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A low-frequency high stability frequency shift oscillator.

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BY

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A LOW-FREQUENCY HIGH STABILITY FREQUENCY SHIFT OSCILLATOR

by

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PREFACE

Because of the regulations of the Federal Communications Commission, and due to the crowded condition of that portion of the radio spectrum normally used for moderate and long range communication, great accuracy is required in adjustment of the frequency of a radio transmitter to be within the allowable frequency error, and high stability is required to maintain this frequency without the need for constant checking and readjustment. These problems have been satisfactorily met for radio telegraph and radio telephone transmitters since the intelligence to be transmitted is rarely if ever associated with the frequency determining source in the apparatus, and therefore problems of frequency stability are divorced from the problems of modulation. Much room for improvement exists however in the case of frequency shift transmitters.

The author spent eleven weeks at the Westinghouse Electric Corporation Plant in Baltimore, Maryland, working on a high stability frequency shift oscillator to be used as the frequency modulated oscillator for the transmission of radio teletype and facsimile signals in a frequency synthesis system.

It is desired to acknowledge the assistance received from the following Westinghouse engineers: M. I. Jacob, in charge of the Government Engineering Communications Section, and A. H. Veiner, with whom the author worked closely during this period.

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TABLE OF SYMBOLS AND ABBREVIATIONS

A-C	alternating current
C	capacitor or capacitance
D.C.	direct current
E	voltage or volts
f	frequency
g_m	transconductance
I	current
I_b	plate current
K	constant
kc	kilocycles per second
L	inductance or inductor
Q	"Quality factor", or approximately $\frac{\omega L}{R}$
R	resistance or resistor
RCA	Radic Corporation of America
R.F.	radio frequency
S	switch
V	vacuum tube
ω	angular frequency
X	reactance
Z	impedance
Δ	incremental or increment
%	per cent
$\pm$	plus or minus
δ	drift voltage

CHAPTER I

INTRODUCTION

1. Summary

This thesis treats briefly the general subject of frequency shift oscillators for use in radio teletype and radio facsimile transmission, but major emphasis is placed upon the design problems encountered, and their solutions, for a particular 100 kilocycle center frequency oscillator.

2. Description of a frequency shift oscillator

A frequency shift oscillator may be described as one whose frequency can be shifted rapidly in a desired predictable manner in accordance with the intelligence contained in the input modulating signal. Due to the nature of the modulating signals which may be employed it is further necessary that the oscillator frequency can be shifted to any frequency within its range and remain at the shifted frequency for an indefinite time determined by the modulating signal. Therefore the system must respond to alternating current signals as low as zero cycles. In this respect it differs from the customary oscillators employed for conventional frequency modulation.

Radio teletype and radio facsimile transmission is usually accomplished by frequency shift modulation.

CHAPTER II

PERFORMANCE CRITERIA

1. Requirements imposed by the specification

A recent U. S. Navy specification (7) for a low frequency, frequency shift transmitter contained, in part, the following data concerning performance required:

(1) Frequency stability is to be within ± 10 cycles at 100 kilocycles under all conditions of voltage variation, vibration, humidity and temperature.

(a) Voltage variation: ± 10% in voltage and ± 5% in frequency. Nominal supply is 220 volts at 60 cycles.

(b) Relative humidity: Up to 95%.

(c) Temperature: 0 degrees Centigrade to + 65 degrees Centigrade.

(2) For facsimile, linearity of frequency shift with respect to input signal voltage is to be within 10%. Ten percent is defined by requiring that the slope of the tangent to any portion of the curve of frequency shift versus input signal voltage shall be no greater than 10% different from the slope of the line connecting the end points of the curve.

(3) Frequency shift for teletype signals shall be continuously variable from 0 to ± 500 cycles from the center, or "carrier" frequency. The amount of shift shall be independent of the keying speed from 0 to 240 dot cycles.

(4) Maximum frequency shift when employing facsimile signals shall be continuously variable from ± 50 cycles to ± 500 cycles. The amount of shift shall be constant within 25% of the maximum shift at zero cycles for any facsimile signal from 0 to 2000 cycles, of maximum amplitude. Maximum amplitude is fixed for any particular transmission, but can be of any value between 10 and 20 volts.

2. Target requirements for the laboratory model.

No specific requirements to be achieved in the laboratory model were established except that all of the requirements of paragraph 1 should be met and bettered. Most difficulty was expected to be encountered in attempting to meet and better the frequency stability requirements. This proved to be correct. A rough target stability was stated as 1 or 2 cycles at room ambient temperature and humidity.

3. Performance achieved previously.

Davey and Matte (4) state:

> A frequency shift exciter......with the crystal oscillator and 200 kc frequency shift oscillator located in the temperature controlled oven, usually has a frequency stability such that the mean R.F. carrier frequency may be held to within ± 50 cycles ...over ordinary periods of operation at any one frequency.

Buff (2) states:

> Present equipments ...... maintain a stability of ± 6 cycles on the 1 to 6 megacycle range over an operating period of 6 hours or more.
>
> A-C line fluctuations of 10% reflect no more than a ± 6 cycle change on the output frequency.

Previous frequency shift oscillators built by the Westinghouse Electric Corporation were not stable enough to meet the requirements of paragraph 1.

CHAPTER III

EARLIER METHODS OF FREQUENCY SHIFT

1. Frequency shift by reactance tubes.

Reactance tubes have the object of injecting reactances into associated networks. If the associated network is an ordinary tube oscillator whose frequency is not stabilized, the injected reactance will change the frequency of oscillation. Hund (5) gives analyses and descriptions of a number of reactance modulators. Repetition of his work here would serve no purpose, however, one point is worth noting, i.e., all of the reactance circuits described or mentioned are high impedance reactance tubes connected in parallel with the high impedance of the tank circuit of the oscillator. A block diagram which would represent any of the oscillators described is shown in Fig. 1.

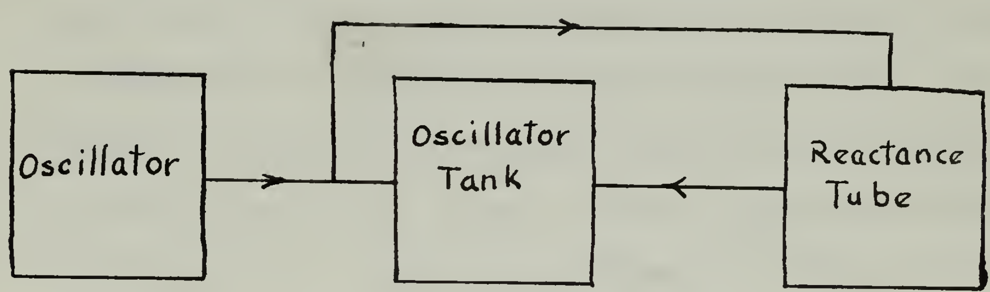
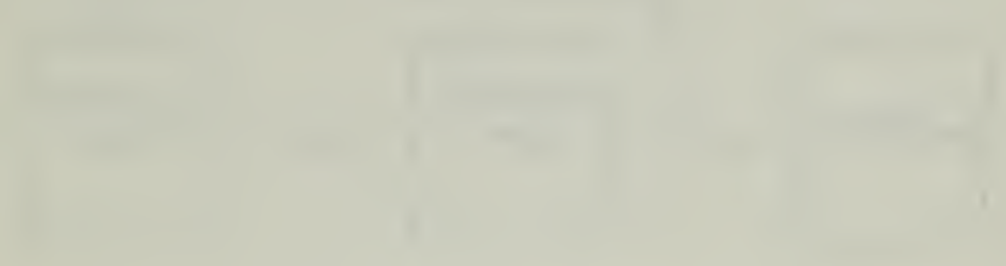


Fig. 1 - Block diagram of a reactance modulated oscillator.

2. Frequency shift by methods other than reactance tubes.

(1) Saturable reactor.

A block diagram of a system utilizing a saturable reactor is shown in Fig. 2.



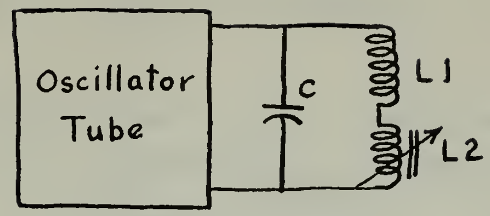


Fig. 2 - Block diagram of a frequency shift oscillator employing a saturable core reactor.

$$f = \frac{1}{\omega \sqrt{(L_1 + L_2)\, C}}$$

Since L_2, the saturable reactor, is a variable reactance it is obvious that f, the resonant frequency, can be shifted by this method (8).

(2) Switching between two independent frequency sources.

An early method of frequency shift involved switching between two independent frequency sources separated in frequency by the desired shift. In such a method the frequency transitions occur at random times with respect to the frequency sources and result in random phase discontinuities of random values. This results in a relatively large bandwidth causing severe interference to adjacent channels. Due to this inherent disadvantage, this method is no longer used.

(3) Balanced modulator method.

Fig. 3 is a block diagram illustrating the principles involved in this method.

f1 = deviation desired

f2 = carrier frequency

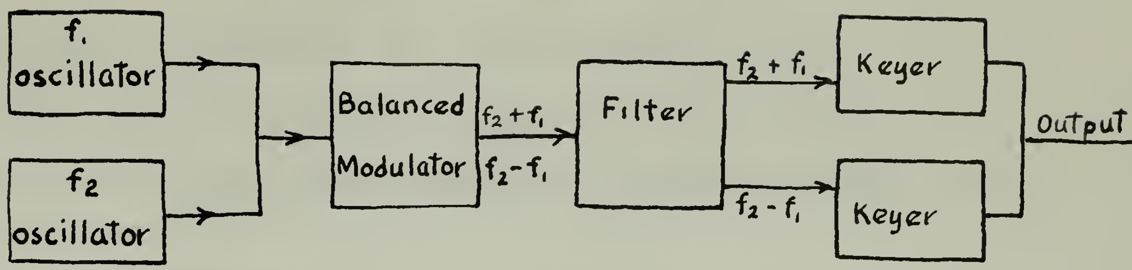


Fig. 3 - Block diagram of a balanced modulator frequency shift keyer.

This method is not suitable for facsimile transmission unless **f1** is varied by the facsimile signal and the output taken through only one of the keyer tubes. This method of frequency shift suffers from the same disadvantage as (2) above.

(4) Direct variation of the effective inductance or capacitance of the frequency determining circuit.

Frequency shift keying of radio telegraph transmitters was used in the early days of radio when arc transmitters were common. Since the arc could not be keyed on and off fast enough to permit reasonable transmitting speeds, the frequency was shifted in accordance with the Morse characters used. Ordinarily this was accomplished by short circuiting part of the tuning coil and shifting the frequency out of audible range of the receiving operator. This method is not used for frequency shift keying for teletype purposes but is an interesting historical note.

CHAPTER IV

METHOD OF FREQUENCY SHIFT EMPLOYED

1. Reactance modulation at a low impedance level.

Fig. 4 is a simplified diagram of a new type of frequency shift oscillator invented by Musk and Healey (6).

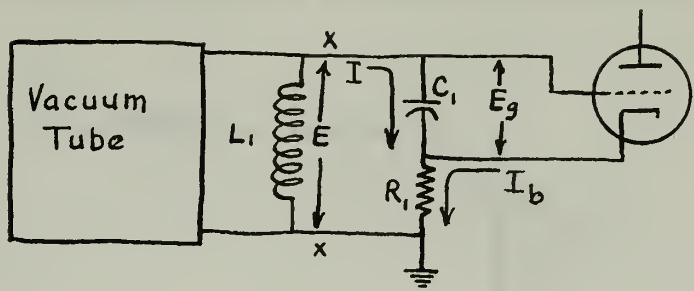


Fig. 4 - Frequency shift modulation at low impedance level.

Looking to the right at points **xx** we can write:

$$E_{xx} = I Z_1 + I Z_2 + I_b Z_2 \qquad \text{where} \begin{cases} Z_1 = \frac{1}{j\omega C_1}, \quad Z_2 = R_1 \\ E_g = I Z_1 \\ I_b = g_m E_g = g_m I Z_1 \end{cases}$$

$$= I(Z_1 + Z_2 + g_m Z_1 Z_2)$$

$$Z_{xx} = \frac{E_{xx}}{I} = Z_1 + Z_2 + g_m Z_1 Z_2$$

Considering the reactance of C_1 as constant over a relatively small frequency range, we see that the net impedance is the sum of two fixed impedances, X_{C_1} and R_1, and a variable impedance $g_m X_{C_1} R_1$. Since the net impedance is variable because of tube V_1, the resonant frequency of L_1, and the apparent capacity appearing across points **xx** is variable depending upon g_m. Thus it is obvious that frequency shift can be accomplished by this circuit.

2. Circuit analysis

A Clapp oscillator(3), (see Fig. 7), was chosen by Westinghouse engineers as the oscillator desired, based upon experience obtained prior to the author's tour of duty at the plant. The resonant circuit determining the frequency of oscillation is represented schematically by Fig. 5.

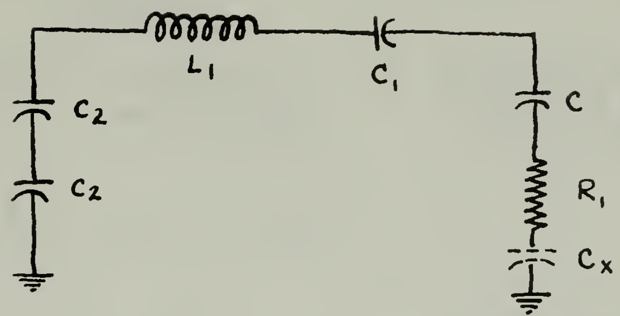


Fig. 5 - Schematic diagram of frequency determining circuit, of Clapp oscillator.

C_x = fictitious capacitor injected by reactance modulator.

f_1 = frequency of oscillation without reactance modulator.

f_2 = frequency of oscillation with reactance modulator.

$$C_3 = \frac{1}{\frac{1}{C_2} + \frac{1}{C_2} + \frac{1}{C_1} + \frac{1}{C}}$$

$$f_1 = \frac{1}{\omega\sqrt{L_1 C_3}}$$

Q of L_1 is high, i.e., R_1 is small

Now looking at the third term on the right hand side of equation IV - 1, *i.e.,* $g_m Z_1 Z_2$ we see that it is the term responsible for the variable injected impedance which has been labled as C_X. Now defining $Z_1 = \frac{1}{j\omega C}$, $Z_2 = R_1$ the following equations can be written

$$Z_{C_x} = \frac{1}{j\omega C_x} = g_m Z_1 Z_2 = g_m \frac{1}{j\omega C} R_1$$

$$C_x = \frac{C}{g_m R_1}$$

$$\therefore f_2 = \frac{1}{\omega\sqrt{L_1 \frac{C_3 C_x}{C_3 + C_x}}} = \frac{1}{\omega\sqrt{L_1 C_3}\sqrt{\frac{C_x}{C_3 + C_x}}} = f_1 \frac{1}{\sqrt{\frac{C_x}{C_3 + C_x}}}$$

$$f_2 = f_1 \left(1 + \frac{C_3}{C_x}\right)^{\frac{1}{2}}$$

Since in practice C_3 will be much less than C_X , the above expression can be simplified by expanding the binomial as a series and then approximating by neglecting all terms of power greater than one.

The binomial series expansion of $(1 + x)^{\frac{1}{2}}$ is given below.

$$(1+x)^{\frac{1}{2}} = 1 + \frac{1}{2}x - \frac{1}{8}x^2 + \frac{1}{16}x^3 - \cdots\cdots$$

Using only the first 2 terms of the expansion we have,

$$f_2 \doteq f_1\left(1 + \frac{C_3}{2C_x}\right)$$

Now substituting for C_X , we have

$$f_2 \doteq f_1\left(1 + \frac{C_3}{2C} g_m R_1\right) \qquad \text{Let} \quad \frac{C_3 R_1}{2C} = K$$

$$\therefore f_2 \doteq f_1(1 + K g_m)$$

$g_{m1} = g_m$ of the reactance tube with no keying signal.

$g_{m2} = g_m$ of reactance tube with keying signal.

$f_3 =$ new frequency of oscillation due to changing

$$\therefore \quad f_2 = f_1(1 + K g_{m1})$$

$$\text{and} \quad f_3 = f_1(1 + K g_{m2})$$

$$\text{Frequency shift} = f_3 - f_2 = f_1 + f_1 K g_{m1} - f_1 - f_1 K g_{m2}$$

$$= f_1 K (g_{m1} - g_{m2})$$

$$= f_1 K \Delta g_m$$

Frequency shift is seen therefore to be proportional to the change in g_m of the reactance tube.

Two convenient methods of varying g_m are:

(1) Variation of control grid bias.

(2) Variation of screen grid voltage in pentodes or tetrodes.

Since frequency shift must be linear (within 10% as defined in Chapter II) the transfer characteristics, or transconductance versus control grid volts and screen grid volts, were compared for linearity. Fig. 6 is a plot of experimental results. It is seen that the curve of g_m versus control grid volts is more linear and therefore more desirable. As a consequence control grid control of g_m was chosen for the first experimental model.

Transfer Characteristics

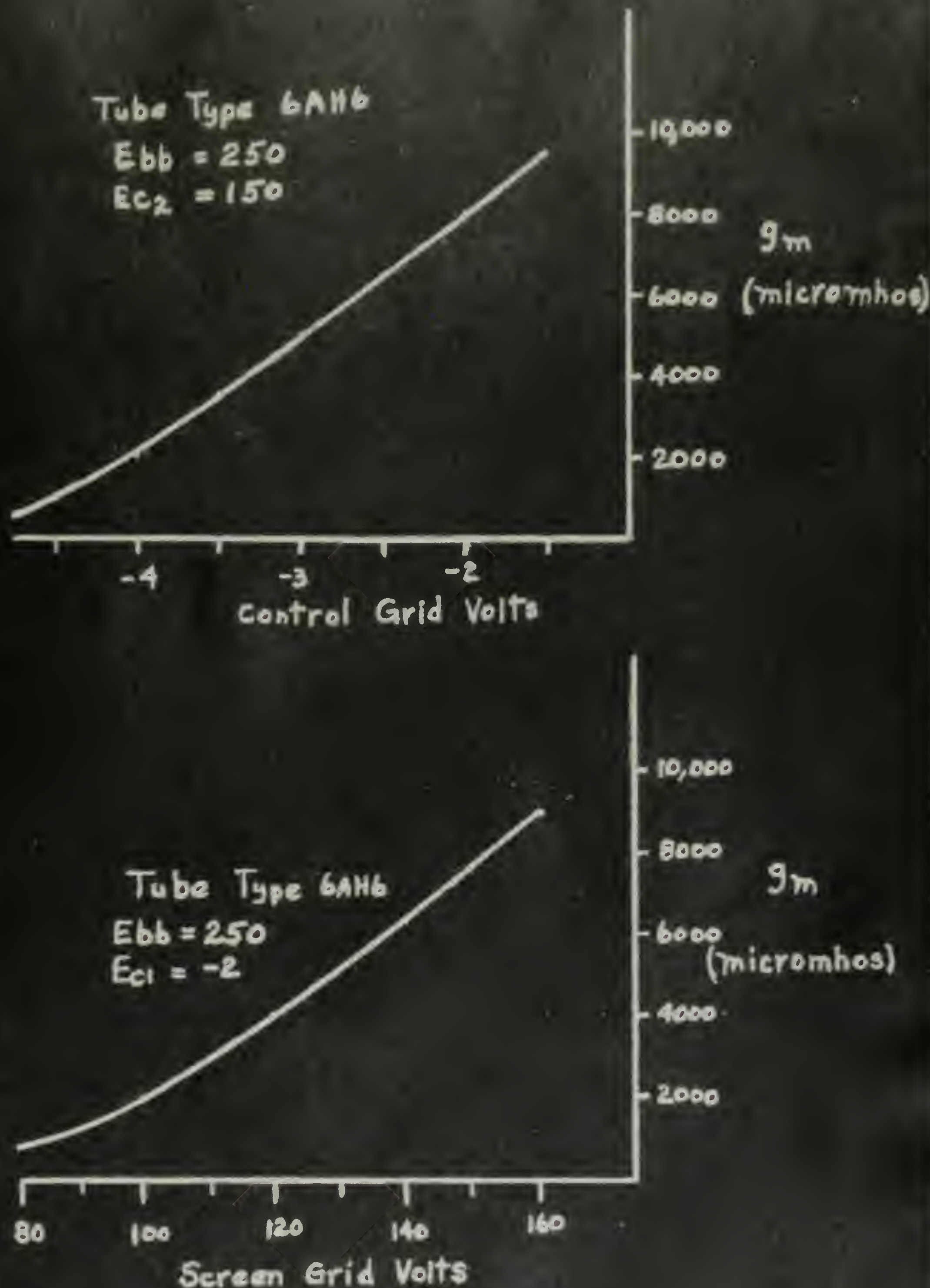


Fig. 6

CHAPTER V

LABORATORY EXPERIMENTAL WORK

1. Reactance modulators.

A simplified schematic diagram of the first oscillator tried is shown in Fig. 7. Measurements revealed that the frequency stability with respect to supply voltage variations was poor. It is well known that the g_m of a tube varies as the supply voltages are changed. Since we found in Chapter IV that frequency shift was proportioned to the change of g_m, the frequency instability encountered is predicted by theory.

As a means of minimizing the effects of supply voltage variations and consequently the undesired variations in g_m, push pull reactance tubes were employed. Fig. 8 is a simplified schematic diagram of the second circuit tested. It is seen that the g_m effective in producing frequency shift is the difference in g_m of tubes V4 and V5. Therefore any changes due to supply voltage variations should be cancelled, provided that the characteristics of V4 and V5 are identical.

In this circuit frequency stability with respect to supply voltage variations was much improved over that obtained with the oscillator shown in Fig. 7. Preliminary measurements indicated that the necessary ± 500 cycle shift was easily attainable and that frequency stability at the mark, space and carrier frequencies was of the order of magnitude required by the specifications. Observation of the wave shape appearing at the control grids of V4 and V5 revealed that the keying

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Single tube reactance modulator

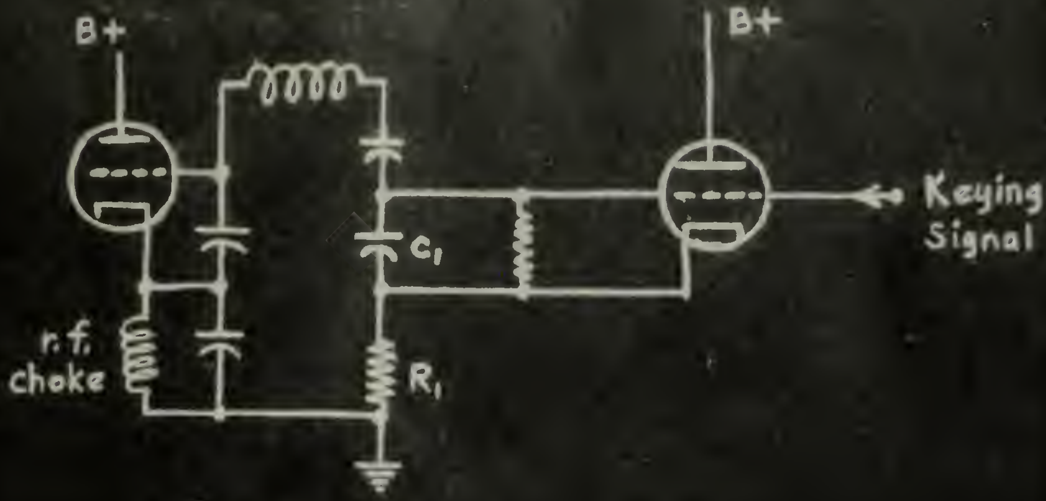

Fig. 7

Push-pull reactance modulator

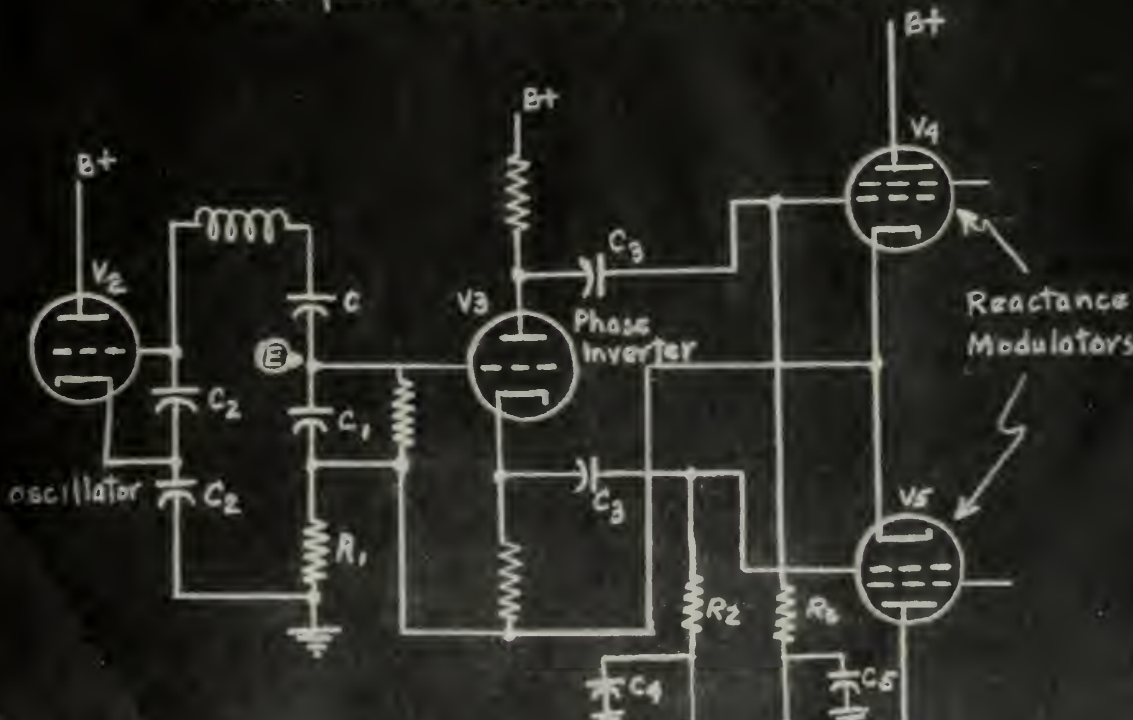

Fig. 8

speed of 240 dot cycles could not be met with acceptable wave shape due to the long time constants of C_3, R_2, C_4 and the internal impedance of the source of keying voltage for V_5, and similarly for V_4.

Since both the radio frequency signal and the keying signal must be combined at the common grid it was considered necessary to decouple the keying circuit from the R.F. input to the reactance modulators. Capacitor C_3 must not introduce appreciable phase shift, otherwise the reactive current would also contain a real component which would vary the effective Q of the frequency determining circuit while keying as well as reduce the magnitude of the frequency shift attained.

Screen grid control appears attractive since isolation of R.F. and keying inputs is inherent. Further, the keying source must be of a relatively low impedance to supply the screen current, therefore the time constant of the R.F. bypass capacitor at the screen terminal and the source impedance is short and does not impair the keying wave shape. The wiring connections to V_4 and V_5 were changed to allow control of g_m by variation of screen grid potential. There appeared to be no significant difference in frequency stability from that obtained with control grid keying. Frequency response with good wave shape was easily obtained with input signals of 240 dot cycles and also with a 2000 cycle sine wave. Fig. 9 is the curve of linearity, i.e., cycles shift from center frequency versus the signal input. This linearity is not within 10% as defined in Chapter II.

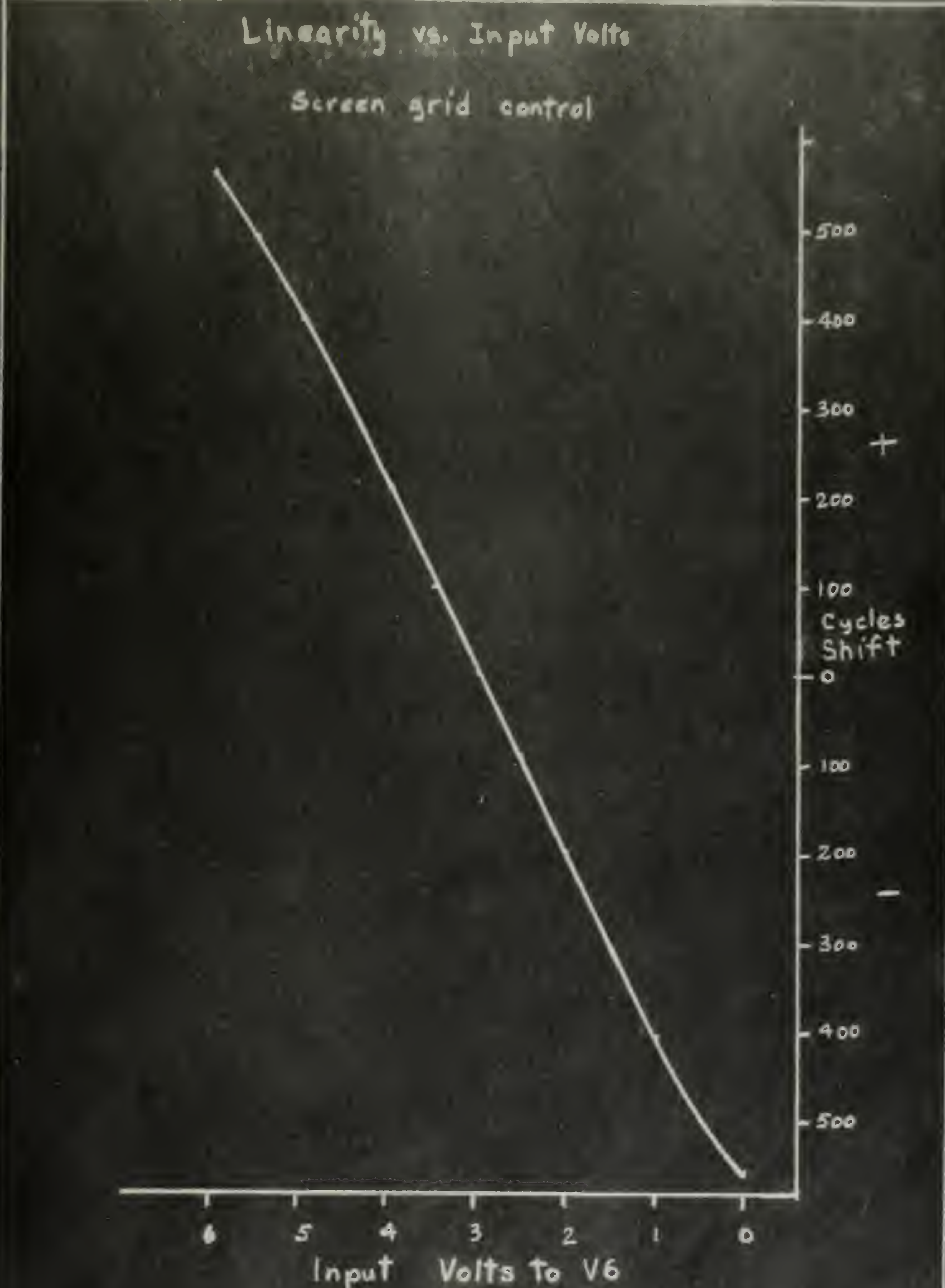
Linearity vs. Input Volts
Screen grid control
500
400
300
+
200
100
Cycles Shift
0
100
200
–
300
400
500
6
5
4
3
2
1
0
Input Volts to V6

Fig. 9

A short term frequency drift was noted to occur immediately after shifting the oscillator frequency from any one to either of the other two frequencies of major significance, i.e., mark, space, or carrier. The magnitude of the drift was about 6 to 10 cycles with most of the drift occurring in the first thirty seconds after switching, but about two minutes were required before drift terminated. All pertinent voltages were measured precisely and it was found that a slight drift in the screen voltage of the reactance tubes occurred, similar in nature to the frequency drift observed. This drift in voltage at the screen was computed to be responsible for only 10% of the drift in frequency. (See Appendix 1) The static plate current was measured and was found to drift about 1.36%. This measurement was made by using a test set up similar to that used for measuring transconductance. The voltage drop across a Tomore low temperature coefficient resistor of ten watts rating used as the plate load resistor was measured. The actual power dissipated in this resistor was 0.34 watts. The change in voltage was measured precisely by using a source of negative battery voltage equal in magnitude to the voltage appearing at the plate and connecting it through two 500,000 ohm resistors to the plate. An RCA Voltohmyst was used to measure the voltage appearing at the junction of the two resistors. The change in voltage appearing at this point is then half the voltage change occurring at the plate. By this method the lowest voltage range of the meter could be employed, i.e.,

0-3 volts, and greatest sensitivity attained. After measuring the voltage drop across the plate load resistor, the plate current was changed by changing the grid bias by a fixed increment and then returning it to the initial value. The drift in plate voltage was read on the 3 volt scale of the RCA Voltohmyst. A triode (type 12AT7) was measured to have about .075% drift in plate current under the same conditions. Drift was exponential in form with the maximum difference in current occurring at the time of switching and returning to the initial value within approximately 5 to 10 seconds.

A circuit analysis, using Laplace transforms, (see Appendix 2), was made to ascertain whether or not control grid keying could be accomplished with acceptable wave shape but utilizing a different input circuit to the reactance tubes. It was found that the decoupling network previously used was unnecessary. Also the maximum values of capacitors to keep the time constant short enough for good keying signal wave shape were determined.

Before returning to triodes for reactance modulators control grid keying of the pentodes was again used. No significant change in short term frequency drift occurred so it was concluded that the drift was not due to the keying method employed.

A twin triode, type 12AT7, with both halves in parallel to obtain the desired transconductance, was substituted for V_4 and also V_5. Short term frequency drift was reduced in magnitude but was still excessive.

The reactance modulators are direct coupled for the keying signal, therefore, methods employed for stabilizing D.C. amplifiers were investigated but no reference to the plate current drift previously mentioned was found in the literature. Artzt (1) shows a number of circuits for improving the stability of D.C. amplifiers. Two of these were employed, slightly modified, to improve the D.C. stability of the reactance modulators. Fig. 10 is a schematic diagram of the reactance modulators, V_4 and V_5 , and their D.C. driving source, V_6 , a differential amplifier. Cathode degeneration, at D.C. and keying speeds, is accomplished by R4 and R5 . Plate degeneration, at D.C. and keying speeds, is accomplished by R6 and R7 . Since R4 through R7 are by-passed for R.F. frequencies, no degeneration occurs at 100 kilocycles. Again some improvement resulted but short term drift was still greater than desired.

Since the drift in plate current was found larger for large changes of grid bias, the possibility of reducing short term frequency drift by utilizing only a small swing of the reactance tube grids was considered. We found, in Chapter IV, that frequency shift was proportional to g_m. It is to be noted however that no amplification occurred between the source of the R.F. signal and the reactance tubes. By introducing amplification between the R.F. source and the reactance tubes, the reactive current, and therefore frequency shift, would be increased in proportion to the magnitude of the amplification, provided that the reactance tubes were not over-driven at R.F.

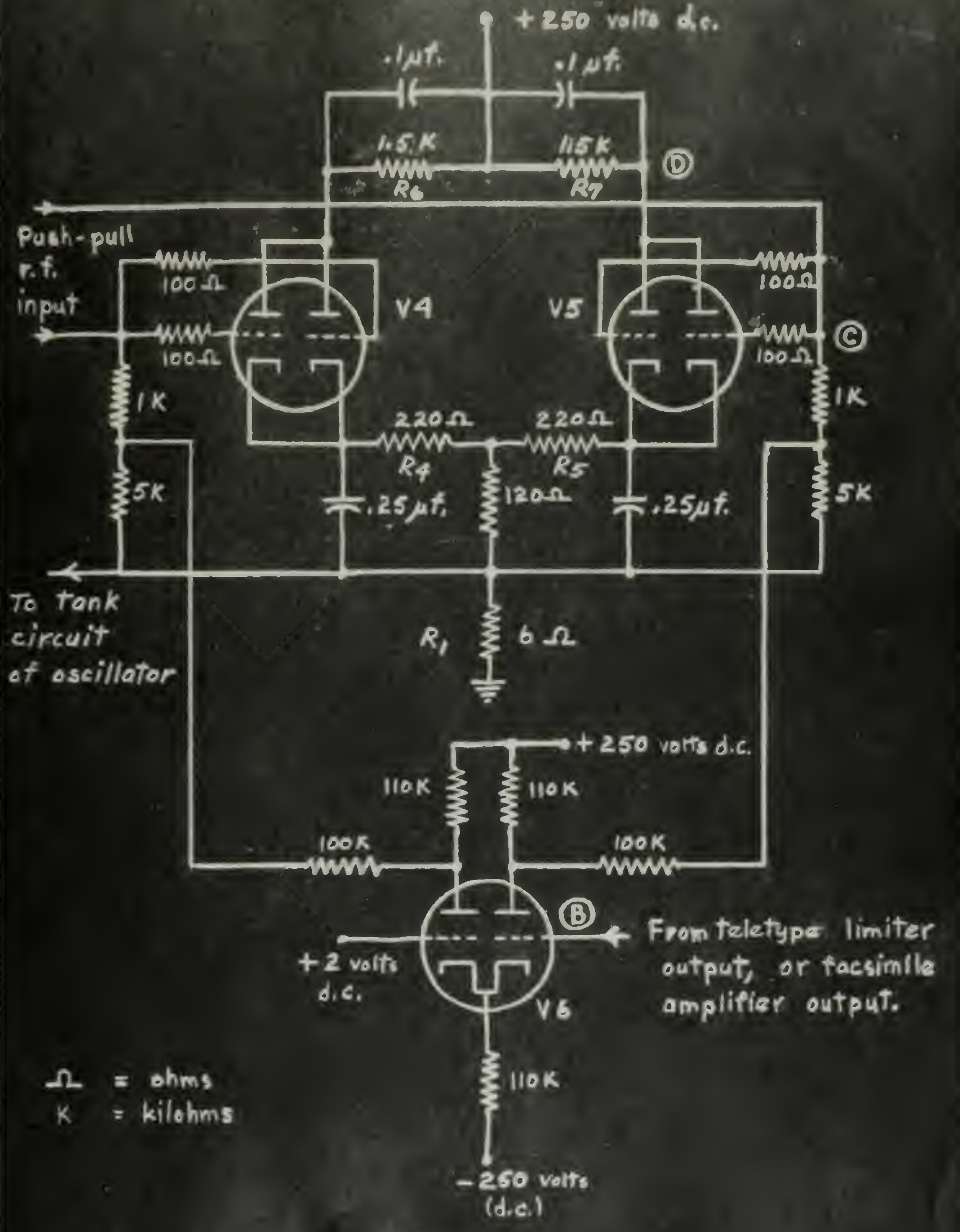

Fig. 10
Triode Reactance Modulator

An amplifier, introducing no significant phase shift, and with a voltage gain of 3 was introduced between point E and V3 of Fig. 8. The gain of the phase inverter is $\frac{1}{2}$, therefore the over-all voltage gain is 1.5.

Short term frequency drift was reduced slightly to approximately $3\frac{1}{2}$ cycles. Lack of time prevented further pursuit of this problem.

Fig. 11 is the curve of frequency shift versus input signal to the differential amplifier, V6. This curve is within the 10% linearity as defined in Chapter II.

2. Teletype limiter circuit.

A D.C. coupled clipper-limiter was designed to accept either neutral or polar keying signals. The design is straight forward and no difficulties in performance were encountered with one exception: the bottom of the rectangular wave output signal was slightly rounded. This can be flattened satisfactorily by increasing, by a factor of approximately ten, the clipping resistor, R8 , at the input of V7 as shown in Fig. 12.

3. Oscillator circuit.

A conventional Clapp oscillator (3) was employed. This circuit is well covered in the literature and will not be discussed here. In order to maintain as nearly constant g_m as possible in the oscillator, a conventional D.C. grid bias feedback loop was used. This consisted of a voltage amplifier driven by the oscillator, and a diode rectifier and filter. The D.C. output of the filter was injected on the control grid of the oscillator thereby maintaining constant bias and there-

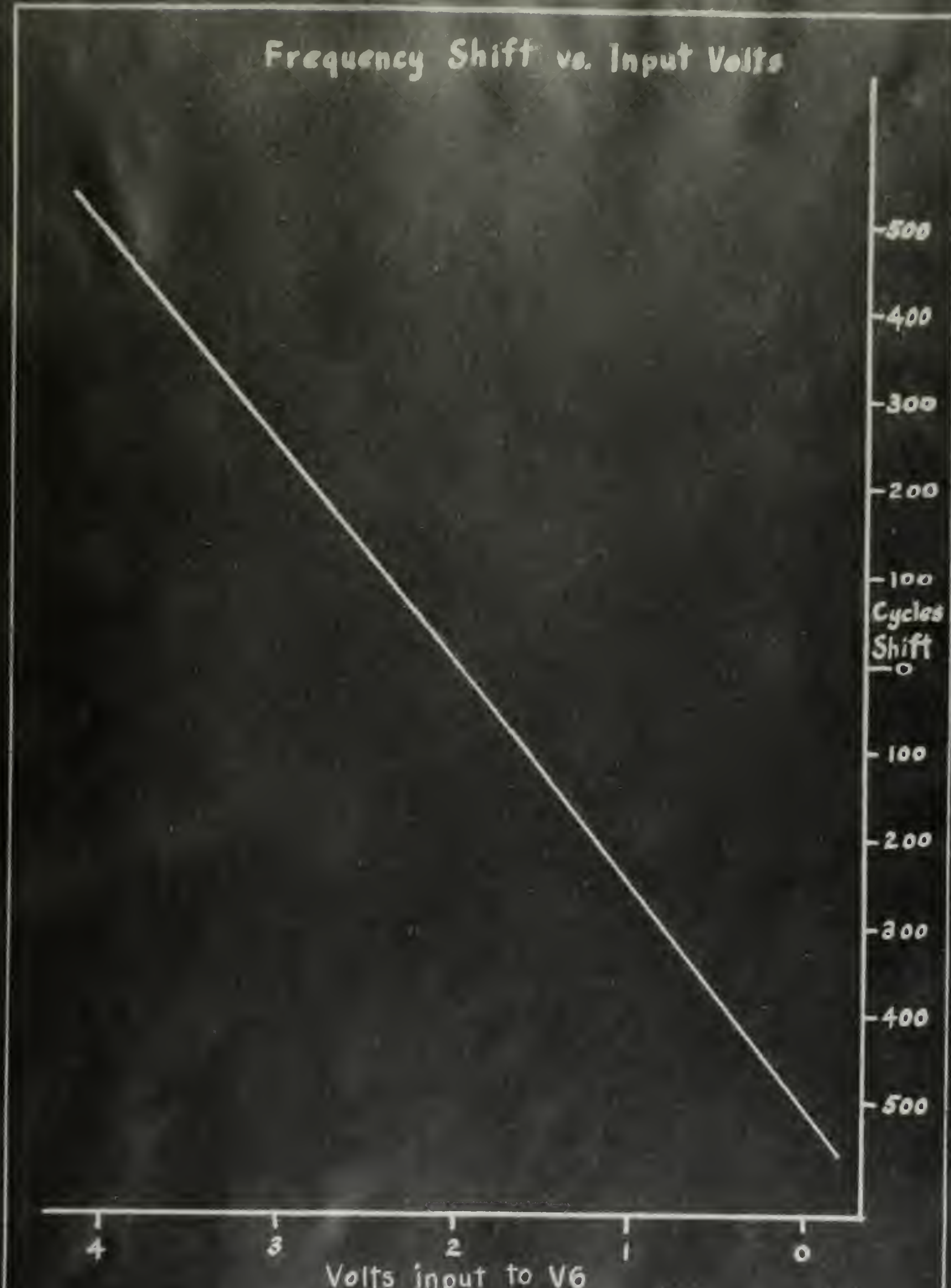
Frequency Shift vs. Input Volts
500
400
300
200
100
Cycles Shift
0
100
200
300
400
500
4
3
2
1
0
Volts input to V6

Fig. 11

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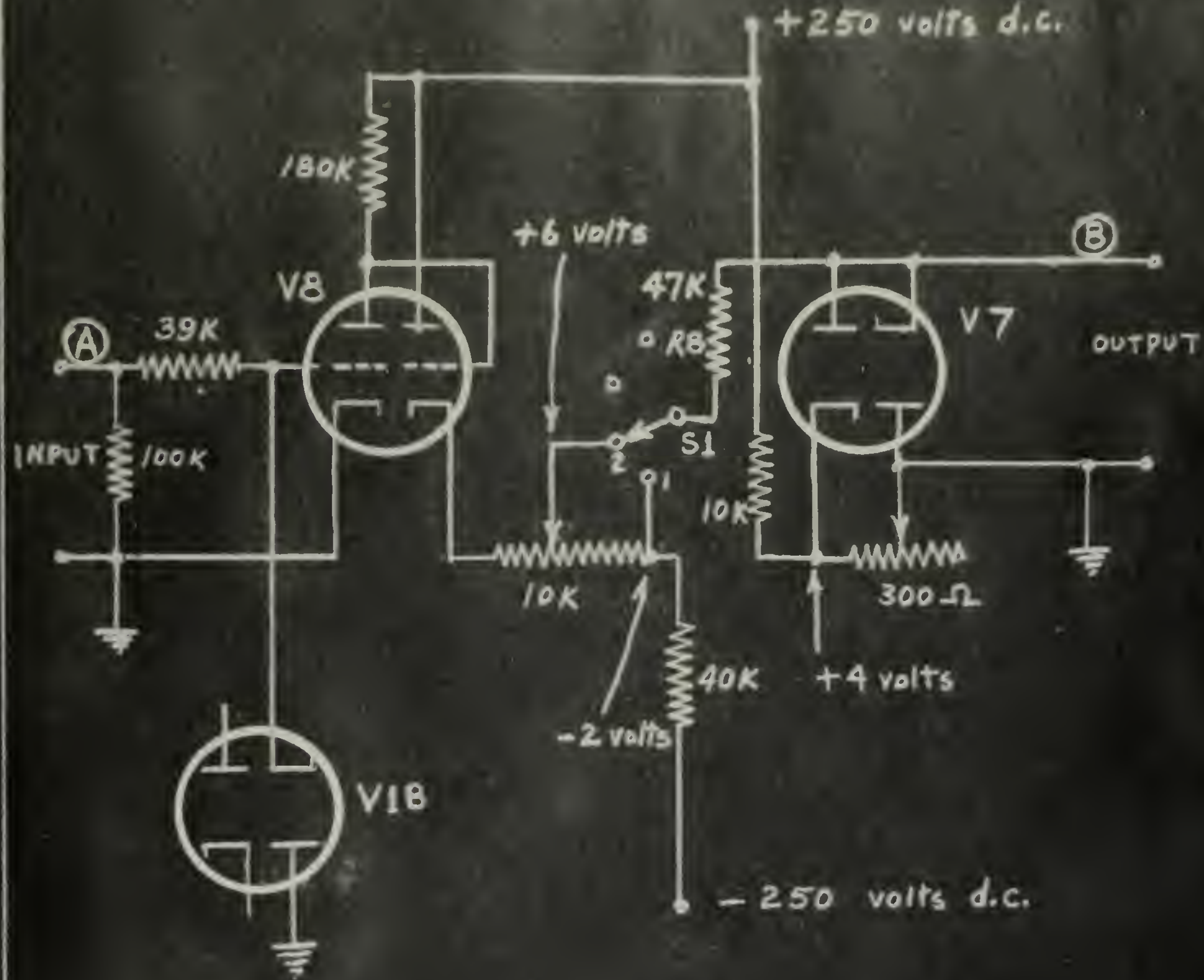
Teletype Limiter Circuit



S1 Position	Function
1	Space
2	Mark and Keying
3	Carrier
4	Facsimile

Fig. 12

fore constant g_m .

4. Facsimile input amplifier.

A direct coupled cathode follower input amplifier was designed to accomodate the range of signals supplied from a facsimile equipment. The output voltage was to be adjustable from 0 to 4 volts (to accomplish full frequency shift) to be fed into the differential amplifier at point B of Fig. 10. Due to the shortage of time this amplifier was not built, but no difficulties should be encountered since its design is conventional.

5. Overall System.

Fig. 13 is a block diagram of the over-all system, each part of which has been covered previously in this Chapter. The facsimile amplifier is not included since it was not built or tested.

SKETCH SHEET

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System Block Diagram

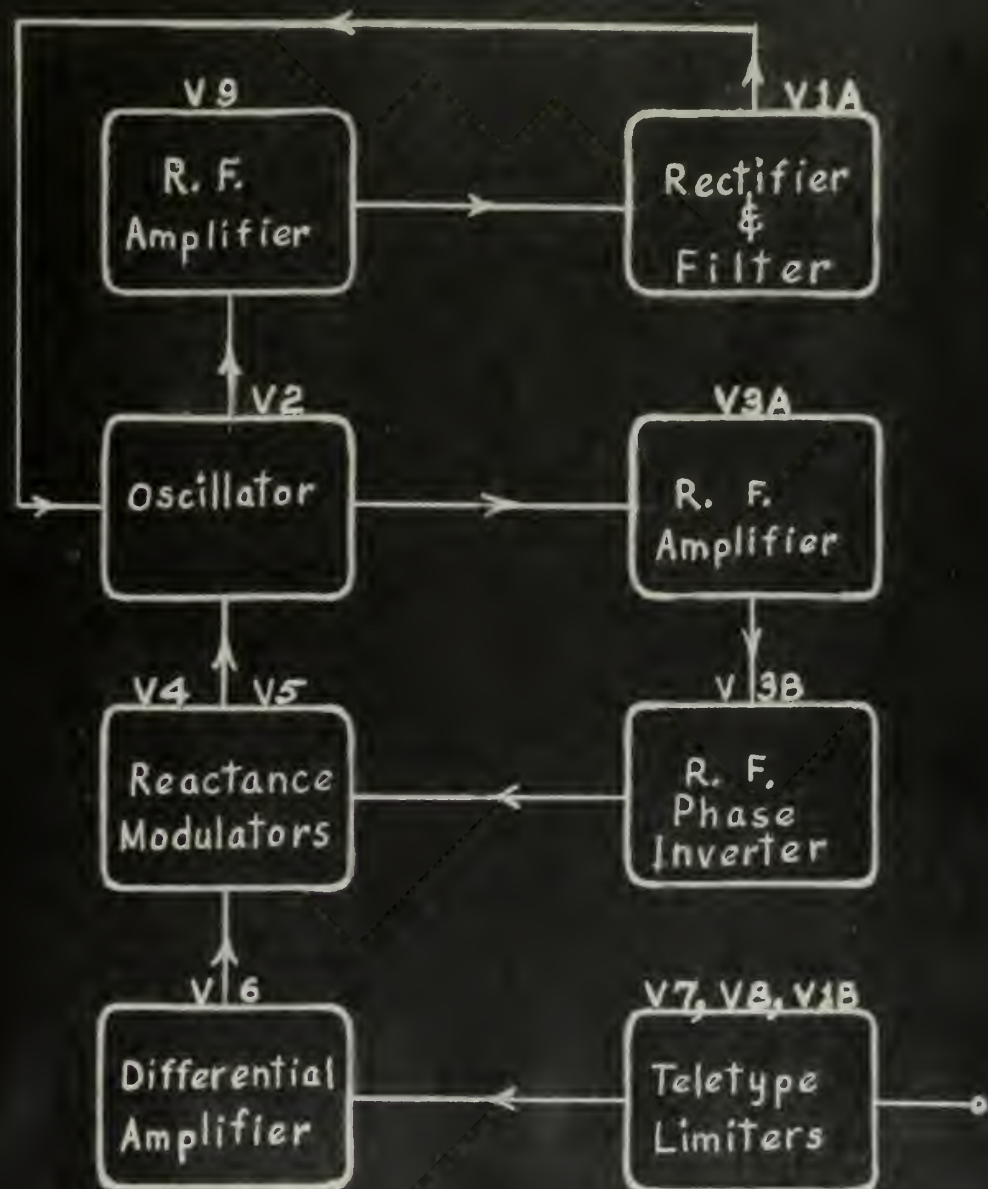

Fig. 13

CHAPTER VI

RESULTS AND CONCLUSIONS

1. Frequency stability.

   (1) Short term drift is approximately $3\frac{1}{2}$ cycles.

   (2) Center frequency drift over a period of 17 hours was measured as 3 cycles.

   (3) Voltage regulated power supplies are required for all supply voltages, including heater voltage. Advertising literature from manufacturers of voltage regulated supplies states that the output voltage can be maintained to within 0.2% for ± 10% variation of input voltage. Assuming this to be true, the maximum frequency error due to changing supply voltages is computed from experimental data to be 0.672 cycles.

   (4) The effects of temperature and humidity variations are not known. In order to minimize the effects of these variations, wire wound resistors were used in all places where variation in resistance could result in a variation of frequency.

   (5) Frequency stability of the oscillator, with the reactance circuits disconnected, was found to be 1 part in 500,000 for a 10% change of plate supply voltage.

2. Linearity of frequency shift for facsimile.

Linearity is well within 10% as required by the specifications. The linearity curve is shown in Fig. 11.

3. Magnitude of frequency shift.

(1) Frequency shift obtained was ± 500 cycles as required. The magnitude of the frequency shift can be varied from zero to maximum by the use of a potentiometer in the grid circuit of the R.F. phase inverter, to vary the R.F. signal voltage from zero to maximum.

(2) The amount of frequency shift as a function of keying speed was not measured directly due to the lack of suitable measuring equipment. The magnitude of the keying voltage applied to the reactance tube grids was found to be independent of the keying speed up to 240 dot cycles and beyond.

4. Keying signal wave forms.

Photographs of wave forms occurring at two significant points at 300 cycle and 2000 cycle input frequencies are shown in Fig. 14.

5. Conclusions.

(1) Some improvement in the flatness of rectangular wave output of the limiters can be effected by increasing the value of R8 as stated in Chapter IV, Section 2.

(2) Neglecting variations of temperature and humidity, whose effects are not known, the stability of the oscillator is sufficient to meet the specifications. It is considered possible that stability is sufficiently good to meet the specifications over the range of temperature and humidity required by the specifications.

(3) Short term drift of plate current, and short term frequency drift as described in Chapter V, Section 1, are interesting phenomena and warrant further investigation.

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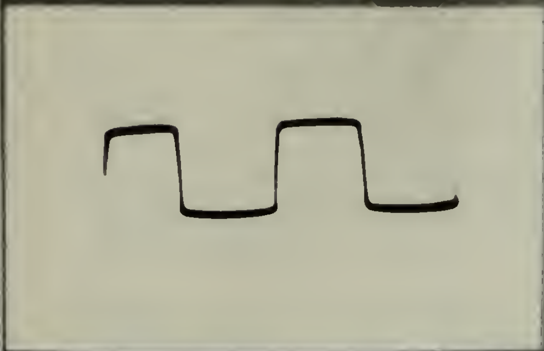
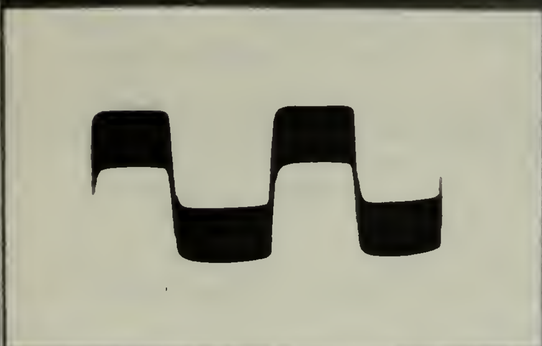
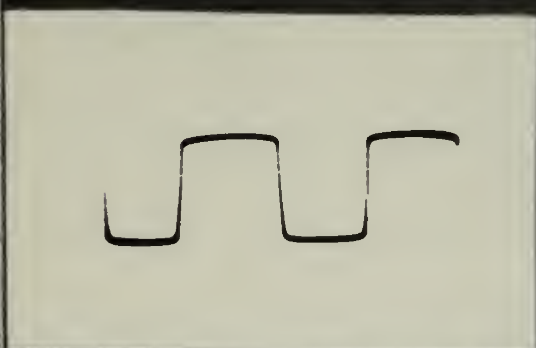
WAVE FORMS



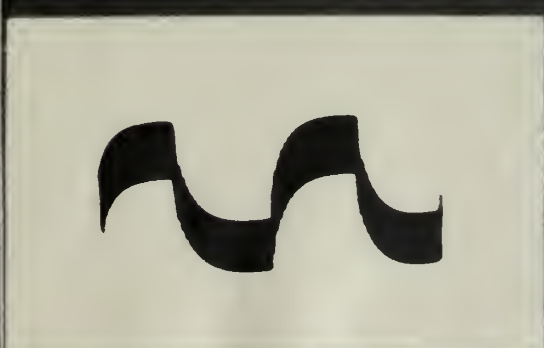
Waveform at point B, Figure 11 with 300 cycle sine wave input at point A, Figure 13.



Waveform at point C, Figure 11 with same input signal as above.



Waveform at point B, Figure 11 with 2,000 cycle sine wave input at point A, Figure 13.



Waveform at point C, Figure 11 with same input signal as above.

Figure 14

BIBLIOGRAPHY

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2. Buff, C. Frequency shift keying techniques. Radio 30:8: 14-17, August 1946.

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APPENDIX 1

Computation of frequency drift due to screen voltage drift of reactance modulator (See page 17)

Δf = total frequency shift = 1000 cycles

ΔE = screen voltage swing to achieve Δf, = 67 volts

sensitivity $= \dfrac{\Delta f}{\Delta E}$

$$\frac{\Delta f}{\Delta E} = \frac{1000}{67} = \text{14.92 cycles per volt}$$

δ = drift voltage at screen = .06 volts

Drift due to $\delta = \delta \dfrac{\Delta f}{\Delta E} = .06 \times 14.92 = 0.9$ cycles

Since the drift is approximately 10 cycles, it is seen that the drift due to screen voltage drift, i.e., 0.9 cycles, is only approximately 10% of the total drift.

APPENDIX 2

Time constant analysis of reactance tube input circuit using Laplace transforms.

Experiments revealed that R_2, R_3 , C_4 , and C_5 of Fig. 8 were not necessary to provide adequate isolation of the reactance modulator from the differential amplifier at 100 kilocycles.

The circuit to be analyzed is given in Fig. 15

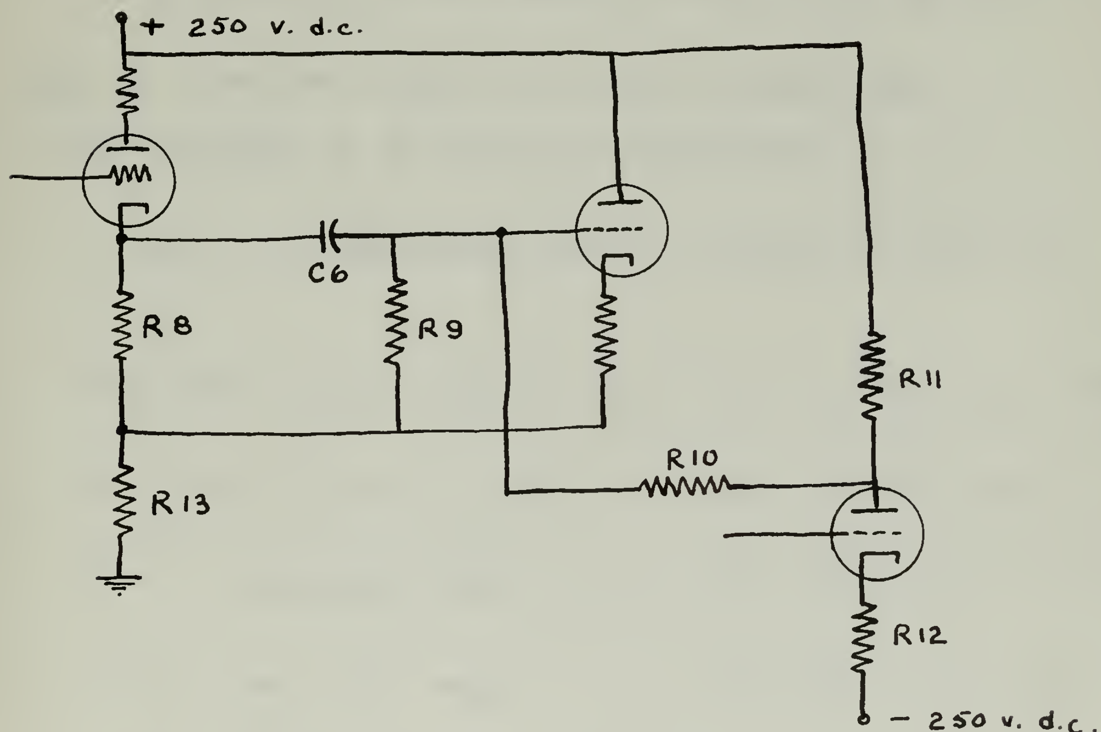
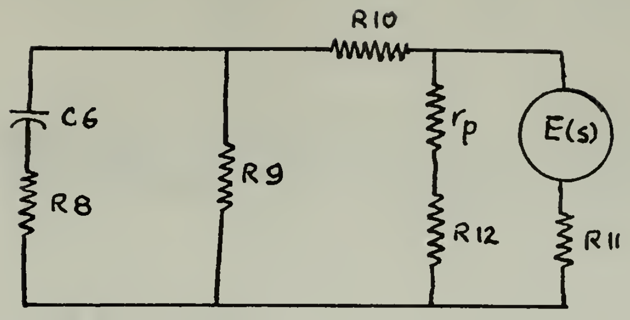


Fig. 15 - Schematic diagram of input circuit to reactance modulator.

Figure 15 can be reduced to the equivalent circuit shown in Figure 16.



$R_{13} \ll$ than any other R and can be neglected

C distributed $\ll C_6$ and can be neglected

Fig. 16 - Equivalent circuit diagram of Fig. 15.

$R_9 = R_8 = 5 \times 10^3$ ohms

$R_{11} = R_{12} = 110 \times 10^3$ ohms

$R_{10} = 100 \times 10^3$ ohms

$r_p \doteq 140 \times 10^3$ ohms

$R_A = R_{11} + R_{12} + r_p = 360 \times 10^3$

$R_B = R_9 + R_{10} + R_{12} + r_p = 365 \times 10^3$

$R_C = R_8 + R_9 = 10 \times 10^3$

$r_p + R_{12} \doteq 250 \times 10^3$

$$R_A I_{1(s)} - (r_p + R_{12}) I_{2(s)} = E(s)$$

$$-(r_p + R_{12}) I_{1(s)} + R_B I_{2(s)} - R_9 I_{3(s)} = 0$$

$$-R_9 I_{2(s)} + R_B + \frac{1}{Cs} = 0$$

$$I_{3(s)} = \frac{(r_p + R_{12})\, R_9 \left(\frac{1}{s}\right)}{R_A\left[R_B\left(R_c + \frac{1}{Cs}\right) - R_9^2\right] + (r_p + R_{12})^2 \left(R_c + \frac{1}{Cs}\right)}$$

$$= \frac{250 \times 10^3 \times 5 \times 10^3 \left(\frac{1}{s}\right)}{360 \times 10^3 \left[365 \times 10^3 \left(10^4 + \frac{1}{Cs}\right) - 25 \times 10^6\right] + 6.25 \times 10^{10} \left(5 \times 10^3 + \frac{1}{Cs}\right)}$$

$$= \frac{1.25 \times 10^9 \left(\frac{1}{s}\right)}{3.60 \times 10^5 \left[3.65 \times 10^9 + \frac{3.65 \times 10^5}{Cs} - 25 \times 10^6\right] + 3.13 \times 10^{14} + \frac{6.25 \times 10^{10}}{Cs}}$$

$$\doteq \frac{1.25}{(s) 3.6^2 \times 10^5 + \frac{3.65^2 \times 10}{Cs} + (s) 3.13 \times 10^5 + \frac{6.25 \times 10}{Cs}}$$

$$= \frac{1.25}{(s)\, 6.75 \times 10^5 + (13.4 + 6.25) \frac{\times 10}{C}}$$

$$= \frac{1.25}{s + \frac{19.65 \times 10^{-4}}{6.75C}}$$

$$= \frac{1.25}{s + \frac{2.9 \times 10^{-4}}{C}}$$

$$i_3(t) = 1.25\, e^{\frac{-2.9 \times 10^{-4}}{C}}$$

Time constant $= \dfrac{C}{2.9 \times 10^{-4}}$

It is desired that the time constant be less than 1/20 of the period of a 240 dot cycle to give a reasonably rectangular wave form at the reactance tube grid

$$T = \frac{1}{f} = \frac{1}{240}$$

$$\frac{T}{20} = \frac{1}{240 \times 20} = \frac{1}{4800}$$

$$\frac{C}{2.9 \times 10^{-4}} = \frac{1}{4800}$$

$$C = \frac{2.9}{4.8} \times 10^{-7}$$

$$= .6 \times 10^{-7}$$

$$= .06 \ \mu f.$$

Closest value available was .05 and this was used. Satisfactory wave shape was obtained as can be seen by Fig. 14.