A COMMUNICATIONS RECEIVER

FOR 9 TO 12 MHZ

EE 399

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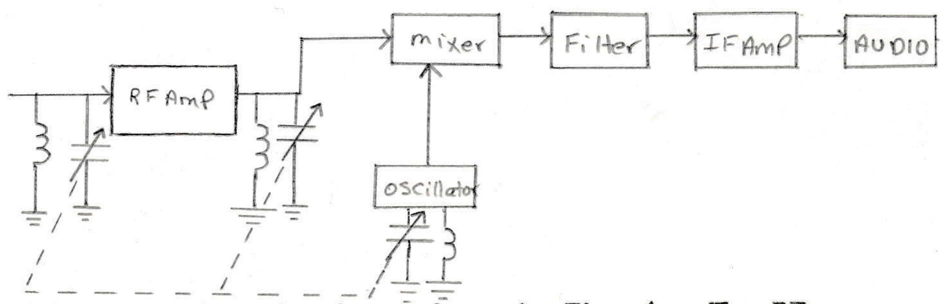
The Superheterodyne Receiver - An Overview

The superheterodyne receiver is generally considered to be the most appropriate circuit for reception and demodulation of electromagnetic waves containing information. The distinguishing characteristic of this receiver type is that all incoming signals are mixed with the output of a local oscillator and the difference frequency is selected and amplified by intermediate frequency (IF) amplifiers. Excellent performance is obtainable because these amplifiers remain at a fixed frequency and only the RF amplifier and local oscillator need be tunable. Additional stability is inherent in the design due to the fact that the gain is concentrated at two (single conversion) or sometimes three (double conversion) different frequencies.

Superheterodyne

Receiver

Fig. 1



A block diagram of the receiver is shown above in Fig. 1. The RF amplifier should have a gain of approximately 10 dB in order to establish the overall noise figure of the receiver. Low noise design of this stage is essential because noise generated here propagates throughout the entire receiver. The RF amplifier also serves as a buffer to prevent local oscillator radiation out to the antenna.

The mixer and local oscillator heterodyne the input rf signal from the RF amplifier down to the IF frequency for further amplification. The nonlinearities of the mixer create numerous intermodulation products, one occuring at the IF frequency.

The mixer may be passive (diode mixer) or active (FET or transistor) and provide gain from -10 dB to as high as 30 dB.

The IF amplifier and filter provide the overall bandwidth and adjacent channel selectivity of the receiver. The majority of the receiver's gain is incorporated here, integrated with some form of automatic gain control to provide good dynamic range for all input signal levels. The frequency of the IF amplifiers is usually low (455 kHz to 10.7 MHz) where high signal gain is possible.

The demodulator recovers the information from the IF signal and is of the necessary form for the signal type being received (AM, FM, SSB). This stage may also provide audio-derived AGC to other receiver stages.

Although the superhet receiver offers exceptional performance when compared to a superregenerative or tuned radio frequency receiver, careful design is necessary throughout. The frequency chosen for the IF influences the image rejection of the receiver. Lower IF's provide lower image rejection, higher signal gain and greater selectivity. Higher IF's provide improved image rejection but sacrifice other criteria to be gained at lower IF's. The mixer and RF amplifier stages are the major contributors of noise in the system so each must be of low-noise design to enhance the receivers signal-to-noise ratio. In double conversion receivers, where a second heterodyne mixer is present, noise generated within the mixers is a significant component influencing receiver performance. In its most widespread application as a communications receiver, its advantages outweigh its disadvantages. With the phase-lock loop technology now available, direct conversion receivers are becoming more popular as alternatives to some applications which had in the past employed superheterodyne receivers.

A COMMUNICATIONS RECEIVER FOR 9 TO 12 MHZ

In an attempt to further my knowledge in my primary area of interest, RF small signal design, this receiver, whose description and development follow, was first conjectured in December, 1981. Considerable effort was made to develop the entire design using the engineering tools I had at that time, as well as investigate other unknown methods which would provide a good engineering approach and a satisfactorily functioning receiver. The most powerful method found applicable to the receiver's design was the use of two-port admittance parameters in the RF small signal design portion of the project.

DESIGN WITH TWO-PORT PARAMETERS

Design of solid state, small-signal RF amplifiers is a systematic, mathematical procedure with the use of two-port parameters. With this technique an exact solution is available for the complete design problem, the only sources of error being due to parameter variations and strays in the physical circuit. Two-port parameters have the advantage of being applicable to any linear active device and so readily describe the MC1350 integrated circuits used in the IF amplifiers of the receiver.

The major factor in the overall design is the potential stability of the device being used. The stability may be determined by evaluation of the Linville Stability Factor, C, by the use of expression (1).

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

When C is less than one the device is unconditionally stable. When C is greater than one the device is potentially unstable, the relative stability being dependent upon the source and load connected to the device. The C-factor is a test for stability under hypothetical worst-case conditions; the input and output open-circuited. If the configuration is potentially unstable, neutralization, unilateralization or the use of restricted source and load values is in order for a stable configuration.

Another stability factor which is used in the development of a relation describing the maximum gain possible with a matched load is the Rollet Stability Factor, (2). Unconditional stability occurs for kr greater than one.

$$kr = \frac{2g_{11}g_{22} - \mathrm{Re}\,(y_{12}y_{21})}{|y_{12}y_{21}|} \qquad (2)$$

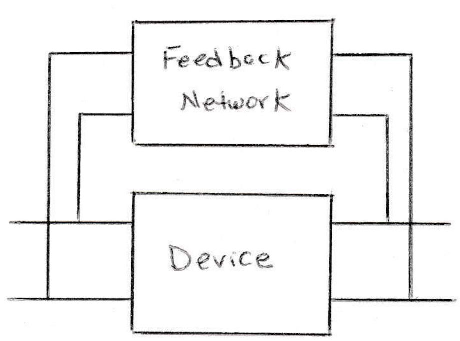
The Stern Stability Factor, which finds usage in equations (17) through (24) denotes unconditional stability when K is greater than one.

$$K = \frac{2(g_{11} + Gs)(g_{22} + Gl)}{|y_{12}y_{21}| + \mathrm{Re}(y_{12}y_{21})} \qquad (3)$$

In cases where unilateralization or neutralization are necessary to realize unconditional stability, composite admittance parameters (Fig. 1) may be calculated and used in all relationships to accurately describe the device and feedback network. Network losses, however, are not taken into consideration.

Composite
Admittance
Parameters

Fig. 1



$$y_{11c} = y_{11d} + y_{11f}$$
$$y_{12c} = y_{12d} + y_{12f}$$
$$y_{21c} = y_{21d} + y_{21f}$$
$$y_{22c} = y_{22d} + y_{22f}$$

A general expression for power gain is

$$G = \frac{\left|y_{21}\right|^2 \mathrm{Re}(Yl)}{\left|Yl + y_{22}\right|^2 \mathrm{Re}\left(y_{11} - \frac{y_{12}y_{21}}{y_{22} + Yl} \right)} \tag{4}$$

Equation (4) does not include any effect due to Ys, the source admittance. A more complete relationship which includes the effects of the degree of input and output matching is the equation for transducer gain, (5).

$$G_t = \frac{4\ \mathrm{Re}(Ys)\ \mathrm{Re}(Yl)\left|y_{21}\right|^2}{\left|(y_{11} + Ys)(y_{22} + Yl) - y_{12}y_{21}\right|^2} \tag{5}$$

The maximum available gain is the theoretical power gain of a linear device with its reverse transfer admittance y_{12} set equal to zero and its source and load admittances conjugately matched to y_{11} and y_{22} respectively.

$$MAG = \frac{\left|y_{21}\right|^2}{4\ \mathrm{Re}(y_{11})\ \mathrm{Re}(y_{22})} \tag{6}$$

The maximum gain possible by choosing only Yl optimally is given by (7) where kr is the Rollet Stability Factor of (2).

$$G_{max} = \left|\frac{y_{21}}{y_{12}}\right| \left[kr - \sqrt{(kr)^2 - 1} \right] \tag{7}$$

Expressions for the input and output admittance of a linear device are described by (8) and (9).

$$Y_{in} = y_{11} - \frac{y_{12}y_{21}}{y_{22} + Yl} \tag{8}$$

$$Y_{out} = y_{22} - \frac{y_{12}y_{21}}{y_{11} + Ys} \tag{9}$$

If the maximum possible power gain without feedback is desired for an amplifier, the following relationships determine the source and load admittances required to achieve maximum transducer gain.

$$G_s = \frac{1}{2\,\mathrm{Re}(y_{22})}\left[\left[2\,\mathrm{Re}(y_{11})\mathrm{Re}(y_{22}) - \mathrm{Re}(y_{12}y_{21})\right]^2 - \left|y_{12}y_{21}\right|^2\right]^{\frac{1}{2}} \quad (10)$$

$$B_s = -\mathrm{Im}(y_{11}) + \frac{\mathrm{Im}(y_{21}y_{12})}{2\,\mathrm{Re}(y_{22})} \quad (11)$$

$$G_1 = \frac{1}{2\,\mathrm{Re}(y_{11})}\left[\left[2\,\mathrm{Re}(y_{11})\mathrm{Re}(y_{22}) - \mathrm{Re}(y_{12}y_{21})\right]^2 - \left|y_{12}y_{21}\right|^2\right]^{\frac{1}{2}} \quad (12)$$

$$B_1 = -\mathrm{Im}(y_{22}) + \frac{\mathrm{Im}(y_{21}y_{12})}{2\,\mathrm{Re}(y_{11})} \quad (13)$$

The case of potential unstability may be handled, as pointed out earlier, by employing an external feedback network to achieve a composite y_{12} of zero. The necessary design equations for the unilateralized case are derived by first evaluating the composite y parameters for the linear device and feedback network combination. Making the proper substitutions give equations (14), (15), and (16).

A general expression for unilateralized powgr gain is

$$G_{pu} = \frac{\left|y_{21} - y_{12}\right|^2 \mathrm{Re}(Yl)}{\left|Yl + y_{22} + y_{12}\right|^2 \mathrm{Re}(y_{11})} \quad (14)$$

The unilateralized power gain with Yl conjugately matched to Yout is given by

$$G_u = \frac{\left|y_{21} - y_{12}\right|^2}{4\,\mathrm{Re}(y_{11} + y_{12})\,\mathrm{Re}(y_{22} + y_{12})} \quad (15)$$

Unilateralized transducer gain is given by

$$G_{tu} = \frac{4\ \mathrm{Re}(Ys)\ \mathrm{Re}(Yl)\left|y_{21}-y_{12}\right|^2}{\left|(y_{11} + y_{12} + Ys)(y_{22} + y_{12} + Yl)\right|^2} \tag{16}$$

Equations (1) through (16) were all that were necessary in the receiver design due to the unconditional stability of the configuration used. In the case of potential unstability, equations for computing the conductance and susceptance of both Ys and Yl for maximum power gain and a particular Stern Stability Factor, k, are listed below.

$$Gs = \sqrt{\frac{k\left[\left|y_{12}y_{21}\right| + \mathrm{Re}(y_{12}y_{21})\right]}{2}} \cdot \sqrt{\frac{g_{11}}{g_{22}}} - g_{11} \tag{17}$$

$$Gl = \sqrt{\frac{k\left[\left|y_{12}y_{21}\right| + \mathrm{Re}(y_{12}y_{21})\right]}{2}} \cdot \sqrt{\frac{g_{22}}{g_{11}}} - g_{22} \tag{18}$$

$$Bs = \frac{(Gs + g_{11})\ Z_o}{\sqrt{k\left[\left|y_{12}y_{21}\right| + \mathrm{Re}(y_{12}y_{21})\right]}} - b_{11} \tag{19}$$

$$Bl = \frac{(Gl + g_{22})\ Z_o}{\sqrt{k\left[\left|y_{12}y_{21}\right| + \mathrm{Re}(y_{12}y_{21})\right]}} - b_{22} \tag{20}$$

Z_o is the real value of Z which results in the smallest minimum of the following relationship.

$$Z^3 + \left[k(L + M) + 2M\right] Z - 2N\sqrt{k(L + M)} = 0 \tag{21}$$

$$Z = \frac{(Bs + b_{11})(Gl + g_{22}) + (Bl + b_{22})k(L + M)/2(Gl + g_{22})}{\sqrt{k\ (L + M)}} \tag{22}$$

$L = y_{12}y_{21}$

$M = \mathrm{Re}(y_{12}y_{21})$

One other alternative available to the designer to obtain a stable configuration involves mismatching Gs to g_{11} and Gl to g_{22} by an equal ratio. If a mismatch ratio, R, is defined as

$$R = \frac{Gl}{g_{22}} = \frac{Gs}{g_{11}} \qquad (25)$$

then R may be computed for any particular circuit stability factor using the equation

$$(1 + R)^2 = k\left[\frac{|y_{21}y_{12}| + \mathrm{Re}(y_{12}y_{21})}{2g_{11}g_{22}}\right] \qquad (26)$$

The power of these mathematical equations is the alternative use of hybrid, impedance, or scattering parameters and the appropriate conversion equations between parameter types. A correct mathematical design approach is made available from very low frequencies through the microwave range.

A program suitable for the HP-41C calculator was written for all relationships thus far developed exclusive of (17) through (24). Extensive use of this program was used to develope the receiver design. The program is included at the end of this paper.

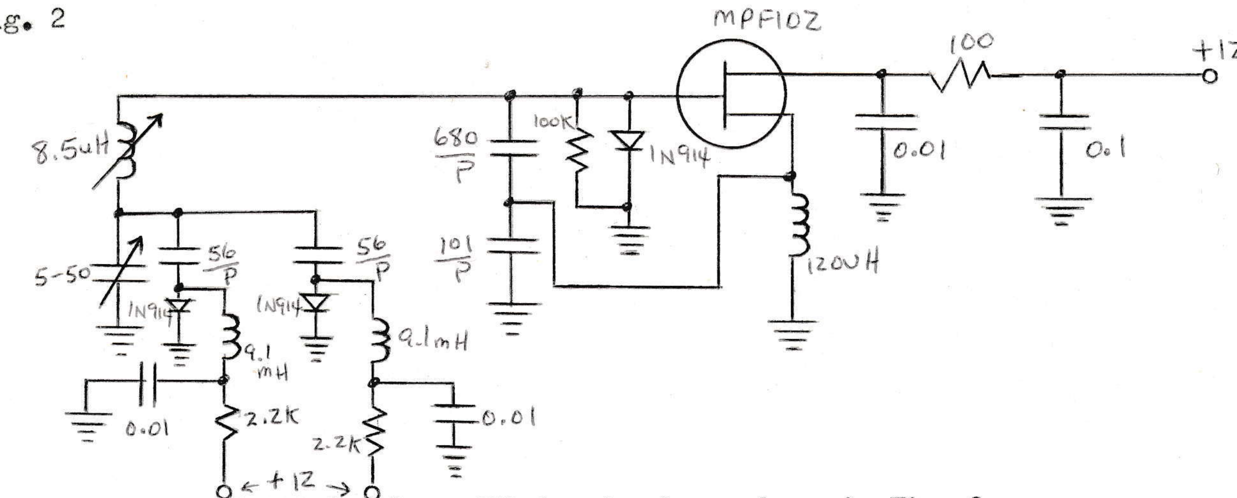
THE ACTUAL DESIGN

A series-tuned Colpitts oscillator with two intermediate stages of buffering is used in the receiver as the local oscillator. Two JFETs, MPF102, and a 2N2222A transistor are used in the configuration with diode switching in the tank circuit for multiple band coverage.

Colpitts Oscillator with

Diode Switching

Fig. 2



The main portion of the local oscillator is shown above in Fig. 2. The equation describing the frequency of operation of the oscillator tank is

$$w_o = \sqrt{\frac{1}{L}\left(\frac{1}{C_1} + \frac{1}{C_2}\right)}$$

C_1 = Total capacitance in series with the inductor

C_2 = Total capacitance of capacitive feedback network

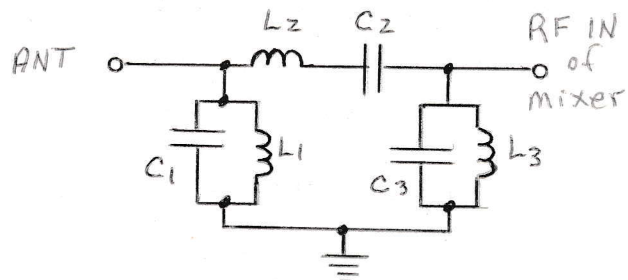
The 9.1 mH chokes in the tank serve to decouple the switching voltage applied to the 1N914 diodes in the tank circuit. The 680 pF and 101 pF capacitors affect the frequency of operation and also establish the proper amount of feedback for stable operation. The source follower has decoupling on its drain in the form of 0.01 and 0.1 uF capacitors and 100 ohm resistor to prevent rf from entering the power supply. Output is developed across the 120 uH inductor and capacitively coupled to a second source follower which provides a light load for the oscillator (see Fig. 9). The 2N2222A in a common emitter configuration develops the output signal which is applied to the local oscillator input of the MC1596G mixer.

The first stage encountered by a received signal is a three-section Chebycheff bandpass filter with a passband of 8 to 12 MHz.

This is incorporated in the receiver's front end to minimize overloading and signal desense due to strong local AM and FM signals. The filter also provides a measure of selectivity to out-of-band signals which would not be present due to the absence of a tuned front end in this receiver's design. The circuit diagram for the filter is shown in Fig. 3 and its passband characteristic in Fig. 4.

Three Section Chebycheff Filter

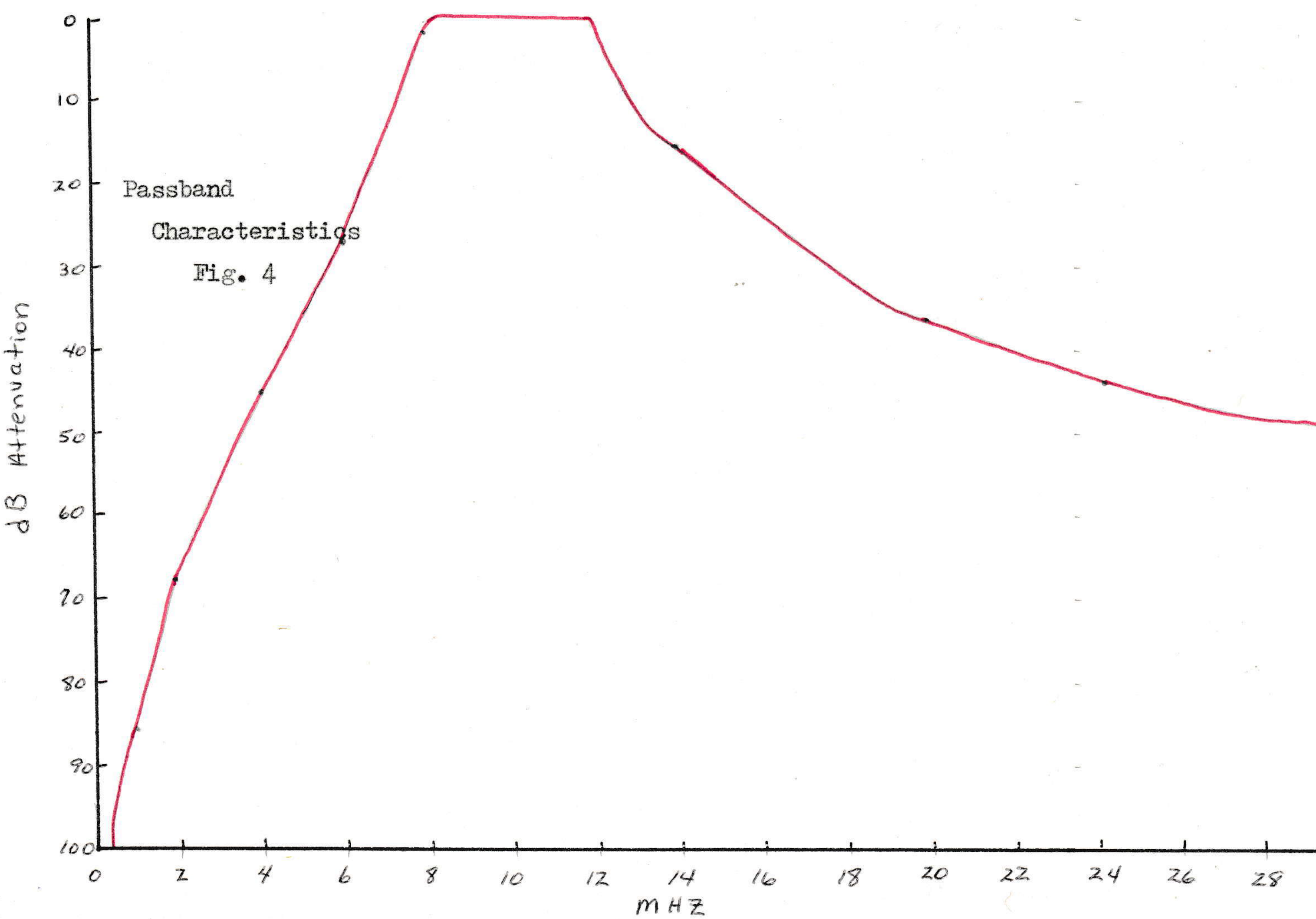
Fig. 3



C1 = 1270 pF
L1 = 0.20 uH

C2 = 116 pF
L2 = 2.182 uH

C3 = 1270 pF
L3 = 0.20 uH



Passband Characteristics

Fig. 4

The MC1596G is an excellent device for use as a balanced modulator or demodulator and provides high common mode rejection in addition to adjustable gain. The output collectors of the device are cross coupled so that full wave balanced multiplication of the two input signals occurs. The direct result is an output spectrum consisting only of the sum and difference of the two input frequencies. The approximate voltage gain for the MC1596 in low-level ac applications is

$$\text{Voltage Gain} = \frac{R_L\, V_c(\text{rms})}{2\sqrt{2}\,\frac{kT}{q}\,(R_E + 2r_e)} \qquad (28)$$

R_E = resistance between pins 2 and three

r_e = 26 mv/I_E (ma)

kT/q = 26 mv at room temperature

The use of (28) gives a voltage gain of 5 dB for a 50 microvolt input signal.

The input admittance of the MC1596 is extremely invariant in the frequency range of interest (Table 1) so no form of matching is necessary. The input capacitance is so low that no attempt is made to cancel it with an equal amount of inductive reactance.

Input Characteristics

Table 1

Frequency	Input Impedance	
5 MHz	240 K ohms	2.05 pF
10 MHz	1 Meg ohm	2.05 pF
15 MHz	240 K ohms	2.05 pF

The necessary biasing for the MC1596 is derived from information provided by an application note on the MC1596 and MC1496 by Motorola. The output of the mixer remains at 455 kHz and must be matched to the next 3000 ohm stage, the 455 kHz filter, for maximum signal transfer.

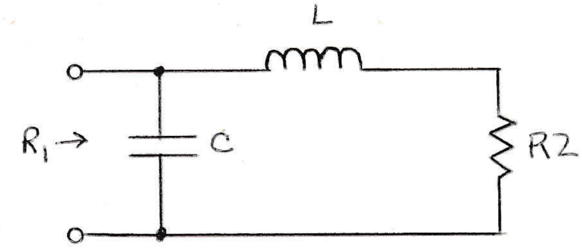
Several widely used impedance matching configurations and their corresponding equations are shown below.

(29A)

$$R_1 > R_2$$

$$X_L = \sqrt{R_1 R_2 - R_2^2}$$

$$X_C = \frac{R_1 R_2}{X_L}$$

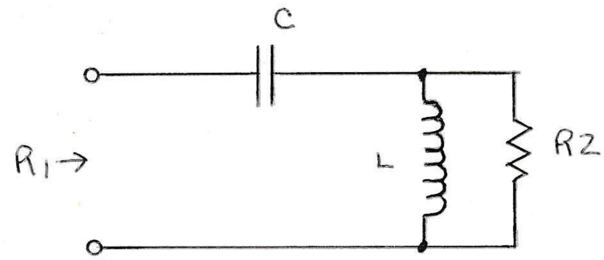


(29B)

$$R_2 > R_1$$

$$X_L = R_2 \sqrt{\frac{R_1}{R_2 - R_1}}$$

$$X_C = \frac{R_1 R_2}{X_L}$$



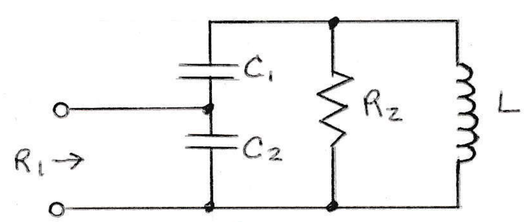
(29C)

$$R_1 < R_2 \text{ , } N > \sqrt{R_2/R_1 - 1}$$

$$X_L = R_2 / N$$

$$X_{C2} = \frac{R_1}{\sqrt{\frac{R_1(N^2 + 1)}{R_2} - 1}}$$

$$X_{C1} = \frac{R_2 N}{N^2 + 1} \left(1 - \frac{R_1}{N X_{C2}} \right)$$

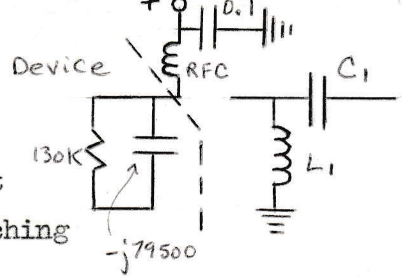


The matching scheme of (29B) is used to match the output of the mixer because it conveniently allows use of one inductor for both power line isolation and impedance matching.

Mixer

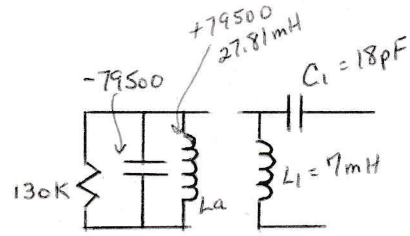
Output

Matching



Match 130 K ohms to 3 K ohms

Fig. 5



La cancels output capacitive reactance

$$X_{L1} = 130\text{ K}\sqrt{\frac{3\text{ K}}{130\text{ K} - 3\text{ K}}} = 19{,}980 \text{ ohms}$$

$$X_{La} = 79{,}500 \text{ ohms}$$

$$L_1 = 7 \text{ mH.}$$

$$La = 27.81 \text{ mH}$$

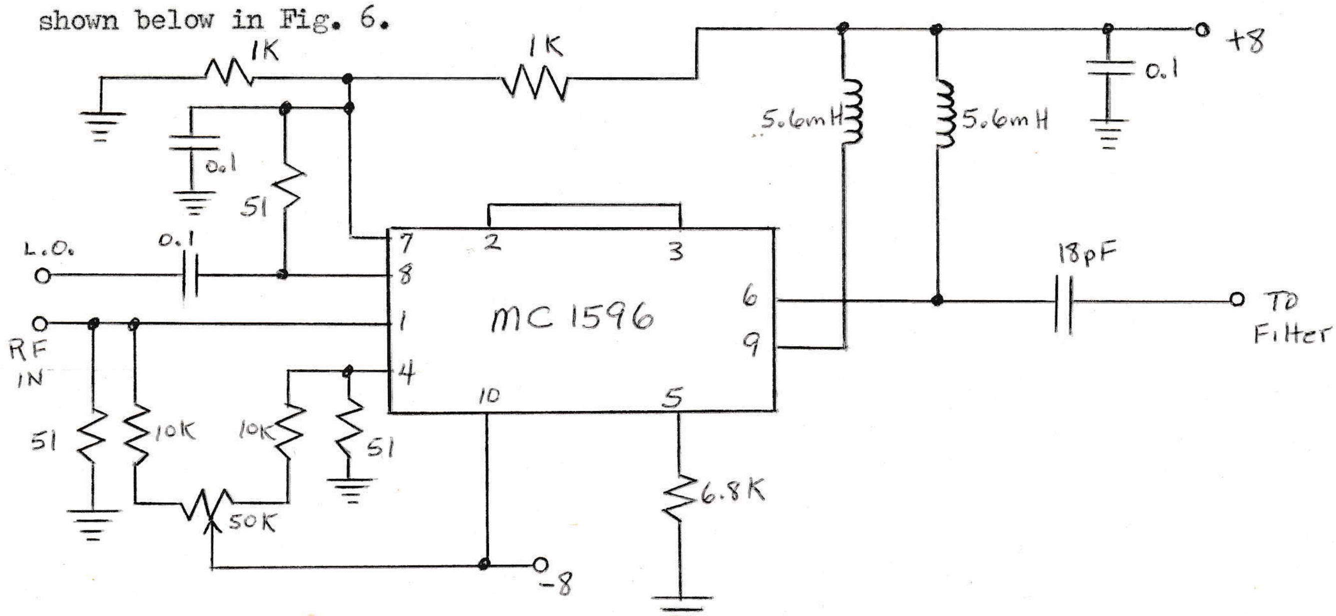
$$X_C = R_1R_2 / X_L = 19520 \text{ ohms}$$

$$C_1 = 18 \text{ pF}$$

Combining the two inductors yields

$$\frac{(27.81)(7)}{27.81 + 7} = 5.6 \text{ mH}$$

The mixer stage with its associated biasing and input-output matching is shown below in Fig. 6.



Mixer Fig. 6

The IF amplifier consists of two Motorola MC1350 integrated circuits in cascade with an automatic gain control signal applied from a CA3160 differential amplifier. The MC1350 has nearly constant input and output admittances over its entire AGC range and has a very low reverse transfer admittance. Admittance values for the device at 455 kHz are listed in Table 2. The reverse transfer admittance, g_{12}, is not known exactly, but it does not significantly affect the gain of the device and has only a minimal effect on b_{11}, the input susceptance.

Admittance Parameters for MC1350 at 455 kHz

Table 2

g_{11}	0.31	mmhos	.00031078 ∠4.0593
b_{11}	0.022	mmhos	
g_{12}	$1(10^{-9})$	mmhos	
b_{12}	0		
g_{21}	0.1594	mmhos	.0001600 ∠-4.99796
b_{21}	-0.01394	mmhos	
g_{22}	$4(10^{-6})$	mmhos	5×10^{-9} ∠36.87
b_{22}	$3(10^{-6})$	mmhos	

Using the admittance values in Table 2 and the appropriate mathematical equations already developed, the following operational characteristics are obtained.

Equation	Data
(2)	Unconditional Stability
(2)	Rollet Stability 14.504
(7)	G_{max} 67.42 dB
(3)	Stern Stability Factor 29.03
(10),(11), (12),(13)	Optimum source and load to achieve maximum power gain

$$Ys = 2.894\ (10^{-4})\ -j\ 2.374\ (10^{-5})\ \text{mhos}$$
$$Yl = 3.734\ (10^{-6})\ -j\ 3.0225\ (10^{-6})\ \text{mhos}$$

The optimum source and load in parallel equivalents are

37.42 dB

$$Zs = 3432 + j\ 282\ \text{ohms}$$
$$Zl = 161.8\ K + j\ 131\ K\ \text{ohms}$$

Coupling from the 455 kHz crystal filter to the input of the first MC1350 is provided by a series 0.05 uF capacitor. No attempt of matching is made because the small mismatch present contributes theoretically to a degradation in gain of only 0.05 dB (5). The output of the first IF stage is transformer coupled by a primary-tuned toroidal circuit to the second stage input. Proper resonance at 455 kHz is provided by an adjustable trimmer capacitor in parallel with the primary. A toroid with a tapped primary is used to provide a load of 30 K ohms for the first stage and also allows dc operating voltages to be applied to the MC1350.

Earlier it was determined that the optimum load for the first stage is 161.8 K + j 131 K ohms. A simple calculation shows that only 3 pF of stray capacitance has nearly 120 K ohms of reactance so, in order to maintain good stability, a lower value of load impedance is required. Use of (5) shows that with a load of 30 K ohms and the Ys previously defined, gain is 63 dB which is more than adequate. The small amount of loss introduced by this mismatch will be very much worth the added stability gained by the lower load impedance.

The actual design of the RF section began with the second IF (progressing toward mixer) and a Z_L for the second stage of 10 K ohms was chosen. Using equations (5) and (8) gives the optimum source for the second IF stage which is

$$Yin = 3.0847\ (10^{-4}) + j\ 2.2178\ (10^{-5})\ \text{mhos}$$

or

$$Zin = 3225 - j\ 232\ \text{ohms}$$

Returning to the IF transformer calculation,

$$\frac{N_p}{N_s} = \sqrt{\frac{Z_p}{Z_s}} \qquad (30)$$

N_p = primary turns Z_p = primary impedance

N_s = secondary turns Z_s = secondary impedance

$$\sqrt{\frac{30,000}{3225}} = \text{Turns Ratio} = 3.05$$

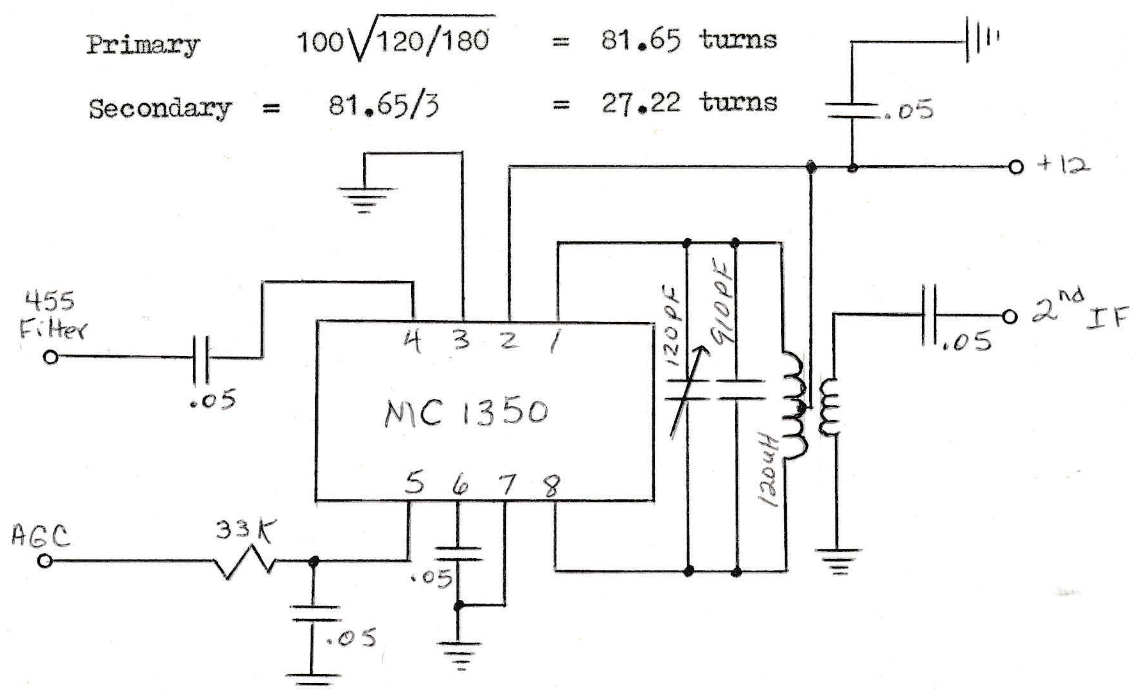
A primary inductance of 120 uH is used and data provided by Amidon Associates with the toroid allows proper determination of primary and secondary turns.

A 15 - mix, T - 68 toroid is used. This material has a permeability u = 25 and a usable frequency range of 0.1 - 2.0 MHz. A_L (uH/ 100 turns) for this core size is 180 uH.

$$\# \text{ turns} = 100 \sqrt{\text{desired L (uH)} / A_L \text{ value (uH)}}$$

Primary $100\sqrt{120/180}$ = 81.65 turns

Secondary = 81.65/3 = 27.22 turns



First IF Stage

Fig. 7

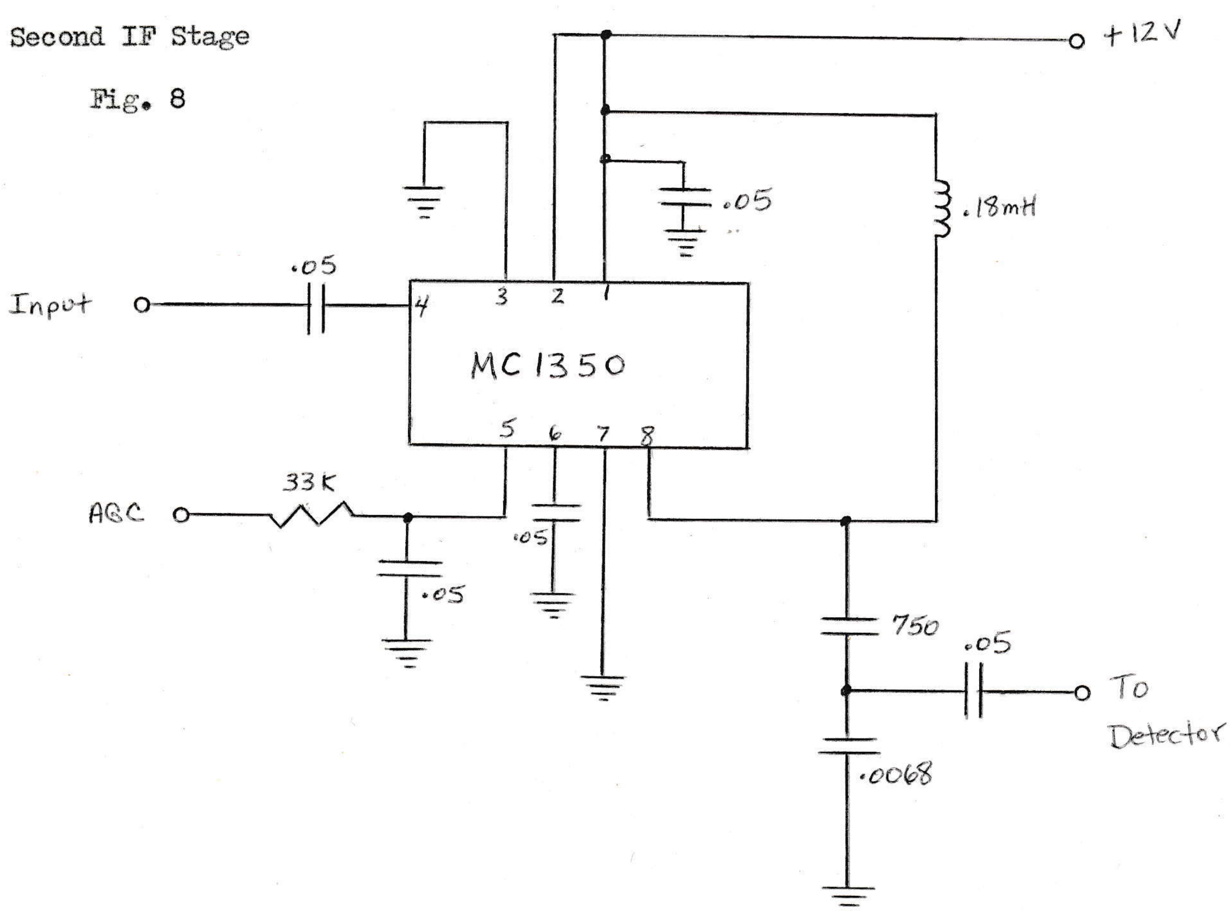
Gain 63 dB

The second stage IF amplifier is designed with the same procedure as just outlined. Once again, no attempt is necessary to match the input of the second stage (3225 -j232 ohms) to the transformer secondary which is very close necessary value. A 0.05 uF coupling capacitor is used to isolate the input from dc ground.

The output of the second IF uses the matching arrangement of (29C). The second IF stage delivers its output to a voltage doubler where the AM signal is demodulated. A Zin for the voltage doubler of 125 ohms is chosen and a load of 10 K ohms for the output of the second MC1350 is used. With Zin = 3225 ohms and Zout = 10 K ohms for the MC1350, the transducer gain is 58.85 dB (5). Several dB of additional gain could be achieved with a higher Zout for the stage but this requires a higher than desired Q for the output matching circuit.

Second IF Stage

Fig. 8



Using (29C) the output matching is found.

Select a Q of 10 $\qquad \frac{455 \text{ kHz}}{45 \text{ kHz}} = 10$ (loaded Q)

$$Q_L \sqrt{\frac{R_L}{R_{in}} - 1} = \sqrt{\frac{10\text{ K}}{125} - 1} = 8.89$$

A minimum loaded Q of 8.89 is necessary for proper operation of the matching circuit. A Q of 10 satisfies this requirement and provides an adequate margin for error

$$C_{total} = \frac{Q_L}{2(pi)f(\frac{R_L}{2})} = \frac{10}{2(pi)(455\text{ kHz})(5\text{ K})} = 700\text{ pF}$$

$$\frac{C_2}{C_1} = \sqrt{\frac{10\text{ K}}{125} - 1} = 7.94 \qquad X_L = X_{C\ tot}$$

The matching network values are C_1 = 788 pF, C_2 = 0.00626 uF, and L = 0.1748 mH. Using these calculations as a guideline, component values which are more standard near the calculated values are chosen.

$$L = 0.18\text{ mH}$$
$$C_1 = 750\text{ pF}$$
$$C_2 = 0.0068\text{ uF}$$

These component values give resonance at 456.4 kHz which is satisfactory for a Q of 10 at 455 kHz.

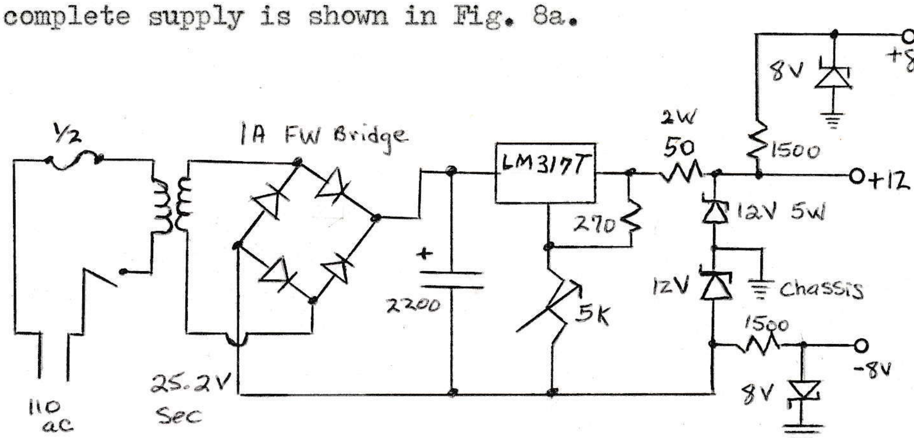
An automatic gain control voltage is applied to the AGC inputs of both MC1350's (Fig. 9). Gain is at a maximum for V_{agc} = 5 volts and a minimum for V_{agc} = 6.7 volts. The actual circuit construction of the AGC portion was found in an IF stage using devices similar to the MC1350 integrated circuits. In its configuration, the CA3160 is set for a voltage gain of 51.

The output of the device is rectified and fed back to the AGC inputs of the two stage IF amplifier system. The MC1350's have an AGC range of at least 60 dB.

The audio stage is a single LM380 audio amplifier chip which is able to deliver one watt into 8 ohms. Bypassing is used as outlined by the manufacturer.

A relatively simple power supply is used. Several different voltage levels are required; $\pm$ 8 and +12 volts. One drawback of the design is that the chassis is not at true power line ground, but is actually at +12 volts with respect to ground. The complete supply is shown in Fig. 8a.

Receiver Power Supply
Fig. 8a



The Colpitts oscillator is contained in a separate shielded area away from the other receiver stages. All circuitry is mounted on printed circuit boards and a minimum of point-to-point wiring is used. Several wire lengths are used, however, and in cases where they are associated with a high-gain stage, bypass capacitors are used to minimize problems due to rf pickup. The complete local oscillator is illustrated in Fig. 9 and the mixer, IF, and audio stages are shown in Fig. 10.

The only problems encountered with the constructed receiver involved noise and low-level audio oscillations due to rf pickup from local AM and FM stations. Additional bypass capacitors in the affected areas resolved the problem. The tuned IF stage is adjusted to 455 kHz by tuning for optimum audio quality and the balance control on the MC1596 mixer adjusted for equal voltages on pins 1 and 4.

Overall receiver performance is more satisfactory than I imagined would be possible with the design. Dynamic performance is excellent and AGC action on the stages superb. A greater degree of selectivity would enhance the receiver's performance, but also influence its price adversely. The entire project cost approximately $100 although several major components were already on hand.

PARTS LIST

(2)	MPF102	(1)	T-68-15 toroid
(1)	2N2222A	(3)	T-30-2 toroid
(9)	1N914	(2)	9.1 mH RFC
(2)	MC1350	(3)	120 uH RFC
(1)	MC1596	(1)	0.18 mH RFC
(1)	CA3160	(1)	5-50 pF variable cap
(1)	9 volt zener, 1 W	(1)	3 position SPST
(2)	8.1 volt zener, 1 W	(1)	200 uA meter
(2)	12 volt zener, 1 W	(1)	13:1 gear reduction drive
(2)	100 K ohm potentiometer	(1)	25.2 V 1 amp transformer
(1)	50 K trim pot	(1)	2200 uF, 50 V electrolytic cap
(1)	1 K trim pot	(1)	SFD455 crystal filter, Murata
(1)	120 pF trimmer cap	(1)	220 UF electrolytic cap, 25 V
(1)	8"x11"x3" enclosure	(1)	47 uF electrolytic cap, 16 V

Other capacitors and resistors as listed on schematic

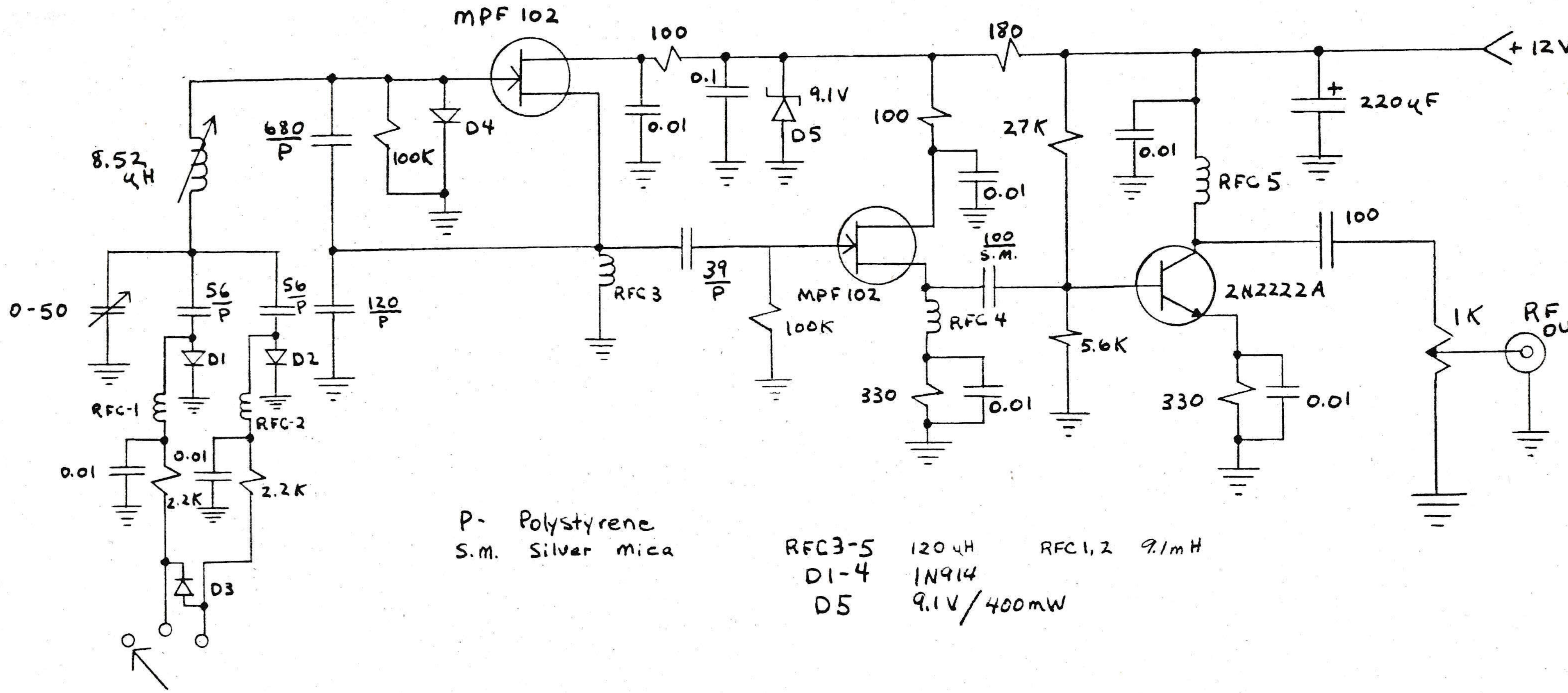


Fig 9
Colpitts Oscillator

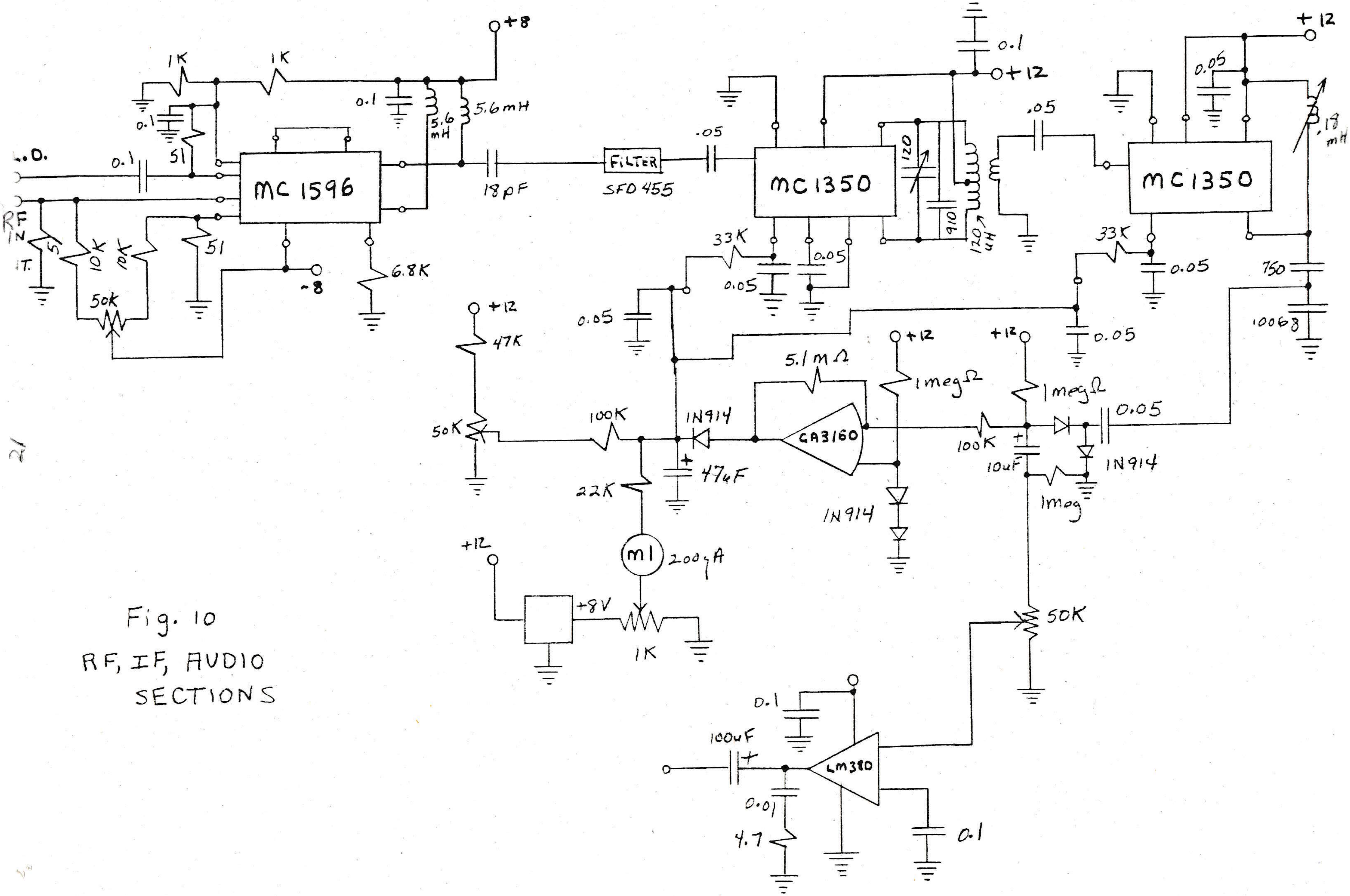
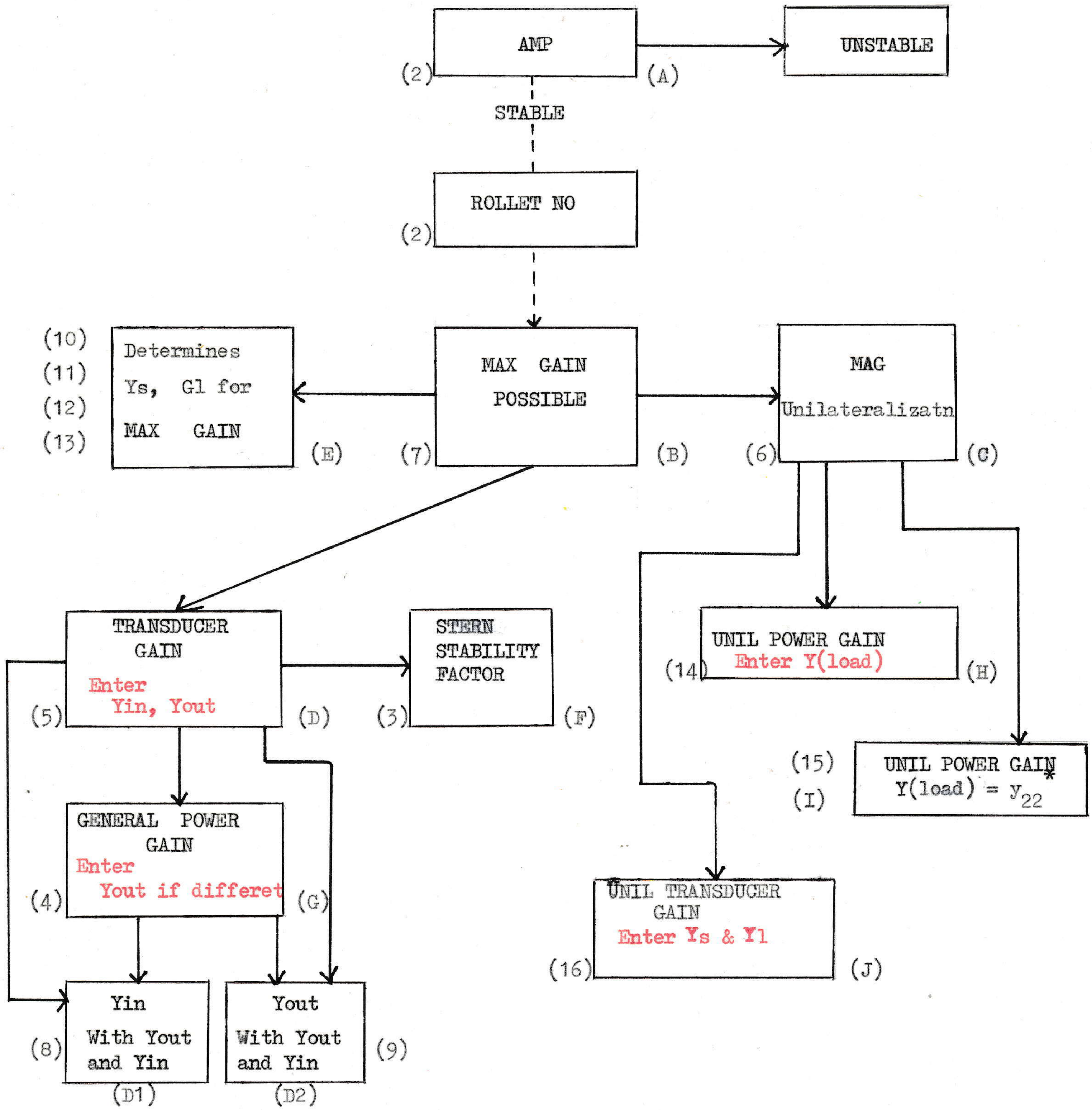


Fig. 10
RF, IF, AUDIO
SECTIONS


AMP
(2)
(A)
UNSTABLE
STABLE
ROLLET NO
(2)
(10)
(11)
(12)
(13)
Determines
Ys, Gl for
MAX GAIN
(E)
MAX GAIN
POSSIBLE
(7)
(B)
MAG
Unilateralizatn
(6)
(C)
TRANSDUCER
GAIN
Enter
Yin, Yout
(5)
(D)
STERN
STABILITY
FACTOR
(3)
(F)
UNIL POWER GAIN
Enter Y(load)
(14)
(H)
(15)
(I)
UNIL POWER GAIN
Y(load) = y_{22}^*
GENERAL POWER
GAIN
Enter
Yout if differet
(4)
(G)
UNIL TRANSDUCER
GAIN
Enter Ys & Yl
(16)
(J)
Yin
With Yout
and Yin
(8)
(D1)
Yout
With Yout
and Yin
(9)
(D2)

HIGH FREQUENCY AMPLIFIER DESIGN

R00	g11	R09	B1	R18	Im(y12y21)
R01	b11	R10	G in	R19	work
R02	g12	R11	B in	R20	work
R03	b12	R12	y12y21		
R04	g21	R13	angle		
R05	b21	R14	k-Rollet		
R06	g22	R15	work		
R07	b22	R16	work		
R08	G1	R17	Re(y12y21)		

```
Lbl AMP
"Parameters"
Aview
Pse
H Yes = 1
Prompt
x=0?
GTO AA
GTO HH
RTN
Lbl AA
G11
Prompt
STO 00
B11
Prompt
STO 01
G12
Prompt
STO 02
B12
Prompt
STO 03
G21
Prompt
STO 04
B21
Prompt
STO 05
G22
Prompt
STO 06
B22
Prompt
STO 07
Lbl A
RCL 03
RCL 02
R-P
STO 12
R !
STO 13
RCL 05
RCL 04
R-P
ST * 12
R !
ST + 13
RCL 13
RCL 12
P-R
CHS
RCL 00
RCL 06
 *
 2
 *
 +
RCL 12
 /
STO 14
 1
X  Y
X  Y?
GTO 01
"POT UNSTABLE"
Aview
STOP
Lbl 01
"STABLE"
Aview
Pse
Pse
RCL 14
K:R=
ARCL.X
Aview
Pse
Pse
Lbl B
RCL 05
RCL 04
R-P
STO 15
RCL 03
RCL 02
R-P
RCL 15
X  Y
 /
STO 15
RCL 14
x²
 1
 -
SQRT
CHS
RCL 14
 +
RCL 15
 *
LOG
10
 *
GMAX=
ARCL.X
 dB
Aview
RTN
Lbl C
RCL 05
RCL 04
R-P
x²
 4
 /
RCL 00
 /
RCL 06
 /
LOG
10
 *
MAG=
ARCL.X
 dB
Aview
RTN
Lbl D
Y in, Re
Prompt
STO 10
Y in, Im
Prompt
STO 08
YL, Im
Prompt
STO 09
Lbl 02
RCL 01
RCL 11
 +
RCL 00
RCL 10
 +
R-P
STO 15
RDN
STO 16
RCL 07
RCL 09
 +
RCL 06
RCL 08
 +
R-P
ST * 15
R !
ST + 16
RCL 16
RCL 15
P-R
STO 15
R !
STO 16
RCL 13
RCL 12
CHS
P-R
ST + 15
R !
ST + 16
RCL 16
RCL 15
R-P
x²
STO 15
Lbl 03
RCL 05
RCL 04
R-P
x²
RCL 08
 *
RCL 10
 *
 4
 *
RCL 15
 /
LOG
10
 **
G-TD=
ARCL.X
 dB
Aview
RTN
Lbl D1
SF 00
GTO 11
Lbl 12
RCL 00
 +
STO 19
R !
RCL 01
 +
RCL 19
Y in, Re
ARCL.X
 MHOS
Aview
Pse
R !
Y in, Im
ARCL.X
 MHOS
Aview
CF 00
RTN
Lbl D2
RCL 01
RCL 11
 +
RCL 00
RCL 10
 +
R-P
STO 15
R !
STO 16
RCL 13
RCL 16
 -
RCL 12
RCL 15
 /
P-R
STO 15
R !
STO 16
RCL 06
RCL 15
 -
Yout, Re
ARCL.X
Aview
Pse
RCL 07
RCL 16
 -
```

```
Yout Im
ARCL.X
Aview
RTN
Lbl E
RCL 13
RCL 12
P-R
STO 17
R !
STO 18
Lbl 04
RCL 12
x^2
STO 15
RCL 00
 2
 *
RCL 06
 *
RCL 17
x^2
RCL 15
-
SQRT
STO 19
 2
 /
RCL 06
 /
GS=
ARCL.X
  MHOS
Aview
Pse
Pse
Lbl 05
RCL 18
 2
 /
RCL 06
 /
STO 20
RCL 01
 -
BS=
ARCL.X
  MHOS
Aview
Pse
Pse
RCL 19
 2
 /
RCL 00
 /
GL=
ARCL.X
  MHOS
Aview
Pse
Pse
RCL 20
RCL 06
 *
RCL 00
 /
RCL 07
 -
BL=
 ARCL.X
  MHOS
Aview
RTN
Lbl F
RCL 00
RCL 10
 +
RCL 08
RCL 06
 +
 *
 2
 *
RCL 12
RCL 17
 +
 /
K.FT.=
ARCL.X
Aview
RTN
Lbl G
Same Load?
Aview
Pse
Yes = 1
Aview
STOP
x=0?
GTO 10
Lbl 11
RCL 09
RCL 07
 +
RCL 08
RCL 06
 +
R-P
STO 15
R !
STO 16
RCL 13
RCL 16
 -
RCL 12
RCL 15
 /
CHS
P-R
FS? 00
GTO 12
RCL 00
 +
RCL 15
x^2
 *
STO 19
RCL 05
RCL 04
R-P
x^2
RCL 08
 *
RCL 19
 /
LOG
10
 *
G.PG.=
ARCL.X
  dB
Aview
STOP
Lbl 10
Yl, Re
Prompt
STO 08
Yl, Im
Prompt
STO 09
GTO 11
RTN
Lbl H
LOAD, Im?
Prompt
STO 09
Lbl 13
RCL 09
RCL 07
 +
RCL 03
 +
RCL 08
RCL 06
 +
RCL 02
 +
R-P
STO 15
R !
STO 16
RCL 15
x^2
RCL 00
 *
STO 19
Lbl 14
RCL 05
RCL 03
 -
RCL 04
RCL 02
 -
R-P
x^2
STO 20
RCL 19
1/x
RCL 20
 *
RCL 08
 *
LOG
10
 *
G.PG=
ARCL.X
  dB
Aview
RTN
Lbl I
RCL 00
RCL 02
 +
 4
 *
RCL 06
RCL 02
 +
 *
1/x
RCL 20
 *
LOG
10
 *
GU.=
ARCL.X
  dB
Aview
```

```
RTN
Lbl J
GS Re
Prompt
STO 10
GS Im
Prompt
STO 11
Same Load?
Aview
Pse
Yes = 1
Prompt
x=0?
GTO 16
GTO 15
Lbl 16
GL Re
Prompt
STO 08
GL Im
Prompt
STO 09
GTO 15
Lbl 15
RCL 01
RCL 03
 +
RCL 11
 +
RCL 00
RCL 02
 +
RCL 10
 +
R-P
RCL 15
 *
x^2
STO 16
Lbl 16
RCL 10
RCL 08
 *
 4
 *
RCL 20
*
RCL 16
 /
LOG
10
 *
G.TU.=
ARCL.X
 dB
Aview
RTN


Lbl HH
H11 Re
Prompt
STO 10
H11 Im
Prompt
STO 11
H12 Re
Prompt
STO 12
H12 Im
Prompt
STO 13
H21 Re
Prompt
STO 14
H21 Im
Prompt
STO 15
H22 Re
Prompt
STO 16
H22 Im
Prompt
STO 17
Lbl 20
RCL 11
RCL 10
R-P
1/x
X  Y
CHS
X  Y
P-R
STO 00
R !
STO 01
Lbl 21
RCL 11
RCL 10
R-P
STO 18
R !
STO 19
RCL 13
RCL 12
R-P
CHS
STO 20
R !
RCL 19
 -
RCL 20
RCL 18
 /
P-R
STO 02
R !
STO 03
Lbl 22
RCL 15
RCL 14
R-P
RCL 18
 /
STO 18
R !
RCL 19
 -
RCL 18
P-R
STO 04
R !
STO 05
Lbl 23
RCL 11
RCL 10
R-P
STO 18
R !
STO 19
RCL 17
RCL 16
R-P
ST * 18
R !
ST + 19
Lbl 24
RCL 13
RCL 12
R-P
STO 20
R !
STO 08
RCL 15
RCL 14
R-P
ST * 20
R !
ST + 08
-1
ST * 20
RCL 19
RCL 18
P-R
STO 18
R !
STO 19
RCL 08
RCL 20
P-R
ST + 18
R !
ST + 19
RCL 19
RCL 18
R-P
STO 18
R !
STO 19
RCL 11
RCL 10
R-P
ST / 18
R !
ST-19
RCL 19
RCL 18
P-R
STO 06
R !
STO 07
GTO AA
RTN
END
```

GLOSSARY OF TERMS

C	Linville Stability Factor
kr	Rollet Stability Factor
K, k	Stern Stability Factor
Gs	Real part of source admittance
Gl	Real part of load admittance
Bs	Imaginary part of source admittance
Bl	Imaginary part of load admittance
g_{ij}	Real part of y_{ij}
b_{ij}	Imaginary part of y_{ij}
Y_l	Complex load admittance
Y_s	Complex source admittance
*	Conjugate
Yin	Input admittance
Yout	Output admittance

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