RADIOTRONICS

IN THIS ISSUE

COVER:
Premium quality valves being manufactured at the AWV plant, Rydalmere, N.S.W.

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3 to 130 WATT AUDIO AMPLIFIERS (PART 1.)

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Preamble

An important quality of an amplifier is its ability to increase the power level of the input signal without adding distortion components or changing the frequency vs amplitude characteristics of the signal.

The output power of the amplifier is the parameter which determines the unit's size in relation to its use, and can vary from about 100 mW to hundreds of watts. The design of an amplifier should be such that it is physically small in size, reliable, of low cost with simple circuitry, efficient in operation with no critical adjustments and tolerant to variations in component parameters.

This report will cover amplifier designs from about 3 watts to greater than 100 watts. At present the output power limit for these designs is about 130 watts, beyond which the techniques illustrated cannot be used due to the limits imposed by the maximum ratings of available power transistors.

Amplifiers

The production of medium to high power, low-distortion audio amplifiers covering a wide range of frequencies has, until recently, been difficult and frequently cumbersome.

Transistors, on the surface, appear to be relatively simple three terminal devices with input and output connections between which power amplification is possible. Their operation may be complicated by the effects of internal dissipation and the resultant increase in internal temperature. Germanium types which as a rule have leakage currents two or more orders of magnitude higher than silicon, can be affected seriously in this respect. The improving availability and cost structure of silicon types is making possible the design of circuits which approach the ideal type of amplifier to a greater degree than ever before.

Figures 1, 2, 3, 4 and 5 show respectively a general circuit and a series of designs covering 3, 10-15, 25-50 and 70-130 watt amplifiers.

In all the designs it is possible to modify individual characteristics to suit particular design criteria. The 3 watt amplifier differs from the other designs in that the load is supplied directly from the complementary stage. This arrangement requires the use of complementary transistor types having power dissipation ratings of approximately 3 watts at 50° C. This same arrangement can be used to provide higher power output levels when complementary transistors having higher dissipation levels become available. In the meantime, for higher power units, the use of a single ended output stage following the complementary drivers provides a satisfactory solution.

In all the designs given in this series. the sensitivity of the amplifier and its distortion can be modified over a very wide range by the simple expedient of changing the amount of negative feed back.

Certain sections of one amplifier circuit can be incorporated into the appropriate section of another circuit to obtain a particular characteristic. Some of the possible modifications have been included in the circuits of the main amplifiers

A tabulation of the major performance characteristics of the amplifiers is given below and the component values for each design are given on the appropriate circuit. Using the general circuit shown in Figure 1 as a reference key, and since all the other circuits in this series of amplifiers and their associated modifications use the same fundamental amplifier design, it is possible to simplify the circuit description by giving each particular component in a circuit the same reference number.

Consequently all components carrying the same reference number perform the same function in each amplifier.

Circuit Description

All the main amplifiers are directly coupled between the input and output sections and incorporate a directly coupled feedback system. The feedback has been arranged to provide a gain reduction of more than 60 dB and virtually produces unity voltage gain. The high value of direct coupled feedback together with the use of silicon transistors produces excellent temperature stability and provides a circuit which is tolerant to wide variations in component values. Since the overall voltage gain virtually is unity and an impedance change occurs between input and output terminals, the power gain is very nearly equal to the current gain of the system. This current gain and the corresponding

TABLE 1—AMPLIFIER SPECIFICATIONS

Circuit (Figure Number)	2	3	3	4	4	4	5	5	5	
Nominal Power Output	3	10	15	25	35	50	70	100	130	W.
Output Load Impedance	15	15	15	12	8.5	6.0	10	6.5	5.0	Ω
Power Output before Clipping	3.3	11.5	16	27	37	51	71	105	130	W.
Distortion before Clipping	1.7	0.1	0.1	0.5	0.5	0.5	0.7	1.0	1.0	%
Sensitivity for Full Output	175	270	325	90	95	100	130	125	117	mV.
Noise below full output	69	80	80	70	70	70	75	75	75	dB

overall direct coupled power gain can be high.

The direct coupled amplifier and the feedback system will exhibit very little frequency discrimination until the signal frequency is high enough to affect the current gain of the transistors and cause phase shifts in the coupling or feedback circuits of the amplifier. Freedom from these effects, by the use of suitable transistors, can be obtained from d.c. up to the megahertz region and thereby allow the use of very stable feedback loops. It is possible to separate the d.c. and a.c. components of the feedback loop and thus enable a very high degree of d.c. feedback to be maintained while suitably modifying the a.c. or signal component.

The amount of a.c. feedback included in the d.c. feedback loop is many orders of magnitude higher than required to satisfy even the most stringent specification. Consequently, any reduction in the a.c. feedback will increase the a.c. voltage gain of the system and thus increase the power gain at signal frequencies. This can be achieved by shunting controlled amounts of signal frequency feedback from the main feedback line. Referring to the general circuit in figure 1, the resistor R7 is added as a controlled shunt for all signals in the feedback line. The d.c. component is removed by the inclusion of the capacitor C3.

Since the overall loop gain is very high, the overall signal frequency voltage gain with the modified a.c. feedback, is determined at mid frequencies by the relationship:

$$\text{Voltage gain } V_g = \frac{R7 + R9}{R7} \qquad (1)$$

The accuracy of equation (1) which is a simplification of a more complex expression, will depend upon the relative values of R7 and R9 and diminishes as R7 is reduced. However, until the gain exceeds a few hundred, i.e., before the feedback factor has dropped too far, the calculation will be sufficiently accurate. For example, when R7 is zero the feedback will be zero, the gain will be a maximum and equation (1) will no longer apply.

The capacitor C3 provides d.c. isolation between the collector of Q2 and the base of Q2 or Q8 as well as separating the d.c. and a.c. components of the feedback loop.

In a subsequent article a brief treatment of each section will be given, with emphasis on any salient points. This will provide an explanation of the operation of the amplifiers as a whole and also help the user to select an appropriate set of units for the required application.

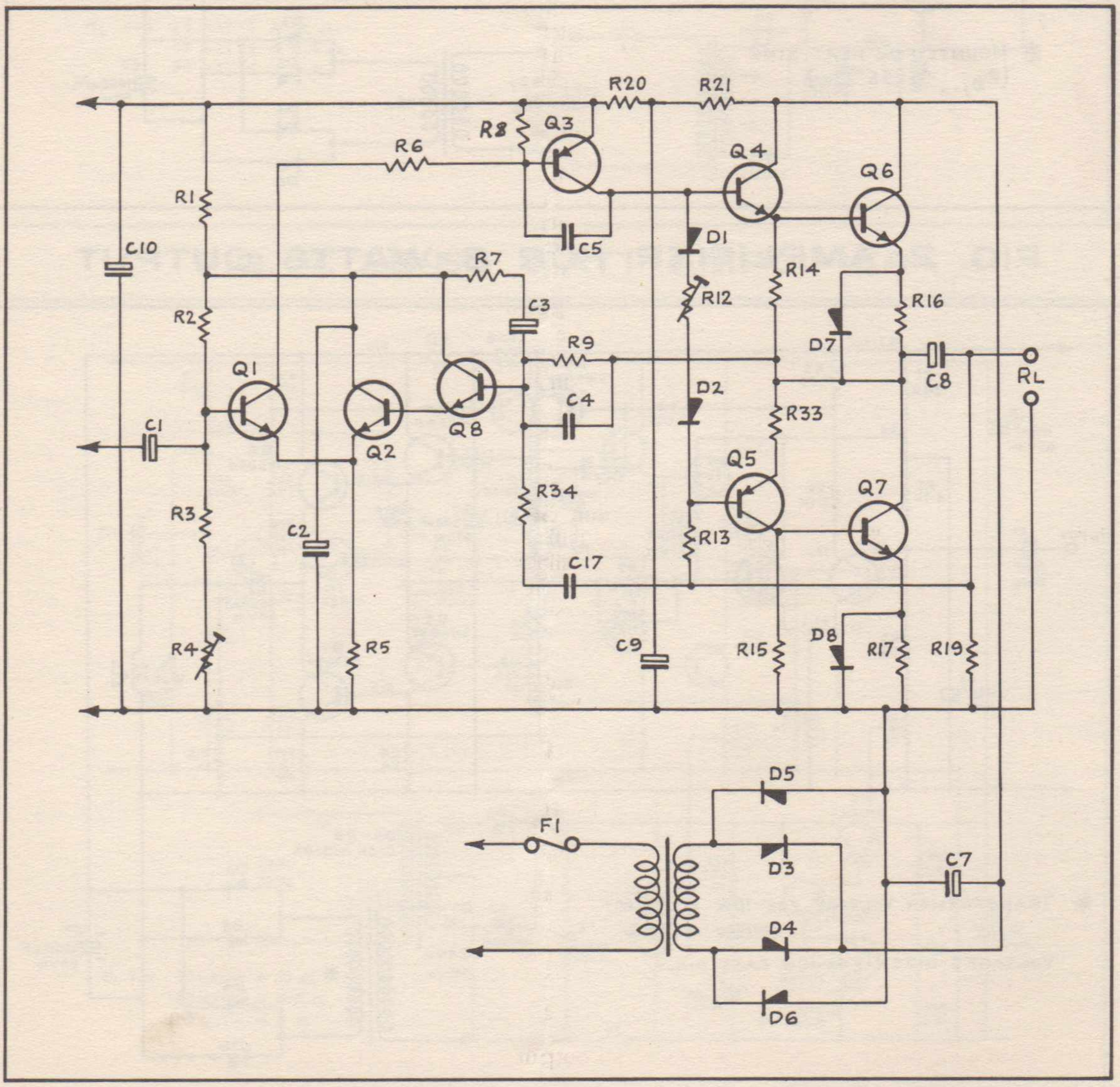


FIG. 1. BASIC AMPLIFIER CIRCUIT

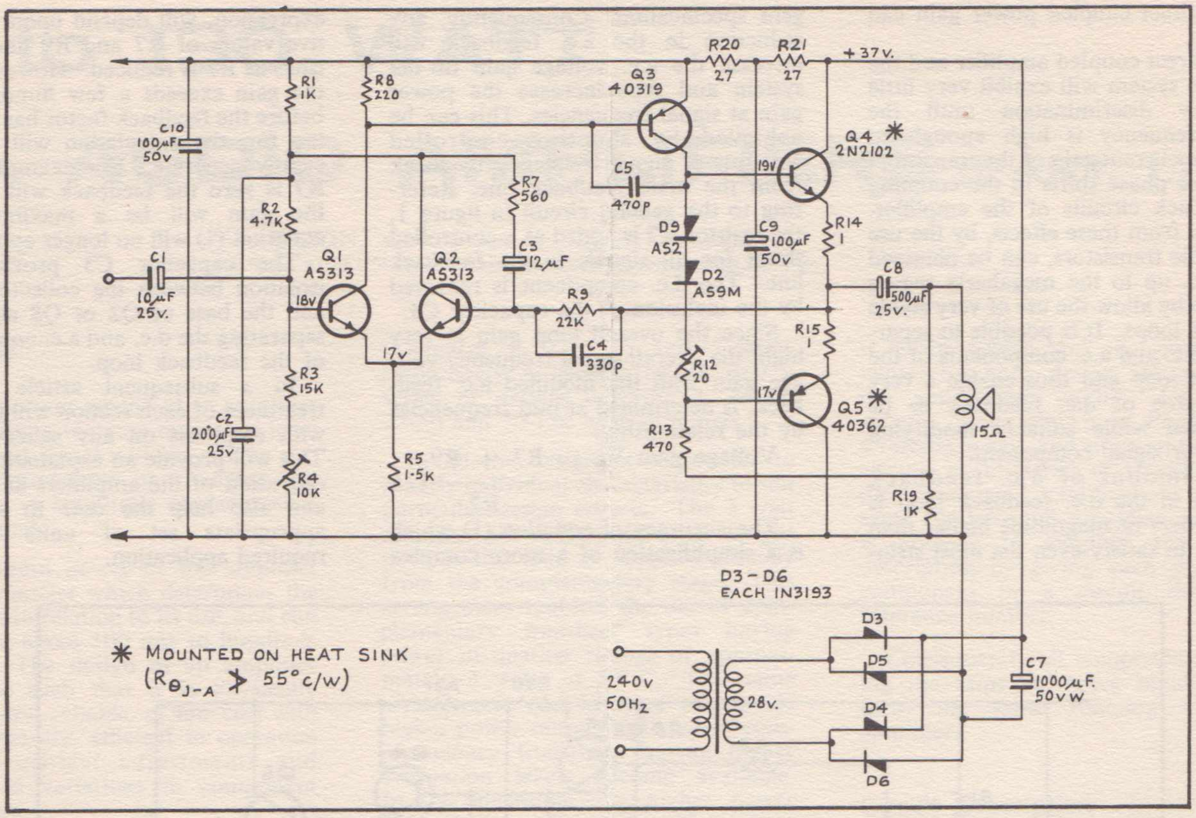


FIG. 2. AMPLIFIER FOR 3 WATTS OUTPUT

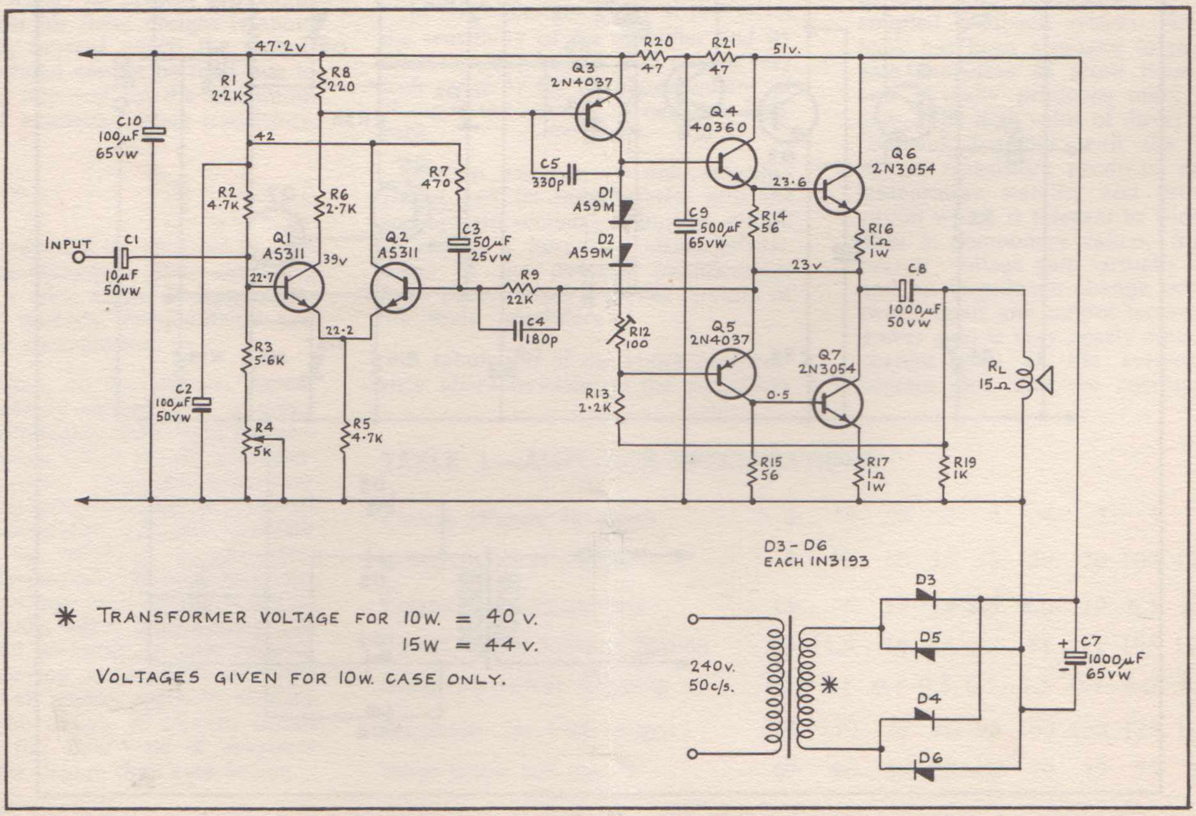


FIG. 3. 10-15 WATT DESIGN

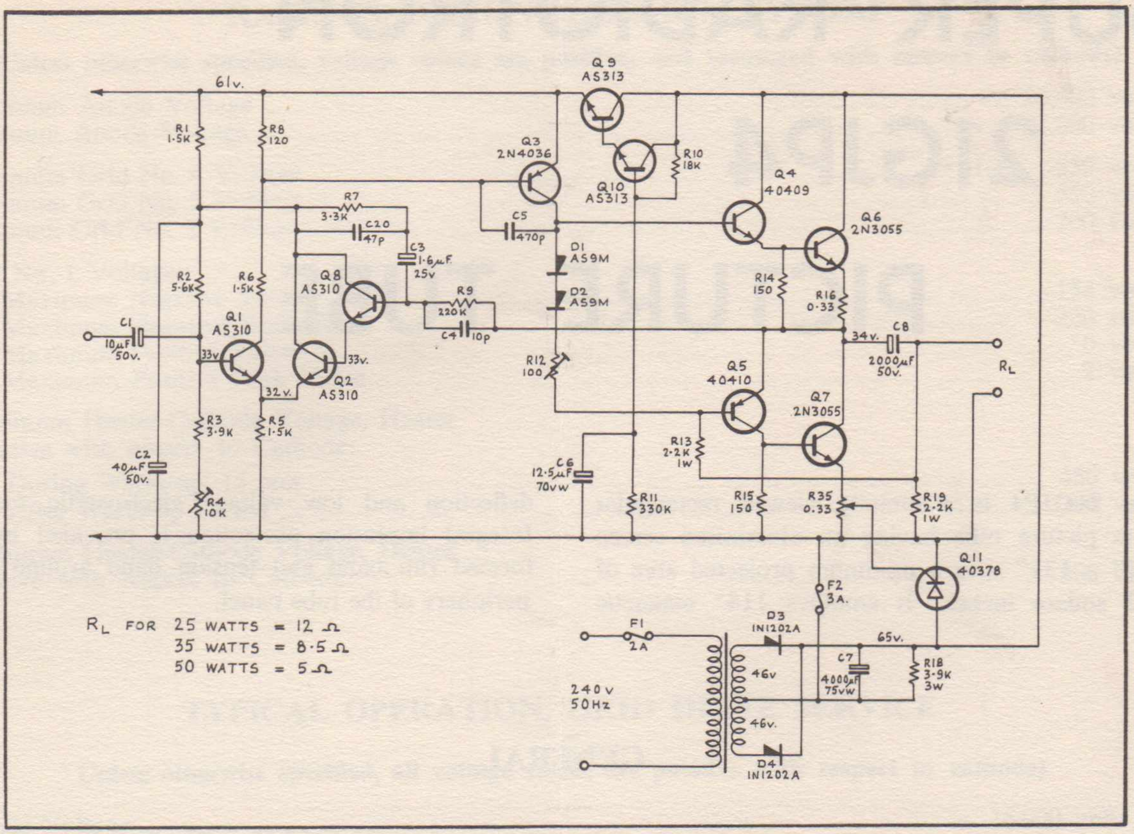


FIG. 4. 25-50 WATT AMPLIFIER

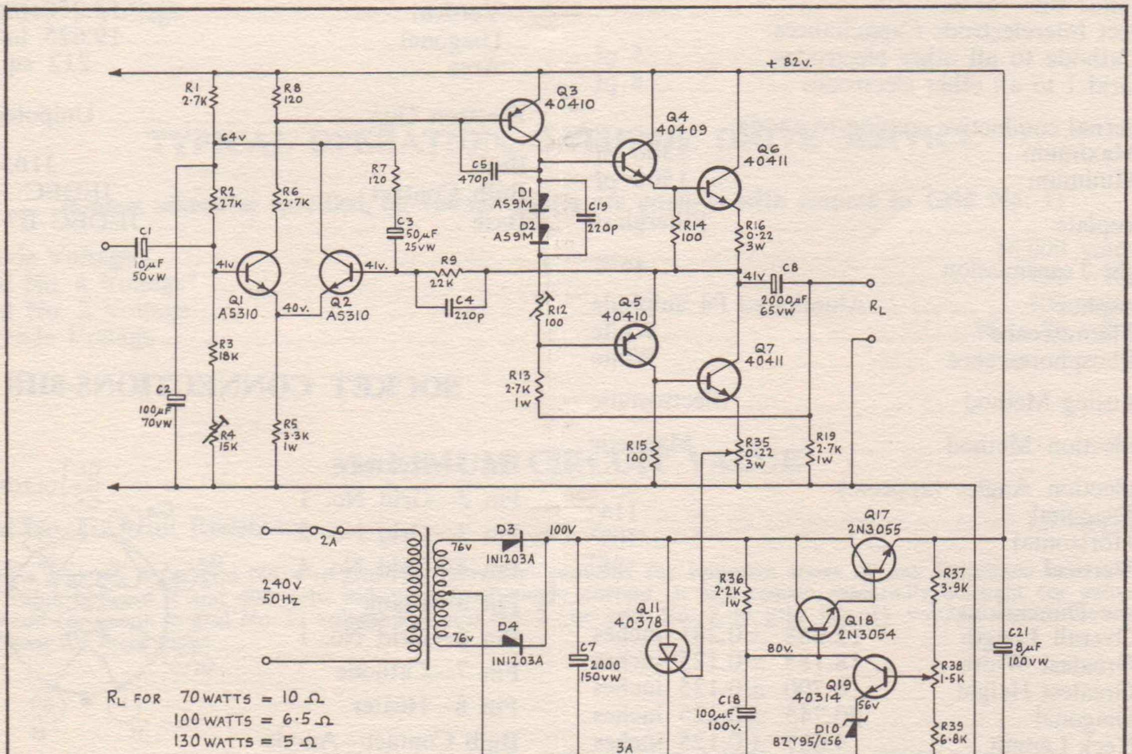


FIG. 5. 70-130 WATT DESIGN

SUPER RADIOTRON
21GJP4
PICTURE TUBE

The 21GJP4 is a directly viewed rectangular glass picture tube having an aluminised screen 16$\frac{7}{8}$″ x 13$\frac{1}{4}$″ with a minimum projected area of 212 square inches. It employs 114° magnetic deflection and low voltage electrostatic focus. Integral implosion protection is provided by a formed rim band and tension band around the periphery of the tube panel.

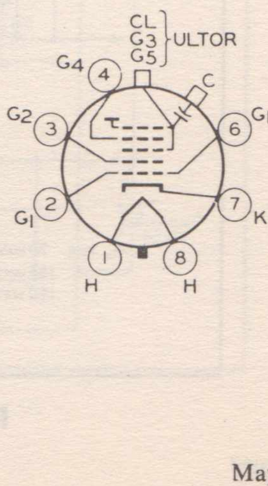
GENERAL

Heater Voltage	6.3 volts
Heater Current	0.6 amp
Direct Interelectrode Capacitances:	
Cathode to all other electrodes	5 pf
Grid 1 to all other electrodes	6 pf
External conductive coating to anode:	
Maximum	2300 pf
Minimum	1500 pf
Faceplate	Filterglass
Light Transmission	46%
Phosphor	Aluminised P4 Sulphide
Fluorescence	White
Phosphorescence	White
Focusing Method	Electrostatic
Deflection Method	Magnetic
Deflection Angles (approx.):	
Diagonal	114°
Horizontal	102°
Vertical	85°
Tube Dimensions:	
Overall Length	12.968 ±0.281 inches
Greatest Width	18.185 ±0.125 inches
Greatest Height	14.700 ±0.125 inches
Diagonal	20.745 ±0.125 inches
Neck Length	4.687 ±0.125 inches

Screen Dimensions (min.):	
Horizontal	16.875 inches
Vertical	13.250 inches
Diagonal	19.625 inches
Area	212 sq. in.
Electron Gun	Unipotential
Bulb	J165$\frac{1}{4}$AI
Bulb Contact	JEDEC J1-21
Base	JEDEC B7-183

SOCKET CONNECTIONS-8HR

Pin 1—Heater
Pin 2—Grid No. 1
Pin 3—Grid No. 2
Pin 4—Grid No. 4
Pin 5—Blank
Pin 6—Grid No. 1
Pin 7—Cathode
Pin 8—Heater
Bulb Contact—Anode



RATINGS, DESIGN MAXIMUM SYSTEM

(Unless otherwise specified, voltage values are positive, and measured with respect to cathode)

Maximum Anode Voltage	20,000 volts
Minimum Anode Voltage	11,000 volts
Maximum Grid No. 4 Voltage	+1100, —550 volts
Maximum Grid No. 2 Voltage	550 volts
Minimum Grid No. 2 Voltage	200 volts
Grid No. 1 Voltage:	
Maximum Negative Value	—154 volts
Maximum Negative Peak Value	—220 volts
Maximum Positive Value	0 volts
Maximum Positive Peak Value	2 volts
Maximum Heater-Cathode Voltage, Heater Negative with respect to Cathode:	
During Warm-up, 15 secs	450 volts
After Warm-up Period	200 volts
Maximum Heater-Cathode Voltage, Heater Positive with respect to Cathode:	200 volts

TYPICAL OPERATION, GRID DRIVE SERVICE

Unless otherwise specified, all voltage values are positive with respect to cathode)

Anode Voltage	16,000 volts dc
Grid No. 4 Voltage*	0-400 volts dc
Grid No. 2 Voltage	400 volts dc
Grid No. 1 Voltage	—36 to —94 volts dc

TYPICAL OPERATION, CATHODE DRIVE SERVICE

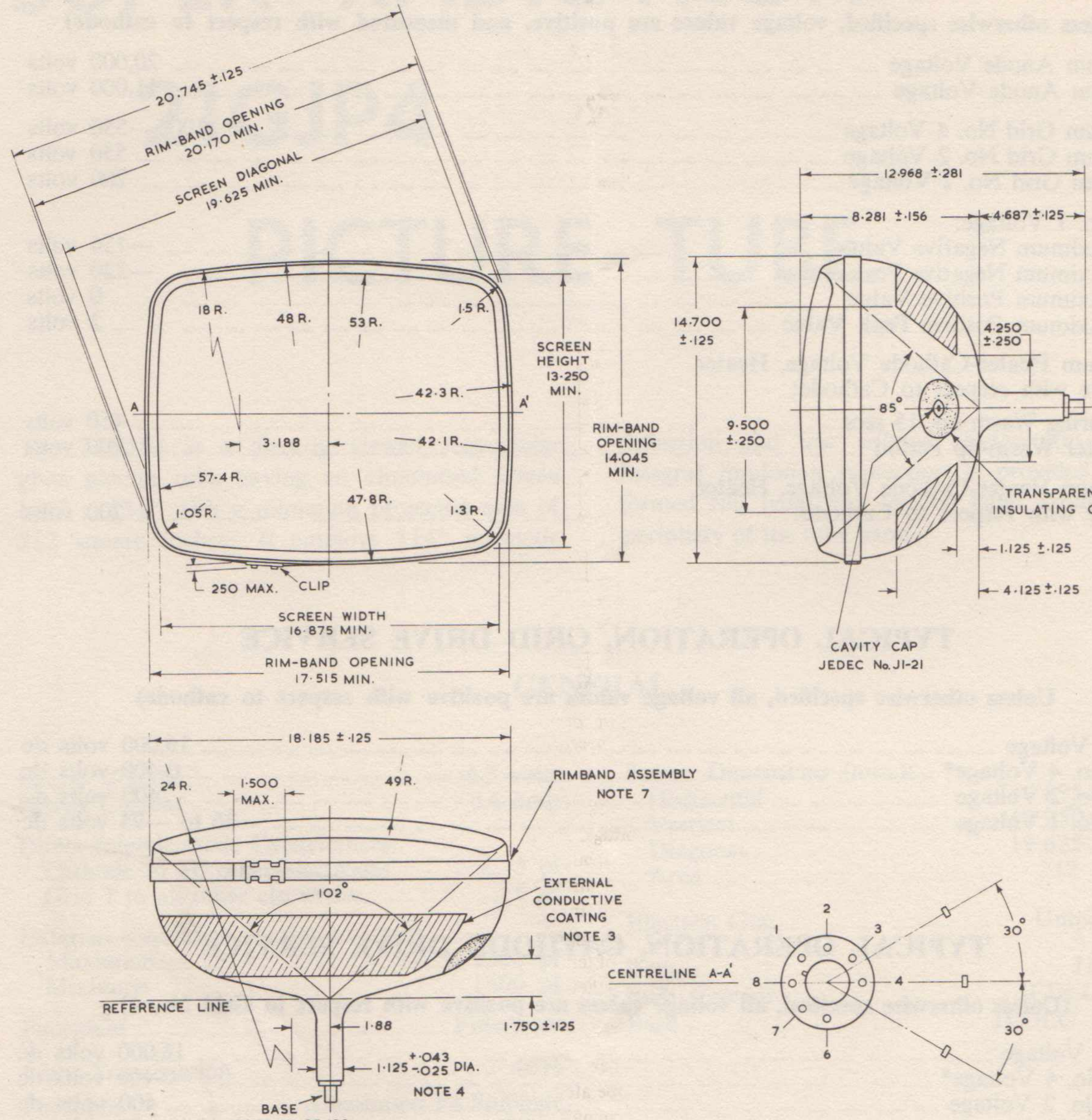
(Unless otherwise specified, all voltage values are positive with respect to Grid No. 1)

Anode Voltage	16,000 volts dc
Grid No. 4 Voltage*	0-400 volts dc
Grid No. 2 Voltage	400 volts dc
Cathode Voltage	36 to 78 volts dc

MAXIMUM CIRCUIT VALUE

Grid No. 1 Circuit Resistance	1.5 megohms

* The grid No. 4 (or grid No. 4 to grid No. 1) voltage required for optimum focus of any individual tube will be a value between 0 and 400 volts independent of anode current. It will remain essentially constant for values of anode (or anode to grid No. 1) voltage and grid No. 2 (or grid No. 2 to grid No. 1) voltage within the ranges shown for these items.



21GJP4

NOTES

NOTE 1. Yoke Reference Line is determined by plane surface of flared end of JEDEC Reference Line Gauge No. 126 when seated on funnel of tube. With minimum neck length tube, the PM centring magnet should extend no more than 2¼″ from Yoke Reference Line.

NOTE 2. Lateral strains on the base pins must be avoided. The socket should have flexible leads permitting movement. The perimeter of the base wafer will be inside a 1¾″ diameter circle concentric with the tube axis.

NOTE 3. External conductive coating forms supplementary filter capacitor and must be grounded

NOTE 4. Neck diameter may be a maximum of 1.168″ at the splice.

NOTE 5. Base pin No. 4 aligns with centreline A-A′ within 30° and is on the same side as anode contact J1-21.

NOTE 6. To clean this area, wipe only with a soft, dry lintless cloth.

NOTE 7. The Rimband assembly must be grounded.

APPLICATION OF THE RCA CA3015 AND CA3016 INTEGRATED-CIRCUIT OPERATIONAL AMPLIFIERS

The RCA integrated-circuit operational amplifiers CA3015 and CA3016 are identical in circuit configuration to the previously announced types CA3008 and CA3010, but have an improved device breakdown voltage that permits operation from ± 12-volt supplies. Operation of the CA3015 and CA3016 from power supplies of ±6 volts or ±3 volts is the same as for the earlier types.[1] This Note describes the operating characteristics of the CA3015 and CA3016 at ±12 volts, and discusses applications that take advantage of the higher gain-bandwidth product and increased output signal swing obtained at the higher voltages.

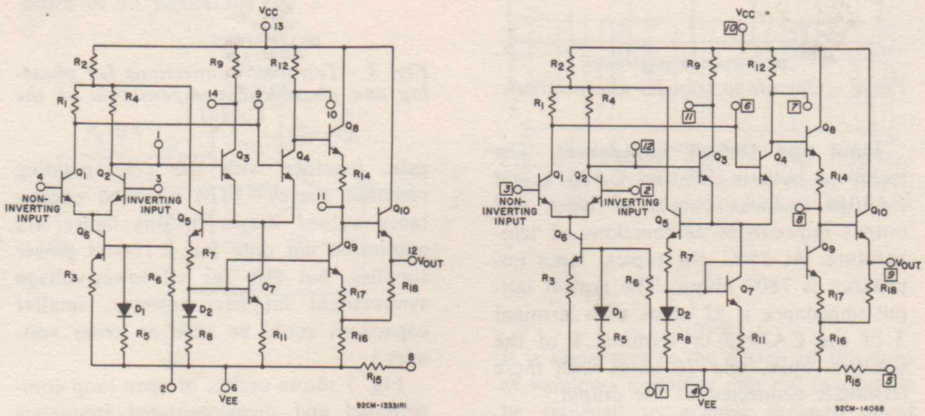


Fig. 1.—Schematic diagrams and terminal connections for the CA3015 and CA3016 integrated-circuit operations amplifiers.

Operating Characteristics

Fig. 1 shows the schematic diagrams and terminal numbers for the CA3015 and CA3016. As in the case of the CA3008 and CA3010, each operational amplifier consists basically of two differential amplifiers and a single-ended output circuit in cascade. The CA3015 is supplied in a 12-lead TO-5 package, and the CA3016 in a 14-lead flat package. Throughout this Note, terminal numbers for the CA3015 are shown in the illustrations; however, the discussion applies to both packages.

DC Characteristics

When operated from ±12-volt power supplies, these operational amplifiers have a typical dissipation of 175 milliwatts with terminal 5 of the CA3015 or terminal 8 of the CA3016 open. If terminals 5 and 9 of the CA3015 or terminals 8 and 12 of the CA3016 are shorted, higher output-current capability can be achieved, but the dissipation increases to a typical value of 500 milliwatts. The input offset voltage is typically 1.4 millivolts, and the variation in input offset voltage is typically less than 200 microvolts per volt for fluctuations of either supply voltage. At 25° C, the input bias current and input offset current are typically 9.6 and 1 microamperes, respectively. (Curves of input offset voltage, input bias current, and input offset current as functions of temperature are given in the technical bulletin for the CA3015 and CA3016.)

Derating. When these operational amplifiers are operated from ±12-volt supplies with terminals 5 and 9 of the CA3015 or terminals 8 and 12 of the CA3016 shorted for greater output capability, the power dissipation is high enough so that temperature derating is necessary. The maximum junction-temperature rating is 150°C, and the thermal resistance is 100°C per watt. The maximum power-dissipation rating is 600 milliwatts at 25°C (with terminals shorted as described above). In this higher-output mode, the circuits can operate safely at ambient temperatures up to 90°C.

AC Characteristics

Transfer Characteristic. The open-loop transfer characteristic is shown in Fig. 2. As in the case of the CA3008 and CA3010, there is no hysteresis effect. The CA3015 and CA3016 technical bulletin includes curves of maximum peak-to-peak voltages as functions of load resistance with terminal 5 of the CA3015 or terminal 8 of the CA3016 open and with terminals 5 and 9 of the CA3015 or terminals 8 and 12 of the CA3016 shorted. The CA3015 and CA3016 can drive a lower-resistance load when these terminals are shorted.

Gain vs Frequency Response. The open-loop low-frequency gain of the CA3015 and CA3016 with ±12-volt supplies is typically 70 dB with a 3-dB bandwidth of 320 kHz. The unity-gain crossover occurs at a frequency of 58 MHz.

Common-Mode Rejection. The common-mode rejection ratio of the CA3015 and CA3016 is typically 104 dB for operation with ±12-volt supplies. A curve of common-mode rejection ratio as a function of frequency is included in the bulletin.

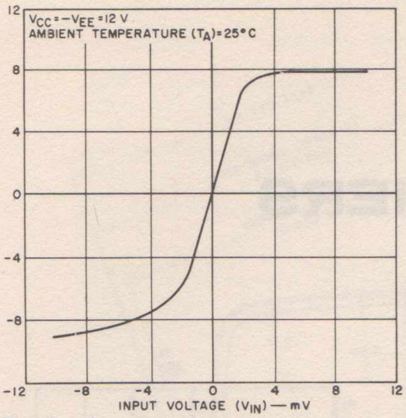


Fig. 2 - Open-loop transfer characteristic.

Input and Output Impedances. The technical bulletin for the CA3015 and CA3016 includes curves of input and output impedances as functions of temperature. At 25°C, the typical input impedance is 7800 ohms. The typical output impedance is 92 ohms with terminal 5 of the CA3015 or terminal 8 of the CA3016 open, and 76 ohms with these terminals connected to the output.

Phase Compensation

The following section describes phase-lag and phase-lead compensation techniques for these operational amplifiers. Fig. 3 shows the various phase-compensation connections for the CA3015.

Phase-Lag Compensation

When the CA3015 and CA3016 are operated from ±6-volt supplies, the phase-compensation techniques described previously for the CA3008 and CA3010 are applicable. When the CA3015 and CA3016 are operated from ±12-volt supplies, corrections must be made in the phase-lag compensation to allow for the shift in frequency at which the second break in the open-loop Bode plot occurs. At ± volts, this second break occurs at a frequency of 10 MHz. For Miller-effect and conventional phase-lag compensation, the series RC combinations must be adjusted so that $1(2\pi RC) = 10$ MHz to correct for the shift in frequency. In addition, the Miller technique requires a larger value of phase-lag capacitance for a non-peaking (±1 dB) response to allow for the higher gain.

Fig. 4 shows a curve of the required phase-lag capacitance as a function of

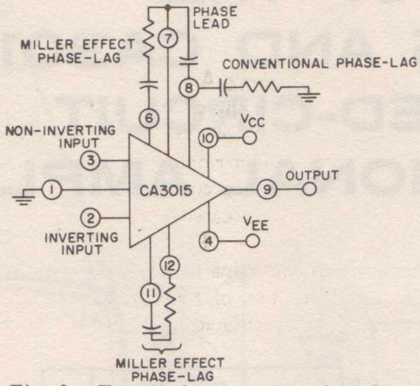


Fig. 3 - Terminal connections for phase-lag and phase-lead compensation of the CA3015.

gain, together with the corresponding response curves. (The required capacitance values shown in this figure are applicable not only for ±12-volt power supplies, but also for all lower-voltage symmetrical supplies; however, smaller capacitors could be used at lower voltages.)

Fig. 5 shows curves of open-loop compensated and uncompensated frequency response with ±12-volt supplies. Al-

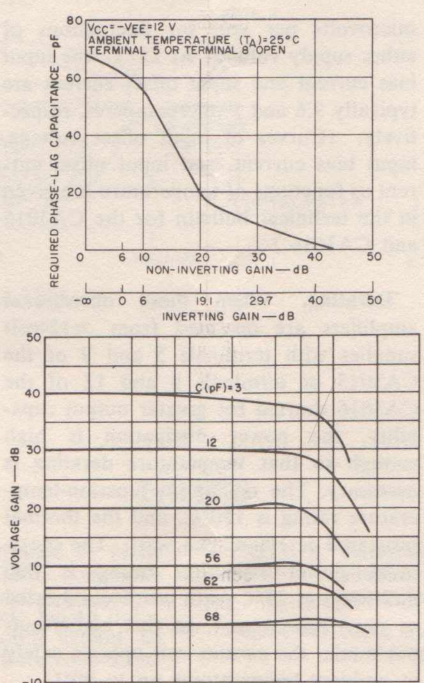


Fig. 4 - Amount of phase-lag capacitance required to obtain a flat (±1 dB) response, and typical response characteristics.

though the phase-lag compensation capacitance of 18 picofarads shown in curve (B) of this figure is sufficient to provide stability in resistive-feedback amplifiers down to unity gain, it is not sufficient to provide flat closed-loop response (±1 dB) below 20 dB.

Phase-Lead Compensation

In addition to the standard phase-lag compensation discussed above, the CA3015 and CA3016 have a phase-lead compensation capability. For this phase-lead compensation, a capacitor is connected between terminals 7 and 8 of the CA3015, as shown in Fig. 3, or between terminals 10 and 11 of the CA3016. The effect of this capacitor is to eliminate the break at 10 MHz in the Bode plot and thus extend the 6-dB-per-octave roll-off. The second break in the Bode plot then occurs at approximately 35 MHz, and the unity-gain crossover occurs at 150 MHz. The phase-lead compensated open-loop response is shown in curves (C) and (D) of Fig. 5 for various values of capacitance. For optimum performance, a minimum phase-lead capacitance of 47 picofarads is recommended.

For flat (±1 dB) responses at closed-loop gains below 30 dB, a small amount of phase-lag compensation is required in addition to the phase-lead compensation. The required phase-lag capacitance for flat (±1 dB) responses and the corresponding response curves are shown in Fig. 6. When phase-lead compensation is used, the series RC combinations should be adjusted so that $1/(2\pi RC) = 35$ MHz.

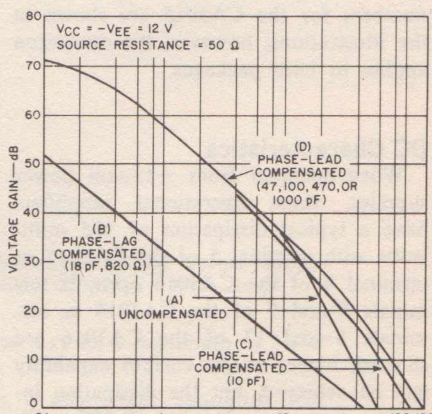


Fig. 5 - Open-loop gain as a function of frequency for compensated and uncompensated amplifiers.

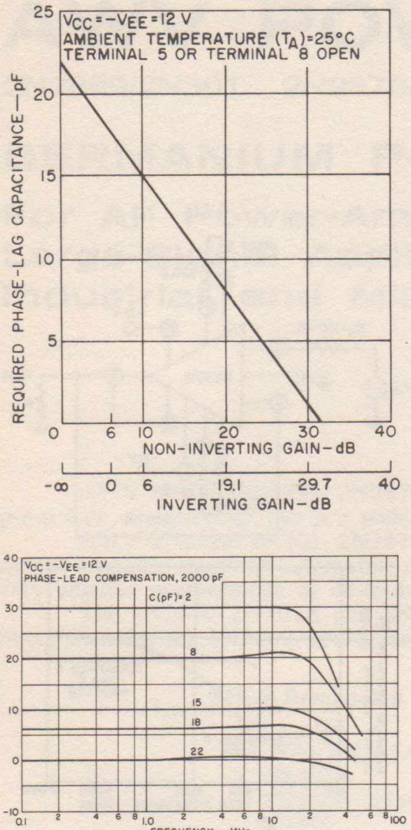


Fig. 6 - Amount of phase-lag capacitance required to obtain a flat (± 1dB) response when phase-lead compensation is used, and typical response characteristics.

The phase-lead compensation is also applicable when ±6-volt power supplies are used, and provides a unity-gain crossover improvement of about one octave as compared to the uncompensated connection. As mentioned earlier, the phase-lag capacitance requirement for ±12-volt supplies shown in Fig. 4 is satisfactory for ±6-volt supplies, although smaller capacitors could be used with the lower voltages.

Applications

For all applications, ac and dc balance at the input must be preserved, i.e., the two inputs must be returned to ground through equal impedances.

50-dB Amplifier

Fig. 7 shows the circuit configuration and frequency response for a non-inverting, 50-dB amplifier employing phase-lead compensation. This amplifier has a 3-dB bandwidth of 3.5 MHz, and a unity-gain crossover at 150 MHz.

10-dB, 42-MHz Amplifier

Fig. 8 shows the circuit diagram and frequency response for a 10-dB, non-inverting amplifier employing both phase-lead and phase-lag compensation. Slight peaking (2 dB) occurs for the phase compensation shown. Flat response with bandwidth reduction to 25 MHz may be obtained by use of a phase-lag capacitance of 15 picofarads.

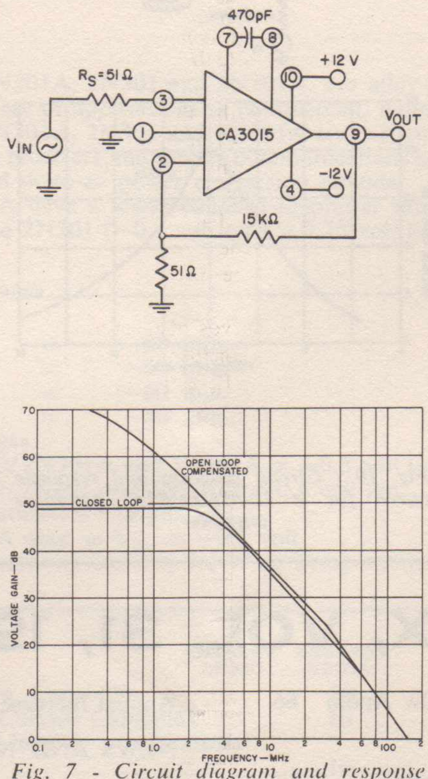


Fig. 7 - Circuit diagram and response curve for a 50-dB, non-inverting amplifier with phase-lead compensation.

Twin-T Bandpass Amplifier

Fig. 9 shows the circuit diagram and frequency response of a bandpass amplifie. using a twin-T network in the feedba:k loop. The difference in resonant frequency between the bandpass-amplifier response and the twin-T network response is caused by device capacitances and loading effects. The unloaded Q (Q_0) of the twin-T network is 14.4; the Q_0 of the bandpass amplifier is 12.8.

The symmetrical twin-T network can be designed by use of the following equations:

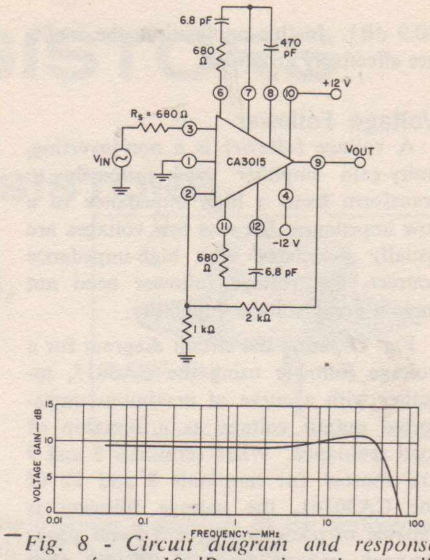


Fig. 8 - Circuit diagram and response curve for a 10-dB, non-inverting amplifier with phase-lead and phase-lag compensation.

$$R_1 = 2 R_2$$
$$C_1 = \tfrac{1}{2} C_2$$
$$f_0 = 1/(2\pi RC)$$

It is important in the design of this type of bandpass amplifier that the two inputs be returned to ground through equal resistances; in this case a value of 2000 ohms is used.

20-dB, 10-MHz Bandpass Amplifier

Fig. 10 shows the circuit diagram and frequency response of an RLC bandpass amplifier. This amplifier is designed to have a Q_0 of about 10 ($R_p = X_c Q_0 = 2200$ ohms) and a gain of about 20 dB at resonance (2200/200 = 11, or

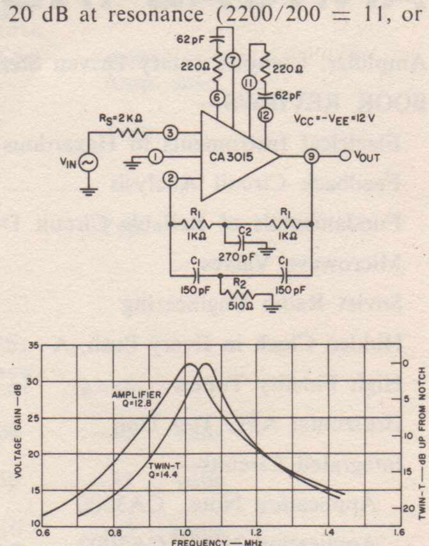


Fig. 9 - Circuit diagram and response curves for a bandpass amplifier using a twin-T network.

20.9 dB). In this application, the inputs are effectively grounded.

Voltage Follower

A voltage follower is a non-inverting, unity-gain amplifier used primarily to transform from a high impedance to a low impedance. Because low voltages are usually associated with high-impedance sources, the voltage follower need not have a great voltage capability.

Fig. 11 shows the circuit diagram for a voltage follower using the CA3015, together with a curve of maximum undistorted output voltage as a function of load resistance. When terminals 5 and 9 are shorted (or terminals 8 and 12 of the CA3016), the voltage follower is capable of transforming a 3.4-volt peak-to-peak voltage from a 100,000 ohm source to a 470-ohm load.

If higher voltage-swing capability is desired, the positive supply voltage (V_{CC}) may be increased. Temperature derating may be necessary, depending on the magnitude of V_{CC} and whether the high- or low-current mode is used.

Reference

1. "Application of the RCA CA3008 and CA3010 Integrated-Circuit Operational Amplifiers", RCA Integrated Circuits Application Note ICAN-5015, November 1965.

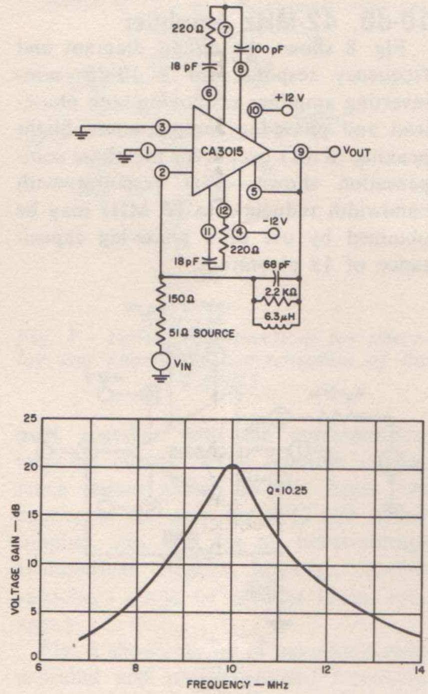


Fig. 10 - Circuit diagram and response curve for a 10-MHz RLC bandpass amplifier.

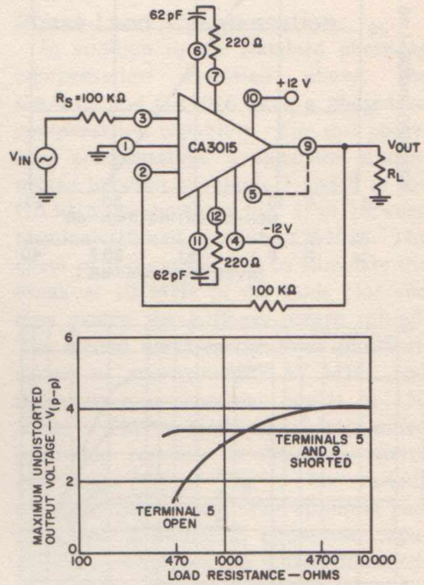


Fig. 11 - Circuit diagram for a voltage follower driven from a 100,000-ohm source, and curve showing maximum undistorted output voltage as a function of load resistance.

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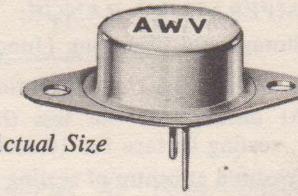
AWV POWER TRANSISTORS

2N2869/2N301 2N2870/2N301A 2N301 2N301A

GERMANIUM P-N-P POWER TRANSISTORS

For AF Power-Amplifier and other Large-Signal Applications in Commercial, Industrial, and Military Equipment



Actual Size

AWV 2N2869/2N301, 2N2870/2N301A, 2N301 and 2N301A, are alloy-junction power transistors of the germanium p-n-p type, designed for use in a wide variety of applications in commercial, industrial, and military equipment.

AWV 2N2869/2N301, 2N2870/2N301A, 2N301 and 2N301A are particularly suitable for use in class A and class B of-output-amplifier stages of automobile radio receivers and mobile communications equipment. They are also capable of providing very efficient performance in dc-to-dc and dc-to-ac power conversion circuits.

The 2N2870/2N301A and 2N301A have a lower collector-to-emitter saturation voltage (−0.3 volt typical, −0.5 volt max.) than for the 2N2869/2N301 and the 2N301 (−0.4 volt typ., −0.75 volt max.).

- High Breakdown Voltages
  BV_{CBO} for
  2870/2N301A = −80 min.
  2N301A = −60 min.
  2N2869/2N301 = −60 min.
  2N301 = −40 min.
- Low Saturation Voltages
  $V_{CE(sat)}$ for 2N2870/2N301A and 2N301A = −0.3 volt typ.
  $V_{CE(sat)}$ for 2N2869/2N301 and 2N301 = −0.4 volt typ.
- Very Low Collector Saturation Current
  $I_{CBO}(_{sat}) = -100\ \mu A$ max. at $V_{CB} = -0.5$ volt
  Assures Excellent Operating Stability for Wide Temperature Variations

---

	2N2869/ 2N301	2N2870/ N301A	2N301 2N301A	
● High Dissipation Capability		30	12.5	watts
● High Collector Current Capability		10	5	Amp. max.
● Excellent Beta Linearity — essentially linear up to		10	5	Amp.
● Power Output—at V_{CC} = 14.4 volts				
Class A_1 single ended		5	5	
Class B push pull		12	10	
● High Large-Signal (DC) Beta = ...		90	90	Typ.
● Hermetically Sealed JEDEC TO-3 Package				

MAXIMUM RATINGS, ABSOLUTE-MAXIMUM VALUES:

	2N2870/ 2N301A	2N301A	2N2869/ 2N301	2N301	
Collector-to-Base Voltage, V_{CBO}	−80	−60	−60	−40	volts
Collector-to-Emitter Voltage V_{CEO}	−50	−32	−50	−32	volts
Emitter-to-Base Voltage, V_{EBO}	−10	−5	−10	−5	amps
Collector Current, I_C	−10	−5	−10	−5	amps
Emitter Current, I_E	+10	+5	+10	−5	amps
Base Current, I_B	−3	−1	−3	−1	amps

TRANSISTOR DISSIPATION:

At Mounting Flange Temperatures*—

	2N2869/2N301 2N2870/2N301A	2N301 2N301A	
Up to −55° C	30	—	watts
−81° C	—	12.5	watts
Above +55° C	*See Fig. 1*	*See Fig. 1*	

TEMPERATURE RANGE:

Storage and Operating (Junction)	−65 to +100		° C

PIN TEMPERATURE (During Soldering):

At distance of not less than $\frac{1}{32}''$ from seating surface for 10 seconds max.	255	255	° C

*Measured at centre of seating surface.

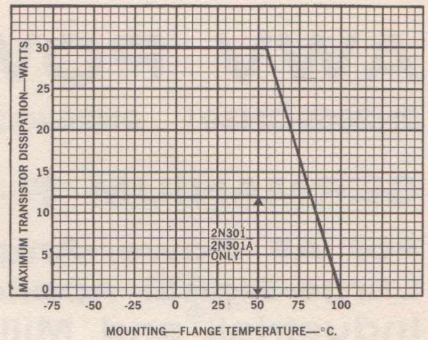


Fig. 1 — Rating Chart for Types 2N2869/2N301, 2N2870/2N301A, 2N301 and 2N301A.

Electrical Characteristics, *at a Mounting-Flange Temperature, T_{MF}*, of 25° C*

Characteristics	Symbols	TEST CONDITIONS					LIMITS						Units
		DC Collector-to-Base Voltage V_{CB}	DC Collector-to-Emitter Voltage V_{CE}	DC Collector Current I_C	DC Emitter Current I_E	DC Base Current I_B	Type 2N2896/ 2N301 2N301			Type 2N2870/ 2N301A 2N301A			
		Volts	Volts	Amp.	mA	Amp.	Min.	Typ.	Max.	Min.	Typ.	Max.	
Collector-to-Base Breakdown Voltage	BV_{CBO}			−0.0005	0		−60 −40	—	—	−80 −60	—	—	Volts
Collector-to-Emitter Breakdown Voltage	BV_{CEO}			−0.6		0	−50 −32	—	—	−50 −32	—	—	Volts
Emitter-to-Base Breakdown Voltage	BV_{EBO}			0	−2		−10	—	—	−10	—	—	mA
Collector-Cutoff Current	I_{CBO}	−30			0		—	—	0.5 3.0	—	—	0.5 3.0	mA
Saturation Collector-Cutoff Current	$I_{CBO(sat)}$	−0.5			0		—	—	0.1	—	—	0.1	mA
DC Forward-Current Transfer Ratio	h_{FE}		−2	−1			50	90	165	50	90	165	—
Base-to-Emitter Voltage	V_{BE}		−2	−1			—	−0.3	−0.5	—	−0.3	−0.5	Volt
Collector-to-Emitter Saturation Voltage	$V_{CE(sat)}$			−10		−1	—	−0.4	−0.75	—	−0.3	−0.75	Volt
Gain-Bandwidth Product	f_T		−2	−1			200	450	—	200	450	—	kHz

* Measured at center of seating surface.

TYPICAL OPERATION FOR TYPES 2N2869/2N301, 2N2870/2N301A, 2N301 AND 2N301A IN THE CLASS B PUSH-PULL AUDIO POWER-AMPLIFIER CIRCUIT OF FIG. 2:

At Mounting-Flange Temperature of 80° C and Signal Frequency of 400 cps Unless otherwise specified, values are for two transistors.

DC Supply Voltage	−14.4	volts
Zero-Signal DC Collector Current□	−0.05	amp
Zero-Signal DC Base-to-Emitter Voltage	−0.13	volt
Peak Collector Current□	−2	amp
Max.-Signal DC Collector Current□	−0.64	amp
Signal-Source Impedance (per base)	10	ohms
Load Impedance (Per Collector)	6	ohms
Power Gain•	30	db
Max.-Signal Power Output•	12	watts
Total Harmonic Distortion (At Power Output of 12 watts)	5% max.	
Circuit Efficiency (At Power Output of 12 watts)•	67	%
Collector Dissipation (Per Transistor at Power Output of 12 watts)	3	watts

•Measured at the primary of the output transformer.
□Per transistor.

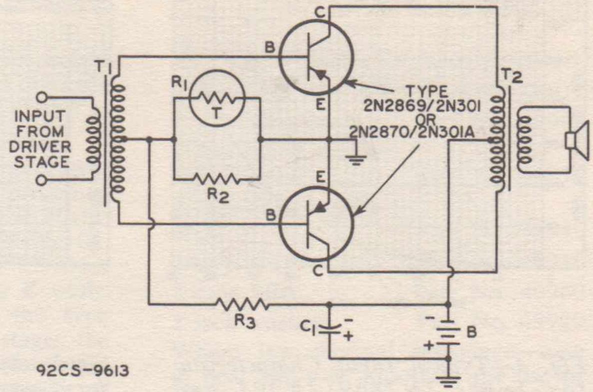


Fig. 2—Class B Push-Pull AF Power-Amplifier Circuit Utilizing Types 2N2869/2N301, 2N2870/2N301A, 2N301 and 2N301A.

B : 14.4-volt supply

C_1: 1000 μf, electrolytic, 25 volts

R_1: Thermistor, 28.25 ohms at O° C, 10 ohms at 25° C, 4.05 ohms at 50° C

R_2: 5.6 ohms, 0.5 watt

R_3: 270 ohms, 1 watt

T_1: Driver Transformer: Primary Impedance determined by large-signal considerations of the driver unit. Secondary Impedance = 40 ohms center tapped.

T_2: Output Transformer: Primary Impedance = 24 ohms center tapped, Secondary Impedance = voice-coil Impedance.

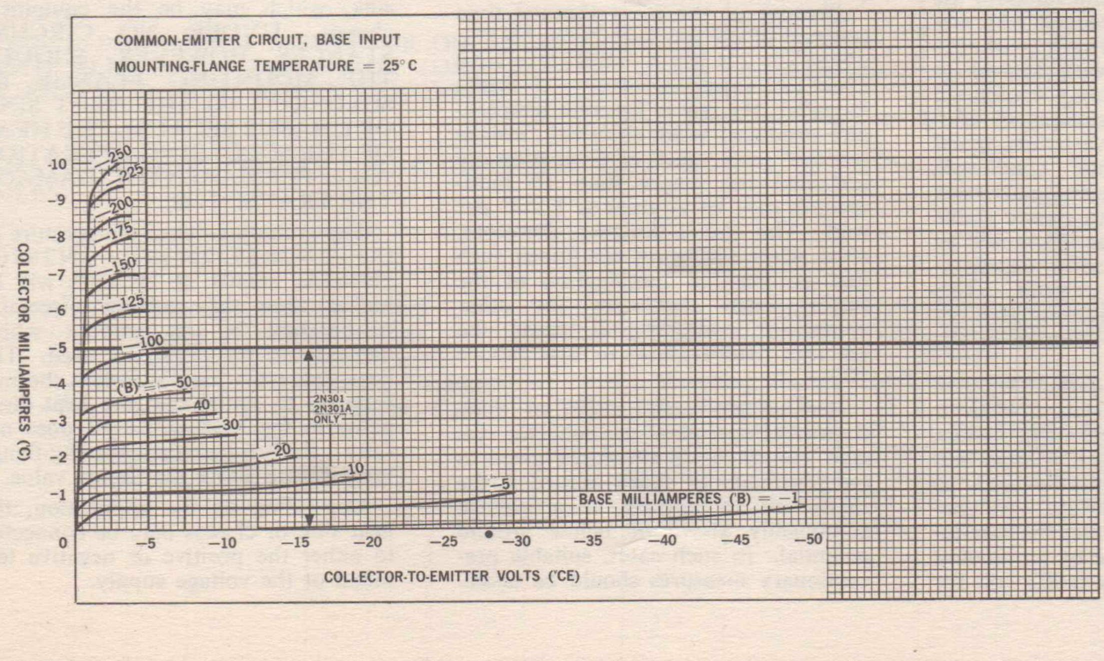


Fig. 3—Typical Collector Characteristics for Types 2N2869/2N301, 2N2870/2N301A, 2N301 and 2N301A.

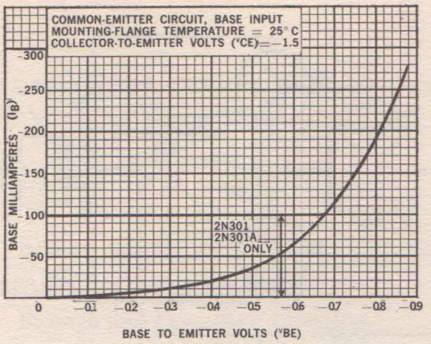


Fig. 4—Typical Input Characteristic for Types 2N2869/2N301 and 2N2870/2N310A.

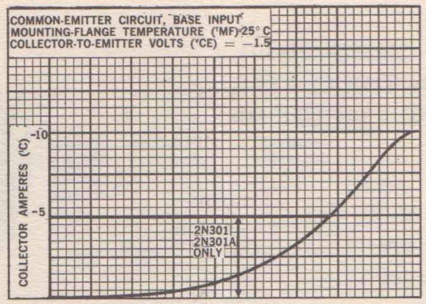


Fig. 5—Typical Transfer Characteristic for Types 2N2869/2N301, 2N2870/2N301A, 2N301 and 2N301A, Types 2N2869/2N301, 2N2870/2N301A.

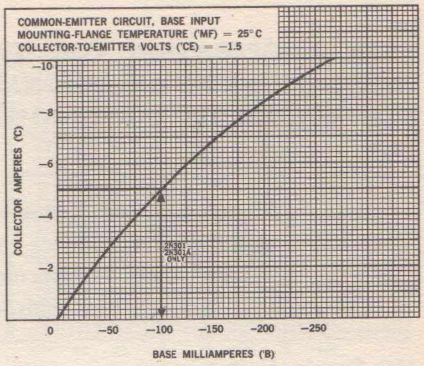


Fig. 6—Typical Current-Transfer Characteristic for types 2N2869/2N301 and 2N2870/2N301A.

OPERATING CONSIDERATIONS

The *maximum ratings* in the tabulated data are established in accordance with the following definition of the *Absolute-Maximum Rating System* for rating electron devices.

Absolute-Maximum ratings are limiting values of operating and environmental conditions applicable to any electron device of a specified type as defined by its published data, and should not be exceeded under the worst probable conditions.

The device manufacturer chooses these values to provide acceptable serviceability of the device, taking no responsibility for equipment variations, environment variations, and the effects of changes in operating conditions due to variations in device characteristics.

The equipment manufacturer should design so that initially and throughout life no absolute-maximum value for the intended service is exceeded with any device under the worst probable operating conditions with respect to supply-voltage variation, equipment component variation, equipment control adjustment, load variation, signal variation, environmental conditions, and variations in device characteristics.

In class A service, to ensure stable operation and low distortion, it will be necessary to provide some degeneration in the emitter circuit. This degeneration may be accomplished by using an unbypassed resistor in the emitter circuit.

In class B service, if the 2N2869/2N301, 2N2870/2N301A, 2N301 or 2N301A is operated near its maximum collector voltage rating, it is important that circuit arrangements be made to prevent thermal runaway. A convenient method is to reduce the base-to-emitter forward voltage by an amount equal to approximately 0.002 volt for each degree centigrade that the mounting-flange temperature is above 25° C.

In the design of circuits using the AWV-2N2869/2N301, 2N2870/2N301A, 2N301 and 2N301A it is extremely important to assure that the maximum junction-temperature rating of 100° C is not exceeded. This consideration is especially important in af-amplifier and other applications involving complex signal waveforms and other factors capable of producing high peak value of dissipation, such as

(a) oscillation or "ringing" due to excessive or improperly neutralized feedback
(b) phase shifts due to circuit capacitances and/or reactive loads
(c) variations in load impedance
(d) overdriving of transistors
(e) high line voltage

Because of the short thermal time constant of the AWV-2N2869/2N301, 2N2870/2N301A, 2N301 and 2N301A (approximately 15 milliseconds), the rise in junction temperature produced when so-called "instantaneous" peak values of dissipation are sustained for longer than 15 milliseconds may be as great as 1.5° C per watt. The circuit designer, therefore, must select operating conditions such that no possible combination of the factors listed above, or any other operating condition will cause the junction temperature to rise above 100° C.

Because the metal shells of these transistors operate at the collector voltage, consideration should be given to the possibility of shock hazard if the shells are to operate at a voltage appreciably above or below ground potential. In such cases, suitable precautionary measures should be taken.

The 2N2869/2N301, 2N2870/2N301A, 2N301 and 2N301A should not be connected into or disconnected from circuits with the power on because high transient currents may cause permanent damage to the transistors.

These transistors can be installed in commercially available sockets. Electrical connection to the base and emitter pins may also be made by soldering directly to these pins. Such connections may be soldered to the pins close to the pin seals provided care is taken to conduct excessive heat away from the seals. Otherwise the heat of the soldering operation will crack the pin seals and damage the transistor.

It is essential that the mounting flange which serves as the collector terminal be securely fastened to a heat sink, which may be the equipment chassis. UNDER NO CIRCUMSTANCES, HOWEVER, SHOULD THE MOUNTING FLANGE BE SOLDERED TO THE HEAT SINK OR CHASSIS BECAUSE THE HEAT OF THE SOLDERING OPERATION WILL PERMANENTLY DAMAGE THE TRANSISTOR.

The mounting-flange temperature of the 2N2869/2N301, 2N2870/2N301A, 2N301 or 2N301A will be higher than the ambient (free-air) temperature by an amount which depends on the heat sink used. The heat sink must have sufficient thermal capacity to assure that the heat dissipated in the heat sink itself does not raise the transistor-mounting-flange temperature above the design value.

Depending on the application, the heat sink or chassis may be connected to either the positive or negative terminal of the voltage supply.

In applications where the chassis is connected to the positive terminal of the voltage supply, it will be necessary to use an anodized aluminium washer having high thermal conductivity, or a 0.002" thick mica insulator between the mounting flange and the chassis. If an aluminium washer is used, it should be drilled or punched to provide the two mounting holes and the clearance holes for the emitter and base pins. The burrs should then be removed from the washer and the washer finally anodized. To ensure that the anodized insulating layer is not destroyed during mounting, it will also be necessary to remove the burrs from the holes in the chassis. Furthermore, to prevent a short circuit between the mounting bolts and the chassis, it is important that a fibre washer be used between each bolt and the chassis as shown in Fig. 7.

An insulated mounting arrangement such as that described in the preceding paragraph and shown in Fig. 7 is also necessary when Type 2N2869/2N301, 2N2870/2N301A, 2N301 or 2N301A transistors are used in class B push-pull af-amplifier stages of the type shown in Fig. 2. In such stages the mounting flanges of the two transistors must be insulated from the chassis or heat sink and from each other to avoid short-circuiting the primary winding of the output transformer.

The following mounting hardware items, are available for these transistors.

1 mica washer	AWV Part No. 49910
2 insulating bushes	Part No. 49900
2 flat metal washers 6BA x 22G	Part No. 49930
2 lugs 5/32 tinned brass	Part No. 49940
2 screws ½" x 6BA	Part No. 49950
2 nuts 6BA	Part No. 49960
2 lock washers	Part No. 49920

Where low thermal resistance—case to heat sink—is important the use of silicone grease (Dow Corning Compound type 340) between the transistor mounting flange and the mica washer and between the mica washer and the heat sink, is advised.

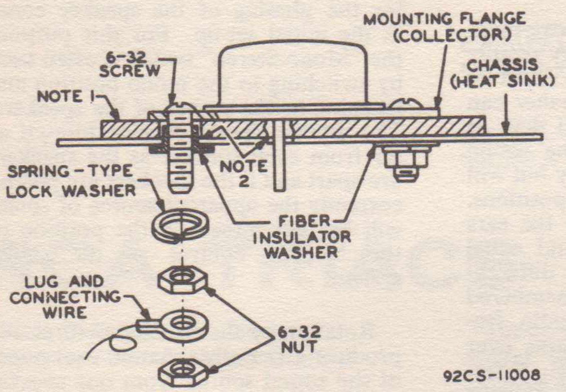
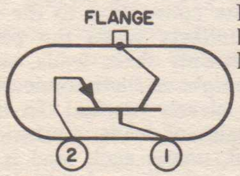


Fig. 7—Suggested Mounting Arrangement for Types 2N2869/2N301, 2N2870/2N301A, 2N301 and 2N301A.

NOTE 1: 0.002" MICA INSULATOR OR ANODISED ALUMINIUM INSULATOR (DRILLED OR PUNCHED WITH BURRS REMOVED).

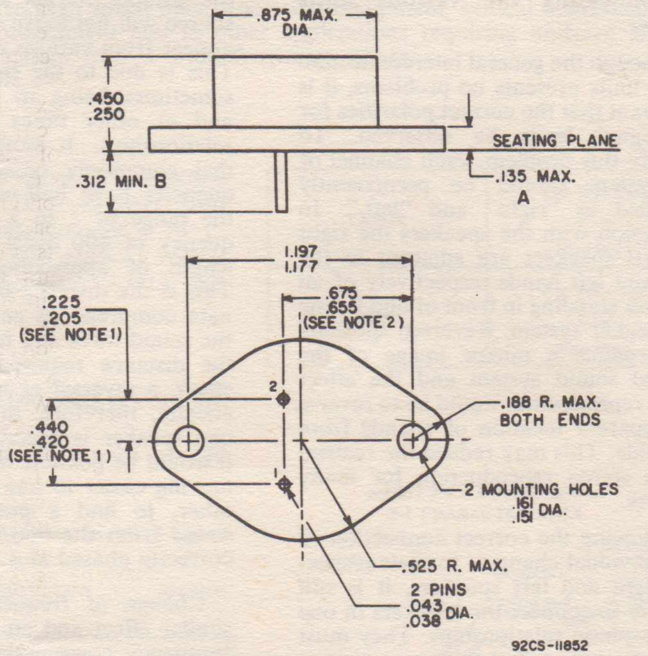
NOTE 2: REMOVE BURRS FROM CHASSIS HOLES.

TERMINAL CONNECTIONS



Pin 1—Base
Pin 2—Emitter
Mounting Flange—Collector. Case

DIMENSIONAL OUTLINE For
Types 2N2869/2N301 and 2N2870/2N301A
JEDEC No. TO-3



ALL DIMENSIONS IN INCHES.
For AWV-2N2869/2N301, 2N2870/2N301A, 2N301 and 2N301A.
Mounting Flange Thickness (A) 0.032 ± 0.003"
Pin Length (B) 0.433 ± 0.012
NOTE 1: THESE DIMENSIONS SHOULD BE MEASURED AT POINTS .050" (1.270MM) to .055" (1.397MM) BELOW SEATING PLANE. WHEN GAGE IS NOT USED, MEASUREMENT WILL BE MADE AT SEATING PLANE.
NOTE 2: TWO LEADS.

PHASING IN STEREOPHONIC SOUND SYSTEMS

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Stereophonic sound reproduction has now been available for about ten years and a good proportion of audio systems sold; whether they be small portable gramophones, radiograms or hi-fi systems, all have facilities for playing stereo records and sometimes stereo tapes. When a stereo system is purchased as a complete unit, the purchaser has no problem in its use, as it is already correctly arranged. But in the case where the system is purchased as individual units, or has been built by the owner, then the user must interest himself in the correct manner of connecting the various units together.

Although the general interconnection of the units presents no problems, it is important that the correct polarities for the signal leads be observed. To simplify this problem, each channel of the system should be permanently identified as "right" and "left." In connection with the speakers the right and left speakers are adjacent to the right and left hands respectively of an observer standing in front of and facing the speaker system. Reversed speakers will produce a mirror image of the original sound system and the effect on the reproduction would be to reverse the apparent location of sounds from each side. This may reduce the realism of the stereo reproduction for many listeners.

Following the correct connection of the individual channels to their respective right and left speakers, it is still possible to connect the speakers in one of two phase relationships. They must be connected so that their cones move in phase, i.e., each speaker in a combination, whether stereo or mono, must move the air column in the same direction at any instant when fed with identical signals.

When the signals fed to each speaker are identical and the speakers have the same acoustic efficiency, a listener in a position equidistant from each speaker will have each ear excited by a signal identical in amplitude and phase or time. The ear primarily senses time differences to identify the source of the sound and, since the above signals would be the same as a point source of the same sound, it is heard as a point source located in the area about midway between the speakers.

Demonstration of the above-mentioned effect can be made very readily with single tone sources. The effect of the phase reversal in one speaker can be noted and it will be found that, as the listener moves about, the sound source will not move smoothly but will appear from various discrete positions. This is due to the signals at the ears sometimes being in phase and equal and at other times having different relationships. It must be remembered that commonly occurring audio frequencies have half-wave lengths over the range of 1 to 2 feet, e.g., a frequency of 400 hertz has a half-wave length of approximately 16 inches. This is the distance between the alternate compressions and rarefactions in the sound wave and for a single tone is the distance required to produce, in effect, a reversal of its phase. Theoretically, therefore, in the case where one speaker is incorrectly connected, it would be possible for the listener, by moving closer to one speaker than the other, to find a position where the sound from the two speakers is again correctly phased at a single frequency.

Change of frequency will have a similar effect and on a programme all frequency components could be confused if speaker phasing is incorrect. Even though one could surmise that with programme material the ear could have difficulties differentiating between a correctly and incorrectly phased speaker system, it is found that normal experience enables the brain to form a pattern for the sound obtained from natural sources. This would not be the case if one speaker was out of phase, since the sound sources would not resemble any natural or familiar type of sound system.

Correct phasing of the speaker cones, therefore, is essential to allow the reproduction of sound as it was received by the recording microphones. This is the intention and is the reason for the phasing of the speaker cones in the initial set-up. For this purpose the "Mono-Stereo" switch is often used by switching to the mono position and listening to the output of the speakers. The listener should position himself as far from each speaker as the speakers are apart and if the speakers are phased correctly the apparent source of sound will be determined by the position of the balance control of the stereo system.

Rotation of the balance control will produce a smooth apparent movement of the sound source from one speaker to the other. A two to one ratio in sound level from the speakers is sufficient to shift the apparent sound source to the higher level speaker. Any difficulties in identifying the sound source will probably indicate incorrect phasing in one speaker.

If identical amplifiers are used a phase reversal can occur either in the speakers or the pickup connections. With the "Mono-Stereo" switch in the "Mono" position, possible incorrect connections, to the pickup can be detected and corrected by reversing the connections to one of the speakers.

Incorrect phasing can occur in making the connection to the pickup. This is unlikely when the pickup is fitted with only three leads, since the common, right and left connections are generally clearly marked. In the case of two pairs of leads, confusion may result and, unless each lead is clearly indicated, reference should be made to the manufacture of the pickup.

NEWS & New Releases

RCA INTEGRATED CIRCUITS

The following linear integrated circuits have recently been announced by RCA.

CA3020	Multistage, Multipurpose AF power amplifier on a single monolithic silicon chip. This circuit has an AF power output capability of 550 mW from a 9V supply.
CA3021 CA3022 CA3023	These are a series of low-power wideband amplifiers suitable for use as gain controlled Linear Amplifiers, AM, FM, IF Amplifiers, Video Amplifiers and Limiters.
CA3028	This circuit features a balanced differential amplifier with a controlled constant current source, it is suitable for operation from DC to 120MHz and has many applications including converter, limiter, mixer and oscillator use.
CA3029 CA3030	These two circuits are operational amplifiers featuring the dual-in-line package which enables easy mounting on printed circuit boards and enables higher packing densities.
CA3031/702A CA3032/702C	Both these circuits are high gain DC amplifiers suitable for use in high speed data processing equipment, critical instrumentation equipment and in other applications requiring uniform amplification over a wide frequency range.

RCA TRIACS

Gated Bidirectional Silicon Thyristors For AC Load Control

RCA Developmental Types TA2918 and TA2919 are gate-controlled full-wave ac silicon switches designed to switch from a blocking state to a conducting state for either polarity of applied voltage with positive or negative gate triggering.

They are intended primarily for the phase control of ac loads in applications such as light dimming, universal and induction motor control, and heater control.

The TA2918 and TA2919 are hermetically sealed in all-welded tin-plated modified TO-5 packages. The small size of this package makes these devices especially suitable for use in equipment where space restrictions are of prime importance. In addition, because they are tin-plated, they can be soldered directly to a heat sink, thereby allowing the use of mass-produced pre-punched parts, and batch soldering techniques to eliminate many of the difficulties associated with mechanical mounting and heat sinking.

These thyristors have a conduction current capability of 6 amperes (rms value) at a case temperature of + 75° C. The TA2918 has a peak blocking voltage rating of 200 volts; the TA2919, 400 volts.

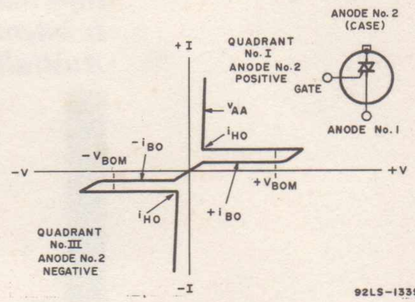
Features

- TA2918 controls 600 watts at 120 volts, 60 Hz
- TA2919 controls 1200 watts at 240 volts, 60 Hz
- Short-emitter design
- Uniform gate sensitivty in all 4 operating modes
- All-diffused construction — assures

Modified TO-5 Package

TA2918	For 120-Volt Line Operation
TA2919	For 240-Volt Line Operation

ANODE-No.2 TO ANODE-No.1 E-I CHARACTERISTICS



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