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ELECTRONIC ASPECTS OF A FOUR CHANNEL TAPE RECORDER FOR USE IN MISSILE DATA ACQUISITION

by

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SUMMARY

A four channel tape recorder for vibration and other measurements within the audio frequency band was developed for use with quarter inch tape. A total of six recorders has been produced; they have been installed in missile rounds for obtaining vibration data during flight. Details of fully transistorized circuitry associated with this recorder are described.

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1. INTRODUCTION

The four channel tape recorder was developed in a form suitable for installation within a missile frame. As the requirement called for a recorder capable of recording vibration and other parameters during the flight of the missile, it was essential that the equipment be designed to withstand the accelerations it would be subject to during flight. Mr. H.A.H. Griffiths, also of the Instrumentation Group, handled the design of the mechanical aspects of the recorder, including the selection of suitable motors for driving the tape. The electronic equipment was mounted within shielded boxes which were attached to the recording deck. As the recording deck was suspended from the frame by vibration reducing mounts, the vibration of the deck itself was much less than that of the missile frame. A printed circuit form of construction was adopted for the various electronic circuits which were fully transistorized. To avoid the risk of movement under vibration all mounting screws and adjustable trimmers were sealed with glyptol prior to the installation of the recorder.

Because a single recording only was required from each recorder no erasing facilities were required. To assist in performance assessment of the recorder in the earlier rounds, a reference signal of known frequency and amplitude was recorded on one channel of the recorder at the same time as data was recorded on the other three channels. A tape speed of $3\frac{3}{4}$ inches per second was adopted for the earlier recorders.

2. RECORDING HEADS

Because of the relatively short time available for development, it was decided to use fairly low impedance heads which would present no major design difficulties in the amplifier output stage. (Ordering of heads had to be done before any detailed design work was started because of the short time schedule). No four track heads for use with quarter inch tape could be obtained in the time required. In any case crosstalk between adjacent tracks may well have proved to be excessive for direct recording using a single 4-track head. It was decided to use two quarter track stereo heads (of "Rola" manufacture) staggered along the tape, as these could be supplied in time. The consequent time displacement between the two pairs of tracks was not an embarrassment in this application.

Rola Co. wound the heads to a nominal inductance of 2mH but the difficulty in producing exactly identical gaps gave rise to an inductance tolerance of approximately $\pm 10\%$ on the nominal figure. The gap used in the

Rola heads was nominally 7/10 mil. The head design was such that each recording track was 40 mil wide and adjacent tracks were separated by a 30 mil wide safety lane (Refer to Fig.1).

3. BIAS ARRANGEMENT

Alternating current bias at a frequency of 55 Kc/s (approximately five times the highest audio frequency of interest) was used. The bias was derived from a conventional Hartley type oscillator employing a separate winding for the output (See Fig.2). Blick[1] uses a similar circuit in a bias supply for a magnetic tape recorder. Adjustment of the oscillator supply voltage, by means of a potentiometer, enabled the amplitude of oscillation to be varied. A trimmer capacitor which formed part of the capacitance which tuned with the oscillator transformer T_1 was used as a means of adjusting the frequency to the right value. The output of the basic oscillator was amplified before being fed to the recording head windings. The transformer T_2 in the output stage was tuned to 55 Kc/s by means of a parallel tuned circuit associated with a separate winding on the transformer. By using a separate winding of more turns than the secondary winding, the value of the tuning capacitance was reduced and the trimmer capacitor enabled the tuning to be varied over a sensible range.

Crosstalk at the higher recording frequencies can be quite a problem if all four channels receive bias from a single transformer. To keep the crosstalk down to an acceptable figure it was necessary to use coupling capacitors (for the recording head windings) which had reactances high compared with the head reactance at the highest audio frequencies to be recorded. In the circuit used the crosstalk, measured by feeding three of the head windings joined in parallel with a signal and monitoring the output across the fourth head winding, was found to be -43db at 10 Kc/s, a figure which was quite acceptable at this frequency. Crosstalk was correspondingly lower at the lower frequencies.

To enable the relative bias amplitude to be varied between channels (a necessity since the head winding inductances differ and hence require different amounts of bias) trimmer capacitors were added in parallel with each of the four coupling capacitors. As the coupling capacitors and head inductances have a considerable effect on the output transformer tuning it is essential that the transformer tuning be checked after any adjustment of the coupling capacitor trimmers.

The optimum bias was determined for a Rola head having an inductance of 2.30mH. Scotch No.200 double play magnetic tape was used for these measurements. A 1000 c/s signal at a constant level of 0.4 mA R.M.S. was recorded at various magnitudes of bias current using the above head. The relative replay response, obtained using a Model 5 Tandberg recorder, has been plotted in Fig.3. For the particular recording track considered, the optimum value of 55 Kc/s bias current was approximately 5 mA R.M.S. To provide current of this magnitude for each of the four channels, the output transformer secondary requirement was approximately 10V R.M.S. at 20 mA R.M.S. As the requisite value of bias current was known only very approximately at the design stage for the bias circuitry, the bias amplifier was designed to be capable of delivering 10 mA R.M.S. to each channel.

It is to be emphasized that the optimum bias value is dependent on the make of tape used. For BASF Type LGS26 double play magnetic tape the optimum value of bias current was somewhat higher (approximately 7.5 mA R.M.S.).

4. RECORD AMPLIFIER

4.1 Output Stage

General requirements for the transistor output stage are that current fed to the head windings be made proportional to the voltage signal at the input to the stage and that the frequency response over the requisite audio band be flat. Because of the wide variation of head reactance with frequency the voltage response across the head will rise with frequency for a flat current response characteristic. The current drive may be obtained from a conventional transistor amplifier which is coupled to the head by a series resistor the value of which is large compared with the head reactance over the frequency range of interest. A.C. coupling must be used in conjunction with the heads as no D.C. can be permitted to flow through the head windings. Two possibilities which immediately arise are transformer coupling from the collector of a conventional amplifier or capacitive coupling from a choke in the collector circuit. To obtain an adequate low frequency response these components are necessarily bulky, an aspect which is a disadvantage in an airborne recorder for which size and weight are to be kept as low as possible. An alternative to an inductive component in the collector circuit is a transistor acting virtually as a constant current source. In such an instance the signal currents would flow through the coupling capacitor to the load.

The general form of the output stage used is depicted in Fig. 4. Resistor R_2 is adjusted so that the voltage at the common collector point is about half the supply voltage. In this condition both transistors will be operating in their active regions. Capacitor C_2 is added so that transistor Q_2 and associated circuitry constitute a common base stage which has a higher output impedance than a common emitter stage. The output resistance of the output stage has been estimated in Appendix I. A value of 5.5 kilohm was obtained.

If a higher output resistance is required the lower transistor may be used in a common base configuration as depicted in Fig. 5. This arrangement was not deemed necessary in the present application.

It is essential that the coupling capacitor to the recording head winding have a very low leakage to ensure that no D.C. flows into the head. To enable the circuit to give the required low frequency performance a fairly large value capacitor is needed, and hence size considerations indicated that an electrolytic type should be used. It was found that a tantalum electrolytic was quite suitable.

Since the bias voltage appearing across the head winding is fairly high (approximately 4 volts R.M.S.) it is necessary to place a bias filter between the output of the amplifier and the head winding so that overloading of the amplifier may be prevented. A high Q parallel tuned circuit was used for this purpose. Mica capacitors were used in this tuned circuit since it was found that polyester type capacitors provided a poor Q at this frequency. The inductor has a tunable slug which allows the frequency of tuning to be varied over a range of about $\pm 2\%$.

4.2 Pre-Emphasis Circuit

The requisite frequency response shaping was achieved in a single stage of pre-amplification by suitable choice of components in the feedback loop of a common emitter amplifier. Blick[1] discusses a similar pre-emphasis circuit design. The recording amplifier response is similar to that employed in a Model 5 Tandberg recorder for a tape speed of $3\frac{3}{4}$ inches per second. The general form of the pre-emphasis circuit is depicted in Fig. 6. Referring to this figure we may write:

$$\text{Gain} = \frac{e_2}{e_1} \simeq -\frac{Z_F}{R_3}$$

where Z_F is the effective impedance of the feedback path.

At mid frequencies the feedback impedance is approximately R_1 and hence:

$$A_{MID} \simeq -\frac{R_1}{R_3}$$

Bass boosting may be achieved by suitable choice of C_2 (lift is increased as C_2 is decreased). At low frequencies the gain is given by

$$A_{LOW} = -\frac{R_1 + \frac{1}{j\omega C_2}}{R_3}$$

$$|A_{LOW}| = \frac{\sqrt{1 + (\omega C_2 R_1)^2}}{\omega C_2 R_3}$$

$$\simeq \frac{1}{\omega C_2 R_3} \quad \text{for} \quad \frac{1}{\omega C_2} \gg R_1$$

The parallel tuned circuit comprising L_1 and C_1 provides the high frequency pre-emphasis. In this instance the circuit is tuned to 11 Kc/s (just above the highest audio frequency of interest which in this application is 10 Kc/s). Resistor R_2 is used to dampen the resonant peak and hence controls the amount of high frequency pre-emphasis which is employed. The gain of the amplifier at high frequencies is given by

$$A_{HIGH} \simeq -\frac{1}{R_3\left(\frac{1}{R_2} + j\omega C_1 + \frac{1}{j\omega L_1}\right)}$$

$$|A_{HIGH}| = \frac{R_2}{R_3\sqrt{1 + \left\{\left(\omega C_1 - \frac{1}{\omega L_1}\right)R_2\right\}^2}}$$

at $f = \dfrac{1}{2\pi\sqrt{L_1 C_1}}$ (about 11 Kc/s for values of L_1 and C_1 chosen)

$$|A_{HIGH}| = \frac{R_2}{R_3}$$

which represents the peak of the frequency response curve.

4.3 The Complete Amplifier

The circuit of the complete amplifier is drawn in Fig.7 and the frequency response of this amplifier is given in Fig.8. An additional stage of amplification has been incorporated between the pre-emphasis stage and the output stage.

Adjustable components were added where needed. Both L_1 and L_2 can be varied by about $\pm 2\%$ by means of a tunable slug. L_1 can be trimmed to provide peaking at 11 Kc/s and L_2 can be trimmed to provide optimum bias filtering. A 2.5K Painton Flat Pot has been incorporated in the output stage to provide a means of balancing this stage.

4.4 Choice of Recording Level

A linearity check was performed on a Rola head having an inductance of 2.30 mH. The 55 Kc/s bias current was set to its optimum value of 5 mA R.M.S. for the Scotch No.200 double play tape used. A 1000 c/s signal was recorded at various levels and the relative replay response was measured using a Tandberg Model 5 as a replay device. Reference to the curve plotted in Fig.9 reveals that non-linearity becomes quite significant above a recording level of 0.5 mA R.M.S. It was decided in this case to set 100% signal at 1000 c/s to 0.45 mA R.M.S.

4.5 Pre-Amplifiers

Pre-amplifiers having a very high input impedance were required for use with piezo-electric transducers in the recording of missile vibrations during flight. To obtain a transducer response with a lower half power frequency of about 30 cycles per second the amplifier input impedance required for a Columbia gauge was about 10MΩ. The design of suitable pre-amplifiers for this purpose was carried out by G. Cadzow of the Systems Dynamics Group.

The following is a summary of some typical data on a Model 402 Columbia accelerometer (the type used in this application).

Crystal capacitance	= 400 PF
Cable capacitance	= 300 PF
Sensitivity (with cable connected)	= 137 mV/g
Resonant frequency	= 16.5 Kc/s

A crystal transducer may be represented approximately by the equivalent circuit of Fig.10(a) at low frequencies, where q is the charge and C_T is the transducer capacitance. The transducer cable capacitance C_c and the amplifier input resistance R have been included in the equivalent circuit of Fig.10(b).

Now $q = K a$

where q is the charge on the transducer

a is the acceleration of the transducer

K is a constant for a particular transducer and is known as the charge sensitivity.

Put $C = C_T + C_c$

We are concerned with the steady state frequency response of the system. Referring to Fig.10(b) we may write

$$e = \frac{iR}{1 + j\omega CR}$$

Assuming $a = A \sin \omega t$

$$i = \frac{dq}{dt}$$

$$= j\omega q \text{ in complex notation}$$

$$= j\omega Ka$$

$$\frac{e}{a} = \frac{j\omega RK}{1 + j\omega CR}$$

$$\left|\frac{e}{a}\right| = \frac{\omega RK}{\sqrt{1 + (\omega CR)^2}}$$

$$= \frac{K}{C} \frac{1}{\sqrt{1 + \left(\frac{1}{\omega CR}\right)^2}} \qquad \text{AMPLITUDE}$$

$$\varphi = \frac{\pi}{2} - \text{artan}\,\omega CR \qquad \text{PHASE}$$

For mid frequencies $\left|\frac{e}{a}\right| \simeq \frac{K}{C}$

The constant $\frac{K}{C}$ is the sensitivity of the transducer which in this example is equal to 137 mV/g.

It is readily seen that the lower half power frequency will be given by

$$\frac{1}{\omega_1 CR} = 1$$

$$f_1 = \frac{1}{2\pi RC}$$

For the particular case considered where C = 700 PF and R = 10 MΩ

$$f_1 = 22.8 \text{ c/s}$$

The maximum acceleration to be recorded was 30g peak. Sensitivities of Columbia gauges varied over quite a broad range. In a sample of 15 gauges the sensitivities varied from 90 mV/g to 147 mV/g. For the peak acceleration of 30g to be recorded, the range of input voltages (maximum) is 1.9 volt R.M.S. to 3.1 volt R.M.S.

A three stage direct coupled amplifier employing feedback from the final stage to the first stage (see Fig. 11) was used. Laakman[2] discusses a similar amplifier circuit employing direct coupling between collector and base of successive stages. The use of direct interstage coupling eliminates the need for many bias components and hence increases circuit simplicity.

Heavy D.C. feedback through the resistors R_5 and R_6 enables adequate D.C. stability to be obtained. Capacitor C_1 is made large enough to effectively bypass any signal frequencies and hence eliminate any signal frequency feedback via the biasing feedback resistor path. Feedback of the signal frequencies is achieved with feedback resistor R_7. Reduction of the overall amplifier gain at high frequencies and the removal of the tendency of the amplifier to become unstable at the higher frequencies are both achieved by virtue of the feedback capacitor C_2. A theoretical analysis of the amplifier enables the following to be determined

$$\text{D.C. Output Voltage } (V_o) = 0.6 + \frac{(V_{cc} - 0.6)(R_5 + R_6)}{\beta_1 R_2}$$

where β_1 is the D.C. current gain of transistor Q_1.

$$\text{Open Loop Voltage Gain} = A_i \frac{R_4}{R_1}$$

(i.e. gain with R_7 infinite)

where A_i is the amplifier current gain (the signal current into R_4 divided by the signal base current of Q_1).

$$\text{Closed Loop Voltage Gain} = \frac{R_7}{R_1}\left\{\frac{1}{1 + \dfrac{R_7}{|A_i|\ R_4}}\right\}$$

$$\text{Input Resistance} = R_1$$

$$\text{Output Resistance} = \frac{R_7}{|A_i| + \dfrac{R_7}{R_4}}$$

The following measurements were made on one of the pre-amplifiers.

D.C. current gain of Q_1 (β_1) :	40
Open loop current gain of amplifier (A_i):	-8.10^3
D.C. output voltage :	2.1V
Output resistance :	250 Ω
Frequency response :	within 2db 10 c/s - 10 Kc/s
Output signal :	1.25V R.M.S. max.

Supply variations :	no effect on closed loop performance -3.5V to -7.5V.
Closed loop voltage gain:	3.75
Noise referred to input :	4 mV R.M.S. (53.5 db below maximum input signal for transducer of lowest sensitivity)
D.C. current requirement:	1.35 mA

4.6 Amplifier Sensitivity Adjustment

The high impedance pre-amplifiers were coupled to the main amplifier via a sensitivity control as in Fig.12(a). The sensitivity was adjusted such that a head current of approximately 0.45 mA R.M.S. flowed at 1000 c/s for maximum signal into the pre-amplifier.

In some of the rounds the auto pilot output voltage was monitored by one channel of the recorder. As the signal level was quite high (70 volts peak to peak maximum) and the output impedance moderately low, no pre-amplifier was required. The arrangement depicted in Fig.12(b) was used as a sensitivity adjustment.

4.7 Reference Signal

To provide a calibrating and checking facility for the earlier rounds a reference tone of known frequency and known amplitude was recorded on one channel of the data recorder. A modified Colpitts type oscillator as drawn in Fig.13 was used to generate a 1000 c.p.s. sinusoidal voltage. Unlike the conventional Colpitts oscillator this circuit has, effectively, a high impedance in series with the base. Melehy, Mahmond, Reed and Myrie[3] state that such an impedance will help to linearise the waveform and improve the frequency stability. Oakes[4] and Gleason[5] discuss the design of this type of oscillator.

A simplified analysis of this oscillator is given in Appendix 2. It is shown in this analysis that the frequency of oscillation is given by

$$f = \frac{1}{2\pi\sqrt{L\dfrac{C_1 C_2}{C_1+C_2}}}$$

where L, C_1 and C_2 are shown in Fig.14(a).

4.8 Complete System Frequency Response

A frequency response curve has been plotted in Fig.15 for a complete record - replay system. For this response the high impedance pre-amplifier as described in Section 4.5 has been used as part of the recording amplifier system. Playback was performed using a Tandberg Model 5 (quarter track stereo recorder) Tape Recorder as a replay device. A recording level of 0.1 mA R.M.S. at 1000 c/s was used. The $\pm$ 3db band-width limits for this particular case are 30 c/s and 8000 c/s.

5. POWER SUPPLY

Power for the complete recorder electronics was derived from a Nickel-Cadmium rechargeable battery pack which was installed in the missile. Charge retention in this type of cell is extremely good and hence the battery can be fitted to the missile and left for a long period without any need for recharging. Twenty four Deac Type DKZ-1000 cells were used. At a current drain of about 1.5A (most of this current was absorbed in the recorder motors) the battery voltage dropped from an initial value of about 27.6V to a value of 26.4V after five minutes recording. As the requisite recording time was only of the order of five minutes such a small battery pack was quite suitable.

A -18V regulator (see Fig.16) was incorporated so as to maintain the supply voltage to the electronics constant as the battery discharged. The bias oscillator and the reference oscillator are both affected by supply voltage variation. A simple series type regulator with one additional stage of amplification was found to be adequate for this application. The 18K resistor connected from collector to emitter of the series transistor ensures that the supply will always "start".

Thermistor CZ2 has a negative temperature coefficient and is used to compensate for the amplitude variations which occur in the bias and reference oscillator. The amplitude of oscillation of both these oscillators increases with both supply voltage and with temperature. Reasonable temperature compensation can be obtained by causing the supply voltage to fall with increase in temperature. The following table compares the performance of the oscillators with and without temperature compensation.

TEMPERATURE RANGE 22.5°C - 50°C

PARAMETER	VARIATION WITHOUT COMPENSATION	VARIATION WITH COMPENSATION
BIAS AMPLITUDE	± 3.5%	± 0.75%
BIAS FREQUENCY	± 0.6%	± 0.75%
REFERENCE AMPLITUDE	± 3%	± 1%
REFERENCE FREQUENCY	± 0.2%	± 0.2%

The total battery drain for the electronics of a four channel data recorder was approximately 175 mA.

6. PHOTOGRAPHS

Photographs of the various units which constitute the electronics of the airborne tape recorder are given in Figs.17 to 22.

7. CONCLUSION

A total of six recorders employing the electronic circuitry described in this paper have been produced. These have been installed in missile rounds mainly to ascertain the levels and frequencies of the vibrations inherent in the missile and also to assess the levels of linear acceleration at boost and at missile impact. Although the recorders showed some evidence of mechanical malfunctioning useful data was obtained during trials. For detailed information on recorder performance under trials conditions reference 7 should be consulted.

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APPENDIX 1

OUTPUT RESISTANCE OF AMPLIFIER OUTPUT STAGE

The effective output resistance of the output stage will be the parallel combination of the output resistance of the common emitter stage employing transistor Q_1 and the common base stage employing transistor Q_2 (Refer to Figs. 4 and 7).

Using expression No. 11 given by Dewitt and Rossoff[6] on p. 126 for the output resistance of a common emitter stage in terms of equivalent tee parameters we may write:

$$R_o = \frac{R_c}{\beta} + (R_e + R_E)\,\frac{R_c + R_g}{R_b + R_g}$$

where R_o = output resistance

R_c = intrinsic collector resistance

β = common emitter current gain

$R_e = \frac{r_e}{2}$ where r_e is the intrinsic emitter resistance

R_E = external emitter resistance (referred to as R in Fig. 4).

$R_b = r_b + \frac{\beta r_2}{2}$ where r_b is the extrinsic base resistance

R_g = "generator" resistance or resistance as seen looking back from the input terminals (i.e. base of Q_1 and lower end of external emitter resistor).

The approximate D.C. operating conditions for transistor Q_1 (BCZ11) are :

I_c (D.C. collector current) = 14 mA

V_c (D.C. collector voltage) = 7 V

Under these conditions the following typical values apply:

r_b = 225 ohm

r_e = 2 ohm

β = 25

R_b = 250 ohm

R_e = 1 ohm (negligible in comparison with R_E, R_E = 120 ohm)

$\frac{R_c}{\beta}$ = 2.4 kilohm (slope of common emitter characteristic)

$\therefore R_c$ = 60 kilohm

R_g = 2 kilohm (note that the value of R_g depends on the setting of the input potentiometer R_2)

Substituting these values in the above equation for the output resistance gives

$$R_o = 5.7 \text{ kilohm}$$

The output resistance of the common base stage comprising transistor Q_2 (2N498) and associated components will be considerably higher than the output resistance of the common emitter stage.

According to Dewitt and Rosoff[6] (Expression No.2) the output resistance of a common base stage is given by :

$$R_o = \frac{R_c \left(\frac{R_e}{R_b} + \frac{1}{\beta} + \frac{R_g}{R_b} \right)}{1 + \frac{R_g}{R_b}}$$

R_g in this case is the external emitter resistance of value 120 ohm.

Since $R_g \gg R_e$ and $R_g \gg \frac{R_b}{\beta}$

$$R_o \simeq \frac{R_c}{1 + \frac{R_b}{R_g}}$$

The D.C. operating current and voltage are approximately the same as for Q_1.

Typical values of R_b and R_c are $R_b = 250$ ohm and $R_c = 430$ kilohm. (These figures are obtained from data for $V_c = 30V$ and $I_c = 30$ mA and hence the value of R_c under the actual operating conditions will be somewhat higher than the figure given here).

Substituting in the simplified equation for the output resistance we obtain :

$$R_o = \frac{R_c}{3.1} = \frac{430}{3.1} = 139 \text{ kilohm}$$

This value of resistance is much higher than that for Q_1.

Hence the output resistance will be the combined resistance made up of 5.7 kilohm in parallel with 139 kilohm.

Output Resistance of Output Stage = 5.5 kilohm.

APPENDIX 2

ANALYSIS OF REFERENCE OSCILLATOR

A small signal equivalent circuit of the oscillator of Fig.14(a) is given in Fig.14(b). The effect of collector resistance has been neglected in this analysis. It has also been assumed that the intrinsic emitter and base resistances are small enough compared with the external resistors to be neglected. The resistance R_2 is equal to $\dfrac{R_3 R_4}{R_3+R_4}$

If a mesh analysis is carried out on the simplified circuit of Fig.14(b) the following equations result.

$$i_1 (R_1 + R_2) + i_2 (-R_1) = 0$$

$$i_1 (-R_1) + i_2\left(R_1 + \frac{1}{j\omega C_2}\right) + i_3 \left(-\frac{1}{j\omega C_2}\right) = 0$$

$$i_1\left(\frac{\beta}{j\omega C_1}\right) + i_2 \left(-\frac{1}{j\omega C_2}\right) + i_3 \left(\frac{1}{j\omega C_1} + \frac{1}{j\omega C_2} + j\omega L\right) = 0$$

For these equations to hold the determinant must equal zero.

$$\begin{vmatrix} R_1 + R_2 & -R_1 & 0 \\ -R_1 & R_1 + \dfrac{1}{j\omega C_2} & -\dfrac{1}{j\omega C_2} \\ \dfrac{\beta}{j\omega C_1} & -\dfrac{1}{j\omega C_2} & \dfrac{1}{j\omega C_1} + \dfrac{1}{j\omega C_2} + j\omega L \end{vmatrix} = 0$$

After simplication we obtain :

$$(R_1 + R_2)\frac{L}{C_2} - \Big\{(\beta + 1) R_1 + R_2\Big\}\frac{1}{\omega^2 C_1 C_2}$$

$$+ j R_1 R_2\left(\omega L - \frac{1}{\omega C_1} - \frac{1}{\omega C_2}\right) = 0$$

Equating of the imaginary part of this equation to zero will give the frequency of oscillation and equating of the real part to zero will give the condition for oscillation to be just possible.

$$\omega L - \frac{1}{\omega C_1} - \frac{1}{\omega C_2} = 0$$

$$\omega^2 = \frac{C_1 + C_2}{L C_1 C_2}$$

$$f = \frac{1}{2\pi\sqrt{L \frac{C_1 C_2}{C_1+C_2}}}$$

From this expression it can be seen that the choke resonates with the effective series capacitance of C_1 and C_2.

Also we have

$$(R_1 + R_2) \frac{L}{C_2} - \left\{ (\beta + 1) R_1 + R_2 \right\} \frac{1}{\omega^2 C_1 C_2} = 0$$

Assuming that $\omega^2 = \dfrac{C_1 + C_2}{L C_1 C_2}$

$$(R_1 + R_2) \frac{1}{C_2} - \left\{ (\beta + 1) R_1 + R_2 \right\} \frac{1}{C_1 + C_2} = 0$$

$$\frac{C_1}{C_2} = \frac{\beta R_1}{R_1 + R_2}$$

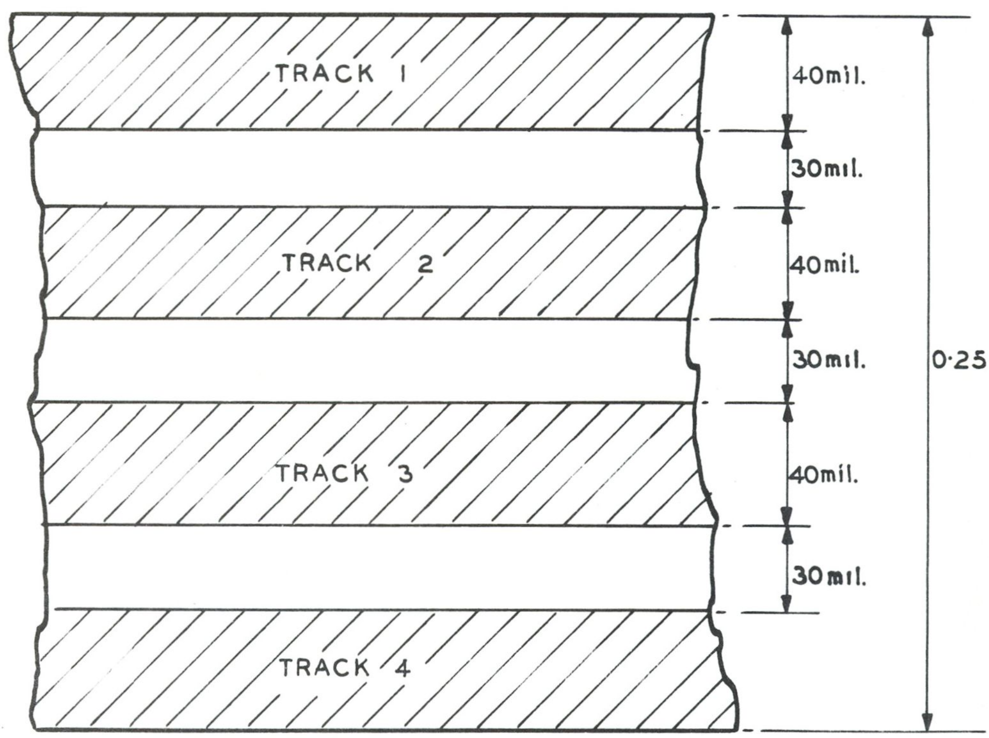
For oscillation to occur

$$\beta \gg \frac{C_1}{C_2} (1 + \frac{R_2}{R_1})$$

In the circuit used $C_1 \simeq C_2 = 0.47\,\mu F$

$$R_1 \simeq 3.8\ K\Omega \qquad R_2 = \frac{27.39}{66}\ K\Omega = 16.0\ K\Omega$$

$\beta > 5.2$ for oscillation to occur.

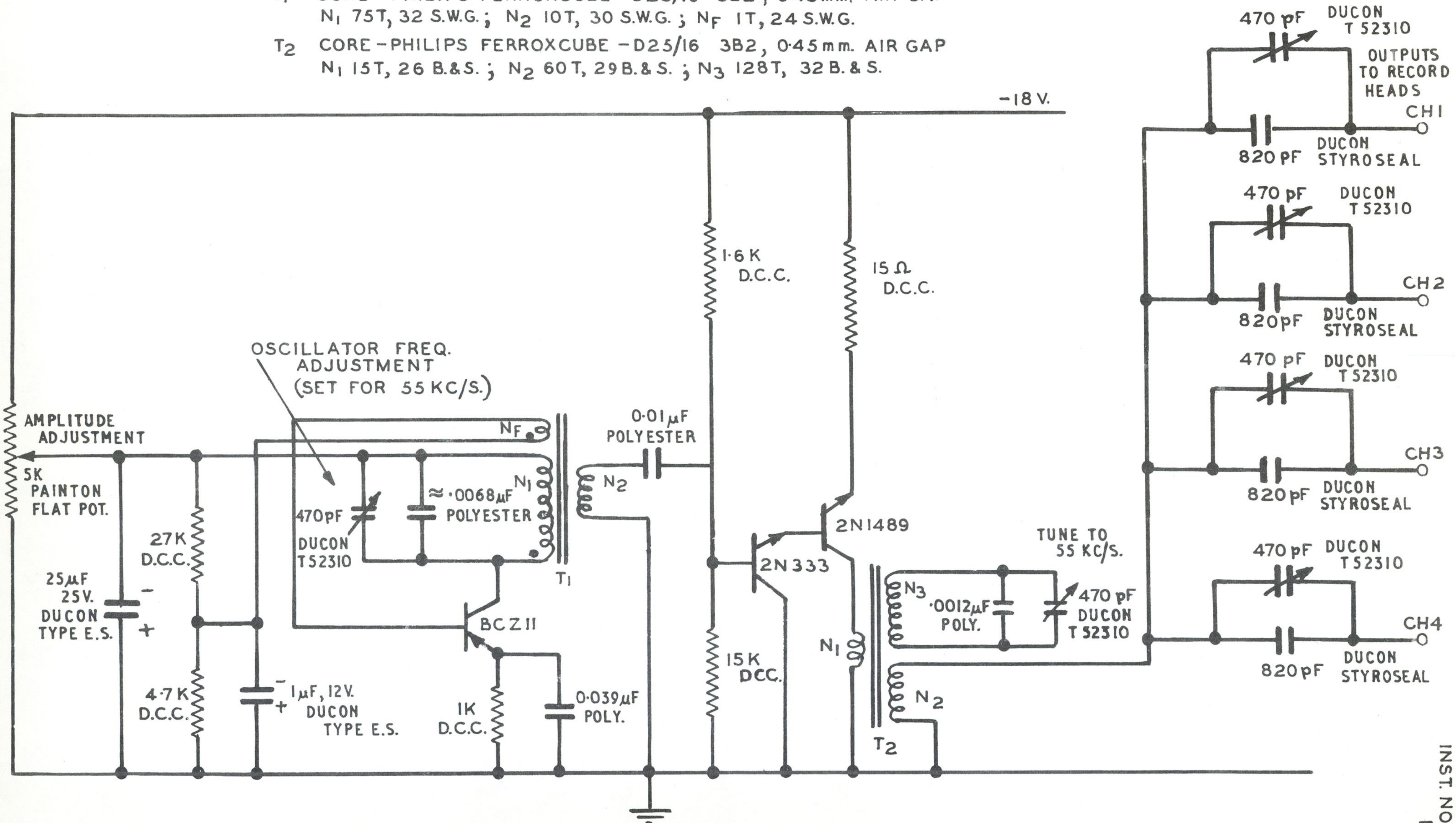


TRACK CONFIGURATION

TRANSFORMER DETAILS :

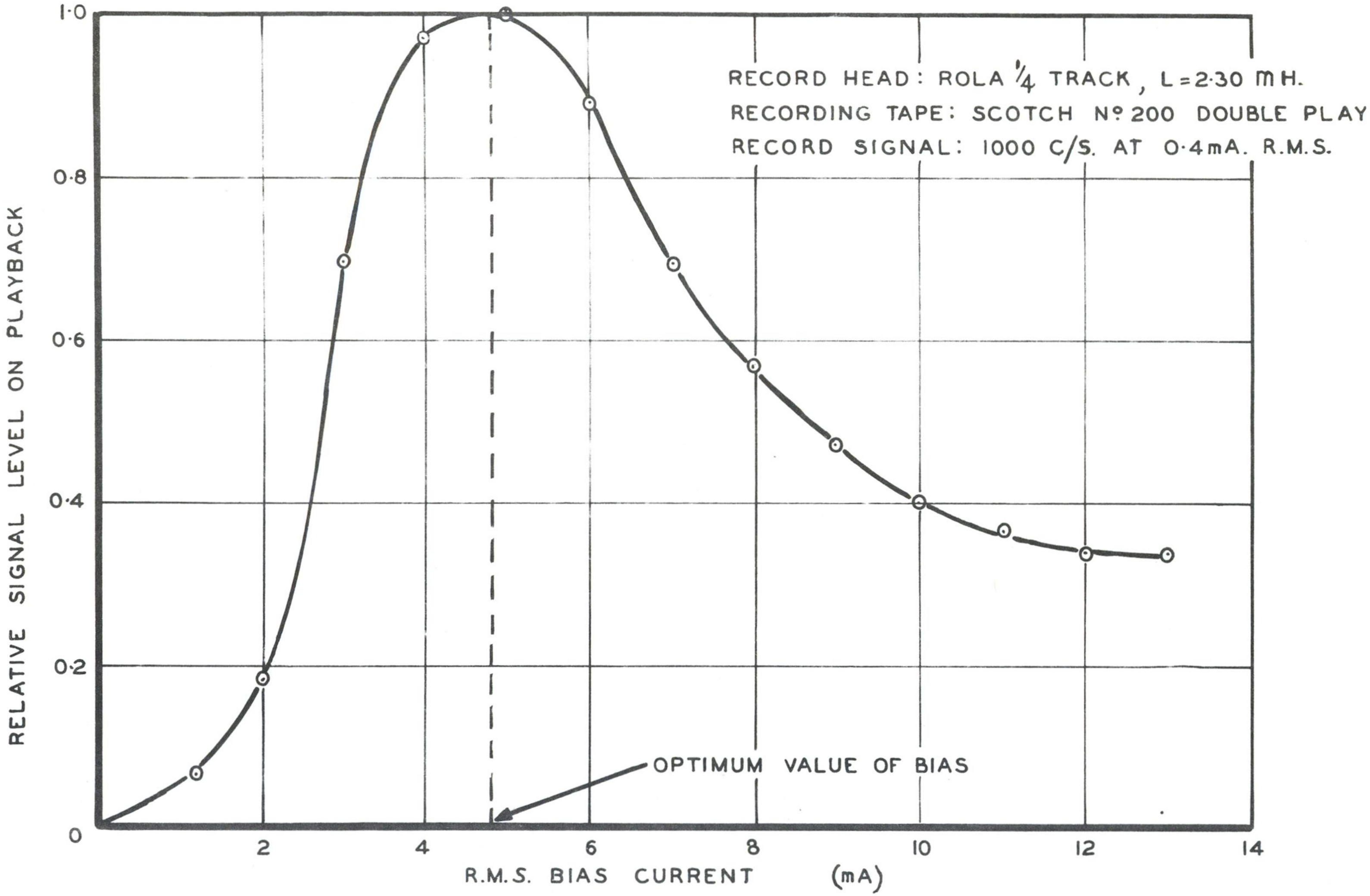
T_1 CORE – PHILIPS FERROXCUBE – D25/16 3B2, 0·45 mm. AIR GAP
N_1 75T, 32 S.W.G.; N_2 10T, 30 S.W.G.; N_F 1T, 24 S.W.G.

T_2 CORE – PHILIPS FERROXCUBE – D25/16 3B2, 0·45 mm. AIR GAP
N_1 15T, 26 B.&S.; N_2 60T, 29 B.&S.; N_3 128T, 32 B.&S.



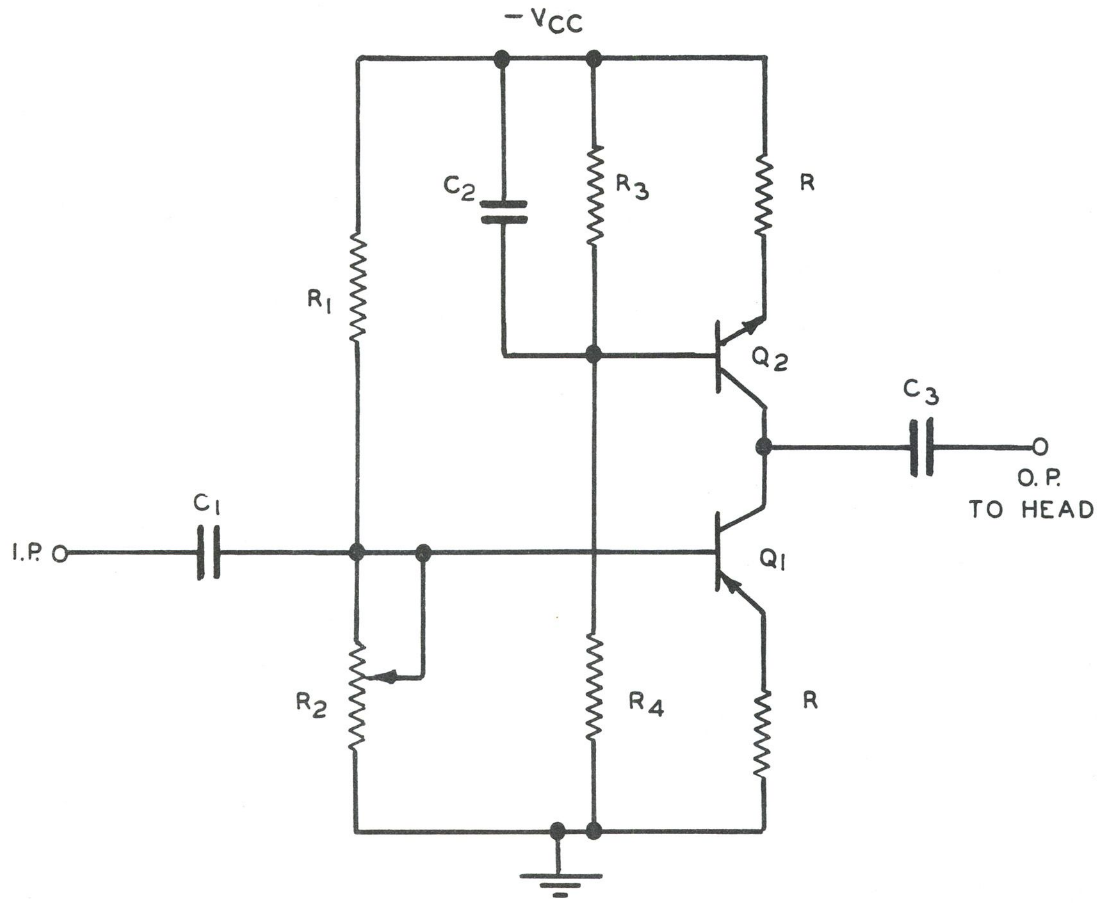
BIAS OSCILLATOR

INST. NOTE 64
FIG. 2

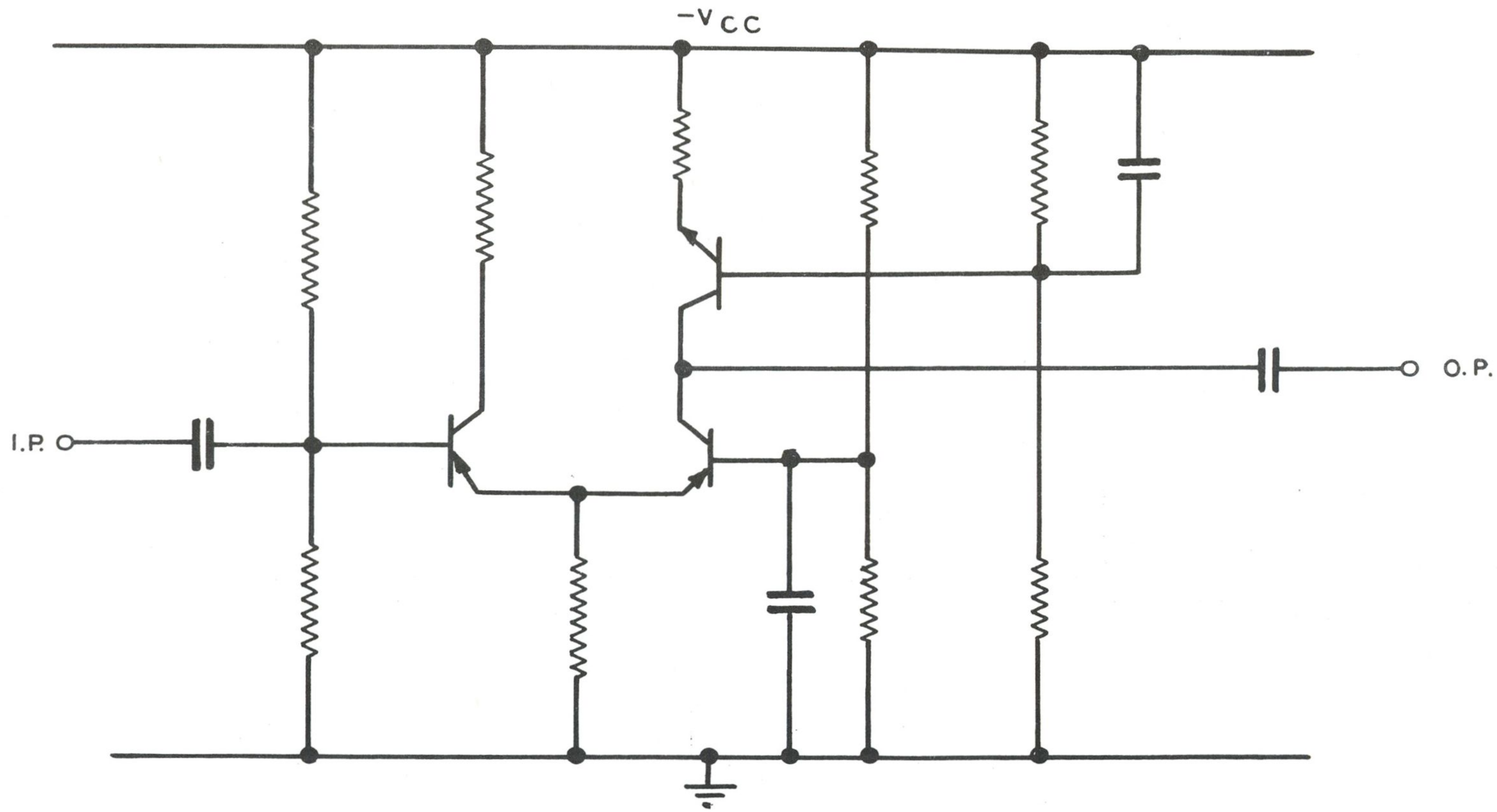


OPTIMUM BIAS CURVE

INST. NOTE 64
FIG. 3

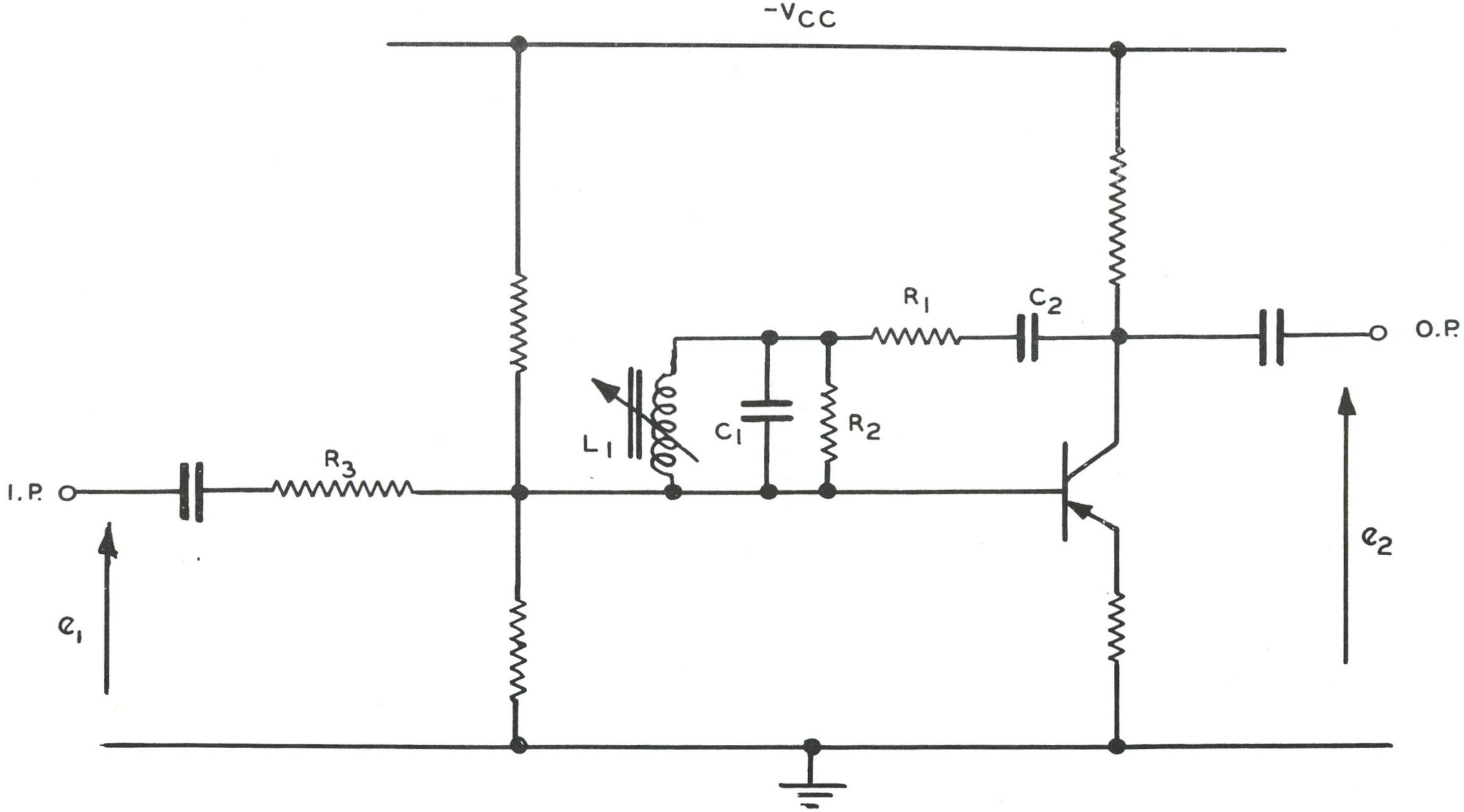


CIRCUIT OF OUTPUT STAGE



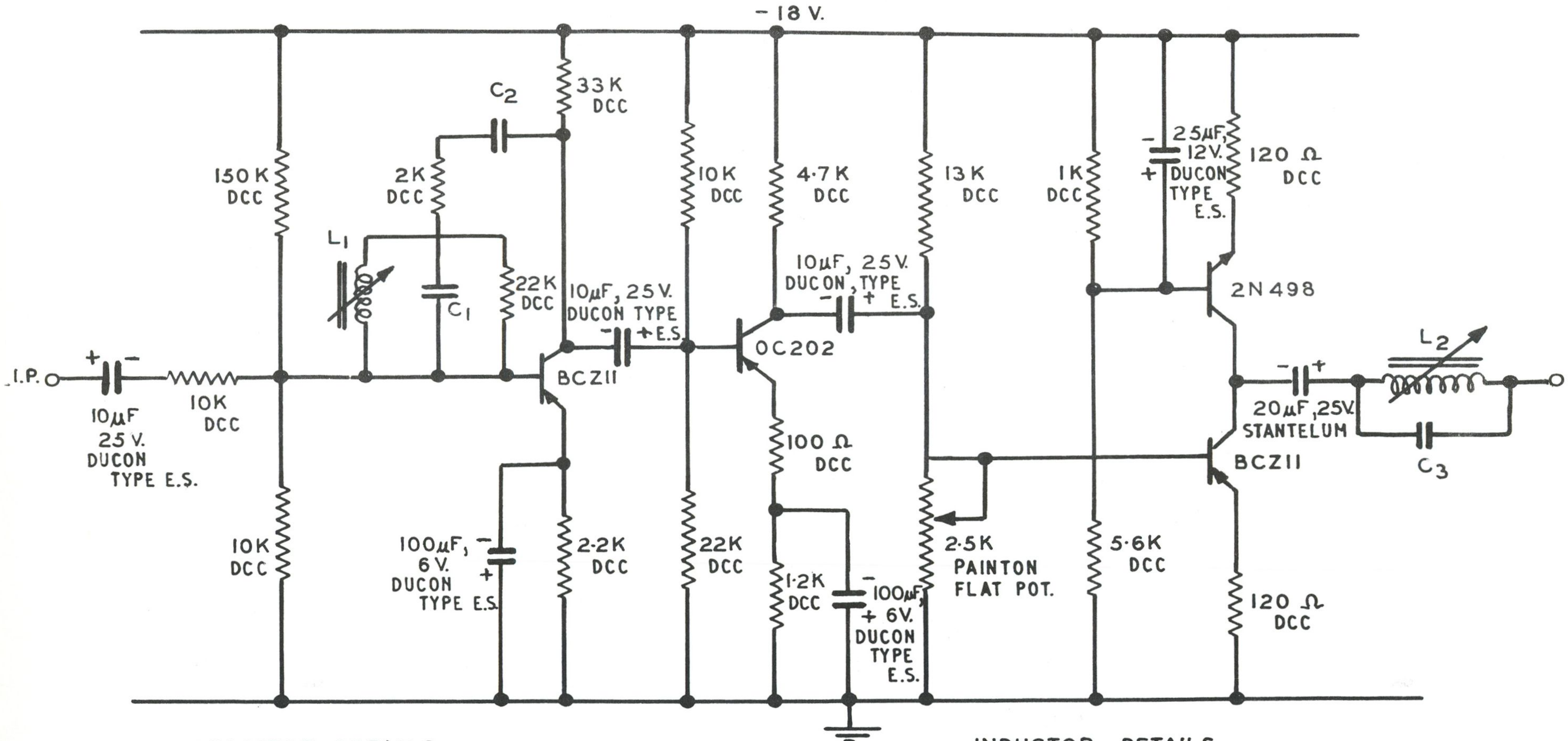
OUTPUT STAGE WITH HIGH OUTPUT IMPEDANCE

INST. NOTE 64
FIG. 5



PRE-EMPHASIS CIRCUIT

INST. NOTE 64
FIG.6



CAPACITOR DETAILS

C_1, C_2 & C_3 TO BE SELECTED TO 1%

C_1 0·0018μF (0·0015μF PHILIPS POLYESTER + DUCON STYROSEAL TRIMMER)

C_2 1·2μF (1·0μF + 0·15μF PHILIPS POLYESTER)

C_3 0·0038μF (0·0022μF DUCON STYROSEAL + 0·0015μF T.C.C. MICA)

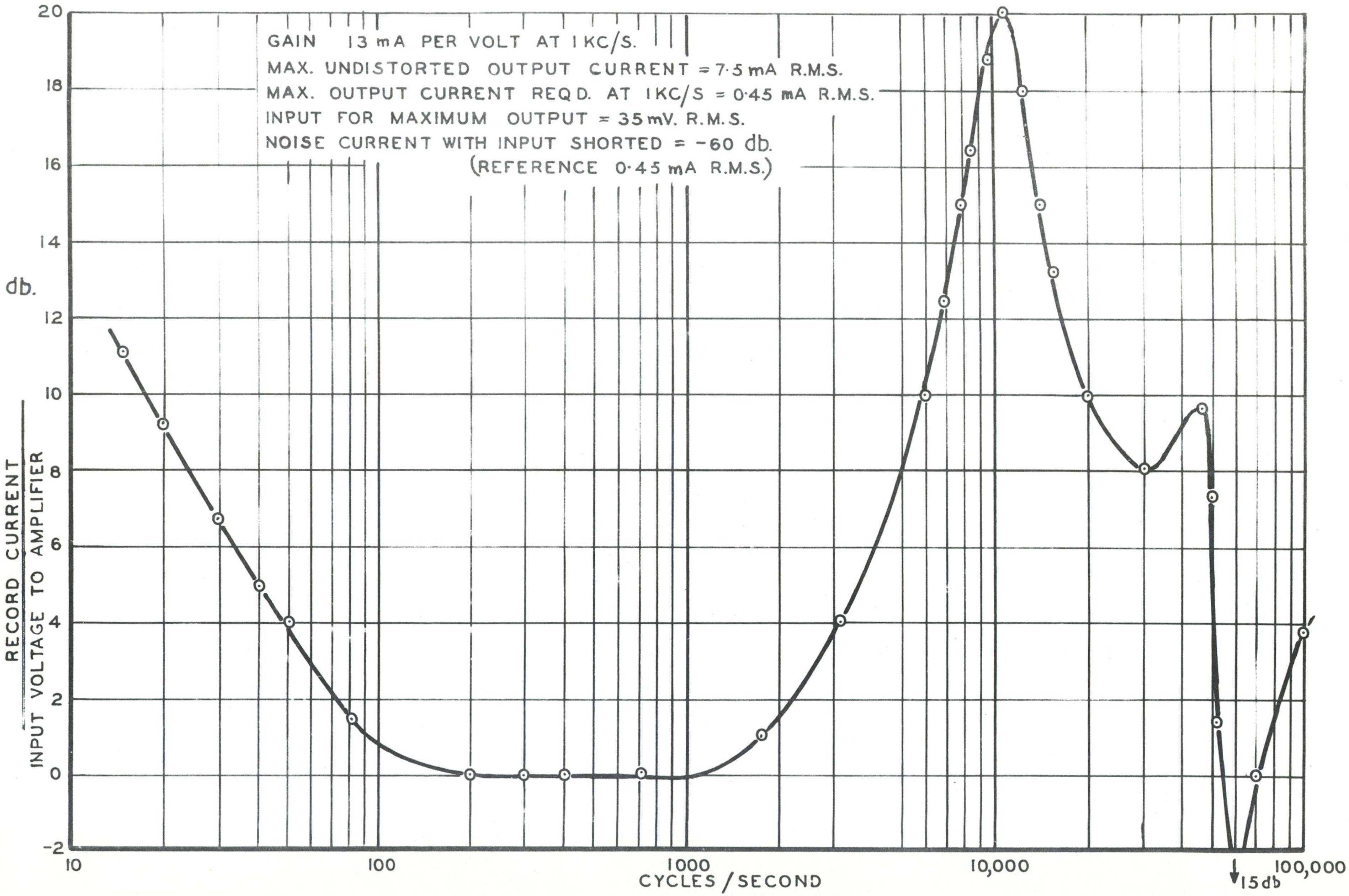
INDUCTOR DETAILS

L_1 CORE; PHILIPS K3-000-46
AIR GAP; 0·3mm.
WINDING; 955 T. 37 B.&S.
INDUCTANCE; 100 mH AT 1KC/S.

L_2 CORE; PHILIPS K3-000-46
AIR GAP; 0·3mm
WINDING; 140 T. 29 B.&S.; INDUCTANCE; 2·1mH AT 1KC/S.

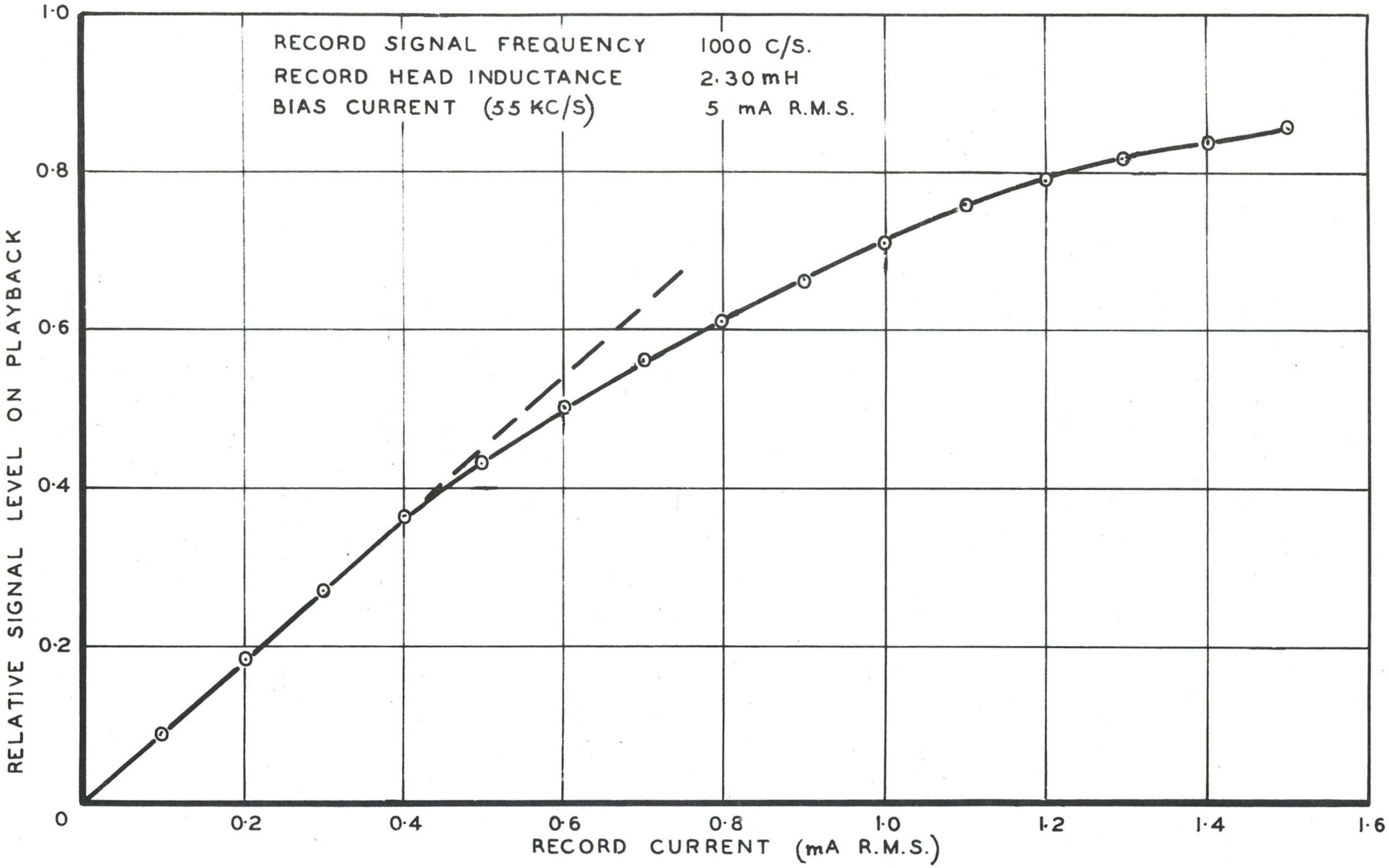
CIRCUIT OF COMPLETE AMPLIFIER

INST. NOTE 64
FIG. 7



FREQUENCY RESPONSE OF COMPLETE AMPLIFIER (EXCLUDING HIGH IMPEDANCE PRE-AMPLIFIER)

INST. NOTE 64
FIG. 8



RECORD HEAD LINEARITY CURVE

INST. NOTE 64
FIG. 9

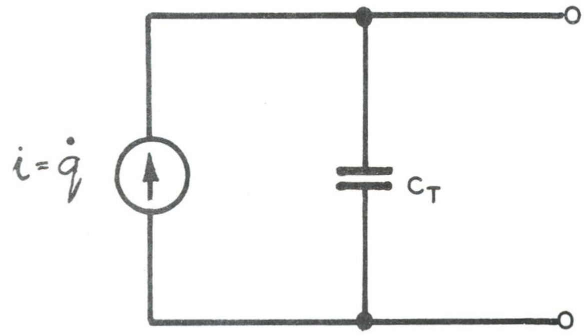


FIG. 10a EQUIVALENT CIRCUIT OF PIEZO-ELECTRIC TRANSDUCER AT LOW FREQUENCIES

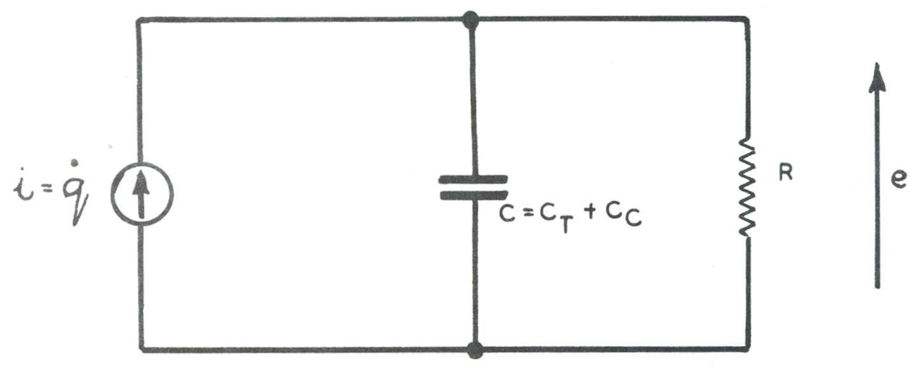
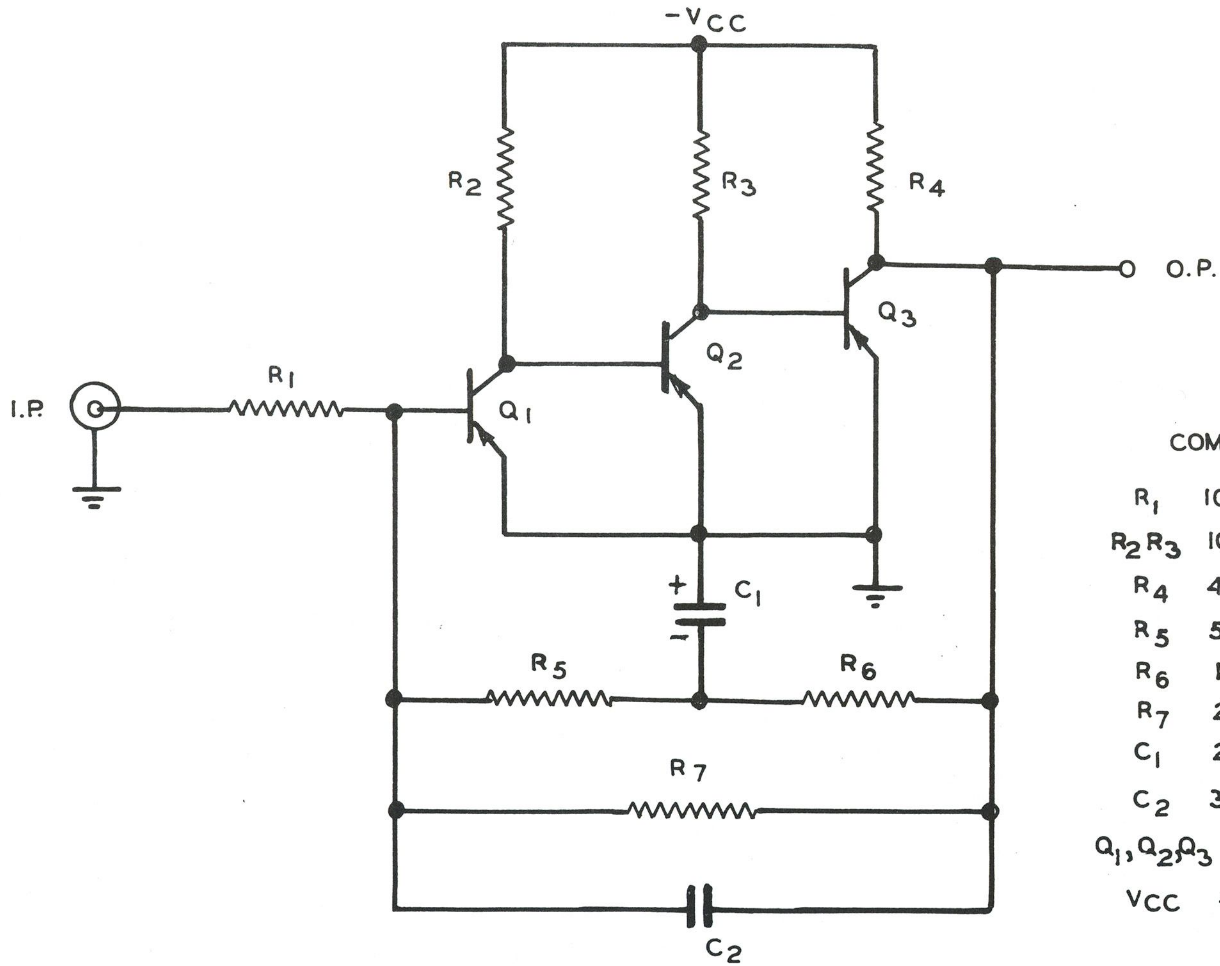


FIG. 10b EQUIVALENT CIRCUIT OF PIEZO-ELECTRIC TRANSDUCER WHICH INCLUDES CABLE CAPACITANCE & INPUT RESISTANCE OF AMPLIFIER



COMPONENT VALUES

R_1	10 MΩ	(RC7-K)
$R_2 R_3$	10 KΩ	(RC7-K)
R_4	4·7 KΩ	(RC7-K)
R_5	51 KΩ	(DCC)
R_6	100 KΩ	(DCC)
R_7	2·4 MΩ	(DCC)
C_1	25 μF, 12 V.	(SME)
C_2	3·3 pF,	(P100K)
Q_1, Q_2, Q_3	OC202	
VCC	4·5 V.	

PRE-AMPLIFIER CIRCUIT

INST. NOTE 64
FIG. 11

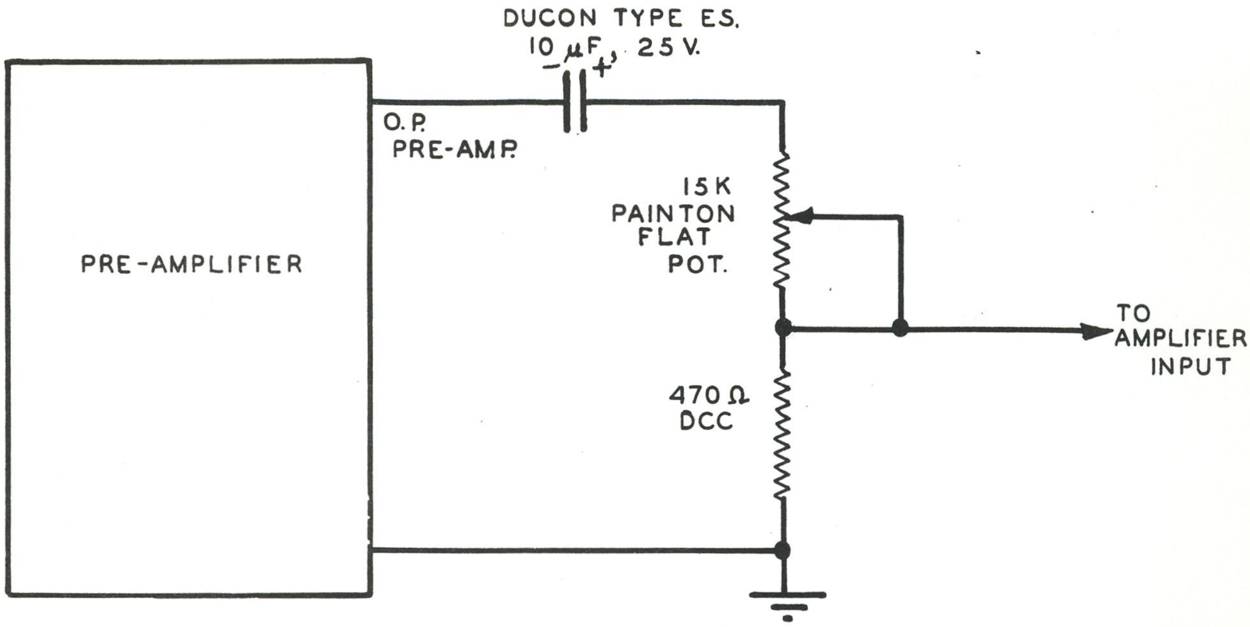


FIG. 12a CONNECTION OF PRE-AMPLIFIER TO MAIN AMPLIFIER

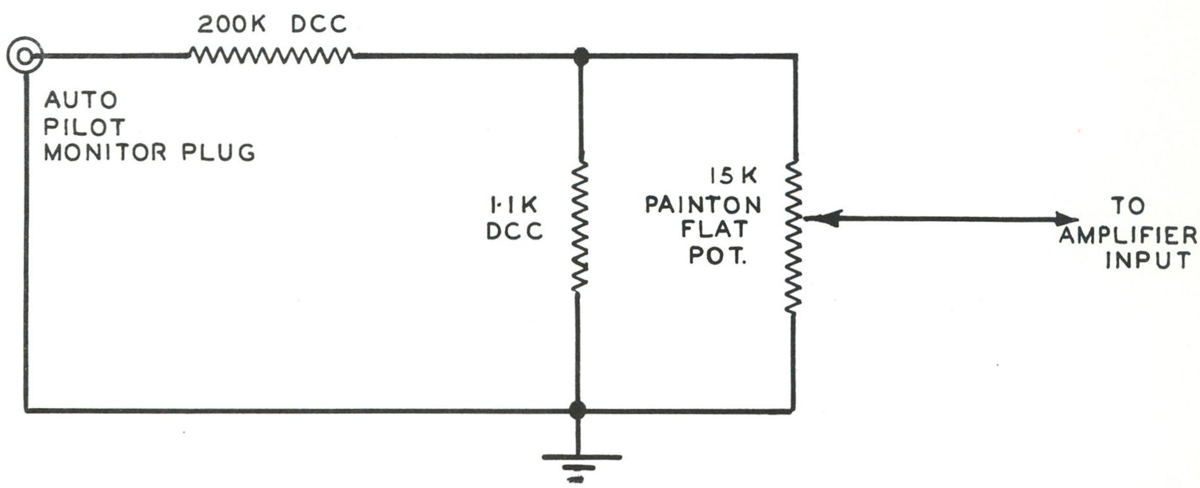
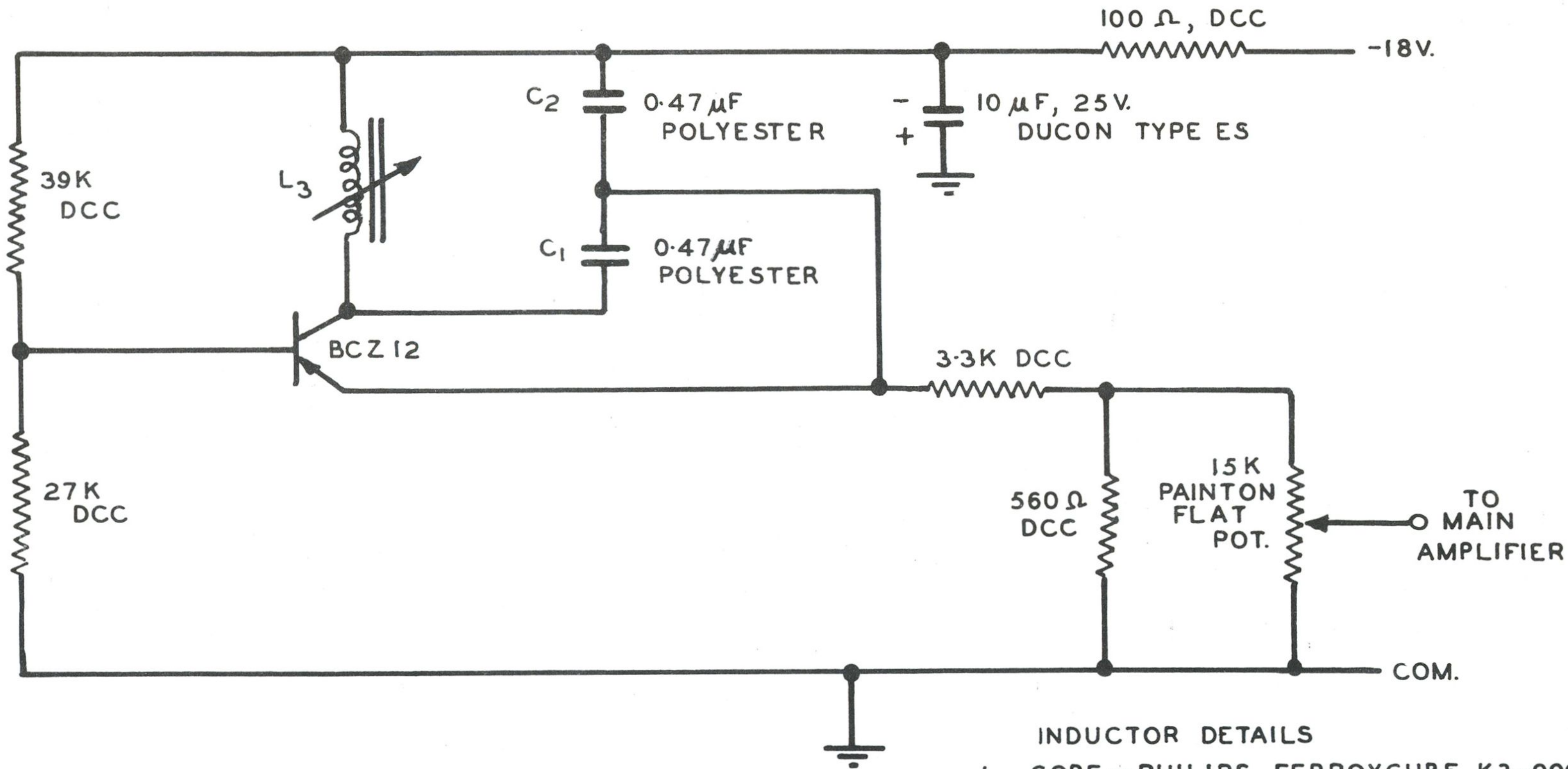


FIG. 12b CONNECTION OF AUTO PILOT MONITOR TO MAIN AMPLIFIER



INDUCTOR DETAILS

L_3 CORE – PHILIPS FERROXCUBE K3-000-46

AIR GAP ; 0·3mm.

955 T , 37 B. & S.

100 mH AT 1KC/S

1 KC/S REFERENCE OSCILLATOR

INST. NOTE 64
FIG. 13

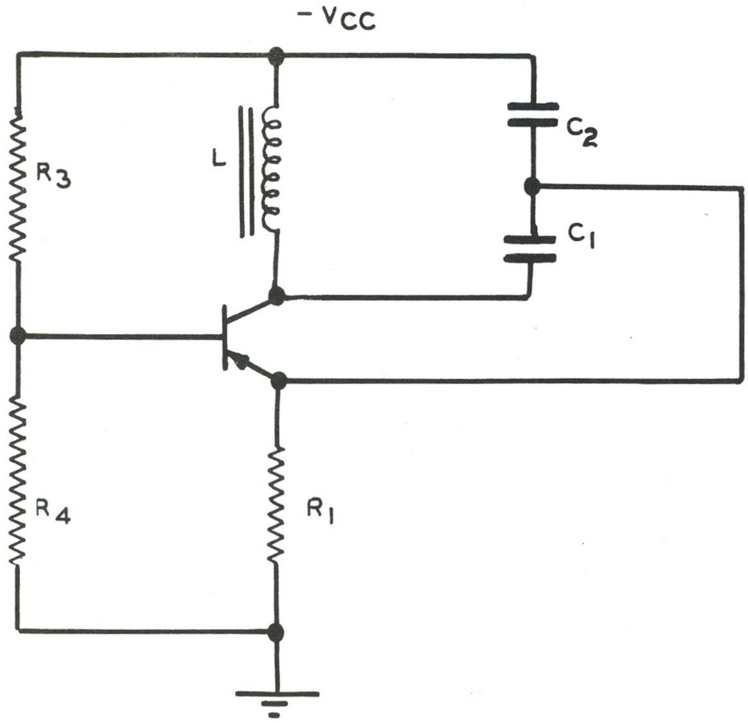


FIG. 14a GENERAL FORM OF OSCILLATOR CIRCUIT

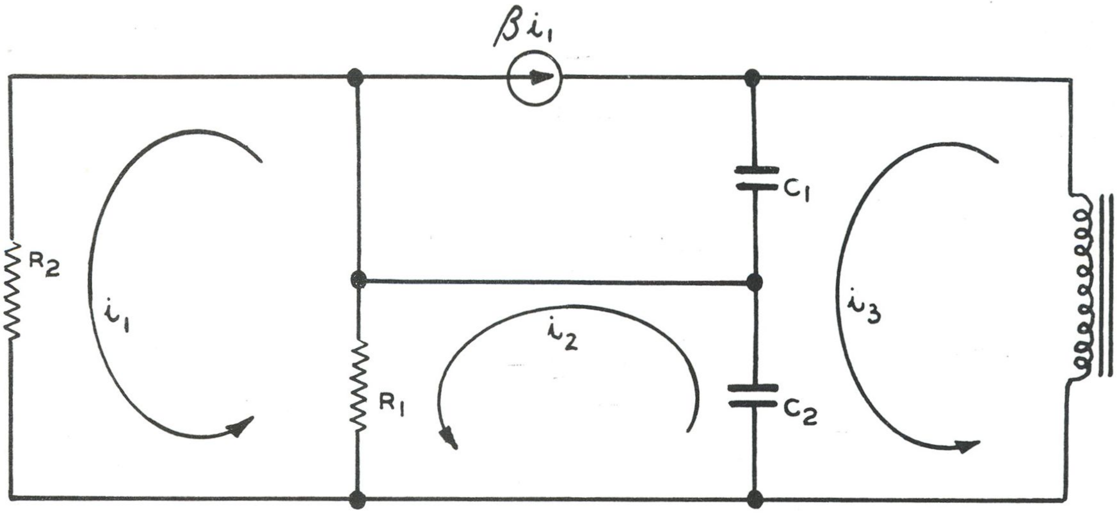
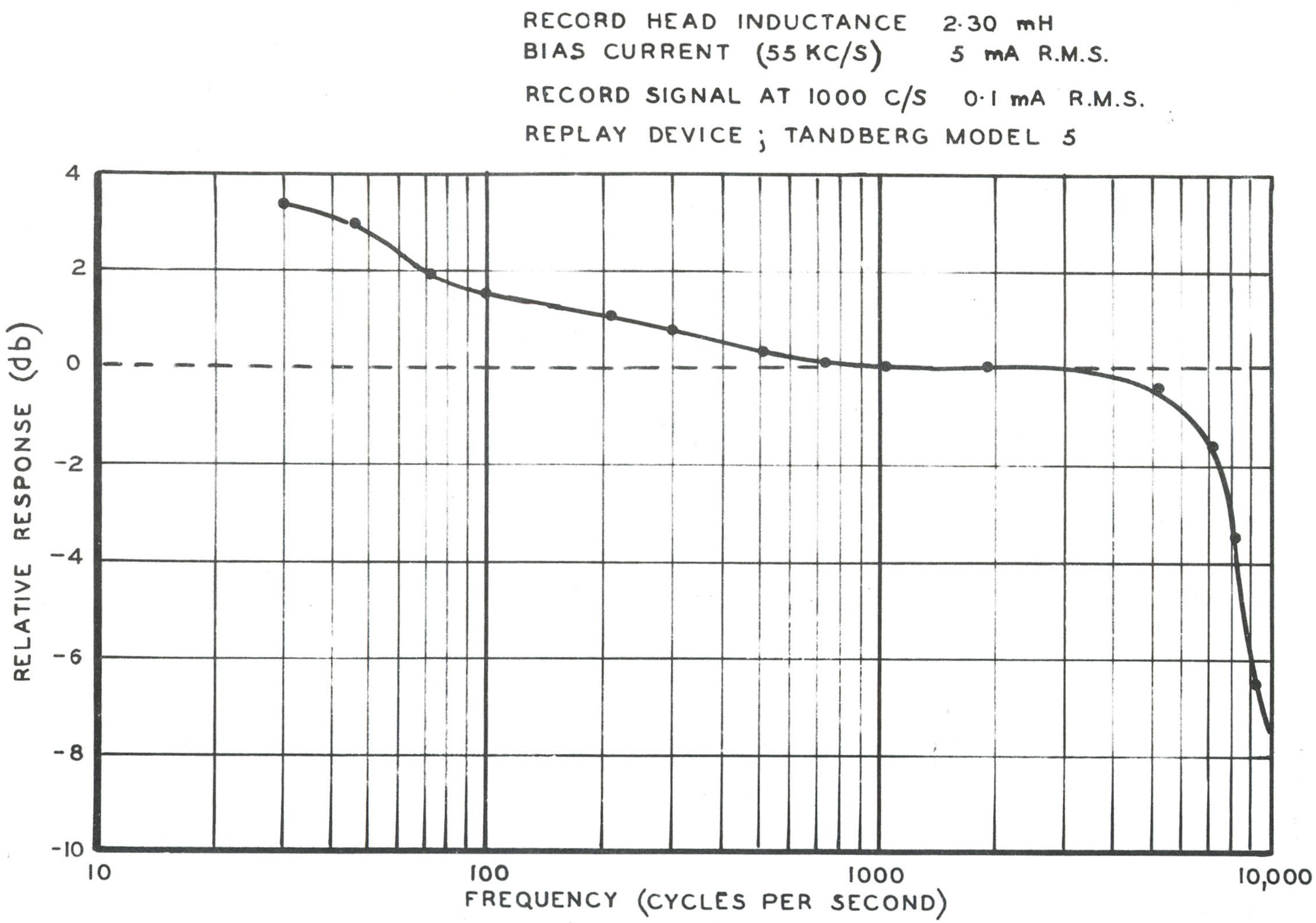
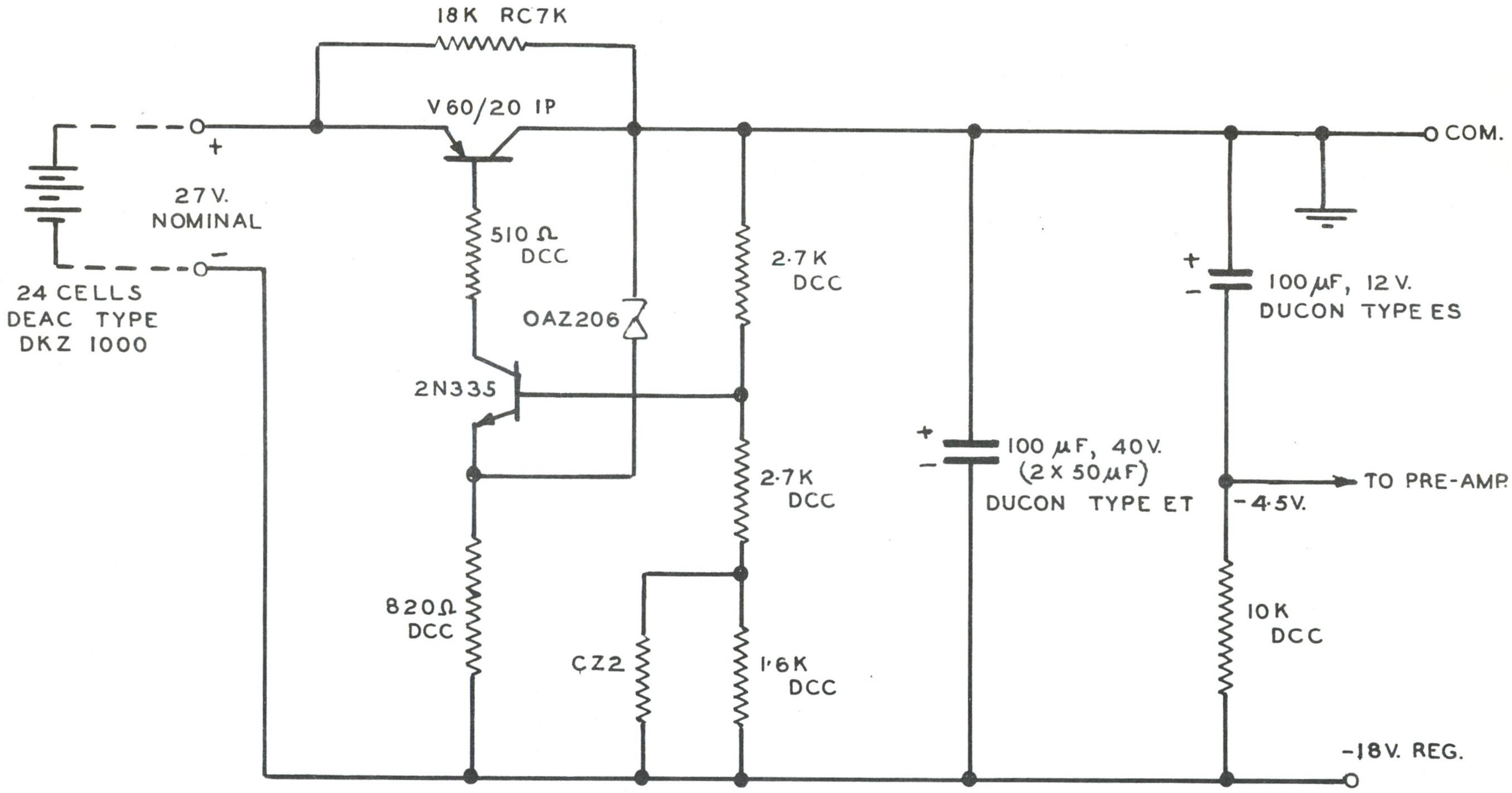


FIG. 14b SIMPLIFIED EQUIVALENT CIRCUIT OF OSCILLATOR



TYPICAL COMPLETE RECORD-REPLAY SYSTEM FREQUENCY RESPONSE

INST. NOTE 64
FIG.15



VOLTAGE REGULATOR

INST. NOTE 64
FIG. 16

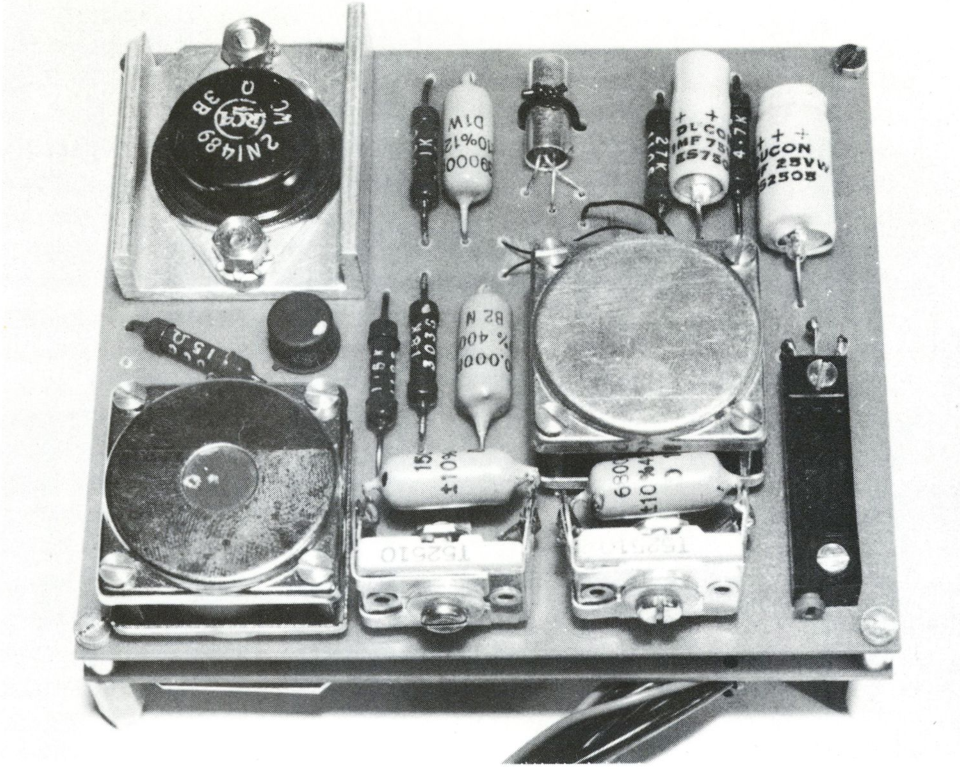


FIG. 17a BIAS OSCILLATOR AND REGULATOR -TOP VIEW

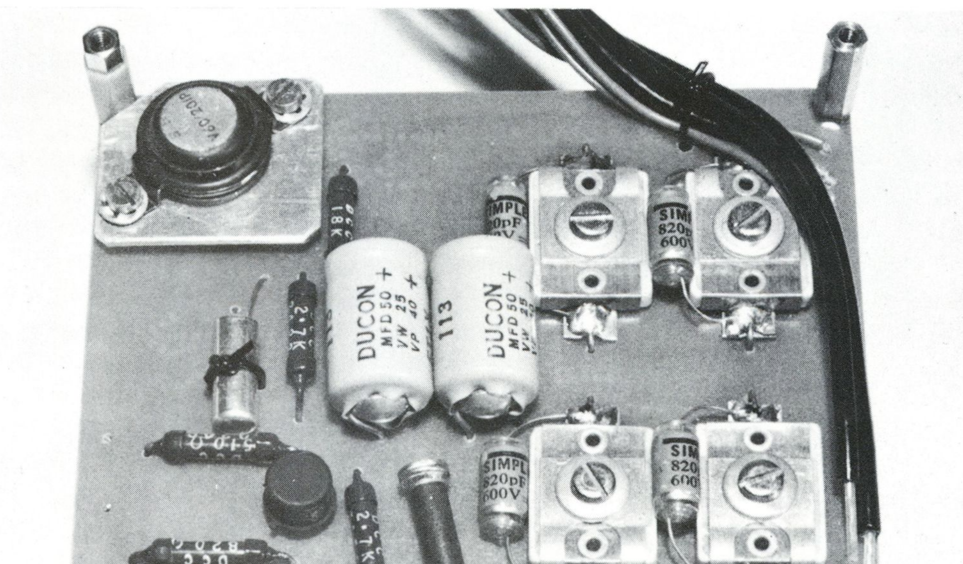
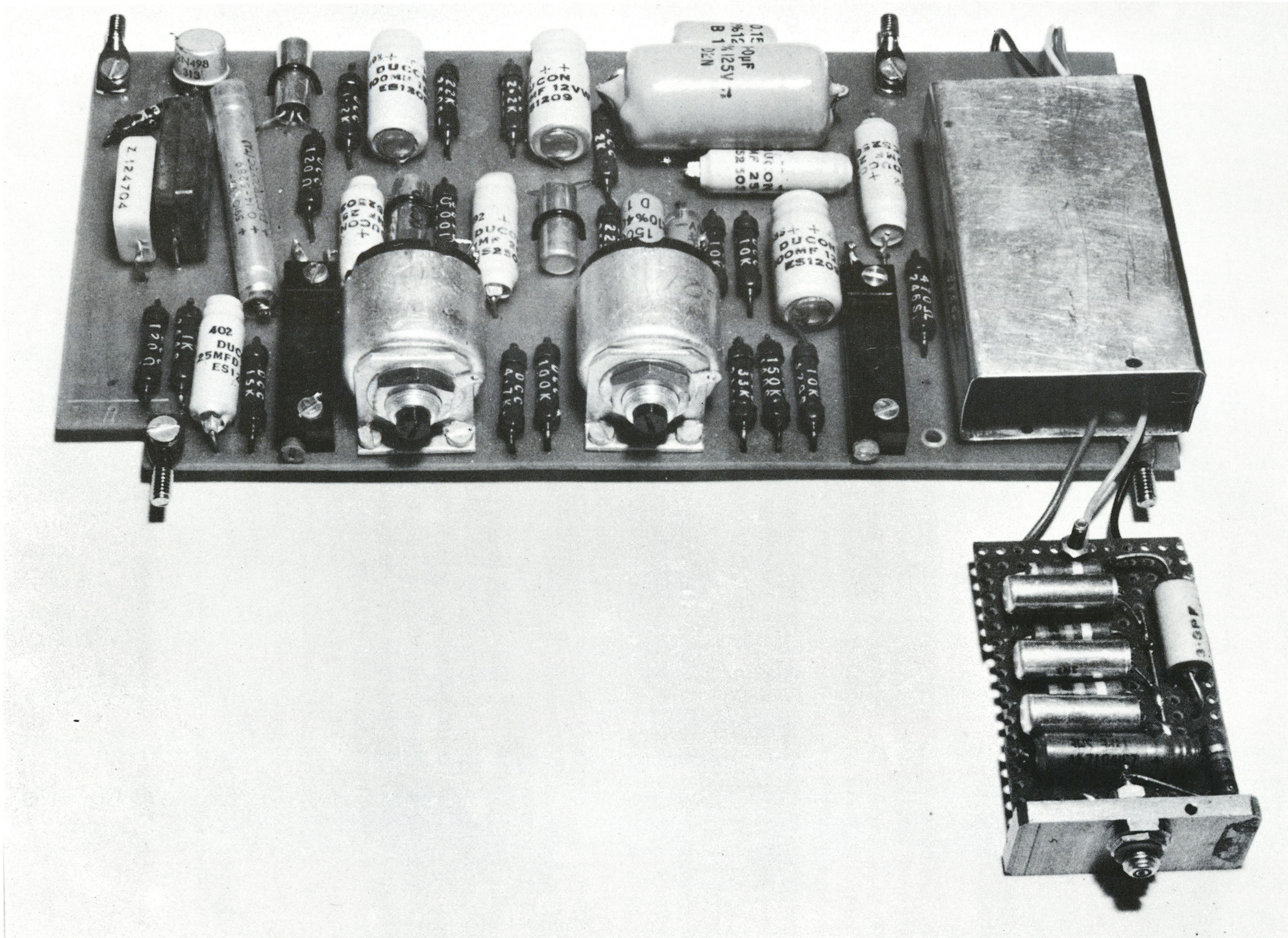
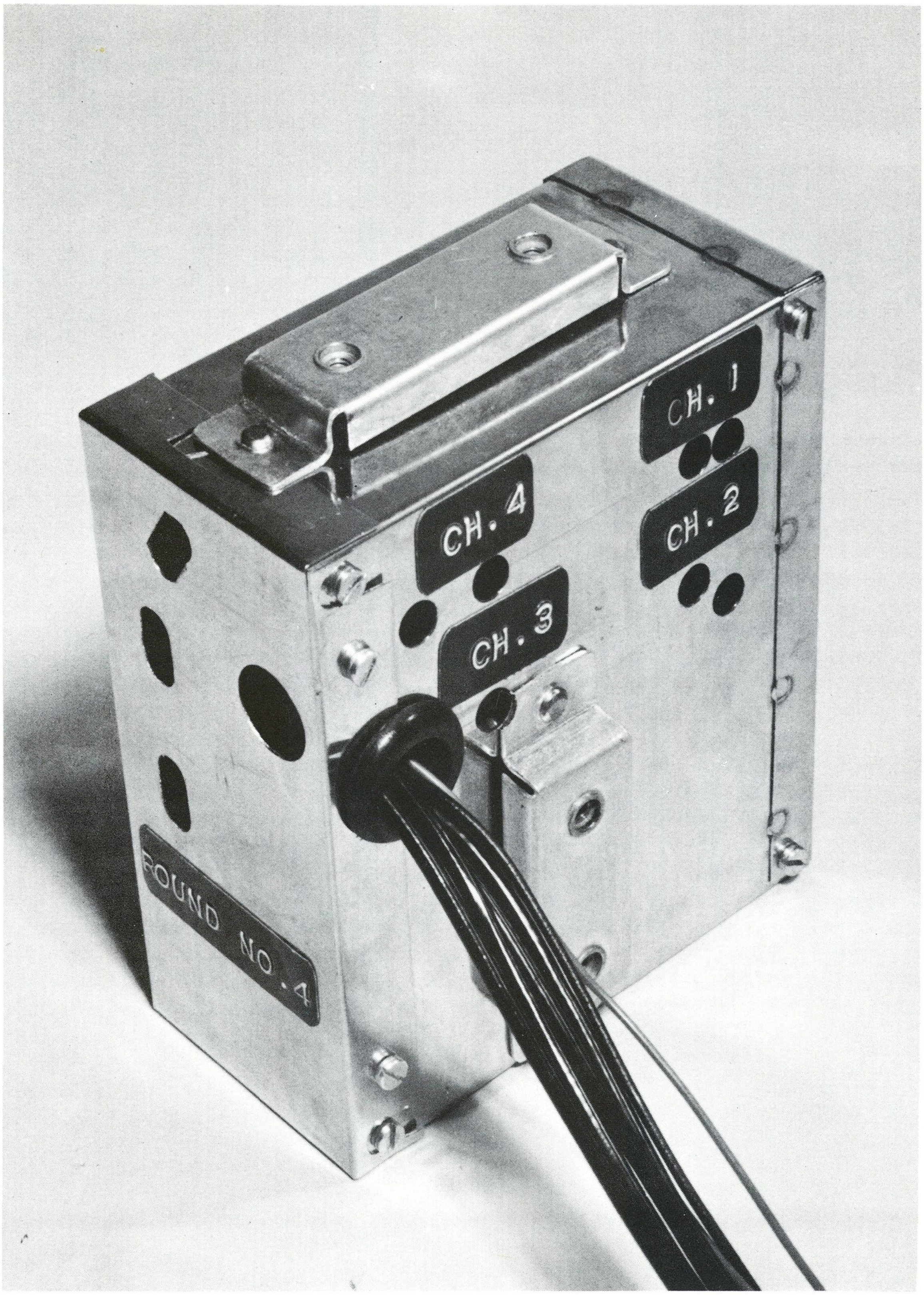


FIG. 17b BIAS OSCILLATOR AND REGULATOR - BOTTOM VIEW

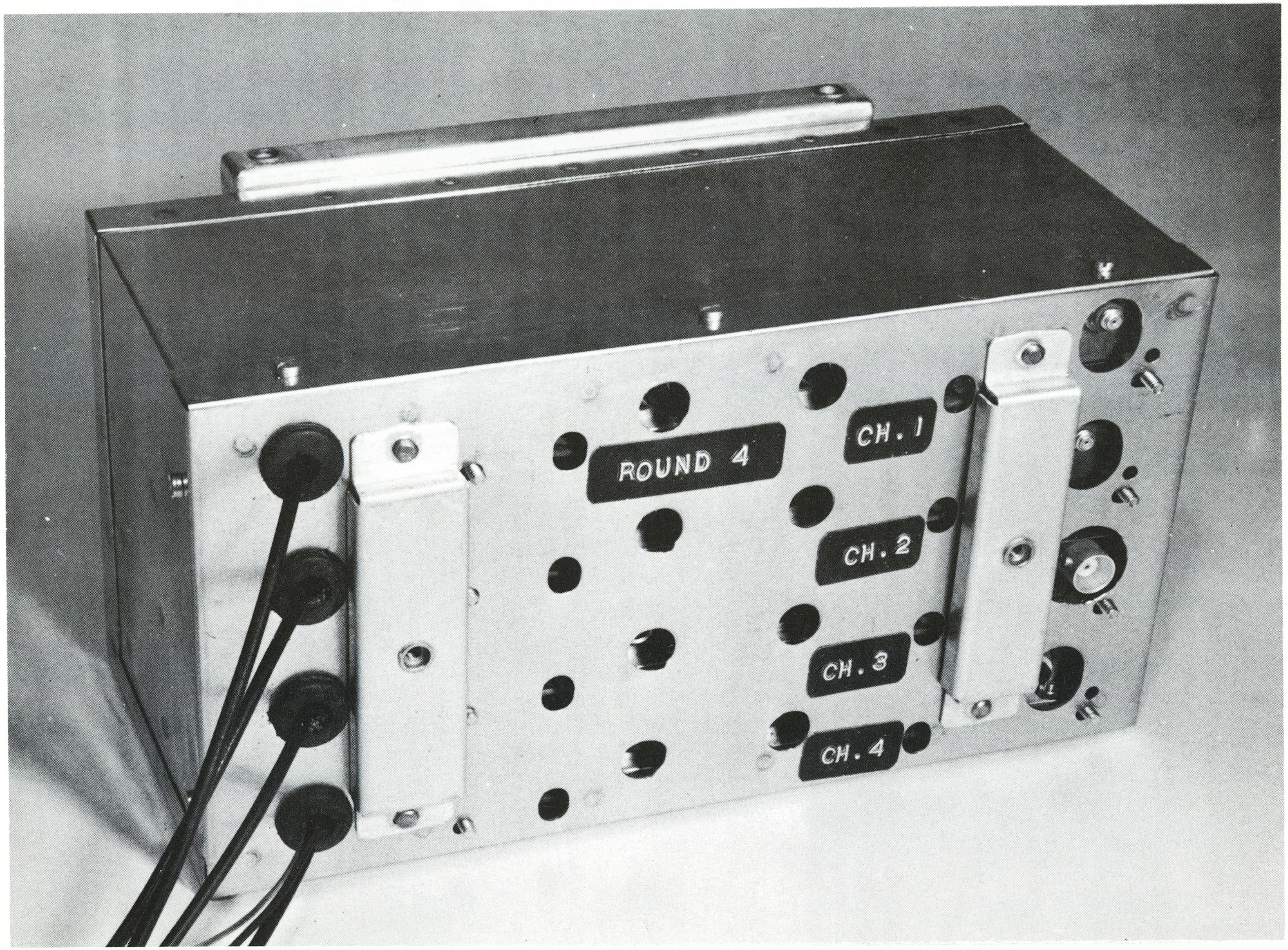


AMPLIFIER CHANNEL SHOWING PRE-AMPLIFIER IN FOREGROUND

INST. NOTE 64
FIG. 18



BIAS OSCILLATOR MOUNTING BOX



AMPLIFIER MOUNTING BOX

INST. NOTE 64
FIG. 20

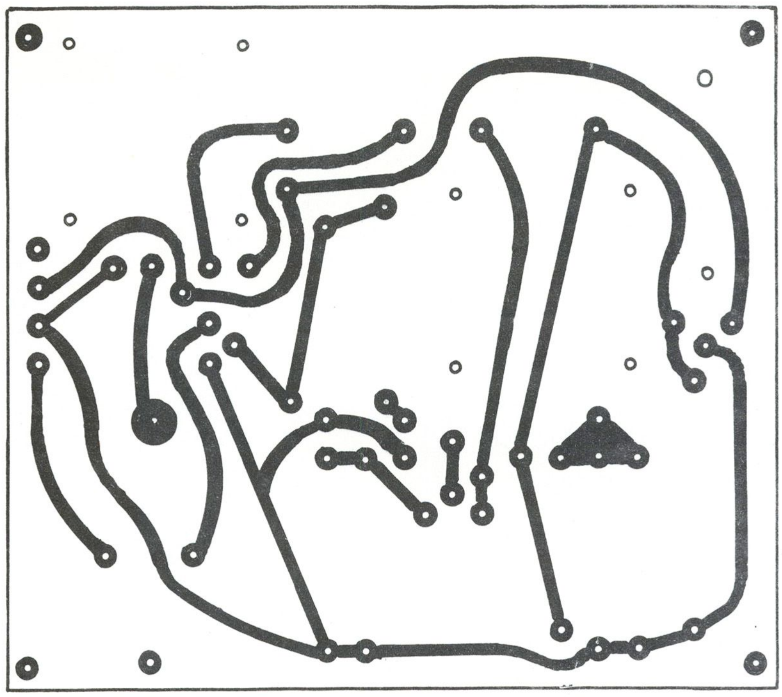


FIG. 21a PRINTED CIRCUIT FOR BIAS OSCILLATOR AND REGULATOR (TOP SECTION)

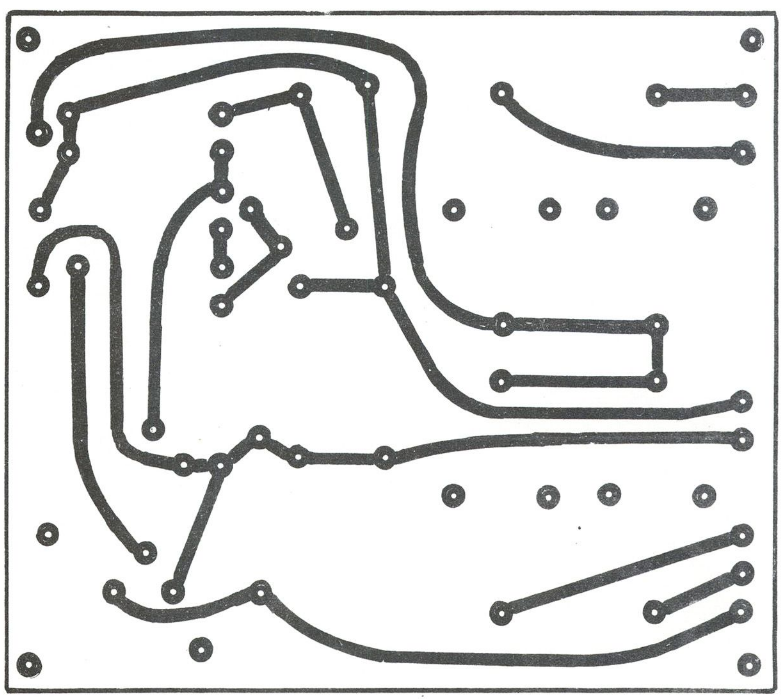
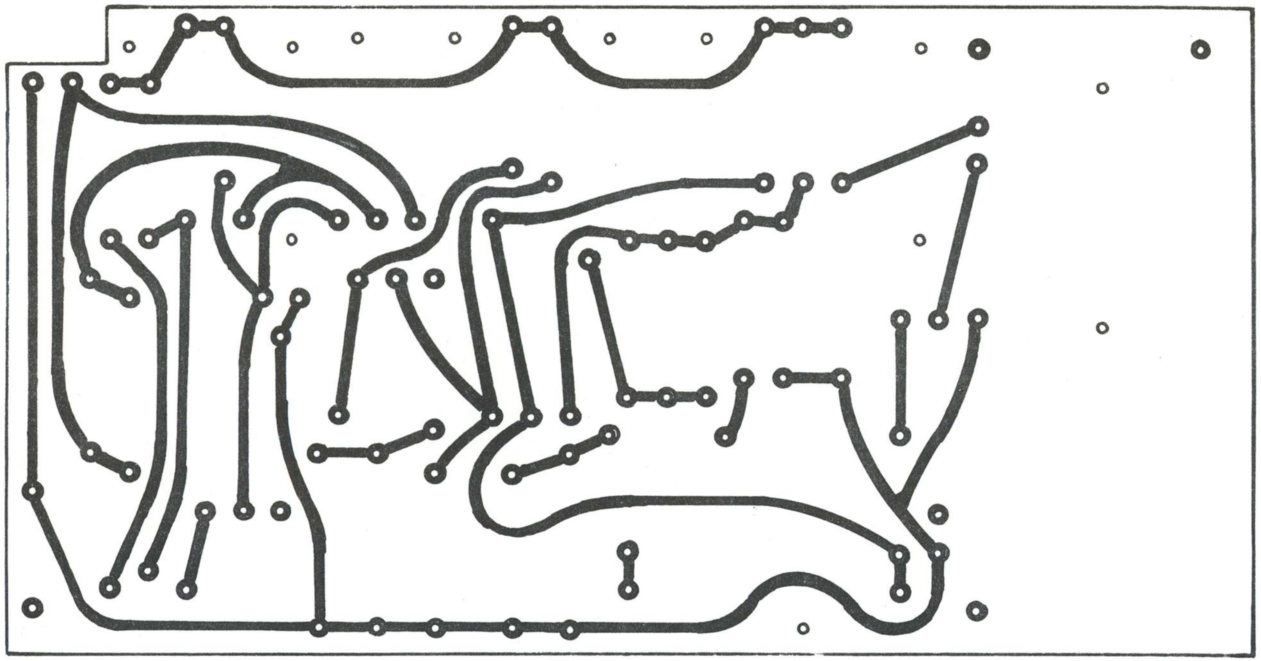


FIG. 21b PRINTED CIRCUIT FOR BIAS OSCILLATOR AND REGULATOR (BOTTOM SECTION)



PRINTED CIRCUIT FOR AMPLIFIER CHANNEL

INST. NOTE 64
FIG. 22