A LOW TIME CONSTANT ELECTRONIC INTEGRATOR FOR TIME DEPENDENT FUNCTIONS

by

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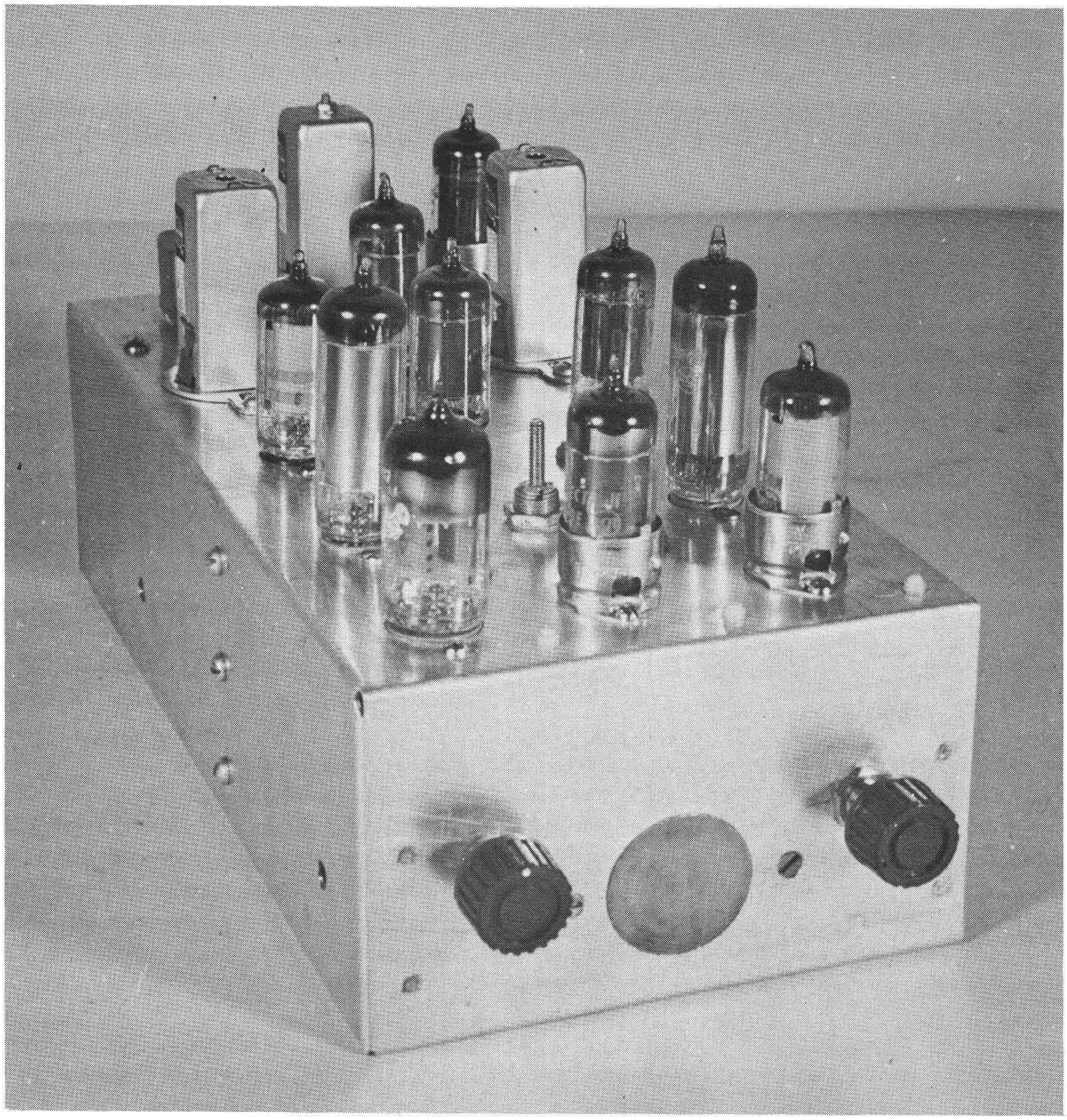


Figure 1. Electronic Integrator

A LOW TIME CONSTANT ELECTRONIC INTEGRATOR FOR TIME DEPENDENT FUNCTIONS

INTRODUCTION

In the analysis of pressure broadening of spectral lines, the luminous energy within the line appears as a function of wave length. By photographic and photoelectric methods this relationship may be transformed into a time variation of electric potential. From a knowledge of the energy distribution within the spectral line, some understanding of the effect of gaseous pressure on light emission may be gained.

One method of obtaining a representation of the intensity variation within a line is to photograph the appropriate portion of the spectrum and apply the resulting photographic plate to microphotometer analysis. Over a small portion of the spectrum the prism or grating in the spectograph produces a linear dispersion on the photographic plate. However, the density of emulsion blackening is not a direct linear function of the light intensity.

The uniform movement of the microphotometer carriage which holds the photographic plate converts any distances traversed into a linear function of time The photomultiplier tube and its associated optical system change the density variations in the photographic plate to current variations in an electrical circuit. These current variations are used to drive the pen of a strip recorder which, by virtue of its uniformly moving paper and current controlled pen, plots the time variation of current on the recording paper. Thus, except for the nonlinearities of the photographic process involved, light intensity so recorded is a function of wave length.

The procedure in the past, at this point, was to re-plot the entire trace manually using a transfer curve which showed the relationship of known intensities to film darkening. Then it was necessary to count squares in order to find the area under the curve, and so obtain a representation of the total energy.

As might well be imagined, such a process was very laborious and time consuming. This prompted the study of faster ways of analyzing the data in order to speed up basic research, and led to the development of a two unit electronic data processing system. The first part is an analog transfer function synthesizer (2) which compensates for the non-linearities of the photographic process, thus providing a linear relationship of spectral intensity to current output. The second unit is a low time constant electronic integrator for time dependent functions for determining the area under the curve. Its design, construction, and testing is the subject of this thesis.

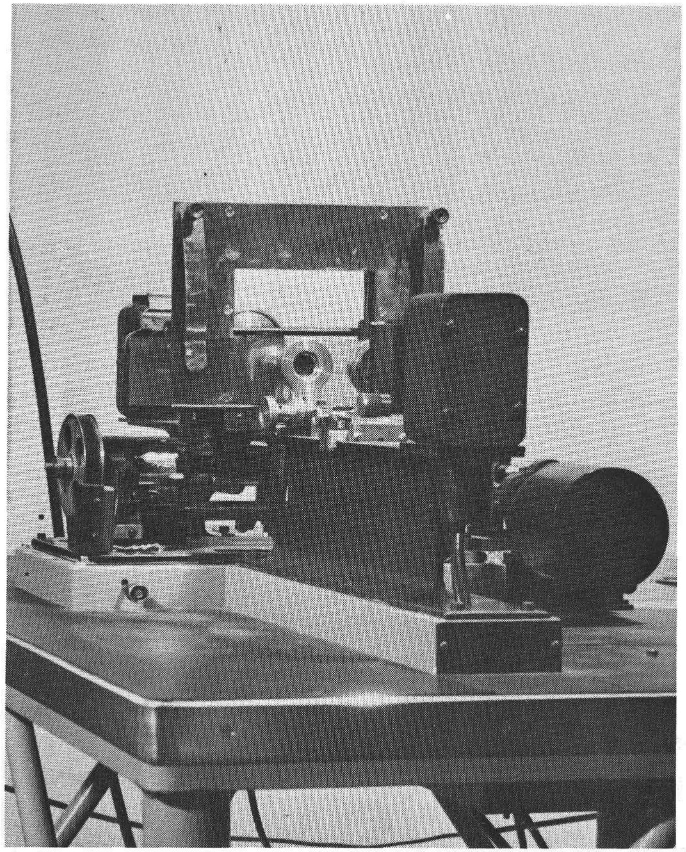


Figure 2a. Microphotometer

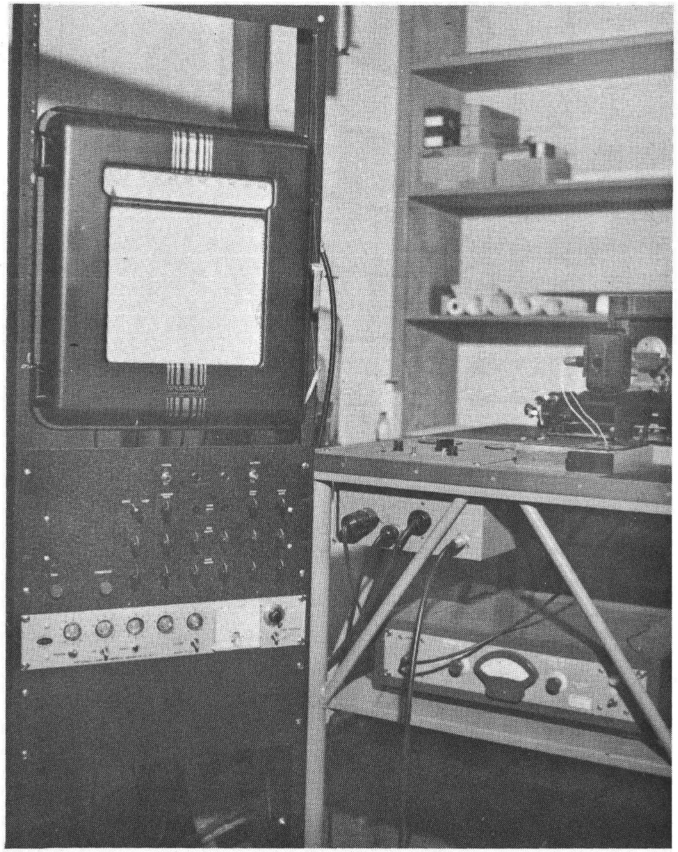
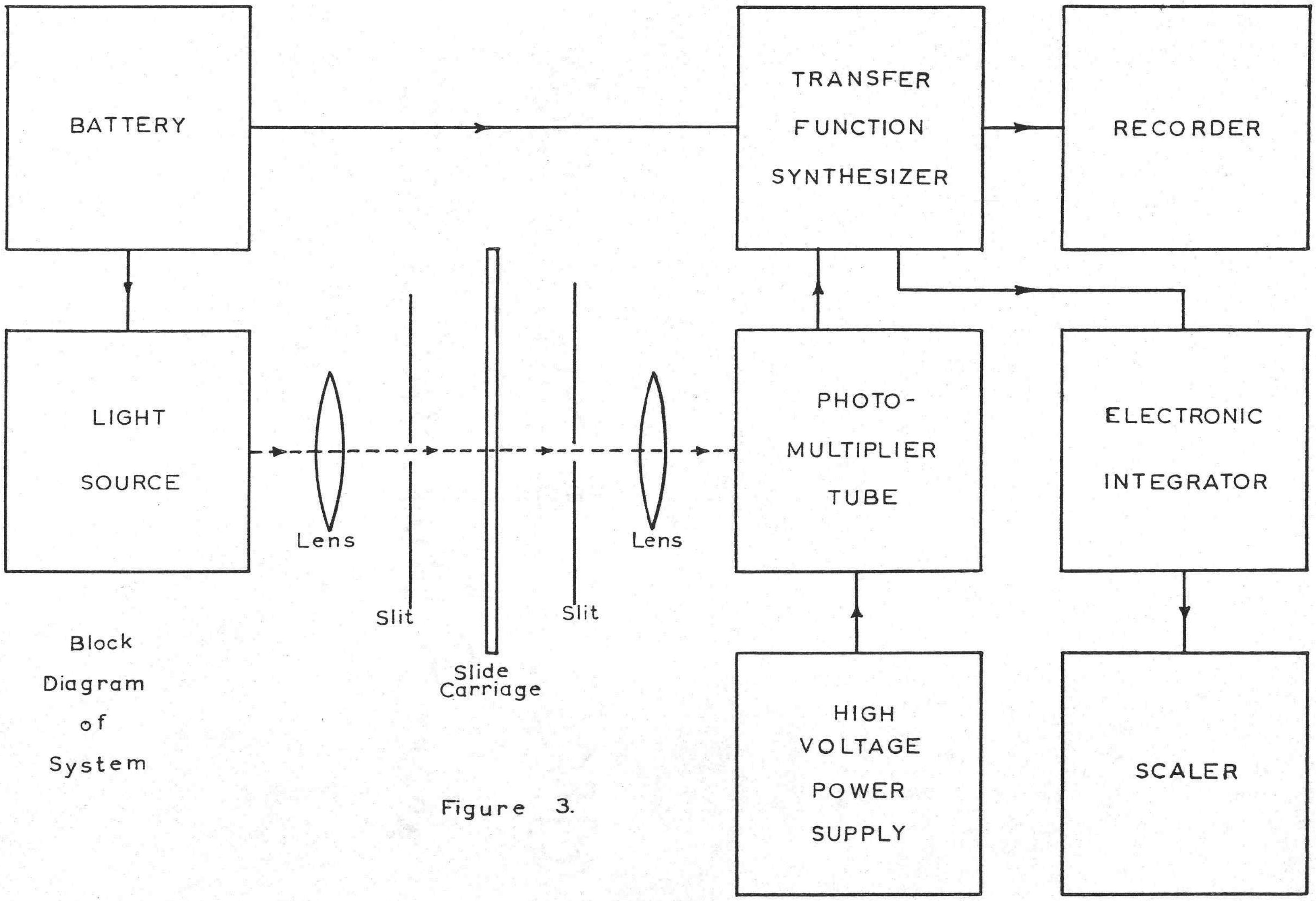


Figure 2b. Data Analyzing System



Block Diagram of System

Figure 3.

METHODS FOR INTEGRATING TIME DEPENDENT ELECTRIC CURRENTS

There are many ways of integrating time dependent functions, but the restrictions set by the microphotometer and other aspects of this application limit the number of possible methods.

In order to operate in conjunction with the microphotometer and the rest of the apparatus the integrator must be able to respond to varying levels of direct current, since the period of the wave form to be integrated is on the order of 15 to 30 minutes. It must accept changing current values without regard to the sign of the first time derivitive at any point on the wave form. It should provide an accurate and rapid means of read-out.

The first and most obvious method of integrating a time dependent function is to use a resistor-capacitor integrating circuit. The time-varying voltage is applied to a resistor-capacitor circuit, and the total charge is read on a device such as a ballistic galvanometer. This method will work on wave forms with fairly short periods if the applied voltage is always well above the capacitor voltage. The problem of capacitor leakage and the fact that the capacitor will discharge if the input voltage falls below the voltage on the capacitor are considerations which discourage its use in an application of this kind.

It is possible to use an electroplating bath to integrate time dependent current functions. In this method the cathode and anode in the electroplating solution are weighed carefully before the current is allowed to deposit metal of known density on the cathode. The electrodes are reweighed accurately after the wave form has run its complete cycle; the difference in weight is used to calculate the total charge that flowed in the circuit, and thus the integrated current function is obtained. This procedure can be used on slowly changing current functions whether increasing or decreasing, but requires as much or more work than does the square-counting method.

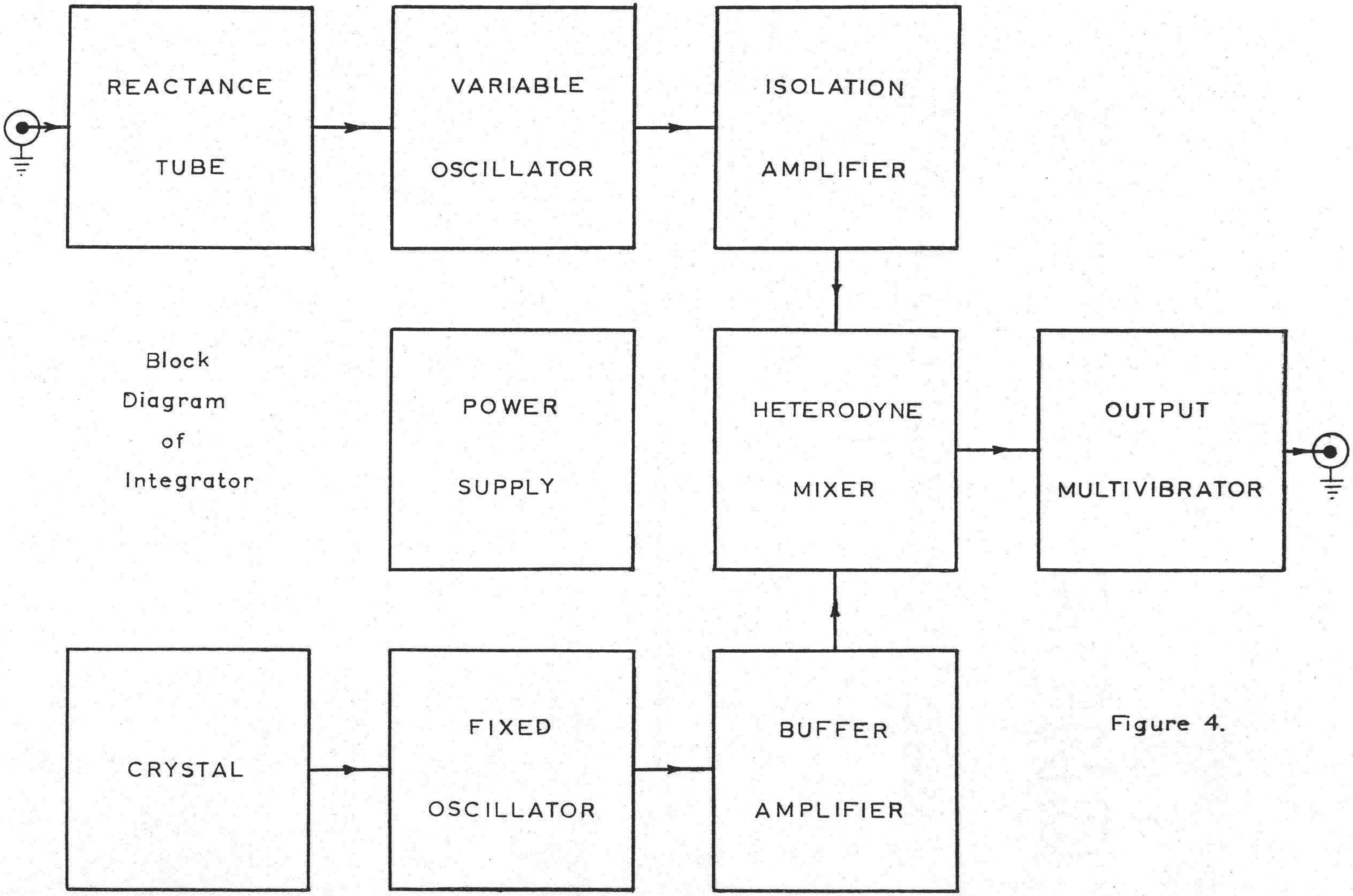
A method of much merit which fulfills all requirements is to use a mechanical system with a gear or ball type integrator operating much like the differential on a car. One of the inputs to the integrator can be controlled by a synchronous motor and thus provide the constant time function. The other shaft input can be driven by a motor-servo-system, whose rate of rotation is controlled by the input voltage or current. The output can be immediately and accurately read from a rotation counter on the output shaft. Other than being cumbersome and expensive this device would be acceptable; in fact it was the electronic analog of this system that led to the development of the all-electronic system which will be described in the following pages.

AN ELECTRONIC METHOD OF INTEGRATING TIME DEPENDENT FUNCTIONS

By careful consideration of the electronic analogs of the mechanical system outlined, an all-electronic system was devised. It was noted that the revolution of the synchronous motor could be replaced by the sine wave output of a crystal controlled oscillator, while the variable speed servo-motor could be superseded by a variable frequency oscillator. The purpose of the rotation counter could be served by an electronic scaler of the type used in nuclear work. In search of an analog for the ball or gear type mechanical integrator, it was found that by mixing the two signals in a hetrodyne frequency converter of the double input type, a multiplicative function was observed at the output which could be converted to an integrated function of the variable frequency with respect to time.

The variable oscillator frequency would necessarily be controlled by an input voltage or current since such was the output of the photomultiplier tube. In order to change the frequency of an oscillator, either the capacitance or the inductance of the resonant circuit must be varied. Several devices and circuits are available for converting a voltage or current change into a capacitance or inductance variation, and all of these were considered before the use of a reactance tube was decided upon.

The integrator, as developed from the mechanical analog, begins with a reactance tube which converts any voltage variation at the input grid to a capacitance change which shifts the resonant frequency of the variable oscillator. This oscillation of variable frequency is



Block Diagram of Integrator

Figure 4.

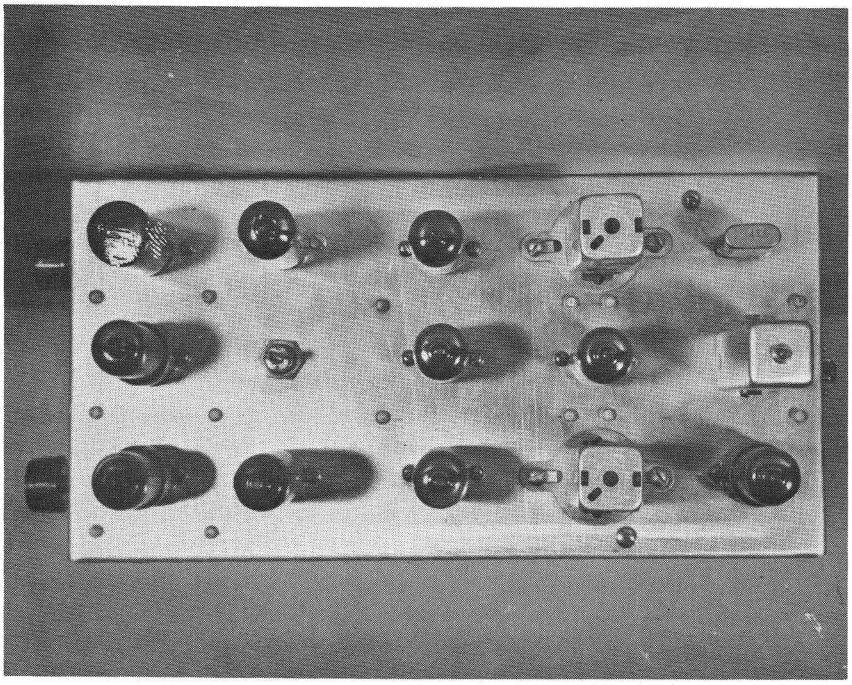


Figure 5a. Top View

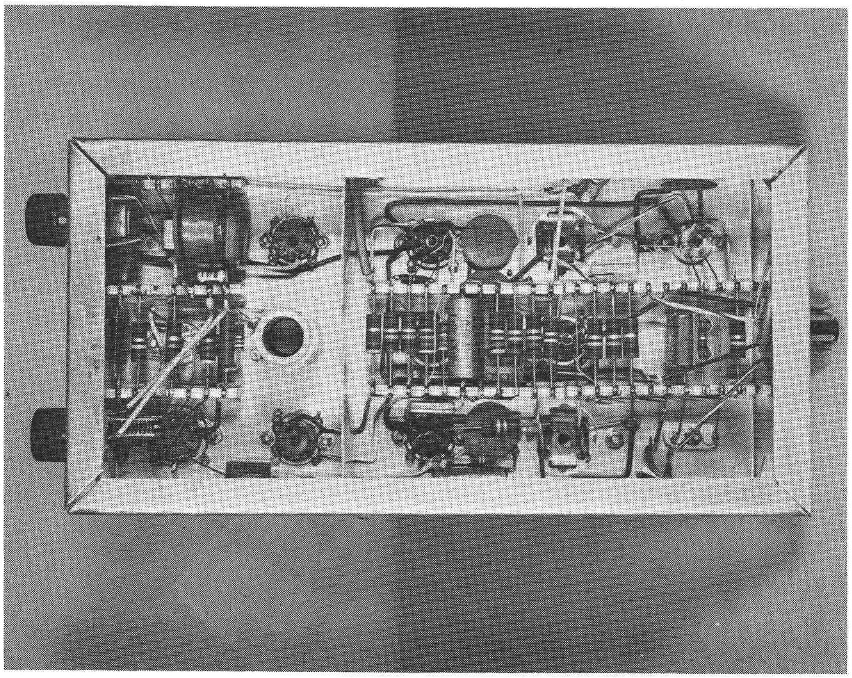


Figure 5b. Bottom View

mixed with the fixed frequency output of a crystal controlled oscillator in a non-linear multigrid vacuum tube. The resulting beat note is used to control a mono-stable multivibrator which provides a suitable output pulse for acceptance by the electronic scaler. The circuit constants are adjusted so that with no input voltage at the grid of the reactance modulator the frequencies of the two oscillators are the same, giving an output of the heterodyne mixer at zero cycles per second. As the potential on the grid of the reactance tube is varied, the frequency of the variable oscillator is changed proportionately, and the number of cycles per second from the heterodyne mixer is in direct ratio to the input voltage over the useful range of the instrument. If the total number of cycles for the duration of the input voltage function are counted on the scaler the reading is directly proportional to the area under a graph representing voltage as a function of time.

This, then is a device which, when presented with the voltage analog of any time dependent function, will integrate the area beneath a time plot of this function. There is no restriction as to polarity, sign of derivative or wave shape of the function to be integrated. The integrator will provide an instantaneous digital result which can be read from any convenient counting apparatus such as a glow tube scaler.

The original device was designed for application to a microphotometer in the analysis of spectral lines, or perhaps in the future development of a direct reading spectrophotometer. However, the utility of the device does not stop here, as it may be used in any application of teaching or research requiring the integration of a function which can be converted to a voltage-time relationship.

THE REACTANCE TUBE AND OTHER FREQUENCY CHANGING METHODS

There are several methods of changing the capacitance of inductance of a resonant circuit and thus the output frequency of an oscillator. The first of these to be considered is a device used extensively in sweep frequency oscillators in the second World War, and is called a "wobbulator". It is nothing more than a capacitor with one fixed plate and a movable plate attached to a driving motor which can be very similar to the driving coil of a speaker. This device is useful for a recurrent phenomenon such as is found in sweep oscillators but is not linear enough and is mechanically too large to provide good carrier frequency stability.

Another method used in commercial sweep oscillators is to vary the inductance by means of a high frequency saturable reactor called an increductor. In this device the control voltage is used to vary the current in the control winding which saturates the common core. This changes the inductance of the high frequency coil which is used as the tank coil of the oscillator. In order to produce a linear voltage to frequency change the associated circuit is precision designed and engineered, and is therefore very expensive as compared to other methods.

A comparatively new way of changing the capacitance in an oscillator circuit is to use a voltage variable capacitor, which is a semi-conductor diode biased in reverse direction. As the bias voltage across the device is increased the stronger electric field causes the transition region of the junction, called the space charge depletion layer, to widen by pulling the free charge carriers farther back into the P and N regions. This leaves a region of fixed charges which may be considered the dielectric of a capacitor. Thus the applied bias voltage causes a change in the thickness of the dielectric and thereby changes the capacitance.

There are two basic types of voltage variable capacitors, the grown junction type and the alloy junction type. From Early (1, p. 1285-1286) the alloy junction type capacitance is given by:

$$C_{pn} = \frac{\Delta Q}{\Delta V} = \frac{K\varepsilon_o A}{d_p + d_n}$$

Where:

$$d_p + d_n = \left[\frac{2K\varepsilon_o V_r}{q} \left(\frac{1}{N_p} + \frac{1}{N_n} \right) \right]^{\frac{1}{2}}$$

Thus the capacitance varies with the voltage as:

$$C_{pn} \approx \frac{1}{\sqrt{V_r}}$$

Where:

V_r	is the potential across the junction.
N_p & N_n	are the donor densities on the P & N sides respectively.
K	is the dielectric constant of the material.
q	is the charge on an electron.
ε_o	is the permittivity of free space.
A	is the effective area of the junction.

For the case of the grown junction:

$$d_p + d_n = 2\left(\frac{3}{2}\, \frac{K\varepsilon_o V_r}{q\,a}\right)^{\frac{1}{3}}$$

Where:

a = the net impurity atom density

Thus for the grown junction:

$$C_{pn} \approx \frac{1}{\sqrt[3]{V_r}}$$

Using the approximate formula for resonance:

$$f = \frac{1}{2\pi\sqrt{LC}}$$

We see that for the alloy type:

$$f \cong \frac{1}{\sqrt{C_{pn}}} \cong \frac{1}{\sqrt{\sqrt{V_r}^{-1}}}$$

$$f \approx \sqrt[4]{V_r}$$

And in the grown junction type:

$$f \approx \frac{1}{\sqrt{C_{pn}}} \approx \frac{1}{\sqrt{\sqrt[3]{V_r}^{-1}}} \approx \sqrt[5]{V_r}$$

This is far from a linear frequency to voltage variation in either case. Thus, even though these devices are small and convenient they cannot be used for critical applications until the semiconductor manufactorers provide a diode with a linear voltage to frequency relationship over a portion of its range. Such diodes have been produced, but only under controlled laboratory conditions.

The most practical method of obtaining a linear voltage-to-frequency change is to use a reactance tube. A reactance tube consists of a vacuum tube connected to the high frequency resonant circuit of an oscillator in such a way as to simulate a variable capacitance or inductance, of a value dependent on the control voltage applied to its grid.

The grid is driven from a resistance-reactance network connected between plate and cathode to provide a voltage almost 90° out of phase with the plate to cathode voltage. This plate-to-cathode voltage is derived from the resonant circuit of the oscillator, and the resulting plate current which is 90° out of phase due to the grid voltage, is injected back into the resonant circuit. Because of the voltage-to-current phase relationship, it is just as though an additional reactance had been added to the resonant circuit, with the important distinction that this injected reactance is voltage controllable.

Figure 6a. shows a simplified reactance tube, and its equivalent circuit appears on Figure 6b. For analysis it is desirable to find the output impedance of the equivalent circuit (3, p. 374-377).

As can be seen from Figure 6b.

$$I_c = \frac{E_{pk}}{R - jX_c}$$

Where:

$$X_c = \frac{1}{\omega C}$$

The grid potential is:

$$E_g = RI_c = \frac{RE_{pk}}{R - jX_c}$$

Then the plate current is:

$$I_p = \frac{E_{pk}}{r_p} + \frac{\mu}{r_p} E_g = \frac{E_{pk}}{r_p} + \frac{\mu}{r_p} \cdot \frac{RE_{pk}}{R - jX_c}$$

And: $g_m = \frac{\mu}{r_p}$

Summing these currents for the total current:

$$I = I_p + I_c = \frac{E_{pk}}{r_p} + g_m \frac{RE_{pk}}{R - jX_c} + \frac{E_{pk}}{R - jX_c}$$

The output admittance then is given by:

$$Y = \frac{1}{Z} = \frac{I}{E_{pk}} = \frac{1}{r_p} + g_m \frac{R}{R - jX_c} + \frac{1}{R - jX_c}$$

or:

$$Y = \frac{1}{r_p} + \frac{1}{R - jX_c} + \frac{1}{\frac{1}{g_m} - j\frac{1}{g_m R \omega C}}$$

This equation can be used as a basis for a new equivalent circuit as shown in Figure 6c. As far as the oscillator resonant circuit is concerned, since r_p and $R - jx_c$ are large compared to the other parallel terms and $\frac{1}{\omega CR}$ is large compared to unity, the output is almost purely capacitive with a value of $g_m RC$.

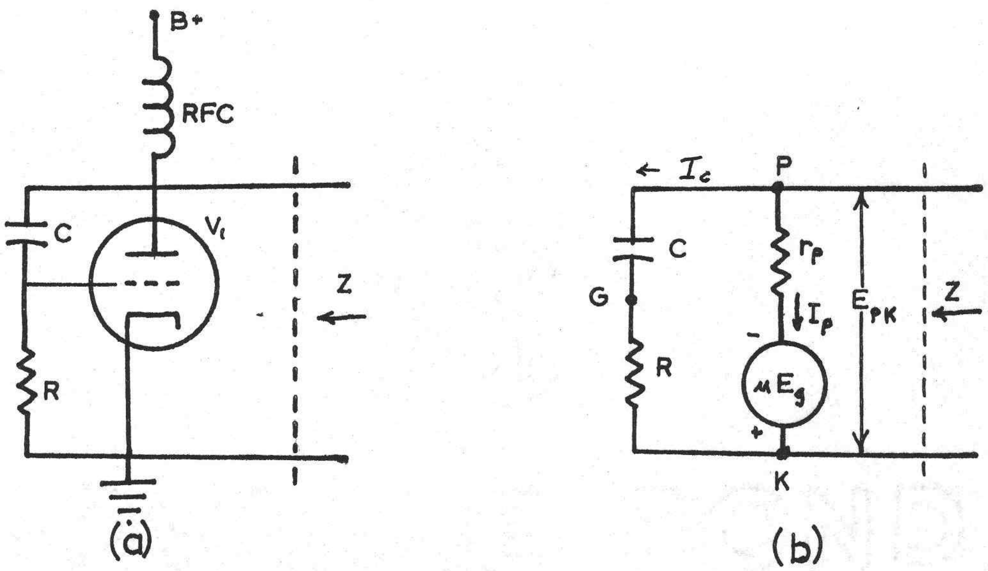
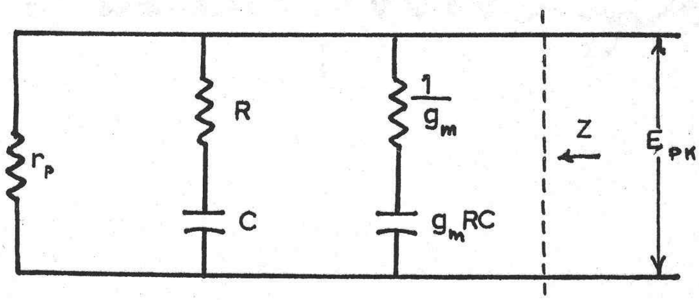
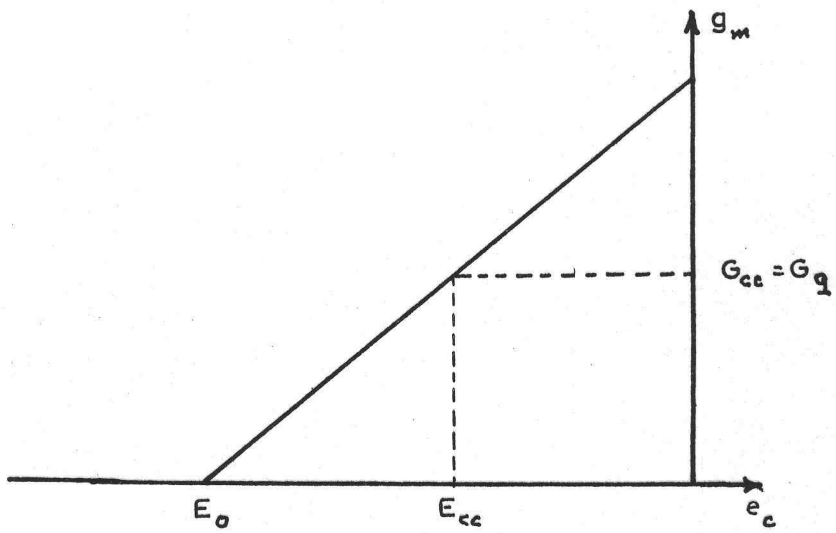

B+
RFC
C
V1
R
Z
(a)
I_c
P
C
r_p
G
I_p
E_PK
Z
R
μE_g
K
(b)

Figure 6.


r_p
R
C
1/g_m
g_mRC
Z
E_PK

(c)


g_m
G_cc = G_q
E_o
E_cc
e_c

(d)

As can be seen from the graph in Figure 7, the mutual conductance (g_m) referred to the first grid in a 6BE6 is a linear function of the third grid voltage over an extensive range. The simplified curve in Figure 6d shows that:

$$g_m = \frac{G_o}{E_o} e_c + G_o$$

Where:

$\frac{G_o}{E_o}$ is the slope of the line in Figure 6d.

$$e_c = E_{cc} + e_s$$

Where:

e_s is the instantaneous signal voltage.

$$g_m = G_o + \frac{G_o}{E_o} E_{cc} + \frac{G_o}{E_o} e_s$$

The effective output capacitance is

$$C_e = g_m RC = G_o RC \left(1 + \frac{E_{cc}}{E_o} + \frac{e_s}{E_o}\right)$$

Then assuming the approximate relationship for the frequency of the oscillator:

$$F = \frac{1}{2\pi\sqrt{L_o(C_o + C_e)}}$$

Where:

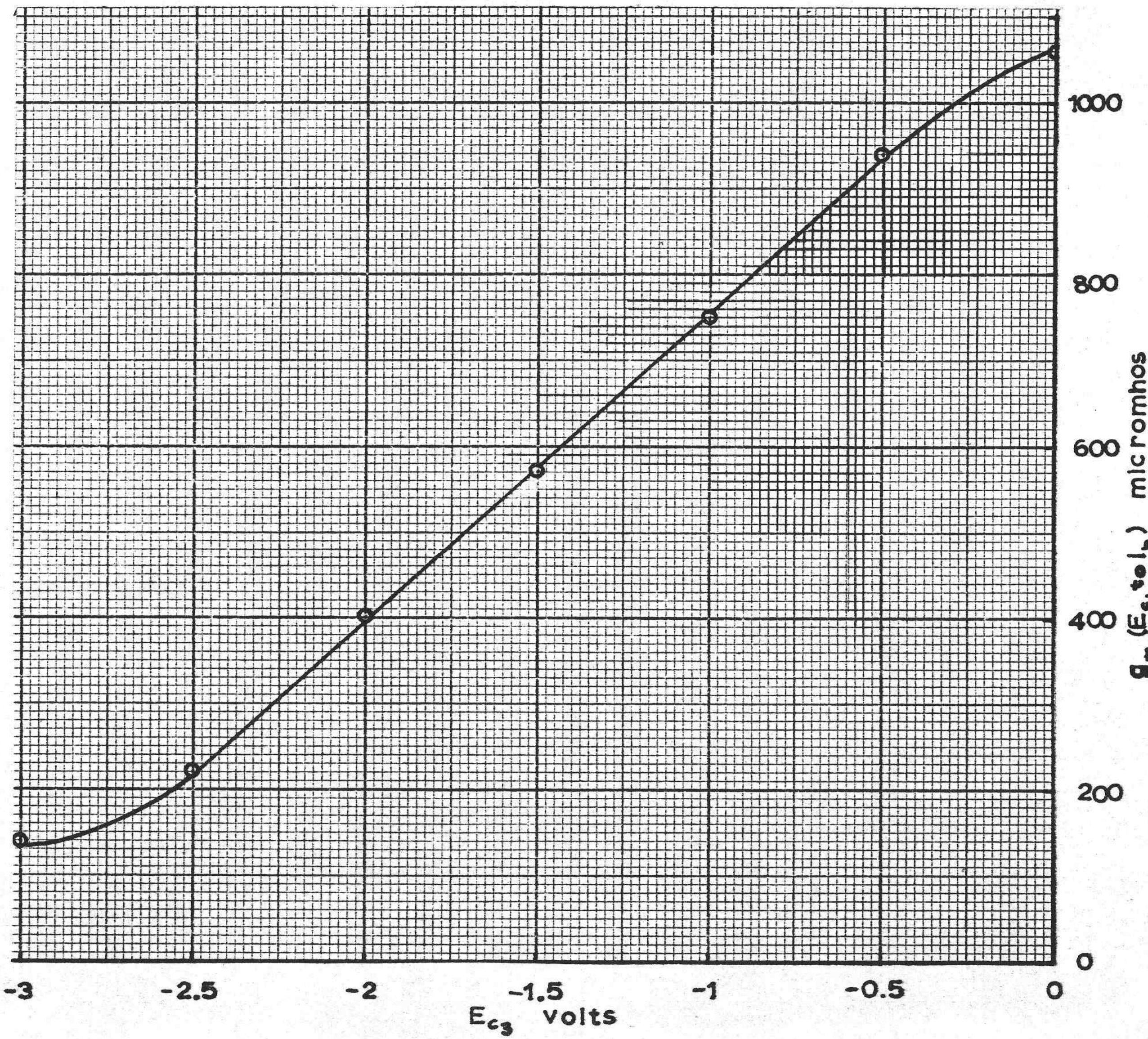
L_o and C_o form the oscillator tank circuit

we have:

$$F = \frac{1}{2\pi\sqrt{L_oC_o + L_oG_oRC(1 + \frac{E_{cc}}{E_o} + \frac{e_s}{E_o})}}$$

Then the quiecent frequency is obviously the frequency with no signal voltage:

$$F_q = \frac{1}{2\pi\sqrt{L_oC_o + L_oG_oRC(1 + \frac{E_{cc}}{E_o})}}$$



Mutual Conductance Curve of 6BE6

Figure 7.

The ratio of these is:

$$\frac{F}{F_q} = \frac{1}{\sqrt{1+\frac{L_oG_oRCe_s}{L_oC_o+L_oG_oRC(E_o+E_{cc})}}}$$

This can be expanded by the binomial theorem, and since the total frequency shift is small compared to the carrier frequency, it is permissible to retain only the first term which is:

$$\frac{F}{F_q} = 1 - \frac{1}{2}\,\frac{e_s}{\frac{C_oE_o}{G_oRC} + E_{cc} + E_o}$$

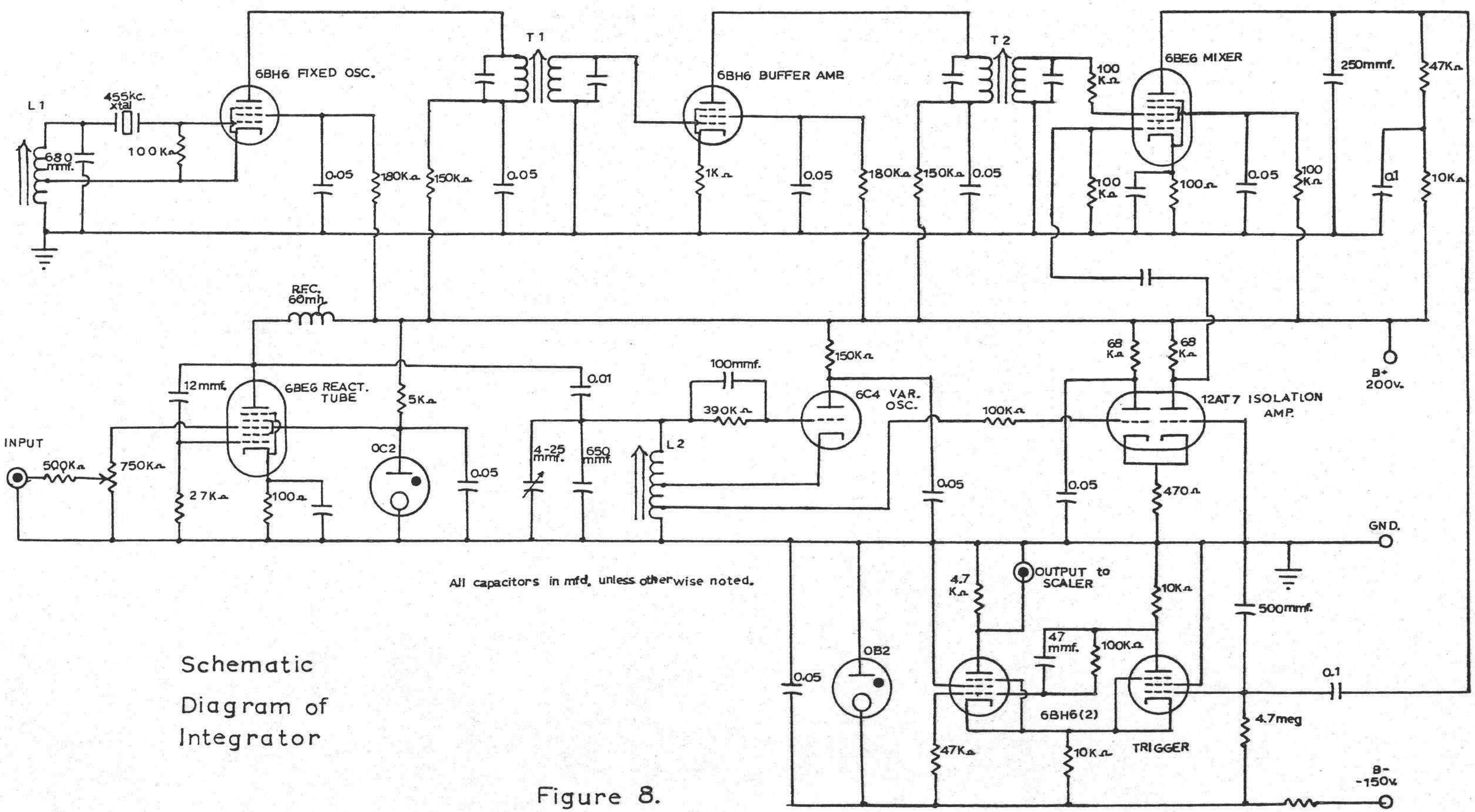
or

$$F = F_q - \frac{F_q\,e_s}{2\left(\frac{C_oE_o}{G_oRC} + E_{cc} + E_o\right)}$$

Thus the frequency of oscillation is a direct linear function of the signal voltage as long as it is within the linear portion of the mutual conductance curve, and further provided the total deviation is small compared to the carrier frequency.

The actual circuit used meets these requirements. As shown in Figure 8, a 6BE6 pentagrid mixer tube is employed with the third grid as a control element and the first grid for voltage injection from the phase shifting network. Grids 2 and 4 and connected directly to a gaseous voltage regulator tube to maintain a constant quiescent mutual conductance and thereby contribute to oscillator stability.

Grid five is a suppressor and is internally connected to the cathode. The plate circuit includes a radio frequency choke connected to B + and coupling capacitors to the phase shift network and oscillator resonant circuit. Bias Voltage is provided by a resistor in the cathode circuit. It can be seen from the equations that the injected capacitance, and therefore the oscillator frequency, is dependent upon various conditions in the reactance tube. Thus it was found that even though the oscillator was quite stable in itself, it was necessary to provide very constant heater and B + voltages to the reactance tube in order to maintain an over all stability of $\pm$ 10 cycles per second for the system.



Schematic Diagram of Integrator

Figure 8.

While it is well known that an electron coupled oscillator is stable in frequency even with supply voltage variations, it is less well known that an unloaded triode oscillator, which is exactly the same as the operational section of an electron coupled oscillator can be just as stable. A further consideration, the practical difficulty of obtaining a pure sine wave voltage from the untuned plate of an electron coupled oscillator, dictates the use of a triode oscillator circuit in this application.

As can be seen from Figure 8, a 6C4 triode is used as the variable oscillator in a Hartley circuit. The specially wound tank coil was designed to provide an output with minimum loading and to give close coupling between grid and plate circuits so that the oscillator might have a low harmonic content in its output wave form.

The coil is wound on a special ceramic form for good temperature stability with a compressed powdered iron core which can be moved for coarse frequency control. It consists of 110 turns of No. 30 formvar insulated magnet wire close wound, with the output tap five turns from the grounded end and the cathode tap for the Hartley oscillator an additional fifty turns from the output tap. The coil was resonated to about 455 KC with a 620 mmf silver mica capacitor and the injected capacitance of the reactance modulator, which appears as a variable capacitance across the tank circuit. A 4-25 mmf ceramic mounted air trimmer capacitor is used as a zero beat control and is mounted on the front panel.

For good oscillator stability the ratio of C to L should be as large as possible. This is in direct opposition to the condition that in order to get a large modulation index from a rectance modulator we must have a small tank capacitance so that the small linear range of the reactance may be used to best advantage. In light of these conflicting criteria a compromise was reached which provided a linear region of 0-1000 c. p. s. with the reactance modulator control. The approximate value of permissible capacitance change in the reactance modulator was known, and this along with the requirement that the ratio of carrier frequency to its fractional change was to be 455 to 1 allowed the total tank capacitance to be calculated; in turn the proper inductance was found which would resonate the tank circuit at about 455 K. C.

Bias voltage in the Hartley circuit is obtained from the grid leak resistor and grid capacitor selected to provide a time constant for best stability under power supply variations. All parts in the oscillator were chosen for frequency stability: capacitors of the NPO ceramic type or silver mica; ceramic coil forms and tube sockets; all parts rigidly mounted on ceramic terminal strips to reduce microphonism and temperature drift.

THE ISOLATION AMPLIFIER

Since the variable oscillator is sensitive to loading changes, the output is tapped off low on the tank coil and fed to a non-resonant resistance-coupled isolation amplifier using a dual triode 12AT7 as shown in Figure 8. The first of the two triode sections is a cathode follower with the plate effectively grounded by a capacitor. Its high input impedance offers a minimum load to the oscillator and provides a low output impedance needed to drive the other triode which operates as a grounded grid amplifier. The signal coupling is provided by a resistor common to both cathodes. The grounded grid of the second triode is an effective shield, preventing any feed-through of the other oscillator frequency from the mixer stage and thus reducing the chance for the oscillators to lock in frequency synchronism.

With a fairly high value of cathode resistance and high plate resistances the total tube current is maintained at a low value for reliable and cool operation. The high cathode resistance also provides effective inter-stage coupling and introduces negative feedback to stabilize the amplifier.

This system provides isolation and amplification for the variable oscillator output without the use of frequency-sensitive traps, and thus assures uniform gain and loading over the entire frequency range of the variable oscillator. It is possible to use this circuit since the very lightly loaded triode oscillator produces a sinusoidal wave form which is relatively free from harmonics.

THE CRYSTAL OSCILLATOR

In order to provide a stable fixed frequency for heterodyne operation a crystal controlled oscillator is needed. The resonant frequency of the quartz crystal is 455 K.C. $\pm$ 0.005% and is relatively constant over a wide temperature range because of the stability of the crystal and also because of the hermetically sealed case.

The circuit as shown in Figure 8 is a Hartley oscillator with an electron coupled 6BH6 pentode. Here the screen grid is used as the oscillator anode with the plate merely an output coupling element. The crystal and a tapped coil with a silver mica capacitor connected in parallel form the oscillator tank circuit which is permeability tuned. Bias for the tube is obtained with a capacitor and grid leak. The output is coupled into the buffer amplifier from the plate of the oscillator tube through a standard 455 K.C. Ferrite shielded, permeability tuned I.F. transformer.

The oscillator is designed to operate at very low plate and screen voltages for several reasons. First, overall reliability is achieved by operating all of the components far below their maximum ratings. Second, by running these parts cool, temperature variations are kept to a minimum and any frequency drift due to temperature sensitive components is minimized. Finally, oscillator inter-action is greatly reduced, a very important item in a device of this type which operates in the zero beat frequency region.

By using low voltages and proper shielding, electrostatic and magnetic coupling of the two oscillators is avoided. These show no tendency to lock together or synchronize as is quite common in oscillators which operate at nearly the same frequency.

THE BUFFER AMPLIFIER

For complete isolation of the crystal oscillator from the heterodyne mixer and for a more nearly pure sinusoidal output, a buffer amplifier is used between the oscillator plate and the mixer grid. It is well known that the output of an electron coupled oscillator is rich in harmonic content. These harmonics, if allowed to reach the mixer grid, will beat with possible harmonics from the variable oscillator, thereby giving spurious outputs which would cause multiple triggering of the output circuit. For example, suppose the 455.000 K.C. output to be rich in second harmonics of 910.000 K.C. and the variable oscillator to be operating at 455.010 K.C. along with its second harmonic of 910.020 K.C. The resulting beat frequency at the output of the heterodyne mixer would not only contain the 10 C.p.s. fundamental but also a 20 C.p.s. component resulting from the non-linear mixing of the two second harmonics, along with all the other sum and difference components of the four frequencies. Sum frequencies are of no consequence here since all but the audio frequency components are filtered out. If the 20 C.p.s. component were of sufficient amplitude it would cause the output device to count more than the actual number of beat cycles and therefore introduce a large error.

Thus the buffer amplifier, along with the tuned elements in its plate and grid circuits, not only provides effective isolation of the oscillator and mixer but also reduces the harmonic content of the oscillator wave form to such a low value than an almost pure 455.000 K.C. is available at the mixer grid.

For isolation purposes and reliability the plate and screen voltages of the 6BH6 pentode buffer amplifier in Figure 8 were set at low values so that all parts would run cool, and shielding would be no problem. Bias voltage is provided by an un-bypassed cathode resistor which introduces a small amount of negative feedback to help keep the stage gain down and maintain stability.

The input and output transformers are special 455 K. C. intermediate frequency transformers. This type has the cup shaped ferrite tuning slug which fits over the outside of the coil, in contrast to the older type inside which the tuning slug moves coaxially. These new transformers were used because the ferrite, together with the aluminum can, provides effective magnetic and electrostatic shielding.

THE HETERODYNE MIXER

The two oscillator frequencies are non-linearly mixed in a type 6BH6 pentagrid converter tube to produce the resultant low frequency output. Referring to Figure 8 it can be seen that the isolation amplifier of the variable frequency oscillator is connected through a suitable capacitor to the No. 1 grid of the converter tube while the buffer amplifier output transformer is connected to the No. 3 grid. Grids No. 2 and 4 are screen grids, capacitor bypassed to ground and connected to B+. Grid No. 5 is a suppressor internally connected to the cathode. Bias voltage is supplied by a bypassed cathode resistor. The plate circuit consists of a plate resistor connected to B+ and a 250 mmf capacitor to ground, forming a low pass filter which eliminates all but the beat note of the two mixed signals from the output. The latter is used to trigger a monostable multivibrator (schmitt trigger).

From Terman (5, p. 448) the principle of operation of the mixer is as follows:

Let:

$$e_1 = E_{cc} + E_1 \sin w_1 t$$

Be the instantaneous voltage on the first grid of the mixer tube and:

$$e_2 = E_{cc} + E_2 \sin w_2 t$$

Be the instantaneous voltage on the third grid of the mixer tube, where:

Ecc is the bias voltage supplied by the cathode resistor, $E_1 \sin w_1 t$ is the signal voltage supplied by the variable oscillator isolation amplifier, and $E_2 \sin w_2 t$ is the signal voltage supplied by the fixed frequency buffer amplifier.

Then, under the assumption that the variable oscillator voltage on grid No. 1 is small compared to the fixed oscillator voltage on grid No. 3, the grid No. 1 trans-conductance may be considered a function only of the grid No. 3 voltage. Thus, the grid No. 1 plate transconductance gm_1 may be considered as varying periodically as a linear function of $E_2 \sin w_2 +$ such that: $g_{m_1} = g_{m_q} + KE_2 \sin \omega_2 t$

Where gm_q is the quiescent value of gm_1 and KE is the amplitude of variation of gm_1.

Then the variational part of the instantaneous plate current useful in this case of heterodyne may be expressed as:

$$i = g_{m_1} E_1 \sin \omega_1 t$$

$$i = (g_{m_q} + KE_2 \sin \omega_2 t)\, E_1 \sin \omega_1 t$$

$$i = g_{m_q} E_1 \sin \omega_1 t + KE_1 E_2 \sin \omega_1 t \, \sin \omega_2 t$$

$$i = g_{m_q} E \, \sin \omega \, t + \frac{K}{2} E_1 E_2 \cos(\omega_1 - \omega_2)t - \frac{K}{2} E_1 E_2 \cos(\omega_1 + \omega_2)t$$

All high frequency components are removed by a low pass filter, since the impedance to ground is very high at low frequencies and very low at high frequencies. The output voltage can be expressed as:

$$e_o = \sum_{n=1} i_n Z_n \quad \text{for n frequency components}$$

Which due to the action of the filter becomes:

$$e_o = \frac{ZKE_1 E2}{2} \cos(\omega_1 - \omega_2)t$$

as long as Z is resistive at the applied frequency.

THE OUTPUT MULTIVIBRATOR

A monostable multivibrator, sometimes called a univibrator or Schmitt trigger circuit, is used in the output to provide sharp even pulses. These are needed to properly trigger the scaler which counts the total number of cycles over the time of the integration. Most scalers are designed to count pulses from nuclear detecting devices and therefore would not operate properly with the sine wave output of the heterodyne frequency mixer, especially at very low beat frequencies. In the present case a sine wave is used to control the univibrator which gives output pulses, the amplitude and duration of which are independent of the frequency and input wave form.

The univibrator is similar to the usual multivibrator except that one of the tubes is biased sufficiently to prevent free running. This gives the device a stable state in which one tube is fully conducting while the other tube is cut off. The circuit will remain in this state until the input wave form raises the grid voltage of the cut-off tube up to the point of conduction. At this point the cut-off tube will conduct fully while the conducting tube will be cut off by the action of the coupling circuit. After a length of time determined only by the circuit constants the tubes will return to the original stable state and remain so until the input wave form again raises the cut-off grid voltage to the point of conduction. Thus for each cycle of input voltage the univibrator is switched from its stable to unstable state and back again, producing an output pulse of amplitude and duration independent of the input wave form.

As shown in Figure 8 the multivibrator comprises two 6BH6 pentodes in a grounded positive configuration. This method of using a negative supply to the cathodes and grids helps in maintaining reliable triggering from a sine wave source. Other than this the circuit is the common one used in most applications which require a sharp trigger pulse of constant amplitude.

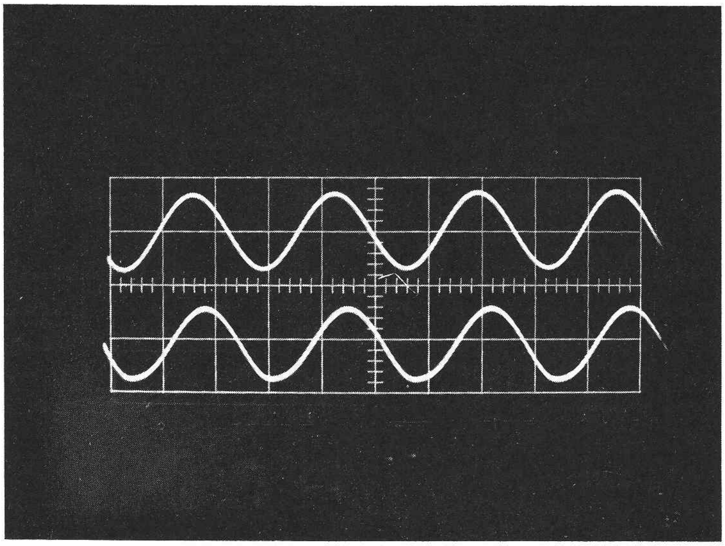


Figure 9a.
Top: Crystal Oscillator
Bottom: Variable Oscillator

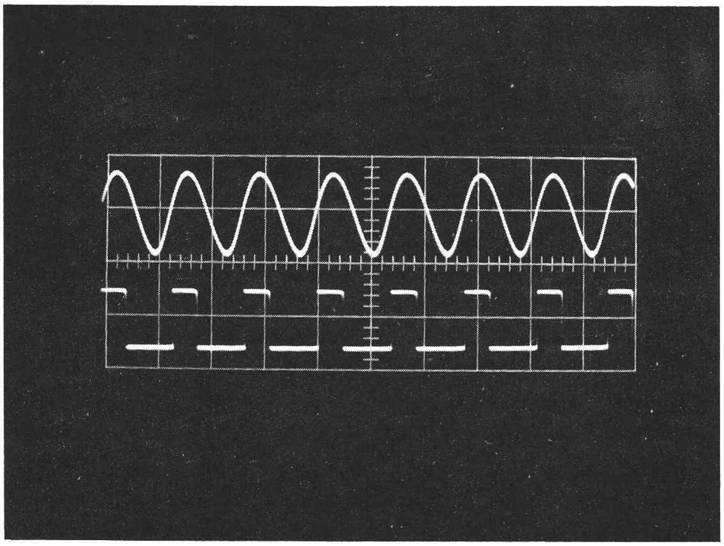


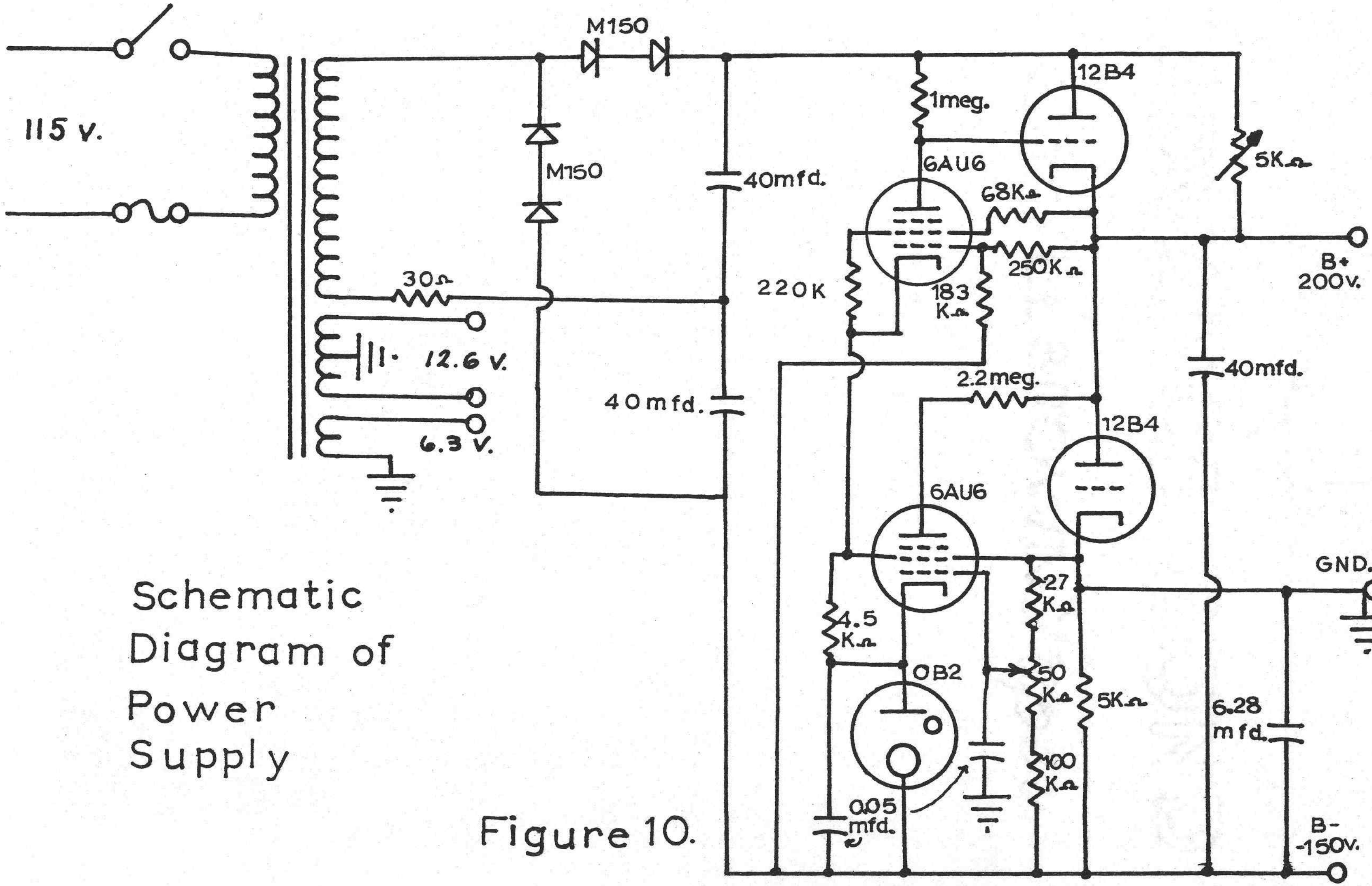
Figure 9b.
Top: Heterodyne Mixer Output
Bottom: Multivibrator Output

THE POWER SUPPLY

For a stable system the power supply variations due to loading and line voltage changes must be held to a minimum. Both oscillators are designed to be stable over a limited range of plate potential but the reactance modulator which controls the final heterodyne frequency is directly dependent on power supply stability to maintain a certain injected reactance and consequent frequency. In order to achieve this stability two electronically regulated supplies are used.

The power supply, which is mounted on the chassis of a companion unit in order to eliminate hum and temperature drift due to the heat dissipated by the high power parts, consists of a transformer, voltage doubler, and electronic regulators.

The transformer is a large type with a power rating far in excess of the current needs of the integrator and its associated equipment mounted on the same rack. The voltage requirements of the equipment made it necessary to increase the rectified 200 volts r. m. s. output of the transformer by use of a voltage doubler. The doubler is a full wave unit using silicon diode rectifiers and two 40 m. f. d. capacitors to provide about 550 volts to the series electronically regulated supplies which use 12B4's as regulators and 6AU6's as control tubes. Since the analog computor section and the output multivibrator require a negative supply and neither side of the voltage doubler is grounded, an arbitrary ground is maintained with an electronic regulator at 150 volts above B-. Since many parts in an oscillator are grounded to the chassis it is necessary to use another voltage regulator to supply 200 volts positive referred to the arbitrary ground of the



Schematic Diagram of Power Supply

Figure 10.

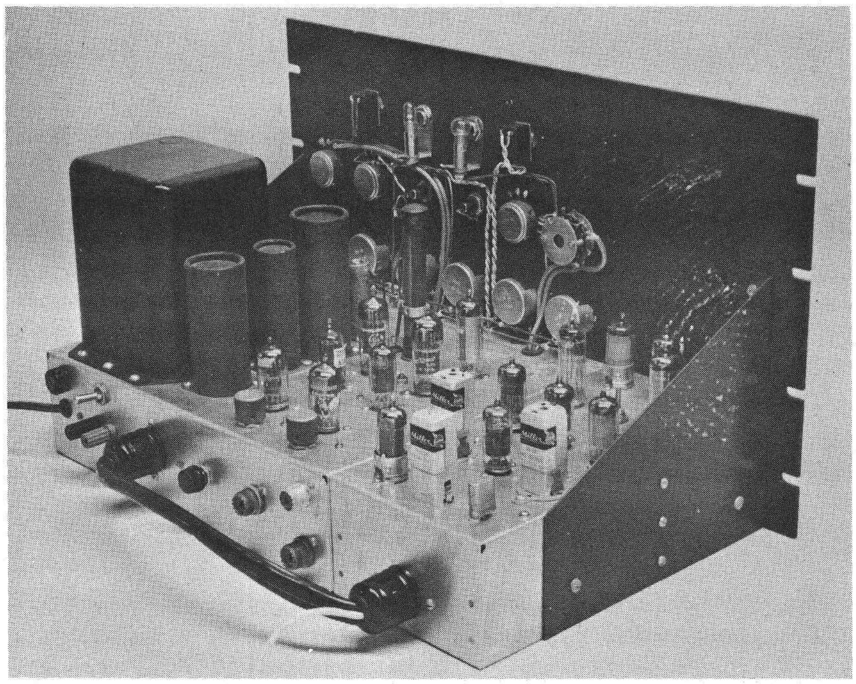


Figure 11a.
Integrator, Power Supply and Transfer Function Device

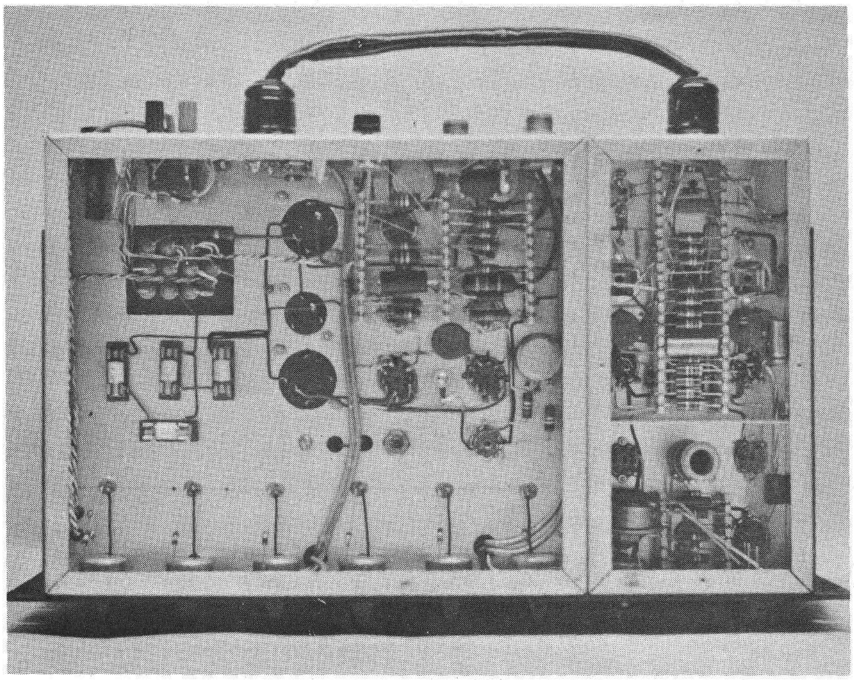


Figure 11b. Bottom View

chassis. The screen voltage of the reactance modulator is further regulated with a 75 volt gas regulator tube.

The output multivibrator supply is maintained at 105 volts by a voltage regulator tube. This provides a low power supply impedance, helping to maintain a flat top on the output pulses and resulting in a constant pulse height regardless of output frequency. Power for the multivibrator and its regulator are obtained from the negative 150 volt supply.

With proper power supply regulation, it was found that the integrator was very stable. The drift from zero beat was no more than 10 c.p.s. over a period of two hours after a suitable warm up period. The low power supply impedance, along with other precautions, reduced oscillator coupling to the point where no locking was observed even down to a beat note of a fraction of a cycle per second.

The integrator was tested to determine whether the stability and linearity were within the limits imposed by the overall accuracy requirements of the system. Preliminary tests showed a rather severe frequency drift which was traced to variations in screen grid voltage on the reactance modulator. This was corrected by using a separate gaseous voltage regulator tube in the screen supply, as stated in the power supply section. This correction improved the stability enough to warrant final testing. Using a Nucleonics Corporation of America model RCG3 scaler on the output with a Precision Scientific Company interval timer it was observed that after at least a half hour warm-up, the system maintained a stability of less than 10 cycles per second drift from zero beat over a two hour period.

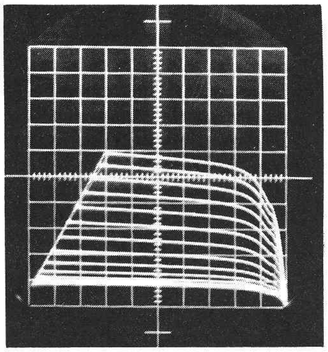
A further test of short term stability was accomplished by connecting an audio amplifier and speaker to the output so that the output pulses were heard as "pops" in the speaker. This test confirmed that the output remained within at least 10 cycles per second of zero beat over a two hour period as judged by ear. To find average drift the scaler was connected to the output and the total number of cycles from zero beat recorded over a thirty minute period. This total number of cycles was divided by the time in seconds and showed the average drift to be slightly over four cycles per second.

The fact that there is a small amount of drift from zero beat demonstrates that there is no tendency for the two oscillators to lock together and remain at the same frequency. The two oscillator wave forms as presented to the mixer are shown in Figure 9a. These were photographed with a polariod land camera on a Beattie oscilloscope mount using a Tektronix type 585 oscilloscope with a type CA Dual-trace plug-in preamplifier. In order to photograph the wave forms it was necessary to temporarily introduce a small coupling capacitor between the grids of the oscillators to synchronize them for a stable trace on the screen. As can be seen from Figure 9a, the variable oscillator has a small amount of harmonic content, while the fixed oscillator wave form is very close to a true sine wave.

Figure 9b shows the relationship of the output of the heterodyne mixer to that of the monostable multivibrator. It also shows that the mixer output is low in harmonic content.

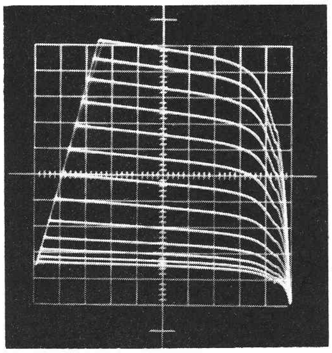
The linearity of the reactance tube is of prime importance in this application. As was shown in the section on the reactance tube, a linear frequency to voltage relationship depends upon the fact that the mutual conductance of the first grid in a 6BE6 is linearly proportional to the potential applied to the third grid. In order to justify this assumption a representative 6BE6 was tested on a Tektronix type 575 tube characteristics curve tracer and photographs were taken for seven different grid No. 3 potentials. The photographs are shown in composite in Figure 12 while the resulting mutual conductance for each different grid potential is plotted in Figure 7. These different figures show that the variation in Grid No. 1 mutual conductance is proportional to a change in Grid No. 3 voltage over the operating range.

Figure 13 shows the circuit and instruments used in checking the overall linearity of the integrator. A Heathkit P. S. 4 power supply was used to vary the input voltage and its value read on a Hewlett-



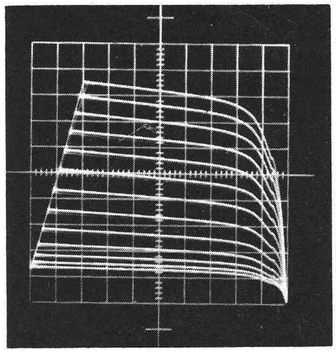
$E_{g_3} = 0v.$

Vert. = 1 ma/div.



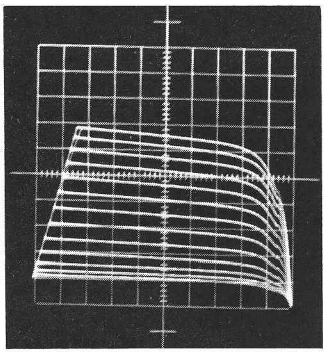
$E_{g_3} = 0.5v.$

Vert. = 0.5ma/div.



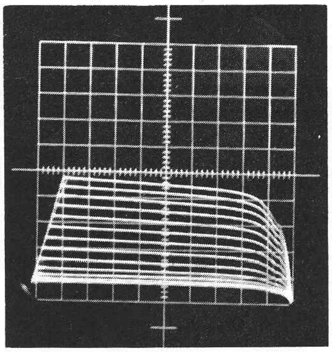
$E_{g_3} = 1.0v.$

Vert. = 0.5ma/div.



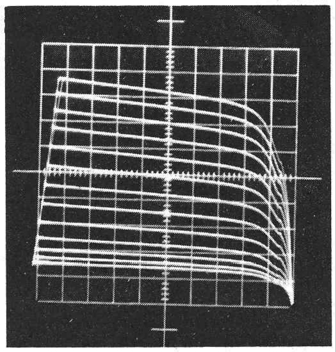
$E_{g_3} = 1.5v.$

Vert. = 0.5 ma/div.



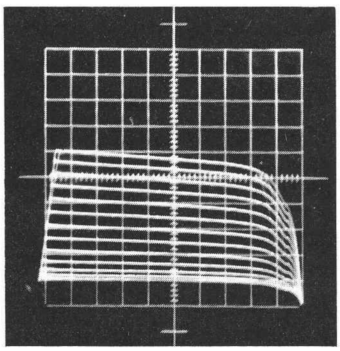
$E_{g_3} = 2.0v$

Vert. = 0.5ma/div.



$E_{g_3} = 2.5v.$

Vert. = 0.2ma/div.



$E_{g_3} = 3.0v.$

Vert. = 0.2ma/div.

For all curves:

$E_{g_1} = 0.5$ v/step

$E_{pk} = 200$ v.

Load = 10kΩ

Hor. = 20 v/div

$E_{g_{2\&4}} = 75$ v.

Htr. = 6.3 v

Figure 12

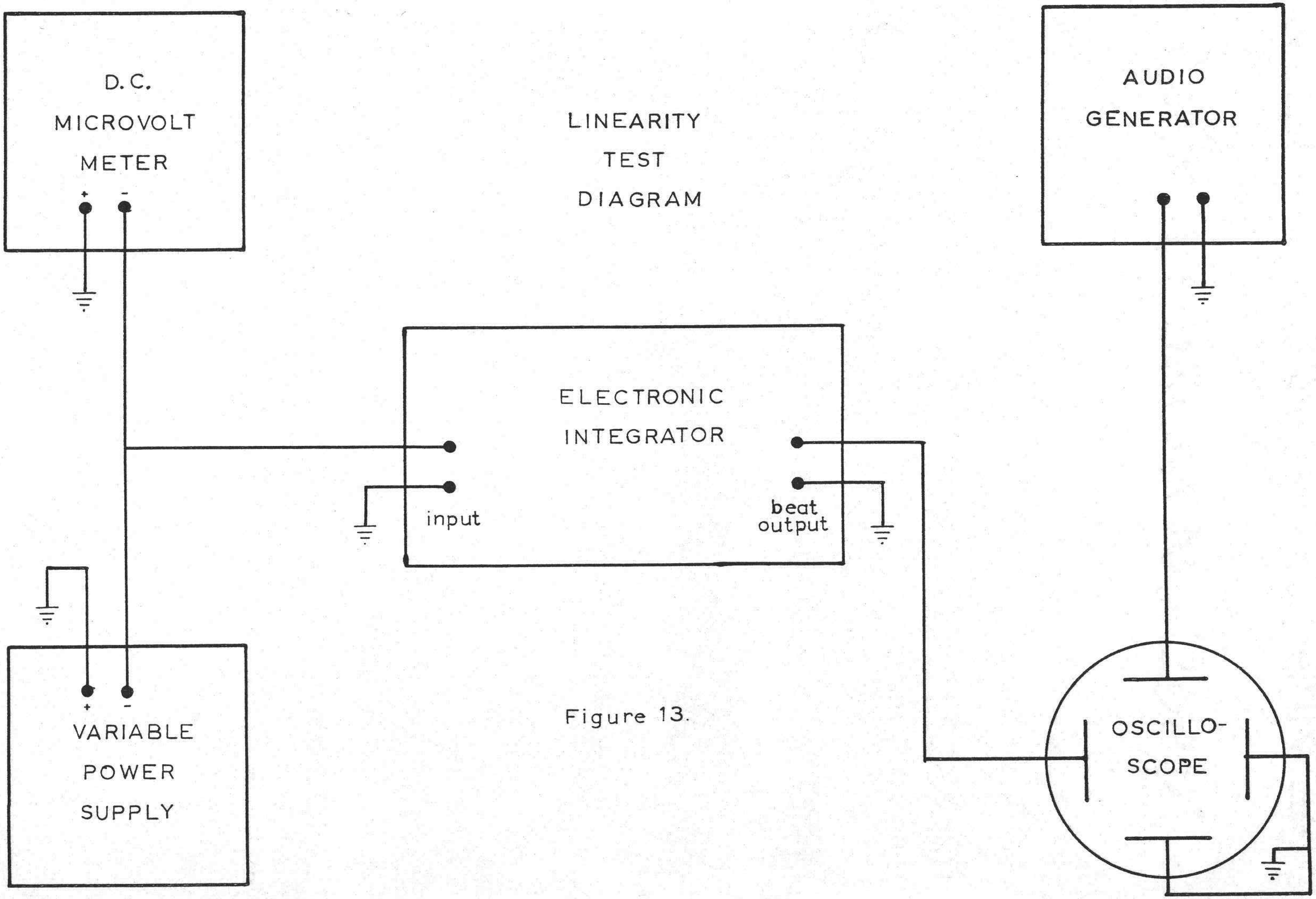
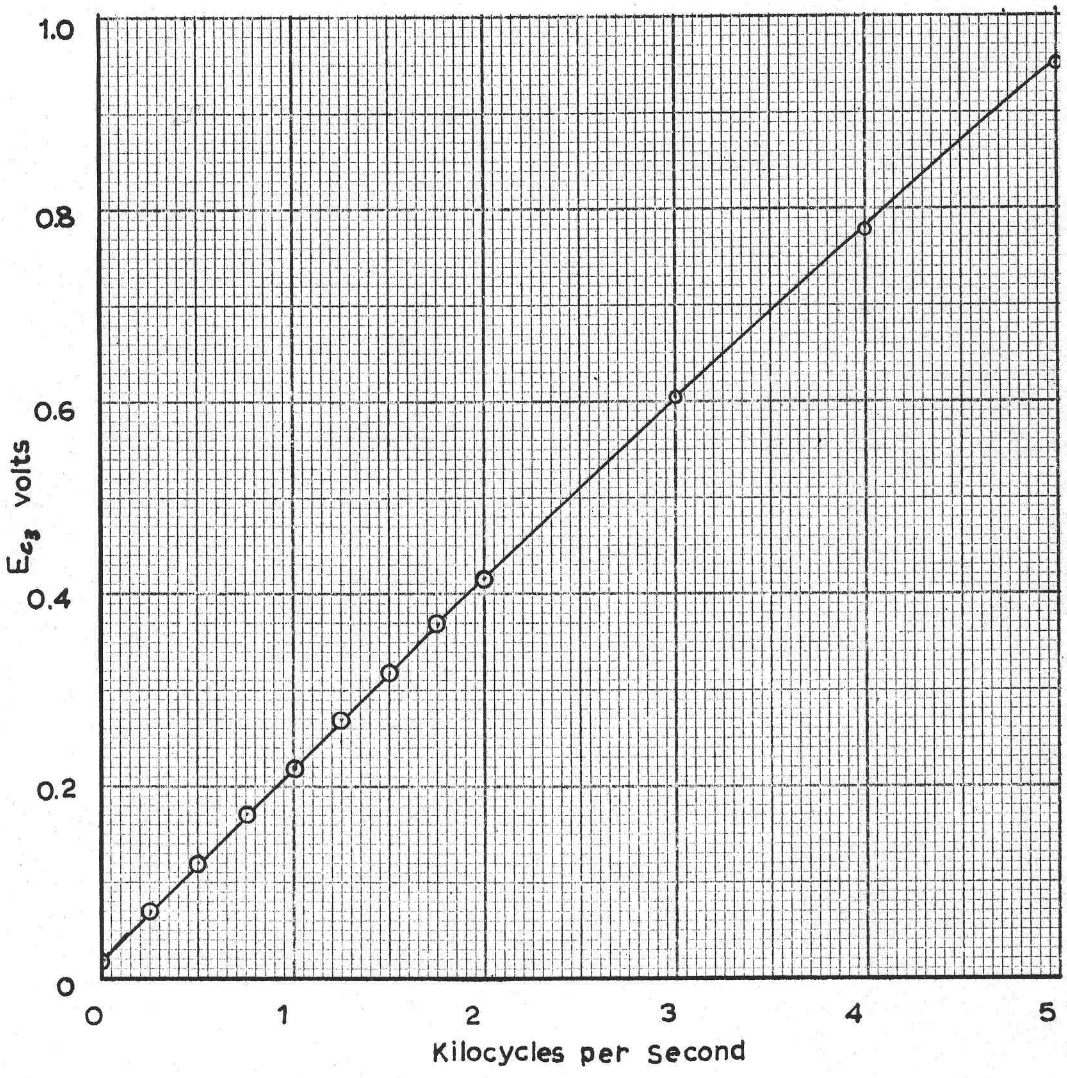

D.C.
MICROVOLT
METER
+
-
LINEARITY
TEST
DIAGRAM
AUDIO
GENERATOR
ELECTRONIC
INTEGRATOR
input
beat
output
+
-
VARIABLE
POWER
SUPPLY
OSCILLO-
SCOPE

Figure 13.



LINEARITY TEST RESULTS

Figure 14.

Packard type 425A micro voltmeter. The beat output of the heterodyne mixer was connected to the horizontal input terminals of a Tektronix type 515" oscilloscope. The vertical signal was provided by a Heathkit Model AG9A audio generator. The resulting Lissajous pattern and the indicated frequency on the audio generator were used to determine the beat frequency; the grid input voltage for that frequency was read directly from the mirror scaled voltmeter. Figure 14 shows the result of the linearity check, which was well within the 1% accuracy of the meter up to 1000 cycles per second and only deviated 5% at 5000 cycles.

CALIBRATION AND OPERATION

The electronic integrator is very simple to calibrate and operate as it requires only two simple controls. The first of these is a variable capacitor in the resonant circuit of the variable oscillator. It is used to adjust the oscillator frequency when zero voltage is applied to the input grid. The only action needed to adjust this control is to observe the output scaler until it reads zero cycles per second or as close to this as is practical. The zero control covers a wide range so that any potential up to about 0.1 volt on the grid of the reactance tube may be used as the basis for zero beat. In this way the versatility of the integrator is increased since both negative and positive signals, superimposed on a bias voltage, can be integrated. This will take care of the case in spectral analysis where the base line of the curve to be integrated is not exactly zero volts, due to the dark current of the photomultiplier tube.

The second control is the calibration control and is set once for each range of the recorder and left in that position until the recorder range is changed. The calibration control is merely a potentiometer connected to the grid of the reactance modulator, so as to provide the proper grid voltage from a given input signal. Its function is to control the number of pulses per second at the output for a given input voltage. To establish the calibration setting, all that is needed is the scaler and some timing device such as a stop watch. A predetermined voltage is applied to the input of the integrator by setting the pen on the

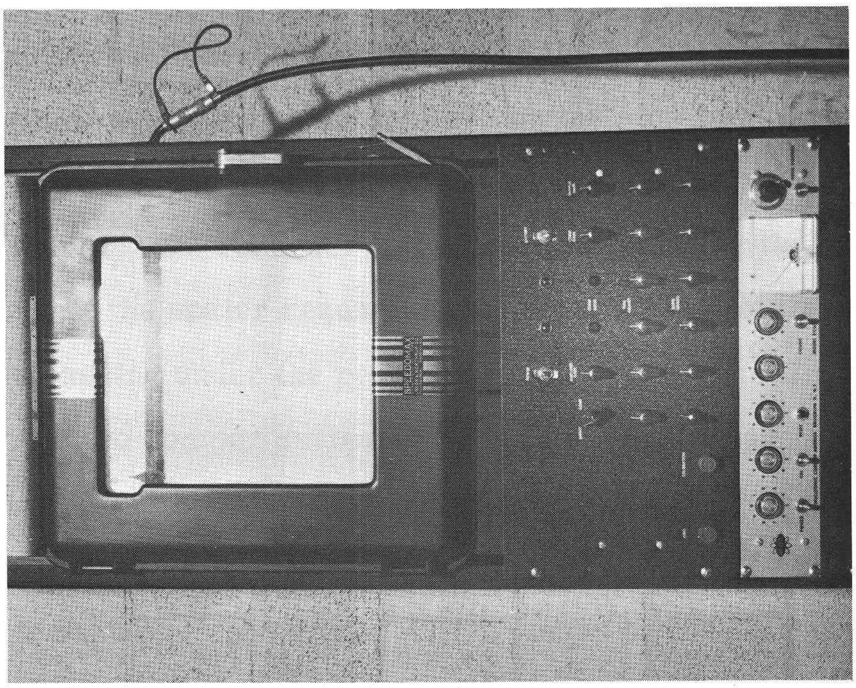


Figure 15a. Rack Mount

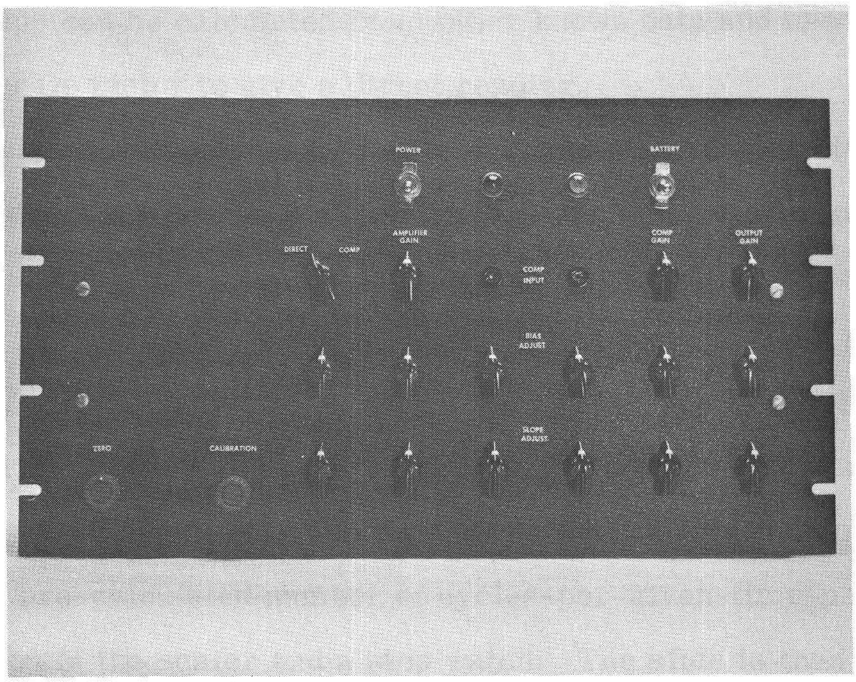


Figure 15b. Front Panel Controls

recorder to some convenient calibration point. Then the calibration control on the integrator is set to provide the proper number of cycles over a given time period as determined by the stop watch. In a simple case the output is adjusted to count the number of squares under a curve by bringing the recorder to full scale and setting the calibration control so that the scaler reads a convenient multiple of the number of squares passing under the pen of the recorder. Thus if one square passes under the pen per second and the full scale voltage is set to give a 1000 c. p. s. count on the scaler, then the number of cycles indicated on the scaler is divided by 10, since there are 100 squares in a full scale reading. The scaler may be made to read out the energy in a spectral line by setting the calibration control in such a way that the relationship between the energy and area under the curve is taken into consideration in the adjustment of the calibration control. This relationship can be calculated from other known data and used along with the scale factor to give a direct reading.

In operation, the slide is placed in the microphotometer and the recorder and transfer function synthesizer are adjusted according to established practice and the instructions provided with these devices. After the two instruments are adjusted a black card is set in front of the photomultiplier so as to provide a zero reference to set the zero beat control. Then the card is removed and the slide adjusted to provide full scale deflection on the recorder. The calibration control is set to the pre-calculated number of cycles-per-given-time-period by observation of the scaler and a stop watch. The slide is then positioned until the approximate edge of the spectral line is near the slit.

Then the scaler is turned on and the entire line allowed to pass in front of the slit, thus recording the luminous energy distribution on the recorder and the total energy in the line on the scaler.

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