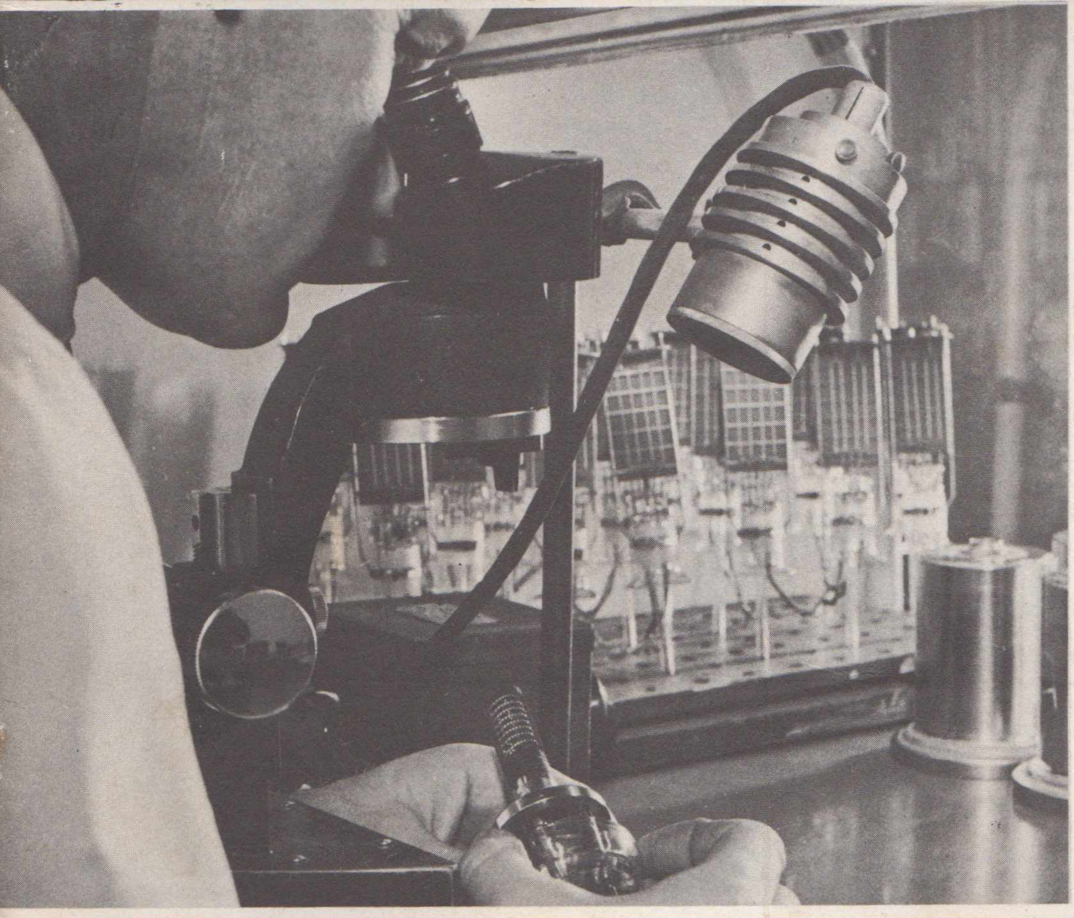
# RADIOTRONICS



## IN THIS ISSUE



## COVER

An operator checking the assembly of an AWV Transmitting Tube.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1


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# HIGH QUALITY PREAMPLIFIER AND TONE CONTROL CIRCUIT Type R105B

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One of the most popular articles published in Radiotronics in recent times was "High Quality Preamplifier and Tone Control Circuit R105" by D.K. Money and E.M. Woodford (July 1965). In the two and a half years that have elapsed since that article was published there have been many advances in the state of the art. The ready availability of silicon transistors suitable for such circuits has been one of the advances.

Using the designs developed by Money and Woodford as a basis, two preamplifiers and a tone control circuit have been designed using silicon transistors. These circuits added to the series of high quality 3 - 130 Watt Audio Amplifiers, described in the May and August 1967 issues of Radiotronics, form excellent high fidelity audio amplifying systems.

Fig. 1

SILICON TRANSISTOR PREAMPLIFIER FOR MAGNETIC CARTRIDGES

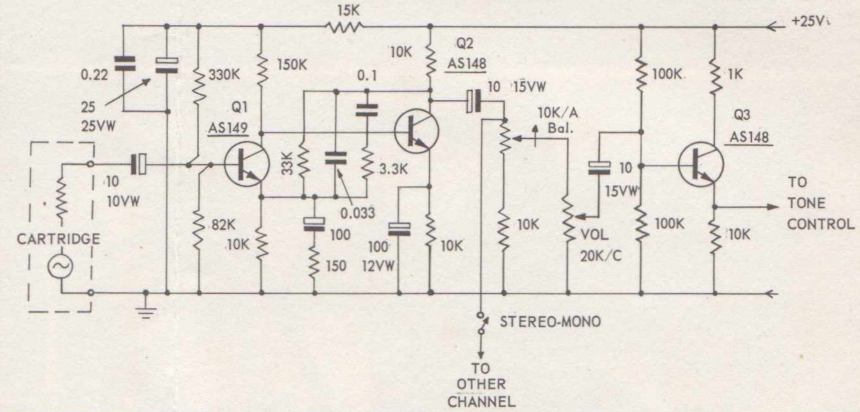
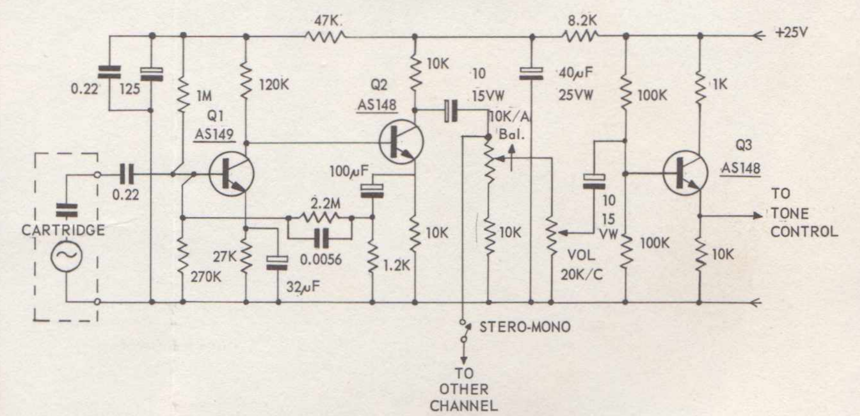



Fig. 2

SILICON TRANSISTOR PREAMPLIFIER FOR CERAMIC CARTRIDGES




## Description:

Two preamplifier circuits are given, the first (figure 1) is for use with magnetic pickups, having an output of approximately 2 millivolts. The second (figure 2) is designed for use in conjunction with ceramic pickups, with a nominal output of 100 millivolts and an internal capacity of approximately 600 pF. These circuits (which include the volume and the balance controls, the balance control being for stereo systems) have been designed to drive the tone control circuit (figure 3).

Both of the preamplifier circuits have a two transistor feedback input system, with inbuilt equalisation which provides the correct type of load for the particular pickup device. The equalisation allows for RIAA recording characteristics and where necessary the inbuilt characteristics of the pickup. Consequently the output signals applied to the balance and volume controls, from the preamplifiers, are substantially constant over the frequency range of the pickup device.

The balance control which immediately follows the preamplifier, consists of a ganged linear potentiometer, with the tracks wired in reverse, so that as the signal of one channel is increased the other is reduced. The range of the balance control is 6dB which is sufficient to provide a smooth transistion from one channel to the other. The insertion loss of this control is 3dB. When used in conjunction with the 'Stereo-Mono' switch, also provided at this point, a desirable silent transition of the effective sound source can be made from one speaker to the other, thus enhancing its use as described in the article "Phasing in Stereophonic Sound Systems" (Radiotronics May 1967).

The circuit of the tone control is also similar to the original, except for the use of silicon transistors and a higher output load providing a higher output source impedance. Although the voltage gain is then more dependent upon the load presented to the amplifier the higher source resistance enables a higher voltage gain to be obtained. The frequency steps and attenuation ratios are the same as the original since no determining components have been changed.

The power supply rails of the original circuit have been reversed to allow the use of the n-p-n silicon transistors. In order to obtain a supply voltage of 25 for the preamplifier and tone control circuits, from the power supply of any of the series of four amplifiers previously described, a suitable series dropping resistance is required. The value of this resistor for each amplifier is given in Table 1.

In order to keep the distrotion of the preamplifier circuits as low as possible, high levels of feedback have been retained. Table 2 gives the performance specifications for the two preamplifiers and the tone control.

The sensitivities stated in Table 2 have been set for the 10-15 watt amplfier series having a high input impedance. For the other amplifiers some change in overall sensitivities will be evident but a small correction can be made by changing R7. Provided that the variation of R7 is not greater than 3:1 the increase in distortion which will result should not effect the overall performance.

TONE CONTROL CIRCUIT
TYPE R105B

Fig. 3

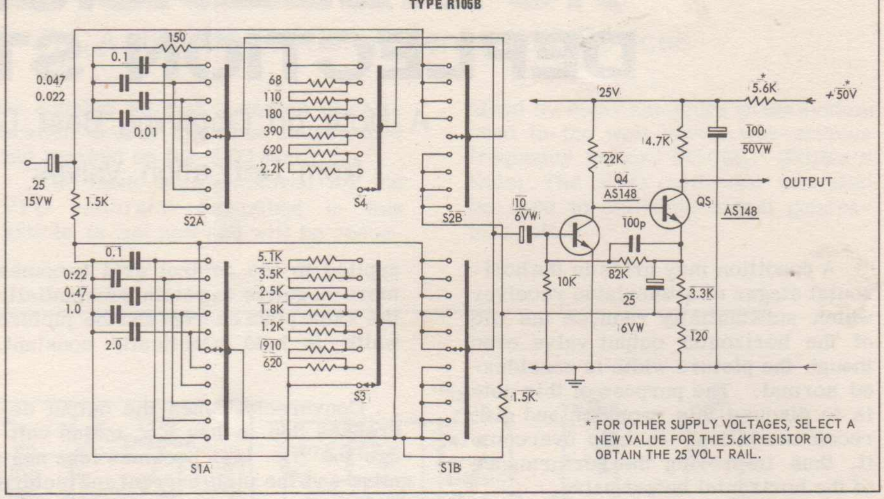



**TABLE 1.**

| Amplifier Design | Series Dropping Resistor | | Total Current |
|---|---|---|---|
| 3 Watts | 2.2 k ohms | 0.5 watt | 4.5 mA |
| 10-15 Watts | 5.6 k ohms | 0.5 watt | 4.5 mA |
| 25-50 Watts | 8.2 k ohms | 0.5 watt | 4.5 mA |
| 70-130 Watts | 12.0 k ohms | 1.0 watt | 4.5 mA |

*TABLE 1 – SERIES DROPPING RESISTOR VALUES TO GIVE A SUPPLY VOLTAGE OF 25*

**TABLE 2.**

| | Magnetic Input | Ceramic Input | Tone Control | |
|---|---|---|---|---|
| Rated Signal Input | 2.0 | 100 | 10 | mV |
| Output Signal | 10.0 | 10.0 | 3000 | mV |
| Harmonic Distortion | 0.02 | 0.1 | 0.05 | % maximum |
| Noise (Curve A) | -65 | -60 | -70 | dB maximum |
| Frequency Response | RIAA Compensation | 30-40K | 10-40K | Hz |
| Output Voltage Maximum | 0.6 | 0.6 | 5.0 | V(R.M.S.) |
| Input Impedance | 50K | * | – | |
| Tone Control Range at 100 Hz an 5KHz | – | – | ±18 | dB |

**Designed to operate with a pickup having an equivalent capacitance of approximately 600pF.*

*TABLE 2 = PERFORMANCE CHARACTERISTICS OF THE PREAMPLIFIERS & TONE CONTROL*

These versatile preamplifiers can be used with other amplifiers, to enable such use the maximum unclipped r.m.s. output voltage is given as well as the current requirement. A supply rail of up to 35 volts can be used. The sensitivity will not change with the higher supply potential but both the maximum output voltage and current consumption will be approximately proportional to the supply voltage.

## Note 1

The use of curve A (ASA Standard S 1.4 - 1961) was discussed in the article "Signal/Noise measurements in Audio Amplifiers" Radiotronics September 1964.

# ANNUAL INDEX VOL. 32, 1967

# TELEVISION DEFLECTION STAGES

## A Note on Obtaining Best Life from Deflection Valves

A condition may arise in the horizontal stages of a television receiver which substantially reduces the life of the horizontal output valve even though the picture width is considered normal. The purpose of this note is to discuss this problem, and give recommendations to help overcome it, thus improving the performance of the horizontal output valve.

Since the introduction of television into Australia, the deflection angle, and screen area of picture tubes have increased. Thus the horizontal output stage has been required to give adequate line scan, when subjected to increased loading conditions, caused by the higher picture tube beam current, and in some instances the increase in e.h.t. voltage.

Therefore the stress on the horizontal output valve became greater, and the valve was required to operate under conditions which approached more closely the permitted maximum plate and screen dissipation.

### OPERATING PRINCIPLES

The majority of modern television receivers use horizontal deflection circuits incorporating automatic stabilisation. Mainly on this type of circuit an adverse and damaging mode of operation can occur. Briefly the function of the stabilising system employed in this stage is as follows: The d.c. bias made available to the grid of the horizontal output valve is derived from high voltage pulses tapped off the primary of the e.h.t. transformer, and rectified by the non-linear characteristics of a voltage dependant resistor. The value of d.c. bias is determined by the amplitude of the high voltage pulses, and therefore the output of the horizontal stage.

Should the output of the stage tend to rise as happens with increased a.c. mains voltage, the d.c. bias applied to the control grid becomes more negative to retain substantially the same plate current and the picture width is held practically constant.

Conversely, when the output decreases due to low a.c. mains voltage the d.c. bias becomes less negative and the plate current and picture width are once again restored to normal.

The function of the stabilised horizontal stage keeps the picture width substantially constant, not only for changes in a.c. mains voltage and associated variations in B+ voltage, but also for changes in e.h.t. under varying amounts of beam current.

The foregoing is true provided the drive voltage waveform applied to the grid of the horizontal output valve has sufficient amplitude. However, should the drive voltage decrease considerably through low output from a defective horizontal oscillator stage, the output of the e.h.t. transformer tends to fall and the d.c. bias shifts sufficiently less negative to return the picture to apparent normal width. The change in bias may be considerable and the valve operating under these conditions may conduct over a much longer part of the operating cycle. As a result the screen voltage decreases and the average screen current increases as the knee plate characteristic is approached, thus the screen dissipation increases considerably. The average plate current and plate dissipation also show a marked increase, particularly under conditions where plate current is still flowing when the high voltage pulse is present on the plate, under normal conditions the valve would be cut off at this stage. This mode of operation is most undesirable and the life of the horizontal output valve is considerably decreased. Naturally, a greatly reduced amplitude of grid drive voltage causes a noticeably narrow picture, while the absence of grid drive voltage causes the plate of the valve to glow red through excessive dissipation.

### SERVICE RECOMMENDATIONS

From the above it can be seen that merely changing the horizontal output valve may "fix" the apparent fault but result in a very limited valve life. To guard against this possibility it is advisable to read the peak to peak drive voltage at the grid of the horizontal output valve using a vacuum tube volt meter, and adjust it if necessary. Approximately 100 volts p-p is sufficient for many receivers. However it is advisable to consult a service manual when in doubt.

Other adjustments to give better operation of the horizontal valve require the setting of the horizontal linearity control for minimum plate current. In unstabilised e.h.t. receivers this is done conveniently by adjusting for minimum voltage across the cathode bias resistor in the horizontal output stage. In stabilised e.h.t. receivers the linearity control should be adjusted to give maximum bias at the grid of the output valve. An alternative method is to observe the brightness of a 12 volt dial light globe fitted in a plate cap adaptor connected in series with the plate of the valve. The brightness of the globe is then proportional to the plate current and the linearity control should be adjusted for minimum intensity.

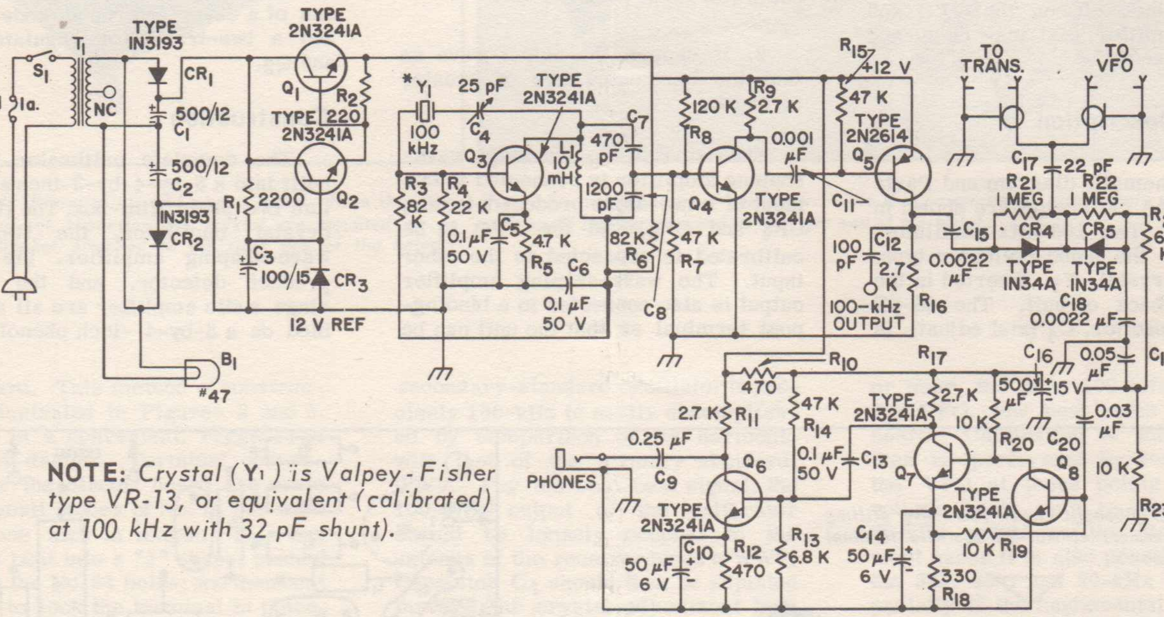
It should also be noted that it is important to observe that television receivers using a vertical output valve biased from the grid of the horizontal output valve are prone to trouble when the horizontal drive is low. The reduction of bias causes increased dissipation of the vertical output valve and difficulties may arise when adjusting for best vertical linearity.

# A VFO CALIBRATOR

George D. Hanchett, R.C.A. Electronic Components and Devices

During recent experiments with a VFO for operation at frequencies other than those tuned by an amateur-band receiver, it became desirable for the writer to have some type of calibrator to which the VFO could be coupled so that a series of calibration points could be determined and marked on the VFO dial.

The basic idea employed for the VFO calibrator described in this article is not new and will be recognized by many amateurs as the system used in the well-known war-surplus frequency meter, BC-221. (Editor's Note: The VFO calibrator can also be used to calibrate signal generators, etc.).




**NOTE:** *Crystal ($Y_1$) is Valpey-Fisher type VR-13, or equivalent (calibrated at 100 kHz with 32 pF shunt).*

$B_1$ — Lamp
$C_1$, $C_2$ — 500 $\mu$F, 12 volts
$C_3$ — 100 $\mu$F, 15 volts
$C_4$ — 25 pF, adjustable padder type, air dielectric
$C_5$, $C_6$, $C_{13}$ — 0.1 $\mu$F, 50 volts, ceramic
$C_7$ — 470 pF, 500 volts, silver mica type
$C_8$ — 1,200 pF, 500 volts, silver mica type
$C_9$ — 0.25 $\mu$F, 200 volts, paper
$C_{10}$, $C_{14}$ — 50 $\mu$F, 6 volts
$C_{11}$ — 0.001 $\mu$F, 1,000 volts, ceramic
$C_{12}$ — 100 pF, 1,000 volts, ceramic
$C_{15}$, $C_{18}$ — 0.0022 $\mu$F, 1,000 volts, ceramic
$C_{16}$ — 500 $\mu$F, 15 volts
$C_{17}$ — 22 pF, 1,000 volts, ceramic
$C_{19}$ — 0.05 $\mu$F, 50 volts, ceramic
$C_{20}$ — 0.03 $\mu$F, 50 volts, ceramic
$CR_1$, $CR_2$ — Diodes, 1N3193
$CR_3$ — Reference diode, 12 volts
$CR_4$, $CR_5$ — Signal diodes, type 1N34A
$F_1$ — Fuse, 1 ampere
$L_1$ — RF choke, 10 mH
$Q_1$, $Q_2$, $Q_3$, $Q_4$, $Q_6$, $Q_7$, $Q_8$ — Transistors, 2N3241A
$Q_5$ — Transistor, 2N2614
(Note: All following resistors ½ watt)
$R_1$ — 2,200 ohms
$R_2$ — 220 ohms
$R_3$, $R_6$ — 82,000 ohms
$R_4$ — 22,000 ohms
$R_5$ — 4,700 ohms
$R_7$, $R_{14}$, $R_{15}$ — 47,000 ohms
$R_8$ — 120,000 ohms
$R_9$, $R_{11}$, $R_{16}$,
$R_{17}$ — 2,700 ohms
$R_{10}$, $R_{12}$ — 470 ohms
$R_{13}$ — 6,800 ohms
$R_{18}$ — 330 ohms
$R_{19}$, $R_{20}$, $R_{23}$ — 10,000 ohms
$R_{21}$, $R_{22}$ — 1 megohm
$R_{24}$ — 68,000 ohms
$S_1$ — Switch, SPST toggle,
$T_1$ — Transformer, 6.3 volts, 1.2 amperes
$Y_1$ — Crystal, Valpey-Fisher type VR-13 or equivalent (calibrated at 100 kHz with 32 pF shunt)

Miscellaneous — 1 crystal socket for HE6/U; 1 head-phone jack; 1 aluminium two-piece box, 5-by-4-by-3 inches; 1 phenolic circuit board, 3-by-4½ inches; 2 UHF coaxial connectors (Amphenol S0239 or equiv.); 1 binding post.

*Figure 1: Schematic diagram and Parts List for VFO calibrator.*

Harmonics of the secondary-standard 100-kHz crystal oscillator are beat with the fundamental, or harmonics, of the VFO to provide audible signals at definite frequencies across the dial. For example, if this unit is used with a 5.0-to-5.5-MHz VFO such as the type widely utilized for SSB operation, the 100-kHz calibration points are the strongest by far. However, the 50-kHz points are also perceptible and, if proper car is exercised, the 33-, 25-, and 20-kHz points can be detected as well. In practice, the calibrator is permanently connected to the RF line between the VFO and the transmitter, and may be turned on when needed.

## Circuit Description

The schematic diagram and Parts List for the calibrator are shown in Figure 1. The 100-kHz oscillator, $Q_3$, is of the tuned-collector type, with the crystal, $Y_1$, inserted in the base feedback circuit. The 25-pF padder capacitor, $C_4$ (xtal adjust), is connected in series with the crystal so that oscillation can be adjusted to exactly 100-kHz. Capacitors $C_7$ and $C_8$ are used as a voltage divider to reduce the coupling to the input of the two-stage wave-shaping amplifier, $Q_4$ and $Q_5$, and thus prevent loading of the secondary-standard oscillator. The two-stage wave-shaping amplifier provides the following advantages:

1. It prevents any reflection of the load from affecting the frequency of the 100-kHz secondary-standard oscillator.

2. It shapes the output wave so that the harmonics are of greater strength.

The output of the two-stage wave-shaping amplifier is connected to one input of a two-diode product detector, $CR_4$ and $CR_5$, and the VFO to be calibrated is connected to the other input. The wave-shaping amplifier output is also connected to a binding-post terminal so that the unit can be used as a conventional 100-kHz crystal calibrator. The values of the components shown in the circuit diagram have been chosen for a peak VFO signal level of 2 to 3 volts. For larger VFO signals, it will be necessary to replace the 22-pF capacitor, $C_{17}$, with some type of capacitive or resistive attenuator.

A three-stage amplifier – $Q_6$, $Q_7$, $Q_8$ – is used to amplify the extremely low audio output to the two-diode product detector to a comfortable head-phone level. The power supply for the complete unit is regulated by use of a zener reference diode, $CR_3$, and a two-transistor regulator, $Q_1$ and $Q_2$.

## Construction

The complete calibrator unit is built into a 5-by-4-by-3-inch aluminium two-piece Mini-box. The 100-kHz crystal oscillator, the two-stage wave-shaping amplifier, the diode product detector, and the three-stage audio amplifier are all assembled on a 3-by-4 -inch phenolic cir-

*Figure 2: Phenolic circuit board after drilling and placement of No. 14 bus-wire terminal connectors.*

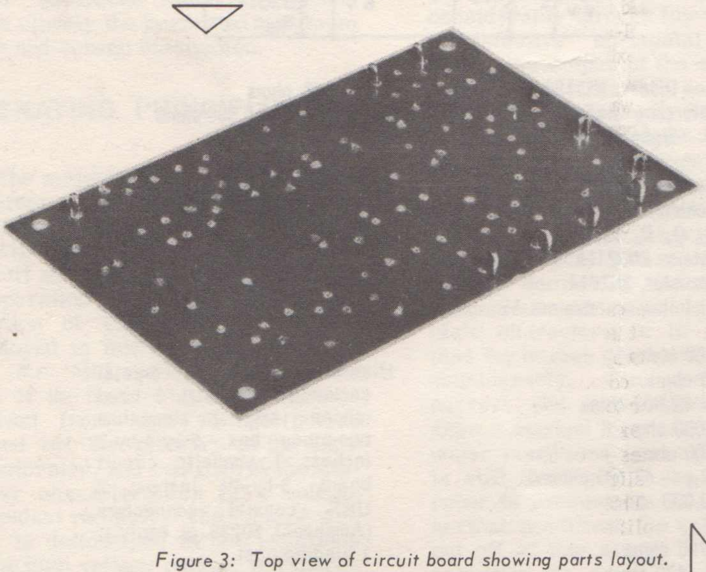
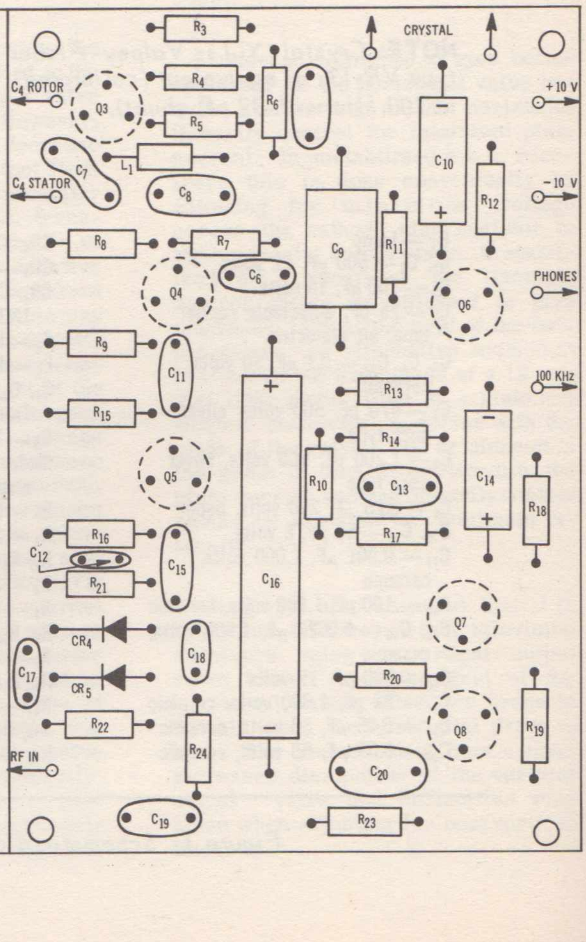


*Figure 3: Top view of circuit board showing parts layout.*



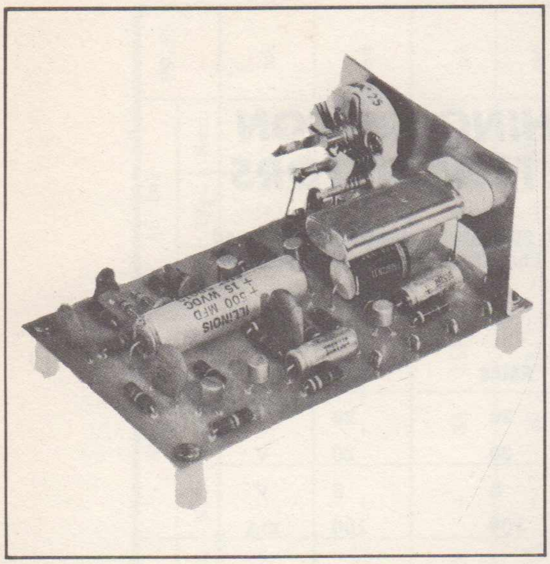


Figure 4: Pictorial view of circuit board. Note that the frequency-set capacitor and crystal are mounted on a small bracket attached to the right end of the board.

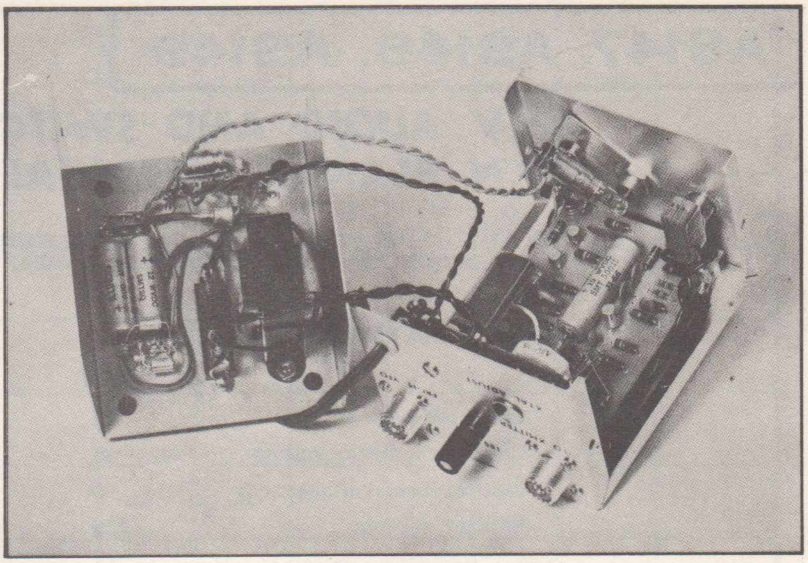


Figure 5: Internal view of calibrator. Note that power supply components are mounted on a small bracket attached to the right end of the board.

cuit board. This method of construction, illustrated in Figures 2 and 3, results in a convenient, rugged, and compact design. Terminal connectors for the circuit board are made from small pieces of No. 14 bus wire about one inch in length. This bus wire is bent into a "J" shape; pushed through the No. 54 holes; and then bent around to lock the terminal in place.

The crystal socket and the air capacitor used for setting the frequency are mounted on a small piece of aluminium which is attached to one end of the circuit board. The circuit board is separated from the Minibox by threaded, -inch, 4-40 spacers and 4-40 screws. RF connections to the VFO and transmitter are accomplished through standard UHF coaxial connectors mounted on the rear of the unit. The 100-kHz output terminal is also mounted on the rear of the unit.

## Adjustment and Operation

The adjustment of the 100-kHz secondary-standard oscillator to precisely 100-kHz is easily accomplished by comparison of its harmonic with that of the primary standard, WWV. For the best beat signal, the 100-kHz output of the calibrator should be loosely coupled to the antenna of the receiver tuned to WWV. Capacitor $C_4$ should then be adjusted through the crystal-adjustment hole until a zero beat exists between the secondary-standard and WWV. It would be well to wait for the quiet period of WWV's transmission (when there is no 440-Hz modulation) to be absolutely certain that the secondary standard is beating with the carrier and not with the modulation.

The use of the calibrator is extremely simple. It is inserted in the RF line of the VFO by connecting the VFO to the input coaxial connector and the transmitter to the output connector. When power is applied to the unit and headphones are inserted in the phone jack, a slight hissing noise should be heard. This noise indicates that the audio amplifier is active. At, or near, the even 100-kHz points on the VFO, low beat notes should be heard. Calibration of the dial can then be performed by zero-beating the VFO at those points. Lower-volume beats may be heard at the 50-kHz points on the dial, and in most cases it is also possible to hear the 33-, 25-, and 20-kHz beats, especially if the fundamental operating frequency of the VFO is below 5 MHz.

With many of today's amateur-radio receivers designed solely for hamband reception, the VFO calibrator is especially applicable to oscillators operating at frequencies outside the hambands. In addition, the unit can prove very useful for calibrating certain types of test equipment and for allowing the VFO to be used as a hamband frequency meter.

• WITH ACKNOWLEDGEMENT TO RCA

# AS147, AS148, AS149

## AWV AUDIO AND SWITCHING SILICON N.P.N PLANAR EPITAXIAL TRANSISTORS

The AS147, AS148 and AS149 are n-p-n silicon planar epitaxial transistors designed for general purpose use in audio and switching circuits. The AS149 is a low noise high-gain unit.

### ABSOLUTE MAXIMUM RATINGS

| | AS147 | AS148 | AS149 | |
|---|---|---|---|---|
| Collector-base voltage | 45 | 20 | 20 | V |
| Collector-emitter voltage | 45 | 20 | 20 | V |
| Emitter-base voltage | 5 | 5 | 5 | V |
| Emitter current | 100 | 100 | 100 | mA |

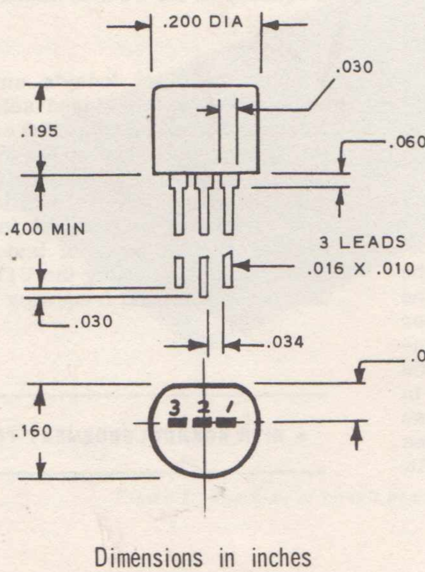
### THERMAL RATINGS

Dissipation in an ambient temperature up to 25°C ... 200mW max.
Derate linearly to zero at 135°C.

During soldering lead temperatures must not exceed 255°C for 10 secs. max. within 1/16" of can.

Storage temperature -25°C to 135°C.

### DIMENSIONAL OUTLINE




Dimensions in inches

### TERMINAL DIAGRAM


Lead 1 – Emitter
Lead 2 – Base
Lead 3 – Collector

## ELECTRICAL CHARACTERISTICS AT TA = 25°C

| CHARACTERISTICS | SYMBOL | TEST CONDITIONS | AS 147 | | | AS 148 | | | AS 149 | | | UNITS |
|---|---|---|---|---|---|---|---|---|---|---|---|---|
| | | | Min. | Typ. | Max. | Min. | Typ. | Max. | Min. | Typ. | Max. | |
| Collector Cut-off Current | $I_{CBO}$ | $V_{CB} = 20V$ | - | - | 15 | - | - | 15 | - | - | 15 | mA |
| Emitter Cut-off Current | $I_{EBO}$ | $V_{EB} = 5V$ | - | - | 100 | - | - | 100 | - | - | 100 | $\mu A$ |
| Base Current | $I_B$ | $V_{CE} = 5V,\ I_E = 10\mu A$ | - | 85 | 250 | - | 85 | 250 | - | - | 100 | nA |
| | | $V_{CE} = 5V,\ I_E = 2mA$ | - | 10 | 18.5 | - | 10 | 18.5 | - | - | 9.6 | $\mu A$ |
| | | $V_{CE} = 5V,\ I_E = 10mA$ | - | 40 | - | - | - | - | - | - | - | $\mu A$ |
| Emitter-Base Floating Voltage | $V_{EBF}$ | $V_{CB} = 20V$ | - | - | 1 | - | - | 1 | - | - | 1 | V |
| Collector-Emitter Voltage (Base Open) | $V_{CEO}$ | $I_C = 2mA$ | 45 | - | - | 20 | - | - | 20 | - | - | V |
| Collector-Emitter Saturation Voltage | $V_{CE(Sat)}$ | $I_C = 10mA,\ I_B = 1mA$ | - | - | 0.25 | - | - | 0.25 | - | - | 0.25 | V |
| Gain Bandwidth Product | $f_T$ | $V_C = 5V,\ I_C = 0.5mA$ | - | 85 | - | - | 85 | - | - | 85 | - | MHz |
| | | $V_{CE} = 5V,\ I_C = 10mA$ | - | 300 | - | - | 300 | - | - | 300 | - | MHz |
| Collector to Base Feedback Capacitance | $C_{b'c}$ | $V_{CB} = 5V,\ I_E = 0$ | - | 2.5 | - | - | 2.5 | - | - | 2.5 | - | pf |
| Noise Figure | N.F. | $V_{CE} = 5V,\ I_C = 0.2mA$<br>$\Delta f = 30$ Hz - 15KHz,<br>$R_S = 2K\Omega$ | - | - | - | - | - | - | - | - | 4 | db |

# NEWS & NEW RELEASES

## AWV MAGNETRON DEVELOPMENT

Recently A.W.V. completed the development of production of a miniature C Band Magnetron under a Department of Supply contract.

The magnetron is a rugged, tunable, integrated magnet, air-cooled device, capable of producing a nominal output power of 1,000 watts. It measures approximately 3.5 inches long, 1.4 inches diameter and weighs approximately 10 ounces.

The techniques involved in obtaining high power in a small volume are relatively new and very exacting. The establishment of an Australian production capability of this kind, therefore, contributes significantly to the rapidly expanding technical resources of the Australian thermionic valve industry.

## SECOND GENERATION INTEGRATED CIRCUITS

RCA have recently announced an improvement in technology resulting in smaller geometry pellets for two of their popular Integrated Circuits. The two devices concerned are the CA3014 and CA3028, the smaller geometry versions will be CA3028A. Both the "A" versions will be electrically interchangeable with their predecessors.

The smaller geometry of the CA3028A pellet gives a reduction in feedback and substrate capacitance which results in improved operation at high frequencies.

## 40575, 40576 TRIACS

The 40575 and 40576 are two triacs which have recently been released by RCA. They have an rms on-state circuit capability of 15 amps at a case temperature of +75°C. The 40575 has a repetitive peak off-state voltage rating at 200 volts and the 40576 400 volts.

The triacs are packaged in a TO-66 case and are intended primarily for the control of ac loads in applications such as space heaters, ovens, and furnace control. The 40576 will control 3600 watt at 240V, 50Hz.

## AWV SILICON TRANSISTORS

A further nine silicon planar transistors have been added to the A.W.V. range. Data on three (3) of the new types is published elsewhere in this issue and the other types will be covered in future issues.

The new types are as follows:-

| | |
|---|---|
| AS147, AS148 | General purpose audio and switching transistors in the A.W.V. epoxy package. |
| AS149 | A low noise high gain transistor designed for audio preamplifier appplications. |
| AS200, AS201 | General purpose H.F. and V.H.F. transistors with a typical $f_T$ of 900 MHz, a typical $R_{bb'}$ $C_{b'c}$ of 8pS, and a TO-104 case. |
| AS204, AS205, AS208, AS209 | General purpose audio types with a maximum dissipation at a case temperature of 75°C of 2 Watt. They are packaged in a modified 3 lead TO-104 can. |

## E.E.V. MINI C IGNITRON

The English Electric Valve Company Pty. Ltd. have just announced the release of a new ignitron type BK542/1081 known as the Mini 'C'.

It has a standard international 'B' size envelope, but can handle 65% more kVA than the 'B' size. This new ignitron is therefore capable of replacing the larger 'C' sized types in many applications; hence the designation MINI 'C'.

The latest development has several advantages:

(1) The tube has a low "take-over" voltage; This minimises misfiring when operating at low load current conditions, and this, in turn, gives improved ignitor life.
(2) For the same operating conditions the life of a Mini 'C' Ignitron will be substantially greater than that of a standard "B" size tube.
(3) There is direct contact between the vacuum envelope and the cooling water, which ensures reliable cooling and rapid action of the thermostat.
(4) The cooling water inlet has been designed to permit a streamline flow pattern which reduced clogging due to sediment.
(5) Quick-release fittings are used for both water connections.
(6) The Mini 'C' when fitted into standard 'B' sized sockets will permit immediate uprating of existing equipment so giving a new intermediate range.


*Mini 'C' Ignitron Type BX542/1081*

# THEORY AND APPLICATION OF THYRISTORS (PART 3)

## Gate Characteristics

SCRs and triacs are specifically designed to be triggered by a signal applied to the gate terminal. The manufacturer's specifications indicate the magnitudes of gate current and voltage required to turn on these devices. Gate characteristics, however, vary from device to device even among devices with the same family. For this reason, manufacturer's specifications on gating characteristics provide a range of values in the form of characteristic diagrams. A diagram such as that shown in Figure 30 is given to define the limits of gate currents and voltages that may be used to trigger any given device of a specific family. The boundary lines of maximum and minimum gate impedance on this characteristic diagram represent the loci of all possible triggering points for thyristors in this family. The curve OA represents the gate characteristic of a specific device that is triggered within the shaded area.

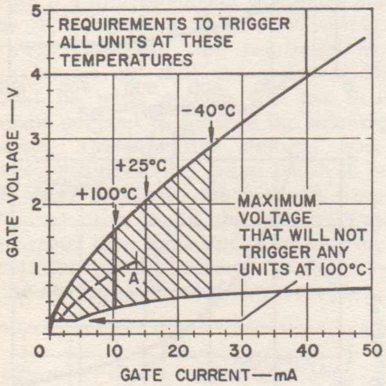



*Figure 30: Gate-characteristics curves for a typical RCA SCR:*

TRIGGER LEVEL - The magnitude of gate current and voltage required to trigger a thyristor varies inversely with junction temperature. As the junction temperature increases, the level of gate signal required to trigger the thyristor becomes smaller. Worst-case triggering conditions occur, therefore, at the minimum operating junction temperature.

The maximum value of gate voltage below the level required to trigger any unit of a specific thyristor family is also an important gate characteristic. At the high operating temperatures, the level of gate voltage required to trigger a thyristor approaches the minimum value, and undesirable noise signals may inadvertently trigger the device. The maximum nontriggering gate voltage at the maximum operating junction temperature of the device, therefore, is a measure of the noise-rejection level of a thyristor.

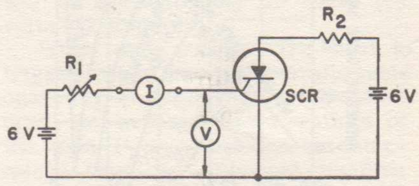



*Figure 31: Test circuit used to determine gate-trigger-pulse requirements of thyristors.*

The gate voltage and current required to switch a thyristor to its low-impedance state at maximum rated forward anode current can be determined from the circuit shown in Fig. 31. The value of resistor $R_2$ is chosen so that maximum anode current, as specified in the manufacturer's current rating, flows when the device latches into low-impedance state. The value of resistor $R_1$ is gradually decreased until the device under test is switched from its high-impedance state to its low-impedance state. The values of gate current and gate voltage immediately prior to switching are the gate voltage and current required to trigger the thyristor.

The GATE NONTRIGGER VOLTAGE $V_{gnt}$ is the maximum dc gate voltage that may be applied between gate and cathode of the thyristor for which the device can maintain its rated blocking voltage. This voltage is usually specified at the rated operating temperature (100°C) of the thyristor. Noise signals in the gate circuit should be maintained below this level to prevent unwanted triggering of the thyristor.

PULSE TRIGGERING - When very precise triggering of a thyristor is desired, the thyristor gate must be overdriven by a pulse of current much larger than the dc gate current required to trigger the device. The use of a large current pulse reduces variations in turn-on time, minimizes the effect of temperature variations on triggering characteristics, and makes possible very short switching times.

In the past, the maximum value of gate signal that could be used to trigger a thyristor was severely restricted by minimum dc triggering requirements and limitations on maximum gate power. The coaxial gate structure and the 'shorted-emitter' construction techniques now used in RCA thyristors, however, has greatly extended the range of limiting gate characteristics. As a result, the gate-dissipation ratings of RCA thyristors are compatible with the power-handling capabilities of other elements of these devices. Advantage can be taken of the higher peak-power capability of the gate to improve dynamic performance, increase di/dt capability, minimize interpulse jitter, and reduce switching losses. This higher peak-power capability also allows greater interchangeability of thyristors in high-performance applications.

The "shorted-emitter" techniques makes use of the resistance path within the gate layer which is in direct contact with the cathode electrode of the thyristor. When gate current is first initiated, most of the current bypasses the gate-to-cathode junction and flows from the resistive gate layer to the cathode contact. When the IR drop in this gate layer exceeds the threshold voltage of the gate-to-cathode junction, the current across this junction increases until the thyristor is triggered.

When an SCR is triggered by a gate signal just sufficient to turn on the device, the entire junction area does not start to conduct instantaneously. Instead, as pointed out in the discussion on SWITCHING CHARACTERISTICS, the device current is

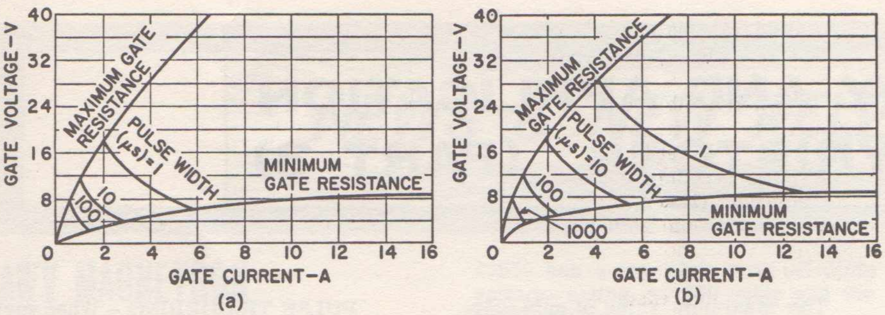



*Figure 32: Forward-gate characteristics for pulse triggering of RCA SCR: (a) low-current types, (b) high-current types.*

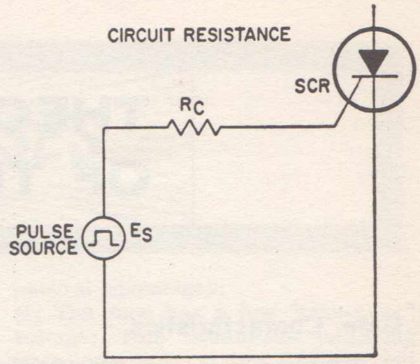



*Figure 34: Equivalent diagram of the basic gate-trigger circuit fo a thyristor.*

confined to a small area, which is usually the most sensitive part of the cathode. The remaining cathode area turns on as the anode current increases. When a much larger-signal is applied to the gate, a greater part of the cathode is turned on initially and the time to complete the turn-on process is reduced. The peak amplitude of gate-trigger currents must be large, therefore, when thyristors have to be turned on completely in a short period of time. Under such conditions, the peak gate power is high, and pulse triggering is required to keep the average gate dissipation within the values given in the manufacturer's specifications. New gate ratings, therefore, are required for this type of application.

The forward gate characteristics for thyristors, shown in Fig. 32, indicate the maximum allowable pulse widths for various peak values of gate input power. The pulse width is determined by the relationship that exists between gate power input and the increase in the temperature of the thyristor pellet that results from the application of gate power. The curves shown in Fig. 32(a) are for RCA SCRs that have relatively small current ratings (2N4101, 2N4102, and 40379 families), and the curves shown in Fig. 32(b) are for RCA SCRs that have larger current ratings (2N3670, 2N3873, and 2N3899 families). Because the higher-current thyristors have larger pellets, they also have greater thermal capacities than the smaller-current devices. Wider gate trigger pulses can therefore be used on these devices for the same peak value of gate input power.

Because of the resistive nature of the "shorted-emitter" construction, similar volt-ampere curves can be constructed for reverse gate voltages and currents, with maximum allowable pulse widths for various peak-power values, as shown in Fig. 33. These curves indicate that reverse dissipations do not exceed the maximum allowable power dissipation for the device.

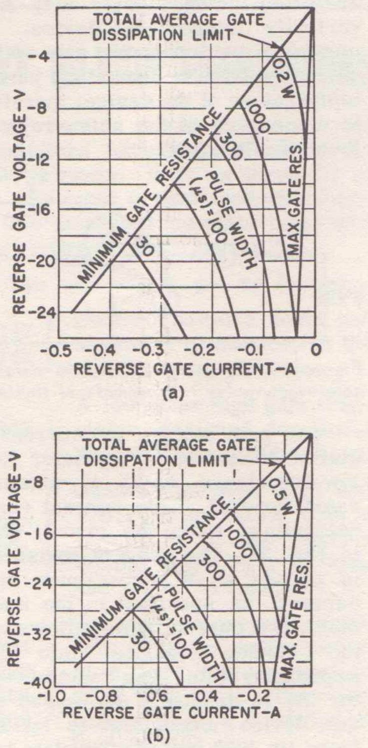



*Figure 33: Reverse gate characteristics of RCA SCR: (a) low-current types, (b) high-current types.*

TRIGGER-CIRCUIT REQUIREMENTS – The basic gate trigger circuit for a thyristor can be represented by a voltage source and a series resistance, as shown in Fig. 34. The series resistance should include both the external circuit resistance and the internal generator resistance. With this type of equivalent circuit, the conventional load-line approach to gate trigger-circuit design can be used. With pulse triggering, it is assumed initially that the turn-on time required to trigger all transistors of the same type is known, and that the maximum allowable gate trigger-pulse widths for specific gate-power imputs are to be determined.

The magnitude of gate-trigger current required to turn on all SCRs of a given type can be determined from the turn-on characteristics shown in Fig. 35. The spread or

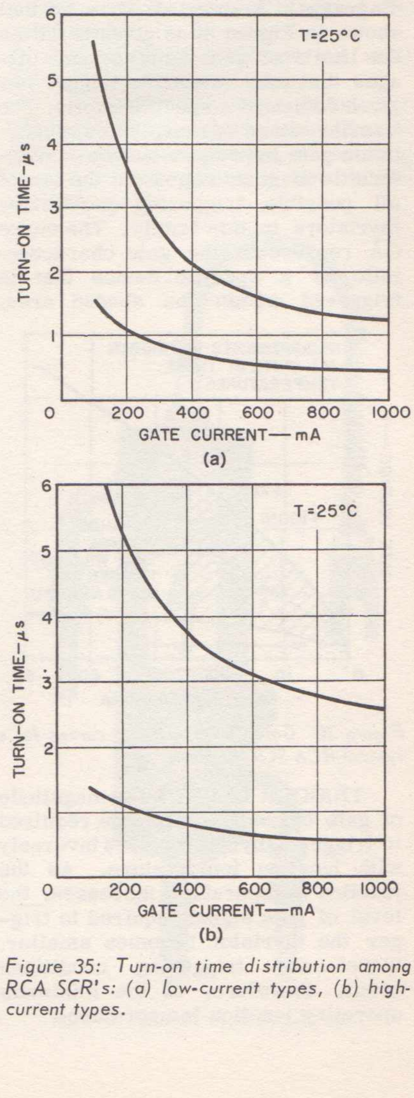



*Figure 35: Turn-on time distribution among RCA SCR's: (a) low-current types, (b) high-current types.*

band of turn-on characteristics for the same gate current results from the variation of gate-trigger characteristics among devices of the same family. Because of the greater overdrive factor involved, the same gate current applied to a device obviously turns on a low-gate-current device in much less time than that required to turn-on a higher-gate-current device. For example, a gate-trigger current of 100 milliamperes overdrives an SCR that requires a trigger current of only 2 milliamperes by a factor of 50 and causes the device to turn on very quickly, while an SCR that requires 10 milliamperes of trigger current is overdriven by a factor of 10 and is turned on more slowly. As the gate current increases, the band of turn-on characteristics becomes narrower, and an increase in gate current does not effectively decrease the turn-on time.

The following example, in which an RCA-2N3873 SCR is to be turned on in 2.5 microseconds, demonstrates the use of the various characteristics in the solution of a typical triggering problem.

The turn-on characteristics shown in Fig. 35 indicate that a gate-trigger current of 1 ampere is required to insure that all devices of this type will turn on in 2.5 microseconds (the 2.5 microsecond ordinate level intersects the upper curve at 1000 milliamperes). In addition, the width of the gate-trigger pulse should be at least 2.5 microseconds to ensure that the SCR remains on after it is triggered. Actually, the minimum requirement is that the pulse width must be wide enough for the SCR anode current to achieve the latching value. Conservative design, however, requires the pulse width to be at least equal to the turn-on time. For inductive loads, the turn-on time is larger than indicated in the characteristics curves because of the slow rise of current through the inductance.

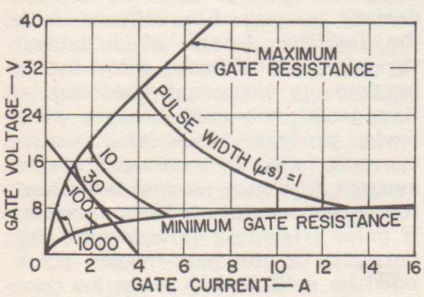



*Figure 36: Forward gate characteristics of typical RCA SCR's showing load-line for a source of 20 volts and a required gate current of 1 ampere.*

A straight load line can then be plotted on the pulse triggering characteristics, as shown in Fig. 36. The two points that determine the position of this line are the source voltage (20 volts) and a point slightly above the intersection of the required gate current (1 ampere) and the curve of maximum gate resistance. The load line should lie below the pulse-width curve required to trigger all SCRs (in this example, the 2.5-microsecond curve). The maximum allowable pulse width is obtained by estimation of the pulse-width curve tangent of the load line. In this example, the pulse width is estimated to be 30 microseconds (the pulse-width curves are logarithmically spaced). The load line intersects the abscissa at the 4 ampere point. The maximum circuit resistance, therefore, is 5 ohms. The peak gate power is the product of gate voltage and gate current at the point of tangency of the pulse-width curve, and is approximately 20 watts (10 volts × 2 amperes).

When gate pulses are used to trigger SCRs, the maximum allowable operating frequency f is dependent upon the average power ratings of the gate $P_{g(avg)}$ and can be determined from the following equation:

$$f = P_{g(avg)} / P_{g(pk)} \times PW_g \qquad (8)$$

where $P_{g(pk)}$ is the peak gate power and $PW_g$ is the gate pulse width.

If it is assumed that only half the total average gate-dissipation rating of the RCA-2N3873, or 0.25 watt, is used to trigger the device, then the maximum allowable operating frequency is determined as follows:

$$f = \frac{0.25\ \text{W}}{20\text{W} \times 2.5 \times 10^{-6}\ \text{second}} = 5000\ \text{Hz}$$

If there is no reverse gate power dissipation, the maximum allowable frequency can be 10,000 Hz. If the maximum allowable pulse width is 30 microseconds, the maximum allowable operating frequency is proportionately reduced to 416 Hz.

The trigger-circuit design is usually fixed by the requirements for reliable triggering, and reverse gate dissipation is considered after the values of source voltage and circuit resistance have been determined. Reverse gate power dissipation results from reverse gate-bias conditions or circuit reaction caused by some switching function. As in the case of the forward gate characteristics, a load-line approach can also be used to determine the reverse gate characteristics. The maximum anticipated value of reverse gate potential is used as the source voltage, and the external circuit resistance is used to determine the slope of the load-line. The load-line on the reverse gate characteristics shown in Fig. 37 represents a reverse gate-source voltage of 24 volts and an external-circuit resistance of 5 ohms. From the relationship that exists among pulse width, average gate power, peak gate power, and frequency, a maximum pulse width can be calculated for the actual operating frequency. For a reverse gate dissipation of 0.25 watt, peak gate power of 10 watts, and a frequency of 5000 Hz, the maximum allowable pulse width PW is calculated as follows:

$$PW = \frac{0.25\ \text{W}}{5000\ \text{Hz} \times 10\text{W}} = 5\ \text{microseconds} \qquad (9)$$

This reverse gate-pulse width should be less than the maximum allowable pulse width, as determined by the curve that lies just below the load line on Fig. 37. In this example, the maximum allowable pulse width for reverse dissipation is 100 microseconds.

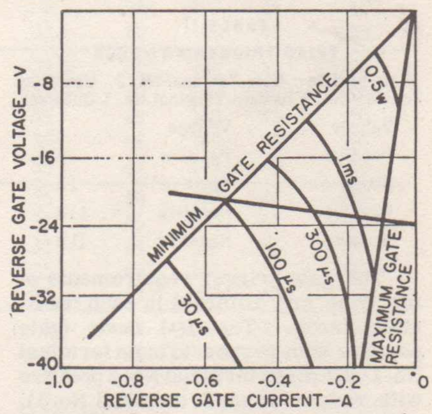



*Figure 37: Reverse gate characteristics for typical RCA SCR showing load-line for a reverse gate-source voltage of 24 volts and an external circuit resistance of 5 ohms.*

The total average dissipation caused by gate-trigger pulses is the sum of the average forward and reverse dissipations. This total dissipation should correspond to the average gate power dissipation shown in the published data for the selected SCR. If the average gate dissipation exceeds the maximum published value, as the result of high forward gate-trigger pulses and transient or steady-state reverse gate biasing, the maximum allowable forward-conduction-current rating of the device must be reduced to compensate for the increased rise of junction temperature caused by the increased gate power dissipation.

The trigger-circuit design considerations described for RCA SCRs also apply to RCA triacs. Although both types of devices are triggered in the same manner, the triac can be triggered by either positive or negative gate-trigger pulses independent of the polarity of the voltage between the main terminals.

In RCA triacs, the gate-trigger-pulse polarity is therefore always measured with respect to main terminal No. 1, which is comparable to the cathode terminal of an SCR. The triac can be triggered by a gate-trigger pulse which is either positive or negative with respect to main terminal No. 1 when main terminal No. 2 is either positive or negative with respect to main terminal No. 1. The triac, therefore, can be triggered in any of four operating modes, as summarised in Table II. The quadrant designations refer to the operating quadrant on the principal voltage-current characteristics, shown in Fig. 2 (either I or III), and the polarity symbol represents the gate-to-main-terminal-No. 1 voltage.

**TABLE II**

**TRIAC TRIGGERING MODES**

| Gate-to-Main Terminal-No. 1 Voltage | Main-Terminal-No. 2 to-Main-Terminal No. 1 Voltage | Operating Quadrant. |
|---|---|---|
| Positive | Positive | I(+) |
| Negative | Positive | II(–) |
| Positive | Negative | III(+) |
| Negative | Negative | III(–) |

The gate-trigger requirements of the triac are different in each operating mode. The I(+) mode (gate positive with respect to main terminal No. 1 and main terminal No. 2 positive with respect to main terminal No. 1), which is comparable to equivalent SCR operation, is usually the most sensitive. The smallest gate current is required to trigger the triac in this mode. The other three operating modes require larger gate-trigger currents. For RCA triacs, the maximum trigger-current rating in the published data is the largest value of gate current that is required to trigger the selected device in any operating mode.

## SERIES AND PARALLEL OPERATION

The voltage or current capabilities of a single thyristor can be extended by use of two or more thyristors of the same type in series or parallel arrangements, respectively. The following paragraphs discuss basic considerations important to the successful operation of thyristors used in multiple connections.

### Series Connections

When thyristors are connected in series for higher-voltage operation, certain procedures should be followed. These procedures usually depend upon the typical electrical characteristics of the thyristors used and the requirements of the circuit application.

The most important consideration in series connections of thyristors is to assure that voltages are divided equally across the individual units in the series string. One technique that may be used to obtain the desired voltage distribution is to select units that are matched with respect to such characteristics as OFF-state voltage breakdown, reverse voltage and current, and temperature effects. The use of a resistor across each unit in the stack is also recommended for improved series operation of thyristors. The value of the resistors should be some fraction of the maximum OFF-state resistance of the thyristor to force equal voltage divison among the devices.

When SCR's are used in series arrangements, differences in the reverse-recovery times of the units have an important bearing on the voltage division. Variation in internal capacitances of thyristors and in stray capacitances between thyristors and ground can also result in an unequal voltage division among the various units. The use of capacitive voltage dividers is recommended to eliminate the effects of such conditions. When capacitive voltage dividers are used, however, a damping resistor should be connected in series with each capacitor to restrict peak-current values when the thyristors are switched to the ON-state while the OFF-state voltage is present on the capacitor.

When thyristors are connected in series, the gate-trigger circuit used to turn on the various units requires special consideration. Because of differences in the delay and rise times of thyristors, gate-trigger currents that have a fast rise time must be used to turn on the units in the series string. The use of large gate-trigger currents minimizes turn-on differences. If large, quick-rising trigger currents are not used, the voltage across units that have longer turn-on times may exceed peak-voltage ratings.

### Parallel Connections

Thyristors are connected in parallel to obtain output currents higher than the current ratings of an individual thyristor. The main consideration for this type of operation is that the current must be divided equally among the parallel thyristors. One technique that is used to assure proper current division is to connect identical balancing resistors in series with each thyristor. The value of these resistors should be several times larger than the maximum ON-state impedance of the thyristors so that the current through each thyristor will be essentially the same even though the ON-state impedances of the thyristors are different. The addition of the balancing resistors to each conduction path, however, increases the power dissipation and, consequently, decreases the over-all efficiency of the circuit. The efficency of the circuit is improved if reactors, rather than resistors, are used to achieve balanced currents.

Another technique used in the parallel connection of thyristors is to select matched devices on the basis of specific conduction characteristics. When this technique is employed, circuit and load impedances must be carefully designed to assure an equal impedance for each conduction path in the parallel array. When factory-matched units are employed, care must be taken to insure that all units are operated at essentially the same case temperature. Because the forward voltage drop of a thyristor is temperature-dependent, differences in case temperature can result in unequal current sharing. All thyristors in the parallel array, therefore, should be mounted on a common heat sink to assure that the operating junction temperature of each device is the same.

When thyristors are connected in parallel, it is preferable to use a continuous gate drive to turn on the devices because of the differences in the latching levels of individual thyristors. Continuous gate drive is particularly important when inductive loads are used becuase such loads produce slow-rising output currents, and the continuous drive assures that gate current is present throughout the full conduction period. If pulse triggering is employed, the duration of the gate-trigger pulse must be sufficient to allow the conduction currents through all the thyristors to build up to values greater than the latching values to assure that all thyristors are completely turned on. The gate-trigger pulses should be quick-rising, high-amplitude pulses to assure good current sharing among the parallel thyristors during the turn-on interval.

Consideration should be given in parallel arrays to the possibility that one thyristor may be inadvertently turned on from some extraneous source, e.g., a high rate of rise of OFF-state voltage (dv/dt). Under such conditions, it is possible that an excessive amount of current may flow through this thyristor.

## TRANSIENT PROTECTION

Voltage transients occur in electrical systems when some disturbance disrupts the normal operation of the system. These disturbances may be produced by various sources (such as lightning surges, energizing transformers, and load switching), and may generate voltages which are well above the rating of the thyristor. Thyristors, in general, will switch from the OFF-state to the ON-state whenever the forward breakover voltage of the device is exceeded, and energy is then transferred from the thyristor to the load. Because the internal resistance of thyristors is high during the OFF-state, the nature of some transients may cause considerable energy to be dissipated in the thyristor before breakover occurs. Also, the transient voltage may exceed the maximum allowable voltage rating and, therefore, may cause irreversible damage to the thyristor. In either case, transient-suppression techniques must be used to prevent device destruction.

The use of thyristors that have a voltage rating greater than the highest transient voltage expected in the system is one way to provide protection against destructive transients. This method, however, is not always the most economical technique. The effects of voltage transients in thyristor circuits can also be decreased by a reduction of the rate at which the energy is dissipated in the device by relocation of the switching elements or by a change in the sequence of switching. Other preventive methods include the use of external circuit components, such as non-linear resistors and RC snubber networks, which limit the peak voltage across the thyristor.

The most common type of transient voltage suppressor is the RC network. This network is connected in parallel with the thyristor, as shown in Fig. 38. The value of the resistor should be selected on the

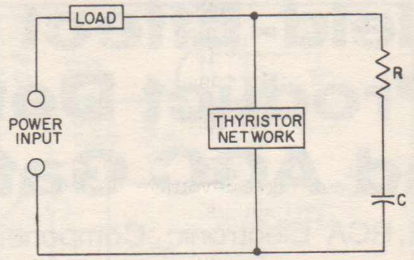



*Figure 38: Diagram showing use of RC network for transient suppression in thyristor circuits.*

basis of the di/dt rating of the thyristor. The size of the capacitor required for suppression of transient voltages is a function of many circuit parameters and is difficult to predict with any degree of accuracy. Actual transient measurements on the equipment will determine the values of circuit elements required. The charging time constant of the capacitor should be greater than the expected duration of the transient so that the increase in capacitor voltage during the transient interval is negligible.

## GENERAL CHARACTERISTICS

Table III gives typical characteristics on the range of Thyristors presently available. The information given in the table should assist circuit designers to select a device suitable for their requirements. Detailed data sheets for these devices are available on request.

*TABLE III. Brief specifications of currently available Thyristors.*

| Type | Package | Off-State (Blocking) Voltage $V_F/V_R$ V | RMS On-State Current $I_{RMS}$ A | $T_C$ °C | Max. $T_j$ °C | $I_{GT}$ mA | Peak Surge Current (1 Cycle) A |
|---|---|---|---|---|---|---|---|
| **SCR's** | | | | | | | |
| 2N3228 | TO-66 | 200 | 5 | 75 | 100 | 15 | 60 |
| 2N3525 | TO-66 | 400 | 5 | 75 | 100 | 15 | 60 |
| 2N3528 | TO-8 | 200 | 2 | 25* | 100 | 15 | 60 |
| 2N3529 | TO-8 | 400 | 2 | 25* | 100 | 15 | 60 |
| 2N3668 | TO-3 | 100 | 12.5 | 80 | 100 | 40 | 200 |
| 2N3669 | TO-3 | 200 | 12.5 | 80 | 100 | 40 | 200 |
| 2N3670 | TO-3 | 400 | 12.5 | 80 | 100 | 40 | 200 |
| 2N3870 | Press-Fit | 100 | 35 | 65 | 100 | 40 | 350 |
| 2N3871 | Press-Fit | 200 | 35 | 65 | 100 | 40 | 350 |
| 2N3872 | Press-Fit | 400 | 35 | 65 | 100 | 40 | 350 |
| 2N3873 | Press-Fit | 600 | 35 | 65 | 100 | 40 | 350 |
| 2N3896 | 1/4"-Stud | 100 | 35 | 65 | 100 | 40 | 350 |
| 2N3897 | 1/4"-Stud | 200 | 35 | 65 | 100 | 40 | 350 |
| 2N3898 | 1/4"-Stud | 400 | 35 | 65 | 100 | 40 | 350 |
| 2N3899 | 1/4"-Stud | 600 | 35 | 65 | 100 | 40 | 350 |
| 2N4101 | TO-66 | 600 | 5 | 75 | 100 | 15 | 60 |
| 2N4102 | TO-8 | 600 | 2 | 25* | 100 | 15 | 60 |
| 2N4103 | TO-3 | 600 | 12.5 | 80 | 100 | 40 | 200 |
| 40378 | Low-Profile TO-5 | 200 | 7 | 60 | 100 | 15 | 80 |
| 40379 | Low-Profile TO-5 | 400 | 7 | 60 | 100 | 15 | 80 |
| **TRIACS** | | | | | | | |
| 40429 | TO-66 | 200 | 6 | 75 | 100 | 25 | 80 |
| 40430 | TO-66 | 400 | 6 | 75 | 100 | 25 | 80 |
| 40431 (with trigger) | Modified TO-5 | 200 | 6 | 75 | 100 | 40[a] | 100 |
| 40432 (with trigger) | Modified TO-5 | 400 | 6 | 75 | 100 | 40[a] | 100 |
| 40485 | Modified TO-5 | 200 | 6 | 75 | 100 | 25 | 100 |
| 40486 | Modified TO-5 | 400 | 6 | 75 | 100 | 25 | 100 |
| 40525 | Modified TO-5 | 100 | 2.5 | 60 | 90 | 3 | 25 |
| 40526 | Modified TO-5 | 200 | 2.5 | 60 | 90 | 3 | 25 |
| 40527 | Modified TO-5 | 400 | 2.5 | 60 | 90 | 3 | 25 |
| 40528 | Modified TO-5 | 100 | 2.5 | 70 | 100 | 10 | 25 |
| 40529 | Modified TO-5 | 200 | 2.5 | 70 | 100 | 10 | 25 |
| 40530 | Modified TO-5 | 400 | 2.5 | 70 | 100 | 10 | 25 |
| 40575 | TO-66 | 200 | 15 | 70 | 100 | 30 | 100 |
| 40576 | TO-66 | 400 | 15 | 70 | 100 | 30 | 100 |

$^*T_{FA}$    a – Trigger Volts.

• WITH ACKNOWLEDGEMENT TO RCA

# The MOS Field-Effect Transistor As a Product Detector and AGC Gate

by W. M. STOBBE, RCA Electronic Components and Devices

The paragons of two worlds of electronics - the vacuum tube and the solid-state device - have been successfully combined in the RCA-developed insulated-gate metal-oxide-semiconductor (MOS) field-effect transistor. This transistor features such useful characteristics as extremely high input impedance, excellent low-noise qualities, high power gain throughout the VHF region, and a square-law transfer characteristic over a substantial current range. Irs transfer characteristic makes the device ideal for use as a product detector. Considerably less BFO voltage (in the order of a few millivolts) is required in this transistor than in the conventional diode or other devices.

The paper discusses the design of a product detector and AGC circuit that incorporates the MOS field-effect transistor as the active element. Figure 1 shows a block diagram of the SSB receiver, indicating the position of the product detector and AGC circuits. These circuits are connected into the receiver for SSB operation by means of the automatic-volume-control (AVC) switch.

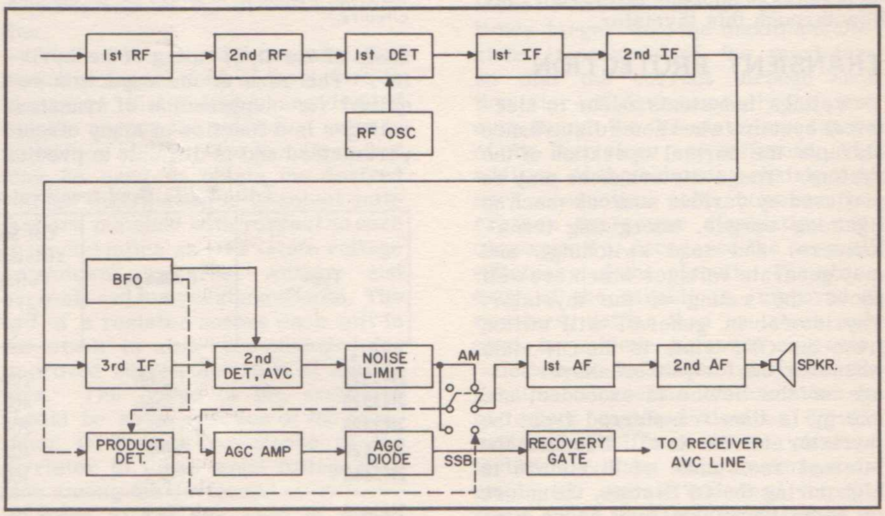



*Figure 1: Block Diagram of a typical general-purpose communications receiver showing additions (dashed lines) of product detector and AGC system.*

## Product Detector

Figure 2 shows the schematic diagram of the product detector. A separate IF-transformer, $T_{100}$, is used to couple the SSB-IF-signal from the grid of the third IF-amplifier tube to the field-effect transistor. Capacitor $C_{101}$ minimizes capacitive loading on the circuit being sampled and also permits tuning of the primary winding of the transformer, $T_{100}$. Capacitor $C_{102}$, which consists of a twisted-wire "gimmick", couples the BFO voltage to the product detector. Any excess BFO voltage blocks the SSB signal and reduces the output level.

As the output of the product detector is switched into the receiver audio-frequency volume control, the output of the conventional AM detector is disconnected from the circuit.

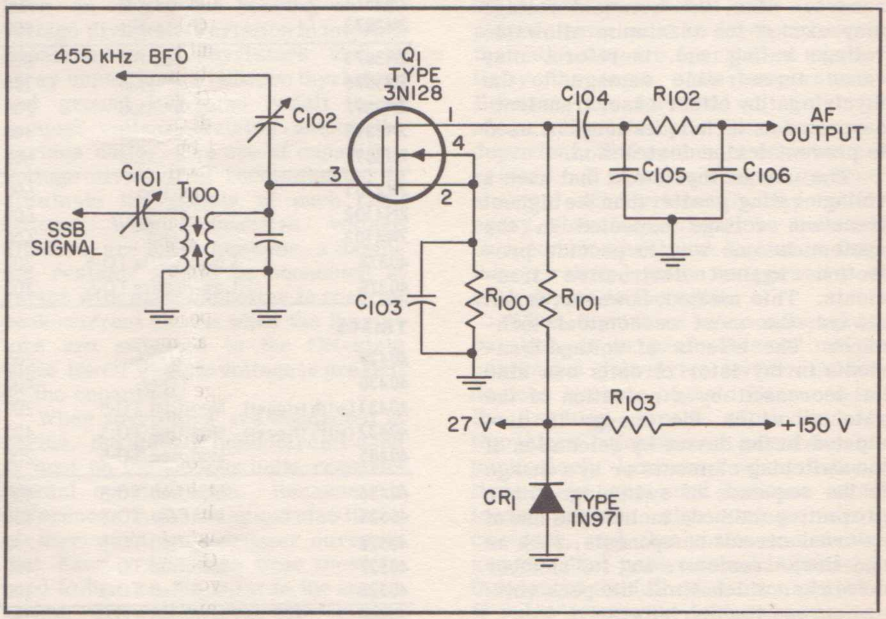



*Figure 2: Schematic diagram of product detector.*

The BFO should be "on" and the AVC system switched to "manual". Gain is then controlled by the RF gain control. If the audio AGC is used, this voltage is connected into the receiver AVC bus. Audio volume is then set to about three-quarters of full "on", and the sensitivity is

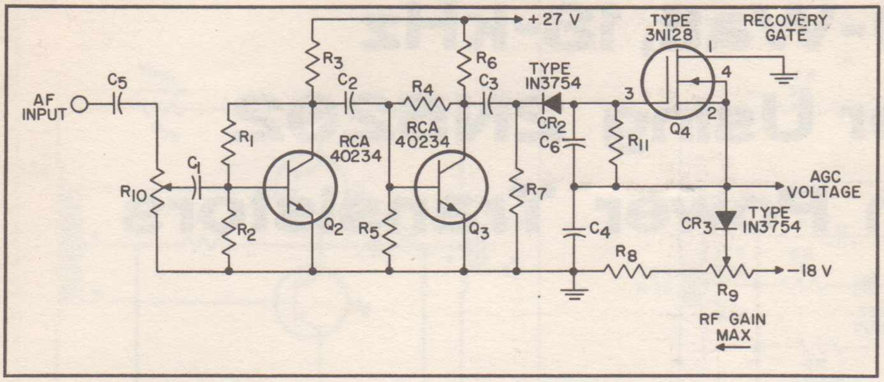



*Figure 4: Schematic diagram of AGC system.*

controlled by the RF gain control. The SSB signal is tuned for maximum intelligibility by means of the main tuning control and the BFO pitch control.

## Audio AGC

Operation on SSB can be greatly enhanced by use of an audio-derived AGC circuit. This circuit permits controlled volume and use of the "S" meter to interpret signal strength.

The AGC circuit shown in Figure 3 consists of a two-stage audio amplifier that uses silicon n-p-n transistors; an AGC diode; and an MOS field-effect transistor for the recovery gate. Proper operation of this circuit requires a completely isolated AVC line in the receiver with infinite resistance to ground, and use of the MOS transistor in conjunction with a time-constant circuit to control recovery to the maximum-gain bias condition.

The AGC-amplifier input signal is obtained from the output of the receiver noise limiter. This arrangement removes noise peaks which might initiate AGC. A good-quality audio-amplifier section is not required at this point because the control voltage is developed by the peaks on the waveform of the average human voice.

After the signal is amplified, it is passed into the AGC diode, $CR_2$. The output of the diode is applied to the time-constant RC network, $R_{11}$ and $C_6$, which controls the decay time of the AGC. The MOS transistor is connected across the AGC output line, and is in a nonconducting state when the signal is applied. The threshold of conduction is determined by the receiver RF gain control. The RF gain control is isolated from the AGC line by means of the receiver AVC diode, $CR_3$. The time-constant RC network discharges at its normal rate until the gate voltage on the field-effect transistor reaches a point with respect to the source that permits conduction to take place between the drain and the source. At this point, the AGC voltage decay is speeded up until another input signal resets the time constant. As a result, the AGC voltage has fast attack and slow decay to the point of conduction, and its "hang" time is determined by the values of $R_{11}$ and $C_6$. The size of the capacitor, $C_6$, may be varied to change the decay time as desired. A switch can also be used to select two or three different time constants.

For the components shown in Figure 3, the maximum-gain bias voltage is about - 2 volts. There must be no leakage to ground on the AGC bus; even a resistance in the order of megohms will interfere with proper operation of the receiver. For example, the load of VTVM is sufficient to discharge the bus and destroy the "hang" feature. The minimum-gain AGC voltage levels off at about - 7 volts as a result of saturation by the amplifying transistors. Additional limiting is obtained by adjustment of $R_{10}$.

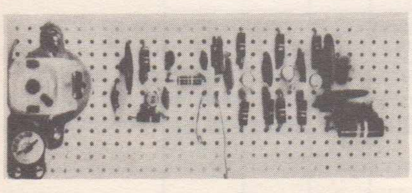


*Figure 4 Layout of product-detector (left) and audio-AGC (right) components on vector board. Short wire in centre of board is the product-detector output; long-wire is the audio input for the AGC system.*

## AGC Operation

For optimum operation, the following parameters of the AGC system are adjusted:

- MAXIMUM-GAIN THRESHOLD. The threshold voltage establishes the conducting state of the recovery gate, $Q_4$, or the maximum-signal fixed-bias level set on the AGC line. This voltage is adjusted by means of the receiver RF gain control.

- MINIMUM-GAIN LIMITING. The amount of AGC bias voltage connected to $Q_2$, $Q_3$, and $CR_2$ depends on the amount of signal voltage present at the base of $Q_2$. This voltage is adjusted by means of $R_{10}$.

- "HANG" TIME CONSTANT. This parameter is defined as the period of time from the initial signal that produces the AGC voltage until the discharge of the RC network to the point where $Q_4$ starts to conduct. This time may be varied by selection of different values of $C_6$.

## Alignment of Product Detector

For alignment of the product detector, a tone-modulated RF signal is applied to the receiver antenna terminals or a local AM signal is tuned in. The function switch is set for AM reception, and the BFO is switched off. An oscilloscope is connected to the output of the product detector. Capacitor $C_{101}$ is set to about half range, and the secondary winding of the 455 kHz IF transformer, $T_{100}$, is adjusted for maximum audio output as indicated on the 'scope. The $T_{100}$ primary is then adjusted for maximum output. If the primary does not peak, capacitor $C_{101}$ is readjusted (starting with minimum capacitance) until peaking occurs. The receiver 455 kHz IF transformer is then readjusted for maximum output to compensate for the loading effect, and all previous adjustments are repeated.

With the BFO on, the receiver is switched for SSB reception. An SSB signal is tuned in, and the BFO is adjusted for normal intelligence. Capacitor $C_{102}$ is then adjusted for maximum audio output. An excess BFO voltage reduces the output.

## Conclusion

Strict attention to details of construction, and careful observance of RCA instructions in the handling of MOS-FET's, should provide the builder with an important additional function for his receiver at a moderate cost.

# A 100-Watt, 18-kHz Inverter Using 2N5202 Silicon Power Transistors

by D.T. De Fino

This Note describes a two-transformer inverter that demonstrates the excellent switching capabilities of the new RCA-2N5202 power transistor. This silicon epitaxial n-p-n device is supplied in the popular TO-66 package. Its fast switching speed makes it especially suitable for use in switching regulators, switching control amplifiers, converters, and inverters. Pertinent characteristics of the 2N5202 are shown in Table I.

Fig. 1 shows a schematic diagram of the two-transistor, two-transformer circuit. A saturable base-drive transformer $T_2$ controls the inverter switching operation. A linearly operating output transformer $T_1$ transfers the output power to the load. The output transformer $T_1$ is not allowed to saturate; therefore, the peak collector current through the transistor is determined principally by the value of the load impedance.

Because no two transistors are perfectly matched, one of the transistors in the inverter circuit conducts more rapidly than the other when the power is turned on. This transistor, $Q_2$ for example, tends toward saturation and causes positive voltages to appear at the dotted ends of the transformers. Thus, there is an effective positive feedback that causes $Q_1$ to switch off and $Q_2$ to switch on. The voltage from the collector of $Q_1$ to the collector of $Q_2$ is then positive and equal to twice the collector supply voltage $V_{CC}$. The voltage $V_{Rfb}$ across the feedback resistor $R_{fb}$ is essentially the product of the resistance $R_{fb}$ and the base current referred to the primary of $T_2$. The voltage across $T_2$ is equal to $2\,V_{CC} - V_{Rfb}$.

At the beginning of the next half-cycle, the voltage across $R_{fb}$ increases very slowly with the slowly increasing magnetizing current through $T_2$. When $T_2$ reaches its saturation flux density, the magnetizing current increases very rapidly and causes a rapid increase in $V_{Rfb}$. As a result, the voltage across $T_2$ decreases rapidly and $Q_2$ comes out of saturation. The collector voltage of $Q_2$ then rises, and regenerative action causes $Q_1$ and $Q_2$ to reverse states. As these processes are repeated during succeeding half-cycles, oscillations are sustained.

Characteristics of the drive transformer and the output trans-

**TABLE I - TYPICAL CHARACTERISTICS OF RCA-2N5202 SILICON POWER TRANSISTOR**

| CHARACTERISTICS | SYMBOLS | TEST CONDITIONS | MIN | MAX | UNITS |
|---|---|---|---|---|---|
| Collector-Cutoff Current | $I_{CEV}$ | $V_{CE}$ = 100 V, $V_{BE}$ = -1.5 V<br>$V_{CE}$ = 100 V, $V_{BE}$ = -1.5 V, $T_C$ = 150°C | -<br>- | 10<br>10 | mA<br>mA |
| Emitter-Cutoff Current | $I_{EBO}$ | $V_{EB}$ = 6 V, $I_C$ = 0 | - | 10 | mA |
| DC Forward-Current Transfer Ratio | $h_{FE}$ | $V_{CE}$ = 1.2 V, $I_C$ = 4 A | 10 | 100 | |
| Collector-to-Emitter Sustaining Voltage | $V_{CER(sus)}$ | $R_{BE}$ = 50 Ω, $I_C$ = 0.2 A | 75 | - | V |
| Base-to-Emitter Voltage | $V_{BE}$ | $V_{CE}$ = 1.2 V, $I_C$ = 4 A | - | 1.9 | V |
| Collector-to-Emitter Saturation Voltage | $V_{CE}(sat)$ | $I_C$ = 4 A, $I_B$ = 0.4 A | - | 1.2 | V |
| Small-Signal Forward-Current Transfer Ratio | $h_{fe}$ | $V_{CE}$ = 10 V, $I_C$ = 0.5 A, f = 10 MHz | 6 | - | |
| Output Capacitance | $C_{ob}$ | $V_{CB}$ = 10 V, $I_E$ = 0, f = 1 MHz | - | 175 | pF |
| Second-Breakdown Collector Current | $I_{S/b}$ | $V_{CE}$ = 40 V (base forward-biased) | 400 | - | mA |
| Second-Breakdown Energy | $E_{S/b}$ | $V_{BB}$ = -4 V, $R_{BE}$ = 50 Ω, L = 50 μH | 0.4 | - | mJ |
| Saturating Switching Times: | | | | | |
| Delay Time | $t_d$ | $V_{CC}$ = 30 V, $I_C$ = 4 A, $I_{B1}$ = 0.4 A | - | 40 | ns |
| Rise Time | $t_r$ | $V_{CC}$ = 30 V, $I_C$ = 4 A, $I_{B1}$ = 0.4 A | - | 400 | ns |
| Storage Time | $t_s$ | $V_{CC}$ = 30 V, $I_C$ = 4 A, $I_{B1}$ = 0.4 A, $I_{B2}$ = -0.4 A | - | 800 | ns |
| Fall Time | $t_f$ | $V_{CC}$ = 30 V, $I_C$ = 4 A, $I_{B1}$ = 0.4 A, $I_{B2}$ = -0.4 A | - | 400 | ns |
| Thermal Resistance, Junction to Case | $\theta_{J-C}$ | | - | 5 | °C/W |

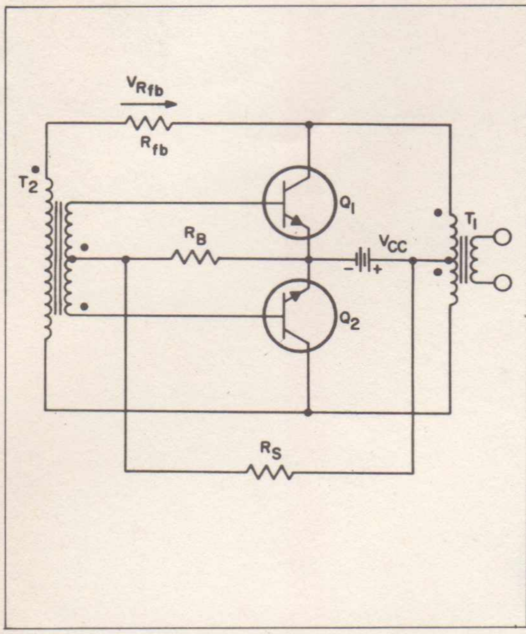



*Fig. 1. – Schematic diagram of two-transistor /two-transformer Inverter*

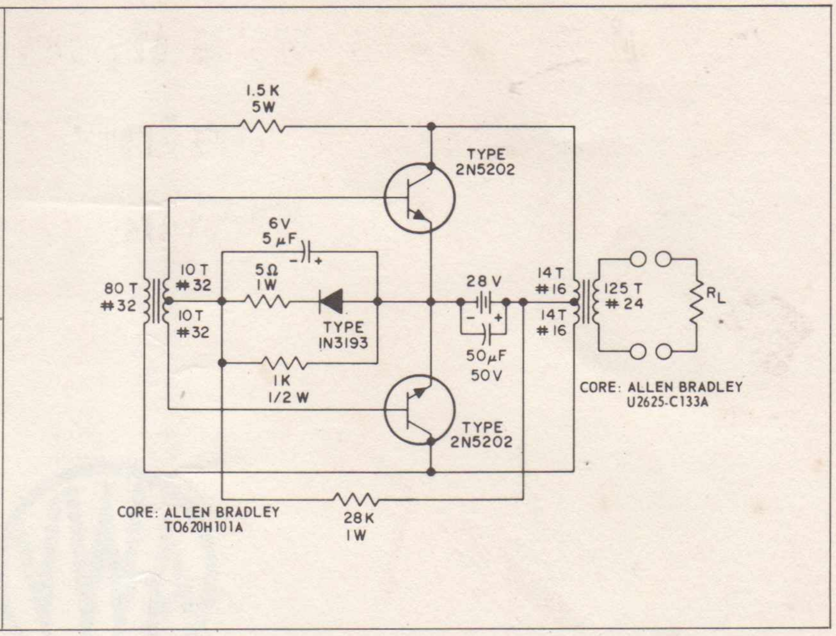



*Fig. 2. – Circuit diagram for 100-watt, 18 kHz inverter*

former used in the circuit of Fig. 1 are determined by means of the following equation:

$$N_p = \frac{V}{4fAB} \times 10^8$$

where $N_p$ is the number of turns in the primary winding, V is the peak voltage across the primary winding, f is the operating frequency in hertz, A is the cross-sectional area of the core in square centimeters, and B is the flux density in gauss. In the design of the drive the transformer $T_2$, the value of flux density B is selected to cause the core to saturate. For the output transformer $T_1$, the value of B is selected to assure that $T_1$ will not saturate. The base resistor $R_B$ is determined by the voltage at the secondary of $T_2$ and the base drive required for the transistor. The resistor $R_B$ is selected so that a voltage of 0.7 volt appears across $R_B$ when the power is turned on initially.*

Fig. 2 shows the current diagram for a practical 100 watt, 18 kHz inverter using RCA-2N5202 transistors. Performance characteristics for this inverter are shown in Fig. 3.

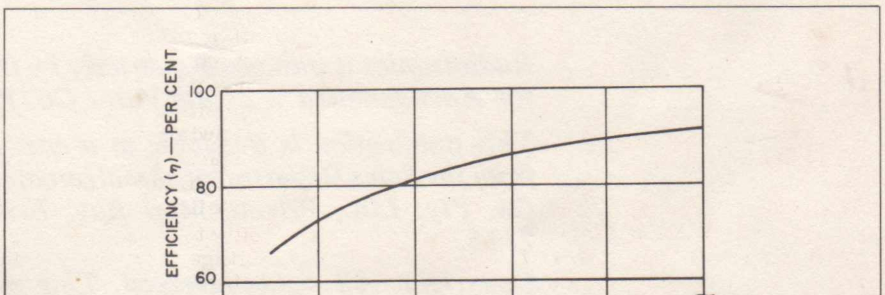



*Fig. 3. – Performance characteristics of inverter shown in Fig. 2.*

*A Complete discussion of inverter design characteristics and design information is given in RCA Application Note SMA-37: "High-Speed Inverters Using Silicon Power Transistors" by H.T. Breece. A copy of this note is available on application.

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