The Design, Construction, and Testing of a Tuned VHF Amplifier.

A Thesis Presented to
the Faculty of Engineering
University of Manitoba

In Partial Fulfillment of
the Requirement for the Degree of
Bachelor of Science in Electrical Engineering.

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And

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Acknowledgment:

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I Introduction:

This thesis deals with the design and construction of a tuned VHF amplifier. The main requirements are that optimum transducer gain must be attempted with a bandwidth of 6 MHz. Theoretical considerations such as stability, determination of optimum terminations and matching networks will be considered.

The design will be carried out using a Fairchild 2N4122 bipolar transistor operating on channel 2 into a 50 ohm coaxial system.

We shall try to include enough qualitative discussion with a minimum of theory so that a reader with only an elementary background in electronics will be able to apply the concepts to a design of his own.

II Design Considerations:

a) Stability:

The internal feedback mechanism of a bipolar transistor may frequently cause a circuit to become unstable and oscillate. Linville[1] has derived a relationship that tests the potential stability of an open circuited active two port at a single frequency. Given the y parameters of the two port the relationship is:

$$C = \frac{|y_{12} y_{21}|}{2 g_{11} g_{22} - Re\{y_{12} y_{21}\}} \qquad (1)$$

$$g_{11}, g_{22} > 0$$

where:

C = Linville Stability Factor.

If C is found to be greater than one then the device is potentially unstable, but if C is less than one the device is stable. In spite of Linville's test a circuit

[1] J. G. Linville & J. F. Gibbons, Transistors and Active Circuits, New York; McGraw-Hill, pp 241-250

containing an unstable active device may be stable for certain passive terminations. Stern[2] has developed a more meaningful relationship which tests the stability of any active two port, at a single frequency, when terminated by load and source admittances Y_s and Y_l. The Stern stability factor, K, is given by:

$$K = \frac{2(g_{11}+G_S)(g_{22}+G_L)}{|y_{12}y_{21}| + Re\{y_{12}y_{21}\}} \qquad (2)$$

The circuit is then stable if K is greater than one and is unstable if K is less than one.

Ghaussi[3] suggests that K should be between 2 and 10 in practical circuit implementation.

[2] Arthur P. Stern, Stability and Power Gain of Tuned Transistor Amplifiers, <u>Proc. IRE</u>, vol. 45, pp. 335-343, March 1957.

[3] M.S. Ghaussi, <u>Principles and Design of Linear Active Circuits</u>, New York, Mc Graw-Hill, p. 468.

b) Gain

At the high frequencies under consideration transmission lines must be used for circuit interconnection and those lines must be matched so that high standing wave ratios are avoided. The fact that only power flow, and not voltage, on a transmission line is meaningful and that high power gain is desirable in an R.F. amplifier to reduce the effect of mixer noise, leads us to the conclusion that the amplifier must be matched for maximum power gain.

If we consider an active two port configuration as shown in fig. 1, then for all frequencies at which the amplifier is to operate efficiently Y_1 and Y_2 must match the line admittances.

Also since we intend to use tuned circuits at both the input and output, the matching networks must provide for all reactive component cancellation at the appropriate resonant frequency. This, combined with the assumption that we are matching for maximum power gain at both ports, gives us the necessary relationships that:

$$Y_{11} = Y_{in}^{*} \qquad Y_{22} = Y_{out}^{*} \qquad (3)$$

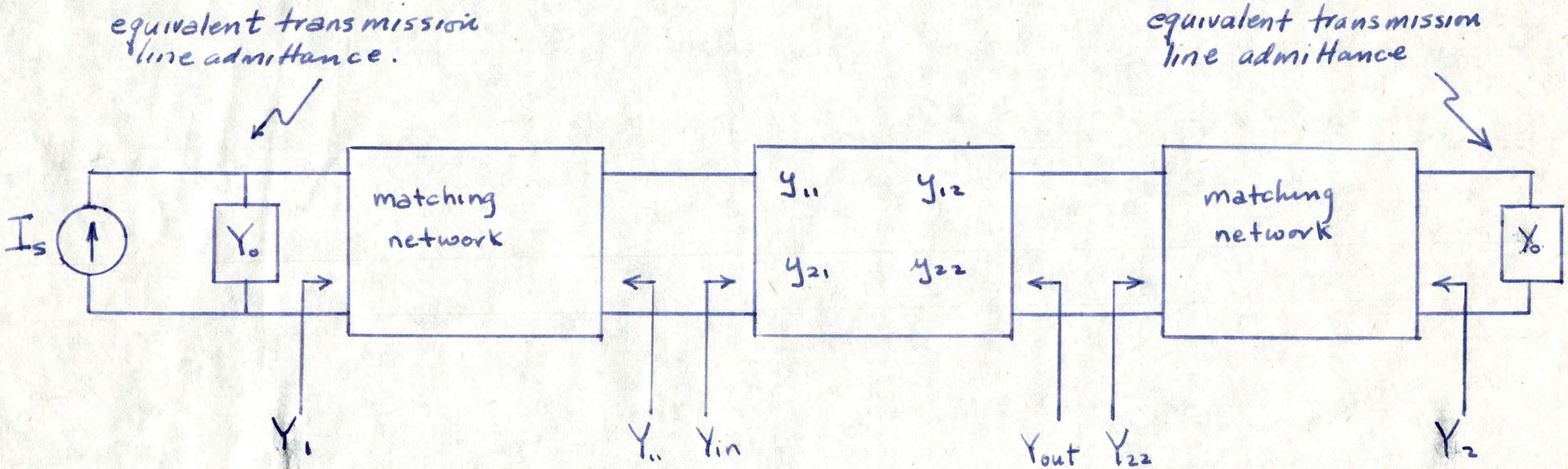

fig. 1

note: the transmission line impedances will be resistive for most frequencies.

Stated in another way we can consider the transmission lines to be the resistive loads of two matching networks whose input admittances are equal to the conjugates of the input and output admittances of the active two port.

This conjugate matching technique allows us to use the concept of transducer power gain[4] (G_T) in our design considerations

We should note here that under conditions of maximum G_T the voltage gain in db. is the same as G_T in db if the input and output admittances are equal.

[4] see Appendix A.

c) Terminations:

For a given two port terminated with load and source admittances, as shown in fig. 2, the following formulae may be derived; .

$$Y_{in} = y_{11} - \frac{y_{12}\, y_{21}}{y_{22} + Y_L} \qquad (4)$$

$$Y_{out} = y_{22} - \frac{y_{12}\, y_{21}}{y_{11} + Y_s} \qquad (5)$$

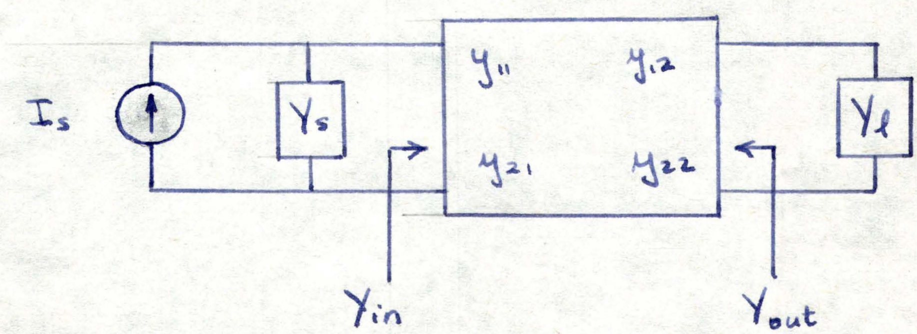

fig. 2.

The dependance of Y_{in} and Y_{out} upon the load and source admittances is due to the internal feedback of the bipolar transistor.

To envoke conjugate matching at the input we must chose $Y_s = Y_{in}^*$. However we do not know Y_L and thus Y_{in} is not available. Unfortunately a similar situation exists at the output.

There are two techniques available to overcome this problem. The first, a method of iteration, is quite lengthy and very messy. The starting point is to assume that the output is conjugately matched to some arbitrary value. Some simplification may be made if this value is assumed to be approximately equal to y_{22}^* since then the iteration will converge at a faster rate. With this assumed value of Y_L we can calculate Y_{in} and pick $Y_s = Y_{in}^*$. Since now we have a value of Y_s we can recalculate Y_{out} and match to its conjugate. If this method is continued through four or five cycles the values obtained will be quite close to the optimum.

The second method, although much more elegant, is much simpler. Kuh and Rohrer[5] have derived the equations to solve for the optimum terminations given the y parameters of the device and the fact that the matching is desired to maximize G_T.

[5] E.S. Kuh and R.A. Rohrer, <u>Theory of Linear Active Networks</u>, San Francisco, Holden-Day Inc. pp. 239-241. (see also Appendix B)

With reference to fig 2 - the necessary relationships are:

$$G_L = \frac{1}{2g_{11}} \sqrt{(2g_{11}g_{22} - M)^2 - L^2} \qquad (6)$$

$$G_S = \frac{1}{2g_{22}} \sqrt{(2g_{11}g_{22} - M)^2 - L^2} \qquad (7)$$

$$B_L = \frac{N}{2g_{11}} - b_{22} \qquad (8)$$

$$B_S = \frac{N}{2g_{22}} - b_{11} \qquad (9)$$

where:

$$y_{12}y_{21} = M + jN = Le^{j\phi} \qquad (10)$$

In some cases however these values of optimum terminations will not yield a stable circuit. Here again we have two alternatives. They are: 1) neutralization 2) mismatch design.

Neutralization consists of using a feedback path to make the circuit unilateral. However since reactive elements are needed the neutralization is good for only a single frequency. As a consequence, mismatch design is considered to be superior. In this method the matching networks are purposely designed so that G_S and G_L are increased enough so that the circuit becomes stable in accordance with Stern's criterion. This mismatch lowers the power gain to increase stability.

Ghaussi[6] has derived the necessary relationships for the determination of the values of the terminations given the two port y parameters and the desired degree of stability. He also includes a formula which calculates the power gain for the mismatch.

[6] M.S. Ghaussi, Principles and Design of Linear Active Circuits, New York, McGraw Hill, pp 467-471

d) Matching Networks

Irrespective of the manner in which the value of the terminations is calculated the matching network must be designed to change the admittance level of the transmission line to this calculated value.

There are a multitude of ways in which this procedure can be carried out. Two of the more common techniques are:

1) utilization of band pass filter techniques:

2) use of the Smith chart.

The only drawback with these techniques is that the circuit Q is sometimes very hard to calculate and thus it is hard to meet the bandwidth requirements. For this reason the use of design charts found in many texts on radio communications is recommended[7].

[7] see for example: K. Henney, Radio Engineering Handbook, 3rd. Ed., New York, Mc-Graw Hill.

[7] F. E. Terman, Radio Engineers Handbook, 1st Ed., New York, Mc Graw Hill.

[7] F. Davis, Matching Network Designs with Computer Solutions, Motorolla application note AN-267.

III Actual Design:

a) Calculation of terminations and G_T.

Channel 2 has a frequency spectrum of 6 MHz bandwidth centred at 57 MHz[8] and therefore all calculations concerning stability and matching will be carried out using the centre frequency.

From the transistor specification sheet we have the following y parameters for a common emitter configuration when biased at 5 volts and 10 ma.:

$$y_{11} = 12 + j14 \quad m\mho$$
$$y_{12} = -0.1 - j0.9 \quad m\mho$$
$$y_{21} = 75 - j150 \quad m\mho$$
$$y_{22} = 0.6 + j1.5 \quad m\mho$$

Using these values Linvill's stability factor is calculated to be[9] 0.97, and since this is quite close to the critical value the circuit stability is in question.

From formulae (6) to (10) the values of the optimum terminations are:[10]

[8] see appendix C for T.V. frequency allocations.

[9,10] see appendix D for calculations.

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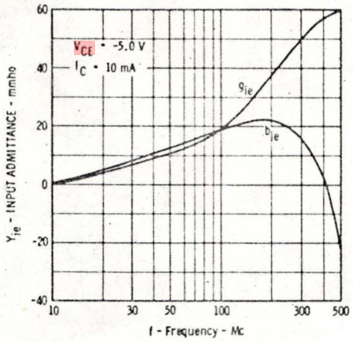
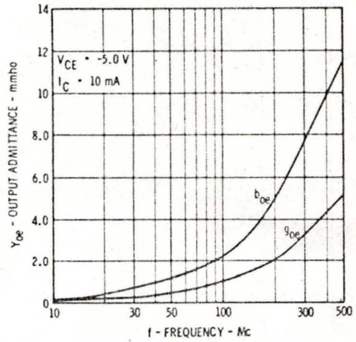
TYPICAL COMMON EMITTER "Y" PARAMETERS

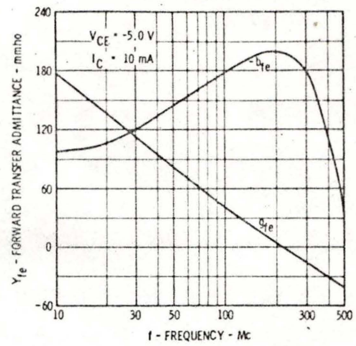
INPUT ADMITTANCE VERSUS FREQUENCY-OUTPUT SHORT CIRCUIT



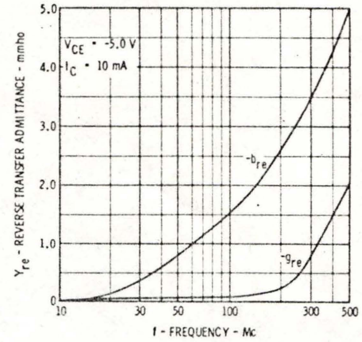
OUTPUT ADMITTANCE VERSUS FREQUENCY-INPUT SHORT CIRCUIT



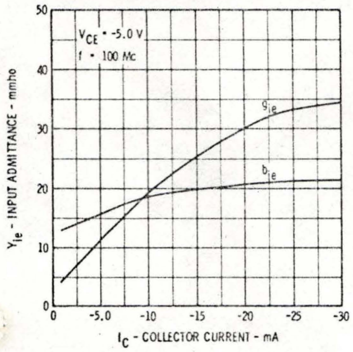
FORWARD TRANSFER ADMITTANCE VERSUS FREQUENCY-OUTPUT SHORT CIRCUIT



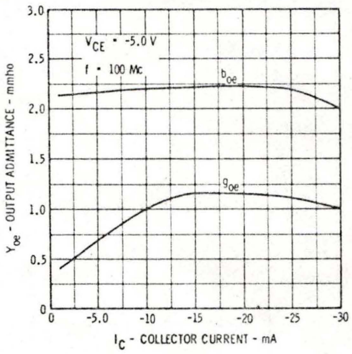
REVERSE TRANSFER ADMITTANCE VERSUS FREQUENCY-INPUT SHORT CIRCUIT



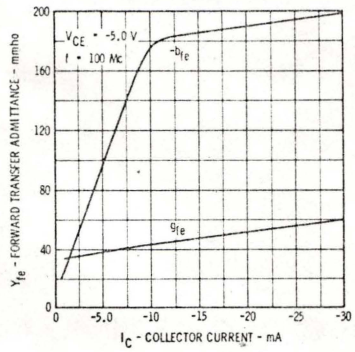
INPUT ADMITTANCE VERSUS COLLECTOR CURRENT OUTPUT SHORT CIRCUIT



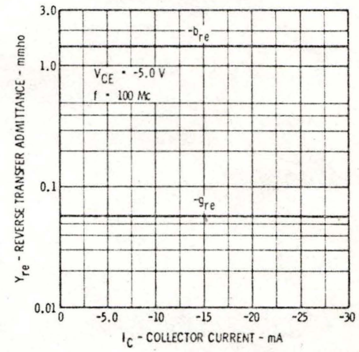
OUTPUT ADMITTANCE VERSUS COLLECTOR CURRENT-INPUT SHORT CIRCUIT



FORWARD TRANSFER ADMITTANCE VERSUS COLLECTOR CURRENT-OUTPUT SHORT CIRCUIT



REVERSE TRANSFER ADMITTANCE VERSUS COLLECTOR CURRENT-INPUT SHORT CIRCUIT

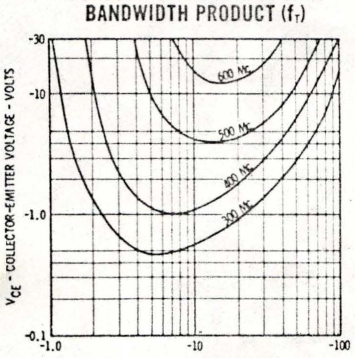


SMALL SIGNAL CHARACTERISTICS (f=1 kHz)

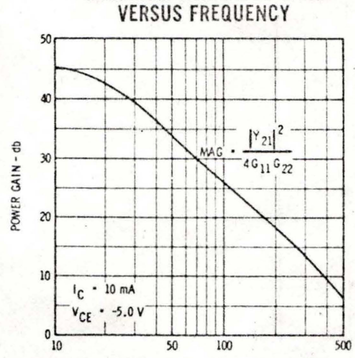
Symbol	Characteristic	2N4121 Min.	2N4121 Max.	2N4122 Min.	2N4122 Max.	Units	Test Conditions
h_{ie}	Input Resistance	1.0	8.0	4.0	12	KΩ	I_C = 1.0 mA V_{CE} = -10 V
h_{oe}	Output Conductance	2.0	24	8.0	40	μmho	I_C = 1.0 mA V_{CE} = -10 V
h_{re}	Voltage Feedback Ratio		3.0		4.0	$x10^{-4}$	I_C = 1.0 mA V_{CE} = -10 V
h_{fe}	Forward Current Transfer Ratio	50	300	150	450		I_C = 1.0 mA V_{CE} = -10 V

TYPICAL ELECTRICAL CHARACTERISTICS

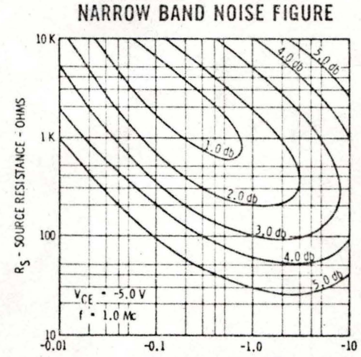
CONTOUR OF CONSTANT GAIN BANDWIDTH PRODUCT (f_T)



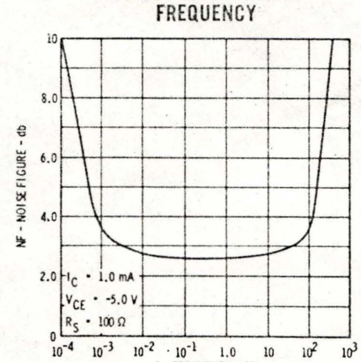
MAXIMUM AVAILABLE GAIN VERSUS FREQUENCY



CONTOURS OF CONSTANT NARROW BAND NOISE FIGURE



NOISE FIGURE VERSUS FREQUENCY



$$Y_L = 1.61 - j3.70 \text{ m}\mho$$

$$Y_S = 32.2 - j57.7 \text{ m}\mho$$

We should note that almost identical values are obtained after 6 cycles of the iteration method.

Now that we have the numerical values of the terminations the circuit stability must be tested using Stern's criterion. Stern's stability factor turns out to be[11] 20.6. This gives us plenty of stability margin and assures that the amplifier will be stable in the region of the passband cuttoff where the amplifier is effectively mismatched.

Since the circuit is stable with the optimum terminations we may design to obtain the optimum transducer gain which is[12] 21.6 db.

[11, 12] see appendix D.

b) Matching Networks

To satisfy the frequency requirements a circuit Q of about 9.5 is necessary. However, since we are tuning both the input and the output we have to account for the interaction of the two tuning circuits. Elementary considerations of cascaded tuned stages[13] shows that the individual Q must be reduced by a factor of 1.56 to obtain the necessary overall Q. This necessitates that each tuned circuit has a Q of approximately 6.1.

Due to the low Q required, simple networks where the Q is readily calculated are not suitable. As a result we have utilized the following networks and formulae;[14]

[13] C.L. Alley and K.W. Atwood, Electronic Engineering, New York, John Wiley and Sons, P 341.

[14] F. Davis, Matching Network Design with Computer Solutions, Motorolla application note AN-267

Output:

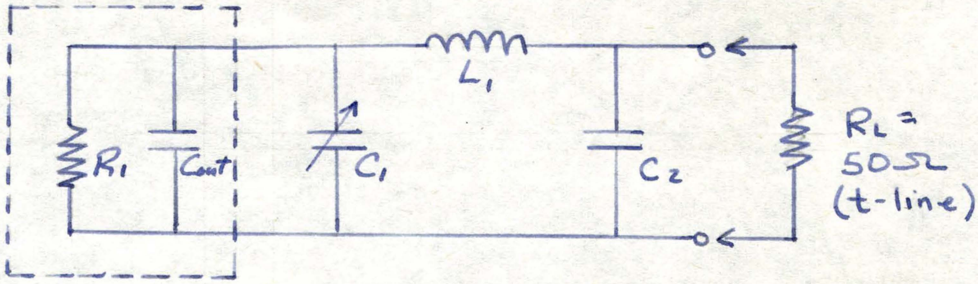


fig. 3

conjugate of Y_L

$$X_{C_1} = \frac{R_1}{Q} \qquad (11)$$

$$X_{C_2} = R_L \sqrt{\frac{R_1/R_L}{(Q^2+1)-(R_1/R_L)}} \qquad (12)$$

$$X_L = \frac{QR_1 + R_1R_L/X_{C_2}}{Q^2+1} \qquad (13)$$

calculated values:[15]

$C_1 = 26.8$ pf

$C_2 = 60.5$ pf

$L_1 = 332$ nH.

[15] see appendix D.

Input:

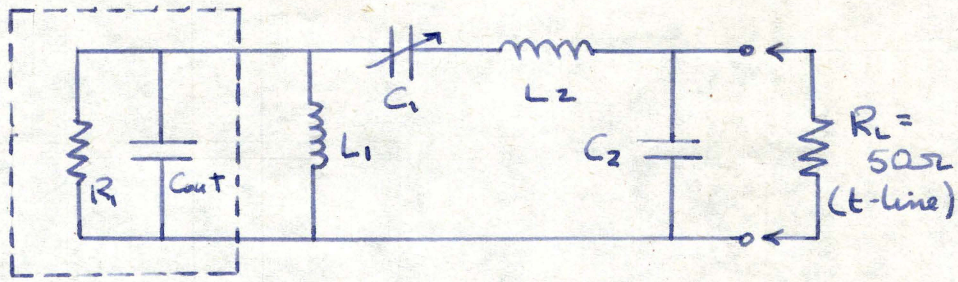


fig. 4

conjugate of Y_s

$$X_{L_1} = X_{C_{out}} \qquad (14)$$

$$X_{C_1} = Q R_1 \qquad (15)$$

$$X_{C_2} = R_L \sqrt{\frac{R_1}{R_L - R_1}} \qquad (16)$$

$$X_{L_2} = X_{C_1} + \frac{R_1 R_L}{X_{C_2}} \qquad (17)$$

calculated values:[16]

$L_1 = 48.3\ nH$

$L_2 = 590\ nH$

$C_1 = 15\ pF$ $\qquad$ $C_2 = 43.5\ pF$

[16] see appendix D.

IV Implementation

Now we must implement the matching networks into an amplifying circuit. It is necessary to meet the biasing conditions specified by the "y" parameters, ie. $V_{CE} = -5.0$ volts and $I_C = 10$ ma.

We chose $V_{CC} = 10.6$ volts and $R_E = 1\,K\Omega$ in order to acheive $I_C = 10$ ma (refer to fig. # 5). To obtain $V_{CE} = -5.0$ volts, we chose to supply $V_{EE} = -4.4$ volts thru the feedthrough capacitor and then thru an RFC whose impedance is far greater than that of the output network which is in parallel with it.

Using the basic equation $\omega_0^2 = \frac{1}{LC}$, we calculated the conjugate of each reactive element when resonating at the center frequency. then we used the Q-meter to find the desired element values, using trial and error until we acheive resonance with the calculated conjugate.

All inductors were hand-wound and adjusted until the desired value of inductance was obtained. ~~For~~ the 590 nH inductor, space limitations caused us to use a very fine insulated wire wound on a resistive carbon core for support.

Plastic cement was painted over the coils so that the original winding configuration and thus the desired measured value of inductance would not change.

An ideal RFC should have a low Q and a very high impedance compared to the output which it is in parallel with. Then we would have no change in the resonant frequency when connecting it into the circuit and also would isolate the power supply from ac signal.

The impedance of the choke was made far greater than the impedance of the output matching network which it looks into. Thus:

choose $Z_{RFC} \gg Z_l$

since $Y_l = 1.61 - j3.70 \doteq 4.0 \text{ m}\mho$

$Z_l = 250\,\Omega \Rightarrow Z_{RFC} = 1\text{K}\Omega$

then $\omega_o L_{RFC} = 1\text{K}\Omega$

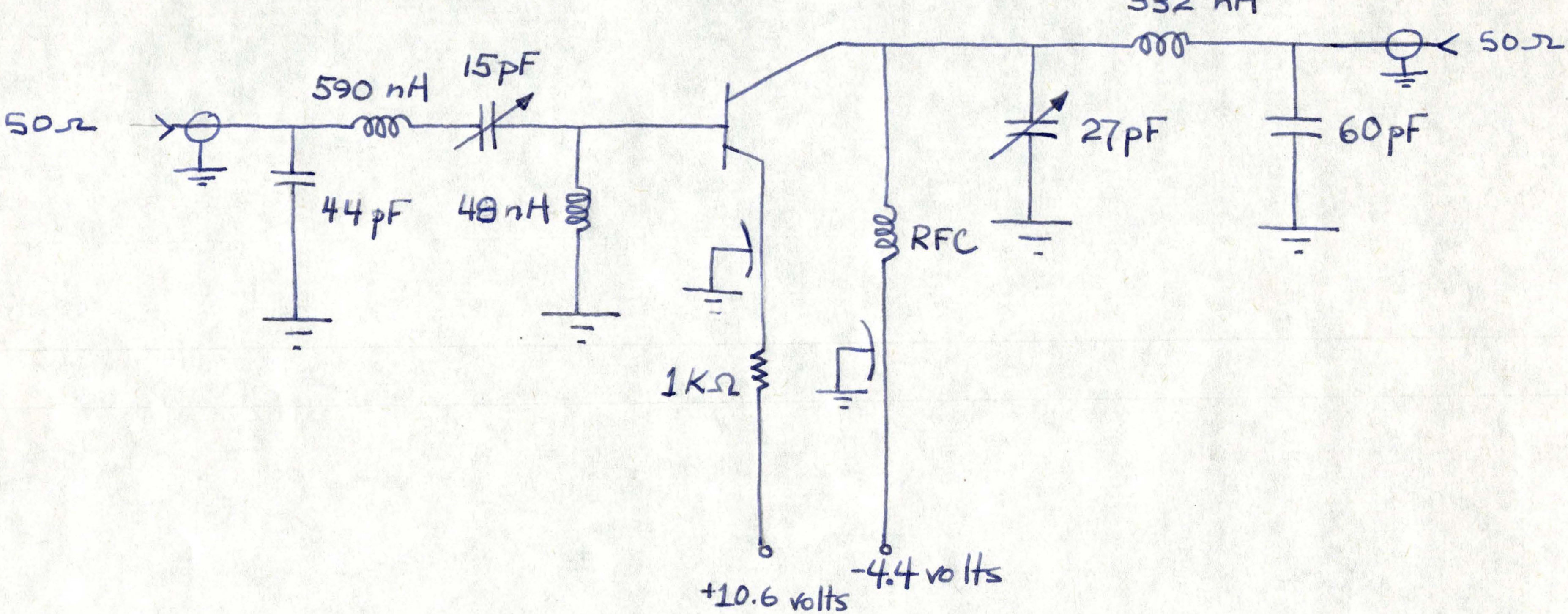
and $L_{RFC} = \dfrac{(10)^3}{2\pi(57)(10)^6} = 0.28\,\mu\text{H}$

The RFC chosen for the amplifier had the desired low Q and an inductance measured to be 0.40 μH.

A note must be made on the critical positioning of the inductor leads, especially for the air-core types. When measuring their values with the Q-meter any fractional variation of lead length during the reading can effect the inductance value by as much as fifty percent. This suggests the possibility of significant error in these element values unless extreme care is taken to place each inductor in the exact coil configuration in which it was measured.

Also, it is necessary to use a minimal amount of solder in each connection in order to avoid the introduction of spurious capacitance into the circuit.

Transistor placement necessitated the extension of one of the leads. Using the base lead was avoided in order to minimize spurious signal injection at the input.



TUNED VHF AMPLIFIER

CHANNEL 2

fig. #5

V Testing.

To evaluate the gain and bandwidth of the amplifier, sweep generator and signal generator - oscilliscope testing circuits were used (refer to fig. #7).

The Telonic Model No. HD-7 sweep generator has a sweep range from 100 KHz - 100 MHz and its circuitry permits precision and stability of the generated signal which we found to be far superior to that of the Marconi signal generator. For this reason we chose to tune the amplifier for center frequency, bandwidth and gain requirements using the sweep generator display.

Tuning was an iterative procedure where we strove to first acheive the maximum gain at the center frequency of 57 MHz and then meet the bandpass requirements of a channel 2 amplifier, ie. a 6 MHz bandwidth about the center frequency. An iterative technique was needed because of the interdependence between input and output impedances due to feedback. Thus tuning was done simultan-eously by adjusting first the output capacitor only slightly and then adjusting the input capacitor - all the time observing the display and attempting to acheive maximum gain at the center frequency with desired bandwidth.

The frequency response observed on the sweep generator display is plotted in fig.# 6.

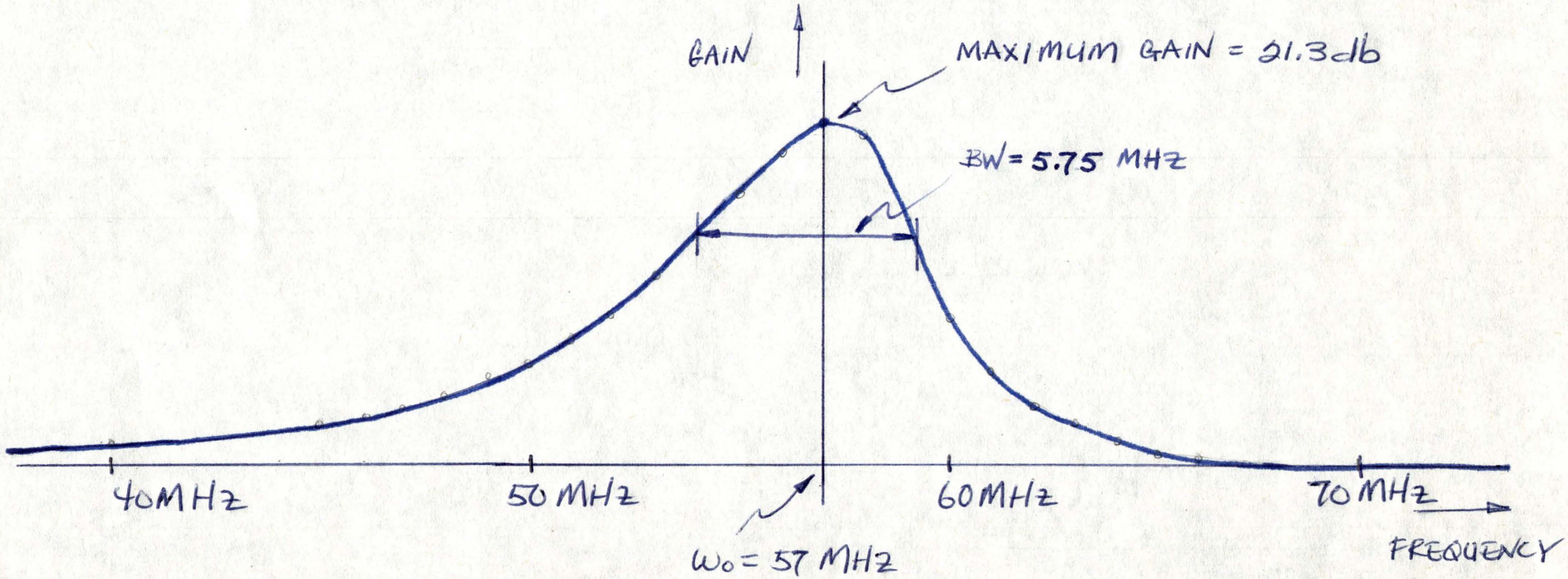
The amplifier's gain was measured using the Marconi signal generator and the Textronix Model 454 oscilliscope (refer to fig.# 7). The generator's output was calibrated on the display screen, and then fed into the RF amplifier whose output was connected into the oscilliscope. By reducing the attentuation on the generator's signal until the initial calibration level was acheived, the gain was measured to be 21.3 db (see fig.# 6).

Measuring the power supply output, they were set at 10.5 and -4.5 volts and current drain was 10.0 ma.

Thus we have met the bias conditions of $V_{CE} = 5$ volts and $I_C = 10$ ma and are assured that the "y" parameters values used in the matching networks' design are given by the transistor manufacturer's specifications.

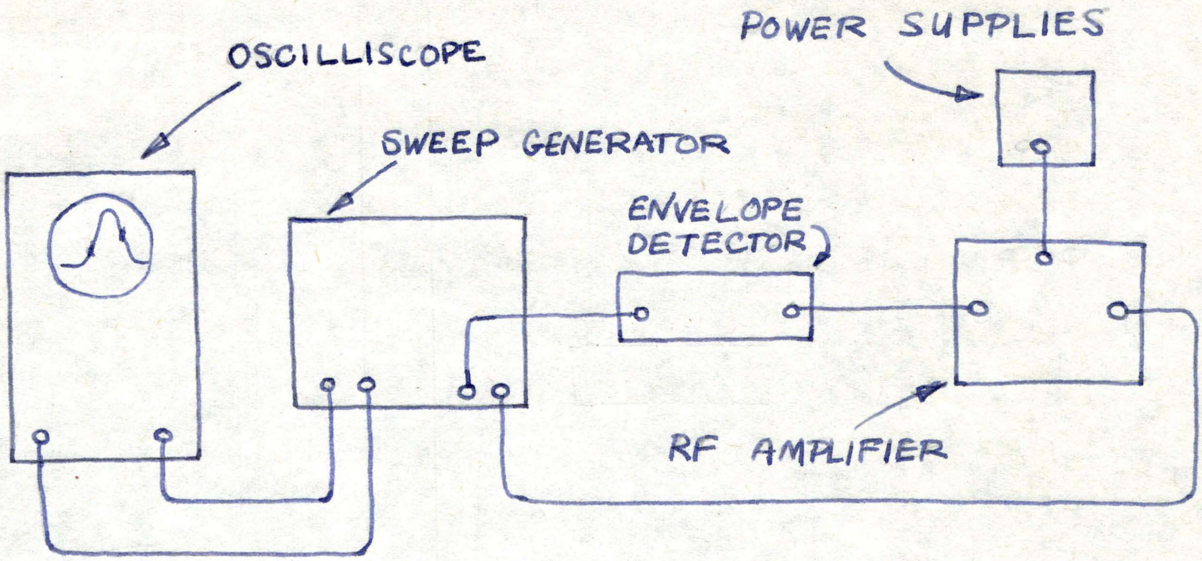
Varying the source voltages effects the output by reducing gain as long as the voltage is less than approximately 15 volts, i.e. as long as we avoid the transistors breakdown region the amplifier will not go unstable.

Also, by mismatching the network in the bandpass region away from the center frequency the amplifier still displayed complete stability.

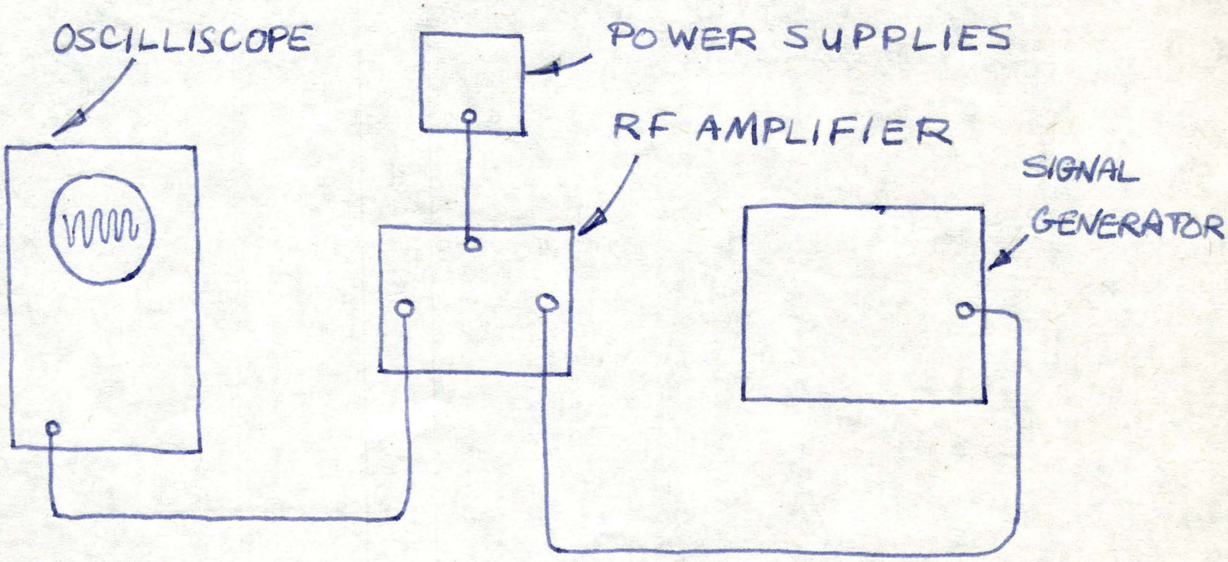


FREQUENCY RESPONSE OF VHF AMFLIFIER
CHANNEL 2

fig. #6



SWEEP GENERATOR TEST CIRCUIT
- VISUAL TUNING
- MEASURE BANDWIDTH



SIGNAL GENERATOR TEST CIRCUIT
- MEASURE GAIN

fig. #7

Amplifier specifications:

center frequency = 57 MHz

bandwidth = 5.75 MHz

gain = 21.3 db

input impedance = 50 Ω

output impedance = 50 Ω

current drain = 10 ma

supply voltages + 10.5 volts
- 4.5 volts

VI Conclusions:

The RF amplifier preformed close to specifications, with maximum gain only 0.3 db below the optimum and with complete stability throughout the passband. The bandwidth, which was considered to be of secondary importance since the selectivity of a receiver is not usually determined by the RF stage, was slightly less than the specified passband.

From the sweep generator display it was observed that when the frequency response was tuned to a completely symmetrical waveform, the bandwidth was only 4.5 MHz. This suggests decreasing the circuit Q to improve bandwidth.

Thus we have shown that the theory discussed can be implemented in the design of a practical amplifier.

Appendix A.

The three commonly used forms of power gain are[17]:

1.) transducer power gain (G_T)
defined as: the ratio of average power delivered to the load to the maximum available average power at the source.

2.) power gain (G_P)
defined as: the ratio of the average power delivered to the load to the average power entering the input port.

3) available power gain (G_A)
defined as: the ratio of the maximum available power at the output port to the maximum available power at the source.

It can be shown that G_T is a function of both the load and source terminations while G_P and G_A are only functions of the load and source terminations respectively. Due to the more restrictive definition

[17] see for example: E.S. Kuh and R.A. Rohrer, Theory of Linear Active Networks, San Francisco, Holden-Day Inc., pp. 236-239

placed upon G_T this form of power gain is the most meaningful and is used most often.

We should add that when both the input and the output are conjugately matched the three forms of power gain are numerically equal. That is:

$$G_T = G_A = G_P \Big| \text{input \& output both conj. matched.}$$

G_T is given by:

$$G_T = \frac{4 |y_{21}|^2 \, Re\, Y_l \, Re\, Y_s}{|(y_{11}+Y_s)(y_{22}+Y_l) - y_{12}\, y_{21}|^2}$$

Appendix B.

Here we shall utilize the fact that when both the input and output are conjugately matched all three forms of power gain are equal.

Since G_P is a function of Y_L we can maximize G_P with respect to Y_L by setting $\frac{\partial G_P}{\partial Y_L}$ equal to zero.

This then gives us the optimum Y_L.

Similarly since G_A is a function of Y_S we can solve for the optimum Y_S.

Then since both the input and output are conjugately matched and G_A and G_P are maximum; G_T is also maximum

A more complete derivation is given in Kuh and Rohrer. Refer to p. 8 of thesis for the page numbers.

Appendix C

T.V. FREQUENCY ALLOCATIONS:

Channel	Frequency (MHz)
2	54-60
3	60-66
4	66-72
5	76-82
6	82-88
7	174-180
8	180-186
9	186-192
10	192-198
11	198-204
12	204-210
13	210-216

Appendix D

A: Linville Stability

$$C = \frac{|y_{12}\, y_{21}|}{2g_{11}g_{22} - Re\{y_{12}\, y_{21}\}}$$

$$= \frac{|-142.5 - j52.5|}{2(12)(0.6) + 142.5}$$

$$= \frac{152}{14.4 + 142.5} = \underline{\underline{.97}}$$

B: Terminations:

a) using formulae (6) to (10):

$y_{12}\, y_{21} = -142.5 - j52.5 = M + jN$

$|y_{12} y_{21}| = L = 152$; $2g_{11} = 24$; $2g_{22} = 1.2$

$$B_L = \frac{-52.5}{24} - 1.51 = -2.19 - 1.51 = -3.70\, m\mho$$

$$G_L = \frac{1}{24}\sqrt{(14.4 + 142.5)^2 - 152^2}$$

$$= \frac{1}{24}\sqrt{(156.9)^2 - (152)^2}$$

$$= \frac{100}{24}\sqrt{.15} = 1.61 \text{ m℧}$$

$$\therefore \underline{Y_L = 1.61 - j3.70 \text{ m℧}}.$$

$$B_s = -\frac{52.5}{1.2} - 14 = -57.7 \text{ m℧}$$

$$G_s = \frac{100}{1.2}\sqrt{.15} = 32.2 \text{ m℧}$$

$$\therefore \underline{Y_s = 32.2 - j57.7 \text{ m℧}}$$

b) Iteration

if we assume $Y_L = y_{22}^*$ we get the following values:

Y_L (m℧)	Y_s (m℧)
$0.6 - j1.5$	$130.8 - j57.7$
$1.3 - j2.2$	$69.1 - j62.6$
$1.61 - j2.76$	$50.3 - j59.6$
$1.69 - j3.14$	$42.3 - j58.6$
$1.69 - j3.36$	$38.4 - j58.3$
$1.68 - j3.5$	$35.9 - j57.9$
$1.67 - j3.58$	$34.6 - j57.8$

C: Stern Stability

$$K = \frac{2(12+32.2)(0.6+1.61)}{152-142.5}$$

$$= \frac{2(44.2)(2.21)}{9.5}$$

$$= \underline{20.6}$$

D: G_T

$$G_T = \frac{4(2.81\times10^4)(1.61)(32.2)}{|(44.2-j\,43.7)(2.21-j\,2.2)+142.5+j\,52.5|^2}$$

$$= \frac{5.83\times10^6}{|144-j\,141.4|^2} = \frac{5.86\times10^6}{4.07\times10^4}$$

$$= 143$$

OR

$$G_{T_{db}} = 10\log 143 = 10(2.16) = \underline{21.6db}$$

E: Output Matching:

$$X_{C_1} = \frac{10^3}{(6)(1.61)} = 104\,\Omega \qquad R_1 = 621\,\Omega$$

$$\therefore\ C_1 = \frac{1}{2\pi \times 57 \times 10^6 \times 104} = \underline{26.8\ pf}$$

$$X_{C_2} = 50\sqrt{\frac{621/50}{37 - \frac{621}{50}}}$$

$$= 50\sqrt{\frac{12.4}{14.6}} = 46.1\,\Omega$$

$$\therefore\ C_2 = \frac{1}{(3.58 \times 10^8)(46.1)} = \underline{60.5\,pf}$$

$$X_L = \frac{6(621) + \frac{(50)(621)}{46.1}}{37}$$

$$= \frac{3726 + 675}{37} = 119\,\Omega$$

$$\therefore\ L = \frac{119}{358 \times 10^6} = \underline{332\,nH}$$

F: Input Matching:

$R_1 = 31.1\,\Omega$

$X_{L_1} = 17.3\,\Omega$ $\qquad X_{C_{out}} = 17.3\,\Omega$

$$\therefore L_1 = \frac{17.3}{3.58\times10^8} = \underline{48.3\,nH}$$

$$X_{C_1} = 6(31.1) = 186.6\,\Omega$$

$$\therefore C_1 = \frac{1}{(186.6)(3.58\times10^8)} = \underline{15\,pF}$$

$$X_{C_2} = 50\sqrt{\frac{31.1}{50-31.1}} = 64.2\,\Omega$$

$$\therefore C_2 = \frac{1}{(64.2)(3.58\times10^8)} = \underline{43.5\,pF}$$

$$X_{L_2} = 186.6 + \frac{(31.1)(50)}{64.2} = 210.8\,\Omega$$

$$\therefore L_2 = \frac{210.8}{358\times10^6} = \underline{590\,nH}$$

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by.

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