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A NETWORK FOR COMPENSATING THE SPECTRUM OF THE 6D4 MINIATURE GAS TRIODE

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A NETWORK FOR COMPENSATING THE SPECTRUM OF THE 6D4 MINIATURE GAS TRIODE

G. C. Page, Jr.

June 29, 1949

Approved by:

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DEPARTMENT OF THE NAVY ★ OFFICE OF NAVAL RESEARCH

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ABSTRACT

A method has been developed for calculating a practical network to permit the use of the 6D4 miniature gas triode as a noise source in jamming transmitters. The compensating network flattens the spectrum of the 6D4 tube to within three decibels, and contains adjustable circuit elements to take care of variations between tubes. It has been used successfully over the frequency range of 50 kc to 5 Mc.

PROBLEM STATUS

This is an interim report; work on the problem is continuing.

AUTHORIZATION

NRL Problem R06-21R (S-1249X-C)

A NETWORK FOR COMPENSATING THE SPECTRUM OF THE 6D4 MINIATURE GAS TRIODE

INTRODUCTION

The 6D4 miniature gas tube has been used with considerable success as a noise source in jamming transmitters. Because of the 30 to 40 db variation in the noise voltage output of this tube over the frequency range of 50 kc to 5 Mc, it was first thought that the modulator, into which the noise was injected, must be equalized. If, however, this noise spectrum could be flattened before it was fed into an amplifier or modulator, much of the trial and error compensation could be avoided, and a flat amplifier could be designed from the beginning.

This flattening was accomplished at the Laboratory by inserting a network[1] between the noise source and the first stage of amplification. Since many difficulties existed in precisely determining the internal impedance of the 6D4 noise generator, logical trial and error methods have heretofore been utilized to obtain an acceptable network. The network is now placed on a calculable basis, is adjustable to counteract differences in 6D4 tubes, and has been successfully used to compensate a number of these tubes.

GENERAL PROBLEM ANALYSIS

The circuit in which the 6D4 is used as a noise generator is shown in Figure 1 and its equivalent network in Figure 2. In order to determine Z on Figure 2 theoretically, it is necessary to obtain the noise spectrum of the generator over the range of 50 kc to 5 Mc, with the aid of a tuned radio-frequency noise analyzer of low input-capacity and high input-resistance. In making the spectrum analysis, C_K (Figures 1 and 2) is purposely very large so that it presents a short circuit to the noise generator over the desired frequency range. A brief consideration of Figure 3, the equivalent circuit of the 6D4 when connected to the noise spectrum analyzer, shows that V, the internal voltage, is unknown; and that Z, which may be complex, is also unknown. If Z is complex, a solution is not possible because there are only two available equations and at least three unknown quantities. Loading the generator with a condenser to obtain another equation is futile because the tube then emits pulses of noise or reacts in a random fashion as a relaxation oscillator. A variable inductance placed in series with the noise generator creates spurious oscillations in the analyzer because of magnetic coupling between the coils in the analyzer and the external inductance.

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[1] *Miles, J. M., "Experimental Jamming Transmitter Using a 200 Watt CWX-Band Developmental Magnetron," NRL Report No. R-3384 (Confidential), pp 5 and 7, 17 November 1948*

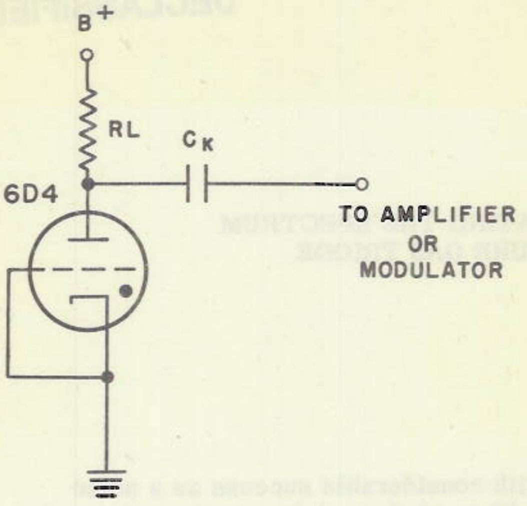
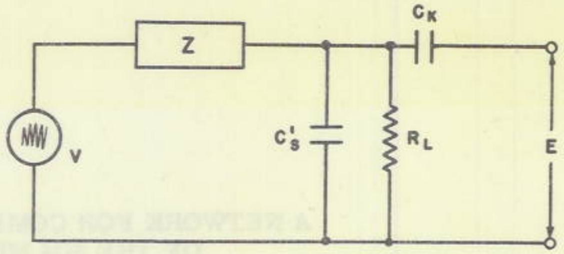


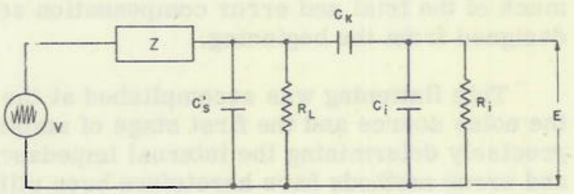
Figure 1 - Circuit for 6D4 as noise generator



Z = INTERNAL IMPEDANCE OF GENERATOR
V = INTERNAL VOLTAGE
C'_S = SOCKET CAPACITY & TUBE CAPACITY
R_L = PLATE-LOAD RESISTOR
C_K = COUPLING CAPACITOR
E = TERMINAL VOLTAGE

Figure 2 - Equivalent circuit of 6D4 as noise generator

It was hoped at first that either the capacitor or the inductor could be resonated with the internal impedance of the generator on the premise that Z was a varying series-resonant circuit; that is, that the R, L, and C of Z varied with frequency but had resonant points detectable at several frequencies. Since the two externally connected parameters introduced unusable oscillatory phenomena, resistive loading was tried with the simplifying assumption that the internal impedance was not complex, but was a pure resistance.



Z = INTERNAL IMPEDANCE OF GENERATOR
C'_S = SOCKET CAPACITY & TUBE CAPACITY
R_L = PLATE-LOAD RESISTOR
C_K = COUPLING CAPACITOR
C_i = INPUT CAPACITY OF NOISE ANALYZER
R_i = GRID RESISTANCE OF NOISE ANALYZER
V = INTERNAL VOLTAGE

Figure 3 Equivalent circuit of noise generator when connected to the noise analyzer

Because C'_S and C_i have negligible effect on the circuit of Figure 3 at the low-frequency end of the noise spectrum, they are omitted from the calculations which follow. C_K, R_i, and R_L are also omitted from computations, for, they too have a trivial consequence on the circuit. For the purpose of obtaining the spectrum and making an internal-impedance (resistance) measurement, R_L may have values form 6 to 100 K without appreciably altering the spectrum.[2]

With the foregoing assumptions in mind, it is possible to reduce the circuit of Figure 4A to that of Figure 4B; and at the same time Z (assumed to be purely resistive) can be determined by adjusting R (Figure 4B) until the voltage is one-half its value on open circuit. Appendix I explains this in more detail. Measurement of R with an ohmmeter is sufficiently accurate, and, for purposes here, this measured value is the internal impedance.

---

[2] *Cobine, J. D. and Curry, J. R. "The Characteristics of the Sylvania 6D4 Miniature Gas Triode as a Noise Source for the Range 0.1-5 Mc" NDRC, Division (15) RRL 411-169, Appendix Graphs H200-A71 and H200-A72, (Unclassified), March 30, 1945*

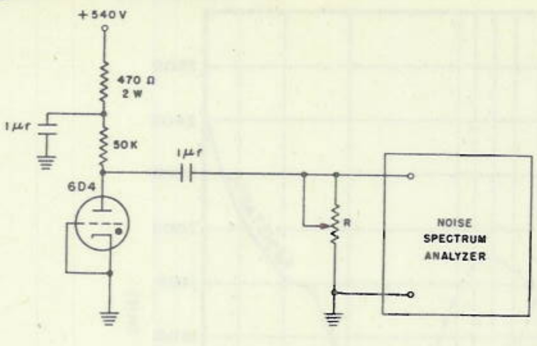


Figure 4A - Setup for measuring internal impedance

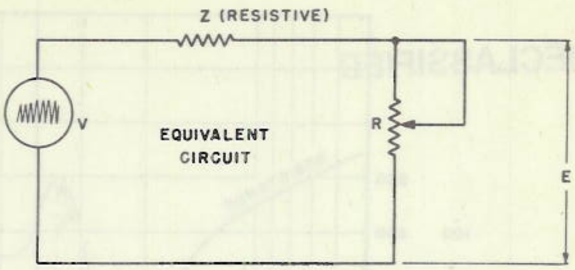


Figure 4B - Equivalent circuit of method for measuring internal impedance of noise generator

Figures 5 and 6 are graphs of the two typical noise spectra obtainable from 6D4's and their accompanying impedance characteristics as measured by the method already outlined. There appear to be at least two types of 6D4's, since two different noise spectra are present, one with a peak at 650 kc, the other with a slightly different shape and a peak at 800 kc.[3] When several tubes were broken open for inspection to ascertain any constructional differences, the position of the grid slot was found to vary with respect to the crimp in the plate, either facing the crimp or the opposite side. Just how these dissimilarities in construction affected the noise spectrum was debatable, since some of each type showed like spectra.

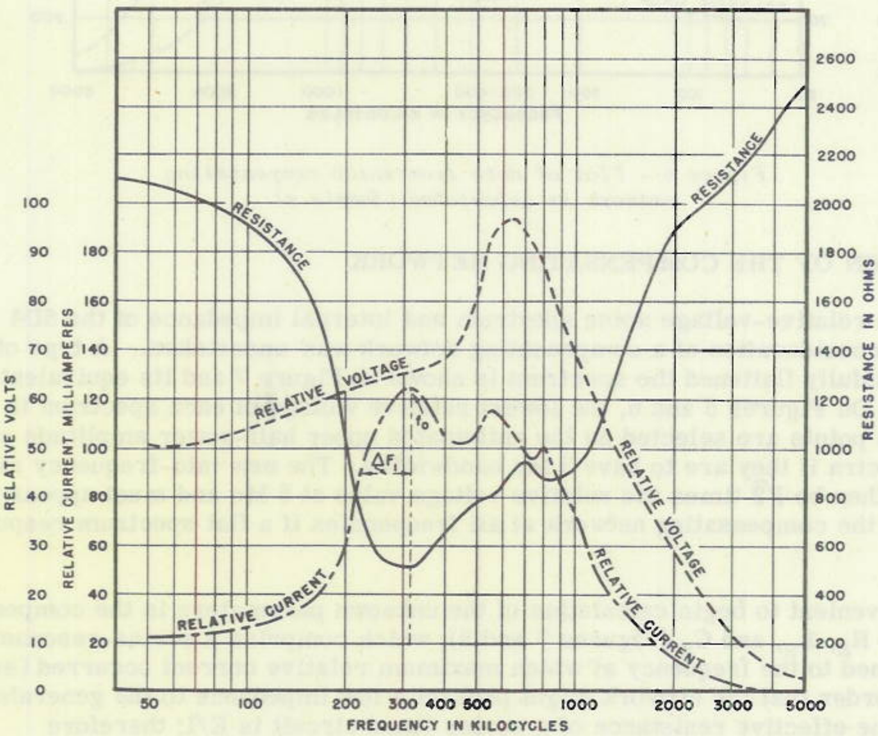


Figure 5 - Plot of data from which compensating network is calculated (Sample 1)

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[3] *Ibid. Graph # 11200-ATE*

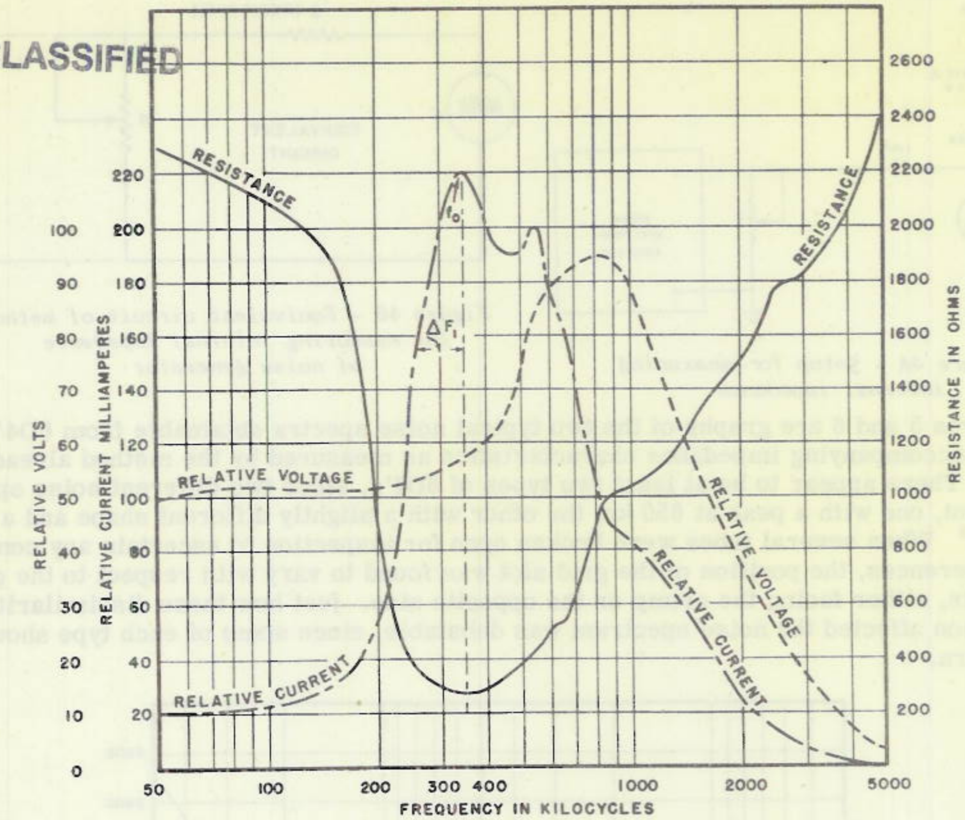


Figure 6 - Plot of data from which compensating network is calculated (Sample 2)

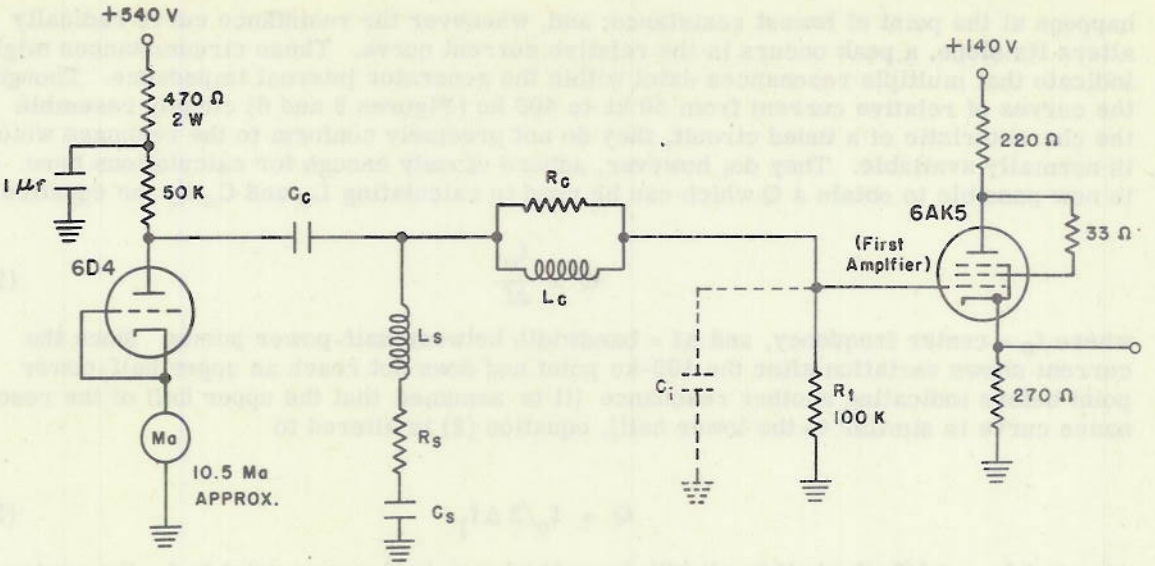
CALCULATION OF THE COMPENSATING NETWORK

Once the relative-voltage noise spectrum and internal impedance of the 6D4 had been determined, consideration of a compensating network was undertaken. A type of network which successfully flattened the spectrum is shown in Figure 7 and its equivalent circuit in Figure 8. On Figures 5 and 6, the lowest relative voltage of each spectrum is 3 to 4 volts. These points are selected as the anticipated upper half-power amplitude of the compensated spectra if they are to have 5 Mc bandwidths. The new mid-frequency response voltage will then be $\sqrt{2}$ times the relative voltage value at 5 Mc and must appear across the output of the compensating network at all frequencies if a flat spectrum response is expected.

It is convenient to begin calculation of the unknown parameters in the compensating network with R_S, L_S, and C_S (Figures 7 and 8), which comprise a series-resonant circuit purposely tuned to the frequency at which maximum relative current occurred (see Figures 5 and 6), in order that the network might present a low impedance to the generator. At resonance, the effective resistance of a series tuned circuit is E/I; therefore

$$R_S = \frac{\sqrt{2} \text{ times the relative voltage at 5 Mc}}{I}, \qquad (1)$$

where I is the maximum current. On Figures 5 and 6 also, the maximum relative current



C_C = COUPLING CAPACITOR IN COMPENSATING NETWORK

L_S, C_S, R_S = SERIES-RESONANT NETWORK

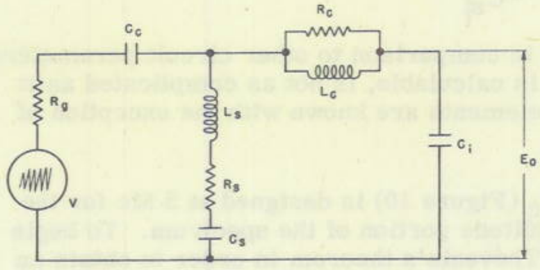
R_C, L_C = PEAKING NETWORK

C_i = INPUT CAPACITY OF AMPLIFIER

R_t = GRID RESISTOR

Figure 7 - Type of network which flattened the noise spectrum

E_O = VOLTAGE OUT OF COMPENSATING NETWORK



R_g = INTERNAL RESISTANCE

V = INTERNAL VOLTAGE

C_C = COUPLING CAPACITOR IN COMPENSATING NETWORK

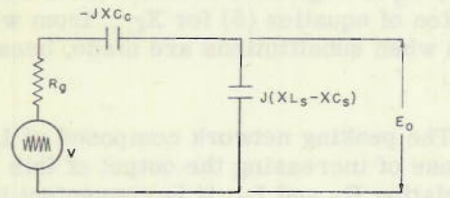
L_s, R_s, C_s = SERIES RESONANT NETWORK

R_C, L_C = PEAKING NETWORK

C_i = INPUT CAPACITY OF ANALYZER

E_O = VOLTAGE OUT OF COMPENSATING NETWORK

Figure 8 - Equivalent circuit of noise generator and compensating network



R_g = MEASURED INTERNAL IMPEDANCE OF 6D4 AT 50 kc (SEE FIGURES 5 AND 6)

$-JXC_C$ = REACTANCE OF UNKNOWN COUPLING CAPACITOR

V = RELATIVE SPECTRUM VOLTAGE AT 50 kc

$J(XL_s - XC_s)$ = REACTANCE OF SERIES-RESONANT SHUNT ARM AT 50 kc

E_o = NEW CENTER-FREQUENCY VOLTAGE OF ANTICIPATED SPECTRUM ($1/\sqrt{2}$ X RELATIVE VOLTAGE AT 5 Mc)

Figure 9 - 50-kc equivalent circuit of compensating network

happens at the point of lowest resistance; and, whenever the resistance curve radically alters its slope, a peak occurs in the relative current curve. These circumstances might indicate that multiple resonances exist within the generator internal impedance. Though the curves of relative current from 50 kc to 400 kc (Figures 5 and 6) closely resemble the characteristic of a tuned circuit, they do not precisely conform to the response which is normally available. They do, however, adhere closely enough for calculations here. It is now possible to obtain a Q which can be used in calculating L_s and C_s by the equation

$$Q = \frac{f_o}{\Delta f} \tag{2}$$

where f_o = center frequency, and Δf = bandwidth between half-power points. Since the current shows variation after the 400-kc point and does not reach an upper half-power point before indicating another resonance (it is assumed that the upper half of the resonance curve is similar to the lower half), equation (2) is altered to

$$Q = f_o/2\,\Delta f_1 \tag{3}$$

where $\Delta f_1 = \Delta f/2$, the half bandwidth from the lower half-power point to f_o, the center frequency. With this Q, C_s and L_s are determinable from

$$Q = \frac{1}{R_s \omega C_s} = \frac{\omega L_s}{R_s}. \tag{4}$$

Next, the value of C_c (Figures 7 and 8) must be calculated at the lowest frequency of the spectrum to be passed by the compensating network; in this case a 50-kc equivalent circuit as in Figure 9 is applicable. To obtain X_{C_c},

$$\left|\frac{V}{E_o}\right| = \frac{\sqrt{R_g^2 + \left[X_{C_c} + \left(X_{C_s} - X_{L_s}\right)\right]^2}}{\left|X_{L_s} - X_{C_s}\right|} \tag{5}$$

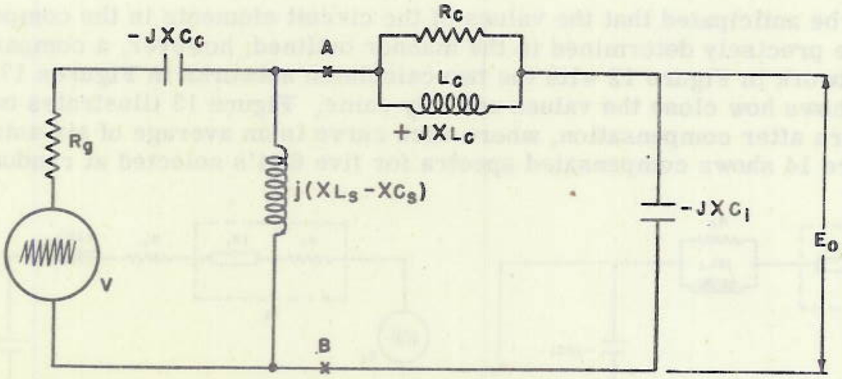
where R_s is neglected because of its small size in comparison to other circuit parameters. Solution of equation (5) for X_{C_c}, from which C_c is calculable, is not as complicated as it looks when substitutions are made, because all elements are known with the exception of X_{C_c}.

The peaking network composed of L_c and R_c (Figure 10) is designed at 5 Mc for the purpose of increasing the output of this low-amplitude portion of the spectrum. To begin calculating R_c and L_c, it is convenient to apply Thévenin's theorem in order to obtain an equivalent generator; so the circuit is opened at A and B (Figure 10) and the open circuit voltage, E_t, is determined as

$$\left|E_t\right| = \frac{V\left|X_{L_s} - X_{C_s}\right|}{\sqrt{R_g^2 + \left(X_{L_s} - X_{C_s} - X_{C_c}\right)^2}}. \tag{6}$$

The next step is to short out the sources of emf and determine the Thévenin impedance looking back into the network from terminals A and B (Figure 10). Then

$$Z_t = \frac{\left(R_g - jX_{C_c}\right)\left[j\left(X_{L_s} - X_{C_s}\right)\right]}{R_g + j\left(X_{L_s} - X_{C_s} - X_{C_c}\right)}. \qquad (7)$$



R_g = MEASURED INTERNAL IMPEDANCE OF 6D4 AT 5 Mc
V = RELATIVE SPECTRUM VOLTAGE AT 5 Mc
$-JX_{C_c}$ = REACTANCE OF COUPLING CAPACITOR AT 5 Mc
$J(X_{L_s}-X_{C_s})$ = REACTANCE OF SERIES-RESONANT SHUNT ARM AT 5 Mc
R_c, JX_{L_c} = UNKNOWN PARAMETERS OF 5-Mc PEAKING NETWORK
$-JX_{C_i}$ = INPUT CAPACITIVE REACTANCE OF FIRST STAGE OF AMPLIFICATION
E_o = NEW CENTER-FREQUENCY VOLTAGE OF ANTICIPATED SPECTRUM ($\sqrt{2}$ X RELATIVE VOLTAGE AT 5 Mc)

Figure 10 - 5-Mc equivalent circuit of compensating network

The equivalent network resulting from the application of Thévenin's theorem is shown on Figure 11A. In order that R_c and L_c be calculated, it is feasible to change the equivalent circuit further by making use of an equivalent series combination of R_c and X_{L_c} designated on Figure 11B by R'_c and X'_{L_c}. Now X_t, the reactive part of Z_t, plus X'_{L_c} is designed to resonate with X_{C_i} at 5 Mc, and since X_t is known,

$$X'_{L_c} = X_{C_i} - X_t . \qquad (8)$$

Because resonance exists, $j\left(X_t + X'_{L_c} - X_{C_i}\right) = 0$, and R'_c may be found by solving

$$\left|\frac{E_o}{E_t}\right| = \frac{\left|X_{C_i}\right|}{R'_C + R_t} . \qquad (9)$$

Enough information is now available so that R'_C and X'_{L_c} may be determined, but they must be converted to the original parallel combination for practical application; hence,

$$R_c = \frac{\left(R'_c\right)^2 + \left(X'_{L_c}\right)^2}{R'_C} \tag{10}$$

and

$$X_{L_c} = \frac{\left(R'_c\right)^2 + \left(X'_{L_c}\right)^2}{X'_{L_c}}. \tag{11}$$

It could not be anticipated that the values of the circuit elements in the compensating network could be precisely determined in the manner outlined; however, a comparison of the practical network in Figure 12 with the two calculated networks in Figures 17 and 18 of Appendix II shows how close the values actually came. Figure 13 illustrates two possibly obtainable spectra after compensation, where each curve is an average of six noise generators; and Figure 14 shows compensated spectra for five 6D4's selected at random.

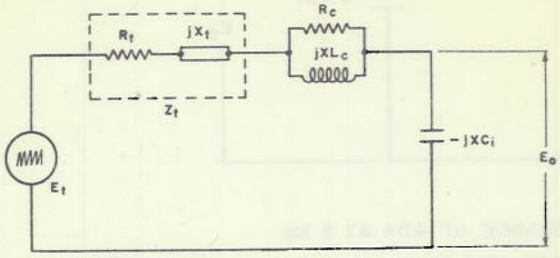


Figure 11A - Thevenin equivalent circuit of 5-Mc network

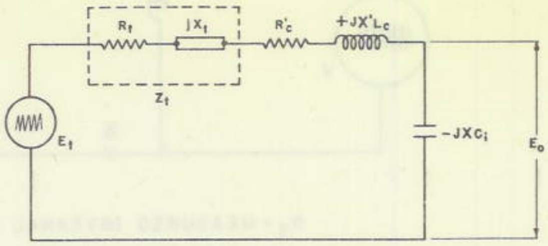


Figure 11B - Thevenin equivalent circuit of 5-Mc network with series transformation

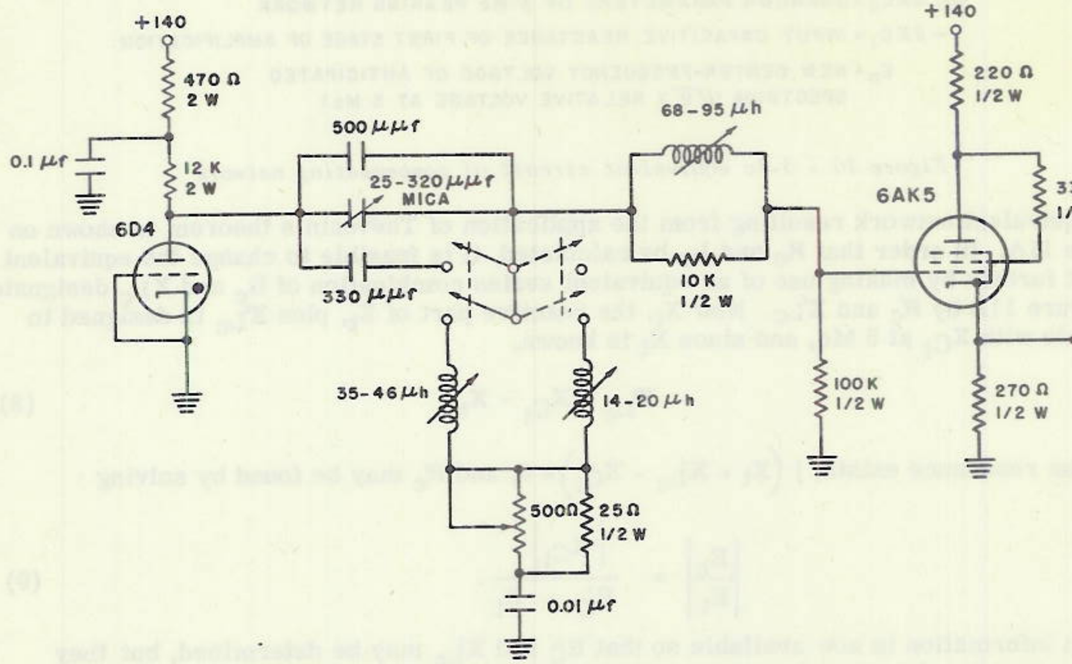


Figure 12 - Practical compensating network

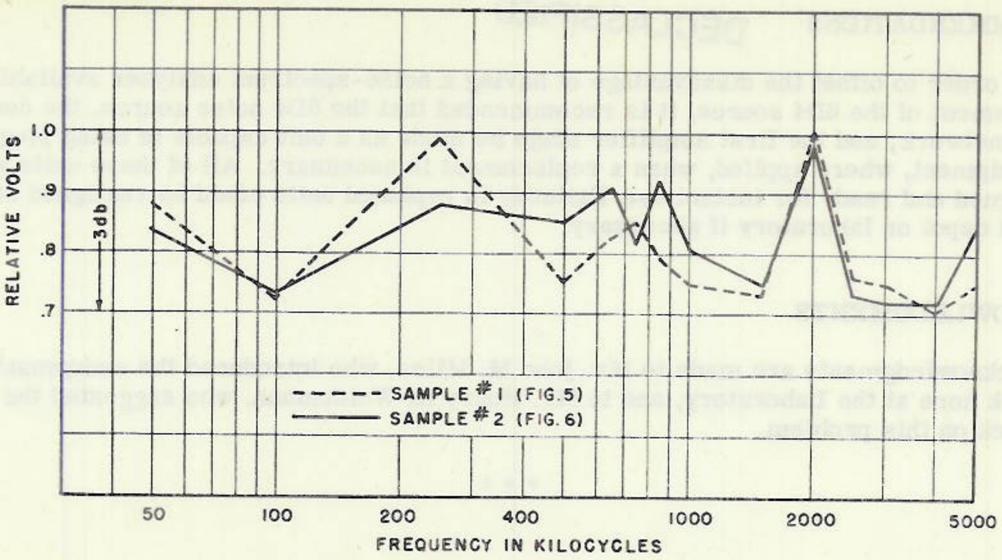


Figure 13 - Typical spectra after compensation

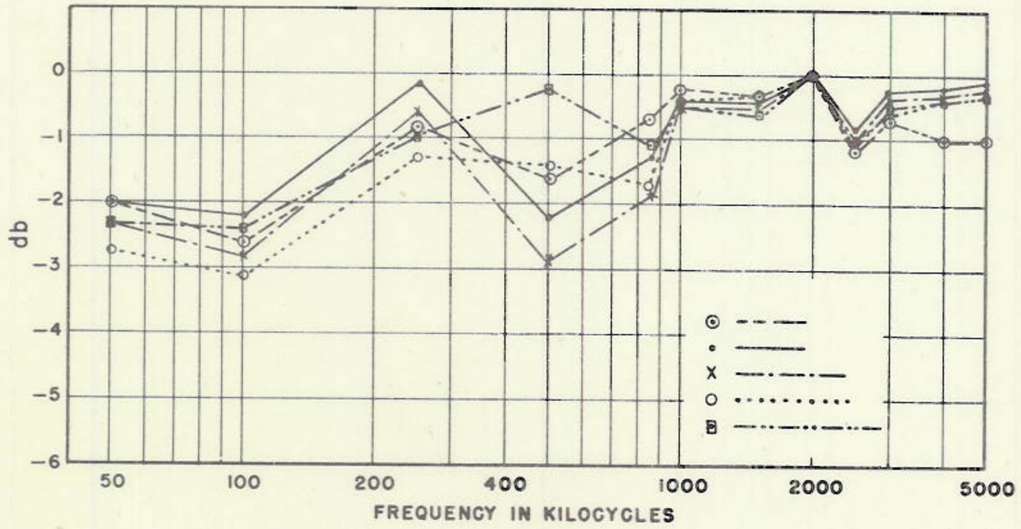


Figure 14 - Typical spectra after compensation

CONCLUSIONS

With the network discussed in this report and its method of determination, it is possible to eliminate much of the tedious trial and error compensation of amplifiers into which the noise spectrum of the 6D4 is fed. Though the spectrum level is attenuated some 35 db there is still a usable noise output somewhat higher in amplitude than that of the 931-A photomultiplier noise source and an entire amplifier stage could possibly be eliminated. The main disadvantage is the necessity of having a noise-spectrum analyzer available when installing a new 6D4 source, since there are differences apparent from tube to tube which could require readjustment of the compensating network.

RECOMMENDATIONS

In order to offset the disadvantage of having a noise-spectrum analyzer available upon replacement of the 6D4 source, it is recommended that the 6D4 noise source, the compensating network, and the first amplifier stage be made as a unit capable of being plugged into the equipment, where applied, when a replacement is necessary. All of these units would be prealigned and ready for instant use. Failures in replaced units could be realigned at some central depot or laboratory if necessary.

ACKNOWLEDGMENTS

Acknowledgments are made to Mr. John M. Miles, who introduced the compensating network here at the Laboratory, and to Mr. Henry K. Weidemann, who suggested the method of attack on this problem.

* * *

APPENDIX I
Explanation of the Method Used to Measure the Internal Impedance of the 6D4 as a Noise Source

In Figures 15 and 16

E_o = Open circuit voltage

E_g = Generator internal voltage

R_g = Generator internal resistance

R = (Externally connected) load resistance

E_r = Voltage across R.

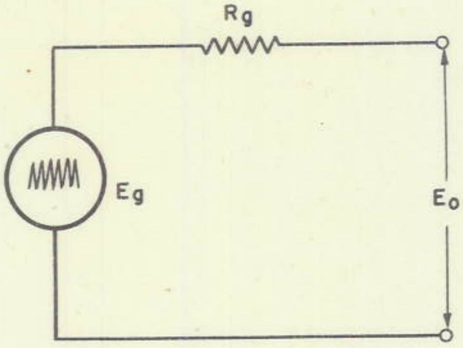


Figure 15 - Generator on open circuit

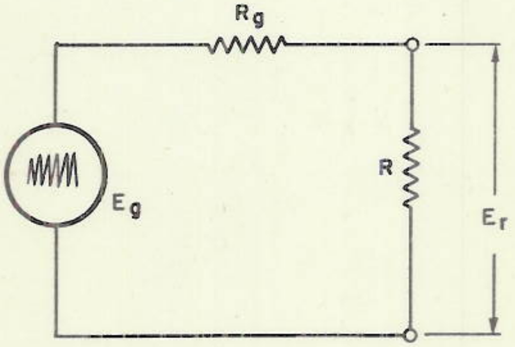


Figure 16 - Generator under load

It is apparent that

$$E_r = \frac{E_g R}{R_g + R}.$$

On open circuit, $E_g = E_o$. Therefore,

$$E_r = \frac{E_o R}{R_g + R}$$

or

$$\frac{E_r}{E_o} = \frac{R}{R_g + R}.$$

If $E_r/E_o = 1/2$, then

$$\frac{1}{2} = \frac{R}{R_g + R}$$

$$R_g + R = 2R$$

$$R_g = R.$$

Hence, the internal resistance is equal to the external resistance.

* * *

APPENDIX II

Sample Calculations Involving Methods Already Discussed, With a Comparison of Calculated Values of Network Parameters and the Practical Compensating Network

Using Figure 5 as the criterion of a typical noise spectrum and internal resistance measurement, equations 1 and 3 through 11 are applied as follows:

$$(1)\quad R_s = \frac{\sqrt{2} \text{ times relative voltage at 5 Mc}}{I}$$

where the relative voltage is 3 taken from the relative voltage curve on Figure 5; and the maximum current, I, is 0.126 ampere taken from the curve of relative current; so

$$R_s = \frac{\sqrt{2} \text{ times } 3}{0.126}$$

$$= 33.7 \text{ ohms.}$$

$$(3)^*\qquad Q = \frac{f_o}{2\,\Delta f_1}.$$

For this particular case $f_o = 315$ kc, taken from the relative current curve at its maximum amplitude. $f_1 = 90$ kc, read on the same curve of relative current; therefore

$$Q = \frac{315}{2 \text{ times } 90}$$

$$= 1.75.$$

$$(4)\qquad Q = \frac{1}{\omega R_s\ C_s} = \frac{\omega L_s}{R_s}.$$

By substituting into (4) values calculated in equations (1) and (3), C_S and L_S may be computed.

$$\omega L_S = 1.75 \text{ times } 33.7$$

$$= 59 \text{ ohms at 315 kc.}$$

Because resonance exists, $X_{L_S} = X_{C_S}$; therefore, $C_S = 0.0086$ microfarad and $L_S = 29.7$ microhenries.

---

* *Equation (2) is not used here.*

In order to continue calculations and apply equation (5), X_{L_S} and X_{C_S} must be converted to their 50 kc values. X_{C_S} at 50 kc = 372 ohms and X_{L_S} at 50 kc = 9.4 ohms.

$$(5) \qquad \left|\frac{V}{E_o}\right| = \frac{\sqrt{R_g^2 + \left[X_{C_c} + \left(X_{C_s} - X_{L_s}\right)\right]^2}}{X_{L_s} - X_{C_s}}.$$

Substituting the 50 kc value of R_g as taken from the curve of resistance (Figure 5) and the 50 kc values of X_{C_S} and X_{L_S}, a solution of equation (5) for X_{C_c} is now possible, where V is the 50 kc relative voltage value and $E_o = \sqrt{2}$ times 3.

$$\frac{51}{4.2} = \frac{\sqrt{(2100)^2 + \left(X_{C_c} + 363\right)^2}}{363},$$

$$X_{C_c} = 3{,}522 \text{ ohms, so}$$

$$C_c = 910 \text{ micromicrofarads.}$$

$$(6) \qquad \left|E_t\right| = \frac{V\left|X_{L_s} - X_{C_s}\right|}{\sqrt{R_g^2 + \left(X_{L_s} - X_{C_s} - X_{C_c}\right)^2}}$$

From Figures 5 and 10, R_g = 2500 ohms at 5 Mc, V = 3 volts relative, X_{L_S}, X_{C_S}, and X_{C_c} are reactance values determined at 5 Mc.

$$E_t = \frac{3(936)}{\sqrt{(2500)^2 + (901)^2}}$$

$$= 1.06 \text{ volts.}$$

$$(7) \qquad Z_t = \frac{\left(R_g - jX_{C_c}\right)\left[j\left(X_{L_s} - X_{C_s}\right)\right]}{R_g + j\left(X_{L_s} - X_{C_s} - X_{C_c}\right)}.$$

The numerical values substituted in (7) are the same as in (6), where X_{C_c} = 35.2 ohms at 5 Mc.

$$Z_t = \frac{(2500 - j35.2)\ (j936)}{2500 + j901}$$

$$= 310 + j824 \text{ vector ohms.}$$

(8) $$X'_{L_c} = X_{C_i} - X_t$$

X'_{L_c} is the series equivalent of the peaking inductance, X_{L_c}, (shown on Figure 11B),

X_{C_i} = 3,183 ohms (assumed input capacity of amplifier at 5 Mc is 10 micromicrofarads),

X_t = 824 ohms from (7), and

$$X'_{L_c} = 3{,}183 - 824$$

$$= 2{,}359 \text{ ohms.}$$

(9) $$\frac{E_o}{E_t} = \frac{X_{C_i}}{R'_c + R_t},$$

$$\frac{4.2}{1.06} = \frac{3{,}183}{R'_c + 310},$$

$$R'_c = 494 \text{ ohms.}$$

Now it remains to convert R'_c and X'_{L_c} to their parallel equivalents.

(10) $$R_c = \frac{\left(R'_c\right)^2 + \left(X'_{L_c}\right)^2}{R'_c},$$

$$R_c = \frac{(494)^2 + (2359)^2}{494}$$

$$= 10{,}562 \text{ ohms.}$$

(11) $$X_{L_c} = \frac{\left(R'_c\right)^2 + \left(X'_{L_c}\right)^2}{X'_{L_c}},$$

$$X_{L_c} = \frac{(494)^2 + (2359)^2}{2359}$$

$$= 2{,}462 \text{ ohms,}$$

$$L_c = 78.4 \text{ microhenries.}$$

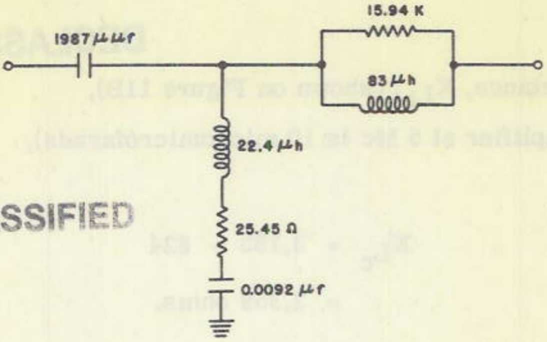


Figure 17 - Sample 1

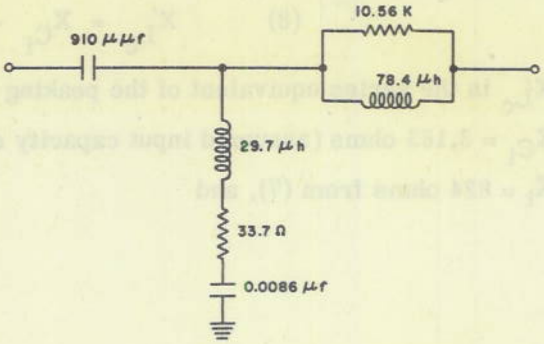


Figure 18 - Sample 2

Calculated compensating networks

Figure 17 shows the calculated compensating network with the values just determined in this appendix. In a similar manner the values of the parameters of Figure 18 were calculated using data obtained from Figure 6.

* * *

LIST OF SYMBOLS USED IN CALCULATIONS

R_L	6D4 plate load resistor
C_k	Coupling capacitor
V	Internal voltage of noise generator
Z	Internal impedance of noise generator
C_s'	6D4 socket capacity plus tube capacity
E	Terminal voltage
C_i	Input capacity of noise analyzer or first amplifier
R_i	Grid resistance of noise analyzer
R	Resistor used for loading noise generator
C_c	Coupling condenser in compensating network
L_s R_s C_s	Series resonant shunt arm of compensating network
R_c L_c	Elements of peaking network
R_T	Grid resistor of first amplifier
R_g	Assumed internal resistance of noise generator
E_o	Voltage out of compensating network (new center frequency voltage)
$-jX_{C_c}$	Reactance of compensating network coupling capacitor at 50 kc
$-j(X_{C_s} - X_{L_s})$	Reactance of series resonant shunt arm
jX_{L_c}	Reactance of peaking inductance
E_t	Thévenin equivalent generator
Z_t	Thévenin equivalent internal impedance
R_t	Resistance component of Thévenin impedance

[redacted]

jX_{L_t}	Reactive component of Thévenin impedance
R_c'	Series equivalent of R_c
jX_{L_c}'	Series equivalent of jX_{L_c}
f_o	Center frequency
Δf	Bandwidth between half-power points
Δf_1	$\frac{1}{2}$ the bandwidth between half-power points

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[redacted]