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Research and Development Technical Report
ECOM-2989

BROADBAND TRANSFORMER DESIGN
FOR RF TRANSISTOR POWER AMPLIFIERS

by

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July 1968

AD 676816

ECOM

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TECHNICAL REPORT ECOM - 2989

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JULY 1968

DA Task No. 1H6-22001-A-056-13

U. S. ARMY ELECTRONICS COMMAND
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ABSTRACT

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 W of output power from 30 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 hy-
brids to operate a pair of these transistors in push-pull operation is
described.

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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 provides 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 $Z_0 = \frac{1}{2} R_L$ 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 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 princi-

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ples 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., I/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 with 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 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. Figure 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. Figure 1d is a photograph of several 4:1 transformers made in this manner. The larger in the upper row have windings from 3.5 to 4.5 in. long. The transformer at the upper left uses a loosely twisted pair of number 20 formvar coated transformer wire having a $Z_o = 50\Omega$. 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 formvar transformer wire about .10 in. wide paired together with contact bond cement gives a $Z_o = 10\Omega$. These transformers have each carried 100 W 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 small toroids and have 2.5 in. lengths of transmission line wound about them. The transformer at the lower right uses RG-196 (50Ω coaxial line) with the

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outer insulation removed. This improves its flexibility and permits winding
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 $Z_0 = 10 \, \Omega$ 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 trans-
former may be informative. Figure 2a shows a 9:1 impedance transformer in
schematic form. Two equal lengths of transmission line are used. Con-
struction 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 in-
dicated. For optimally matched conditions and negligible phase shift:

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

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

$$Z_0 = \frac{E/3}{I} = \frac{R_{IN}}{3} = 3 R_L \tag{3}$$

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 trans-
former. 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^2:1
(n always an integer). Also the optimum Z_0 choice will always be given by:

$$Z_0 = \sqrt{R_L R_{IN}} \tag{4}$$

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 individ-
ual sources. The hybrid provides excellent isolation between the input
ports when separate sources of power are to be added (or alternately isola-
tion between two output ports acting as separate sources when power split-
ting). 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 character-
istics will effectively eliminate chain type failures common in multiple
transistor output stages with transistors directly paralleled at their out-
puts.

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. Figure 3 illustrates schematically in transmission line form one such hybrid together with voltage, current and impedance relationships. The component R_p 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_p and half is delivered to the load R_L.

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

Figure 4a illustrates the interconnections and operating relationships for a 180° phase hybrid being used for power adding. Figure 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 input at the working port is dissipated in R_p. In general for this hybrid:[6]

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

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 $Z_0 = R_L$, 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 impractical to exactly match Z_0 to R_L; the length of this section must be maintained quite short relative to $\lambda/8$ if reactive effects due to mismatch are to be avoided.

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APPLICATION OF BROADBAND TRANSMISSION LINE STRUCTURES TO THE DESIGN OF UN-
TUNED BROADBAND RF TRANSISTOR AMPLIFIERS

Requirements for broadband untuned VHF transistor amplifiers often de-
mand operation over two or more bands representing over an octave of fre-
quency 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 de-
sign 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 fix-
ed 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 V d.c. supply. Operation is specified to be 30 W
of RMS output power from 30 to 80 MHz. The antenna load is 50 Ω .

Output Matching

The optimum load line for producing 30 W of output power from this
transistor operating from 28 V d.c. is 13 Ω to 15 Ω . One 4:1 transformer
will step down the 50 Ω antenna load to 12.5 Ω . 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.[7] At the low impedance port ($R_{IN} \approx \frac{R_L}{4}$) i.e., the load is
stepped down:

$$Z_{IN} = Z_0 \left[\frac{Z_L \cos \theta l + j Z_0 \sin \theta l}{2 Z_0 (1 + \cos \theta l) + j Z_L \sin \theta l} \right] ; \tag{6}$$

at the high impedance port ($R_{IN} \approx 4 R_L$) i.e., the load is stepped up:

$$Z_{IN} = Z_0 \left[\frac{2 Z_L (1 + \cos \theta l) + j Z_0 \sin \theta l}{Z_0 \cos \theta l + j Z_L \sin \theta l} \right] \tag{7}$$

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where Z_L is the load impedance, θ 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 $\theta\ell$) and as a function of Z_0.

Figures 5, 6, and 7 display R_{IN}, X_{IN}, and phase respectively, at the low impedance port (i.e., transformer used for load step down) as a function of length and Z_0 of the transmission line. The R_{IN} and X_{IN} values can be unnormalized by multiplying by $R_L/4$. The various Z_0 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_0 equal to Z_0 optimum, R_{IN} drops off gradually, from a value of $\frac{R_L}{4}$, 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_0 optimum the inductive character of the input impedance is accented. The behavior closely approaches a fixed R-L series circuit when Z_0 is Z_0 optimum or greater. With Z_0 values less than Z_0 optimum, the input impedance behaves as a parallel R-C circuit.

Figure 8 shows curves of insertion loss as a function of line length for various values of Z_0. 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 ℓ 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 $R_g = \dfrac{R_L}{4}$ the value of R_{IN} at $\ell = 0$ for load step down or

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

Note that in Fig. 8 pairs of Z_0 values, related by KZ_0 optimum and $\dfrac{Z_0 \text{ optimum}}{K}$, where K can be any given constant, yield identical insertion loss.

For each of the curves described above the abscissa is given in 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:

$$\ell \leq \frac{7200\,n}{f_u} \quad \text{inches.}$$

(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:

$$\ell \geq \frac{20\,R_L}{\left(1 + \frac{u}{u_o}\right) f_L} \quad \text{inches}$$

(10)

where R_L is the load resistance in ohms, f_L the lowest operating frequency in MHz, and u/u_o the relative permeability of the ferrite at f_L. This expression relates a length corresponding to the input impedance having a $10°$ to $15°$ 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 u/u_o of 15. With $R_L = 50\,\Omega$, and $f_u = 30$ MHz, equation (9) gives $\ell = 2.08$ in. Reference to the curves for the low Z port (Fig. 7) with $Z_O = Z_O$ optimum = $25\,\Omega$ at $.08\lambda$, R_{in} is about 6% down from the ideal $12.5\,\Omega$ value and X_{in} is essentially zero.

Using equation (8) with $f_u = 80$ MHz and $n = 0.08$ it follows that $\ell = 7.5$ in. Therefore, any length between 2.5 in. and 7.5 in. wound on a ferrite toroid will offer excellent performance. Arbitrarily let the choice be a length of 4 in. It is apparent that for this frequency range of 30 MHz to 80 MHz the design of the transformer is quite 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 $25\,\Omega$. 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 $12.5\,\Omega$, X_{in} is about $5\,\Omega$ 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 gives more specific meaning to the effect of the $5\,\Omega$ inductive reactance. Figure 9 illustrates the output matching circuit in block diagram.

Measurements

To verify calculated results, measurements were made over a frequency range of 1 through 108 MHz using an HP 4815A Vector Impedance Meter. A 9-in. length of RG-58 cable ($Z_O = 50\,\Omega$ and $\ell = 0.25\lambda$ at 200 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, covered the entire transmission line section. With this fixed $Z_O = 50\,\Omega$ line terminated in R_L of $50\,\Omega$, $100\,\Omega$ and $200\,\Omega$, 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

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compared in Fig. 9. Differences between measured and calculated were less than 0.2 dB at 100 MHz for the 0.5 and 2.0 Z_0 optimum conditions. The Z_0 optimum simulation gave performance hardly distinguishable from the theoretical from 10 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 lower 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 to 80 MHz. Large-signal base input impedance measurements of this transistor at 30, 40, 60, and 80 MHz indicate the resistance $R_s = 2\Omega$, the inductance $L_s = 3nH$ and the capacitance $C_b = 2500$ pF. The L_s 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. Figure 15 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_0 (i.e., Z_0 optimum or less) would lend an increasingly capacitive phase with frequency. This would offset the increasingly inductive attitude of the transistor input impedance with frequency. Unfortunately, this specifies Z_0 values from 1 to 4Ω. Such unusually low values have been achieved with homemade coaxial and strip transmission line, but have not been useful because of the unavoidably bulky construction. The minimum interconnection lengths contribute parasitic inductance that more than offsets the improvements anticipated over higher Z_0 lines with compact interconnections.

The lowest practical values of Z_0 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 $Z_0 = 12\Omega$ is about 2 in. Figures 10 and 11 show the variation of R_{IN} and X_{IN} with $Z_c = 12\Omega$ transmission line of lengths 2.0, 2.5 and 3.0 in. The differences in performance are seen to be relatively minor. These curves were developed from computer calculations using equations (7) with Z_1 being the transistor input impedance.

The transistor input was carefully simulated with a ˙ 1 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 in. length of $Z_O = 12 \Omega$ 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 resistance of 0.6Ω and 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 in. 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 to 80 MHz range rather than growing increasingly inductive about 50 MHz. For zero phase at 80 MHz:

$$C = \frac{X_{IN1}}{\omega \left(R_{IN1}^{2} + X_{IN1}^{2} \right)} \quad \text{farads}$$

where ω, X_{IN1}, and R_{IN1} are determined at 80 MHz. This gives 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 be defined easily by $V_{pk}/2R_L$ where R_1 is the equivalent parallel load resistance.

R_{IN2} represents the equivalent parallel resistance looking in at point 2. Figure 15 gives a clear representation of this. Figure 14 shows the behavior of R_{IN2} and the phase of Z_{IN2} with frequency for two values of Z_O. The transmission line length is 2.5 in. A Z_O of 8Ω gave results hardly distinguishable from the 16Ω value plotted. From Fig. 14 the better choice of Z_O 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 32 and 40Ω) is somewhat low. An additional 4:1 transformer

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would give a load line resistance too large to develop sufficient drive power for the output stage (P_{in} = 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 72 and 90 Ω. This load line would permit delivering the 4 W 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 W. Two 180° 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. Figure 16 shows a typical arrangement.

Output Matching

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

Input Matching

The input hybrid is constructed of the lowest practical characteristic impedance line Z_0 = 12.5 Ω. The shortest practical length of 2.5 in. 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_s = 2 Ω each the input would be approximately 4 Ω. 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. Figure 15 indicates R_{in} could be between 6 and 10 Ω typically. The 4:1 transformer would step this up to a 25 to 40 Ω load line for the driver. This is a suitable load line for developing 10 W 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 to 80 MHz was described. A means for transforming this input impedance to a non-inductive load line was

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demonstrated. Finally the push-pull power adding of two transistors through the use of the 180° hybrid was illustrated.

ACKNOWLEDGEMENTS

The authors wish to acknowledge the assistance of Mr. Dave Huisman for the necessary computer programming for this work. Mr. Robert Horn and Mr. Dennis Bowman contributed noteworthy ideas and observations during the course of several stimulating discussions.

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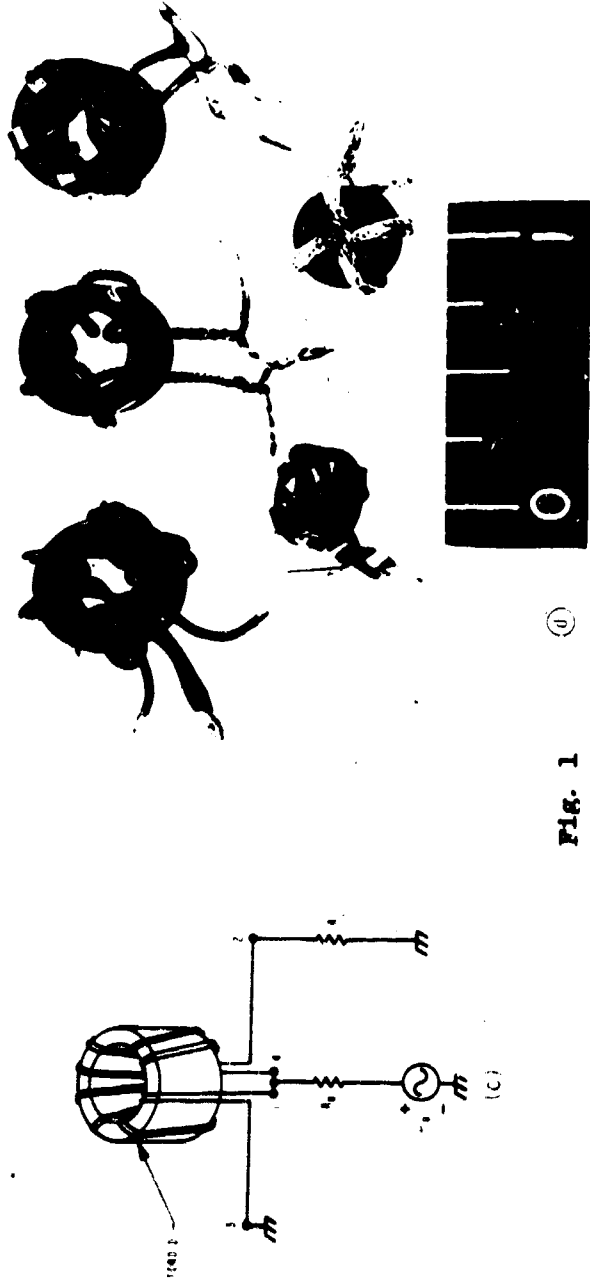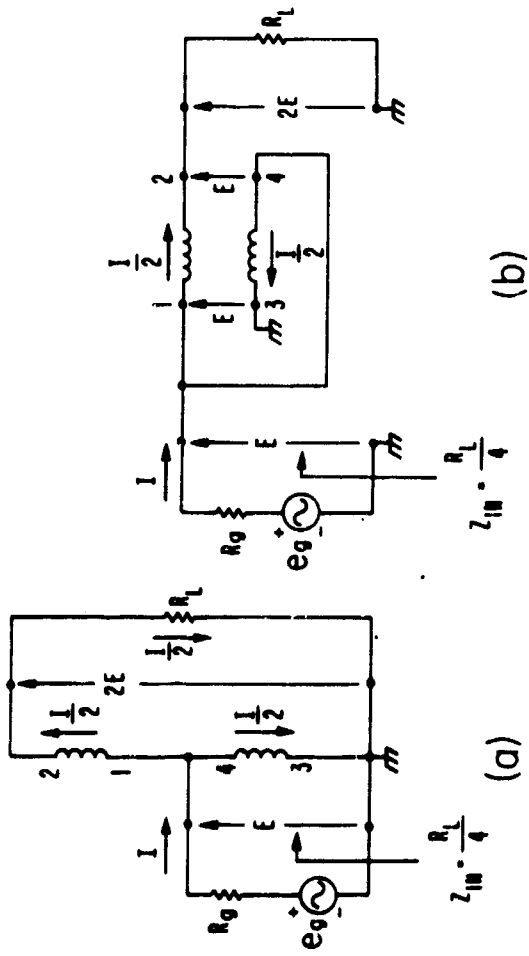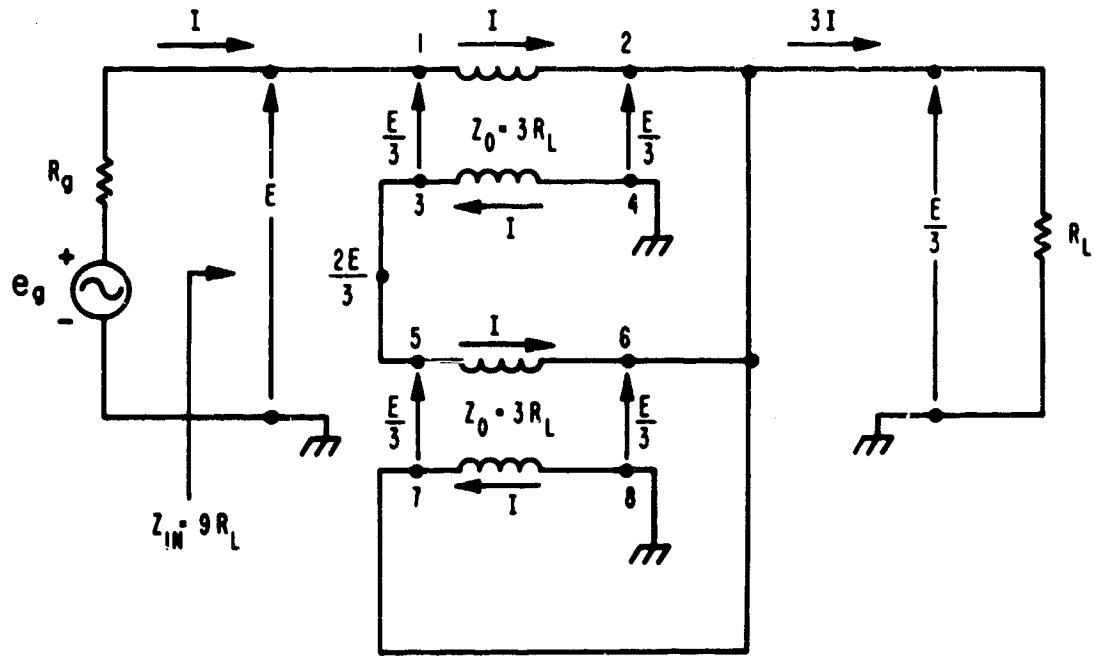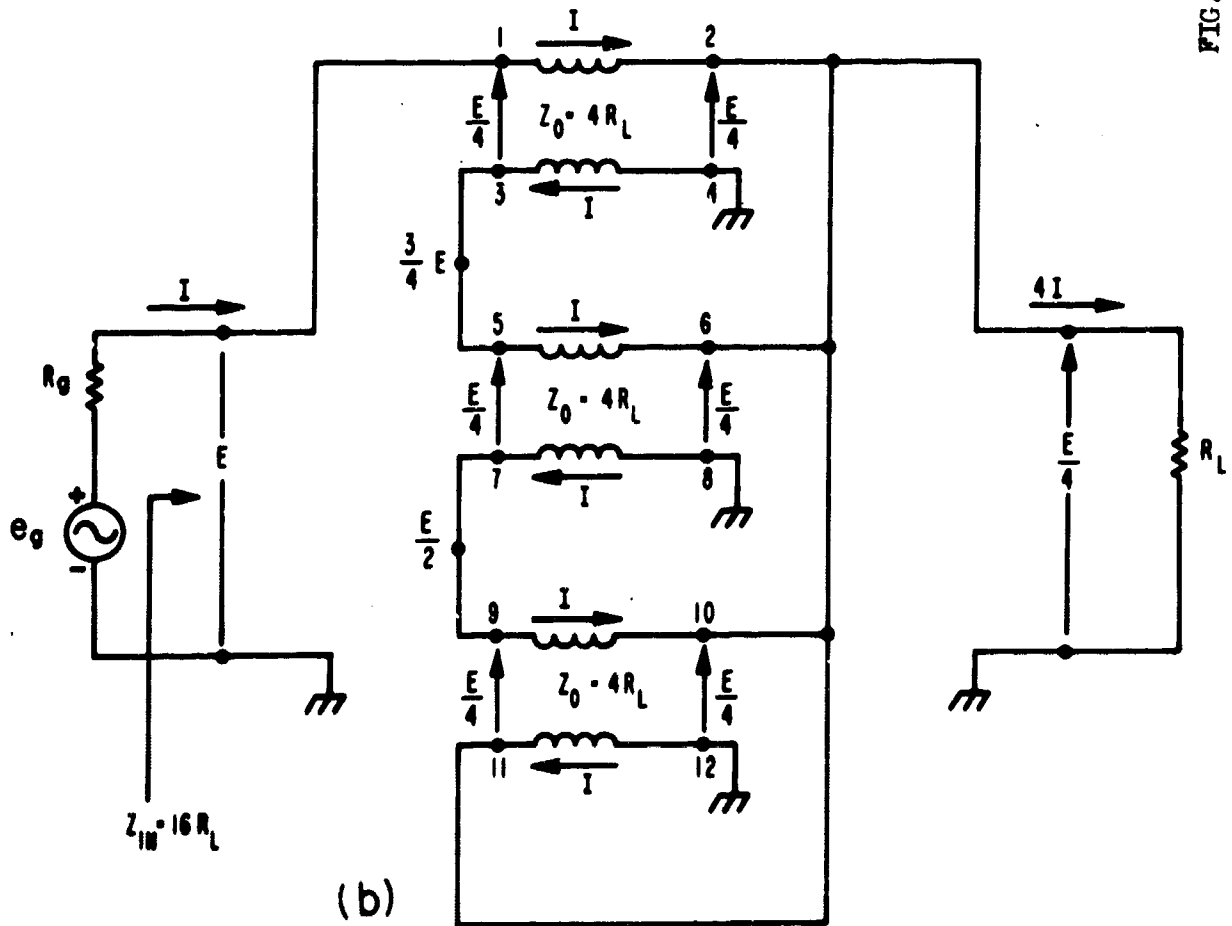7. W. C. Johnson, Transmission Lines and Networks, N. Y., Mc Graw-Hill, 1950.

Fig. 1

(a)

FIG. 2

(b)

13

FIG. 3

14

(a)
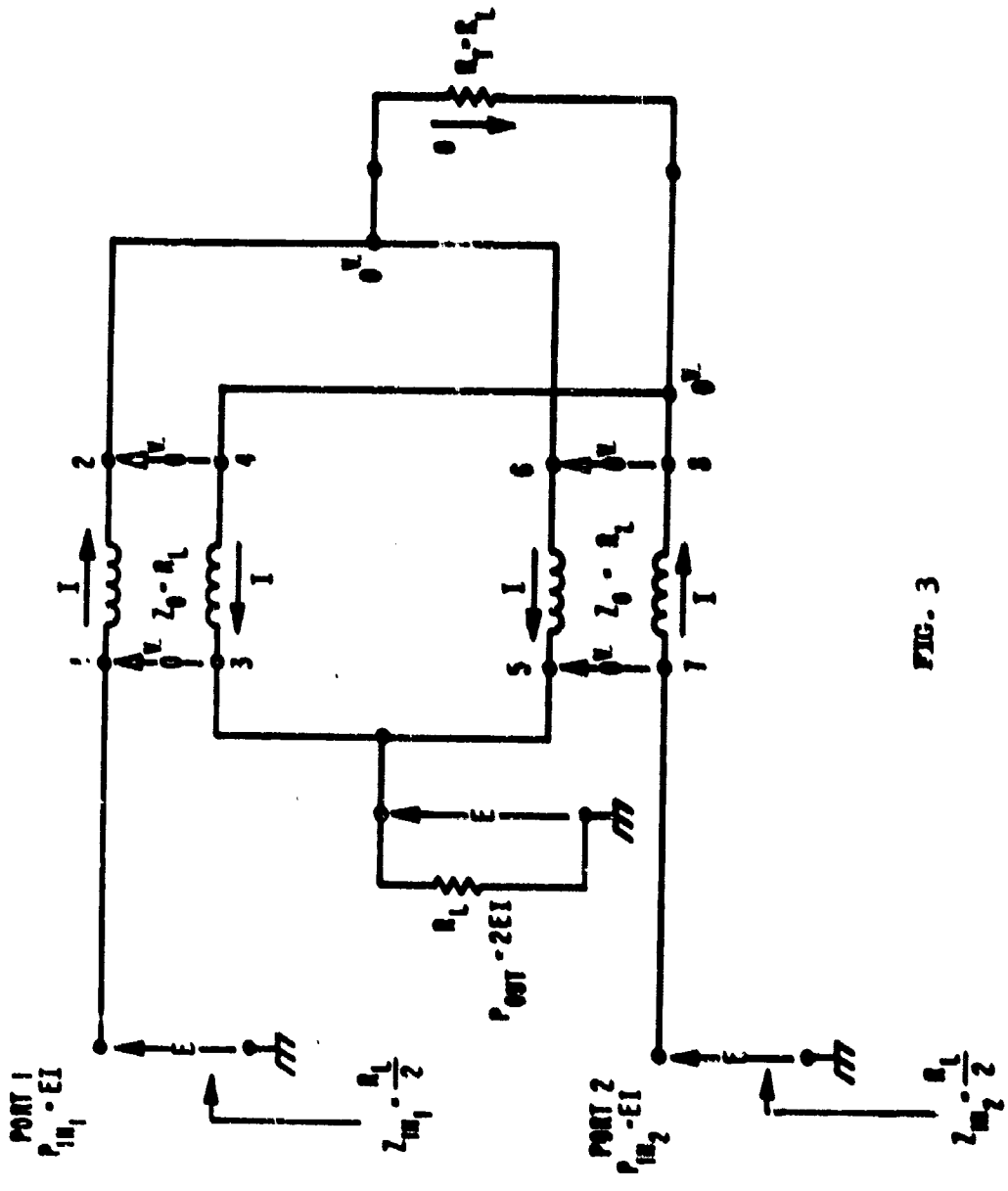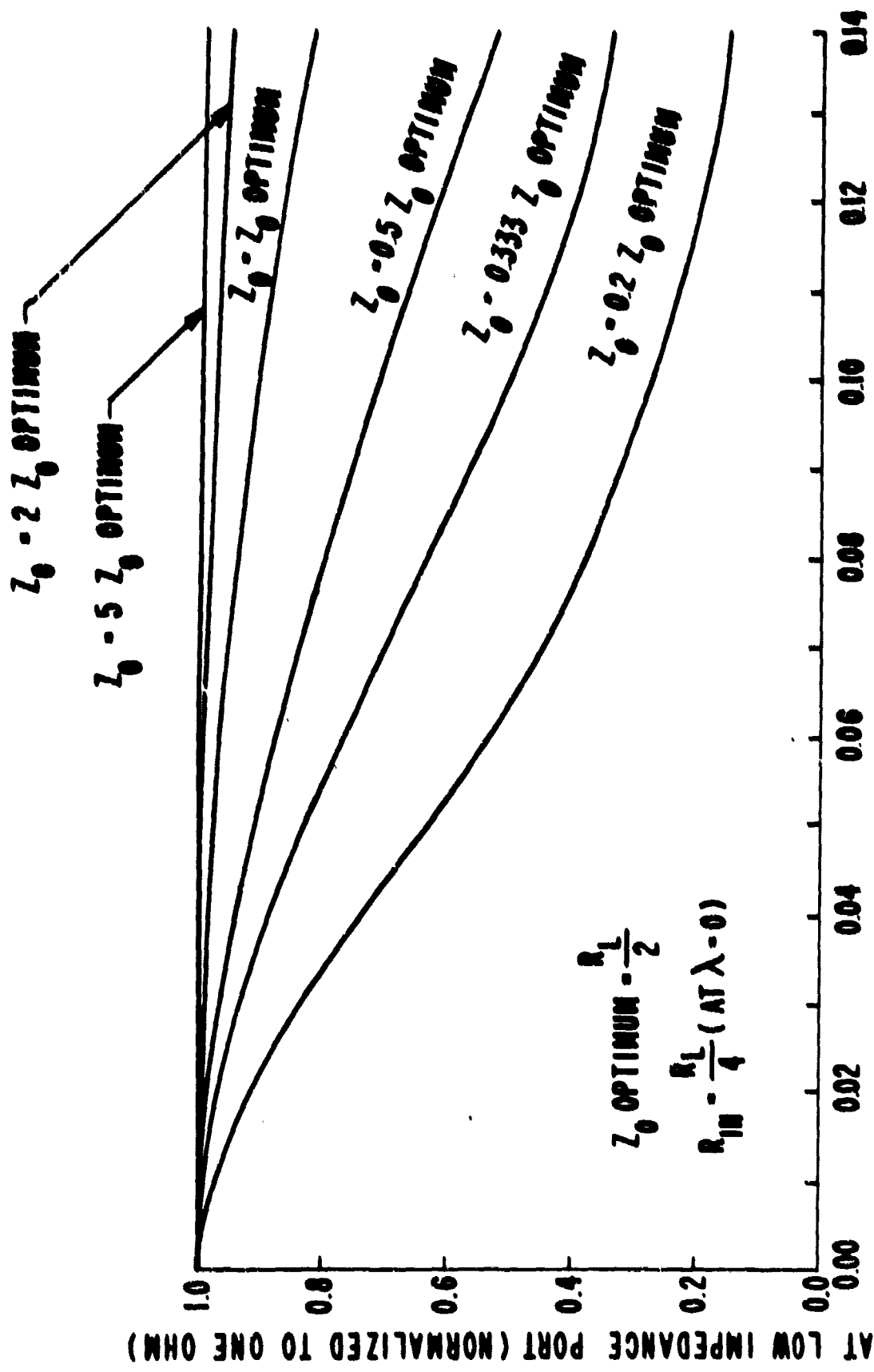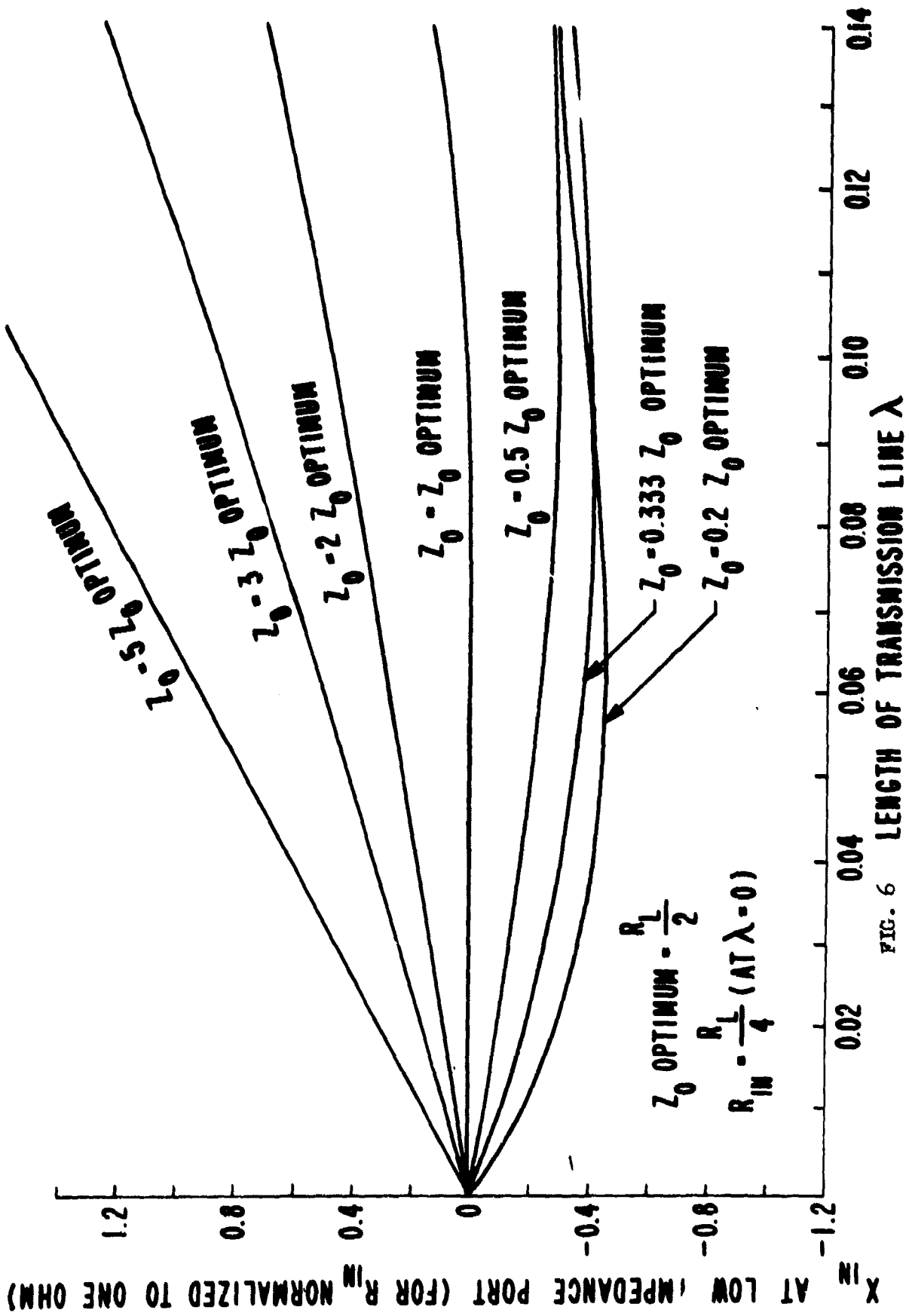
(b)

FIG. 4

15

FIG. 5 LENGTH OF TRANSMISSION LINE λ

16

FIG. 6 LENGTH OF TRANSMISSION LINE λ

Z_0 OPTIMUM = $\dfrac{R_L}{2}$

$R_{IN} = \dfrac{R_L}{4}$ (AT λ = 0)

Z_0 = 5 Z_0 OPTIMUM

Z_0 = 3 Z_0 OPTIMUM

Z_0 = 2 Z_0 OPTIMUM

Z_0 = Z_0 OPTIMUM

Z_0 = 0.5 Z_0 OPTIMUM

Z_0 = 0.333 Z_0 OPTIMUM

Z_0 = 0.2 Z_0 OPTIMUM

X_{IN} AT LOW IMPEDANCE PORT (FOR R_{IN} NORMALIZED TO ONE OHM)

17

Z₀ - 5 Z₀ OPTIMUM

Z₀ - 3 Z₀ OPTIMUM

Z₀ - 2 Z₀ OPTIMUM

Z₀ - Z₀ OPTIMUM

Z₀ - 0.5 Z₀ OPTIMUM

Z₀ - 0.333 Z₀ OPTIMUM

Z₀ - 0.2 Z₀ OPTIMUM

Z_0 OPTIMUM $= \dfrac{R_L}{2}$

$R_{IN} = \dfrac{R_L}{4}$ (AT $\lambda = 0$)

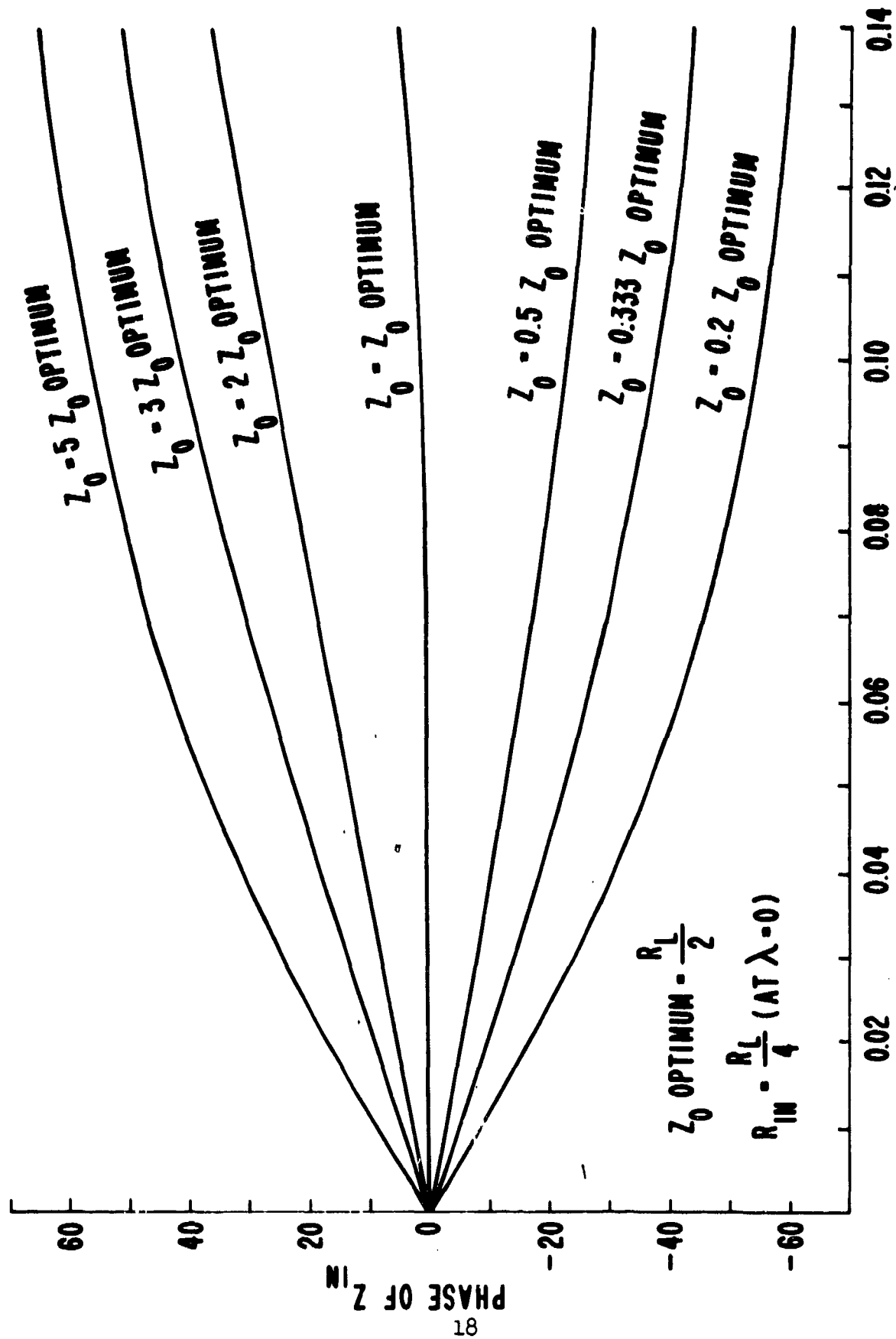PHASE OF Z_{IN}

LENGTH OF TRANSMISSION LINE λ

FIG. 7 LENGTH OF TRANSMISSION LINE λ
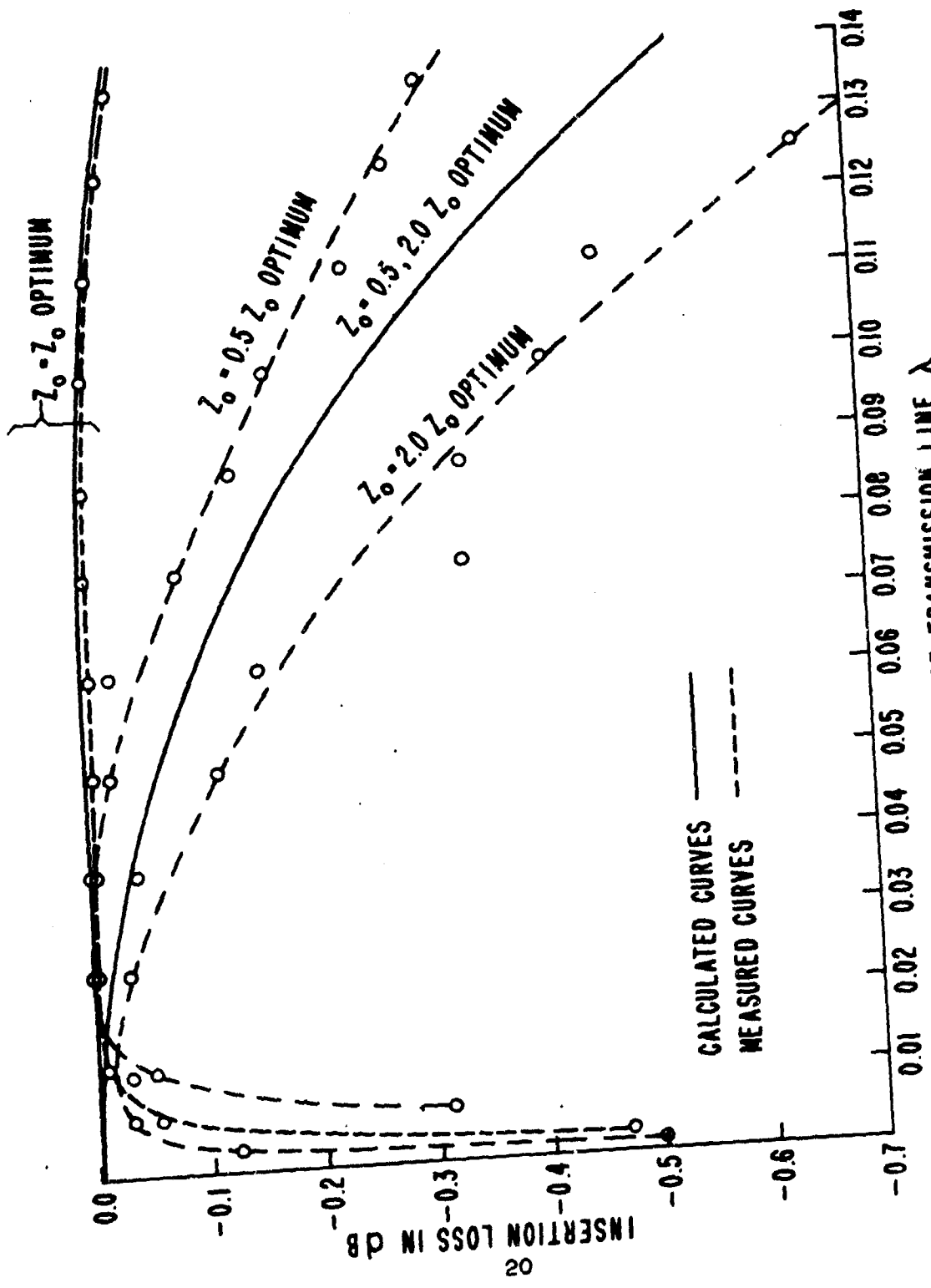
18

LENGTH OF TRANSMISSION LINE λ

Fig. 8

$Z_o = Z_o$ OPTIMUM

$Z_o = 0.5, 2.0 \ Z_o$ OPTIMUM

$Z_o = 0.4, 2.5 \ Z_o$ OPTIMUM

$Z_o = 0.33, 3.0 \ Z_o$ OPTIMUM

$Z_o = 0.2, 5.0 \ Z_o$ OPTIMUM

INSERTION LOSS IN dB

19

Z₀ = Z₀ OPTIMUM

Z₀ = 0.5 Z₀ OPTIMUM

Z₀ = 0.5, 2.0 Z₀ OPTIMUM

Z₀ = 2.0 Z₀ OPTIMUM

LENGTH OF TRANSMISSION LINE λ

Fig. 9

INSERTION LOSS IN dB

CALCULATED CURVES ———

MEASURED CURVES – – –

20

FIG. 10 FREQUENCY (MHZ)

21

FREQUENCY (MHZ)

Fig. 11

X_{IN_1} (Ω)
22

$Z \cdot 16\Omega$
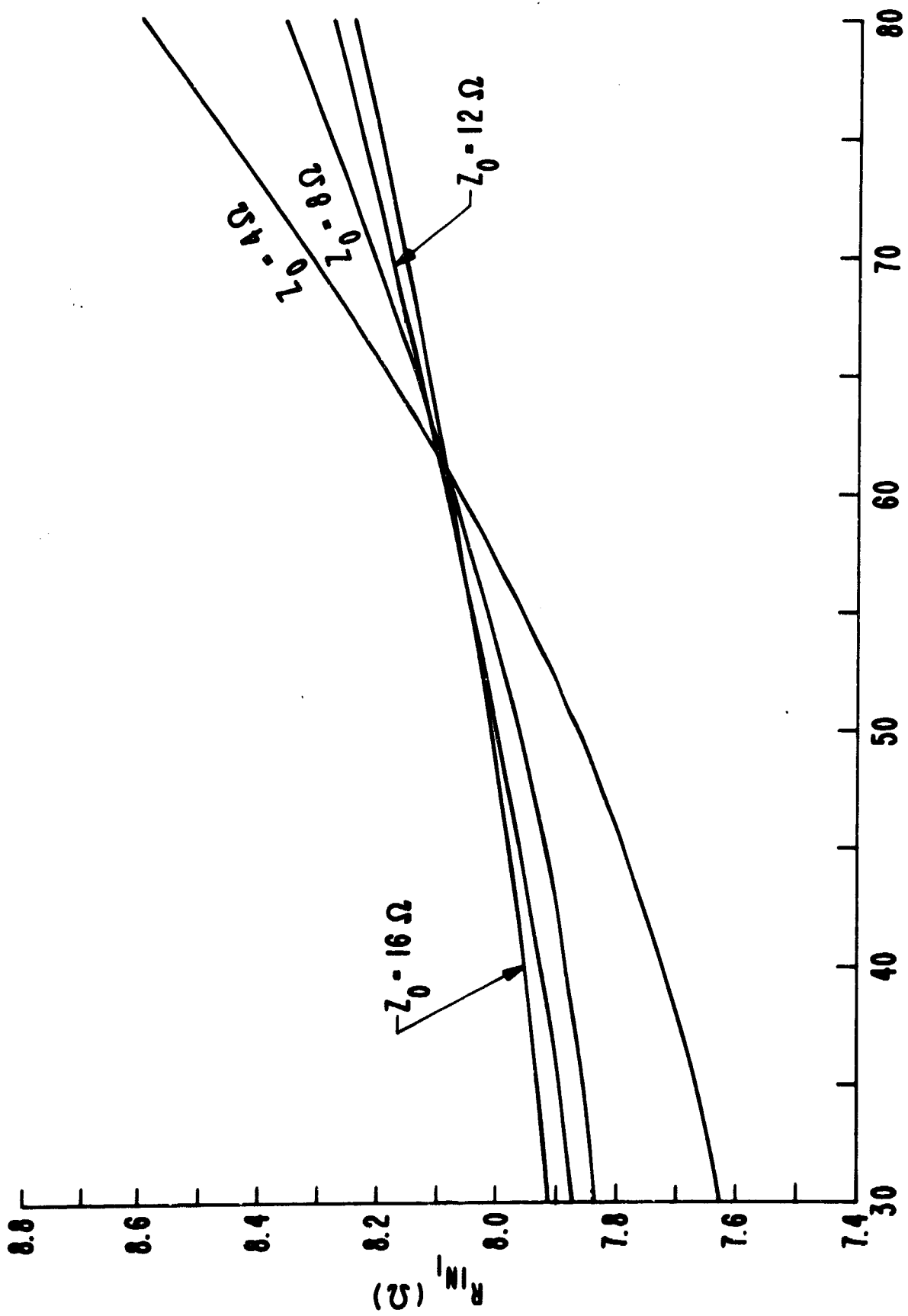$Z \cdot 12\Omega$
$Z \cdot 8\Omega$
$Z \cdot 4\Omega$
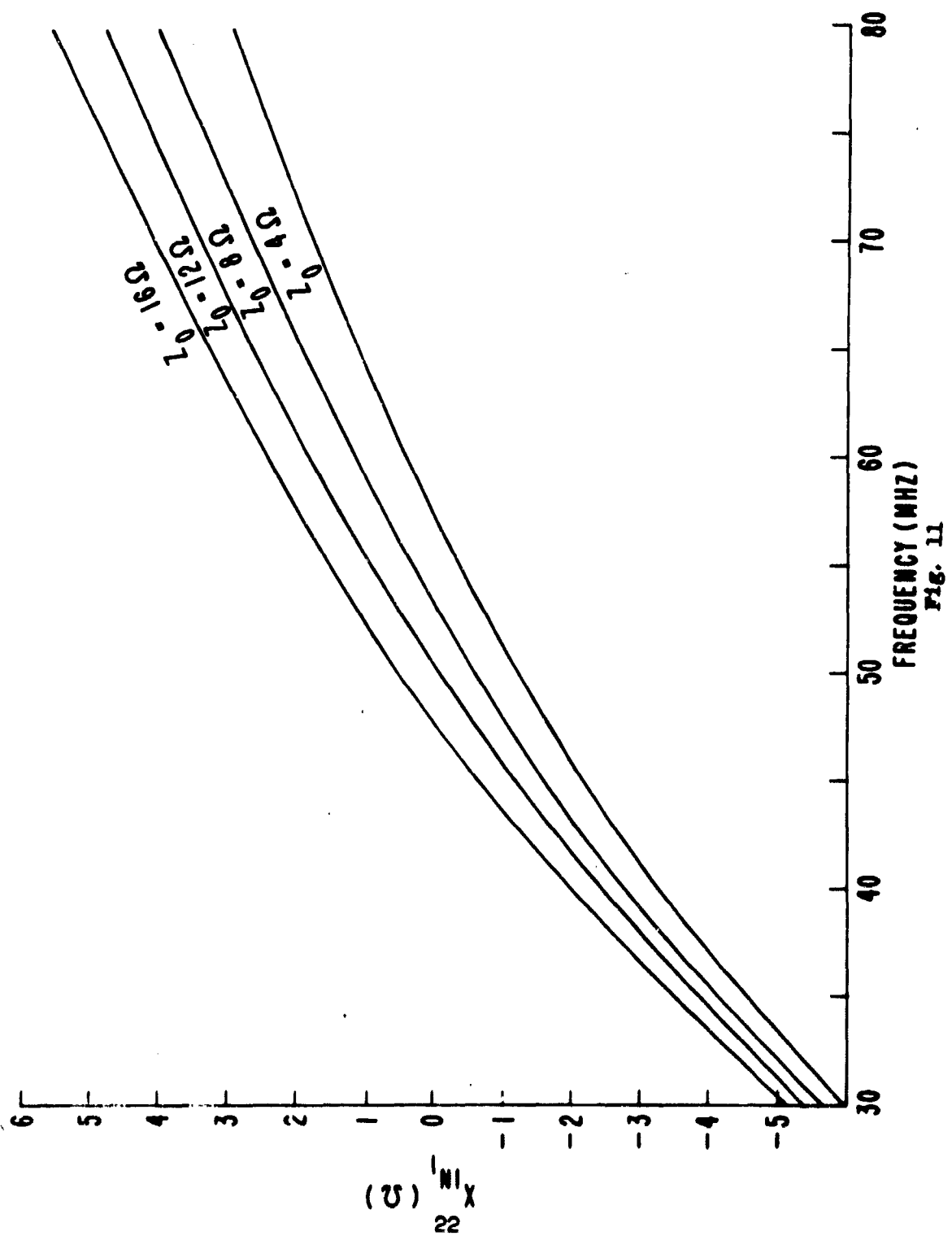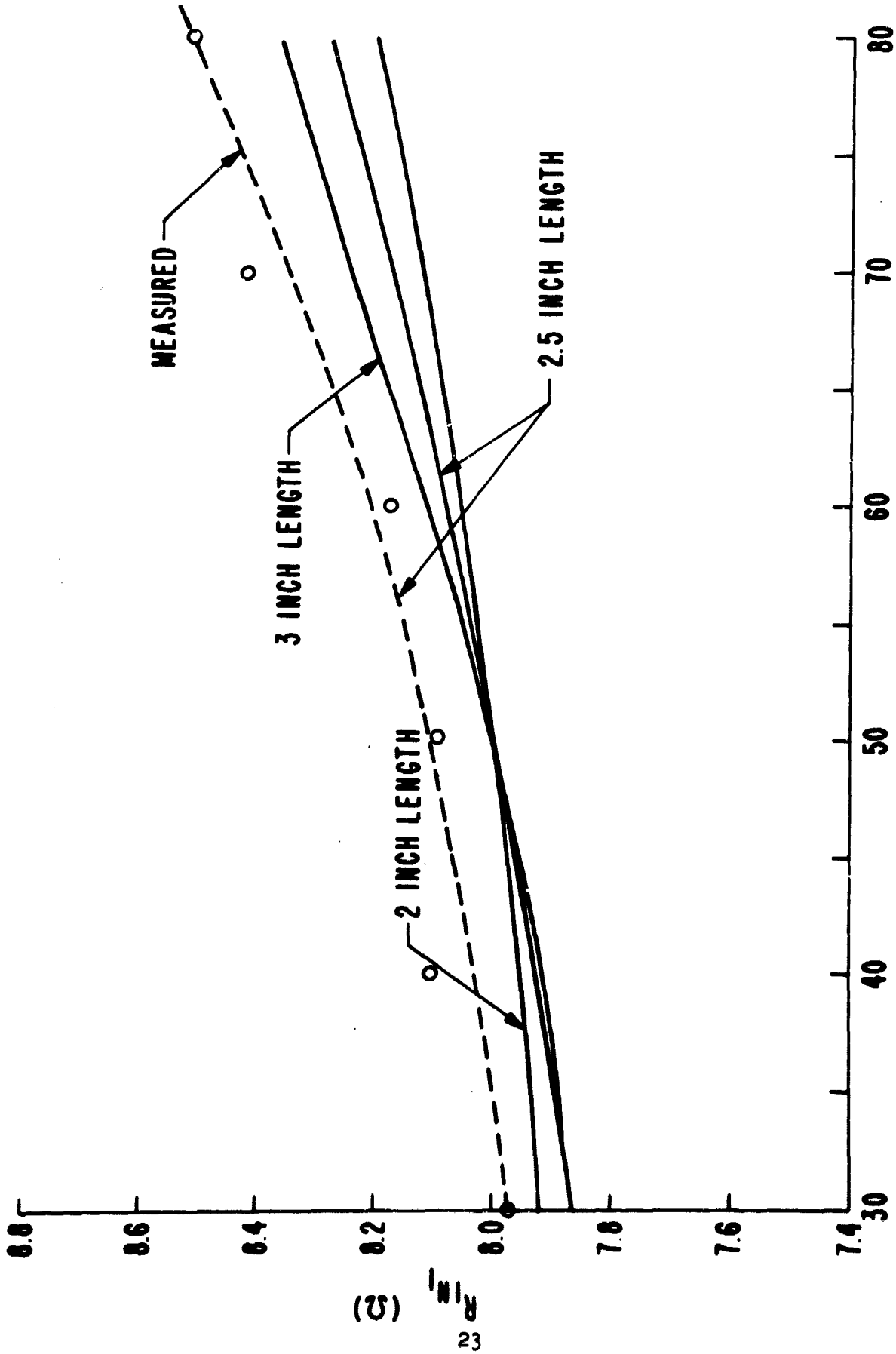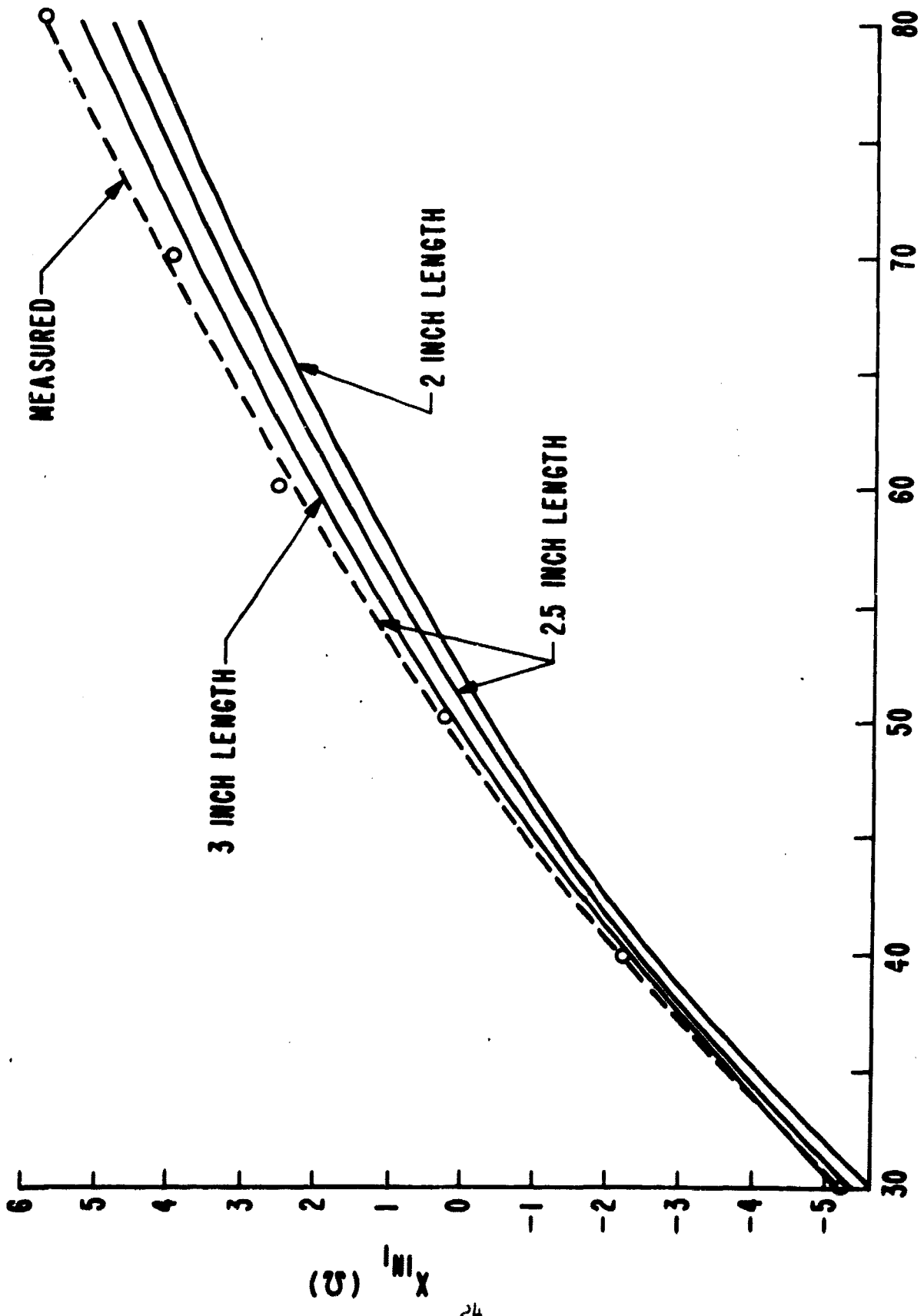
FIG. 12 FREQUENCY (MHZ)
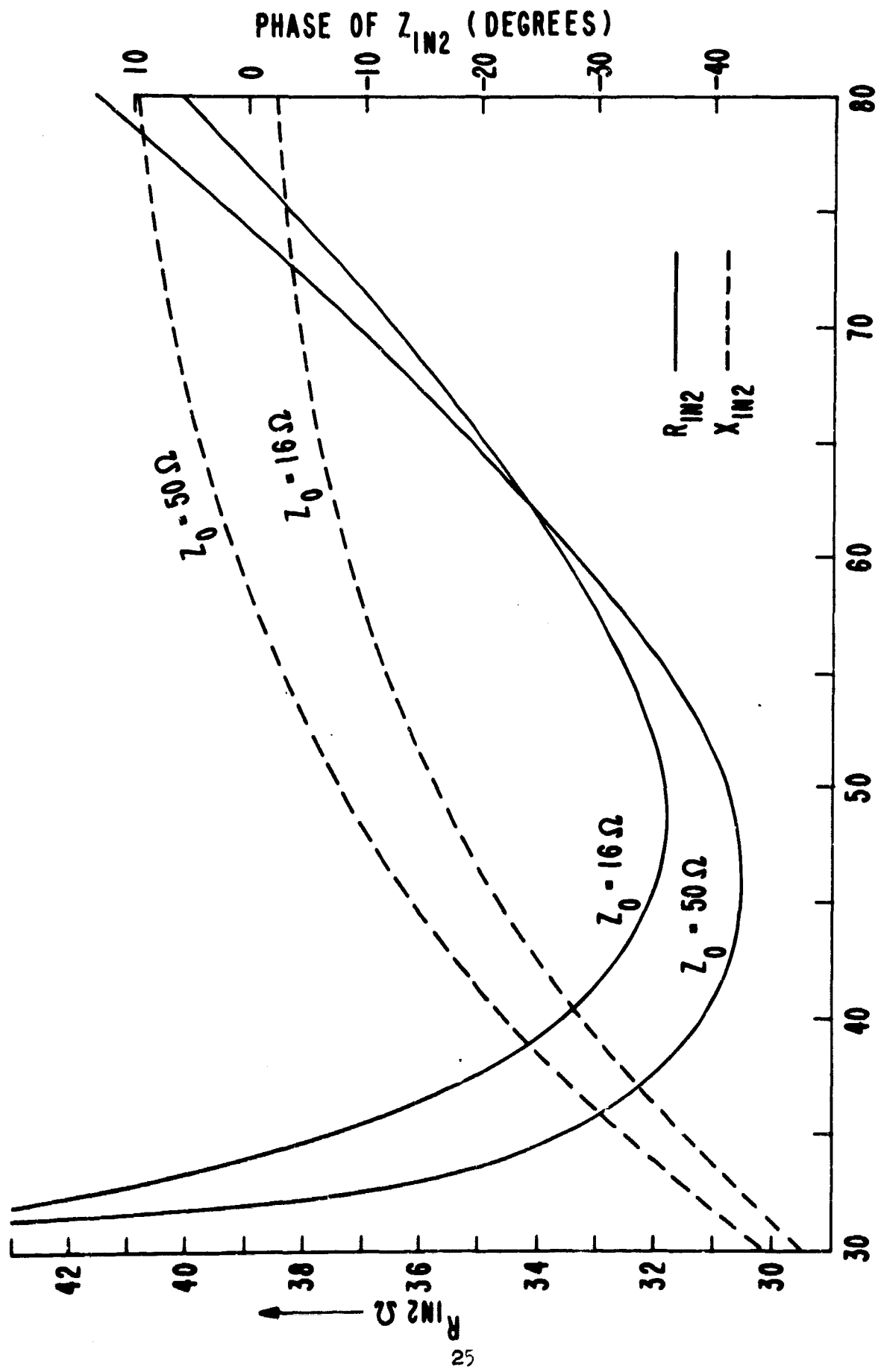
23

FIG. 13 FREQUENCY (MHZ)
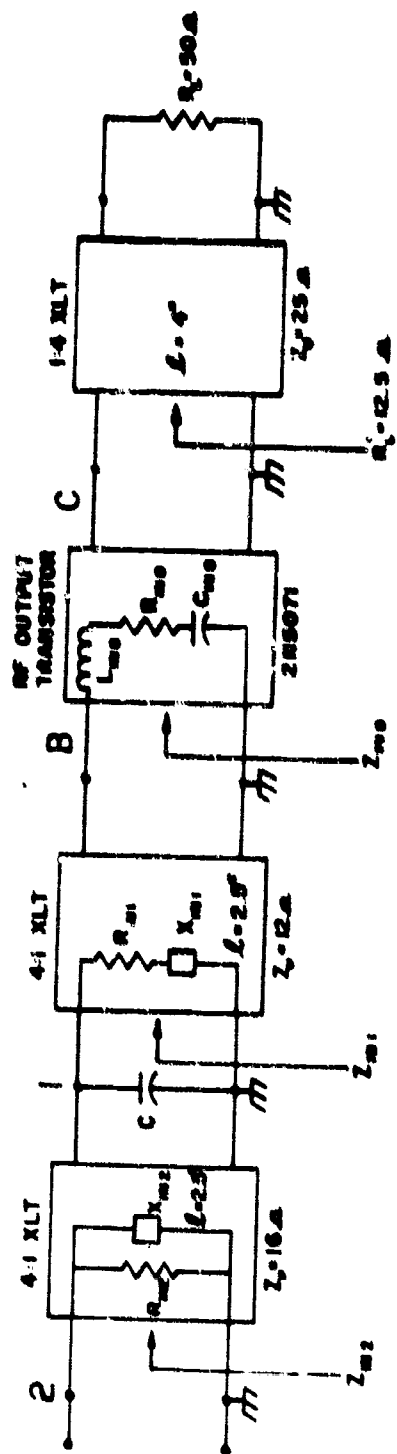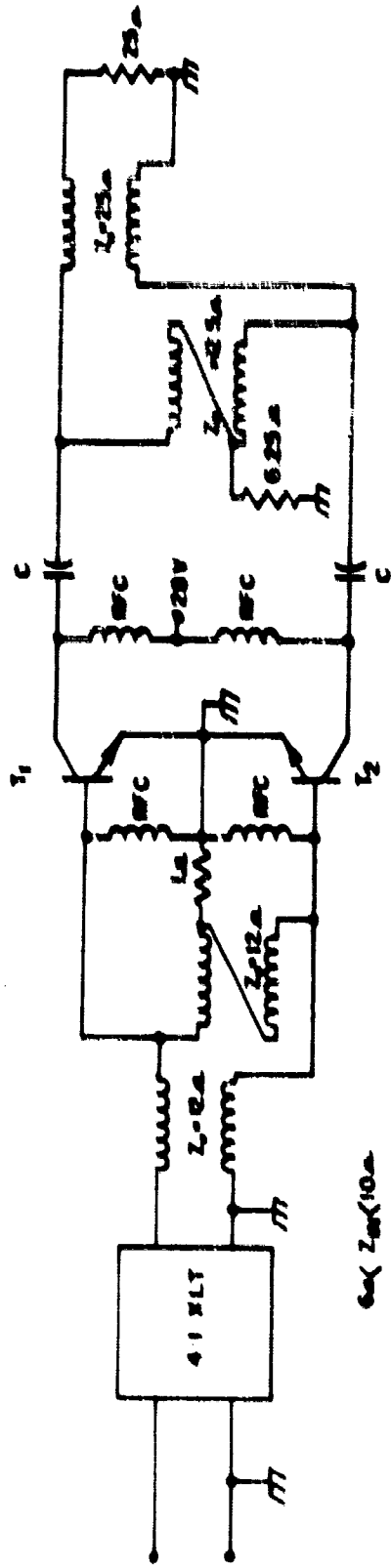
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FIG. 14   FREQUENCY (MHZ)

25

FIG. 15

FIG. 16

27

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1. ORIGINATING ACTIVITY *(Corporate author)*	2a. REPORT SECURITY CLASSIFICATION
U. S. Army Electronics Command Fort Monmouth, N. J.	UNCLASSIFIED
	2b. GROUP

3. REPORT TITLE

Broadband Transformer Design for RF Transistor Power Amplifiers

4. DESCRIPTIVE NOTES *(Type of report and inclusive dates)*

Technical Report

5. AUTHOR(S) *(First name, middle initial, last name)*

Octavius Pitzalis, Jr.
Thomas P. Couse

6. REPORT DATE	7a. TOTAL NO. OF PAGES	7b. NO. OF REFS
July 1968	27	7

8a. CONTRACT OR GRANT NO.	9a. ORIGINATOR'S REPORT NUMBER(S)
b. PROJECT NO. 1H6-22001-A-056	ECOM 2989
c. Task No. -01	9b. OTHER REPORT NO(S) *(Any other numbers that may be assigned this report)*
d. Subtask No. -13	

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13. ABSTRACT

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 W of output power from 30 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 operate a pair of these transistors in push-pull operation is described.

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Broadband transformers						
Impedance matching transformers						
RF transformers						
Untuned VHF transistor power amplifiers						
Broadband hybrids						

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(2)