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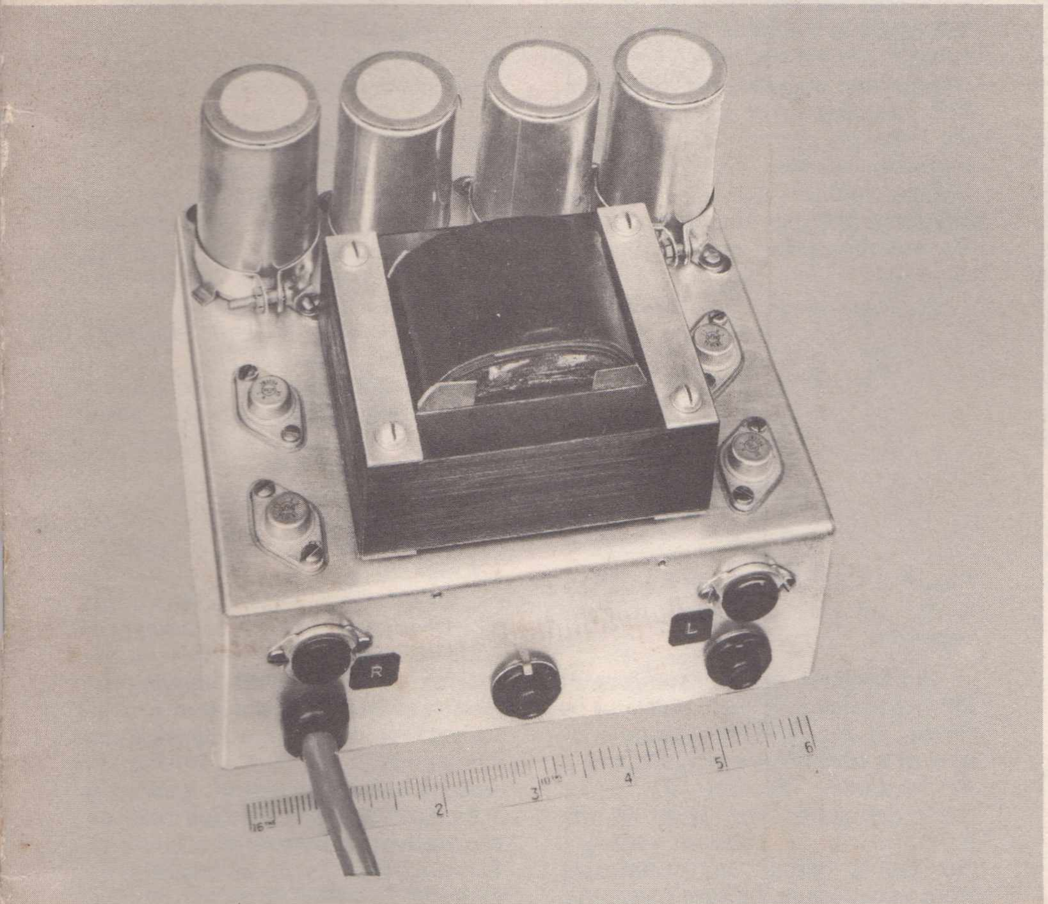

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AWV Stereo 15 + 15-watt Amplifier.

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SILICON TRANSISTORS FOR HIGH-VOLTAGE APPLICATION

by D. T. DeFino RADIO CORPORATION OF AMERICA

This note discusses several new applications for RCA high-voltage silicon transistors (2N3583, 2N3584, 2N3585, 2N3439 and 2N3440). These devices are triple-diffused n-p-n types featuring high frequency response, fast switching speeds, and low cost. Electrical characteristics are listed in Table I.

The advent of these types has made possible many new applications for transistors. Among these applications are circuits in which, until now, the use of transistors was restricted because of high operating voltages (horizontal-deflection circuits, for example). Other applications include those in which the use of a higher supply voltage can enhance circuit design, performance, and economy. Several other important applications are illustrated.

SERIES VOLTAGE REGULATOR

A voltage regulator provides a constant output voltage when the input voltage and/or output current is varied over a limited range. As shown in Fig. 1, the pass transistor, acting on a signal from the control circuit, prevents the output voltage V_{out} from varying. The control circuit receives a sample of the output voltage, compares it with a reference voltage, and amplifies the difference. The resulting error signal corrects the collector current I_C of the pass transistor so that the collector-to-emitter voltage V_{CE} is always the difference between the input voltage V_{in} and the desired output voltage.

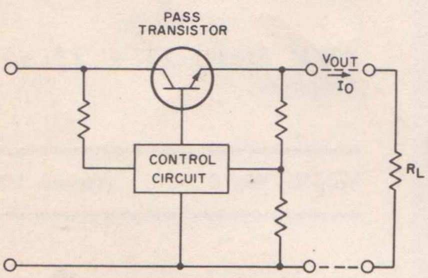


Figure 1. Basic form of a transistorised series voltage regulator.

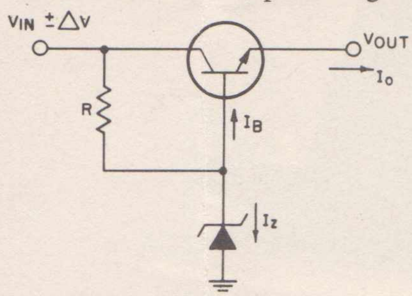


Figure 2. Simplest circuit arrangement for a transistor voltage regulator.

The simplest circuit arrangement for a transistor voltage regulator is shown in Fig. 2. The circuit consists of a transistor, a resistor, and a zener diode. Because the zener diode maintains the base of the transistor at a constant voltage, changes in output can result only from variations in the base-to-emitter voltage V_{BE} with current and temperature. A zener diode having a high current rating is required if large currents are drawn from the transistor.

The maximum value of resistance R which can be used in the circuit is determined as follows:

$$R = \frac{V_{in} - \triangle V - V_{out}}{I_B\ (max)}$$

Because the maximum base current $I_B(max)$ is equal to $I_o(max)/h_{FE}(min)$ where I_o is the output current and h_{FE} is the dc forward-current transfer ratio, the resistance equation can be rewritten as follows:

$$R = \frac{V_{in} - \triangle V - V_{out}}{I_o\ (max)} \times h_{FE}(min)$$

The zener diode must be capable of handling a peak current I_z given by

$$I_z = \frac{V_{in} + \triangle V - V_{out}}{R} = \frac{[V_{in} + \triangle V - V_{out}]\ [I_o(max)]}{[V_{in} - \triangle V - V_{out}]\ [h_{FE}(min)]}$$

In the series regulator, the pass transistor must remain always in the active region. For this reason, the pass transistor must be chosen carefully to avoid dc forward-bias second breakdown. As shown in Fig. 3, under the worst-case condition $I_o(max)$, $V_{in}(min)$, the bias point of the transistor must be within the dc forward-bias second-breakdown rating $P_{s/b}$, or the dc power-dissipation rating P_{dc}, whichever is the limiting factor. From the equations given

Maximum Ratings, Absolute-Maximum Values

	2N3583	2N3584	2N3585	2N3439	2N3440	
COLLECTOR-TO-BASE VOLTAGE, V_{CBO}	250	375	500	450	300	Volts
COLLECTOR-TO-EMITTER VOLTAGE, V_{CEO} (sus)	175	250	300	350	250	Volts
EMITTER-TO-BASE VOLTAGE, V_{EBO}	6	6	6	7	7	Volts
CONTINUOUS COLLECTOR CURRENT, I_C	2	2	2	1	1	Amp
PEAK COLLECTOR CURRENT	5	5	5	—	—	Amp
BASE CURRENT, I_B	1	1	1	0.5	0.5	Amp
TRANSISTOR DISSIPATION, P_T	35	35	35	5	5	Watts

Table I. Electrical chacteristics of RCA high-voltage silicon transistors.

above, it is obvious that near the operating point h_{FE} should be as high as possible. In general, leakage current and saturation voltage are not important.

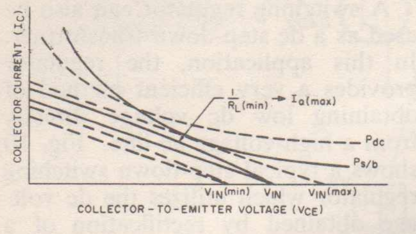


Figure 3. Transistor load line.

DESIGN EXAMPLE

The following conditions are specified for a series voltage regulator:

$$V_{out} = 100\text{ V}$$
$$I_o(\max) = 400\text{ mA}$$
$$V_{in} = 135 \pm 15\text{ V}$$
$$h_{FE}(\min) = 20$$

Circuit values are then determined as follows:

$$R = \frac{(135-15-100)\,20}{0.4} = \frac{400}{0.4} = 1\text{ k}\Omega \text{ at } 2.5\text{ W}$$

$$I_B(\max) = \frac{0.4}{20} = 20\text{ mA}$$

$$I_Z = \frac{135+15-100}{1000} = \frac{50}{1000} = 50\text{ mA}$$

Therefore, the zener-diode requirements are $V_Z = 100$ V, $I_Z = 50$ mA, $P_Z = 5$ W. Under worst-case conditions, the transistor must be capable of handling 400 milliamperes at 50 volts, or a dissipation of 20 watts. In addition, the point 50 V and 400 mA must be within the dc second-breakdown rating of the transistor. Fig. 4 shows the circuit values for this regulator.

The power-dissipation rating of the resistor and zener diode can be reduced by addition of another transistor (usually much smaller in dissipation) in a configuration such as that shown in Fig. 5. This arrangement effectively increases the overall minimum gain. The two transistors can be regarded as one in which the effective h_{FE} (approximately the product of the gain of the two transistors) can be substituted for h_{FE} in the previous equations. Because the 2N3440 has a minimum gain of 40 at 20 mA, the minimum effective gain is (40)(20) = 800. From this value, the new resistor and zener diode requirements can be calculated as follows:

$$R = \frac{(135-15-100)\,800}{0.4} = 40\text{ k}\Omega \text{ at } 0.062\text{ W}$$

$$I_Z = \frac{135+15-100}{40000} = \frac{50}{40000} = 1.25\text{ mA}$$

$$P_Z = 125\text{ mW}$$

The maximum power dissipated by the 2N3440 transistor in this circuit is (20 mA) (50 V) = 1 W.

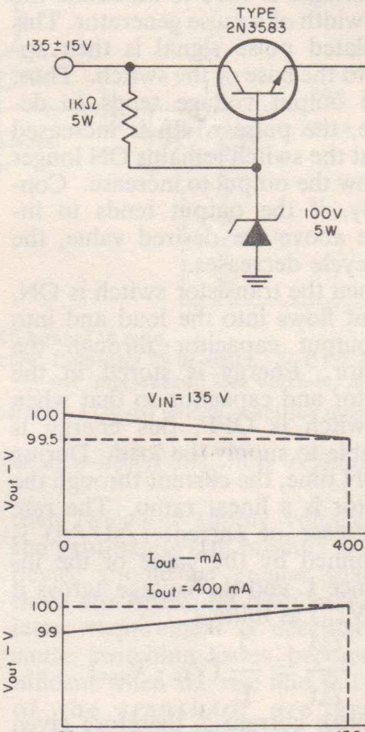


Figure 4. Schematic diagram of a simple transistor voltage regulator.

The disadvantage of the circuit of Fig. 5 as compared with that of Fig. 4 is that voltage regulation is less sensitive because there are two junctions to create V_{BE} variations with current and voltage changes.

Fig. 6 shows a feedback arrangement designed to improve regulation. In this circuit, the output is sampled and compared with a very stable reference voltage. The resulting error signal is used to adjust the bias on the pass transistor. The requirements for Q3 are determined in the same manner as those for the zener diode in the preceding circuits. The zener-diode current $I_Z(\max)$ is equal to the collector current $I_C(\max)$ of Q_3 divided by the minimum gain of Q_3 at $I_C(\max)$.

In general, the full load voltage need not be fed back. Instead, a voltage divider can be used to reduce the voltage requirement on the zener diode. Although the voltage divider also degrades the performance, this method must be used if a variable output voltage is required. Fig. 7 shows a typical high-voltage regulator that provides an output variable from 175 to 225 volts and delivers up to 150 mA. Performance curves for this circuit are shown in Fig. 8.

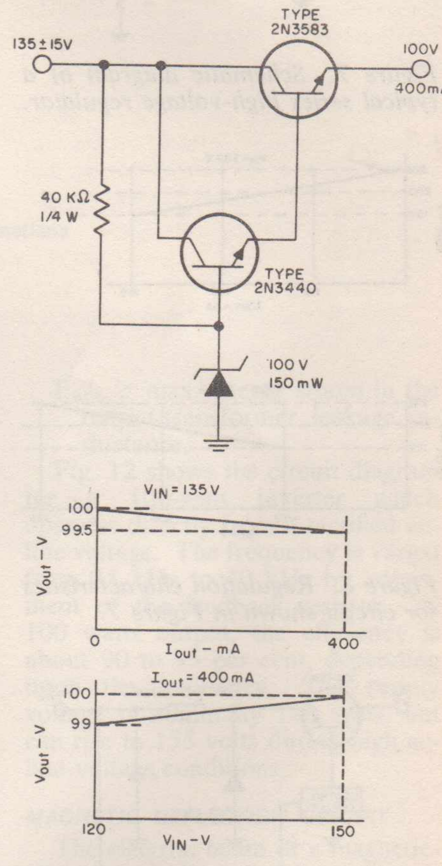


Figure 5. Schematic diagram of a series voltage regulator using darlington driver.

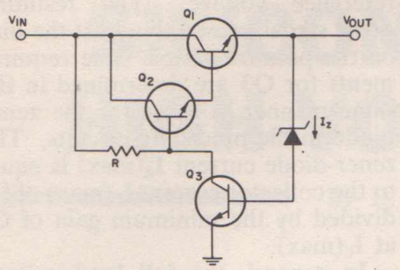


Figure 6. Schematic diagram of a series voltage regulator employing feedback amplifier.

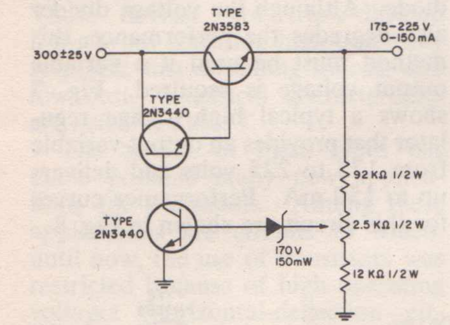


Figure 7. Schematic diagram of a typical series high-voltage regulator.

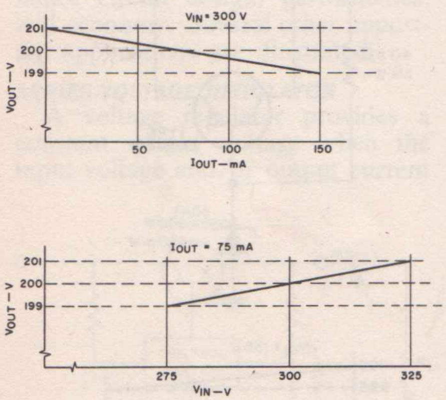


Figure 8. Regulation characteristics for circuit shown in Figure 7.

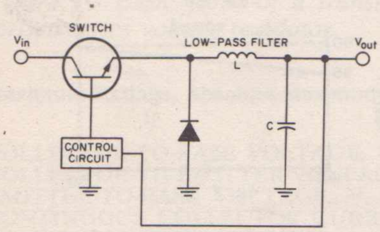


Figure 9. Simplest form of a transistor switching regulator.

SWITCHING REGULATOR

The advantage of a transistorized switching regulator, such as that shown in Fig. 9, is its extremely high efficiency. It does not, however, provide the excellent regulation obtainable from a series-type regulator. For this reason, a switching regulator is normally used as a coarse or preregulator preceding a series regulator. The switching regulator is highly efficient because the transistor switch is either saturated or cut off. Because both of these conditions are states of low dissipation, very little power is lost in the transistor.

The function of the feedback circuit is to sample the output voltage and compare it with a reference voltage. The difference between these two voltages is used to modulate the pulse width of a pulse generator. This modulated pulse signal is then applied to the base of the switch. Thus, if the output voltage tends to decrease, the pulse width is increased so that the switch remains ON longer to allow the output to increase. Conversely, if the output tends to increase above the desired value, the duty cycle decreases.

When the transistor switch is ON, current flows into the load and into the output capacitor through the inductor. Energy is stored in the inductor and capacitor so that when the switch is OFF, this energy is available to supply the load. During the ON time, the current through the inductor is a linear ramp. The rate of increase of current ($\triangle I/\triangle t$) is determined by the value of the inductance L and the voltage across it (V_{in}-V_{out}) as follows:

$$\frac{\triangle I}{\triangle t} = \frac{1}{L}(V_{in} - V_{out})$$

The peak current is therefore given by

$$I_p = \frac{V_{in} - V_{out}}{L}(t_{on})$$

The transistor chosen for this application must provide sufficiently fast switching times, i.e., rise time t_r and fall time t_f. For good regulation over a wide range of input voltage and output current, the duty cycle must be variable from 10 to 90 per cent. Consequently, the minimum pulse width should be one-tenth of the period (1/10 f). For low switching losses, the rise and fall times should be about one-fifth of the minimum pulse width, or one-fiftieth of the frequency of the pulse generator (1/50 f).

A switching regulator can also be used as a dc step-down transformer. In this application, the regulator provides a very efficient method of obtaining low dc voltage directly from a high-voltage ac line. Fig. 10 shows a typical step-down switching regulator which utilizes the dc voltage obtained by rectification of a 117-volt ac line source to provide a regulated 60-volt supply. Performance characteristics for the circuit are shown in Fig. 11.

INVERTERS

An inverter is used to transform dc power to ac power. If the ac output is rectified and filtered to provide dc again, the over-all circuit is referred to as a converter. A converter is normally employed to change the magnitude of an available dc supply.

A transistorized inverter can be made very light in weight and small in size. It is a highly efficient circuit and, unlike its mechanical counterpart, has no moving components. The output from the inverter can be used to drive any equipment which requires an ac supply (motors, ac radios, television receivers, fluorescent lights, and the like). Another very important application of an inverter is in driving the electromechanical transducers used in ultrasonic equipment (such as ultrasonic cleaners and sonar detection devices).

The operating frequency of an inverter is usually fixed between 60 Hz and 100 kHz, depending upon the application. For applications in which the operating frequency can be chosen by the designer, the highest possible frequency should be selected.

In general, the size and weight of the inverter can be decreased as the supply voltage and frequency are increased. This relation results mainly from the decreasing size of the transformer needed. The upper frequency and supply voltage are limited by the transistors used. The collector-to-emitter breakdown voltage, for example, must be greater

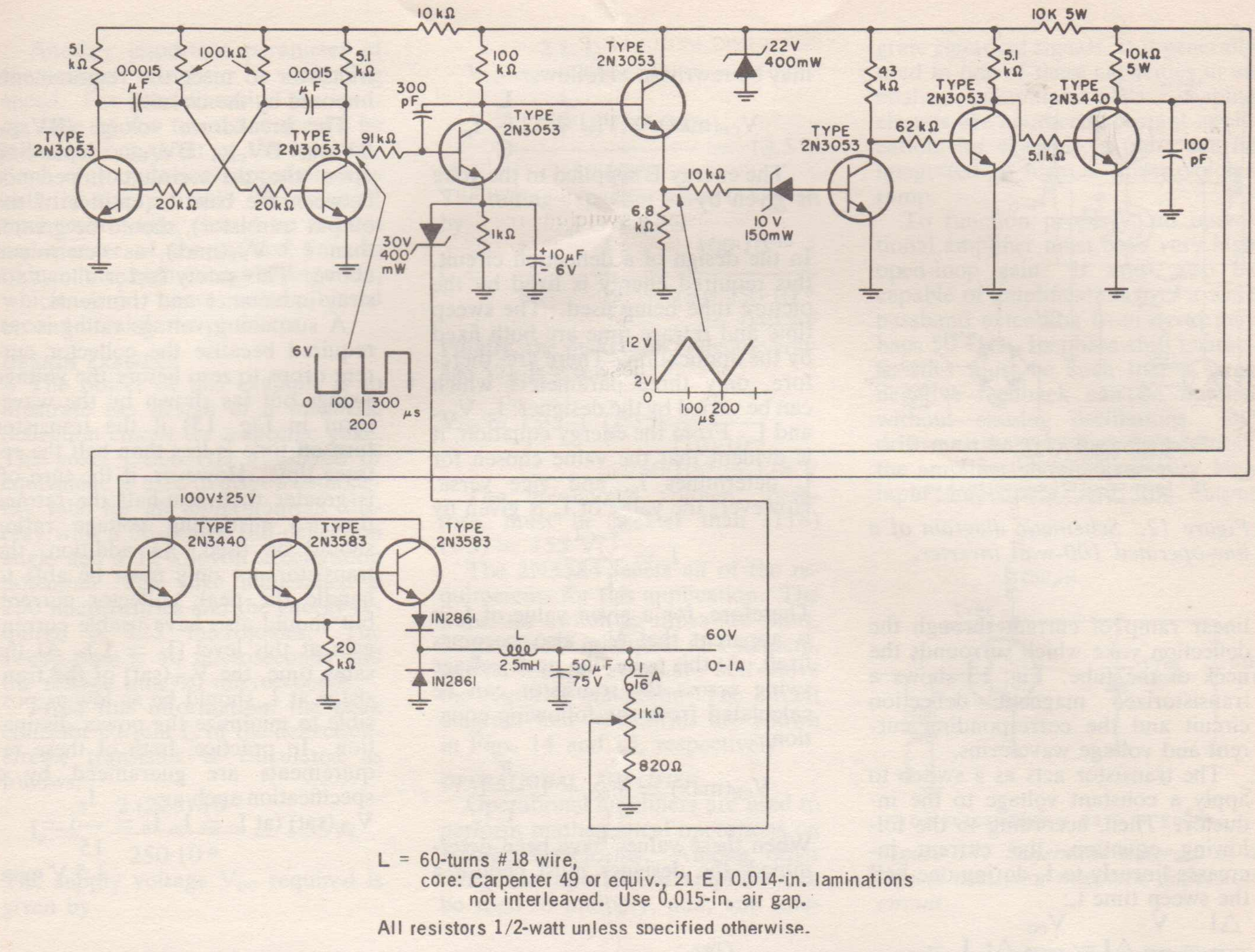


Figure 10. Schematic diagram of a typical step-down switching regulator.

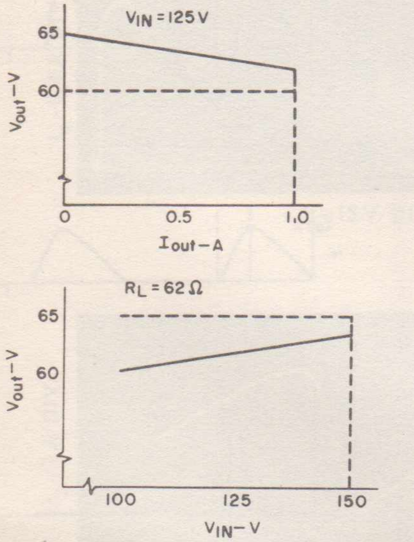


Figure 11. Performance curves for circuit shown in Figure 10.

than twice the supply voltage, and the gain-bandwidth product f_T of the device should be greater than ten times the operating frequency. The latter requirement is necessary because switching losses become significant when the rise and fall times of the transistor are greater than about one-fifth of the pulse width.

The important parameters to be considered in the selection of a transistor for an inverter circuit are summarized below:

$V_{CER}(sus) \geq 2V_{CC}$ + leakage reactance spikes

High gain (to reduce feedback power and increase efficiency)

$f_T \geq 10$ f (to reduce switching losses)

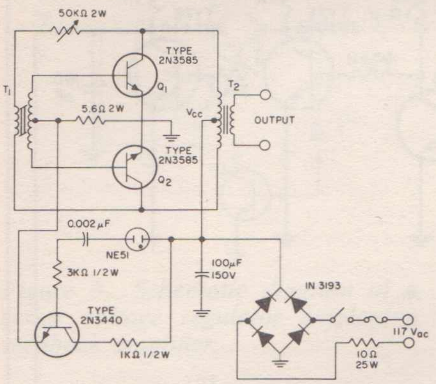
$I_S/_b \geq$ highest starting bias current at V_{CC}

$E_S/_b \geq$ max. energy stored in the output-transformer leakage inductance.

Fig. 12 shows the circuit diagram for a 100-watt inverter which operates directly from a rectified ac-line voltage. The frequency is varied from 25 kHz to 40 kHz by adjustment of the feedback resistor. At 100 watts output, the efficiency is about 90 to 95 per cent, depending upon the frequency. The supply voltage is nominally 140 volts, but can rise to 155 volts during high ac-line-voltage conditions.

MAGNETIC DEFLECTION CIRCUIT

The electron beam of a magnetically driven display tube is swept across the face of the tube by a linearly changing magnetic field. This deflecting field is produced by a



T_1 = Allen Bradley RO-3 (E I 102 H 142 A) or equiv.
primary: 160-turn #32 wire;
secondary: each 3-turns #32 wire.

T_2 = Indiana General C2 material (CF216) or equiv.
primary and secondary: 80-turns #28 wire.

Figure 12. Schematic diagram of a line-operated 100-watt inverter.

linear ramp of current through the deflection yoke which surrounds the neck of the tube. Fig. 13 shows a transistorized magnetic deflection circuit and the corresponding current and voltage waveforms.

The transistor acts as a switch to apply a constant voltage to the inductor. Then, according to the following equation, the current increases linearly to I_p during one-half the sweep time t_s:

$$\frac{\triangle I}{\triangle t} = \frac{V}{L} \quad \triangle I = \frac{V_{CC}}{L} \triangle t, \; I_p = \frac{V_{CC}}{L}\frac{t_s}{2}$$

When the transistor is turned off, LC forms a tuned circuit in which the yoke current decreases very rapidly (retrace time t_r) through zero to $-I_p$. At this point capacitor C has a negative voltage across it, the diode is forward-biased, and the yoke current begins to increase toward zero. At this point the cycle begins again.

During the retrace time, when the yoke current is decreasing from I_p to $-I_p$, the voltage across the transistor becomes quite high. The collector-to-emitter voltage is given by

$$V_{CE}(max) = V_{CC} + I_p \, \omega \, L$$

The term ω can be expressed as follows:

$$\omega = \frac{1}{LC} = \frac{\pi}{t_r}$$

Therefore, the equation for $V_{CE}(max)$ may be rewritten as follows:

$$V_{CE}(max) = V_{CC} + \frac{L}{C} I_p.$$

The energy E supplied to the yoke is given by

$$E = \tfrac{1}{2} L I_p^2$$

In the design of a deflection circuit, this required energy is fixed by the picture tube being used. The sweep time and retrace time are both fixed by the application. There are, therefore, only three parameters which can be varied by the designer: I_p, V_{CC}, and L. From the energy equation, it is evident that the value chosen for L determines I_p, and vice versa. However, the value of I_p is given by

$$I_p = \frac{V_{CC}}{L}\frac{t_s}{2}$$

Therefore, for a given value of I_p it is apparent that V_{CC} also becomes fixed. At this point, the peak voltage swing across the transistor can be calculated from the following equation:

$$V_{CE}(max) = V_{CC} + I_p \frac{\pi}{t_r} L$$

When these values have been determined, the designer must choose a transistor to meet the requirements imposed by the circuit.

The breakdown voltage (BV_{CEO}, BV_{CER}, BV_{CES}, BV_{CEX}, depending upon the drive-circuit impedance between the base to emitter of the output transistor), should be greater than 1.3 $V_{EE}(max)$, as determined above. This safety factor allows for stray inductance and transients.

A sustaining voltage rating is not required because the collector current drops to zero before the voltage swings out (as shown by the waveform in Fig. 13) if the transistor turn-off time is less than half the retrace time. However, if the turn-off is greater than one-half the retrace time, a sustaining voltage rating should be used. In addition, the transistor not only must be able to handle the peak collector current, but should also have usable current gain at this level ($I_c = I_p$). At the same time, the $V_{CE}(sat)$ of the transistor at I_p should be as low as possible to minimize the power dissipation. In practice, both of these requirements are guaranteed by a specification such as:

$$V_{CE}(sat) \; (\text{at } I_C = I_p, \; I_B = \frac{I_p}{15}) = 1.5 \text{ V max.}$$

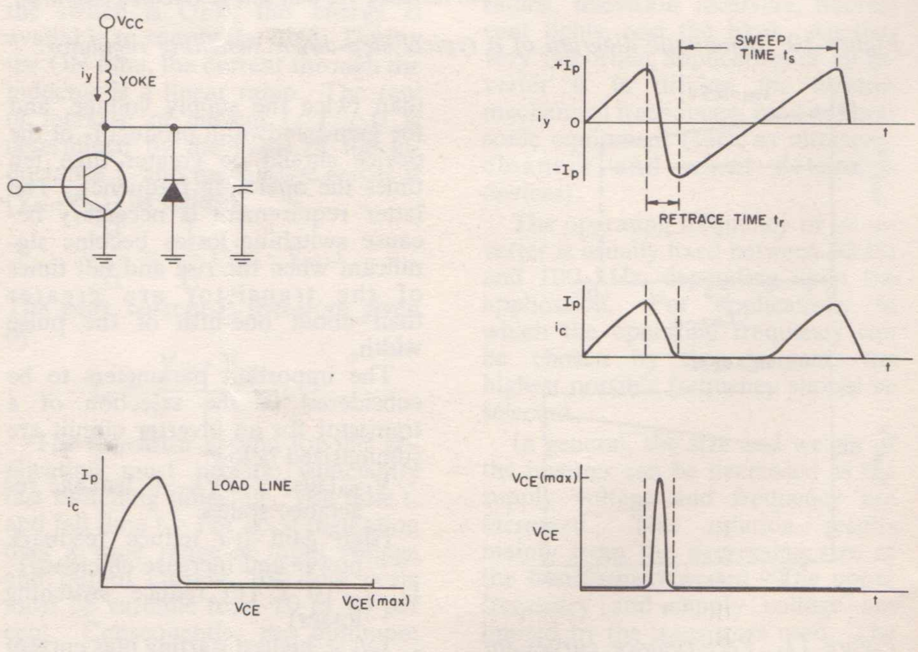


Figure 13. Basic configuration for a transistor magnetic deflection circuit showing corresponding current and voltage waveforms.

Another important parameter of the output transistor is switching speed. For good linearity, the turn-on time of the transistor should be less than one-tenth of the total on-time of the device (approximately half the sweep time). The turn-off time, meanwhile, should be at least one-quarter of the retrace time to reduce the high-energy dissipation, which could cause reverse-biased second-breakdown problems.

DESIGN EXAMPLE

The object of this example is to illustrate the design of a magnetic deflection circuit for a specific yoke. The yoke, Celco HD 428-S560 or equivalent, is used to drive a cathode-ray tube for an alpha-numeric display with a 36-degree full-deflection angle and a 12-kilovolt acceleration potential. The yoke inductance is 250 microhenries and the energy required is 225 microjoules. The sweep time is 50 microseconds and the retrace time 10 microseconds.

From this information, the peak collector current I_p of the deflection-circuit transistor is calculated as follows:

$$I_p = \frac{2\,(225)\,10^{-6}}{250 \cdot 10^{-6}} = 1.35 \text{ A}$$

The supply voltage V_{CC} required is given by

$$V_{CC} = \frac{2\,L\,I_p}{t_s} = \frac{2\,(250 \cdot 10^{-6})\,(1 \cdot 35)}{50 \cdot 10^{-6}} = 13.5 \text{ V}$$

The tuning-capacitor value C s given by

$$C = \left(\frac{t_r}{\pi}\right)^2 \frac{1}{L} = \frac{100 \cdot 10^{-12}}{(\pi)^2\, 250 \cdot 10^{-6}} = \cdot 040\ \mu\text{F}$$

Finally, the maximum collector voltage V_{CE} is given by

$$V_{CE} = 13 \cdot 5 + (1 \cdot 35)\,\frac{\pi}{(10) \cdot 10^{-6}}\, 250 \cdot 10^{-6} = 118 \text{ V}$$

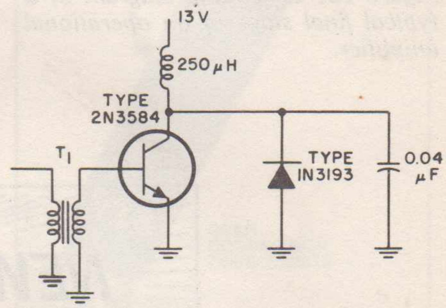
The breakdown voltage, therefore, must be greater than (118) (1·3) = 155 V.

The 2N3584 meets all of the requirements for this application. The transistor switching times are short, its gain is 25 minimum at 1 ampere, and its voltage ratings are well above the required minimum. The circuit diagram and waveforms are shown in Figs. 14 and 15, respectively.

OPERATIONAL AMPLIFIER

Operational amplifiers are used to perform mathematical operations on voltage waveforms. Among other things, an operational amplifier can be used to multiply, add, and integrate electrical signals. It is generally used in one of these capacities in an analog computer. Wave-shaping circuits are another important application; for example, a pulse can be integrated to form a linear voltage ramp.

To function properly, an operational amplifier must have very high open-loop gain. It must also be capable of amplification over a wide passband extending from dc to perhaps 50 kHz. Its phase-shift characteristics must be such that a large negative feedback can be applied without causing oscillations. DC drift must be very low. In addition, the amplifier should have very high input impedance and low output



T_1 = C.P. ELECTRONICS X-9370 OR EQUIV.

Figure 14. Schematic diagram of a typical transistor magnetic deflection circuit.

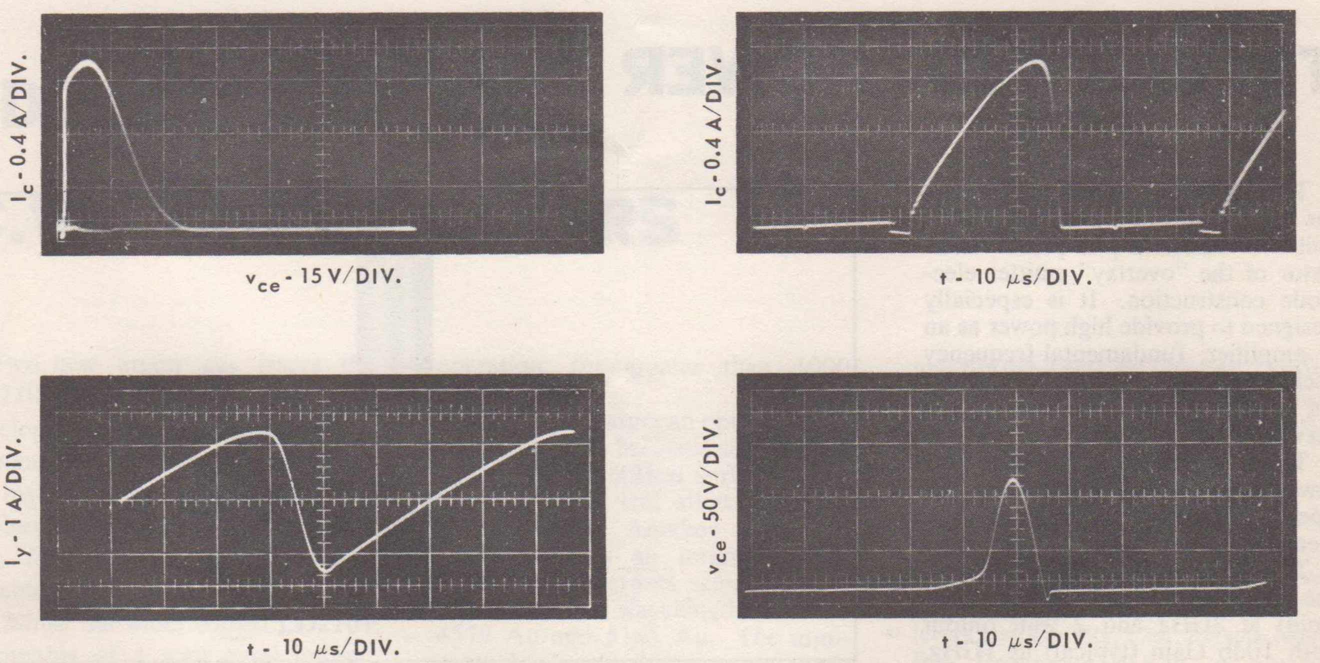


Figure 15. Current and voltage waveforms produced by circuit shown in Figure 14.

impedance, or vice versa. Generally, the high-input-impedance type is used.

To meet all of these requirements, an operational amplifier normally utilizes a chopper amplifier and other stabilizing circuits. This portion of the amplifier can be designed to operate at low supply voltages. The final stage, however, requires a high supply voltage because it must provide a large voltage swing to drive the high input impedance of the next operational amplifier. A typical final stage that meets this requirement and also provides the necessary low output impedance is shown in Fig. 16. Fig. 17 shows the performance curves for this circuit.

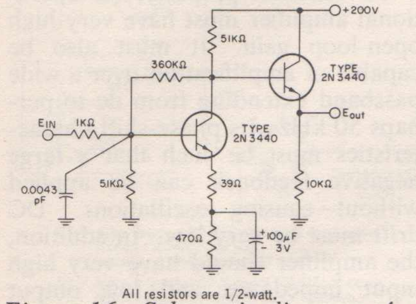


Figure 16. Schematic diagram of a typical final stage of an operational amplifier.

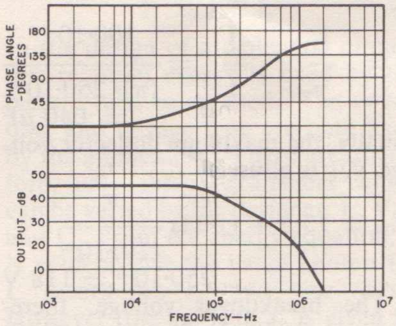


Fig. 17. Performance curves for circuit shown in Figure 16.

In general the transistor requirements for an operational amplifier output are the same as for a class A audio amplifier. These requirements are summarized below:

$V_{CER}(sus) > 2\ V_{CC}$

h_{FE}: must be linear over the operating-current range.

$P_S/_b/P_D$: the dc bias point must be within the safe operating region.

f_T: the gain-bandwidth product should be as high as possible; a rule-of-thumb minimum is 10 MHz.

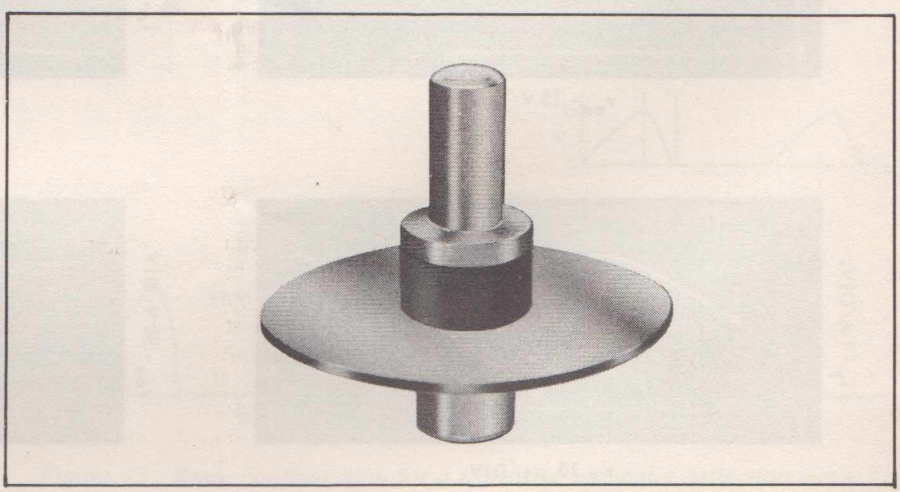
NEWS & NEW RELEASES

TA7003 UHF POWER TRANSISTOR

The RCA developmental transistor type TA7003 is a microwave epitaxial silicon n-p-n planar transistor of the "overlay" emitter electrode construction. It is especially designed to provide high power as an rf amplifier, fundamental-frequency oscillator, and frequency multiplier for military and communications service.

The transistor features an all new low-inductance coaxial package for operation of the TA7003 at UHF frequencies, into S-band.

The power output capabilities of the device are 1 watt with 5db Gain (min) at 2GHz and 2 watt output with 10db Gain (typical) at 1GHz.



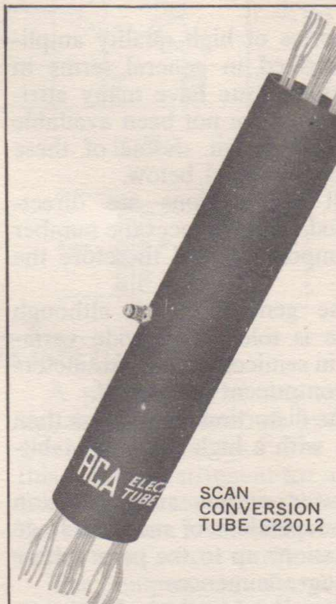
C22012 SCAN CONVERSION TUBE ASSEMBLY

The recently announced RCA Developmental Scan Conversion Tube Assembly Type C22012 is a very small, ruggedized, electrical-input, electrical-output cathode-ray charge-storage tube having integral reading-gun focusing coil, reading and writing gun deflecting yokes, electromagnetic shielding, as well as encapsulated leads. This assembly is suited for use in equipment meeting the environmental specifications of MIL-E-5400 and is evaluated in accordance with the requirements of MIL-STD-810A. The C22012 assembly is designed especially for use in military and aerospace systems where space conservation, weight, and low-power consumption are of importance.

The C22012 employs an EBIC (Electron Bombardment Induced Conductivity) type of target having a minimum useful diameter of 1 inch, a magnetic-focus, magnetic-deflection type reading gun, and an electrostatic-focus, magnetic-deflection type writing gun. The deflection angle of each gun is about 35° edge-to-edge.

Factory alignment of the integral deflecting yokes and the focusing coil assures optimum overall performance from the C22012 and simplifies set-up procedure.

Despite its small size (10″ long, 2″ diameter) the C22012 assembly has high resolution capability and is intended for applications employing television monitors utilizing 500 to 1000 scanning lines.



SCAN CONVERSION TUBE C22012

LD2100, LD2101 C.W. ARGON LASERS

Two new argon gas lasers the LD2100 and LD2101 have been developed by RCA. They are continuous duty, dc excited argon lasers for high density coherent light, with predominant blue and green lines. The forced aircooled LD2100 is capable of 100 milli-watts total output whilst the water cooled LD2101 is capable of 1 watt total output, both types are capable of stable operation, for greater than 1000 hours.

The lasers feature an optical cavity with removable hermetic seals, thereby minimizing optical surface maintenance, whilst still allowing access to the cavity. Another feature of these lasers is an internal cavity prism which permits selection of a minimum of six wavelengths between 4579 Au and 5145 Au. The minimum fundamental transverse mode power output for the LD2100 is 50 mw for the 4880 Au line and 25mW for the 5145 Au line whilst that for the LD2101 is 0.5W and 0.3W respectively.

A regulated DC power supply is included as an integral part of each unit.

For further information on these lasers contact Amalgamated Wireless Valve Co. Pty. Ltd., Private Mail Bag, Ermington, N.S.W., or any of their interstate offices

3-130 WATT AUDIO AMPLIFIERS (PART 2)

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The series of high quality amplifiers described in general terms in the previous issue have many attributes which have not been available in previous design. Some of these attributes are listed below.

1. All the sections are direct-coupled. This reduces the number of components and therefore the cost.
2. The general circuit although simple is tolerant to wide variations in semiconductor parameters and component tolerances.
3. The distortion level is less than 0.1% with a high order of stability.
4. Power output and distortion are independent of mains voltage fluctuations up to the point where clipping commences.
5. Class B operation of the output stage keeps the units efficiency high, which is particularly important for high power amplifiers.
6. The noise level of the amplifiers is low and in multichannel systems the channel separation will be comparable with the noise over a large part of the frequency range.
7. More than one amplifier can be operated on the one simple power supply with negligible interaction between them.
8. The general design can be used for many applications other than audio: these include servo amplifiers, motor controls, high speed power supplies etc.
9. The high powered amplifiers are protected against the possibility of component damage due to output load short circuits.
10. The power supply is quite simple, as the performance of the amplifier is unaffected by supply ripple.
11. The frequency and power response of the amplifiers are not dependent upon the output power level.
12. As the damping factor is greater than 100 the load impedance is very low thus extremely effective damping of the speaker is produced. This of course is a desirable feature in high quality amplifiers.
13. Safe operation in ambient temperatures up to 65° C (149° F) is possible.
14. Tone burst tests indicate excellent overload characteristics particularly for the smaller amplifiers and low level versions of the high power systems.
15. Tests have indicated that the intermodulation distortion is better than 0.5%.

Further details can be obtained from the tabulated performance table and graphs included in the latter part of this article.

TABLE 2 **COMPLETE ELECTRICAL SPECIFICATION OF AMPLIFIERS**

Circuit number	2	3	3	4	4	4	5	5	5	
Nominal power output	3	10	15	25	35	50	70	100	130	W
Output load impedance	15	15	15	12	8.5	5.0	10	6.5	5.0	Ω
Unloaded supply voltage	37	51	60	65	65	65	82	82	82	V
Power output at 10% distortion	4.7	14	21	37	50	62	95	130	150	W
Power output at 1% distortion	2.7	12	17	30	39	52	75	105	130	W
Power output at clipping	3.3	11.5	16	27	37	51	71	105	130	W
Distortion before clipping	1.2	0.1	0.1	0.5	0.5	0.5	0.7	1.0	1.0	%
Overall feedback	27	40	40	25	25	25	25	25	25	dB
Sensitivity for full output	175	270	325	90	95	100	130	125	117	m V
Input impedance	3.0	3.0	3.0	3.0	3.0	3.0	10	10	10	KΩ
Noise below full output	69	80	80	70	70	70	75	75	75	dB
Low frequency at –3db (3)	35	12	12	50	50	50	15	15	15	Hz
High frequency at –3db (3)	30	30	30	20	20	20	10	10	10	KH_z
Output heat sink thermal Resistance $R_{\Theta(J-A)}$	55[1]	22[2]	16[2]	9.5[2]	7.0[2]	5.0[2]	6.5[1]	4.5[1]	3.0	°C/W
Quiescent current cold	10	10	10	20	20	20	30	30	40	mA

NOTES

1. Thermal resistance junction to air for each output transistor.
2. Thermal resistance junction to air for both output transistors.
3. The frequency response can be modified to suit by adjustment of the values of C_3 and C_5

CIRCUITS

Examination of the amplifier circuits figures 2 to 5 will show that there are five basic sections i.e.

(a) The differential input amplifier.
(b) The driver stage.
(c) The complementary driver stage.
(d) The single ended push-pull output stage.
(e) The power supply.

In the case of the 3 watt amplifier the complementary driver stage is in fact the output stage.

The circuits for each stage are relatively conventional and therefore do not require detailed explanation, there are however several factors that are worth mentioning. The input stage has two transistors Q1 and Q2 which are coupled as a differential pair. The difference between the input signal (fed to Q1), and the feedback signal (fed to Q2) provides the drive for the driver stage (Q3).

The frequency response of the amplifier is very wide but can be controlled by varying the values of the capacitors C3 and C4. The low frequency roll off is determined by C3 and R7 e.g. when the reactance of C3 equals the resistance R7 the low frequency output is down by 3dB. The components C4 and R9 have a similar effect on the high frequencies.

The complete specifications of the four basic amplifiers (figures 2 to 5) along with their variations, are given in table 2 below.

10 and 15 watt Amplifiers

As the two most popular amplifiers of the series given in this article would be the 10 and 15 watt designs two complete stereo versions have been constructed. The 15 watt design is pictured on the front cover. In addition to the performance details listed in table 2 for these two amplifiers several other characteristics have been measured and details of these tests are given below.

(a) Frequency and Power Response—Figure 6

This graph indicates the frequency response of the two amplifiers for a constant input signal. Graphs A & B give the response of the amplifiers

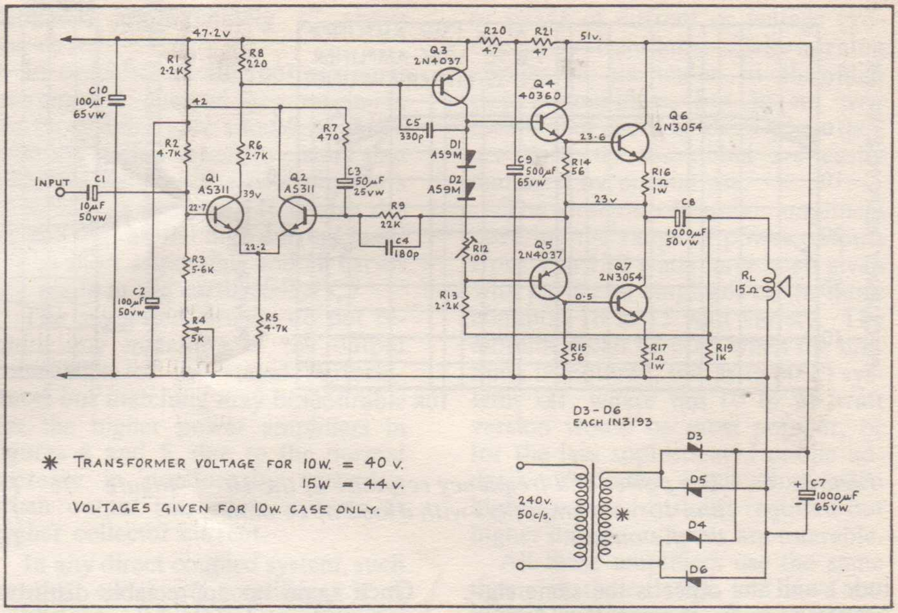

Fig. 3. 10-15 Watt Design

operating at their nominal power outputs whilst curve C represents the response of either amplifiers operating at a level of 100 mW.

Due to the excellent linearity of the amplifiers there is no appreciable difference between the response measured at maximum or very low output powers.

(b) Total Harmonic Distortion—Figure 7

This curve indicates the relation between total harmonic distortion and output power at a frequency of 1 KHz. Both the 10 and 15 watt amplifiers exhibit low distortion between the clipping levels (approximately 12 and 17 watts respectively). The rapid rise in distortion that occurs immediately after the commencement of signal clipping is shown and is due to the rapid onset of signal "flattening" attempts to exceed the supply voltage.

(c) Tone Burst Characteristic—Figure 8

Tone burst tests are designed to demonstrate the amplifiers' ability to reproduce transients. The test signal consists of a number of complete cycles of a tone at a predetermined level followed by a similar number of complete cycles of the tone at a much lower level. The particular part of the cycle of the tone at which the "burst" ceases or is reduced in level can be adjusted precisely by means of the accurate triggering circuitry of the generator.

A photograph of this tone burst signal using a tone of 1000 Hz is given in Figure 8A. The form of this test most stringent for an amplifier is to operate the amplifier so that the maximum level of the tone burst drives it to severe overload.

The output waveform is then watched using a high quality c.r.o. Particular attention should be paid to the waveform during the low level part of the signal.

The important factors in the response of an amplifier to this type of signal are (1) similarity of input and output waveforms and (2) the avoidance of any disturbance such as ringing, undershoot, overshoot or a delayed return to normal level after the completion of the burst.

These characteristics determine the ability of the amplifier to recover from overloads and transients in programme material and contribute largely to the "goodness" or otherwise of the amplifiers' "sound".

In the photograph shown in Figure 8A the c.r.o. has been adjusted to intensify a part of the tone burst input waveform which is three cycles wide immediately before and after the change in level.

Figure 8B is a photograph of two signals superimposed one on top of the other. One is the signal from the tone burst generator shown as the full sine curve (stepped in ampli-

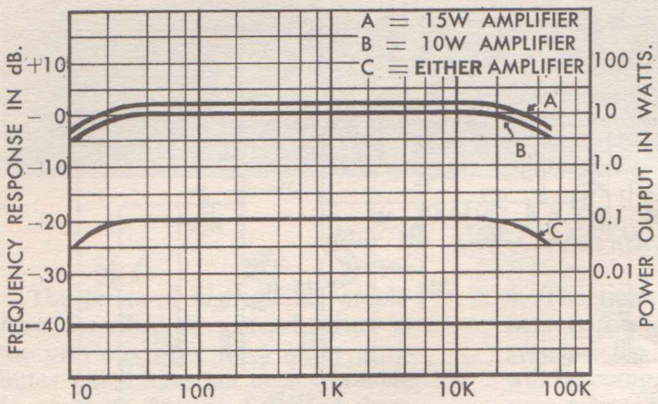

Figure 6. The power and frequency response of the 10 and 15 watt amplifiers with a load of 16 ohm on the output.

FIGURE 7

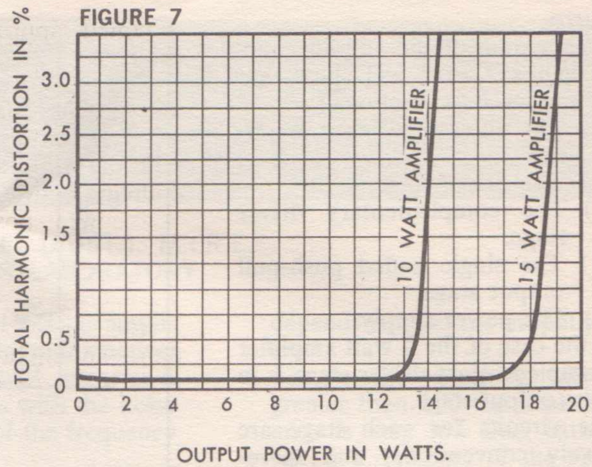

Figure 7. The total harmonic distortion of the 10 and 15 watt amplifiers versus output levels at 1000 hertz.

tude) and the other is the same signal after passing through the 15 watt amplifier at an amplitude sufficient for the high level signal to cause a flat topped overload of approximately 6 dB. For this display the time base of the c.r.o. has been adjusted to expand the section of the burst shown intensified in Figure 8A so that it filled the screen. The accuracy with which the amplified output i.e. up to the square topped overload point, duplicates the input signal can be seen in the photograph. The tilt evident on the horizontal sections of the clipped output is due mainly to the frequency roll off at the low frequency end caused by the coupling capacitor at the output of the amplifier. For the amplifier tested this roll-off produced a drop in response of 3 dB at 13Hz (see Figure 6).

In figure 8C input and output conditions are the same as for 8B but the tone frequency is now 10kHz. Once again no appreciable disturbance at the end of the burst is visible. The shift to the right evident in the output waveform of the amplifier is due to phase shift and the corresponding fall in frequency response at the frequency the tone i.e. 10KHz. The frequency and/or power response for the amplifier shows a drop of approximately 0.7 dB at this frequency (see Figure 6) and a measured phase shift of 10 μS at 10KHz.

*(d) **Separation—Figure 9***

The ratio of the signal appearing in one channel to that in the other when it is operating at full unclipped output power is defined as the separation and is expressed in dB.

Figure 9 shows the curve of separation versus frequency for the 10 watt amplifier. For reference purposes the noise level of the amplifier relative to 10 watts is also shown. The actual measured separation curve showed a number of sharp peaks or frequencies of reduced separation at the mains frequency and its low order harmonics. Since the amplitude of the speakers is always ←50 dB relative to 10 watts they have no detectable affect on the amplifier's performance and have been ignored in preparing the curve.

*(e) **Total Harmonic Distortion—Figure 10***

Figure 10 shows the distortion versus frequency at the 10 watt output level for both the 10 and 15 watt amplifiers. The flat portion of the curve between 50 and 1000 Hz is due mainly to the distortion present in the input signal and also the fact that for levels of distortion <0.05% a degree of inaccuracy exists in the measurement. For example Figure 7 was prepared using a frequency approximately 1000 Hz where a sharp filter was available to reduce the source distortion to approximately 0.03%, is with a resultant 10 watt

Figure 8A *Figure 8B* *Figure 8C*

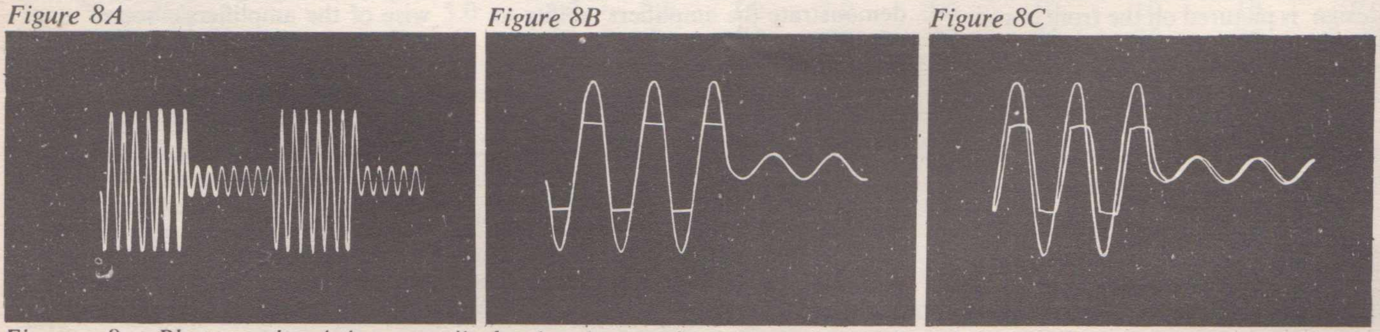

Figure 8. Photographs of the c.r.o. display for the tone burst input signal (A) and the waveform at the output of the 15 watt amplifier for two frequencies 1000 Hz (B) and 10 kHz. (C) Both B and C show the input signal superimposed on the output from the amplifier.

distortion of approximately 0.07%. The dashed part of the curve probably shows a more accurate indication of the true distortion.

Construction

As in all projects of this type neat wiring and component layout is desirable to help stability and reduce the possibility of wiring errors which due to the direct coupled circuitry of these amplifiers may be difficult to locate.

The output transistors of the 3 watt amplifier should be connected to a heat sink consisting of a piece of 16 gauge aluminium about 3 square inches in size. The heat sink for the 10 to 15 watt amplifier could be the aluminium chassis, as shown in the cover photo. For the higher powered amplifiers separate extruded aluminium heat sinks are required. In most instances insulation of the output transistors from the heat sinks would be necessary.

To help cooling and to keep induced currents to a minimum the power transformers should be mounted approximately ⅛" above the chassis.

In order to obtain the maximum amount of channel separation in stereo systems care should be taken with the wiring layout to ensure that coupling between the two amplifiers is as low as possible. It is also desirable that all the high current leads are as short as possible and all earths are made at the earth end of C7.

The suggested circuits do not require any matching of the output transistors for the stated performance, but matching may be desirable for the higher power amplifiers in figures 4 and 5, due to the normal increase in nonlinearity occurring when output transistors operate at higher collector current.

In any direct coupled system, such as this series of amplifiers, it is extremely dangerous to remove, replace, shunt or otherwise modify the circuit while it is operating. Any fault conditions produced under such operation could destroy the semiconductors by overheating or peaks of current or voltage produced by transients. This warning applies in particular to the high power amplifiers, but in no way means that the amplifiers when they are correctly assembled are easily damaged by normal use.

The designs of audio amplifiers covering the range of power outputs from 3 to 130 watts have been given with particular emphasis on the more common 10 to 15 watt variety. The amplifiers can be used when the ultimate is required, that is in Hi Fi systems etc. where the 10 to 15 watt version would be most popular, or for the less sophisticated public address or similar application when higher power is usually required but higher distortion levels are tolerable.

All these amplifiers use the same theme in their design and have excellent performance characteristics. **It is proposed to develop circuits for low level preamplifiers, tape, microphone and pickup systems using silicon transistors and these will be published in a future issue.**

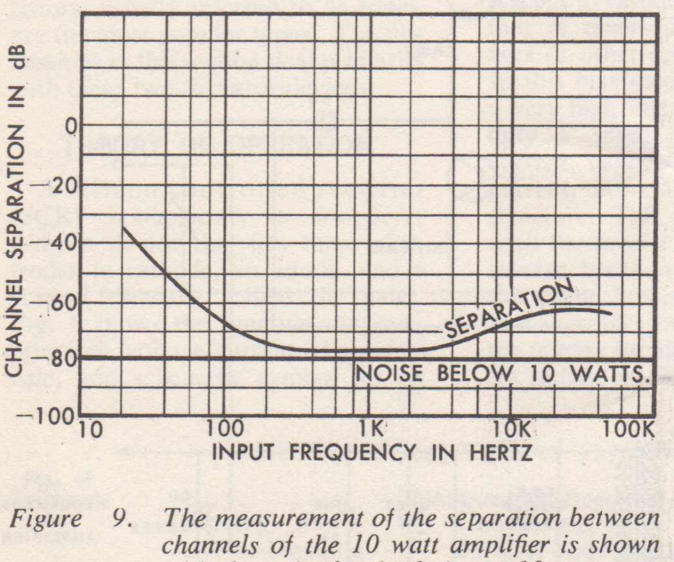

Figure 9. The measurement of the separation between channels of the 10 watt amplifier is shown with the noise level relative to 10 watt.

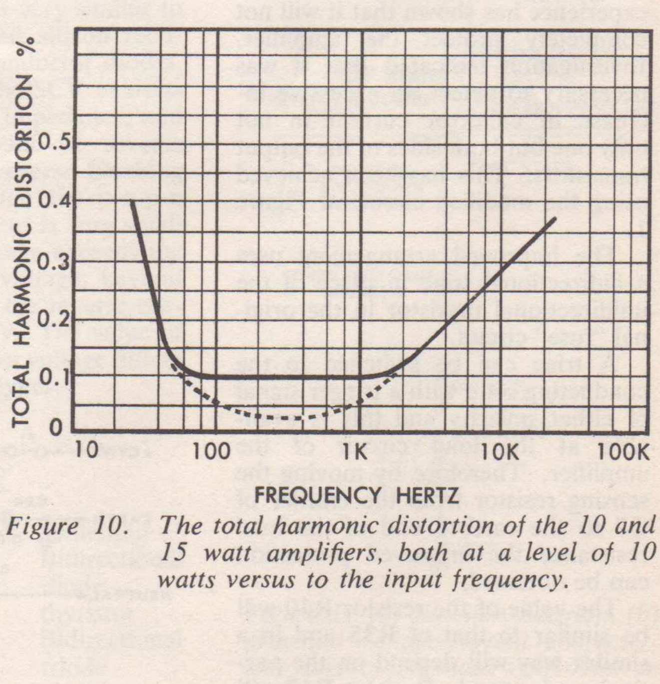

Figure 10. The total harmonic distortion of the 10 and 15 watt amplifiers, both at a level of 10 watts versus to the input frequency.

The Protection of Transistor Power Amplifiers against a Short Circuited Load Condition.

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Circuits shown in our high power amplifiers (see Radiotronics May, 1967—Figures 4 and 5) for the short circuit protection of the output transistors are of the electronic fuse type. They incorporate circuitry which, when the collector current is higher than a preset maximum, e.g., as occurs with a low value of output load impedance, rapidly will disconnect the power supply from the output stage.

In the circuits listed the current which detects a reduced load resistance is that flowing in only **one** of the output transistors, i.e., for one half cycle of the output wave form. Although this circuit can be set to operate close to and lower than the safe maximum for the transistors used and after this condition is reached will operate in approximately 1 μS, experience has shown that it will not completely protect the amplifier. Investigation indicated that it was necessary to detect an excessive increase in collector current in not only one but both sides of the output transistors. This has been achieved using the modified circuit of Figure 1.

The improved arrangement uses a bidirectional triac in place of the unidirectional thyristor in the original "fuse" circuit.

A triac can be switched to the conducting state with a trigger signal of either polarity and this is available at the load circuit of the amplifier. Therefore by moving the sensing resistor from the emitter of Q7 to the earthed end of the load resistance the improved protection can be obtained.

The value of the resistor R40 will be similar to that of R35 and in a similar way will depend on the particular triac used. Resistor R17 will now replace R35 and have a value equal to R16 which is its counterpart in the other side of the amplifier.

The new circuit has several advantages over the older design as indicated below:—

(1) Full protection for both output transistors is provided.
(2) Protection can also be obtained against spurious signals appearing across the output load.
(3) The protection circuit is not part of the main amplifier and consequently can be adapted to any design.
(4) The polarity of the supply rail is unimportant since the triac will operate on a voltage of either polarity.

During tests on these protection circuits it was found that the transients produced by the power switch or others that occur occasionally on any mains supply could operate the triac and open the fuse. The effects of these transients can be avoided by connecting a filter capacitor of 0.1 μF (600 volt) across the transformer primary and 1.0 μF (200 volt) across the secondary as shown in Figure

The modified circuit of Figure 1 has been used in a large number of tests in the higher powered amplifiers and has been completely satisfactory. It will provide protection less than 2 μS. after the development of a fault.

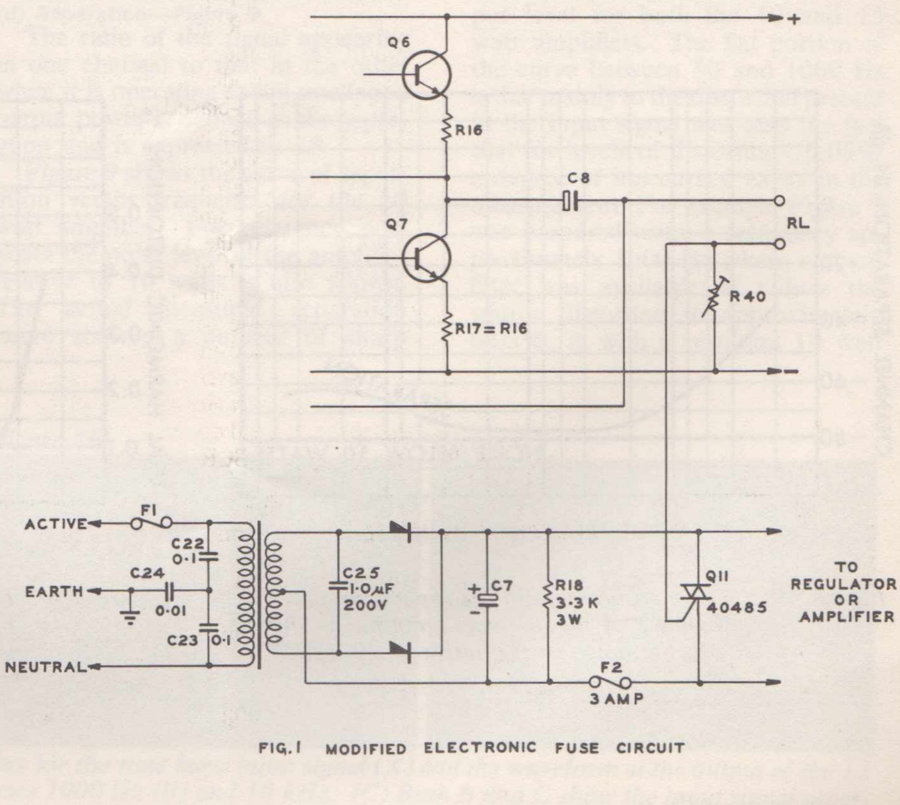

FIG. 1 MODIFIED ELECTRONIC FUSE CIRCUIT

THEORY AND APPLICATION OF THYRISTORS (PART 1)

The term **thyristor** is the generic name for semiconductor devices that have characteristics similar to those of thyratron tubes. Basically, this group includes bistable semiconductor devices that have three or more junctions (i.e., four or more semiconductor layers) and that can be switched between conducting states (from OFF to ON or from ON to OFF) within at least one quadrant of the principal voltage-current characteristic. There are several different types of thyristors, which differ primarily in the number of electrode terminals and in their operating characteristics in the third quadrant of the voltage-current characteristic, as shown in Table 1. Reverse-blocking triode thyristors, commonly called **silicon controlled rectifiers (SCR's),** and bidirectional triode thyristors, usually referred to as **triacs,** are the most popular types. The discussions in this section deal primarily with these two thyristor devices.

THEORY OF OPERATION

A silicon controlled rectifier (SCR) is basically a four-layer p-n-p-n device that has three electrodes (a cathode, an anode, and a control electrode called the gate). Fig. 1 shows the junction diagram, principal voltage-current characteristic, and schematic symbol for an SCR. A triac also has three electrodes (main terminal No. 1, main terminal No. 2, and gate) and may be considered as two parallel p-n-p-n structures oriented in opposite directions to provide symmetrical bidirectional electrical characteristics. Fig. 2 shows the junction diagram, voltage-current characteristic, and schematic symbol for a triac. An analysis of the voltage-current characteristics of SCR's and triacs and of the charge-carrier interactions that make possible the switching transitions indicated by these characteristics provides useful information concerning the operation and possible applications of these devices.

Voltage-Current Characteristics

As shown in Fig. 1 (b), the operation of an SCR under reverse-bias conditions (anode negative with respect to cathode) is very similar to that of reverse-biased silicon rectifiers or other semiconductor diodes. In this bias mode, the SCR exhibits a very high internal impedance, and only a slight amount of reverse current, called the **reverse blocking current,** flows through the p-n-p-n structure. This current is very small until the reverse voltage exceeds the reverse breakdown voltage; beyond this point, however, the reverse current increases rapidly. The value of the reverse breakdown voltage differs for individual SCR types.

During forward-bias operation (anode positive with respect to cathode), the p-n-p-n structure of the SCR is electrically bistable and may exhibit either a very high impedance (forward-blocking or OFF state) or a very low impedance (forward-conducting or ON state). In the forward-blocking state, a small forward current, called the forward OFF-state current, flows through the SCR. The magnitude of this current is approximately the same as that of

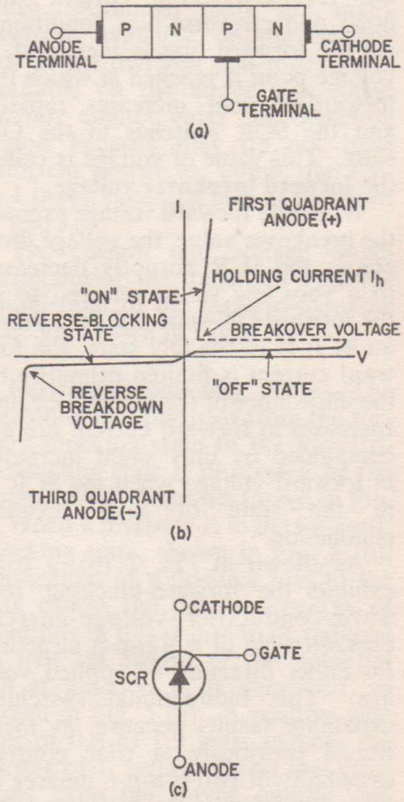

Figure 1. (a) Junction diagram, (b) principal voltage-current characteristic, and (c) schematic symbol for an SCR thyristor.

Table 1—Different Types of Thyristors

No. of Terminals	Third-Quadrant Operation		
	Blocking	Conducting	Switching
2	Reverse-blocking diode thyristor	Reverse-conducting diode thyristor	Bidirectional diode thyristor
3	Reverse-blocking triode thyristor	Reverse-conducting triode thyristor	Bidirectional triode thyristor

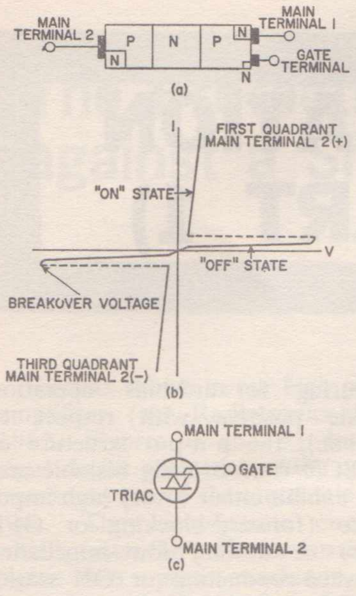

Figure 2. (a) Junction diagram, (b) principal voltage-current characteristic, and (c) schematic symbol for a triac thyristor.

the reverse-blocking current that flows under reverse-bias conditions. As the forward bias is increased, a voltage point is reached at which the forward current increases rapidly, and the SCR switches to the ON state. This value of voltage is called the **forward breakover voltage.**

When the forward voltage exceeds the breakover value, the voltage drop across the SCR abruptly decreases to a very low value, referred to as the forward ON-state voltage. When an SCR is in the ON state, the forward current is limited primarily by the impedance of the external circuit. Increases in forward current are accompanied by only slight increases in forward voltage when the SCR is in the state of high forward conduction.

As shown in Fig. 2 (b), a triac exhibits the forward-blocking, forward-conducting voltage-current characteristic of a p-n-p-n structure for either direction of applied voltage. This bidirectional switching capability results because, as mentioned previously, a triac consists essentially of two p-n-p-n devices of opposite orientation built into the same crystal. The device, therefore, operates basically as two SCR's connected in parallel, but with the anode and cathode of one SCR connected to the cathode and anode, respectively, of the other SCR. As a result, the operating characteristics of the triac in the first and third quadrants of the voltage-current characteristics are the same, except for the direction of current flow and applied voltage. The triac characteristics in these quadrants are essentially identical to those of an SCR operated in the first quadrant. For the triac, however, the high-impedance state in the third quadrant is referred to as the OFF state rather than as the reverse-blocking state. Because of the symmetrical construction of the triac, the terms forward and reverse are not used in reference to this device.

Thyristors are ideal for switching applications. When the working voltage of a thyristor is below the breakover point, the current through the device is extremely small and the thyristor is effectively an open switch. When the voltage across the main terminals increases to a value exceeding the breakover point, the thyristor switches to its high-conduction state and is effectively a closed switch. The thyristor remains in the ON state until the current through the main terminals drops below a value which is called the **holding current.** When the source voltage of the main-terminal circuit cannot support a current equal to the holding current, the thyristor reverts back to the high-impedance OFF state.

The breakover voltage of a thyristor can be varied, or controlled by injection of a signal at the gate, as indicated by the family of curves shown in Fig. 3. Although this family of curves is shown in the first quadrant typical of SCR operation, a similar set of curves can also be drawn for the third quadrant to represent triac operation. When the gate current is zero, the principal voltage must reach the breakover value $V_{(BO)}$ of the device before breakover occurs. As the gate current is increased, however, the value of breakover voltage becomes less until the curve closely resembles that of a rectifier. In normal operation, thyristors are operated with critical values well below the breakover voltage and are made to switch ON by gate signals of sufficient amplitude to assure that the device is switched to the ON state at the instant desired.

After the thyristor is triggered by the gate signal, the current through the device is independent of gate voltage or gate current. The thyristor remains in the ON state until the principal current is reduced to a level below that required to sustain conduction.

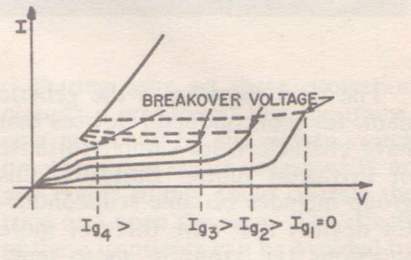

Figure 3. Curves showing breakover characteristics of a thyristor for different values of gate current.

CHARGE-CARRIER INTERACTIONS

The electron-hole interactions that make possible the switching transitions in p-n-p-n semiconductor structures are represented graphically by the potential-energy diagrams shown in Figs. 4, 5 and 6. These diagrams show the potential energies of holes and electrons as a function of distance through the crystal. The upward direction indicates increasing levels of electron energy, and the downward direction indicates increasing levels of hole energy. The dots in the diagrams represent free electrons, and the circles represent free holes.

The electrons in a solid can occupy only specific energy levels or electron states. Each existing state can be occupied by only one electron. The **Fermi energy level E_F** is the dividing line above which most of the existing electron states are empty and below which most states are full. Conduction in a solid occurs only by movement of free charge carriers, i.e., free electrons or free holes. A free electron is an electron which is at an energy level for which most of the existing states are empty, and a free hole is an empty state at an energy level for which most of the existing states are filled. Free electrons exist therefore, only at energy levels above E_F, and free holes exist only at levels below E_F.

Because electrons and holes tend to seek the lowest available energy levels, they both move toward E_F, which is the zero-energy level for both types of charge carriers. On the potential-hill diagrams, electrons always tend to "fall", and holes always tend to "rise". If the charge carriers were not affected by outside influences, therefore, all free electrons and holes would eventually reach the Fermi level and disappear. A distribution of free electrons above E_F, however, is maintained by thermal energy of the lattice which constantly agitates the electrons to non-zero energy levels.

In the metal contact regions, there is a continuous distribution of electron states about the Fermi energy level so that free holes and free electrons exist simultaneously side by side. In the semiconductor regions, there is a band of energy, called the **forbidden region,** in which no electron states exist. As a free carrier tries to move through the system of metal to semiconductor to metal, it finds that it can move freely through the metal, but when it reaches the semiconductor it encounters an obstacle, the forbidden-energy region, which it must go over or under depending upon whether it is a free electron or a free hole. The carrier must obtain sufficient energy so that it is displaced far enough from the Fermi level to go over or under the forbidden-energy region. If sufficient energy, such as thermal agitation or an applied voltage, is not available, the carrier is reflected back to its origin.

In silicon crystal, the forbidden region is wide enough so that, at ordinary temperatures, there is not sufficient thermal energy available to distribute carriers both above and below the band. If the Fermi energy level is close to the top of the band, thermal energy is sufficient to lift electrons into states on top of the forbidden region, but is not sufficient to push holes into states below this region. As a result, the material contains many free electrons, but very few free holes, and is referred to as an n-type semiconductor because it contains mostly negative-charge carriers.

Similarly, a p-type region in the semiconductor, which contains mostly positive-charge carriers, results when the Fermi level is close to the bottom of the forbidden band. For this condition, thermal energy is sufficient to excite holes into states below the band, but is not sufficient to excite electrons into states above the band.

The position of the Fermi level in the forbidden region is determined by the carrier concentration. This concentration, in turn, is determined by both the dopant concentration and the concentration of injected carriers.

At the metal-to-semiconductor interfaces, the dopant concentration is very high. At such interfaces, the carrier concentration cannot be changed significantly by injected carriers, and the position of the Fermi level in the forbidden region is firmly fixed. In the semiconductor regions and near the junctions, the dopant concentration is relatively low so that the total carrier concentration and, therefore, the position of the Fermi level in the forbidden region can be changed by injection of carriers from surrounding regions. These factors make the forbidden-energy region appear flexible within the body of the semiconductor but rigid at the metal-to-semiconductor contacts. This rigidity of the potential hill at the contacts prevents electrons in the metal from entering the p-type semiconductor, but allows holes to circulate freely between the metal and the p-type region; similarly, electrons can circulate freely between metal and the n-type region but holes cannot cross the metal-to-n-type semiconductor interface.

Forward-Blocking State — The sequence of diagrams in Fig. 4 illustrates the transition of the thyristor from the equilibrium (zero-bias) condition to the forward-blocking state. In the equilibrium condition, the concentration of charge carriers (electrons and holes) is determined primarily by dopant concentrations. For this condition, which is represented by the potential-hill diagram shown in Fig. 4 (b), there is approximately one free carrier for each dopant atom.

When the cathode side of the thyristor is biased negatively with respect to the anode side, the potential energy of the electrons is increased in the cathode region and that of the holes is increased in the anode region. Because of the difference in energy level from cathode to anode, the shape of the forbidden-energy region is altered in the most lightly doped section (i.e., the n-type base) so that the height of the potential hill of the central junction is increased. As shown in Fig. 4 (c), any electrons that exist in this region "fall down" the resultant hill, and any holes in this region "rise" to the top of the hill. In this way, all free charge carriers are removed, and the hill becomes a depletion region, as shown in Fig. 4 (d).

The movement of charge carriers with an increase in the forward voltage results in a charging, or displacement, current similar to the current ($i = Cdv/dt$) that charges a capacitor. This displacement current ceases when the forward voltage reaches a steady value because there are no additional carriers for the field to move. Although there are many electrons available on the cathode side of the thyristor and many holes available on the anode side, these carriers cannot enter the depletion region because they do not have sufficient energy to "climb" the 0.8- to 1.0-volt potential hills at junctions J_1 and J_3.

The current and voltage waveforms during the transition from the equilibrium to the forward-blocking state are shown in Fig. 4 (e).

Forward-Conducting State—The transition in a thyristor from the forward-blocking state to the forward-conducting state is illustrated by the potential-hill diagrams shown in Fig. 5. When a thyristor is in the forward-blocking state, shown in Fig. 5 (b), application of a positive bias to the gate causes the potential energy of electrons in this region to be reduced so that the height of the potential hill at junction J_1 is decreased, as shown in Fig. 5 (c). A positive gate bias of 0.8 to 1.9 volts reduces the barrier of J_1 sufficiently so that electrons from the n-type emitter can move across the p-type base into the depletion region. The electric field then sweeps them across this region, as indicated in Fig 5 (c).

Electrons accumulate in the "well" at the bottom of the depletion region until their combined negative charge increases the potential electron energy sufficiently to cause the potential hill at junction J_3 to disappear. Holes can then move from

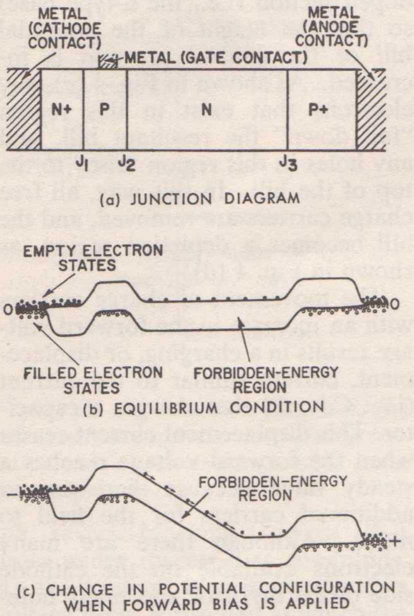

Figure 4. Potential-hill diagrams for various stages of thyristor transition from equilibrium condition to forward-blocking condition (electron energy increases upward, hole energy increases downward).

the p-type emitter across the n-type base into the depletion region. These holes then immediately "climb" the potential hill at J_2, as shown in Fig. 5 (d).

The increased supply of holes to the p-type base further depresses the potential hill at J_1 so that the n-type emitter can inject an even greater number of electrons into the depletion layer. This action, in turn, increases the injection of holes from the p-type emitter. As a result of these regenerative effects, the current through the thyristor increases rapidly, and the depletion region collapses to complete the transition to the forward-conducting state. Fig. 5 (e) illustrates this condition. In this state, the concentrations of both holes and electrons are greatly increased over the equilibrium concentrations. The thyristor can be sustained in the forward-conducting state by an anode-to-cathode forward-voltage drop of approximately 1 volt, and the thyristor current is limited only by the external circuit.

The current and voltage waveforms during the transition from the forward-blocking to the forward-conducting state are shown in Fig. 5 (f).

Turn-Off—The transition in the thyristor from the forward-conducting state back to the forward-blocking state is illustrated in Fig. 6. This transition is accomplished either by momentary reduction of the anode current to zero, or by momentary reversal of the anode-to-cathode voltage.

In the conducting state, carrier concentrations far in excess of the equilibrium level are injected into the n- and p-type regions. These excess carriers remain for a finite time after the anode current is reduced to zero. If the forward bias is re-applied before these excess carriers are removed, the device simply returns to the conducting state and does not switch to the blocking condition. After the excess carriers are removed and the device is returned to equilibrium, the potential hills rebuild, and the device can return to the forward-blocking state, as shown in Fig. 4.

The removal of the excess carriers can be accomplished if the anode current is reduced to zero until the excess carriers recombine or move out of the depletion region. This removal corresponds to a direct transition from the conditions shown in Fig. 6 (c) to those shown in Fig. 6 (g). The potential hill at junction J_1 rebuilds first because it is in the more heavily doped region of the device,

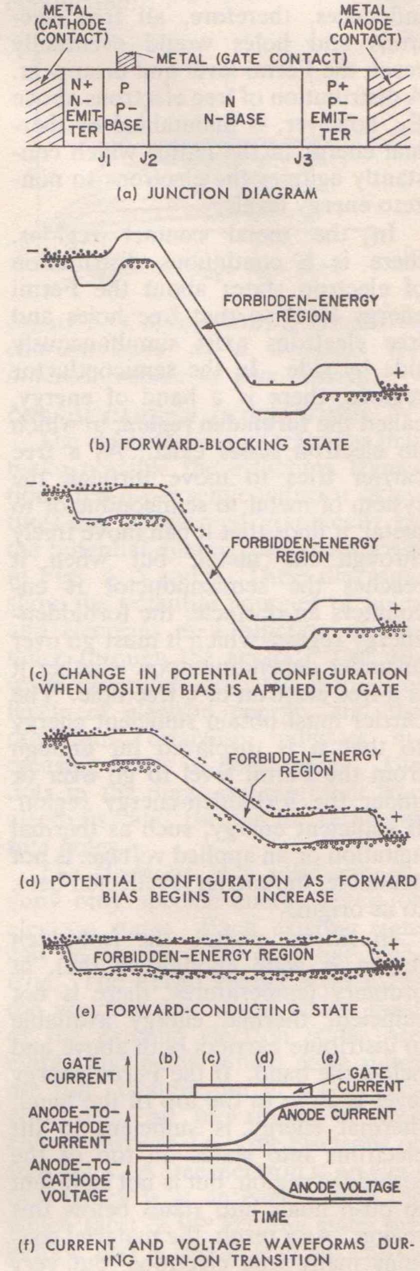

Figure 5. Potential-hill diagrams for various stages of thyristor transition from forward-blocking state to forward-conducting state.

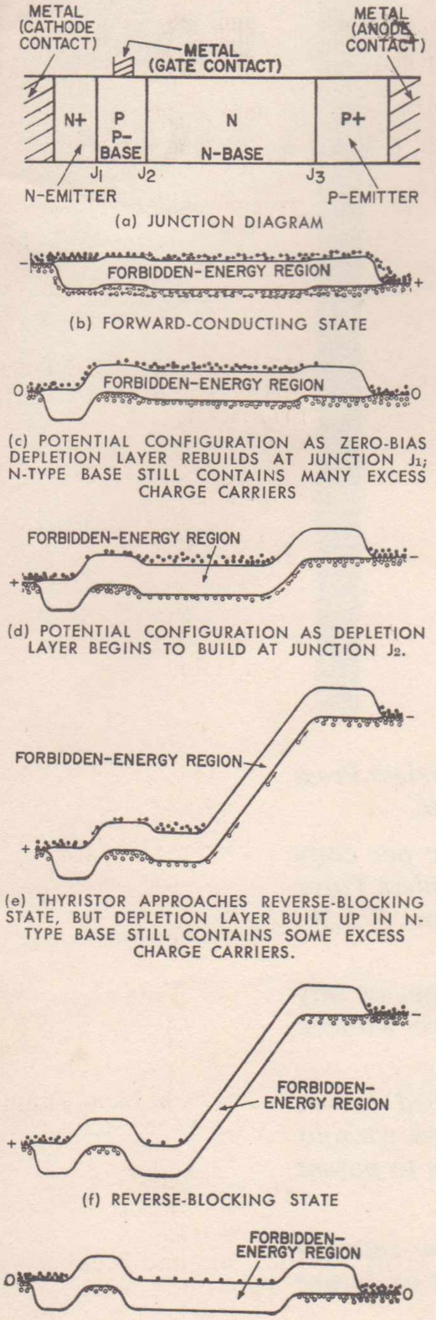

Figure 6. Potential-hill diagrams for various stages of thyristor transition from forward-conducting state to turn-off.

but the hills at J_2 and J_3 also rebuild as the excess carriers disappear during the zero-anode-current condition.

A more rapid removal of the excess carriers can be accomplished by a momentary reversal of the anode-to-cathode voltage. This transition is shown in Figs. 6 (d) through 6 (f). As the reverse voltage increases, carriers are pulled out of the device in the direction opposite to that in which they were injected so that a substantial reverse current results.

The removal of carriers is aided as a potential hill and a depletion region begin to build at junction J_3, as shown in Fig. 6 (d). As the remaining quantity of excess carriers is reduced, the reverse current decreases, and reverse voltage builds up. At the stage shown in Fig. 6 (e), the reverse depletion region has built up, but the undepleted n-type base region still contains some excess carriers which prevent the potential hill at J_2 from rebuilding, and which continue to flow out as reverse current. At the stage shown in Fig. 6 (f), the excess carriers have all been removed, and device has reached its steady-state reverse-blocking condition. In Fig. 6 (g), the reverse bias has been removed, all regions return to the equilibrium zero-bias carrier concentrations, and the device is ready for return to the forward-blocking condition.

Current and voltage waveforms corresponding to the various conditions described in Figs. 4, 5 and 6 are shown in Fig. 7.

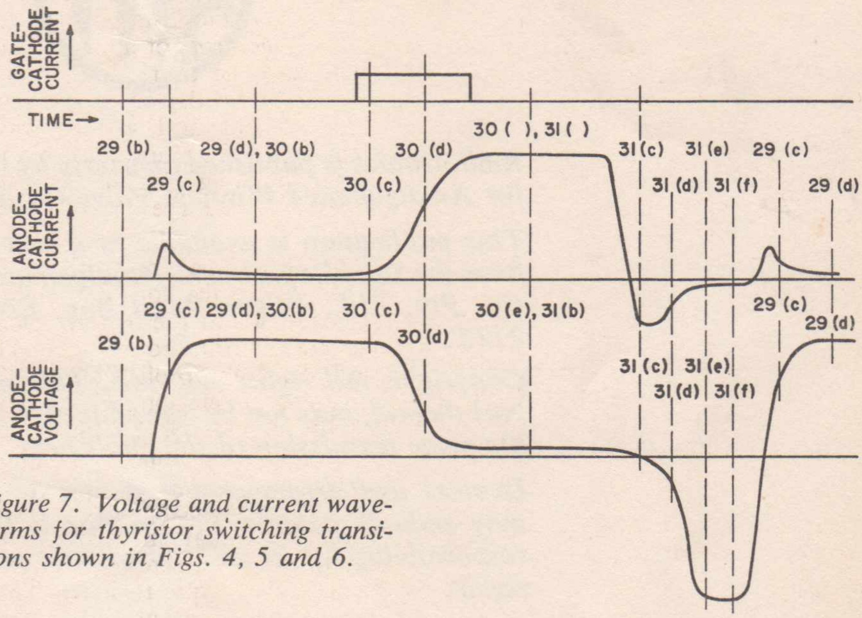

Figure 7. Voltage and current waveforms for thyristor switching transitions shown in Figs. 4, 5 and 6.

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