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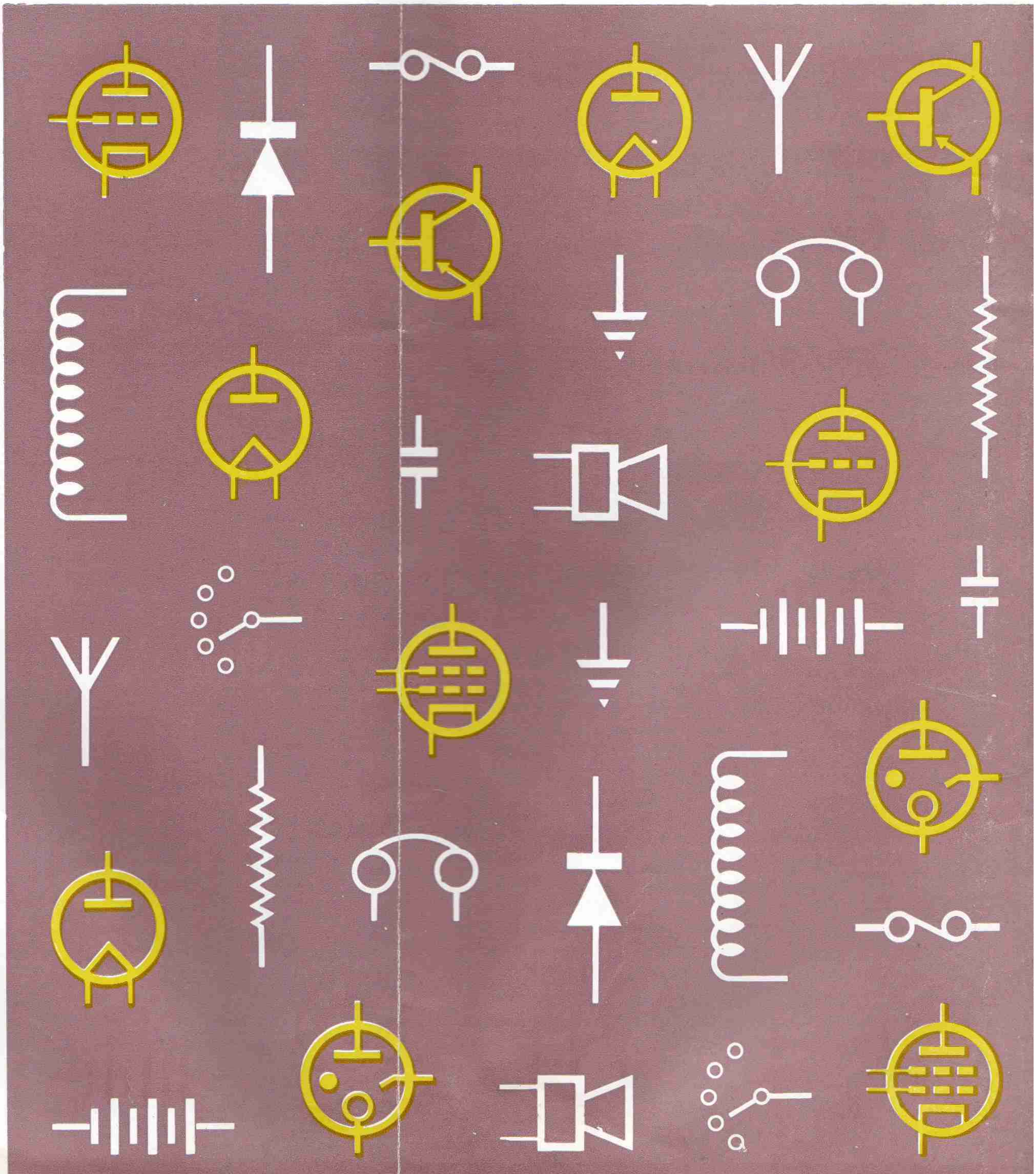

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RADIOTRONICS

Vol. 26, No. 4, 1961

Editor, Bernard J. Simpson

IN THIS ISSUE

USING 7558 AND 7551 58

This article discusses the use of the 7558 and 7551 beam power valves in class C rf amplifier, frequency-multiplier and af modulator applications.

THE IMPORTANCE OF SHIELDING 61

SILICON POWER TRANSISTORS - AN APPLICATION GUIDE 62

This is the third and final part of this interesting and informative article from RCA. This part discusses actual circuit applications of silicon power transistors.

EEV VIDICONS AT WORK 71

VAPOUR-COOLED VALVES FOR INDUSTRIAL HEATERS 71

20-AMPERE SILICON RECTIFIERS 72

Announcing fourteen new stud-mounted silicon power rectifiers with an average forward current rating of 20 amperes and PIV ratings ranging from 55 to 600 volts.

NEW RELEASES 76

New releases mentioned this month are 2054, water-cooled super power triode; 4017, 4019, 4020, intermediate-noise-factor TW tubes; 4604, quick-heating beam power valve; 6939, miniature twin power pentode; 7842, 7843, 7844, small conduction cooled beam power valves.

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USING 7558 AND 7551

VHF BEAM POWER VALVES

IN RF AMPLIFIER, FREQUENCY-MULTIPLIER AND AF MODULATOR SERVICE

This article discusses the application of the 7558 and 7551 beam power valves in class C rf-amplifier and frequency-multiplier service at frequencies up to 175 megacycles and in af modulator and power-amplifier service. Circuits illustrating the use of these valves in a "straight-through" amplifier and in various types of frequency multipliers are described.

The 7558 and 7551 are beam power valves of the nine-pin miniature type. Although primarily designed for vhf service, both valves also provide excellent performance in audio-frequency modulator and power-amplifier applications. In typical operation at 175 megacycles, both valves can deliver useful power output of 8.5 watts under CCS conditions and 10 watts under ICAS conditions. In af modulator service, two valves in push-pull can provide 20.5 watts of audio power with less than 5 per cent total harmonic distortion.

Valve Design Features

The 7558 and 7551 have identical electrical characteristics except for heater-voltage requirements. The 7558 has a 6.3-volt heater and is designed primarily for continuous service in fixed-station communications equipment in which the normal heater-voltage variation does not exceed ±10 per cent. The 7551 has a 12-to-15-volt heater and provides reliable service in mobile communications equipment operating from a 6-cell storage battery system.

Although the heater of the 7551 is designed to operate over a voltage range from 12 to 15 volts, it will withstand momentary excursions from 11 to 17 volts. During manufacture the 7551 is subjected to rigid controls and tests for heater-cathode leakage, interelectrode leakage, low-frequency-vibration performance, and intermittent shorts, as well as a 500-hour intermittent life test at a heater voltage of 15 volts and a heater-cycling life test at a heater voltage of 17 volts. As a result, the rugged heater design and the careful testing ensure dependable performance in mobile equipment under operating conditions encountered during battery charging and discharging.

Features which contribute to the efficient performance of both these tubes at 175 megacycles are low lead inductances, low interelectrode capacitances, and low rf losses. Low lead inductances for both cathode and grid No. 2 (screen grid) are achieved by the use of two pin connections for each of these electrodes. If the two pin connections for the cathode are arranged so that the rf input current flows through one and the rf output current through the other, degeneration in the cathode circuit is reduced. The two base-pin connections for the grid No. 2 facilitate rf by-passing and aid in cooling. The low rf losses and high input resistance permit the use of relatively high values of grid-No. 1-circuit resistance

(100,000 ohms) to minimize loading of the driver stage. In addition, high grid-No. 1-circuit resistance makes it possible to obtain the high value of grid bias required for good frequency-multiplier plate-circuit efficiency with low values of dc grid current.

The maximum permissible temperature of the bulb is 225 degrees centigrade. Within reasonable ambient temperatures, no artificial means of cooling are required for most applications. However, a heat-dissipating shield may be used as an aid in reducing bulb temperature when necessary.

RF-Power-Amplifier Service

Fig. 1 shows a circuit in which the 7551 or 7558 is used as a "straight-through" amplifier at 175 megacycles. In this single-ended amplifier, C_1 and C_8 effectively simulate the input and output capacitances, respectively, of the other section of a push-pull circuit. Thus, the valve input and output capacitances are in series with C_1 and C_8. This arrangement optimizes performance at higher frequencies because it provides a higher L/C ratio in the resonant circuits than would be possible with a conventional parallel-tuned tank circuit.

At vhf, the rf grounding of the heater circuit is important because of its effect on the stability of the amplifier. The most satisfactory method of grounding the heater circuit is achieved when one of the heater pins is connected directly to the chassis at the socket and the other is bypassed to the chassis through a low-inductance capacitor.

As mentioned previously, degeneration in the cathode circuit is minimized when the two pin connections, 1 and 9, are used to separate the input and output rf currents. These pins should be grounded by the shortest possible connections to the chassis to reduce the external cathode-lead inductance.

If a socket shield is required to prevent rf feedback from the plate circuit to the grid-No. 1 circuit (usually at frequencies above 100 megacycles), the shield should be placed across the socket in the plane of pins 4 and 9 so that the heater (pin 4) and cathode (pin 9) can utilize the shield as a low-inductance ground connection.

The circuit shown in Fig. 1 was found to be very stable when tested for oscillation at reduced plate and screen-grid voltages without fixed bias. With the plate circuit unloaded and no grid drive, the amplifier did not oscillate with any combination of settings of the plate and grid-No. 1-tank capacitors.

When more rf power is required than a single 7558 or 7551 can deliver, two valves may be used. At frequencies below 60 megacycles, the valves may be connected in parallel; at higher frequencies (up to 175 megacycles), a push-pull circuit is recommended to reduce the effective input and output capacitances.

Grid-Driving Power

The grid-No. 1-driving power given in the published data for the valve is the power which must be delivered by the previous stage. This

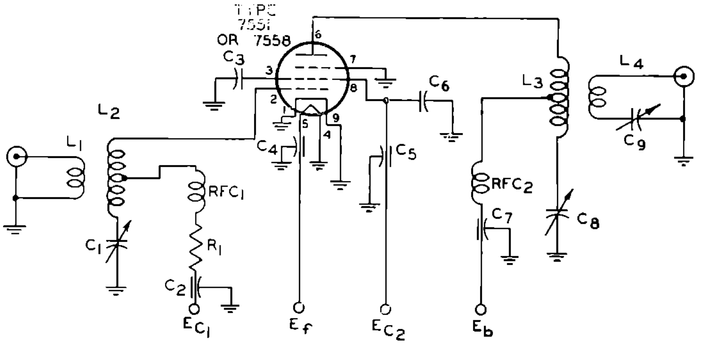

Fig. 1—Basic circuit for 175 Mc amplifier, 87.5/175-Mc doubler, and 58.5/175-Mc tripler. C_1, C_9: 7-45 pf, ceramic disc. (C_1 for doubler: 4-30 pf ceramic disc). C_2, C_4, C_5, C_7: 1000 pf, feed-through, silver button mica. C_3, C_6: 1000 pf, bypass, silver button mica. C_8: 3.6-15 pf, midget, double spaced. L_1: 2 turns of No. 18 Enam. wound on $\frac{1}{2}''$ diameter form. L_2: No. 18 Enam. wound on $\frac{1}{2}''$ diameter form: 5 turns centre tapped for amplifier; 7 turns centre tapped for doubler; 8 turns centre tapped for tripler. L_3: 4 turns centre tapped No. 18 Enam. wound on $\frac{1}{2}''$ diameter form. L_4: 3 turns No. 18 Enam. wound on $\frac{1}{2}''$ diameter form. R_1: 22,000 ohms, 0.5 watt for amplifier; 47,000 ohms, 0.5 watt for doubler, 68,000 ohms, 0.5 watt for tripler. RFC_1, RFC_2: 1.8 μh., 1000 ma., 80-200 Mc. (RFC_1 for doubler and tripler: 7.0 μh., 1000 ma., 35-110 Mc.)

value is the actual total power input required by the grid-No. 1 circuit, and includes the power delivered to the grid No. 1, the power dissipated in the grid-bias resistor, the losses in the valve caused by transit-time loading, and the rf losses in the valve, tank circuit, socket, and wiring. The driver stage should be designed to provide reserve power to allow for variations in line voltage, components, and valve characteristics.

If the valve is to be operated under conditions differing from the suggested typical operating conditions, the performance can be checked as follows: First, load the amplifier to the desired value of plate current. Then vary the grid-driving power slowly (tank-circuit tuning remains unchanged) and note the change in output. If the change in output is approximately proportional to the change in grid-No. 1 drive, the stage is underdriven. The drive should be increased until very little increase in power output results from a large increase in drive. However, the maximum rated value of dc grid-No. 1 current should not be exceeded.

If the stage is underdriven, low power output and low efficiency will result. In plate-modulated service, underdriving causes severe distortion at high levels of modulation. This condition is readily recognizable as downward modulation; it is evidenced by a decrease in the average plate current as the modulation level is increased.

Because overdriving the 7558 or 7551 may cause the grid-No. 2 input rating to be exceeded before the maximum control-grid rating is reached, it is recommended that the grid-No. 2 current be metered to determine whether the power input is within ratings.

The grid-No. 1 bias voltage may be obtained from a grid-No. 1 resistor or from a combination of grid-No. 1 resistor with either a cathode resistor or fixed supply. The use of a bias supply or a cathode resistor will provide protective bias in the event of loss of excitation.

Plate-Modulated RF Power-Amplifier Service

For high-level modulation in plate-modulated telephony service, both the plate and grid No. 2 must be modulated. The rf impedance between grid No. 2 and cathode must be kept low by means of a suitable bypass capacitor. The value of bypass capacitor used should be kept as small as possible to avoid excessive af bypassing. If the capacitance value is too small, however, feedback may occur between the plate and grid No. 1, with subsequent instability. If the grid-No. 2 bypass capacitor is replaced by a series-resonant circuit tuned to resonate at the operating frequency, the rf impedance between grid No. 2 and ground is low at this frequency and high at audio frequencies.

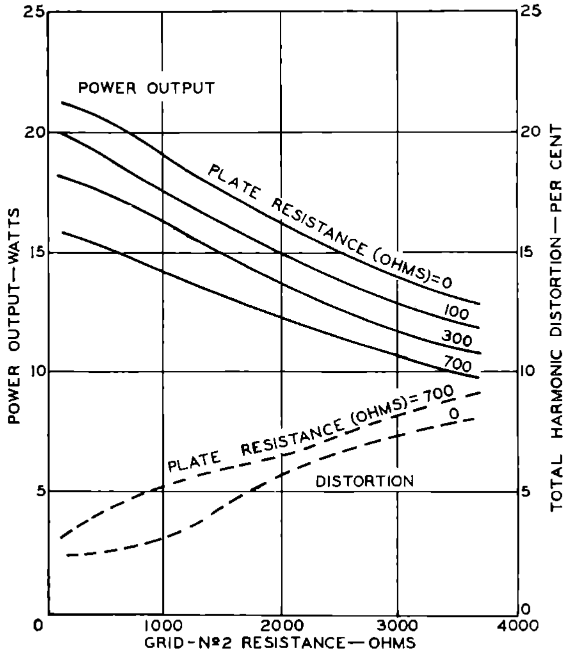

Fig. 2—Power output and harmonic distortion as a function of internal power-supply resistance.

For linear modulation characteristics, the use of a combination of grid resistor and fixed supply in the grid-No. 1 circuit has the advantage of providing the desired variations in bias voltage as the modulating voltage varies.

Frequency-Multiplier Service

Doubler Operation: With the input modifications given in the parts list, the circuit shown in Fig. 1 may be used for frequency-doubler service to provide a power output of three watts CCS and 4.5 watts ICAS at an output frequency of 175 megacycles. A pair of valves in "push-push" may be used to obtain greater power output in doubler service. In this case, the plates are connected in parallel and the inputs in push-pull. With this arrangement, the even-order harmonics are accentuated and the odd-order harmonics suppressed.

Tripler Operation: The input of the basic circuit shown in Fig. 1 can also be modified as shown in the parts list to permit frequency-tripler operation. This circuit provides a power output of 1.4 watts CCS and 2.3 watts ICAS at an output frequency of 175 megacycles. For high-order multiplication, a single valve used as a frequency tripler can provide more than adequate driving power for the push-push frequency doubler or straight-through amplifier. For additional power output, a pair of valves may be used in a push-pull arrangement. This circuit accentuates the odd-order harmonics and suppresses even-order harmonics, making it suitable as a frequency tripler. No socket shield is necessary for either doubler or tripler operation.

AF Power Amplifier and Modulator—Class AB Service

Two 7558 or 7551 valves connected as a class AB_1 amplifier with 300 volts applied to the plates and 250 volts to grid No. 2 can furnish a power output of approximately 20 watts with 5 per cent distortion. The power output and distortion also depend on the characteristics of the driver valve, the input transformer, and the regulation of the power-supply units.

Because the source of plate and grid-No. 2 supply voltage of a class AB amplifier has internal resistance, the dc voltages applied to the plates and screen grids decrease as power output increases. However, when plate and grid-No. 2 voltages are reduced, power output is also reduced, and a change in load resistance may be necessary to maintain a reasonably low level of distortion.

Fig. 2 shows power output as a function of the internal power-supply resistance over a range of 0 to 700 ohms in series with the plate supply and 0 to 3500 ohms in series with the grid-No. 2 supply. Fig. 2 also includes a typical curve (dashed line) of total harmonic distortion as a function of internal power-supply resistances. The curves indicate that with zero plate and grid-No. 2 supply resistance, the typical power output is 20.5 watts with a total distortion of 2.5 per cent. With the poorest power-supply regulation measured (i.e., 700 ohms of series plate-circuit resistance and 3500 ohms of series grid-No. 2 resistance), the power output drops to 10.5 watts and total distortion increases to 9.3 per cent. Consequently, good voltage-supply regulation is essential for efficient operation.

Transformers having poor audio-frequency response characteristics will degrade the over-all performance of the circuit. The transformer characteristics which are important in modulator applications include the primary and secondary leakage inductance, the capacitances between windings and from each winding to ground, core loss, and winding resistance. Excessive values of any of these factors can cause loss of frequency response, phase distortion, generation of non-linear distortion products, and reduced power output.

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THE IMPORTANCE OF SHIELDING

Don't underestimate the importance of replacing valve and circuit shields in the tuner unit, picture intermediate frequency and sound sections of a television receiver. Carelessness in replacing or securing shields properly can lead to receiver performance problems such as interference beats in the picture, if oscillations, degraded pictures, distorted sound, and critical fine tuning.

Shields have an appreciable effect on the alignment of a receiver and are important to receiver performance. Always check the circuit and valve shielding after servicing a television receiver. Make sure all shields are properly installed and properly grounded.

SILICON

POWER TRANSISTORS

AN APPLICATION GUIDE

(CONCLUSION)

APPLICATIONS

This section gives typical schematic diagrams and design procedures for a variety of circuits using silicon power transistors. The design procedures include both analytic and empirical equations. They also include approximations which apply specifically to RCA silicon power transistors.

200-Watt DC-to-DC Converter Circuit

Specifications:

Input Voltage (V_{in})	42 volts
Output Voltage (V_{out})	200 volts
Efficiency	77%
Frequency	3 Kc
Input Current (I_{in})	6.0 amp
Maximum Output Current [I_{out}(max)]	0.97 amp
Maximum Transistor-Case Operating Temperature	100°C

Design Procedure:

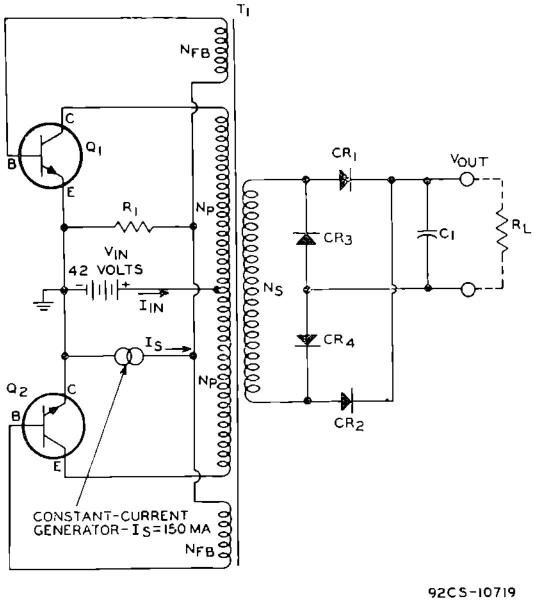
A. Equations for determining number of turns on transformer windings, N_p, N_{FB}, and N_s.

$$N_p = \frac{V_{in} \times 10^8}{4f\,B_{sat}A}$$

where B_{sat} is the saturation flux density of the core material in gausses, A is the cross-sectional area of the core in square centimetres, f is the frequency of oscillation.

$$N_{FB} = \left(N_p\right)\left(\frac{1{\cdot}5\,V_{BE}(\max)}{V_{in}}\right)$$

where V_{BE}(max) is the maximum value of base-to-emitter voltage at I_{in}.



C: |1 μf

CR_1, CR_2, CR_3, CR_4: RCA-1N1763 silicon rectifiers

Q_1, Q_2: RCA-2N1490 high-power silicon transistors

R_1: 8 ohms

T: Core: ferrite toroid, Allen-Bradley Part No.T3750-107B, type RO3 material, or equivalent; primary: 44 turns No.14 wire, bifilar wound (N_p = 22 turns): secondary (N_s) = 105 turns No.24 wire; feedback winding = 16 turns No.20 wire, bifilar wound (N_{FB} = 8 turns)

Fig. 7—DC-to-DC Converter Circuit.

$$N_S = \left(N_p\right)\left(\frac{V_{out}}{V_{in}}\right)$$

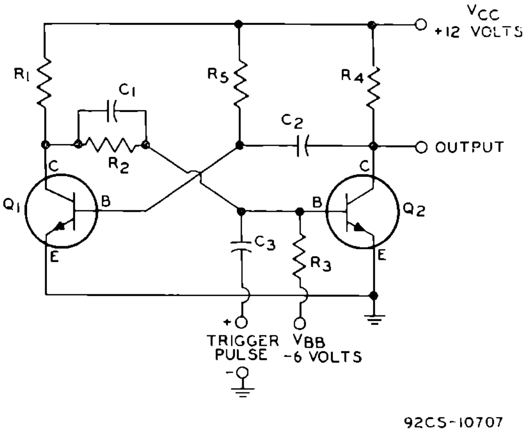
B. Values of starting current I_s are determined from the requirement for starting bias V_{BE}. For RCA silicon power transistors, V_{BE} should be approximately 0·6 volt to insure easy starting. This starting bias can be provided by means of a resistance-voltage-divider network.

C. Notes :

(1) To minimize the number of turns on the transformer windings the frequency of oscillation should be as high as possible, without introducing large core losses or high switching-transient power losses in the transistors.

(2) The value of V_{BE}(max) at a collector current I_{in} for the transistor type used can be measured, or its limit value found by extrapolation of the published limit at specified current from a typical curve of V_{BE} vs. I_C. Such a curve can be obtained from published typical curves of h_{FE} vs. I_C, and V_{BE} vs. I_B.

(3) This circuit can be operated as an inverter by omitting the components connected across the secondary winding. The peak-to-peak square-wave output, under these conditions, is approximately 2 × V_{out} dc.



C_1: 0.005 μf
C_2: 0.01 μf
C_3: 0.05 μf
Q_1,Q_2: RCA-2N1481 medium-power silicon transistors
R_1,R_4: 60 ohms
R_2: 820 ohms
R_3: 5000 ohms
R_5: 1000 ohms

Fig. 8—Monostable Multivibrator Circuit.

Monostable-Multivibrator Circuit for Pulse Generation

Specifications:

Output Pulse :

Amplitude..........	12 volts across 60 ohms
Width	5 μsec
Rise time (t_r)......................	0·8 μsec
Fall time (t_f)......................	0·6 μsec

Trigger Pulse :

Amplitude	5 volts
Width	1 μsec
DC Supply Voltage..............	12 volts
Current Drain	200 ma

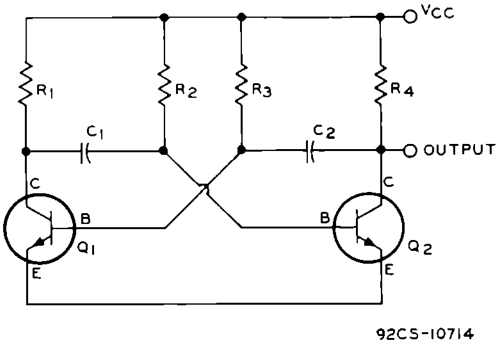


Fig. 9—Astable Multivibrator Circuit.

Design:

A. Values of R_1 and R_4 are determined by pulse requirements, supply voltage, I_{C1}, and I_{C2}. ($I_{C1} \approx V_{CC}/R_1$).

B. Choose R_3 such that

$$R_3 < \frac{V_{BB}}{5} \cdot \frac{1}{I_{CBO_2}\,(\max)}$$

C. The range for R_2 is given by :

$$R_3\left(\frac{V_{CEsat_2}}{V_{BB} - I_{CBO}\,(r_{bb'} + R_3)}\right) < R_2 <$$

$$R_3\left(\frac{V_{CC} - V_{BE}}{V_{BB} + V_{BE_2} + I_{B_2}R_3}\right) - R_1, \text{ where}$$

V_{CEsat_2} is the saturation voltage of Q_2 at I_{C_2}

V_{BE_2} is the base-to-emitter voltage of Q_2 at I_{C_2}

I_{B_2} is the base current of Q_2 at I_{C_2}

V_{BB} is the bias supply with a range such that : 2 volts $< V_{BB} < V_{CC}$

$r_{bb'}$ is the extrinsic base resistance (see Table II).

D. $R_5 < R_1 h_{FE_1}(\min)$ [usually selected such that $R_5 \approx 0{\cdot}8\ R_1 h_{FE_1}(\min)$] where : $h_{FE_1}(\min)$ is the minimum value of dc current transfer ratio of Q_1 at I_{C_2}.

E. The value of C_1 (" Speed-up " capacitor) is determined by trigger-sensitivity requirements and switch-off time constant R_1C_1.

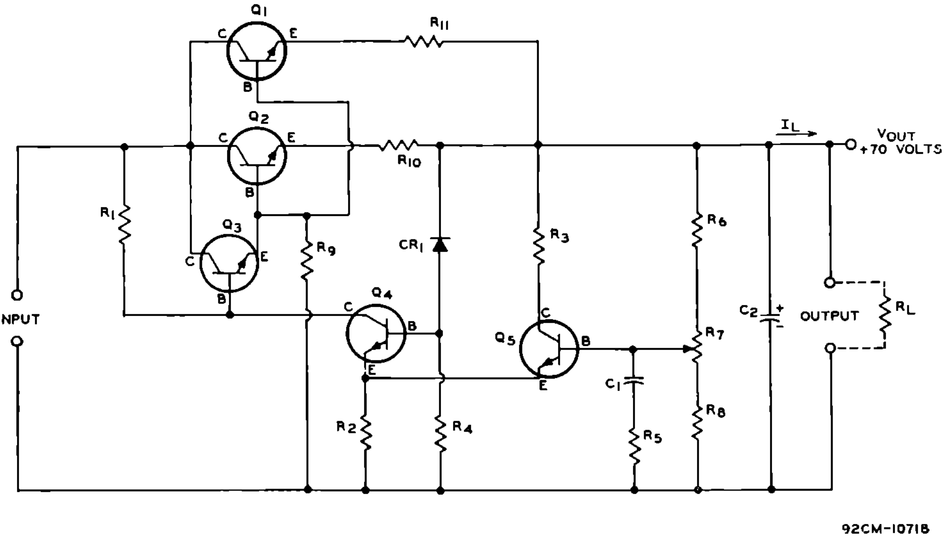
F. The quasistable period is given by the approximate expression :

$$T \approx 0{\cdot}69\ C_2\ R_5$$

assuming $V_{CEsat} = V_{BE}$, and $I_{co}R_5 << E_2$

Astable Multivibrator:

Similar methods of analysis may be used to determine quasistable periods and values for C_2,



C_1: 0.5 μf
C_2: 25 μf, 100 v
CR_1: 1N1363-R silicon reference diode
Q_1, Q_2: RCA-2N1489 high-power silicon transistors
Q_3: RCA-2N1485 intermediate-power silicon transistor
Q_4, Q_5: RCA-2N1482 medium-power silicon transistors

R_1: 620 ohms
R_2: 1000 ohms
R_3: 750 ohms
R_4: 4000 ohms
R_5: 100 ohms
R_6: 2500 ohms
R_7: Potentiometer, 1000 ohms
R_8: 3500 ohms
R_9: 20000 ohms
R_{10}, R_{11}: 0.5 ohm

R_5, and R_1 for the astable multivibrator circuit shown in Fig. 9.

Series-Type Voltage-Regulator Circuit

Specifications:

Regulated Output Voltage	70 volts
Load Current	0-4 amperes
Input Voltage	86 volts ± 15%
Regulation (no load to full load)	2·5%
Ripple Rejection	40 db
Maximum Transistor-Case Temperature	75°C

Design:

A. Select a voltage value (V_{D1}) for reference diode CR_1 such that

$$0{\cdot}1\ V_{out} < V_{D1} < 0{\cdot}9\ V_{out}$$

B. Find I_B(max) for transistor Q_3 :

$$I_{BQ3}(\max) = \frac{I_L(\max)}{[h_{FEQ1}\ (\min)]\ [h_{FEQ3}\ (\min)]}$$

C. Determine I_{CQ4}, R_2, R_3, and R_4 :

$$I_{CQ4} = 4 \max I_{bQ3}$$

$$R_2 = \frac{V_{out}\ V_{D1}}{2I_{CQ4}}$$

$$R_4 = \frac{\dfrac{V_{out}\ V_{D1}}{20I_{CQ4}}}{h_{FEQ4}(\min)}$$

$$R_3 = R_2 \text{ approximately}$$

D. Find values for bleeder network R_6, R_7, R_8 :

$$R_6 + R_7 + R_8 = \frac{\dfrac{V_{out}}{20I_{CQ4}}}{h_{FEQ5}(\min)}$$

where $I_{CQ5} = I_{CQ4}$

Select values for R_6, R_7 and R_8 such that the voltage at the base of Q_5 is approximately equal to $V_{out} - V_{D1}$ when potentiometer R_7 is set at the centre of its range.

E. The component values for network C_1, R_5, and C_2 are arbitrarily selected to reduce the differential amplifier gain at higher frequencies and prevent oscillations. For the phase shifts encountered in the RCA silicon-transistor types shown the over-all loop gain should be reduced to unity at 50 Kc.

F. Values for resistors R_{10} and R_{11} are chosen to provide proper current sharing by Q_1 and Q_2. R_9 provides the I_{co} path for Q_1 and Q_2.

6-Watt Servo-Amplifier Circuit

Specifications:

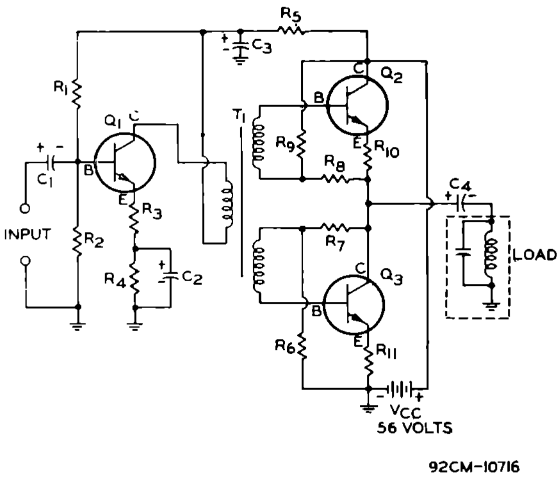
Power Output	6 watts
Signal Frequency	400 cps
Over-all Power Gain	60 db
Over-all Efficiency	50%

Class B Power Amplifier and Driver Design: Output Stage

A. Because no output transformer is used for this type of amplifier design, the supply voltage and peak collector-current requirements are determined directly from the load characteristics. For a given power output (P_o) and load resistance (R_L) the minimum supply voltage [V_{CC}(min)] and peak collector current (I_C peak) are given by :

$$(V_{CC})\text{ min} \approx \sqrt{8\,P_oR_L}$$

$$(I_c)\text{ peak} \approx \sqrt{\frac{2P_o}{R_L}}$$



C1: 10 μf, 15 v.
C2: 47 μf, 15 v.
C3: 20 μf, 50 v.
C4: 500 μf, 50 v.
Q1: RCA-2N1481 medium-power silicon transistor
Q2, Q3: RCA-2N1485 intermediate-power silicon transistors
R1: 68000 ohms
R2: 5600 ohms
R3: 56 ohms
R4: 560 ohms
R5: 3300 ohms
R6, R8: 400 ohms
R7, R9: 18000 ohms
R10, R11: 4 ohms
T: Driver transformer: Core material = 0.014-inch "Crystalligned" (Magnetic Metals Co., Camden, N.J.) or equivalent; primary = 1500 turns; secondary = 450 turns, bifilar wound (each section = 225 turns)
Load: Control phase 6-watt control motor, Kearfott Type R110-5C, or equivalent

Fig. 11—6-Watt Servo-Amplifier Circuit.

B. Q_2 and Q_3 must be selected with a minimum value of $V_{CC}/2$ under $V_{CEO(sus)}$ conditions and a value of V_{CC} under V_{CEX} conditions. They must also be able to handle (I_C) peak and remain within dissipation ratings.

C. Select values of R_7, R_8, R_9, R_{10}, and R_{11} to provide the necessary idling forward bias for the output stages [$(I_C)_o$, $(V_{BE})_o$].
As a guide :

$$(I_C)_o \approx \frac{(I_C)\text{ peak}}{20}$$

$$(V_{BE})_o \approx 0{\cdot}5 \text{ to } 0{\cdot}7 \text{ volt}$$

also $R_{10}\,(I_C)_o = R_{11}\,(I_C)_o = 0{\cdot}1$ to $0{\cdot}8$ volt.

D. Determine the equivalent input resistance (R_{in}') of the output stage at full load.
For Q_2 or Q_3 during conduction this resistance is approximately :

$$R_{in}' = \left[r_{in} + (h_{FE} + 1)\left(R_{10} + \frac{0{\cdot}026}{I_E}\right)\right]$$

where :

$I_E = (I_c)$ peak
$r_{in} = h_{ie}$ measured at $I_C = (I_C)$peak

E. Determine the required driver power (P_i)

$$P_i = \frac{P_o}{P_G} \text{ where } P_G \approx \frac{(h_{FE}^2)R_L}{R_{in}'}$$

Driver Stage and Driver Transformer:

A. Assuming the driver transformer is 90% efficient determine the driver output power (P_d), from :

$$P_d = \frac{P_i}{0{\cdot}9}$$

and select a driver supply voltage (V_{CC}).

B. The primary-to-secondary turns ratio for the driver transformer is

$$\frac{N_p}{N_s} = \frac{R_{Lac}}{R_{in}'}$$

where R_{Lac} is the large-signal ac load resistance for the driver stage and has the value :

$$R_{Lac} = \frac{V_{CC}}{2P_d}$$

R_{in}' is the input resistance of the output stage, as previously determined.

C. The dc operating point for the driver transistor is :

$$V_{CE} \approx V_{CC}' \text{ and } I_c \approx \sqrt{\frac{2P_d}{R_L}}$$

D. Choose an appropriate driver transformer having separate secondary windings. If none is available then the transformer can be designed using basic transformer design data and knowing the following quantities :

(1) $\frac{R_{out}}{R_{Lac}}$ where $R_{out} = h_{oe}$ measured

at $I_c = \sqrt{\frac{2P_d}{R_L}}$

(2) Maximum permissible phase shift at the lowest useful frequency. These quantities determine the required primary mutual inductance and the number of primary turns.

20-Watt Series-Type Audio-Amplifier Circuit

Specifications:

Power Output	20 watts
Distortion at 20 watts output (f = 1000 cps)	1·7%
Bandwidth	20 cps to 25 Kc
Over-all Efficiency	58%
Over-all Power Gain	36 db

Design Procedure:

A. The minimum value of supply voltage [V_{CC}(min)] and peak collector current (I_c peak) in Q_2, Q_3, Q_4, and Q_5 are given by :

$$V_{CC}(\text{min}) = \sqrt{8\ P_o R_L}$$

where : P_o is the maximum output power
R_L is the load resistance

The peak collector current for the output stage is :

$$(I_c)\text{peak}_{out} = \sqrt{\frac{2P_o}{R_L}}$$

The peak collector current for the driver stage is :

$$(I_c)\text{peak DR} = \frac{1}{h_{FE}}\sqrt{\frac{2P_o}{R_L}}$$

h_{FE} measured at $I_C = (I_c)$peak.

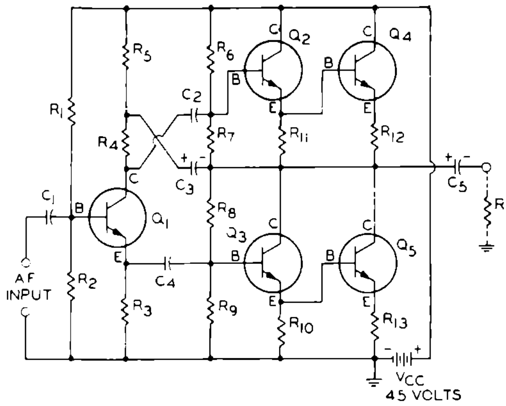
B. Q_2, Q_3, Q_4, and Q_5 must have $V_{CEO(sus)}$ rating greater than $V_{CC}/2$ and V_{CEX} ratings greater than V_{CC}. They must also be capable of handling (I_c) peak without exceeding their dissipation ratings.

C. Select values for R_{12} and R_{13} such that :

$$R_{12},\ R_{13} \approx \frac{R_L}{20}$$

Select idling currents for the output and driver stages such that :

$$(I_C)_{o(out)} = (I_C)_o\ \text{DR} \approx \frac{(I_c)\text{peakDR}}{10}$$



C_1,C_4: 100 μf, 6 v.
C_2: 500 μf, 12 v.
C_3: 500 μf, 25 v.
C_5: 1000 μf, 25 v.
Q_1: RCA-2N1485 intermediate-power silicon transistor
Q_2,Q_3: RCA-2N1481 medium-power silicon transistors
Q_4,Q_5: RCA-2N1489 high-power silicon transistors
R_1: 20000 ohms
R_2: 2700 ohms
R_3,R_4: 39 ohms
R_5: 110 ohms
R_6,R_8: 3600 ohms
R_7,R_9: 200 ohms
R_{10},R_{11}: 120 ohms
R_{12},R_{13}: 0.5 ohm

Fig. 12—20-Watt Series-Type AF amplifier.

Select idling base-to-emitter voltages for the output and driver stages such that :

$$(V_{BE})_{(out)} \approx (V_{BE})_{oDR} = 0{\cdot}5 \text{ to } 0{\cdot}7 \text{ volt}$$

Select values for R_{10} and R_{11} such that :

$$R_{10},\ R_{11} \approx \frac{(V_{BE})_{o(out)}}{(I_C)_{oDR}}$$

D. Design of Phase-Inverter Stage Bias point for phase-inverter stage :

$$V_E > (V_{BE})\text{peak } Q_3 \approx (V_{BE}) \text{ peak } Q_5$$

where V_E is the dc emitter-to-ground voltage of Q_1.

(V_{BE})peak Q_5 can be found from the transconductance characteristics for Q_5 at $I_C = (I_C)\text{peak}_{out}$.

For best performance use (V_E) ≈ 2(V_{BE})peak $Q_5 = V_{R3} = V_{R4}$ and let $(V_{CE})_o\ Q_1 \approx V_{CC}/2$, where $(V_{CE})_o$ is the idling collector-to-emitter voltage.

R_5 should have a value of 20 × R_L because $V_{R5} = (V_{CE})_o\ Q_1 - V_{R3} - V_{R4}$. The bias current for Q_1 is :

$$(I_C)_{o\ Q1} \approx \frac{V_{R5}}{R_5} \approx (V_{CE})_{o\ Q1} - 2V_{R3}$$

$$R_3, R_4 = \frac{V_{R3}}{(I_C)_o Q_1}$$

Select R_1, R_2 to provide the proper operating point for Q_1. Select transistor (Q_1) having adequate dissipation, and current ratings.

E. Determination of Bias Network for the Driver Stages

The bleeder current through R_6, R_7, and R_8, R_9 should be approximately equal to the idling current of Q_2 and Q_3 which is approximately equal to $(I_c)_{o\,DR}$.

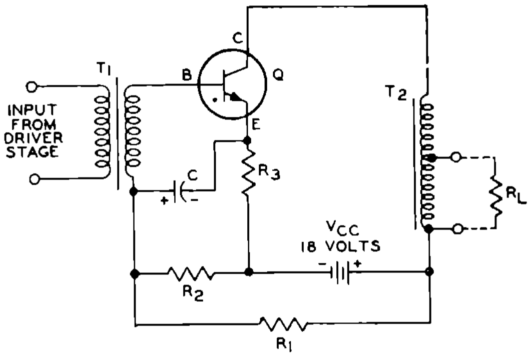
Select values for R_7, R_9 such that : R_7 and R_9 are greater than R_3 and R_4 and less than the input resistances of Q_2 and Q_3, also V_{R7} and $V_{R9} \approx 2(V_{BE})_{o\ out}$.

Select values for R_6 and R_8 such that :

$$R_6, R_8 = \frac{V_{CC}{}^{-2}(V_{BE})_o}{(I_C)_o}$$

F. Selection of Coupling Capacitors

C_1, C_2, C_3, C_4, and C_5 should be large enough to present a coupling impedance no greater than one-tenth of the effective resistance into which these capacitors couple at the lowest frequency of operation. In addition, the ratio C_2/C_4 should be approximately 5 because of the different source impedances seen by Q_2 and Q_3.



C: 1000 μf, 3 v.

Q: RCA-2N1485 intermediate-power silicon transistor

R_1: 110 ohms, 5 watts

R_2: 8.2 ohms

R_3: 1 ohm

T_1: Driver transformer for matching input impedance of Q (12 ohms) to output impedance of driver stage.

T_2: Output autotransformer, Columbus Process Co., Columbus, Indiana, Type X-5383, or equivalent.

Fig. 13—Class A AF-Power-Amplifier Circuit.

Class A Audio-Frequency Power-Amplifier Circuit

Specifications:

Power Output..............	3 watts into 8 ohms
Bandwidth	200 cps to 6 Kc
Distortion	less than 5% at rated power output at 1000 cps
Maximum Transistor-Case Operating Temperature	100° C
Supply Voltage (V_{CC})	18 volts

Design Procedure:

A. Calculate I_C :

$$2I_C^2\,(R_3 + R_{sat}) - V_{CC}\,I_C + 2P_0 = 0$$

where : R_{sat} is the saturation resistance of transistor Q_1, P_o is the desired power output divided by the output-transformer efficiency : i.e., power into the transformer. Use the smaller of the two values of I_C found by solving the equation.

B. Calculate the impedance to be reflected by output transformer.

$$R_L' = \left[\frac{V_{CC} - 2I_C(R_3 + R_{sat})}{2P_1}\right]^2$$

C. Select or design the output transformer on the basis of the reflected impedance, load impedance, direct current in winding, power level, efficiency, and frequency response desired. If an autotransformer is used the dc resistance of the portion of the winding shunted across the load should be very low to minimize direct current in the load.

D. Design Bias Network

It is desirable to use a low value of resistance, such as 1 ohm, for the emitter resistor (R_3). The values of the other two resistors in the bias network (R_1 and R_2) should be chosen to provide the desired I_C and the desired stability factor (S). Considerations involved in choosing a stability factor are discussed elsewhere in this article.

When the values of the emitter resistor (R_3) and the stability factor (S) have been selected, the bias network may be calculated as follows :

$$R_p = \frac{R_3(1 + h_{FE})\,S - 1)}{1 + h_{FE} - S} - r_b$$

where : R_p is the equivalent parallel combination of R_1 and R_2, S is the stability factor, and r_b is the internal base resistance of the transistor, which is approximately equal to the reciprocal of the maximum slope of the typical-input-characteristics curve.

R_1 is given by :

$$R_1 = \frac{V_{CC}R_p}{V_{BE} + I_b(R_p + R_3 + r_b) + I_cR_3}$$

where V_{BE} is the base-to-emitter voltage required for the desired I_b.

The value of R_2 is given by :

$$R_2 = \frac{R_1R_p}{R_1 - R_p}.$$

E. Calculate the required driving power (P_D) from :

$$P_D = \frac{P_o}{P_{gain}} = \frac{P_o\ r_{in}}{(h_{FE})^2\,R_L}$$

where r_{in} is the input impedance of the transistor (found by measuring the slope of a line tangent to the typical-input-characteristics curve at the operating point).

F. Calculate the value for the bias-network bypass capacitor (C).

The response of the bias network will be less than 3 db down at the frequency where the capacitive reactance equals r_{in} ; i.e., when

$$C = \frac{1}{2\pi f r_{in}}$$

Class C Power-Oscillator Circuit

Specifications:

Power Output	10 watts
Frequency	100 KC
Efficiency	Greater than 50%
Supply Voltage (V_{CC})	28 volts

Design Procedure:

A. Choose a value for the resonant load resistance R_L which will absorb the desired output power P_o :

$$R_L = \frac{V_{CC}^2}{2P_o}$$

B. Choose loaded Q of tuned circuit. A value of 10 or greater is desirable if a good output waveform is required. If the waveform is not an important consideration a value less than 10 may be used with consequent higher efficiency (efficiency improves as the ratio of unloaded Q to loaded Q is increased).

C. Choose values of C_2 and L for the tuned circuit :

$$C = \frac{Q}{2\pi f R}$$

where R is the parallel combination of R_L, the equivalent parallel resistance of the output impedance of the transistor, and the impedance reflected from the feedback winding.

$$L = \frac{1}{4\pi^2 f^2 C}$$

D. Determine the number of turns on the feedback (base) winding by experiment. As a start the feedback winding may have 1/4 as many turns as the output winding.

E. Determine the values for the bias network by experiment. Fourier analysis shows that at the angle of collector-current flow normally used in class C operation the average collector current differs very little from the rms collector current. Because the bias is developed partly by the signal (i.e., by the resulting charge stored on the capacitor), the average collector-to-base voltage is greater than V_{CC}. As a starting point a value for the bias resistor may be determined from :

$$R_1 = \frac{10V_{CC}\sqrt{P_oR_L}.}{P_o}$$

provided it may be assumed that

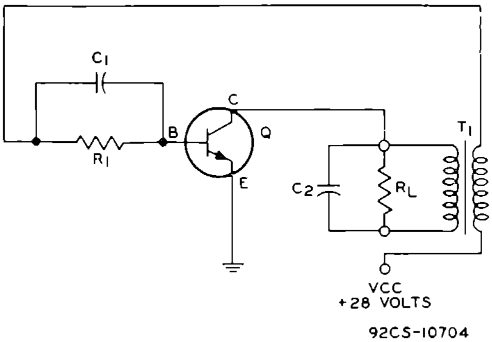
$$h_{FE} = 10$$

$$I_c(RMS) = I_C(AV.)$$

$$V_{CB} = V_{CC}$$

The bias capacitor should not have too large a value or the oscillator will operate intermittently, due to the inability of the capacitor to discharge sufficiently during the cycle. A good starting point is to make

$$C_1 = \frac{5}{fR_1}$$



C_1: See Design Procedure

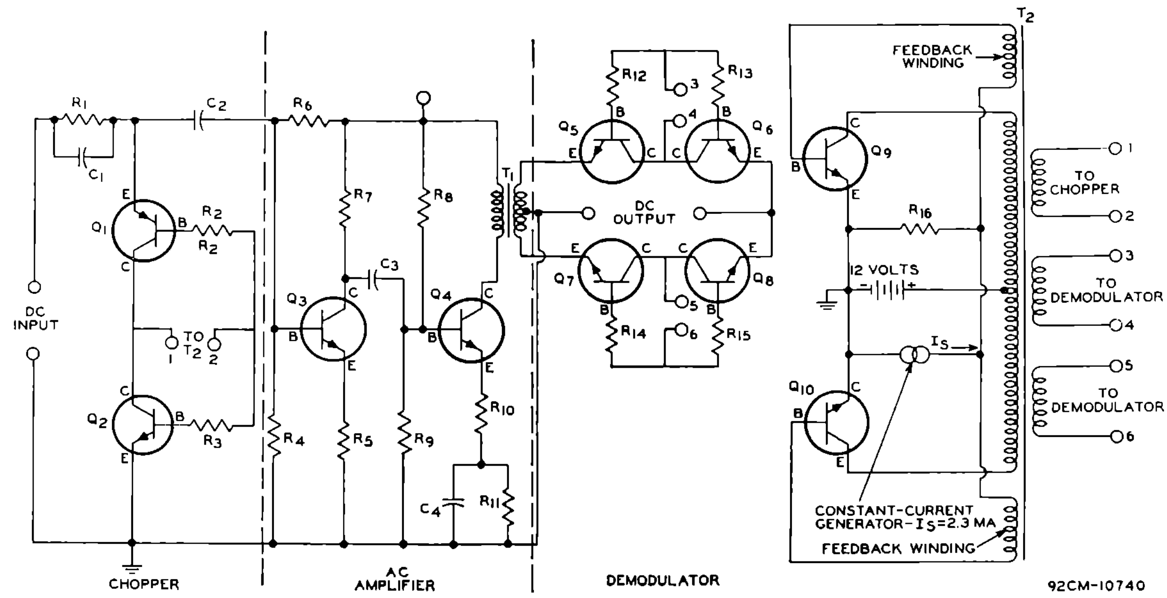
C_2: 0.33 μf

Q: RCA-2N1490 high-power silicon transistor

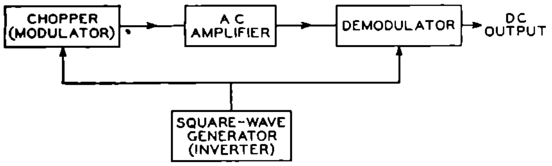
R_1: 510 ohms

T: Air-core rf transformer: collector winding = 19 turns No.10 wire; base winding = 5 turns No.22 wire; inside diameter of windings = 0.88 inch. (A ferrite-rod-core transformer may also be used.)

Fig. 14—Power-Oscillator Circuit.



C1: 5 μf
C2: 10 μf
C3: 10 μf
C4: 10 μf
Q_1-Q_{10}, inclusive: RCA-2N1481 medium-power silicon transistors
R_1: 510 ohms
R_2, R_3, R_9: 1000 ohms
R_4, R_7: 2000 ohms
R_5: 91 ohms
R_6: 20000 ohms
R_8: 5000 ohms
R_{10}: 120 ohms
R_{11}: 50 ohms
R_{12}, R_{13}, R_{14}, R_{15}: 500 ohms
R_{16}: 510 ohms



(FREQUENCY=4 Kc)

92CS-10700

T_1: Demodulator transformer, primary, 400 turns No.38 wire; secondary, 1200 turns center-tapped (bifilar wound); core, Allen Bradley Ferrite Pot Core No.C1300-215A (Type RO-3 material) or equivalent.

T_2: Inverter transformer, primary, 70 turns No.26 wire; center-tapped (bifilar wound); chopper winding, 15 turns No.30 wire; demodulator windings, each 30 turns No.30 wire; feedback winding, 30 turns No.30 wire, center-tapped (bifilar wound); core, Allen Bradley Ferrite Toroid No.T-1312-103A, (Type RO-3 material) or equivalent.

Fig. 15 — Low-Level DC-Amplifier Circuit

F. When the oscillator has been constructed, the values of C_1, R_1 and the number of turns on the feedback winding should be adjusted until the desired power output is obtained with minimum average (dc) collector current. An efficiency of 60% was obtained with the circuit shown. Efficiency is calculated as follows :

$$\eta = \frac{P_O}{V_{CC}I_C(AV.)}$$

Low-Level DC-Amplifier Circuit

Specifications:

Input-Voltage Range	±5 to ±50 millivolts
Output Voltage Range	±1 to ±10 volts
Load Impedance	500 ohms
Voltage Gain	200

Design Procedure:

A. Saturable-Core Square-Wave Oscillator—Inverter

1. The procedure for the design of transformer T_2 is the same as that outlined in the design procedure for the DC-to-DC Converter.

2. Select R_{16} which will provide a bias point which will assure self-starting of the inverter.

3. Select an operating frequency suitable for the transients the amplifier must handle (a frequency between 1 and 5 Kc is usually satisfactory).

B. Modulator—Series Pair Chopper

1. To insure good low-level operation transistors Q_1 and Q_2 should be matched for null voltages and currents in the inverted operating mode ; i.e., emitter and collector terminals interchanged.

2. R_1 should have a value smaller than R_{in} of the ac amplifier section and greater than

the desired input resistance of the entire amplifier.

3. Choose a peak voltage for the chopper secondary winding greater than the maximum dc input voltage.
4. Choose values for R_2 and R_3 approximately twice the value of R_1.

C. AC Amplifier

1. The design procedure for the ac amplifier is the same as that given for class A amplifiers. The procedure for the design of T_1 is also the same as that for the design of the class A amplifier output transformer. T_1 must be designed to transmit the square-wave carrier with negligible change in its rise-time and fall-time characteristics.
2. The turns ratio of T_1 is given by :

$$N_s/N_p = \frac{4V_{out}\ DC}{V_{out}\ Ampl.}$$

where : V_{out} DC is the required maximum dc output voltage. V_{out} Ampl. is the peak-to-peak square-wave voltage developed across the output of the amplifier stage.

3. Although the input impedance of the ac amplifier should be reasonably high, the overall stability factor S must be low to insure a high degree of temperature stability. Therefore, a compromise must be made in selecting base bias resistor values R_4 and R_6.
4. The load resistance for the amplifier is given by :

$$R_L\ Ampl. \times \frac{N_p^2}{4N_s^2} R_L\ DC$$

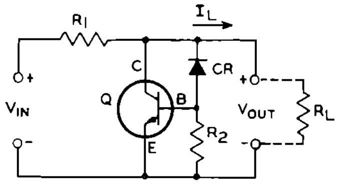
where : R_L DC is the load resistance seen by the dc output of the amplifier.

D. Demodulator—Chopper

1. The demodulator transistors Q_5, Q_6, Q_7, and Q_8 should be selected in matched pairs so that their inverted offset voltages cancel.
2. Base resistors R_{12}, R_{13}, R_{14}, and R_{15} should be approximately equal to R_L DC.
3. The two demodulator windings of the inverter transformer (T_2) must be bifilar wound, and the peak-to-peak voltages developed across these windings must be greater than twice the maximum dc output voltage. The number of turns for each of these windings is given by :

$$N_s\ Demod. \geqq \frac{2V_{out}\ DC}{V_{in}} N_p$$

where : V_{in} is the dc input voltage of the inverter. N_p is the number of turns on the primary winding of T_2.



92CS-10700

CR: 1N1823 voltage-regulator diode
Q: RCA-2N1485 intermediate-power silicon transistor
R_1: 28 ohms　　R_2: 50 ohms

Fig. 16—28-Volt Shunt-Type Regulator Circuit.

28-Volt Shunt-Regulator Circuit

Specifications:

Regulated Output Voltage	28 volts
Load Current Range	0 to 0·5 ampere
Input Voltage Range	42 to 56 volts
Regulation	1·5%
Output Resistance	0·8 ohm

Design Procedure:

A. Select a transistor type capable of satisfying the requirements of the regulator for I_C(max), P_D(max) and V_{CE}(max).
where :

$$I_c(max) \approx I_L(max)$$
$$V_{CE}(max) \approx V_{out}$$
$$P_D(max) \approx I_c(max)\ V_{CE}(max)$$

B. Select R_2 such that the bleeder current through the voltage-reference diode CR is sufficient to break down the diode where :

$$I_{R2} \approx \frac{0{\cdot}7}{R_2} \text{ amp}$$

C. The maximum permissible value of the reference-diode resistance (R_D) can be found from the required output resistance (R_o) for the regulator.

$$R_o \approx \frac{R_D + \dfrac{h_{ie}R_2}{h_{ie} + R_2}}{1 + h_{FE}\left(1 + \dfrac{h_{ie}}{R_2}\right)}$$

where : h_{ie} is the input resistance and h_{FE} is the dc current transfer ratio of Q_1 at $I_C = I_L(max)/2$.

D. Select diode CR to provide a reference voltage $V_R = V_{out} - V_{BE}$, where V_{BE} is the dc base-to-emitter voltage of Q_1 for $I_C = I_C(max)$ (obtained from published data for Q_1).

CR must handle a maximum current :

$$I_{R2} + \frac{I_C(max)}{h_{FE}}$$

have a resistance $\leqq R_D$, and have a dissipation rating $\geqq$

$$V_R \left(I_{R_2} + \frac{I_C(\text{max})}{h_{FE}} \right)$$

E. The value of R_1 depends upon the variation of the input voltage (V_{in}). The maximum and minimum values of V_{in} given in terms of R_1 are :

$$V_{in}(\text{max}) = V_{out} + R_1[I_L(\text{max}) + I_C(\text{max})]$$

$$V_{in}(\text{min}) = V_{out} + R_1 I_C(\text{max})$$

For the usual case $I_C(\text{max}) = I_L(\text{max})$.

(With acknowledgements to RCA)

EEV VIDICONS AT WORK

Underwater television equipment has been used very successfully in the past for search and recovery duties and at the recent Second Industrial Photographic and Television Exhibition (in England) a new underwater television system using EEV 1″ Vidicon pick-up tubes was demonstrated.

The complete Marconi-Siebe, Gorman camera is only 3ft. by 2ft. 6ins. including the lamps and casing, which means that the camera can be used in quite confined spaces. The buoyancy of the camera can be balanced when under water so that it becomes virtually weightless and as it is principally a hand held equipment, the camera thereby provides maximum operating facility to the diver.

The Vidicon pick-up tubes supplied by EEV are approximately 1 inch diameter by 6 inches long and these small dimensions contribute significantly to the compactness and portability of the camera. The lighting equipment comprises 2 x 150-watt lamps which will provide an acceptable illumination level for the sensitive vidicon pick-up tube even in water containing a high percentage of suspended solids.

Another novel application of closed circuit television using English Electric vidicon pick-up tubes is the public televiewing platform on the site of the new Midland Bank Overseas Branch at the Junction of Gracechurch Street and Fenchurch Street, London. A 21″ monitor has been installed by Taylor Woodrow Construction Ltd., with controls for remotely panning and tilting a Marconi industrial television camera. In this case the EEV 1-inch vidicon pick-up tube presents pictures of work in progress which would otherwise be hidden from view. The public can move the camera in bearing and elevation and with this aid the "Sidewalk Superintendents" will be able to watch developments very much more comprehensively than ever before.

VAPOUR-COOLED VALVES FOR INDUSTRIAL HEATERS

English Electric forced air-cooled triodes have proved reliability in valve oscillator service, but the polluted atmospheres encountered in foundries and forge shops etc. involve prohibitive cost in air filtration for this type of cooling. This problem is overcome by using vapour cooled valves in completely enclosed units.

With the type of vapour cooling used with EEV valves, the anode of the valve sits in a tank of water and the heat, generated by the valve, boils the water at atmospheric pressure. The steam then given off has the heat extracted by a condenser; a compact cooling arrangement is provided by incorporating the boiler and condenser in one tank. This system is particularly suitable for industrial heaters where the small quantity of water required to cool the work coils can also be utilised to extract the heat from the condenser. Distilled water is used in the boiler to prevent the formation of scale on the valve anode.

Induction heaters operating in the frequency range of 1 Kc to 10 Kc have traditionally been powered by motor alternators. Recent improvements in valve design have, however, led to an increase in valve efficiency which now makes the valve oscillator a serious competitor. Moreover, valve oscillators have other advantages in that output power can be varied to close limits and are self compensating with load power factor changes.

A variation of the cooling system described above is used in transmitter and similar applications. In transmitter applications there is a reservoir, a boiler in which the valve is mounted and a condenser which is often cooled by natural convection—the water in some cases being circulated by thermosiphon action—thus eliminating all moving parts. Very little water is lost. Using additional equipment, the waste heat extracted from the condenser can be used for building heating, which may amount to a significant saving.

20-AMPERE STUD-MOUNTED RECTIFIERS

FOR INDUSTRIAL AND MILITARY APPLICATIONS

STUD CATHODE	REVERSED POLARITY VERSIONS
1N248-C	1N248-RC
1N249-C	1N249-RC
1N250-C	1N250-RC
1N1195-A	1N1195-RA
1N1196-A	1N1196-RA
1N1197-A	1N1197-RA
1N1198-A	1N1198-RA

Fourteen new types of silicon power rectifier, designed to meet stringent military, mechanical and environmental specifications, these rectifiers are manufactured by the diffused-junction process, which, with the hermetic sealing and welded construction, provides exceptional uniformity of characteristics. These rectifiers, using the JEDEC DO 5 outline, feature low thermal resistance, low leakage current, and low forward voltage drop. The high output current available provides output currents of up to 84 amperes, using 6 rectifiers in a three-phase full wave bridge circuit, and up to 60 amperes using four rectifiers in a single-phase full wave bridge circuit.

These rectifiers are intended for use in power supplies of mobile equipment, dc-to-dc converters, battery charges, dynamic braking systems, aircraft and missile power supplies, high-power transmitter and rf-generator power supplies, machine-tool controls, dc motor power supplies, and in other heavy industrial and military equipment.

HALF-WAVE RECTIFIER SERVICE

Maximum Ratings

Absolute-Maximum Values for Supply Frequency of 50-60 cps, Single-Phase Operation, with Resistive or Inductive Load.

Peak Inverse Volts:

1N248-C	55	1N1196-A	400
1N249-C	110	1N1197-A	500
1N250-C	220	1N1198-A	600
1N1195-A	300		

RMS Supply Volts:

1N248-C	39	1N1196-A	284
1N249-C	77	1N1197-A	355
1N250-C	154	1N1198-A	424
1N1195-A	212		

DC Blocking Volts:

1N248-C	50	1N1196-A	400
1N249-C	100	1N1197-A	500
1N250-C	200	1N1198-A	600
1N1195-A	300		

Forward Average DC Amperes at $T_c = 150°C$	20
Peak Recurrent Amperes	90
Peak Surge Amperes,* One half cycle, sine wave	350
Case Temperature, Operating and Storage	−65 to +175°C

Characteristics, at $T_c = 150°C$

Max. Forward Voltage Drop† (volts)	0.6

Max. Reverse Current† (ma):

1N248-C	3.8	1N1196-A	2.5
1N249-C	3.6	1N1197-A	2.2
1N250-C	3.4	1N1198-A	1.5
1N1195-A	3.2		

---

* Superimposed on device operating within the maximum specified voltage, current, and temperature ratings and may be repeated after sufficient time has elapsed for the device to return to the presurge thermal equilibrium conditions.

† At maximum peak inverse voltage, average forward amperes = 20, and averaged over one complete cycle.

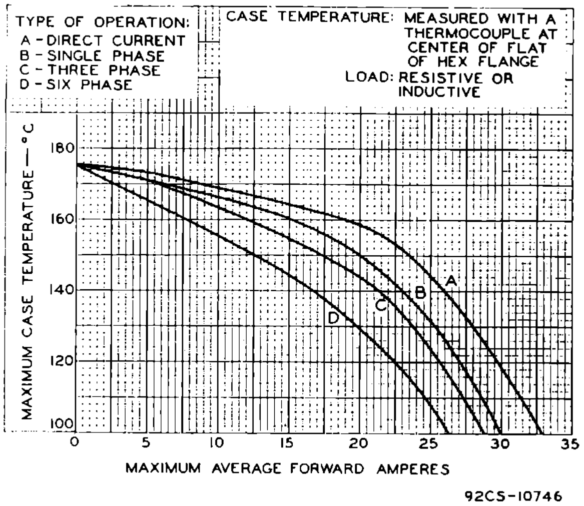
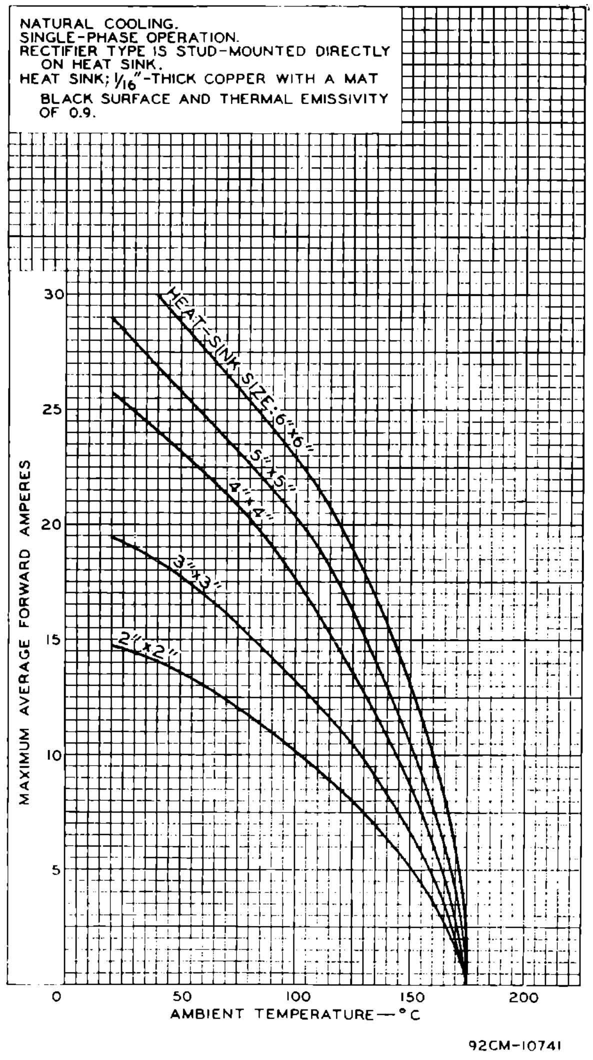


Fig. 1—Rating Chart I for Types 1N248-C, 1N249-C, 1N250-C, 1N1195-A, 1N1196-A, 1N1197-A, 1N1198-A, and corresponding reverse-polarity versions.



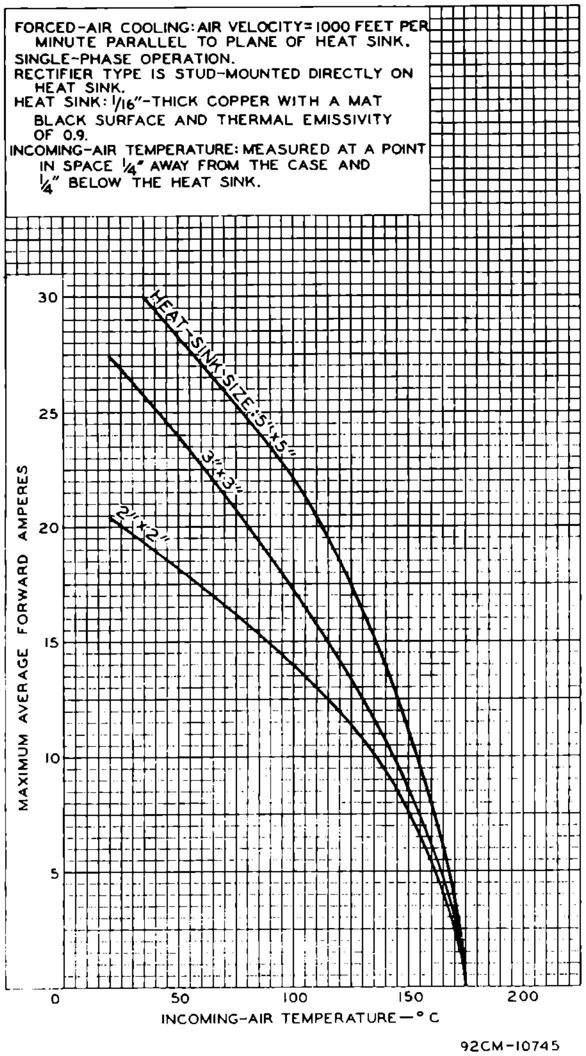


Fig. 3—Rating Chart III for Types 1N248-C, 1N249-C, 1N250-C, 1N1195-A, 1N1196-A, 1N1197-A, 1N1198-A, and corresponding reverse-polarity versions.

Fig. 2—Rating Chart II for Types 2N248-C, 1N249-C, 1N250-C, 1N1194-A, 1N1196-A, 1N1197-A, 1N1198-A, and corresponding reverse-polarity versions.

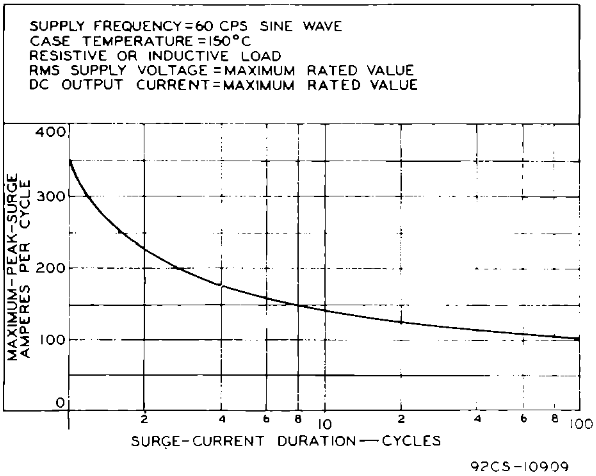


Fig. 4—Rating Chart IV for Types 1N248-C, 1N249-C, 1N250-C, 1N1195-A, 1N1196-A, 1N1197-A, 1N1198-A, and corresponding reverse-polarity versions.

OPERATING CONSIDERATIONS

Because these rectifiers may operate at voltages which are dangerous, care should be taken in the design of equipment to prevent the operator from coming in contact with the rectifier.

The recommended installation torque is 26 to 36 inch-pounds applied to a ¼-24 UNF-2A hex nut assembled on thread.

The applied torque during installation should not exceed 75 inch-pounds.

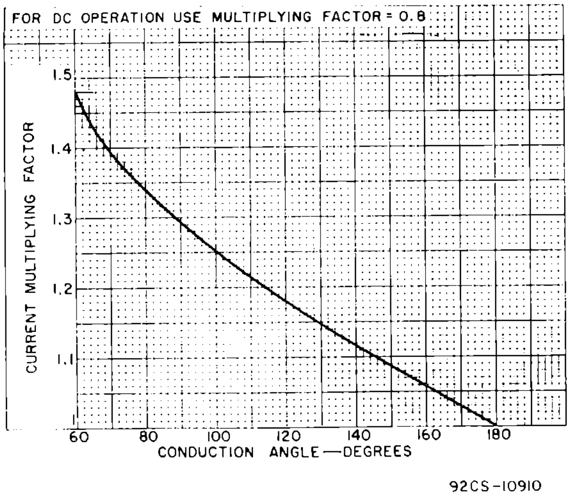


Fig. 5—Chart V for Types 1N248-C, 1N249-C, 1N250-C, 1N1195-A, 1N1196-A, 1N1197-A, 1N1198-A, and corresponding reverse-polarity versions.

Use of Rating Charts

Chart V is used in conjunction with Rating Charts II and III to determine maximum average forward amperes per rectifier unit for polyphase operation and dc operation. The procedure for the use of Chart V is as follows:

Step 1: From Chart V determine the current-multiplying factor for the applicable conduction angle. (For dc operation use current multiplying factor of 0.8.)

Step 2: Divide the required load current in amperes by the number of rectifier circuit branches—as shown in the accompanying table—to determine average forward amperes per rectifier element.

TYPE OF OPERATION	NUMBER OF CIRCUIT BRANCHES
Single-phase, Full-wave:	
Centre Tapped	2
Bridge	2
Three-phase:	
Wye	3
Double-wye	6
Bridge	6
Six-phase, Star	6

Step 3: Multiply average forward amperes established in Step 2 by the current multiplying factor established in Step 1 to determine adjusted average forward amperes per rectifier element, for use with Rating Chart II or Rating Chart III.

Step 4: Using the product obtained in Step 3, determine from Rating Chart II or Rating Chart III either (a) the maximum allowable incoming-air temperature or ambient temperature for a given heat-sink size, or (b) the minimum heat-sink size for a given incoming-air temperature or ambient temperature.

EXAMPLE

Conditions:

(a) Three-phase, half-wave operation; conduction angle = 120°.
(b) Desired output current = 45 amperes.
(c) Forced-air cooling; incoming-air temperature = 90°C.

Problem:

Determine minimum heat-sink size.

Procedure:

Step 1: From Chart V, the current multiplying factor for a conduction angle of 120° is 1.18.

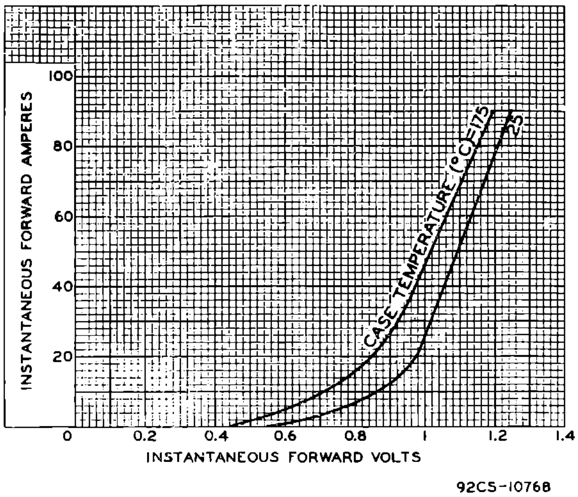


Fig. 6—Typical Forward Characteristics for Types 1N248-C, 1N249-C, 1N250-C, 1N1195-A, 1N1196-A, 1N1197-A, 1N1198-A, and corresponding reverse-polarity versions.

Step 2: For three-phase half-wave operation the number of rectifier circuit branches is three. The average forward current through each rectifier element is, therefore, 45/3, or 15 amperes.

Step 3: Multiplying average forward amperes (15) obtained in Step 2 by the current multiplying factor (1.18) obtained in Step 1 yields 17.7 adjusted average forward amperes.

Step 4: From Rating Chart III, for forced-air cooling, the minimum heat-sink size for the conditions shown in Step 3 is 3″ x 3″.

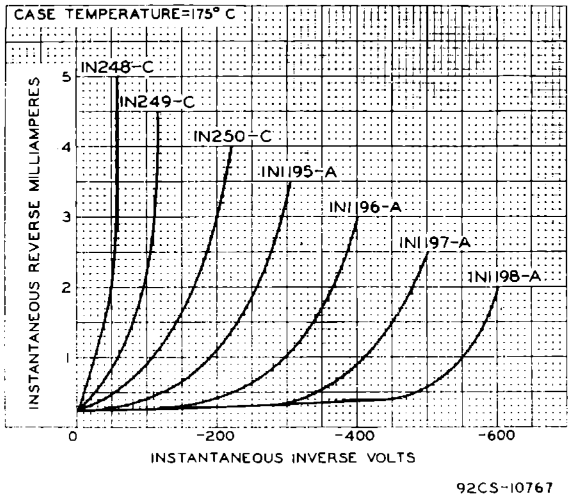


Fig. 7—Typical Reverse Characteristics for Types 1N248-C, 1N249-C, 1N250-C, 1N1195-A, 1N1196-A, 1N1197-A, 1N1198-A, and corresponding reverse-polarity versions.

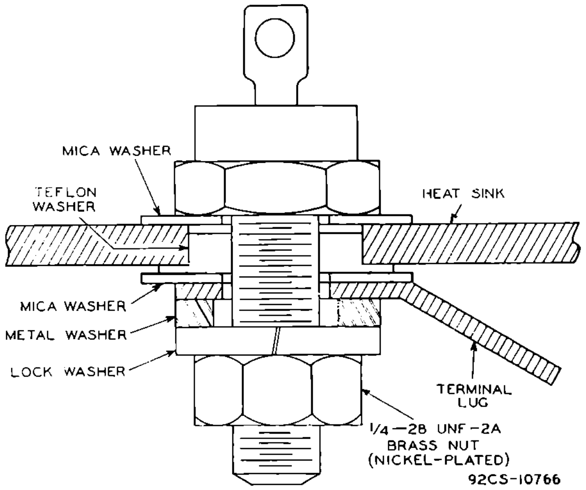
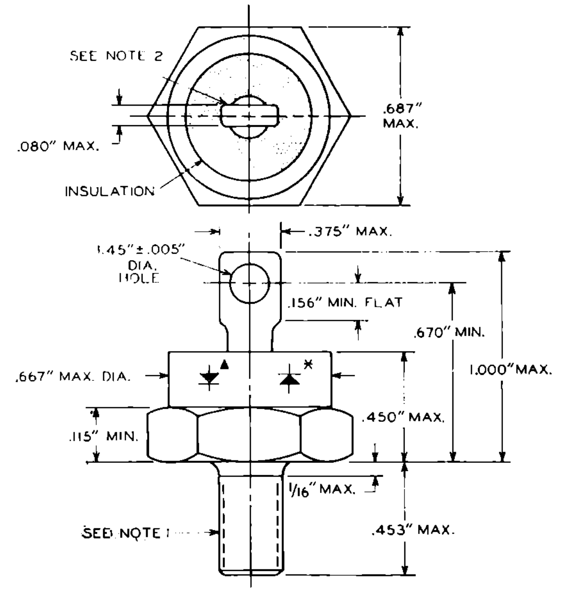


Fig. 8—Suggested Mounting Arrangement.

DIMENSIONAL OUTLINE



NOTE 1: Must withstand torque of 30 inch-pounds applied to 1/4-28 UNF-2A nut assembled on thread.

NOTE 2: Angular orientation of this terminal undefined.

NOTE 3: Device can be used in any position.

POLARITY: Triangle shows polarity symbol for stud cathode types; polarity symbol for reversed polarity types indicated by asterisk.

NEW RELEASES

2054

Five megawatts of useful peak power output is the capability of the new 2054 when used as a plate-pulsed rf power amplifier in a cathode-drive circuit at a frequency of 440 Mc, at a pulse duration of 2000 microseconds, and at a duty factor of 0.06. This water-cooled super-power triode is intended for applications such as long-range search radar, pulsed transmission in communications service, and particle accelerator service.

4017, 4019, 4020

These are three new travelling-wave tubes with periodic permanent-magnet focusing. They are intermediate noise factor amplifier tubes intended respectively for S-band (2000-4000 Mc), L-band (1000-2000 Mc), and C-band (4000-7000 Mc) microwave systems. These tubes feature a minimum CW power output of 10 milliwatts, with typical small-signal gains of 35 db (4017) and 32 db (4019, 4020). Typical noise factors are 15 db for the 4017, 16 db for the 4019, and 17 db for the 4020.

4604

In push-to-talk mobile and emergency communications equipment where standby power comes at a premium, the new quick-heating beam power type 4604 provides instant power. Intended for CW and FM telephony service the 4604 can be operated at full input up to 60 Mc and with reduced input to 175 Mc. At 175 Mc it can deliver a power output of about 30 watts with a gain of 6.5. The quick-heating filament in the 4604 takes less than a second to reach operating temperature. Designed to operate at 6.3 volts, the coated filament draws only 0.65 ampere. Because the filament is operated only during transmission periods, and not during standbys, battery power is conserved. Additional features of the 4604 include: sturdy structure, small size, and a maximum plate dissipation of 25 watts.

6939

The 6939 is a twin power pentode of the 9-pin miniature type. This valve, which has built-in neutralizing capacitors, is intended especially for use as a push-pull rf-power-amplifier or as a frequency-multiplier in communications equipment operating at frequencies up to 500 Mc. Features which contribute to the excellent performance of this valve in communications equipment include 14 watts CW input (ICAS) up to 500 Mc, and 6 watts (ICAS) useful power output as an rf power amplifier up to 500 Mc.

7842, 7843, 7844

The 7842. 7843, and 7844 are three new, very small, conduction-cooled uhf beam power valves for missiles, satellites, or mobile equipment where the use of air cooling may not be practical. They may be used as an rf power amplifier, oscillator, frequency multiplier, af power amplifier, or modulator. With a maximum CW plate input of 180 watts, they are designed for operation with full ratings at frequencies up through the Aeronautical Radio-Navigation Band of 960 to 1215 Mc, and are useful up through 2000 Mc and above. The 7844 has a 6.3-volt/2.1-ampere heater, with oxide-coated-type cathode; the 7843 is similar but has 26.5-volt/0.52-ampere heater. The 7842 is similar to the 7844, but has 6.3-volt/3-ampere heater, with matrix-type cathode. The 7842 is designed especially for applications where dependable performance under severe shock and vibration is essential. The 7842 withstands vibrational accelerations of 10g and shock impact as great as 500g.

CORRECTION

Most readers will already have realised that the diagrams for Figs. 3 and 4 of "Photoconductivity" in the February issue, pages 25 and 26, were transposed.


AWV