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ANTENNA LABORATORY

Technical Report No. 64

ANTENNA IMPEDANCE MATCHING BY MEANS OF ACTIVE NETWORKS

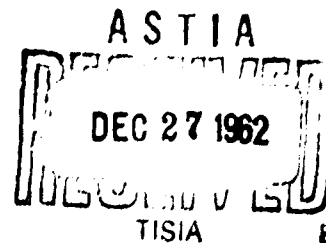
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by
S. Laxpati
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Contract AF33(657)-8460
Hitch Element Nr. 62403454
760D-Project 4028, Task 402824
Aeronautical Systems Division
Project Engineer E. Turner, ASRNCF-1

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ELECTRICAL ENGINEERING RESEARCH LABORATORY
ENGINEERING EXPERIMENT STATION
UNIVERSITY OF ILLINOIS
URBANA, ILLINOIS

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ABSTRACT

The paper introduces several schemes for wideband matching of impedances using active elements in the matching network. It is shown that the design procedure is a straightforward one when two active elements such as negative impedance converters are used. An alternate scheme using one active element is also discussed and it is shown that a RC matching circuit using Kharariwala's synthesis technique usually obtains a rather complicated and sometimes impractical type of network. However, a simple LC matching network may be designed using one active element if certain approximation of the load impedance function is made. Illustrative designs using one and two active elements are described in the paper.

The noise performance of an active matching circuit which has an infinite bandwidth in an ideal sense is compared with a simple active padding network. It is found that there is little relative advantage of one over the other. It is shown, however, that there is a definite advantage of the former circuit over the latter one if one compares their power performance.

Experimental verification of the theoretical designs are included in the paper.

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1. INTRODUCTION

The input impedance of an antenna is a complicated function of frequency and as such cannot be put into an analytic form for all frequencies. However it can be approximately represented over a frequency band by a lumped element lossless network terminated by a resistance R_r as shown in Figure 1. The complexity of this two terminal pair lossless network is dependent on the degree of approximation of the impedance curve of the antenna in the desired frequency range.

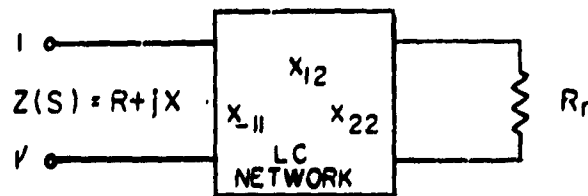


Figure 1. Antenna Equivalent Impedance.

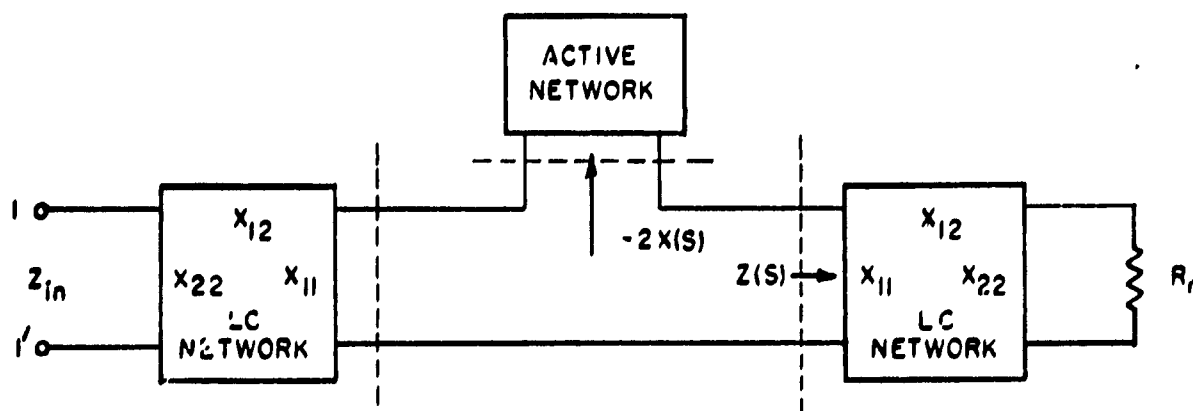
When the antenna is electrically short, it has a high Q and is a highly frequency sensitive device. It can be approximated by a large capacitive reactance in series with a resistance. In order to use these antennas efficiently over a band of frequencies, various impedance matching techniques are employed.

This problem of matching of antennas using passive networks, based on different matching criteria has been treated extensively by various authors^{1, 2}. Fano³, in his paper, has discussed limitations on passive network matching of arbitrary impedance. Not much work has been done on the matching problem using active networks.

This report is a study of the possibility of using active networks for antenna impedance matching.

2. GENERAL SCHEME OF MATCHING

The proposed scheme (Figure 2) is for a general type of matching circuit for an antenna load which admits an equivalent representation of the type shown in Figure 1. It can be easily shown that the impedance Z_{in} , input impedance of the composite network is a pure resistive, and is equal to R_r .



$$Z(S) = R + jX$$

Figure 2. Scheme for Matching Arbitrary Impedance

We can thus represent the equivalent generator load as shown in Figure 3. If the generator impedance R_g is taken as R_r , it is obvious that the power transferred to the load R_r is a maximum and constant.

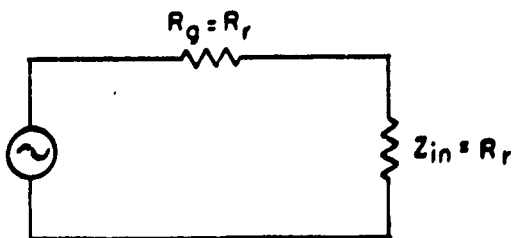


Figure 3. Equivalent Load on the Generator

Since the network in Figure 2 has input impedance R_r , and has all its elements lossless except the load resistance R_r , the power transfer to this load is constant and maximum. This R_r in case of the antenna equivalent network is the radiation resistance of an antenna.

When R_r , the even part of $Z(S)$ (real part of $Z(\omega)$) is constant a simple matching scheme of the type shown in Figure 4 may be used.

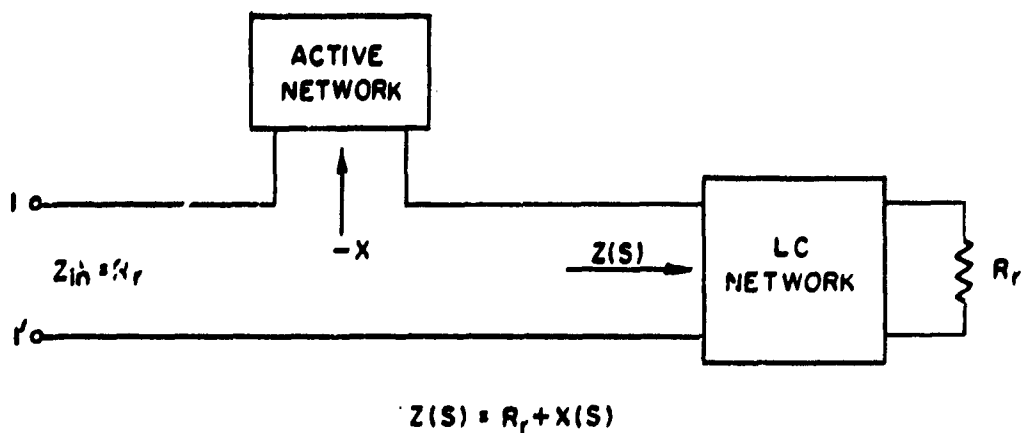


Figure 4. Scheme for Matching Simple Load Impedances

The bandwidth of the system in Figure 2 would be dependent on the bandwidth of the active network and also the frequency band over which the equivalent representation of the antenna impedance is employed. Thus the bandwidth of the system would be lower than either one.

3. DESIGN OF THE SYSTEM

3.1 Determination of Antenna Equivalent Network

The process of determining the lumped element equivalent network representation is largely an empirical one. If the impedance variation as a function of frequency is available, to obtain rational function of the form of quotient of polynomials is a standard process. It will be assumed here that either the quotient of polynomials representation of antenna impedance or a complete equivalent network is available and our discussion will only pertain to the design of a matching network.

3.2 Design of Matching Network $[-2X(S)]$

If the form of the impedance function $Z(S)$ is known, finding $X(S)$ is just a matter of working out the algebra. Then the major problem is to synthesize an active network which will produce an impedance of $-2X(S)$ at its input terminals. Various schemes for doing this are described below.

3.2.1 Design by Means of Two Active Elements

We shall first describe a simple way of synthesizing an impedance function $[-2X(S)]$. Figure 5 shows this method wherein it is required to use two active elements. One of these is a negative impedance converter while the other is a negative resistance. This negative resistance can very well be a negative impedance converter terminated by resistance R_r . The two four terminal lossless networks shown are identical and they are the same as the four terminal lossless networks employed in equivalent antenna impedance representation in Figure 1.

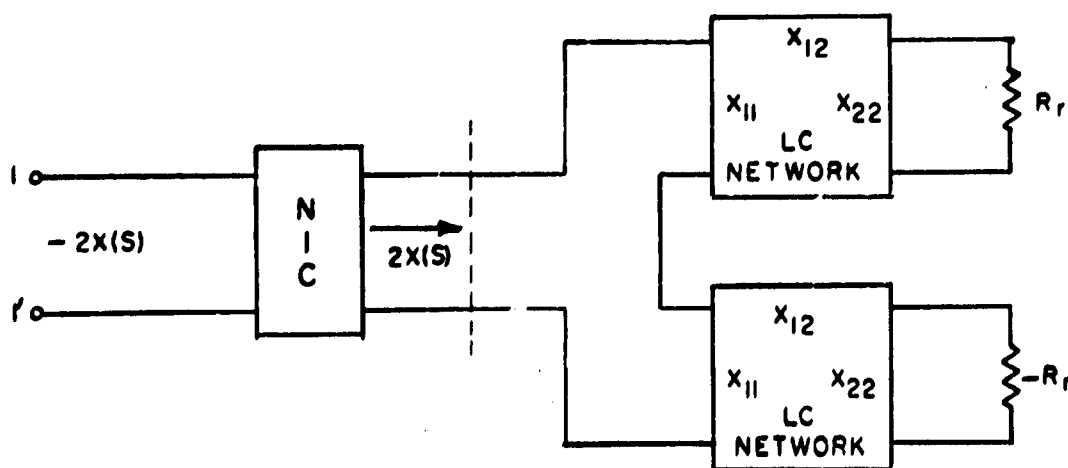


Figure 5. Matching Network Using Two Active Elements

3.2.2 Design by Means of One Active Element.

We shall now study the possibility of designing this network by means of a single active element.

Kinariwala⁴ has shown that "any driving point immittance function can be realized by a transformerless RC structure in which is embedded only one active element. The only restriction on the immittance function is that it is a ratio of two polynomials with real coefficients." A more practical method has been suggested by him in the same paper but with added restrictions on the immittance function. This results in a cascade network as shown in Figure 6. The restriction on the immittance is that in addition to its having real coefficients, the function is positive on some interval of σ -axis ($s = \sigma + j\omega$).

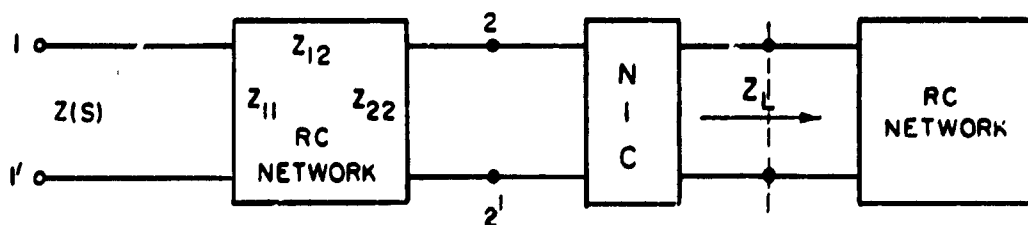


Figure 6. Cascade Matching Network Using One Active Element

Carlin and Youla⁵ have shown in their paper that "any arbitrary real, rational driving point immittance function whose zeros and poles are completely unrestricted as to multiplicity and location in the complex S -plane, may be realized as a lumped network consisting of reciprocal lossless elements and at most one positive and one negative resistor."

Rohrer⁶ has used a method of synthesis similar to that used by Kinariwala, but for LC networks. This network structure is the same as in Figure 6, where the two terminal network and four terminal network are LC networks.

The function $-2X(s)$ that we wish to design has indeed its representation as a quotient of polynomials of real coefficients. Thus by means of any of the above techniques one is able to design this impedance using either only RC or LC elements and one active element. The active element is either a NIC

or a single negative resistor. Use of transformers in synthesis of a four terminal network can be avoided by use of a method suggested by Fialkow and Gerst⁷ in case of RC networks, where as the same can be achieved in case of LC networks by performing an impedance level change similar to loop impedance level change on a Darlington network.

Examples of synthesis using Kinariwala's synthesis method for RC networks and synthesis using LC elements are worked out in the next section of this report.

It should be noted that these methods of synthesis do not always lead to a practically realizable network. This is largely due to the arbitrariness in the choice of certain factors.

4. EXAMPLES

4.1 Antenna Equivalent Impedance $Z(S)$.

Consider an antenna equivalent impedance as shown in Figure 7.

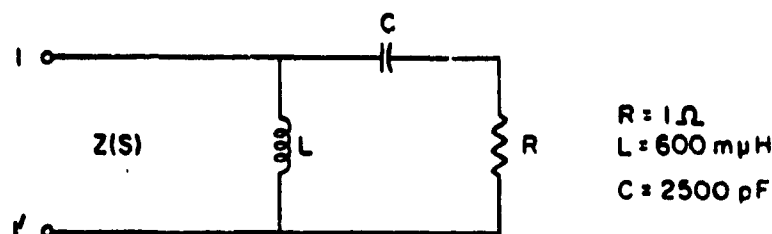


Figure 7. A Simple Equivalent Impedance of an Antenna

Then $Z(S)$ of the network is given by

$$Z(S) = \frac{LCR S^2 + LS}{LC S^2 + CR S + 1}$$

Substituting for R , L , and C and then frequency scaling by 10^8 , magnitude scaling by $\frac{1}{120}$; the impedance $Z(S)$ is

$$Z(S) = \frac{.125 S^2 + .5 S}{15 S^2 + .25 S + 1}$$

hence,

$$-2X(S) = 2(\text{negative of odd part of } Z(S))$$

$$\approx \frac{-14.94 S^3 - S}{225 S^4 + 29.94 S^2 + 1}$$

The complete network for matching is shown in Figure 8, where $Z_1(S) = -2X(S)$ is still undetermined.

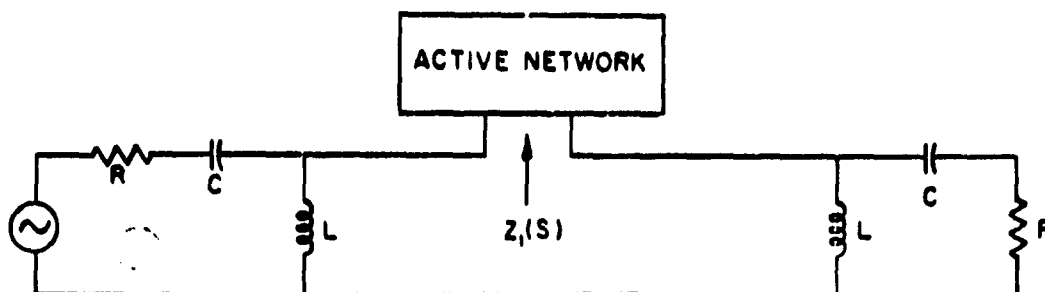


Figure 8. Matching Network for the Impedance of Figure (7)

4.2 Synthesis of $Z_1(s)$ using RC elements

4.2.1 General Procedure

We shall summarize here the synthesis method suggested by Kinarivala⁴ using RC elements. We shall work with admittance function $Y_1(s)$ rather than impedance function $Z_1(s)$ as exemplified in his paper, with relevant modifications in the procedure.

Let us write $Y_1(s) = \frac{1}{Z_1(s)} = \frac{N}{D}$.

Selecting $B = P_2 P_4$ where B has only negative real roots, we can write

$$Y_1(s) = \frac{N}{D} = \frac{\frac{N}{B}}{\frac{D}{B}} = \frac{\frac{P_1}{P_2} - \frac{P_3}{P_4}}{\frac{Q_1}{P_2} - \frac{Q_3}{P_4}} \quad (1)$$

where $N = P_1 - P_3$; $D = Q_1 - Q_3$; and P_1, P_3, Q_1, Q_3 all have negative real roots.

Rearranging Equation (1) we have

$$Y_1(s) = \frac{P_1}{Q_1} \cdot \frac{\frac{P_3}{P_1} - \frac{P_4}{P_2}}{\frac{Q_3}{Q_1} - \frac{P_4}{P_2}} \quad (2)$$

Also for cascade structure of Figure 9 we can write

$$Y_1(s) = Y_{11} \frac{\frac{1}{Z_{22}} - Y_L}{Y_{22} - Y_L} \quad (3)$$

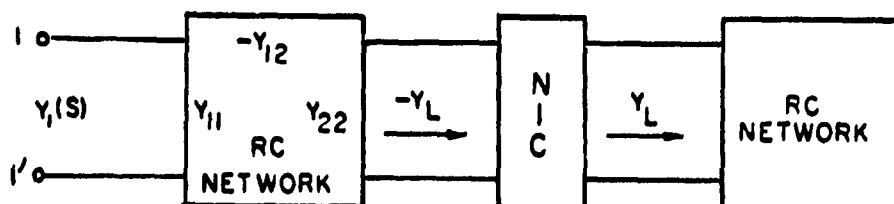


Figure 9. Block Diagram for Cascade Synthesis of $Y_1(s)$

Comparing Equations (2) and (3); we obtain,

$$Y_{11} = \frac{P_1}{Q_1} ; \quad Y_{22} = \frac{Q_3}{Q_1} ; \quad (4)$$

$$\frac{1}{Z_{22}} = \frac{P_3}{P_1} ; \quad Y_L = \frac{P_4}{P_2} ;$$

We can rearrange Equation (3) and also Equation (4) to obtain the following

$$Y_1(s) = Y_{11} - \frac{(-Y_{12})^2}{Y_{22} - Y_L} = \frac{P_1}{Q_1} - \frac{\frac{P_1^2}{Q_1^2}}{\frac{Q_3}{Q_1} - \frac{P_4}{P_2}} \quad (5)$$

Using Equation (1) in above we have

$$Y_1(s) = \frac{N}{D} = \frac{DP_1 + s^2 P_2}{DQ_1}$$

$$\text{i.e. } s^2 P_2 = NQ_1 - DP_1 = P_2 R \text{ (say)} \quad (6)$$

We now give the procedure to obtain the Y parameters of the four terminal network and Y_L ; using the above information from Equations (4) and (6).

- (a) Choose P_1 and Q_1 of degrees equal to the rank of $Y_1(s)$ such that $\frac{P_1}{Q_1}$ is a RC admittance.
- (b) Evaluate $Y_1(s)$ at some point on the negative real axis farther away from the origin than the root of Q_1 farthest away from origin.
- (c) Make $Y_1(s) = Y_{11} = \frac{1}{Q_1}$ at this point by merely multiplying $Y_{11}(s)$ by an appropriate constant.
- (d) Determine $P_2 R$ and find its roots.
- (e) Assign appropriate roots to P_2 from steps (c) and (d).
- (f) Determine R . Find $s^2 = R^2$, $N' = NR$; $D' = DR$.
- (g) Express $\left\{ -\frac{D'}{Q_1 P_2} \right\}$ in partial fraction to obtain $\left\{ \frac{Q_3}{Q_1} - \frac{P_4}{P_2} \right\}$
- (h) All admittances are obtained by noting that $Y_{12} = \frac{R}{Q_1}$.

4.2.2 Computations for the $Z_1(s)$

We have from previous computations in Section 4.1;

$$Y_1(s) = \frac{225 s^4 + 29.94 s^2 + 1}{-14.94 s^3 - s}$$

Let us consider

$$P_1 = K (s+1) (s+3) (s+5) (s+7)$$

$$Q_1 = (s+2) (s+4) (s+6) (s+8)$$

and equating $Y_{11} = \frac{P_1}{Q_1}$ to $Y_{11}(s)$ at $s = -10$; we obtain $K = 61.2375$.

Note that the above choice of roots of P_1 , Q_1 and the point $S = -10$ are the ones that effect the resulting networks. It is essential that these be chosen judiciously.

We now obtain $P_2 R = NQ_1 - DP_1$, which gives after considerable algebra work,

$$P_2 R = 225(S + .0666)(S + .53)(S + 2.481)(S + 4.467)(S + 6.522) \\ (S + 10)(S^2 - .00023 S + .0671)$$

select

$$P_2 = 225(S + 2.481)(S + 4.467)(S + 6.522)(S + 10)$$

$$R = (S + .0666)(S + .53)(S^2 - .00023 S + .0671)$$

We now obtain from above,

$$\left\{ -\frac{D'}{Q_1 P_2} \right\}$$

and then put into its partial fraction expansion which is given by

$$-\frac{D'}{Q_1 P_2} = \frac{Q_3}{Q_1} - \frac{P_4}{P_2} = \frac{1}{225} \left\{ \frac{.34085 S}{S + 2} + \frac{306.49 S}{S + 4} + \frac{3500.13 S}{S + 6} + \frac{1313.62 S}{S + 8} \right\} \\ - \frac{1}{225} \left\{ \frac{3.181 S}{S + 2.481} + \frac{736.58}{S + 4.467} + \frac{4120 S}{S + 6.522} + \frac{223.554 S}{S + 10} \right\}$$

Now we identify

$$Y_{11} = \frac{61.237 (S + 1)(S + 3)(S + 5)(S + 7)}{(S + 2)(S + 4)(S + 6)(S + 8)} \quad (7)$$

$$\begin{aligned}
 -Y_{12} &= \frac{(s + .0686)(s + .53)(s^2 - .00023 s + .0671)}{(s + 2)(s + 4)(s + 6)(s + 8)} \\
 Y_{22} &= \frac{1}{225} \left\{ \frac{.341 s}{s + 2} + \frac{306.5 s}{s + 4} + \frac{3500 s}{s + 6} + \frac{1313.6 s}{s + 8} \right\}
 \end{aligned} \tag{7}$$

and

$$Y_L = \frac{1}{225} \left\{ \frac{3.181 s}{s + 2.48} + \frac{736.5 s}{s + 4.467} + \frac{4120 s}{s + 6.522} + \frac{223.55 s}{s + 10} \right\}$$

We synthesize the four terminal network with the above Y-parameters by Guillemin's method⁸. The two ladders have Y_{11} and Y_{12} parameters as given below, but for a scaling factor.

$$Y_{11\alpha} = Y_{11} = Y_{11\beta};$$

$$-Y_{12\alpha} = \frac{s^4 + .5964 s^3}{(s + 2)(s + 4)(s + 6)(s + 8)}, \quad \& \quad -Y_{12\beta} = \frac{.1022 s^2 + .04 s + .00237}{(s + 2)(s + 4)(s + 6)(s + 8)}$$

The two ladders (α and β) synthesized from the above Y_{11} and Y_{12} parameters, which give exact Y_{11} but Y_{12} within a constant multiplier, after unscaling in frequency and magnitude are shown in Figure 10.

The α and β ladders when connected in parallel has the Y-parameters as given in Equation (8).

$$\begin{aligned}
 Y_{11}' &= \frac{61.2375 (s + 1)(s + 3)(s + 5)(s + 7)}{(s + 2)(s + 4)(s + 6)(s + 8)} \\
 -Y_{12}' &= \frac{(s + .066)(s + .53)(s^2 - .00023 s + .067)}{300.6 (s + 2)(s + 4)(s + 6)(s + 8)} \\
 Y_{22}' &= \frac{.2464 (s^4 + 13.93 s^3 + 57.45 s^2 + 86.1 s + 33.48)}{72 (s + 2)(s + 4)(s + 6)(s + 8)}
 \end{aligned} \tag{8}$$

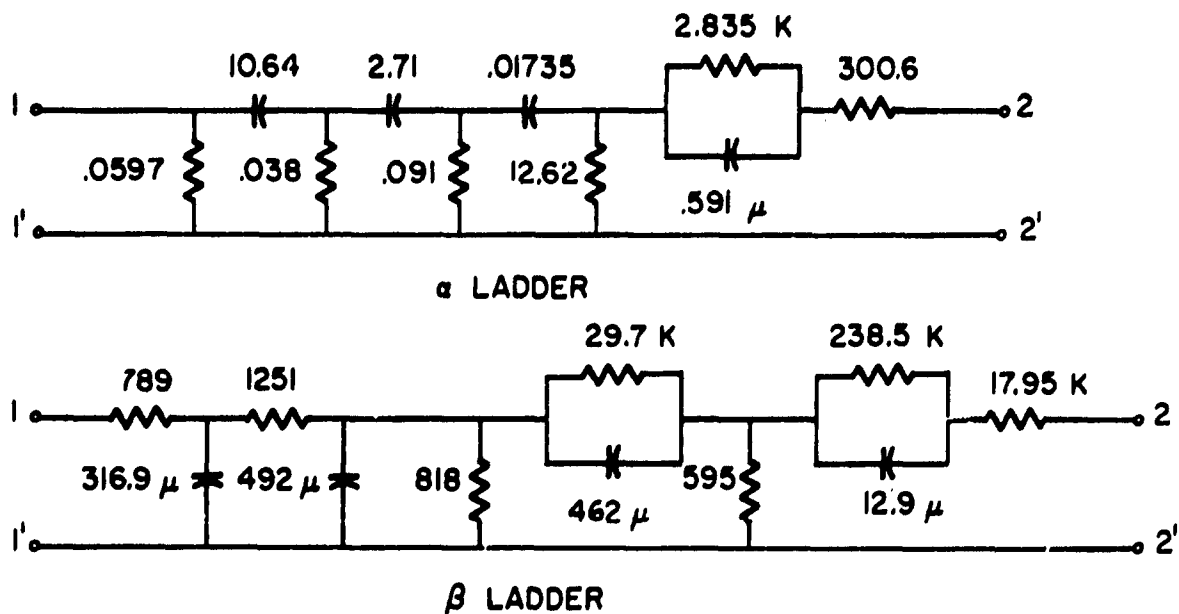


Figure 10. α and β Ladders of the Four Terminal Network

Since we need the Y-parameters of this four terminal network as in Equation (6), we use a method of scaling analogous to that of impedance scaling Darlington network; as shown below.

We have from Equation (5)

$$\begin{aligned}
 Y_1(s) &= Y_{11} - \frac{(-Y_{12})^2}{Y_{22} - Y_L} \\
 &= Y_{11} - \frac{(-k Y_{12})^2}{k^2 (Y_{22} - Y_L)} \\
 &= Y_{11}' - \frac{(-Y_{12}')^2}{(Y_{22}' - Y_L')}
 \end{aligned}$$

where

$$Y_{11}' = Y_{11}, Y_{12}' = k Y_{12}, Y_{22}' - Y_L' = k^2 (Y_{22} - Y_L)$$

Using the above in Equations (7) and (8) we have $k = 1/300.6$; and

$$\begin{aligned}
 Y_{22}' - Y_L' &= \left\{ \frac{1.675 s}{s+2} 10^{-8} + \frac{.1504 s}{s+4} 10^{-4} + \frac{1.72 s}{s+6} 10^{-4} + \frac{.645 s}{s+8} 10^{-4} \right\} \\
 &\quad - \left\{ \frac{1.56 s}{s+2.48} 10^{-7} + \frac{.362 s}{s+4.467} 10^{-4} + \frac{2.025 s}{s+6.522} 10^{-4} + \frac{.11 s}{s+10} 10^{-4} \right\}
 \end{aligned}
 \tag{9}$$

But we have Y_{22}' as in Equation (8), hence we obtain using Equation (9);

$$\begin{aligned}
 Y_L' &= \left\{ \frac{1.726 s}{s+2} 10^{-3} + \frac{.376 s}{s+4} 10^{-3} + \frac{.226 s}{s+6} 10^{-3} + \frac{.227 s}{s+8} 10^{-3} + \frac{1.56 s}{s+2.48} 10^{-7} \right. \\
 &\quad \left. + \frac{.362 s}{s+4.467} 10^{-4} + \frac{2.025 s}{s+6.522} 10^{-4} + \frac{.11 s}{s+10} 10^{-4} \right\}
 \end{aligned}$$

The two terminal network with input admittance Y_L' is shown in Figure 11.

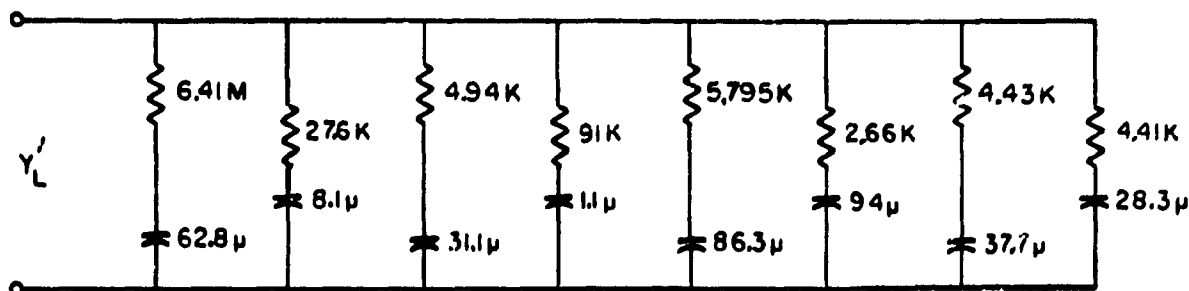


Figure 11. Two Terminal Network with Input Impedance Y_L^1 .

Unscale the two networks of Figures 10 and 11 by 10^8 in frequency and $1/120$ in magnitude to obtain the complete network as shown in Figure 12 to give the required $Y_1(s) = 2X(s)$. This network must be used in turn in complete matching system as shown in Figure 8.

The scaling of each element is achieved using the following relations, where * denotes the scaled elements.

$$R = 120 R^* ; \quad C = \frac{10^{-8}}{120} C^*$$

4.3 Synthesis of $Z_1(s)$ using LC elements

We now synthesize the impedance $Z_1(s)$ as obtained in Section 4.1, by means of LC elements based on the method used by Rohrer⁶.

We have

$$Z_1(s) = \frac{-14.94 s^3 - s}{225 s^4 + 29.94 s^2 + 1}$$

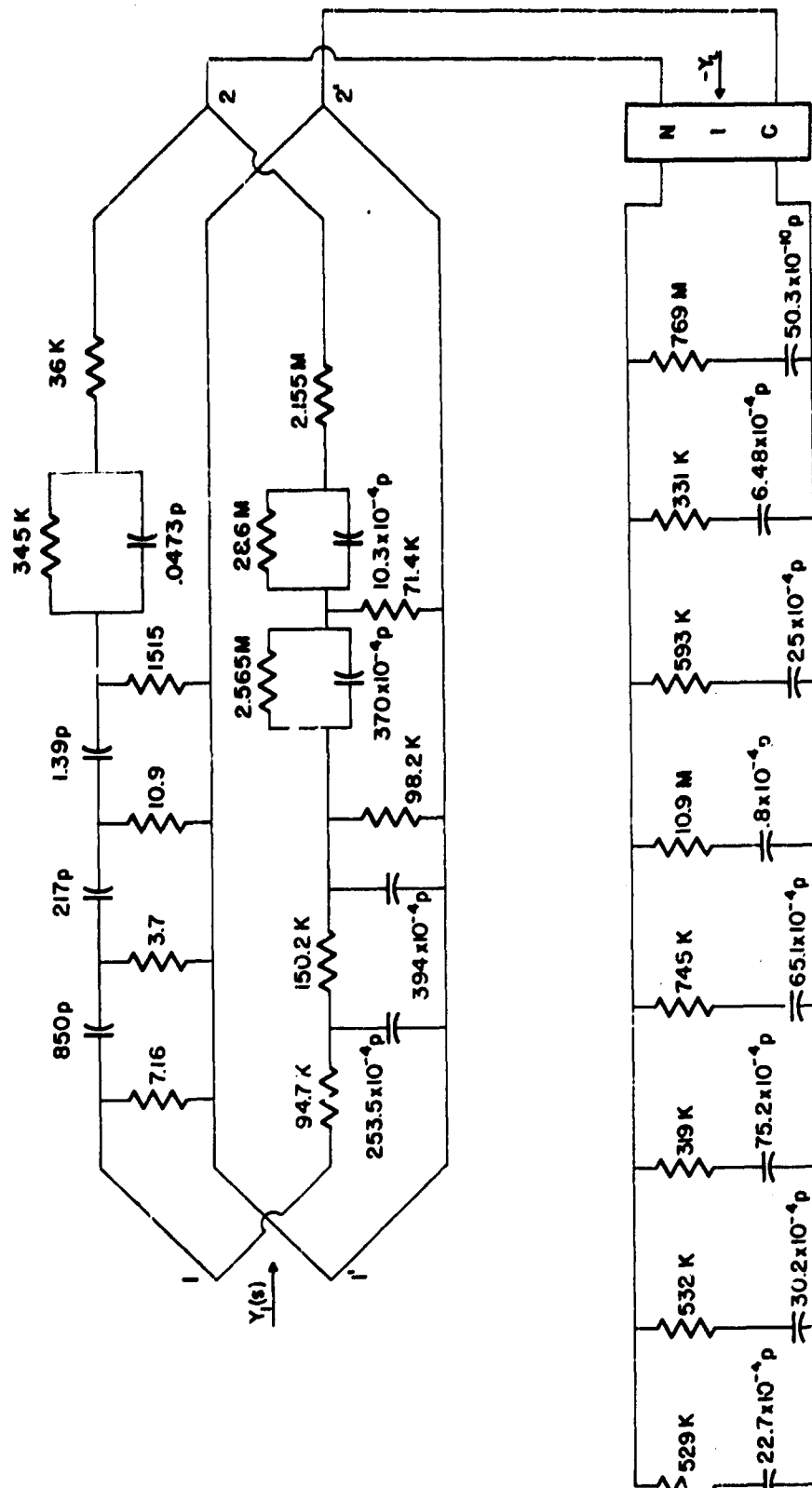


Figure 12. Complete RE matching network for the $Z_1(s)$

which is twice the odd part of impedance $Z(s)$, after frequency scaling by 10^8 and magnitude scaling by $\frac{1}{200}$ where,

$$Z(s) = \frac{.125s^2 + .5s}{15s^2 + .25s + 1}$$

Determine the positive real function $Z'(s)$ associated with $Z_1(s)$ related to it by $Z_1(s) = \text{odd } Z'(s)$, to obtain

$$Z'(s) = \frac{61.052s^2 + .0014s + 4.0865}{15s^2 + .245s + 1} = \frac{m_1 + n_1}{m_2 + n_2}$$

using the condition that $(m_1 m_2 - n_1 n_2)$ is a perfect square, where m is an even function and n an odd function.

Then we have

$$Z_1(s) = \frac{m_1}{n_2} \cdot \frac{\frac{n_1}{m_1} - \frac{n_2}{m_2}}{\frac{m_2}{n_2} - \frac{n_2}{m_2}}$$

$$= Z_{11} \frac{\frac{1}{Y_{22}} - Z_L}{Z_{22} - Z_L}$$

where we realize $Z_1(s)$ as a four-terminal network terminated by a negative impedance Z_L as in RC elements case, but now using LC elements.

Thus we identify

$$Z_{11} = \frac{m_1}{n_2} ; \quad Z_{22} = \frac{m_2}{n_2} ;$$

$$Z_{12} = \frac{\sqrt{m_1 m_2 - n_1 n_2}}{n_2} ; \quad Z_L = \frac{n_2}{m_2}$$

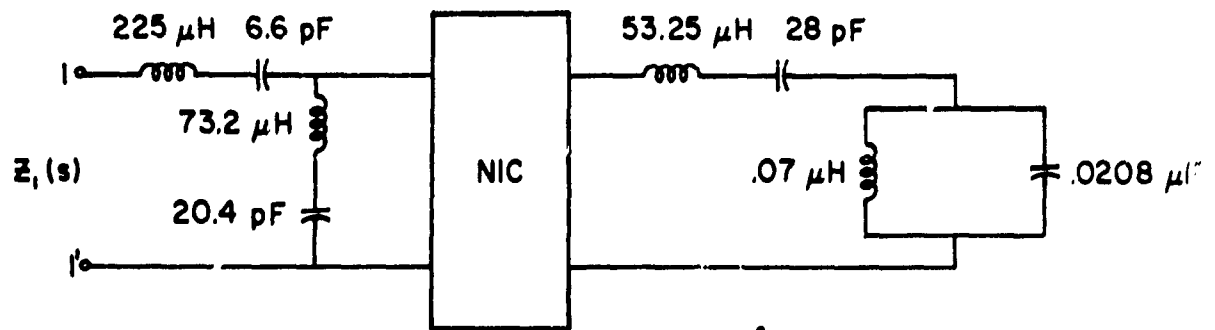


Figure 13. Complete LC Matching Network for the $Z_1(s)$

Hence we have the Z parameters of the cascade structure as follows:

$$\begin{aligned}
 Z_{11} &= \frac{61.052S^2 + 4.0865}{.245S} \\
 Z_{12} &= \frac{2.0215 (14.94S^2 + 1)}{.245S} \\
 Z_{22} &= \frac{15S^2 + 1}{.245S} \\
 Z_L &= \frac{.245 S}{15S^2 + 1}
 \end{aligned} \tag{10}$$

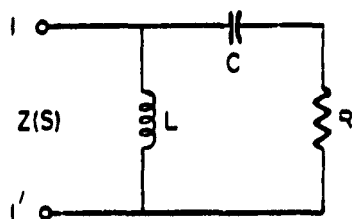
Unscaled cascade network realizing $Z_1(s)$ is shown in Figure 13, where the scaled network has Z parameters of Equation (10).

This network is in turn used in system of Figure 8 to obtain the complete system used for matching the impedance $Z(s)$.

4.4 Simplification of Network by means of Approximation

We find in Sections 4.2 and 4.3 that the networks realizing $Z_1(s)$, which is twice the negative of odd part of $Z(s)$, are very complicated and sometimes lead to impractical element values. We find it possible to simplify this network considerably if we approximate the $Z_1(s)$, as shown below in case of a $Z(s)$ different from that used in Sections 4.1, 4.2, and 4.3.

Consider the antenna equivalent impedance $Z(s)$ as shown in Figure 14.



IMPEDANCE AT 10 Mc/s

$R = 1 \Omega$
 $C = 250 \text{ pF}$
 $L = 10 \mu\text{H}$

$R = 1 \Omega$
 $X_L = 628 \Omega$
 $X_C = 65 \Omega$

Figure 14. A Second Antenna Equivalent Impedance

Hence we have

$$Z(s) = \frac{s^2 + 400s}{s^2 + .01s + 4}$$

After frequency scaling of 10^7 , and

$$X_{eq} = \frac{399.99s^3 + 1600s}{s^4 + 7.9999s^2 + 16} \quad (11)$$

and

$$Z_1(s) = -2X_{eq}(s) = \frac{-799.98s^3 - 3200s}{s^4 + 7.9999s^2 + 16}$$

If the matching network for this $Z_1(s)$ is synthesized on the line of Section 4.3, the network has quite impractical element values. But if we approximate X_{eq} by X'_{eq} as in Equation (12), we find the network a very simple one.

$$X'_{eq} = \frac{400s^3 + 1600s}{s^2 + 8s^2 + 16} = \frac{400s}{s^2 + 4} \quad (12)$$

and

$$Z'_1(s) = -2X'_{eq} = \frac{-800s}{s^2 + 4}$$

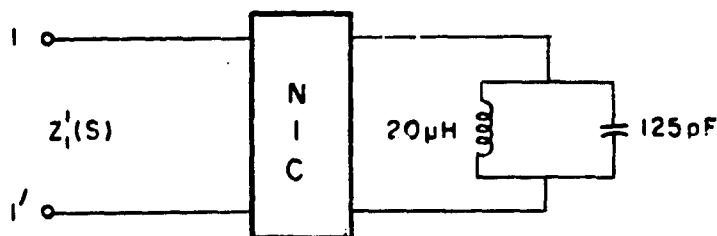
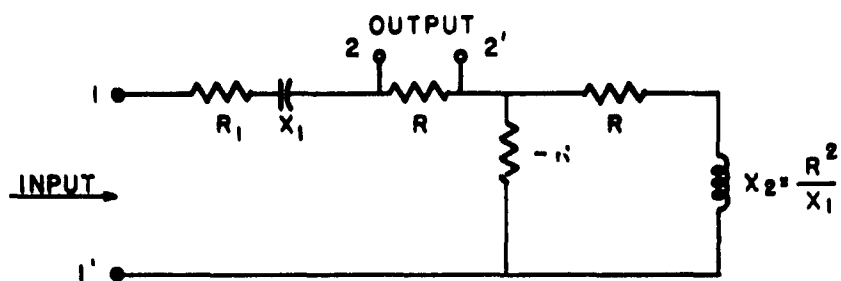


Figure 15. A Simplified LC Matching Network

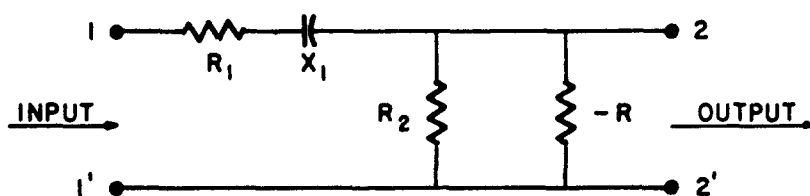
Figure 15 shows the network realizing $Z_1'(s)$. We now check the error involved in this approximation and the limitation on the frequency band over which it is valid. To determine this one must consider the ratio of difference in X_{eq} 's to the R_{eq} of the known impedance $Z(s)$ since this affects the error in the impedance faced by the source. Tabulated in Table I is the factor $\frac{X_1'_{eq} - X_{eq}}{R_{eq}}$ at different frequencies over the range of 10 to 1. It is clear that this ratio is less than 1 over this frequency, and is very small for most of the frequencies. Thus we can conclude that over the frequency range for which the antenna equivalent impedance of the form shown in Figure 14, the matching network is exceedingly simple if an approximation of neglecting R is used.

Table I

Frequency in MC/S	x_{eq}	x'_{eq}	R_{eq}	$\left(\frac{x'_{eq} - x_{eq}}{R_{eq}} \right)$
1	69.706	69.711	0.012	.1814
2	207.63	207.635	0.426	-0.001
3	1679.709	1686.7651	62.915	+0.111
4	-433.881	-433.963	7.434	-0.0115
5	-214.076	-214.091	2.827	-0.005
6	-147.654	-147.662	1.937	-0.004
7	-114.618	-114.653	1.589	-0.003
8	- 94.542	- 94.545	1.412	-0.002
9	- 80.843	- 80.846	1.306	-0.005
10	- 70.837	- 70.839	1.238	+1.688
20	- 32.658	- 32.6582	1.05	-0.0002



(a) NEGATIVE IMPEDANCE INVERTER MATCHING NETWORK.



(b) IMPEDANCE MATCHING BY MEANS OF PADDING RESISTER.

Figure 16. (a) Negative Impedance Inverter Matching Network
(b) Network Matching Impedance by Means of Padding Resistor

where

$$e_{N1} = \text{Thermal noise of resistance } R_1 = \sqrt{4kTR_1\Delta f}$$

$$e_N = \text{Thermal noise of resistance } R = \sqrt{4kTR\Delta f} \quad (13)$$

$$i_n = \text{noise current of tunnel diode} = \sqrt{2e I'_{dc} \Delta f}$$

k is Boltzmann's constant.

The noise figure defined on the basis of short circuit current at the output terminals is

Noise Figure = N.F.

$$= \frac{\text{Sum of the square of short circuit noise current at the output terminals due to all the noise sources}}{\text{Sum of the square of short circuit noise current at the output terminals due to external noise sources alone}}$$

We have in this case a negative impedance inverter and associated matching network as the system under consideration for the noise figure analysis. Hence the noise figure of network of Figure 16(a) is given by

$$\begin{aligned} \text{N.F.} = 1 + & \left\{ \frac{R(R_1 - jX_1 + j\frac{R^2}{X_2} - R)}{R^2 - X_1X_2 + jX_2(R - R_1)} \right\}^2 \left(\frac{e_N}{e_{N1}} \right)^2 + \left\{ \frac{(R_1 - jX_1 + j\frac{R^2}{X_2} - R)}{R} \right\}^2 \left(\frac{e_N}{e_{N1}} \right)^2 \\ & + \left\{ j \frac{R(R + jX_2)(R_1 - jX_1 + j\frac{R^2}{X_2} - R)}{X_2(R_1 - jX_1 + j\frac{R^2}{X_2} - R)} \right\}^2 \left(\frac{i_n}{e_{N1}} \right)^2 \end{aligned}$$

Algebraic simplification using Equation (13) gives

$$\text{N.F.} = 1 + \frac{R}{R_1} + \frac{X_1^2}{R_1 R} + \frac{R^2 + X_1^2}{R_1} \left(\frac{e}{2kT} I'_{dc} \right) \quad (14)$$

5.2.2 Noise Figure of Network of Figure 16(b)

Figure 16(b) shows a matching network using a padding resistor formed by the parallel combination of R_2 and $(-R)$. $(-R)$ is a negative resistance amplifier, and again we consider it to be a tunnel diode.

Shown in Figure 17(b) is the network of Figure 16(b) with all its noise sources present.

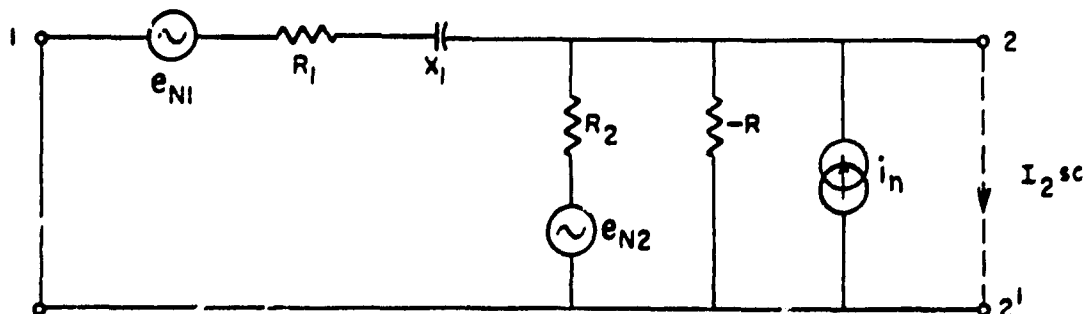


Figure 17(b). Network Matching Impedance by Means of Padding Resistor With All Noise Sources

Where we have

$$\begin{aligned} e_{N1} &= \sqrt{4kTR_1\Delta f} \\ e_{N2} &= \sqrt{4kTR_2\Delta f} \\ i_n &= \sqrt{2eI'_{dc}\Delta f} \end{aligned} \quad (15)$$

Analyzing along the lines of Section 5.2.1 for noise figure on the basis of output short circuit current we obtain, after algebraic simplification using Equation (15),

$$N.F. = 1 + \frac{R_1^2 + X_1^2}{R_1 R_2} + \frac{R_1^2 + X_1^2}{R_1} \left(\frac{e}{2kT} I'_{dc} \right) \quad (16)$$

From Equation's (14) and (16), one can readily verify that under conditions $X_1 \gg R$; and $R_2 \simeq R$ (necessary for large bandwidth), these two expressions are approximately identical. Thus for all practical purposes the noise perfor-

mance can be considered about equal for the two circuits.

5.3 Power Considerations

To compare the performance of the two systems of Figure 16 for power requirements, we determine the power input to the systems for the same power dissipated in the antenna equivalent radiation resistance R_1 . The main power source shown here as in Figure 18 is assumed to have an internal impedance R_g .

Let I be the current flowing through the resistance R_1 . A simple analysis leads to the following results in the two cases under the assumption that the current delivered to the load is the same in each case.

Case I. Negative impedance inverter matching network of Figure 18(a).

$$(a) \text{ Power output to } R_1 = P'_{out} = I^2 R_1$$

$$(b) \text{ Power input by the generator} = P'_{in1}$$

$$= I^2 (R_g + R_1)$$

$$(c) \text{ Power input by the negative resistor } (-R) = P'_{in2}$$

$$= I^2 \left(R + \frac{X_1^2}{R} \right)$$

$$(d) \text{ Total power input } P'_{in} = P'_{in1} + P'_{in2}$$

$$= I^2 \left\{ R_g + R_1 + R + X_1^2/R \right\} \quad (17)$$

Case II. Padding resistor matching network of Figure 18(b)

$$(a) \text{ Power output to } R_1 = P_{out} = I^2 R_1$$

$$(b) \text{ Power input by the generator} = P_{in1}$$

$$= I^2 \left\{ R_g + R_1 + R_e \right\} - j I^2 X_1$$

$$(c) \text{ Power input by the negative resistor } (-R) = P_{in2}$$

$$= I^2 \left(\frac{R_e^2}{R} \right)$$

$$(d) \text{ Total power input} = P_{in} = P_{in1} + P_{in2}$$

$$= I^2 \left\{ R_g + R_1 + R_e \left(\frac{R+R_e}{R} \right) \right\} - j I^2 X_1 \quad (18)$$

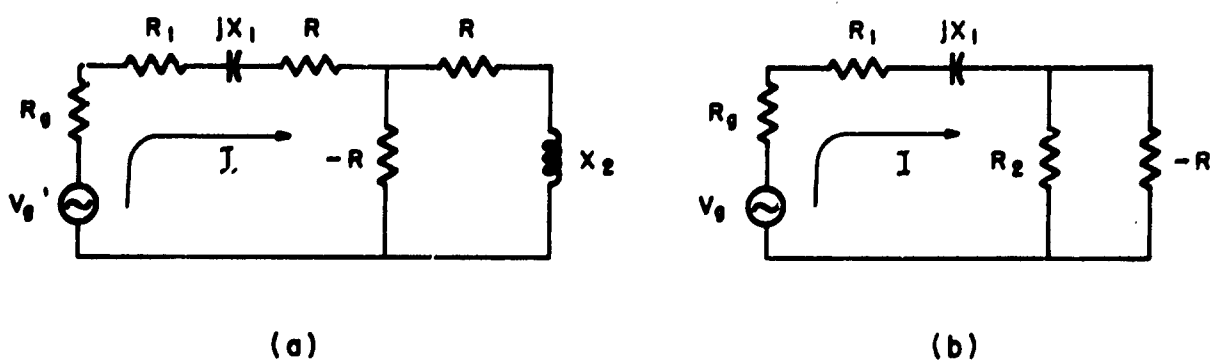


Figure 18. (a) Negative Impedance Inverter Matching Network for Power Considerations
 (b) Padding Resistor Matching Network for Power Considerations

$$\text{where } R_0 = \frac{RR_2}{R-R_2}$$

Comparing the two expressions for the total power input in two cases given by Equations (17) and (18); we readily notice that whereas we need a large reactive power input for Case II, the input power required in Case I is only real power. In Case II R_0 is of the order of X_1 at the lowest frequency of matching. Hence it can be easily verified that the input power is less in Case I. Thus the system of Figure 16(a) offers a definite saving in power requirement.

5. EXPERIMENTAL WORK

The experimental work conducted during this investigation has been concerned primarily with the matching networks for simple antenna equivalent circuits, and using negative impedance converters as active elements. The N.I.C.'s as a positive feedback network has, as a rule, a small useful frequency band. Thus a number of NIC's were built to function in different frequency bands from 1 KC/S to about 10Mc/S. NIC's using tubes were built to function in the lower frequency range of up to 10 KC/S and successfully tested. Later on, transistors have been used all throughout in this investigation.

The basic design of these NIC's was based on the circuits given by Linvill⁹ and Bonner¹⁰ in their papers. Figures 19, 20, 21 show three different types of NIC's using transistors. These NIC's were used in matching simple equivalent impedances of antenna. The results of these are shown in Figures 22, 23, and 24. In each figure is shown the signal across the antenna equivalent resistance with and without the matching networks compensation.

It is observed that there is a limitation set up at high frequencies, and this is the limitation in the performance of NIC at these frequencies due to the appreciable phase shift introduced by the transistors used. The low frequency limitation observed is due to the large ratio of reactive to real impedance of the antenna equivalent impedance. This ratio is about 25 to 35 at the lowest frequency.

Figures 25, 26 show the results obtained using two NIC compensations.

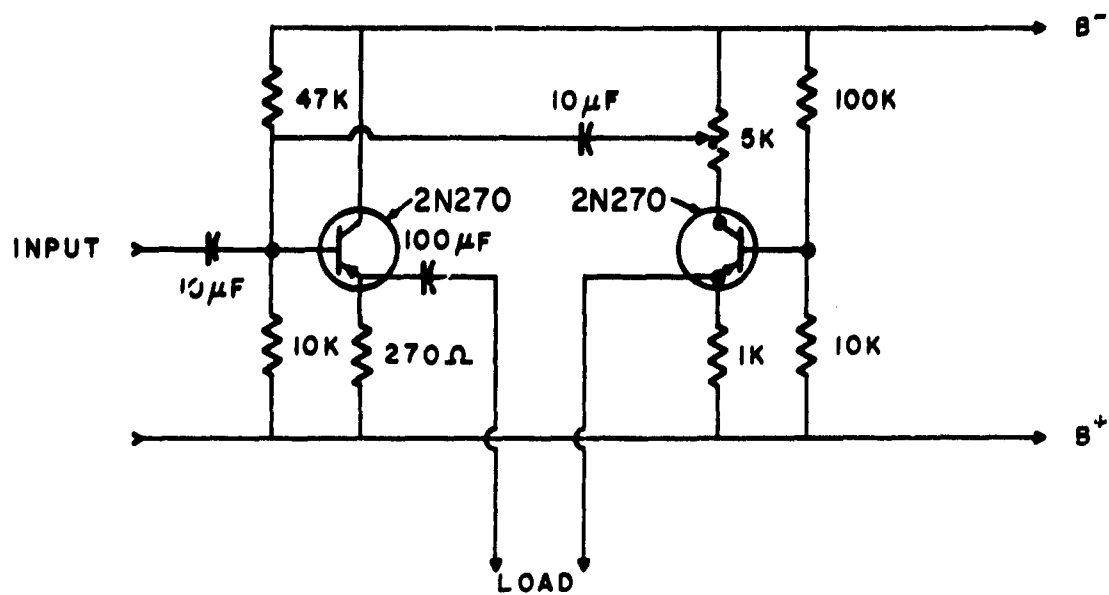


Figure 19. Negative Impedance Converter Number 1.

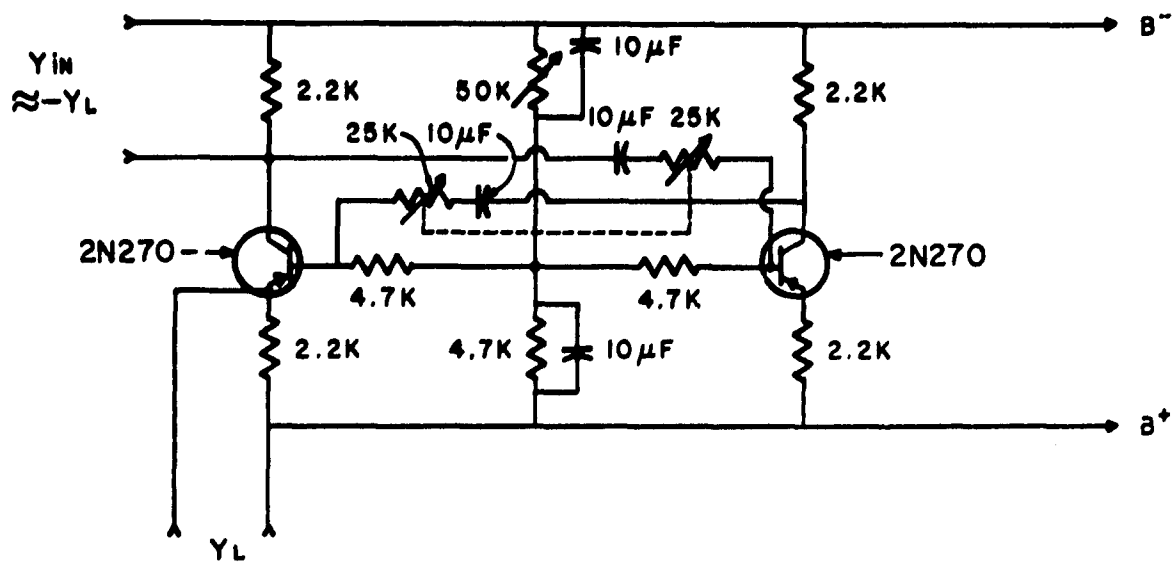


Figure 20. Negative Impedance Converter Number 2.

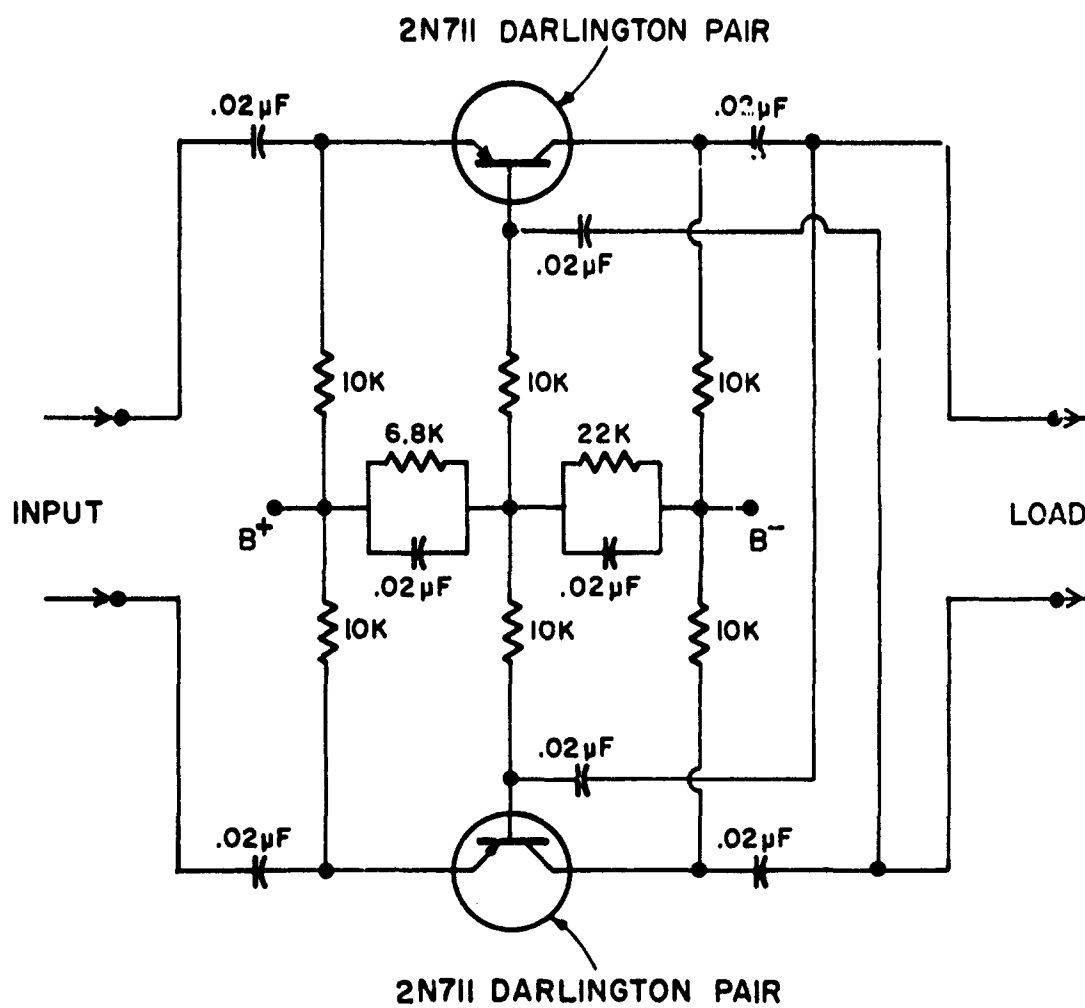


Figure 21. Negative Impedance Converter Number 3.

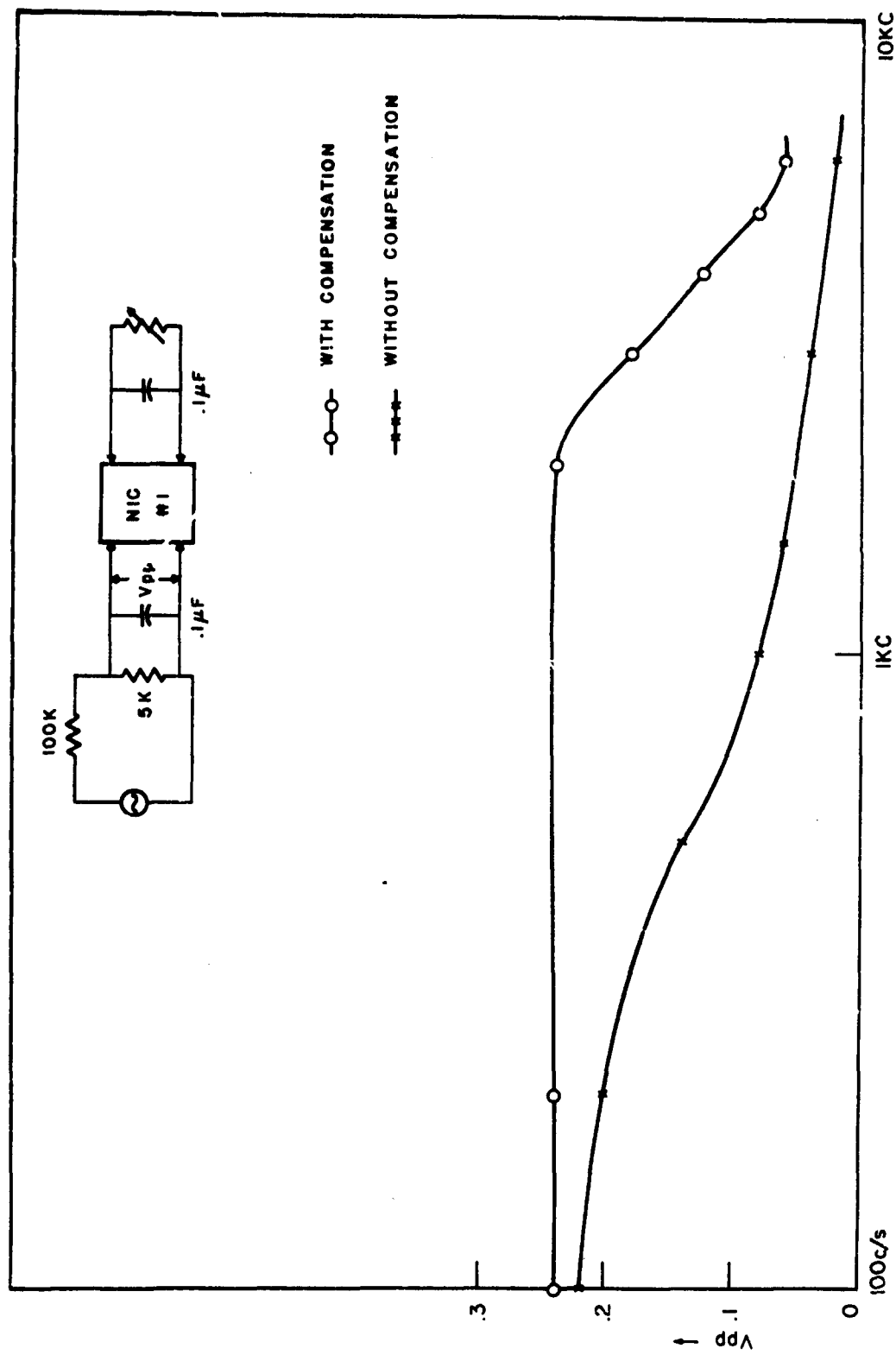


Figure 22. Compensation Curve Using NIC Number 1 for Shunt Compensation

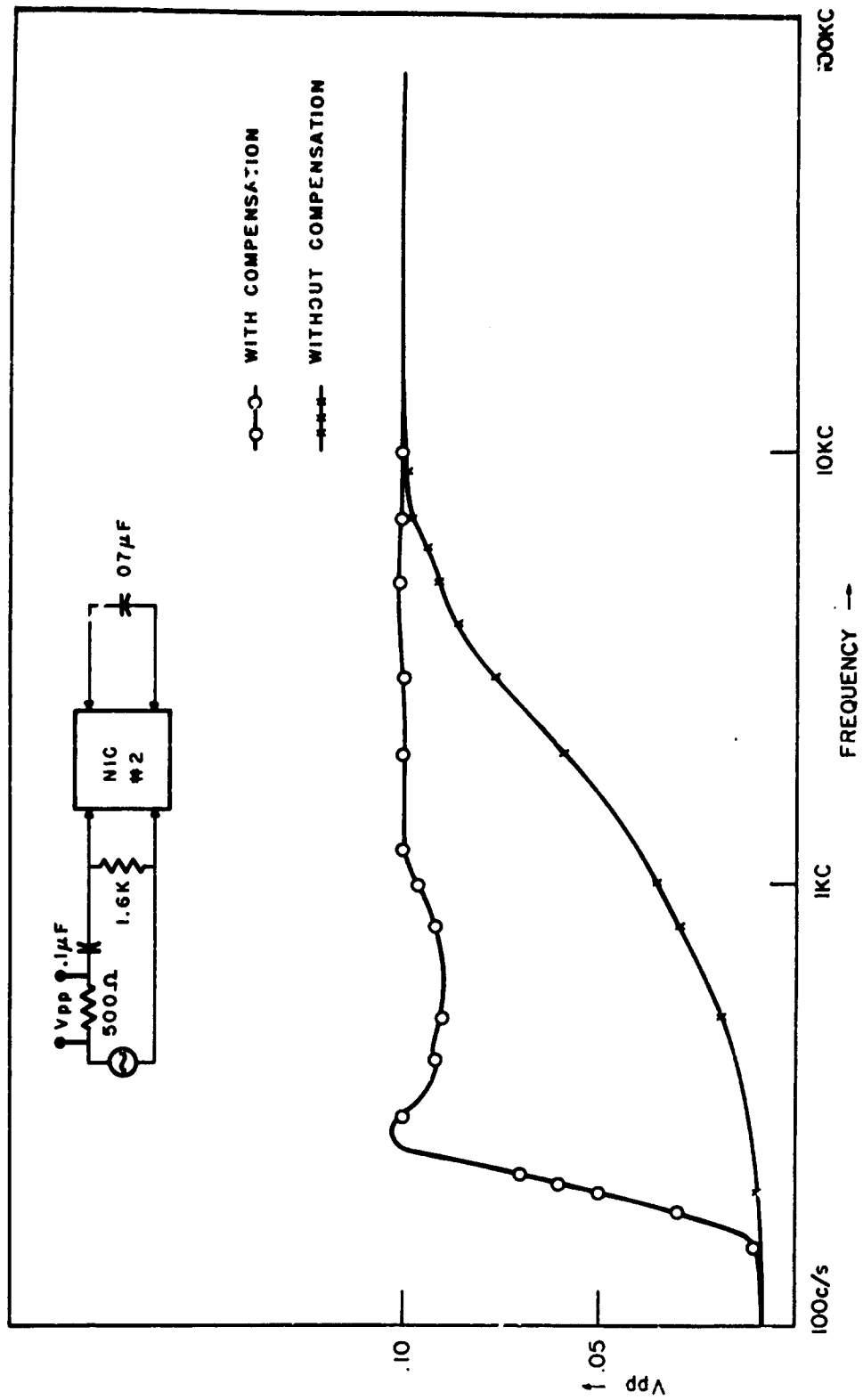


Figure 23. Compensation Curve Using NIC Number 2 for Series Compensation

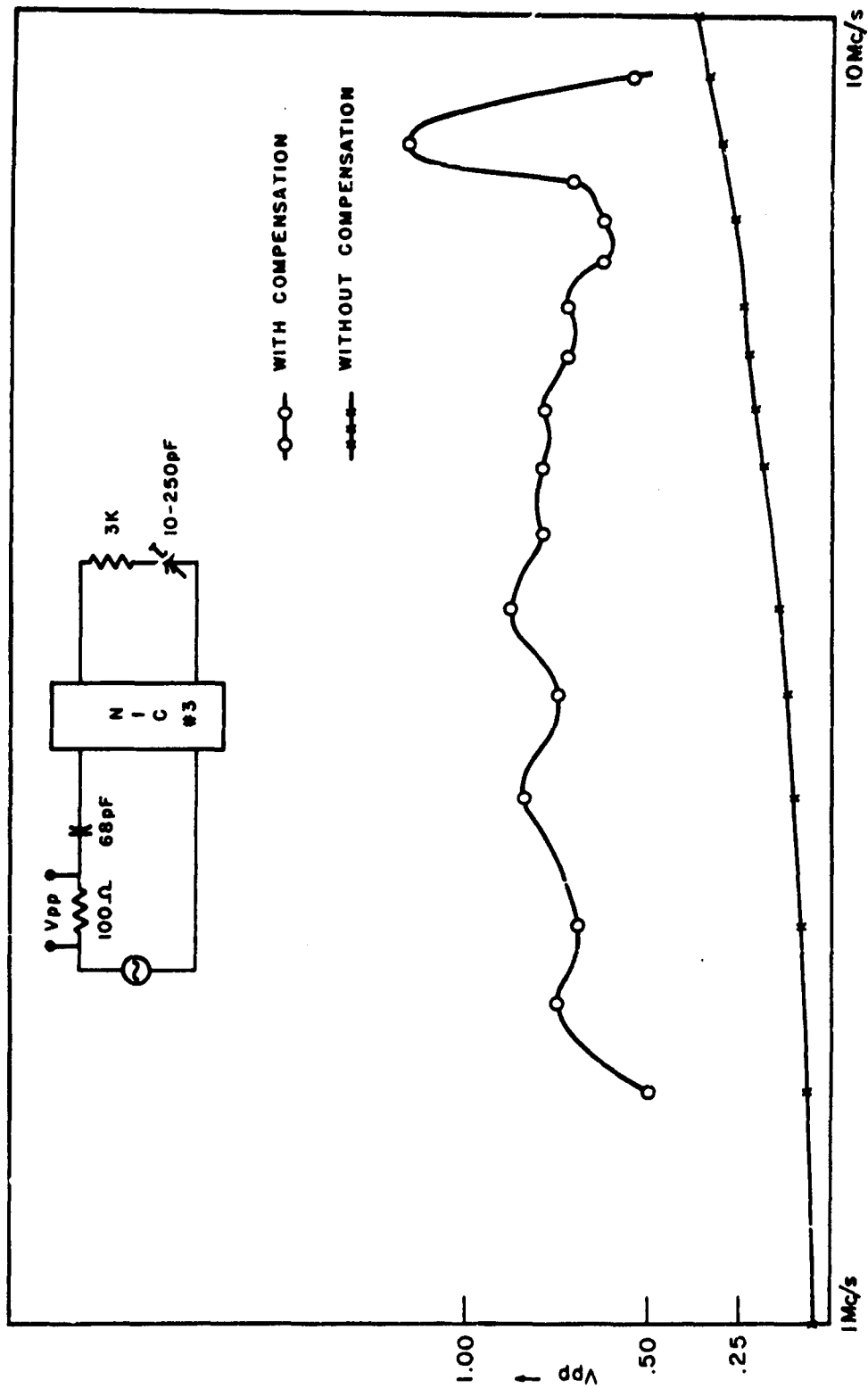


Figure 24. Compensation Curve Using NIC Number 3
for Series Compensation

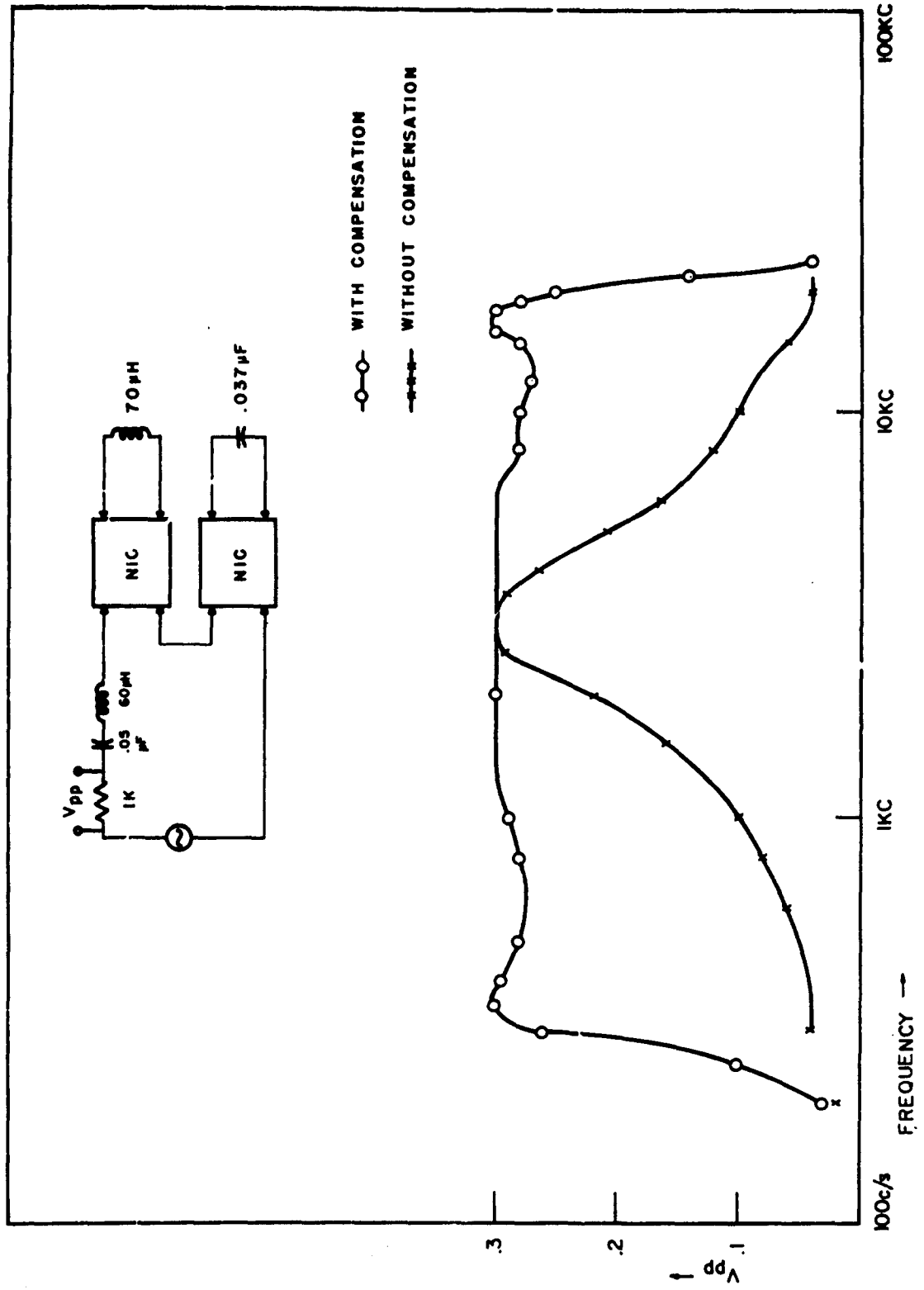


Figure 25. Compensation Curve for Series Resonance Equivalent Impedance

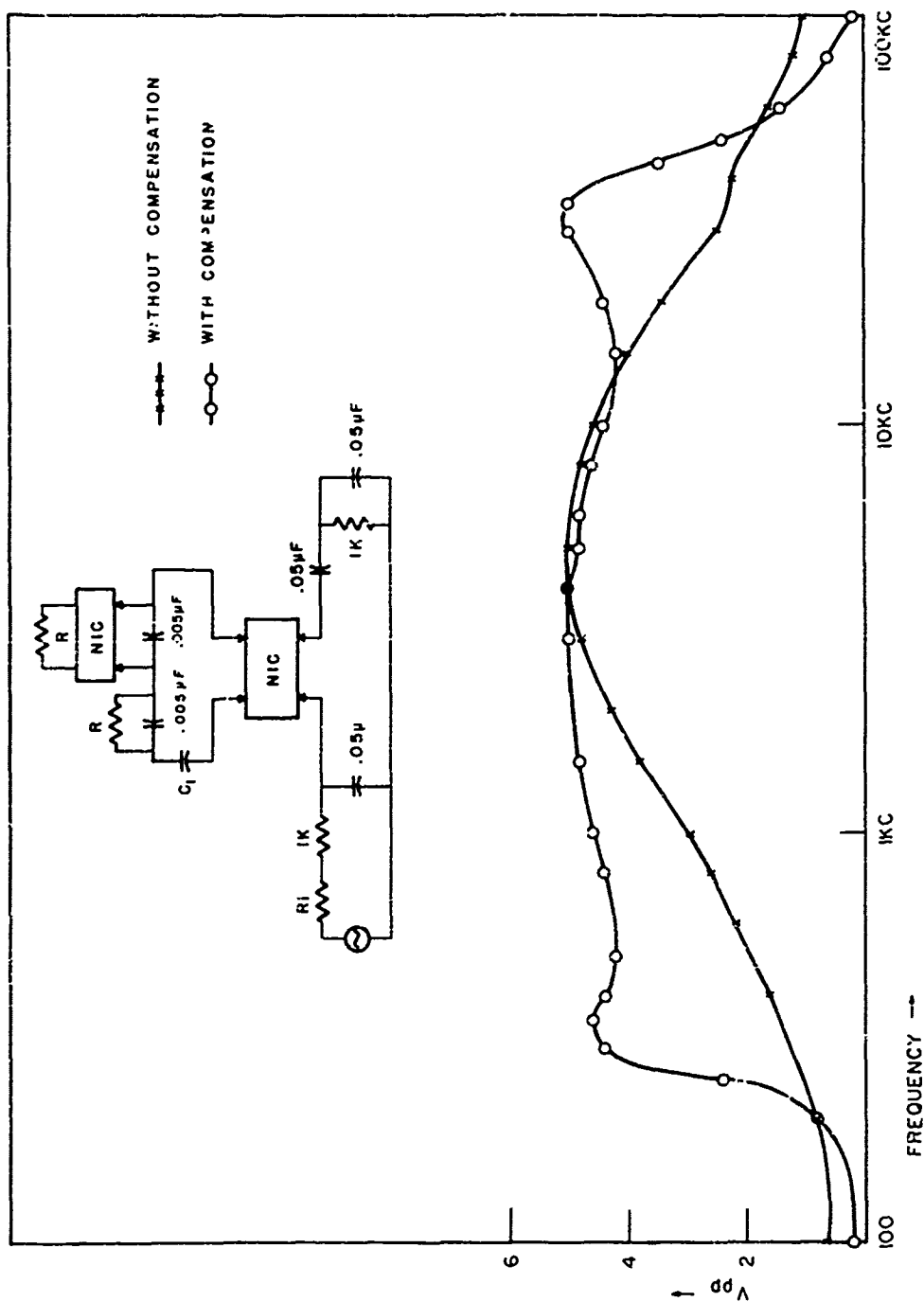


Figure 26. Compensation Using Two NIC s.

7. CONCLUSIONS

(1) Active networks consisting of negative impedance converters or negative resistances may be successfully used to design wideband matching circuits for arbitrary load impedances. The design procedure using two active elements is very straightforward. An R-C matching circuit using one active element only may be designed using a procedure outlined by Kinariwala, but the method does not always lead to practical values of circuit components. L-C circuits using one active element may also be designed to approximately match a given load. One usually pays the price for reduced number of active elements in terms of increased complexity of design.

(2) It is difficult to form a general conclusion in regard to the comparative advantages of active matching networks, over a simple active padding network from the point of view of the noise performances because there are various factors including the nature of the load impedance which determines this behavior. For a particular case it was found that there is little relative advantage of the active matching circuit over the padding network, insofar as the noise performance is concerned. However, for the same case it was found that from the point power performance the active circuit had a definite advantage.

(3) Experimental studies demonstrate the feasibility of building the theoretically designed circuits and show that active matching circuits perform satisfactorily in the design frequency band limited only by the bandwidth of the active elements and their signal handling capacity.

The general conclusion is that wideband matching may be accomplished through the use of active elements. However, the complexity of design, the noise and power performance etc., depend strongly on the nature of the load impedance to be matched and the relative bandwidth desired. A study of some of these aspects has not yet been completed and a continuation of the study along this line is recommended.

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Material Laboratory, Code 932
New York Naval Shipyard
Brooklyn 1, New York

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Diamond Ordnance Fuse Laboratories
Attn: 240
Washington 25, D. C.

Director
U. S. Navy Electronics Laboratory
Attn: Library
San Diego 52, California

Adams-Russell Company
200 Sixth Street
Attn: Library (Antenna Section)
Cambridge, Massachusetts

Aero Gen Astic
Attn: Security Officer
1200 Duke Street
Alexandria, Virginia

NASA Goddard Space Flight Center
Attn: Antenna Section, Code 523
Greenbelt, Maryland

Airborne Instruments Labs., Inc.
Attn: Librarian (Antenna Section)
Walt Whitman Road
Melville, L. I., New York

American Electronic Labs
Box 552 (Antenna Section)
Lansdale, Pennsylvania

Andrew Alfred Consulting Engineers
Attn: Librarian (Antenna Section)
299 Atlantic Ave.
Boston 10, Massachusetts

Amphel-Borg Electronic Corporation
Attn: Librarian (Antenna Section)
2801 S. 25th Avenue
Broadview, Illinois

Bell Aircraft Corporation
Attn: Technical Library
(Antenna Section)
Buffalo 5, New York

Bendix Radio Division of
Bendix Aviation Corporation
Attn: Technical Library
(For Dept. 162-1)
Baltimore 1, Maryland

Boeing Airplane Company
Aero Space Division
Attn: Technical Library
W/F Antenna & Radomes Unit
Seattle, Washington

Boeing Airplane Company
Attn: Technical Library
W/F Antenna Systems Staff Unit
Wichita, Kansas

Chance Vought Aircraft Inc.
TDRU: ECAER Representative
Attn: Technical Library
W/F Antenna Section
P. O. Box 5907
Dallas 22, Texas

Collins Radio Company
Attn: Technical Library (Antenna
Section)
Dallas, Texas

Convair
Ft. Worth Division
Attn: Technical Library (Antenna
Section)
Grants Lane
Fort Worth, Texas

Convair
Attn: Technical Library (Antenna
Section)
P. O. Box 1050
San Diego 12, California

Dalmo Victor Company
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Section)
1515 Industrial Way
Belmont, California

Dorne & Margolin, Inc.
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30 Sylvester Street
Westbury, L. I., New York

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Attn: Technical Library (Antenna
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Orlando, Florida

Electronic Communications, Inc.
Research Division
Attn: Technical Library
1830 York Road
Timonium, Maryland

Fairchild Engine & Airplane Corporation
Fairchild Aircraft & Missiles Division
Attn: Technical Library (Antenna
Section)
Hagerstown 10, Maryland

Georgia Institute of Technology
Engineering Experiment Station
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M/F Electronics Division
Atlanta 13, Georgia

General Electric Company
Electronics Laboratory
Attn: Technical Library
Electronics Park
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General Electronic Labs., Inc.
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Section)
18 Ames Street
Cambridge 42, Massachusetts

General Precision Lab., Division of
General Precision Inc.
Attn: Technical Library (Antenna
Section)
63 Bedford Road
Pleasantville, New York

Goodyear Aircraft Corporation
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M/F Dept. 474
1210 Massillon Road
Akron 15, Ohio

Granger Associates
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Johns Hopkins University
Applied Physics Laboratory
8621 Georgia Avenue
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Hughes Aircraft Corporation
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Section)

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ITT Laboratories
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Section)
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U. S. Naval Ordnance Lab.
Attn: Technical Library
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Lincoln Laboratories
Massachusetts Institute of Technology
Attn: Document Room
P. O. Box 73
Lexington 73, Massachusetts

Litton Industries
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4900 Calvert Road
College Park, Maryland

Lockheed Missile & Space Division
Attn: Technical Library (M/F Dept-
58-4th, Plant 1, Bldg. 130)
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The Martin Company
Attn: Technical Library (Antenna
Section)
P. O. Box 179
Denver 1, Colorado

The Martin Company
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Section) Mail No. T-3C
Baltimore 3, Maryland

The Martin Company
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Microwave Laboratory)
Box 5837
Orlando, Florida

W. L. Maxson Corporation
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460 West 34th Street
New York 1, New York

McDonnell Aircraft Corporation
Attn: Technical Library (Antenna
Section)
Box 516
St. Louis 66, Missouri

Melpar, Inc.
Attn: Technical Library (Antenna
Section)
3000 Arlington Blvd.
Falls Church, Virginia

University of Michigan
Radiation Laboratory
Willow Run
201 Catherine Street
Ann Arbor, Michigan

Mitre Corporation
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ronic Warfare Dept. D-21)
Middlesex Turnpike
Bedford, Massachusetts

North American Aviation Inc.
Attn: Technical Library (M/F
Engineering Dept.)
4300 E. Fifth Avenue
Columbus 16, Ohio

North American Aviation Inc.
Attn: Technical Library
(M/F Dept. 56)
International Airport
Los Angeles, California

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NORAIR Division
1001 East Broadway
Attn: Technical Information (3924-3)
Hawthorne, California

Ohio State University Research
Foundation
Attn: Technical Library
(M/F Antenna Laboratory)
1314 Kinnear Road
Columbus 12, Ohio

Philco Corporation
Government & Industrial Division
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(M/F Antenna Section)
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Philadelphia 44, Pennsylvania

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Air Arms Division
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P. O. Box 746
Baltimore 3, Maryland

Wheeler Laboratories
Attn: Librarian (Antenna Lab)
Box 561
Smithtown, New York

Electrical Engineering Research
Laboratory
University of Texas
Box 8026, Univ. Station
Austin, Texas

University of Michigan Research
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Attn: Dr. J. A. M. Lyons
Ann Arbor, Michigan

Radio Corporation of America
RCA Laboratories Division
Attn: Technical Library
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Princeton, New Jersey

Radiation, Inc.
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Antenna Section
Drawer 37
Melbourne, Florida

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1700 Main Street
Santa Monica, California

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23999 Ventura Blvd.
Calabasas, California

Raytheon Corporation
Equipment Division
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P. O. Box 520
Waltham 54, Massachusetts

Republic Aviation Corporation
Applied Research & Development
Division
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95 Canal Street
Nashua, New Hampshire

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State College, Pennsylvania

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P. O. Box 1828
Clearwater, Florida

Sperry Gyroscope Company
Attn: Librarian (Antenna Lab)
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Stanford, California

Stanford Research Institute
Attn: Librarian (Antenna Lab)
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Microwave Lab)
100 First Street
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Technical Research Group
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2 Aerial Way
Syosset, New York

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Physical Science Laboratory
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Bell Telephone Laboratories, Inc
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Attn. Technical Reports Librarian
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Aeronautical Systems Division
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National Bureau of Standards
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201 Lowell Street
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Missile and Space Division
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Bedford, Mass.

American Systems Incorporated
Attn. Technical Library Antenna Section
Hawthorne, California

National Research Council
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518 Franklin Avenue
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Radio Corporation of America
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