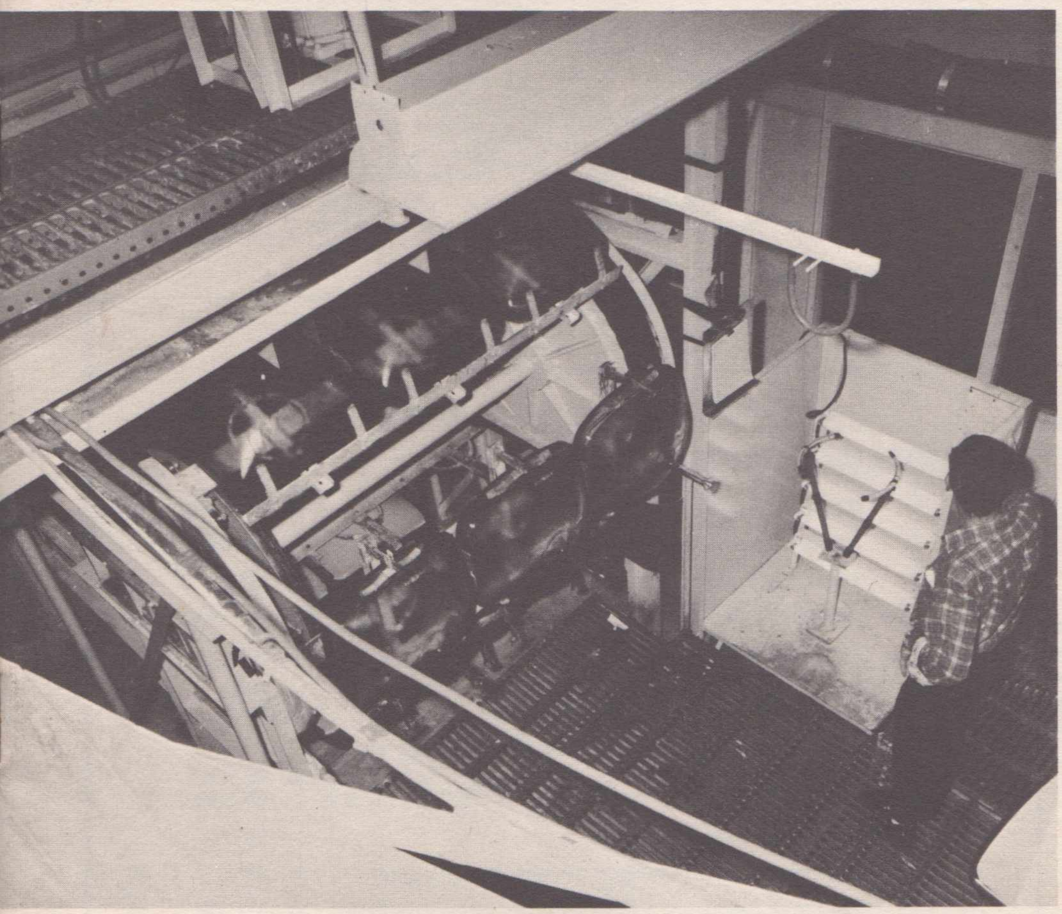
# RADIOTRONICS



## IN THIS ISSUE



## COVER

Picture Tube Settling Belt at AWV Plant Rydalmere N.S.W.


Vol. 34, No. 2, May 1969

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AN AWV PUBLICATION

REGISTERED IN AUSTRALIA FOR TRANSMISSION BY POST AS A PERIODICAL

# 30 WATT STEREO AUDIO AMPLIFIER

by P.B. Kirkwood AWV Applications Laboratory

**This amplifier has been designed to produce a total dynamic power output of 30 watts and can be driven by a standard ceramic pickup cartridge. The function controls included in the design are for volume, balance and a continuously variable bass and treble tone system.**

A low cost audio amplifier using a single ended push pull output stage has been developed for the use in stereophonic radiograms and portable record players. The amplifier is capable of producing 10 watts continuous or 15 watts peak at 10% distortion of audio power per channel into a suitable speaker network of 8 ohms impedance. The amplifier also contains bass and treble tone controls, a balance control and a volume control at the input of the preamplifier.

The design may be changed to provide a reduction in power output by either increasing the output load impedance or by reducing the supply voltage. The latter method will also reduce the overall cost by allowing the use of lower cost transistors in the driver and output stages.

One of the major problems associated with the use of power output transistors in single ended systems is that the collector to base leakage current, $I_{cbo}$, doubles for approximately every 10° C rise in junction temperature. If the driver transistors for the output stage are to be kept reasonably small, then their circuit impedance will be too high for satisfactory temperature stability of the output stage. The temperature sensitive leakage current, $I_{cbo}$, of the output transistors must not be allowed to affect the base to emitter junction of the same transistor, as this will, by transistor action, increase the collector current by the product of $I_{cbo}$ and the direct current gain, $h_{FE}$, at $I_{cbo}$. This will, if the temperature is allowed to increase sufficiently, cause destructive thermal runaway.

## Performance Specifications

| | | | |
|---|---|---|---|
| Dynamic Power output (two channel operation) | | = | 30 watts |
| Single Channel Operation | | | |
| Dynamic Power output (1) | | = | 15 watts |
| Rated power output (at onset of clipping) | Pc | = | 10 watts |
| Total Harmonic Distortion at Pc | | = | 0.5% |
| Sensitivity for Pc (2) | | = | 700 mV |
| Input impedance | | = | 1.2 MΩ |
| Noise below 10 watts - open circuit input | | = | 80dB |
| - short circuit input | | = | >80 dB |
| Frequency Response (relative to 6 watts at 1 KHz) | | | |
| | L.F. - 3dB | = | 40 Hz |
| | H.F. - 3dB | = | 27 KHz |
| Quiescent current (cold) | | = | 15 mA |
| Maximum Permissable Ambient Temperature | | = | 45°C |
| Thermal Resistance (junction to ambient) | 40050* | = | 8°C/W |
| | AS208 | = | 150°C/W |

*Output transistors to be mounted on a piece of 16 gauge aluminium having a total surface area of 32 sq. inches per channel.

Note

(1) Dynamic power output measured at 10% total harmonic distortion with a zero impedance power supply.

(2) If a higher sensitivity is required, then the feedback resistor and capacitor ($R_f$ and $C_f$) can be adjusted to suit. Reducing the value of $R_f$ increases the sensitivity by reducing the feedback, but the overall distortion will increase. The value of $C_f$ is chosen so that its reactance at the -3dB low frequency point is equal to $R_f$.

## ERRATA

**Volume 34 No. 1**
**New Generation Audio Amplifiers.**

(i) Figure 1 – Caption should read "Full size template of copper side of board as viewed through board from component side."

(ii) Figure 2 – R8 should be R47 and vice versa.

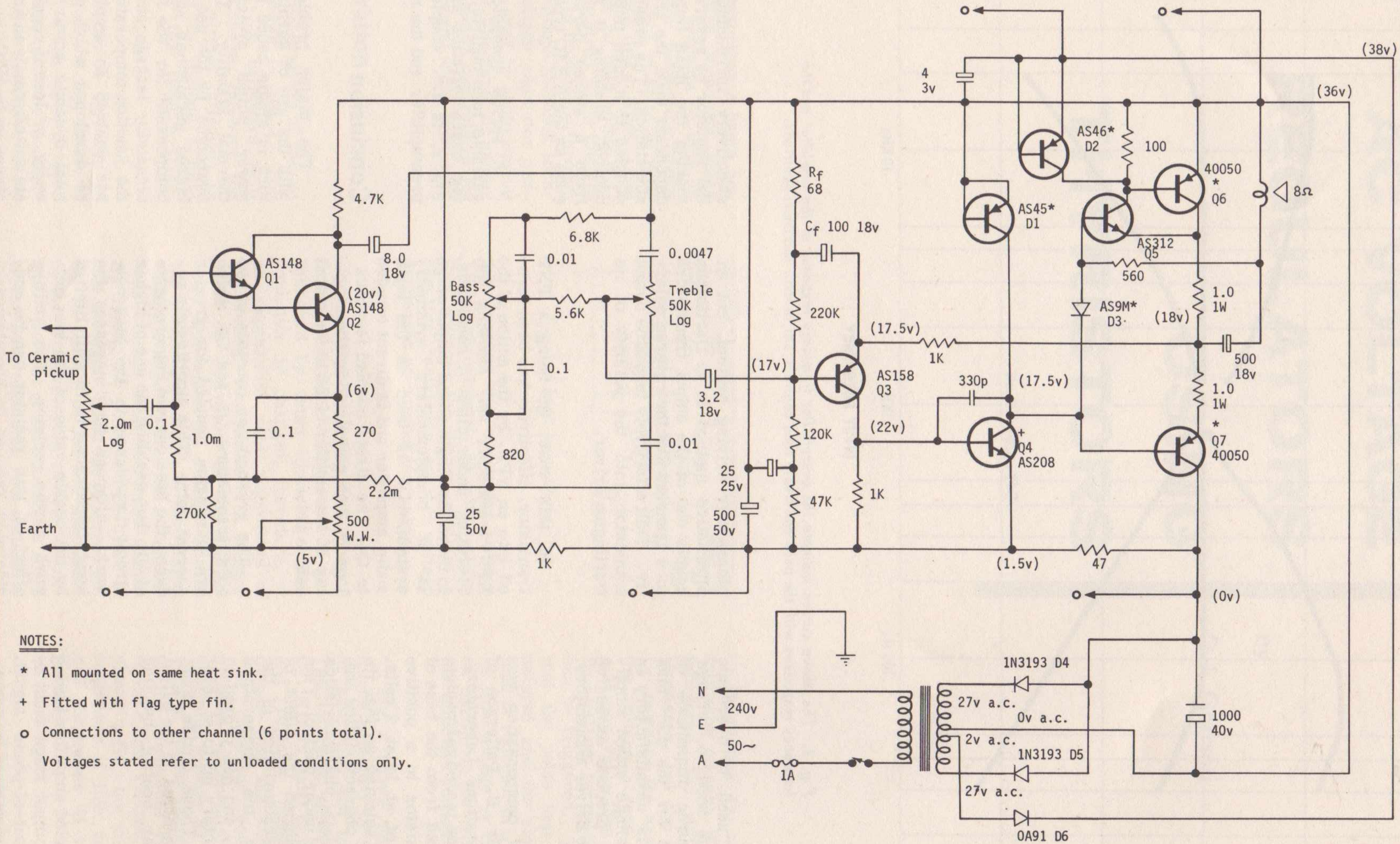
(iii) Figure 4C – R8 should be R47.

(iv) Table 4 is for each output transistor.

(v) Figure 7 – 2N5320 should be 2N2102 – C3 should be 1.6 F.

Suitable circuit boards are available from R.C.S. Radio Pty. Ltd., 651 Forest Road, Bexley ... 2207

## 30 WATT STEREO AUDIO AMPLIFIER




NOTES:

* All mounted on same heat sink.

\+ Fitted with flag type fin.

o Connections to other channel (6 points total).

Voltages stated refer to unloaded conditions only.

*Fig. 1. The above circuit diagram is for one channel only and indicates the six connections to the second channel.*

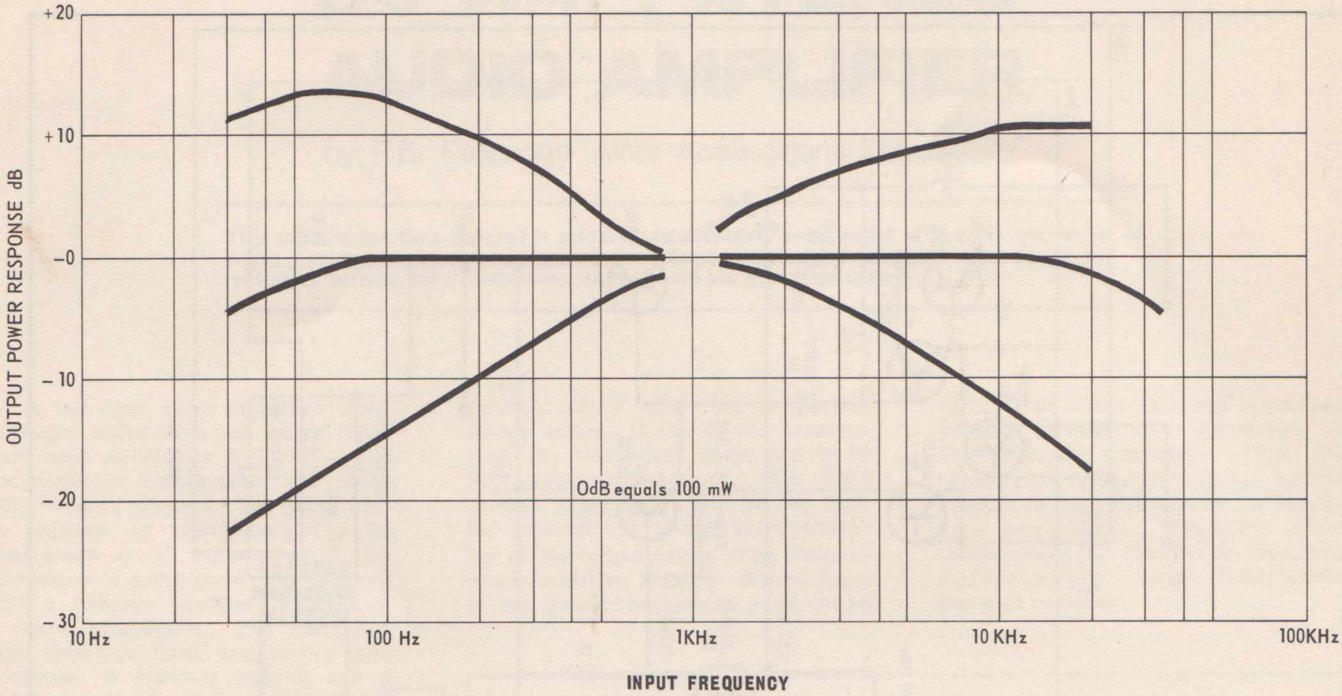



*Fig. 2. The above curves indicate the overall flat frequency response of the amplifier and the frequency responses with the bass and treble tone controls set in their maximum positions.*

The method used to overcome this problem has been to arrange a very low source impedance of 8 ohms or less for the d.c. bias network. This is unsatisfactory in direct coupled single ended amplifiers as it will produce excessive dissipation in the driver transistors.

A method of compensation was developed by Mr. B. Patterson of the AWV Applications Laboratory, whereby a reverse biased germanium diode is connected from the base of the output transistor to a positive voltage source of at least 2 volts, the diode being selected so that its reverse leakage characteristics are similar to the $I_{cbo}$ characteristics of the transistor under control. This considerably reduced the effect of $I_{cbo}$ on the characteristics of the transistor over the temperature range concerned, and has proved to be quite satisfactory up to at least 45°C (113°F). These diodes must be thermally connected to the output transistors under control by flag heat sinks.

A direct coupled amplifier having a single ended output stage must be driven by an out-of-phase signal. This is normally obtained by a complementary driving system, but an alternative method is to Darlington couple one of the output transistors to a transistor of the reverse polarity. This effectively turns the output transistor into the polarity of its Darlington driver.

In this case, Q6 being a PNP transistor, effectively has been turned into an NPN, by the action of the NPN transistor, Q5, allowing Q4, operating under class A conditions, to drive the single ended output stage as a complementary symmetry system. Q3 is used as the input audio amplifier and is direct coupled to Q4. The direct coupled feedback loop is then completed by the coupling of the emitter of Q3 to the output centre point.

The preamplifier consists of two NPN transistors Q1 and Q2, which are Darlington coupled for greater current gain. This significantly reduces the base current requirements of Q1, thus enabling the use of higher circuit impedance in the bias network. Although this increases the input impedance, it is still limited by the usable value of the bias network series resistor. The limiting effect of this resistor can be considerably reduced by bootstrapping, effectively increasing the value of the resistor by several times and so making the input impedance mainly dependent upon the transistor characteristics. The preamplifier is a.c. coupled to a high impedance volume control, enabling it to be operated from a stereo ceramic cartridge such as the B.S.R. type C1.

A series connected continuously variable tone control, whose operating characteristics are indicated in Fig. 2, is a.c. coupled between the preamplifier and the main amplifier.

## Conclusion

The major problem associated with the use of germanium transistors in single ended push pull audio power output systems has been thermal stability. This has been overcome by the use of a reversed biased germanium junction diode to compensate for the increase in the transistor leakage current, $I_{cbo}$, as the junction temperature rises. This has enabled an amplifier design to be developed which will provide a total dynamic power output of 30-watts, or less if required, and operate satisfactorily in ambient temperatures up to 45°C.

# AC VOLTAGE REGULATORS USING THYRISTORS

by

G.J. Granieri

This Note describes a basic ac-voltage regulating technique using thyristors that prevents ac rms or dc voltage from fluctuating more than ± 3 per cent in spite of wide variations in input line voltage. Load voltage can also be held within ± 3-per cent of a desired value despite variations in load impedance through the use of a voltage-feedback technique. The voltage regulator described can be used in photocopying machines, light dimmers, dc power supplies, and motor controllers (to maintain fixed speed under fixed load conditions).

## Circuit Operation

The schematic diagram of the ac regulator is shown in Fig. 1. For simplicity, only a half-wave SCR configuration is shown; however, the explanation of circuit operation is easily extended to include a full-wave regulator that uses a triac.

The trigger device $Q_1$ used in Fig. 1, a diac such as the 40583, is an all-diffused three-layer trigger diode. This diac exhibits a high-impedance, a low-leakage-current characteristic until the applied volt-

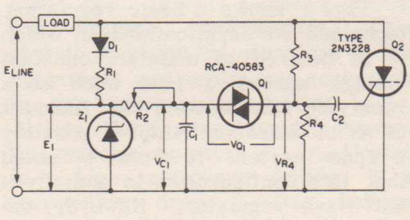



*Fig. 1 A basic ac regulator.*

age reaches the breakover voltage $V_{BO}$, approximately 35 volts. Above this voltage, the device exhibits a negative resistance so that voltage decreases as current increases.

Capacitor $C_1$ in Fig. 1 is charged from a constant-voltage source established by zener diode $Z_1$. The capacitor is charged, therefore, at an exponential rate regardless of line-voltage fluctuations. A trigger pulse is delivered to the 2N3228 SCR, $Q_2$, when the voltage across capacitor $C_1$ is equal to the trigger voltage of diac $Q_1$ plus the instantaneous voltage drop developed across $R_4$ during the positive half-cycle of line voltage. When $Q_1$ is turned on, $Q_2$ is turned on for the remainder of the positive cycle of source voltage. Control of the conduction angle of the SCR regulates rms voltage to the load.

Regulation is achieved by the following means: When line voltage increases, the voltage across $R_4$ increases, but the charging rate of $C_1$ remains the same; as a result, the voltage across $C_1$ must attain a larger value than required without line-voltage increase before diac Q1 can be triggered. The net effect is that the pulse that triggers $Q_2$ is delayed and the rms voltage to the load is reduced. In a similar manner, as line voltage is reduced, $Q_2$ turns on earlier in the cycle and increases the effective voltage across the load.

Fig. 2 shows the voltage waveforms exhibited by the ac regulator at both high and low line voltage. The charging voltage for capacitor $C_1$, $E_1$, is equal to the zener voltage and remains constant up to the instant that the SCR is turned on. The capacitor voltage, $V_{C1}$, increases exponentially because the charging voltage $E_1$ is constant. The voltage across resistor $R_4$ conforms to the sinusoidal variations of the 60-Hz line voltage. At any given phase angle, the voltage across $R_4$ increases if line voltage increases and decreases if line voltage decreases.

The diac and SCR both trigger

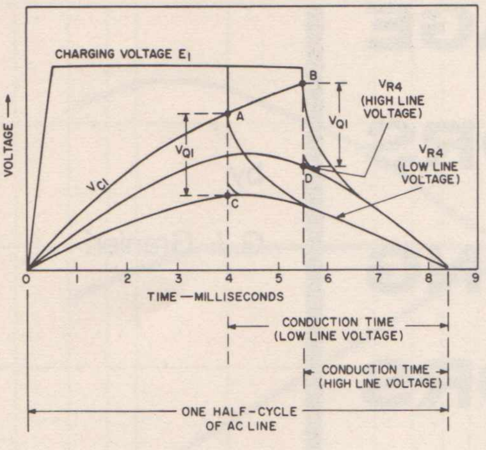



*Fig. 2. Voltage waveforms exhibited by the ac regulator in Fig. 1.*

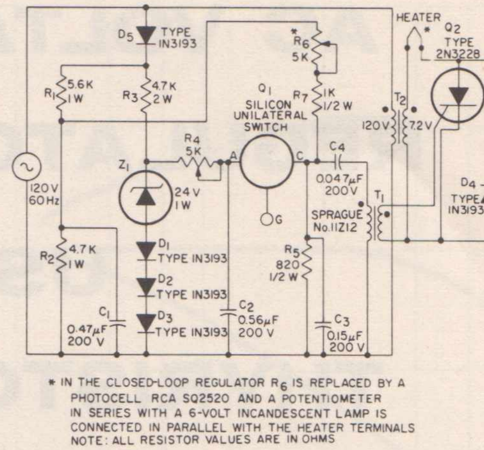



*Fig. 3. A circuit using a regulator to maintain voltage constant across a load.*

when the capacitor voltage, $V_{C1}$, equals the breakdown voltage of the diac plus the instantaneous value of voltage developed across $R_4$ during the positive half-cycle of line voltage. This capacitor voltage is represented by points A and B for the low and high line-voltage conditions, respectively. The instantaneous voltages across $R_4$ just before the SCR is triggered are represented by points C and D for the low and high line-voltage conditions, respectively. The voltage difference between points A and C and between points B and D is equal to the breakdown voltage of the diac.

Fig. 2 illustrates that the conduction time of the SCR is decreased as line voltage increases, and is increased when the line voltage decreases. By proper selection of the values of the voltage-divider-ratio resistors $R_3$ and $R_4$, it is possible to prevent the load voltage from varying more than 3 per cent with a 30 per cent (approximate) change in line voltage.

It should be mentioned that during measurements of load voltage careful consideration must be given to the measuring instruments. Most of the circuits described in this Note produce a non-sinusoidal voltage across the load; the rms value of this voltage can be measured only with a true rms meter, such as a thermocouple meter. It is possible, however, that in certain applications the low input impedance of the thermocouple meter might load down the circuit being measured. In such cases, a high-input-impedance rms meter may be required.

## Heater Regulation

Fig. 3 shows a basic regulating technique for applications in which it is desired to maintain constant voltage across a load such as a receiving-tube heater, the filament of an incandescent lamp, or possibly a space heater. It should be noted that this configuration is actually a half-wave regulator. However, the circuit of Fig. 3 differs from the circuit of Fig. 1, in which one half-cycle is phase-controlled to provide regulation. In Fig. 3, essentially full voltage is applied to the load for one half-cycle by means of $D_4$; the other half-cycle is phase-controlled by the SCR to provide regulation.

The circuit in Fig. 3 is an open-loop regulator that features a high degree of safety; i.e., an open- or short-circuited component does not result in an excessive load voltage. Phase-controlled voltage regulation is provided by a silicon unilateral switch $Q_1$* and a control circuit, as follows: Capacitor $C_2$ is charged from a voltage source that is maintained constant by zener diode $Z_1$; diodes $D_1$, $D_2$, and $D_3$ compensate for the change in zener voltage with temperature. The voltage across $C_2$ increases until the sum of the breakover voltage of $Q_1$ and the instantaneous voltage across $R_5$ is exceeded. At this point, a positive pulse is coupled into the gate of $Q_2$ by means of the pulse transformer $T_1$. The SCR $Q_2$ then switches on for the remainder of the positive cycle of line voltage. Control of the conduction angle of the SCR varies rms voltage to the heater.

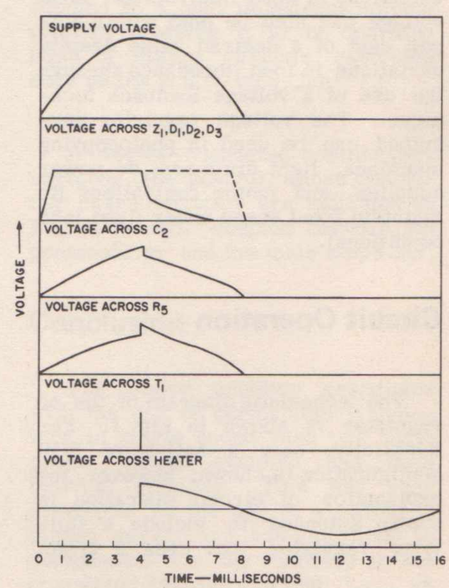



*Fig. 4. Voltage waveforms exhibited by the circuit of Fig. 3.*

As line voltage increases, the voltage across $R_5$ also increases; because $C_2$ charges along the same

* A silicon unilateral switch is a silicon, planar, monolithic integrated circuit that has thyristor electrical characteristics closely approximating those of an ideal four-layer diode. The device shown switches at approximately 8 volts.

exponential curve, however, the voltage across $C_2$ must attain a larger value before $Q_2$ is turned on. The net effect is a delay in the trigger pulse and reduced rms voltage across the heater. In a similar manner, as line voltage is reduced, the SCR turns on earlier in the cycle and increases the effective voltage across the heater. By proper adjustment of potentiometer $R_6$ in conjunction with potentiometer $R_4$, it is possible to obtain excellent heater-voltage compensation over a range of line voltages. Fig. 4 shows the waveforms associated with the heater-regulator circuit.

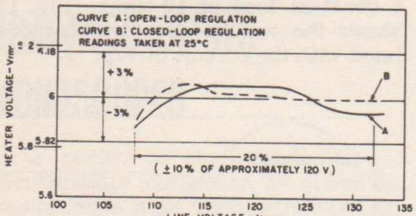



*Fig. 5. Heater voltage as a function of line voltage of the open and closed-loop regulators.*

Curve A in Fig. 5 shows heater voltage as a function of line voltage for the open-loop regulator circuit shown in Fig. 3. Curve B in Fig. 5 shows a similar curve for a closed-loop regulator using a lamp-photocell module. The lamp, in series with a limiting resistor, is connected across the heater terminals, and the photocell replaces $R_6$. The lamp unit senses the phase-controlled true rms heater voltage. Changes in lamp brightness produced by heater-voltage variations change the photocell resistance in reverse proportion to the lamp voltage. The remainder of the circuit functions as previously described except that regulation is obtained not only through the monitoring of the instantaneous magnitude of line voltage, but also through the sensing of the true rms voltage across the heater. This characteristic identifies the circuit as an ac voltage regulator with closed-loop feedback control. The closed-loop regulator produces less error, is more resistant to the drift effects of components, and is easier to adjust than the open-loop regulator.

The lamp used in the closed-loop regulator is rated at 6 volts, but the series resistor limits the voltage to approximately 2 volts so that extremely long lamp life can be expected. An additional advantage at low voltage is that the light intensity varies linearly with the voltage across the lamp so that a small increase in voltage increases brightness markedly; near rated voltage the intensity does not vary linearly and the variation in brightness is not very apparent. A loss in sensitivity would result if the lamp were operated at its rated voltage.

The open-loop regulator can regulate 6 volts to within ± 3 per cent within a temperature range from 10 to 40°C with an input-voltage swing of ± 10 per cent. The closed-loop regulator can regulate 6 volts to within ± 2 per cent within a temperature range from 0 to 60°C with an input-voltage swing of ± 10 per cent.

## Light Dimmer with Over-Voltage Clamp

Light-dimmer circuits are becoming increasingly popular for home use. Fig. 6 shows a typical light-dimmer configuration. This circuit provides the advantage of low hysteresis and continuous control up to the maximum conduction angle. At low illumination levels, however, the variable resistor $R_p$ is adjusted to a high resistance setting. If a momentary drop in line voltage occurs at this condition, the high breakover voltage of the diac in conjunction with the high resistance could result in a circuit misfire; i.e., the light could be extinguished and remain so until the circuit is reset by readjustment of the control to a high illumination setting.

A natural successor to the circuit of Fig. 6 might consist of a configuration which not only provides the light-dimming function but also extends the life of the lamp being controlled. One of the major causes of reduced lamp life can be directly attributed to line-voltage fluctuations and in particular to periods of over-voltage. Nominal line voltage is approximately 120 volts ± 10 per cent; it is the + 10 per cent variation that causes lamps to reach end-of-life prematurely.

A technique for limiting or clamping the lamp voltage, without sacrificing any of the desirable

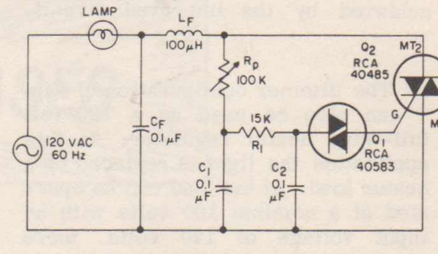



*Fig. 6. A typical light-dimmer circuit.*

features of the dimmer of Fig. 6, is shown in Fig. 7; $L_F$ and $C_F$ suppresses rf interference. Fig. 7 employs the basic regulating circuit described earlier; however, in the configuration shown, the switching voltage of $Q_1$, a silicon bilateral switch*, is reduced by steering diodes $D_1$ and $D_2$ in conjunction with resistor R. This arrangement not only makes it possible to achieve larger conduction angles, but also prevents the circuit from misfiring at low illumination levels when it is subjected to dips in line voltage. The light-dimmer circuit in Fig. 7 is capable of clamping the high-line-voltage condition to within + 3 per cent of its nominal value; as a result, the lamp is subjected to voltages of 120 volts plus 3 per cent and minus 10 per cent. The -10 per cent line dip has little effect on lamp-life reduction.

The circuit also regulates lamp voltage for various settings of potentiometer $R_p$. Fig. 8 shows line voltage as a function of lamp voltage for two settings of $R_p$ for the circuits of Figs. 6 and 7. These curves illustrate the increased regulation

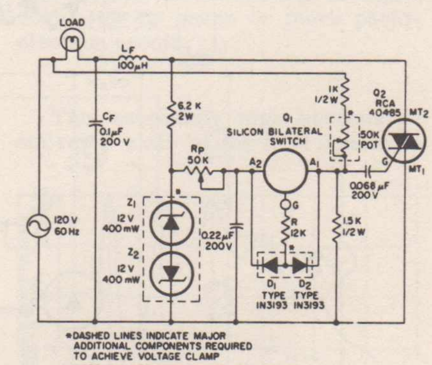



*Fig. 7. A light-dimmer circuit that includes clamping.*

* A silicon bilateral switch is a silicon, planar, monolithic integrated circuit that switches at approximately 8 volts in both directions.

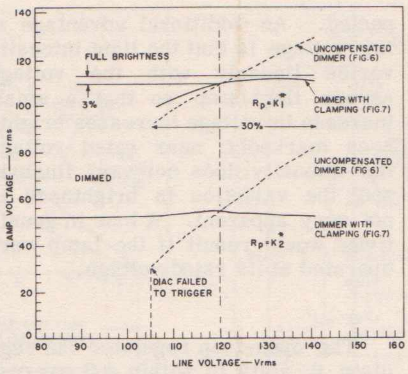
achieved by the improved circuit.

The dimmer configuration of Fig. 7 can also be used as a 120-volt full-wave heater regulator. In this application the light is replaced by a heater load. If the load can be operated at a nominal 100 volts with an input voltage of 120 volts, more symmetrical regulation can be realized; i.e., ± 3 per cent regulation can be achieved with a line variation of ± 10 per cent. In the full-wave heater-regulator application, diodes $D_1$, $D_2$, and resistor R in Fig. 7 can be eliminated because a wide conduction angle is not required.

Such a control might also be used in colorimetry, an application in which it is necessary to match the color (and temperature) of a lamp with a standard; in this application line-voltage fluctuations can create a measurement error. Other areas of application, such as photography, heater control, and hot-plate and solder-pot control, can also make effective use of the dimmer circuit with over-voltage clamp.

## Voltage-Regulated DC Supply

A simple but stable dc power supply using thyristors is shown in

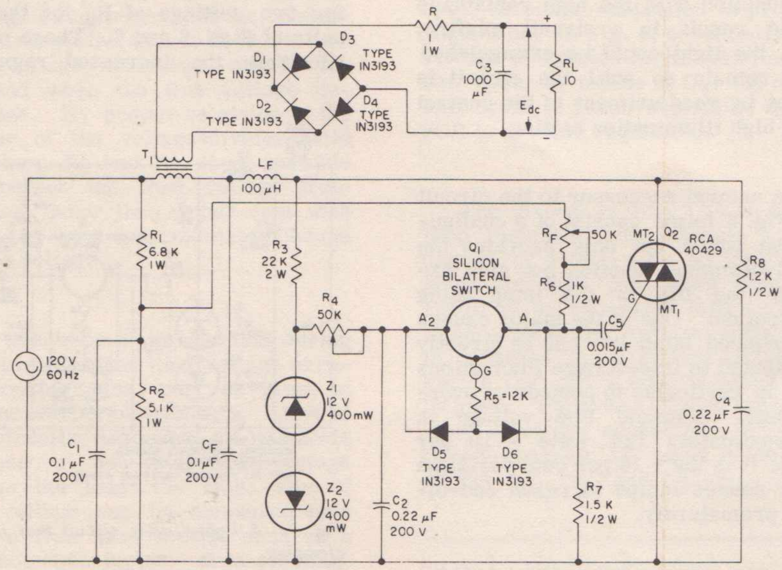



*$K_1$ AND $K_2$ ARE ARBITRARY BUT DIFFERENT VALUES

*Fig. 8. Lamp voltage as a function of line voltage for two values of $R_p$ in the circuits.*

Fig. 9. The power-supply section consists of the well known full-wave bridge with RC filter. A line-voltage transformer is employed to step-down the supply voltage of 120 volts rms to approximately 12.5 volts rms. If a dc output voltage greater than 10 volts is desired, a transformer with a lower primary-to-secondary turns ratio should be employed.

The heart of the regulator shown in Fig. 9 is the phase controlled triac on the primary side of the line transformer. Because the load presented to the triac is somewhat inductive, an RC network is used to assure proper commutation; $L_F$ and $C_F$ suppress rf interference. The




NOTE: ALL RESISTOR VALUES ARE IN OHMS

*Fig. 9. A voltage-regulated dc supply.*

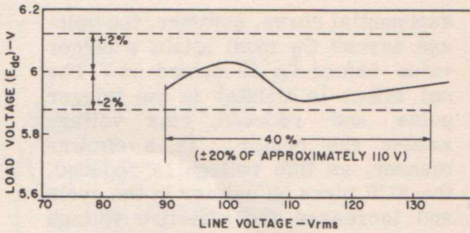



*Fig. 10. Load voltage as a function of line voltage for the circuit of Fig. 9; load resistance is constant at 10 ohms.*

circuit automatically compensates for wide variations in line voltage. Fig. 10 shows a curve of line voltage as a function of load voltage, $E_{dc}$, for a constant load of 10 ohms. Fig. 11 shows the voltage waveforms associated with the circuit of Fig. 9.

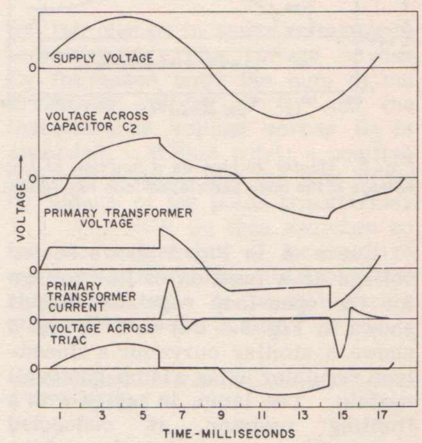



*Fig. 11. Voltage waveforms exhibited by the circuit of Fig. 9.*

If increased line, temperature, and load compensation is desired in the regulated dc supply of Fig. 9, a closed-loop type of control can be obtained by use of a photocell in place of $R_F$ and connection of a lamp across the output terminals of the supply in such a way that the light from the lamp can impinge on the photocell surface.

## Selection of Control Device

Other thyristors than those shown in this Note can also be used for voltage regulation. The selection of an SCR or triac for a particular regulating circuit depends on the voltage and current requirements of the application. Suitable devices are readily available for 240V ac line operation.


• WITH ACKNOWLEDGEMENTS TO RCA.

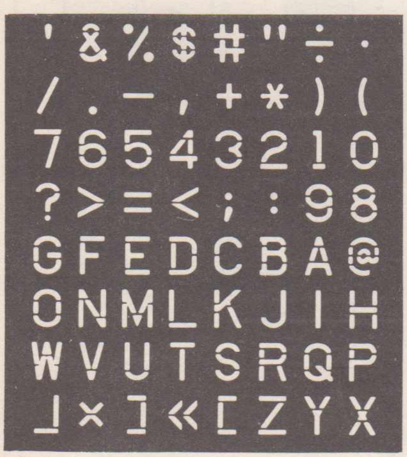
# NEWS & NEW RELEASES

## RCA-4560 SERIES MONOSCOPES

A series of electrostatic focus, electrostatic deflection 2" diameter monoscopes have been introduced by RCA.

The Tubes in the RCA-4560 series are small, 2" diameter electrostatic-deflection monoscopes employing a metallic stencil electrode having a useful area of 1.1" × 1.1". These tubes are designed for use as alpha-numeric character generators in conjunction with display cathode-ray tubes in computer data terminal display systems.

The stencil pattern used in a given tube is individually styled to meet specific customer requirements. A typical 4560 stencil pattern having alpha-numeric characters is shown below. Almost any customised stencil pattern can be readily fabricated and incorporated in the 4560.

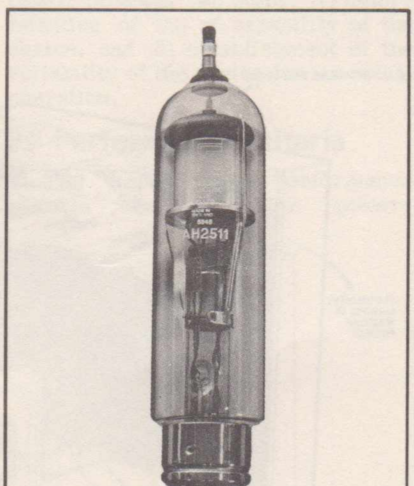



Typical Stencil Electrode Pattern of a 4560 Tube.

## EEV AH2511

The latest addition to the range of hot cathode mercury vapour rectifier manufactured by the English Electric Valve Company is the AH2511.

It is a high-voltage, Half-wave rectifier of similar physical size to the well known AH211A, but with a 50% higher mean output current (3A) and a 5V low current (11.5A) heater.




Maximum ratings for the AH2511 are 15kV peak inverse voltage and 12A peak current. In a three-phase full-wave circuit it will provide a d.c. output ranging from 14.4kV 9A to 2.3kV 15A according to the operating conditions.

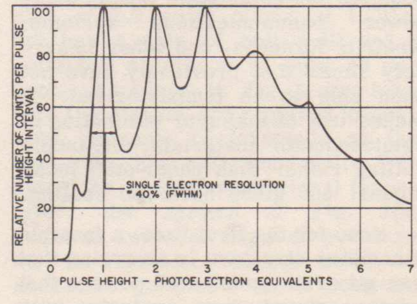
The overall dimensions are length 12.126 in (308 mm) and diameter 2.835 in (72 mm). Net weight is 1 lb. (450 g).

## CD3100D PHOTOMULTIPLIER TUBE

The RCA Developmental Type C31000D is a 12-stage head-on type of photomultiplier tube designed for low-light level measurement applications such as photon and low-energy scintillation counting. It cesiated gallium-phosphide first dynode followed by high-stability copper-beryllium dynodes in the succeeding stages. It will also employ a high quantum efficiency bialkali photocathode deposited on a pyrex entrance windown.

The first dynode of the C31000D provides up to an order of magnitude increase in secondary-emission ratio over conventional dynode materials. This high ratio provides a pulse height resolving capability that permits discrimination of single, double-up to seven or more photoelectron events.

The extremely high secondary-emission ratio of the first dynode is




Typical Photoelectron Pulse Height Spectrum instrumental in providing a decrease in noise induced in signal current by 18 per cent. Noise in signal due to the secondary-emission amplification mechanism of the first dynode is proportional to a factor M, where

$$M = \left[1 + \frac{B}{m - 1}\right]^{1/2}$$

B = A Statistical factor having a value of approximately 1.6

m = The secondary-emission ratio of the dynode, typically 5 for conventional dynode materials, 30 for gallium phosphide at 600 volts.

These characteristics make the C31000D especially suited for the counting of radioactive materials releasing low-energy particles when used in conjunction with suitable scintillators.

# DEVELOPMENTAL 800W TRANSISTOR

A new laminated construction technique has been used to build experimental transistors that, for the first time, rival large electron tubes in power output.

Although the technique, developed by RCA, is still in advanced laboratory development, one of the new superpower transistors has already generated radio waves oscillating at 1-million cycles per second with a power of 800 watts, and considerably higher powers and frequencies are expected.

A new RCA technology responsible for these transistors is a result of a long-term development program, sponsored by the U.S. Air Force and Navy, that may eventually make possible all-solid-state sonar, high-power communications systems, electric furnaces, and other heavy-duty items that previously have not been able to be transistorized. It makes use of fusing or laminating of semiconductor materials, ultrasonic cutting rather than photo-etch techniques, and glass hermetic sealing.

Now, for the first time, a feasible transistor structure is emerging that can generate substantial powers that have previously been exclusive with electron tubes.

# SLEE CLEAN ROOMS

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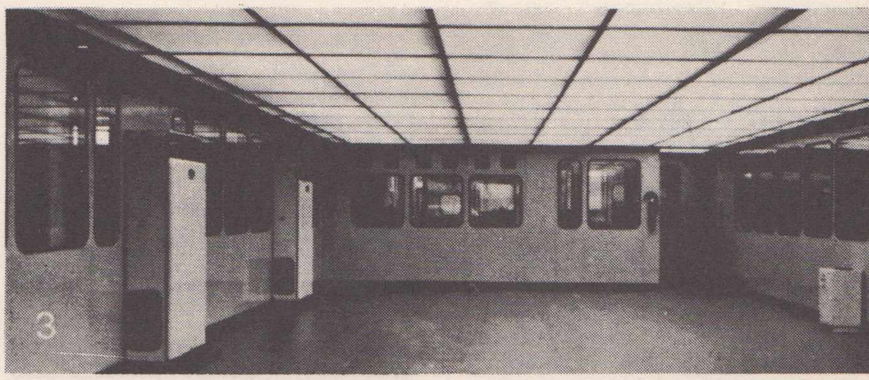
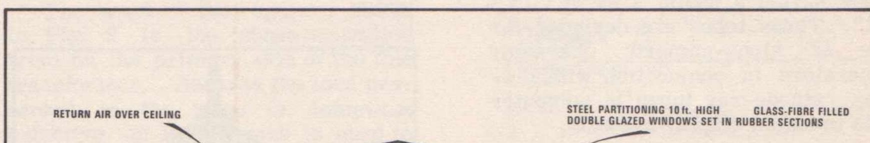



*Typical SLEE Clean Room with AB2 Ultronaire Conditioning*




*Shown above is a typical SLEE Laminar Clean Room.*

# UHF POWER GENERATION USING RF POWER TRANSISTORS

by H.C. Lee

One major usage of rf power transistors is in uhf/microwave power generation. RF power transistors are widely used for both narrowband and broadband power amplification. Transistors suitable for power amplification must be capable of delivering power efficiently with sufficient gain at the frequency band of interest. The usefulness of an rf power transistor is not measured by its power-frequency product or its emitter geometry, but rather by its ability to meet cost limitations and over-all performance objectives including reliability requirements in a given application or circuit.

This Note discusses the use of rf power transistors in high-power generation that uses multiple transistcrs, pulse operation, and broadband power amplifiers. Operational principles and design approaches for these applications are presented, and practical and reliability aspects are discussed. The selection of an rf power transistor for a given application involves two steps: (1) determination of the rf capability of the device, and (2) establishment of the reliability of the device for its actual operation.

## PF Performance Criteria

The important rf performance criteria in transistor power-amplifier circuits are power output, power gain, efficiency, and bandwidth. State-of-the-art single overlay transistors, as shown in Fig. 1, can now produce cw power as follows:

| Frequency (MHz) | Power (W) | Gain (dB) | Efficiency (%) |
|---|---|---|---|
| 76 | 100 | 7 | 90 |
| 400 | 50 | 6 | 70 |
| 1200 | 10 | 10 | 50 |
| 2300 | 7 | 6 | 40 |

When transistor performances are compared, it is important to consider gain and efficiency, as well as power output and frequency, because additional gain can be achieved only at the expense of collector efficiency with the use of additional transistors. For example, Fig. 2 demonstrates the use of two transistors which have the same power output, but different gain and collector efficiency. The high-gain unit shown in Fig. 2(a) is capable of delivering an output of 2.5 watts at 1 GHz with a gain of 10 dB and a collector efficiency of 50 per cent. The low-gain unit shown in Fig. 2(b) is also capable of 2.5 watts output at 1 GHz, but has a gain of only 5 dB and a collector efficiency of only 30 per cent. As shown in Fig. 2, two low-gain transistors are required to provide the same performance as the high-gain, high-efficiency unit. Besides the use of an additional transistor, the system of Fig. 2(b) requires twice as much dc power as that of Fig. 2(a). In this case, the additional gain of 5 dB is achieved at the expense of 5.9 watts of dc power. From the practical point of

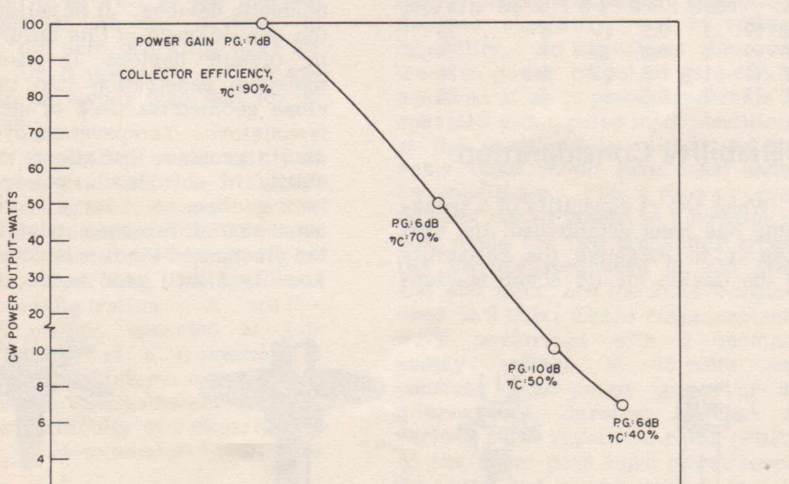



*Fig. 1. State-of-the art power output of single rf power transistor as a function of frequency.*

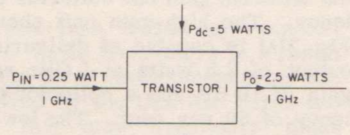
view, the system of Fig. 2(b) is more complex, and the dissipation of the output transistor is higher.

**TABLE 1 – Inductances of Packages shown in Fig. 3.**

| Package | Lead Inductances - nH $L_e$ | $L_b$ |
|---|---|---|
| TP - 39 | 3 | 3 |
| TO - 60 (isolated emitter) | 3 | 3 |
| TO - 60 (grounded emitter) (2N5016) | 0.6 | 2 |
| HF - 19 (hermetic stripline) | Approx. Same | |
| HF - 11 (coaxial case) (2N5470) | 0.1 | 0.1 |

## Package Considerations

The package is an integral part of an rf power transistor. A suitable package for uhf applications should have good thermal properties and low parasitic reactance. Package parasitic inductances and resistive losses have significant effects on circuit performance characteristics such as power gain, bandwidth, and stability. The most critical parasitics are the emitter and base lead inductances. Table I gives the inductances of some of the more important commercially available rf power-transistor packages. Photographs of the packages are shown in Fig. 3. The TO-60 and TO-39 packages were first used in devices such as the 2N3375 and the 2N3866. The base and emitter parasitic inductance for both TO-60 and TO-39 packages is in the order of 3 nanohenries; this inductance represents a reactance of 7.5 ohms at 400 NHz. If the emitter is grounded internally to a TO-60 package (as in the 2N5016), the emitter lead inductance can be reduced to 0.6 nanohenry.



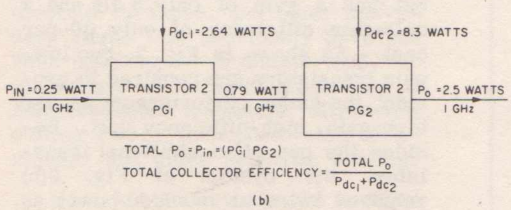




*Fig. 2. A comparison of one-and two-transistor systems that have the same output power but different gain and collector efficiencies.*

The plastic stripline package (used in the 2N5017) has an emitter lead inductance of 0.4 nanohenry and a base lead inductance of 0.6 nanohenry. The main advantage of the rf plastic package is that a substantial reduction in parasitic inductance is achieved because the emitter and base leads can be placed closer to the transistor chip. Hermetic low-inductance radial-lead packages are also available. The HF-19 package introduced by RCA utilizes ceramic-to-metal seals and has rf performance comparable to that of an rf plastic package. The parasitic inductances can be reduced further in a hermetic coaxial package. The HF-11 package used in the 2N5470 has parasitic inductances in the order of 0.1 nanohenry.

Table II compares the performance of the TO-39 package, the HF-19 hermetric stripline package, and the HF-11 coaxial package with the same transistor chip. At a frequency of 1 GHz and an input power of 0.3 watt, the coaxial package performs significantly better than either the stripline or the TO-39 package. The coaxial package results in an increase of output power by a factor of two as compared to the TO-39 package. In addition, the coaxial-package transistor is capable of delivering an output of more than 1 watt with a gain of 5 dB at 2 GHz. A well-designed coaxial package outperforms any other rf package currently available.

**TABLE II Package Inductances with same transistor chip.**

Using Same Transistor Chip

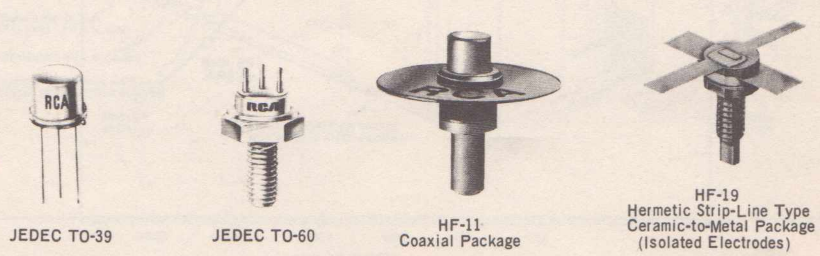
| | f-GHz | Pin-W | Po-W | P.G.-dB | c(28 V)-% |
|---|---|---|---|---|---|
| TO-39 | 1 | 0.3 | 1 | 5 | 35 |
| HF-19 | 1 | 0.3 | 1.5 | 7 | 45 |
| HF-11 | 1 | 0.3 | 2.2 | 8.6 | 50 |
| HF-11 | 2 | 0.3 | 1 | 5 | 35 |

## Reliability Consideration

When the rf capability of a transistor has been established, the next step is to establish the reliability of the device for its actual application. The typical acceptable failure rate for transistors used in commercial equipment is 1 per cent per 1,000 hours (10,000 MTBF); for transistors used in military and high-reliability equipment, it is 0.01 to 0.1 per cent per 1000 hours. Because it is not practical to test transistors under actual use conditions, dc or other stress tests are normally used to simulate rf stresses encountered in class B or class C circuits at the operating frequencies. Information derived from these tests is then used to predict the failure rate for the end use equipment. The tests generally used to insure reliability include high-temperature storage tests, dc and rf operating life tests, dc stress step tests, burn-in, temperature cycling, relative humidity, and high humidity reverse bias. The end-point measurement for thest tests should include collector-to-emitter voltage $V_{CEO}$, in addition to the common end points collector-to-emitter current $I_{CEO}$, collector-to-base voltage $V_{CBO}$, collector-to-emitter saturation voltage $V_{CE}(sat)$, power output, and power gain.

One of the common failure modes in uhf/microwave power transistors is degradation of the emitter-to-base junction. The high-temperature storage life test and the dc and rf operating life tests can excite this failure mode. The failure mode can be detected by measurement of $V_{EBO}$, which is not included in most life-test end-point specifications.

Plastic uhf power transistors are more sensitive to emitter-to-base-junction degradation than similar hermetic devices. It is believed that the enhancement of this failure mode in plastic devices is caused by moisture penetration into the very close geometries used in uhf power transistors. Temperature cycling is also a problem that affects the reliability of uhf plastic power transistors because large thermal-expansion differences exist between the plastic and the fine bonding wires (usually 1 mil) used in the devices.




JEDEC TO-39 JEDEC TO-60 HF-11 Coaxial Package HF-19 Hermetic Strip-Line Type Ceramic-to-Metal Package (Isolated Electrodes)

*Fig. 3. Commercially available rf power transistor packages.*

UHF power transistors are complex electrical, thermal, chemical, and mechanical systems. The well-designed uhf power transistor is a systems solution to the integration of these parameters. It appears that the plastic environment is a less viable solution to this systems problem than a hermetic approach. Although a plastic environment has been an excellent systems solution for low-frequency and vhf power transistors, in which much larger bonding wires, metallic strips, and rugged device geometries are used, it is not a completely satisfactory solution for uhf power transistors.

## Safe-Area Curves for RF Operation

The important parameters of a transistor which are directly related to reliability and rf performance include rf breakdown voltages, thermal characteristic, and load-mismatch capability.

Although a safe-area curve to avoid second breakdown on the collector-current—vs—collector-to-emitter voltage ($I_C$ - $V_{CE}$) plane can be established for forward-bias or class A operation, such a curve for class B, class C, or pulsed operation is difficult to define because the breakdown voltages under rf conditions are considerably higher than the dc breakdown voltages, and the thermal resistance is a function of $V_{CE}$ and $I_C$. The safe operating area for class B or C conditions at rf frequencies is a function of these parameters, as well as the thermal time constant of the device. In general, the safe operating area for class C or B operation can be expected to be higher than that for dc conditions.

VSWR capability, or the ability of an ef power transistor to withstand a high VSWR load, is another important consideration. VSWR capability is a function of frequency of operation, operating voltage, and circuit configuration. A well-designed circuit operated at low supply voltage at a frequency at which power gain is not excessive is less prone to VSWR mismatch. Four modes of difficulty are experienced in the load-mismatch test, as follows:

(1) slow thermal failure as a result of low rf swing and very poor efficiency;

(2) high-speed failure as a result of the high positive peak value of rf swing;

(3) an instability (non-destructive) which occurs because the high value of $V_{CE}$ causes avalanching (such a condition in the common-emitter configuration produces a negative resistance characteristic and results in a spurious signal generator);

(4) an instability caused by the negative overswing which can severely forward-bias the collector-base junction and trigger a low-frequency oscillation which resembles a motor-boating or squelched oscillation.

Additional work is required for further characterization of transistor parameters, as related to VSWR capability, rf breakdown, and safe operating area.

## Pulse Operation of RF Power Transistors

A large potential application for rf power transistors is in pulse equipments such as DME (distance measuring equipment), CAS (collision avoidance system), and radar. The ratio of peak to average or cw power obtainable with a transistor is much less than that which can be obtained with a vacuum tube because a transistor is a current - amplification device, while a vacuum tube is a voltage-amplification device. The ability of an rf power transistor to deliver higher pulsed output power than cw power depends on the transistor current-handling capability, thermal capability, and rf voltage capability. No significant improvement in power output or gain can be achieved if an rf power transistor is operated under pulse input conditions at the same supply voltage and the same input power level used under cw conditions. Fig. 4 shows curves of peak output power as a function of duty cycle for two transistor types: the 2N5016 measured at 225 MHz and 400 MHz, and the 2N5470 measured at 2 GHz. These measurements were performed with a constant supply voltage of 28 volts and constant input-power pulses of 5-microsecond duration applied at various pulse repetition rates (PRR). At the same peak input power level, the gain and power output remain approximately the same for duty cycles ranging from 100 per cent (cw) down to 0.1 per cent.

Fig. 5 shows the 2 GHz amplifier shown in Fig. 4. The 2N5470 transistor is placed in series with the center conductor of the line, or cavity, and its base is properly grounded to separate the input and output cavities. The input section consists of a 20 ohm line section and a capacitance $C_1$. The output section consists of a 36 ohm line section and capacitances $C_2$ and $C_3$. Direct coupling is used at both input and output. Fig. 6 shows the 400-MHz lumped-element amplifier circuit used for the 2N5016 pulse measurements.

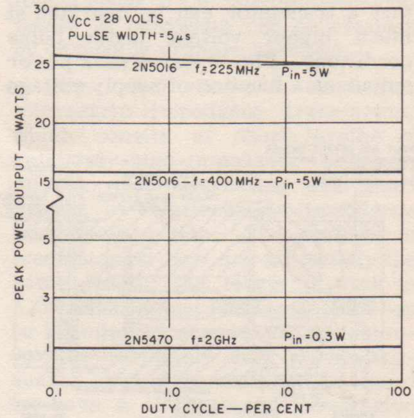



*Fig. 4. Peak output power as a function of duty cycle for the 2N5016 and 2N5470 transistors at selected frequencies.*

circuit used for the measurements

The major difference between cw and pulse operation, however, is that the input drive level can be increased substantially under pulsed input conditions. Fig. 7 shows peak power output as a function of duty cycle for the 2N5470 at a frequency of 2 GHz and a constant supply voltage of 28-volts with input power as a parameter. Under cw operation in the 2-GHz amplifier circuit shown in Fig. 5, an increase of input power from 0.3 to 0.5 watt does not result in an increase of power output, i.e., the power output seems to be saturated at 1.1 watts. However, under pulsed input conditions of 5 microsecond pulse duration and 10 per cent duty cycle, the output power increases substantially from 1.1 watts to 1.9 watts as the input power increases from 0.3 to 0.7 watt. These requirements indicate that the power input to the 2N5470 transistor at 2 GHz under cw conditions is limited by thermal capability rather than by peak current or periphery. This transistor appears to be capable of operating at much higher peak current under pulse conditions than would be permissable under cw conditions. This improvement is possible because the pulse duration of 5-microseconds is probably smaller than the thermal time constant of the transistor, and the junction temperature is more a function of average device dissipation than of peak dissipation. A similar improvement in peak power output and gain can be obtained by pulse operation of the 2N5016 at 225 MHz, as shown in Fig. 8, but the improvement is not as great as that obtained for the 2N5470.

A second major difference between cw and pulse operations is that a transistor can be operated at much higher voltage under pulse conditions. Fig. 9 shows peak power output as a function of supply voltage

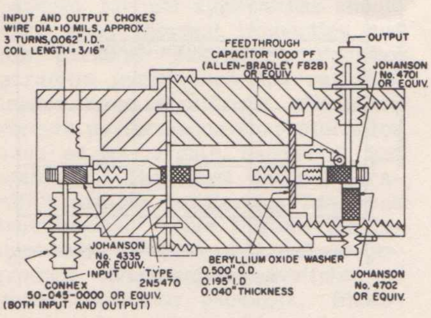
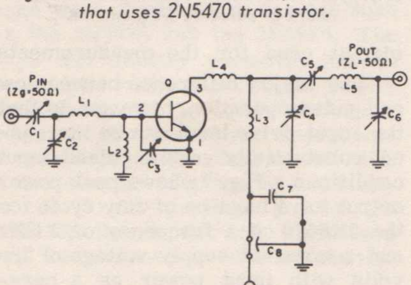



*Fig. 5. A 2-GHz coaxial amplifier circuit that uses 2N5470 transistor.*




$C_1$ = 1 to 10 pF, piston capacitor
$C_2$,$C_3$,$C_4$,$C_5$,$C_6$ = 1 to 30 pF, piston capacitors.
$C_7$ = 0.01 F, disc, ceramic
$C_8$ = 1000 pF, feedthrough
$L_1$ = 1/4 inch O.D. copper tubing; 1-1/4 inches long
$L_2$ = 12 H, choke
$L_3$ = 0.27 ohm, wire wound
$L_4$ = 1/8 by 1/32 by 5/8 inch long copper strip.
$L_5$ = 1/4 inch O.D. copper tubing, 2-1/4 inches long

Note 1 - $L_1$ and $L_5$ are mounted coaxially within a 1-5/8 by 1-5/8 by 6 inch box.

Note 2 - For optimum performance $C_8$ should be mounted between emitter and base with minimum lead lengths.

*Fig. 6. A 400-MHz amplifier circuit that uses a 2N5016 transistor.*

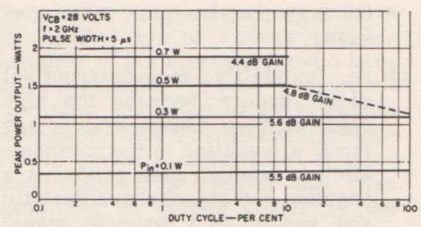



*Fig. 7. Peak output power as a function of duty cycle for the 2N5470 transistor operating at 2 GHz.*

$V_{CC}$ for the same transistor types (the 2N5016 measured at 225 MHz and 400 MHz, and the 2N5470 measured at 2 GHz). These measurements were performed with constant peak input power pulses at 1-per-cent duty cycle and 5 microsecond pulse duration. At an input power level of 0.5 watt, the 2 GHz power output of the 2N5470 increases from 1.9 watts at 28 volts to 2.5 watts at 45 volts. At an input power of 9 watts, the 400-MHz power output of the 2N5016 increases from 25.5 watts at 28 volts to 40 watts at 45 volts. At 225 MHz, the increase in power is even greater. These results indicate that rf power transistors can be operated at much higher voltage under pulse conditions, and, consequently, can deliver more pulsed power. It appears that rf power transistors can withstand much higher voltage under short-pulse conditions without operating in the second-breakdown region. The average current resulting from short-pulse operation is much lower than that of cw operation.

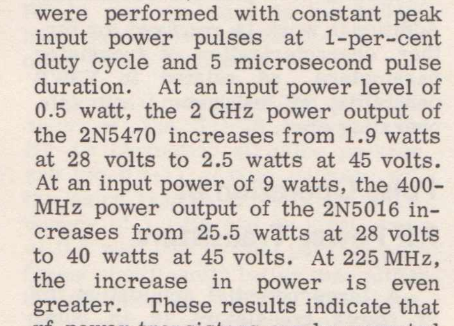



*Fig. 8. Peak output power as a function of duty cycle for pulse operation of the 2N5016 transistor at 225 MHz.*

## Broadband Power Amplifier

RF power transistors are often used in broadband amplifier circuits for commercial and military applications. Transistor transmitters are superior to tube transmitters with respect to broadband capability, re-

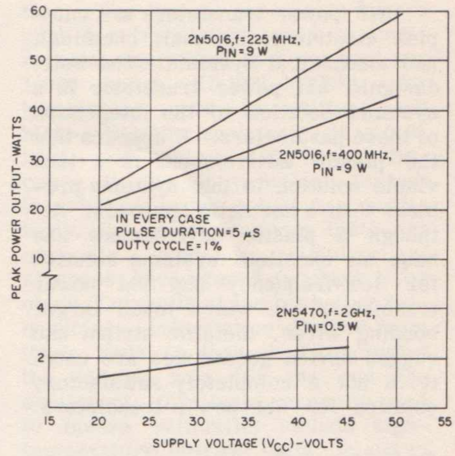



*Fig. 9. Peak output power as a function of supply voltage $V_{CC}$ for the 2N5470 and 2N5061 transistors at selected frequencies.*

liability, size, and weight. The aircraft communication bands of 116 to 152 MHz and 225 to 400 MHz are of interest for both military and commercial applications. Another area of interest is ECM (electronic counter-measures) applications. Transistors suitable for broadband applications must be capable of providing both the required power output within the entire frequency range of interest and constant gain within the bassband. The bandwidth of a transistor power amplifier is limited by the following: intrinsic transistor structure, transistor parasitic elements, and external circuits such as input and output circuits.

## Intrinsic Transistor Structure

The parameters which determine the inherent bandwidth of a transistor intrinsic structure are the emitter-to-collector transit time, the collector depletion-layer capacitance, and the base-spreading resistance. The emitter-to-collector transit time, which represents the sume of the emitter-capacitance charging delay, the base transit time, and the collector depletion-layer transit time, affects the over-all time of response to an input signal. Of particular importance is the emitter-capacitance charging delay, which is current-dependent and equal to $1/f_T$, where $f_T$ is the gain-bandwidth product of the transistor. A high $f_T$ is essential for broadband operation; in addition, a constant $f_T$ with current level is required for large-signal operation. The ratio of the $f_T$ to the product of the base-spreading resistance and the collector depletion-layer capacitance ($r_bC_c$) comprises the gain function of a transistor.

Under conjugate-matched input and output conditions, the power gain as a function of frequency (which is equal to $f_T/8\pi f^2 r_b C_c$) falls off at a rate of 6 dB per octave. In a power amplifier, the power gain usually decreases by less than 6 dB per octave, as shown in Fig. 10(a), because the load resistance $R_L$ presented to the collector is not equal to the output resistance of the transistor, but is dictated by the required power output and the collector voltage swing. The curve in Fig. 10(a) indicates that one approach for achieving a broadband transistor amplifier is to optimize the matching at the higher end of the frequency band and to introduce mismatch in the input, or output, or both at the lower end of the band so that a constant power output is obtained from $f_1$ to $f_2$, as shown in Fig. 10(b). The power output that can be obtained in a transistor broadband amplifier is comparable to that measured at the high end of the band in a narrowband amplifier; efficiency and power gain are slightly lower than in a narrowband amplifier because the load and source impedance cannot be ideally matched to the transistor over a broad frequency band.

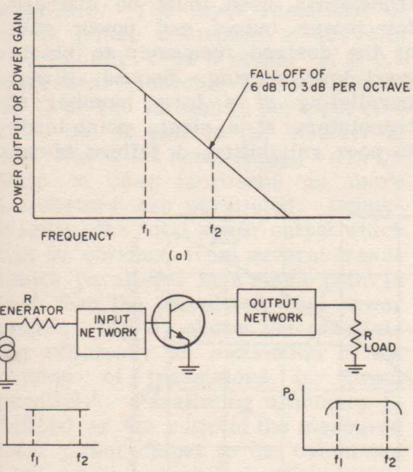



*Fig. 10(a) Output power as a function of frequency in a power amplifier; (b) equivalent broadband amplifier.*

The disadvantage of this approach for broadbanding is the relatively high input VSWR at the low end of the band. A more sophisticated approach for achieving broadband performance is to consider the intrinsic transistor structure, the transistor parasitic elements, and the external circuits as part of the over-all band-pass structure, in which the input and output circuits are coupled together by the transistor feedback capacitance. This combined structure reproduces the power-output or power-gain curve of Fig. 10(a) from $f_1$ to $f_2$. External feedback is then applied to control the input drive to flatten the power output over a broad frequency band.

## Parasitic Limitations

Any discrete transistor contains parasitic elements which impose further limitations on bandwidth. The most critical parasitics are the emitter lead inductance $L_e$ and the base inductance $L_b$. These parasitic inductances range from 0.1 to 3 nanohenries in commercially available rf power transistors. In the simple representation of a common-emitter equivalent transistor input circuit at high frequency shown in Fig. 1, the inductance $L_{in}$ represents the sum of the base parasitic inductance and the reflected emitter parasitic inductance, and $R_{in}$ is the dynamic input resistance. The real part $R_{in}$ is inversely proportional to the collector area and, therefore, the power-output capability of the device; the higher the power output, the lower the value of $R_{in}$. A low ratio of the reactance of $L_{in}$ to $R_{in}$ is important as the first step in broadbanding and for ease of circuit design. Unless the reactance of $L_{in}$ is appreciably lower than the input resistance $R_{in}$, the reactance must be tuned out and thus the bandwidth limited.

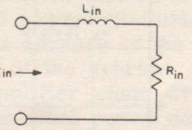



*Fig. 11. Equivalent input circuit of an rf power transistor.*

## External Circuits

For a broadband amplifier circuit to deliver constant power output over the frequency range of interest, a proper collector load must be maintained to provide the necessary voltage and current swings, and the input matching network must be capable of transforming the low input impedance of the transistor to a relatively high source impedance.

Suitable output circuits for broadband amplifiers include constant-K low-pass filters, Chebyshev filters (both transmission-line and lumped-constant), baluns, and tapered lines. Fig. 12(a) shows a conventional constant-K low-pass filter. The input impedance $Z_{11}$ is substantially constant at frequencies below the cut-off frequency $\omega_c = 1\sqrt{L_K C_K}$. A constant collector load resistance can be obtained if the shunt arm (1-1) of $C_K$ is split into two capacitances, as shown in Fig. 12(b); part of the capacitance represents the $C_{ob}$ of the transistor, and the other part has a value which makes the total capacitance equal to $C_K$. Further improvements of bandwidth can be obtained by cascading of more sections.

Fig. 12(c) shows a short-step microstrip impedance transformer which consits of short lengths of relatively-high-impedance transmission line alternating with short lengths of relatively-low-impedance transmission line. The sections of transmission line are all exactly the same length; the length of each is $\lambda/16$. A constant load resistance can be maintained across the collector-emitter terminals over a wide frequency band if the circuit is designed to have a Chebyshev transmission characteristic.[1,2] Fig. 12(d) shows a lumped-equivalent Chebyshev impedance transformer which consists of a ladder network using series inductances and shunt capacitances.

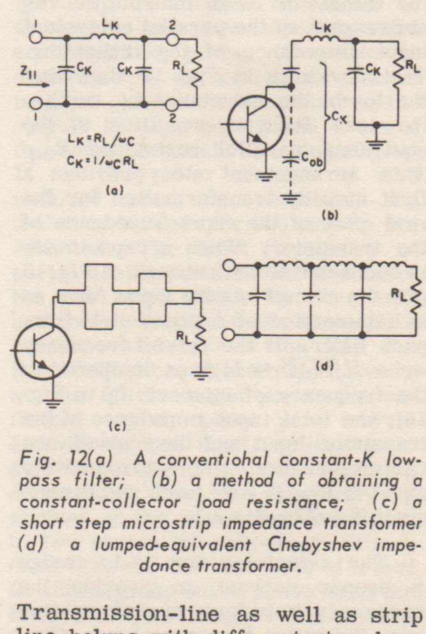



*Fig. 12(a) A conventiohal constant-K low-pass filter; (b) a method of obtaining a constant-collector load resistance; (c) a short step microstrip impedance transformer (d) a lumped-equivalent Chebyshev impedance transformer.*

Transmission-line as well as strip line baluns with different step-down ratios (4:1, 9:1, 16:1) can also be used in the output to provide the broadband impedance transformation.

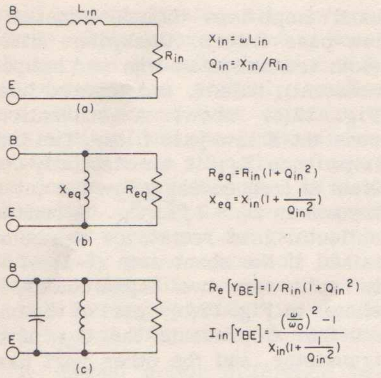



*Fig. 13(a) Series equivalent input circuit of an rf power transistor; (b) equivalent parallel input; (c) equivalent parallel input circuit with external base-emitter capacitance.*

One difficulty in broadbanding a transistor power amplifier is to maintain the desired bandwidth in an input circuit which provides the required impedance transformation from the extremely low input impedance of a transistor to a relatively high source impedance. The design of the input circuit depends on the approach chosen: optimizing the matching at the high end only, or using the transistor parasitic elements as part of a low-pass structure. A simple way of optimizing the matching at the high end is to introduce a capacitance between the base and the emitter terminals of the transistor or to tune out the reactive part of the parallel equivalent input impedance of the transistor. The newworks in Fig. 13 show that the lower the inductance $L_{in}$ or $Q_{in}$, the less fequency-sensitive is the equivalent parallel resistance $R_{eq}$. This arrangement also provides a first step-up transformation for the real part of the input impedance of the transistor. When a capacitance is connected to the network of Fig. 13 (a), the circuit has the same form as a half-section of a constant-K low-pass filter. If the cut-off frequency $\omega_C = 1/\sqrt{L_{in}C}$ is high as compared to the frequency of interest ($f_2$ in Fig. 10), the total input impedance of the transistor input and the capacitance C combination is approximately equal to $R_{in}/(1 - \omega^2/\omega_C^2)$ and is constant if $(\omega^2/\omega_C^2) \angle\angle 1$.

The remaining step is to design a proper network to provide the necessary impedance transformation over the entire frequency band. Circuits suitable for the imput include multi-section constant-K filters, Chebyshev filters, and tapered lines. A more sophisticated approach to obtain a broadband transformation in the input is to treat the parasitic inductance $L_{in}$ of Fig. 1 as part of the transformation network. For example, $L_{in}$ can be considered as one arm of the Chebyshev low-pass filter of Fig. 12(d). For a given bandpass characteristic, the number of sections increases with the value of $L_{in}$. Again, therefore, low package parasitic inductance is important.

## The RCA Dev. No. TA7344 Transistor

At present, plastic uhf power transistors are used exclusively in 225-to-400-MHz broadband applications. UHF plastic packages have substantially lower parasitic inductances than either TO-60 or TO-39 packages, as discussed previously.

The introduction of the RCA hermetic low-inductance stripline package makes it possible to design broadband power amplifiers without compromising reliability. This new radial-lead package utilizing ceramic-to-metal seals is superior to uhf plastic packages in two respects: it has lower parasitic inductances, and it is hermetically sealed. For example, the RCA Dev. No. TA7344 transistor, first in a series of hermetic radial-lead devices, has a dynamic input impedance of 1.5 + j1.2 at 400 MHz. Fig. 14 shows typical curves of power output and efficiency as a function of input power for the TA7344 at a frequency of 400 MHz and a collector-to-emitter voltage of 28 volts. This transistor is capable of delivering an output of 19 to 20 watts with gain of 6.5 dB and collector efficiency approaching 70 per cent at 400 MHz. One important feature of this device is that the power gain is linear within 1.6 dB at power levels between 7 and 20 watts. The TA 7433 is also capable of an output of 20 watts with gain of more than 10 dB at 225 MHz, as shown in Fig. 15.

## High-Power Generation

When more rf power is required than can be provided by a single transistor, combining techniques must be used. Two of the more commonly used methods of combining transistors to obtain high power are: (1) the "brute-force" method of paralleling several transistors at a single point, and (2) the use of hybrids to combine several individual amplifier chains or modules.

RF power transistors can be directly paralleled at a single point, as shown in Fig. 16. All collectors and bases are connected together, and a single input matching circuit and a single output matching circuit are used. Although this arrangement offers circuit simplicity, it has several disadvantages. First, the transistors used must be matched for power output and power gain at the desired frequency to obtain good load sharing. Second, direct paralleling of a large number of transistors at a single point leads to poor reliability; a failure of one

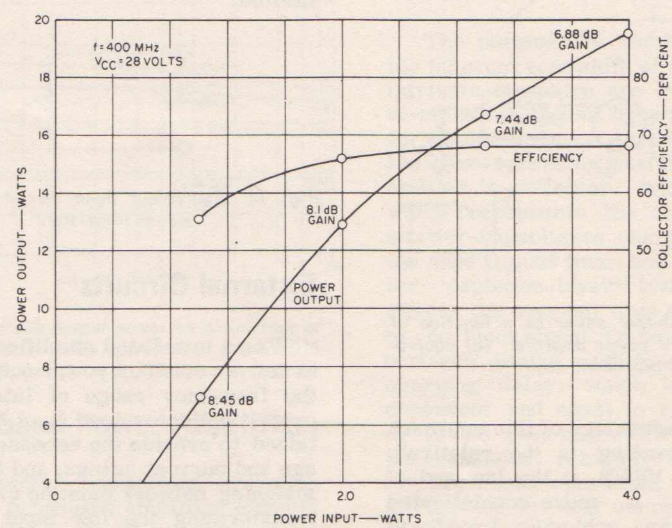
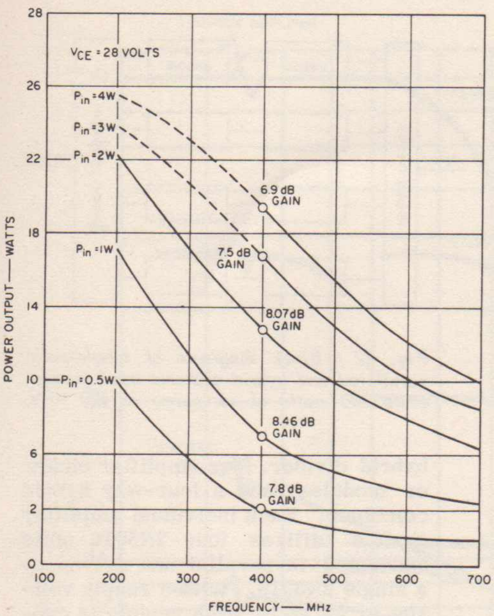



*Fig. 14. Output power and efficiency as functions of input power for the RCA Dev. No. TA7344 at 28 volts.*



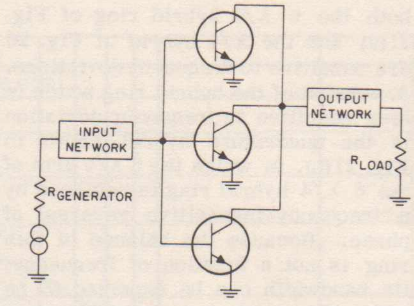




*Fig. 16. A method of paralleling rf power transistors at a single point.*

*Fig. 15. Output power as a function of frequency in the RCA Dev. No. TA344 at 28 volts.*

transistor usually causes a total failure of the over-all amplifier circuit.

Of particular importance is the reduction in both input and output impedances resulting from parelleling transistors. The impedance level can be of the same order as the rf losses in the input and output elements. The input resistance of an rf power transistor at 400 MHz is typically 1 to 5 ohms. If a 0.1 microhenry inductor with an unloaded Q of 150 is used in the input circuit, the rf loss in the inductor at 400 MHz is 1.6 ohms ($R_{loss} = \omega L/Q$). This rf loss increases as more transistors are paralleled. Consequently, the total power output which can be obtained from several transistors paralleled at a single point is less than the calculated total power output. Fig. 17 shows the paralleling efficiency as a function of the number of transistors in direct parellel.[3] Paralleling efficiency is defined as the ratio of the measured total power output to the calculated total power output (i.e., the number of units multiplied by the power output of an individual unit). The paralleling efficiency decreases rapidly as the number of transistors increases. For example, when the 2N5016 is used at a frequency of 400 MHz and a collector-to-emitter voltage of 28 volts, the paralleling efficiency is 95 per cent for two transistors connected in parallel, 90 per cent for three transistors, 85 per cent for four units, and 55 per cent for eight units.

Most of the disadvantages of the "brute-force" direct-paralleling method can be avoided by a more sophisticated approach, shown in Fig. 18, in which several amplifier modules or chains are combined by the use of an input hybrid divider and an output hybrid combiner. This arrangement provides a reliable and efficient method of achieving high vhf/uhf power. Reliable operation results because of the isolating properties of the hybrid. A failure of one amplifier chain or module reduces the total power output, but does not cause failure of the other amplifier chains or modules. In addition, this arrangement provides a highly efficient method of combining vhf/uhf power. Reliable operation results because of the isolating properties of the hybrid. A failure of one amplifier chain or module reduces the total power output, but does not cause failure of the other

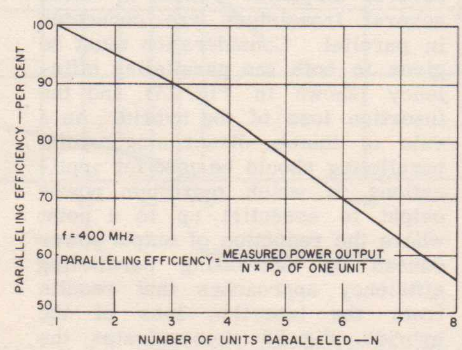



*Fig. 17. Efficiency as a function of the number of transistor in parallel.*

amplifier chains or modules. In addition, this arrangement provides a highly efficient method of combining vhf/uhf power because the insertion loss of a hybrid is small.

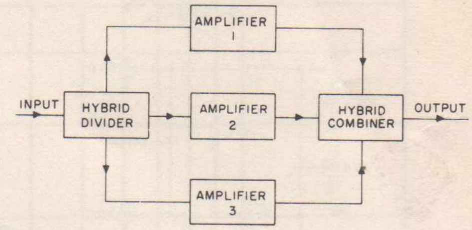



*Fig. 18. Use of hybrids to combine several individual amplifiers.*

A hybrid is an n-port network used as a constant-impedance circuit for power summing and dividing. It maintains phase and amplitude equality between any number of outputs, and also provides isolation between matched outputs. Fig. 19(a) shows a two-way transmission-line hybrid power divider which consists of two quarter-wave transmission lines, each having a characteristic impedance of $Z_o = 2\ R_o$.[4] The generator port 1 and distribution ports 2 and 3 are terminated by resistors $R_o$. A lumped resistor of value $R_o$ is connected from each of the distribution poarts to a common point. When a signal is fed into the power divider (port 1), it divides by virtue of symmetry into two equiphase and equiamplitude ports. No power is dissipated by the resistance R when matched loads are connected to the outputs because port 2 and 3 are at the same potential. The input (port 1) of the power divider is also matched when the conditions for isolation between the two outputs are satisfied. The input impedance of port 1 is the parallel combination of the two output loads $R_o$ after each has been transformed through a auarter-wavelength of the line $Z_o$. If a reflection or mismatch occurs at one of the output ports, the reflected signal splits; part travels directly to the input, splits again, and then returns to the remaining output port. Thus, the reflected wave arrives at the remaining output port in two parts; the path-length difference between the two paths of travel is 180 degrees out of phase when the lines are $\lambda/4$ in length. The value of the resistor R is properly chosen ($R = R_o$) so that the two parts of the reflected wave are equal in amplitude and 180 degrees out of phase; thus, complete cancellation occurs.

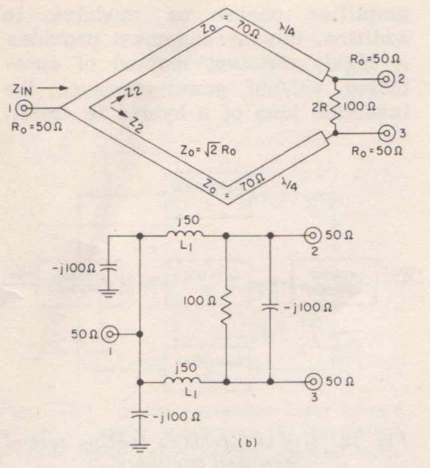



*Fig. 19(a) A two-way, transmission-line, hybrid power divider; (b) a lumped-constant equivalent of this power divider.*

The hybrid shown in Fig. 19(a) can also be used as a two-way combiner (i.e., power introduced at ports 2 and 3 will combine or add at port 1). The lumped equivalent of the quarter-wave transmission-line hybrid is shown in Fig. 19(b).

The technique illustrated in Fig. 19 can be extended to an n-way power divider or combiner, as shown in Fig. 20.[4] The characteristic impedance of each quarter-wave line should have a characteristic impedance of $Z_O = \sqrt{n R_O}$, and the resistor R should have a value of $R_O$.

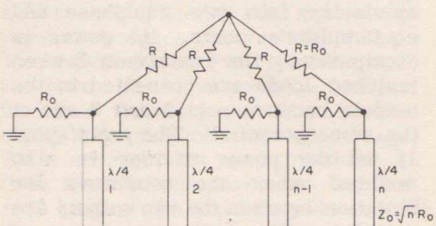



*Fig. 20. N-way, quarter-wave hybrid.*

Fig. 21(a) shows another hybrid, the 6 λ/4 ring. Each port is separated from the adjacent port by a λ/4 section, except for the 3 λ/4 section between ports 3 and 4. Because of this arrangement, power introduced at port 1 appears at equal levels at the adjacent ports (2 and 4), but does not appear at the opposite port 3. In a similar way, power introduced at ports 2 and 4 combines or adds at port 1.

The VSWR and the isolation of both the 6 λ/4 hybrid ring of Fig. 21(a) and the λ/4 hybrid of Fig. 20 are sensitive to frequency deviations. A version of the hybrid ring which is less sensitive to frequency deviation is the quadrature hybrid, shown in Fig. 21(b), in which the 3 λ/4 arm of the 6 λ/4 hybrid ring is replaced by a frequency-insensitive reversal of phase. Because the balance of this ring is not a function of frequency, its bandwidth can be expected to be wide. The quadrature hybrid accepts an input signal at any of its four ports, and distributes half to a second port and half to a third port with 90-degree or quadrature phase difference. The fourth port is isolated.

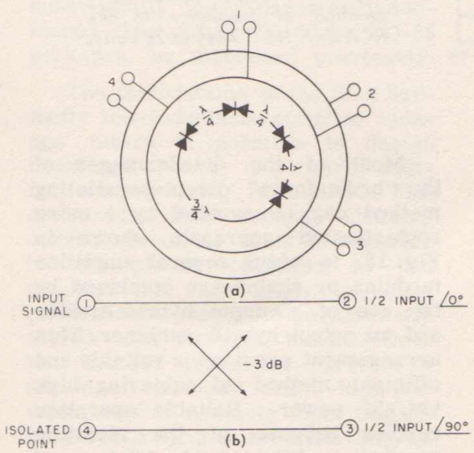



*Fig. 21(a) A 6λ/4 ring hybrid; (b) a quadrature hybrid.*

The choice between hybrids and single-point paralleling for high-power generation depends on the required over-all performance, size, and cost. The most effective system usually employs hybrids to combine several amplifier chains in which several transistors are connected in parallel. Consideration must be given to both the paralleling efficiency (shown in Fig. 17) and the insertion loss of the hybrid. As a rule of thumb, direct single-point paralleling should be used for applications in which maximum power output is essential up to a point where the reduction of output power caused by decreasing paralleling efficiency approaches that results from the insertion loss of the hybrids. Fig. 22 demonstrates the use of such techniques to generate cw power of 200 watts at 400 MHz. The system consists of a four-to-one hybrid divider, four amplifier chains or modules, and a four-way hybrid combiner. Each individual amplifier module utilizes four 2N5016 units connected in parallel and driven by a single 2N5016. With a supply voltage of 28 volts, each module is capable of delivering output power of 54 watts at 400 MHz with gain of 12.4 dB and collector efficiency of 50 per cent. The four-to-one hybrid combines the output of four modules to produce cw power of 200 watts at 400 MHz.

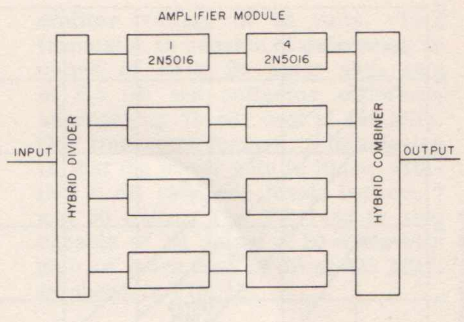



*Fig. 22 Block diagrams of single-point paralleled and hybrid systems used to generate 200 watts of cw power at 400 MHz.*

A similar technique has been used successfully to generate cw power of more than 1000 watts at 400 MHz by use of sixty-four 2N5016 units, and power of 10 watts at 2.3 GHz by use of sixteen 2N5470's.[5] The use of hybrids in conjunction with single-point paralleling has become an accepted technique for generating vhf/uhf high power. Such techniques are now found in practical systems that deliver output power up to 300 watts in the low uhf range.

**FIGURE 23 IS A SELECTION CHART WHICH DETAILS THE PRESENT STATE OF THE ART OF RF POWER TRANSISTOR DEVELOPMENT.**

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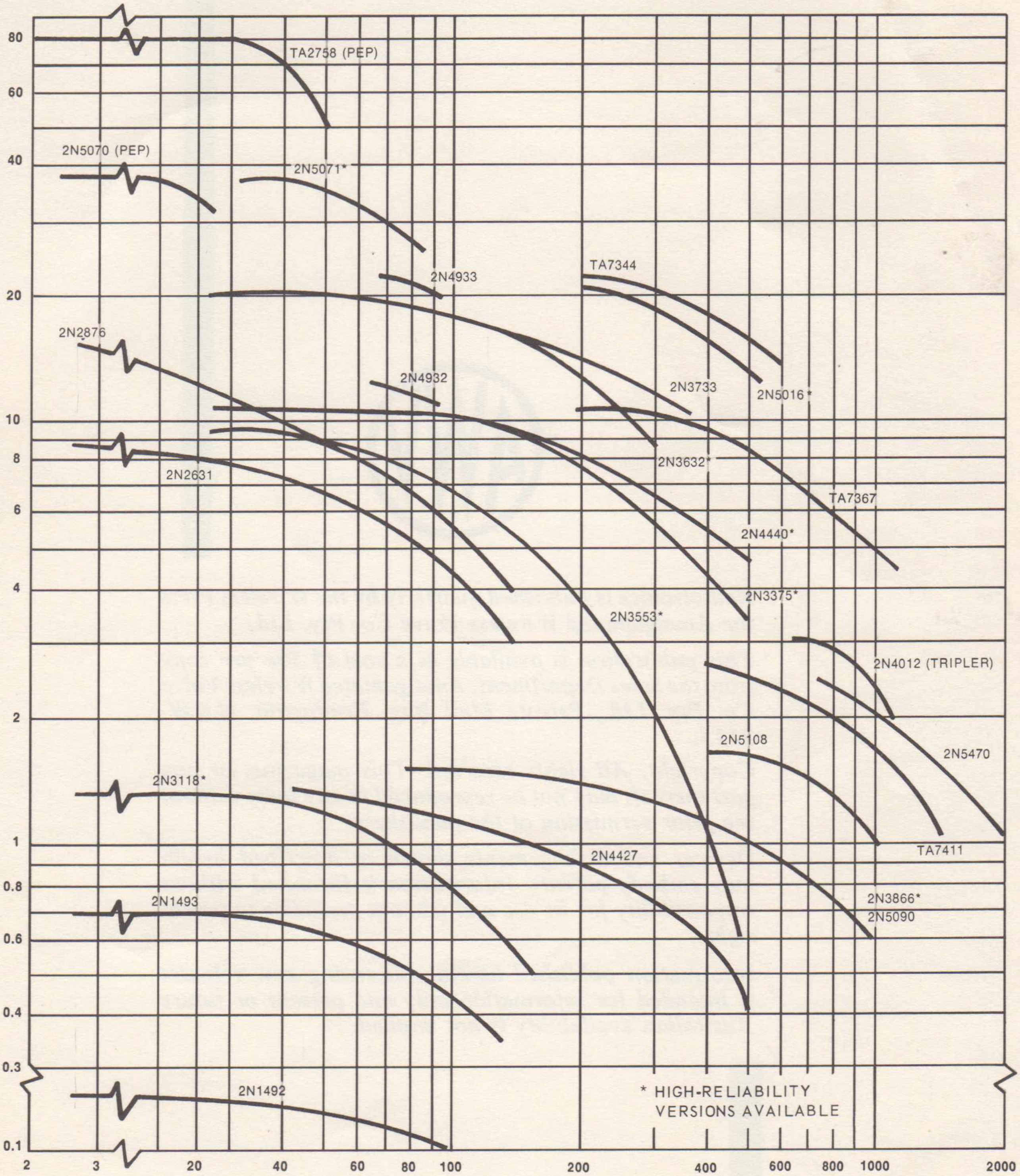


*FIG. 23. – R.F. POWER TRANSISTOR SELECTION CHART.*

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