GENERATION OF SHORT ELECTRICAL IMPULSES

FOR GATING MINIATURE TYPE

VACUUM TUBES

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SUMMARY

A particular application requires a voltage pulse of 1/100th microsec. duration, 30 volts amplitude and 100 kc/sec. recurrence frequency. Methods usually employed for low-frequency, microsecond duration impulses are found inadequate. The author analyzes, using the La Place transform theory, the transient response of two circuits. The differentiating circuit is found impractical. Class C operation of a pentode, having sinusoidal input, with a parallel LC plate circuit can give the desired pulse.

For analysis an idealized plate current-grid bias characteristic is used. The transient response of a 6 mc/sec. circuit to a 700 kc/sec. input was measured. Definite values of G_m and E_{co} can be obtained from measured $I_p - E_g$ curves which give correlation between calculated and experimental transient response. The calculated pulse output at 33 mc/sec. based on the ideal $I_p - E_g$ curve is verified by experiment.

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PART I

Some Theoretical Aspects

1. INTRODUCTION

Radar equipments could not exist without the ability to generate short duration electrical impulses. As a result then of the colossal effort expended in the development of methods of radiolocation, a wealth of information has been made available regarding the generation of electrical pulses. Because the manner in which electrical signals and components are employed in these circuits is often unusual, a new vocabulary has been added to the communication engineer's dictionary. In most cases these additions are picture words whose meanings are obvious. If doubt exists reference can be made to a glossary of terms included at the end of the text.

For most purposes pulse recurrence frequencies (p.r.f.) greater than 10 kc/sec. and pulse durations less than one-half microsecond were of little value in radar detection systems. Nevertheless applications do exist for pulses having recurrence frequencies of 100 kc/sec. and of millimicroseconds duration. Little information, however, is available on techniques employable for producing them.

It is the purpose of this investigation to discover a simple means of generating 10 millimicrosecond pulses having a recurrence frequency of 100 kc/sec. and having sufficient amplitude, approximately thirty (30) volts, to gate the control grid of a vacuum tube. In the interest of portability and economy increasing emphasis is being placed, in both civilian and military design, on the use of miniature techniques. In appreciation of this fact only miniature type vacuum tubes will be considered for the actual pulse generating circuits and as the tube to be gated.

2. AN APPLICATION FOR THE PULSE GENERATOR

2.1 General

As the facilities which radio communication systems provide have improved there has been an increasing demand for frequency allocations. As one approach to satisfying this requirement the tolerance on carrier-frequency variations has become more stringent. For fixed station, fixed frequency operation the quartz-crystal oscillator provides adequate stability. As a means of simultaneously giving the reliability of the crystal oscillator and the flexibility of a variable-frequency oscillator many automatic frequency control systems have been devised. The reactance-tube-controlled multiplier to be described is an example of such a system, which requires a pulse generator possessing the apparently arbitrary conditions established in the introduction.

2.2 A Reactance-Tube-Controlled Frequency Multiplier (1)

Figure 1 is a block diagram of a reactance-tube-controlled synchronizer, and Fig. 2 is a circuit which employs this principle to multiply a 100 kc/sec. sine wave by 10 to obtain a 1 mc/sec. output. In Fig. 2 the input is a low level 100 kc/sec. sine wave driving the amplifier V_1. The tube V_2 is a Class C amplifier with a conduction angle of approximately 90^o. The delay line in the plate circuit generates short square voltage pulses at the leading and the trailing edges of the conduction period. Tube V_3 is a time discriminator with inputs of the

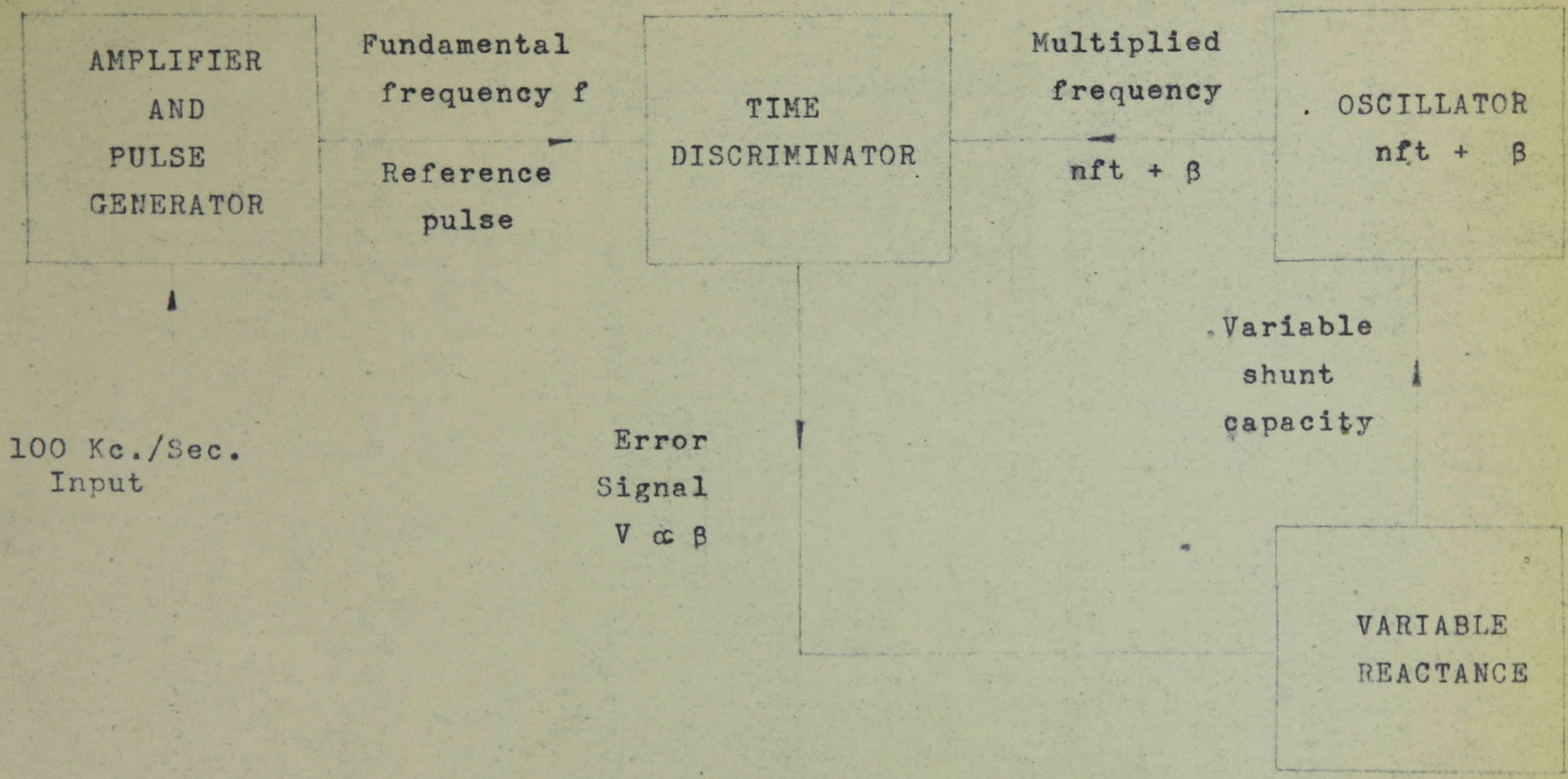


Fig.1 Block diagram of reactance-tube-controlled frequency synchronization.

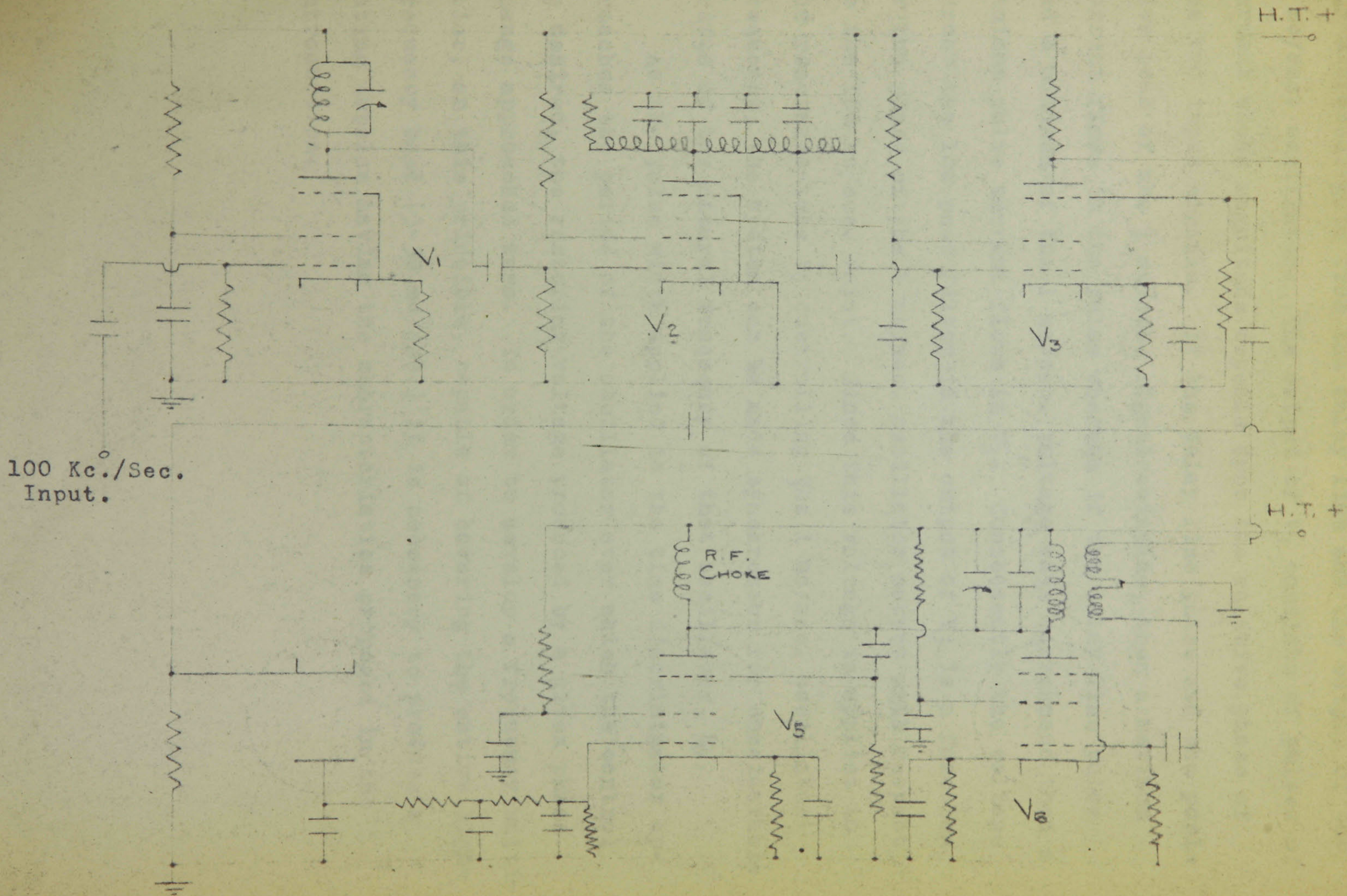


Fig.2 Reactance-tube-controlled Multiplier.

0.3 microsec. gate from the delay line and the output of the 1 mc/sec. oscillator. The output of V_3 consists of pulses of current whose magnitude depends upon the relative phase of the two input signals. If the delay line pulse and the positive peak of the 1 mc/sec. signal coincide, then a maximum current flows in the pulse whereas if the delay line pulse and a minimum of the 1 mc/sec. voltage wave correspond then a minimum pulse current flows in V_3. Consequently the voltage across the low pass filter in the output of V_3 is a function of the phase of the 1 mc/sec. oscillator output with respect to the 100 kc/sec. input. Since this voltage is applied to the reactance-tube V_5 controlling the 1 mc/sec. oscillator frequency, the system can be made synchronous for considerable drifts of the natural frequency of that oscillator, V_6.

As the pulse width applied to the time discriminator approaches the period of the oscillator over which the control is desired, the restoring voltage produced by a given phase change approaches zero. In order to develop a frequency multiplier, on this principle, capable of covering the entire high-frequency band (3-30 mc/sec.) it is necessary to produce a gating impulse having the characteristics proposed in the introduction.

3. PULSE GENERATORS

3.1 General

A pulse generator is a device which will produce an electrical potential or current, of an arbitrary shape, for a time which is short with respect to the period with which the condition recurs. The means by which this may be accomplished may be placed in two categories.

There are systems, Figure 3, in which the energy expended during the pulse is slowly stored during the interpulse period to be released entirely or partially into the load by a switch action.

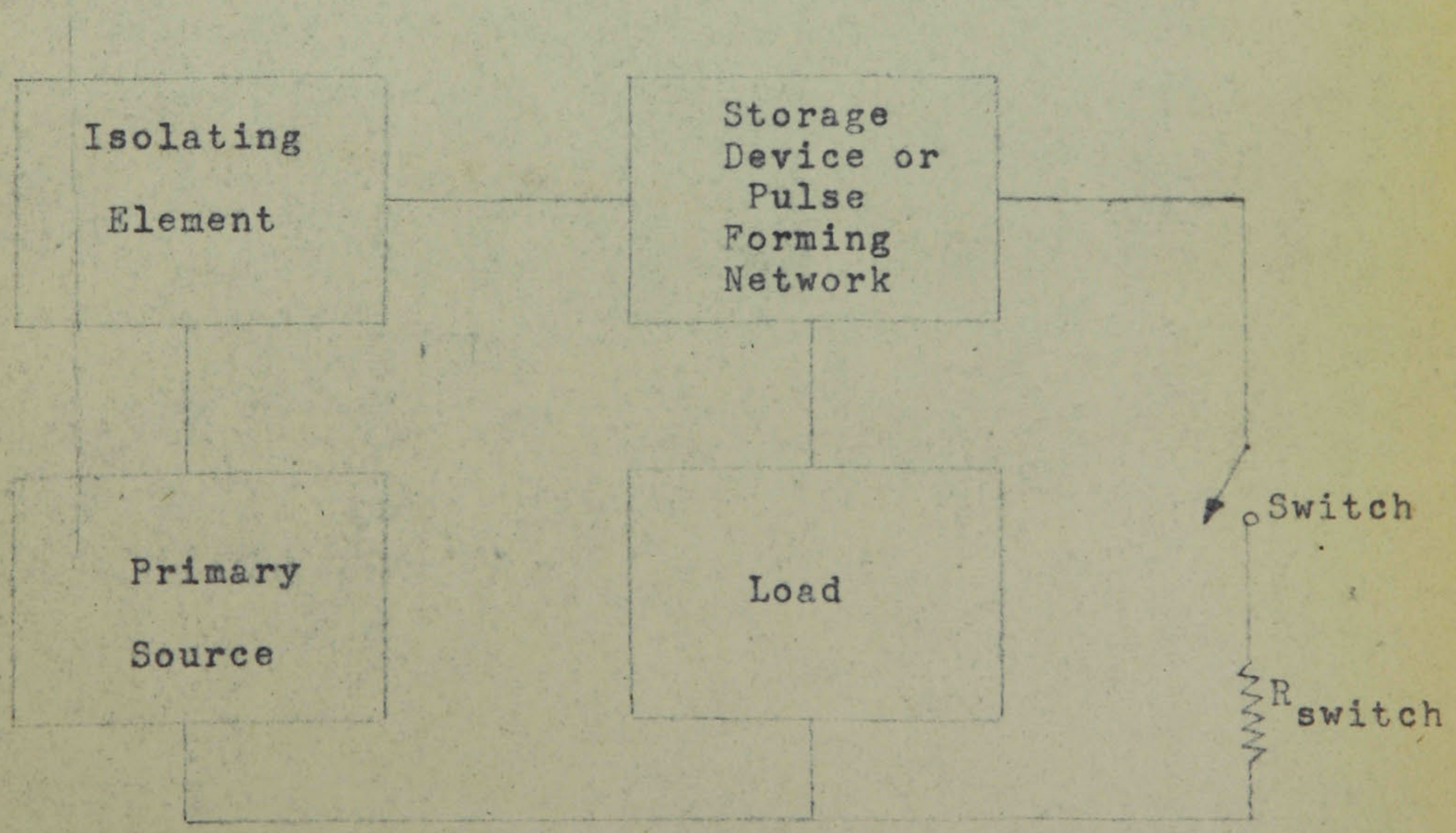


Fig.3 Storage Type Pulse Generator.

In the case in which the energy is completely dissipated the storage device usually determines the shape of the pulse and is called the pulse forming network. In the latter situation where only part of the stored energy is expended the pulse shape is determined primarily by the manner in which the switch is opened and closed. If, however, the storage device does not have an infinite capacity it determines to some extent the pulse shape as will be shown presently. This method of storage and discharge is usually employed where considerable energy is involved in the pulse since the primary voltage or current source may then have a much lower instantaneous power capacity.

Alternatively all of the energy involved may be supplied from the primary source as the pulse is produced. This method, a block diagram of which is shown as Figure 4, is particularly adaptable to low level generators where the instantaneous power requirements are within the capabilities of an economical source.

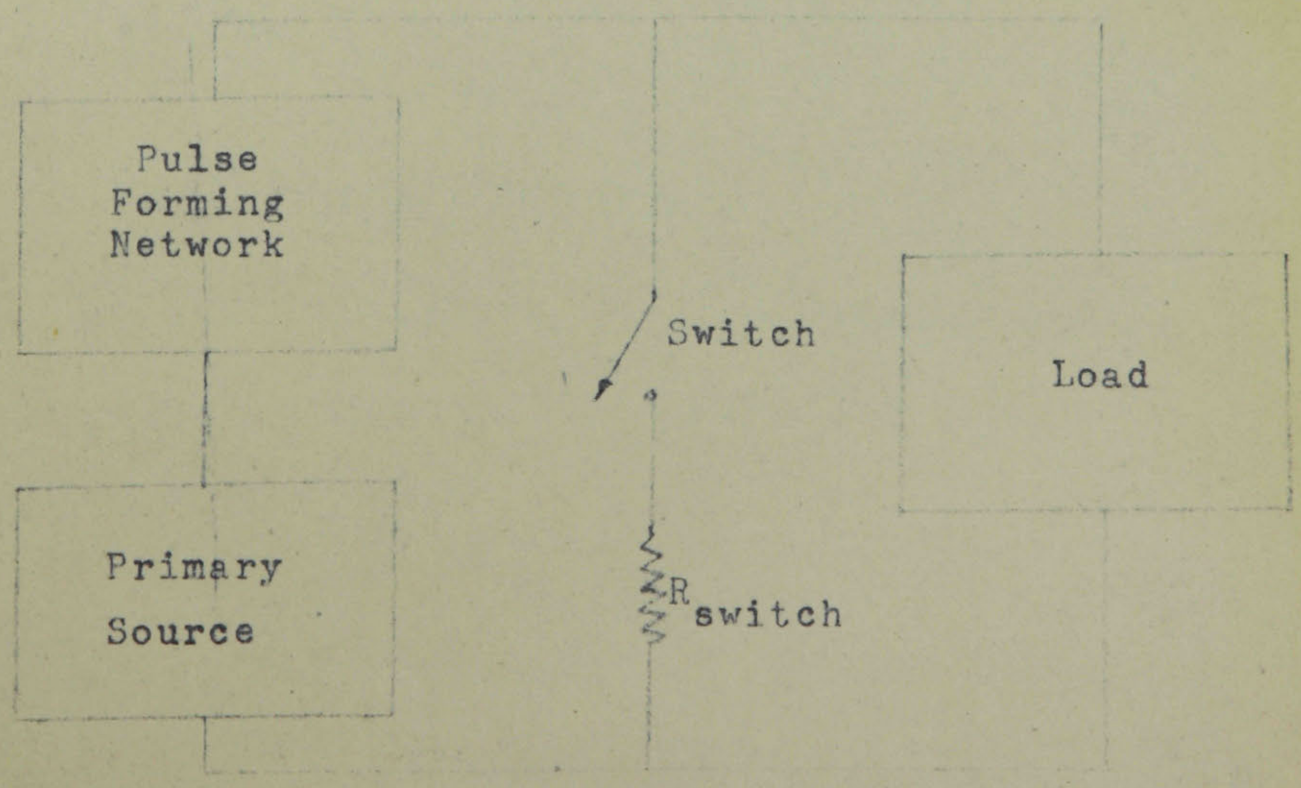


Fig. 4 Direct Type Pulse Generator.

3.2 Phase Inversion

3.2.1 General

It may be discovered, as their analysis progresses, that some pulse generator circuits can produce conveniently only negative going pulses. Since the ultimate concern here is to obtain a positive pulse it will be well to establish whether phase inversion of pulses of the duration desired is a feasible process.

Inversion may be accomplished by either vacuum tube amplification (8) or by use of an electromagnetic transformer (2). In the former case the load impedance determined by the pulse generator may be coupled to the grid of the amplifier tube while in the transformer coupling the correct load impedance as seen from the pulse generator and the desired output impedance may be simultaneously realized by proper choice of turns ratio.

3.2.2 Vacuum Tube Amplification

To discover the effect, on a rectangular pulse, of amplification using RC coupling the simplified circuit of Figure 5 may be used.

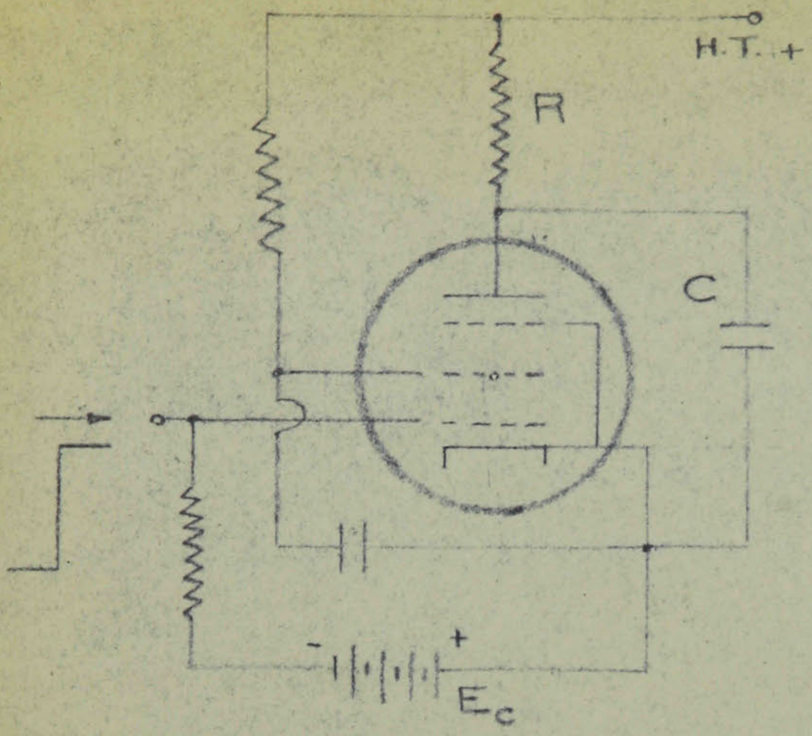


Fig.5.

RC Coupled Amplifier
Simplified Schematic
For Determination of
Step Function
Response

There R is the plate load resistor of the pentode and C is the capacity from plate to ground which includes C_{pk} of the amplifier, C_{gk} of the following tube and stray wiring capacity. The voltage amplification is

$$Gain = G_m R \qquad \underline{1}$$

where G_m is the transconductance of the tube. The voltage response to a step function of current, initiated by a rectangular voltage pulse applied at the grid, is an exponential curve free from overshoot. Now the rise time of an exponential curve as measured between the 10 and 90 percent final amplitude points is given by

$$Rise\ Time\ (T) = 2.2\ RC \qquad \underline{2}$$

Dividing E_q.1 by E_q.2 produces a convenient figure of merit for a tube with respect to pulse amplification.

$$\frac{\text{Gain}}{\text{Rise Time }(T)} = \frac{G_m}{2.2C}$$

3

Now for an output tube the chief concern is with the peak voltage obtainable. Multiplying both sides of Eq.3 by E_g yields as a figure of merit for an output stage the relationship

$$\frac{E_p \text{ (Max.)}}{\text{Rise Time }(T)} = \frac{I_P}{2.2C}$$

where E_p(max.) and I_p are the amplitudes of the output signal and the current pulse in the valve respectively. The maximum output voltage realizeable for a given rise time would be obtained if the input pulse operated the tube from cut-off to zero bias for a positive signal and vice versa for a negative input.

Now when considering the amplification of negative pulses having a small duty cycle, it must be recognized that the average current in the tube is essentially the zero signal value. Usually in order to stay within the power dissipation rating of the tube the grid must be operated with considerable negative bias. Consequently the amplitude of the current pulse for a negative going input signal is much lower than that obtainable for a positive going pulse.

A table of E_p(max.)/T, Fig. 6, has been compiled (8) for those valves normally employed as pulse amplifiers. It is assumed for this tabulation that the input signal is a negative going pulse of small duty cycle and that the capacity beyond the amplifier itself is 20 micro-microfarads (pf.).

Tube Type	I_p (ma.)	C (μμf.)	$E_{max.}$ / T .
6AK5	10	23	200
2-6AK5's (parallel)	20	26	350
6AN5	45	24.5	800
6AQ5	80	26	1400

Fig.6: Relationship between max'm plate volts and rise-time for a pulse output amplifier.

Now if a pulse having an exponential rise to 90% of the steady state and an exponential fall to zero is acceptable, then for a pulse 1/100 microsec. wide at the 10% point T = 1/200 microsec. From the table above it is seen that the maximum amplitude obtainable is with a 6AQ5 and equals 7 volts.

It is possible by use of reactive networks inserted in the plate of the RC coupled amplifier to decrease the rise time by a factor of two without introducing overshoot or oscillations. Even this improvement will not bring the pulse amplitude up to the required level.

The system known as distributed amplification (9) is capable of producing the required output voltage. The complexity of this method and its excessive use of vacuum tubes are major deterrents to its use.

It is apparent that elementary forms of vacuum tube amplification are not capable of providing the 30 volts of amplitude required in the original statement of the problem. They must therefore be rejected as a means of phase inversion.

3.2.3 The Pulse Transformer

A discussion of the properties of transformers useful in the inversion of pulses of very short duration is quite beyond the scope of this thesis. It has been concluded from the literature available (1)(2)(10) that at present they are not valuable for pulse lengths shorter than 0.1 microsec. unless distortion of the pulses by oscillatory transients is permissible. Essentially they would act as tuned circuits whose natural period is determined by the leakage inductance and the interturn and interwinding capacities. Since, in any event, both positive and negative signals with respect to the quiescent state may be obtained from a tuned circuit the use of a pulse transformer to produce the transient oscillations is superfluous.

3.3 Switches

3.3.1 Switch Types

The switch is an essential part of both the systems referred to in (3.1). It may be either mechanical or electronic depending upon the rate at which the switching action occurs and upon the power requirements of the system. In consideration of the high recurrence frequency and the size limitations involved in the proposed application mechanical switches will not be discussed here. The electronic switches available will be considered under two headings, 3.3.2 high vacuum tubes and 3.3.3 gas tubes.

3.3.2 High Vacuum Tubes

This class of tubes is commonly referred to as hard tubes. Under normal operating conditions as indicated by the range of values given in tube manuals miniature hard tubes are in general incapable of handling the high peak currents required in pulse generating circuits. The possibility that the current and voltage ratings established for conventional operation might be exceeded was studied (3) and promising results obtained. The tubes under test were operated as Class C amplifiers. The applied signal was a pulse, essentially rectangular, of variable width, recurrence frequency and amplitude. When the tubes were operated in the positive control grid region values of peak current approaching the temperature limited current - from 0.3 to 0.5 amps. per watt of heating power in oxide-coated cathode tubes - were obtained. No reduction in this current during the normal life of the tube was observed except where the average power dissipated by the elements exceeded the ratings as defined for conventional operation.

Many of the tubes tested were not intended for operation in the positive control grid region. When worked in that region it is stated as a rough generalization that the power-dissipating ability of the control grid is one-tenth that of the plate of the tube.

The plate current in a pentode is given by the relationship (4).

$$I_p = K\left(E_{g1} + \frac{E_{sg}}{\mu_{sg}}\right)^{\frac{3}{2}}$$

While it is apparent that the operation of the control grid in the positive region is the most effective way of increasing peak current it should be noted that an increase of screen-grid voltage will accomplish the same purpose. In cases where loading of the driving circuit may be undesirable this latter approach to increased peak current may be employed providing the restrictions on power dissipation are respected.

The chief advantage of the "hard" tube is that it may be turned on and off at will by change of control-grid voltage. In most pulser circuits this switching process should occur in a minimum time and "hard" tubes with sharp cut-off characteristics can accomplish this satisfactorily.

3.3.3 Gas Tubes

The thyratron gas tube is capable of providing peak currents far greater than those in a "hard" valve for the same heater power. The extremely low resistance of this tube used as a switch is a real advantage in circuits having low impedance loads. The fact that the control electrode of this tube is effective only in the initiation of the conduction period prohibits its use except in that group of storage type circuits which dissipate their energy completely during the pulse. In this case the voltage from plate to cathode of the thyratron falls below the critical value required for conduction and the tube extinguishes. Another deterrent to its use is the length of time required for ionization and de-ionization to take place. The former value determines the minimum pulse duration produceable, the latter value establishes an upper limit on the recurrence frequency of the pulse generator.

The thyratrons available in miniature envelopes are filled with argon. These tubes have an ionization time of approximately 0.5 to 5 microsec. and de-ionization times from 35 to 75 microseconds.

Some experimental work (5) has been conducted on the use of thyratron valves for frequencies greater than the reciprocal of the de-ionization time. It is found, for frequencies up to about 200 kc/sec., that a critical ratio of grid-cathode triggering voltage to plate-cathode voltage exists. The large negative grid potential required in this mode of operation apparently removes positive ions from the interelectrode spaces thus hastening de-ionization. While stable operation is obtained the peak currents are considerably lower than the rated currents for the tubes.

Of the thyratrons commercially produced those filled with hydrogen (6) are the most satisfactory for pulse work. Their de-ionization time is dependant upon the plate voltage and can be reduced for operation at frequencies of 100 kc/sec. providing the relationship $e_{py}^2 \times f_r = 2.6 \times 10^{11}$ maximum is observed (e_{py} = peak forward anode volts, f_r = pulse recurrence frequency). Unfortunately no tubes of this design are available in a miniature envelope.

It may be concluded that for operation at 100 kc/sec. gas tubes are at present much less satisfactory than hard valves for use as a switch in a miniature pulse generator circuit.

3.4 Storage Devices

While generally referred to as pulse generators those circuits in which the storage device acts only as a reservoir might be more aptly called pulse power amplifiers. Since the ultimate purpose here is to achieve a pulse shape at a low power level only, no recourse to circuits of this type will be necessary. However, certain pulse-forming networks do employ capacitive or inductive storage of energy. To outline then the principle of storage, before adding the complication of pulse formation, a circuit employing a condensor for this purpose is shown in Fig. 7.

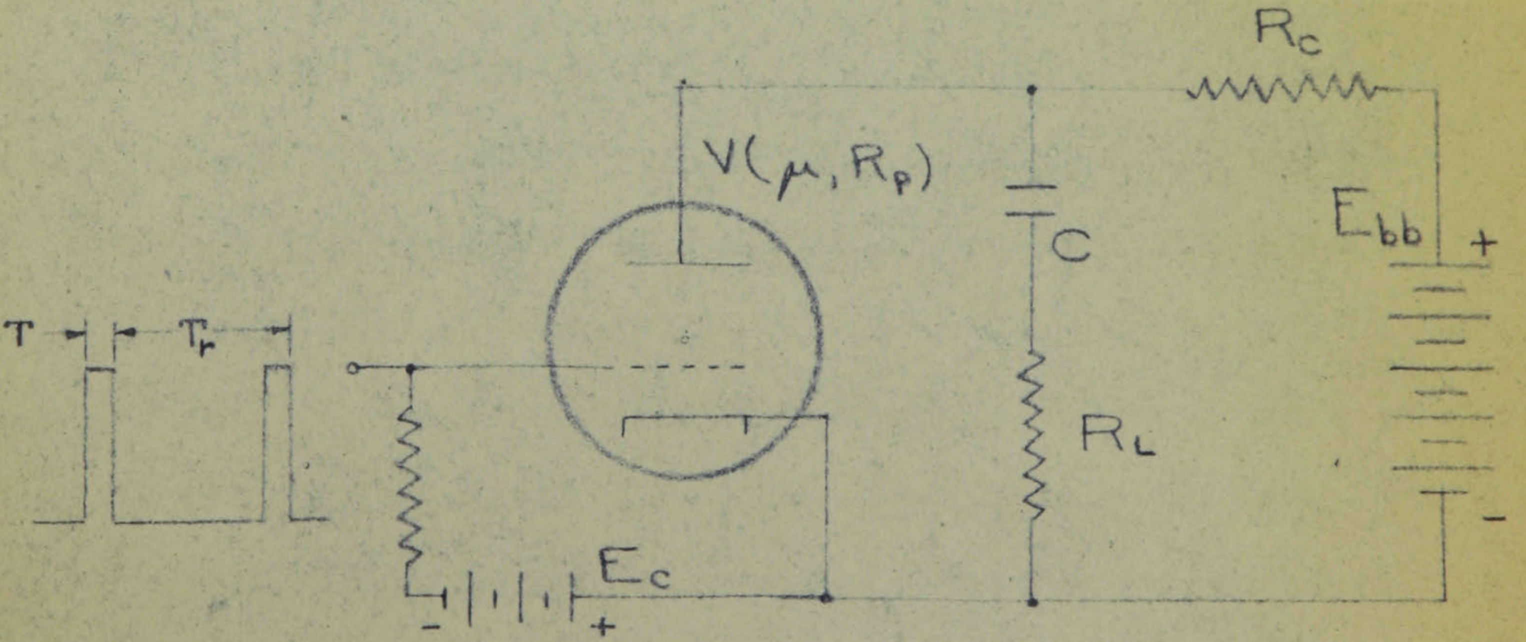


Fig. 7 Voltage Storage Pulse Power Amplifier

In this system the tube V operated as a Class C amplifier acts as the switch. It is closed periodically at intervals T_r

for a time of duration T. During the interpulse period $(T_r - T)$ the condensor C is charging to the potential E_{bb} through R_c and R_L. Resistor R_c acts as the isolating element shown in the block diagram Fig. 3 and is chosen to be large with respect to the load impedance R_L. A negligible voltage then appears across the load during the charging period and a minimum of primary source current flows in the switch tube during the pulse.

If as is shown in the illustration the switch is a triode then, for a rectangular pulse applied to the grid, the tube may be replaced by a switch in series with a resistor having a value equal to the dynamic plate resistance of the triode. Under these circumstances the current in the circuit during the pulse will be $i(t - t_0) = \frac{E_c(t_0)}{R} e^{-\frac{(t-t_0)}{RC}}$ where $R = (R_L + R_p)$. It is evident that the shape of the pulse top is determined by the circuit components. This circuit is usually designed so that $RC \gg T$. Consequently the pulse current is essentially constant during the pulse. In the case where $RC < T$ the output will no longer have vertical sides but will rise abruptly and decay exponentially to zero volts. With this circuit it should be possible to obtain very short pulses by using triodes with low plate resistance or pentodes operated in the coalescent region of the tube characteristic (i.e. for plate voltages below the bend in the I_p-E_p curves). The disadvantage lies in the practical difficulty of obtaining a suitable position in the circuit for the load impedance when positive going pulses are desired. A somewhat similar circuit will be discussed more elaborately in the section on pulse forming networks.

It is well to note that in the circuit just discussed current flows in the switch tube for the pulse duration only. This is an important advantage of voltage storage over current storage systems.

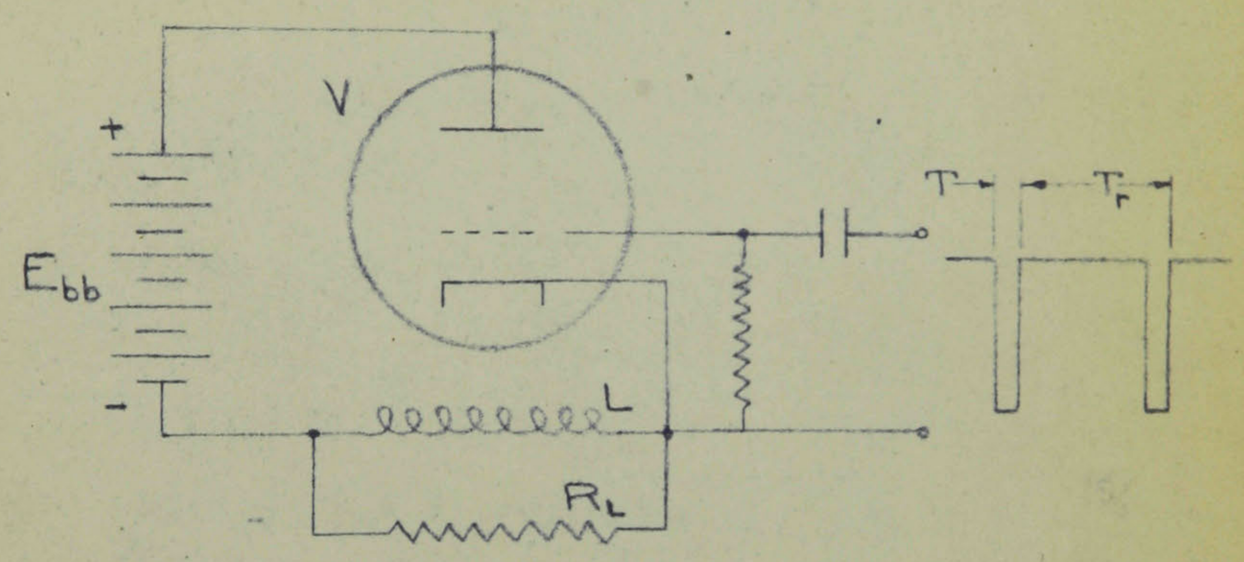


Fig. 8 : Current-storage Pulse Amplifier.

Fig. 8 is that for a circuit in which energy is stored because of the current flow in the inductance L and the switch tube V. The energy is discharged through the load impedance R_L when the tube V is cut off. The tube in this case must conduct during the interpulse period and will normally require much larger power handling capacity than in the former circuit. Furthermore in the current storage method the switch tube is often placed in an awkward position for the application of control signals.

3.5 Pulse Forming Networks

3.5.1 The Matched Transmission Line

Of all the circuit arrangements which have been used to shape the voltage or current impulse that most commonly employed has been the transmission line. Mr. H.J. White (2) shows most eloquently using the La Place transform theory (7) why this should be the case. Since the limitations of this circuit for the present application must be seriously considered and because the La Place theory will be used extensively in the analysis of further more complex problems it will be well to repeat here the work referred to above. Referring to the discharge circuit only of the pulse generator Fig. 3 and assuming that all the energy in the pulse forming network is stored in one condensor, we obtain the circuit of Fig. 9.

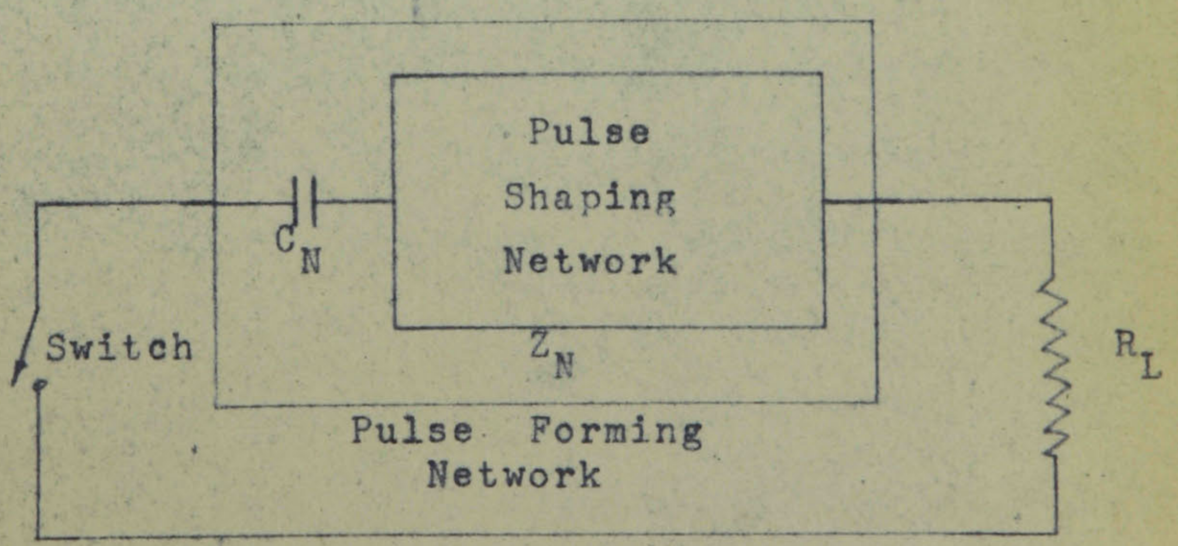


Fig.9 Generalized Circuit to Obtain a Rectangular Current Pulse.

The current in the circuit is given implicitly by

$$i(t) R_L + \frac{1}{C_n}\int_0^t i(t)\,dt + \frac{Q_n}{C_n} + Z_n(t)\cdot i(t) = 0 \qquad \underline{4}$$

where Q_n is the initial charge on the storage condensor, is the integro-differential operator which describes the elements, as yet unknown, contained in pulse-shaping network, and R_L is the load resistance.

Now the definition of the La Place transform of a function $i(t)$ is

$$L\left[i(t)\right] = \int_0^\infty i(t)\, e^{-st}\, dt = I(s) \qquad \underline{5}$$

So the transform of E_q. 4 is

$$\left(R_L + \frac{1}{C_n s} + Z_n(s)\right) I(s) = \frac{Q_n}{C_n s} = \frac{V_n}{s} \qquad \underline{6}$$

where V_n is a constant - the initial voltage on the network. E_q. 6 may be solved for the network impedance giving

$$Z_n(s) = \frac{V_n}{s\, I(s)} - R_L - \frac{1}{C_n s} \qquad \underline{7}$$

Now it is necessary for good microwave oscillator operation and highly desirable for the frequency control application to have an approximately rectangular pulse. It will be appropriate to attempt to discover the pulse shaping network impedance which will achieve this result. The transform of a current pulse of amplitude I_L and duration T is from E_q. 5

$$I(s) = \frac{I_L}{s}\left(1 - e^{-sT}\right)$$

Substituting in E_q. 7

$$Z_n = \frac{sV_n}{s\, I_L \left(1 - e^{-sT}\right)} - R_L - \frac{1}{C_n s}$$

or rearranging

$$Z_n + \frac{1}{C_n s} = R_L \left[\frac{\left(\frac{V_n}{I_L R_L} - 1\right) + e^{-sT}}{1 - e^{-sT}} \right]$$

If the numerator and denominator are multiplied by $e^{\frac{sT}{2}}$ and V_n is chosen equal to $2I_L R_L$ then

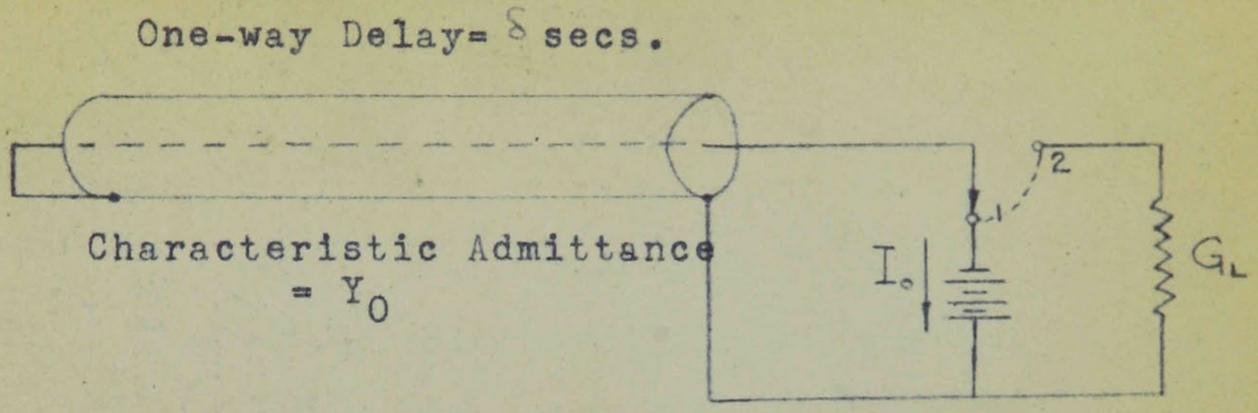
$$Z_n + \frac{1}{C_n s} = R_L \left[\frac{e^{sT/2} + e^{-sT/2}}{e^{sT/2} - e^{-sT/2}} \right] = R_L \coth \frac{sT}{2} \qquad \underline{8}$$

The right-hand side of Eq. 8 is recognized as the impedance function of an open-circuited lossless transmission line of characteristic impedance $Z_o = R_L$ and transmission time $\delta = \frac{T}{2}$. Hence in this case for a rectangular pulse the storage condensor in conjunction with the shaping network must be the electrical equivalent of the transmission line or alternatively the transmission line, with the energy stored in the distributed capacity of the line, may be employed as a pulse forming network. The latter method is the one usually employed.

By a similar analysis it may be shown that a lossless transmission line of delay time δ, short-circuited at one end and carrying an initial current I_o will, when connected to a load of impedance equal to the characteristic impedance Z_o of the line, produce across the load a voltage pulse of magnitude $\frac{I_o Z_o}{2}$ and of duration 2δ.

3.5.2 The Transmission Line with a General Resistive Load

In the event that the transmission line is not operating into a matched load a series of reflections will occur at the mismatch. The nature of the transient produced by discharge of the short-circuited line into a resistance load, Fig. 10, may be studied by application of the La Place transform method of analysis for the general case.



(Switch moves from position 1 to 2 at time t=0)

Fig.10 Circuit for Analysis of Voltage Transient on Unmatched Load.

The transmitting end a.c. admittance of the transmission line (Y_0, δ) with the receiving end short-circuited is given by elementary transmission line theory as

$$Y(\omega) = Y_0 \coth j\omega\delta$$

and the La Place transform of this admittance by

$$Y(s) = Y_0 \coth s\delta$$

Analysing the circuit on a one node basis

$$V(s)\left[Y_0 \coth s\delta + G_L\right] = I(s) = \frac{I_0}{s}$$

whence

$$V(s) = \frac{I_0}{s(G_L + Y_0 \coth s\delta)} = \frac{I_0}{s(G_L + Y_0)} \cdot \frac{1 - e^{-2s\delta}}{1 + \frac{Y_0 - G_L}{Y_0 + G_L} e^{-2s\delta}}$$

$$= \frac{I_0(1 - e^{-2s\delta})}{(Y_0 + G_L)\, s}\left[1 - \frac{Y_0 - G_L}{Y_0 + G_L} e^{-2s\delta} + \left(\frac{Y_0 - G_L}{Y_0 + G_L}\right)^2 e^{-4s\delta} + \dots\right]$$

The inverse transform gives the voltage across the load as

$$v(t) = \frac{I_0}{Y_0 + G_L}\left\{1 - U(t-2\delta) - \frac{Y_0 - G_L}{Y_0 + G_L}\left[U(t-2\delta) - U(t-4\delta)\right] + \left(\frac{Y_0 - G_L}{Y_0 + G_L}\right)^2\left[U(t-4\delta) - U(t-6\delta)\right] - \cdots\right\}$$

where $U(t-a) = 1$ for $(t-a) > 0$
$U(t-a) = 0$ for $(t-a) < 0$

If $Y_0 = G_L$ then all terms except the first two vanish and the voltage consists, as indicated previously, of a single rectangular pulse of amplitude $\frac{I_0}{2Y_0}$ and duration 2δ. In the proposed applications the steps, due to reflection on unmatched load, will be harmless providing that the difference between the initial amplitude and these subsequent steps exceeds the pulse amplitude desired. In the case $Y_0 < G_L$ the magnitude of the initial pulse exceeds that after the addition of the first reflected voltage wave by

$$V_{Y_0 < G_L} = \frac{I_0}{Y_0 + G_L}\left\{1 + \frac{Y_0 - G_L}{Y_0 + G_L}\right\} = \frac{2 I_0 Y_0}{(Y_0 + G_L)^2}$$

So the pulse amplitude of width 2δ is always less than that obtainable for matched load. In the case where $Y_0 > G_L$ the magnitude of the initial impulse exceeds that after the addition of the second reflection - since in this case the first reflection produces - ve total amplitude - by the voltage

$$V_{Y_0 > G_L} = \frac{I_0}{Y_0 + G_L}\left\{1 - \left(\frac{Y_0 - G_L}{Y_0 + G_L}\right)^2\right\} = \frac{I_0 \times 4 Y_0 G_L}{(Y_0 + G_L)^3}$$

For this amplitude to equal or exceed the matched load amplitude

$$\frac{I_0 Y_0}{2 G_L^2 \left(\frac{1}{2} + Y_0/2G_L \right)^3} \geq \frac{I_0}{2Y_0}$$

or

$$8 \geq \left(\frac{G_L}{Y_0}\right)^2 \left(1 + \frac{Y_0}{G_L}\right)^3$$

A graph of the right-hand side of this expression is shown as Fig. 11.

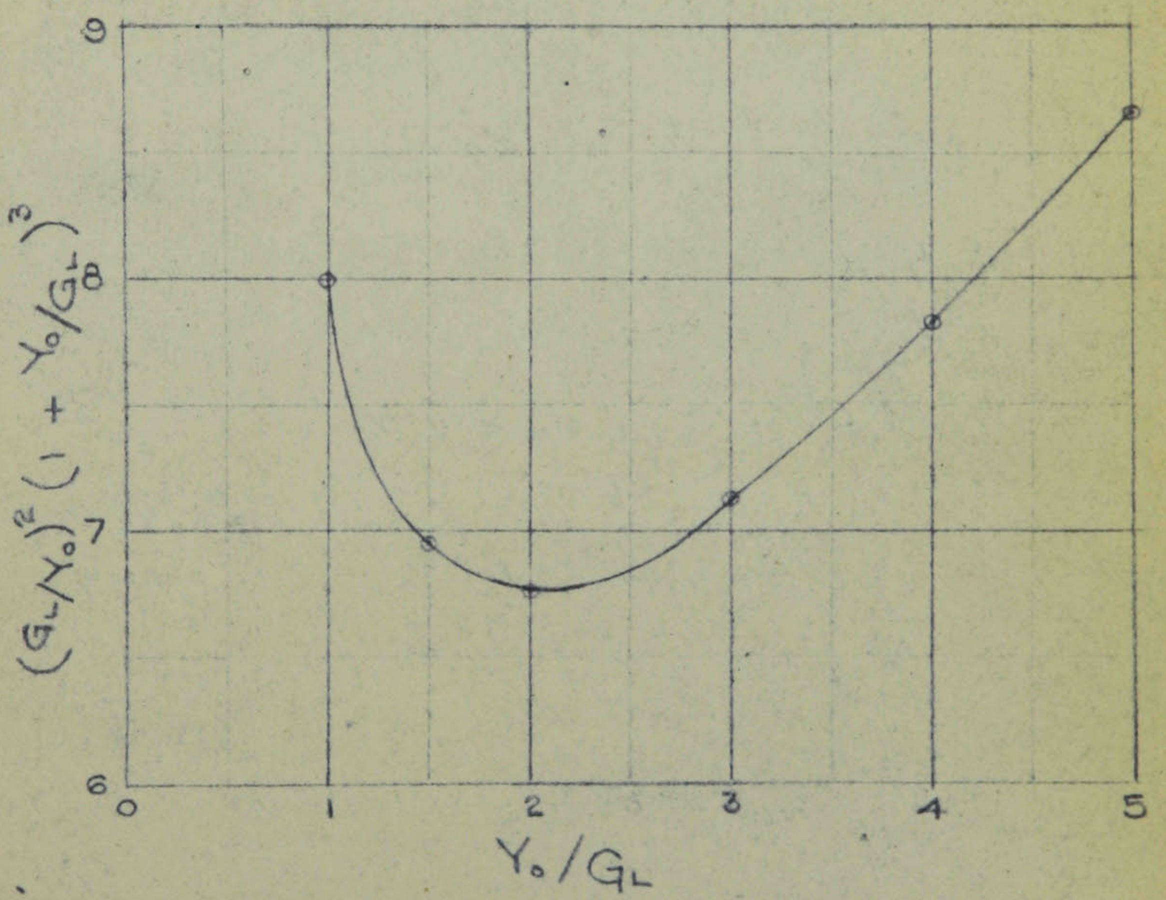


Fig. 11 Graph for Establishing Range of Ratio Y_0/G_L for Pulse Output Greater Than for Matched Load.

It is apparent that maximum amplitude = $\frac{I_0}{1.7 Y_0}$ is obtained for $Y_0 \doteq 2 G_L$ and that for $G_L < Y_0 < 4.2\, G_L$ the amplitude is at least as large as the matched load value.

Furthermore for a lossless transmission line it may be determined that $Y_0 = \sqrt{\frac{C}{L}}$ and that $\delta = \sqrt{LC}$ where L, C, and δ

are the total distributed inductance, capacity and transmission time respectively. It is possible now to determine, assuming a step-function of applied grid voltage and of plate current, whether with available hard valve switches a transmission line having parameters practically attainable will provide the required output pulse.

3.5.3 The Transmission Line in a Practical Pulse Generator Circuit

The miniature tube most suitable for this application is the 6AN5. It has a peak current capacity, in the negative control-grid region, of 100 ma. and an output capacity of 4.5 pf. Now for a pulse amplitude of 30 volts across the load in the circuit of Fig. 12

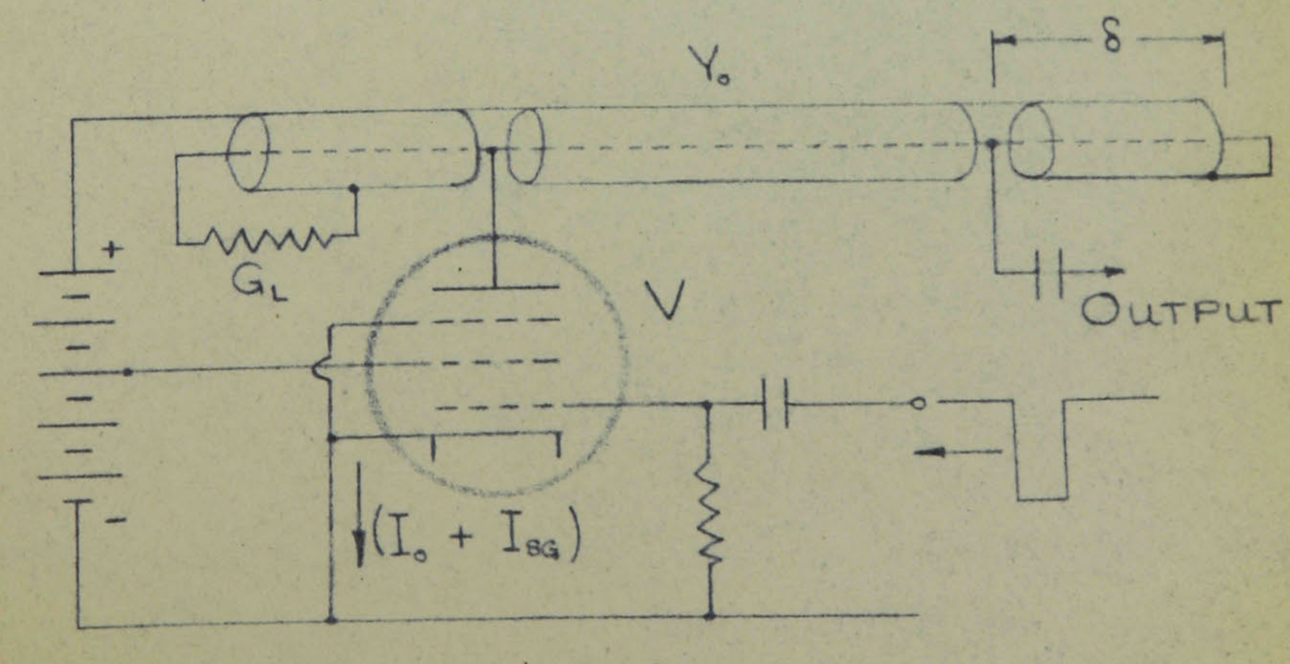


Fig. 12 The Practical Transmission Line Circuit For Pulse Generation

using the matched line voltage output as a design value.

$$30 \text{ Volts} = \frac{I_o Z_o}{2} = \frac{10^{-1} Z_o}{2} \quad \text{whence } Z_o = 600 \text{ ohms}$$

For a pulse width of 1/100 microsec. δ = 1/200 microsec. So

$$\delta = \sqrt{LC} = Z_0 C = 5 \times 10^{-9} \text{ secs.}$$

whence

$$C = \frac{5 \times 10^{-9}}{600} = 8.3 \text{ pf.}$$

This capacity is of the same order of magnitude as the sum of the output capacity of the switch valve and the input capacity of a driven valve. Consequently the transmission lines properties will be vitiated unless some satisfactory means can be found to incorporate these circuit capacities into that of the transmission line. This may be done by use of a synthetic transmission line consisting of lumped inductances and capacitances connected as in a Constant-K type low-pass filter, Fig. 13.

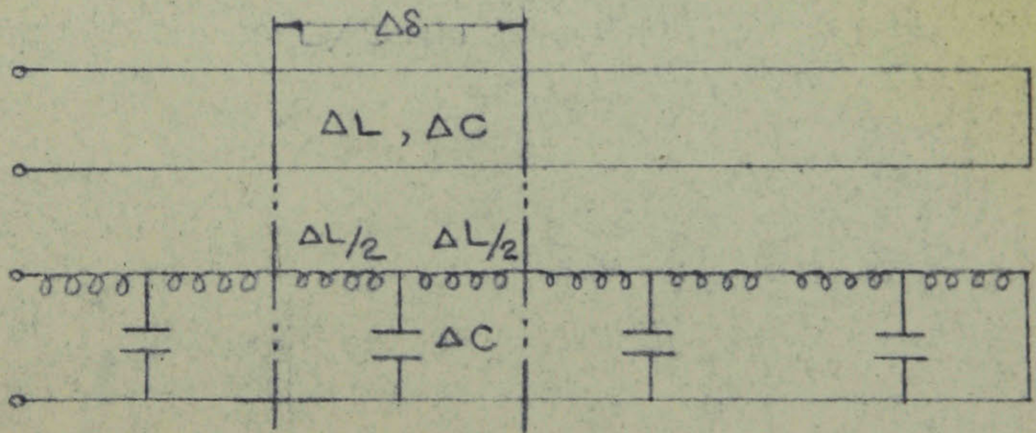


Fig.13 Synthesization of a Uniformly Distributed Constants Transmission Line

It can be shown that the synthesized line approximates the actual line for pulse generation only if the transmission time ($\sqrt{LC}$) per lumped section is short with respect to the pulse it is desired to generate. If the output capacity of the switch presents the only capacity in the lumped section in which it is incorporated, then the transmission time for that section $\delta_s = \sqrt{L_s C_s} = Z_0 C_s = 600 \times 4.5 \times 10^{-12} = 2.7 \times 10^{-9}$ a value which is not small compared to 10^{-8} sec.

A further deterrent to the use of a transmission line as a pulse forming network is found if the idealization that a step function of current is applied to the line is modified.

A more realistic condition would be a linear rise from zero current to the peak value. The effect of this on pulse amplitude and shape may be most readily studied by use of the series of waveforms in Fig. 14. These represent the incident and reflected voltage waves and their resultant at the load, on a lossless transmission line short-circuited at one end and matched at the other by its characteristic impedance.

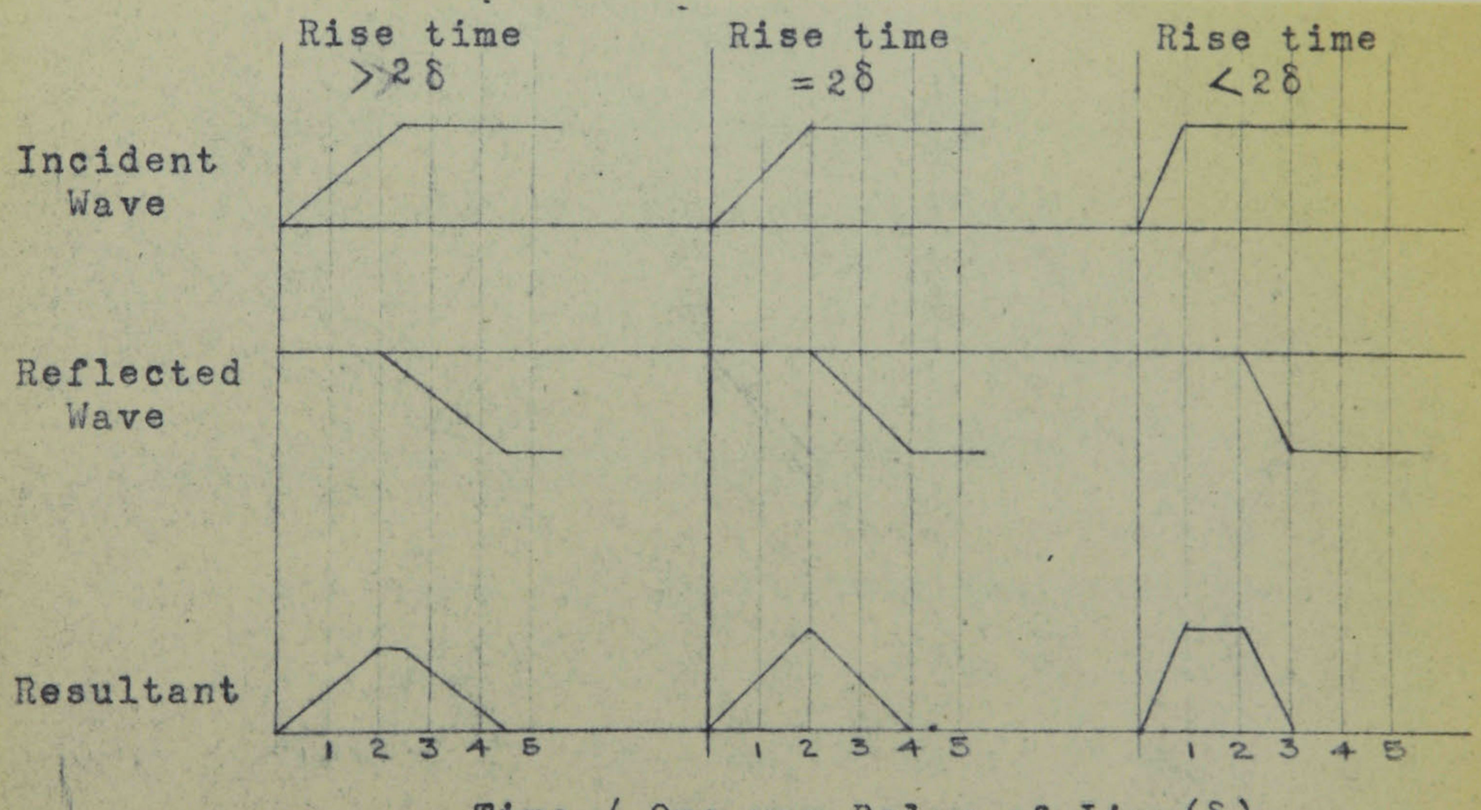


Fig. 14 Effect of Rise time of Current Wave on the Pulse Shape Across the Load.

From these diagrams the pulse amplitude and base width can be deduced. The pulse amplitude is given by:

$$V_{pk} = \frac{I_o Z_o \times \delta}{\text{Rise Time}} \qquad \text{(for rise time } > 2\delta)$$

$$= \frac{I_o Z_o}{2\delta} \qquad \text{(for rise time } \leq 2\delta)$$

and the pulse width at the base is given by

$$T = 2\delta + \text{Rise Time} \qquad \text{(for all rise times)}$$

It is apparent that, as the rise time increases from zero, if δ remains constant the peak signal required to produce a given amplitude above the point at which the pulse has a duration of 1/100 microsec. also increases. The only way of providing the larger output is by increasing the impedance of the line. This will aggravate the previously established difficulty of obtaining an artificial line satisfactorily approximating the properties of the uniformly distributed transmission line.

The conclusion must be drawn that a transmission line is not a desirable pulse forming network under the proposed conditions.

3.5.4 The Differentiating Circuit

On numerous occasions it has been suggested to the author that the circuit configuration Fig. 7 should provide a simple means of generating a short duration pulse if R_L is approximately equal to R_c and $(R_c + R_L)C$ is small with respect to T_r. The circuit now falls in the class of those of Fig. 4.

While commonly called a differentiating circuit the system, when a step function of voltage or current is applied, will produce a voltage response across the load of the form $Ve^{-\frac{t}{RC}}$. In this case if RC is chosen equal to T, the required pulse width, the amplitude of the pulse of that width will by inspection be

$$V_{pk} = \left\{ E_{bb} - E_{COND.}(t=0) \right\} \frac{R_L}{R_L + R_c} \left(1 - \frac{1}{\exp.}\right)$$

The response of the practical circuit Fig. 15(a) in which the interelectrode capacities are included cannot be so readily deduced. Recourse is made to the La Place transform equations of the system for this solution. Fig. 15(b) is an equivalent circuit for the analysis.

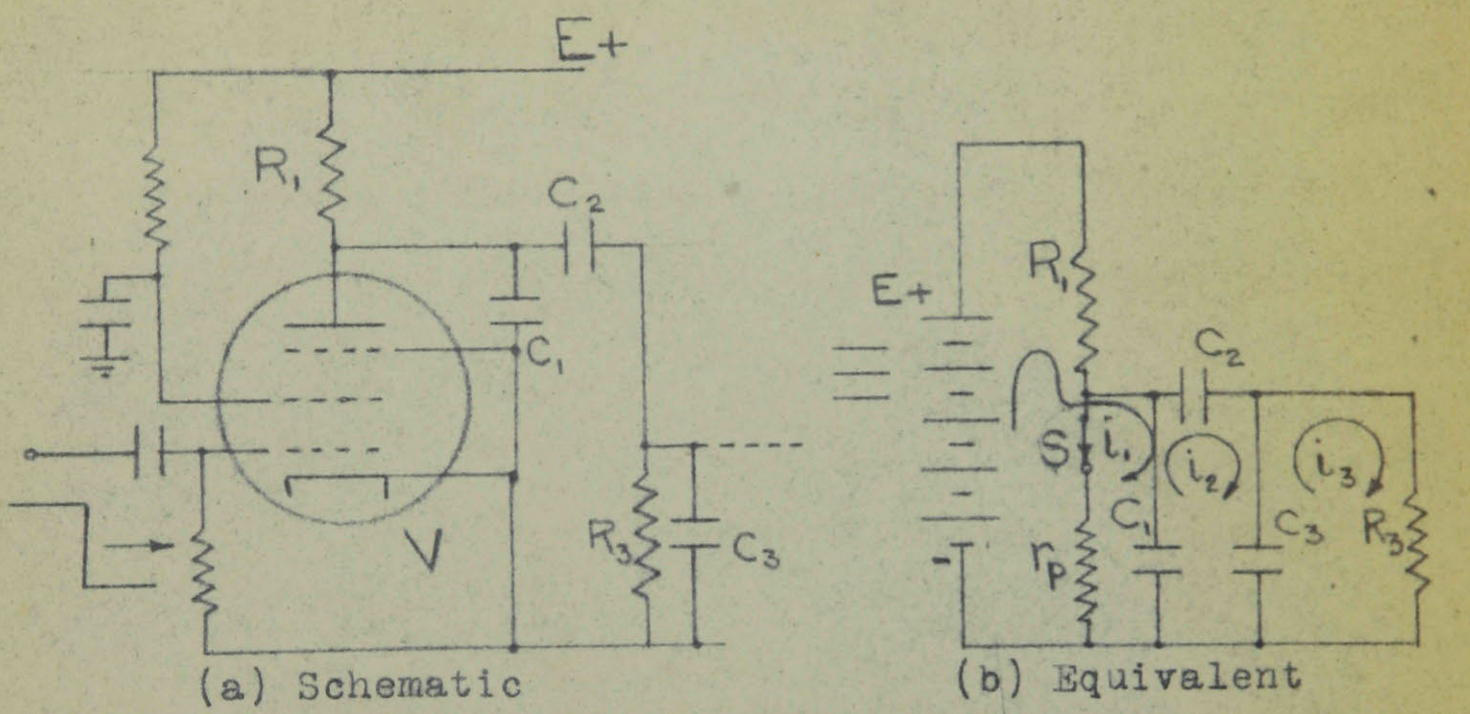


Fig. 15 The Practical Differentiating Circuit

The mesh equations are:

$$i_1 R_1 + \frac{1}{C_1}\int (i_1 - i_2)\,dt = E$$

$$\frac{1}{C_1}\int (i_2 - i_1)\,dt + \frac{1}{C_2}\int i_2\,dt + \frac{1}{C_3}\int (i_2 - i_3)\,dt = 0$$

$$\frac{1}{C_3}\int (i_3 - i_2)\,dt + i_3 R_3 = 0$$

Taking the La Place transforms of this system of equations gives:

$$I_1(s)\left\{R_1 + \frac{1}{C_1 s}\right\} - \frac{I_2(s)}{C_1 s} + \frac{E}{s}\frac{r_p}{(r_p + R_1)} = \frac{E}{s}$$

$$-\frac{I_1(s)}{C_1 s} + I_2(s)\left\{\frac{C_1C_2 + C_2C_2 + C_1C_3}{C_1C_2C_3 s}\right\} - \frac{I_3(s)}{C_3 s} = 0$$

$$-\frac{I_2(s)}{C_3 s} + I_3(s)\left\{R_3 + \frac{1}{C_3 s}\right\} = 0$$

So:

$$I_3(s) = \frac{\begin{vmatrix} R_1 + \frac{1}{C_1 s}, & -\frac{1}{C_1 s}, & \frac{E}{s}\frac{R_1}{R_1 + r_p} \\ -\frac{1}{C_1 s}, & \frac{1}{Cs}, & 0 \\ 0, & -\frac{1}{C_3 s}, & 0 \end{vmatrix}}{\begin{vmatrix} R_1 + \frac{1}{C_1 s}, & -\frac{1}{C_1 s}, & 0 \\ R_1, & \frac{1}{C_2 s}, & R_3 \\ 0, & -\frac{1}{C_3 s}, & R_3 + \frac{1}{C_3 s} \end{vmatrix}}$$

$$= \frac{ER_1C_2}{(R_1 + r_p)\alpha} \; \frac{1}{s^2 + \frac{\beta}{\alpha}s + \frac{1}{\alpha}}$$

Where:

$$\alpha = R_1R_3(C_1C_2 + C_2C_3 + C_1C_3)$$

$$\beta = (R_1C_1 + R_3C_3 + R_3C_2 + R_1C_2)$$

$$\frac{1}{C} = \frac{C_1C_2 + C_2C_3 + C_1C_3}{C_1C_2C_3} = \frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3}$$

$$\therefore\; i_3(t) = \frac{ER_1C_2}{(R_1 + r_p)\alpha\sqrt{\left(\frac{\beta}{\alpha}\right)^2 - \frac{4}{\alpha}}}\left\{e^{-\left(\frac{\beta}{\alpha} - \sqrt{\left(\frac{\beta}{\alpha}\right)^2 - \frac{4}{\alpha}}\right)\frac{t}{2}} - e^{-\left(\frac{\beta}{\alpha} + \sqrt{\left(\frac{\beta}{\alpha}\right)^2 - \frac{4}{\alpha}}\right)\frac{t}{2}}\right\}$$

And $E_{R_3}(t) = R_3\, i_3(t)$

This expression is too complex to obtain the amplitude above the level at which the curve has a prescribed pulse width. However, by assuming a set of values which might be met in practice a comparison with the ideal case can be made.

$E_{bb} = 300$ volts, $C_1 = C_3 = 5$ pf., $C_2 = 10$ pf., $R_1 = R_2 = r_p = 500$ ohm

In the ideal case neglecting C_1 and C_3:

$$T = RC = 10 \times 1000 \times 10^{-12} = 10^{-8} \text{ secs.}$$

$$V_{pk} = (300 - 150) \times \frac{500}{2 \times 500} \times 0.63 = 47 \text{ Volts}$$

With the ideal circuit a pulse of the desired characteristics would be easily realized.

In the practical case:

$$\alpha = .31 \times 10^{-16} \; ; \quad \beta = 15 \times 10^{-9} \; ; \quad \frac{\beta}{\alpha} = 48 \times 10^{7} \; ; \quad \frac{1}{\alpha} = 3.2 \times 10^{16}$$

$$\frac{ER_1C_2}{(R_1 + r_p)\alpha} = 48 \times 10^{6} \; ; \quad \sqrt{\left(\frac{\beta}{\alpha}\right)^2 - \frac{4}{\alpha}} = 3.2 \times 10^{8}$$

So

$$E_{R_3} = 75 \left\{ e^{-8 \times 10^{7} t} - e^{-40 \times 10^{7} t} \right\}$$

A plot of this function is shown as Fig. 16. From the graph it may be seen that the amplitude above the 1/100 microsecond width is only 13 volts. Consequently the differentiating circuit one meets in practice is of no value in the generation of a pulse of the short duration intended.

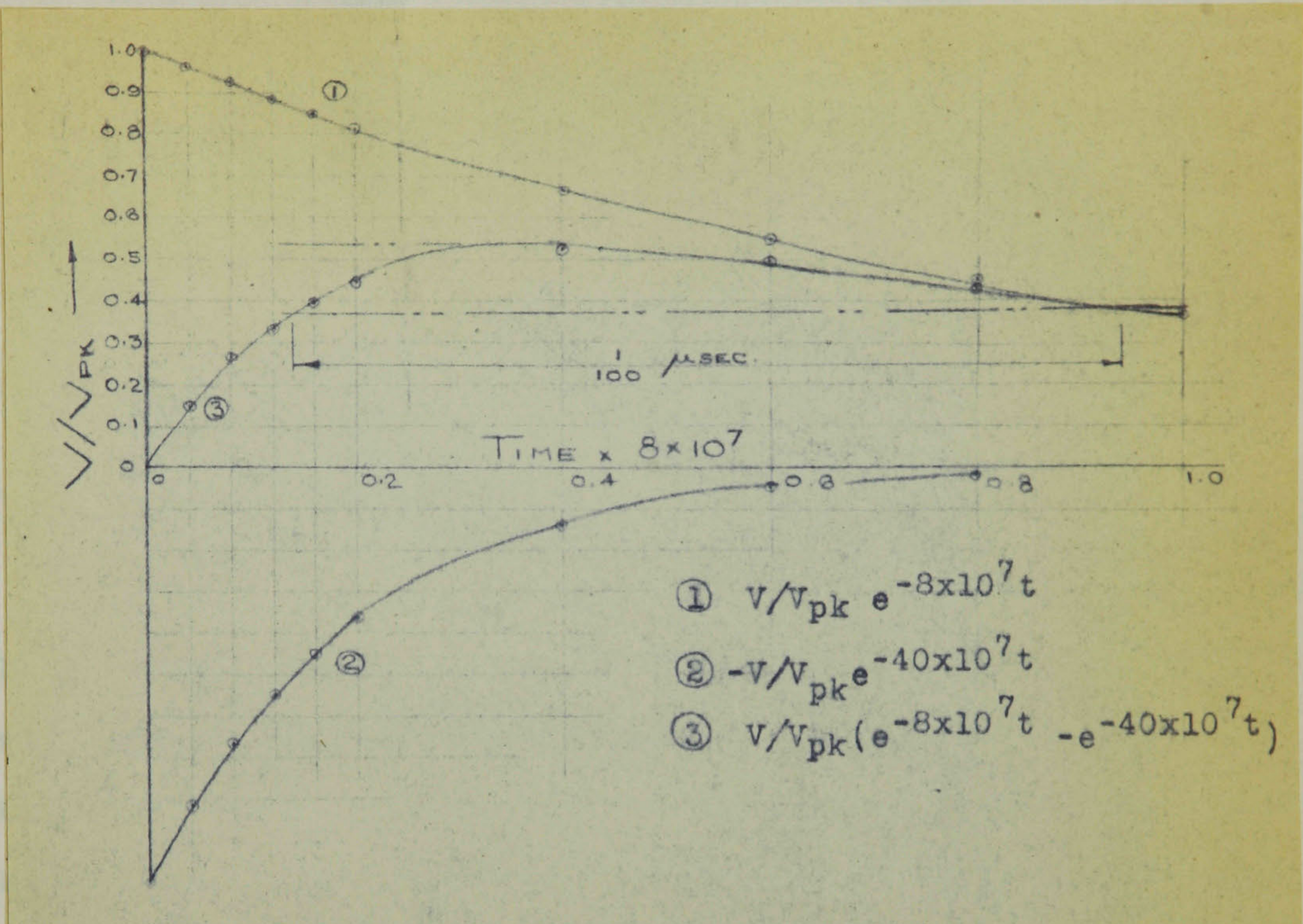


Fig.16 Output of Practical Differentiating Circuit.

3.5.5 <u>The Blocking Oscillator</u>:

The blocking oscillator (1)(2)(10)(12)(13) has been widely used as a source of fast rise time short duration pulses. The circuit, Fig. 17, is a transformer-coupled feedback oscillator in which plate current is allowed to flow for one-half cycle after which cut-off bias is imposed on the grid to prevent further oscillation.

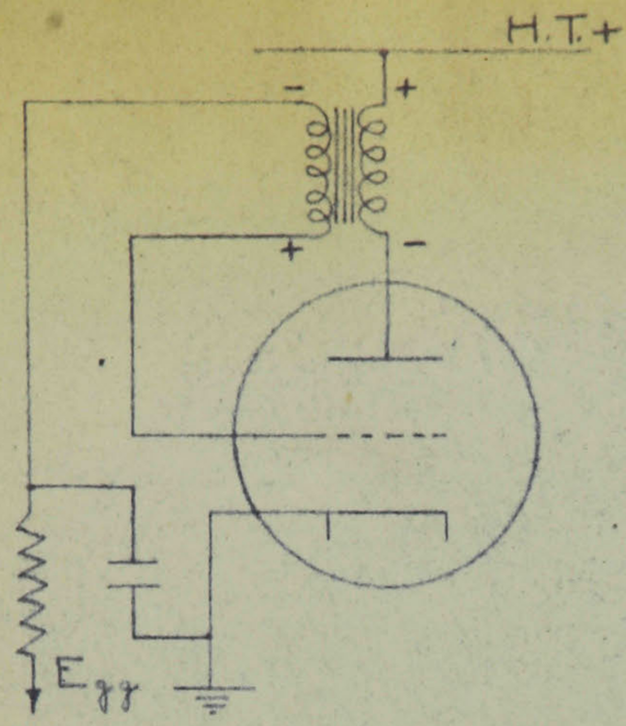


Fig. 17

The Blocking Oscillator.

(Plate-to- Grid Feed-back Type)

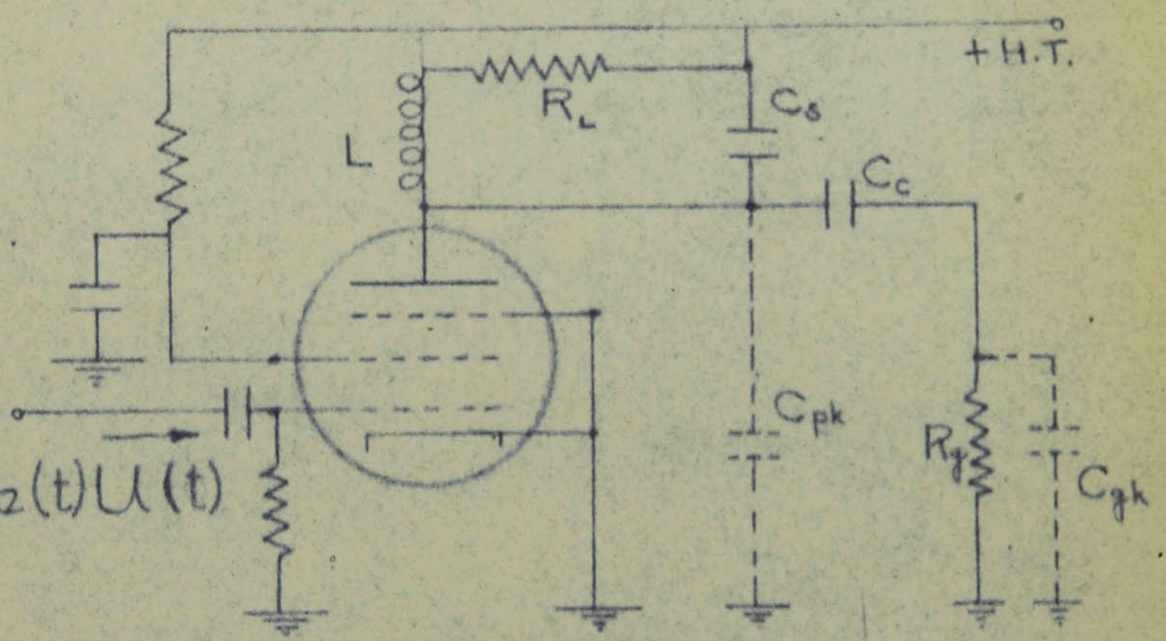
An exact analysis of the operation of this device has not been given. However, from the discussions to which reference was made a number of conclusions may be drawn. For proper operation the resonant period of the transformer and associated stray capacitance should be short with respect to the duration of the pulse desired. The actual duration of the pulse is determined by the low-frequency response of the transformer although a delay line may be incorporated in the circuit to reduce that period.

The transformers employed here are of the same class as those referred to in Sec. 3.2.3. The most satisfactory of them are suitable in blocking oscillator circuits for the generation of pulses, as in the former application, of a minimum duration of 0.1 microsec. Therefore at the present state of the art this method is not adaptable to the present problem.

3.5.6

<u>The Shock-Excited Tuned Circuit</u>

A parallel combination of inductance and resistance, and capacitance shown in the plate circuit of Fig. 18 will produce a damped sinusoidal oscillation when acted on by a suddenly applied current or voltage. The output signal can be determined by solution of the integro-differential equations of the network.



e(t) An arbitrary shape input wave
U(t) = 0 t<0
= 1 t≥0

Fig. 18 Circuit for Shock-Excitation of a Parallel Tuned (LC) Circuit.

For the simple case of the valve conducting a constant current I for t<o being cut off at t = o the response (14) is given by

$$E(t) = I\sqrt{\frac{L}{C}}\, e^{-\frac{R}{2L}t} \sin \omega t \qquad \text{where } \omega \doteq \frac{1}{\sqrt{LC}}$$

This output may be very nearly doubled by establishing a current I in the valve at time t = o and then cutting off the valve when the oscillatory signal produced across the tuned circuit reaches the first potential minimum. This corresponds to the case of a system having an initial current I through the inductance and an initial charge on the condensor. Thus the peak voltage realizable using a positive going pulse input is $E_{pk} \doteq 2I\sqrt{\frac{L}{C}} = \frac{I}{\pi f C}$ where $f = \frac{1}{2\pi\sqrt{LC}}$

Now for a 6AN5 as the switch in a circuit in which the input and output capacities of the valves associated with the tuned circuit plus wiring represent the total capacitance, the following values may be reasonably assumed.

I = 100 ma. C = 15 pf. f = 33.3 mc/sec.

whence $E_{pk} = \frac{10^{-1}}{\pi \times (33.3\times10^{6})\times(15\times10^{-12})} = 64$ volts.

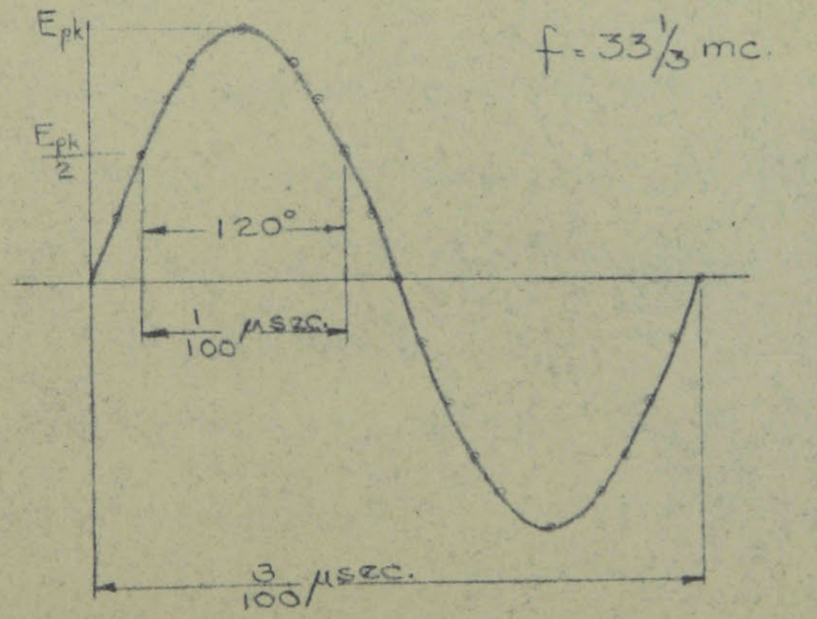
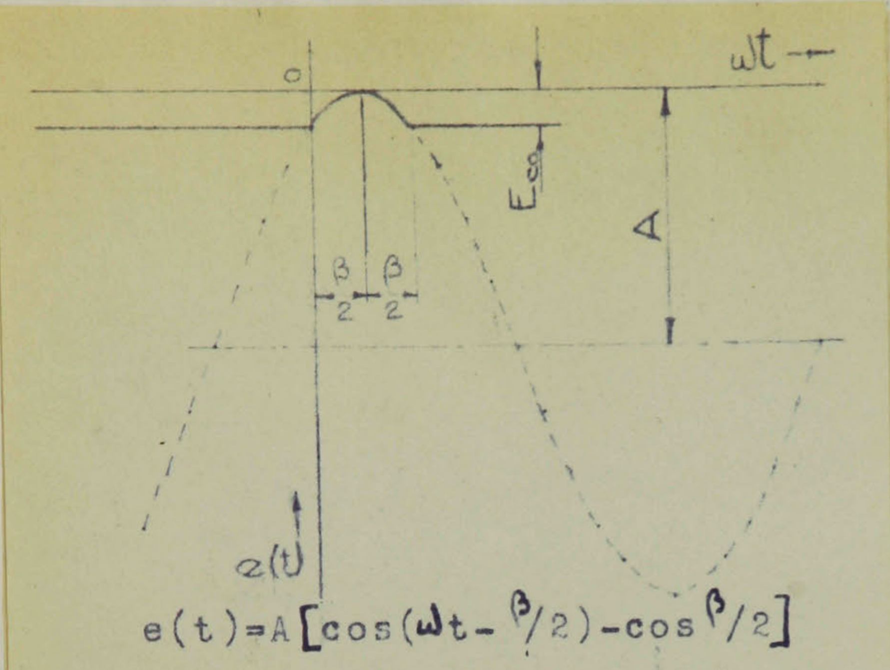


FIG. 19

From the diagram Fig. 19 it is seen that the amplitude of that portion of a 33 1/3 mc/sec. sine wave having a base width of 1/100 microsec. is $E_{pk}/2$ = 32 volts. With this circuit it is possible then to realize the prescribed conditions for the pulse if positive alternations other than the first one are suppressed.

An alternative method of shocking the circuit would be to use the peak of a sinusoidally varying voltage as the grid driving function. Such a signal is depicted in Fig. 20. It can be effectively realized when the time constant of the



$e(t) = A\left[\cos(\omega t - \beta/2) - \cos \beta/2\right]$

Fig. 20

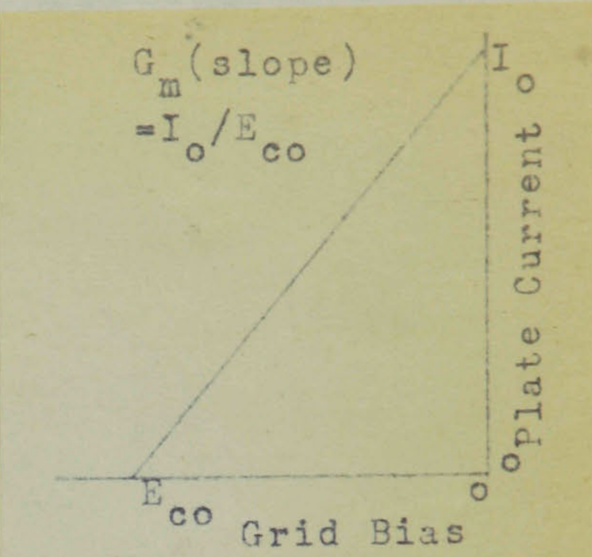


Fig. 21

Ideal $I_p - E_g$ Curve

resistance and capacitance in the grid circuit of the valve, Fig. 18, is large with respect to the period of the sine wave input. If the initial current and charge in the circuit are assumed to be zero the response to a single current pulse of the shape described can be readily obtained from the relationship I(s) = V(s) Y(s) where:

V(s) is the La Place transform of the output voltage desired.

Y(s) is the admittance operator for the plate load, of the pulse generator, considered as a single node-pair. In the case of Fig. 18 with $G = G_V + 1/R$ and $C = C_S + C_{pk} + C_{gk}$.

$$Y(s) = G + Cs + 1/Ls + R$$

$$= \frac{LC\left\{ s^2 + \frac{GL+RC}{LC} s + \frac{1+RG}{LC} \right\}}{Ls + R}$$

$$= \frac{LC\left\{ s - (-\alpha + j\theta)\right\}\left\{ s - (-\alpha - j\theta)\right\}}{Ls + R} \qquad \underline{9}$$

where

$$-\alpha \pm j\theta = -\frac{GL+RC}{2LC} \pm \sqrt{\left(\frac{GL+RC}{2LC}\right)^2 - \left(\frac{1+RG}{LC}\right)}$$

and $I(s) = L \; -gm\, e(t)$ is the La Place transform of the current pulse in the valve, for the $I_p - E_g$ curve Fig. 21

$$= -gmA\, L\left[\cos(\omega t - \beta/2) - \cos \beta/2\right]$$

Now: $L\left[\cos(\omega t - \beta/2) - \cos \beta/2\right]$

$$= \int_0^{\beta/\omega} \frac{e^{j(\omega t - \beta/2)} + e^{-j(\omega t - \beta/2)}}{2} \cdot e^{-st}\, dt - \cos\frac{\beta}{2}\int_0^{\beta/\omega} e^{-st}\, dt$$

$$+ \int_{\beta/\omega}^{\infty} 0 \cdot e^{-st}\, dt$$

$$= \int_0^{\beta/\omega} \frac{e^{-(s-j\omega)t - j\frac{\beta}{2}} + e^{-(s+j\omega)t + j\frac{\beta}{2}}}{2}\, dt + \cos\frac{\beta}{2}\left[\frac{e^{-\frac{\beta}{\omega}s} - 1}{s}\right]$$

$$= \left[\frac{e^{-(s-j\omega)t - j\frac{\beta}{2}}}{-2(s-j\omega)} + \frac{e^{-(s+j\omega)t + j\frac{\beta}{2}}}{-2(s+j\omega)}\right]_0^{\beta/\omega} + \cos\frac{\beta}{2}\left[\frac{e^{-\frac{\beta}{\omega}s} - 1}{s}\right]$$

$$= -\left[\frac{e^{-\frac{\beta}{\omega}s + j\frac{\beta}{2}} - e^{-j\frac{\beta}{2}}}{2(s-j\omega)} + \frac{e^{-\frac{\beta}{\omega}s - j\frac{\beta}{2}} - e^{+j\frac{\beta}{2}}}{2(s+j\omega)}\right] + \cos\frac{\beta}{2}\left[\frac{e^{-\frac{\beta}{\omega}s} - 1}{s}\right]$$

$$= -\left[\frac{-s\left\{\left(e^{-j\frac{\beta}{2}} + e^{j\frac{\beta}{2}}\right) - e^{-\frac{\beta}{\omega}s}\left(e^{j\frac{\beta}{2}} + e^{-j\frac{\beta}{2}}\right)\right\} + j\omega\left\{\left(e^{j\frac{\beta}{2}} - e^{-j\frac{\beta}{2}}\right) + e^{-\frac{\beta}{\omega}s}\left(e^{j\frac{\beta}{2}} - e^{-j\frac{\beta}{2}}\right)\right\}}{2(s^2 + \omega^2)}\right]$$

$$+ \cos\frac{\beta}{2}\left[\frac{e^{-\frac{\beta}{\omega}s} - 1}{s}\right]$$

$$= \frac{s\cos\beta/2\,(1-e^{-\frac{\beta}{\omega}s}) + \omega\sin\beta/2\,(e^{-\frac{\beta}{\omega}s}+1)}{(s^2+\omega^2)} - \frac{\cos\beta/2\,(1-e^{-\frac{\beta}{\omega}s})}{s}$$

Combining Terms and Multiplying by $-G_mA$ Gives:

$$I(s) = G_mA\left[\cos\frac{\beta}{2}(1-e^{-\frac{\beta}{\omega}s})\frac{\omega^2}{s(s^2+\omega^2)} - \sin\frac{\beta}{2}(1+e^{-\frac{\beta}{\omega}s})\frac{\omega}{s^2+\omega^2}\right] \qquad \underline{10}$$

Substituting $\underline{9}$ and $\underline{10}$ in $I(s) = V(s)Y(s)$ Gives:

$$V(s) = \frac{G_mA}{LC}\left[\cos\frac{\beta}{2}(1-e^{-\frac{\beta}{\omega}s})\frac{\omega^2\quad(Ls+R)}{s(s-j\omega)(s+j\omega)[s-(-\alpha+j\theta)][s-(-\alpha-j\theta)]}\right.$$

$$\left. - \sin\frac{\beta}{2}(1+e^{-\frac{\beta}{\omega}s})\frac{\omega\quad(Ls+R)}{(s-j\omega)(s+j\omega)[s-(-\alpha+j\theta)][s-(-\alpha-j\theta)]}\right]$$

Expanding in Partial Fractions Yields:

$$\frac{V(s)LC}{G_mA} = \cos\frac{\beta}{2}(1-e^{-\frac{\beta}{\omega}s})\left[\frac{R}{s(\alpha^2+\theta^2)} + \frac{K_1\angle\gamma_1}{2(s-j\omega)}\right.$$

$$\left. + \frac{K_1\angle-\gamma_1}{2(s+j\omega)} + \frac{K_2\angle\gamma_2}{2[s-(-\alpha+j\theta)]} + \frac{K_2\angle-\gamma_2}{2[s-(-\alpha-j\theta)]}\right]$$

$$- \sin\frac{\beta}{2}(1+e^{-\frac{\beta}{\omega}s})\left[\frac{K_1\angle\gamma_1+90^\circ}{2(s-j\omega)} + \frac{K_1\angle-\gamma_1-90^\circ}{2(s+j\omega)}\right.$$

$$\left. + \frac{K_3\angle\gamma_3}{2[s-(-\alpha+j\theta)]} + \frac{K_3\angle-\gamma_3}{2[s-(-\alpha-j\theta)]}\right] \qquad \underline{11}$$

Where:

$$K_1 \angle \gamma_1 = \frac{-(R + jL\omega)}{[\alpha + j(\theta + \omega)][\alpha - j(\theta - \omega)]}$$

$$K_2 \angle \gamma_2 = \frac{\omega^2 [L(-\alpha + j\theta) + R]}{(-\alpha + j\theta)[-\alpha + j(\theta - \omega)][-\alpha + j(\theta + \omega)](j\theta)}$$

$$K_3 \angle \gamma_3 = \frac{(-\alpha + j\theta)}{\omega} K_2 \angle \gamma_2$$

The L^{-1} Transform of Equation 11 Gives the Desired Voltage Response:

$$v(t) = \frac{GmA}{LC} \Bigg| \Bigg[\cos \frac{\beta}{2} \Big\{ \frac{R}{\alpha^2 + \theta^2} + K_1 \cos(\omega t + \gamma_1) + K_2 e^{-\alpha t} \cos(\theta t + \gamma_2) \Big\}$$

$$- \sin \frac{\beta}{2} \Big\{ K_1 \cos(\omega t + \gamma_1 + 90^\circ) + K_3 e^{-\alpha t} \cos(\theta t + \gamma_3) \Big\} \Bigg] U(t)$$

$$- \Bigg[\cos \frac{\beta}{2} \Big\{ \frac{R}{\alpha^2 + \theta^2} + K_1 \cos(\omega t + \gamma_1 - \beta) + K_2 e^{-\alpha(t - \frac{\beta}{\omega})} \cos(\theta t + \gamma_2 - \frac{\beta\theta}{\omega}) \Big\}$$

$$+ \sin \frac{\beta}{2} \Big\{ K_1 \cos(\omega t + \gamma_1 + 90^\circ - \beta) + K_3 e^{-\alpha(t - \frac{\beta}{\omega})} \cos(\theta t + \gamma_3 - \frac{\beta\theta}{\omega}) \Big\} \Bigg] U(t - \frac{\beta}{\omega}) \Bigg| \qquad \underline{12}$$

Where:

$U(t) = 0 \quad t < 0$ And $U(t - \frac{\beta}{\omega}) = 0 \quad t < \frac{\beta}{\omega}$

$\quad\;\; = 1 \quad t \geq 0 \qquad\qquad\qquad\qquad = 1 \quad t \geq \frac{\beta}{\omega}$

Expression 12 is of course far too cumbersome to be of much practical value. If a number of approximations, valid in practice, are made considerable simplification will be realized.

It will then be possible to extract from the equation those terms which determine the peak amplitude obtainable at the natural frequency of the tuned circuit.

For $Q > 50$, $R_g \gg QL\theta$ and $\theta < 10\omega$ it can be shown that $\alpha \ll \theta$ and $R \ll L\omega$. Hence

$$K_1 \angle \gamma_1 \doteq \frac{L\omega \angle -90^\circ}{(\theta+\omega)(\theta-\omega)}$$

$$K_2 \angle \gamma_2 \doteq \frac{L\omega^2 \angle 90^\circ}{\theta(\theta+\omega)(\theta-\omega)} = \frac{\omega}{\theta} K_1 \angle 90^\circ$$

$$K_3 \angle \gamma_3 \doteq \frac{L\omega \angle 180^\circ}{(\theta-\omega)(\theta+\omega)} = K_1 \angle 180^\circ$$

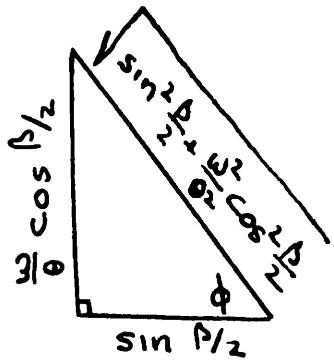
By substituting these values in Eq. 12, and combining terms wherever possible, the expression for the transient response is reduced to

$$v(t)_{0<t<\frac{\beta}{\omega}} = \frac{G_m A}{LC}\left[\cos\frac{\beta}{2}\frac{R}{\theta^2} + K_1 e^{-\alpha t}\left\{\sin\frac{\beta}{2}\cos\theta t - \frac{\omega}{\theta}\cos\frac{\beta}{2}\sin\theta t\right\} + K_1\sin\left(\omega t - \beta/2\right)\right]$$

$$= \frac{G_m A}{LC}\left[\cos\frac{\beta}{2}\frac{R}{\theta^2} + K_1 e^{-\alpha t}\sqrt{\sin^2\frac{\beta}{2} + \frac{\omega^2}{\theta^2}\cos^2\frac{\beta}{2}}\left\{\frac{\sin\frac{\beta}{2}\cos\theta t}{\sqrt{\sin^2\frac{\beta}{2}+\frac{\omega^2}{\theta^2}\cos^2\frac{\beta}{2}}} - \frac{\frac{\omega}{\theta}\cos\frac{\beta}{2}\sin\theta t}{\sqrt{\sin^2\frac{\beta}{2}+\frac{\omega^2}{\theta^2}\cos^2\frac{\beta}{2}}}\right\} + K_1\sin\left(\omega t - \frac{\beta}{2}\right)\right]$$

$$= \frac{G_m A}{LC}\left[\cos\frac{\beta}{2}\frac{R}{\theta^2} + K_1\left\{e^{-\alpha t}\sqrt{\sin^2\frac{\beta}{2}+\frac{\omega^2}{\theta^2}\cos^2\frac{\beta}{2}}\;\cos(\theta t + \phi) + \sin\left(\omega t - \frac{\beta}{2}\right)\right\}\right] \quad \underline{13}$$

Where $\phi = \tan^{-1}\dfrac{\omega}{\theta}\dfrac{\cos\beta/2}{\sin\beta/2}$



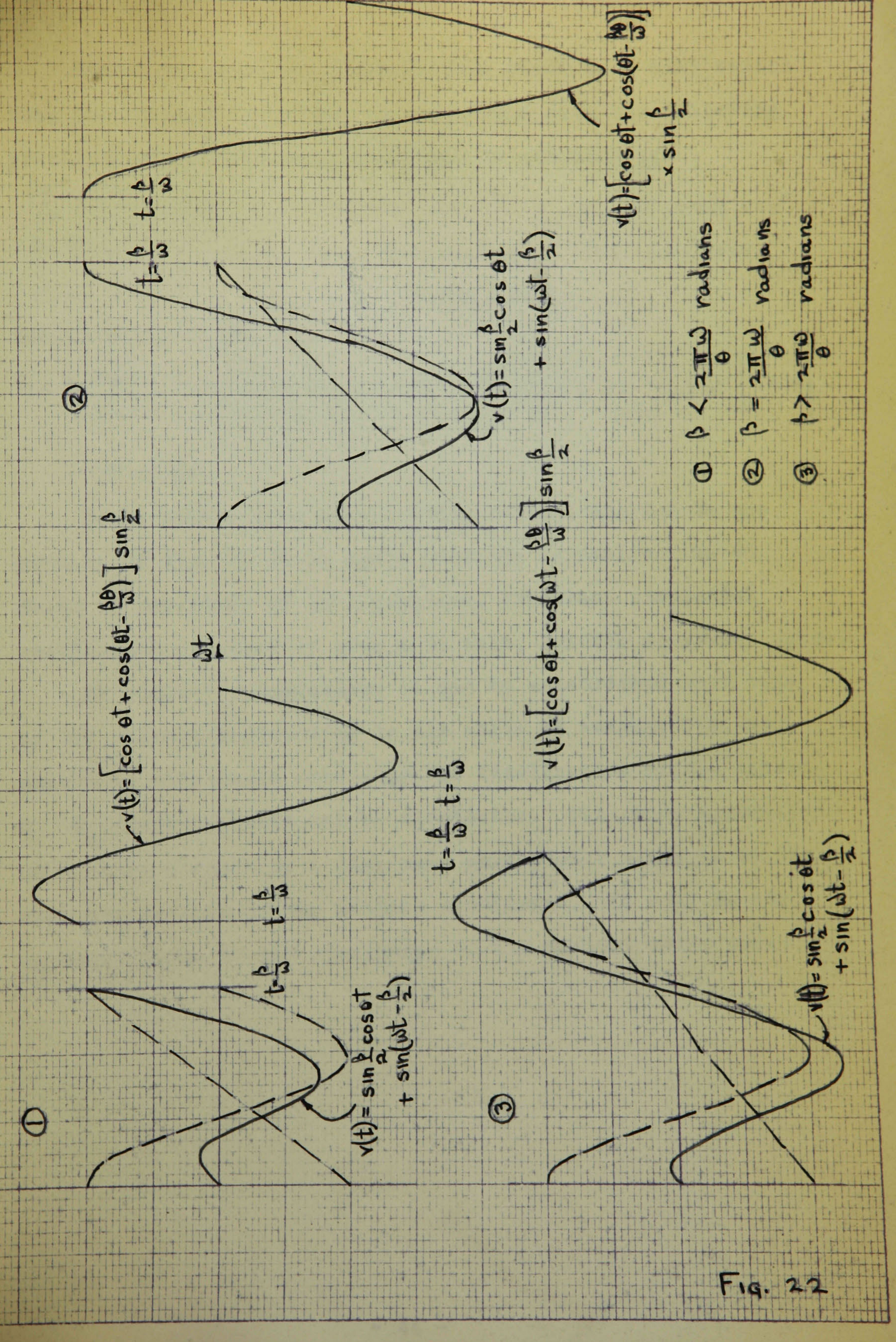
Using the same procedure it has been established that the output for $\frac{\beta}{\omega} \leq t < \infty$ is given by

$$v(t)_{\frac{\beta}{\omega} \leq t < \infty} = \frac{G_m A K_1 e^{-\alpha t}}{LC} \sqrt{\sin^2\frac{\beta}{2} + \frac{\omega^2}{\theta^2}\cos^2\frac{\beta}{2}} \left\{ \cos(\theta t + \phi) + e^{\frac{\alpha \beta}{\omega}} \cos\left(\theta t - \phi - \frac{\beta\theta}{\omega}\right) \right\} \quad \underline{14}$$

Since it is the absolute peak amplitude in which one is interested the problem remains of discovering which of the relationships 13 and 14 provides that information. The accomplishment of this mathematically leads to hopelessly involved trigonometrical expressions. A graphical consideration of the equations yields a satisfactory solution.

If it is assumed for convenience in plotting that $\frac{\omega^2}{\theta^2}\cos^2\frac{\beta}{2} \ll \sin^2\frac{\beta}{2}$ then the nature of the wave may be examined, Fig. 22, for conduction angles β greater than, equal to, and less than one cycle of the resonant frequency of the plate load-impedance. Thus it is seen that the peak amplitude is determined respectively by $v(t)_{0<t<\frac{\beta}{\omega}}$, $v(t)_{0<t<\frac{\beta}{\omega}}$ or $v(t)_{\frac{\beta}{\omega}<t<\infty}$ (a necessary condition for voltage continuity at this point), and for $v(t)_{\frac{\beta}{\omega}<t<\infty}$.

Furthermore the maximum peak amplitude in this much simplified approach is that for the conduction angle equal to one cycle of the transient oscillation. If the accurate expression 14 is used this maximum occurs for $\frac{\beta\theta}{\omega} + 2\phi = 360^\circ$ which is for a conduction angle a little less than one cycle.



①
v(t) = sin β/2 cos θt + sin(ωt − β/2)
t=β/ω
t=β/ω
v(t)=[cos θt + cos(θt − βθ/ω)] sin β/2
ωt
②
v(t) = sin β/2 cos θt + sin(ωt − β/2)
t=β/ω
t=β/ω
v(t)=[cos θt + cos(θt − βθ/ω)] x sin β/2
③
v(t) = sin β/2 cos θt + sin(ωt − β/2)
t=β/ω
t=β/ω
v(t)=[cos θt + cos(ωt − βθ/ω)] sin β/2
① β < 2πω/θ radians
② β = 2πω/θ radians
③ β > 2πω/θ radians

FIG. 22

Resorting again to a calculation based on a 6AN5 switch tube the maximum amplitude is computable with the following practical values for the parameters in equation 14:

$\frac{p\theta}{\omega} + 2\phi = 360^\circ$; $\omega = 2\pi \times 6 \times 10^6$; $\theta = 2\pi \times 33\frac{1}{3} \times 10^6$

$p \doteq 60^\circ$; $G_m = 8000\,\mu$mhos ; $A = 200$ Volts

$C = 25$ pf ; $L = 0.92\,\mu$henries.

Substituting in Eq. 14:

$$V_{pk} = \frac{8000 \times 10^{-6} \times 200 \times L \times 2\pi \times 6 \times 10^6}{L \times 25 \times 10^{-12} \times (33\frac{1}{3} - 6)(33\frac{1}{3} + 6) \times 4\pi^2 \times 10^{12}} \sqrt{\sin^2 30^\circ + \left(\frac{6}{33\frac{1}{3}}\right)^2 \cos^2 30^\circ} \left[1 + 1\right]$$

$= 60$ volts.

Hence with a sine wave driving voltage applied to the circuit of Fig. 18 it is again possible to realize the 60 volts of amplitude of 33 1/3 mc/sec. transient sinusoidal oscillation required to give a 30 volt, 1/100th microsec. pulse.

By similar calculations it can be established that each of the signals represented in the chain of stages, Fig. 23, should be attainable providing the idealization of the valves $I_p - E_g$ characteristic can be vindicated. The shock-excitation of a tuned circuit by a sine-wave grid signal seems to be the most practicable of the methods considered. It will be the approach taken to the attainment of the desired short pulse.

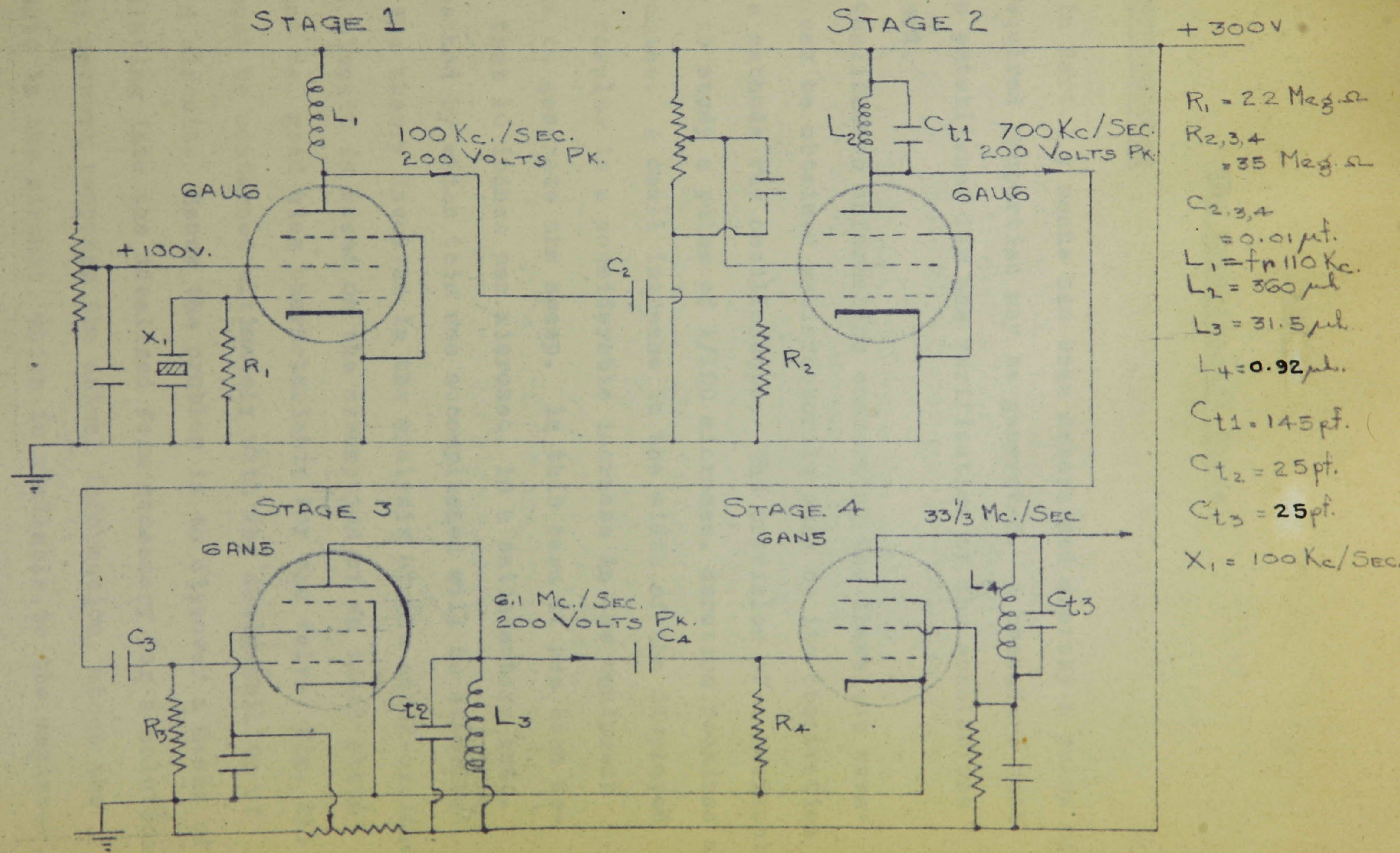


FIG. 23: PULSE GENERATOR CHAIN

PART II

The Experimental Phase

4. INTRODUCTION

In Part I a means has been determined whereby a pulse of the required properties may be generated. A procedure must now be established for the verification of the conclusions obtained.

Qualitative information concerning the transient wave-shape can be obtained satisfactorily only by its observation with a cathode ray oscilloscope. The provision of a horizontal sweep to study a pulse of 1/100 microsec. duration requires a compromise. A small increase in the width of the displayed pulse results in a considerable increase in the equipment needed to generate the sweep. In this case it has been decided that 10 inches per microsec. is a satisfactory rate. The method by which this was accomplished will be described.

The theory involved in the analysis of the shock-excited tuned circuit is based on the assumption of an ideal plate current vs. grid bias characteristic for the valve, Fig. 21. One must be concerned ultimately with the designability of such a circuit. Hence the problem is to discover a means of translating into the idealized form necessary for a solution of the network response the actual information, about the elements in the circuit, which is available to the engineer.

In this case the data provided are the measured $I_p - E_g$ curve for a valve and the constants of the linear elements, inductance, capacitance and resistance, incorporated in the circuit of Fig. 18. By a process of judicious guessing various ideal $I_p - E_g$ curves related to the measured curve in a reproduceable manner can be obtained. Based on the parameters G_m and E_{co} of each of these ideal curves the peak output voltage of the tuned circuit can be calculated in terms of the amplitude of the sine wave applied to the valve's grid.

Using the valve and elements employed for the calculations, the transient voltage across the tuned circuit can be measured by a vacuum-tube voltmeter for the same range of input signals. Comparisons of the measured and the various calculated outputs must then be made. A reasonable correlation between the measured and a given calculated response may be taken to confirm the suitability of the process by which the idealized $I_p - E_g$ curve, used for that calculation, was derived.

5. THE CATHODE-RAY OSCILLOSCOPE

5.1 The Tube and Power Supply

No commercial oscilloscope, to which the author had access, provides the facilities outlined in the introduction. The REL Type 106A High Speed Oscilloscope provides a basic unit which with the addition of an external sweep generator and beam-intensity gate permits the necessary observations.

Actually use was made of the aforementioned oscilloscope to avoid unnecessary construction. The power supply is used as originally designed but of the facilities for the display and adjustment of the pattern only the cathode-ray tube, and the focussing, brilliancy and the vertical centring systems are employed.

The vertical deflection sensitivity of the cathode-ray tube 5GP1 with the electrode potentials employed is 37 volts/inch. Hence for reasonable observation of a signal having 30 volts amplitude no amplification is necessary. The leads to the vertical deflection plates have therefore been brought directly to a terminal board on the rear end of the oscilloscope chassis in order to keep to a minimum the capacitive loading on the circuit under study.

The cut-off potential of the 5GP1 control grid is approximately 35 volts. Therefore a beam intensifying pulse having a magnitude of nearly 50 volts must be provided. Direct connection to the terminal board is again made to reduce capacity and so decrease the rise time of the pulse.

The horizontal deflection sensitivity of the tube is 75 volts/inch. Consequently for the required sweep speed of 10 inches/microsec., balanced to ground signals having a linear rise of 375 volts/microsec., must be applied to the horizontal deflection plates. While not required because of circuit considerations, connections are again made through the rear terminal board.

5.2 The Horizontal Sweep Generator

The recurrence period of the horizontal sweep must of course be that of the pulse under observation 10 microsec. Now if the centre of the tube is taken as the origin for horizontal displacement and sinusoidal voltage

$$e(t) = A \sin 2\pi f t = A \sin (360^\circ \times 10^5 \times t) \qquad f = 100\text{ kc/sec.}$$

is applied to the right-hand horizontal deflecting plate and

$$e(t) = A \sin(2\pi f t + 180^\circ) = A \sin (360^\circ \times 10^5 \times t + 180^\circ)$$

to the other plate then the displacement x(in.), from the centre of a 5GP1 having the stated 75 volts/in. deflection sensitivity, will be

$$x(\text{in.}) = \frac{2A}{75} \sin 360 \times 10^5 \times t$$

Hence for full screen deflection of 2 1/2 inches in 1/4 microsec. the required amplitude of the sinusoidal signals is

$$A = \frac{75 \times 2\,1/2}{2 \sin 360 \times 10^5 \times 1/4 \times 10^{-6}} = 600\text{ Volts}$$

and the angle of these sine waves through which they are sweeping the beam across the tube face is -9° to $+9^\circ$ and 171° to 189° respectively.

The difference between the value of y = sin x and that of a tangent to the sine wave at the point x = 0 is given by

$$\Delta x = x - \left(x - \frac{x^3}{3!} + \frac{x^5}{5!} - + -\right) \doteq \frac{x^3}{3!}$$

so the % error in linearity $\Delta x \% = x^2 \times 16.6\%$

Over the range required above this error amounts to

$$\Delta x \% = \left(\frac{9 \times 2\pi}{360}\right)^2 \times 16.6 = 0.41\%$$

which is negligible for the purpose intended for the sweep.

Since all the tubes in the proposed pulse generator, Fig. 23, are operated as Class C amplifiers and furthermore since the conduction angles are small and the first positive peak at each plate occurs at the termination of the conduction period, the relative phase of the 100 kc/sec. driving wave and the final 33 mc/sec. transient in Fig. 24(a) and (b) respectively is approximately correct. The waveshapes Fig. 24 (c)(d) and (e) then represent the shape and relative phase of the signals applied to the two horizontal deflection plates and to the beam intensifying grid respectively for observation of the transient (b) applied to the vertical deflection system.

The horizontal sweep generator must provide these facilities. Furthermore, since the relative phase of the final transient output and the 100 kc/sec. oscillator of the pulse generator will vary as the conduction angle of any stage is varied, provision must be made to shift the phase of signals (c)(d) and (e) equally with respect to (a) about the mean value represented.

The circuit of Fig. 25 is that of a sweep generator constructed by the author allowing these operations. Valve V_1

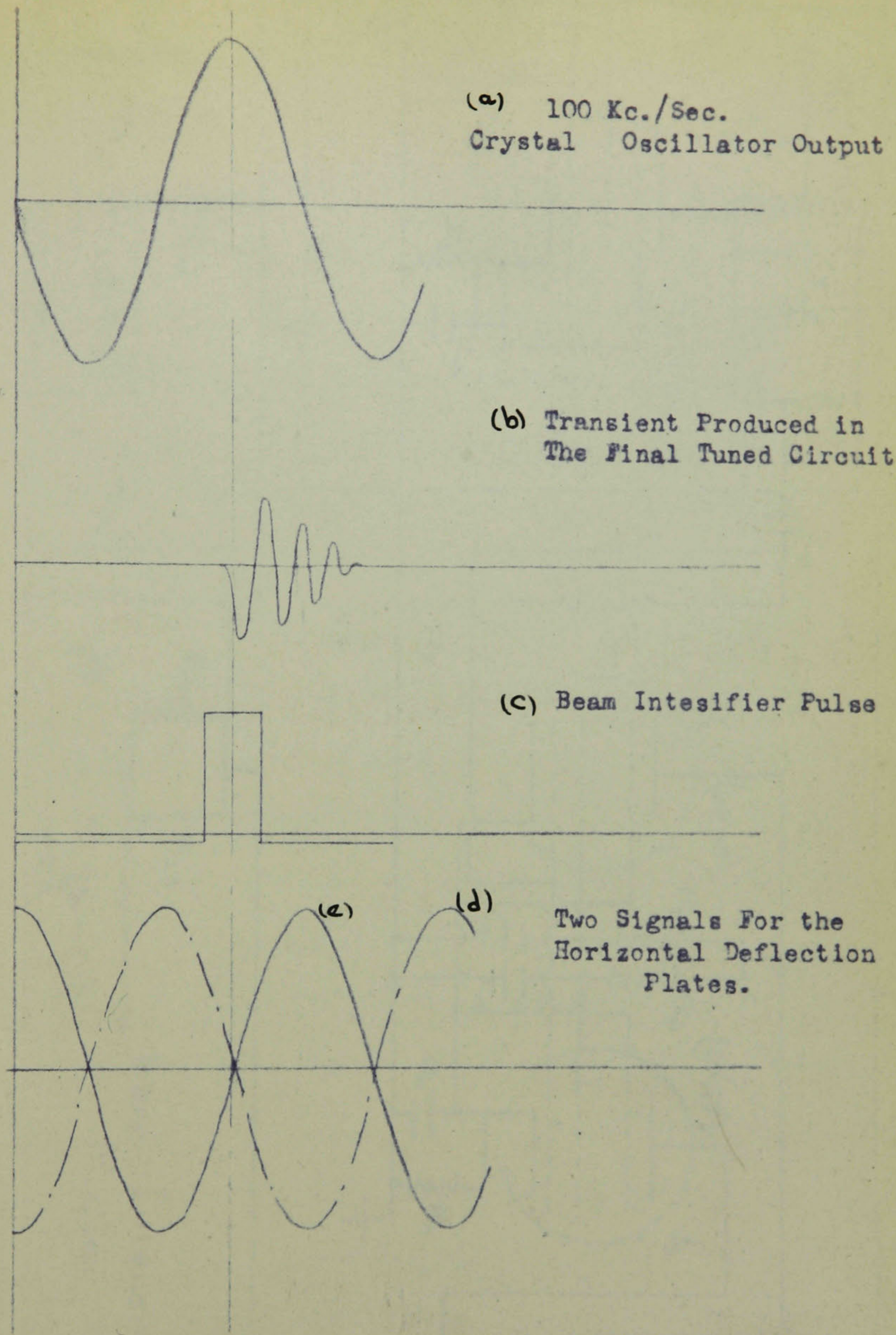
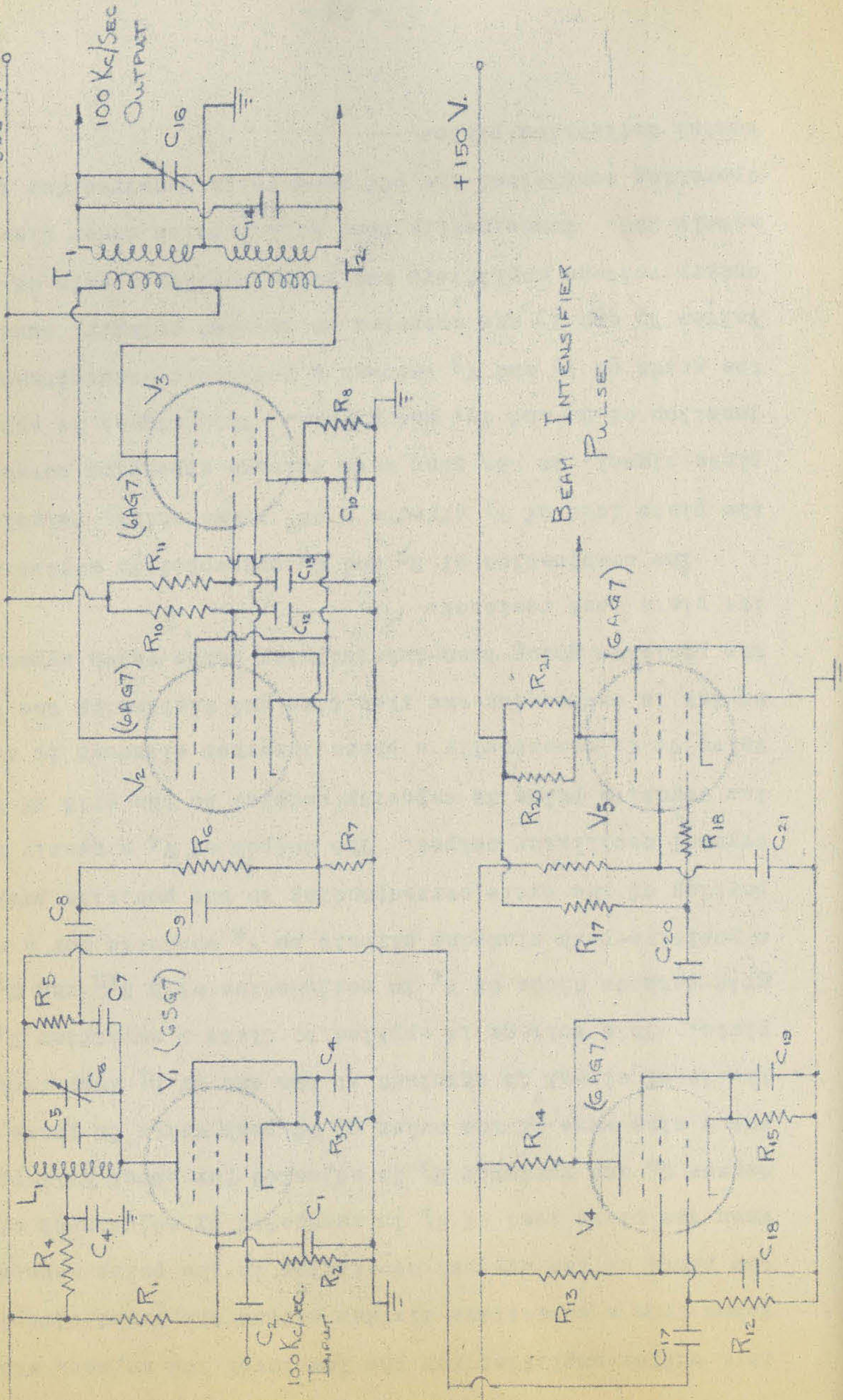


Fig.24 : Phase Relationship of the Signals Required From The Horizontal Sweep Generator.



) pf.
pf.
fd. paper
) pf.
pf.
pf.
pf.
pf.
f.
mf. paper
21,18
mf. paper
pf.
pf. variable.
pf.
mf. paper
) pf.

1 watt
K ½ watt
2 watt
ume control
K 2 watt
½ watt
K ½ watt
½ watt
Ohms ½ watt
24K 2 watt
Meg. ½ watt
2 watt
2 watt
ohms ½ watt
½ watt
½ watt
2 watt
.7 K 2 watt

imary L 19 mh.
condary L 10 mh.
tual L 4 mh.

is a buffer amplifier for the low level 100 kc/sec. sine wave taken from a capacitive divider across the tuned circuit at the plate of the crystal oscillator in the pulse generator. When the plate load of V_1 is resonated by adjustment of condensor C_6 and resistor R_3 is adjusted for about half-peak output a sine wave of the order of 60 peak volts in phase with the input signal is obtained at the end of L_1 away from the plate. This voltage is applied to Class C amplifier V_4. The grid-cathode diode of V_4 in conjunction with R_{12} and C_{17} form a positive-peak clamping circuit so V_4 conducts for a small portion of the cycle corresponding to the positive peak of the crystal oscillator output. The output of V_4 a nearly rectangular negative pulse is capacity coupled to the grid of V_5. Valve V_5 is essentially a phase inverter although it does assist in decreasing the rise time and flattening the top of the positive going beam-intensifying pulse which appears across its plate load resistor.

The combination of R_5 and C_7 connected in series across the plate load of V_1 effects a 90° phase shift, relative to the plate signal, in the sine wave voltage appearing between the junction of R_5 and C_7, and ground. This signal is applied to the grids of V_2 and V_3 through a resistance-capacitance divider. Valves V_2 and V_3 are operated as untuned primary, tuned secondary voltage amplifiers and each provides a gain of approximately 220. Consequently they will furnish under Class A operating conditions the 600 peak volts required for the horizontal deflection plates.

It was initially assumed that C_6 was adjusted for resonance of the tank circuit. If it is used to detune the plate load of V_1 the relative phase of the grid signals of V_2, V_3 and V_4 will be unaltered but their phase with respect to the input will be varied. However, the magnitude of these grid signals will be reduced. Variable resistor R_3 is provided to control the stage gain of V_1 and so permit readjustment for the correct operating voltages. By this means any necessary correction of the relative phase of the transient and the horizontal sweep may be accomplished.

6. MEASUREMENT OF PLATE CURRENT-GRID BIAS CHARACTERISTICS

6.1 General

As discussed previously the plate current-grid bias characteristic, of the valve to be incorporated in a design of the type being considered, is the main piece of information available to the engineer. In practice of course the designer will generally employ published characteristics and will apply to his final result the tolerances necessitated by production variations in the tube characteristics.

Even Joint Navy-Army-Air Force Specifications, which are probably the most stringent of those applied to production vacuum tubes, allow in most cases a variation in plate current of 33 1/3% about the nominal value for given test conditions. Consequently to use the manufacturer's average characteristics in attempting to determine a rule by which they may be idealized would be ridiculous. A method for obtaining these characteristics for the particular tubes used in the experiment is desired.

In Sec. 3.3.2 the possibility was considered of using vacuum tubes as switches carrying peak currents considerably in excess of those normally obtained. It is that mode of operation which will be employed in the experiments. The use of a static method of plotting the desired characteristics is not feasible since the average power ratings of the tube elements would be exceeded for even high values of grid bias.

6.2 The Plate Current-Grid Bias Tester

In the report considered (3) a dynamic tube tester, which used a recurring pulse of low duty cycle, is described. A tester using the basic ideas of that circuit but of more elementary form has been designed and is considered adequate for the present problem. Figure 26 shows the circuit.

Valve V_1 operates in an asymmetrical multivibrator circuit having a frequency of 500 cycles/sec. The positive pulse appearing across R_5 has a duty cycle of 1/10th. Diode V_2(1) in conjunction with C_4 and R_5 acts as a peak clamping circuit which insures that the voltage which the grid of the tube (V_3) under test reaches is that measured by D.C. Voltmeter No. 1.

The peak swing in voltage appearing across the reference resistor R in the plate circuit of the tested tube is a direct measure of the current in the valve. Diode V_2(2), resistor R_8, condensor C_5 and D.C. Voltmeter No. 2 (Hewlett-Packard 410 A - 1 megohm/volt.) form a negative-peak-reading vacuum-tube voltmeter which indicates the plate drop.

Dynamic tube characteristics can be obtained by plotting the ratio peak plate volts/reference resistance against values of grid bias given by the clamping potential.

6.3 Characteristics of the Valves Employed

For electrode potentials on a test valve, such that a static curve could be obtained without destroying it, static and dynamic plate current-grid bias characteristics were measured. Figure 27 shows the results obtained. An error of

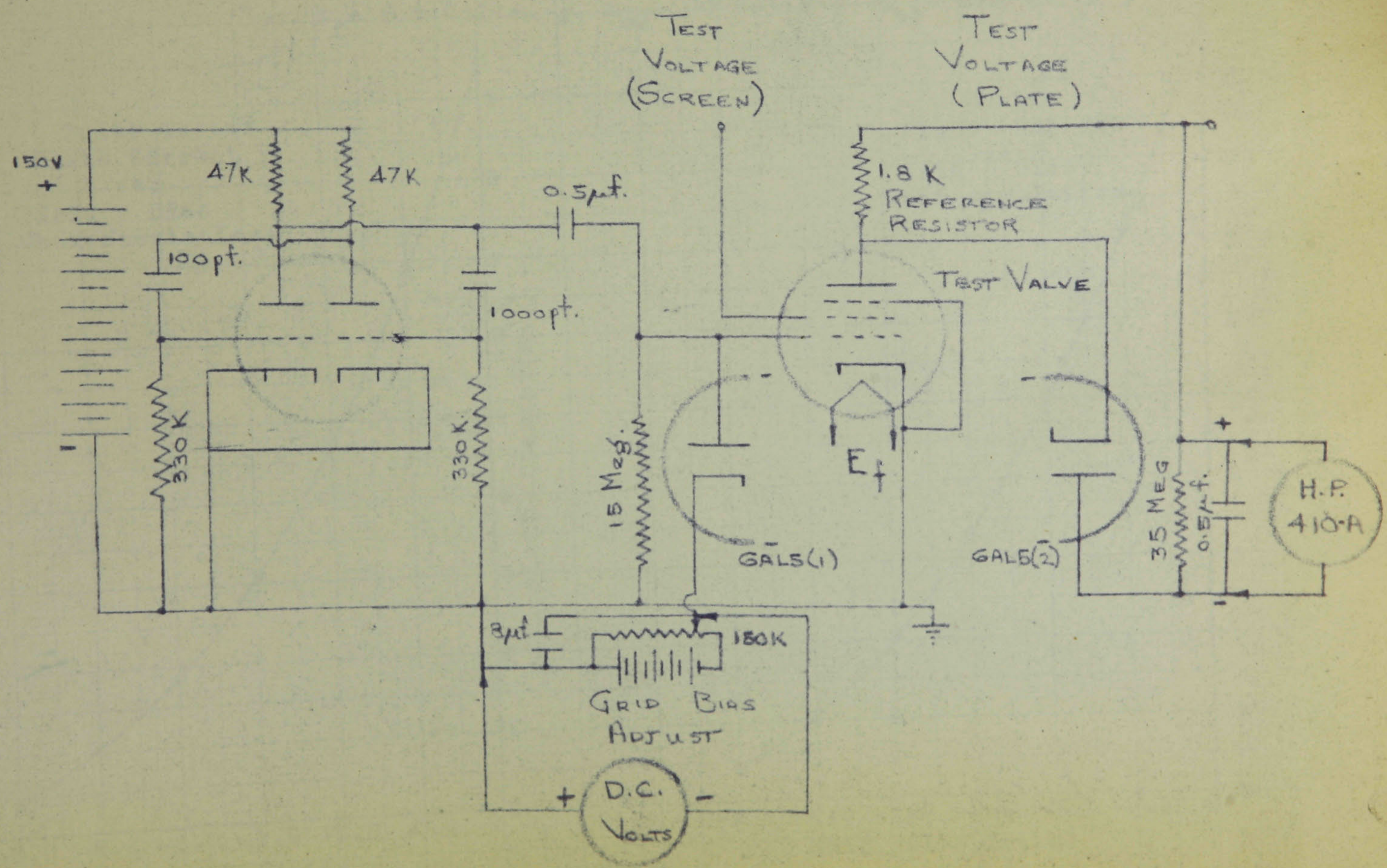
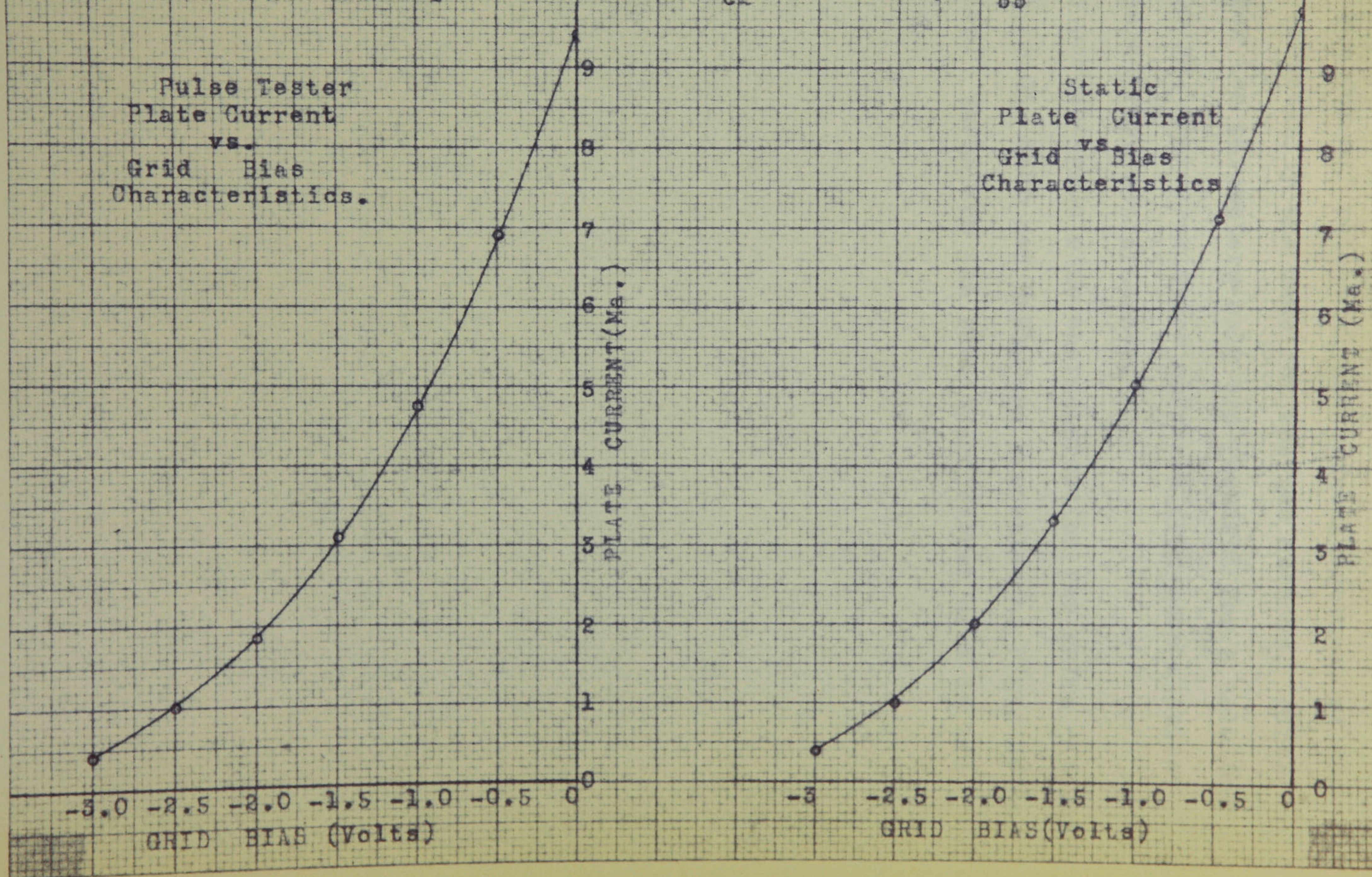


Fig. 26 : Schematic Diagram of Plate Current - Grid Bias Tester.

Fig. 27

E_f = 6.3 Volts ; E_{c2} = 90 Volts ; E_{bb} = 300 Volts .



approximately four percent in the peak current value is introduced but the cut-off potential and transconductance of the valve are essentially the same for the two methods. Since these latter two parameters only are required for a solution of the problem, characteristics obtained from such a pulse-tester should prove satisfactory.

Plate current-grid bias characteristics for valve - types: 6AN5, 6AK5, 6AG5, 6AU6 and 6AH6, are presented as Figs. 28 to 37 inclusive. These are the particular valves used in the experiments.

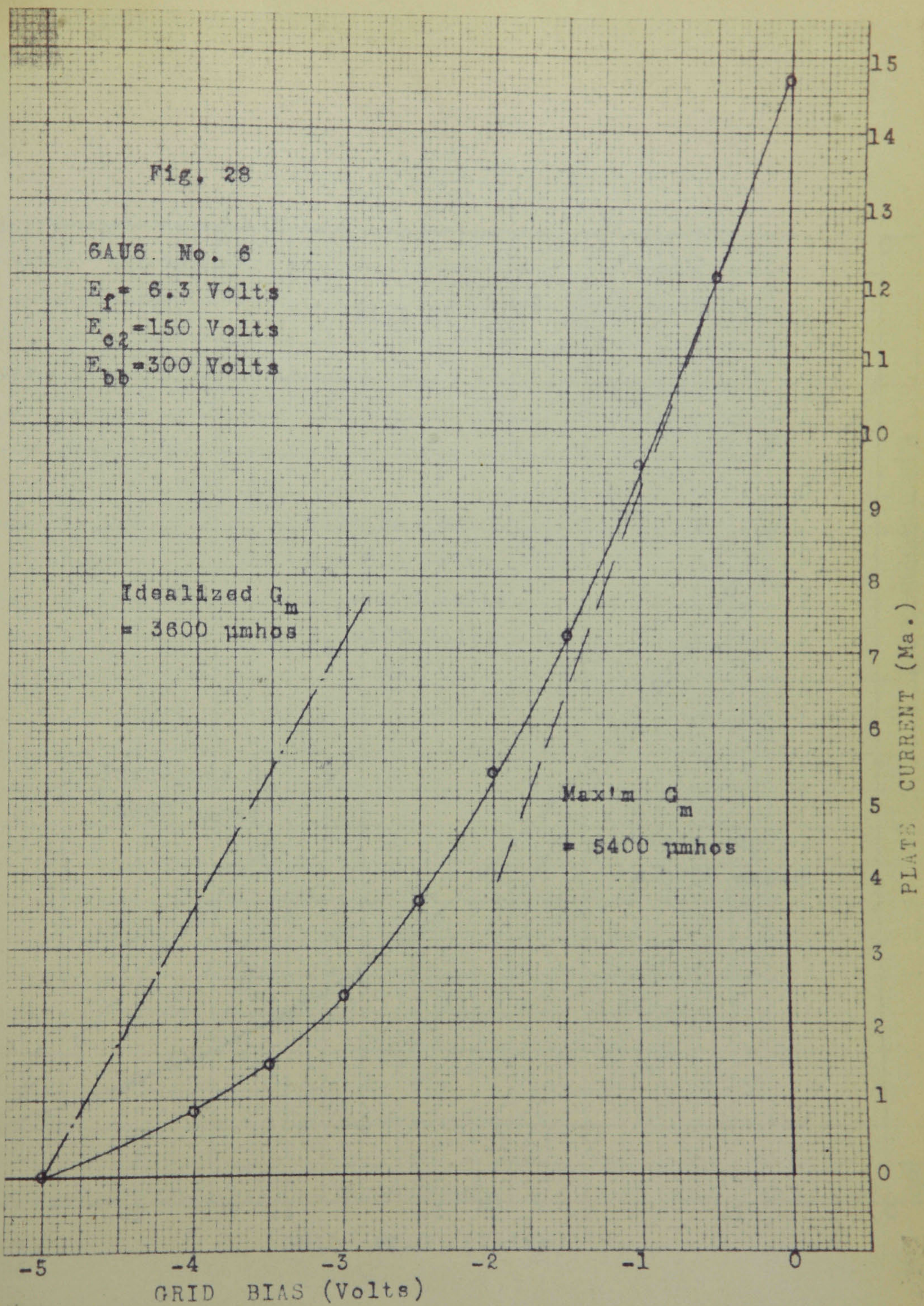


Fig. 28
6AU6 No. 6
E_f = 6.3 Volts
E_{c2} =150 Volts
E_{bb} =300 Volts
Idealized G_m
= 3600 µmhos
Max'm G_m
= 5400 µmhos
15
14
13
12
11
10
9
8
7
6
5
4
3
2
1
0
PLATE CURRENT (Ma.)
-5
-4
-3
-2
-1
0
GRID BIAS (Volts)

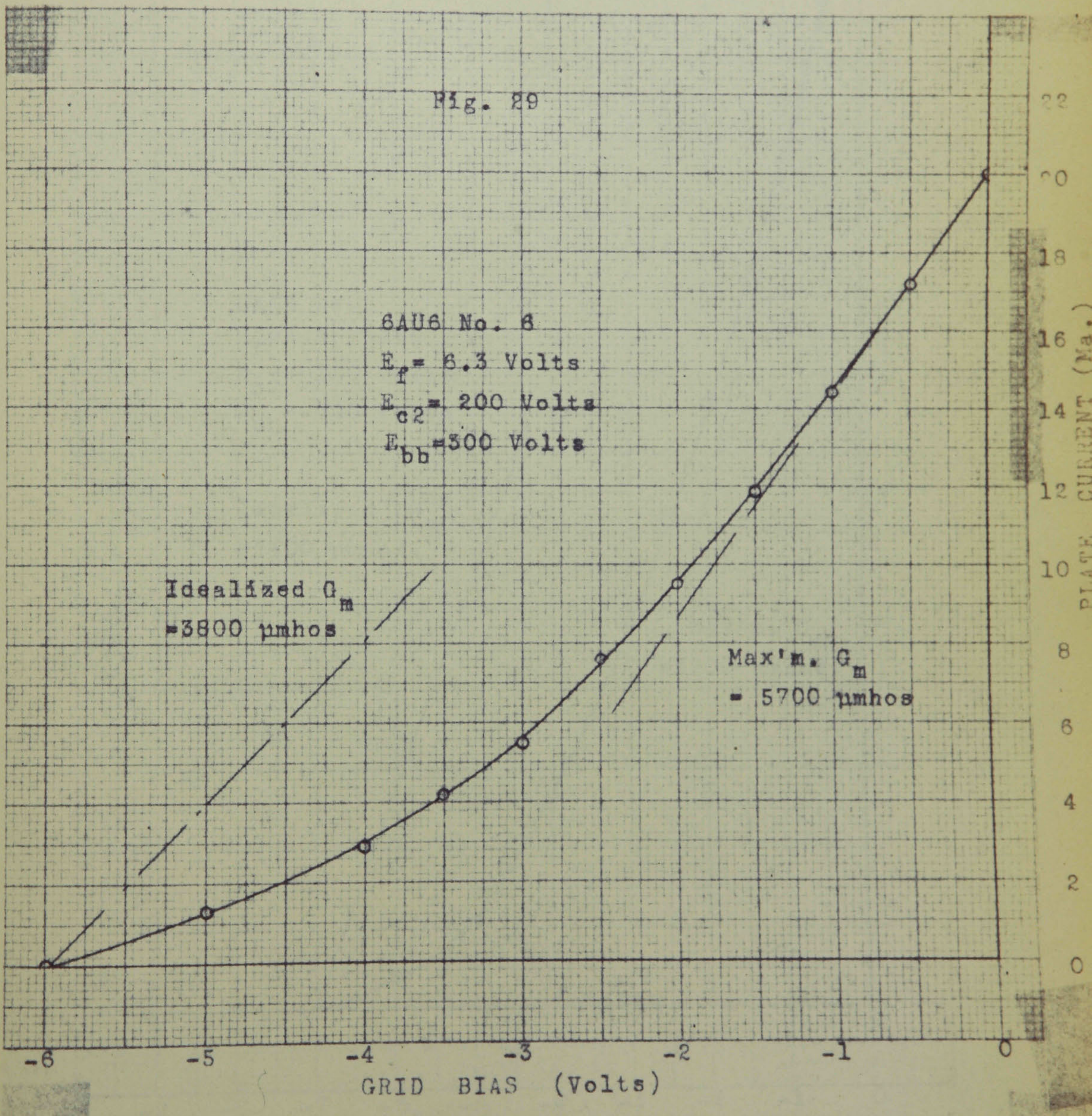


Fig. 29
6AU6 No. 6
E_f = 6.3 Volts
E_{c2} = 200 Volts
E_{bb} = 300 Volts
Idealized G_m
= 3800 µmhos
Max'm. G_m
= 5700 µmhos
PLATE CURRENT (Ma.)
GRID BIAS (Volts)
-6
-5
-4
-3
-2
-1
0
22
20
18
16
14
12
10
8
6
4
2
0

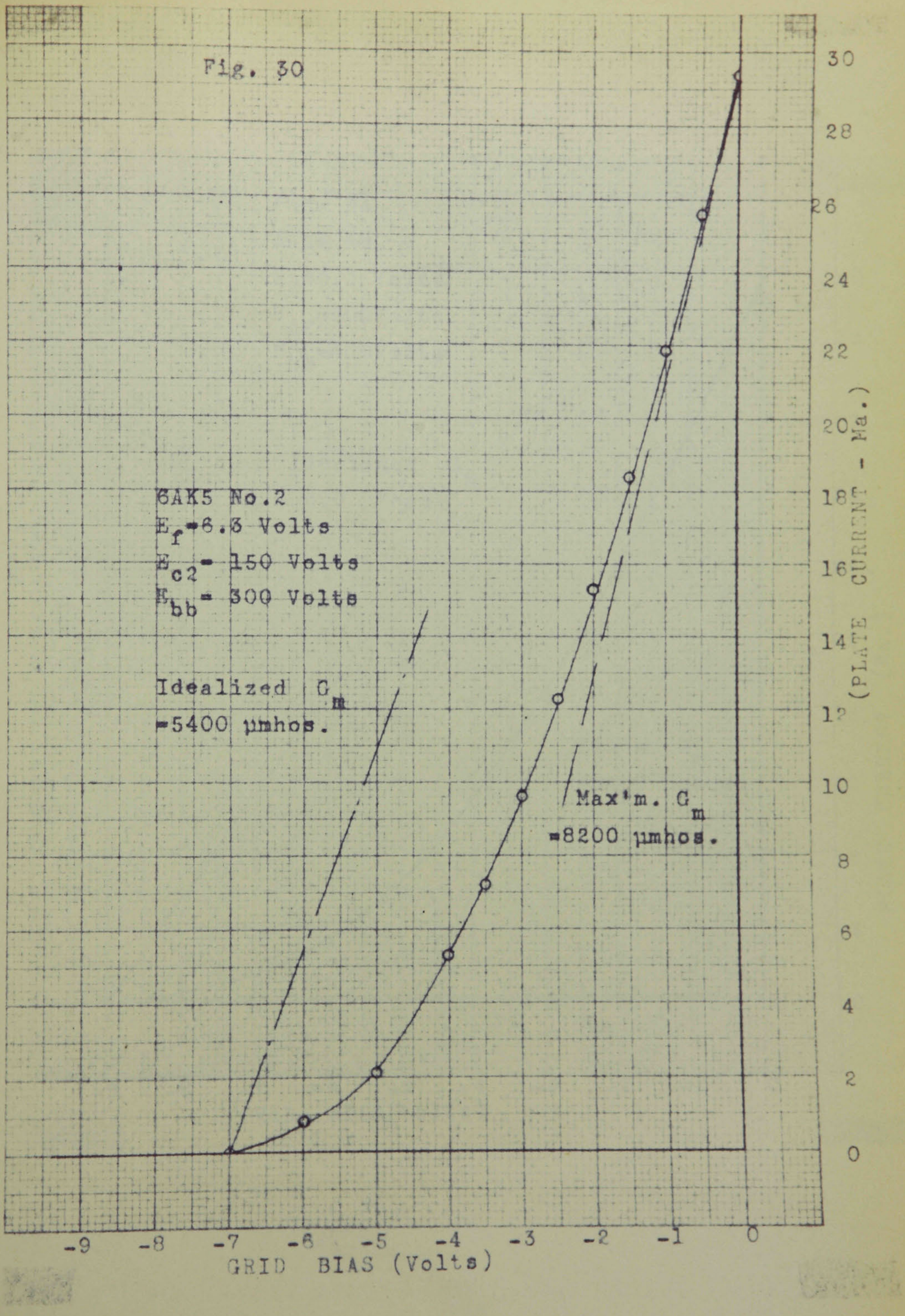


Fig. 30
6AK5 No.2
E_f=6.3 Volts
E_{c2}= 150 Volts
E_{bb}= 300 Volts
Idealized G_m
=5400 µmhos.
Max'm. G_m
=8200 µmhos.
(PLATE CURRENT - Ma.)
30
28
26
24
22
20
18
16
14
12
10
8
6
4
2
0
-9
-8
-7
-6
-5
-4
-3
-2
-1
0
GRID BIAS (Volts)

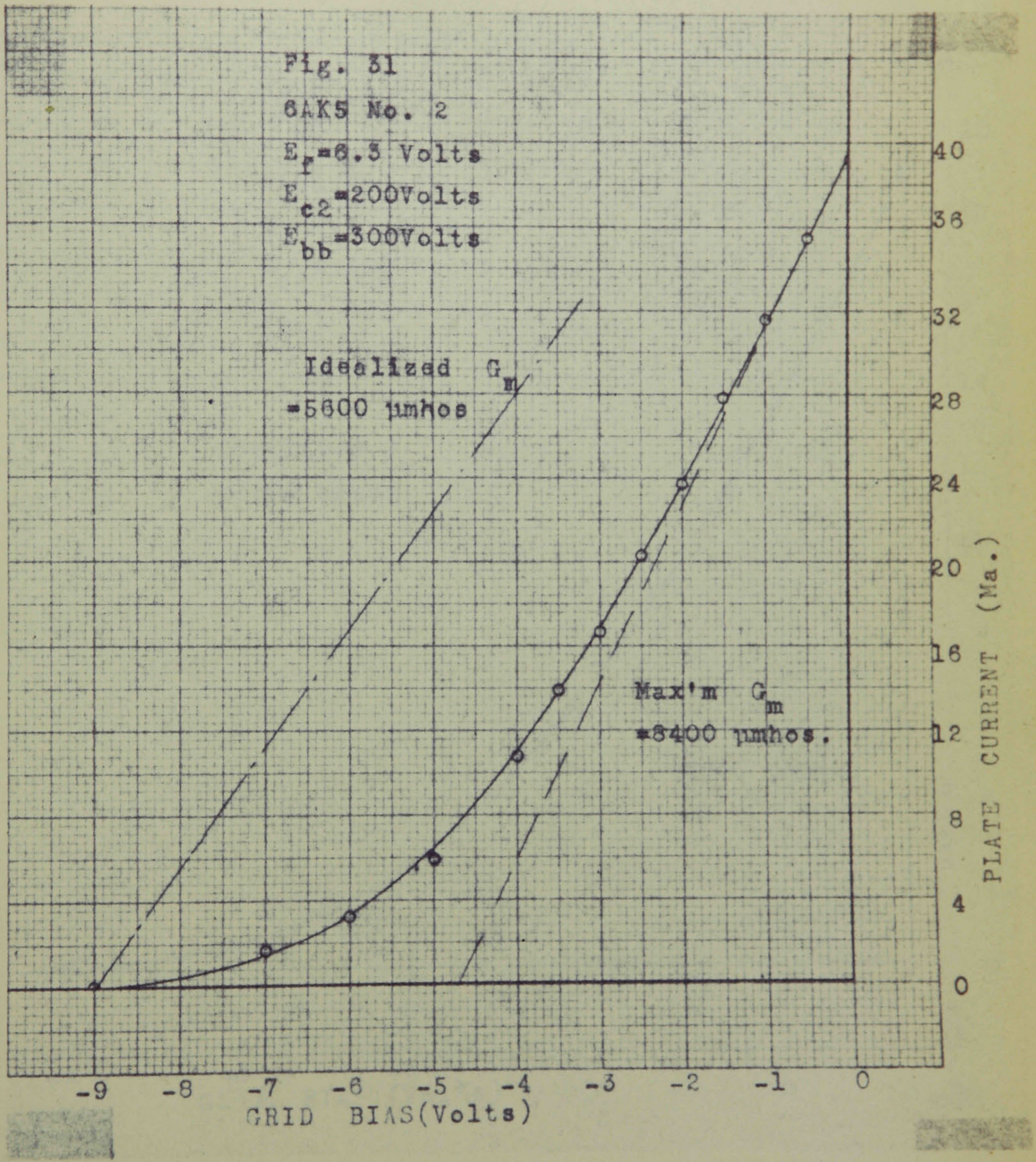


Fig. 31
6AK5 No. 2
E_f=6.3 Volts
E_{c2}=200Volts
E_{bb}=300Volts
Idealized G_m
=5600 µmhos
Max'm G_m
=8400 µmhos.
PLATE CURRENT (Ma.)
0
4
8
12
16
20
24
28
32
36
40
-9
-8
-7
-6
-5
-4
-3
-2
-1
0
GRID BIAS(Volts)

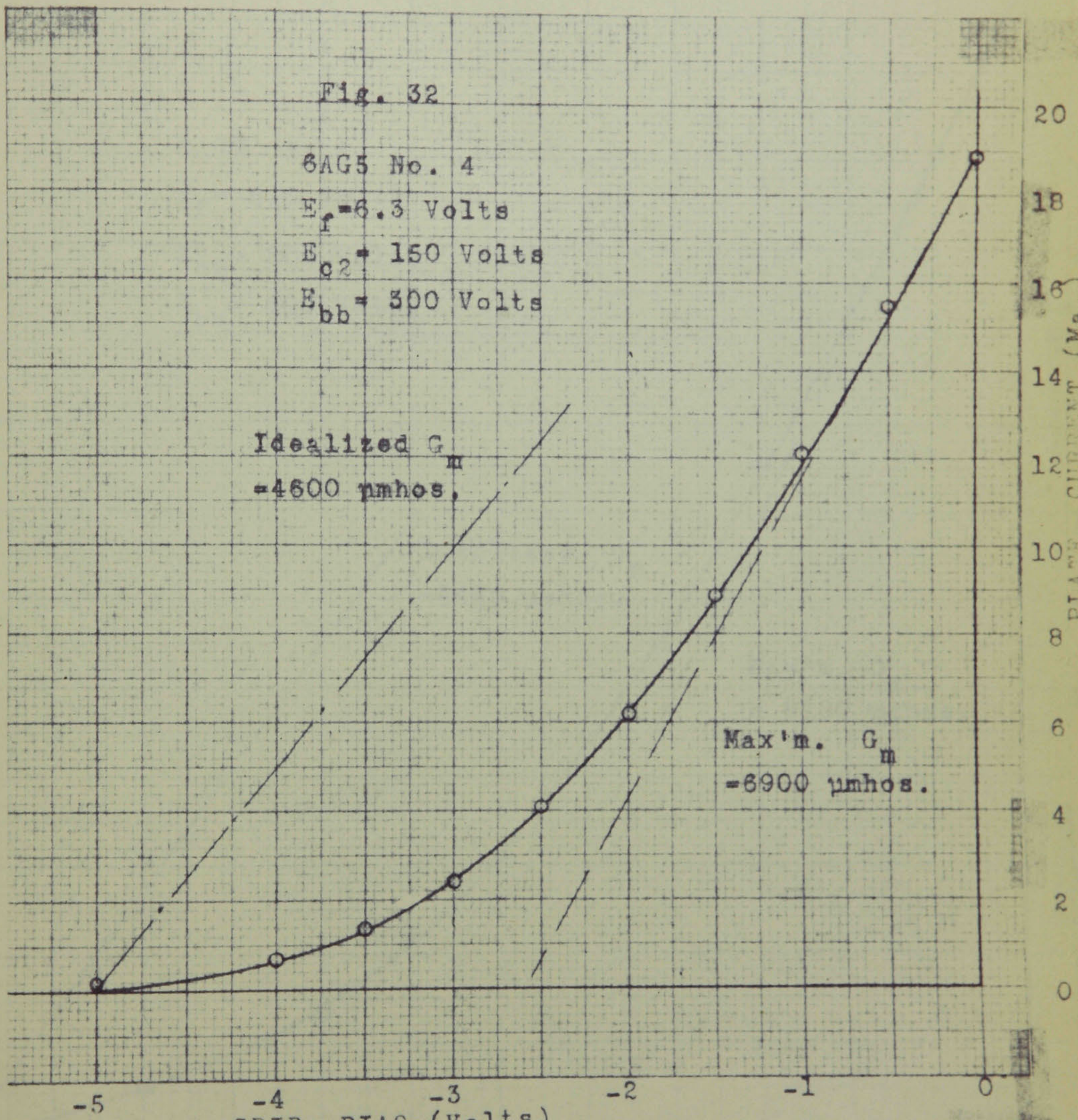


Fig. 32
6AG5 No. 4
E_f=6.3 Volts
E_{c2}= 150 Volts
E_{bb}= 300 Volts
Idealized G_m
=4600 µmhos.
Max'm. G_m
=6900 µmhos.
PLATE CURRENT (Ma.)
20
18
16
14
12
10
8
6
4
2
0
-5
-4
-3
-2
-1
0
GRID BIAS (Volts)

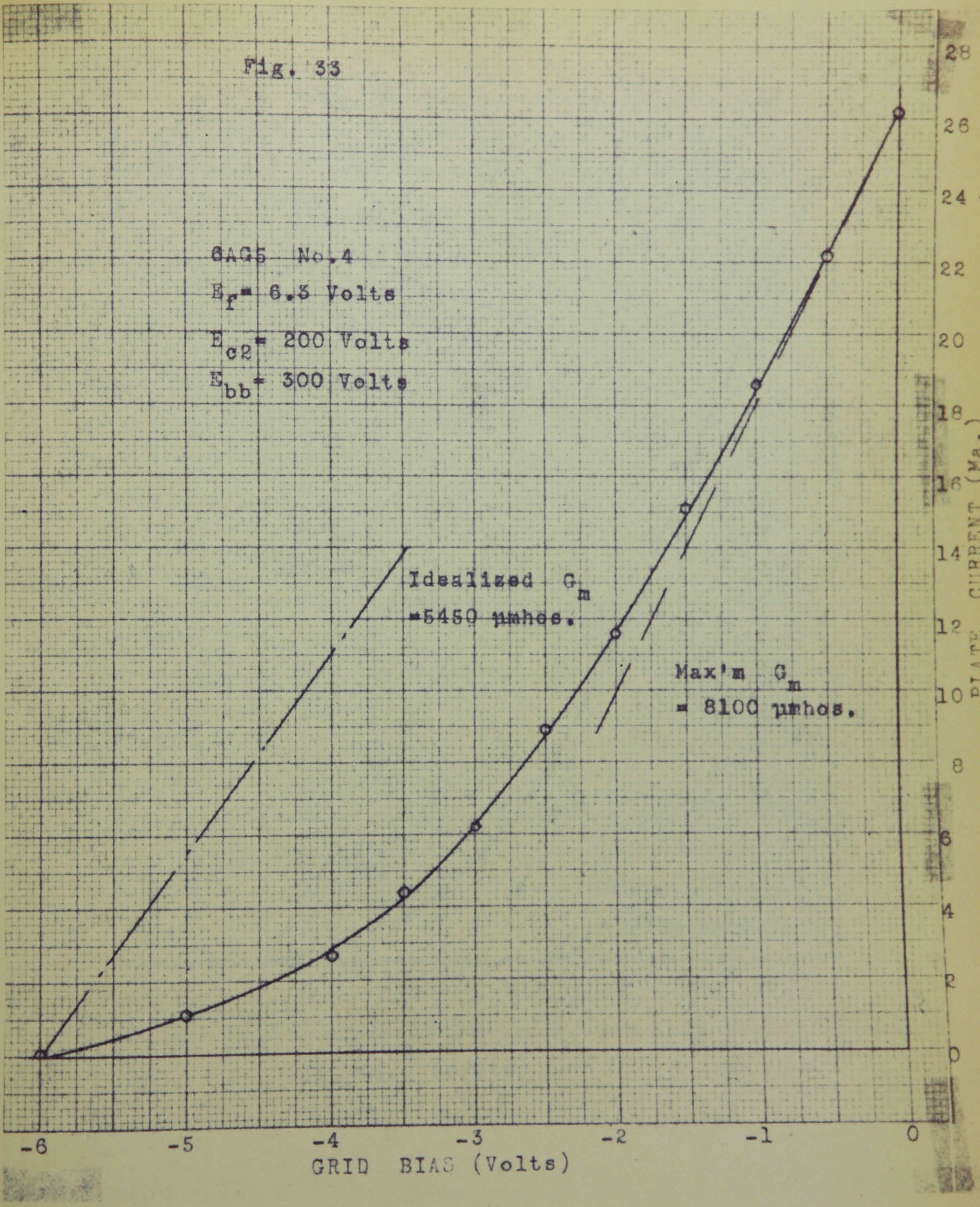


Fig. 33
6AG5 No. 4
E_f = 6.3 Volts
E_{c2} = 200 Volts
E_{bb} = 300 Volts
Idealized G_m
=5450 μmhos.
Max'm G_m
= 8100 μmhos.
PLATE CURRENT (Ma.)
28
26
24
22
20
18
16
14
12
10
8
6
4
2
0
-6
-5
-4
-3
-2
-1
0
GRID BIAS (Volts)

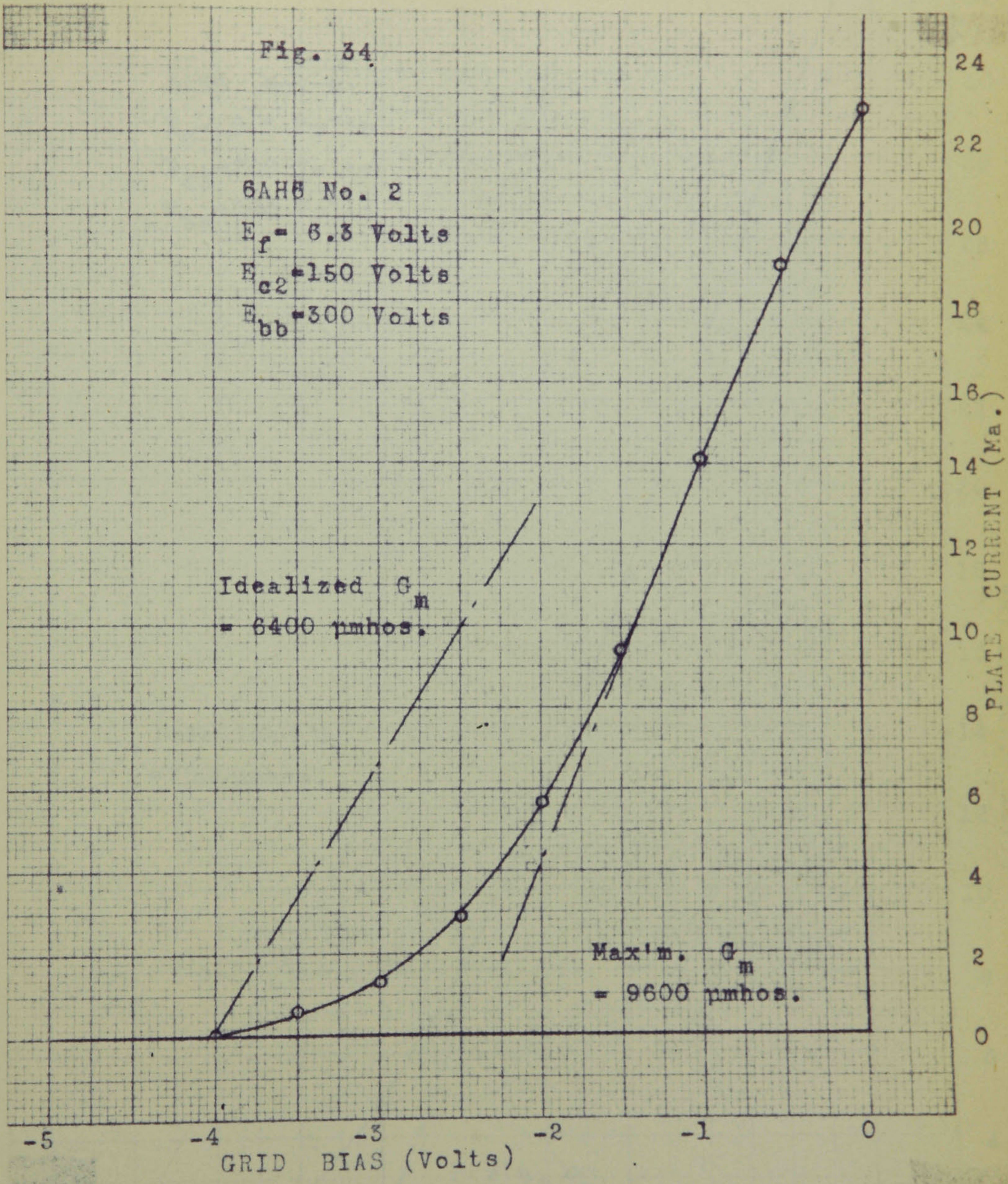


Fig. 34
6AH6 No. 2
E_f = 6.3 Volts
E_c2 = 150 Volts
E_bb = 300 Volts
Idealized G_m
= 6400 μmhos.
Max'm. G_m
= 9600 μmhos.
24
22
20
18
16
14
12
10
8
6
4
2
0
PLATE CURRENT (Ma.)
-5
-4
-3
-2
-1
0
GRID BIAS (Volts)

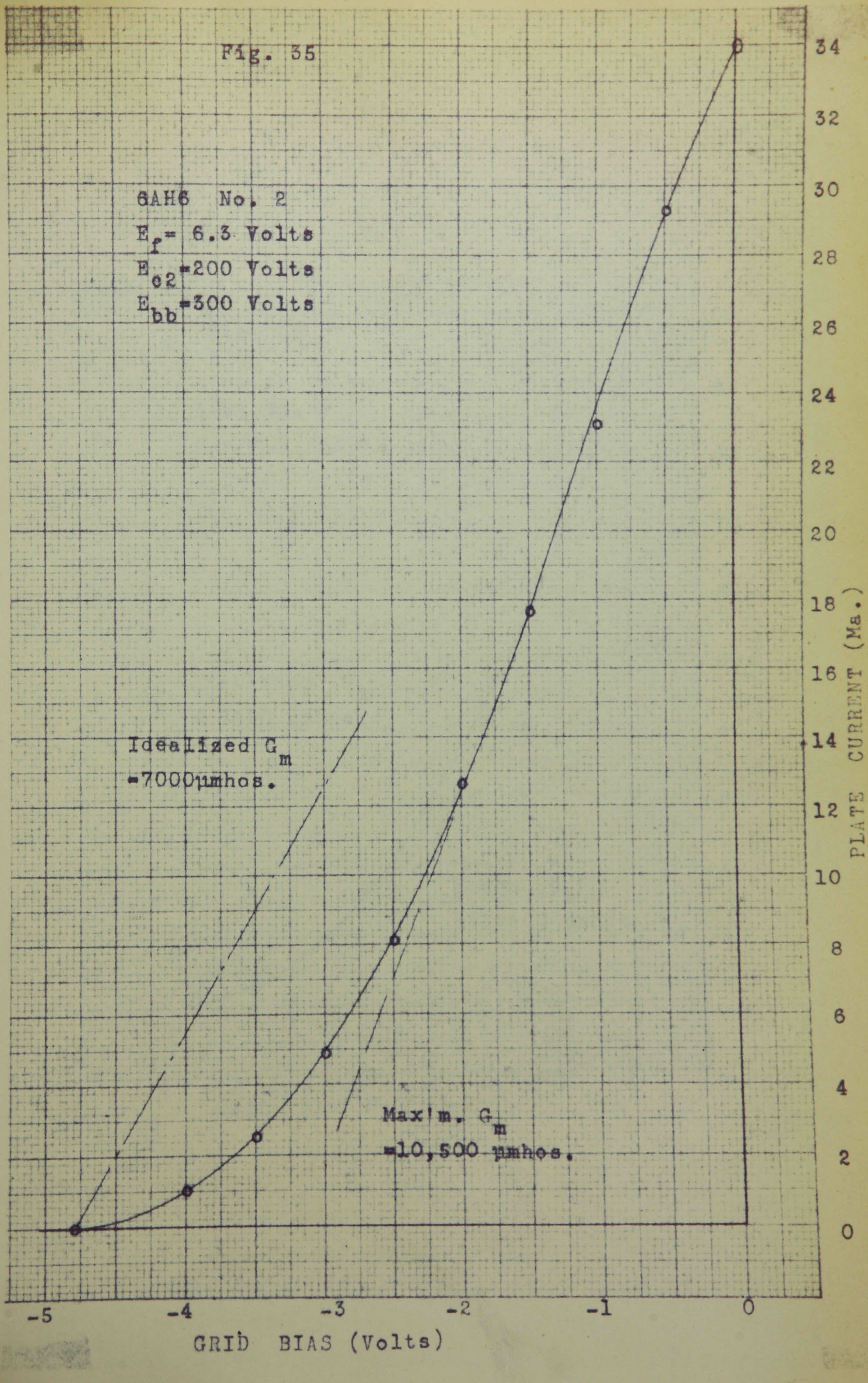


Fig. 35
6AH6 No. 2
E_f= 6.3 Volts
E_c2=200 Volts
E_bb=300 Volts
Idealized G_m
=7000µmhos.
Max'm. G_m
=10,500 µmhos.
34
32
30
28
26
24
22
20
18
16
14
12
10
8
6
4
2
0
PLATE CURRENT (Ma.)
-5
-4
-3
-2
-1
0
GRID BIAS (Volts)

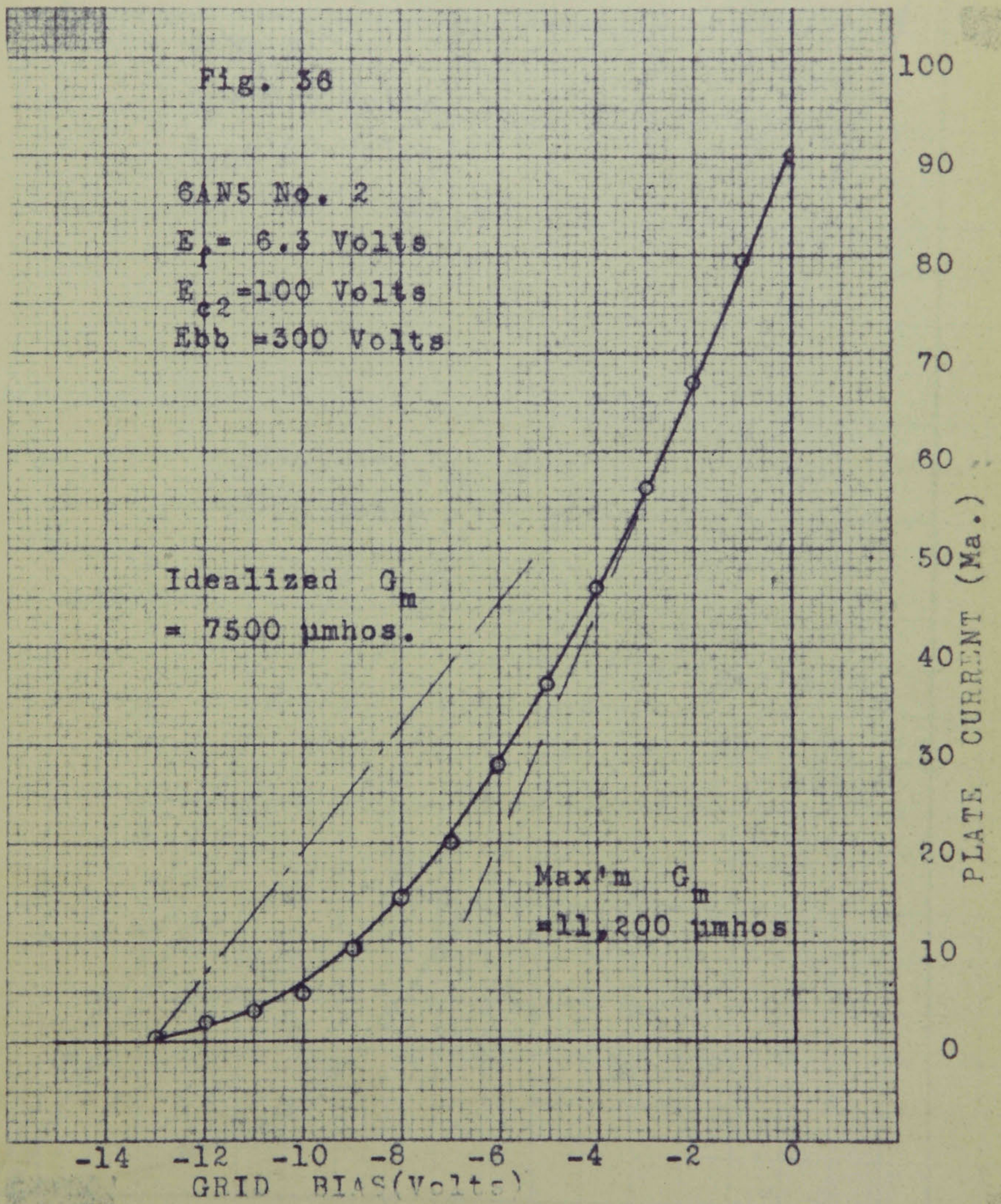


Fig. 36
6AN5 No. 2
E_f = 6.3 Volts
E_{c2} =100 Volts
Ebb =300 Volts
Idealized G_m
= 7500 µmhos.
Max'm G_m
=11,200 µmhos
100
90
80
70
60
50
40
30
20
10
0
PLATE CURRENT (Ma.)
-14
-12
-10
-8
-6
-4
-2
0

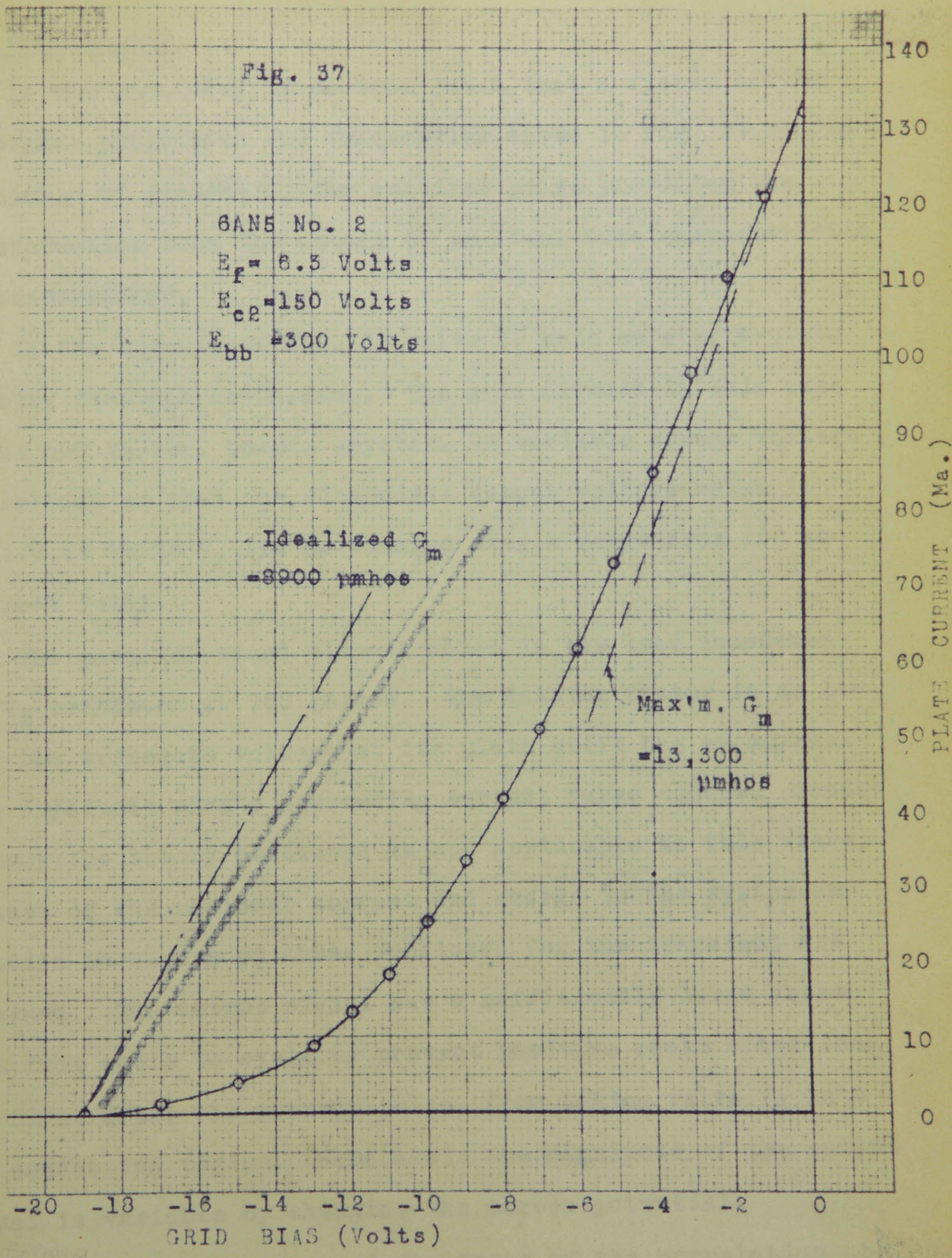


Fig. 37
6AN5 No. 2
E_f= 6.3 Volts
E_{c2}=150 Volts
E_{bb} =300 Volts
Idealized G_m
=8900 μmhos
Max'm. G_m
=13,300
μmhos
PLATE CURRENT (Ma.)
140
130
120
110
100
90
80
70
60
50
40
30
20
10
0
-20
-18
-16
-14
-12
-10
-8
-6
-4
-2
0
GRID BIAS (Volts)

7. MEASUREMENT OF THE TRANSIENT RESPONSE

7.1 The Circuit

It was concluded in section 3.5.6 that a system having the signal amplitudes and frequencies shown in Fig. 23 should be capable of generating the required pulse providing the transformation from the actual to an ideal tube characteristic was accomplished.

In the circuit of Fig. 23 valve V_1 acts as a grid-plate capacity feedback oscillator. The grid circuit in this case is the 100 kc/sec. quartz crystal. A variable screen voltage is provided so that the sinusoidal output voltage which is applied to valve V_2 can be changed over a range from 100 to 300 peak volts.

The parallel tuned circuit which is the load impedance for V_2 resonates at 700 kc/sec. The damping factor is so low that the transient voltage at its plate still has an appreciable amplitude when a succeeding current pulse occurs. Consequently the theory developed is not applicable to this stage. Because of the residual current and charge in the system amplitudes in excess of those predicted for the transient are attained. The output signal has a sufficiently large decrement from cycle to cycle to prevent positive peaks other than the first after the current pulse, from taking valve V_3 into the conducting region. Here again the magnitude of the output signal is variable by changing the screen potential.

Stage 3 is the one used for tests of the correlation obtainable between theory and practice for the transient response derived. In this case the damping factor is so high that the output has decayed to zero before application of another current pulse. For experimental purposes the valves referred to in section 6.3 were successively placed in this stage.

The grid-cathode diode of the valve under test, in conjunction with the very long time constant of the grid resistor and coupling capacity to the 700 kc. source, provides a positive-peak-reading vacuum-tube-voltmeter. The peak applied signal was determined by measuring the average d.c. potential at the grid using the D.C. Voltmeter portion of an Hewlett-Packard 410 A instrument.

Values of the peak positive transient output at the frequency of the plate tuned circuit, 6 mc/sec., were obtained using the a.c. voltmeter portion of the same instrument. This procedure avoided the necessity for two a.c. voltmeters having the very low input capacity possessed by the high frequency probe of the one used. It also avoided the alternative procedure of moving the probe (for each measurement) from input circuit to output circuit with the possible detuning of the circuits which would have resulted.

7.2 The Measurements

The values of the linear circuit elements of the 700 kc/sec. and the 6 mc/sec. stages were determined by use of a Q-Meter (Boonton Radio Corporation Type 160-A). The natural resonant frequency was found in both cases by obtaining the frequency at which no detuning, of a resonant circuit on the

Q-Meter, occurred when the impedance from plate to ground of valves V_2 and V_3 respectively were connected in parallel with the Q-Meter capacity. The dynamic resistance can be found from the values of the original Q, the final Q, and the capacity at the resonant frequency when the previous connection is accomplished. The circuit capacity was determined by finding the change in resonant frequency when a known reference capacity was inserted in parallel with the tuned circuit.

With the information so obtained, the values of the elements L, C and R used in the analysis can be evaluated. In the measurements it was intended to adjust G, the conductance in parallel with the tuned circuit, to the value determined by the lowest dynamic plate resistance of the five pentode types used. A printing error in the tube characteristic sheets from which that information was derived has inadvertently led to use of a shunting resistance (12,500 ohms) much lower than necessary. The value of G in any calculations has been based on this resistance.

The measurements and subsequent calculations lead to the following values for the circuit elements:-

700 kc/sec.: $L = 360$ microh. $C = 145$ pf. $R_D = 140$ kohms.
$f_r = 700$ kc/sec.

6 mc/sec.: $L = 31.5$ microh. $C = 22$ pf. $G = 0.8 \times 10^{-4}$ mhos.
$f_r = 6.1$ mc/sec.

The peak positive output of the 700 kc/sec. tuned circuit has been obtained using a valve type 6AU6 with a screen potential of 200 volts d.c. and for 100 kc/sec. grid drives from 30 to 200 volts peak. The response curve is shown in Fig. 38.

The peak positive output of the 6 mc/sec. tuned circuit has been obtained for two screen potentials for each of the valve types whose plate current-grid bias characteristics were measured. The input signal in this case was the 700 kc/sec. sine wave varied in steps between 100 and 300 volts peak. The ten response curves so obtained are plotted in Figs. 39 to 48 inclusive.

8. COMPARISON OF MEASURED AND CALCULATED RESPONSE

8.1 The Output of the 700 kc/sec. LC Circuit

In the discussion of the circuit it was pointed out that the damping factor of this stage is sufficiently low that the decay of the peak voltage amplitude between current pulses is incomplete and a steady state condition exists. In actual fact the response should be determined by combining the transient and steady state solutions for the network. A reasonable approximation to the response can be obtained if only the product of the dynamic-resistance and the seventh harmonic component of the Fourier analysis (4) of the current pulse in the valve are considered. The increased tabulation involved in getting a response by the former procedure is out of proportion to the improved satisfaction that can be derived.

A comparison between the measured response and a response obtained by the harmonic analysis, based on the actual E_{co} and the zero bias plate current of the 6AU6 employed, is made in Fig. 38.

8.2 The Output of the 6 mc/sec. LC Circuit

In section 3.5.6 it was seen that peak amplitude was determined by expression 13 if $\frac{2\pi\omega}{\theta} \leq \beta$ and by Eq. 14 if $\frac{2\pi\omega}{\theta} \geq \beta$. In the former case the peak amplitude, obtained from Eq. 13, is

$$V_{pk} = \frac{G_m A}{LC}\left[\cos\frac{\beta}{2}\,\frac{R}{\theta^2} + K_1\left\{e^{-\alpha T}\sqrt{\sin^2\frac{\beta}{2} + \frac{\omega^2}{\theta^2}\cos^2\frac{\beta}{2}} + \sin\left(\omega T - \frac{\beta}{2}\right)\right\}\right]$$

where $T = \frac{(2\pi \cdot \phi)}{\theta}$ is the time to the first maximum and in the latter case from Eq. 14

$$V_{pk} = \frac{G_m A K e^{-\alpha T}}{LC} \sqrt{\sin^2\frac{\beta}{2} + \frac{\omega^2}{\theta^2}\cos^2\frac{\beta}{2}} \left\{ \text{Magnitude} \left(1 + e^{\frac{\alpha\beta}{\omega}} \angle \frac{\beta\theta}{\omega} + 2\phi\right) \right\}$$

The values of $\frac{V_{pk}}{G_m A}$ determined by these two equations and the elements listed in section 7.2 for the 6 mc/sec. circuit were tabulated for conduction angles β from 10° to 40° and from 40° to 70° respectively. For grid cut-off potentials from 3 to 18 volts and grid drives of 100, 125, 150, 175, 200, 225, 250 and 300 volts peak the corresponding conduction angles β were determined. Tables were prepared from these, for each cut-off potential, relating the factor $\frac{V_{pk}}{G_m}$ to the grid drive. To evaluate the output now requires the assumption of specific values of E_{co} and G_m for the valve in the circuit.

8.3 Idealization of the Plate Current-Grid Bias Characteristic

A considerable number of processes for idealization were attempted before one was obtained which appears to be uniformly satisfactory for the valves tested. Intuition led the author to believe that a cut-off potential lower than the value given by measurement and a transconductance approximating the average value for the valve should provide the desired response. The curves so obtained did not justify the assumption. By gradual evolution of assumed values it was discovered that the actual cut-off potential and a transconductance equal to two-thirds of the maximum transconductance for the valve gave results closely resembling those obtained by measurements.

The idealized $I_p - E_g$ characteristics so derived are shown, with the actual characteristics, in Figs. 28 to 37 inclusive. Response curves based on this assumption are plotted in Figs. 39 to 48, inclusive, for comparison with the experimental values.

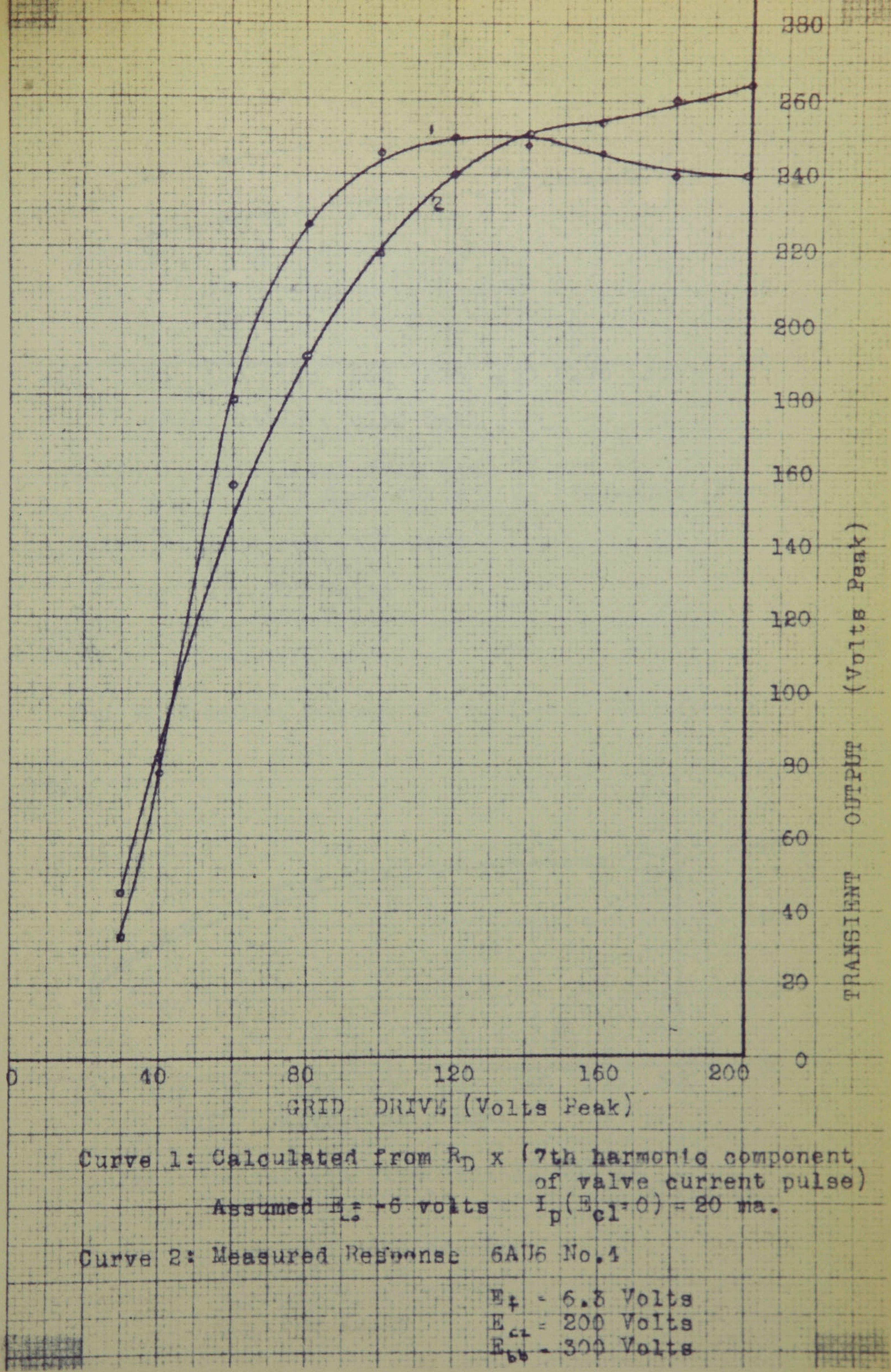


Fig. 38: Transient Response of 700 kc/sec LC Circuit (6AU6)

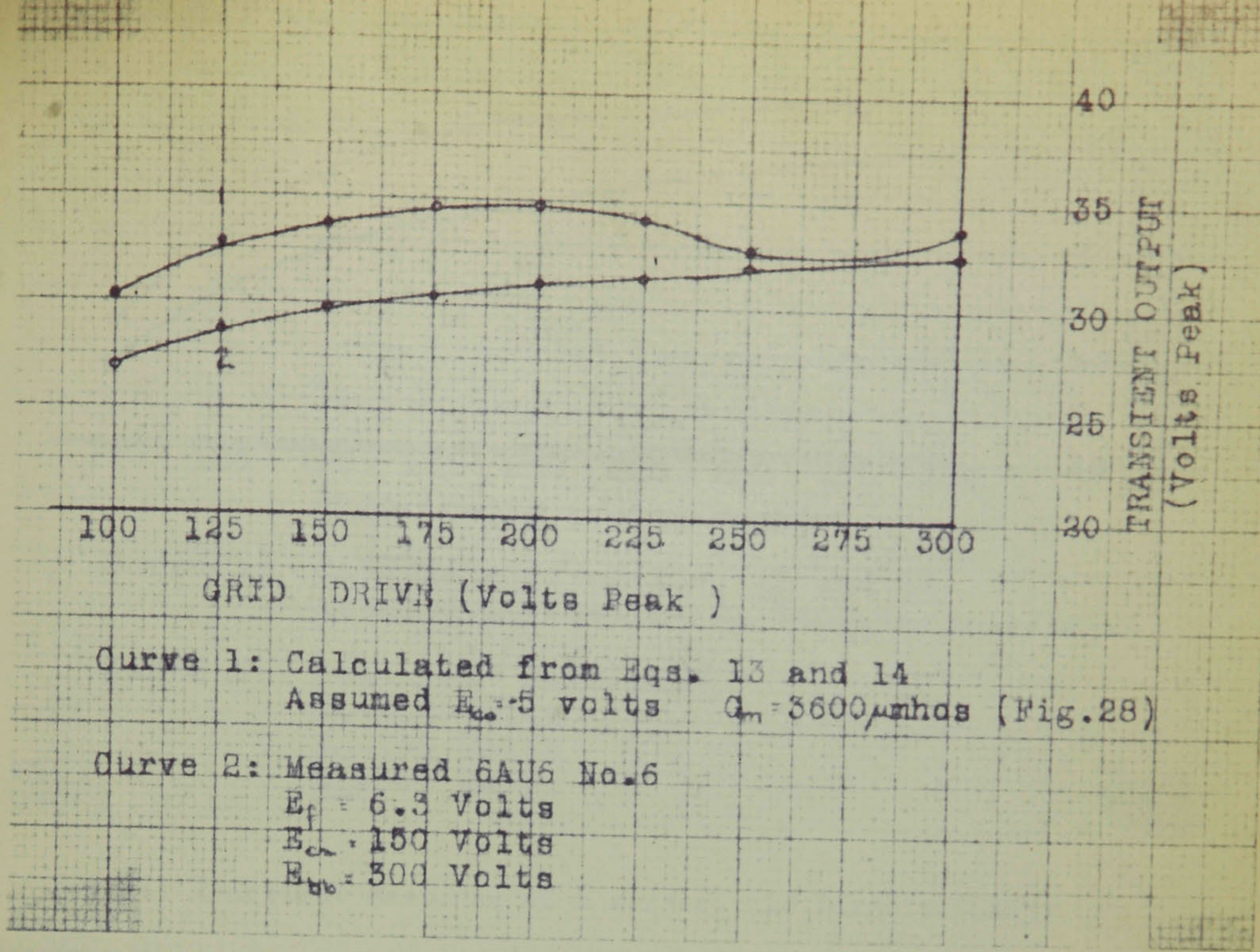


Fig. 39: Transient Response of 6 mc./sec LC Circuit(6AU6)

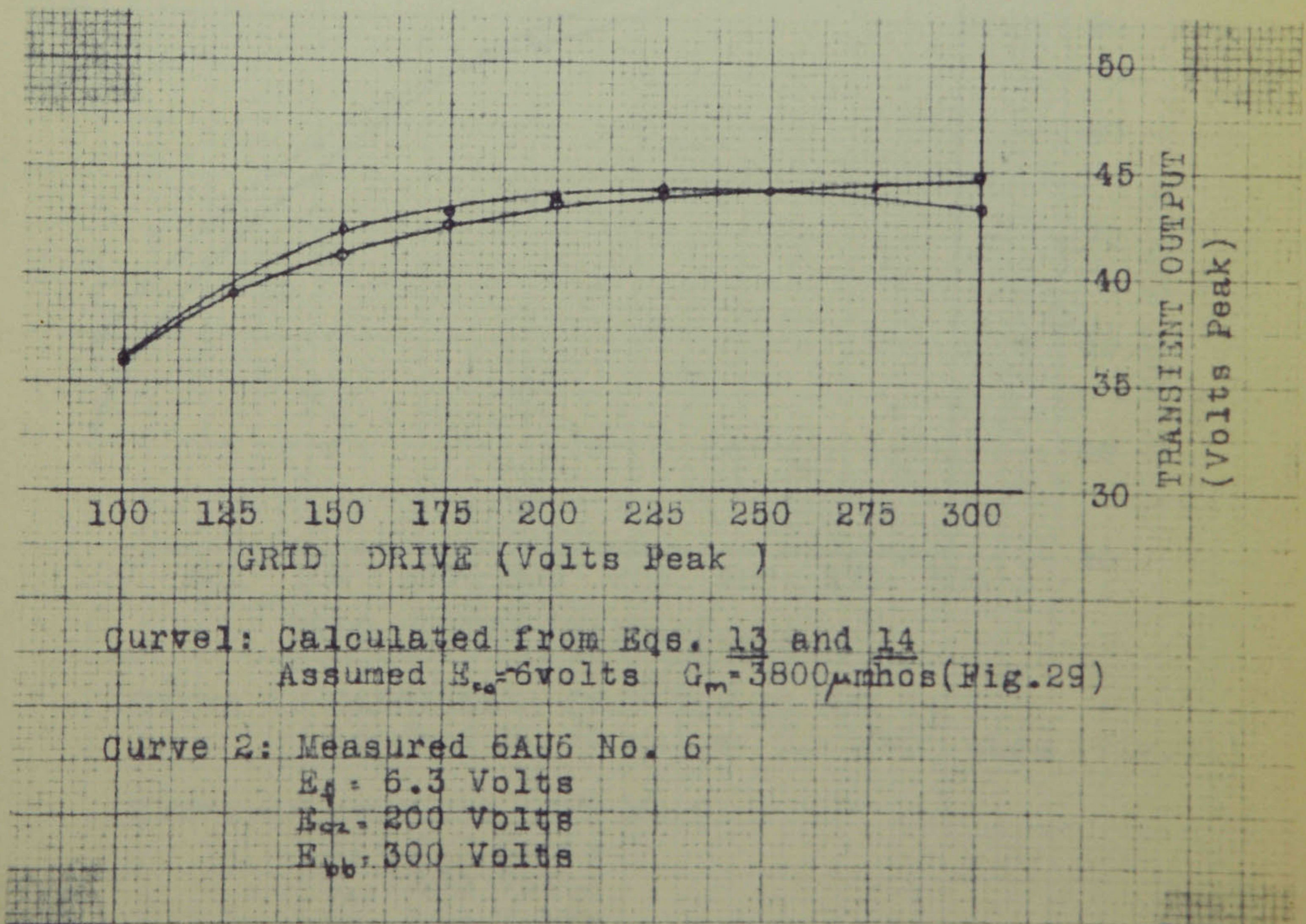


Fig. 40: Transient Response of 6 mc/sec. LC Circuit(6AU6)

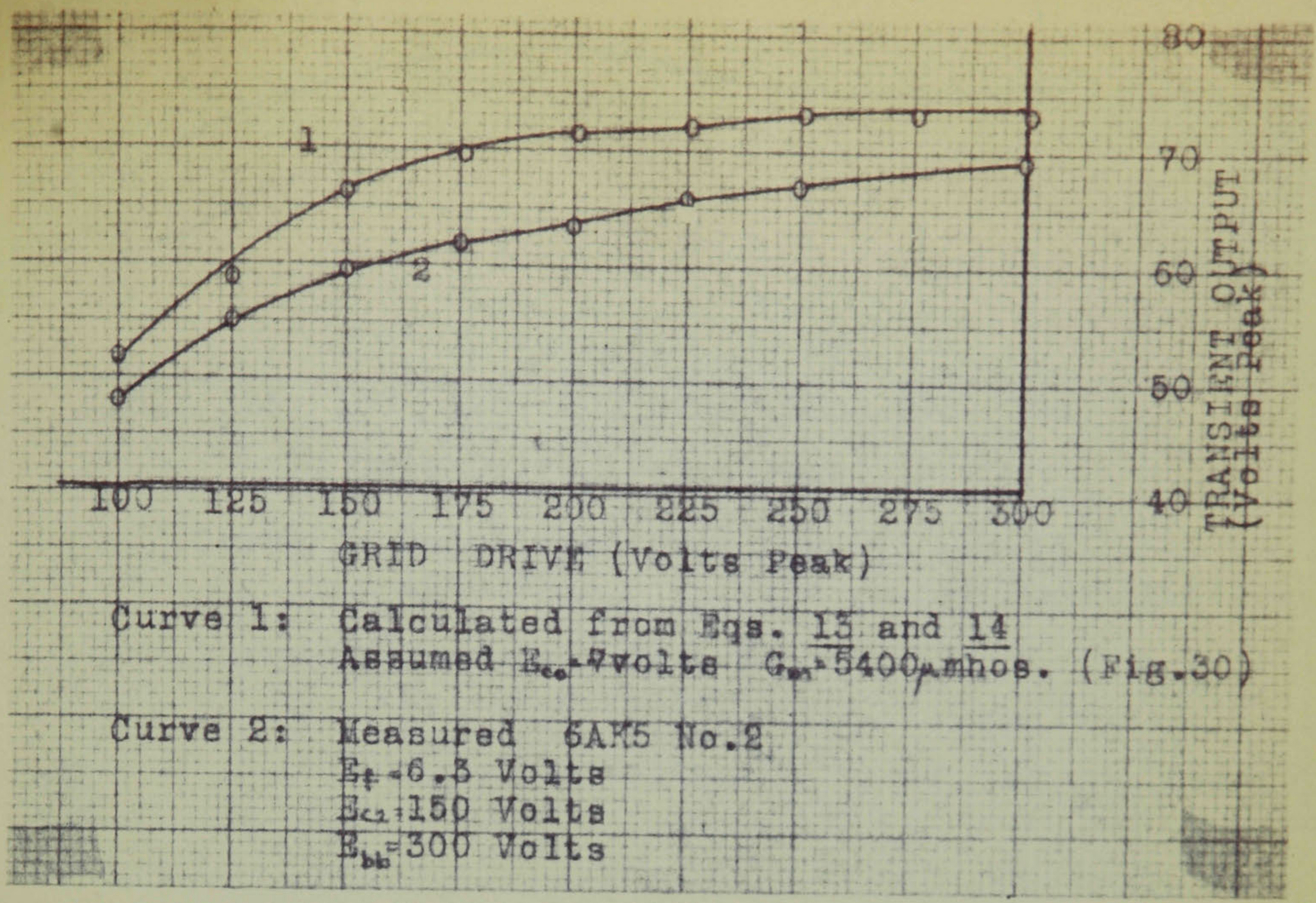


Fig.41: Transient Response of 6 mc/sec. LC Circuit.(6AK5)

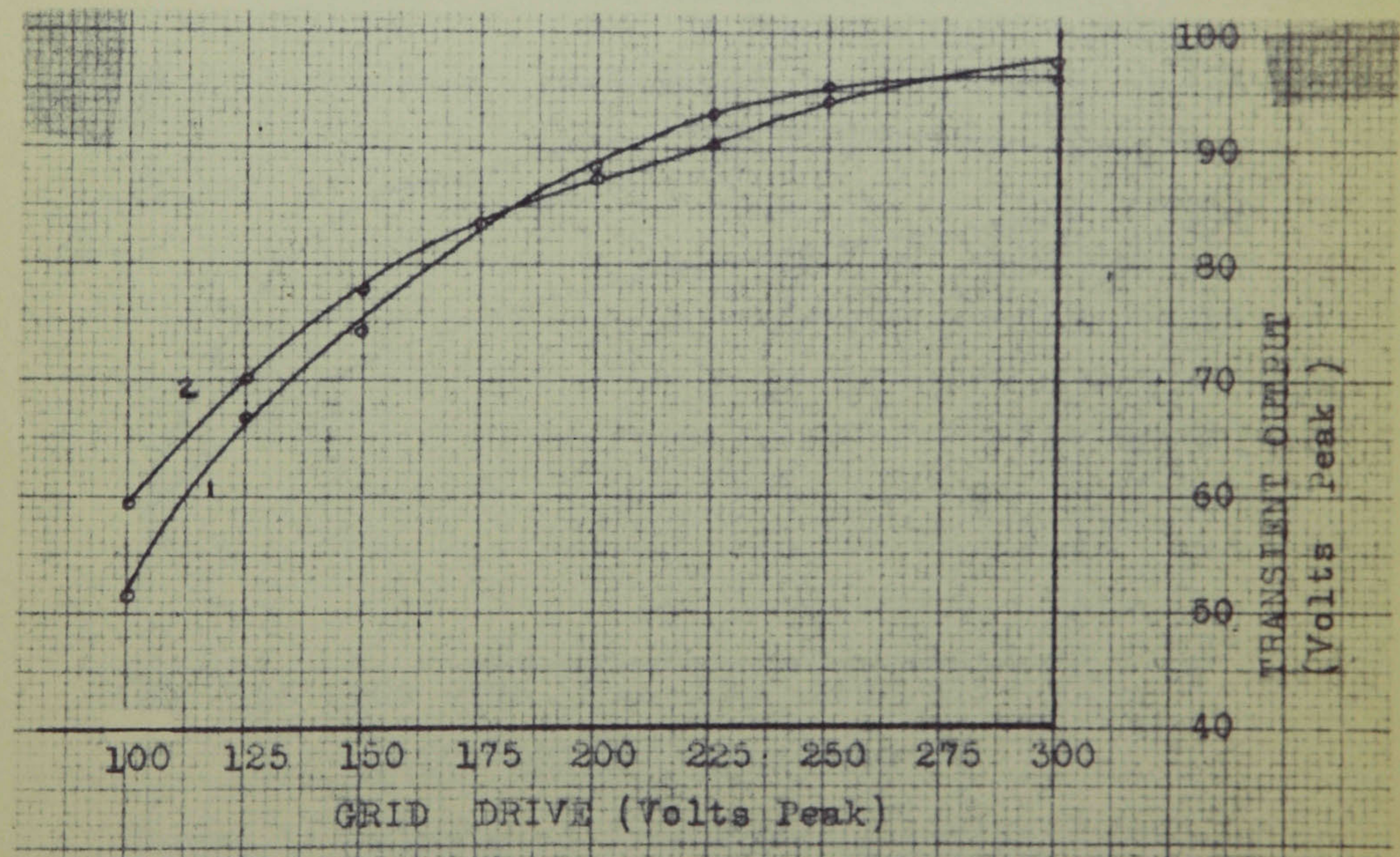


Fig.42: Transient Response of 6 mc/sec. LC Circuit.

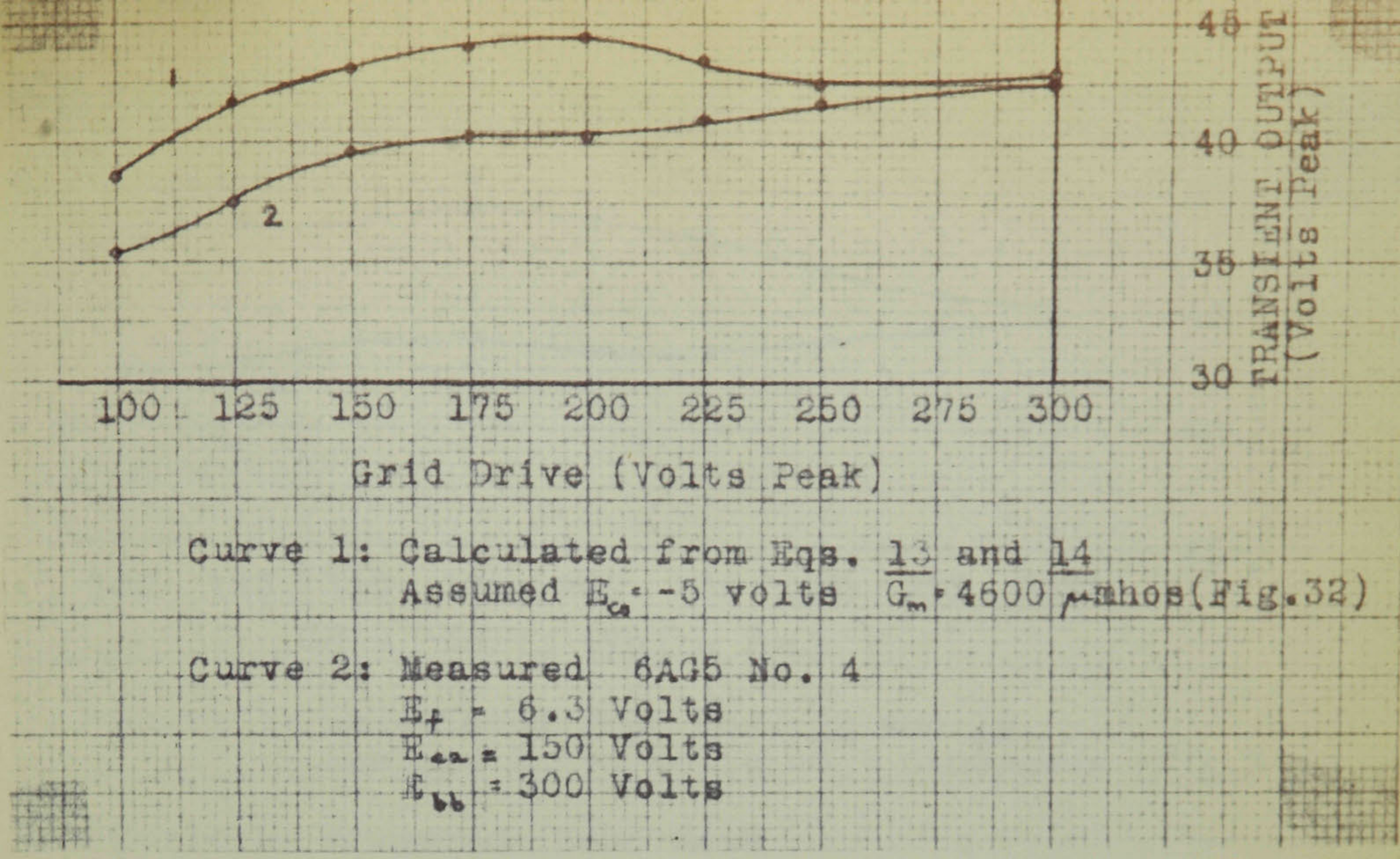


Fig .43: Transient Response of 6 mc/sec. LC Circuit(6AG5)

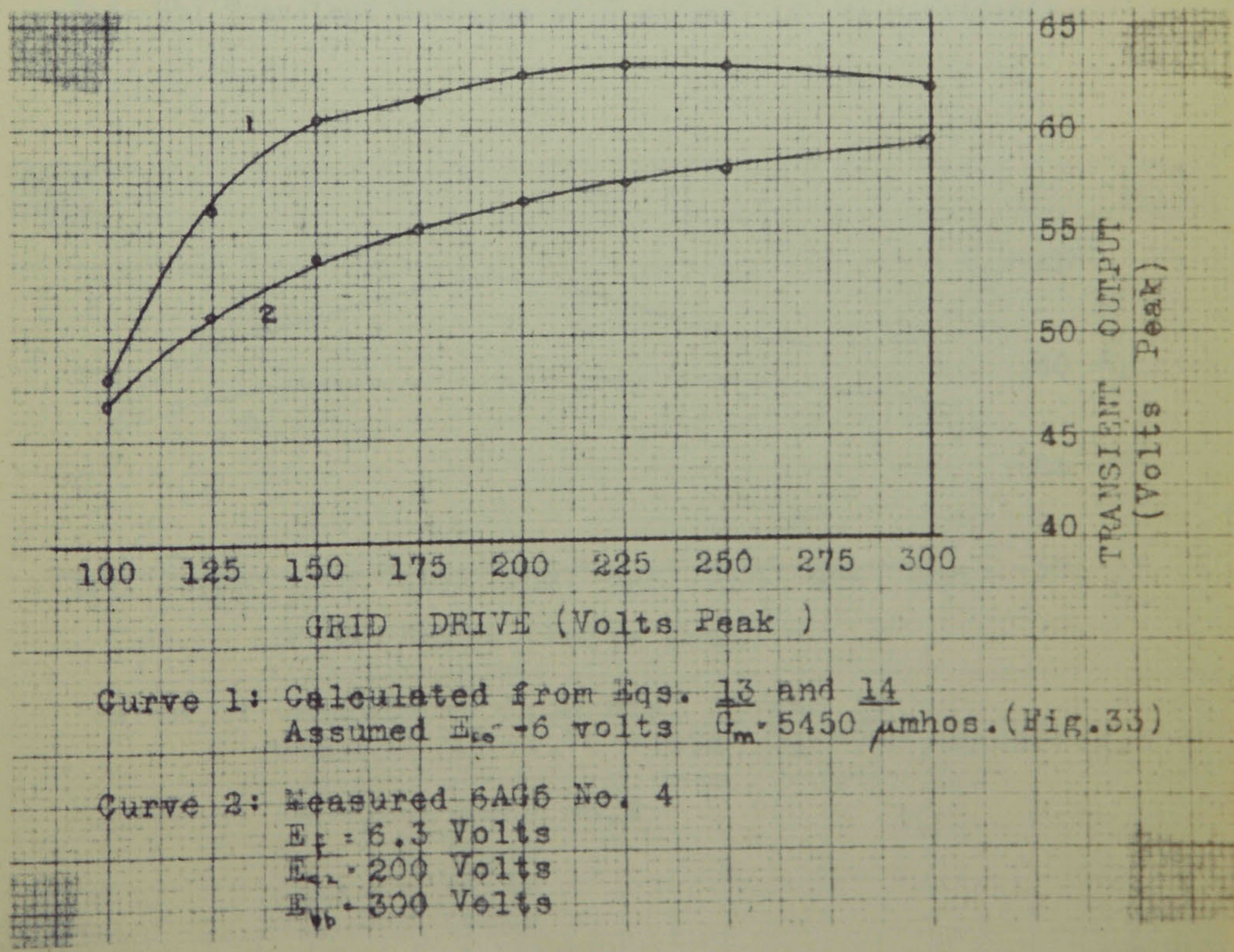


Fig. 44: Transient Response of 6 mc/sec. LC Circuit(6AG5)

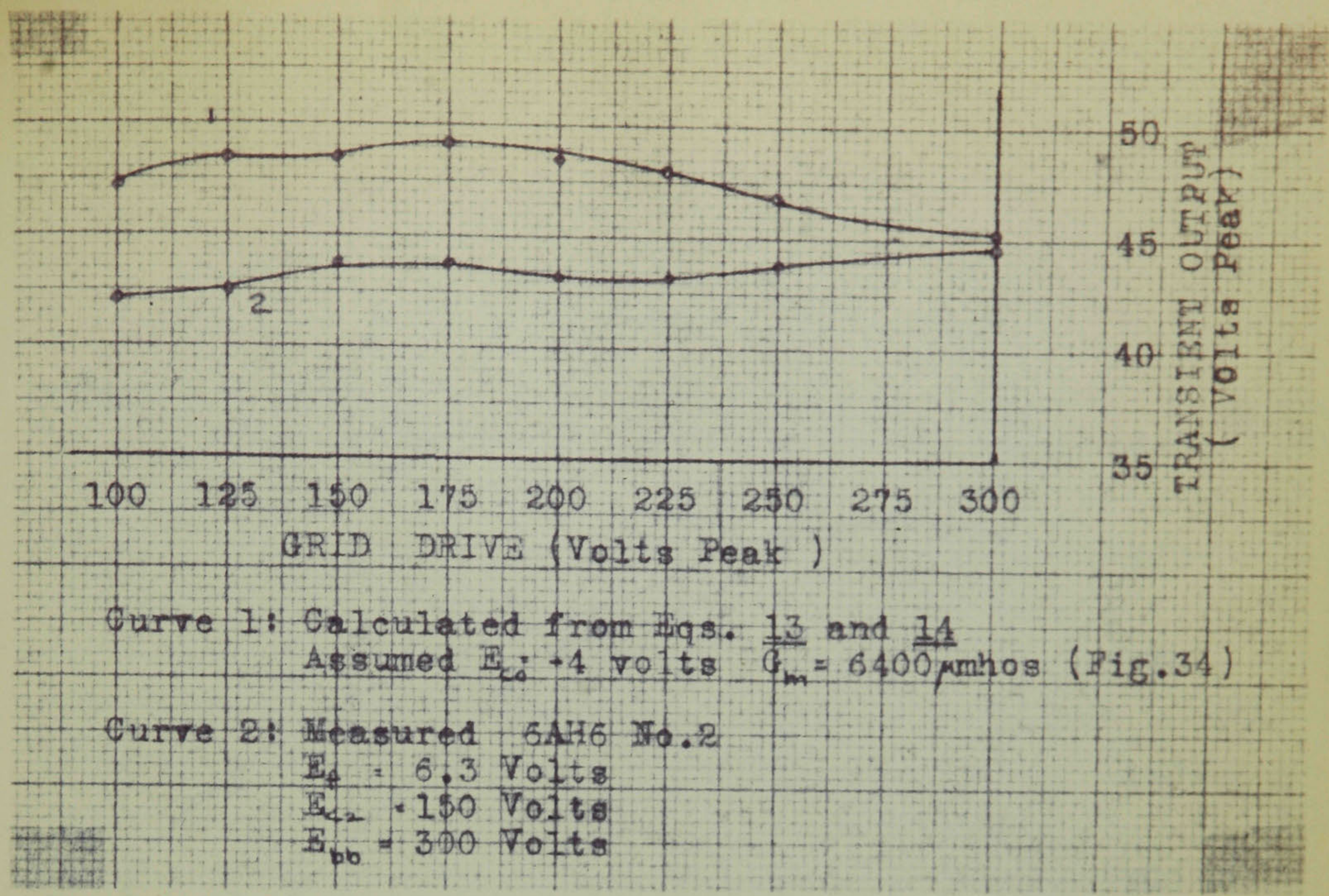


Fig. 45: Transient Response of 6 mc/sec. LC Circuit(6AH6)

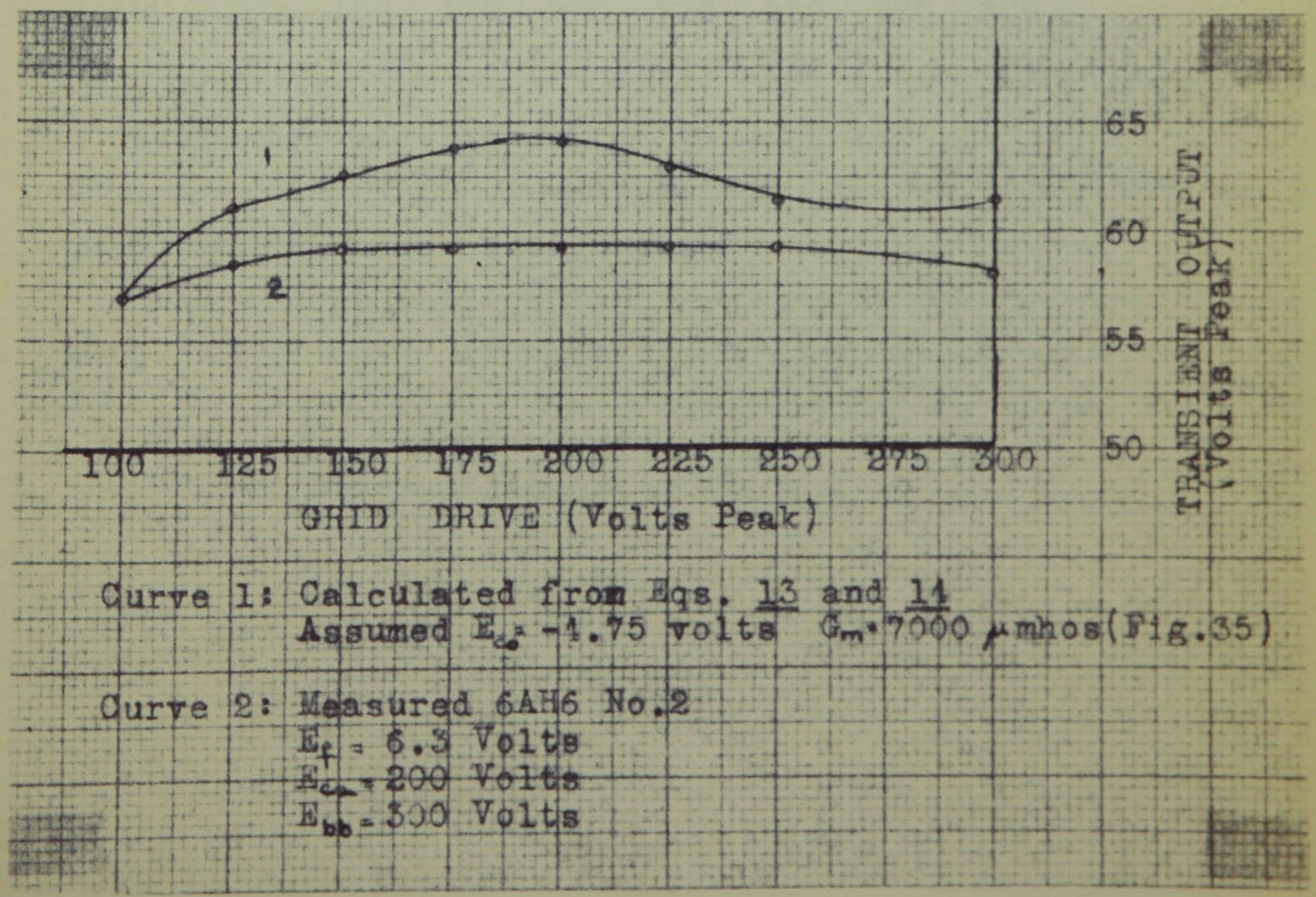


Fig. 46: Transient Response of 6 mc/sec. LC Circuit(6AH6)

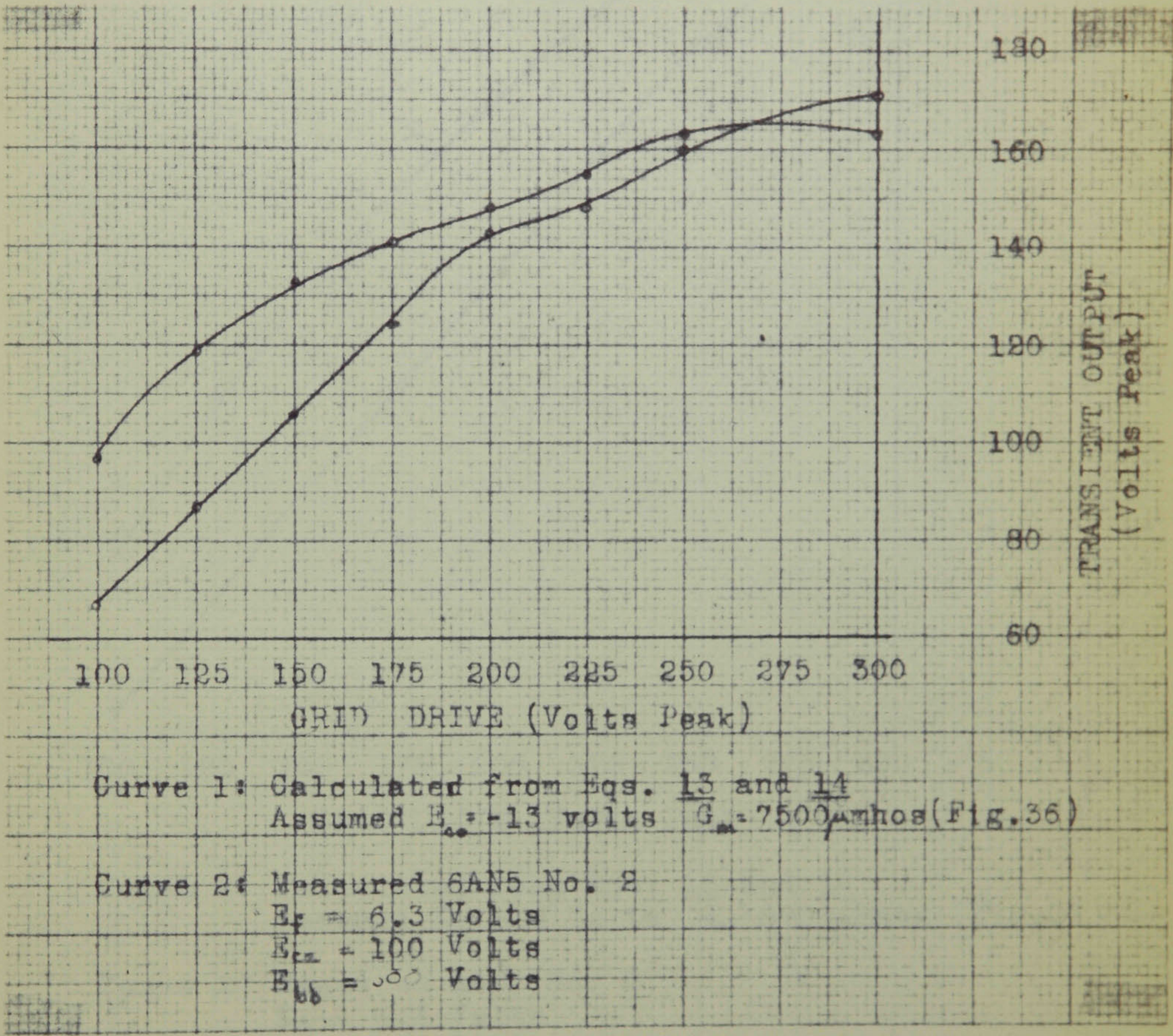


Fig. 47: Transient Response of 6 mc/sec. LC Circuit (6AN5)

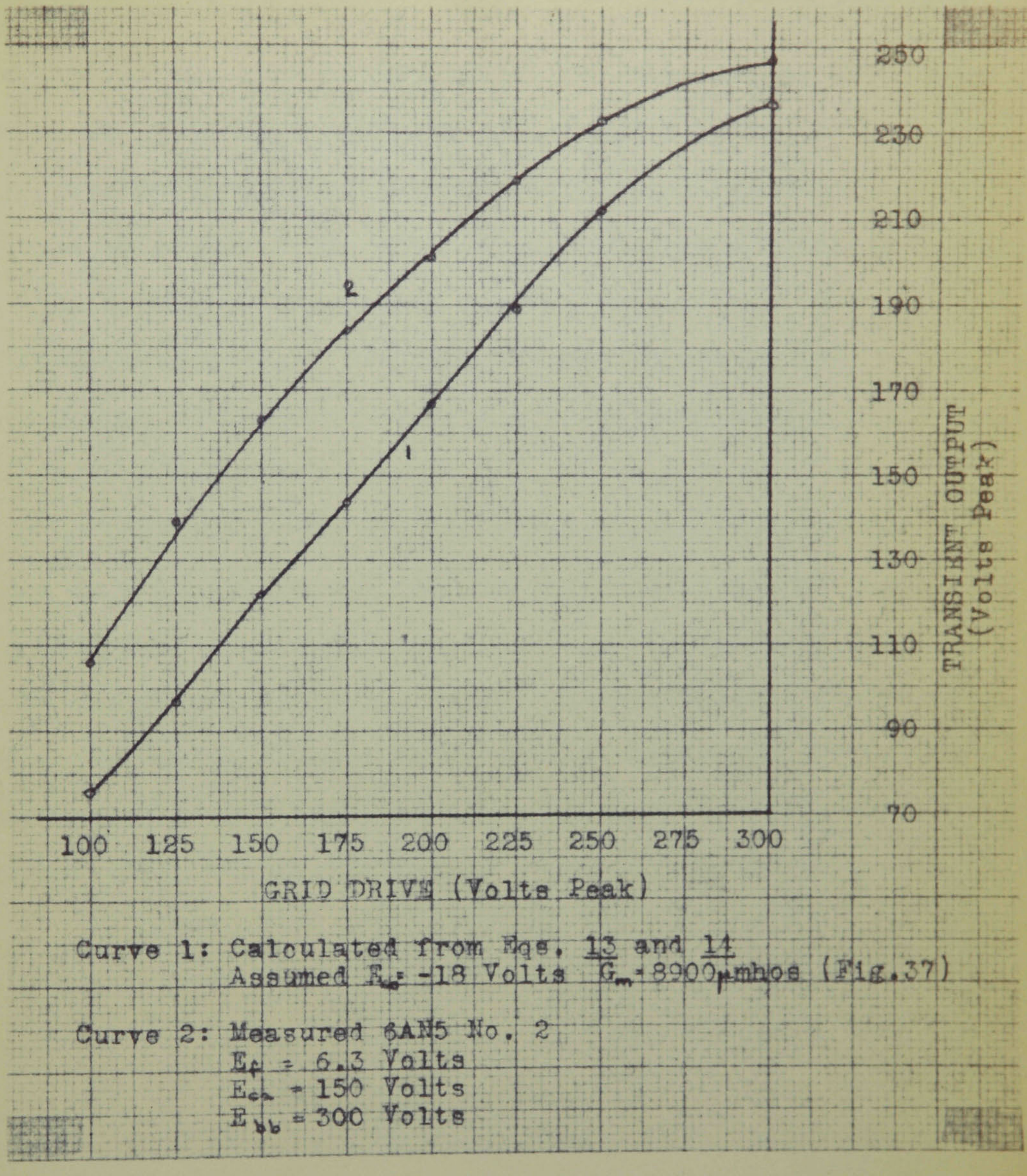


Fig. 48: Transient Response of 6 mc/sec. LC Circuit(6AN5)

9. THE PULSE GENERATOR

The final stage, number four in Fig. 23, was constructed as shown there to operate with a screen potential of 250 volts. Now from the 6AN5 plate current-grid bias characteristics shown as Figs. 36 and 37, and from the expression for the plate current in a pentode as given on page 15, two conclusions can be drawn. The grid cut-off potential of the valve is directly proportional to the screen potential and the maximum transconductance of the valve is not seriously altered by change in screen potential. Extrapolating the values for E_{co} and G_m, obtained from the curves referred to above, to cover the case where E_{sg} is 250 volts the new values are $E_{co} = 30$ volts and $G_m = 8{,}000$ micromhos. On this basis the conduction angle for a 200 volts input signal is $\frac{\rho}{2} = 32^\circ$ which is approximately the angle upon which the calculation on a 33 1/3 mc/sec. transient in section 3.5.6 was based. Furthermore the values used for the circuit elements in that calculation are those of the actual elements in the 33 1/3 mc/sec. tuned circuit as determined by the methods described in section 7.2.

When the circuit was operated under the conditions outlined the amplitude obtained at the desired frequency was 56 volts peak.

Because of a time jitter of the order of 1/200 the of the oscilloscope trace the light intensity and definition of the display were considered unsatisfactory for photographic recording. The waveshape observed resembled in every respect that shown in Fig. 22 for $\frac{2\pi\omega}{\theta} = \beta$ in which the first negative peak is almost identically one half of the subsequent positive peak.

10. CONCLUSIONS

Idealization of an actual plate current-grid bias characteristic has been accomplished, for the purpose of analysing the transient response of a tuned circuit, by assuming a control-grid cut-off potential equal to that determined by measurements and a transconductance equal to two-thirds of the maximum value for the valve.

The process defined is justified within the range studies by the satisfactory correlation between measured and calculated responses.

It is possible, with the idealized characteristic, to "design" from available engineering data a circuit for producing transient sinusoidal oscillations. Furthermore in the case considered it is possible to obtain a pulse generator, having a minimum of circuit complication, capable of producing a pulse having a base width of a hundredth of a microsecond and an amplitude of thirty volts.

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