in R_{IN2} and phase. These variations must be accepted as a consequence of the low impedance R-L-C input of the transistor. The use of the capacitor has preserved the non-inductive character of this load line.

If the efficiency of the driver stage is critical the value of this load line (between 32Ω and 40Ω) is somewhat low. An additional 4:1 transformer would give a load line resistance too large to develop sufficient drive power for the output stage ($P_{IN} \approx 4$ watts). The use of a 9:1 transformer in place of the second 4:1 between points 1 and 2 would be suitable. The resultant load line would have very similar trends to Fig. 14 but with the approximate resistive value between 72Ω and 90Ω. This load line would permit delivering the 4 watts of input power to the 2N5071 transistor at good efficiency.

Push-Pull Design

Assume it is desired to operate two 2N5071 transistors for an output power of 60 watts. Two 180 degree phase hybrids as previously described are necessary: One for splitting single ended input power source into two out of phase sources driving the transistor inputs, the other for adding the outputs from the two transistors. Fig. 16 shows a typical arrangement.

Output Matching

The output matching offers no particular problems. Assuming the antenna load to be 50Ω it must be partially transformer 2:1 by the low pass filter to either a 25Ω or 100Ω value for efficient output matching. Fig. 16 shows the load as 25Ω. The hybrid can easily be constructed with the optimum values of Z_0 as indicated. Under the conditions shown each transistor operates into a 12.5Ω load line suited to 30 watts output power from each. The transmission line lengths are not critical. A length of 4 inches for each winding is adequate.

Input Matching

The input hybrid is constructed of the lowest practical characteristic impedance line $Z_0 = 12.5\Omega$. The shortest practical length of 2.5 inches will minimize mismatch effects. During push-pull operation one transistor is conducting while the second is biased off. The conducting transistor has a low impedance as described

previously. The off transistor appears as a relatively high impedance due to the reverse biased base emitter junction. This affects the impedance of the input port of the power splitter. With balanced loads of $R_B = 2\Omega$ each the input would be approximately 4Ω. However, with one output port terminated in an open circuit the input impedance approaches three times this value. For transistor loads the input impedance is somewhere between these extremes. Fig. 15 indicates R_{IN} could be between 6Ω and 10Ω typically. The 4:1 transformer would step this up to a 25Ω to 40Ω load line for the driver. This is a suitable load line for developing 10 watts of drive power with good efficiency.

Conclusions

A group of transmission line transformers and hybrids useful for untuned broadband RF transistor amplifiers has been presented together with the fundamental principles underlying their operation. Design curves and expressions for realizing predictable performance from 4:1 transformers terminated in resistive loads were presented. The typical behavior of RF power transistor input impedance over the VHF range of 30 MHz to 80 MHz was described. A means for transforming this input impedance to a non-inductive load line was demonstrated. Finally the push-pull power adding of two transistors through the use of the 180 degree hybrid was illustrated.

Key Words

Transmission line transformers, broadband transformers, impedance matching transformers, RF transformers, untuned VHF transistor power amplifiers, broadband hybrids.

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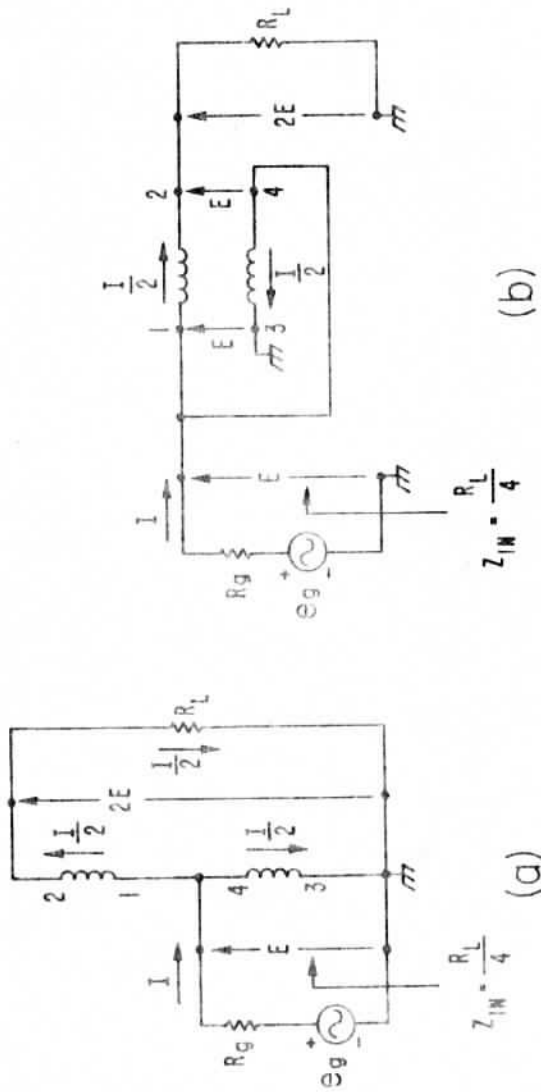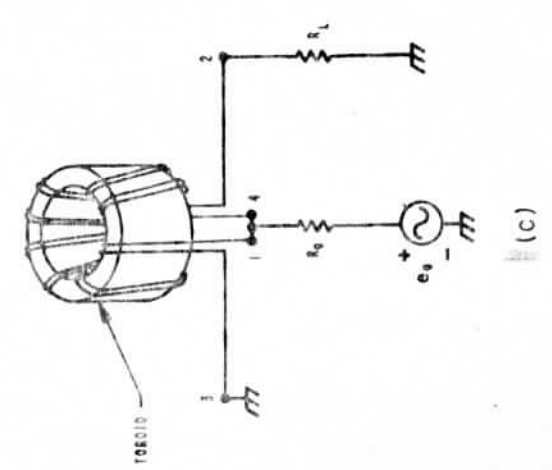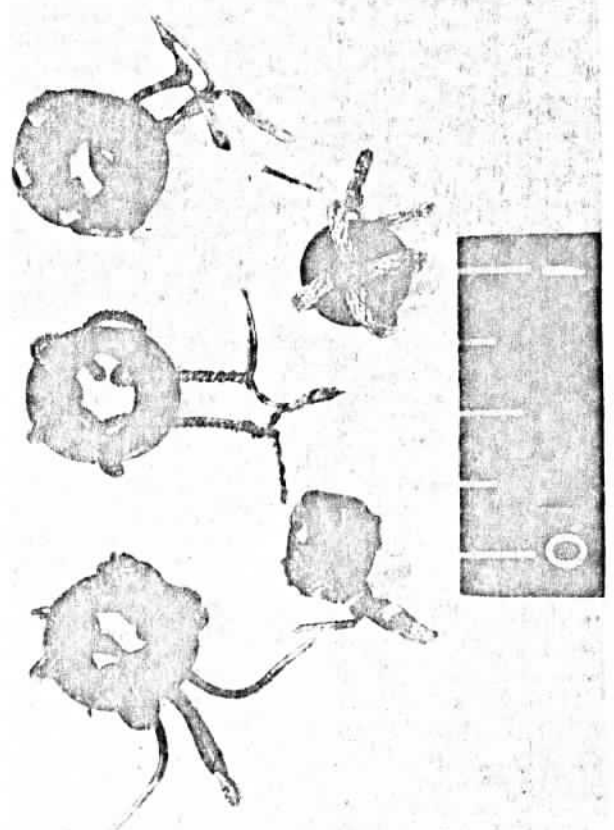N. Y., McGraw Hill, 1950.

Figure 1. 4:1 Transformer (a) Autotransformer analogue. (b) Equivalent trans-
mission line form. (c) Pictorial representation of construction
using twisted pair transmission line. (d) Actual units of twisted
pair, strip, and coaxial transmission line.

Cirkit koppartejp 60µ
kladd 95µ
teflon 250µ
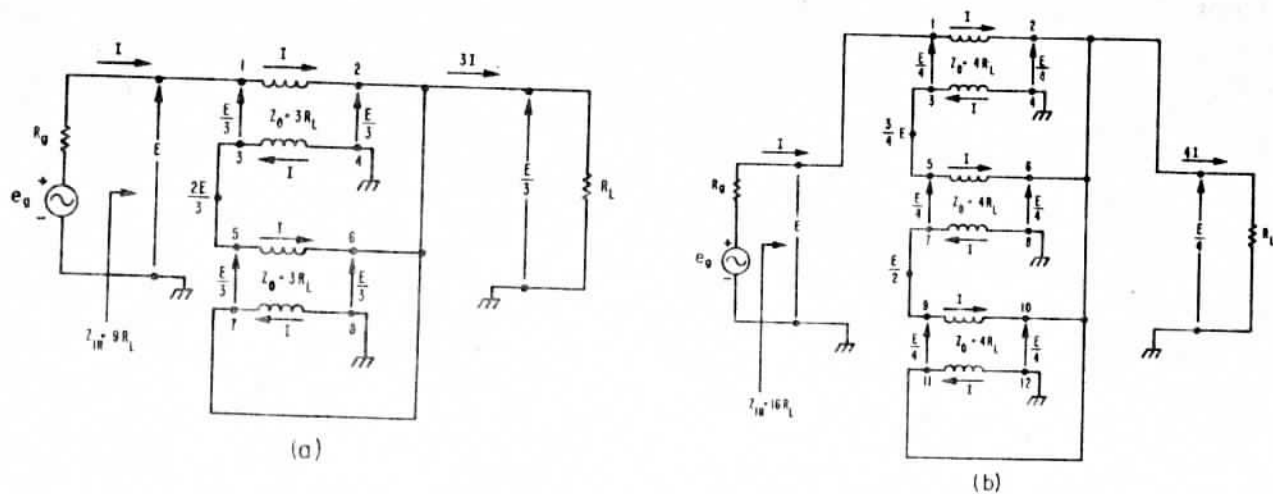
Figure 2.   9:1 impedance transformer.   (b) 16:1 impedance transformer.
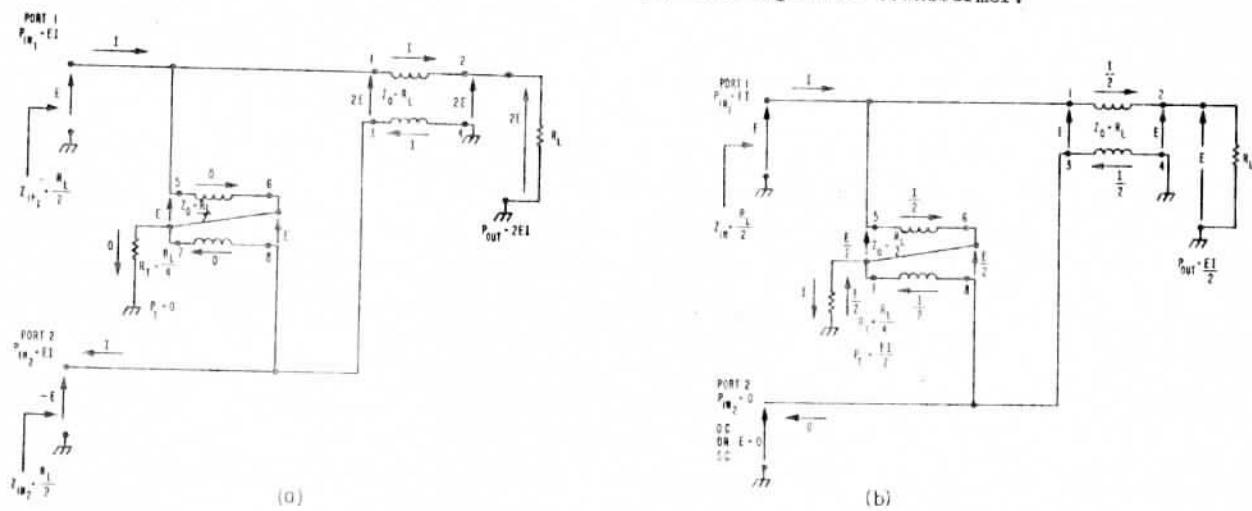
Figure 4.   Broadband hybrid for power adding or splitting signals 180° out of phase.   (a) Normal balanced conditions for power adding ($P_{IN1} = P_{IN2}$). (b)   Severe mismatch operation for power adding ($P_{IN1} = EI$, $P_{IN2} = 0$).
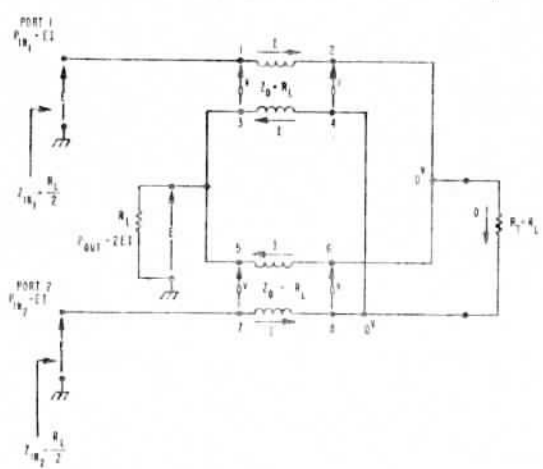
Figure 3.   Broadband hybrid for power adding or splitting of in-phase signals.
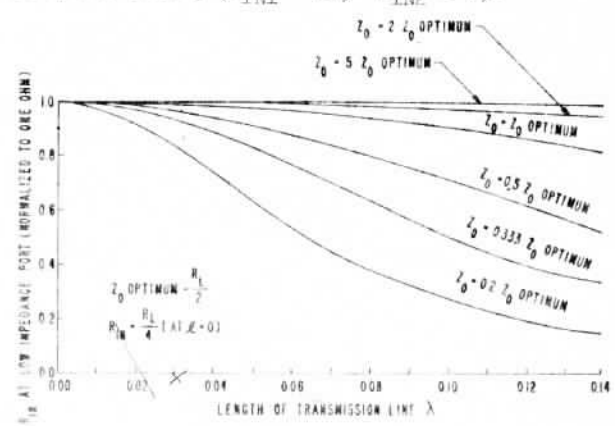
FIG. 5 Normalized input resistance R_{IN} at the low impedance port of the 4:1 transformer with real load R_L versus length of transmission line in wavelengths for various values of characteristic impedance Z_0.
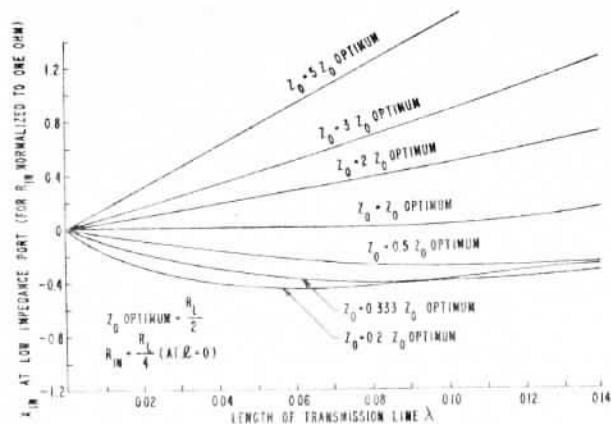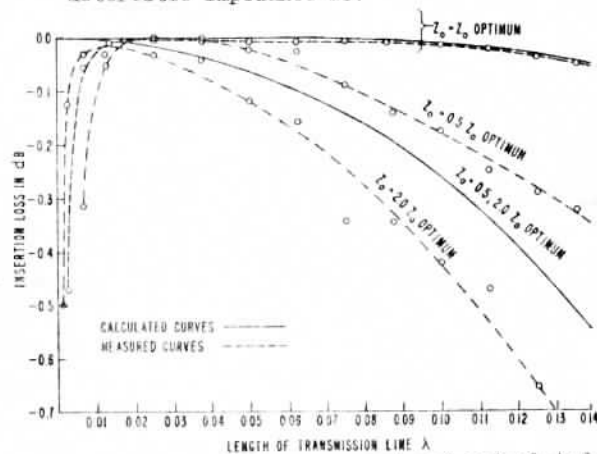
FIG. 6 Normalized input reactance X_{IN} at the low impedance port of the 4:1 transformer with real load R_L versus length of transmission line in wavelengths for various values of characteristic impedance Zo.
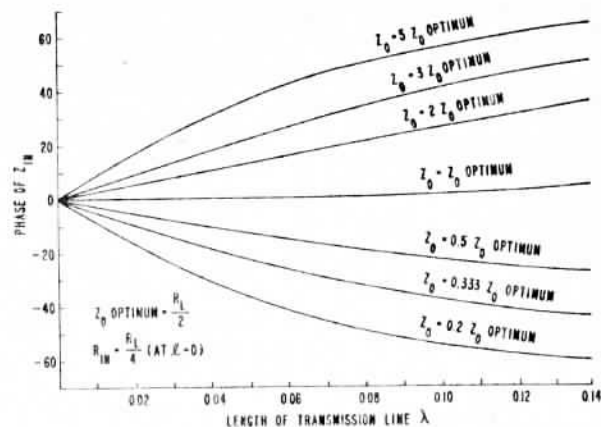
FIG 7 Phase of input impedance at the low impedance port of the 4:1 transformer with real load R_L versus length of transmission line in wavelengths for various values of characteristic impedance Zo.
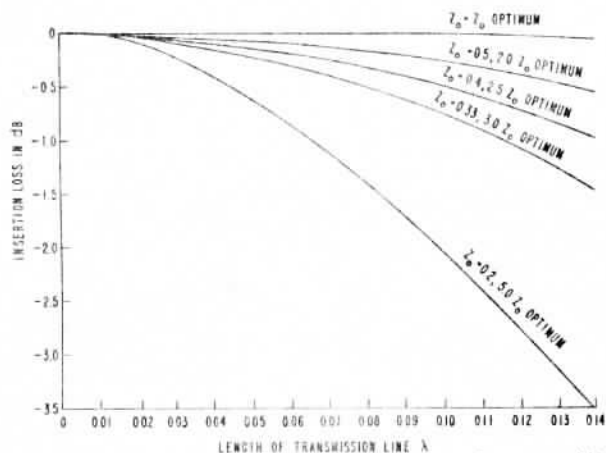
FIG 8 Insertion loss of the 4:1 transformer with real load R_L versus length of transmission line in wavelengths for various values of characteristic impedance Zo.

FIG 9 Curves comparing measured and calculated insertion loss for a 4:1 transformer with real load R_L versus length of transmission line in wavelengths for several values of characteristic impedance Zo.
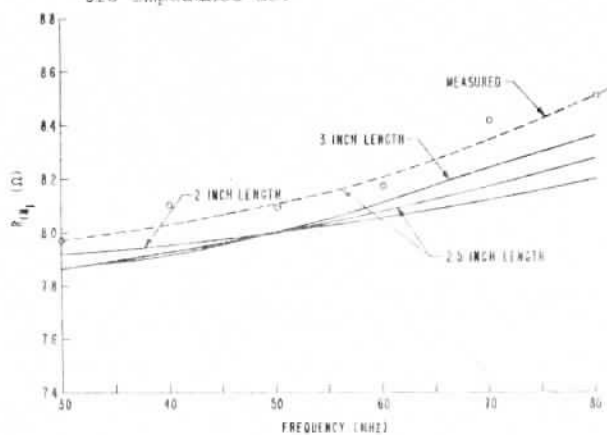
FIG. 10 Calculated and measured curves of input resistance R_{INI} (at the high impedance port of a 4:1 transformer with transistor load) versus frequency for various lengths of transmission line with Zo equals 12 Ω
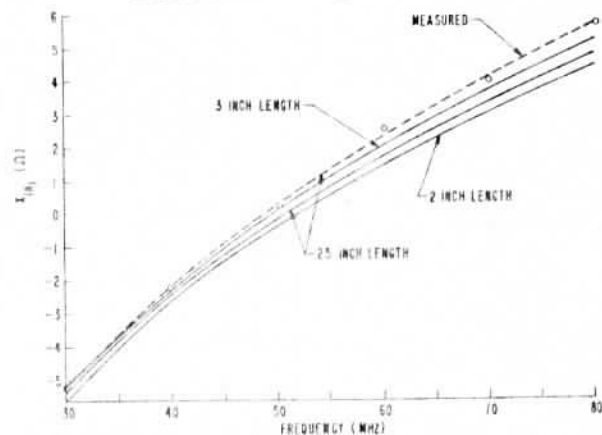

FIG 11 Calculated and measured curves of input reactance X_{INI} (at the high impedance port of a 4:1 transformer with transistor load) versus frequency for various lengths of transmission line with Zo equals 12 Ω
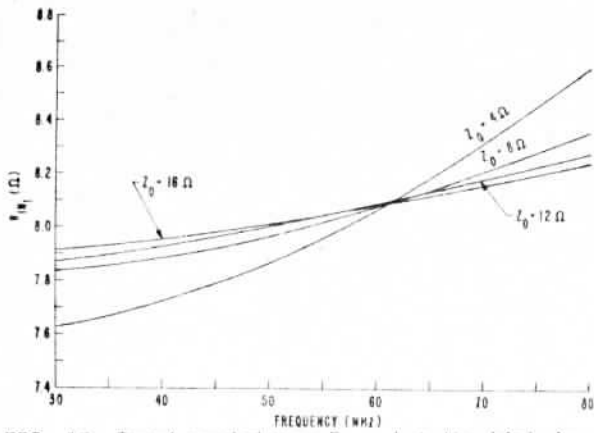
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FIG. 12 Input resistance R_{IN1} (at the high impedance port of a 4:1 transformer with transistor load) versus frequency for various values of characteristic impedance Zo. (Line length constant at 2.5 inches).
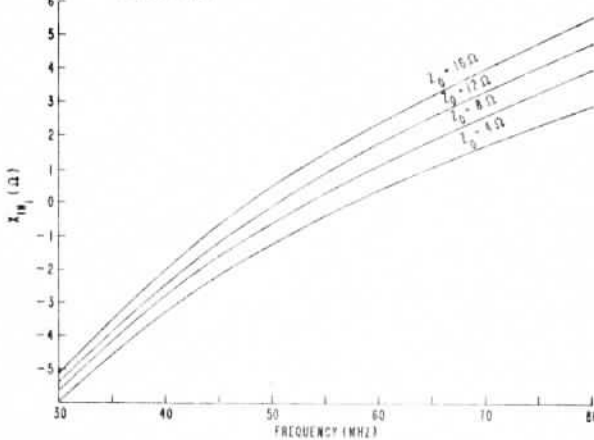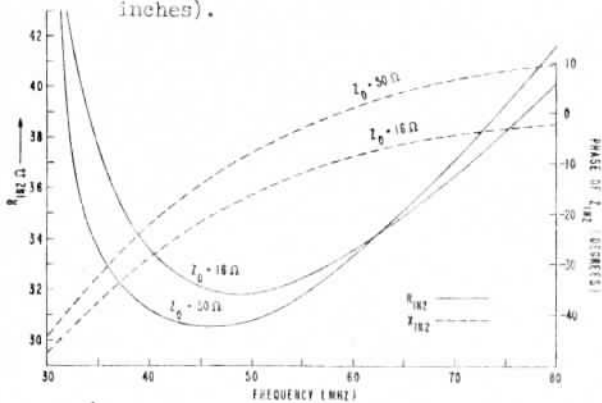

FIG. 13 Input reactance X_{IN1} (at the high impedance port of a 4:1 transformer with transistor load) versus frequency for various values of characteristic impedance Zo. (Line length constant at 2.5 inches).

FIG.14 Curves of input resistance R_{IN2} and phase of input impedance Z_{IN2} (at the high impedance port of two cascaded 4:1 transformers with transistor load) versus frequency.
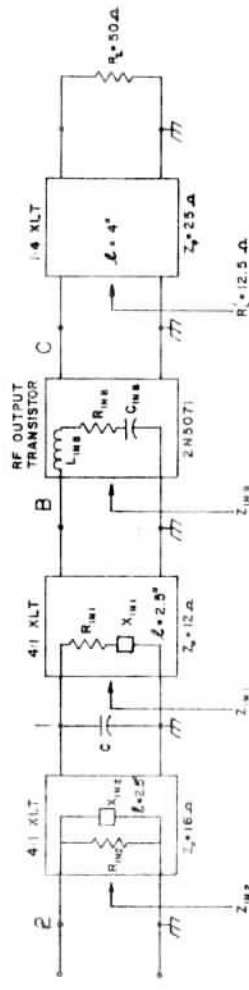

Figure 15. Block diagram of the 30 watt, 30-80 MHz power amplifier with input and output matching transformers.
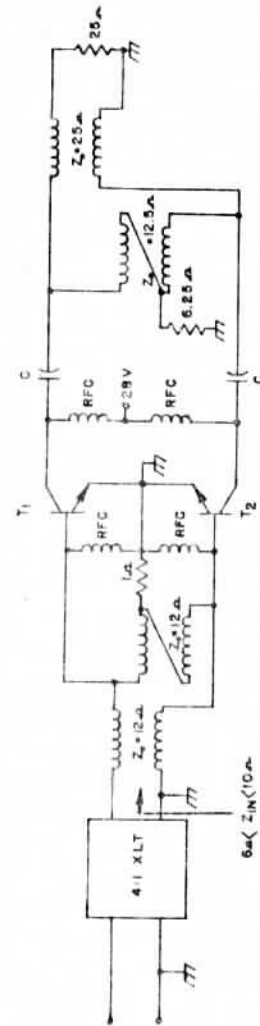

Figure 16. Basic schematic for a pair of 2N5071 transistors operating in a broadband push-pull amplifier with 180 degree phase hybrids for power splitting and power adding.