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TRANSISTORIZED VOLTAGE REGULATORS

Transistorized voltage regulators offer a number of advantages over other types because of their small size, low cost, reliability, accuracy and range of control.

There are three basic types of regulating systems: series regulators in which a voltage-controlled element is placed in series with the load, shunt regulators in which a current-controlled element is placed in shunt with the load, and series-shunt regulators in which both series and shunt elements are used. Regulators employing any of these three types of system can provide constant voltage, constant current, or constant impedance across the load.

This article discusses transistorized voltage regulators of both the series and shunt types. Included are design considerations, step-by-step design procedures, and the solutions to sample design problems. An appendix contains the derivation of design equations. Whilst a few silicon transistors are used in the examples, the information is of course applicable to any suitable types, either silicon or germanium.

(With acknowledgements to RCA)

SERIES-TYPE VOLTAGE REGULATORS

In series-type voltage regulators, a cascaded dc amplifier amplifies an error or difference signal obtained from a comparison between a portion of the output voltage and a reference voltage. This amplified error signal forms the input to a regulating element in series with the load across which a controlling voltage is developed. The amplifier and series elements, which are treated here as generalized elements, may include any number of individual transistor stages necessary to fulfill design requirements.

A typical transistorized series voltage regulator is shown in Fig. 1. In this circuit,

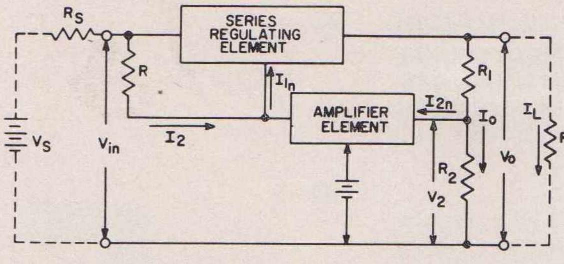


Fig. 1 - Typical transistorized series voltage regulator.

R_L is the load resistance, R_S is the resistance of the regulator input voltage or source (i.e., transformer resistance, rectifier resistance, or the like), V_S is the open-circuit input voltage to the regulator, ΔV_S is the input-voltage variation, V_R is a reference voltage, V_o is the output voltage, and V_2 is the error signal detected by the dc amplifier element.

The series regulating element consists of one or more transistors connected in a common-collector configuration, as shown in Fig. 2.

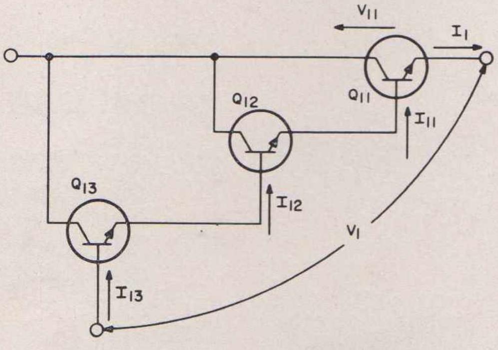


Fig. 2 - Series regulating element consisting of two or more transistors in common-collector configuration.

The number (ln) of additional transistor stages connected in a Darlington connection is determined by the current requirement I_{1n}. The maximum voltage V_{in} appearing across the series elements is equal to or greater than the maximum change in voltage at the regulator input (ΔV_S) due to changes in either the open-circuit input voltage (V_S) or the load current (I_L). The transistors chosen for a series element must have a minimum forward-bias collector-to-emitter voltage rating greater than this maximum voltage.

Typical dc amplifier elements such as those shown in Fig. 3 include an output stage working

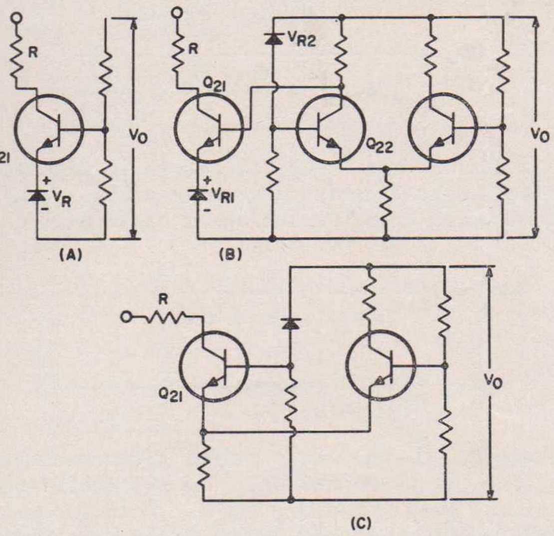


Fig. 3 - Three typical dc amplifier elements.

into a load resistor R and the necessary number (2n) of dc cascaded stages to provide the required amount of gain for a given condition of line-voltage or load-current regulation. The reference-voltage source (V_R) is placed in one of the cascaded stages in a manner such that an error or difference signal between V_R and some portion of V_O is developed and amplified. Some form of temperature compensation must be included to insure stability of the dc amplifier.

Frequency stability is especially important in the design of transistorized series regulators. Because negative feedback is used in the amplifier element, a total phase shift of 180 degrees around the entire loop (amplifier and series element) at high frequency results in oscillations unless the closed loop gain is less than unity. Therefore, at the frequency where total phase shift is 180 degrees, provision should be made to reduce closed loop gain to less than unity. The use of shunt capacitance in the output or elsewhere in the amplifier section produces the necessary gain "roll-off" with frequency.

Design Procedure

The following step-by-step procedure is recommended for the design of transistorized series-type voltage regulators. The equations used in the procedure are derived in the Appendix.

1. List input requirements, load conditions, and output-voltage requirements in terms of the following parameters:
   input voltage (V_S),
   input-voltage variation (ΔV_S),
   source resistance (R_S),
   output load resistor (R_{L_o}),
   output-load-resistance variation (ΔR_L),
   output voltage (V_o),
   output-voltage variation (ΔV_o), and
   load current (I_{L_o}).

The terms V_S and R_{L_o} are design-center values: ΔV_S and ΔR_L are maximum deviations from these values.

2. Select an appropriate value and source of reference voltage V_R (e.g., battery, voltage-reference diode, bleeder-resistance network, etc.). The value of V_R should be chosen somewhere between 0.2 and 0.9 times the value of V_o, and preferably as high as practical. A typical value for V_R is 0.5 V_o.

3. Select values for V_2 and I_2 based on the following considerations:

(a) For a single dc stage, $V_2 = V_R$. If additional stages are used, $V_2 = V_R + \frac{1}{2}(V_o - V_R)$.

(b) The value of I_2 should be less than the maximum current rating ($I_{VR_{max}}$) of the reference source. The following relation may be used as a guide to selecting the value of I_2:

$$I_2 < \frac{I_1}{20}, \text{ where } I_1 = I_L = \frac{V_o}{R_{L_o}}$$

4. Select the series transistor Q_{11}. The maximum collector-to-emitter voltage ($V_{CE_{max}}$) rating and the maximum collector-to-emitter saturation voltage ($V_{CE_{sat_{max}}}$) rating* of this transistor must satisfy the following conditions:[1]

$$V_{CE_{max}} = V_S + \Delta V_S - V_o\left(1 + \frac{R_S}{R_{L_o} + \Delta R_L}\right) \qquad (1)$$

* Maximum rated saturation voltage can be calculated from the rated collector-to-emitter saturation resistance and the specified current.

1 See Appendix, equations (A-12) and (A-13).

$$V_{CE_{sat_{max}}} = V_S - \Delta V_S - V_o\left(1 + \frac{R_S}{R_S - \Delta R_L}\right) \quad (2)$$

$$\text{at } I_c = \frac{V_o}{R_{L_o} - \Delta R_L}$$

The maximum collector-current rating $I_{C_{max}}$ of transistor Q_{11} must be greater than $I_{1_{max}}$:[2]

$$I_{C_{max}} \geq \frac{V_o}{R_{L_o} - \Delta R_L} \quad (3)$$

In addition, the transistor must be capable of dissipating the maximum power required by the regulator at a given operating temperature. Maximum power dissipation PD_{max} is given by the following equations:[3]

$$\text{If } I_{1_{max}} > \frac{V_S + \Delta V_S - V_o}{2R_S},$$

$$\text{then } PD_{max} = \frac{(V_S + \Delta V_S - V_o)^2}{4R_S} \quad (4)$$

$$\text{If } I_{1_{max}} < \frac{V_S + \Delta V_S - V_o}{2R_S},$$

$$\text{then } PD_{max} = (V_S + \Delta V_S - V_o)(I_{1_{max}}) - I_{1_{max}}^2 R_S \quad (5)$$

5. Let $I_{1n} = I_2/10$, and select the number n-1 of additional series-regulator stages to satisfy the following condition for I_{1n}:

$$I_{1n} = \frac{I_1}{h_{FE_{11}} \cdot h_{FE_{12}} \cdot h_{FE_{13}} \cdot \cdot \cdot h_{FE_{1n}}}$$

where $h_{FE_{1n}}$ is the dc forward-current transfer ratio of the Q_{1n} stage when $I_C = I_{1(n-1)}$.

Select the transistor type for each stage, based on the collector-current requirement for that stage.

6. Determine the voltage drop between the output and the control input of the series regulating element. This voltage drop (V_1) is the sum of the base-to-emitter voltages of the n stages of the series element at their respective collector currents.

7. Determine the design-center value for the collector-to-emitter voltage V_{11} across the transistor Q_{11} by means of equations (1) and (2):

$$V_{11} = \frac{V_{CE_{max}} + V_{CE_{sat_{max}}}}{I_2}$$

8. Select the value for R to satisfy the following condition:

$$R = \frac{V_{11} - V_1}{I_2}$$

9. The following equation for the output-voltage variation ΔV_o includes the equivalent transconductances g_{m1} of the series element and g_{m2} of the amplifier element:[4]

$$\Delta V_o = \frac{\Delta V_S + \frac{V_o}{R_{L_o}^2}\left(R_S + \frac{1}{g_{m_1}}\right)\Delta R_L}{1 + g_{m_2} R \frac{V_2}{V_o} + \frac{R_S}{R_{L_o}} + \frac{1}{R_{L_o} g_{m_1}}} \quad (6)$$

Because the number of stages in the series element is fixed, the value of g_{m1} can be determined from the following equation:[5]

$$g_{m_1} = 1 \div \left(\frac{1}{g_{m_{11}}} + \frac{1}{g_{m_{12}} h_{fe_{11}}} + \dots + \frac{1}{g_{m_{1n}} h_{fe_{11}} h_{fe_{12}} \dots h_{fe_{1(n-1)}}}\right) \quad (7)$$

where $g_{m_{1n}}$ is the small-signal transconductance of the Q_{1n} stage and $h_{fe_{1n}}$ is the small-signal current gain of the Q_{1n} stage. Both parameters are measured at the rated collector current for stage Q_{1n}.

A value of equivalent transconductance (g_{m_2}) for the entire dc amplifier element can then be found by substitution of fixed values for all other parameters in equation (6). This value of g_{m_2} governs the number of stages required to provide the desired output-voltage control.

10. Select the transistor for the first stage in the amplifier Q_{21}. The forward-bias common-emitter breakdown-voltage rating of the transistor should not be exceeded in the type of amplifier configuration chosen (see Fig. 3). Q_{21} is designed to operate at a collector current I_2 into a load R. The over-all small-signal voltage gain of the amplifier must be at least g_{m_2} x R. Because the bleeder current I_o through R_1 and R_2 is of the same order of magnitude as I_2 (usually $I_2 > I_o$), and I_{2n} is required to be much smaller than I_o, I_{2n} is also much smaller than I_2. Besides providing a voltage gain of g_{m_2} x R, therefore, the amplifier element must also provide a current gain.

[2] See Appendix, equation (A-14).
[3] See Appendix, equation (A-17) and (A-18).
[4] See Appendix, equation (A-5).
[5] See Appendix, equation (A-9).

11. Choose the amplifier type and number of stages. If temperature stability is required, a differential amplifier should be used at the input. When the maximum amount of gain per stage is desired, a reference diode should be used in the emitter circuit of each transistor rather than an emitter resistance.

12. For protection against overload, the following arrangements can be used:

(a) protection by limiting resistance. A resistance R_x placed in series with the source resistance R_S limits the current to a safe maximum value; the short-circuit current

$$I_{SC} = \frac{V_S}{R_S + R_x}.$$

(b) protection by current limiting. A reference diode V_D used in series with a resistance R_A (as shown in Fig. 4) supplies a

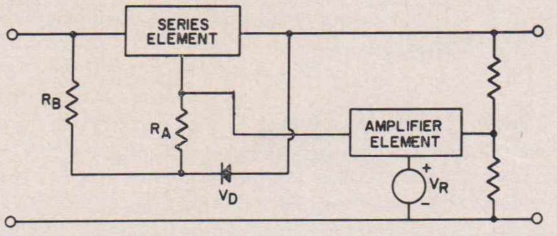


Fig. 4 - Circuit using current limiting to provide protection against overload.

constant current to the control input of the series element if the output voltage falls to some predetermined level. The output current is limited to the control current times the current gain of the series element.

Sample Design Problem

1. Conditions and Requirements:

V_o = 28 volts

V_S = 45 volts

R_S = 8 ohms

R_{L_o} = 42 ohms

I_{L_o} = 0.67 ampere

Chassis Temperature = 55° C

Circuit - shown in Fig. 1

ΔV_o = ±0.1 volt (requirement)

ΔV_S = ±5 volts

ΔR_L = ±14 ohms

$I_{L_{min}}$ = 0.5 ampere; $I_{L_{max}}$ = 1 ampere

2. Select a silicon reference diode which will supply a voltage V_R approximately equal to 0.5 V_o (in this case, use 12 volts) and have a power-dissipation capability of 400 milliwatts at 25 degrees centigrade or 300 milliwatts at 55 degrees centigrade. The maximum diode current $I_{VR_{max}}$ will then be 400/12, or 33 milliamperes, at 25 degrees centigrade, and 300/12, or 25 milliamperes, at 55 degrees centigrade.

3. Select V_2 and I_2.

(a) $V_2 = V_R + \frac{1}{2}(V_o - V_R) = 12 + \frac{1}{2}(28-12) = 20$ volts.

(b) $I_2 < I_1/20 = 0.75/20 = 37.5$ milliamperes

Select $I_2 = 10$ milliamperes

4. Select a series transistor Q_{11} having the following voltage ratings:

$$V_{CE_{max}} \geqq V_S + \Delta V_S - V_o \left(1 + \frac{R_S}{R_{L_o} + \Delta R_L}\right)$$

$$= 45 + 5 - 28\ (1 + 8/56)$$

$$= 50 - 1.14\ (28) = 50 - 32$$

$$= 18 \text{ volts}$$

$$V_{CE_{sat_{max}}} \leqq V_S - \Delta V_S - V_o \left(1 + \frac{R_S}{R_{L_o} - \Delta R_L}\right)$$

$$= 45 - 5 - 28\ (1 + 8/28)$$

$$= 40 - 1.29\ (28) = 40 - 36.2$$

$$= 3.8 \text{ volts}$$

at $I_C = (V_o)/(R_{L_o} - \Delta R_L) = 1$ ampere

$I_{C_{max}} > (V_o)/(R_{L_o} - \Delta R_L) = 28/28 = 1$ ampere

Therefore, $I_{C_{max}} \approx I_{1_{max}}$

$$\text{Because } \frac{V_S + \Delta V_S - V_o}{2\ R_S} = \frac{45 + 5 - 28}{2\ (8)} = 22/16$$

$= 1.37$ amperes $> I_{1_{max}}$, then

$$PD_{max} = (V_S + \Delta V_S - V_o)\ (I_{1_{max}}) - (I_{1_{max}})^2\ R_S$$

$$= (50 - 28) - (1.0)^2\ (8) = 22 - 8$$

= 14 watts at a case temperature of 55 degrees centigrade

The RCA-2N1489 silicon power transistor meets these requirements. The following design-center values can be obtained from the published data for the 2N1489:

I_C = 0.67 ampere h_{FE} = 50 V_{BE} = 0.8 volt

h_{fe} = 30 g_m = 5 mhos

5. Let $I_{1n} = I_2/10 = 10/10 = 1$ milliampere.

$$h_{FE_{11}} \cdot h_{FE_{12}} \cdot \; . \; . \; h_{FE_{1n}} = \frac{I_1}{I_{1n}} = \frac{0.67}{1.0 \times 10^{-3}} = 670$$

$$\frac{I_1}{I_{1n}} \cdot \frac{1}{h_{FE_{11}}} = \frac{670}{50} = 13.4$$

This requirement can be met by use of a single additional series-regulator stage having a minimum dc current gain $h_{FE_{12}}$ of 13.4. The current in transistor Q_{11} is then given by

$$I_{11} = \frac{I_1}{h_{FE_{11}}} = \frac{0.67}{50} =$$

0.0134 ampere or 13.4 milliamperes

An RCA-2N1481 satisfies this current requirement. The following design-center values can be obtained from the published data for the 2N1481 for a collector current of 13.4 milliamperes:

$V_{BE} = 0.7$ volt $\qquad h_{FE} = 40$

$g_m = 0.75$ mho $\qquad h_{fe} = 50$

6. Solve for V_1 as follows:

$$V_1 = V_{BE_1} + V_{BE_2} = 0.8 + 0.7 = 1.5 \text{ volts.}$$

7. The design-center value for V_{11} is given by

$$V_{11} = \frac{V_{11_{min}} + V_{11_{max}}}{2}$$

$$V_{11} = V_{S_o} - V_o - \frac{V_o R_S R_{L_o}}{R_{L_o}^2 - \Delta R_L^2} =$$

$$45 - 28 - \frac{(28)\ (8)\ (42)}{(42)^2 - (14)^2}$$

$$= 17 - \frac{9400}{1760 - 195} = 11 \text{ volts}$$

8. Solve for R using the following relation:

$$R = \frac{V_{11} - V_i}{I_2} = \frac{11 - 1.5}{10 \text{ ma}} = \frac{9.5}{10 \times 10^{-3}} = 950 \text{ ohms}$$

(Use a 1000-ohm resistor.)

9. Solve for g_{m2} as follows:

$$\Delta V_o = \frac{\Delta V_S + \frac{V_o}{R_{L_o}^2}\left(R_S + \frac{1}{g_{m1}}\right)\Delta R_L}{1 + g_{m2} R \frac{V_2}{V_o} + \frac{R_S}{R_{L_o}} + \frac{1}{R_{L_o} g_{m1}}}$$

where

$$g_{m1} = \frac{1}{\frac{1}{g_{m11}} + \frac{1}{g_{m12}\, h_{fe11}}} = \frac{1}{\frac{1}{5} + \frac{1}{0.75 \times 30}} =$$

$$\frac{1}{244} = 4.1 \text{ mhos}$$

$V_S = 5$ volts $\qquad R_S = 8$ ohms

$\Delta V_o = 0.1$ volt $\qquad V_B = 14$ volts

$V_o = 28$ volts $\qquad \Delta R_L = 14$ ohms

$R_{L_o} = 42$ ohms

Substitution of these values in the above equation produces the following result:

$$0.1 = \frac{5 + \frac{28}{1769}(8 + 0.24)\ 14}{1 + g_{m2}\frac{(1000)\ (14)}{28} + \frac{8}{42} + \frac{1}{(42)\ (4.1)}}$$

$$= \frac{5 + 1.87}{1 + 500\ g_{m2} + 0.19 + 0.06}$$

$$= \frac{6.87}{1.25 + 500\ g_{m2}} = 0.1$$

$$50\ g_{m2} = 6.65,\ g_{m2} = 0.132 \text{ mho (minimum)}$$

10. An RCA-2N1481 transistor should provide the collector current of 10 milliamperes for Q_{21}. The following design-center values can be obtained from the published data for the 2N1481 for this 10-milliampere current:

$g_m = 0.4$ mho $\qquad h_{FE_{min}} = 20$

Because this value of g_m is greater than the calculated g_{m2}, only one stage of amplification is required.

The circuit for the amplifier element is shown in Fig. 5. The values of resistors R_1 and

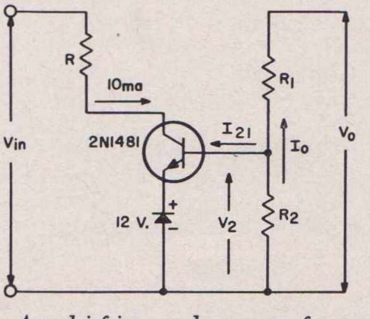


Fig. 5 - Amplifier element for series voltage regulator.

R_2 are determined as follows:

$$I_{21} = \frac{I_2}{h_{FE21}} = \frac{10 \times 10^{-3}}{20} = 0.5 \text{ milliampere} = 500 \text{ microamperes}$$

Because $I_o >> I_{21}$, select I_o = 10 milliamperes.

$$V_2 \approx V_R = 12 \text{ volts}$$

$$R_2 = \frac{V_2}{I_o} = \frac{12}{10} = 1200 \text{ ohms}$$

$$R_1 = \frac{V_o - V_2}{I_o} = \frac{28 - 12}{10} = \frac{16}{10} = 1600 \text{ ohms}$$

(Use 1500 ohms.)

The complete regulator circuit is shown in Fig. 6.

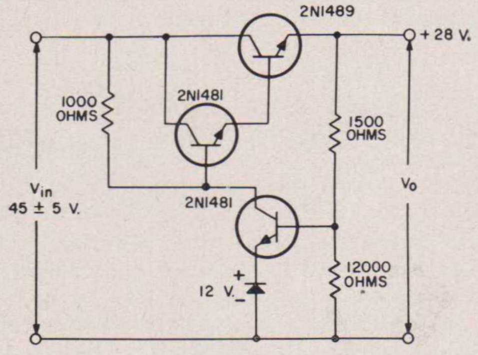


Fig. 6 - Complete circuit for transistorized series voltage regulator.

SHUNT VOLTAGE REGULATORS

Although shunt regulators are not as efficient as series regulators for most applications, they have the advantage of greater simplicity. The shunt regulator includes a shunt element and a reference-voltage element. The output voltage remains constant because the shunt-element current changes as the load current or input voltage changes. This current change is reflected in a change of voltage across the resistance R_1 in series with the load. A typical shunt regulator is shown in Fig. 7.

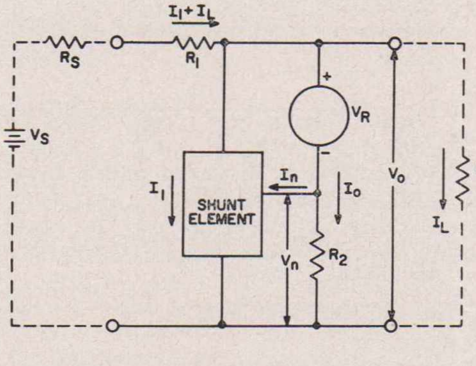


Fig. 7 - Typical transistorized shunt voltage regulator.

The shunt element contains one or more transistors connected in the common-emitter configuration in parallel with the load. A typical transistor shunt element is shown in Fig. 8.

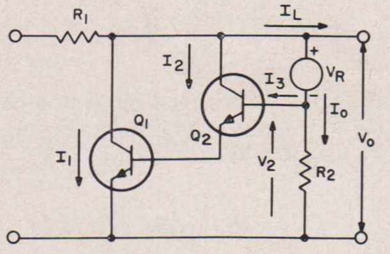


Fig. 8 - Typical transistor shunt element.

Design Procedure

The following step-by-step procedure is recommended for the design of transistorized shunt-type voltage regulators. Equations used in this procedure are derived in the Appendix.

1. List input requirements, load conditions, and output-voltage requirements in terms of the following parameters:
   - input voltage (V_S),
   - input-voltage variation (ΔV_S),
   - source resistance (R_S),
   - output load resistor (R_{L_o}),
   - output voltage (V_o), and
   - output-voltage variation (ΔV_o).

The terms V_S and R_{L_o} are design-center values; ΔV_S and ΔR_L are maximum deviations from these values.

2. Select a transistor type which will be within ratings for the following values of

V_{1max}, I_{1max}, and maximum dissipation PD_{1max} across the shunt element when both line and load regulation are required:

$$I_{1max} = I_{Lmax} = \frac{V_o}{R_{L_o} - \Delta R_L}$$

$$V_{1max} = V_o \text{ (under forward-bias conditions)}$$

$$PD_{1max} = I_{1max}\, V_{1max}$$

3. Select a value for resistance R_2 to provide a current I_o greater than the minimum value required to maintain the value of the reference voltage (i. e., to break down a voltage-reference diode, for example). The following relation may be used as a guide:

$$R_2 = \frac{n}{I_o}$$

where n is the number of stages in the shunt element.

4. The output resistance R_o of the regulator is given by

$$R_o = \frac{2\Delta V_o}{V_o} \cdot R_{L_o}$$

Assume a value of series resistance for the reference R_f; then the equation for output resistance can be solved for h_{fe}:[6]

$$R_o = \frac{R_f + \dfrac{h_{ie}\, R_2}{h_{ie} + R_2}}{1 + h_{fe}\dfrac{R_2}{R_2 + h_{ie}}} \qquad (8)$$

where $h_{fe} = h_{fe_1} \cdot h_{fe_2} \cdot \cdot \cdot h_{fe_{1n}}$, and h_{fe_n} and h_{ie} are the ac current transfer ratio and the input impedance, respectively, of the Q_n stage.

5. Determine I_n and V_n for the shunt element, as follows:

$$I_n = \frac{I_1}{h_{FE_1} \cdot h_{FE_2} \cdot \cdot \cdot h_{FE_{n-1}}}$$

where $h_{FE_{n-1}}$ is the dc current gain of the Q_{n-1} stage of the shunt element measured at a collector current of I_{n-1}.

$$V_n = V_1 + V_2 + \cdot \cdot \cdot V_{n-1}$$

where V_n is the base-to-emitter voltage of the Q_{n-1} stage at a collector current of I_{n-1}.

6. Select a voltage reference source which has a resistance less than the value that had been assumed for R_f (or recompute h_{fe} using a new value of R_f), a voltage $V_R = V_o - V_n$, a maximum current greater than $I_o + I_n$, and a maximum dissipation rating greater than $V_R\,(I_o + I_n)$.

7. Determine the value of series resistance R, including both source resistance R_S and external resistance R_1. R is dependent on the value of the input voltage V_S and its variation ΔV_S; R may also be expressed in terms of V_S, as follows:

$$V_S + \Delta V_S = V_o + R\,(I_{Lmax} + I_{1max})$$

$$V_S - \Delta V_S = V_o + R\, I_{1max}$$

For the usual case, $I_{1max} = I_{Lmax}$.

Sample Design Problem

1. Conditions and requirements:

Circuit - shown in Fig. 7

R_S = 10 ohms

R_{L_o} = 110 ohms

ΔR_L = ±55 ohms

V_S = 49 volts

ΔV_S = ±7 volts

V_o = 28 volts

ΔV_o = ±0.0125 volt

Maximum transistor case temperature = 55° C

2. $$I_{1max} = I_{Lmax} = \frac{V_o}{R_{L_o} - \Delta R_L} = 0.5 \text{ ampere}$$

$V_{1max} = V_o$ = 28 volts under forward-bias conditions

$PD_{1max} = V_{1max}\, I_{1max}$ = (28) (0.5) = 14 watts

Select the RCA-2N1485 to meet all the above requirements.

3. Select a voltage reference diode for V_R, and let I_o = 2 milliamperes. If two stages are used for the shunt element, the value of R_2 is given by

$$R_2 = 2/2 \times 10^{-3} = 1000 \text{ ohms.}$$

4. The output resistance R_o is given by

$$R_o = (2\Delta V_o/V_o)\,(R_{L_o}) = (0.025/28)\,(110) = 0.10 \text{ ohm}$$

[6] See Appendix equation (A-27).

If R_f has a maximum value of 5 ohms, and the input impedance h_{ie} of the Q_n stage has a typical value of 50 ohms, equation (8) may be solved for h_{fe} as follows:

$$0.10 = \frac{5 + \dfrac{(50)\ (1000)}{50 + 1000}}{1 + h_{fe}\ \dfrac{1000}{1000 + 50}}$$

$$0.1 + 0.095\ h_{fe} = 5 + 48 = 53$$

$$h_{fe} = \frac{52.9}{0.095} = 560$$

Consequently, two stages are required for the shunt element, with a product $h_{fe_1} \times h_{fe_2} = 560$. The 2N1485 selected in step 2 for the first stage Q_1 has the following design-center values:

$$I_c = I_1 = V_o/R_{L_o} = 250 \text{ milliamperes}$$

$$h_{fe_1} = 56$$

$$h_{FE_1} = 50$$

$$V_{BE} = 0.8 \text{ volt}$$

For the second stage Q_2, therefore, the following values are required:

$$h_{fe_2} = 560/56 = 10$$

$$I_c = I_1/h_{FE_1} = 250 \text{ ma}/50 = 5 \text{ milliamperes.}$$

Select an RCA-2N1481 transistor to meet these requirements. The following design-center values can be obtained from published data for the 2N1481 for a collector current of 5 milliamperes:

$h_{fe_2} = 20$ $\qquad$ $h_{ie} = 50$

$h_{FE_2} = 25$ $\qquad$ $V_{BE} = 0.7$ volt

This value of $h_{ie} = 50$ was determined by taking the slope of the typical V_{BE} vs I_B curve at the 5-milliampere collector-current operating point, where $I_B = I_C/h_{FE_2} = 5/25 = 0.2$ milliampere. If actual measurements indicate a different value from that assumed above in step 4, the new value is used in equation (8) and h_{fe} is recomputed.

5. The current I_2 and voltage V_2 are calculated as follows:

$$I_2 = \frac{T_1}{h_{FE_1} \cdot h_{FE_2}} = \frac{250}{50 \cdot 25} = 0.20 \text{ milliampere}$$

As listed in step 4, the base-to-emitter voltage of the 2N1485 for the design-center collector current of 250 milliamperes is 0.8 volt. For the 2N1481, the base-to-emitter voltage for the design-center collector current of 5 milliamperes is 0.7 volt. Therefore, V_2 is given by

$$V_2 = 0.8 + 0.7 = 1.5 \text{ volts.}$$

6. Select the RCA-1N1781 silicon voltage reference diode to meet the following design conditions:

$R_f = 5$ ohms

$V_R = 28 - 1.5 = 26.5$ volts

$I_{max} = I_o + I_2 = 2 + 0.2 = 2.2$ milliamperes

$PD_1 = 26.5\ (2.25) = 60$ milliwatts

7. Let $I_{1_{max}} = I_{L_{max}}$, and solve for R as follows:

$$49 - 7 = 28 + R\ (0.5)$$

$$42 = 28 + 0.5\ R$$

$$0.5R = 14$$

$$R = 28 \text{ ohms}$$

Therefore,

$$R_1 = R - R_S = 28 - 10 = 18 \text{ ohms.}$$

The complete circuit for the transistorized shunt voltage regulator is shown in Fig.9.

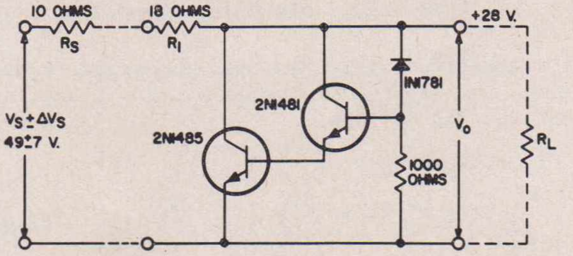


Fig. 9 - Complete circuit for transistorized shunt voltage regulator.

APPENDIX

Series Regulator Design Equations

Derivation of ΔV_o in terms of ΔV_S and ΔR_L

From the circuit of Fig. 1, the following assumptions can be made:

$$I_{1n} \ll I_2 \ll I_1$$
$$I_{1n} \ll I_o \ll I_1$$
$$I_1 \approx I_L$$

and the following parameters can be defined:

Total Series Element:

$$g_{M_1} = I_1/V_1; \quad g_{m_1} = \Delta I_1/\Delta V_1$$

Total Amplifier Element:

$$g_{M_2} = I_2/V_2; \quad g_{m_2} = \Delta I_2/\Delta V_2$$

The equation for V_o may be expressed as follows:

$$V_o = V_S - g_{M_2} R(V_2 - V_R) - I_1 (R_S + 1/g_{M_1}) \qquad \text{(A-1)}$$

As assumed above,

$$I_1 \approx I_L = V_o/R_L$$

Therefore,

$$V_o = V_{S_2} - g_{M_2} R(V_2 - V_R) - (V_o/R_L)(R_S + 1/g_{M_1}) \qquad \text{(A-2)}$$

This equation can be modified as follows to include incremental changes in V_2:

$$V_o + \Delta V_o = V_S + \Delta V_S - g_{M_2} R(V_2 - V_R) - g_{m_2} R(V_2/V_o)\,\Delta V_o \qquad \text{(A-3)}$$
$$- \left(\frac{V_o}{R_L + \Delta R_L} + \frac{V_o}{R_L + \Delta R_L}\right)\left(R_S + \frac{1}{g_{M_1}} + \frac{1}{g_{m_1}}\right)$$

If synthetic division is used and higher-order terms are neglected, equation (A-3) can be rewritten as follows:

$$V_o + \Delta V_o = V_S + \Delta V_S - g_{M_2} (V_2 - V_R) - g_{m_2} R(V_2/V_o)\,\Delta V_o - (V_o/R_L)(R_S + 1/g_{M_1}) - (\Delta V_o/R_L)(R_S + 1/g_{m_1}) + (V_o/R_L^2)(R_S + 1/g_{m_1})\,\Delta R_L \qquad \text{(A-4)}$$

All the non-constant terms in this equation can then be combined and the equation solved for ΔV_o:

$$\Delta V_o = \frac{\Delta V_S + (V_o/R_L^2)\,(R_S + 1/g_{m_1})\,R_L}{1 + g_{m_2} R(V_2/V_o) + (R_S/R_L) + (1/R_L g_{m_1})} \qquad \text{(A-5)}$$

Equations for Series Regulator Elements

For the series element shown in Fig. 10, the voltage V_1 is given by

$$V_1 = V_{BE}\,(Q_{11} \text{ at } I_c = I_1) + V_{BE}\,(Q_{12} \text{ at } I_c = I_{11}) + \ldots + V_{BE}\,(Q_{12} \text{ at } I_c = I_{1(n-1)}) \qquad \text{(A-6)}$$

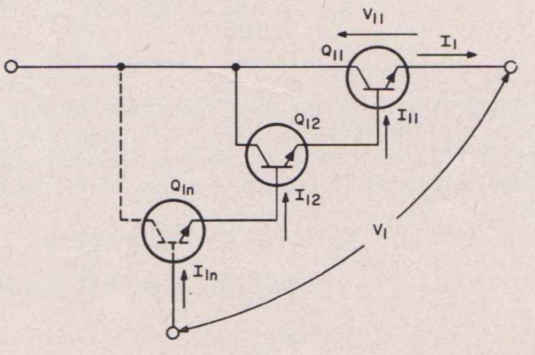


Fig. 10 - Typical series regulator element.

Incremental changes in V_1 and I_1 are expressed as follows:

$$\Delta V_1 = \frac{\Delta I_1}{g_{m_{11}}} + \frac{\Delta I_{11}}{g_{m_{12}}} + \ldots + \frac{\Delta I_{1(n-1)}}{g_{m_{1n}}} \qquad \text{(A-7)}$$

where $g_{m_{1n}}$ is the ac transconductance of the Q_{1n} stage measured at a collector current of $I_{1(n-1)}$.

The current variations in the different stages are given by

$$\Delta I_{11} = \frac{\Delta I_L}{h_{fe_{11}}}, \quad \Delta I_{1n} = \frac{\Delta I_{1(n-1)}}{g_{m_{1n}}}, \text{ etc.}$$

where h_{fe} is the ac current transfer ratio for the Q_{1n} stage at a collector current of $I_{1(n-1)}$.

The voltage V_1 may then be expressed as follows:

$$V_1 = \frac{\Delta I_1}{g_{m_{11}}} + \frac{\Delta I_1}{h_{fe_{11}}\, g_{m_{12}}} + \ldots + \frac{\Delta I_1}{\left[h_{fe_{11}}\, h_{fe_{12}} \cdots h_{fe_{1(n-1)}}\right] g_{m_{1n}}} \qquad \text{(A-8)}$$

Because $g_{m_1} = \Delta I_1/\Delta V_1$, the terms in equation (A-8) may be combined as follows to solve for g_{m_1}:

$$g_{m_1} = 1 \div \left(\frac{1}{g_{m_{11}}} + \frac{1}{g_{m_{11}}\ h_{fe_{12}}} + \ldots + \frac{1}{\left[h_{fe_{11}} \cdot h_{fe_{12}} + \ldots + h_{fe_{1(n-1)}}\right] g_{m_{1n}}} \right) \qquad \text{(A-9)}$$

The maximum and minimum input voltages to the series regulator are defined as follows:

$$V_{i_{max}} = V_{S_o} + \Delta V_S - \frac{V_o}{R_{L_o} + \Delta R_L} \cdot R_S \qquad \text{(A-10)}$$

$$V_{i_{min}} = V_{S_o} - \Delta V_S - \frac{V_o}{R_{L_o} - \Delta R_L} \cdot R_S \qquad \text{(A-11)}$$

The collector-to-emitter voltage ratings for all the transistor stages in the series element under forward-bias conditions must be greater than $V_{11_{max}}$, where

$$V_{11_{max}} = V_{i_{max}} - V_o = V_{S_o} + \Delta V_S - V_o \left(1 + \frac{R_S}{R_{L_o} + \Delta R_L}\right) \qquad \text{(A-12)}$$

The saturation-voltage ratings for Q_{11} must be less than $V_{11_{min}}$ at a collector current of $I_{1_{max}}$:

$$V_{11_{min}} = V_{i_{min}} - V_o = V_{S_o} - \Delta V_S - V_o \left(1 - \frac{R_S}{R_{L_o} - \Delta R_L}\right) \qquad \text{(A-13)}$$

and

$$I_{1_{max}} = \frac{V_o}{R_{L_o} - \Delta R_L} \qquad \text{(A-14)}$$

The power dissipation PD of Q_{11} is given by

$$PD_{11} = I_1 \cdot V_{11} = I_1\ (V_S - V_o - I_1 - R_S)$$
$$= (V_S - V_o)\ I_1 = I_1^2\ R_S$$

$PD_{11_{max}}$ occurs when $dPD/dI_1 = 0$, and

$$V_S = V_{S_o} + \Delta V_S$$

$$dPD/dI_L = d/dI_1 \left[(V_{S_o} + \Delta V_S - V_o)\ I_1 - I_1^2\ R_S\right]$$
$$- V_S + \Delta V_S - V_o - 2I_1\ R_S = 0$$

Therefore,

$$PD_{11_{max}} \text{ occurs at } I_1 = \frac{V_S + \Delta V_S - V_o}{2\ R_S} \qquad \text{(A-15)}$$

$$PD_{11_{max}} = (V_S + \Delta V_S - V_o\ ) \frac{(V_S + \Delta V_S - V_o)}{2\ R_S} - \frac{R_S\ (V_S + \Delta V_S - V_o^2)}{2\ R_S} \qquad \text{(A-16)}$$

Equation (A-16) can be simplified as follows:

$$\text{If } \frac{V_S + \Delta V_S - V_o}{2\ R_S} \leq I_{1_{max}},$$

$$\text{then } PD_{11_{max}} = \frac{(V_S + \Delta V_o - V_o)^2}{4\ R_S} \qquad \text{(A-17)}$$

$$\text{If } \frac{V_S + \Delta V_S - V_o}{2\ R_S} \geq I_{1_{max}},$$

$$\text{then } PD_{11_{max}} = (V_S + \Delta V_S - V_o)\ (I_{1_{max}}) - I_{1_{max}}^{\ 2}\ R_S \qquad \text{(A-18)}$$

Derivation of Expressions for Cascaded DC Amplifier Elements

In the cascaded dc amplifier shown in Fig. 11, the current fluctuation ΔI_2 is given by

$$\Delta I_2 = \Delta V_{22}\ g_{m_{21}}$$

where

$$\Delta V_{22} = (V_2/V_{22}) \cdot \Delta V_2\ (Av_{22} \cdot Av_{23} \cdot \cdot \cdot Av_{2n})$$

Av_{2n} is the voltage gain for the nth stage.

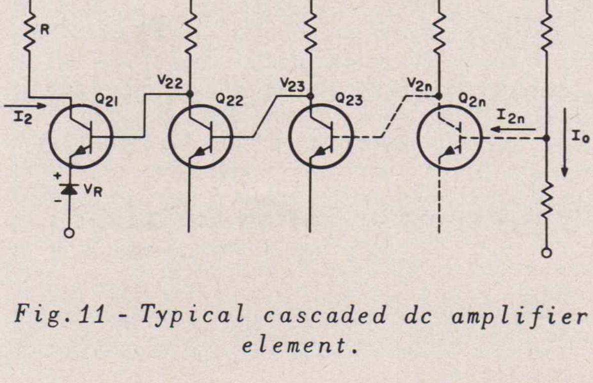


Fig. 11 - Typical cascaded dc amplifier element.

The ac transconductance g_{m_2} for the entire amplifier element is given by

$$g_{m_2} = \Delta I_2/\Delta V_2$$

Therefore,

$$g_{m_2} = (g_{m_{21}}/\Delta V_2)\ (V_2/V_{22})\ (\Delta V_2)\ (Av_{22} \cdot Av_{23} \cdot \cdot \cdot Av_{2n}) \qquad \textbf{(A-19)}$$

$$g_{m_2} = (g_{m_{21}})\ (V_2/V_{22})\ (Av_{22} \cdot Av_{23} \cdot \cdot \cdot Av_{2n}) \qquad \textbf{(A-19a)}$$

Both n-p-n and p-n-p types of transistors can also be connected in tandem to produce high gain and a high degree of temperature stability, as shown in Fig. 12. In this circuit,

$$\Delta I_{c_{2(n-1)}} = \Delta V_2 \times g_{m_{2n}} \qquad \textbf{(A-20)}$$

$$\Delta I_2 = \Delta I_{c_{2(n-1)}}\ (h_{fe_{2n}} \cdot \cdot \cdot h_{fe_{23}} \cdot h_{fe_{22}} \cdot h_{fe_{21}}) \qquad \textbf{(A-20a)}$$

Equations (A-20) and (A-20a) can be combined as follows:

$$\Delta I_2 = g_{m_{2n}}\ (h_{fe_{2n}} \cdot \cdot \cdot h_{fe_{23}}\ h_{fe_{22}} \cdot h_{fe_{21}})\ \Delta V_2 \qquad \textbf{(A-20b)}$$

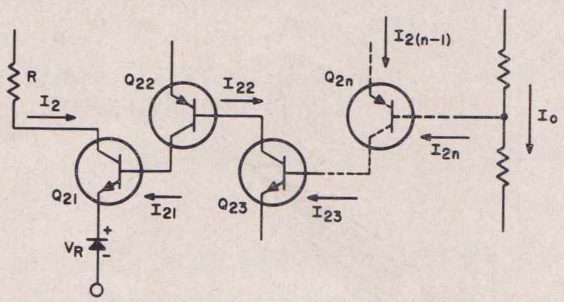


Fig. 12 - Tandem connection of transistors used to produce high gain and a high degree of temperature stability.

This equation can then be solved for g_{m_2}, as follows:

$$g_{m_2} = g_{m_{2n}}\ (h_{fe_{2n}} \cdot \cdot \cdot h_{fe_{23}} \cdot h_{fe_{22}} \cdot h_{fe_{21}}) \qquad \textbf{(A-21)}$$

Shunt Regulator Design Equations

Derivation of Output Resistance

The output resistance for a single-stage shunt regulator such as that shown in Fig. 13 is determined as follows: First, the voltage reference V_R is replaced by a battery and internal resistance R_f. The current variations for the ac circuit are then given by

$$\Delta I_2 = (R_2/R_2 + h_{ie})\ \Delta I_o$$

$$\Delta I_1 = (R_2/R_2 + h_{ie})\ \Delta I_o\ h_{fe} \qquad \textbf{(A-22)}$$

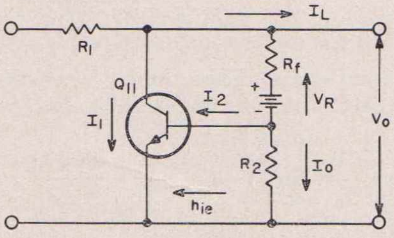


Fig. 13 - Typical single-stage shunt regulator.

The variation in I_o is given by

$$\Delta I_o = \frac{\Delta V_o}{R_f + \dfrac{R_2\ h_{ie}}{R_2 + h_{ie}}} \qquad \textbf{(A-23)}$$

Equations (A-22) and (A-23) can be combined as follows:

$$\Delta I_1 = \frac{R_2\ h_{ie}}{R_2 + h_{ie}} \times \frac{V_o}{R_f + \dfrac{R_2\ h_{ie}}{R_2 + h_{ie}}} \qquad \textbf{(A-24)}$$

The output resistance R_o is then defined as follows:

$$R_o = \Delta V_o/(\Delta I_1 + \Delta I_o) \qquad \textbf{(A-25)}$$

$$R_o = \frac{V_o}{\dfrac{V_o R_2\, h_{fe}}{(R_2 + h_{ie})\ \dfrac{(R_f + R_2\ h_{ie})}{(R_2 + h_{ie})}} + \dfrac{V_o}{R_f + \dfrac{R_2\ h_{ie}}{R_2 + h_{ie}}}} \qquad \textbf{(A-26)}$$

$$R_o = \frac{R_f + \dfrac{R_2\ h_{ie}}{R_2 + h_{ie}}}{\left(1 + \dfrac{R_2}{R_2 + h_{ie}}\right)(h_{fe})} \qquad \textbf{(A-27)}$$

THE TR24

A 24-watt Transistorized High Fidelity Amplifier

Introduction

This amplifier uses a new type of drift power transistor which is just becoming available in Australia, a type which to date offers the best answer to the audiophile seeking transistors for his equipment. The transistor is the 2N2147, which is a p-n-p diffused collector graded base type, designed primarily for high power audio amplifier applications where wide frequency response and good linearity are required. The 2N2147 is particularly well suited for use in low distortion class B amplifiers because of its excellent Beta linearity, low intrinsic base resistance and high cut-off frequency.

The 2N2147 uses the JEDEC TO-3 outline, which is mechanically very similar to the familiar outline of the 2N301 and similar types. With a maximum collector to emitter voltage rating of 40 volts with emitter conducting, and a maximum collector to base voltage rating of 50 volts with emitter cut off, the 2N2147 is rated for a maximum collector current of 10 amperes and a maximum collector dissipation of 12.5 watts at a mounting flange temperature of 80°C. The dc forward current transfer ratio with a collector current of 1 ampere and a collector to emitter voltage of —2 volts, is 150. The alpha cut-off frequency is 50 Kc.

This article describes the first of a series of amplifiers using these new transistors. This unit is a nominal 24-watt stereo unit (12 watts per channel), mains driven, with preamplifier for ceramic or crystal pickup.

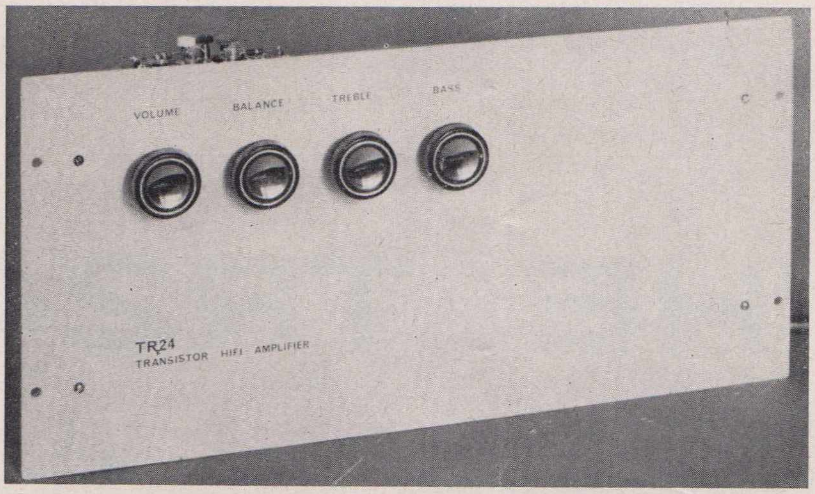
Acknowledgment

This amplifier is based on one of a series developed by Messrs. H. Kleinman and C. Wheatley, of the RCA Semiconductors and Material Division at Somerville, N.J., using both the 2N2147 and the 2N2148, a similar but lower dissipation type. These amplifiers were developed to exploit the possibilities of the new types, and the interest in them made us decide to develop Australian versions of them, using local parts. In this way the use of these amplifiers could be opened up for local enthusiasts. It is therefore with full acknowledgment to the work of Kleinman and Wheatley, and to RCA, that this material is presented here.

The Amplifier

The development of the drift power transistor has made possible the construction of high performance audio amplifiers which can equal the performance of the best valve amplifiers. The advantages of transistorized equipment are by now so well known as hardly to need repetition. They include low overall dissipation in the equipment, largely due to the absence of heater requirements, instant play, freedom from microphonics and some types of hum, greater ruggedness and smaller size, among others. But until the introduction of these drift power transistors, the benefits enumerated have had to be obtained at the sacrifice of performance or at a significantly higher cost.

The drift-field power transistor, with a current gain flat to 50 Kc, has therefore opened the door to economical high performance audio amplifiers.



View of the completed TR24.

The current gain of the transistor is linear to 5 amperes or more, permitting the achievement of low distortion figures. Further, with this linearity at high currents, it is possible to couple a low-impedance loudspeaker system directly to the amplifier without the need for an output transformer.

The output stage in each channel consists of two 2N2147's in a single-ended push-pull arrangement. These two transistors are transformer-driven by a 2N591 in each channel, preceded by a 2N408. Negative feedback is provided over the complete main amplifier from the output or loudspeaker connection point to the base of the 2N408 predriver stage.

The load impedance of the amplifier is 16 ohms, and it should be noted that because there is no output transformer, one of the loudspeaker leads in each channel will be live. Caution must therefore be exercised to see that this lead is not shorted to ground.

The Power Supply

The power supply in the original design was used at first, but although the hum level was low, it seemed that some improvement could be made. Here, of course, we are using 50 cps power supplies, which would slightly worsen the results compared with a 60 cps supply.

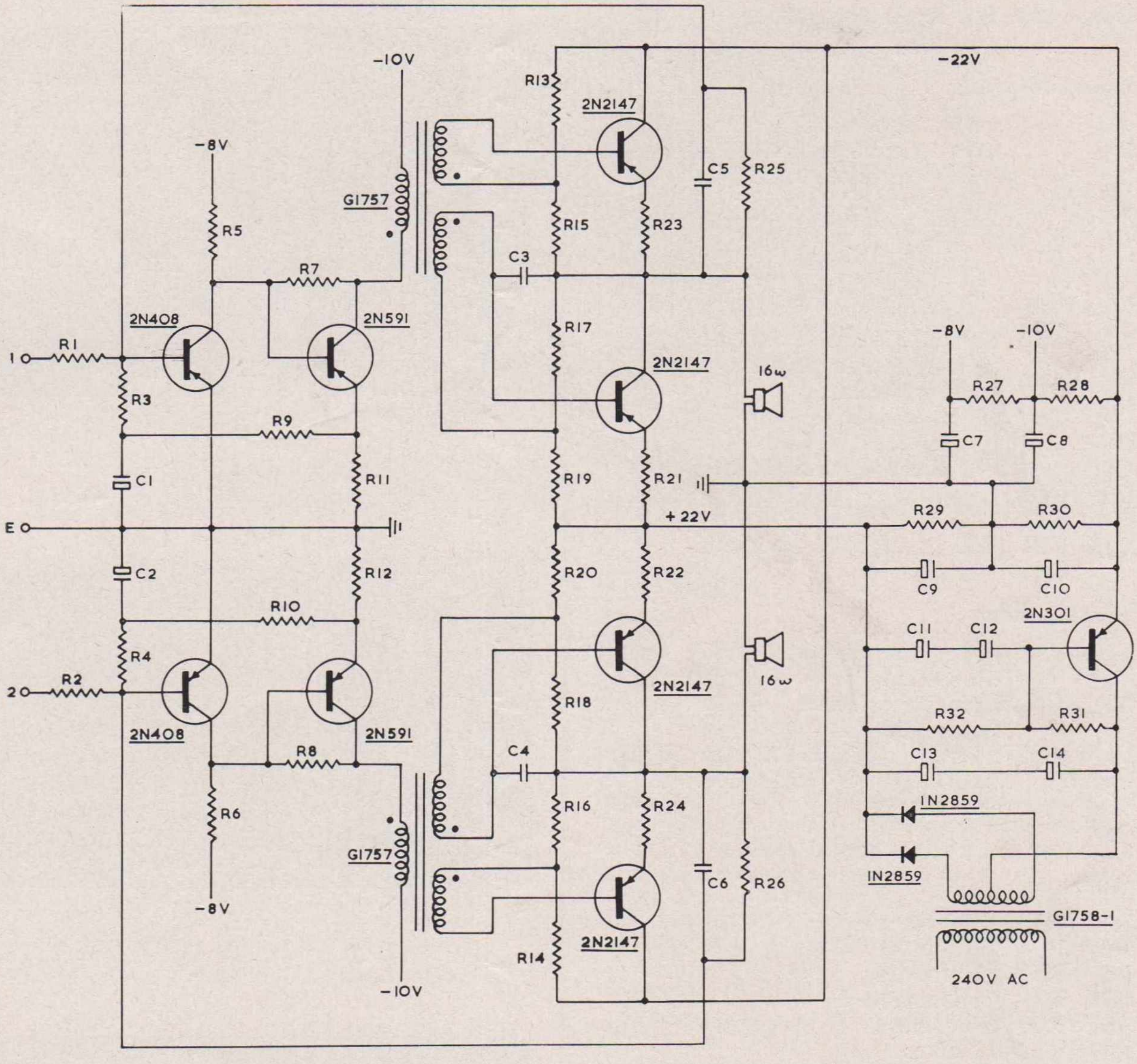
A completely new power supply was therefore evolved, using an electronic filter or "capacity multiplier." Before describing the arrangement, it may be appropriate here to state our standards of acceptability for hum level in an audio amplifier. There are, of course, measurements that can be made of the hum and noise output referred to the signal output under full gain conditions. These figures need to be treated rather critically, however, as they are usually taken into a resistive load.

Further, with amplifiers having a fairly high output level, even hum and noise figures which appear to be very good can still represent an appreciable, and audible, hum and noise output. A subjective test is therefore always made. This consists of setting up the amplifier with one or two vented enclosure loudspeakers, as the case may be, and running the amplifier with no input in completely silent surroundings. This test is usually applied late at night in a quiet suburban living room. Under these conditions it is expected that there will be no audible hum and noise at one foot or more from either speaker.

As readers will realise, this test, whilst perhaps not very scientific and entirely subjective, is in fact the acid test as far as hum and noise are concerned. It is surprising, reverting to the remarks above on measured hum and noise levels, how little hum and noise output will be audible under these conditions. To get back to the amplifier we are discussing in this article, the unit as now presented produced no audible hum under these conditions, and the barest perceptible background noise was detected with the ear pressed close to the front of the loudspeaker. This test was carried out under conditions in which the mains transformer lamination hum could be detected at the amplifier itself, i.e., under extremely quiet conditions. This test is mentioned here not as a substitute for normal testing, but because it tells the amplifier's story in a very simple way, requires no external equipment, and is therefore easily carried out by anyone.

Returning to the power supply, to provide the correct dc connection conditions for the loudspeakers, the power supply requirement is a 44-volt collector supply voltage with the centre point (+22 volts) grounded; i.e., voltages of +22 and −22 about ground are required.

The negative side of the power supply also feeds the driver and preamplifier stages through



Circuit diagram, TR24 main amplifier.

suitable voltage-dropping resistors. In the original design, a full-wave rectifier system was followed by an LC pi filter, the output of which fed the output channels and the driver stages. An electronic filter was then used to smooth the supply for the predriver and preamplifier stages. In the new design shown here, the entire output of the power supply is electronically filtered, and it is estimated that the cost of the power supply in each case would be about the same.

The capacitance multiplier or electronic filter is a very useful device to obtain efficient smoothing in cases such as this where the impedance of the power supply is so low. The problem of smoothing low impedance power supplies is being met more frequently as the quantity of transistorized equipment increases, and is perhaps therefore worth a few words. As most readers will know, it is basically true to say that the lower the impedance of a power supply, the larger the values of capacitor required to provide a certain amount of smoothing. This follows from the fact that the impedance of the capacitor used must be appreciably lower than the impedance of the power supply to have any worthwhile effect. This raises the bogey of enormous values of smoothing capacitor, with attendant size and cost.

In many cases, and it applies here, the answer is the electronic filter, in which a conventional-sized capacitor can be used. In this system, the

Parts List for Main Amplifier

Semiconductors

4	AWV transistors	2N2147
2	AWV transistors	2N408
2	AWV transistors	2N591
1	AWV transistor	2N301
2	AWV diodes	1N2859

Capacitors

C1, C2	200 μf, 3 V.W. electrolytic (2 x 100 μf in parallel)
C3, C4	0.005 μf, ceramic
C5, C6	33 ρf, ceramic
C7	100 μf, 12 V.W. electrolytic
C8	500 μf, 18 V.W. electrolytic
C9, C10, C11, C12, C13, C14	1000 μf, 25 V.W. electrolytic

Transformers

T1, T2	Driver transformer. M.S.P. sample No. G1757
T3	Mains transformer. M.S.P. sample No. G1758-1 Primary 0-220-240 volts 50 cps. Secondary 75 volts C.T. at 1.5 amperes.

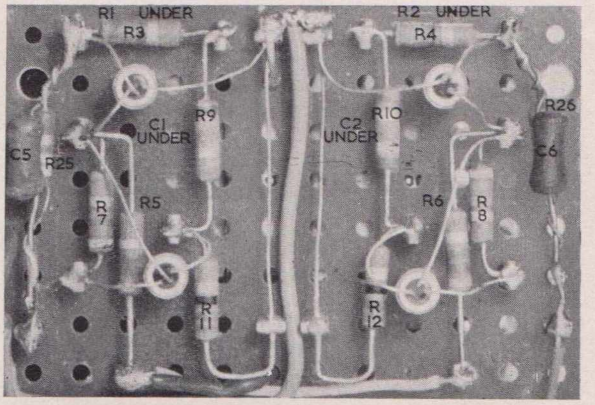
Resistors

R1, R2	1800 ohms
R3, R4	2700 ohms
R5, R6	8200 ohms
R7, R8	47K ohms
R9, R10	680 ohms
R11, R12	22 ohms
R13, R14	220 ohms, 5 watt, W.W.
R15, R16	2.7 ohms
R17, R18	220 ohms, 5 watt, W.W.
R19, R20	2.7 ohms
R21, R22, R23, R24	0.68 ohm
R25, R26	390K ohms
R27	330 ohms, 1 watt
R28	330 ohms, 2 watt
R29, R30	820 ohms, 1 watt
R31	330 ohms, 2 watt
R32	47K ohms, 1 watt

All resistors ½ watt rating, 10% tolerance except where stated.

Miscellaneous

Aluminium plate, matrix board and pins, type 7000 heat sinks, aluminium angle, hardware.



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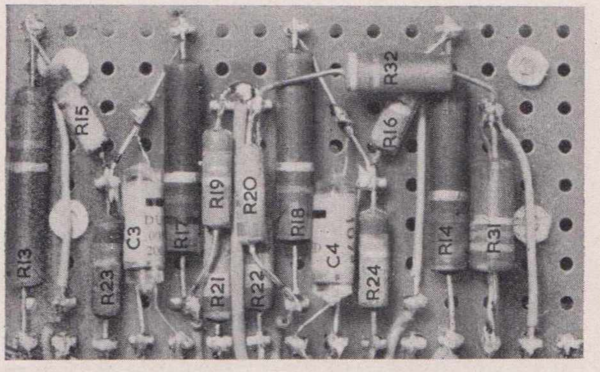

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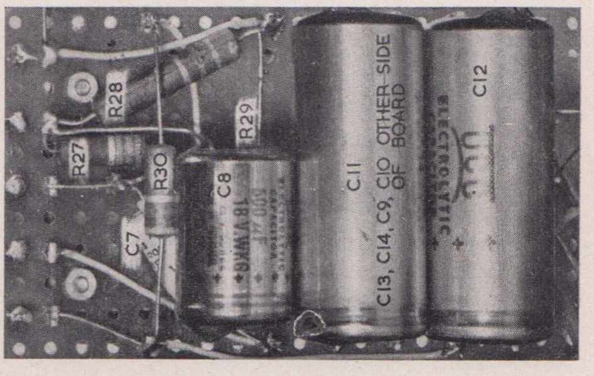
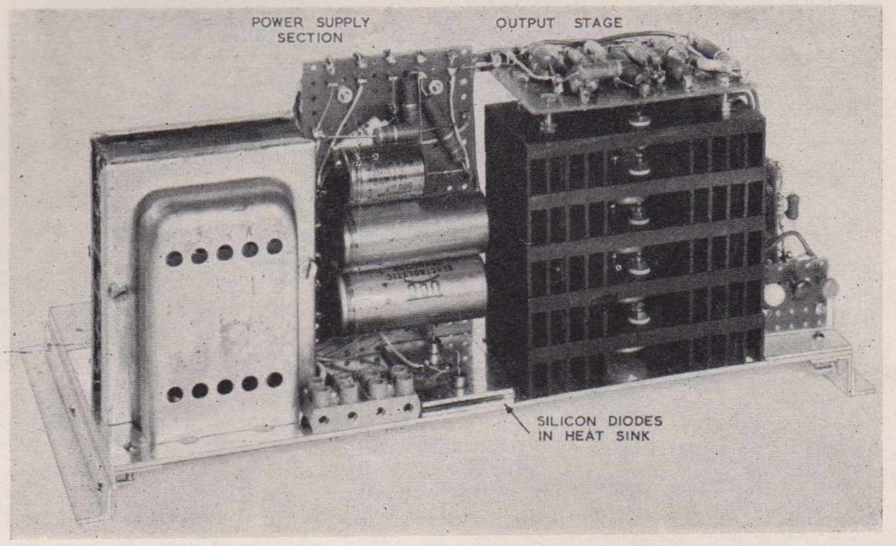


C

Component layouts in (A) driver and predriver section, (B) output stages, and (C) power supply components.



View of the TR24 main amplifier.

effective capacitance available is the physical value of the component used, multiplied by the current gain of the transistor. For example, in this case, a 2N301 is used as the capacity multiplier, and the physical capacitance provided is 500 microfarads. Across the output of the filter, that is, between the emitter and the positive bus of the power supply, the effective capacitance is 500 multiplied by the current gain of the 2N301. The current gain of the transistor here is of the order of 66 times, the effective capacitance across the output of the filter is of the order of 33,000 microfarads. The result speaks for itself.

In the circumstances, it may seem paradoxical to place another 500 microfarads across the output of the filter. There is a reason for this. The use of the electronic filter effectively increases the impedance of the power supply. The variation in current drawn from the power supply is from about 400 ma at no signal to about 1.5 amperes at maximum signal. This is a ratio of nearly 4:1, very high to those used to thermionic valve work, and too high in relation to the new value of power supply impedance. The addition of the extra 500 microfarads overcomes this difficulty.

Construction

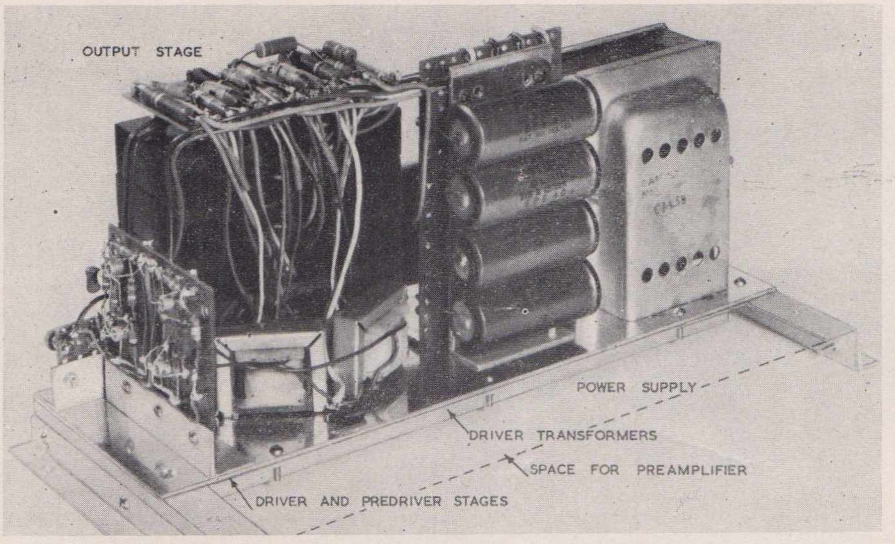
This is an unusual amplifier, and an unusual method of construction was used. But although unusual, construction is very simple. Most of the general details can be seen from the accompanying photographs and diagrams.

The complete amplifier, with preamplifier, is constructed on a 6″ x 12″ 14-gauge aluminium plate which forms the front of the amplifier proper. To the front of this plate is later fixed a front panel carrying the control labels and unit mounting holes. In this way a clean front appearance is presented, and the unit can be fitted in a modified rack fashion, either in an individual case or with other equipment.

The four 2N2147 transistors and the 2N301 are mounted on five type 7000 4″ x 2″ finned aluminium heat sinks. Colour-coded leads are connected to each transistor and the five units are then stacked up as shown in the illustrations. The matrix board carrying components for the output stages is then mounted on top of the stack and the wiring completed. This method shows a great saving of space and is justified with the great reliability of transistors.

The matrix board assembly with the driver and predriver stages is completely assembled before being mounted on the main amplifier, interconnecting leads to the power transistors and driver transformers being left long enough to reach when the unit is finally mounted. A further section of matrix board carries the power components, filter capacitors and a few miscellaneous parts. This unit also can be assembled before mounting on the amplifier.

It will thus be seen that the method of assembling the main amplifier is very simple. In addition, it results in a great saving of space and provides maximum cooling potential for the power transistor heat sinks. It will not be denied that a little dexterity is needed in making the



A further view of the TR24 main amplifier.

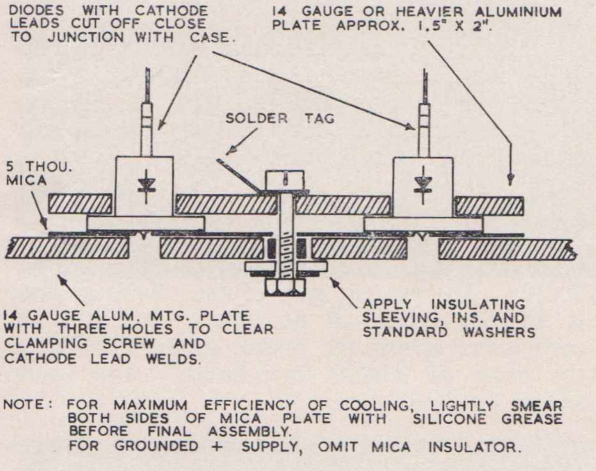
interconnections between the driver assembly and the power transistors and driver transformers. These connections should be made before the preamplifier is mounted, so that the unit can be worked on from both sides. Given this, and the fact that a small pencil-type soldering iron is a "must" for this type of unit anyway, there will be no difficulty.

There will no doubt be some readers who would like to vary the mechanical arrangement of the amplifier. In general, the layout does not appear critical, always with the usual precautions, so that some latitude is available. Adequate heat sinks must be provided for the power transistors and the filter transistor, and reasonably good construction techniques will be required throughout. It is felt that this is not an amplifier for the beginner.

Performance

Because we are dealing this month only with the main amplifier section of the unit, remarks on performance will in general be restricted to that section. The frequency response of the main amplifier is flat from 20 cps to 25 Kc within ± 1 db, 2 db down at 30 Kc and 8 db down at 40 Kc. The response in the supersonic region is quoted to show that there is no undesirable rise of response in that area.

The power output of the amplifier is 12 watts per channel, measured with sine wave input at 1 Kc. This output is obtained at better than 1% total harmonic distortion. In fact, on the model, the total harmonic distortion for full output was measured at 0.6%. The noise figure is better than —60 db. Full output is obtained for a signal input of 750 millivolts.



Method of arranging the two silicon diodes with a common positive connection, and heat sinking them.

Summary

It will be seen from the data so far presented that the promise of a transistorized high fidelity amplifier with characteristics comparable with some of the best valve amplifiers has been fulfilled. Because of the simple mains transformer used and the absence of two expensive output transformers, the cost of building this amplifier would bear favourable comparison with the cost of a similar valve amplifier, whilst the use of transistors gives a useful advantage in the matter of size and heat dissipation.

Next month we will deal with a suitable preamplifier to use with this amplifier, intended for use with a crystal or ceramic cartridge.

NEW RELEASES

2N699

The 2N699 is a silicon n-p-n triple-diffused planar transistor, intended for a wide variety of small-signal applications in military and industrial equipment. The 2N699 has a minimum gain-bandwidth product of 50 Mc; low output capacitance, low saturation voltages, high breakdown voltage (BV_{CBO}) of 120 minimum volts at $I_C=0.1$ ma, and utilizes the JEDEC TO-5 package.

NEW GERMANIUM SWITCHING TRANSISTORS

RCA now brings the mechanical ruggedness of its four-poster, double-welded-base tab structure to twelve medium-speed germanium n-p-n and p-n-p switching transistors. This tough construction — introduced in the Premium 3907/2N404 Switching Transistor—has now been applied to these new types: 2N388, 2N388-A, 2N1605, 2N1605-A, 2N1302, 2N1303, 2N1304, 2N1305, 2N1306, 2N1307, 2N1308 and 2N1309. All twelve types utilize the JEDEC TO-5 package with base internally connected to case. The p-n-p types 2N1303, 2N1305, 2N1307, 2N1309 and their n-p-n complements types 2N1302, 2N1304, 2N1306, 2N1308 feature a high collector-current rating of 300 milliamperes. The two n-p-n types 2N1605 and 2N1605-A are complements to the industry preferred p-n-p types 2N404 and the 2N404-A. Two n-p-n types, 2N388 and 2N388-A, feature a high operating (junction) temperature of 100°C and a high typical alpha-cutoff frequency of 14 Mc.

These 12 types provide the equipment designer with a wide range of characteristics values for greater design flexibility, and are tested in strict accordance with military specification MIL-S-19500B.

2N2205, 2N2206

These types are two new very high-speed switching types, both silicon n-p-n double-diffused planar epitaxial transistors. The 2N2205 and the 2N2206 are especially suited for use in industrial and military equipment requiring high reliability and high packaging densities. The 2N2205 is electrically identical to the 2N1708, but utilizes the popular JEDEC TO-18 package. The 2N2206 is mechanically identical and electrically similar to the 2N1708, but has a higher minimum beta (h_{FE}) of 40. It is intended for use in saturated-switching applications where a transistor having high minimum beta, short storage time, and extremely small size (JEDEC TO-46 package) are primary design requirements.

2N2270

The 2N2270 is a silicon triple-diffused planar transistor for small-signal and medium-power applications in industrial and military equipment.

The planar construction ensures exceptionally low noise and low leakage characteristics, whilst the triple-diffused junction design ensures high

EEV FX294

EEV announce the introduction of a new triode hydrogen thyratron, the FX294, of extremely rugged construction. It has a short recovery time and is therefore suitable for pulse operation at repetition frequencies greater than 10,000 cps. The maximum anode current is 60 ma mean and 35 amps peak. The valve is capable of withstanding a shock of 40g for 10msec., a sustained acceleration of 20g for three minutes, and a vibration acceleration rising from 3g at 20 cps to 11g to 50 cps, then levelling to 20g from 150 cps to 5 Kc. Of all glass construction, the valve is equipped with flying leads for direct connection to associated components; mounting is achieved by means of clamps round the bulb.

4037

The 4037 is a pencil-type high-mu, octal-based triode. This type features the coaxial pencil-type structure with an octal base and a large cathode cylinder. The 4037 replaces the 2C40 in most applications. The 4037 has an amplification factor of 56 and is intended for use in cathode-drive or grid-drive service as a CW local oscillator up to 3500 Mc, an rf power amplifier or mixer up to 1500 Mc, or a frequency multiplier up voltage and dissipation ratings. The 2N2270 can operate at junction temperatures up to 200°C. It has a minimum gain bandwidth product of 60 Mc, useful in applications from dc to 20 Mc. to 2000 Mc. It can be operated at altitudes up to 100,000 feet without pressurization. The 4037 can deliver a useful power output of 0.1 watt as an unmodulated class C oscillator at 3000 Mc, and 5 watts as an unmodulated class C power amplifier at 500 Mc.

6263A

The 6263A is a pencil-type medium-mu triode for military and critical industrial applications. This A-version retains the desirable characteristics of its prototype and, in addition, is designed to meet special tests for low-pressure breakdown, low- and high-frequency vibration, and 1-hour stability life performance. The 6263A has an amplification factor of 27, and is intended for use particularly as an rf power amplifier and oscillator in mobile equipment and in aircraft transmitters. It can be operated at altitudes up to 60,000 feet without pressurization. The 6263A can be operated with full ratings at frequencies up to 500 Mc, and with reduced ratings at frequencies as high as 1700 Mc.

7905

The 7905 is a quick-heating beam power valve of the 9-pin miniature type, designed primarily for use in mobile and emergency-communications equipment. In such equipment, the 7905 is particularly useful in radio-frequency power amplifier, oscillator and frequency-multiplier service at frequencies up to 175 Mc. Features which contribute to the exceptional performance of this new beam power valve at very high frequencies are 18 watts CW input at 175 Mc, and 7 watts useful power output (ICAS) at 175 Mc.

STOP PRESS

Transistors type 2N2147, 2N2148 are now in local production. The local product uses the same case as the 2N301 instead of the TO-3 case; mounting details are the same for both cases.

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