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POWER SUPPLY FOR THE A.R.L. FLIGHT MEMORY AIRBORNE EQUIPMENT

by

K.F. Fraser

SUMMARY

The Flight Memory Airborne Equipment, developed at the Aeronautical Research Laboratories (A.R.L.) was required to be powered normally from the aircraft 28 V.D.C. supply, or, in the case of aircraft power supply failure, from a self-contained battery pack for a short period. The synchronous motor for driving the wire recording deck required an inverter to supply 400 cycle per second power of stable frequency. D.C. regulated voltages of +5V, +24V, +36V and -24V, and an unregulated voltage of 250V, were required for the electronics of the recording system. Details of the complete power supply design are given in this paper.

CONTENTS

INTRODUCTION

The power supply detailed in this paper has been successfully used in conjunction with the A.R.L. Flight Memory Airborne Equipment. This equipment was developed for the purpose of recording cockpit speech and various flight parameters such as airspeed, altitude, cabin pressure, pitch and roll. During flight, continuous recording and erasure of this data was performed on a miniature wire recording deck. In the event of an air disaster it was envisaged that the wire recording could be salvaged and played back to provide valuable information as to the cause of the disaster. To enable recording to be continued for, say, a few minutes following a power failure in flight, it was decided that the equipment incorporate its own emergency battery pack. Therefore it was deemed convenient to operate normally from the standard 28 volt aircraft D.C. supply, and not to use the aircraft 115V 400 c/s supply. The latter supply would have been suitable for powering the recording deck synchronous motor, but an inverter would still have been needed to drive the motor from the battery pack following a supply failure.

It is convenient to sub-divide the power supply design into three distinct aspects namely:

(1) Power supply Switching

(2) Inverter for Recording Deck Motor

(3) Supplies for Electronics Unit

The design considerations for each of these aspects will be examined in turn.

PART 1. POWER SUPPLY SWITCHING

1. SYSTEM OF SWITCHING FROM AIRCRAFT TO EMERGENCY SUPPLY

The system of switching from aircraft to emergency supply is depicted in general form in Fig.1. The comparator (which will be discussed later) has low hysteresis and will energise relay P_1 provided the aircraft supply voltage is above 21V. Below 21V output the subsequent voltage regulators will cease to regulate. If P_1 is energised the aircraft supply is connected to the output via the contacts of P1.

If switch S is open, relay P2 cannot be energised and the emergency battery remains open circuit. As soon as switch S is closed the emergency battery voltage energises relay P2. The emergency battery is a secondary type which can be kept fully charged by floating it across the main aircraft supply. The diode enables charging current to flow into the emergency battery (assuming that switch S is closed) but prevents any discharge back into the aircraft supply, should this drop for any reason. If the voltage from the aircraft supply drops below 21V, for which condition it is assumed that the aircraft supply has failed, relay P1 will drop out, hence connecting the emergency supply and isolating the aircraft supply from the output terminals.

Obviously if switch S is closed and the aircraft supply is switched off, the emergency battery will provide power for the recorder and hence discharge. It is therefore essential that switch S be opened after a normal landing. The general requirement that the equipment be installed on a "fit and forget" basis rules out the possible solution that the pilot manually operates switch S. In that instance the system would be subject to human error, for if the pilot forgot to operate the switch after landing, the emergency battery would discharge. One possible solution for S would be to have a microswitch which operates from the undercarriage when the weight of the aircraft is down. Such a switch would necessarily switch off before the aircraft came to rest and it is possible that in rare cases some valuable information may be lost. In the case of a power failure in flight the requisite information should be on the magnetic wire before the emergency landing was attempted. The case of an unexpected crash would be accommodated as the aircraft supply would provide power at least until the moment of impact, after which damage to the recording equipment may render further recording impossible.

If the emergency battery has low capacity, the recorder will stall after a certain period of time determined by the emergency battery capacity and load current. This period can be chosen within wide limits by suitable choice of the battery. A disadvantage of this system is that some recording may be made after the regulators have ceased to regulate (output below 21V) and hence any information so obtained would be unreliable. A more sophisticated system could incorporate another comparator across the output which would only enable the relay P2 to pull in if the output voltage is above 21V. Adjustment of the switching level of this comparator to some value above 21V could also be considered as an adjustment on the running time after power failure. Such a system is depicted in Fig.2, but the simple system of Fig.1 is the one that has been employed to date. Note that, in both systems, discharge of the emergency battery cannot occur if switch S is open.

2. CHOICE OF EMERGENCY BATTERY

To be sure that the emergency battery is fully charged in the event of a power supply failure it was considered necessary to use a chargeable battery pack which could be floated across the aircraft D.C. supply under normal conditions. It was decided to use a Nickel-Cadmium sintered plate type of secondary cell. Some of the advantages of this type of cell are:

(1) Charge retention is very good (about 6 times as good as for a Lead-acid cell).

(2) Difference between charge and discharge voltages is not very great. (For the Silver-zinc cell the difference is quite high).

(3) Performs well over a wide range of temperatures. It has, in particular, a good low temperature charge and discharge performance.

(4) It is very rugged. Cycle life is better than Lead-acid, Nickel-Iron, Silver-Zinc and Cadmium-Zinc type cells.

The Nickel-Cadmium cell is however a relatively expensive cell, and the energy per unit mass of this cell is inferior to the Silver-Zinc cell by about a factor of 3.

A hermetically sealed Nickel-Cadmium cell type 1000 DKZ of Deac manufacture was chosen. The capacity of this cell is 1 ampere hour at the 10 hour rate. In physical format the cell is a cylinder

5 cm. in diameter and 1 cm. in depth. For floating operation the manufacturer recommended a charge voltage of 1.35V to 1.45V per cell. Hence the batch of 20 cells chosen would conveniently float on a supply voltage of 27V to 29V. Although the normal discharge of these cells is about 100mA, much heavier currents may be drawn for short periods, although the cell capacity is somewhat reduced under these conditions.

The total current demand of the A.R.L. Flight Memory equipment was 1.4 amp. At this rate of discharge the fall in voltage was rapid, 21 volt being reached in about 10 minutes from the fully charged state.

To equalize the emergency battery discharge voltage and the aircraft D.C. supply one possibility would be to charge a higher voltage battery pack (more cells) from a higher voltage source derived from a D.C. to D.C. converter. Such a system was not used in the power supply for the airborne equipment as the additional complexity was not considered to be warranted.

3. COMPARATOR SWITCH

The comparator switch represented in block form in Fig.1 is an active type circuit using a Schmitt trigger (Ref.29 pp.164,475) as the basic comparator. The relays used are of Leach manufacture (type M200) having two sets of changeover contacts rated at 2 amps resistive. To energise the 600 ohm coils in these relays about 18 volt maximum is required. Drop out voltage for this relay is in the vicinity of 5 volt.

The basic form of the comparator circuit is shown in Fig.3. At low levels of input voltage V (zener diode not conducting) Q_1 will be "on" and Q_2 will be "off", provided that $\frac{R_2}{R_1} < \frac{R_4}{R_3}$. When $V = (1 + \frac{R_3}{R_4})\, V_Z$ the zener diode will conduct hence holding the common emitter point fixed relative to the negative supply line. Q_1 will continue to conduct until reverse bias appears on its emitter base junction. This will occur when $V = (1 + \frac{R_1}{R_2})\, V_Z$. At this level of input voltage the Schmitt circuit will change state, Q_2 switching "on" and Q_1 "off". A positive going step of voltage approximately V_Z volts in amplitude will appear on the collector of Q_2. This pulse is fed via emitter follower Q_3 to transistor Q_4 which acts as a switch. The low voltage zener diode in the emitter circuit of Q_4 ensures that Q_4 is "off" until the "on" pulse appears. If the supply voltage drops below $(1 + \frac{R_1}{R_2})\, V_Z$ the circuit switches back abruptly and the relay drops out.

The complete circuit with component values is drawn in Fig.4. Adjustment of the level at which switching occurs is made possible using a potentiometer as part of the input divider. For the input divider values chosen (as in Fig.4), $0.81 < \frac{R_1}{R_2} < 1.64$. Using the nominal 10 volt zener diode shown, the voltage level at switching could therefore be adjusted between 18.1V and 26.4V. It was found necessary to decouple the comparator from the supply input as shown in Fig.4 in order to prevent relay chattering at full load and for V set at the level for switching to occur. Chattering of the relay was caused by virtue of the finite source impedance of the supply. When the load was switched on a small drop of input voltage occurred which caused the comparators to switch back again. The whole process was regenerative. Chattering of the relay stopped when the input voltage was raised above the switching level; and stopped under any condition when the decoupling network was added. Transistor Q_4 was protected from large voltage transients at switch-off by connecting a diode across the relay coil.

PART 2. INVERTER FOR RECORDING DECK MOTOR

1. GENERAL FORM OF INVERTER CIRCUIT EMPLOYED

An inverter was required to drive the 400 cycle per second 115 volt hysteresis type synchronous motor (Smiths HMIZ/2) used in the recording deck from either the aircraft D.C. supply or the emergency supply. Although this motor was a three phase type it was possible to run from a single phase supply by using a capacitor to produce some phase shift. Some experiments using square wave drive indicated that a synchronous motor could be driven from a square wave source. Moreover operation of the three phase motor from a single phase supply was possible using a capacitor to provide phase shift in a similar manner to that for sinusoidal supplies. In general, square wave inverters have a higher efficiency than sinusoidal ones, and for this reason a square wave inverter was used. Now it is to be appreciated that a 115V R.M.S. sine wave has a peak value of 162.5V whereas the R.M.S. and peak values of a square wave are the same. A square wave output of about 140V at full load was chosen, this figure being in the vicinity of midway between the R.M.S. and peak value of the appropriate sine wave for driving this motor. It was shown experimentally that the motor performed satisfactorily over a fairly wide variation in voltage, and hence voltage stability was not a prime consideration in the inverter design.

The two transformer inverter described in A.R.L. Inst. Tech. Memo. No. 60 ("Design Considerations for a Two Transformer Square Wave Inverter") was found to be unsuitable for the present application mainly because the frequency stability was inadequate when this inverter was used in conjunction with an inductive type load such as an hysteresis motor. Frequency stability was the most important consideration in the inverter design as the speed of the spools carrying the magnetic recording wire was proportional to the speed of the motor. To obtain the requisite frequency stability it was found necessary to use a separate oscillator to drive the inverter output stage. Because the inverter had to function down to 21V when drawing power from the emergency supply it was deemed necessary to interpose a 20V regulator between the inverter and the supply.

2. OUTPUT STAGE AND SERIES REGULATOR

The full load requirement was approximately 20 volt-amp at 0.8 power factor (inductive) for a 400 c.p.s. sinusoidal drive. For the square wave output the requisite output was somewhat higher, of the order of 25 volt-amp. A typical class B push pull output stage employing a phase splitting drive transformer is depicted in Fig.5. The output transformer used in this circuit was designed for linear operation and is identical to the output transformer designed for the two transformer inverter referred to in Inst. Tech. Memo. No. 60 As a separate oscillator was to be used in this inverter, the drive transformer was also designed for linear operation. A 0.51 μ F capacitor was found experimentally to be the optimum value of phase shift capacitor for driving the three phase motor.

A conventional series regulator incorporating three amplifying stages (Fig.6) was set to provide 20V input to the inverter for supply variations of 21V to 30V. In order to stabilize the collector current in the transistor (OC77) of the first amplifying stage a full wave rectifier was placed across the primary of the output transformer so that a negative supply of approximately -20V was available for the collector load resistor of this transistor. In order to prevent the regulator from being adversely affected by reverse currents during switching a very large reservoir capacitor was needed on the output of the regulator. A capacitor was connected between the collector of the first amplifying transistor (OC77) and the positive supply rail to remove the tendency of the regulator to oscillate. The regulated output was maintained within 0.5% from 21V to 28V input voltage and from 10% load to full load. Circuit details of the regulator and the inverter output stage are shown in Fig.6.

3. OSCILLATOR

The basic oscillator circuit employed was a relaxation type using a unijunction transistor. In order to provide a square wave drive for the output stage, the output pulses from the oscillator were used to trigger a binary. As the binary divides the frequency by two the basic oscillator was designed to operate at 800 cycles per second.

A typical unijunction transistor relaxation oscillator is drawn in Fig.7. The theory of operation of the unijunction transistor is given in Ref. Nos. 5, 6 & 7. Assuming the capacitor C is initially uncharged, then charging will be continued until the emitter-base junction becomes forward biased. This voltage is known as the peak point voltage V_P

$$V_P = \eta V_{BB} + V_D \quad \text{(Ref.5)}$$

where η is the intrinsic stand-off ratio and is a constant for a given unijunction transistor over a wide range of temperature and interbase voltage.

V_D is about 0.7 volt at 25°C and decreases with temperature at about 3mV/°C.

V_{BB} is the interbase voltage which is approximately V for R_2 and R_3 small compared with the interbase resistance.

When the capacitor potential reaches V_P the emitter becomes forward biased, the emitter to base one resistance drops to a very low value, and the capacitor rapidly discharges into the emitter circuit. The circuit returns to the cut-off condition when the emitter current has fallen to the valley current (see Ref. Nos. 5, 6 & 7) and the process is repeated.

Neglecting the discharge time of the capacitor, since this will be small compared with the charging time, the period of oscillation is given approximately by the time taken for the capacitor to charge to the peak point voltage. If V_D is neglected then,

$$V_P \simeq \eta V = V\left(1 - e^{-\frac{T}{R_1 C}}\right)$$

where T is the period

$$T \simeq R_1 C \ln\left(\frac{1}{1-\eta}\right)$$

$$f \simeq \frac{1}{R_1 C \ln\left(\frac{1}{1-\eta}\right)}$$

where f is the frequency of oscillation, and is essentially independent of supply voltage.

For the type 2N1671 unijunction transistor used $\eta \approx 0.5$. The value of capacitance C chosen was nominally 0.12 μF. Hence

$$R_1 = \frac{1}{f C \ln\left(\frac{1}{1-\eta}\right)}$$

$$= \frac{1}{800 \times 0.12 \times 10^{-6} \ln 2}$$

$$= 15{,}000 \text{ ohm.}$$

Actually about 18,000 ohm was needed. A 16,000 ohm resistor in series with a 5,000 ohm potentiometer for frequency adjustment was used. Resistor R_2 is used for temperature compensation and the requisite value is given by

$$R_2 \simeq \frac{0.70\, R_{BB}}{\eta V} + \frac{1-\eta}{\eta} R_3 \qquad \text{(See Ref.5 \& Fig.7)}$$

where R_{BB} is the interbase resistance which was about 7,500 ohm for the unijunction transistor used.

For $V \approx 24$ volt and $R_3 = 100$ ohm (see bwlow)

$$R_2 \approx 538 \text{ ohm}$$

Actually a 470 ohm resistor was used.

A low impedance output, in the form of short duration positive going pulses, was taken across resistor R_3. The value of R_3 was not very critical and was made 100 ohm. For improved stability the supply for the basic oscillator was taken from the +24 volt regulator (to be described in Part 3).

The oscillator output was coupled to a symmetrically triggered binary. As the binary supply was taken from the nominal 28 volt supply a square wave of about 24 volt amplitude appeared on each collector. The input for the drive transformer in the output stage was taken between the collectors of each half of the binary, thereby doubling the square wave amplitude in comparison to that appearing at each collector. The complete oscillator circuit is drawn in Fig.8.

PART 3. SUPPLIES FOR ELECTRONICS UNIT

1. DESIGN REQUIREMENTS

The following regulated supplies were required for operation of the A.R.L. Flight Memory Airborne electronics unit:

(i) +36V at 25mA

(ii) +24V at 25mA

(iii) -24V at 55mA

(iv) + 5V at 50mA (for transducers)

It was desirable that provision be made for isolation of the common terminal of these regulators from the aircraft supply common. In addition a floating unregulated supply of 250V at 0.6mA was required.

Regulation of the order of $\pm$0.25% was required on the +24V and the +5V supplies, whereas $\pm$1% was adequate for the +36V and -24V supplies.

Power for these supplies was to be derived from either the aircraft D.C. supply or the emergency supply. To be assured of reliable operation from the emergency supply the regulators were to accommodate an input voltage variation of 21V to 28V.

2. D.C. TO D.C. CONVERTER

A D.C. to D.C. converter with multiple windings on the output transformer secondary for the various supplies, seemed to be the obvious device to precede the regulators. Summation of the requisite D.C. outputs referred to above indicates a power requirement of 3.2 watt. Allowing for about 30% loss in regulators at 28V input would mean a converter output requirement of about 4.2 watt at this voltage. To make provision for any increased power demand due to possible further developments in recording equipment a 50% increase in power was allowed for, bringing the requisite output power from the converter to 6.3 watt. If a converter efficiency of 75% is assumed, then an input power of 8.4 watt at 28V would be required.

A single transformer saturating core type of square wave inverter employing a common base connection for the switching transistors was chosen (see Refs.9,12,14 & 20). Square wave inverters are to be preferred to sinusoidal ones because of the higher efficiency and the lower degree of filtering required after rectification of the output. As the requisite output is D.C. it is of little consequence if the frequency of operation of the inverter varies with input voltage, load and temperature. For saturating core converters the frequency will be very nearly proportional to input supply voltage. However to obtain good frequency stability under loaded conditions a square hysteresis loop material such as HCR would be required. As these conditions did not have to be met in the present application a ferroxcube type 3B2 core was used in spite of the fact that saturation was not particularly sharp and also changed markedly with temperature.

2.1 Frequency of Operation of Inverter

The basic circuit of the inverter is drawn in Fig.9. Each transistor alternatively switches "on" and "off" producing a square voltage waveform across the transformer. As the collector and emitter windings are connected in anti-phase the effective number of primary turns per half must be taken as $N_1 - N_F$. Using Lenz's equation

$$\frac{d\phi}{dt} = \frac{V}{N_1 - N_F}$$

where V is the supply voltage and ϕ is the flux in the core. The voltage drop across the transistor during saturation is assumed to be negligible for this analysis. During a half cycle ϕ goes from $-\phi_M$ to $+\phi_M$ where ϕ_M is the peak flux in the core

$$2\,\phi_M = \int_0^{\frac{1}{2f}} \frac{V\,dt}{N_1 - N_F}$$

$$f = \frac{V}{4\,(N_1 - N_F)\,\phi_M}$$

where f is the frequency of oscillation

$$\phi_M = B_m\,A$$

where B_M is the peak flux density in the core and A is the minimum core area

$$f = \frac{V}{4\,(N_1 - N_F)\,B_M A} \quad \text{m k s units}$$

$$f = \frac{V\;10^8}{4\,(N_1 - N_F)\,B_M\,A} \quad \text{c g s units}$$

A type 3B2 ferroxcube core of Philips manufacture using size D36/22 with two rings was chosen. According to the manufacturer's data the saturation flux density of this type of ferroxcube is 4.5 kilogauss at H = 2000 oersted and T = 20°C. It would be impracticable to obtain such a high value of magnetising force (H). From the B-H curve reproduced in Fig.10 it is clear that the slope of the curve drops quite markedly above 5 oersted. Switching of the transistors will occur when the collector current rises sufficiently to bring the "on" transistor out of saturation. The value of flux density at this instant will be somewhat below the value quoted above.

Consider the simplified circuit for the conducting transistor drawn in Fig.11. The value of the turns ratio $\frac{N_1}{N_F}$ was made equal to 6 where $N_1 = 120$ and $N_F = 20$. Referring to this circuit we may write

$$V_F = I_b R_b + V_{be}$$

where V_{be} is the base to emitter voltage during saturation. V_{be} at saturation depends on both collector and base current. For $R_b = 220$ ohm and taking a figure of $V_{be} = 0.5$ volt gives

$$I_{b_{max}} = \frac{V_F - 0.5}{R_b} = \frac{5.6 - 0.5}{220} 10^3 \text{ mA} = 23.2\text{mA}$$

(Refer to Fig.10). At this value of base current the transistors used will come out of saturation at approximately 1.2 amp collector current.

If it is assumed that 200 mA is the load component of the primary current (5.6 watt input at 28V) then it may be assumed that the transistors will switch when the magnetising component of the primary current reaches 1A. The magnetising force at this value of magnetising current may be calculated from the well known expression

$$H = \frac{4\pi (N_1 - N_F) I_m}{10L}$$

where $N_1 - N_F$ = nett primary turns = 100

I_m = magnetising current = 1A

L = magnetic path length = 8.2 cm

H = magnetising force in oersted

$$H_{max} = 15.3 \text{ oersted}$$

From the B-H curve of Fig.10, assuming zero air gap in the core, the corresponding maximum flux density will be 3.5 kilogauss at 20°C.

Reverting to the expression for frequency

$$f = \frac{V.\ 10^8}{4\ (N_1 - N_F)\ B_M\ A}$$

and substituting the following values

$$V = 28V$$

$$N_1 - N_F = 100$$

$$B_M = 3{,}500 \text{ gauss}$$

$$A = 1.29 \text{ cm}^2$$

it follows that the frequency of operation should be about 1550 cycles per second at 20°C. At this frequency filtering of the rectified output should present no problems.

2.2 Complete Inverter Circuit

The complete circuit of the inverter is drawn in Fig.12. Type 2N1160 transistors used had ratings which amply covered the requirements. The 18K resistor between the common base connection and ground was necessary to provide some positive bias for the transistors to ensure that oscillations would always commence. If this resistor is omitted it is possible for both transistors to remain cut-off. The 0.1 μF capacitor was added to reduce dangerously large spikes on transistor collector to base voltage during switching. When one transistor switches off the collector of the other transistor tends to go positive beyond the supply voltage. The capacitor absorbs this sudden voltage surge and discharges through the 18K biasing resistor in readiness for the next voltage spike.

2.3 Rectifier-Filter Circuits

For each of the regulated supplies full wave rectification using a bridge rectifier was employed on the secondary of the inverter transformer. Type OA5 diodes were used in each of these bridges. A single electrolytic capacitor across the D.C. arms of the bridge provided adequate filtering of the unregulated D.C. for the regulators. For circuit details refer to the regulator circuit diagrams Figs.14, 17, 19 and 21. The rectifier filter circuit for the 250V supply will be discussed in the next section.

3. 250V UNREGULATED SUPPLY

A voltage doubler employing high voltage selenium rectifiers type 36 EHT 13 was used. The 36 EHT 13 will take 2 mA continuous current. To reduce starting currents in the rectifiers to within safe limits the capacitors must not be too large. This requirement was fulfilled using 0.22 microfarad capacitors. A 3.3 megohm bleed resistor was placed across the output terminals.

The following measurements were made on the supply:

D.C. voltage at 28V converter input	=	275V
Output resistance	=	37,000 ohm
Ripple voltage at 600 μA current	=	2.3V R.M.S.

For circuit details of this supply refer to Fig.13.

4. +24V REGULATOR

Cancellation of some of the errors produced in the recording electronics could be achieved if all the regulators drifted in unison. For this reason it was desirable to use a single zener diode as reference for all the regulators. As good regulation was required on the +24V supply it was decided to use the zener reference diode in this regulator and then use the +24V regulated output from this supply as a reference for all the other regulators.

A series type regulator employing a two stage amplifier to drive the main series transistor was used in this and all the other regulated supplies. A 6.8 volt zener diode type Z2A68 was chosen as voltage reference. At this voltage the Z2A series of zener diodes have approximately zero temperature coefficient. The complete circuit is drawn in Fig.14. If the collector load resistor of the BCZ10 was connected to the negative side of the unregulated input poor regulation under conditions of varying input voltage would result by virtue of the large excursions of collector current and hence of base current for this transistor. Since a negative 24V regulated output was required, it was convenient to connect the collector resistor of the BCZ10 to that supply and hence stabilize the collector current with respect to input voltage variations. The 0.22 μF capacitor added between the collector of the BCZ10 and the positive rail reduced a very definite tendency for the circuit to oscillate by reducing the high frequency loop gain.

The following measurements were made on the supply:

Output Resistance	= 0.50 ohm
Change in Output Voltage as Supply Voltage Changes from 21V to 30V at 30 mA Output Current	= 24 mV
Ripple Voltage at 30 mA Output Current	= 1.1 mV R.M.S.
Drift over 1 hour	= 5 mV

5. -24V REGULATOR

The -24V regulated supply can be conveniently referenced by the +24V supply as depicted in the block diagram of Fig.15. One of the difficulties associated with this supply is maintaining good regulation under changing input voltage conditions by virtue of current changes in the first amplifying transistor (BCZ10). For the positive supplies this difficulty could readily be overcome by employing the output of the -24V regulator, but for this supply a more negative source of stabilized voltage could be used to advantage. As the provision of a more negative source of stabilized voltage would have required a separate transformer winding, rectifiers and zener diode it was felt that this extra circuitry was not warranted in this instance. However a system of compensation was employed to improve the regulation under conditions of varying input voltage. Referring to Fig.16 it can be seen that if any change of current occurs through the adjusting arm of the output divider a change in output voltage will result. If R_2 is omitted it is obvious that any change in the unregulated input voltage will result in a change in current through the adjusting arm thus resulting in poor regulation. Now a change in input voltage of ΔV will result in a collector current change of approximately $\frac{\Delta V}{R_1}$. The requisite change in base current will be $\frac{\Delta V}{\beta R_1}$ where β is the transistor current gain. If R_2 is made equal to βR_1 then the required base current changes should flow via R_2. For the BCZ10 used the current gain was about 40. R_1 was made equal to 12K and R_2 was made equal to 470K.

The complete circuit is drawn in Fig.17. The following measurements were made on the supply:

Output Resistance	= 0.22 ohm at 28V to 0.70 ohm at 21V
Change in Output Voltage as Supply Voltage Changes from 21V to 30V at 60mA Output Current	= 54mV
Ripple Voltage at 60 mA Output Current	= 18.5mV R.M.S.
Drift over 1 hour	= 46mV

6. +36V REGULATOR

The system of referencing the +36V supply from the +24V supply is shown schematically in Fig.18. In this supply the emitter of the first amplifying transistor (OC77) is connected to the +24V line. If a circuit arrangement similar to that used for the +24V and the -24V supplies were used in this instance it can readily be seen that the feedback would be positive and hence the circuit would not regulate. To overcome this difficulty an NPN transistor having its emitter connected to the -24V supply was used in place of the emitter follower transistor used in the other supplies.

The complete circuit is drawn in Fig.19. The following measurements were made on the supply:

Output Resistance	= 2.3 ohm
Change in Output Voltage as Supply Voltage Changes from 21V to 30V at 30 mA Output Current	= 54 mV
Ripple Voltage at 30 mA Output Current	= 3.5 mV R.M.S.
Drift over 1 hour	= 75 mV

7. +5V REGULATOR

The +5V regulated supply can be referenced from the +24V supply as depicted in Fig.20. Basically the **circuit** is very similar to that used in the +24V regulator. Current for the first amplifying transistor (BCZ10) was derived from the -24V supply as in the +24V regulator.

The complete circuit is drawn in Fig.21. The following measurements were made on the supply:

Output Resistance	= 0.20 ohm
Change in Output Voltage as Supply Voltage Changes from 21V to 30V at 60 mA Output Current	= 7 mV
Ripple Voltage at 60 mA Output Current	= 0.35 mV R.M.S.
Drift over 1 hour	= 3 mV

PHOTOGRAPHS

A photograph of the complete airborne equipment with the exception of the wire recording deck is given in Fig.22. The location of the various units which constitute the power supply has been marked. In Fig.23 a different view of the airborne equipment is shown and the wire recording deck is also included.

CONCLUSION

The power supply equipment described in this paper has been successfully employed for powering the recording electronics at the wire recording deck of the A.R.L. Flight Memory Airborne Equipment. A prototype model of the Flight Memory Airborne Equipment has been made and in-flight data has been recorded. The data so recorded has been decoded in detail using the A.R.L. Flight Memory Ground Station Equipment. An overall accuracy of about 1% was achieved.

DRAWINGS

The following A.R.L. drawings give details of circuits and inter-connections relevant to the power supply for the A.R.L. Flight Memory Airborne Equipment:

Title	A.R.L. Drawing No.
Aircraft and Emergency Power Switching	SK12400
Power Supply for Recorder Motor (115V, 400 c/s)	SK12387
Wiring of A.C. Model Recording Deck	SK12401
Power Supply for Airborne Electronics Unit	SK12619
Electronics Unit Wiring	SK12380

ELECTRICAL COMPONENTS

In addition to component values, component types have been included with the circuit drawings in this text.

The fixed resistors used were of two varieties, having tolerances as specified hereunder.

D.C.C. - I.R.C. manufacture, deposited carbon, $\frac{1}{2}$ watt, 1% tolerance, high stability.

RC7K (Services classification) - composition, $\frac{1}{4}$ watt, 5% tolerance.

Two varieties of potentiometer were used:

MORGANITE - Type BJ linear, composition, 0.1 watt.

COLVERN - Wire wound, Type CLR 1106/22, 1 watt.

One variety of non-electrolytic type capacitor was used, namely:

POLYESTER - Philips manufacture, polyester dielectric. Unless otherwise specified the Type 296 AC/A with 400 VW rating was used. Type C296 AA/A refers to the 125 VW series.

The working voltage, manufacturer and type of electrolytic capacitor have been specified on the circuit drawings.

REFERENCES

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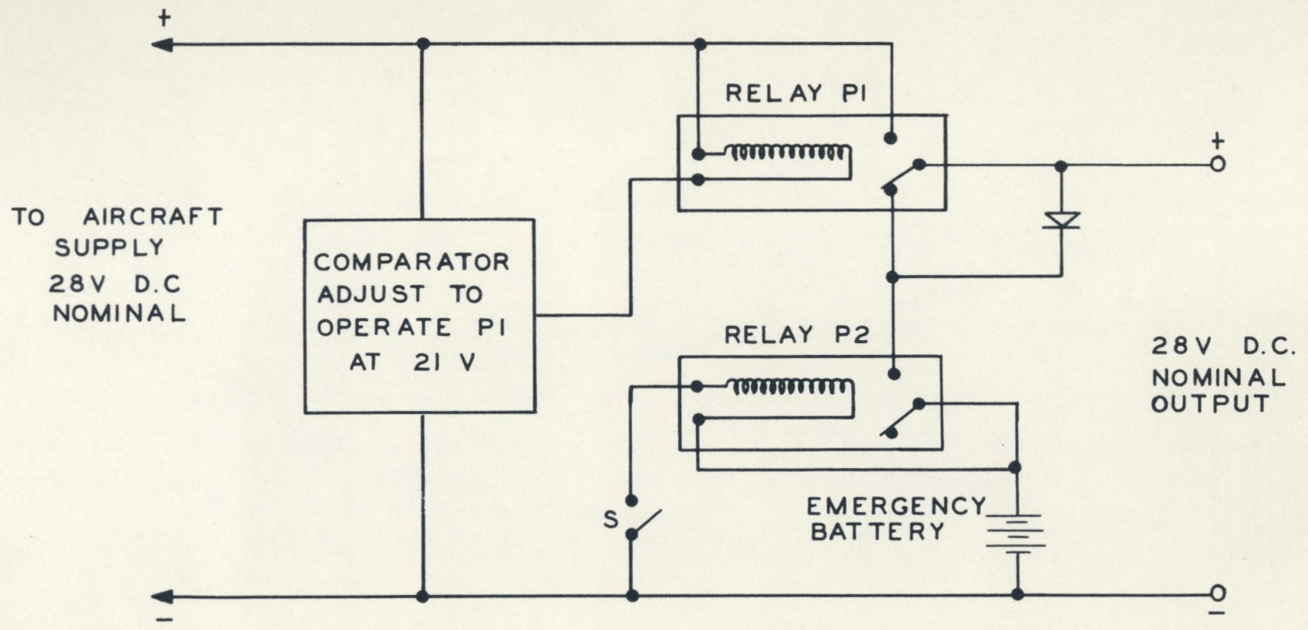
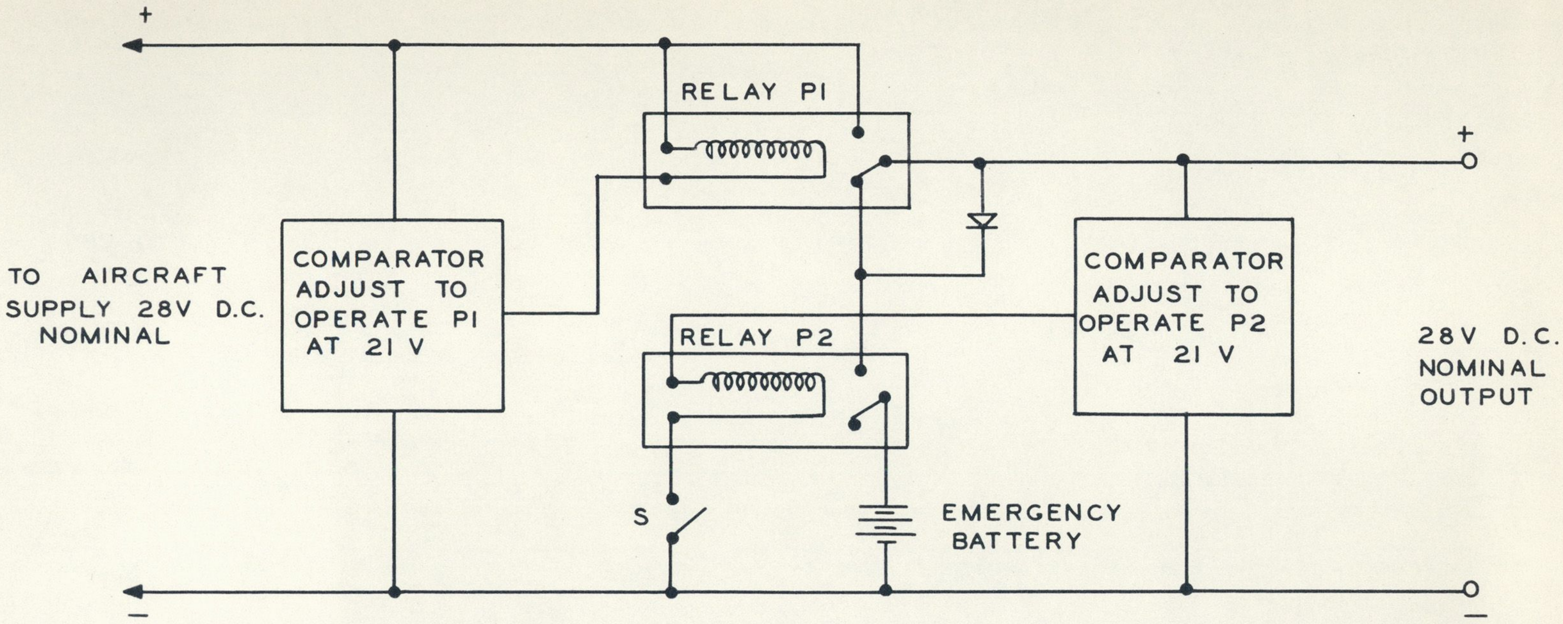
Postal Address: Chief Superintendent,
Aeronautical Research Laboratories,
P.O. Box 4331,
Melbourne, Victoria, 3001,
Australia.

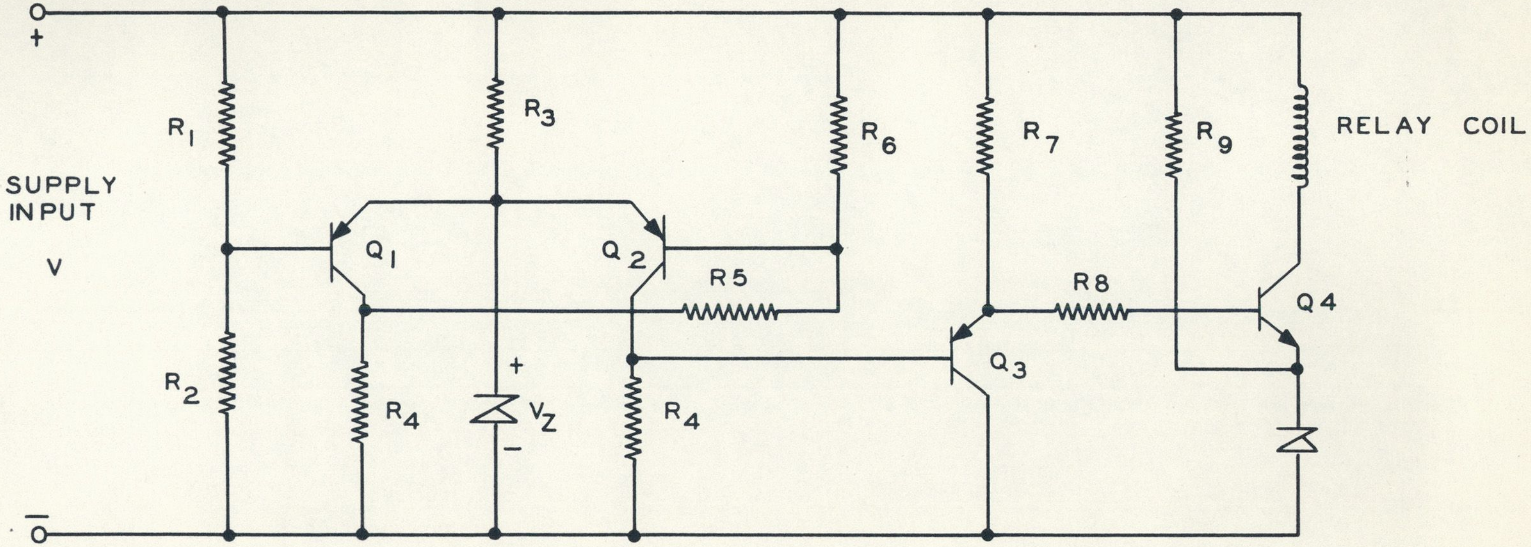


SYSTEM OF SWITCHING FROM AIRCRAFT TO EMERGENCY SUPPLY



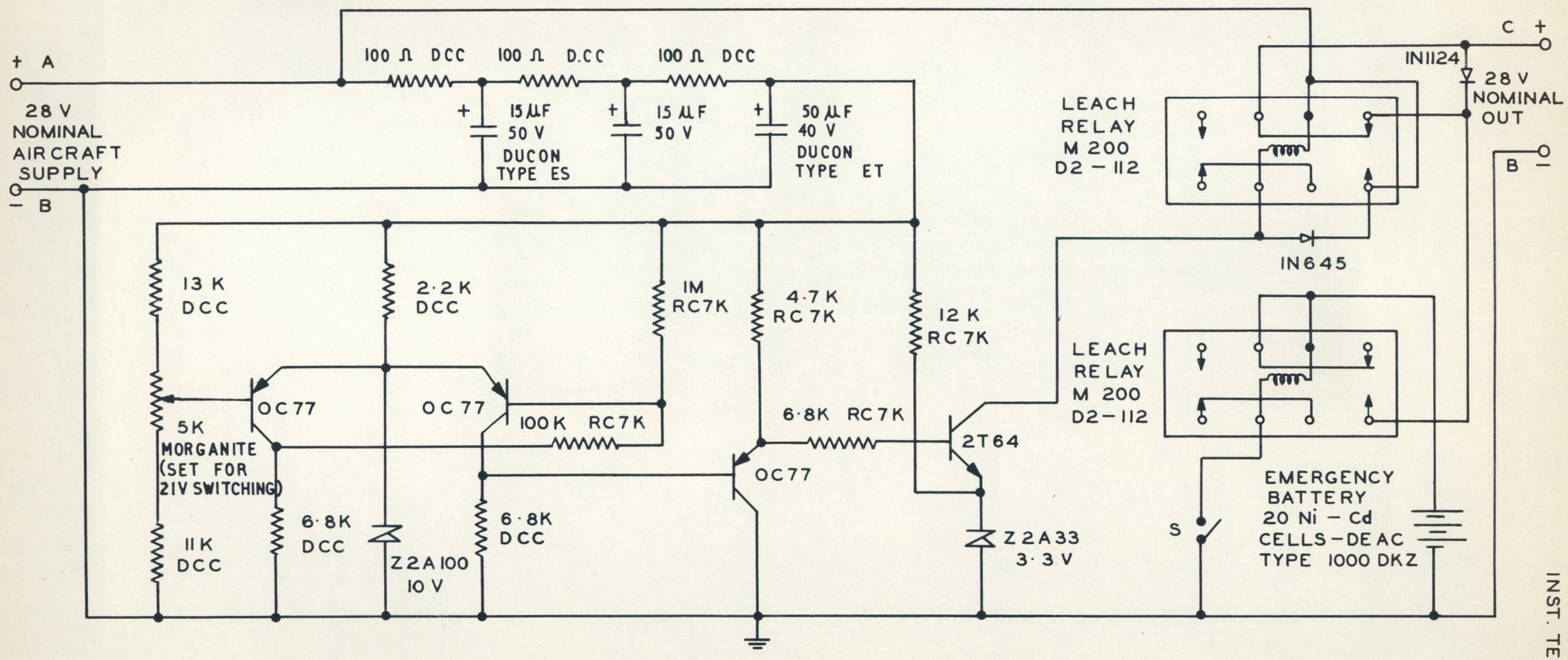
SWITCHING SYSTEM WHICH PREVENTS OUTPUT FROM DROPPING BELOW 21 V

INST. TECH. MEMO. 49
FIG. 2



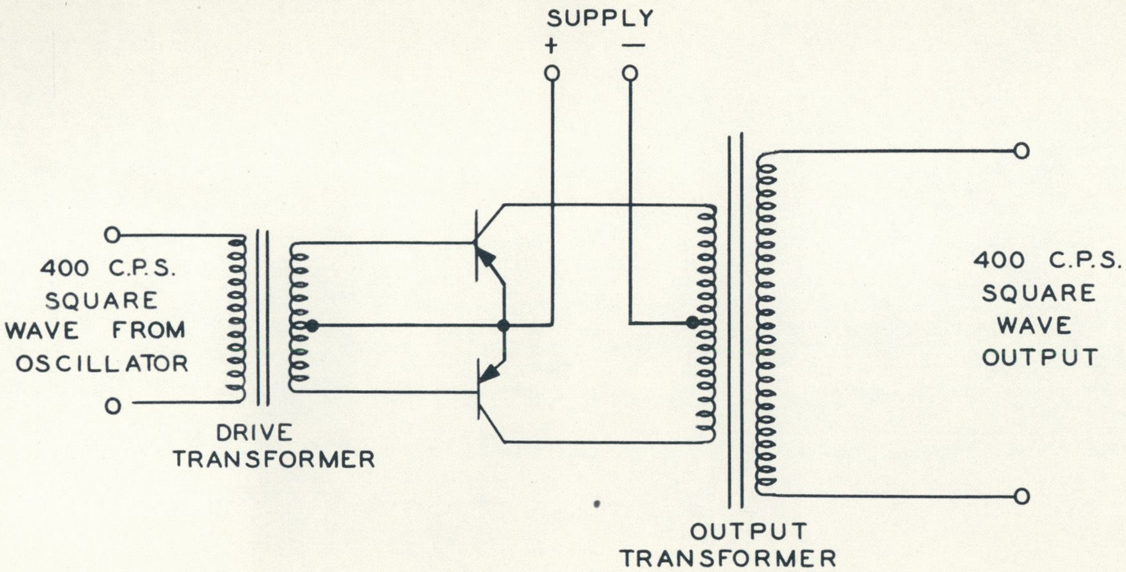
BASIC COMPARATOR CIRCUIT

INST. TECH. MEMO. 49
FIG.3



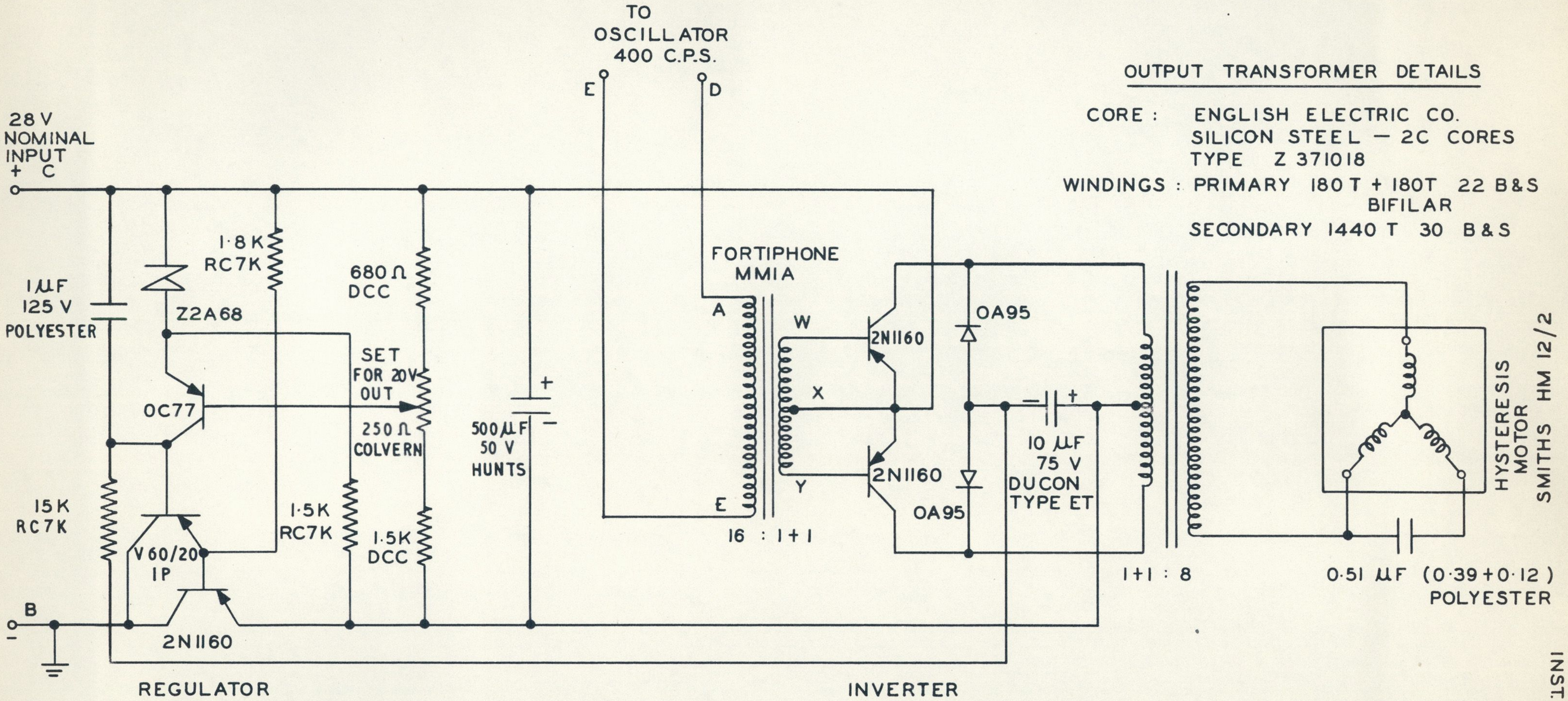
COMPLETE SUPPLY SWITCHING CIRCUIT

INST. TECH. MEMO. 49
FIG. 4



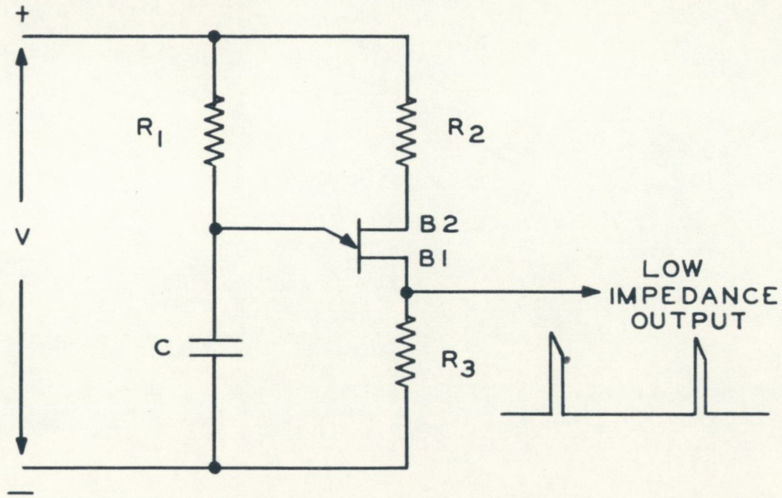
TYPICAL CLASS B PUSH PULL OUTPUT STAGE

INST. TECH. MEMO. 49
FIG. 5



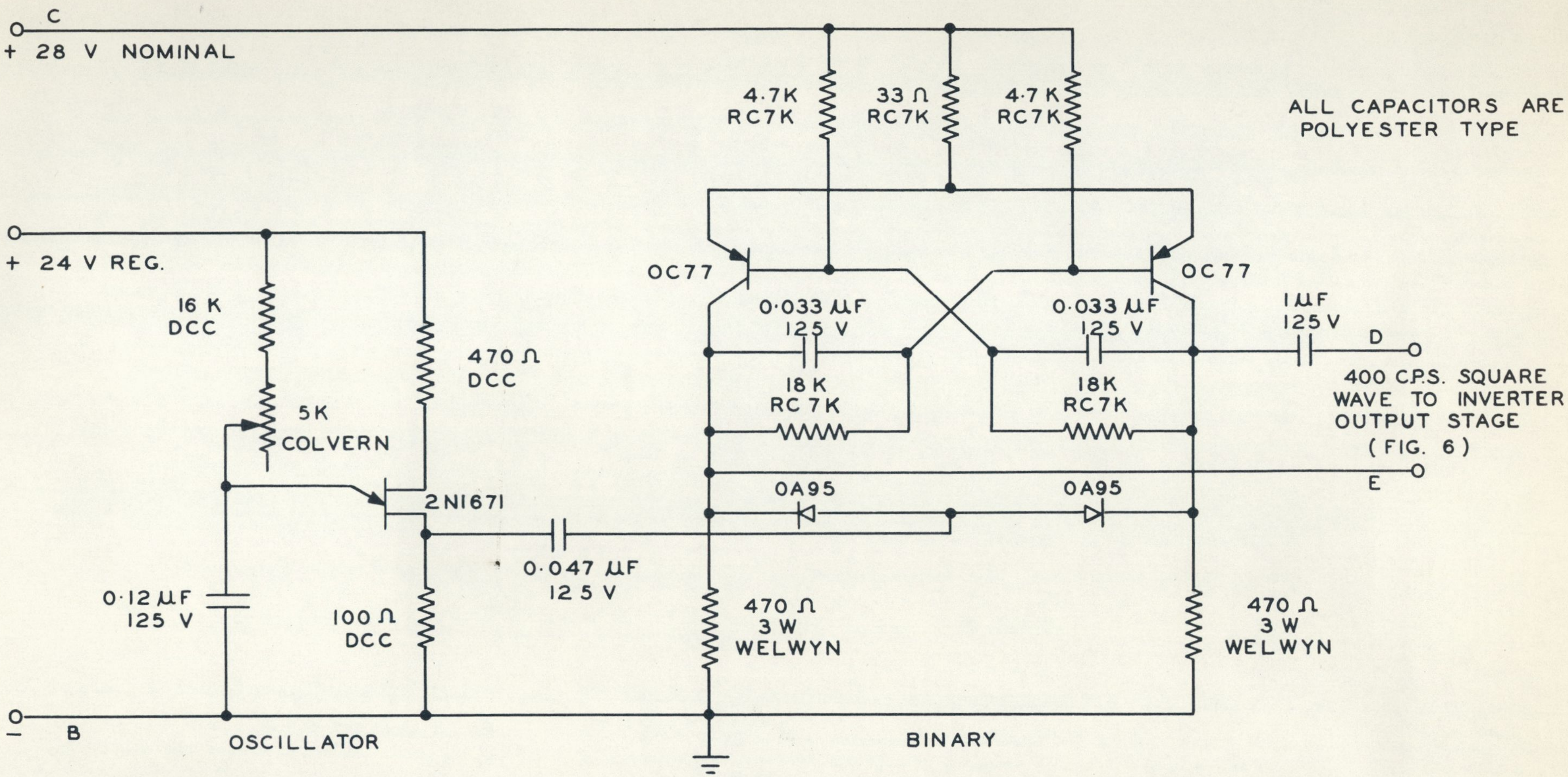
CIRCUIT DETAILS OF SERIES REGULATOR & INVERTER OUTPUT STAGE

INST. TECH. MEMO. 49
FIG. 6



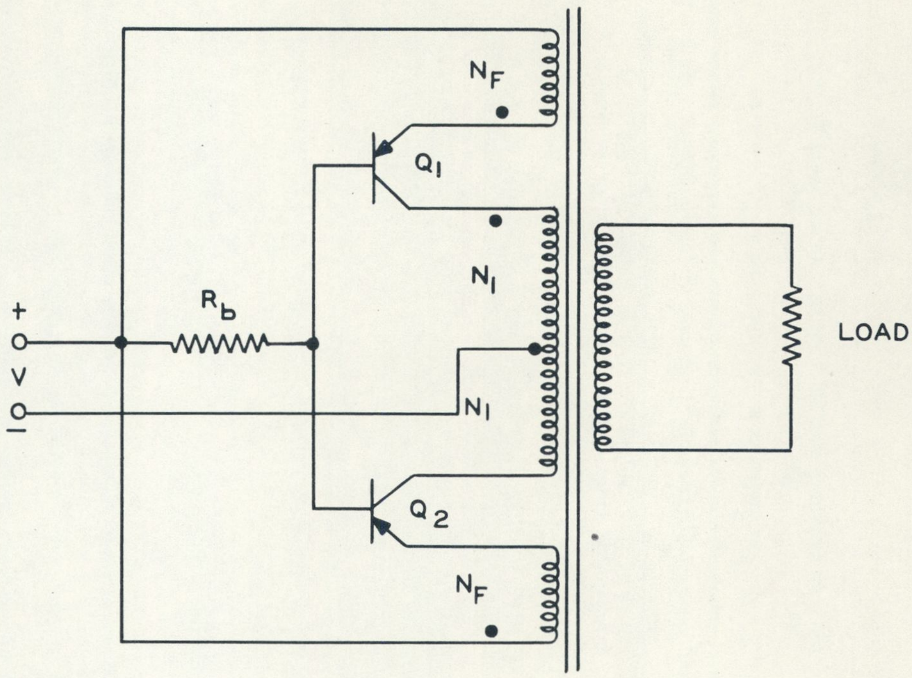
BASIC UNIJUCTION TRANSISTOR RELAXATION
OSCILLATOR CIRCUIT

INST. TECH. MEMO. 49
FIG. 7



CIRCUIT DETAILS OF RELAXATION OSCILLATOR WITH SQUARE WAVE OUTPUT

INST. TECH. MEMO. 49
FIG. 8



COMMON BASE INVERTER CIRCUIT

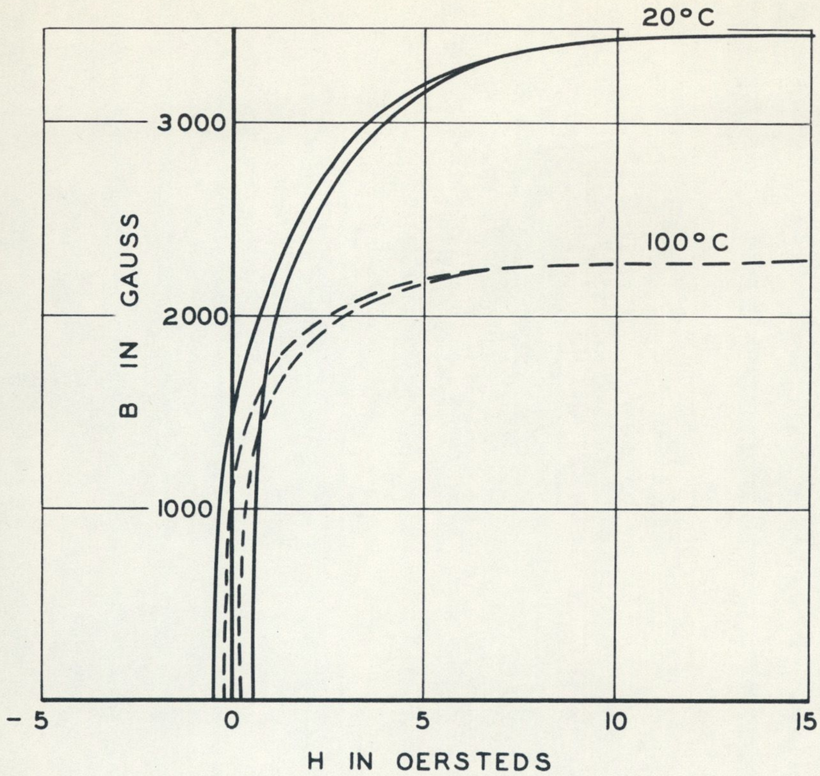


FIG. 10
B−H CURVES FOR FERROXCUBE TYPE 3B2

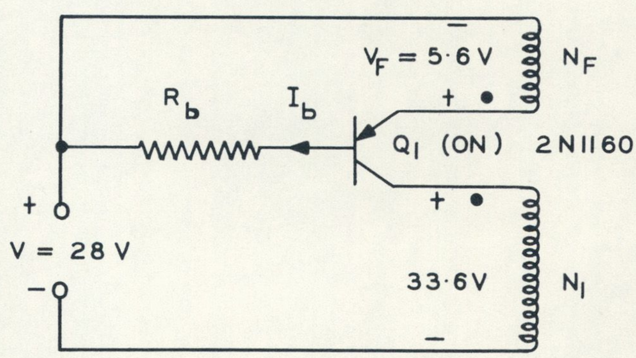
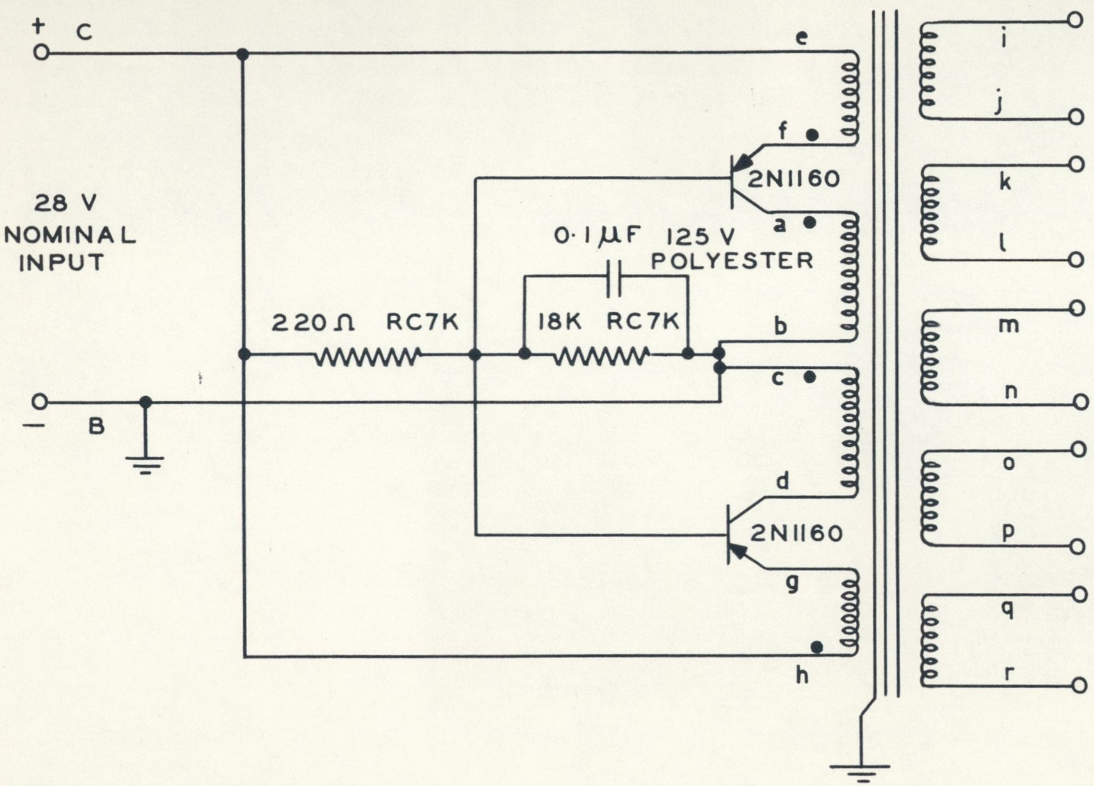


FIG. 11
SIMPLIFIED CIRCUIT FOR CONDUCTING TRANSISTOR



TRANSFORMER DETAILS

CORE : PHILIPS FERROXCUBE TYPE 3B2
DOUBLE STACK POT CORE SIZE
D36/22

AIR GAP — AS CLOSE TO ZERO
AS PRACTICABLE

WINDINGS

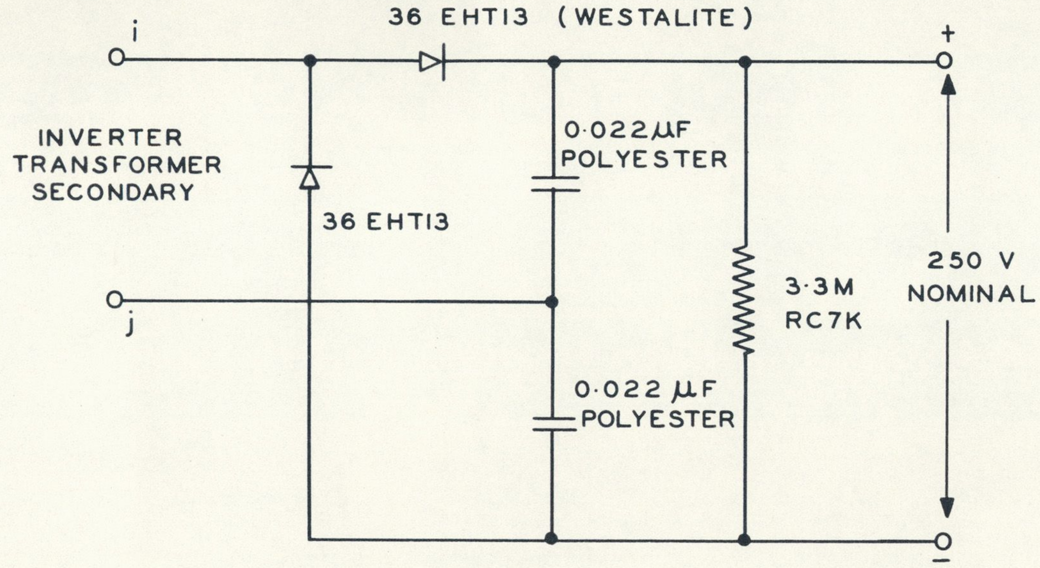
PRIMARY FORMER

ab/cd	120 TURNS [EACH] BIFILAR	27 B & S
ef/gh	20 TURNS [EACH] BIFILAR	27 B & S

SECONDARY FORMER

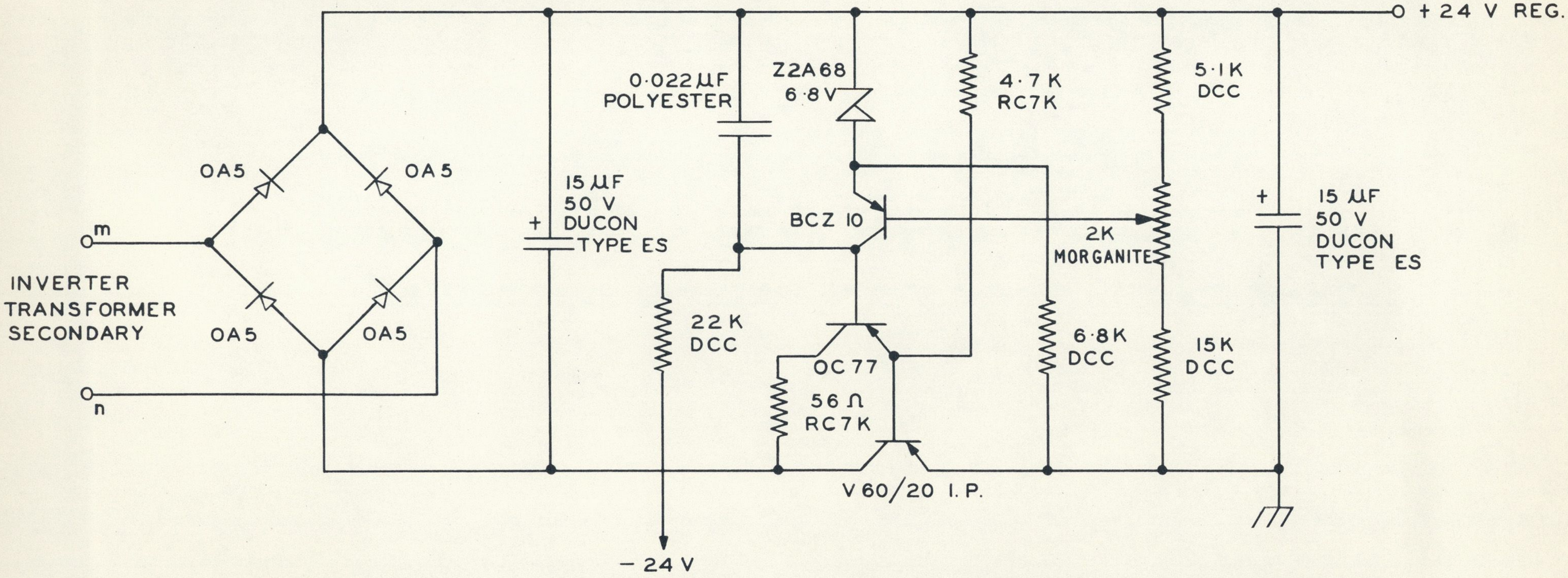
ij	625 TURNS	44	S.W.G.
kl	200 TURNS	32	B&S
mn	140 TURNS	32	B & S
op	40 TURNS	31	S.W.G.
qr	140 TURNS	31	S.W.G.

INVERTER CIRCUIT DETAILS



250 V UNREGULATED SUPPLY

INST. TECH. MEMO. 49
FIG. 13



+ 24 VOLT REGULATOR

INST. TECH. MEMO. 49
FIG. 14

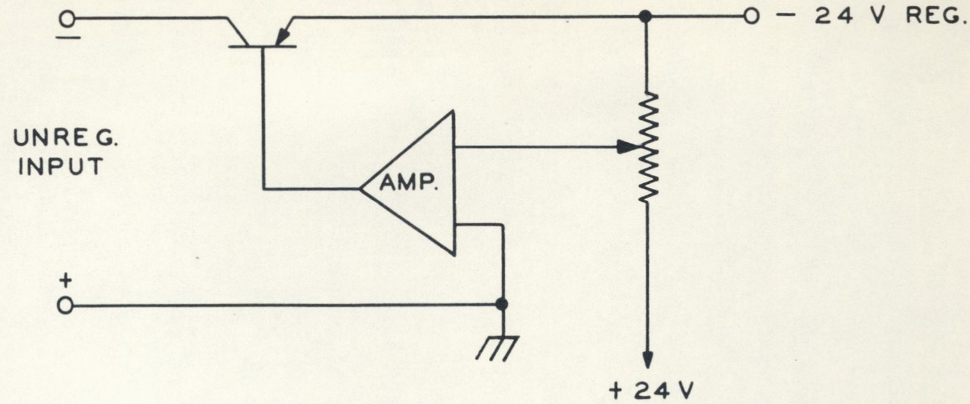


FIG. 15 SCHEMATIC OF — 24 V REGULATOR

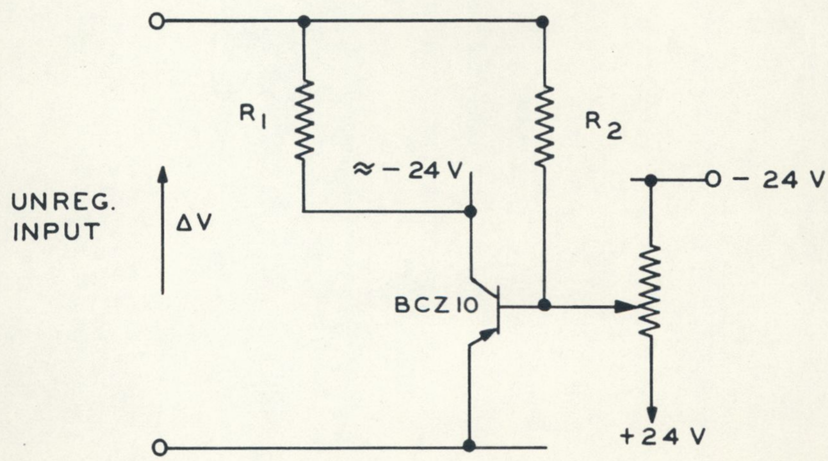
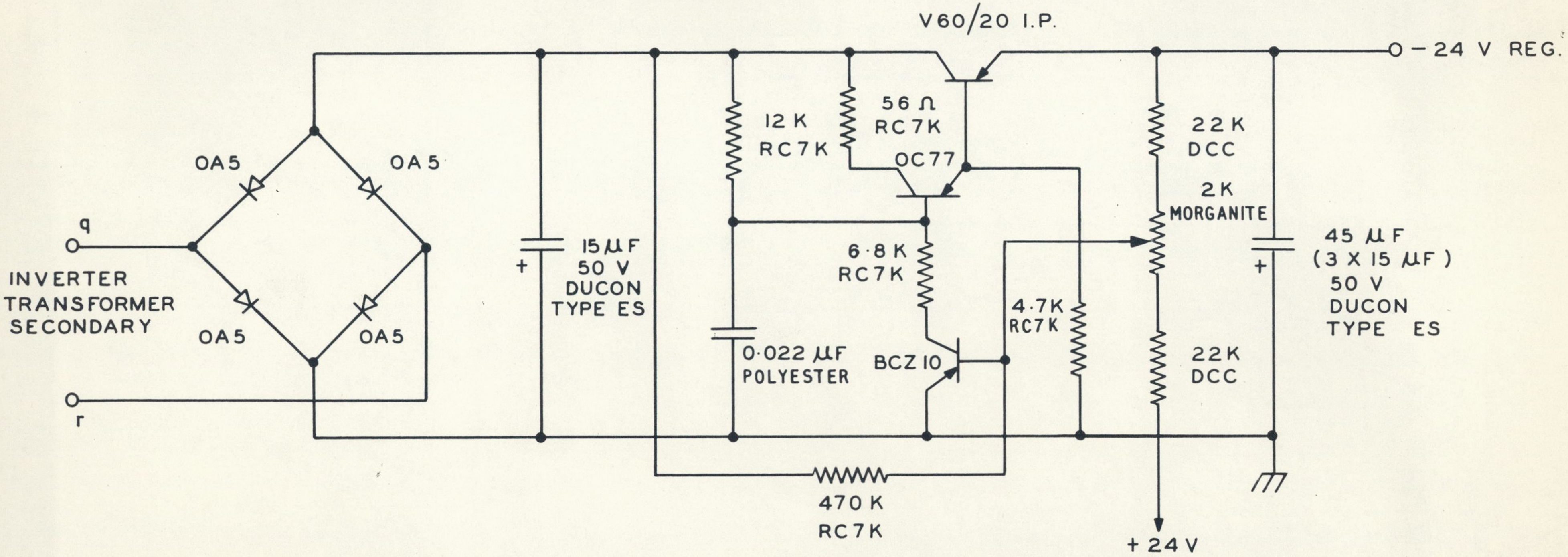
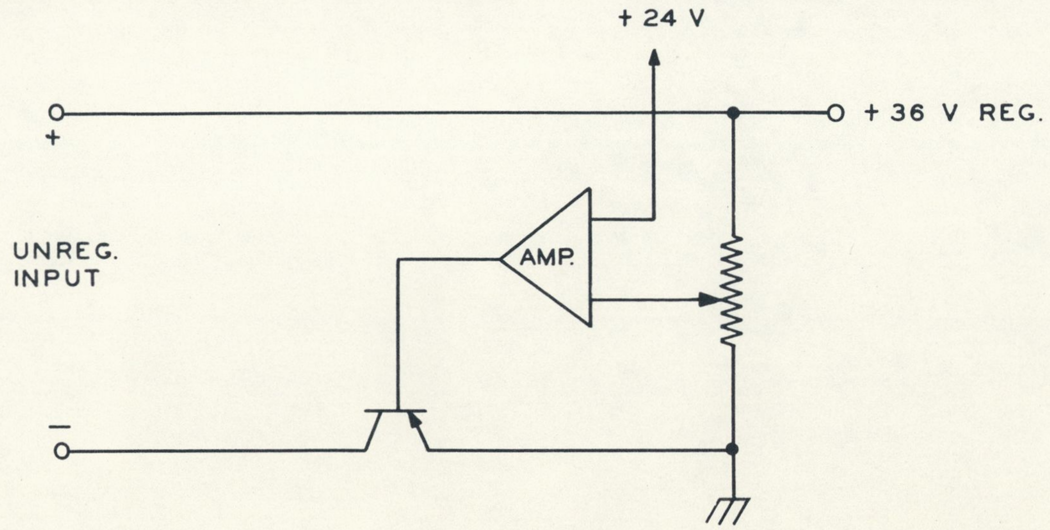


FIG. 16 COMPENSATING CIRCUIT FOR — 24 V REGULATOR



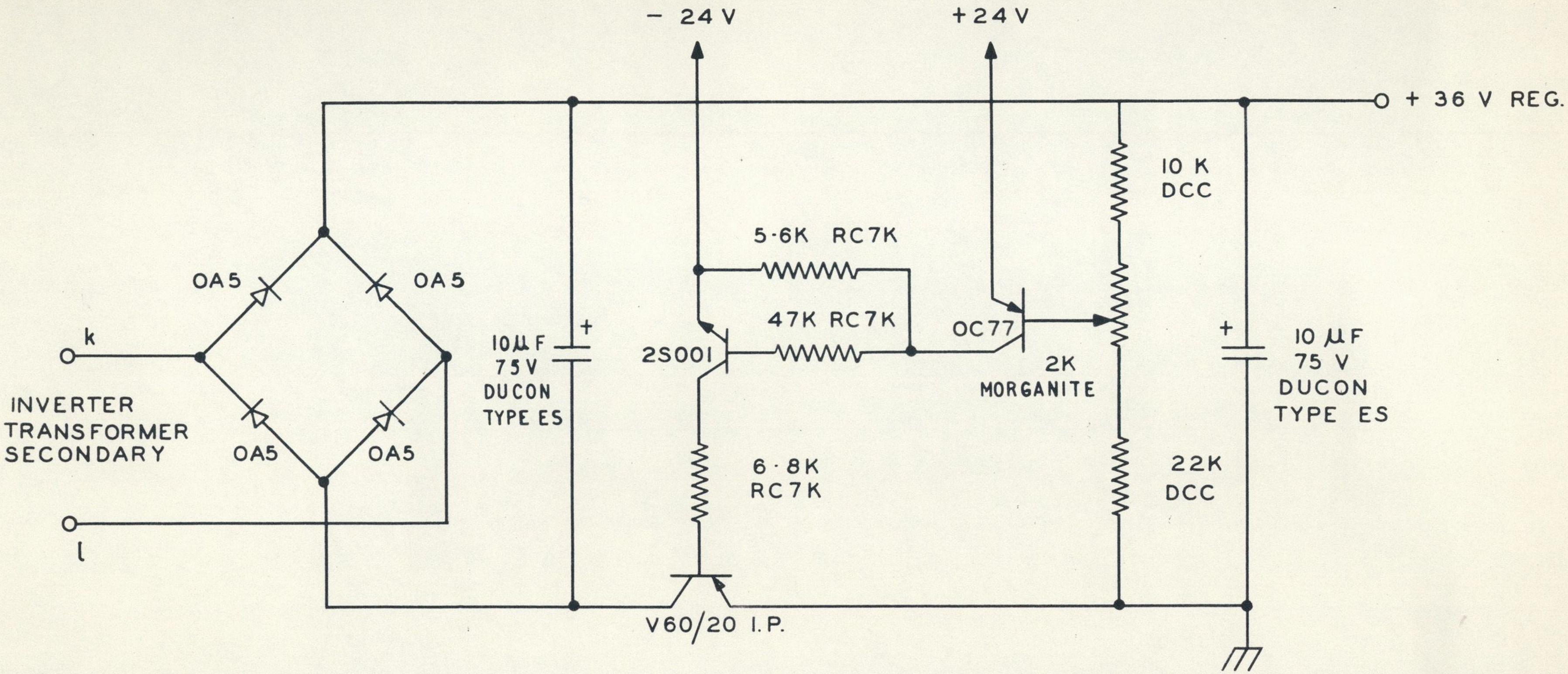
− 24 VOLT REGULATOR

INST. TECH. MEMO. 49
FIG. 17

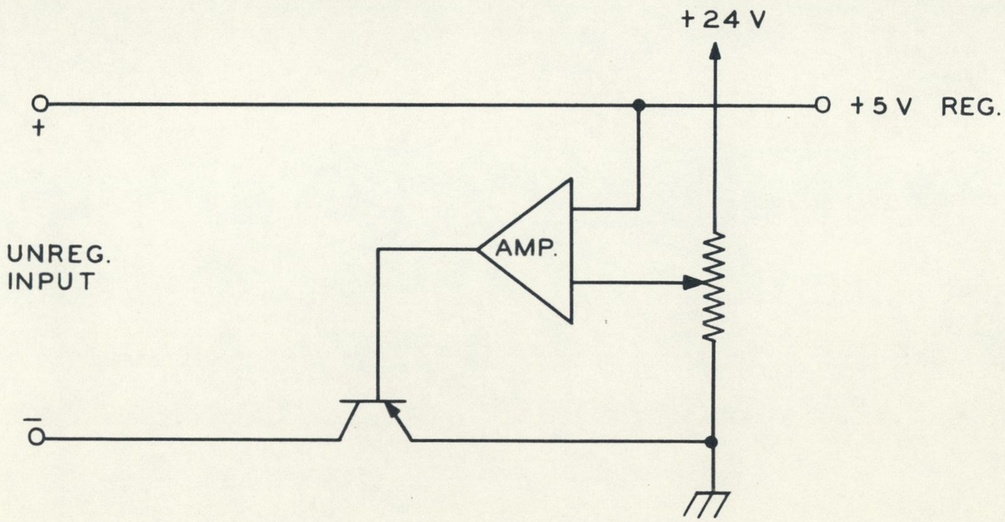


SCHEMATIC OF + 36 V REGULATOR

INST. TECH. MEMO. 49
FIG. 18

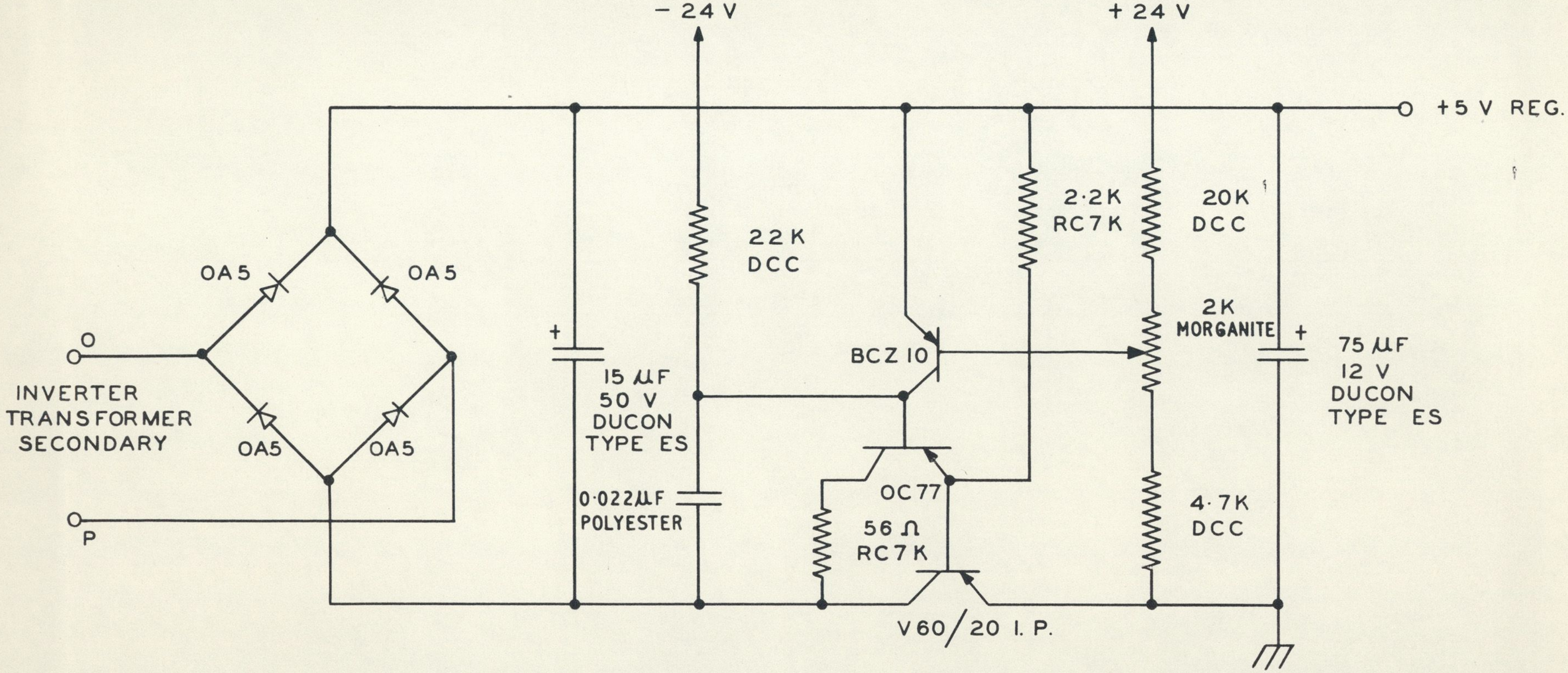


+ 36V REGULATOR



SCHEMATIC OF +5 V REGULATOR

INST. TECH. MEMO. 49
FIG. 20



+ 5V REGULATOR

INST. TECH. MEMO. 49
FIG. 21