DEVELOPMENT OF THE SIGNAL DETECTION ELECTRONICS FOR AN AUGER SPECTROMETER

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AUGER SPECTROMETRY

Physical Basis

An atom, ionized by the removal of inner shell electrons returns to its ground state through an outer electron dropping into the vacant position. The energy released in this transition appears either as an X-ray or an Auger electron which is released from the outer shells. Therefore, the Auger electron originates from a radiationless process, and its energy as derived from the re-arrangement of the orbital electrons, is characteristic of the parent atom.

The two de-excitation processes compete, with the probability of Auger electron emission decreasing with increasing atomic weight. For the lighter elements such as Carbon the probability for de-excitation through Auger electron emission is approximately 95%.

Since Auger electron emission energy is a characteristic of the parent atom, it is clear that a spectrum of emission energies from a material sample will give information as to the atoms contained within it. One way to initiate the ionization process is by the bombardment of the surface with electrons. Since penetration into the sample will fall off rapidly beneath the surface, then this particular excitation method will give mainly information about the atoms at or near the surface when one examines the reflected energy spectrum. It is therefore for the particular purpose of examining the surfaces of materials that Auger Spectroscopy has been developed. This paper describes the development method used in the realization of a practical Auger Spectrometer.

Operating Principals of the Auger Electron Optics System[4]

In Figure 1 a primary electron beam is used to excite the target while the hemispherical electron energy analyzer measures the energy spectrum of the electrons emitted from the excited target. Electrons striking the target are either elastically scattered or transfer energy to the atoms of the sample. It is the de-excitation process of these atoms which yield the characteristic Auger electron spectrum.

The primary electron beam serves only to produce Auger electrons in the target. The main requirement of the beam is that its energy be sufficient to produce the Auger electrons. The higher the beam current the more Auger electrons are produced; while the beam usually impinges upon the target at a grazing angle of approximately 15° to localize the excited electrons to the surface.

The grid structure sorts electrons according to energy by interposing a potential barrier between the collector and target. Electrons with energies greater than the potential barrier reach the collector. In order to get an accurate measurement of the total energy of the emitted electrons, the electron trajectories must be parallel to the direction of the retarding field. It is also desired that as many of the emitted electrons as possible be collected. This requires that the grid structure have as large an angle of acceptance as possible. These requirements are fulfilled by using a small target at the centre of curvature of a spherical collector. The particular commercially available grid structure used for the spectrometer in this report is the Physical Electronics

Industries "PHI-15-180". It is a four-grid hemispherical unit with the two centre grids tied together to form G_{23}. The outermost and innermost grids act as electrostatic shields. By tying the two central grids together the region between them becomes an equipotential which achieves a more definite height and shape to the potential barrier. This allows for resolutions in electron energies of approximately 0.5%, and the maximum negative excursion which is used as an indicator of an Auger transition can be determined within the range of 1.0 volt. The collector is biased positively such that any electrons surmounting the potential barrier will be attracted to it. Figure 3 shows the potential variation through the analyzer.

A better understanding of the operation of the optics can be obtained by expanding the total collector current in a Taylors Series about the retard grid potential E_0. In general we can write[1]

$$I(E) = I(E_0) + \left.\frac{\partial I_c}{\partial E_c}\right|_{E_G} \Delta E_c + \left.\frac{\partial I_c}{\partial E_G}\right|_{E_c} \Delta E_G$$

$$+ \frac{1}{2!} \left.\frac{\partial^2 I_c}{\partial E_c^2}\right|_{E_G} (\Delta E_c)^2 + \frac{1}{2!} \left.\frac{\partial^2 I_c}{\partial E_G^2}\right|_{E_c} (\Delta E_G)^2 \qquad (1)$$

$+ \ldots\ldots$

If the collector potential E_c is kept a constant as is the case in Auger spectrometry, then the expansion becomes simply

$$I(E) = I(E_0) + \left.\frac{\partial I_c}{\partial E_G}\right|_{E_c} \Delta E_G + \frac{1}{2!} \left.\frac{\partial^2 I_c}{\partial E_G^2}\right|_{E_c} (\Delta E_G)^2 \qquad (2)$$

$+ \ldots\ldots\ldots$

where $\Delta E_G = E - E_0$. The term $I(E_0)$ is the total collector current and represents the number of electrons with energies greater than E_0 and less than or equal to the energy of the primary incident beam. The term $\frac{\partial I(E)}{\partial E_G}$ gives the energy distribution of the electrons, and the term $\frac{\partial^2 I(E)}{\partial E_G^2}$ is the first derivative of the energy distribution and the curve usually plotted for Auger spectrum analysis. These derivative terms are obtained electronically by superimposing a sinusoidally varying term onto the D.C. retard voltage applied to the grid G_{23}. If we let the term $\Delta E_G = A \sin \omega t - E_0$ then the collector current may be expressed as

$$I_c(E) = I_c(E_0) + A \left.\frac{\partial I_c}{\partial E_G}\right|_{E_c, E_G = E_0} \sin \omega t$$

$$+ \frac{A^2}{4} \left.\frac{\partial^2 I_c}{\partial E_G^2}\right|_{E_c, E_G = E_0} \left(\frac{1}{2} - \frac{1}{2} \cos 2\omega t\right) + \ldots. \qquad (3)$$

Therefore we see that the desired derivative term is proportional to $\cos 2\omega t$ which is at a first harmonic frequency of the sinusoidally varying collector current. As the retard potential is swept the magnitude of the derivative $\frac{\partial^2 I_c}{\partial E_G^2}$ will vary according to the number of Auger electrons with transitions at the particular retard potential.

Equivalent Circuit of Electron Analyzer

Figure 4(b) represents the electronic equivalent of the hemispherical grid structure. The basic transfer function from target to collector is represented as a voltage source $A(T.S)V_{G23}$ in series with an effective collector resistance r_c. This is the same as the equivalent of the ideal triode except for the term A(T.S.) which is included to remind one of the importance of the Taylors Series expansion in Auger Spectroscopy. The collector resistance is given as

(1)

$$r_c = \left.\frac{\partial I_c}{\partial E_c}\right|_{E_G=\text{constant}} \tag{4}$$

A plot of the D.C. transfer function is given in Figure 5 from which one can see the large value of r_c.

Also included in the equivalent circuit are three interelectrode capacitances which will exist in any such tube structure as we are considering. Attempts are usually made in designing the grid structure to minimize the values of these capacitances, since they provide undesired coupling between grids. In the case of the grid structure used here the two grids G_1 and G_4 are kept at ground potential ω.r.t. a.c. variations. The most important of the interelectrode capacitances is the capacitor C_{G23-c} since it provides a means of coupling signals from G_{23} to the collector independent of the beam current; and it is the variation in beam current that provides the desired Auger information.

A simple analysis of the effect of this capacitance can be obtained by considering it as a simple triode amplifier. The output voltage between the terminals C and g is simply given by V_0 = IZ where I is the short-circuit current and Z is the impedance seen between the terminals C and g. In determining Z[1] we consider voltage sources as short circuits, and therefore we see that Z is the parallel combination of the impedances corresponding to Z_L, C_{c-g}, r_c and C_{G23-c}.

Thus

$$Y = \frac{1}{Z} = Y_L + Y_{c-g} + Y_{G23-c} + g_c \tag{5}$$

where $Y_L = 1/Z_L$ = admittance of Z_L

$Y_{c-g} = j\omega C_{c-g}$ = admittance of capacitance C_{c-g}

$Y_{G23-c} = j\omega C_{G23-c}$ = admittance of capacitance C_{G23-c}

$g_c = 1/r_c$ = admittance of collector resistance.

The current from collector to ground for a short circuit between these terminals is given by

$$I = -g_m V_i + V_i\, Y_{G23-c} \tag{6}$$

where $g_m = \left(\frac{\partial I_c}{\partial V_{G23}}\right)_{V_c} \equiv$ mutual conductance.

The corresponding output voltage will then be

$$V_0 = IZ = I/Y = \frac{-g_m V_i + V_i\, Y_{G23-c}}{Y_L + Y_{c-g} + Y_{G23-c} + g_c} \tag{7}$$

Now, if the interelectrode capacitances were zero then V_0 would become

V'_0 where

$$V'_0 = \frac{-g_m V_i}{Y_L + g_c} \quad . \tag{8}$$

From the last two equations above it is made clear that the effect of the interelectrode capacitance C_{G23-c} is to add a component to the output voltage which is independent of the current reaching the collector from the target. Note that this "feedthrough component" from Y_{G23-c} is at an angle of 90° with respect to the desired component.

The value of the interelectrode capacitances was measured using a Wayne-Kerr Autobalance bridge, and special low capacity leads in order to get good accuracy. The measured values were

$$C_{G23-c} = 1.3 \text{ pf}$$

$$C_{G23-g} = 62 \text{ pf}$$

$$C_{c-g} = 158 \text{ pf}$$

Although the offending capacitance C_{G23-c} is very small (due to the screening action of the grid G_4 which is tied to ground), its effect upon operation of the Auger spectrometer can be very detrimental. This can be seen from the equation for V_0. The magnitude of V_0 is

$$|V_0| = \sqrt{\frac{g_m^2 + (\omega C_{G23-c})^2}{(\frac{1}{Z_L} + \frac{1}{r_c})^2 + \left[\omega(C_{G23-c} + C_{c-g})\right]^2}} \; |V_i| \tag{9}$$

Now the mutual conducance $g_m = \left(\dfrac{\partial I_c}{\partial V_{G23}}\right)_{V_c}$

$\cong \left.\frac{\Delta I_c}{\Delta V_{G23}}\right|_{V_c}$ (10) and in average operation of the tube the term ΔI_c will be in the range of .01μA for a change of V_{G23} of 1 volt. Therefore the approximate value of g_m^2 will be in the range of 10^{-16} to 10^{-14}. Most spectrometers have modulating signals superimposed on the G_{23} retard voltage with frequencies in the range of 500 Hz to 4 KHz. This frequency range is determined by two factors. The lower end is dictated by the desire to keep bothersome 60 - 120 Hz hum from interfering in the measurements. The second consideration is to get as good of an average as possible for the output readings in any given energy region. This requires higher modulating frequencies. In the spectrometer used in this report the value of the fundamental of the modulating frequency was chosen as 3 KHz. For this frequency the magnitude of $(\omega C_{G23-c})^2$ is 6.0×10^{-16}. Therefore the feedthrough component is of the same approximate magnitude as the desired component from the target current. This is to be contrasted with the usual manner in which tubes are operated where $g_m >> \omega C_{G23-c}$. Currents leaving the target in an Auger spectrometer are typically 10 ∿ 100μA, whereas cathode currents in thermionic emission tubes are ∿ 1-10MA. The smaller currents in Auger spectrometry are necessitated by the need to minimize surface damage of the surface under study. This being the case it is usually necessary to rid the analyzer of this undesired feedthrough component by a neutralizing circuit. This was found to be definitely the case for the analyzer used in this report, and the major effort needed to realize an operational spectrometer was based on the development of a proper neutralization circuit.

Another consideration of primary importance in the design of an

Auger spectrometer is that associated with the distortion of the modulating source. This can easily be seen as follows. As indicated earlier, the basic method used for detecting Auger transitions is by selecting the component of current at the first harmonic (i.e. the component at twice the frequency) of the modulating frequency. This component of current exists due to the manner in which the electronics is theoretically supposed to work (i.e. by making a Taylor's series expansion of the current about a given D.C. retard potential) and has nothing to do with distortion from the oscillator used to produce the modulation. If the oscillator distorts then the signal applied to the grid is given by

$$V_i = A \sin \omega t + B \sin 2\omega t + C \sin 3\omega t + \ldots. \tag{11}$$

This first harmonic component of voltage due to the distortion will feed through to the collector by means of C_{G23-c}. This component will be at the same frequency as the desired component due to Auger transitions. It's phase relative to the Auger component will be dictated by the phase difference between the fundamental and first harmonic of the modulating signal. Since the amplitude of the first harmonic component due to the current will vary with the Auger transition, it is thus clear that certain of the weaker transitions may be completely masked by the distortion feedthrough component.

Thus, there exists two basic considerations in designing the electronics used in Auger Spectrometry. The most important one is the design of a proper neutralizer circuit for the capacitance C_{G23-c}, and the second is the design of a low distortion oscillator to produce the modulating signal for the grid G_{23}. The next section describes the approach and the circuits derived to fulfil these requirements.

CHAPTER II

AUGER SPECTROMETER ELECTRONICS DESIGN

1. The Neutralizer Unit

In order to rid the collector signal of the feedthrough created by the finiteness of the capacitance C_{G23-c}, a neutralizing circuit was developed. Figure #6 shows the schematic diagram of the unit.

The circuit presented here accomplishes neutralization through a so-called "inductive-arm" network. The fundamental requirement for proper operation of this unit is the constancy of the capacitance C_{G23-c}, as a function of applied plate voltage, retard voltage and analyzer current. Experiments carried out indicated that the value remained a constant within 1% for plate voltages of 0 to 300, retard voltages to 1500 volts, and beam currents from 0 to 60μA. Neutralization is accomplished by generating an external signal which is identical to the feedthrough voltage on the collector. This signal is one input to a differential amplifier. The other input to the amplifier is the signal from the collector. A differential amplifier produces an out-put signal given by $e_0 = A(2_2-e_1)$ where A is the gain of the amplifier and e_2 and e_1 are the signals at the input terminals. From the discussion above we can write

$$e_2 = e(A.S) + e(F.T.) \tag{12}$$

$$e_1 = e(F.T.) \tag{13}$$

$$\therefore \; e_0 = A \;\; e(A.S) \qquad (14)$$

where

$e(A.S)$ = collector signal due to Auger spectrum

$e(F.T.)$ = collector signal due to capacitive feedthrough from G_{23} to collector.

The feedthrough capacitance was given earlier as 1.3 pf. To obtain a commercially available capacitor of this exact value would be very difficult. This problem can be easily surmounted by using the impedance transformation properties of transformers. The basic method[(3)] making use of these properties is shown in Figure 7. Figure 7(A) shows a capacitance inserted in the primary of a N:1 step-down transformer. The circuit is excited by an alternating signal V and the output is taken across the secondary as V_0. Figure 7(B) indicates that an equivalent output signal V_0 can be obtained by a voltage source of strength V/N exciting a capacitance N^2 times as large as the capacitance in 7(A). The situation in 7(B) is easily obtained by placing a capacitance N^2 times the capacitance in 7(A) in series with the secondary of the stepdown transformer and exciting the primary by the voltage V. Figure 7(C) shows this schematically and indicates the equivalency of the output signal V_0. In the case of the Auger spectrometer the signal V is that from the modulating oscillator.

Thus, for a feedthrough capacitance of 1.3 pf the required capacitor in the secondary of the second step-down transformer will be $N^2(1.3)$pf. In the circuit of the neutralizer built the value of N is 20, so the capacitor is 400(1.3) = 520 pf. Because of the non-idealness of the transformers (i.e. shunt resistances and capacitances associated with the windings) it is not possible to produce a practical working circuit exactly as indicated in Figure 7(C). In the actual neutralizer circuit of Figure 6 a more sophisticated network is connected to the secondary of the inductive-arm transformer. The network consists of a 100KΩ variable resistor in series with a 365 pf variable capacitor. Shunting the capacitor is a series connection of a 975 KΩ resistor and a 50KΩ ten-turn potentiometer. The tap on the potentiometer supplies the neutralization signal supplied to one of the inputs of the differential amplifier. This network allows complete freedom in varying the magnitude and phase of the neutralizing signal. The differential amplifier used in the neutralizer is of a standard design and is built to supply a gain of 100. This circuit, when connected to the Auger spectrometer and properly balanced was found to be capable of reducing the feedthrough effect to below 1μV with a time drift measured in the range of 2μV/week.

2. <u>The Modulating Oscillator</u>[5]

As indicated earlier, the range of modulating frequencies commonly used in Auger spectrometry is 500 Hertz to 5 Killohertz. In the spectrometer designed in this report, the frequency chosen was 3 KHz. For these low frequencies the Wien Bridge Oscillator is found to be an excellent sine-wave generator from the standpoint of low distortion, cost and size. The former consideration is the most important in this application as harmonics will produce intermodulation and masking signals at the collector. One of the major problems associated with this type of oscillator is amplitude stability, and many circuits with automatic gain control in the feedback loop have been proposed. One suggested in reference 5 was used to build the modulating oscillator. Following the presentation offered there the basic operating principal of the oscillator is as follows.

Referring to Figure 8[5] and assuming that the output signal e_0 is a sinusoid, then the feedback ratio of the bridge is by the simple voltage divider Z_2/Z_1+Z_2 where

$$Z_1 = R_1 - j/\omega C_1 \qquad \text{and} \qquad Z_2 = R_2(1+j\omega R_2C_2) \tag{15}$$

Following normal operational amplifier theory the two inputs will try to maintain a 0 volt potential difference between them and hence we may write for the signal at the non-inverting input

$$e_1 = \frac{Z_2}{Z_1+Z_2} e_0 \tag{16}$$

The oscillation condition is given by a 0 net phase shift between e_0 and e_1. Therefore the ratio e_1/e_0 must be real and the imaginary component of the feedback ratio will be equal to zero. Therefore we have

$$\omega_0 R_1R_2C_2 - \frac{1}{\omega_0 C_1} = 0, \text{ and solving for } \omega_0 \text{ we have}$$

$$\omega_0 = \frac{1}{\sqrt{R_1R_2C_1C_2}} \tag{17}$$

Usually $R = R_2$, $C_1 = C_2$ and for this case $\omega_0 = \frac{1}{R_1C_1}$ and the corresponding value for $\frac{Z_2}{Z_1+Z_2}$ is 1/3. If the feedback is greater than 1/3 oscillations will grow and if it is less than 1/3 they will decay away. Thus it is most important to maintain this ratio at the required value.

Figure 9 shows the rather sophisticated oscillator of reference 1 which provides excellent control of the feedback ratio.

The actual Wien-Bridge oscillator is composed of R_1, C_1, R_2, C_2 and operational amplifier A_1. This signal is further amplified by A_2 which supplies the oscillator output. The output level is sensed by the absolute value circuit of A_3 and A_4. The amplifier A_4 acts as an error integrator and becomes stable at the point when the input signal is equal to the reference signal. The diode bridge is used for controlling the negative feedback to A1. The gain of the integrator is set by the capacitor C_3. The choice of C_3 is a tradeoff between response time and distortion. Small values of C_3 will allow the circuit to stabilize quickly and large values of C_3 reduce the distortion of the oscillator. The 47 µf capacitor used in the circuit allowed for second harmonic distortion of approximately -64 db for output levels from 1 volt to 15 volts R.M.S.

The Set-up and Bandpass Amplifier[5]

When the spectrometer was originally run with the neutralization unit, it was found that there was sufficient 60 cycle feedthrough from the electron gun control unit to saturate the input stages of the lock-in amplifier used to detect the Auger signals. This being the case, it was decided that a filter arrangement should be connected between the neutralizer output and lock-in amplifier to rid the input signal of the high amplitude 60 cycle hum. The arrangement used to accomplish this is shown in Figure 9. It is essentially two separate circuits on one board. The upper circuit is a two-stage cascaded high pass active filter. The lower circuit is a two-stage cascaded active bandpass filter, and is the circuit which is used in the system when actual measurements are made.

The high pass filter is characterized by a high pass function of the form

(5)

$$H(s) = \frac{H_0\omega^2}{s^2 + \alpha\omega_0 s + \omega_0^2} \tag{18}$$

with a resultant magnitude function

(5)

$$G(\omega) = \left(\frac{H_0^2}{1 + (\alpha^2-2)(\frac{\omega_0}{\omega})^2 + (\frac{\omega_0}{\omega})^4} \right)^{1/2} \tag{19}$$

At $\omega = \omega_0$ the gain is down $1/\alpha$ from its value at $\omega = \infty$. The value of ω_0 chosen for this circuit was 1.471×10^4 rads/sec which corresponds to a frequency of approximately 2.3 KHz. The value of α for this circuit is chosen as 1 so that the frequency ω_0 represents the -3 db point. The frequency ω_0 is determined through the equation

$$\omega_0 = 1/\sqrt{R_1 R_2 C_1 C_2} \tag{20}$$

where in the circuit used here $R_1=R_2=6.8$K and $C_1=C_2=.01$ µf. The high frequency gain of the circuit can be seen to be unity by noting that the capacitors are a short circuit for $\omega = \infty$.

This function is realized by the first 310 amplifier in the circuit. The 308 amplifier in cascade with the 310 serves to add gain and curb the high frequency response through the use of the 100 pf feedback capacitor shunting the 100K feedback resistor. This arrangement constitutes a low-pass filter with a transfer function given by

$$H(s) = \frac{H_0\omega_0}{s+\omega_0} \tag{21}$$

(5)

with a resultant magnitude function

$$G(\omega) = \left(\frac{H_0^2\omega_0^2}{\omega^2+\omega_0^2}\right)^{1/2} \tag{22}$$

(5)

In terms of circuit components

$$H(s) = \frac{\dfrac{-R_1}{C_1R_1R_2}}{s + \dfrac{R_2}{C_1R_1R_2}} \tag{23}$$

with R_1 , C_1 representing respectively, the feedback resistor and capacitor and R_2 the input series resistor. The gain at D.C. is given by H_0 where $H_0 = -R_2/R_1$ and the -3 db point set at a frequency $\omega=\omega_0=1/R_1C_1$. For this circuit $f_0=2\pi\omega_0=16$ KHz and the low frequency gain is 10=20 db. Thus we see that the first 310 and 308 amplifier arrangement form a high-low pass filter with a constant gain of 10 for frequencies between 3 KHz and 15 KHz. This range covers both the fundamental and first harmonic of the oscillator frequency (the oscillator fundamental frequency was set at 3 KHz). The second 310 - 308 combination is identical to the first. Thus, the overall gain of the circuit is 100. Since the high pass filters have second-order transfer functions, the roll-off will be 12 db/octave starting at f_0 = 2.3 KHz. Therefore, for the total arrangement the 60 cycle suppresion will be approximately 120 db provided no external 60 cycle field reaches the input pins. The output of this circuit is then connected to an oscilloscope and can be used to check for proper tube operation prior to taking an Auger spectrum.

The second circuit is a bandpass filter[5] with its centre frequency set at the first harmonic of the oscillator frequency. This is done since it is the frequency component of the output signal usually desired in displaying Auger spectrums. The particular circuit arrangement used here is a two-stage cascaded infinite-gain multiple feedback circuit.

The transfer function for a bandpass circuit is given as

(5)

$$H(s) = \frac{H_0 \alpha \omega_0 s}{s^2 + \alpha \omega_0 s + \omega_0^2} \tag{24}$$

with a gain function given by

$$G(\omega) = |H(s)| = \left(\frac{H_0^2 \alpha^2}{\left(\frac{\omega}{\omega_0}\right)^2 + (\alpha^2 - 2) + \left(\frac{\omega_0}{\omega}\right)^2} \right)^{1/2} \tag{25}$$

The maximum gain H_0 occurs at the frequency $\omega = \omega_0$ and in the case of the bandpass filter the α term may be recognized as $\alpha = \frac{1}{Q}$ where $Q = \frac{\omega_0}{\omega_2 - \omega_1}$ and ω_2 and ω_1 are the -3 db points on the response curve.

In terms of circuit components the transfer function is written as

(5)

$$\frac{E_0}{E_{in}}(s) = \frac{s(K/R_1C_4)}{s^2 + (s/R_5C_4)(1 + C_4/C_3 - KR_5/R_6) + (1/C_3C_4R_5)(1/R_1 + 1/R_2 + 1/R_6)} \tag{26}$$

The circuit parameters are

(5)

$$H_0 = \frac{1}{R_1} \frac{1}{(1/KR_5)(1 + C_4/C_3) - 1/R_6} \tag{27}$$

(5)

$$\omega_0 = \left[\frac{1}{R_5 C_3 C_4} \left(\frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_6} \right) \right]^{1/2} \tag{28}$$

(5)

$$\alpha = \sqrt{\frac{1}{R_5(1/R_1+1/R_2+1/R_6)}}\sqrt{\frac{C_3}{C_4}} \left(1 + \frac{C_4}{C_3} - \frac{KR_5}{R_6}\right) \tag{29}$$

In making use of this circuit in the spectrometer the following design simplifications and procedures were used in accordance with reference 1.

Given: α, $\omega_0 = 2\pi f_0$

H_0 is a free parameter

Choose $C=C_3=C_4$; $R=R_1=R_5$

For this arrangement ω_0 is given as

$$\omega_0 = \frac{1}{R_7 C \sqrt{\alpha}} \tag{30}$$

from which the value of R_7 may be obtained. The value of R_6 is then obtained from the relationship.

$$R_6 = \frac{R_8}{2-\sqrt{\alpha}} \tag{31}$$

and the value for R_2 is determined from the relationship

$$\frac{1}{R_2} = \frac{1}{R_7}\left[\frac{1-\alpha}{\alpha} - \frac{1}{k}(2-\sqrt{\alpha}) \right] \tag{32}$$

For this arrangement $H_0 = K/\sqrt{\alpha}$ (33)

The value of α was chosen as 0.1 and the value of $K = R_8/R_7$ was $\sqrt{10}$ giving the total value for the gain at centre frequency as $H_0 = 10$. Thus, the cascaded pair offer an overall gain of 100 and a Q of 10. The 60-cycle signal is 2 decades down in frequency from the centre value f_0 of the filter. Since the filter is a cascaded second order type the altenuation/decade is 40 db resulting in an 80 db rejection of the hum.

Results:

As pointed out at the beginning of this report, the major requirement for proper operation of the spectrometer was considered to be the neutralization of the energy analyzer. The neutralizer unit built as described in this report was connected to the Auger spectrometer. The output from this unit was fed to PAR Model 121 Lock-In amplifier. It was used in the f/2 mode with a detection frequency of 6.0 KHz. The sensitivity was set at 50μV for full-scale internal meter deflection. The time constant was set at 3 seconds in accordance with the retard voltage sweep rate of 0.3 volts/second. The electron beam gun providing the primary beam was set for an accelerating potential of 3 KeV. A beam current of 10.0μA was established and focused upon the sample targets located at the centre of the hemispherical grid structure system. For these runs the oscillator used was a Hewlett-Packard tube-type Wien-Bridge oscillator. It's operation was considered poor at best since its first harmonic distortion was only 45 db below the fundamental. However, due to the fact that the high quality oscillator specifically designed for the system had been assembled in a lock-in amplifier designed by the author, it was the only one available. The bandpass amplifier was not used either, since it was found that careful operation of the focus control on the electron beam gun could reduce 60-cycle hum to the acceptable level.

The samples used in the test runs were silicon. Samples 1 and 2 were for silicon prepared with standard peroxide-amonium hudroxide and peroxide-hydrochloric solutions. Samples 3 and 4 were given additonal surface preparation by argon-ion bombardment by using an additional built-in electron gun in the system to provide the ionizing source for the argon gas.

As is evident from the response curves obtained from the spectrometer, the neutralizer system operated in accord with expectation. Signal magnitudes are in accord with expectation.

APPENDIX 1

DESIGN OF A LOCK-IN AMPLIFIER

As indicated in Figures 1 and 2 and the results section the major electronics unit used for processing the Auger Spectrometer output is the lock-in amplifier. This device is a phase sensitive detector which detects signals coherent in frequency and phase with an ac reference voltage, rejecting essentially all other frequency components, even noise many times larger than the signal of interest. It is this last point which makes it such an attractive device for Auger analysis since in general the signals of interest have a negative signal to noise ratio.

Dectection occurrs in a mixer circuit. One input, the reference, is operated in the saturated mode in order to produce a switching effect. The other input is operated in a signal linear mode. Thus, the mixer output for these input signals will depend upon phase on frequency differences between them. If the two signals are of the same frequency and are in phase, then the mixer output signal will represent the full-wave rectified spectrum of the linear mode signal. This spectrum contains a D.C. component, and it is this component which is used as the output signal of the lock-in amplifier. The a.c. components of the mixer output are removed by passing the signal through a low-pass filter D.C. filter.

In Figure 15 is shown a block diagram of the necessary sub-structures of a lock-in amplifier. Accompanying the diagram are waveforms representing the proper input and output signals for the different sub-structures.

The operation of the device is quite straightforward. A high quality, low distortion oscillator is used to provide the primary driving signal to the instrument whose characteristics are to determined. The oscillator output is also sampled by the harmonic selector. The harmonic selector will either pass the oscillator signal multiplied by some amplification or attenuation factor) or provide frequency doubling at its output. This feature is to allow for the convenience of measuring test instrument output signals at either the fundamental or first harmonic of the driving signal. The signal from the harmonic selector is then fed to a phase shifter. This unit allows phase variations from 0° to 360°. It's purpose in a lock-in amplifier is to provide phase matching between the two input signals to the mixer. Since this condition provides maximum correlation between the two signals, it also provides the greatest lock-in output signal (for a given input signal to the mixer provided from the selective amplifier). Also, by allowing for phase shifts one is able to look at vector components of an input signal by examining lock-in outputs for 90° phase shift changes. The sharper unit simply provides square wave pulses of the proper D.C. level and at the frequency of the phase-shifter output. These square wave pulses are used to provide the switching action for the mixer unit. The other input to the mixer comes from the selective amplifier. This unit is a bandpass active filter. It's centre frequency may be set either at the frequency of the fundamental of the oscillator or at its first harmonic. It's purpose is to select the frequency component of the output signal from the test instrument which is desired. It provides rejection of the other frequency signals along with 60-cycle hum. The selective amplifier output drives the signal input pins of the mixer. With this input signal and the sharper pulses both supplied to the mixer, the mixer output will consist of a full wave rectified representation of the input signal. A rectified signal will contain a D.C. level. This D.C. level is extracted by the D.C. amplifier which is simply a low-pass filter with gain. The output of the D.C. amplifier is then used to drive a chart recorder.

It can be seen that the D.C. level of the lock-in amplifier output will be directly proportional to the magnitude of the frequency component being measured by the amplifier. This characteristic makes it the perfect device for examining Auger spectrums since the magnitudes of the frequency components at the collector of the Auger optics system vary with the Auger electron emissions.

Figures 16 through to 19 provide schematic diagrams for the various building blocks. These circuits have all been tested in breadboard configurations to ascertain their technical correctness. In the following sections a brief description of the circuits is provided in order to shed light onto the operational characteristics of this particular lock-in amplifier realization.

Figure 16 shows the schematic of the selective amplifier. It is a two-stage cascaded realization of the infinite gain state-variable type. This configuration makes use of operational amplifiers in the same way they would be used in an analogue computer realization of transfer functions (i.e. using integrators and summers). The general methematical form of a bandpass transfer function was given in Chapter II, Section 3. In terms of circuit components the transfer function is given as

(5)

$$\frac{E_{0_1}}{E_{in_1}} = \frac{-s \dfrac{1}{R_1C_1} \dfrac{1+R_6/R_5}{1+R_3/R_4}}{s^2 + s \dfrac{1}{R_1C_1} \dfrac{1+R_6/R_5}{1+R_3/R_4} + \dfrac{R_6}{R_5} \dfrac{1}{R_1R_2C_1C_2}} \tag{34}$$

Comparing this with the mathematical form of the bandpass transfer function we see that

$$H_0 = R_4/R_3 \tag{35}$$

$$\omega_0 = \left(\frac{R_6}{R_5R_1C_1R_2C_2}\right)^{1/2} \tag{36}$$

$$Q = \frac{1}{\alpha} = \frac{1+R_4/R_3}{1+R_6/R_5} \left(\frac{R_6R_1C_1}{R_5R_2C_2}\right)^{1/2} \tag{37}$$

In normal design procedures one is given Q and ω_0, $R_5=R_6=R_3$ and $C_1=C_2$. Under these conditions

$$\omega_0 = \frac{1}{R_1C_1} \tag{38}$$

and $R_4=R_3(2Q-1)$. For more detail on this particular bandpass realization I refer the reader to reference 5. In implimenting this circuit two identical bandpass units were built and connected together in the cascade arrangement. This arrangement allowed for a very high degree of rejection of unwanted signals plus the advantages of accurately controlled gain and Q values. The gain per stage of this unit is 20 and the Q value is set at 9.5. Therefore the overall cascaded gain and Q value is 400 and 9.5 respectively. The capacitors C_1 and C_2 can each take on two values thus allowing for selection of the particular frequency component required. For this arrangement the two allowable centre frequencies are 3KHz and 6KHz corresponding to the fundamental and first harmonic of the oscillator driving signal. The LM310 which drives E_{in_1} is simply used for buffering and the output 725 adds variable gain ranging from 1 to 100 in three steps. Therefore total gain of the selective amplifier can be set between a value of 400 and 40,000.

Figure 17 shows the schematic of the mixer circuit and associated driving circuitry. The mixer circuit used is the MC 1596 manufactured by Motorola Semiconductor. It is referred to in their literature as a balanced modulator. The schematic (to be found in reference 2) reveals it to essentially be a differential cascade-connected amplifier with constant current sources. In the particular realization used here the unit is used with differential input-differential output conditions. The input arrangement allows for doubly-balanced operation and the use of the differential output essentially doubles the gain of the unit. The upper pair (U1,U2) are driven from the shaper circuit which is located at the lower left of the diagram. In operation, the shaper unit provides differential pulse outputs switching from a D.C. level of 2 volts to 3 volts. These level changes are more than sufficient to provide saturation or cutoff conditions for the upper pair, and in this way provide the switching action for the mixer.

The input signal to be measured is supplied to the mixer in a differential mode by the phase splitter located at the upper left portion of the diagram. The input for the splitter is supplied by the selective amplifier. The use of the variable 470K pots on the phase-splitter outputs is to provide for amplitude adjustments needed to allow for calibration of the unit.

Figure 18 gives the schematic diagram of the harmonic select and phase shift module. It is, as far as I am aware, a completely novel arrangement for providing the functions. It consists of 2 CD4046AE phase lock loops and one D-type flip-flop. These particular components are manufactured by RCA CMOS division. This particular phase-lock-loop has the feature of offering two compacators designated as compacator #1 and compacator #2. Phase compacator #1 is an exclusive or type and produces a 90° phase shift between signal in and V.C.O. out when the signal in is locked at the centre frequency of the V.C.O. Within the lock range of the phase lock loop if either the V.C.O. centre frequency or the signal in frequency is adjusted the result will be a change in relative phase shift other than 90° between signal in and V.C.O. out. Since the unit is operated within its lock range the V.C.O. frequency will not change. At the extremes of the lock-range the relative phase shifts will be 0° and 180^{0}. Linearity of phase shift with variation of V.C.O. centre frequency is found to be much better than 1%. Hence, this unit, using phase comparator #1 provides excellent phase shift capabilities. The signal in can be either pulse or sinusoidal, and the V.C.O. output is of course pulse. If the signal is of sufficient amplitude to cause the comparator #1 to change logic states then it can be coupled directly into the signal input. If not, by placing a capacitor in series with the input lead one can make use of an internally supplied amplifier to provide the overdriven state for the phase comparator #1. The centre frequency of the V.C.O. is set by R_1 and C_1, using tables supplied by the manufacturer. The 1 MΩ resistor designated R_2

in the diagram simply offsets the minimum frequency of operation of the V.C.O. for maintenance of a lock condition with the input signal. By varying the resistor R_1 through its range the phase shift between signal in 2 and V.C.O. output 2 will vary from between 0° and 180°.

The phase lock loop shown in the upper left of the diagram is used in conjunction with the D-type flip-flop to provide harmonic selection. In this circuit the phase comparator #2 is used. This comparator has the feature of adjusting itself to a 0° phase difference between the input signal and V.C.O. output when the phase-lock loop is in the locked condition. Again, the centre frequency of the V.C.O. is set by R_1 and C_1. In this case the centre frequency is adjusted to 10KHz, and the lock range for this arrangement is given by the manufacturer as 0 Hz to 20KHz. The loop between the V.C.O. output and the comparator input is either direct or contains the flip-flop. The direct connection allows the V.C.O. to operate at the frequency of the signal input. If the flip-flop is inserted the V.C.O. frequency is divided by 2 before being applied to the comparator input. This will cause the V.C.O. to double in frequency in order for it to reachieve the locked condition. Thus, this type of connection provides the required frequency doubling needed for harmonic analysis.

Figure 19 simply shows the D.C. amplifier. The first 741 circuit arrangement is a differentially connected single pole low pass filter. The -3 db point is set by the parallel resistor-capacitor combination in the feedback loops. For the values given this point is at 10Hz. The D.C. gain provided by the stage is set by the ratio of the feedback resistors in the 741 to the load resistors in the mixer. The feedback resistors are 500K and the load resistors are 5 KΩ. Therefore, the D.C. gain is 100. Also, since the stage is operated differentially, the effective gain will be 200, if one uses the D.C. level present at either load resistor for the value of the input signal. The gain through the mixer was adjusted to be 1, so that the input signal to the mixer has to be 43 millivolts R.M.S. to provide for a 10 volt D.C. level at the low-pass filter output. The variable pot arrangement attached to the inverting input of the filter is used to provide D.C. offsets. The second 741 low-pass filter is used to provide additional a.c. rejection. The total rejection provided by the combination is 12 db per decade. The D.C. gain provided by the second 741 is set at 1. The output signal is sent to a meter and a recorder.

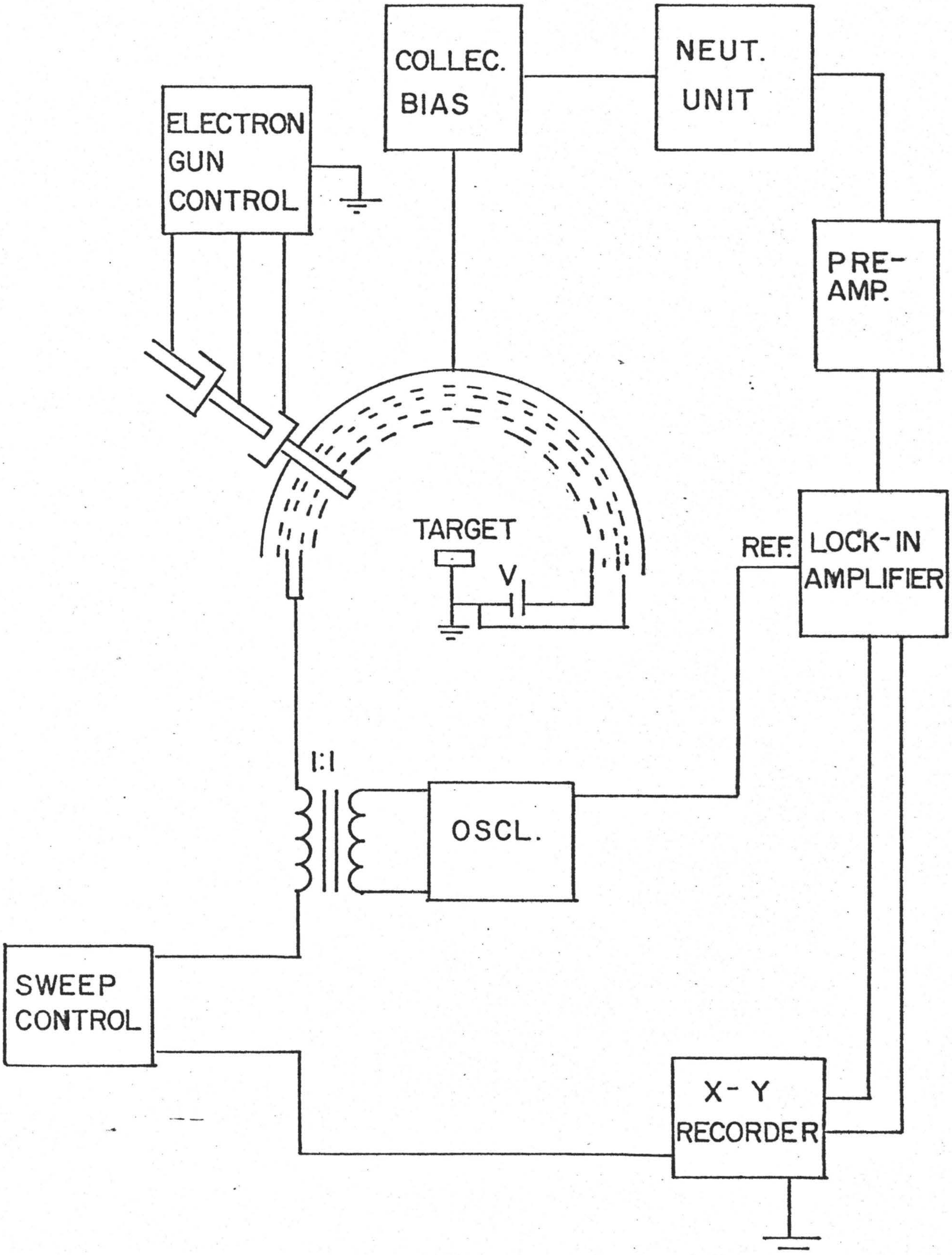
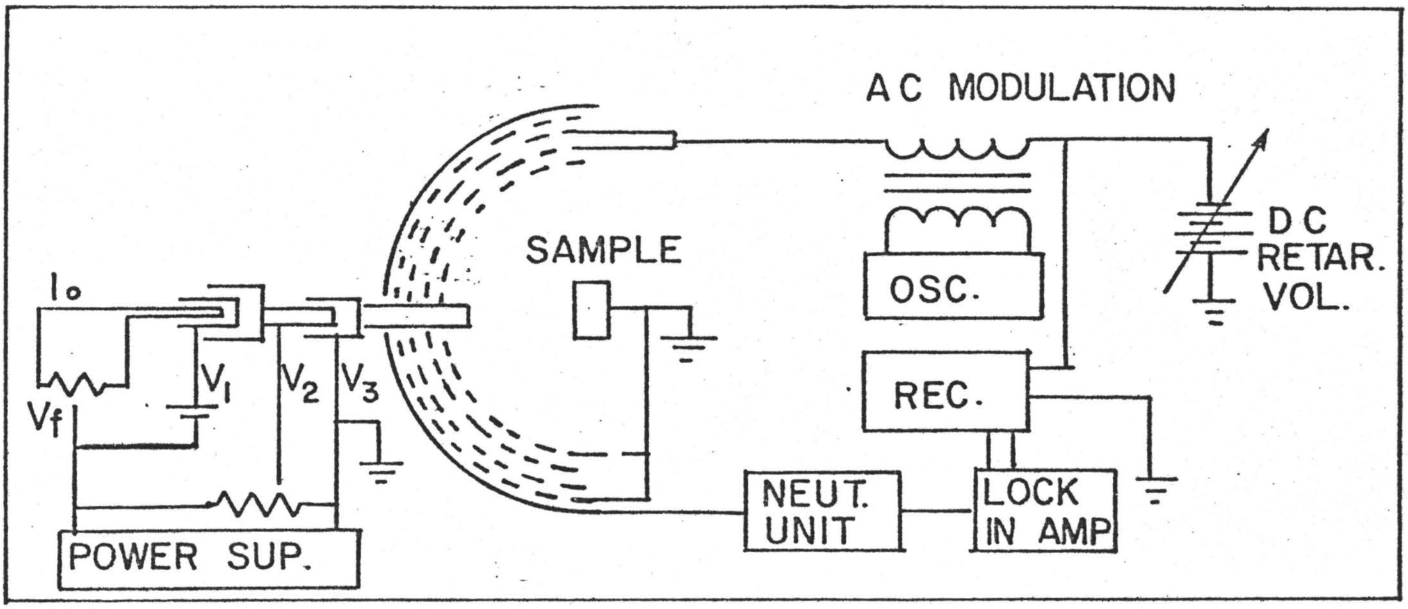
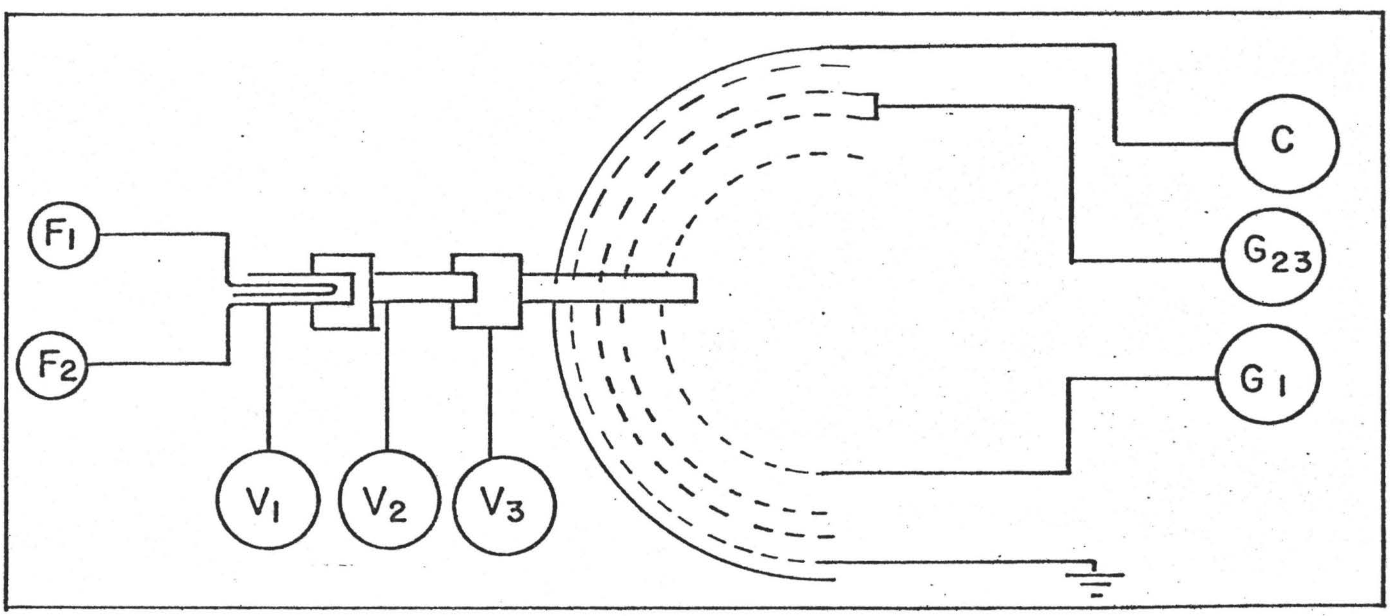


Figure 1 Schematic of Auger Spectrometer.



(A)



(B)

Figure 2 (A) Simplified schematic of circuit.

(B) Connection Diagram.

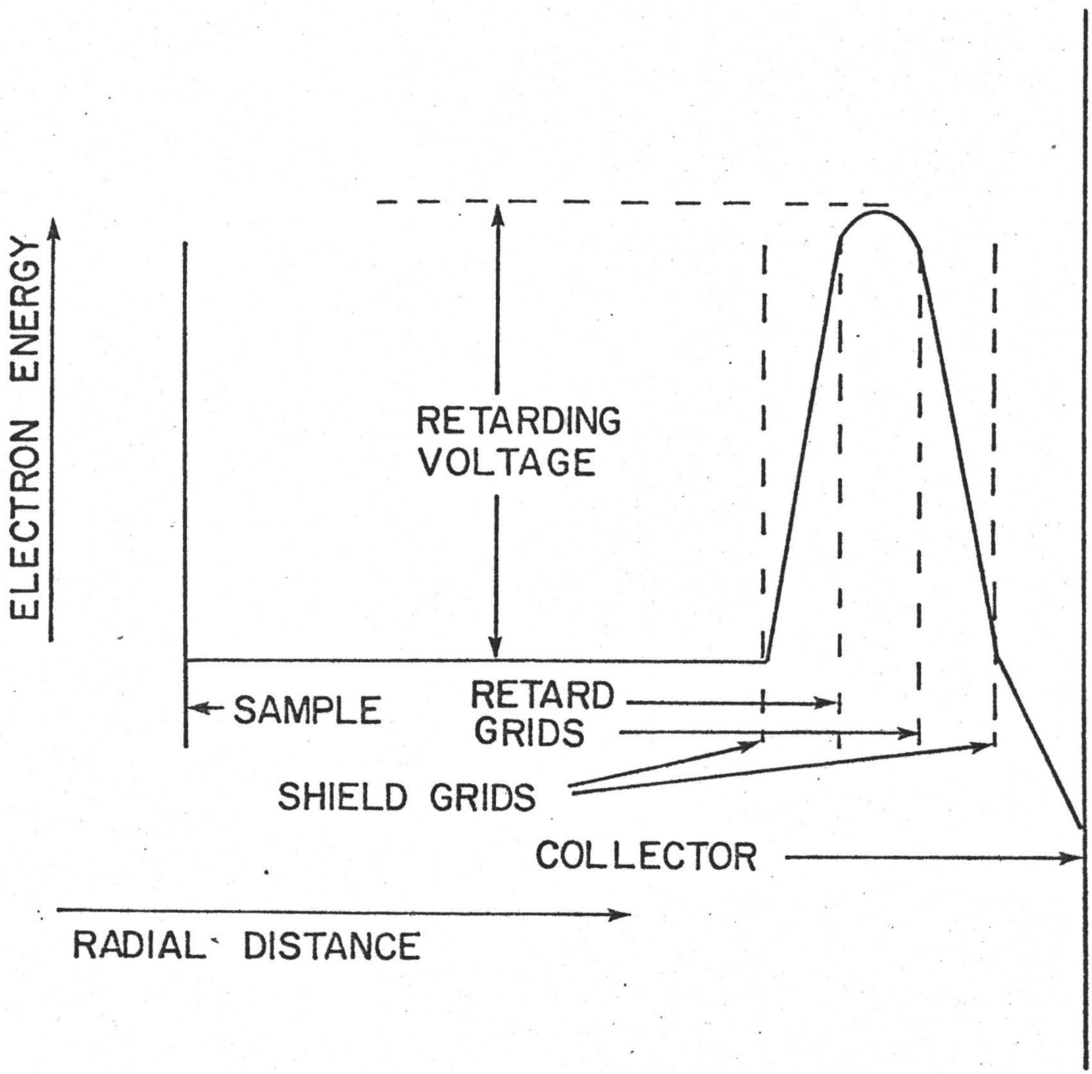
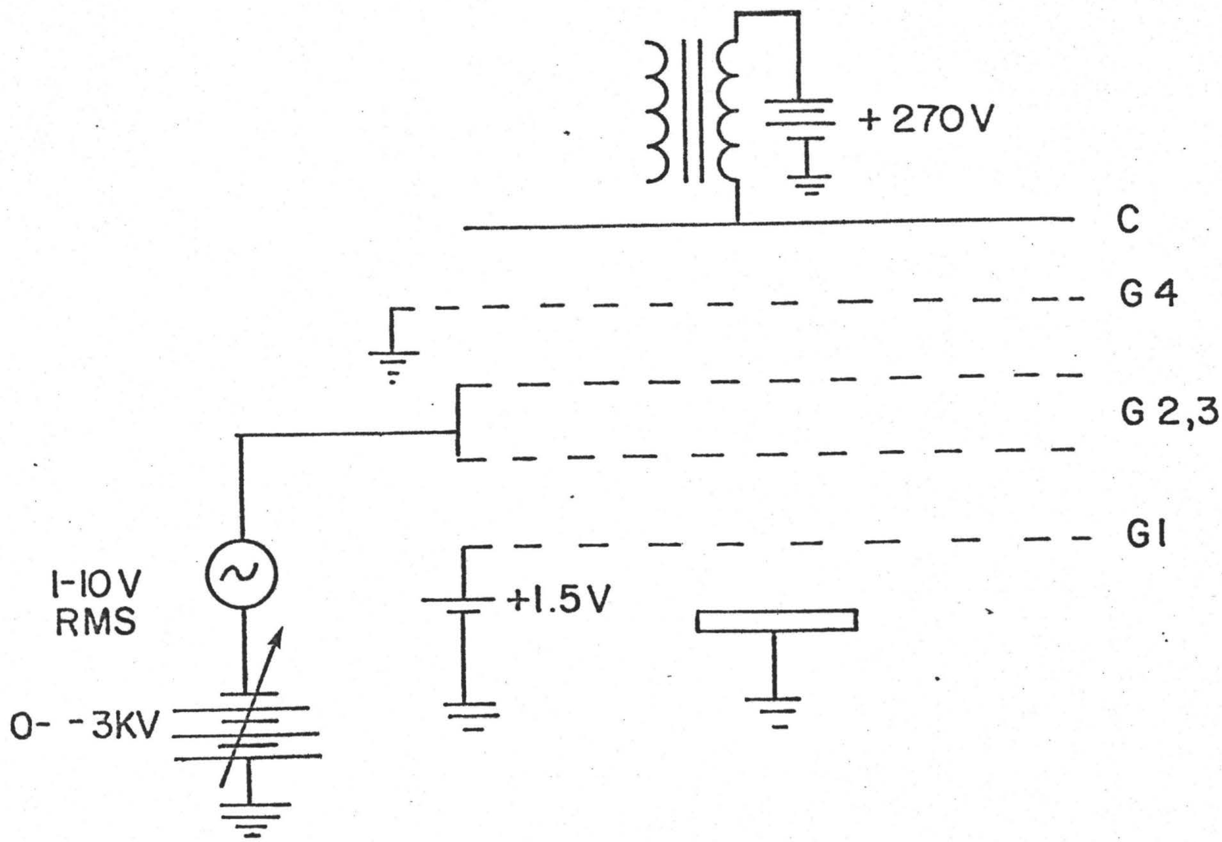
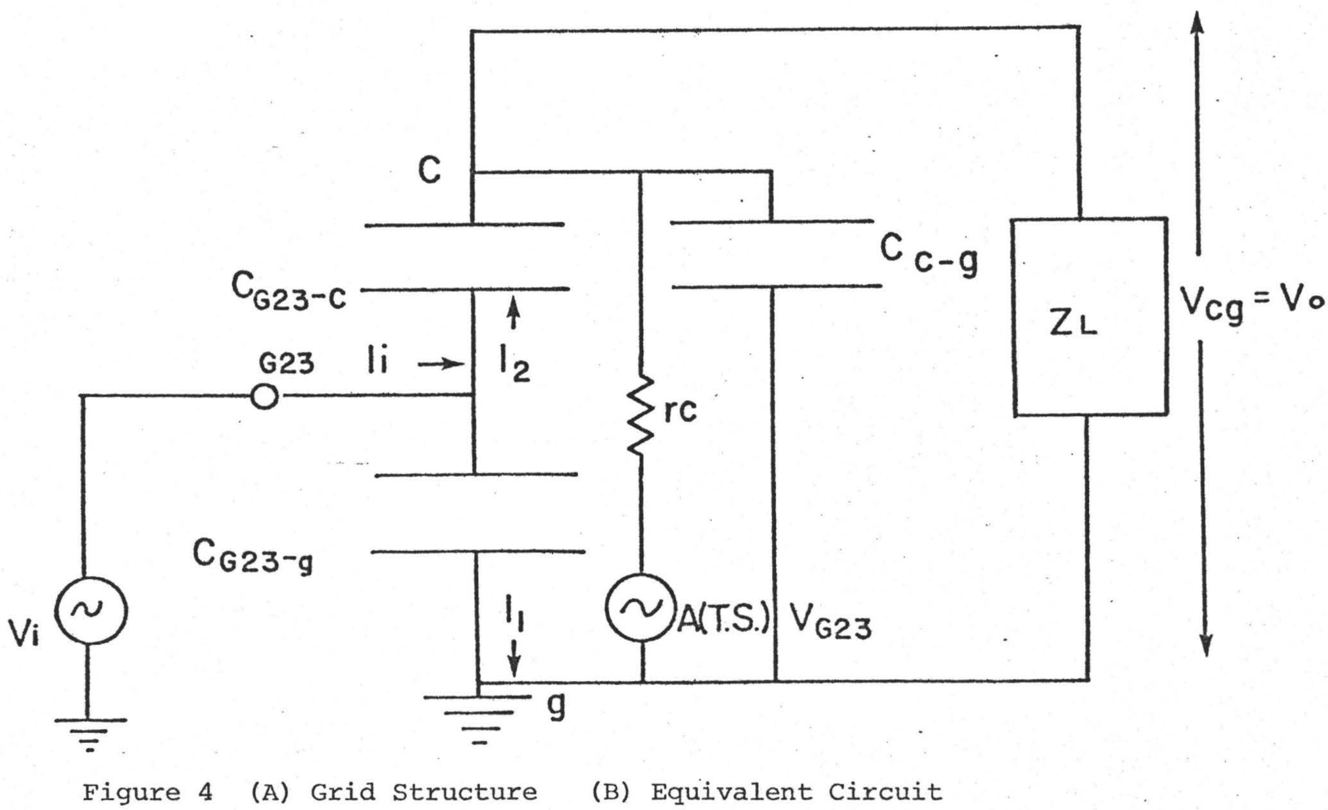


Figure 3 Energy variation through analyzer.



(A)



(B)

Figure 4 (A) Grid Structure (B) Equivalent Circuit

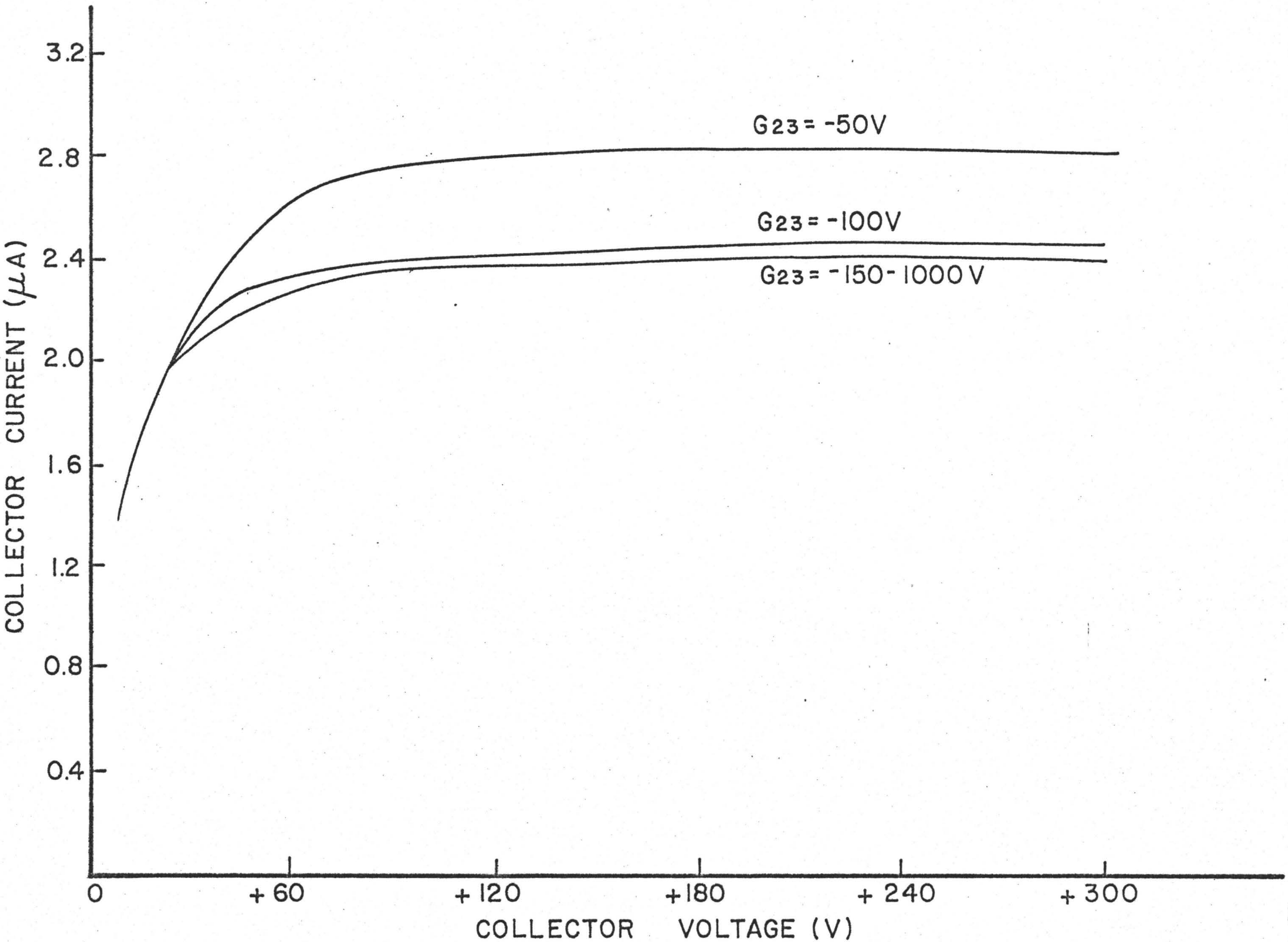


Figure 5 Collector current vs. voltage for primary beam current of 10μA.

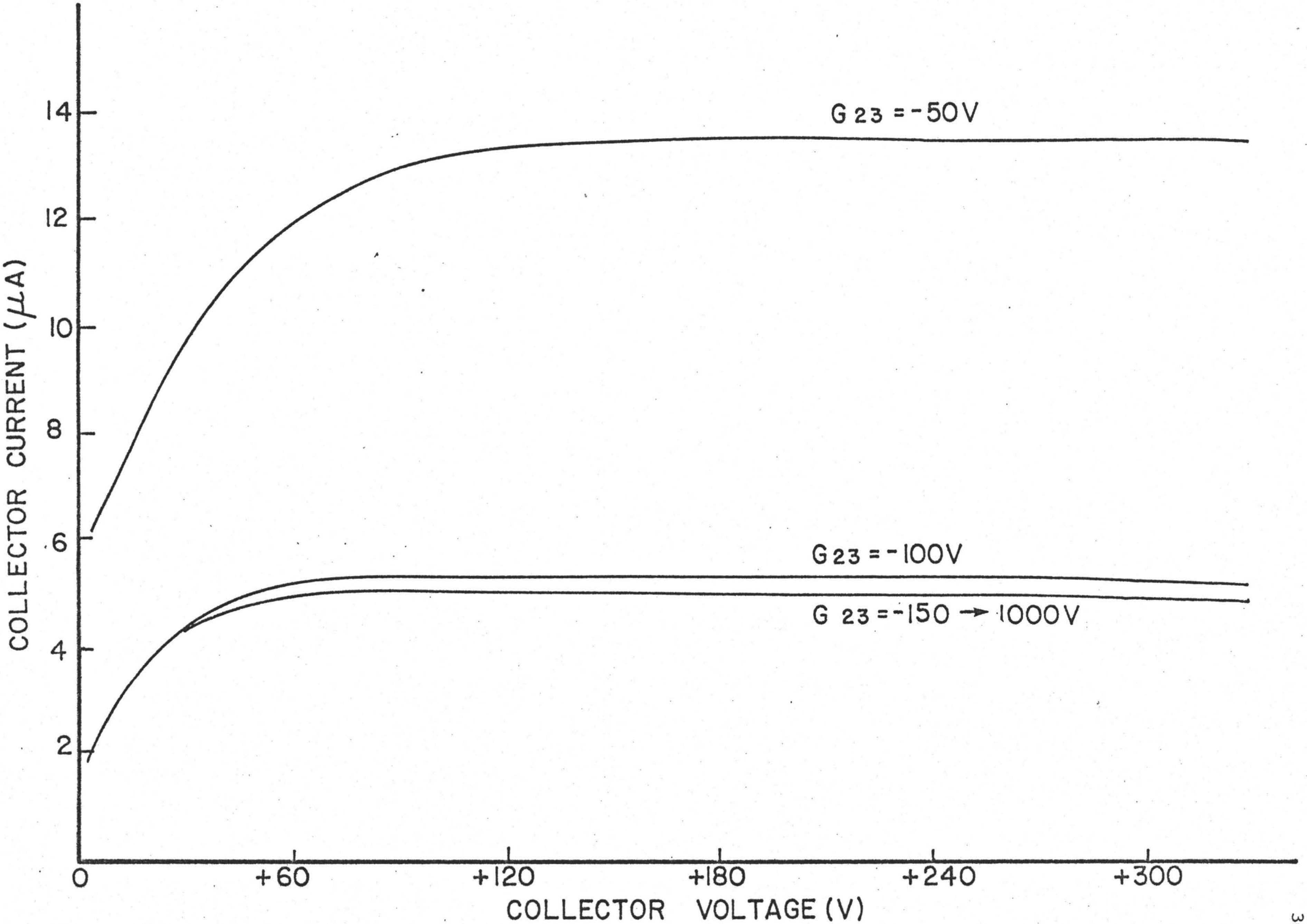


Figure 6 Collector current vs. voltage for primary beam current of 25μA.

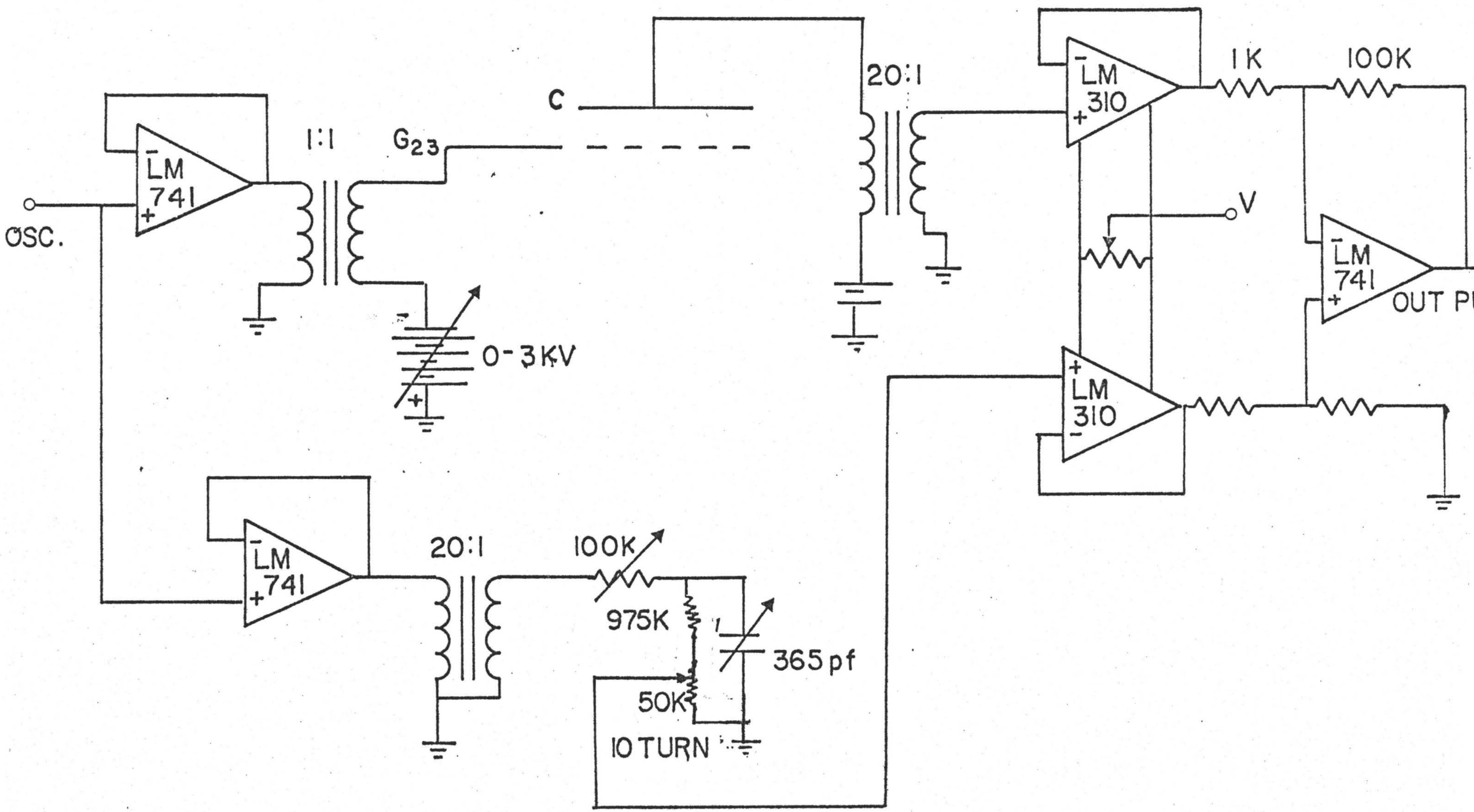


Figure 7 Neutralizer Circuit.

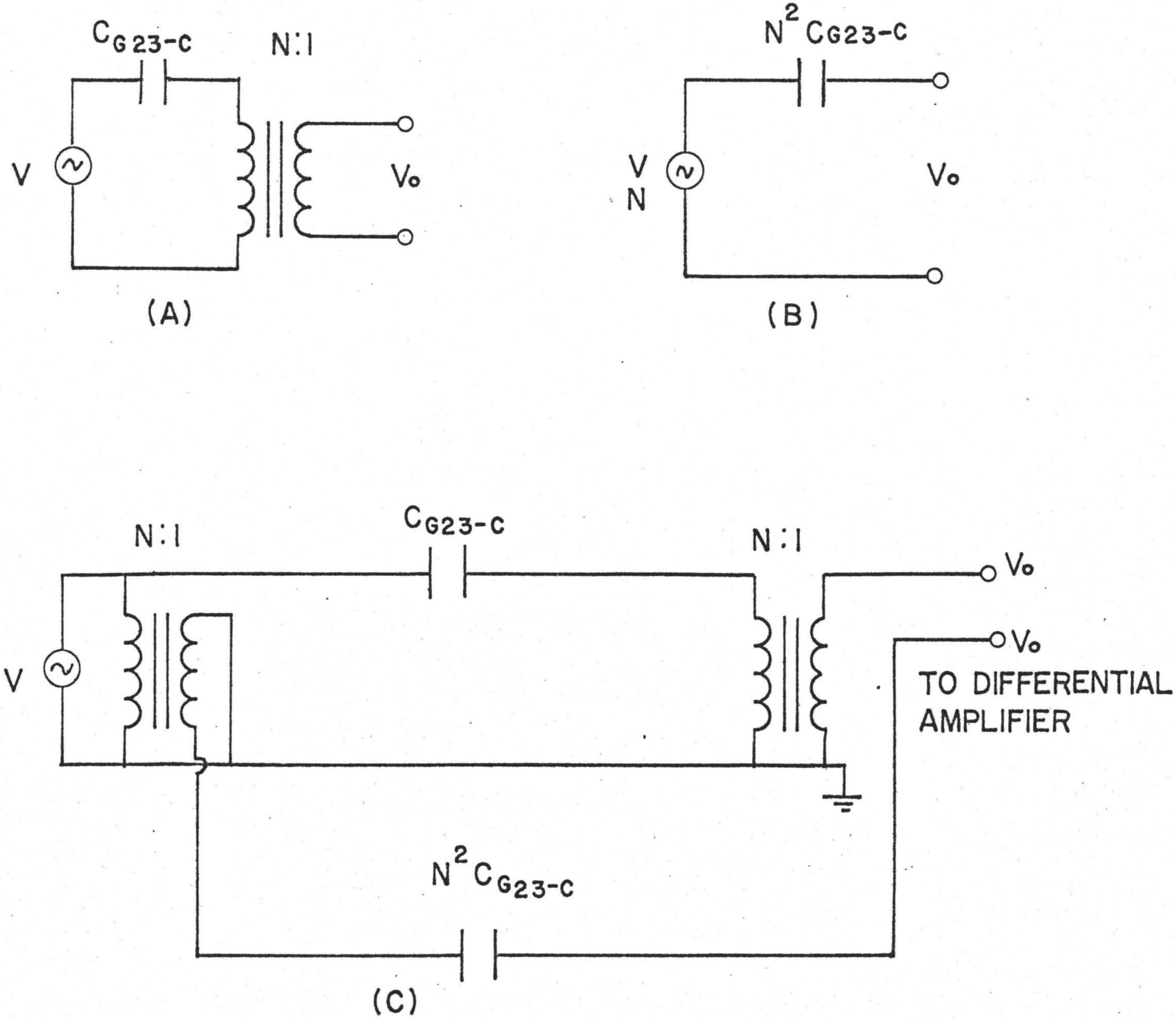


Figure 8 Operation of Neutralizer.

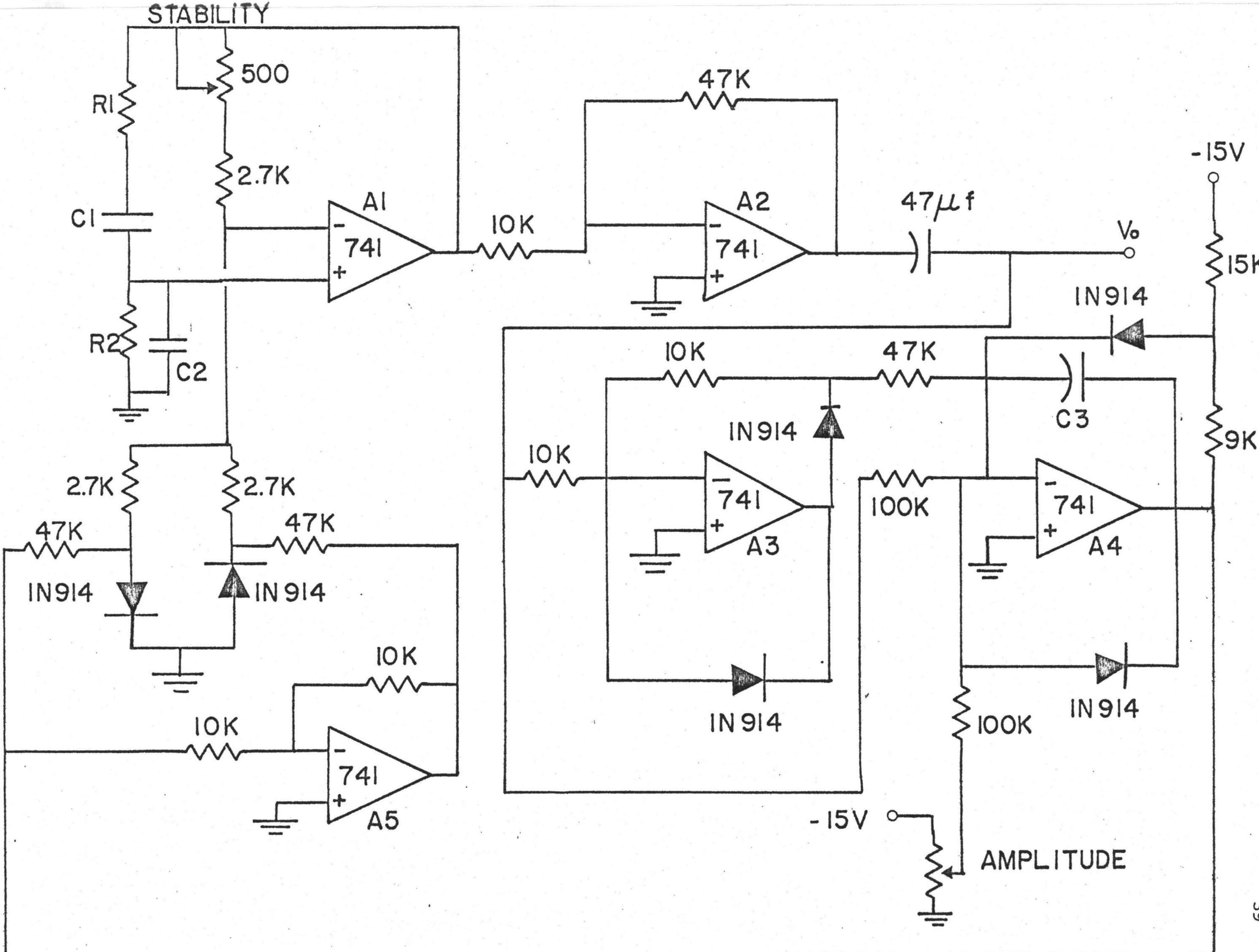


Figure 9 Modulating Oscillator.

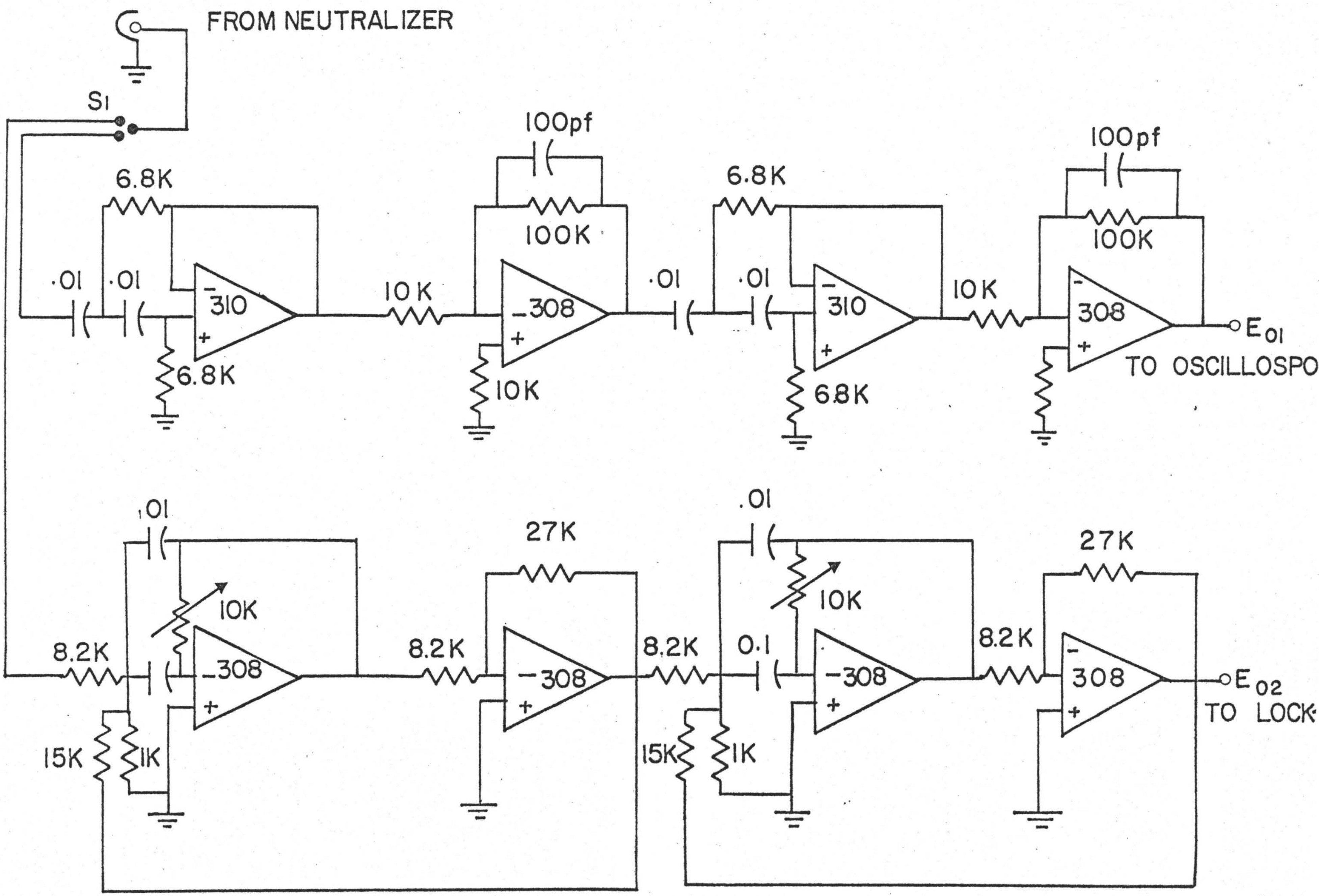


Figure 10 Setup Unit and Bandpass Filter.

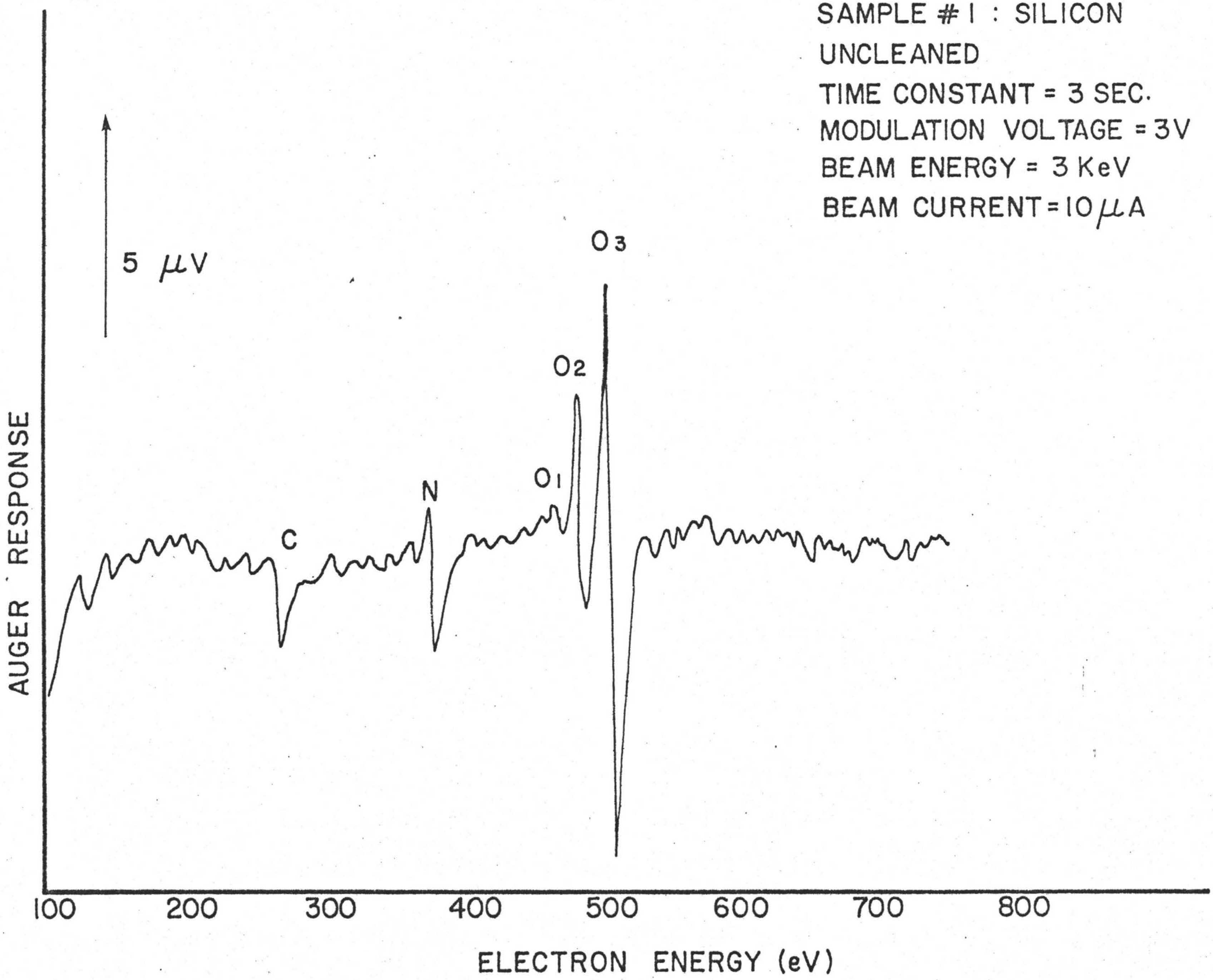


Figure 11 Auger spectrum of sample #1.

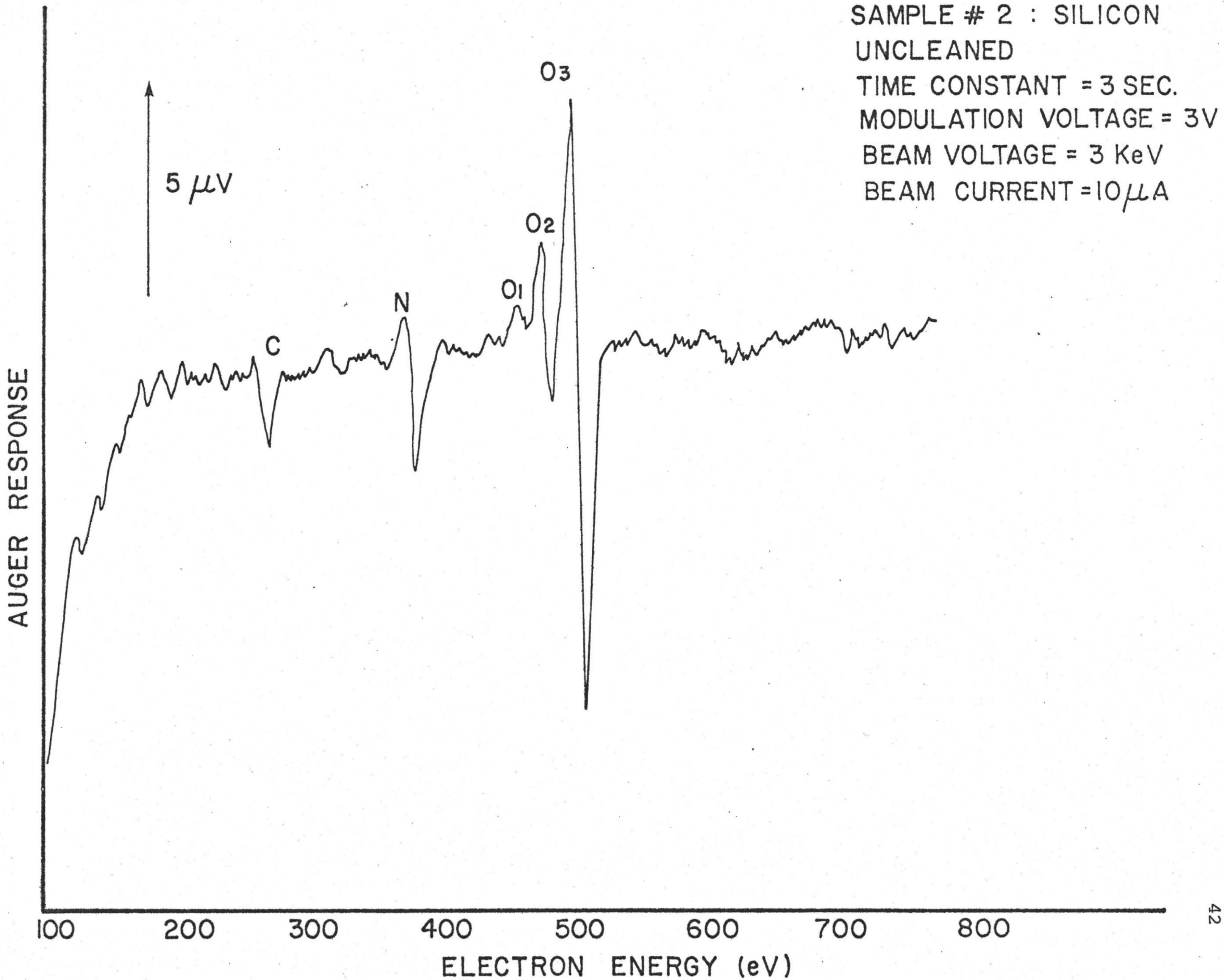


Figure 12 Auger spectrum of sample #2.

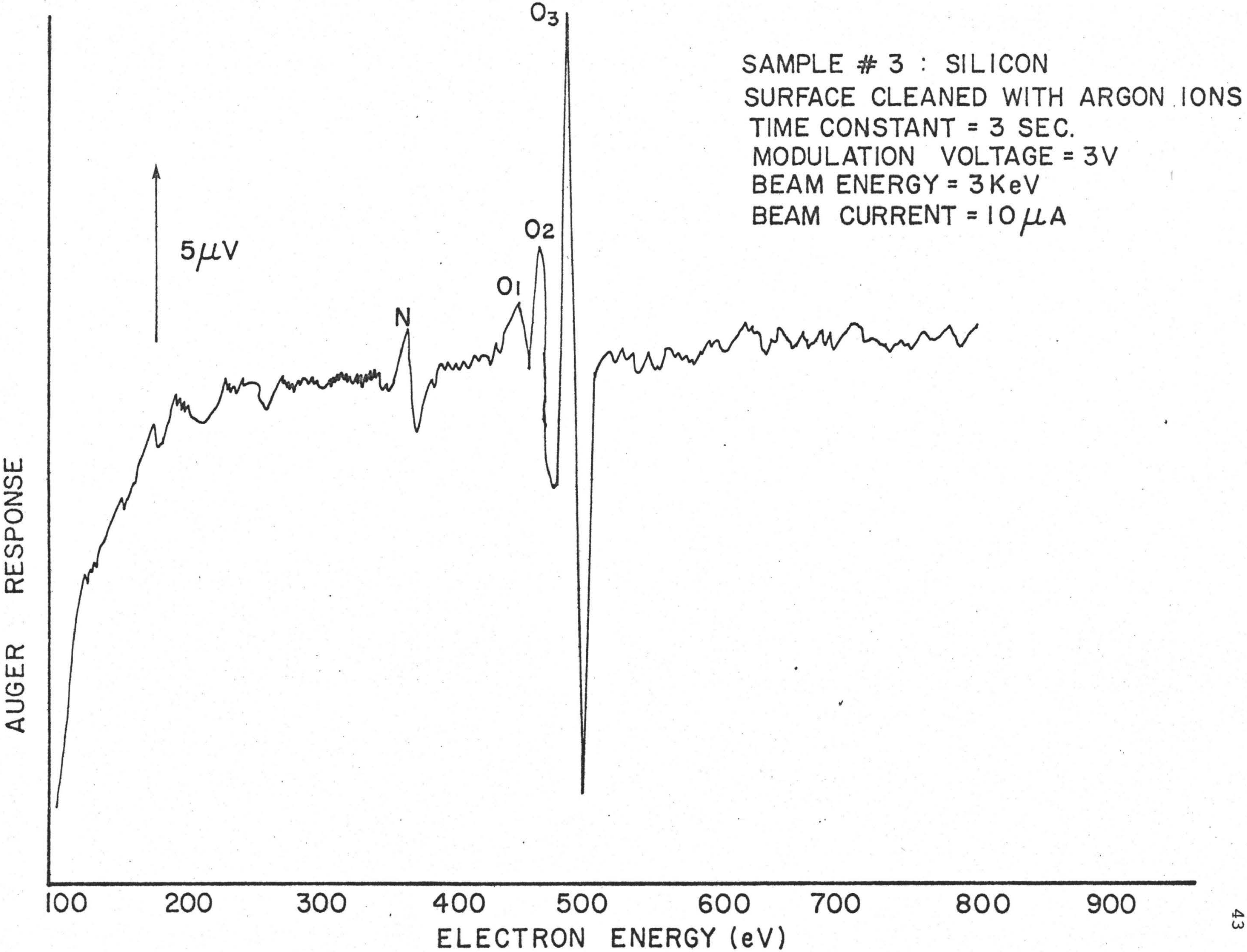


Figure 13 Auger spectrum of sample #3.

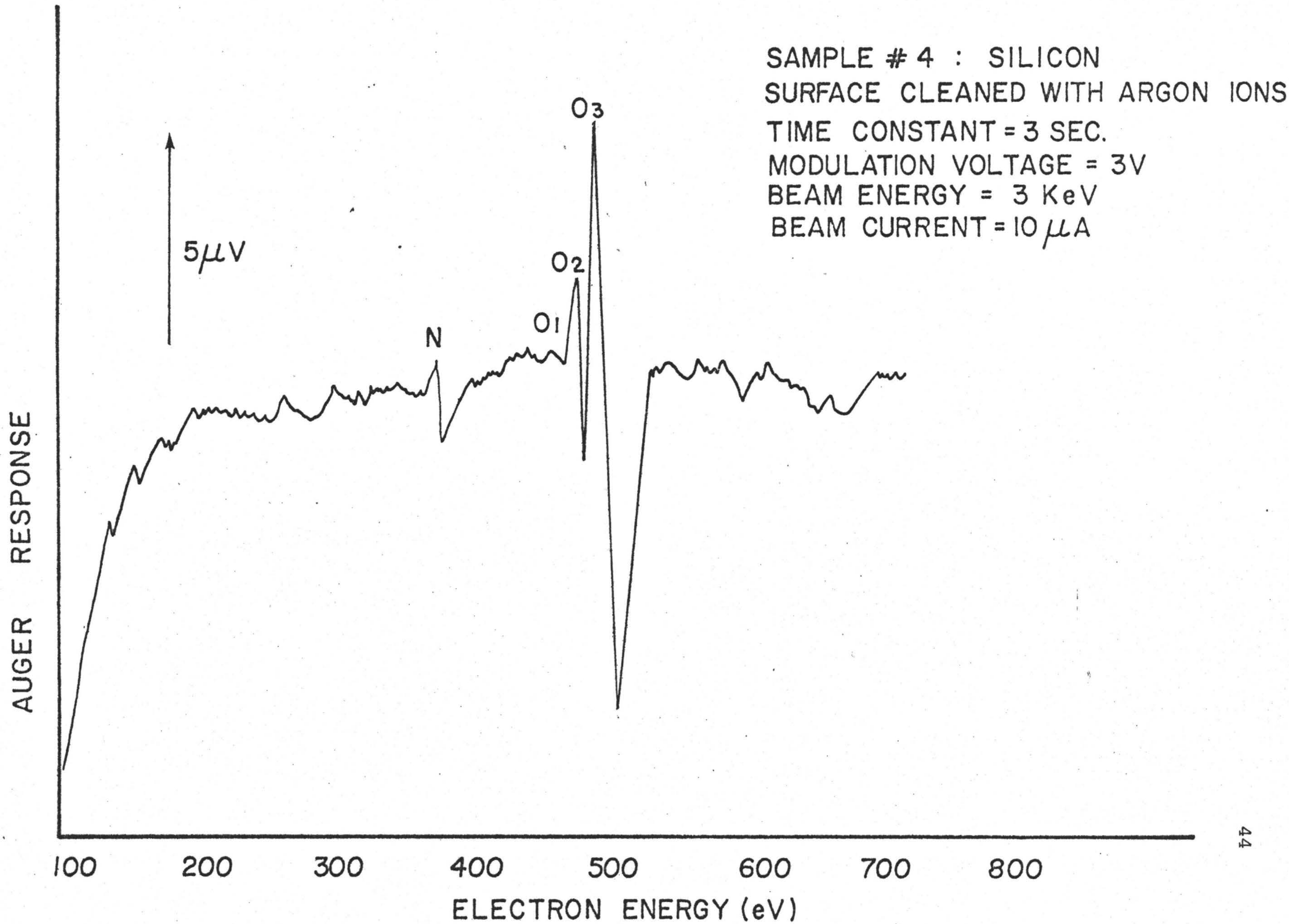


Figure 14 Auger spectrum of sample #4.

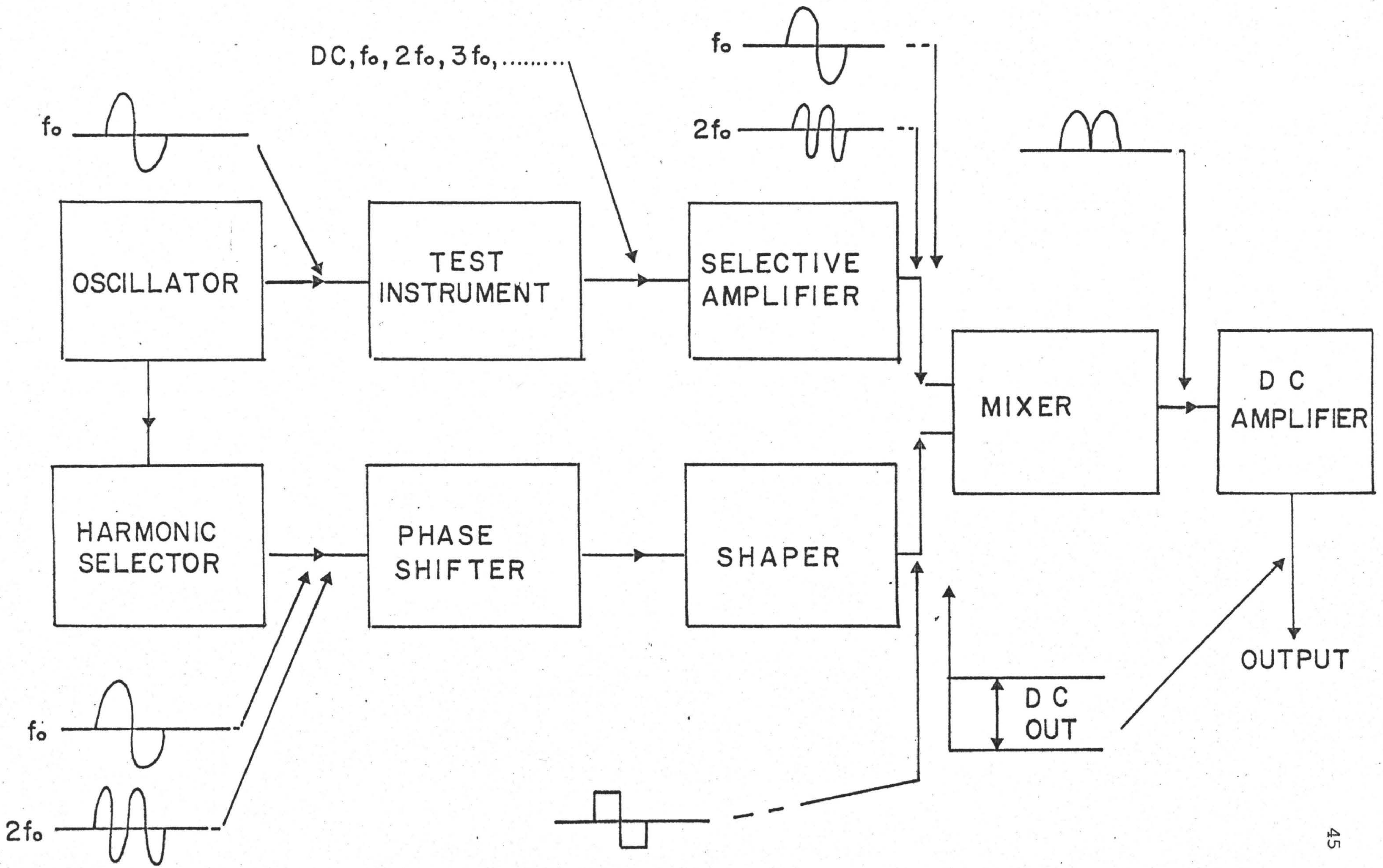


Figure 15 Block diagram of lock-in amplifier.

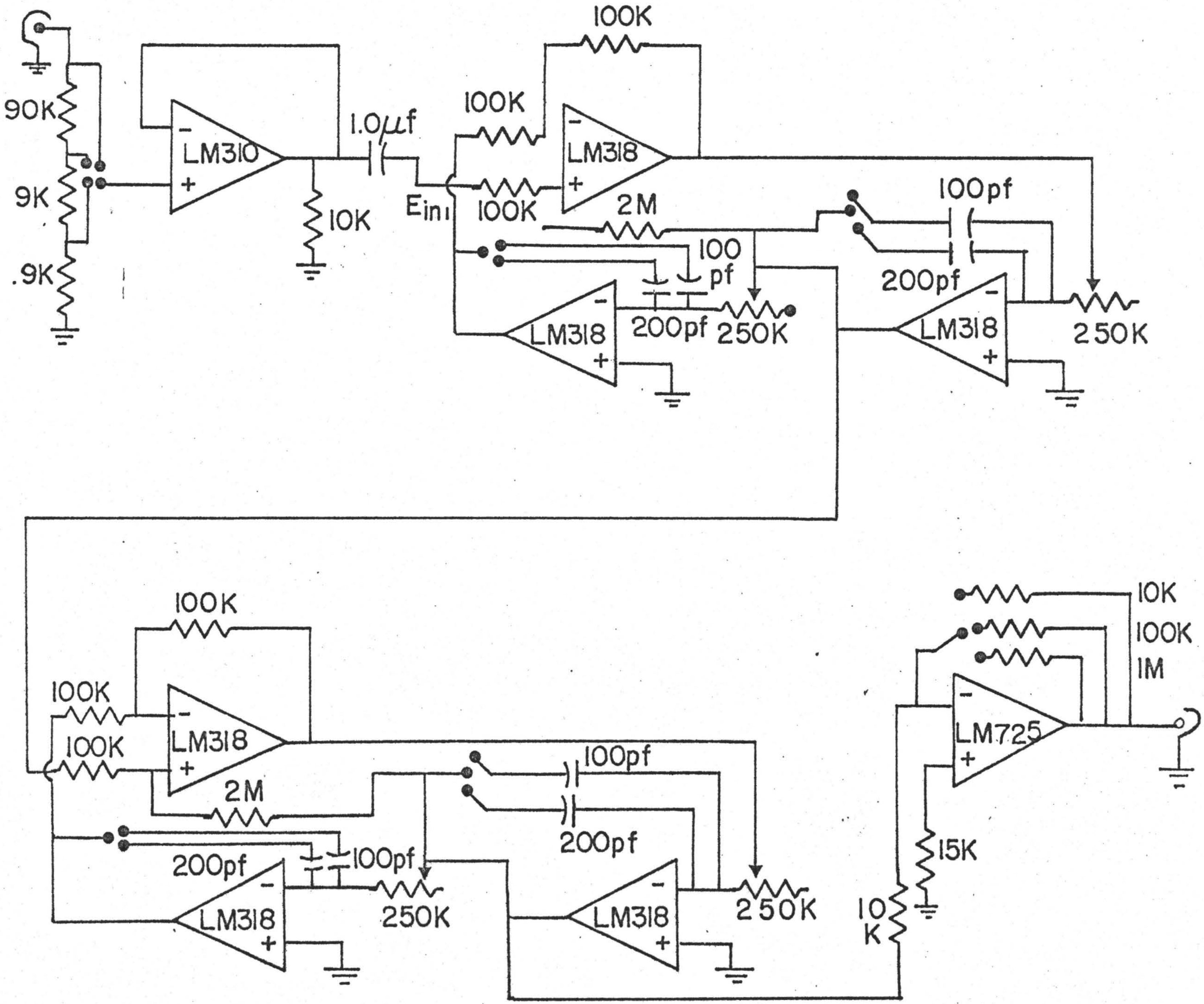


Figure 16 Schematic of selective amplifier.

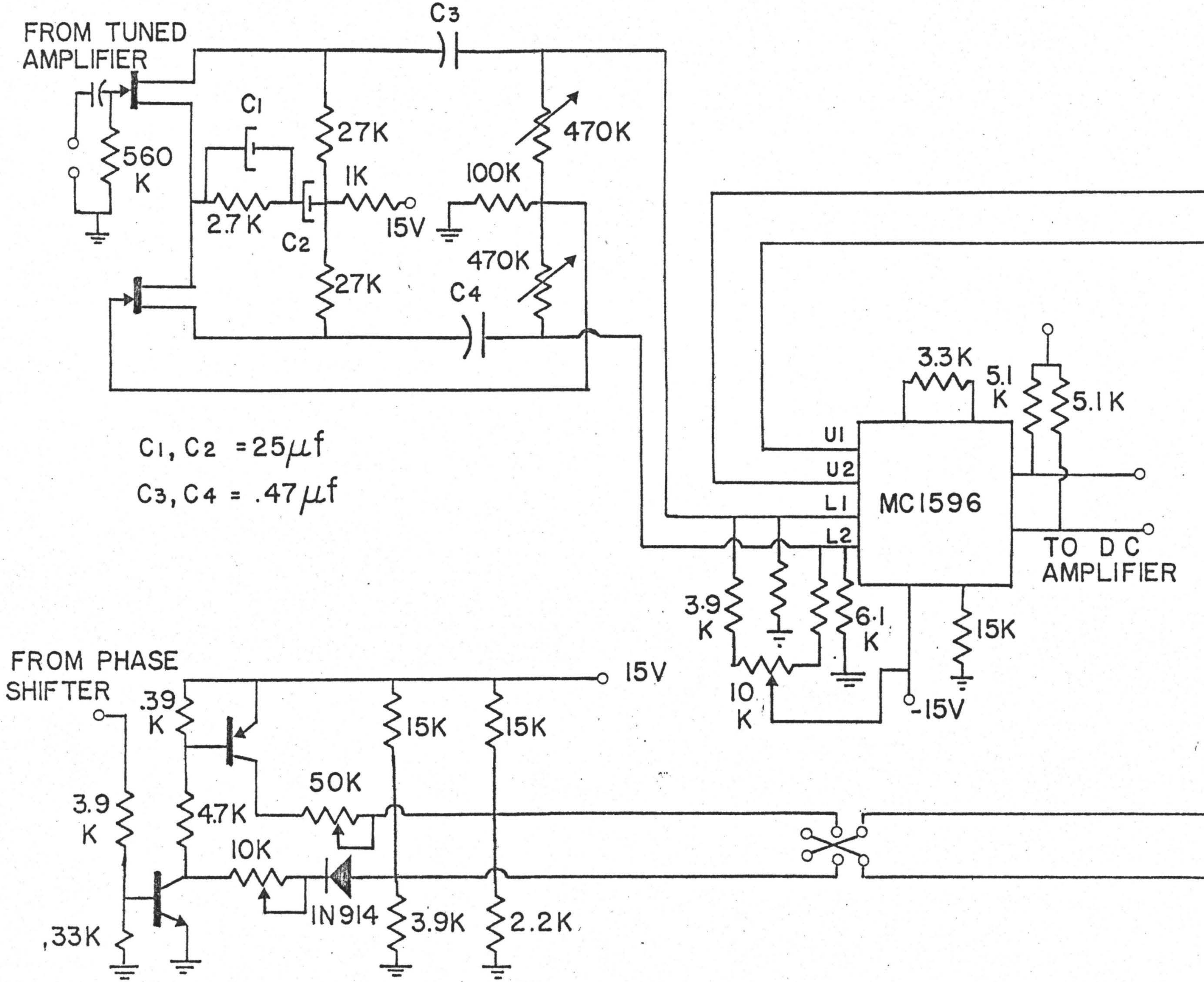


Figure 17 Schematic of phase splitter, reference shaper and mixer circuit.

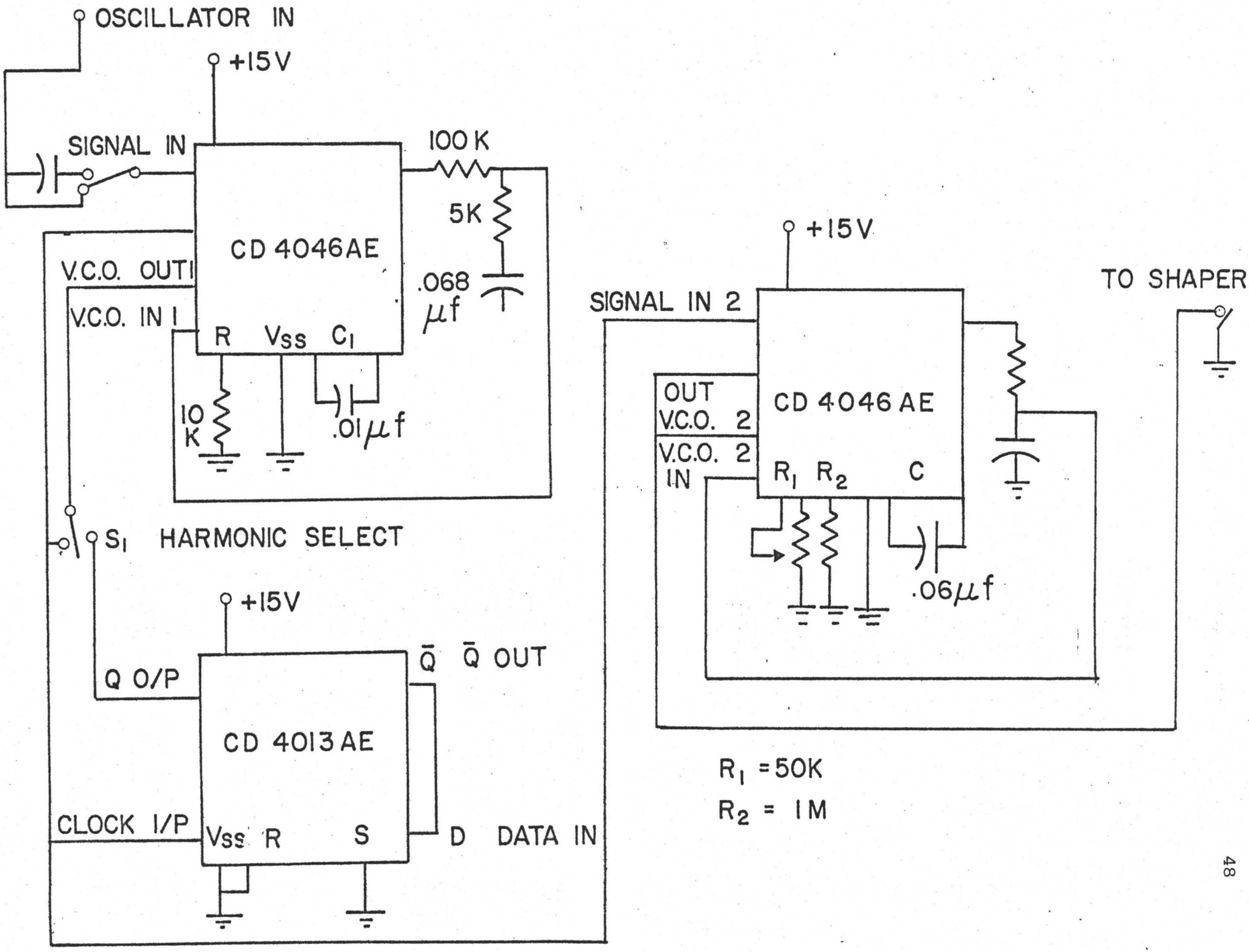


Figure 18 Schematic of harmonic selector and phase shifter circuit.

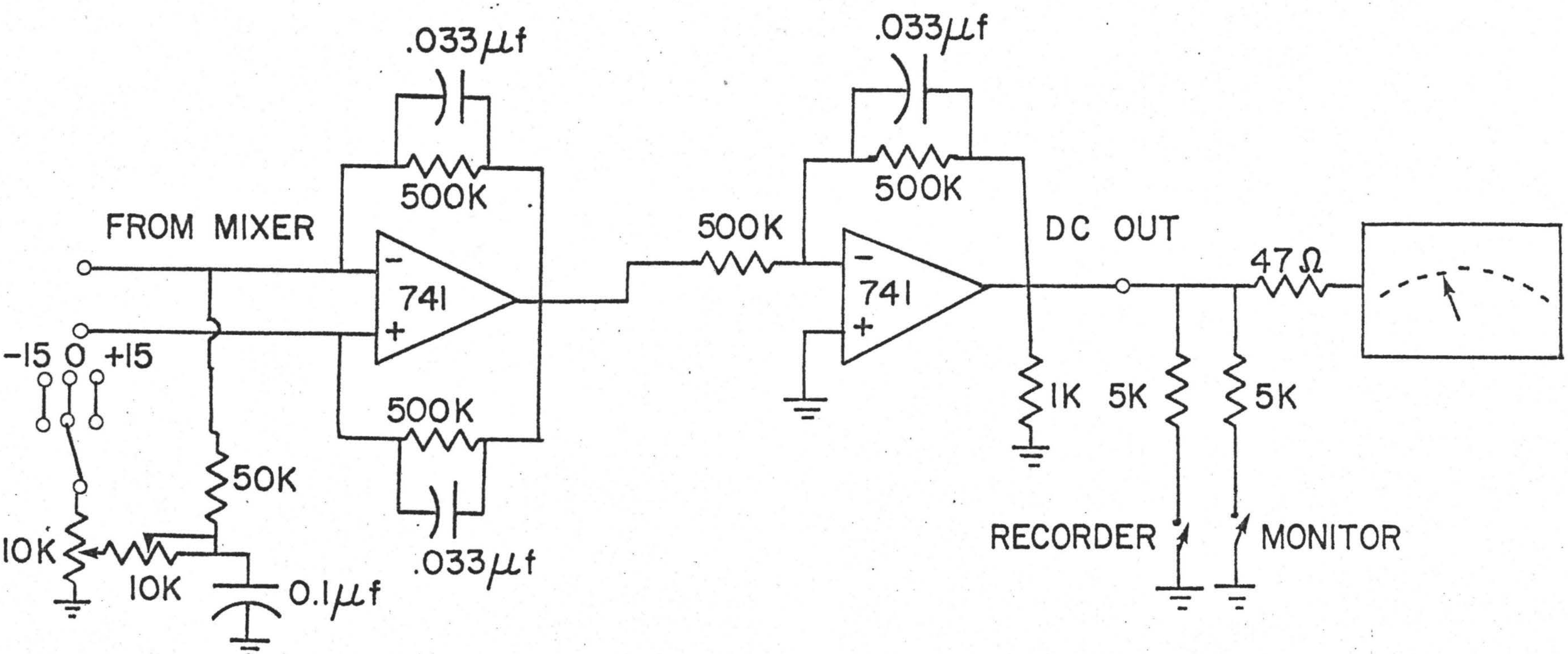


Figure 19 Schematic of D.C. amplifier and output circuit.

References

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