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12-VOLT AUDIO AMPLIFIER AND CONVERTER DESIGNS USING SILICON TRANSISTORS

By: W. D. WILLIAMS and E. F. DUDECK

INTRODUCTION

This Note describes the application of types 40250 and 40251 diffused-junction, n-p-n, silicon transistors in 12-volt (mobile and portable) circuits. The 40250, specifically designed for the output stage of medium-power amplifiers, may also be used in dc-to-dc converter and dc-to-ac inverter applications. The 40251 was specifically designed for 12-volt inverter and converter applications but may also be used as a high-level audio output amplifier or as a voltage regulator.

The superior frequency response and higher safe operating region plus the mass production potential of these units makes them suitable for many applications where silicon transistors had not previously been considered.

AMPLIFIER APPLICATIONS

In audio amplifiers, the 40250 and 40251 offer several advantages. The high maximum junction temperature (200°C) allows operation in high ambient temperatures or with very small heat sinks. The high F_T of these units allows simple design of amplifiers having frequency response over the entire audio spectrum. The techniques used in their construction result in a unit with high transient power handling capability and freedom from second breakdown in the operating region.

Medium Power Audio Amplifier

Typical performance data for the medium power audio amplifier shown in Figure 1 is presented below.

1. Total harmonic distortion at 1000 cps = 2.0 percent.
2. Power Gain = 87 db
3. Input Impedance = 500 ohms
4. Current Drain = 525 ma

Curves showing distortion versus power output, distortion versus supply voltage, and current drain versus supply voltage for this circuit are shown in Figures 2, 3, and 4.

The amplifier was tested with one 40250 and two 2N3053 units having betas very close to the lower limits. Results were as follows:

1. Distortion = 6.0 percent at 2.0 watts output
2. Power Gain = 80 db
3. Input Impedance = 330 ohms

Substitution of a higher beta unit in any or all positions gave improved performance for all characteristics.

The operation of the amplifier with all high-limit (for beta) units was tested at 100°C ambient with the following results:

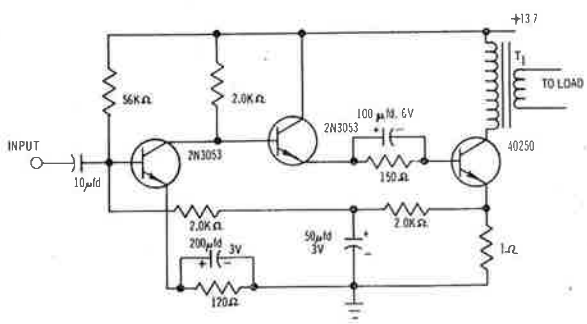


Fig. 1 — Medium power amplifier. Output transformer primary impedance, 24 ohms, secondary impedance to match load, 0.52 amps unbalanced dc current in primary.

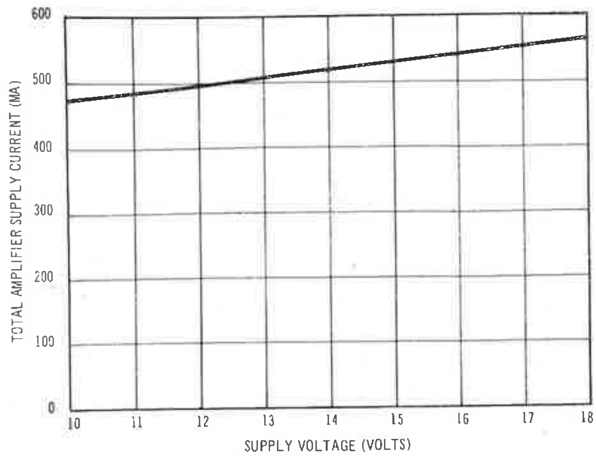


Fig. 2 — Total amplifier current versus supply voltage, medium power audio amplifier 40250.

1. Distortion = 2.4 percent at 2.0 watts output
2. Power Gain = 86 db
3. Amplifier current increased over 25°C current by 12 percent (580 to 650 ma)

The negative dc feedback used in this circuit produces a very stable circuit with temperature changes and parameter variations. This circuit should prove useful wherever a low-cost, medium quality amplifier with 2.0 to 3.0 watt capability is needed.

Public Address Amplifier

Typical performance data for the circuit shown in Figure 5 is listed below.

1. Total harmonic distortion at 1000 cps with 50-watts output = 10 percent

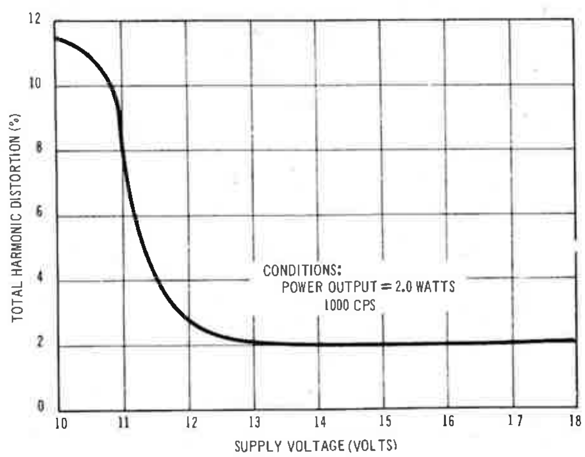


Fig. 4 — Distortion versus power output, medium power audio amplifier 40250. power audio amplifier 40250.

2. Total harmonic distortion at 1000 cps with 40 watts output = 4.0 percent
3. Total harmonic distortion at 1000 cps with 10 watts output = 1.5 percent
4. Power Gain = 63 db
5. Current drain for 50-watts output = 8.1 amps
6. Current drain for no output = 3.0 amps
7. Stable for all resistive input and output terminations

Distortion versus output power and frequency response curves for this circuit are shown in Figures 6 and 7.

This circuit operates satisfactorily with case temperatures as high as 125°C. Substituting transistors with betas very close to the lower limit into each socket did not appreciably degrade performance.

The dc negative feedback around the first three stages stabilises these stages for temperature and parameter variations. The 1N1199A diode in the bias network provides compensation for heat sink temperature and parameter variations. This unit should be bonded thermally to the same heat sink as the output units.

Approximately 15 db of ac negative feedback is used around the entire circuit to linearise the output and provide a flat frequency response. A capacitor is used from the collector of the input 2N3053 to ground for reducing the gain at frequencies above the pass band and thereby avoiding oscillations at high frequencies.

DC-TO-DC Converter Applications

The 40250 and 40251 are ideally suited for application in dc-to-dc converters for mobile applications. The low saturation resistance and

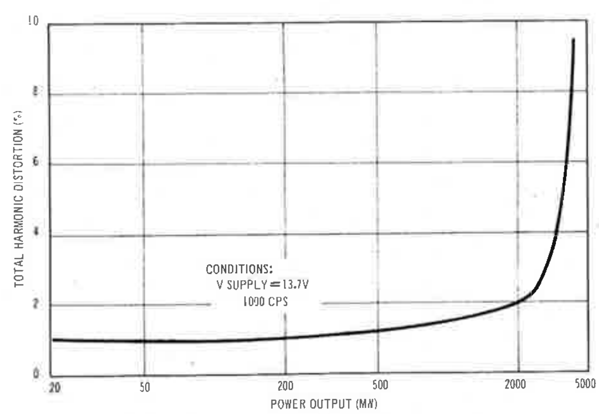
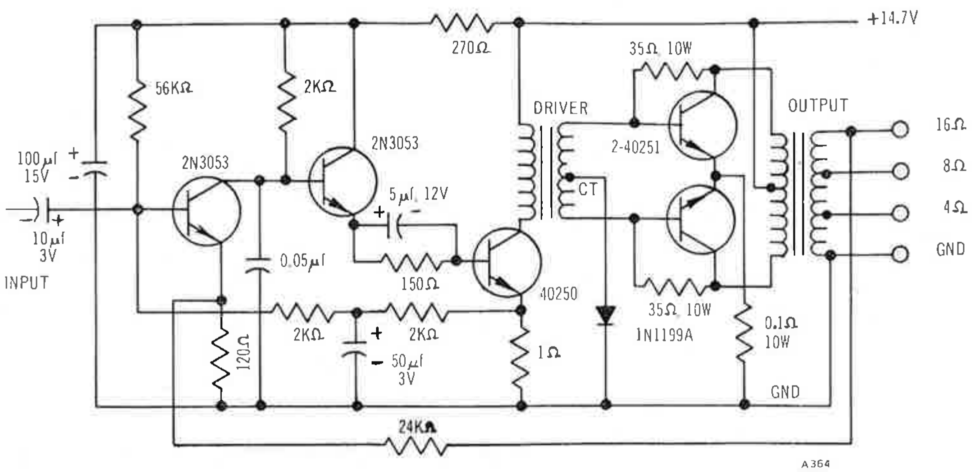


Fig. 3 — Distortion versus supply voltage, medium power audio amplifier 40250.



DRIVER TRANSFORMER:

CRYSTLYNED 100 EI 0.014" 1:1 INTERLEAVED
PRIMARY: 180 TURNS #22 WIRE
SECONDARY: 90 TURNS CENTER TAPPED #20 WIRE BIFILAR WINDING

OUTPUT TRANSFORMER:

CRYSTLYNED 150 EI 0.014" 1:1 INTERLEAVED
PRIMARY: 68 TURNS CENTER TAPPED #16 WIRE BIFILAR WINDING
SECONDARY: 120 TURNS #20 WIRE TAPPED AT 80 AND 60 TURNS

Fig. 5 — Public address amplifier.

fast switching times of these devises allows design of efficient high frequency converters. The capability of these units to operate at very high ambient temperatures (up to 160°C) allows flexibility in placement of the converter.

The basic design formulae for the transformer of a push-pull transformer-coupled converter are:[1]

$$N_P = \frac{V_{in}\, 10^8}{4fAB_{sat}}$$

$$N_{FB} = N_P \left(\frac{V_{be\ max}}{V_{in}}\right)$$

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

[1] "Design of Transistorized DC-to-DC Converters" by C. R. Turner, Electronic Design, Sept. 16 and 30, 1959.

where:

B_{sat} = Flux density in guasses at saturation

A = Cross sectional area of core

f = Frequency of operation

N_P = Number of primary turns

N_{FB} = Number of feedback turns

N_S = Number of secondary turns

Two converter circuits are shown in Figure 8. The circuit of Figure 8a provides 100-watts of output at 110 volts. A typical efficiency of 81.5 percent is realised at full power output. Frequency of operation is 3.5 Kc. When units with betas very close to the lower limit were used

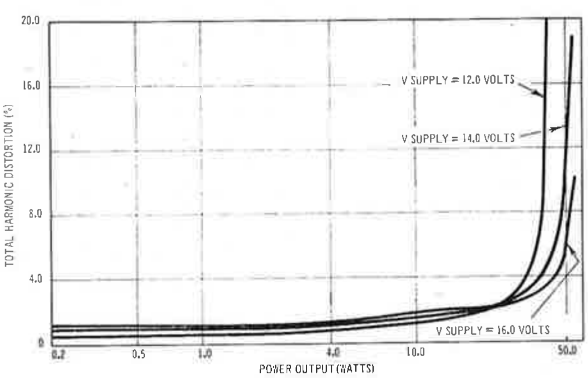


Fig. 6 — Typical operation of public address amplifier utilising 40250 and 40251.

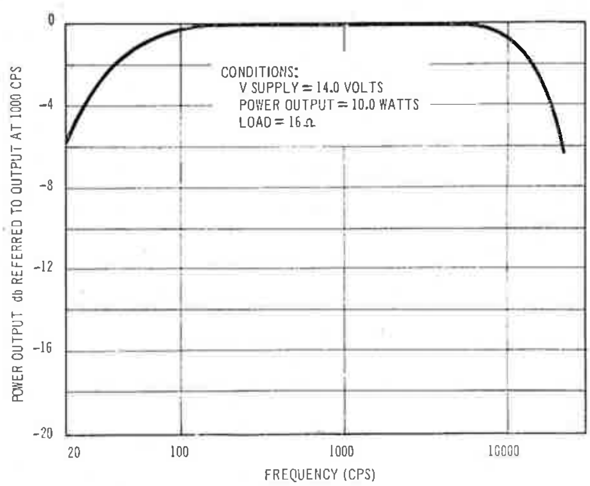


Fig. 7 — Frequency response of public address amplifier utilising 40250 and 40251.

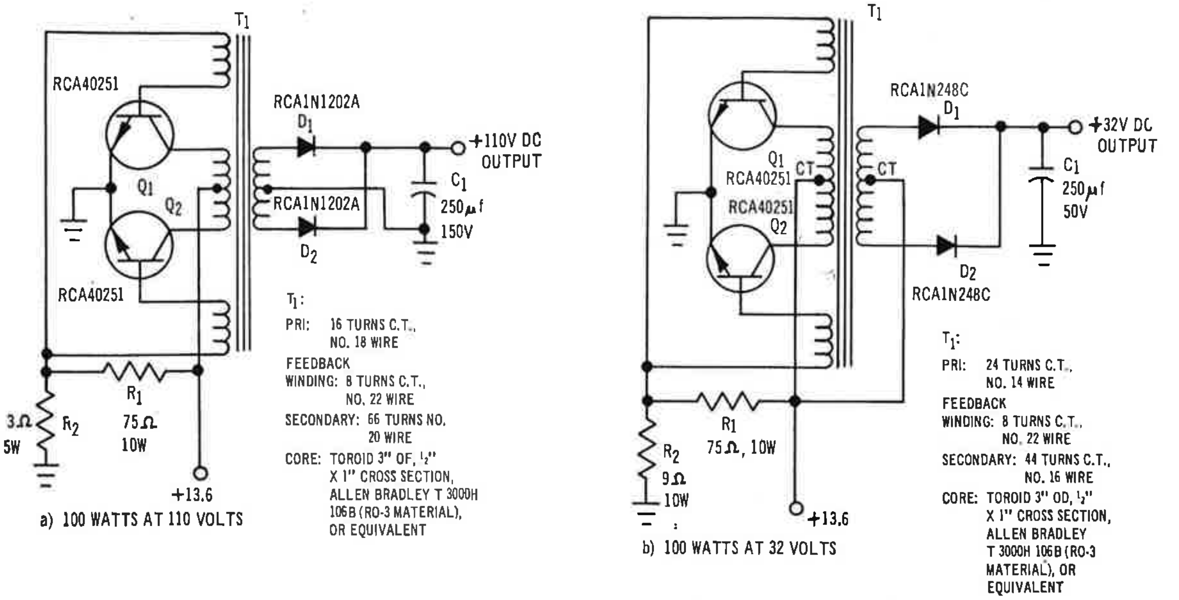


Fig. 8 — 12-volt converter designs employing silicon transistors.

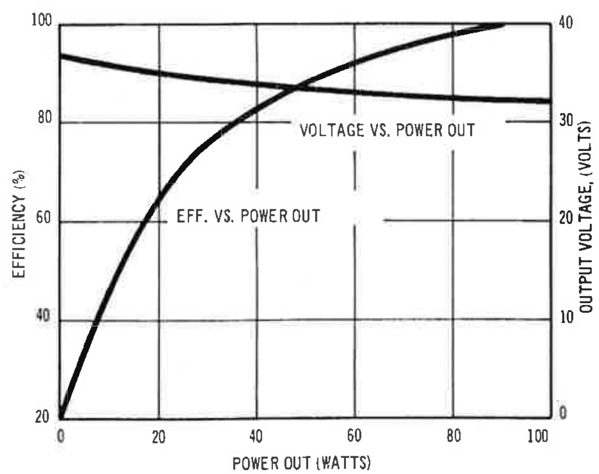


Fig. 9 — 110-volt converter, output voltage versus power output and efficiency versus power output.

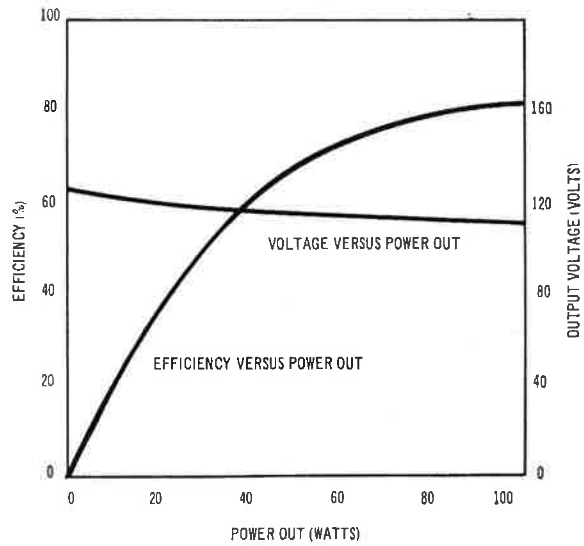


Fig. 10 — 32-volt converter, output voltage versus power output and efficiency versus power output.

a 2 percent decrease in efficiency occurred. Operation at 100°C resulted in a 6 percent decrease in efficiency and a 9 percent drop in output voltage.

The converter circuit shown in Figure 8b provides 100-watts output at 32 volts and operates at a frequency of 2.5 Kc. The output of the converter is connected in series with the available 12-volt supply to increase efficiency. An efficiency of 80.5 percent is achieved at full power output despite the low output voltage which lowers rectification efficiency. Operation at 100°C ambient resulted in a decrease in efficiency of 4 percent and a decrease in output voltage of 7 percent.

Voltage versus power output and efficiency versus power output for these circuits are shown in Figures 9 and 10.

(With acknowledgements to RCA.)

THERMAL RUNAWAY IN HIGH VOLTAGE SWITCHING TRANSISTORS

by D. W. McChesney

Introduction

The objectives of this paper are threefold: First, to describe the condition known as "thermal runaway"; second, to analyze mathematically thermal runaway in terms of the pertinent transistor and circuit parameters, from a switching circuit aspect, and finally, to discuss the effects of thermal runaway on the operating conditions of the RCA 2N398 family of germanium alloy, high voltage switching transistors.

One of the fundamental problems encountered in transistor circuitry design is the change in transistor operating characteristics with temperature. One of these characteristics, I_{CBO}, is a direct result of the thermally generated hole-electron pairs in the semiconductor material. The hole-electron pairs generated in the depletion region of the reverse biased collector-base semiconductor junction cause a reverse saturation current (I_{CBO}). The magnitude of this current increases exponentially with increasing temperature.

When a transistor is operating in a circuit, an increase in ambient temperature will cause an increase in saturation current. In transistors operating in common emitter configuration in the active region, the variation in reverse saturation current with temperature can result in an exaggerated change in collector current, depending on the stability factor of the circuit. The stability factor (S) is defined as $\partial I_C/\partial I_{CBO}$ and may vary from $1/1-\alpha$ to approximately 1 depending on the type of biasing used. If the increase in saturation current and the resulting increase in collector current causes an increase in power dissipation in the transistor junction, the junction temperature will rise increasing the saturation current even further. If conditions were such that this cumulative thermal regeneration continued, the transistor might be destroyed by excessive junction temperature or at least a transistor which was initially biased in the active region might become biased deep in saturation leaving it useless as a linear amplifier. This undesirable sequence of events is called "thermal runaway".

In switching circuits, the transistor does not normally operate in the active region, it merely passes through the active region when switching back and forth from saturation to cutoff. With the absence of a "turn-on" signal at the base of the transistor, the base-emitter junction is usually reverse biased through a series resistor and voltage supply. The series combination should be able to supply reverse current equal to the maximum saturation current of the transistor at the highest expected operating temperature without allowing the base-emitter diode to become forward biased.

With such biasing, the collector current is equal to the saturation current and the stability factor (S) is equal to one. The condition is equivalent to a transistor having a reverse biased collector-base diode with its emitter open-circuited. However, even at this stablest of reverse saturation conditions, thermal runaway can still occur within the ambient operating temperature range of the transistor, particularly when high collector voltage levels are employed.

The following equations indicate the relationship between saturation current, transistor power dissipation, and junction temperature. These equations can be used to determine if a transistor is being operated in a thermally stable condition.

$$(1)\quad T_J = T_A + \theta_{JA} P$$

$$(2)\quad I_{CBO_{TJ}} = I_{CBO_{TR}}\, e^{(T_J - T_R)/\Phi}$$

where

T_J = the transistor junction temperature in °C

T_A = the free air ambient temperature in °C

θ_{JA} = the thermal resistance of the transistor from the junction to free-air ambient in °C/watt.

P = the power dissipated in the transistor in watts.

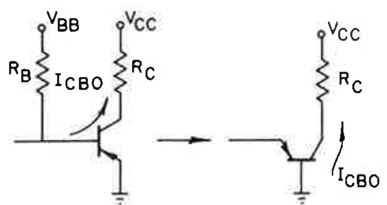
I_{CBO} = the collector-base reverse saturation current with the emitter open-circuited. The subscript T_J following I_{CBO} signifies a variable value of I_{CBO} dependent on T_J. Any other subscript following I_{CBO}, such as T_R signifies the particular value of I_{CBO} at $T_J = T_R$.

T_R = a reference temperature, in this case $T_R = 25°C$

Φ = the exponential temperature coefficient of the saturation current, in this case taken as a constant for a given semiconductor material.

If it is considered that a common emitter transistor switch operating in the cutoff region is equivalent to a reverse-biased collector-base diode with emitter open, then:

$I_C = I_{CBO}$ and the circuit conditions are as shown on page 2.



The transistor power dissipation (P) in equation (1) becomes $P = (V_{CC} - I_{CBO_{T_J}} R_C)\, I_{CBO_{T_J}}$ If equation (2) is substituted for I_{CBO} in equation (1) then:

(3)
$$T_J = T_A + \left[\theta_{JA} V_{CC} I_{CBO_{TR}} e^{-T_R/\Phi}\right] e^{T_J/\Phi} - \left[\theta_{JA} I_{CBO_{TR}}^{\ 2} R_C\, e^{-2T_R/\Phi}\right] e^{2T_J/\Phi}$$

Mathematically, thermal runaway can be defined as the condition existing when $\partial T_J/\partial f \to \infty$ where "f" is any of the variables effecting T_J

Let T_A be the independent variable since the other parameters are fixed once the circuit and transistors are chosen.

If equation (3) is differentiated with respect to T_A and the $\partial T_J/\partial T_A$ found is allowed to approach ∞, the following values of T_J are found to satisfy the conditions. (See Appendix 2).

$$T_J \text{ critical} = \Phi \ln \left[\frac{V_{CC} e^{T_R/\Phi}}{4 R_C I_{CBO_{TR}}} \left(1 \pm \sqrt{1 - \frac{8\Phi R_C}{\theta_{JA} V_{CC}^{\ 2}}}\right)\right]$$

There are two critical values of T_J at which $\partial T_J/\partial T_A \to \infty$. The lower value critical T_J is the one important to thermal runaway.

$$T_{J_{C_1}} = \Phi \ln \left[\frac{V_{CC} e^{T_R/\Phi}}{4 R_C I_{CBO_{TR}}} \left(1 - \sqrt{\frac{1 - 8\Phi R_C}{\theta_{JA} V_{CC}^{\ 2}}}\right)\right]$$

The second critical value of T_j comes about because there is an R_C in the circuit which limits the current after thermal runaway. If in equation (4), R_C is allowed to become zero:

(4)
$$\lim_{R_C \to 0} \left\{T_{J_{C_1}}\right\} = \Phi \ln \left[\frac{\Phi e^{T_R/\Phi}}{\theta_{JA} I_{CBO_{TR}} V_{CC}}\right]$$

$$\lim_{R_C \to 0} \left\{T_{J_{C_2}}\right\} = \infty$$

Curves of T_J versus T_A (Figures 1 and 2) are plotted using circuit and transistor parameters typical with those of the 2N398 family of transistors. The curves having the parameter $I_{CBO} = 3\mu a$ at 25^oC are typical of the 2N398A. The curves of Figure 1 are for $R_C = 0$; those of Figure 2 are for $R_C = 10K$. Notice that there are 2 points where $T_J/T_A \to \infty$ when $R_C = 10K$; however, after the first point, there is a discontinuity in the characteristics so the second point is not realized.

The ambient temperature at which thermal runaway occurs (T_A critical) may be found by solving equation (3) for T_A when $T_J = T_{JC_1}$.

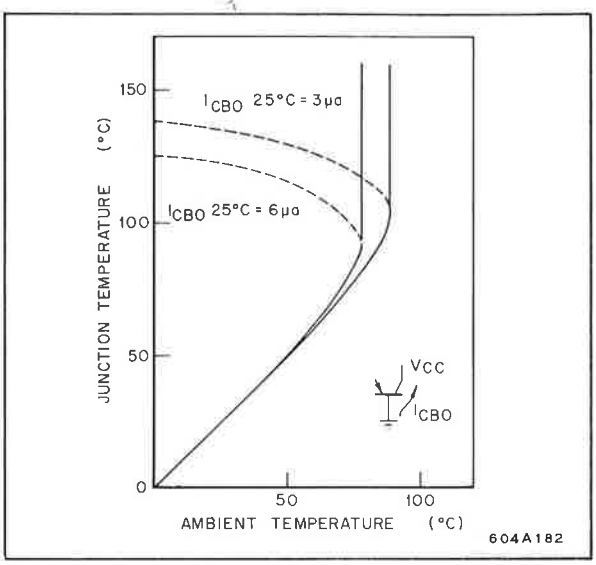


Figure 1

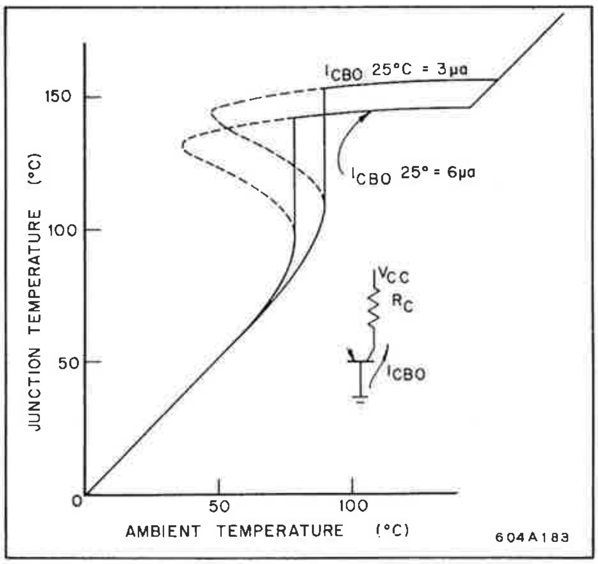


Figure 2

The curves of Figures 1 and 2 show that units having high I_{CBO} go into thermal runaway at lower ambient temperature than units having low I_{CBO}, all other parameters being equal. Equation (4) shows that maximum protection against thermal runaway is obtained with transistors having low I_{CBO} (25°C), low θ_{JA}, and high Φ. The exponential temperature coefficient of saturation current (Φ), is a "constant" of the semiconductor material. Φ, does, however, vary with temperature because the saturation current is not a simple exponential function of junction temperature. The value of Φ for the 2N398 family of germanium alloy transistors was found experimentally to be approximately 16°C over the temperature range important to thermal runaway. Typical values of I_{CBO} versus temperature for the 2N398 and 2N398A are plotted in Figure 3. Φ may be found from the slopes of these curves.

The remaining two transistor parameters, I_{CBO} (25°C) and θ_{JA}, can be altered to improve thermal stability.

The saturation current is chiefly a function of the crystal resistivity. Thermal resistance is a measure of the resultant resistance of the transistor housing to heat flow away from the junction, and is a function of diameters, lengths, and materials of the various heat paths through the connector wires, base tab, and potting compound to the header, can, and leads, and on to the ambient air.

The increased thermal stability attained in the 2N398A was achieved by a change in junction geometry dimensions which allowed a lower resistivity crystal to be used without degrading the transistor current gain. The collector breakdown distribution of the product does center at values lower than the original 2N398 because of this resistivity reduction, but the breakdown voltages still remain well above the minimum allowable values. The 2N398B incorporates the 2N398A design and resistivity changes plus a much improved thermal resistance to yield a unit having excellent thermal stability.

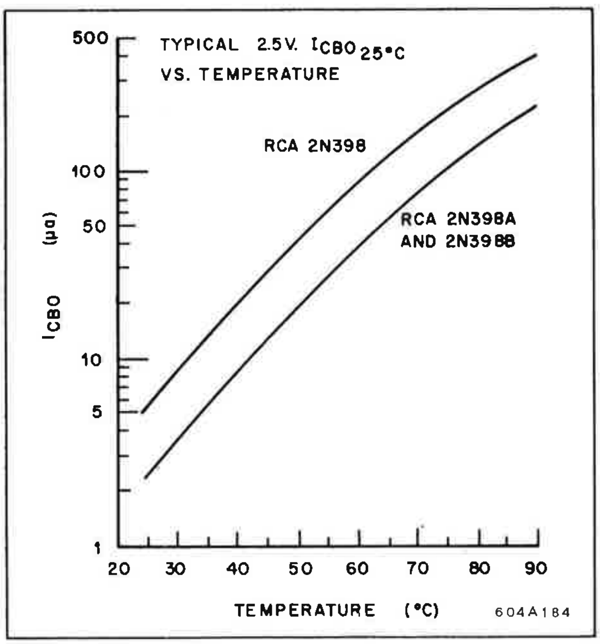


Figure 3

Theoretical T_J versus T_A characteristics are plotted in Figure 4 for two different values of θ_{JA} (350°C/watt and 250°C/watt); these being respectively typical thermal resistance values for the 2N398A and 2N398B. I_{CBO} 25°C in both cases is 3 μa. A V_{CB} versus T_A curve (Figure 5) depicting thermal runaway, can be constructed by transposing the V_{CB} versus T_J curve through the T_J versus T_A characteristic. V_{CB} versus T_J is merely the curve of

$$V_{CB} = V_{CC} - R_C\, I_{CBO_{25^\circ C}}\, e^{(T_J - 25^\circ C)/\Phi},$$

which is independent of θ_{JA}.

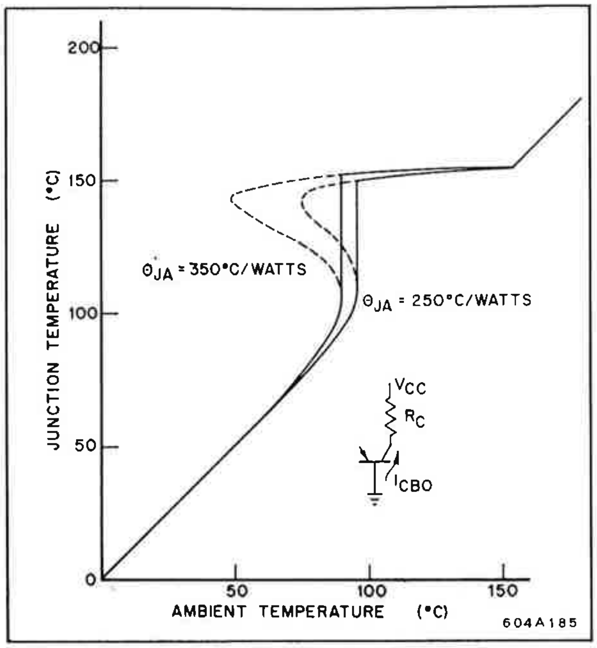


Figure 4

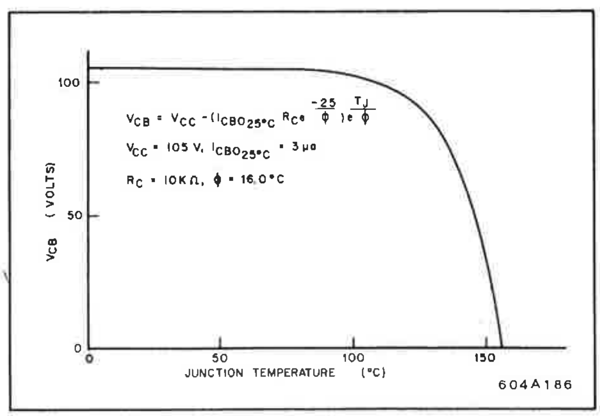


Figure 5

The V_{CB} versus T_A curves (Figure 6) show the thermal runaway characteristics for two transistors that differ only in thermal resistance. The curves do not emphasize the true thermal stability differences between the 2N398A and 2N398B. As stated previously, the 2N398A and 2N389B product are of the same geometry design and crystal resistivity, therefore the median points of their 2.5v $I_{CBO_{25^{\circ}C}}$ distributions before testing and selection are equivalent. The big difference appears when one looks at the allowable upper limits of the 2.5v $I_{CBO_{25^{\circ}C}}$ values.

The maximum allowable 2.5v $I_{CBO_{25^{\circ}C}}$ for the 2N398B is 6μa. The maximum allowable 2.5v $I_{CBO_{25^{\circ}C}}$ for the 2N398A is 14μa, no better than the 2N398.

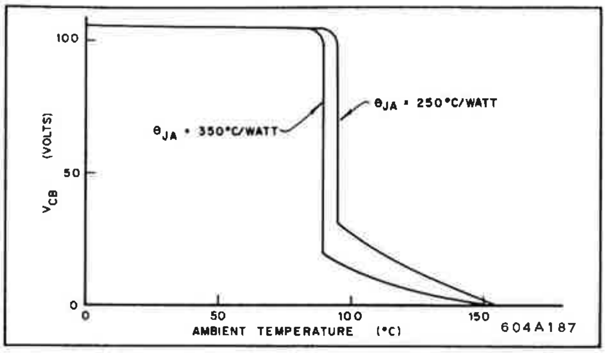


Figure 6

Figure 7 shows the T_J versus T_A curves for thermal runaway at worst-case conditions for the 2N398A and 2N398B. Figure 8 shows V_{CB} versus T_A curves for thermal runaway at worst case conditions for the 2N398A, I_{CBO} maximum = 14μa and θ_{JA} maximum = 500°C/watt. For the 2N398B, I_{CBO} maximum = 6μa and θ_{JA} = 300°C/watt.

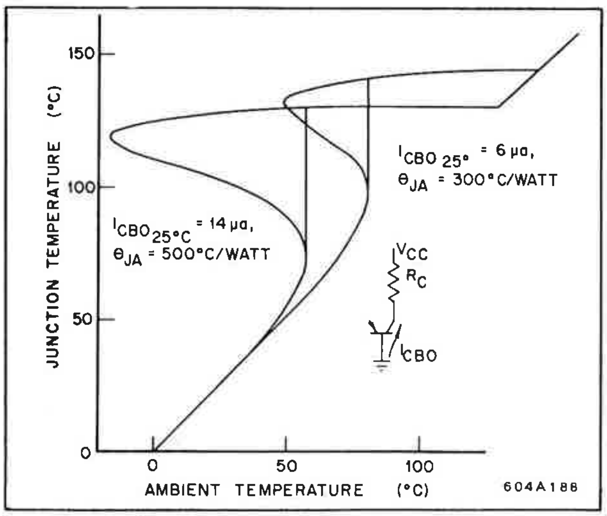


Figure 7

These curves emphasize the extreme differences in thermal stability of the two types. However, for designing high reliability circuits, these conservative ambient tempreature limitations should be observed. Several V_{CB} versus T_A curves from actual data on the 2N398A and 2N398B transistors are shown in Figure 9. All of the units plotted had 2.5v $I_{CBO_{25^{\circ}C}}$ less than 3 μa. Note that the 2N398A curves agree quite well with the theoretical results shown in Figure 6.

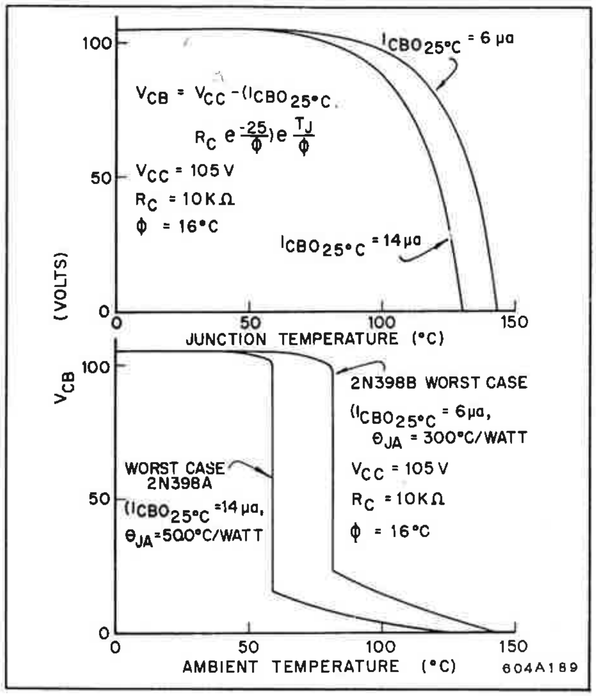


Figure 8

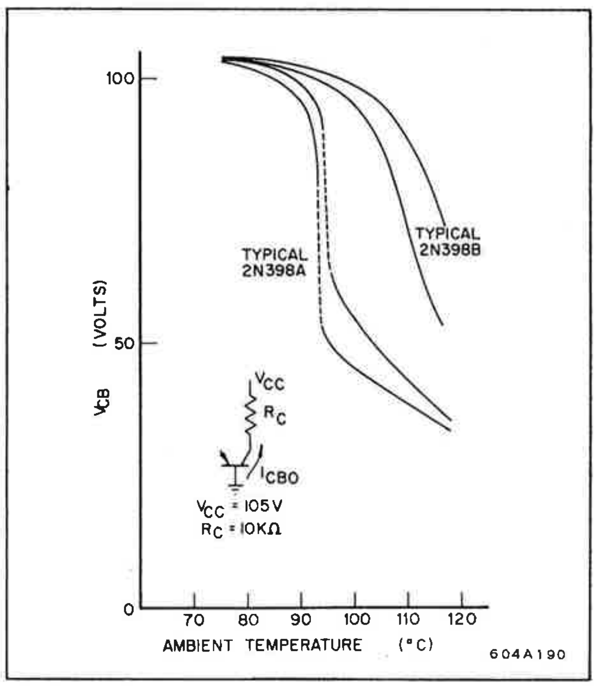


Figure 9

Conclusion

In conclusion, three statements may be made: (1) thermal runaway can occur in switching circuits, especially where high voltages are employed; (2) when the circuit and transistor parameters are known, the ambient temperature at which thermal runaway will occur can be predicted; and (3) when selecting transistors for high thermal stability, one should choose units having thermal stability specified into them through low I_{CBO} maximum limits and low thermal resistance values.

Appendix

The two equations which govern thermal runaway when the transistor emitter-base diode junction is reverse biased are:

$$T_J = T_A + \theta_{JA}P \qquad (1)$$

$$I_{CBO_{T_J}} = I_{CBO_{T_R}} e^{(T_J - T_R)/\Phi} \qquad (2)$$

where T_J = the transistor junction temperature in °C.

T_A = the free air ambient temperature in °C.

θ_{JA} = the thermal resistance of the transistor from the junction to free air ambient in °C/watt.

P = the power dissipated in the transistor in watts.

I_{CBO} = the collector-base reverse saturation current with the emitter open circuited. The subscript T_J following I_{CBO} signifies a variable value of I_{CBO} dependent on T_J. Any other subscript following I_{CBO}, such as T_R signifies the particular value of I_{CBO} at $T_J = T_R$.

T_R = a reference temperature, in this case $T_R = 25°C$.

Φ = the exponential temperature coefficient of the saturation current, in this case taken as a constant for a given semiconductor material.

With the emitter-base diode reverse biased Pin equation (1) = $(V_{CC} - I_{CBO_{T_J}} R_C) I_{CBO_{T_J}}$ or $T_J = T_A + \theta_{JA} (V_{CC} - I_{CBO_{T_J}} R_C) I_{CBO_{T_J}}$; substituting equation (2) for $I_{CBO_{T_J}}$ in equation (1) yields:

$$T_J = T_A + \theta_{JA} (V_{CC} - I_{CBO_{TR}} e^{(T_J - T_R)/\Phi} \quad I_{CBO_{TR}} \quad (T_J - T_R)/\Phi$$

$$T_J = T_A + \left(\theta_{JA} V_{CC} I_{CBO_{TR}} e^{-T_R/\Phi}\right) e^{T_J/\Phi} - \left(\theta_{JA} I_{CBO_{TR}}^{\,2} R_C e^{-2T_R/\Phi}\right) e^{2T_J/\Phi} \qquad (3)$$

Mathematically thermal runaway can be defined as the condition occuring when $\frac{\partial T_J}{\partial f} \longrightarrow \infty$ where f is any one of the variables effecting the junction temperature.

Let T_A be f, the independent variable, since the other parameters are fixed for any one transistor and circuit condition.

Partially differentiating equation (3) with respect to T_A yields:

$$\frac{\partial T_J}{\partial T_A} = 1 + \left[\frac{\theta_{JA} V_{CC} I_{CBO_{TR}} e^{-T_R/\Phi}}{\Phi}\right] e^{T_J/\Phi} \frac{\partial T_J}{\partial T_A} - \left[\frac{2\theta_{JA} I_{CBO_{TR}}^2 R_C e^{-2T_R/\Phi}}{\Phi}\right] e^{+2T_J/\Phi} \frac{\partial T_J}{\partial T_A}$$

$$\text{so: } \frac{\partial T_J}{\partial T_A} = \frac{1}{1 - \left[\frac{\theta_{JA} V_{CC} I_{CBO_{TR}} e^{-T_R/\Phi}}{\Phi}\right] e^{T_J/\Phi} + \left[\frac{2\theta_{JA} I_{CBO_{TR}}^2 R_C e^{-2T_R/\Phi}}{\Phi}\right] e^{2T_J/\Phi}}$$

or: $$\frac{\partial T_J}{\partial T_A} = \frac{1}{1 - K_1 e^{T_J/\Phi} + K_2 e^{2T_J/\Phi}}$$

where: $$K_1 = \frac{\theta_{JA} V_{CC} I_{CBO_{TR}} e^{-T_R/\Phi}}{\Phi}$$ and $$K_2 = \frac{2\theta_{JA} I^2_{CBO_{TR}} R_C e^{-2T_R/\Phi}}{\Phi}$$

thermal runaway occurs when $\frac{\partial T_J}{\partial T_A} \longrightarrow \infty$

or: $$K_2 \left(e^{T_J/\Phi}\right)^2 - K_1 \left(e^{T_J/\Phi}\right) + 1 \longrightarrow 0$$

solving this for values of $e^{T_J/\Phi}$ by the quadratic formula yields

$$e^{T_J/\Phi} = \frac{+K_1 \pm \sqrt{K_1^2 - 4K_2}}{2K_2}$$

or: T_{JC} (T_J critical, the value of T_J where $\partial T_J/\partial T_A \longrightarrow \infty$) =

$$T_{JC} = \Phi \ln \left[\frac{+K_1 \pm \sqrt{K_1^2 - 4K_2}}{2K_2}\right]$$

substituting for the K's yields

$$T_{JC} = \Phi \ln \left[\frac{V_{CC}\, e^{T_R/\Phi} \left(1 \pm \sqrt{1 - \frac{8\Phi R_C}{\theta_{JA} V_{CC}^2}}\right)}{4 R_C I_{CBO_{TR}}}\right]$$

There are two values of T_J at which $\frac{\partial T_J}{\partial T_A} \longrightarrow \infty$; the lower value of T_J is the one which thermal runaway occurs.

$$T_{JC_1} = \Phi \ln \left[\frac{V_{CC}\, e^{T_R/\Phi}}{4R_C\, I_{CBO_{TR}}} \left(1 - \sqrt{1 - \frac{8\Phi R_C}{\theta_{JA}\, V_{CC}^2}}\right)\right]$$

The second value of T_{JC} occurs at a temperature above the thermal runaway point. This second T_{JC} is a result of the current limiting resistor R_C; however, the T_J vs. T_A characteristic is discontinuous at T_{JC}, so that T_{JC_2} is not realized physically. (See curves 2 and 4, Appendix 1).

If the limits of T_{JC_1} and T_{JC_2} are taken as $R_C \longrightarrow 0$, it is found that:

$$\lim_{R_C \to 0} \left\{T_{JC_1}\right\} = \Phi \ln \left[\frac{\Phi\, e^{T_R/\Phi}}{\theta_{JA}\, I_{CBO_{TR}}\, V_{CC}}\right] \quad \text{but} \quad \lim_{R_C \to 0} \left\{T_{JC_2}\right\} = \infty$$

$$\lim_{R_C \to O} \left\{ T_{JC_1} \right\} = \lim_{R_C \to O} \left\{ \Phi \ln \left[\frac{V_{CC} e^{T_R/\Phi}}{4 I_{CBO_{TR}} R_C} \left(1 - \sqrt{1 - \frac{8\Phi R_C}{\theta_{JA} V_{CC}^2}} \right) \right] \right\}$$

$$= \Phi \ln \left[\lim_{R_C \to O} \left\{ \frac{V_{CC} e^{T_R/\Phi}}{4 I_{CBO_{TR}} R_C} \left(1 - \sqrt{1 - \frac{8\Phi R_C}{\theta_{JA} V_{CC}^2}} \right) \right\} \right]$$

$$\lim_{R_C \to O} \left\{ \frac{V_{CC} e^{T_R/\Phi}}{4 I_{CBO_{TR}} R_C} \left(1 - \sqrt{1 - \frac{8\Phi R_C}{\theta_{JA} V_{CC}^2}} \right) \right\} = \frac{O}{O}$$

By applying L'Hopital's rule differentiating both the numerator and the denominator of T_{JC_1} with respect to R_C and again taking the limit as $R_C \to O$ a limit value may be found.
Differenting yields.

$$\lim_{R_C \to O} \left\{ \frac{V_{CC} e^{TR/\Phi}}{4 I_{CBO_{TR}}} \left[-1/2 \left(1 - \frac{8\Phi R_C}{\theta_{JA} V_{CC}^2} \right)^{-1/2} \left(\frac{-8\Phi}{\theta_{JA} V_{CC}^2} \right) \right] \right\} =$$

$$\lim_{R_C \to O} \left\{ \frac{\Phi e^{TR/\Phi}}{\theta_{JA} V_{CC} I_{CBO_{TR}}} \left(\frac{1}{\sqrt{1 - \frac{8\Phi R_C}{\theta_{JA} V_{CC}^2}}} \right) \right\} = \frac{\Phi e^{TR/\Phi}}{\theta_{JA} V_{CC} I_{CBO_{TR}}}$$

$$\text{so: } \lim_{R_C \to O} \left\{ T_{JC_1} \right\} = \Phi \ln \left[\frac{\Phi e^{TR/\Phi}}{\theta_{JA} V_{CC} I_{CBO_{TR}}} \right]$$

$$\lim_{R_C \to O} \left\{ T_{JC_2} \right\} = \lim_{R_C \to O} \left\{ \Phi \ln \left[\frac{V_{CC} e^{TR/\Phi}}{4 I_{CBO_{TR}} R_C} \left(1 + \sqrt{1 - \frac{8\Phi R_C}{\theta_{JA} V_{CC}^2}} \right) \right] \right\}$$

$$= \Phi \ln \left[\lim_{R_C \to O} \left\{ \frac{V_{CC} e^{T_R/\Phi}}{4 I_{CBO_{TR}} R_C} \left(1 + \sqrt{1 - \frac{8\Phi R_C}{\theta_{JA} V_{CC}^2}} \right) \right\} \right]$$

$$\lim_{R_C \to O} \left\{ \frac{V_{CC} e^{T_R/\Phi}}{4 I_{CBO_{TR}} R_C} \left(1 + \sqrt{1 - \frac{8\Phi R_C}{\theta_{JA} V_{CC}^2}} \right) \right\} = \frac{1}{O} = \infty \text{ therefore } \lim_{R_C \to O} \left\{ T_{JC_2} \right\} = \infty$$

T_{AC}, the value of ambient temperature at which thermal runaway occurs may be found by substituting T_{JC} for T_J in equation (3) and solving for T_A.

$$\text{This yields: } T_{AC} = T_{JC_1} - \left[\theta_{JA} V_{CC} I_{CBO_{TR}} e^{-TR/\Phi} \right] e^{T_{JC_1}/\Phi} + \left[\theta_{JA} I_{CBO_{TR}}^2 R_C e^{-2TR/\Phi} \right] e^{+2T_{JC_1}/\Phi}$$

$$\text{this may be reduced to } T_{AC} = T_{JC_1} - \frac{V_{CC}^2 \theta_{JA}}{8R_C} \left(1 - \sqrt{1 - \frac{8\Phi R_C}{\theta_{JA} V_{CC}^2}} \right) - \Phi/2$$

in the case where $R_C \rightarrow 0$

$$\lim_{R_C \rightarrow 0} \left\{ T_{AC} \right\} = \lim_{R_C \rightarrow 0} \left\{ T_{JC_1} \right\} - \lim_{R_C \rightarrow 0} \left\{ \frac{V_{CC}{}^2 R_T}{8R_C} \left(1 - \sqrt{1 - \frac{8R_C \Phi}{\theta_{JA} V_{CC}{}^2}} \right) \right\} - \Phi/2$$

it has been previously shown that $\lim_{R_C \rightarrow 0} \left\{ \frac{1 - \sqrt{1 - \frac{8\Phi R_C}{\theta_{JA} V_{CC}{}^2}}}{R_C} \right\} = \frac{4\Phi}{\theta_{JA} V_{CC}{}^2}$

so $$\lim_{R_C \rightarrow 0} \left\{ T_{AC} \right\} = \lim_{R_C \rightarrow 0} \left\{ T_{JC_1} \right\} - \frac{V_{CC}{}^2 \theta_{JA}}{8} \left(\frac{4\Phi}{\theta_{JA} V_{CC}{}^2} \right) - \Phi/2$$

$$\lim_{R_C \rightarrow 0} \left\{ T_{AC} \right\} = \lim_{R_C \rightarrow 0} \left\{ T_{JC_1} \right\} - \Phi$$

On curve 4 in Appendix 1 the V_{CB} vs. T_J characteristic is transposed through the T_J vs. T_A plot to yield a curve of V_{CB} vs. T_A, depicting the thermal runaway conditions.

V_{CB}vs. T_J is independent of θ_{JA} so two transistors having different θ_{JA} but equal $I_{CBO_{T_R}}$'s would yield the same V_{CB} vs. T_J characteristic.

$$V_{CB} = V_{CC} - I_{CBO} R_C$$

$$V_{CB} = V_{CC} - \left(I_{CBO_{TR}} R_C \epsilon^{-TR/\Phi} \right) e^{T_J/\Phi}$$

(With acknowledgements to RCA)

NEW RELEASES

NEW ECONOMY VHF/UHF 'OVERLAY' TRANSISTOR

A new general purpose vhf/uhf "overlay" transistor, with a typical 5 watt power output at 50 Mc and 20 db gain, has been released by RCA. This new economy "overlay" transistor, the 2N3553, offers exceptionally high gain in the 27-to-260 Mc frequency range for rf applications as a driver or power output stage and also performs well as a high-gain frequency multiplier and as an oscillator up to 500 Mc. Its fast rise and fall time makes it ideally suited for pulse amplifier applications. The wide-band capabilities open a variety of video amplifier applications. Designed for 12-to-28 volt operation, the new transistor will have broad applications in the mobile, aircraft, marine, citizens' band, portable, and military electronic equipment fields.

The 2N3553 is mounted in a standard JEDEC TO-39 package and is designed to assure freedom from second breakdown in class A, —B and —C operation at maximum ratings. This silicon n-p-n epitaxial planar transistor utilises a unique emitter electrode structure known as an "overlay." This overlay structure is a tiny silicon mosaic made up of 156 individual emitter sites which are microscopic in size.

These emitters are paralleled through use of a diffused grid structure and an overlay of metal, which is applied on the silicon wafer by means of a photo-etch process. The overlay structure is designed to increase the emitter-base periphery for a given emitter area far beyond that achieved in circle, ring, or comb-type transistor pellet structures. This increase in emitter periphery-to-area ratio greatly improves the transistor performance in the vhf/uhf range.

QUICK-HEATING POWER VALVE 8462

A new conduction-cooled beam power tube requiring a warm-up time of less than 1 second has been developed by the ECD Industrial Tube and Power Division, Lancaster, Pa. Designed for low voltage mobile or stationary equipment, the 8462 may be used as an rf power-amplifier, oscillator, regulator, distributed amplifier, or linear rf power amplifier.

The fast warm-up of the 8462 eliminates stand-by filament power in "push to talk" emergency equipment. For extremely rapid warm-up of less than 100 milleseconds, the tube may be used in a suitably designed over-voltage control circuit.

6247

This subminiature of triode was developed for cases where very small size, tight tolerances and reliability under stringent conditions — such as high temperature, mechanical shocks and vibrations — are specified. The valve thus belongs to the family of types including the 5702WB, 5703WB and 5744WB. At an anode voltage of 250 volts and with a cathode resistor of 570 ohms, the 6247 features a transconductance of 2.65 ± 0.65 ma/v and an amplification factor of 60. The maximum anode load amounts to 1.6 watts. The valve meets the requirements of MIL-specification E1/515 A.

5894 (QQE06/40)

The power rating of this small transmitting valve is about twice as high as the output of the 6252 (QQE03/20). In class C telegraphy operation it supplies 90 watts output at 200 Mc, and 60 watts at 500 Mc, under continuous commercial service conditions. Socket and accessories are uniform with those of the 6252.

NEW SCR 40216

A new silicon controlled rectifier, 40216, designed for high speed pulse duty at currents to 900 amperes and voltages to 600 volts has been developed by the ECD Industrial Tube and Semiconductor Division of R.C.A.

The new device, intended initially for radar pulse modulator applications, can fill a broad range of requirements where extremely fast turn-on of high currents is important. Principal areas of application will probably be in the range of 200 to 900 amperes with a pulse width of 4 to 16 microseconds at repetition rates of 500 to 3500 pulses per second.

6080WA

This valve is a further development of the twin triode 6080. It features the same technical data as the 6080, but is distinguished by a reduction in the tolerances of the electrical values. At an anode supply voltage of 135 volts and cathode resistance of 250 ohms, the anode currents in the two sections differ by maximum 25 ma. The minimum rating of the transconductance tolerance was increased slightly from 5.8 to 6 ma/v. In the 6080WA the maximum envelope temperature is 230°C, in comparison with 260°C, in the 6080. The valve satisfies the requirements of MIL-specification E1/510 D.

BOOK REVIEWS

"ELECTRONICS DATA HANDBOOK", M. Clifford, Gernsback Library Inc. size $8\frac{1}{2}''$ x $5\frac{1}{2}''$, 160 pages.

This book is interesting in more ways than one. To the reviewer, used to dealing with the mechanics of the printing industry, the book is interesting in having been set up in Linofilm. That is, it was set up on film rather than directly into typemetal. However, to the casual observer, the modern technique produces a book that looks like any other. To the electronics user, the book presents a large collection of formulae and other data in handy reference form.

With the growth in size and sophistication of electronics, the wealth of references and other data has become greater than the average user can easily carry in his mind. When the data is required, it is generally scattered through many different volumes on our shelves, and often takes some finding. Books of this kind therefore, whilst they may at times contain nothing new, serve a real need in saving us precious time in searching for information. The handy collection of data and formulae is also the sort of thing any serious student would find invaluable.

"ELECTRONIC DIGITAL INTEGRATING COMPUTERS", F. V. Mayorov, edited by Dr. Yaohan Chu. 382 pages. 273 text diagrams. Size 8¾" x 5½". Iliffe Books Ltd.

General-purpose-digital computers played the principal role in the first stage of computer development. It was subsequently seen that for solving a specific class of problem, a special-purpose computer is simpler, more effective and faster. For practical applications, where the problems are solved in accordance with a previously prepared programme, a bulky and expensive general-purpose computer which requires lengthy programming is not necessary.

On the other hand, the cheapness, simplicity, compactness, low-weight, high operating speed, reliability and computing accuracy of the digital integrating computer, also known as a digital differential analyser, commend it for many applications in process control, scientific engineering computations, simulation in research and development, and in inertial navigation and guidance systems. Machines of this type have been developed which have almost zero memory-access time and are already 100,000 times faster in operation than general-purpose high speed computers. The present book gives the principles, organisation, applications, circuits, memories and input-output devices of electronic digital integrating computers. It describes both serial and parallel machines as well as both binary and decimal types.

This is the first work to be published on the subject in the English language, and is based on a book by F. V. Mayorov, who is an Associate of the Solodovnikov of the All Union Scientific Research Institute of Complex Automation of Moscow. The translation has been edited and adapted for Western readers by Dr. Yaohan Chu, Manager of Digital Systems at the R.C.A. Centre, lecturer in electrical engineering at the University of Maryland, and author of books and papers on computer design and techniques. He has added examples of American, British and Japanese practice, and also included a bibliography of a number of Western sources.

"SIMPLIFIED MODERN FILTER DESIGN", P. R. Geffe. Iliffe Books Ltd., 182 pages, including 206 diagrams in the text. Size 8¾" x 5½".

Normally the synthesis of any except the most simple networks requires an extensive knowledge of higher mathematics and the carrying out of long and tedious calculations. In this book the difficult part of filter design — the calculations—has already been performed by specialists, the results of whose work is embodied in an extensive series of tables and graphs. To design filters as sophisticated as the present state of the art allows, an industry requires, it is only necessary to use the tables of designs and the modifications which are appropriate to them. The design of highpass, bandpass and bandstop filters can be carried out by transforming the tabulated lowpass networks.

This book deals in considerable detail with the effects of parasitic elements such an incidental dissipation (finite Q) and straight capacitance. Other important chapters cover measurement problems peculiar to electrical wave filters, the design of practical high-Q inductors, attenuation equalisers and delay equalisers. This book then enables the engineer to perform his calculations not only with ease, but with the assurance that he will produce a design that will perform satisfactorily in practice. Moreover, it shows him how to predict the performance of an actual filter confidently and without the need to resort to experimental, or breadboard, methods to make it work.

"THE DESIGN AND USE OF ELECTRONIC ANALOGUE COMPUTERS", C. P. Gilbert, Chapman and Hall Ltd., 5¾" x 9", 528 pages.

To apply the analogue computer to new and unusual situations, and to use it efficiently in any situation, a knowledge of the principles of its design is helpful, and a knowledge of the basic theory of its operation is invaluable. The author had in mind during the writing of this book the user of the small to medium-sized computer, who often does not have some of the sophisticated ancillary equipment associated with larger installations, and therefore has greater need of the basic theoretical approach in order to get the best out of his equipment.

This book contains a considerable amount of design information, concentrating on those items that the average user is more likely to want to build himself. The design of non-linear circuits and transfer function circuits is described in detail, and a whole chapter is devoted to DC amplifiers. Overall computer construction is described in terms of the essential functions and relative costs of the various components, with many illustrative examples. Practical aspects, such as fault-finding, are also dealt with.

"COLD CATHODE CIRCUIT DESIGN", D. M. Neale, Chapman and Hall Ltd., 5¾" x 9", 259 pages.

With the spread of electronics into industrial and other spheres over recent years, the use of cold cathode tubes has increased manyfold. The term "cold cathode", whilst self-explanatory in itself, covers a wide field of devices, ranging through stabilising and reference tubes, corona stabilisers, relay tubes, stepping tubes, to register and display tubes. This is the first book your reviewer has seen covering practical aspects of the use of these tubes. A certain amount of disappointment has been experienced in the past by certain users of these devices, and the author feels that many of these cases are due to insufficient information on the devices and their application. This book goes just about all the way to remedy this situation; the treatment is thorough, and the liberal use of design summaries and worked examples must prove useful and instructive to any reader.

"DESIGN OF LOW-NOISE TRANSISTOR INPUT CIRCUITS", W. A. Rheinfelder. Iliffe Books Ltd., 160 pages, including 107 diagrams in the text. Size $8\frac{3}{4}''$ x $5\frac{1}{2}''$.

To the electronic engineer noise is every unwanted signal: the objective in low-noise design is to obtain the maximum signal-to-noise ratio. So much importance does the author attach to a clear understanding of the noise factor concept in the design of the highest quality equipment that he has made a special study of the subject. The results of his research and experience in this field are embodied in the present book, originally published in the U.S.A. and, as far as the publishers are aware, the only one of its kind. In it, he develops a clear step-by-step method in which signal and noise factors are calculated separately and he also devotes a complete chapter to the measurement of the noise figure. This approach allows the most complicated circuits to be dealt with.

The book first develops the theory, and then deals with the practice. Simple derivations of all important formulae are given to help the reader obtain a deeper insight into the fundamentals of practical low-noise concept, and throughout the book there are a multitude of time-saving graphs and design curves for the practical circuit designer.

"TRANSISTOR CIRCUITS IN ELECTRONICS", S. S. Hakim and R. Barrett, Iliffe Books Ltd. 341 pages including 329 diagrams in the text. Size $8\frac{3}{4}''$ x $5\frac{1}{2}''$.

This book has been specially written for students taking electronics as their main subject in degree, Dip. Tech. and H.N.C. courses. Its principal features are that it is written for electronic rather than telecommunication engineering, and that it deals in almost equal proportion with the use of transistors in amplifiers and oscillators on the one hand and for linear and switching circuits on the other.

"TRANSISTOR CIRCUITS IN ELECTRONICS" commences with a discussion of transistor characteristics and the permissible working region of various types, and follows on with an explanation of graphical analysis — stabilisation of operating points, loadlines, power output and non-linear distortion.

The authors then deal at considerable length with small signal equivalent circuits and parameters, and then go on to discuss dc amplifiers, wideband, tuned and power amplifiers and waveform distortion. A further chapter deals with feedback amplifier and oscillator circuits, stability and automatic gain control.

The second half of the book starts by considering the transistor as a switch and then leads on to a consideration of regenerative switching circuits. The next chapter serves as an admirable introduction to logic circuits. The authors lead the reader gently, but firmly, into Boolean algebra, the algebra of sets, and binary arithmetic, and show how these are applied in the study of computer and other logical circuits. The book finishes with a discussion of modulator and demodulator circuits.

No mathematical knowledge higher than that of elementary calculus is assumed, but the reader is also expected to be familiar with Kirchoff's Laws. Numerical examples have been introduced wherever necessary to illustrate the principles developed.

1000-Mc OSCILLATORS AND AMPLIFIERS

by P. E. Kolk and T. J. Robe

Electronic Components and Devices, RCA.

OSCILLATORS

Transistor oscillators using the 2N2708 and 2N2857 rf silicon transistors have provided good performance well into the uhf range. A 120-Mc crystal-controlled two-doubler local oscillator using these two transistors provided 300 millivolts across 50 ohms at 480 Mc[1].

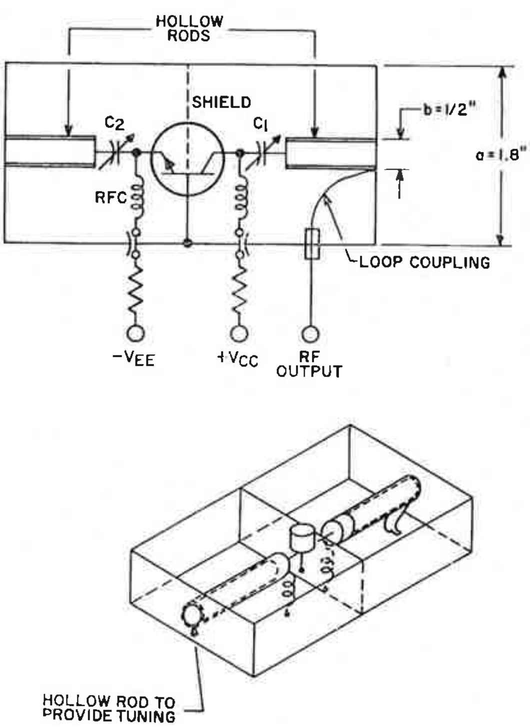


Fig. 1—1000-Mc cavity oscillator using the 2N28

A cavity oscillator using the 2N2857 at 1 Gc is shown in Fig. 1. The box consists of two cavities, one for the input and one for the output. The transistor is connected in the common-base configuration. The variable capacitors resonate with the shorted hollow rods at 1 Gc. The geometry of the rods and the value of tuning capacitance determine the frequency of oscillation. The hollow rods provide access for tuning the variable capacitors, which are fastened to the rods in the enclosed cavity. The design of a cavity oscillator using common-base y parameters at 1 Gc is described below.

Typical common-base y parameters for the 2N2857 at 1 Gc with a collector-to-base voltage of 12 volts and a collector current of 12 milliamperes are as follows :

$$y_{ib} = (6 + j2) \text{ mmhos}$$
$$y_{fb} = (18 + j15) \text{ mmhos}$$
$$y_{rb} = -(1 + j7) \text{ mmhos}$$
$$y_{ob} = (5 + j16{\cdot}5) \text{ mmhos}$$

The use of Johanson JMC 2902 (1-to-10-picofarad) capacitors in the collector cavity dictates the use of a hollow rod having an inner diameter of $\frac{3}{8}$ inch and an outer diameter of $\frac{1}{2}$ inch. The ratio a/b (refer to Fig. 1) must be at least 3·6, where b is the outer diameter of the rod. Therefore, dimension a must be 1·8 inches.

The characteristic impedance Z_o is given by

$$Z_o = \frac{377}{2\pi} \log_e \ 3{\cdot}6 = 76{\cdot}6 \text{ ohms}$$

If the susceptance term in y_{ob} is used, the transistor shunt capacitance in the collector tank is 2·5 picofarads ($j\omega C_b = j\ 16{\cdot}5$ mmhos or

X_{Cb} = 60·6 ohms at 1 Gc). The effective tank capacitance is the series combination of the transistor capacitance and C^1. Because C^1 is approximately equal to the transistor capacitance, the effective tank capacitance is about 1·25 picofarads.

Then,

$$\omega L = 1/\omega C = 120 \text{ ohms}$$
(reactance of 1·25 picofarads at 1 Gc)

and

$$\frac{\omega L}{Z_o} = \frac{120 \text{ ohms}}{76 \cdot 6 \text{ ohms}} = 1 \cdot 57$$

The free-space wavelength λ at 1 Gc is 30 centimeters. The length l of the shorted output rod is calculated as follows :

$$\omega L = X_L = Z_0 \tan \frac{2\pi}{\lambda} l \qquad (1)$$

and

$$l = \frac{\lambda}{2\pi} \arctan \frac{\omega L}{Z_o} = \frac{30 \text{ cm}}{360^\circ} \arctan 1 \cdot 57$$

$$= \frac{30}{360^\circ} (57 \cdot 5^\circ) = 4 \cdot 8 \text{ cm or } 1 \cdot 89 \text{ inches.}$$

The input cavity is calculated in the same way. From y_{ib}, $j\omega C_{ib}$ = 5 mmhos or 200 ohms (0·9 picofarad at 1 Gc). The effective tank capacitance is one-half this value, or 0·45 picofarad. This value requires a Johanson JMC 1802 capacitor (0·4 to 8 picofarad). The input hollow rod has an inner diameter of $\frac{3}{8}$ inch and an outer diameter of $\frac{1}{2}$ inch. As in the output circuit, these dimensions yield a box height of 1·8 inches.

The inductive reactance ωL is 4000 ohms. From Eq. (1), the length of the shorted input rod is calculated to be 2·6 inches. This oscillator will provide 20 milliwatts at 1 Gc with a collector-to-base voltage of 14 volts and an emitter current of 14 milliamperes. The transistor case is clipped to the shield to provide a heat sink.

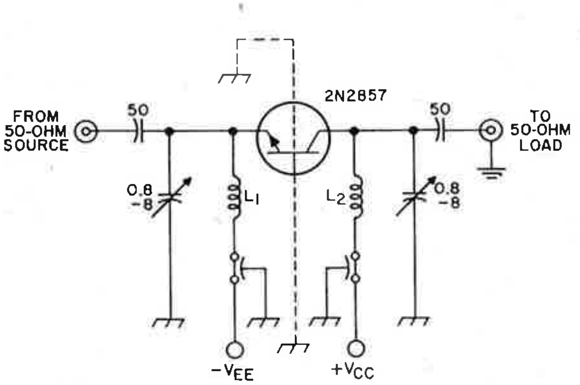


Fig. 2—1000-Mc common-base amplifier using the 2N2857.

1000-Mc COMMON-BASE AMPLIFIERS

In certain amplifier applications, the common-base configuration may be preferable to the common-emitter configuration, particularly at frequencies approaching the upper limits of practical usage of a given transistor. The common-base transistor circuit generally has lower reverse transadmittance at very high frequencies than the common-emitter circuit, and can be operated without neutralization at frequencies well beyond those at which neutralization is required in common-emitter operation. In addition the common-base circuit is normally regenerative when operated without neutralization, and provides much higher gains than the common-emitter circuit. Of course, care must be used to insure stable operation of the circuit. The most convenient method of insuring stability at ultra-high frequencies is mismatching of the transistor terminals.

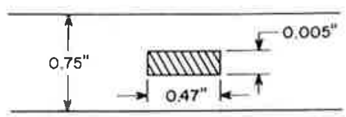


Fig. 3—Strip transmission line having characteristic impedance of 77 ohms.

The following design of a 1000-Mc common-base amplifier using the 2N2857 serves as an example of the design procedure. At a collector-to-base voltage of 6 volts and an emitter current of 1·5 milliamperes, the common-base short-circuit input admittance of the 2N2857 is typically 500 ohms in parallel with 2·5 picofarads. Stability calculations showed that the transistor terminals would have to be heavily mismatched. This loading was accomplished by placing 50-ohm pads at both the input and output terminals of the transistor. Short sections of strip transmission line shorted at one end were used to form the inductive component of the input and output resonant circuits. These lines were tuned with piston trimmer capacitors, as shown in Fig. 2.

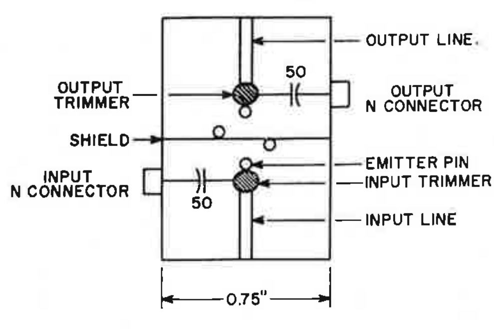


Fig. 4—Top view of 1000-Mc amplifier chassis.

The sections of strip transmission line (L_1 and L_2) were each designed to tune with a capacitance of approximately 5 picofarads. A strip transmission line having the dimensions shown in Fig. 3 has a characteristic impedance of 77 ohms (from curves on p. T-80 of the 1963 Microwave Engineers Handbook). The length of the shorted section of the line is determined by the value of reactance which it must reflect at 1000 Mc. To tune 5 picofarads, this reactance must be 32 ohms, or 0·42 on a normalized basis with respect to the 77-ohm Z_0 of the line. The peripheral coordinates of a conventional Smith chart show that a line length of 0·062 λ is required to reflect a normalized reactance of 0·42 from a short. Since the wavelength at 1000 Mc is 11·8 inches, the line length must be

$$0{\cdot}062\lambda \times (11{\cdot}8 \text{ in}/\lambda) = 0{\cdot}73 \text{ inch}$$

In order to reduce stray wiring reactances to an absolute minimum, the emitter and collector leads of the transistor socket were soldered directly to the two edges of the transmission-line sections, and the base and case leads of the socket were soldered to the shield passing between the collector and emitter. A sketch of the chassis viewed from the top is shown in Fig. 4.

The insertion gain of the circuit shown in Fig. 2 was invariably higher than the transistor maximum available gain (MAG) calculated from the y parameters. This difference indicated that, in spite of the very low terminations of 50 ohms, the circuit was still regenerative, although good-quality responses were noted when the circuit was examined with a sweep generator.

1 P. E. Kolk and T. J. Robe, "Design of a Low-Noise VNF Receiver Using RCA Silicon Planar Transistors", IEEE *Transactions on Vehicular Communications*, March 1964.

(With acknowledgements to RCA)

CORRECTION

Data given last month on the 11LP4 and 25LP4 TV picture tubes is ambiguous in respect of the front panel and implosion cap. The 11LP4 faceplate is filterglass, with no bonded cap. The 25LP4 faceplate is filterglass, with a clear implosion window bonded onto the face of the tube.

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