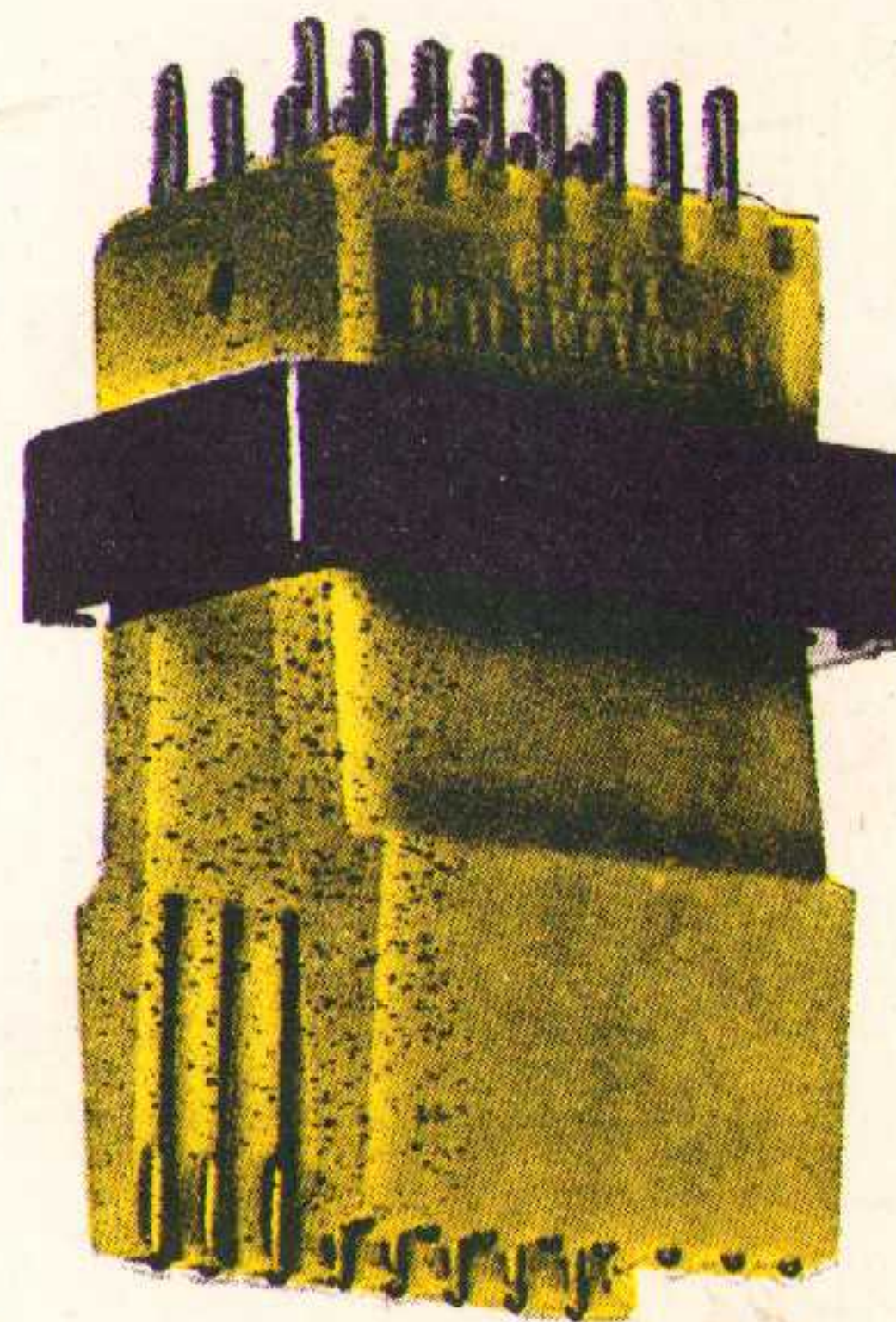
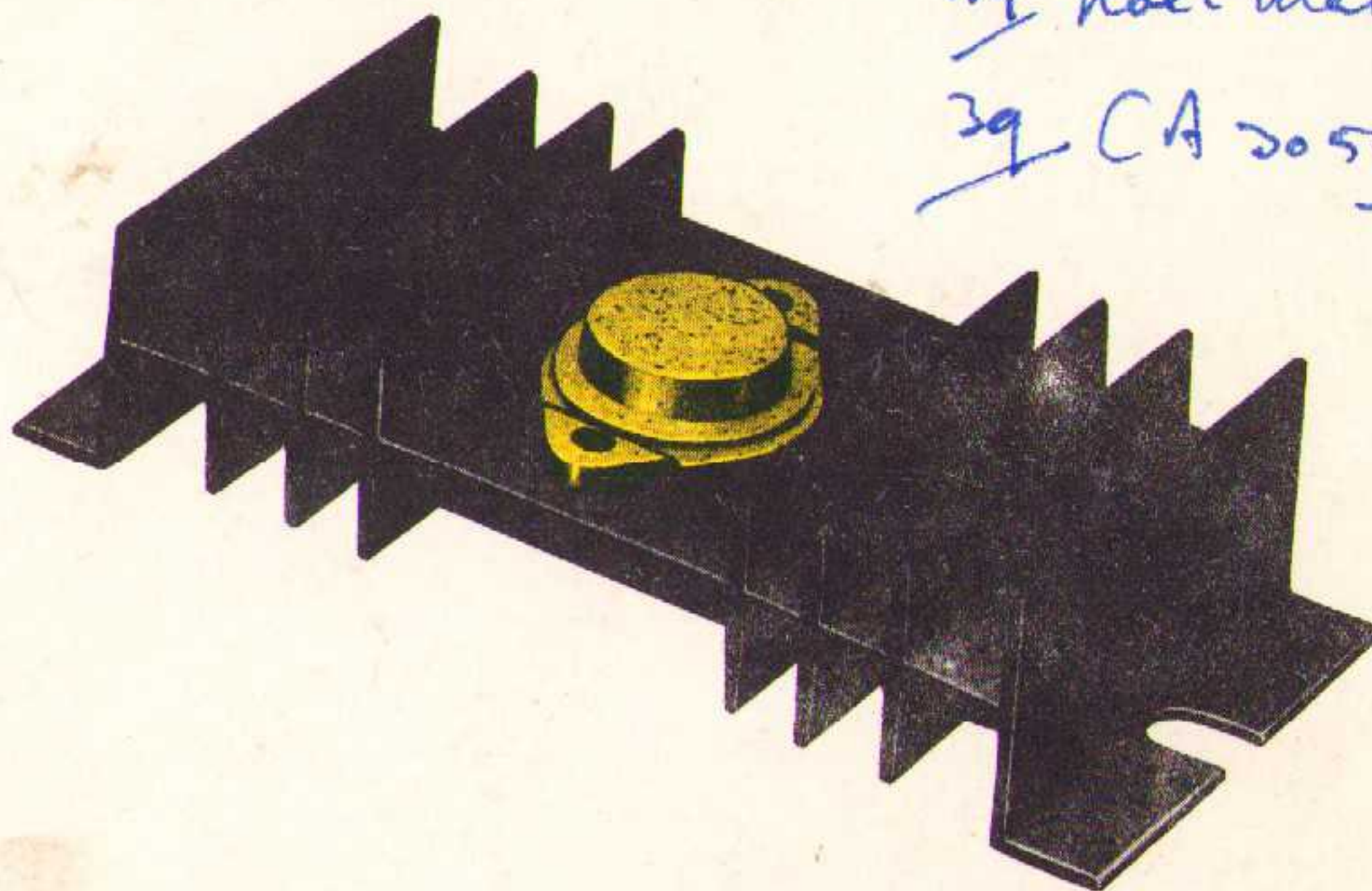
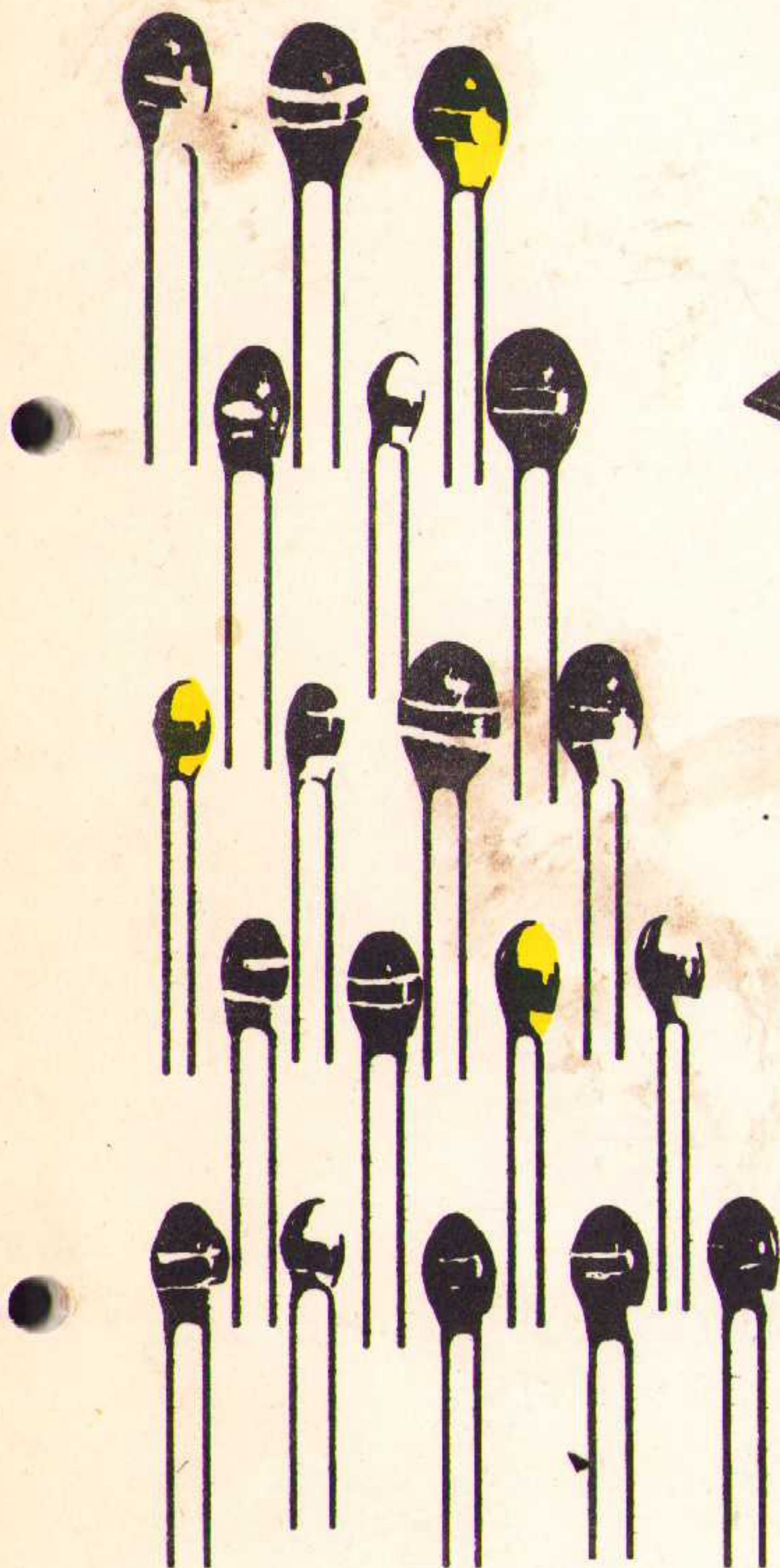


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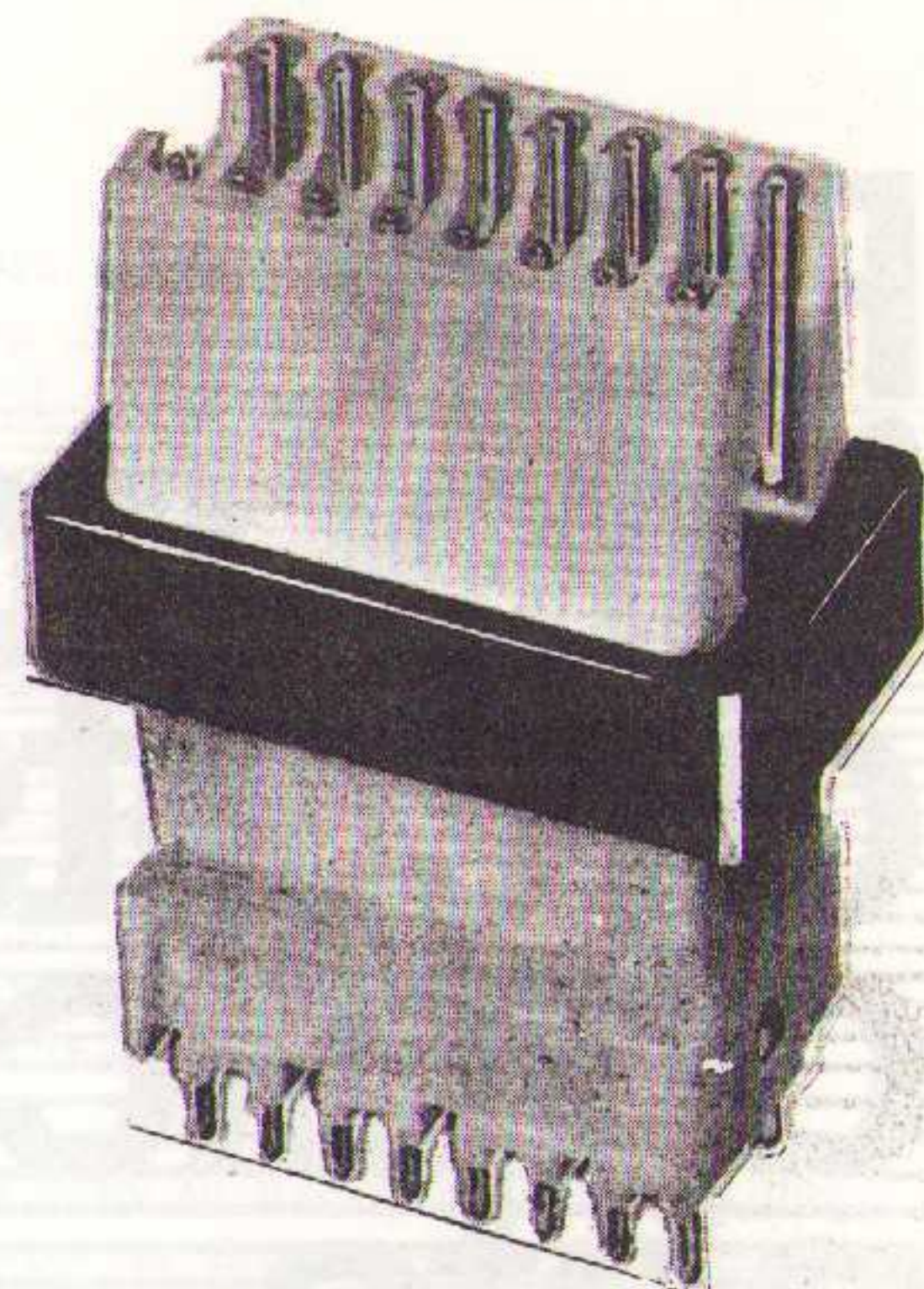
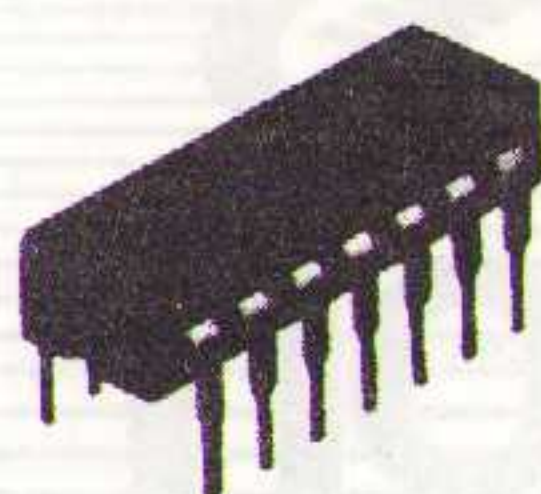
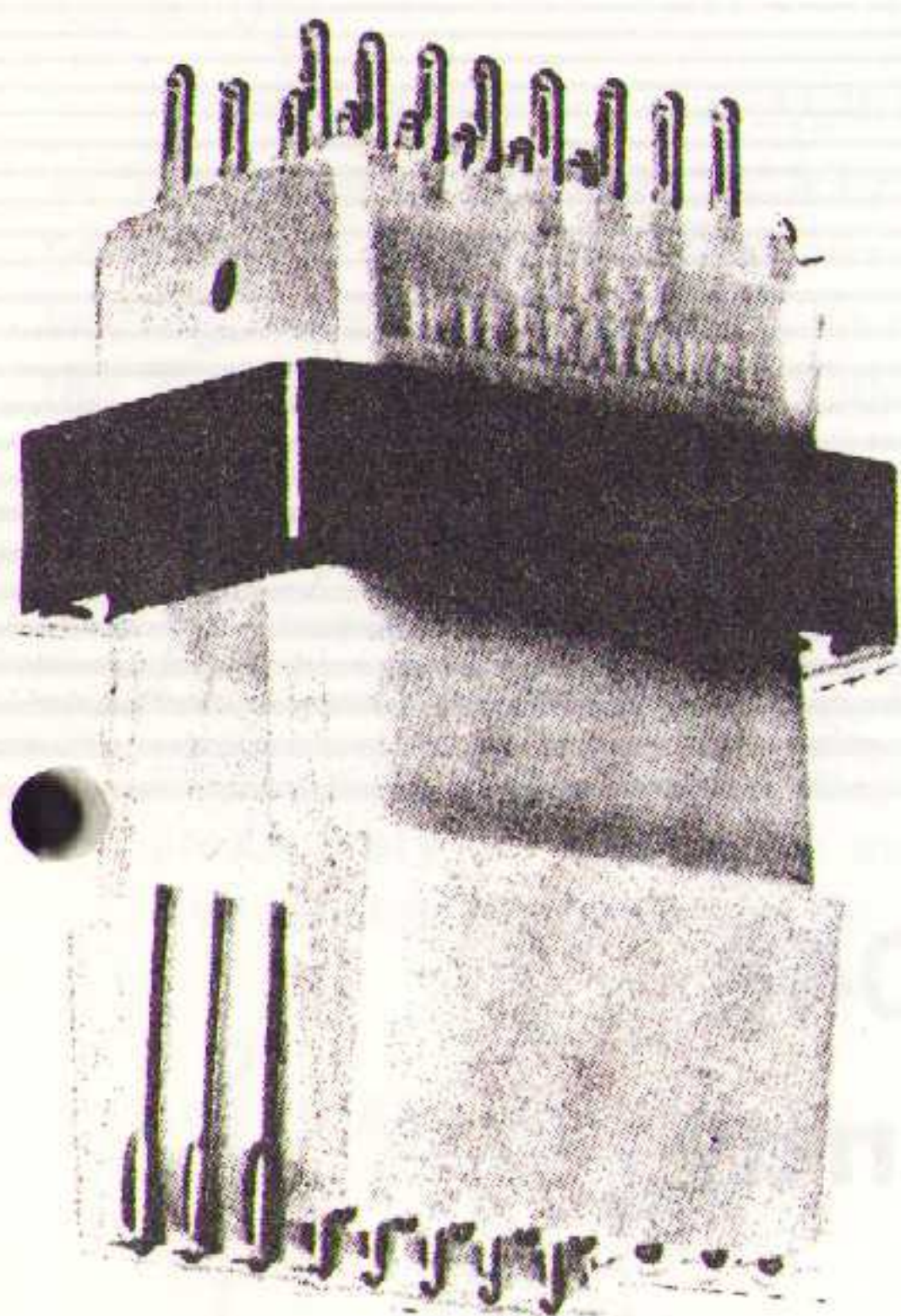
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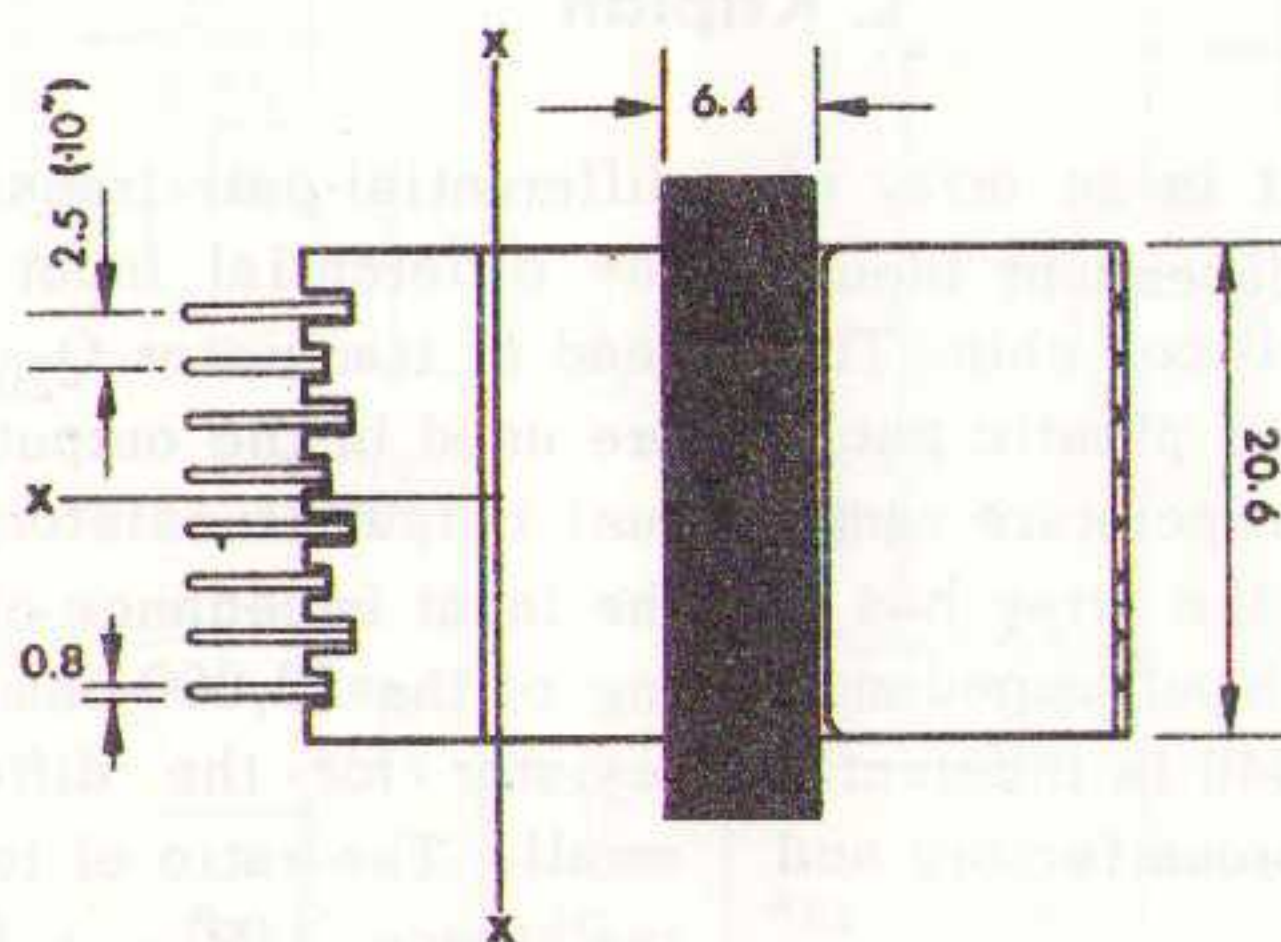
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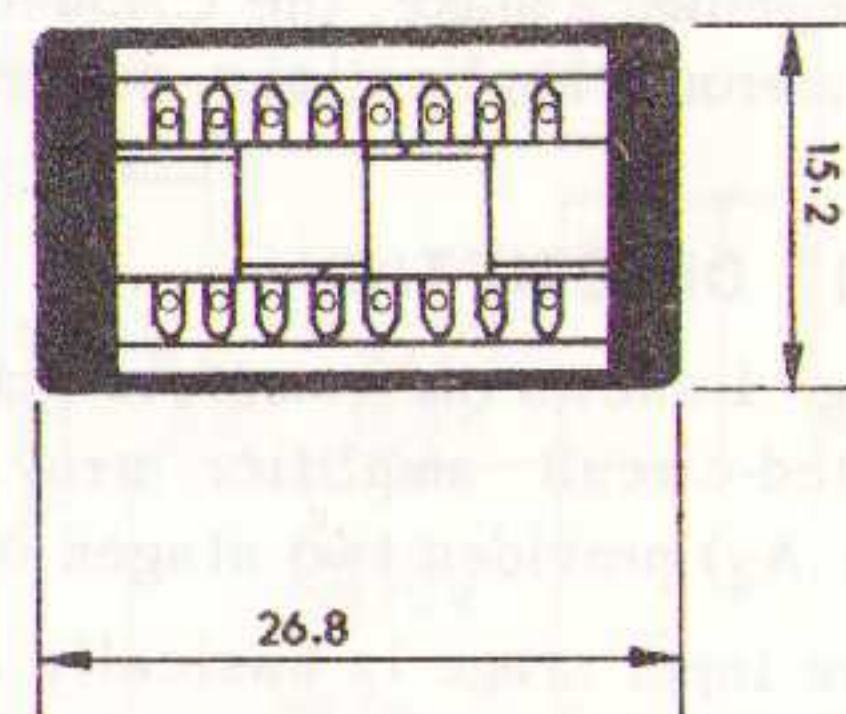
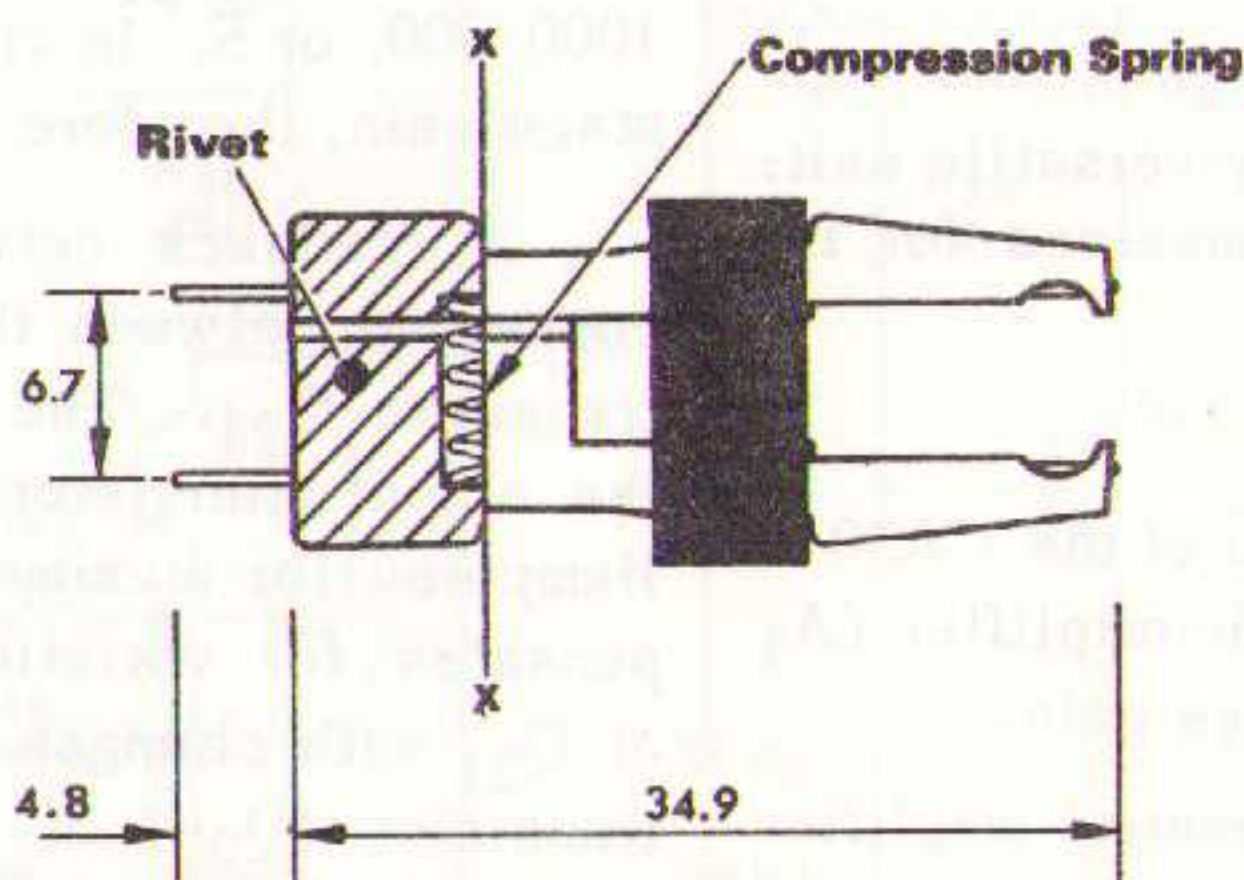
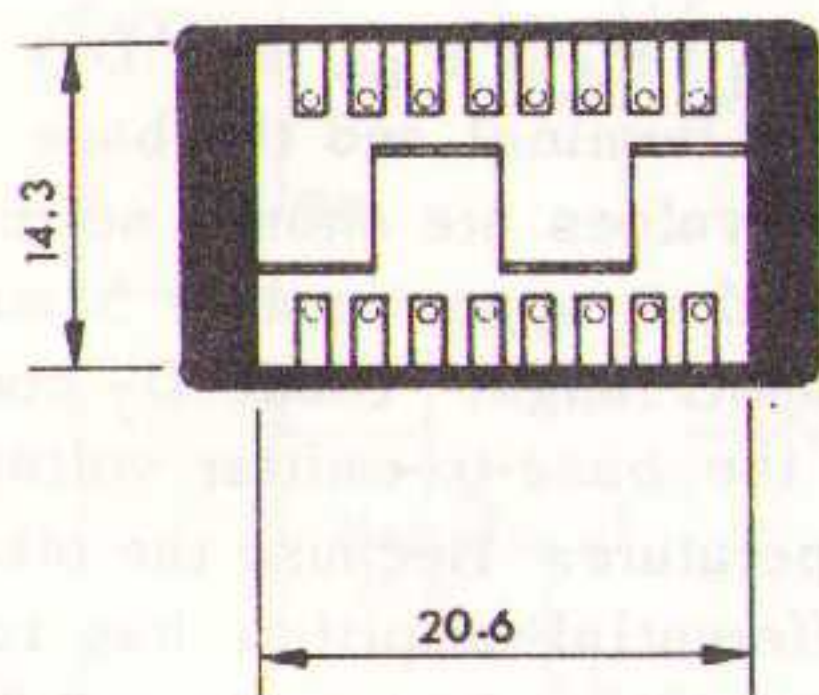
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Pins: Phosphor bronze with 1µm hard gold on nickel plate finish.

Body: moulded in yellow acetal resin

Ring: moulded in black nylon



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RCA

ICAN-4072

integrated circuits application note



Applications of the RCA CA3048 Integrated-Circuit Amplifier Array

by
L. Kaplan

The RCA CA3048 integrated circuit is an array of four identical amplifiers, each with independent inputs and outputs, all on a single monolithic silicon chip. The circuit is housed in a 16-lead dual-in-line plastic package. It has an operating and storage temperature range of -25°C to $+85^{\circ}\text{C}$. Each amplifier in the array has a typical open-loop gain of 58 dB and input impedance of 90,000 ohms. The noise in the CA3048 is inherently very low and is tightly controlled in rigorous factory and quality-control testing.

The combination of low noise, high gain, and high input impedance make the CA3048 a very versatile unit, and numerous applications suggest themselves for its use.

CIRCUIT DESCRIPTION

Fig. 1 shows the complete schematic of the CA3048 integrated-circuit amplifier array. Each amplifier (A_1 through A_4) provides two stages of voltage gain.

The input stage is basically a differential amplifier with a Darlington transistor added on the one side. The output stage consists of a combination of three transistors and associated resistors connected in an inverting configuration. For example, in amplifier A_3 , Q_{19} is the Darlington input transistor, and Q_{20} and Q_{21} are the

differential-pair transistors. The load resistor R_{29} for the differential input stage is located in the collector lead of transistor Q_{20} . Transistors Q_{13} , Q_{14} , and Q_{17} are used in the output stage. Transistor Q_{17} is the actual output transistor; transistors Q_{13} and Q_{14} raise the input impedance of the output stage so that the loading of the 30,000-ohm source resistance R_{29} (i.e., load resistor for the differential-amplifier input stage) is small. The ratio of total collector resistance to emitter resistance $[(R_{31} + R_{38})/R_{50}]$ in the output stage is 1000/200, or 5. In view of the small source loading, the stage gain, therefore, is essentially equal to 5.

A feedback network (R_{41} , R_{42} , R_{46} , and D_7) is connected between the output terminal and the base of transistor Q_{21} . The resistor values are chosen so that the output transistor is biased at approximately 5 milliamperes for maximum dynamic range. Diode D_7 compensates for variations in the base-to-emitter voltage of Q_{21} with changes in temperature. Because the other transistor (Q_{20}) of the differential amplifier has two emitter-base junctions in series, two diodes, D_3 and D_4 , are required for temperature compensation. Diodes D_3 and D_4 also provide temperature compensation for the differential-pair transistor Q_1 in amplifier A_2 (similarly diodes D_5 and D_6 are shared by amplifiers A_1 and A_4).

RCA Electronic
Components

Diodes D_3 and D_4 and diodes D_5 and D_6 are connected to their respective inputs through a relatively stiff voltage divider (for amplifier A_3 , the divider consists of R_{27} and R_{28}). The input to amplifier A_3 is normally applied to the base of the Darlington transistor Q_{19} . The 100-kilohm resistor R_{37} supplies bias current to this transistor. The voltage drop across resistor R_{37} is small because of the very small base current of transistor Q_{19} .

Each amplifier of the CA3048 may be viewed as an ac operational amplifier in which a fixed resistance is permanently connected between the output and the inverting input. The built-in feedback resistor delimits the characteristics of the CA3048 amplifiers in the following ways:

1. The impedance as viewed from the noninverting input terminal consists mainly of the 100 kilohm input-bias resistance (R_{13} , R_{15} , R_{37} , or R_{39}). This resistance is shunted by the input capacitance of approximately 10 picofarads and the additional resistive loading presented by the input impedance of the

Darlington input pairs. When the amplifier is operated under open-loop conditions (inverting input at ac ground), the total input impedance consists of 90 kilohms in shunt with the input capacitance. When the built-in feedback loop is allowed to function (by insertion of an unbypassed resistance in the noninverting input lead), then the loading caused by the Darlington input pairs is reduced, and the input resistance rises asymptotically towards 100 kilohms.

2. The impedance as viewed from the inverting input terminal is small (in the order of 40 to 50 ohms.)
3. When the CA3048 is used in its normal mode of operation, each amplifier in the array may be represented by the equivalent circuit shown in Fig. 2. (The capacitances shown are the sum of the device capacitances, socket capacitances, and stray capacitances.) The transconductance G_m , which is equal to the product of the voltage gain and the output conductance (10^{-3} mho), is typically 0.8 mho at midband.

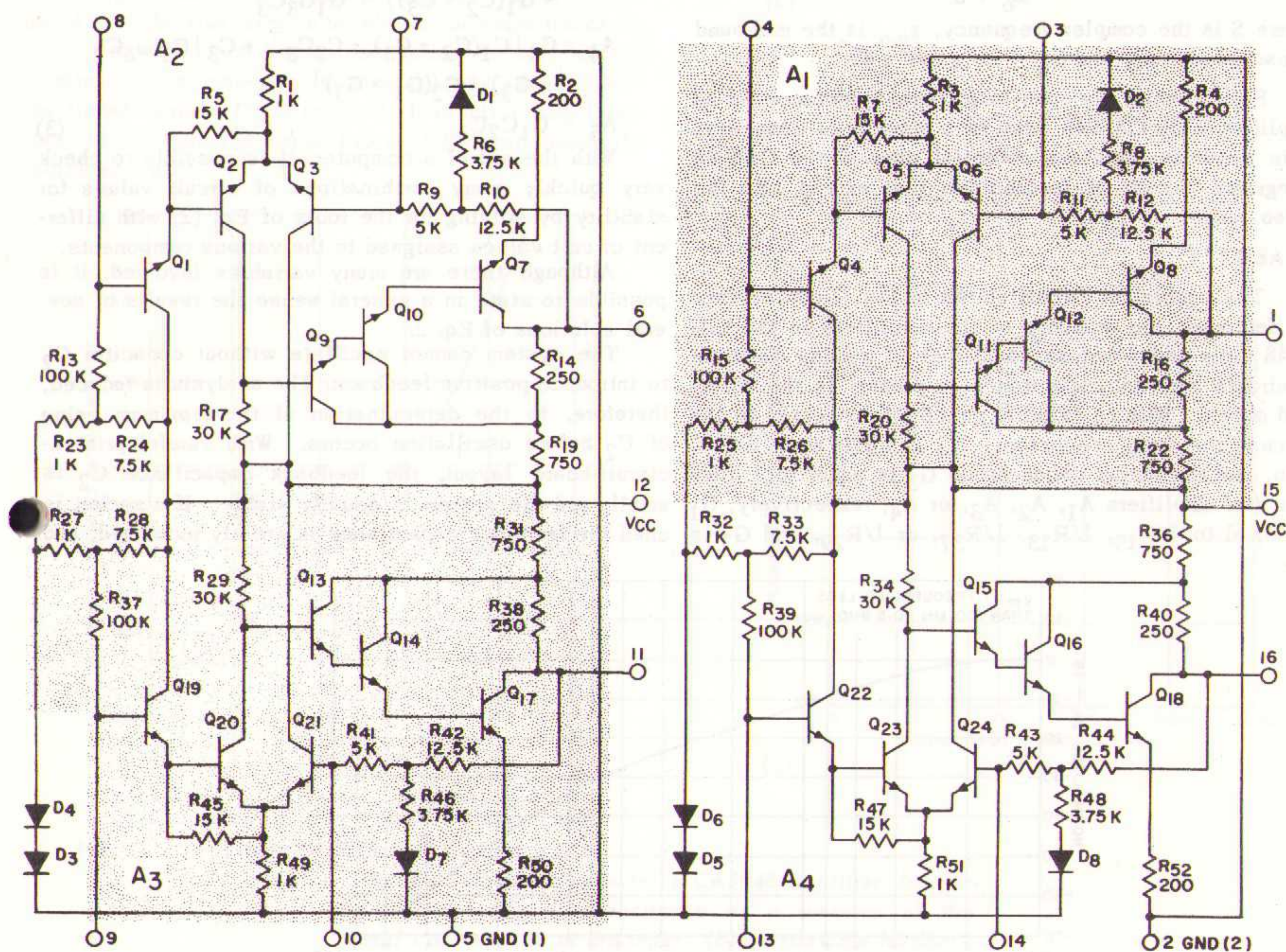


Fig. 1 - Schematic of the CA3048 integrated-circuit amplifier array.

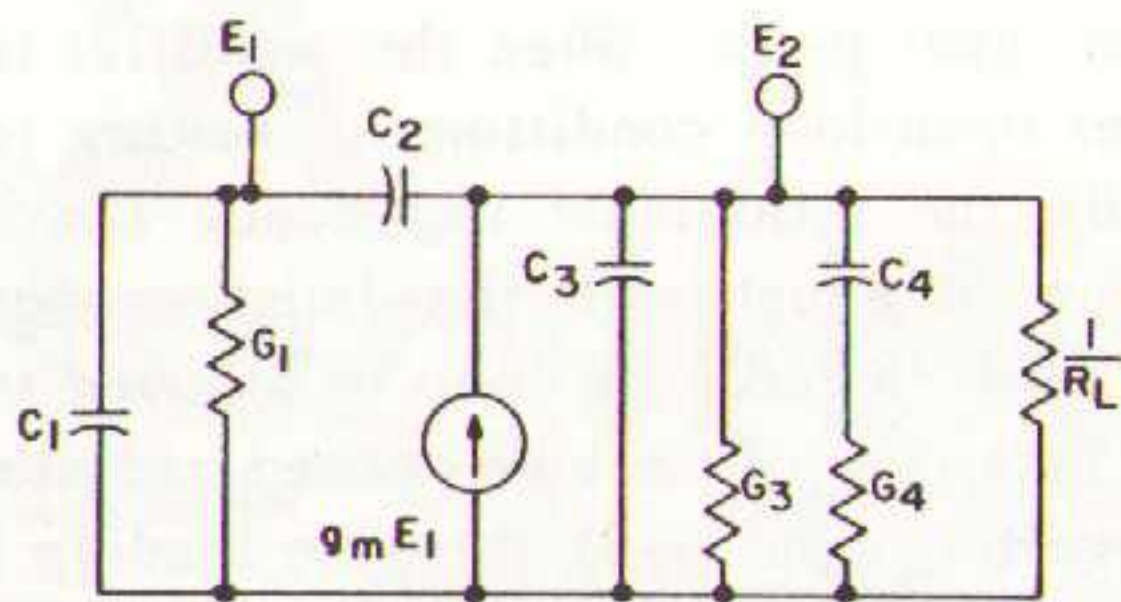


Fig. 2 - Equivalent circuit of a CA3048 amplifier.

GAIN-FREQUENCY RESPONSE

Curves of the transconductance of any amplifier in the CA3048 array as a function of frequency up to 30 MHz show two break points. At frequencies above the first break point, which occurs at 300 kHz, the transconductance rolls off at a rate of 6 dB per octave to 12 MHz. At frequencies above 12 MHz, the rate of roll-off increases to 12 dB per octave. At frequencies up to 12 MHz, therefore, the transconductance of any amplifier in the CA3048 array is expressed by the following equation:

$$g_m = \frac{\omega_o g_{mo}}{\omega_o + S} \quad (1)$$

where S is the complex frequency, g_{mo} is the mid-band transconductance, and $\omega_o = 2\pi \times 300 \times 10^3$.

Fig. 3 shows the open-loop transconductance for an amplifier in the CA3048 array as a function of frequency. This response indicates potential uses of the CA3048 integrated circuit at frequencies that extend into the video range.

STABILITY

The equivalent circuit shown in Fig. 2 can be used to determine the stability of the amplifiers in the CA3048 array under various conditions of loading when undesirable external capacitance is present in the wiring and socket. With no external generator connected to the circuit, the input conductance G_1 is equal to $1/100000$ mho, and the output conductance G_3 is equal to $1/1000$ mho (for amplifiers A_1 , A_2 , A_3 , or A_4 , respectively, G_1 is equal to $1/R_{15}$, $1/R_{13}$, $1/R_{37}$, or $1/R_{29}$, and G_3 is

equal to $1/(R_{16} + R_{22})$, $1/(R_{14} + R_{19})$, $1/(R_{31} + R_{38})$, or $1/(R_{36} + R_{40})$. The capacitance C_4 and the conductance G_4 shown in Fig. 2 represent an external damping network which can be varied or deleted as demanded by stability or gain-bandwidth requirements.

A necessary and sufficient condition for a system to be stable is that the roots of the characteristic equation of the system have no positive real parts. The characteristic equation for the circuit of Fig. 2 is obtained by expansion of the circuit determinant and collection of the coefficients of the complex frequency S . The equation assumes the following form:

$$A_1 + A_2S + A_3S^2 + A_4S^3 + A_5S^4 = 0 \quad (2)$$

After much tedious algebra, the coefficients are determined as follows:

$$A_1 = \omega_o G_1 G_3 G_4$$

$$A_2 = \omega_o G_4 [G_3(C_1 + C_2) + G_1(C_2 + C_3) - g_{mo} G_2] + G_1 G_3 (C_3 \omega_o + G_4)$$

$$A_3 = \omega_o \{ C_2 C_3 (G_1 + G_3 + G_4 - g_{mo}) + C_1 [C_3 (G_3 + G_4) + C_2 G_4] \} + G_4 [G_3 (C_1 + C_2) + G_1 (C_2 + C_3)] + G_1 G_3 C_3$$

$$A_4 = G_4 [C_1 (C_2 + C_3) + C_2 C_3] + C_3 [C_1 (\omega_o C_2 + G_3) + C_2 (G_1 + G_3)]$$

$$A_5 = C_1 C_2 C_3 \quad (3)$$

With the aid of a computer, it is possible to check very quickly many combinations of circuit values for stability by solving for the roots of Eq. (2) with different circuit values assigned to the various components.

Although there are many variables involved, it is possible to state in a general sense the results of several solutions of Eq. 2.

The system cannot oscillate without capacitor C_2 to introduce positive feedback. The analysis is reduced, therefore, to the determination of the maximum value of C_2 before oscillation occurs. With careful printed-circuit-board layout, the feedback capacitance C_2 is small, and the system is usually stable. If a socket is used the feedback capacitance is greatly increased, and

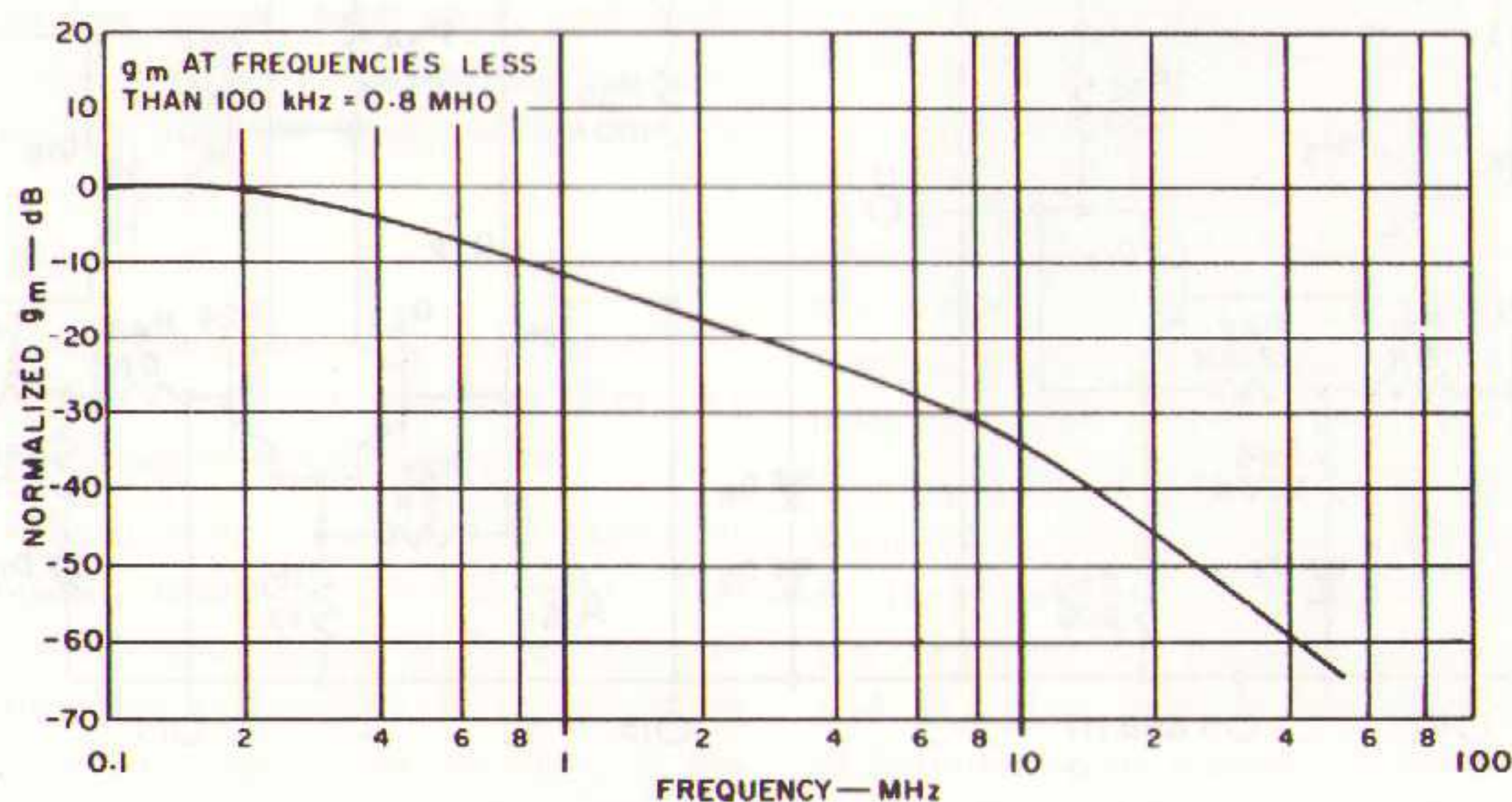


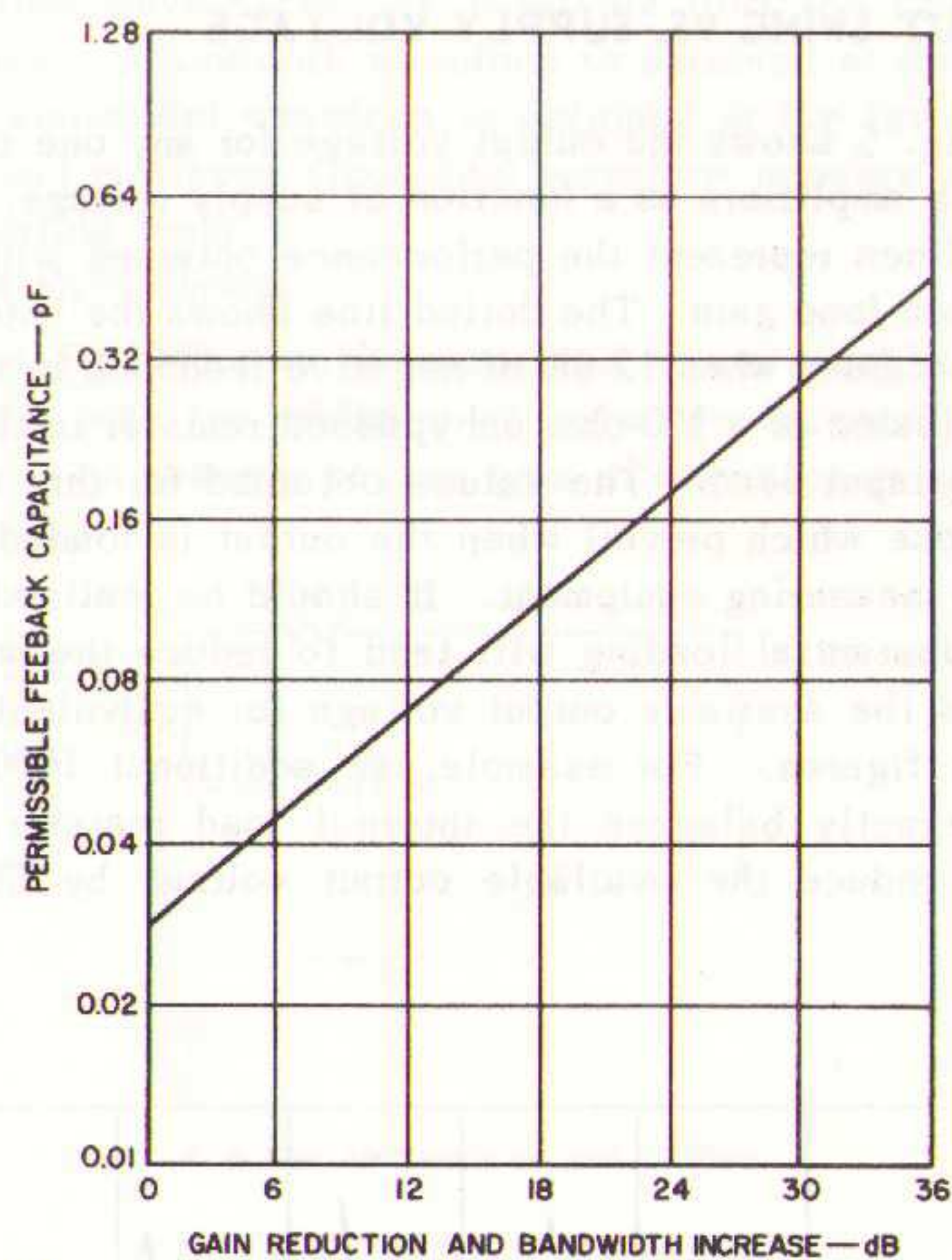
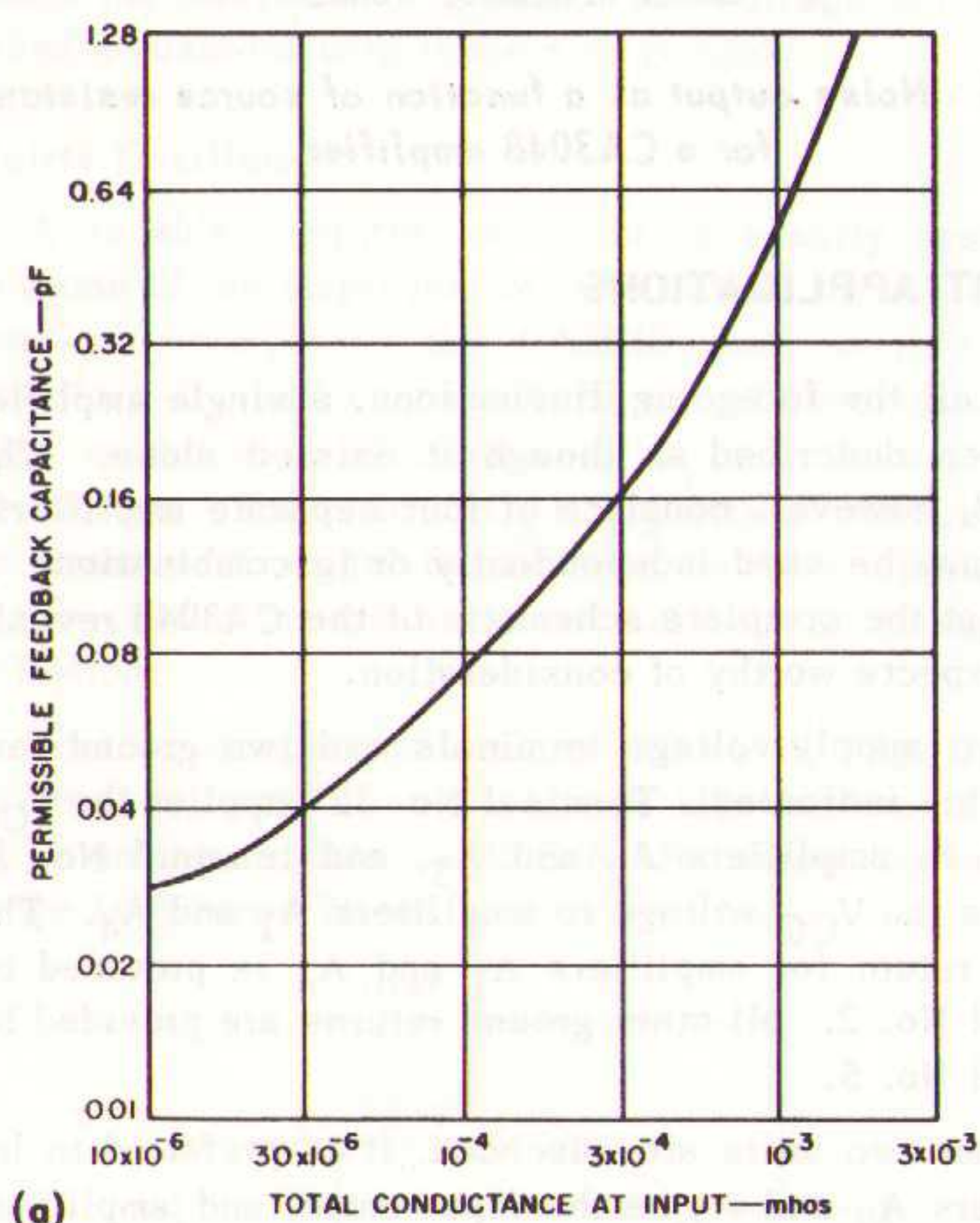
Fig. 3 - Typical gain-frequency response for a CA3048 amplifier.

stabilization of the circuit is generally required.

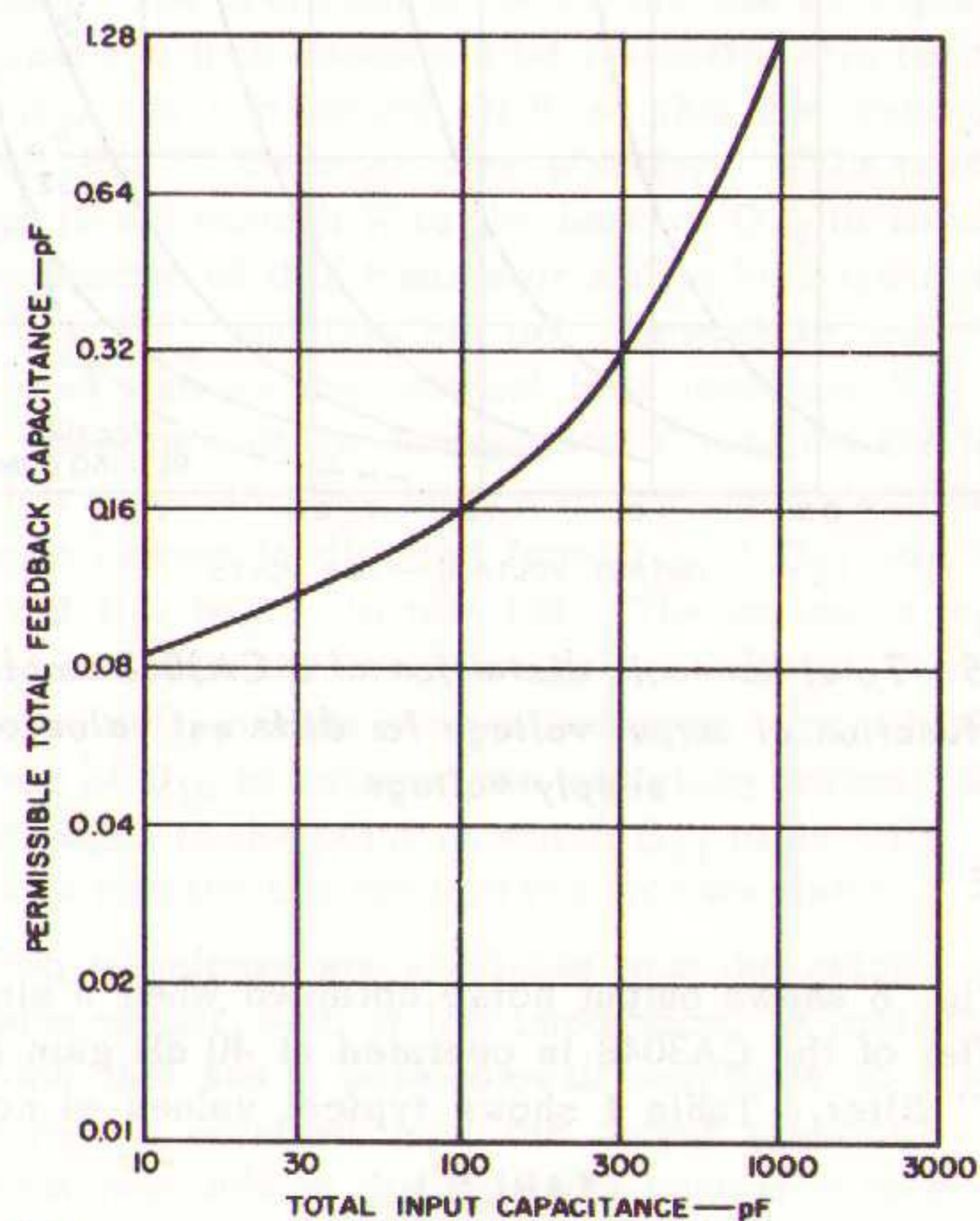
As with any two-port network, any increase in source or load conductance aids stability. In addition, analyses of Eq. (2) show that addition of shunt capacitance at the input is a very effective stability technique when the source impedance is high. Introduction of negative feedback into the circuit [which is simulated in Eq. (2) by a decrease in the value assigned to the transconductance g_m and an increase in the cutoff frequency] also improves circuit stability.

Another stability method, which is effective for any source impedance or gain value, is the addition of a damping network such as that formed by capacitance C_4 and conductance G_4 in Fig. 2. In this method, the value of C_4 is chosen so that its reactance is equal to the parallel combination of R_3 and R_L at the highest frequency of desired amplification. The value of R_4 is made small so that the gain is reduced at high frequencies and is typically 1/10 or 1/20 the value of the parallel combination of R_3 and R_L .

The series of curves in Fig. 4 show the results of the computation for the roots of Eq. (2). It should be noted that the maximum value shown for capacitance C_2 is that obtained just before oscillation occurs. Severe peaking of the response (or ringing) may result before the listed value of C_2 is reached. It is advisable, therefore, to maintain the capacitance of C_2 well below the indicated value.



(b)



(c)

Fig. 4 - Stability curves for a CA3048 amplifier: (a) permissible feedback capacitance as a function of the total conductance at the input; (b) permissible feedback capacitance as a function of gain reduction and of bandwidth increase; (c) permissible feedback capacitance as a function of the total input capacitance.

OUTPUT SWING VS. SUPPLY VOLTAGE

Fig. 5 shows the output voltage for any one of the CA3048 amplifiers as a function of supply voltage. The solid lines represent the performance obtained with the full open-loop gain. The dotted line shows the improvement obtained when 12 dB of negative feedback is added by inclusion of a 150-ohm unbypassed resistor in the inverting-input lead. The values obtained for this curve are those which prevail when the output is loaded only by the measuring equipment. It should be realized that any substantial loading will tend to reduce the magnitude of the available output voltage for equivalent distortion figures. For example, an additional 1000-ohm load exactly balances the internal load resistor, and would reduce the available output voltage by 50 per cent.

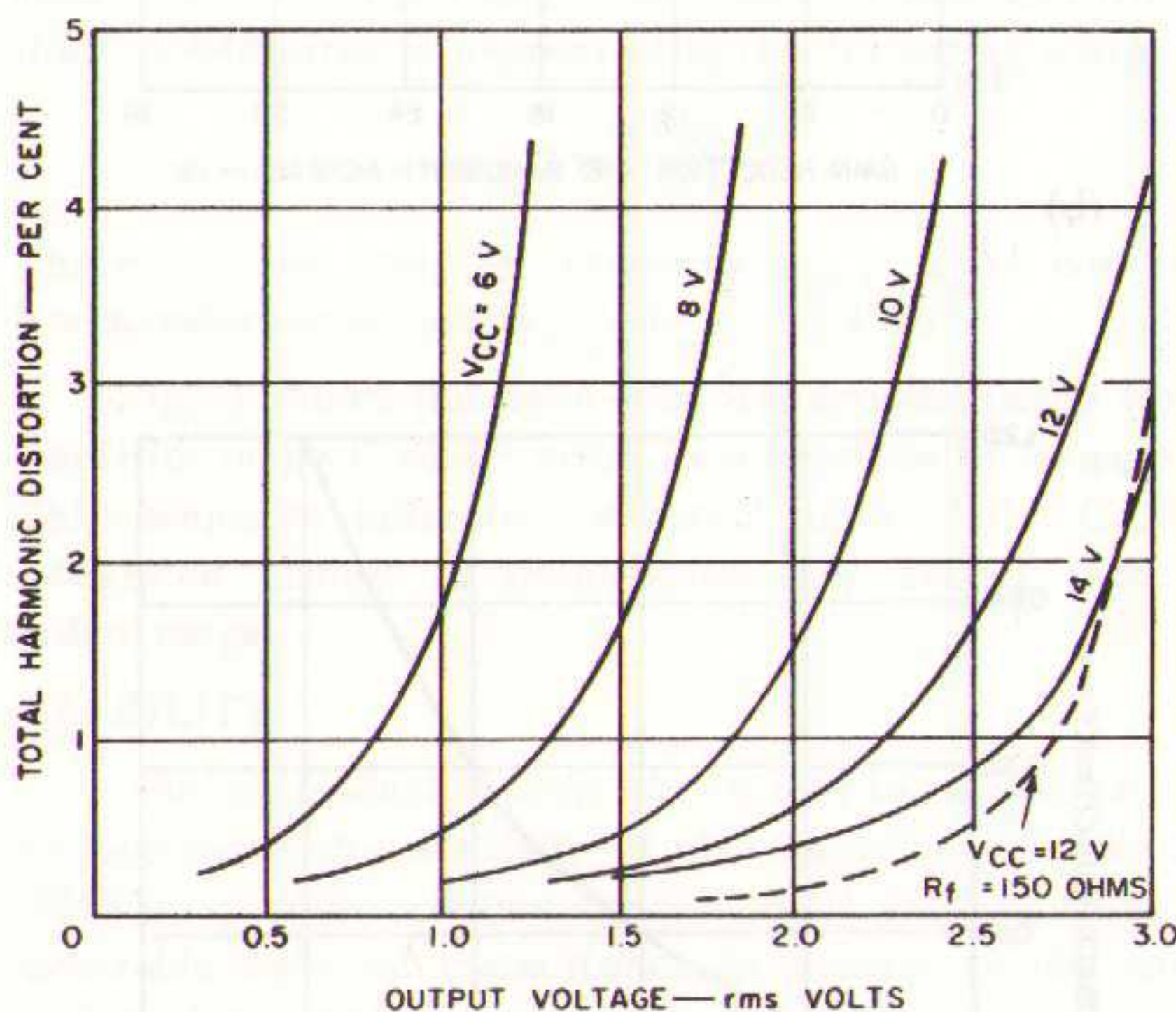


Fig. 5 - Total harmonic distortion of a CA3048 amplifier as a function of output voltage for different value of dc supply voltage.

NOISE

Fig. 6 shows output noise obtained when a single amplifier of the CA3048 is operated at 40 dB gain into a "C" filter. Table I shows typical values of noise

TABLE I
TYPICAL NOISE VOLTAGE AND CURRENT FOR AN AMPLIFIER IN THE CA3048 ARRAY

Frequency (Hz)	E_{noise} (volts)	I_{noise} (amperes)
10	30.5×10^{-9}	7.5×10^{-12}
100	17×10^{-9}	4.3×10^{-12}
1000	8×10^{-9}	1.2×10^{-12}
10000	6×10^{-9}	0.5×10^{-12}
100000	4×10^{-9}	0.3×10^{-12}

voltage (E_{noise}) and current (I_{noise}) for the CA3048 at spot frequencies of 10, 100, 1000, 10000, and 100000 Hz. From these values, the equivalent input noise voltage for any value of source resistance may be computed by use of the following equation:

$$E_{\text{equiv}} = \sqrt{(E_{\text{noise}})^2 + (I_{\text{noise}} R_{\text{source}})^2} \quad (4)$$

Laboratory measurements have shown that the noise performance of the CA3048 is not significantly affected by variation of the supply voltage. The values shown in Fig. 6, therefore, may be used with supply voltages down to about 2 volts if it is remembered that the open-loop gain decreases to about 35 dB at a supply voltage of 2.5 volts.

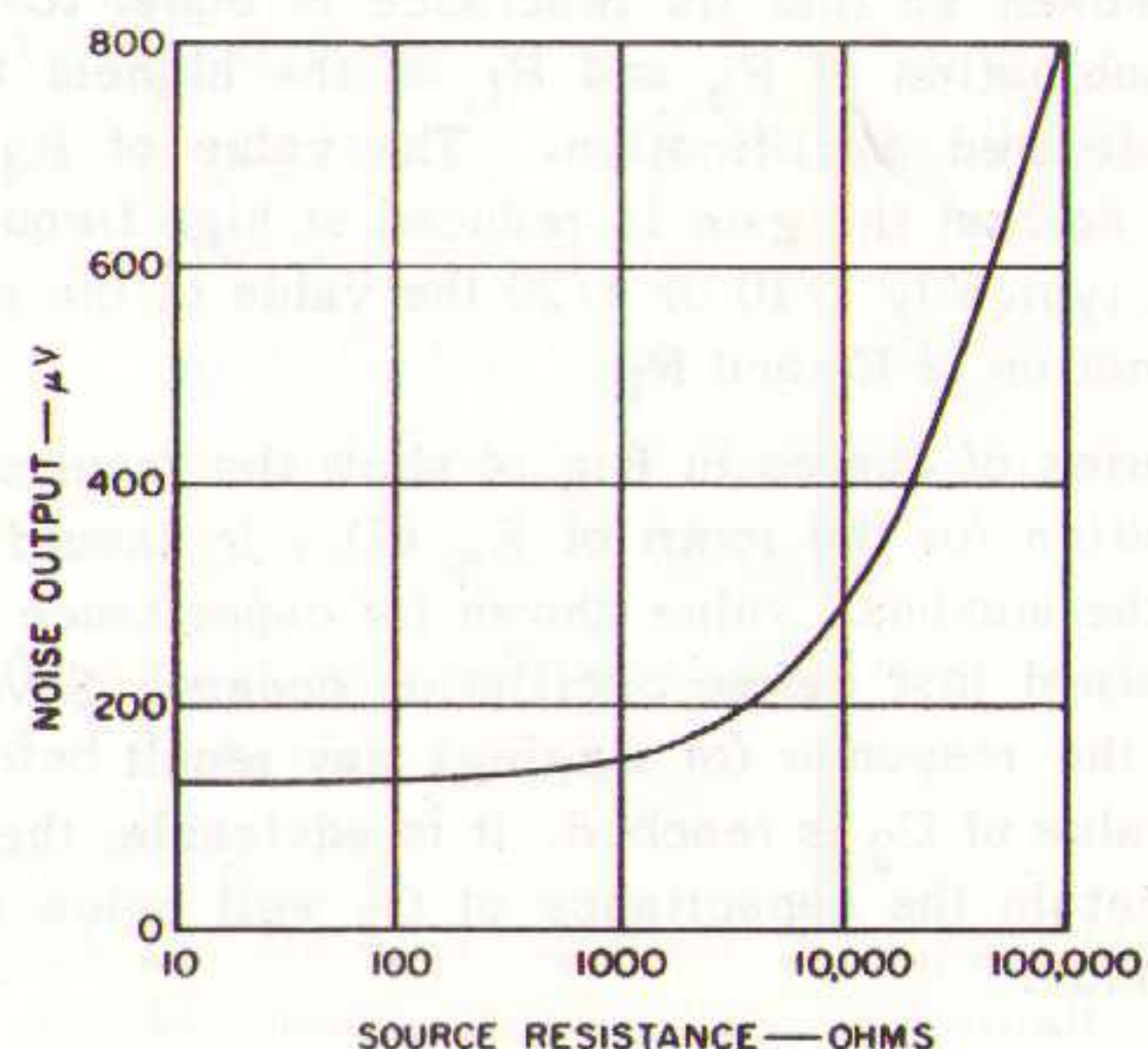


Fig. 6 - Noise output as a function of source resistance for a CA3048 amplifier.

CIRCUIT APPLICATIONS

In all the foregoing discussions, a single amplifier has been described as though it existed alone. The CA3048, however, consists of four separate amplifiers, which may be used independently or in combination. A glance at the complete schematic of the CA3048 reveals other aspects worthy of consideration.

Two supply-voltage terminals and two ground terminals are indicated. Terminal No. 12 supplies the V_{CC} voltage to amplifiers A_2 and A_3 , and terminal No. 15 supplies the V_{CC} voltage to amplifiers A_1 and A_4 . The ground return for amplifiers A_1 and A_4 is provided by terminal No. 2; all other ground returns are provided by terminal No. 5.

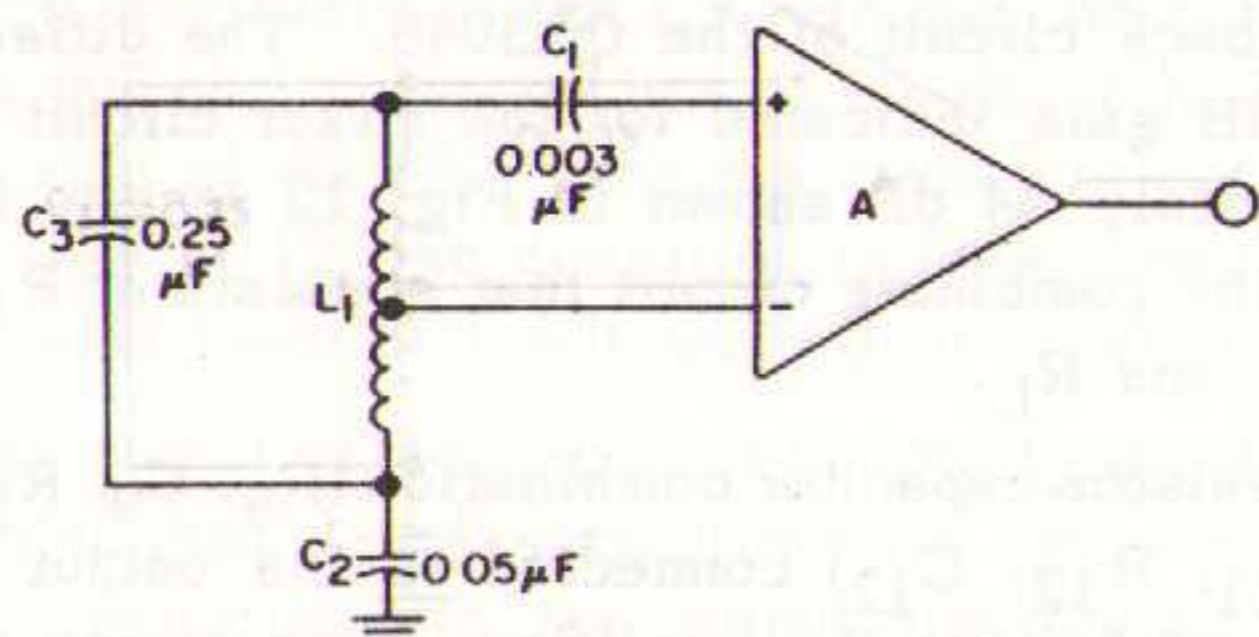
When two units are cascaded, it is preferred to let amplifiers A_2 and A_3 be the input units, and amplifiers A_1 and A_4 be the output units. This arrangement permits separation of both the V_{CC} and ground lines for low- and high-level signals.

If resistive decoupling is used, amplifiers A_2 and A_3 can be operated at lower V_{CC} voltages to effect a savings in current consumption.

Hartley Oscillator

The Hartley oscillator is easily designed and constructed using the CA3048 amplifier. No feedback capacitor is required, and it is possible to extract "square", sawtooth, or sinusoidal waveshapes.

In the circuit shown in Fig. 7, the tap on the coil is located at one-fourth the total turns, capacitors C_1 and C_2 provide dc blocking, and capacitor C_3 tunes with inductor L_1 ($\omega_0 = 1/\sqrt{L_1 C_3}$). When the circuit is oper-



A IS ANY AMPLIFIER OF THE CA3048

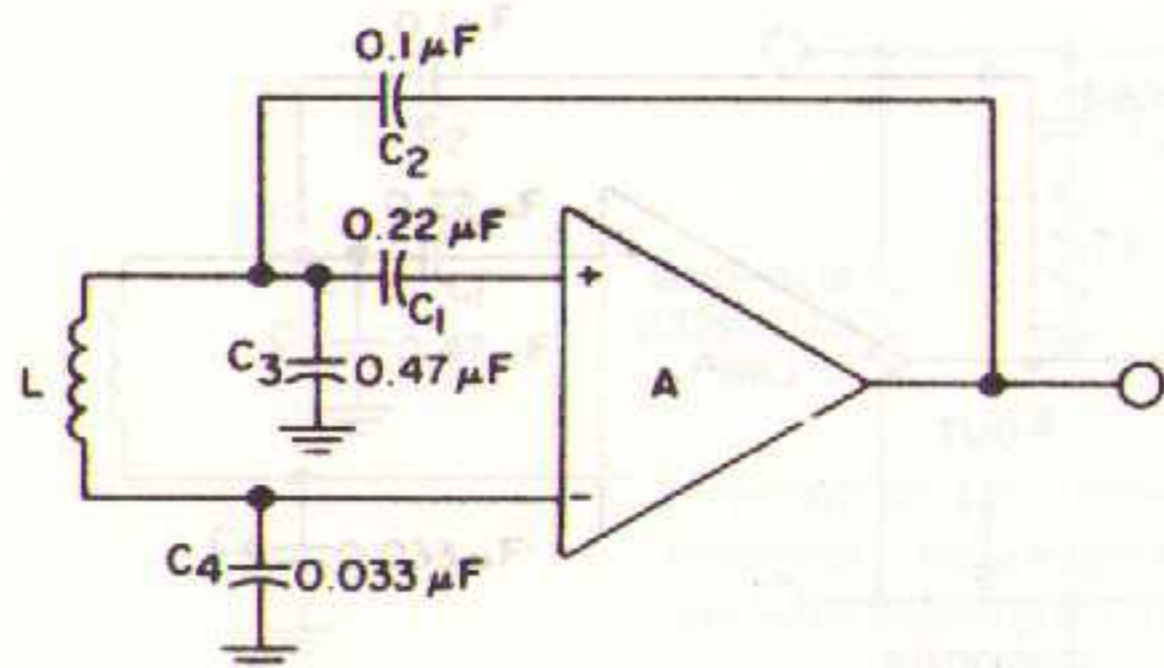
Fig. 7 - Hartley oscillator.

ated from a 12-volt supply, the output voltage is a clipped sine wave that has a peak-to-peak value of about 7 volts. The voltage at the inverting input is a sawtooth that has a peak-to-peak value of about 0.300 volt. If an unclipped sine wave is desired, it is available across the coil L_1 . A sine wave can be obtained in the single-ended connection if the value of C_2 is made large with respect to C_3 so that it effectively bypasses the sawtooth to ground; the voltage across L_1 is then sinusoidal with respect to ground.

Colpitts Oscillator

A tunable Colpitts oscillator is readily designed using one of the amplifiers of the CA3048 array. Fig. 8 shows an example of the CA3048 used in this way. Capacitors C_1 and C_2 are dc blocking capacitors; the series combination of capacitors C_3 and C_4 resonates with coil L . The ratio of C_3 to C_4 determines the relative amounts of signal fed back to the two inputs, and may be chosen on the basis of stability or strength of oscillation.

For the component values shown in Fig. 8, the frequency of oscillation is 33.536 kHz with a 12-volt supply and decreases to 33.546 kHz when the supply voltage is reduced 25 per cent to 9 volts.



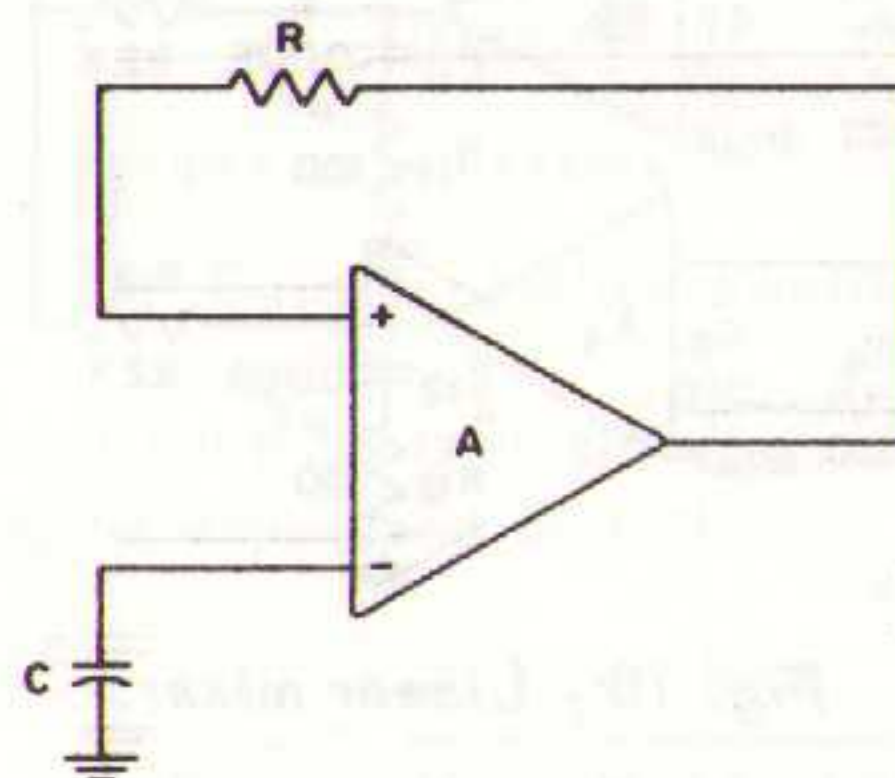
A IS ANY AMPLIFIER OF THE CA3048

Fig. 8 - Colpitts oscillator.

Three waveshapes are available from the Colpitts oscillator. A sawtooth waveform is obtained at the output, a sinusoidal waveform is obtained at the inverting input, and a clipped sinusoidal waveform appears at the noninverting input.

Astable Multivibrator

The CA3048 may be connected as an astable multivibrator with the addition of only two external components. An example of this type of operation is shown in Fig. 9.



A IS ANY AMPLIFIER OF THE CA3048

Fig. 9 - Astable multivibrator.

The resistor R introduces positive feedback into the circuit, and the capacitor C sets the period of the waveform. The operation of the circuit can be explained more easily if it is assumed that transistor Q_{17} (of amplifier A_3) has just turned OFF so that the voltage at terminal No. 11 becomes very positive. This positive voltage is fed through R to the base of Q_{19} to maintain the conduction of this transistor and to hold transistors Q_{21} , Q_{13} , Q_{14} , and Q_{17} cut off. Meanwhile, capacitor C charges through the internal bias resistors R_{41} and R_{42} . When the voltage on capacitor C reaches the level at which transistor Q_{21} begins to become forward-biased, some current is diverted from Q_{20} to Q_{21} and Q_{13} , Q_{14} , and Q_{17} begin to turn ON. The action is regenerative because the negative-going voltage from the collector of Q_{17} feeds a negative-going signal back to the base of Q_{19} to enhance the switching action. When C discharges to the point at which Q_{21} turns OFF, Q_{20} begins to turn ON and the process repeats itself.

Two waveforms are available from the astable multivibrator circuit, both at low impedance. A rectangular waveform that has a peak-to-peak amplitude of 7 volts or greater is obtained from the output terminal. The waveform available at the inverting input is an isosceles triangle that has a peak-to-peak amplitude of approximately 0.220 volt.

With the circuit as shown, reliable oscillation is obtained for values of the resistor R in the order of 2.2 megohms, with supply voltages as low as six volts.

4-Channel Linear Mixer

Fig. 10 illustrates the use of the CA3048 as a linear mixer. Each input is connected to its own CA3048

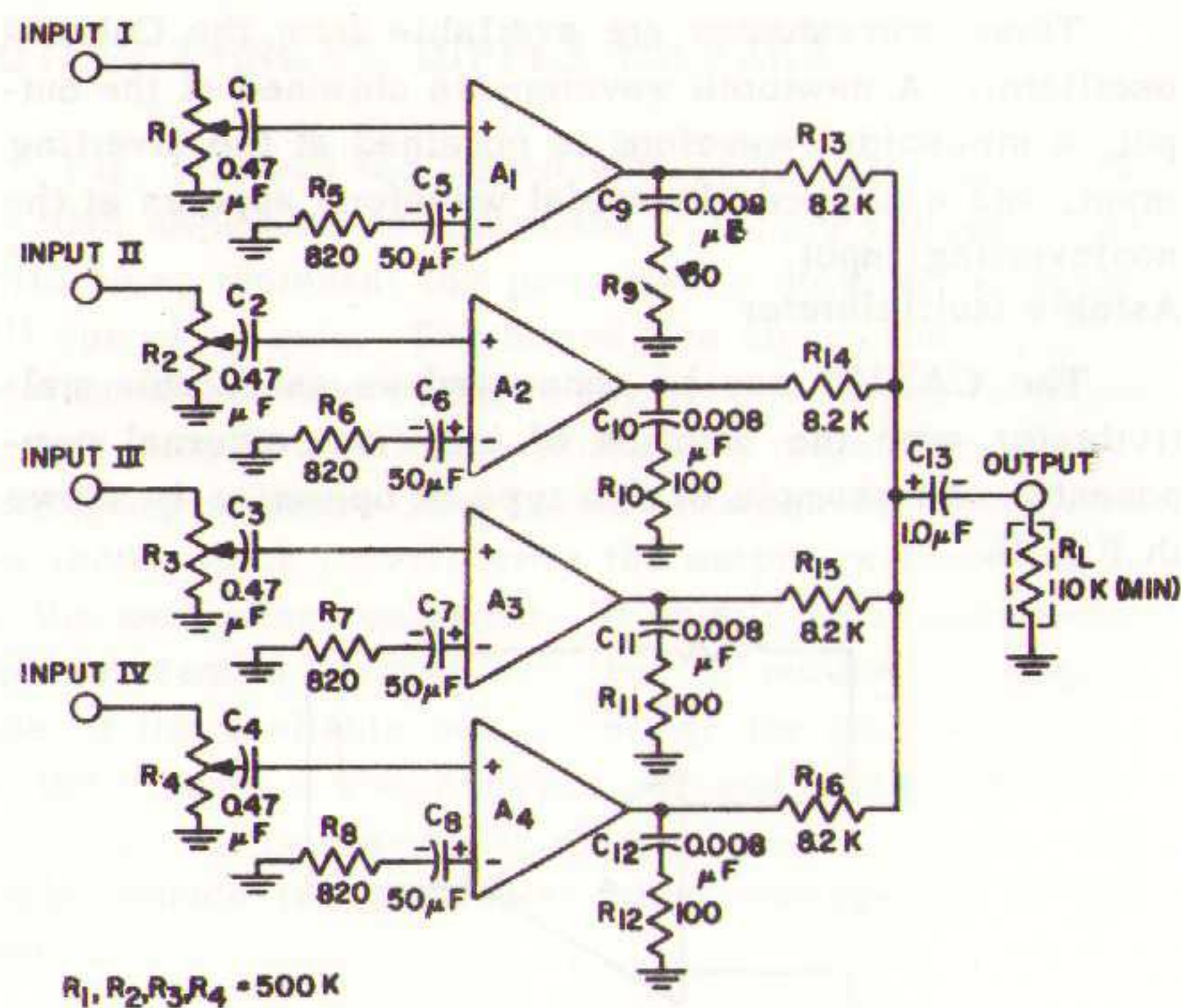


Fig. 10 - Linear mixer.

amplifier through the gain-control potentiometers R_1 , R_2 , R_3 , and R_4 . Capacitors C_1 , C_2 , C_3 , and C_4 block the dc voltage at the inputs.

The gain of any input to the corresponding output is 20 dB for the circuit values shown and a load impedance of 10000 ohms or greater. Resistors R_5 , R_6 , R_7 , and R_8 program the gain of the system, and may be varied to provide more or less gain, depending on the requirements of the application. The curve in Fig. 11 illustrates the effect of variation in the resistance in the feedback circuit of the CA3048. The difference in the 20 dB gain indicated for the mixer circuit and the approximately 34 dB shown in Fig. 12 results from the loss in the combining circuit that consists of R_{13} , R_{14} , R_{15} , R_6 , and R_L .

A resistor-capacitor combination (R_9 , C_9 , R_{10} , C_{10} , R_{11} , C_{11} , R_{12} , C_{12}) connected to the output of each amplifier stabilizes the amplifiers when source and load conductances are too small to provide adequate damping.

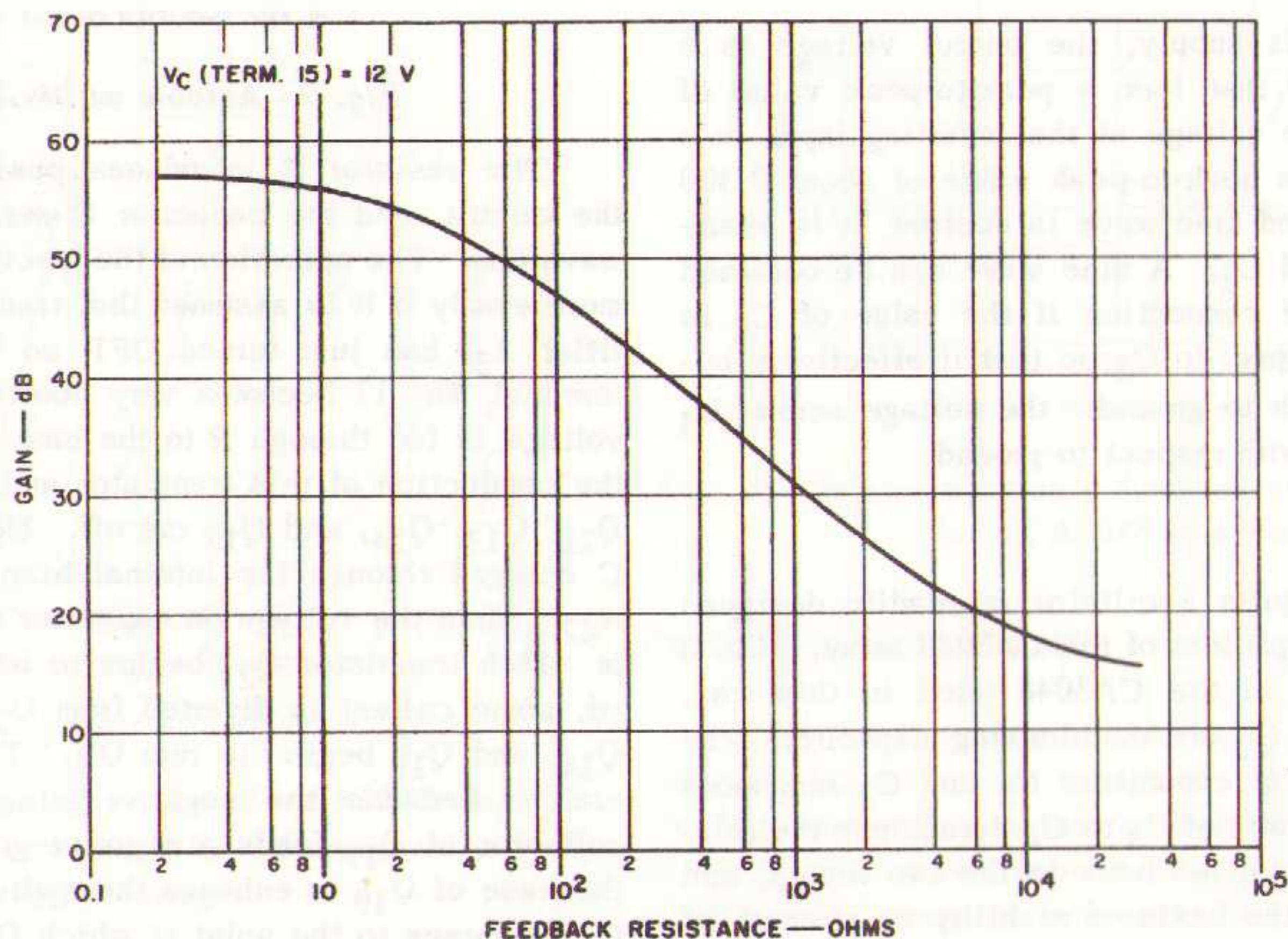


Fig. 11 - Gain of a CA3048 amplifier as a function of feedback resistance.

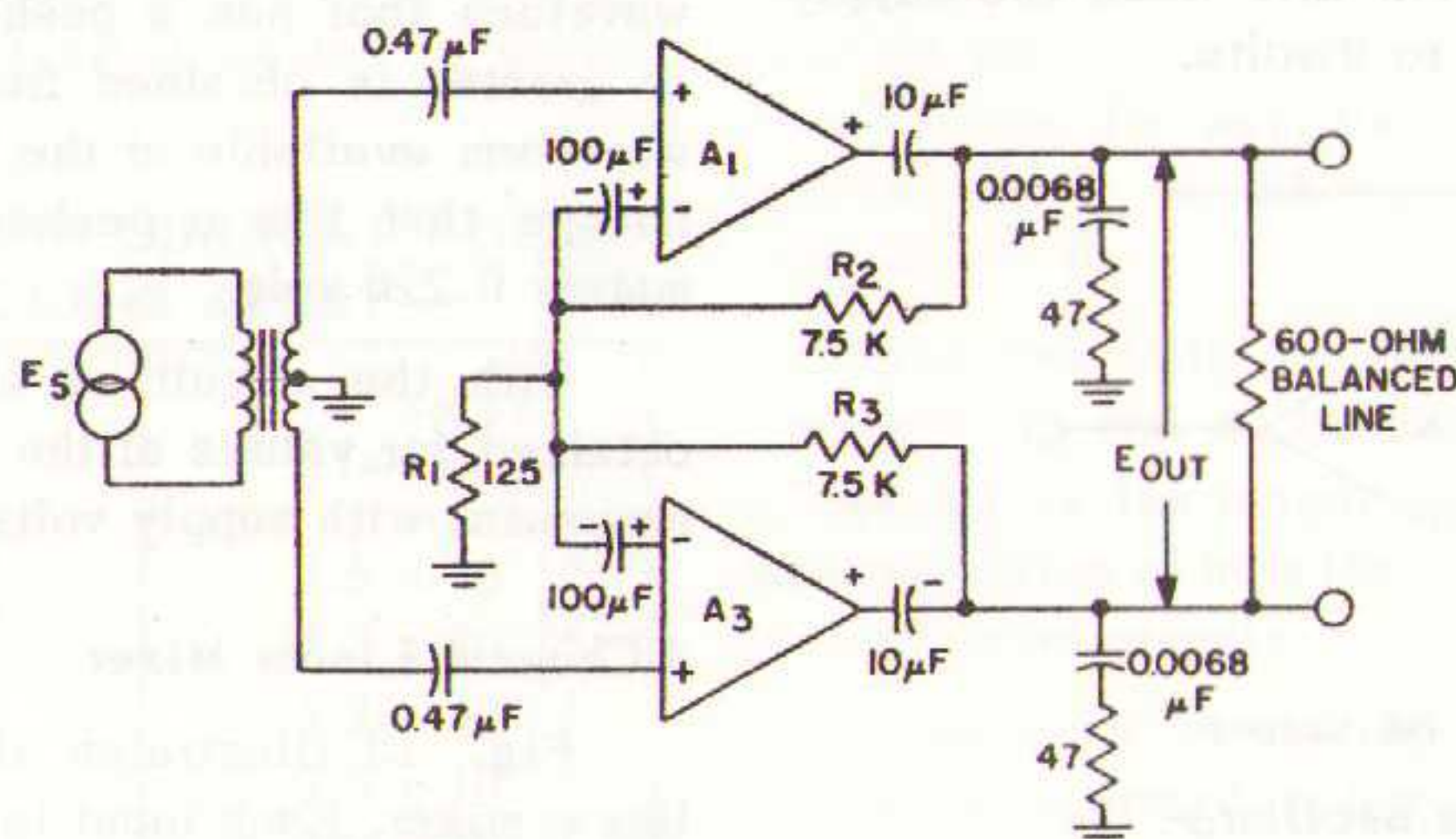


Fig. 12 - Balanced-line driver.

The input impedance of each amplifier of the CA-3048 is nominally 90 kilohms. In the linear mixer, however, the input potentiometers R_1 through R_4 are 500 kilohms. The effective impedance presented to the device, therefore, is quite high except when the circuit is adjusted for maximum gain. At this time, the impedance decreases to about 75 kilohms.

Driver for 600-ohm Balanced Line

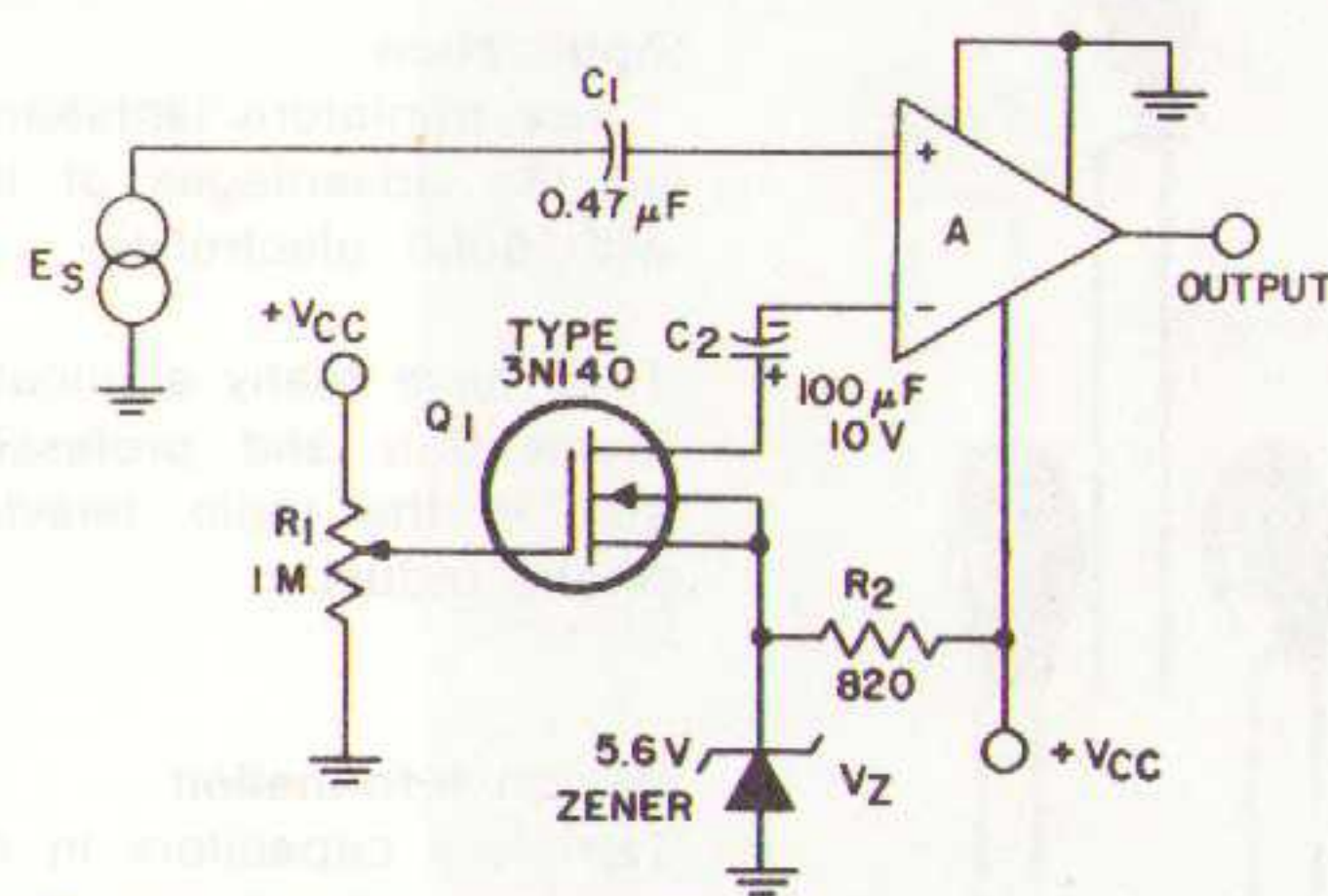
Two amplifiers of the CA3048 may be connected to drive a 600-ohm balanced line at levels up to 1 volt rms with a gain of 40 dB. When the circuit is connected as shown in Fig. 12, the distortion is less than 1 per cent at an output level of 1 volt and a gain of 40 dB.

The output of the circuit is limited to a value slightly greater than 1 volt rms, primarily because of drive-current limitations to the output transistors. In this respect, it is self-protecting. Should a short circuit develop across the line, the circuit will not destroy itself.

Resistor R_1 in Fig. 12 is common to the output and input circuits of both amplifiers A_1 and A_3 . Should a gain unbalance exist, or should the input signals be of unequal amplitude, then the outputs would tend to become unbalanced with respect to ground. For example, if amplifier A_1 had the larger output, a signal in phase with the output at A_1 would be developed across resistor R_1 . In this event, the voltage developed at R_1 would tend to reduce the output of amplifier A_1 because this voltage is applied to the inverting input. At the same time, the voltage at R_1 is applied to the inverting input of amplifier A_3 and tends to increase the effective input voltage of that amplifier and, in this way, help to restore balance. The balancing effect takes place regardless of the cause of the initial unbalance.

Gain-Controlled Amplifier

Any amplifier of the CA3048 may be used as a gain-controlled amplifier in order to accommodate a wide range of input signal amplitude. Fig. 13 shows one amplifier of the CA3048 used in this type of configuration. By variation of the dc potential at the gate No. 1 of the MOS transistor Q_1 , the gain of the amplifier may be varied from 14 dB to 49 dB. In this circuit, the MOS transistor Q_1 acts as a variable impedance in the feedback loop of the CA3048.



"A" IS ANY AMPLIFIER OF THE CA3048

Fig. 13 - Gain-controlled amplifier.

In a circuit such as that shown in Fig. 13, the voltage gain may be expressed as follows:

$$\frac{E_o}{E_s} = \frac{A(R_a + R_f)}{R_f + R_a(A + 1)} \quad (5)$$

where R_a is the equivalent resistance of the MOS transistor and R_f is the feedback resistance, which includes the internal feedback resistance of the integrated circuit together with any paralleled external feedback resistance.

As the value of resistor R_a approaches zero, the gain of the circuit approaches the open-loop gain of the amplifier (A). For very large values of resistor R_a the gain of the circuit approaches $A/(A + 1)$, or approximately unity. The maximum theoretical agc range then is A.

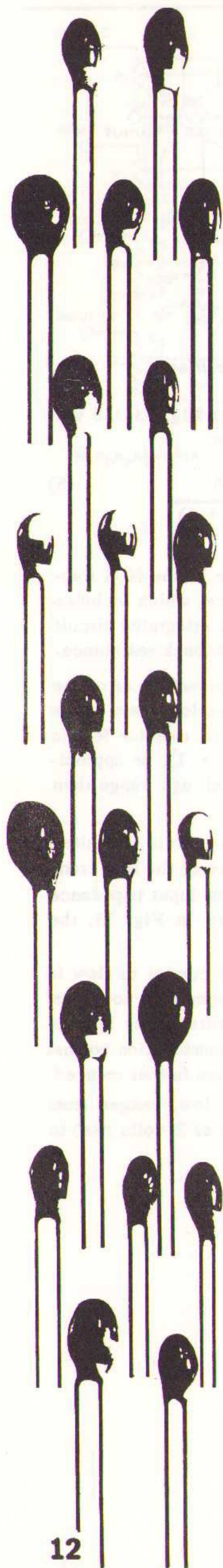
The practical agc range of this circuit is limited on the high end by the "ON" resistance of the MOS transistor, and on the low end by the finite input impedance of the CA3048. In the circuit shown in Fig. 13, the range of control is 35 dB.

There is no necessity for direct current to flow in the MOS transistor, and if a low impedance source of about five volts is available, this voltage may be substituted for the Zener diode-resistor combination so that the power requirements of the circuit are further reduced.

Distortion, which is inherently low, ranges from 0.65 per cent at minimum gain (output of 2 volts rms) to 0.4 per cent at maximum gain.

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Tantalum Capacitors TAG



Application

These miniature tantalum capacitors have all the advantages of the normal types with solid electrolyte.

They have many applications in the communication and professional fields, and also in the radio, television and gramophone industry.

Design Information

Tantalum capacitors in drop form are of polar construction with sintered anode, tantalum oxide as dielectric and a solid cathode. The capacitors are coated with a synthetic resin. The terminal wires are brought out on one side. For indicating the anode, the capacitor has been provided with a spot. When this spot is viewed with the terminals downwards the anode is at the right of the spot.

Electrical Characteristics

Tantalum capacitors in drop form correspond in their electrical and mechanical characteristics to "VDE 0560 Part 16, Construction SF".

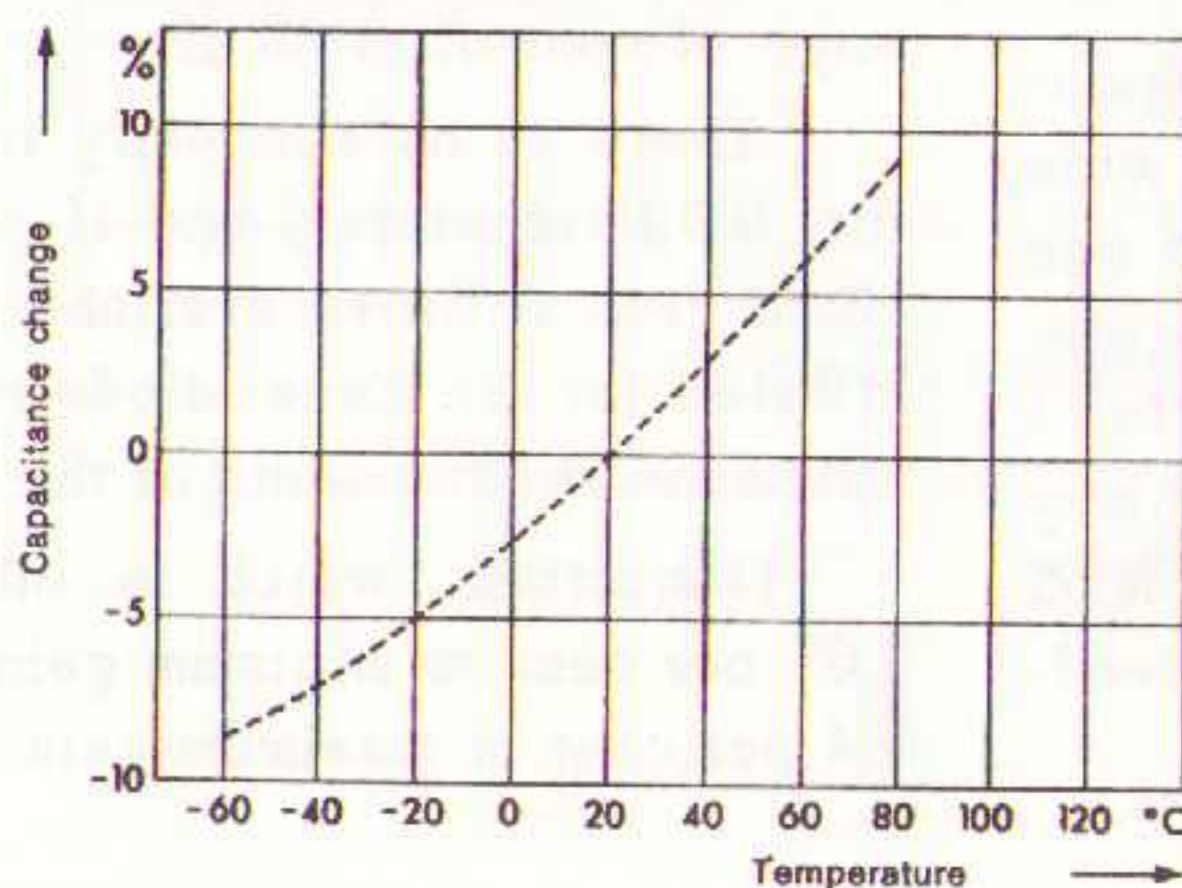
Nominal capacitance

The nominal capacitance is measured with a bridge arrangement at +20 °C with 50 Hz alternating voltage, free from harmonics, and is the capacitance of the equivalent series circuit.

Capacitance tolerance

+50% to -20%

Variation of capacitance with temperature



Rated voltage V_N

This voltage applies for a capacitor surface temperature up to 85 °C.

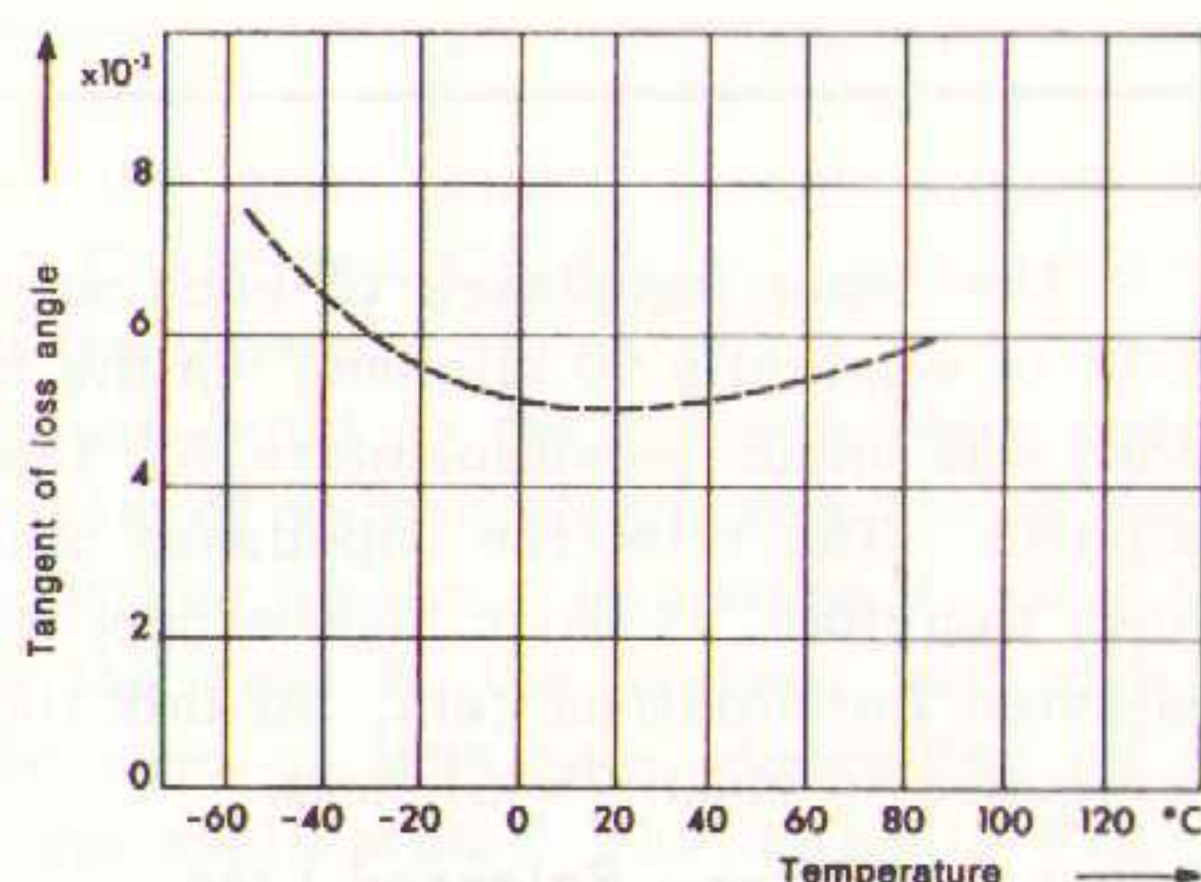
The rated voltages are:
3/6/3/10/16/20/25/35 V d. c.

Tangent of loss angle $\tan \delta$ (dissipation factor)

Maximum value : $\tan \delta = 0,1$.

Measurement is performed with alternating voltage, free from harmonics, of ≤ 0.3 V, 50 Hz at 20 °C.

Variation of $\tan \delta$ with temperature



Leakage current I_R

Maximum value = $0,05 C_N \cdot V_N$ or 2 μA , which ever is the greatest.

Measurement is performed at +20 °C with a series resistance of 1 k Ω .

Measurement is made 3 minutes after application of the measuring voltage.

Reverse voltage

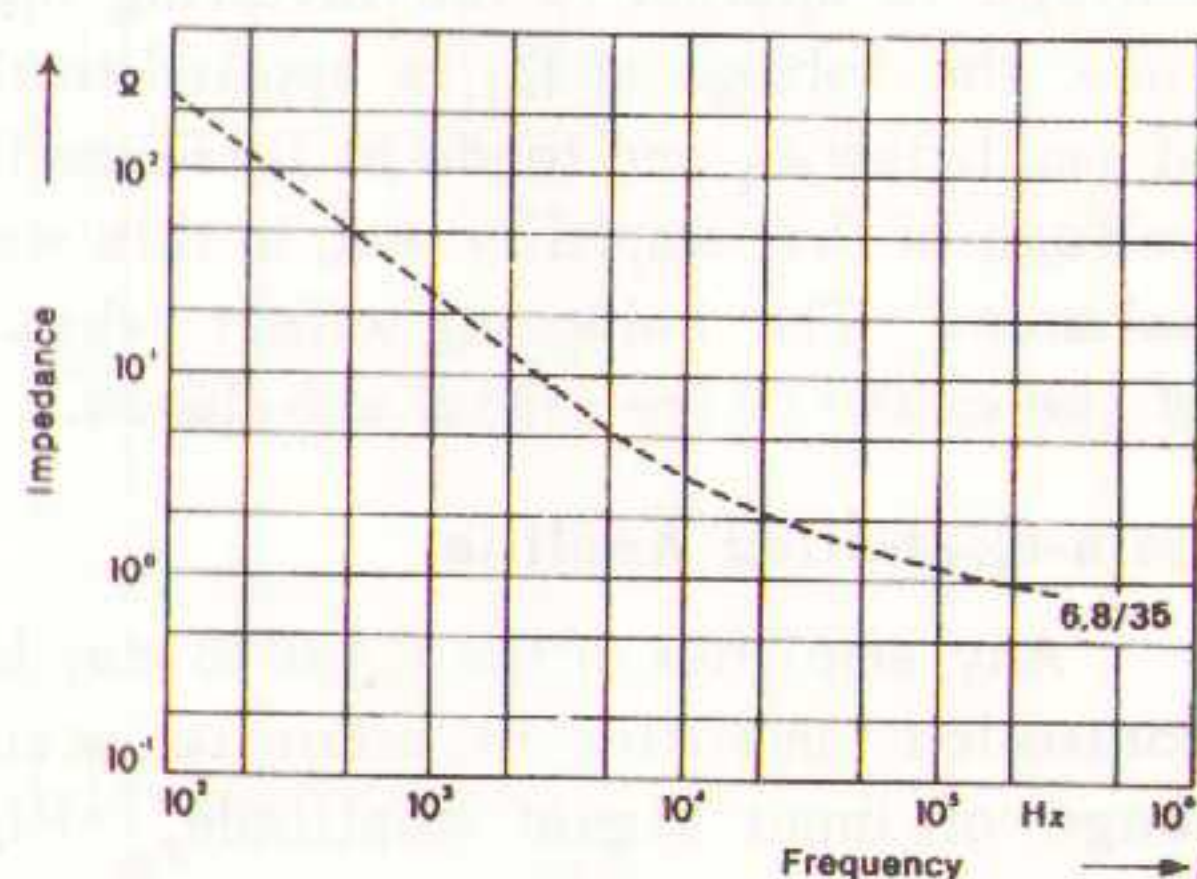
Maximum 0.5 V for rated voltage

$V_N \geq 6$ V and 0.3 V for $V_N < 6$ V

Superimposed alternating voltage

The sum of the d. c. voltage and of the superimposed peak alternating voltage must not exceed the rated voltage, and no inadmissible reverse polarity must occur.

Variation of impedance with frequency



The maximum impedance values, as given in table on page 7 are lower than those for comparable professional grade aluminium electrolytic capacitors.

Variation of impedance with temperature

At -55 °C, the impedance (measured at 10 kHz) will be twice that at room temperature.

Climatic properties

Temperature range for operation and storage -55 °C to +85 °C. Humidity according to IEC 68 test C, 21 days. Meets environmental test conditions of Military-Specification MIL-C-26655 B.

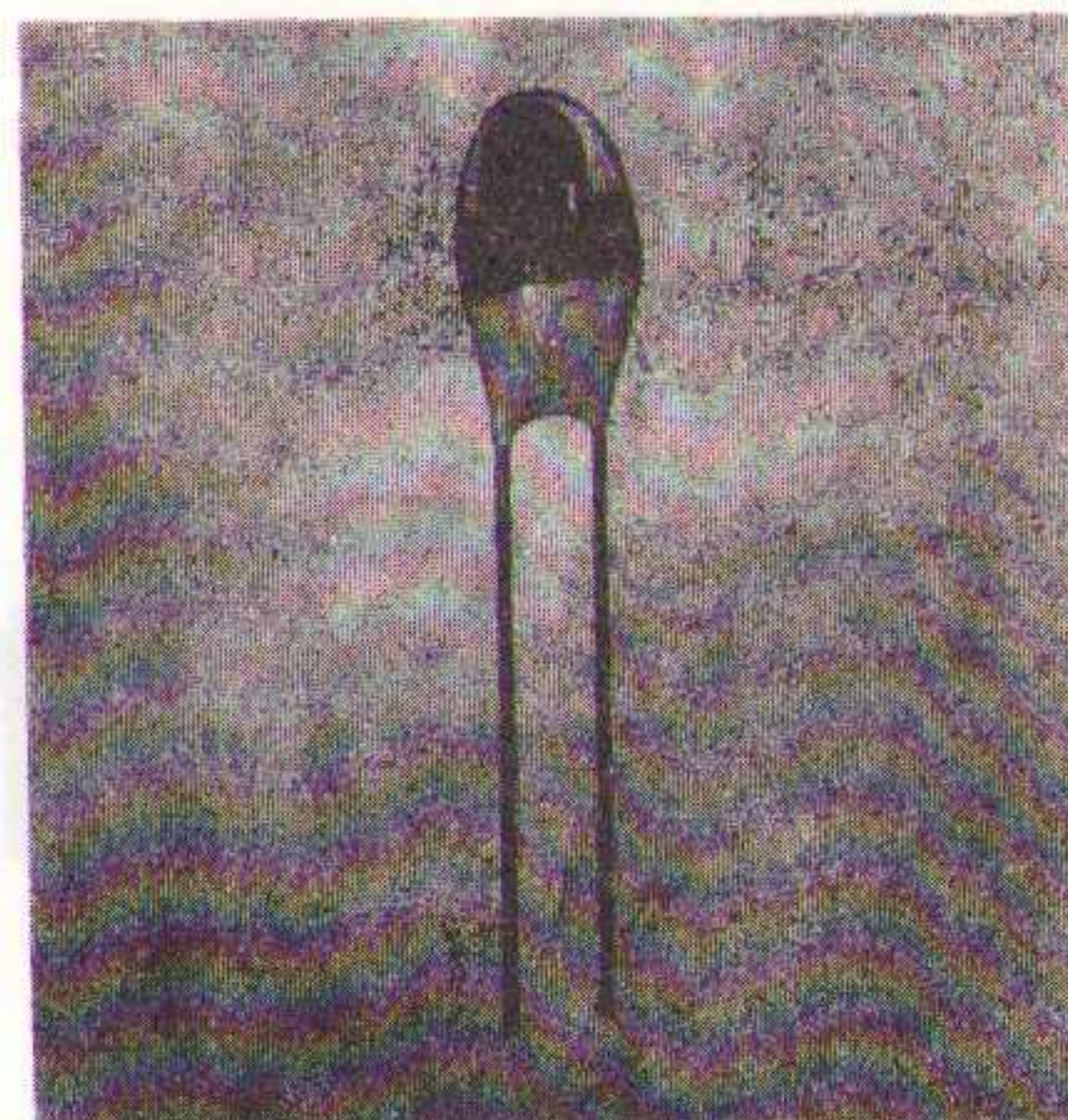
Mechanical properties in accordance with IEC 68 test U

Tensile strength of the terminal wires: 0.5 kg. Bending strength of the terminal wires: 2 bends of 90°. Dip-solderable.

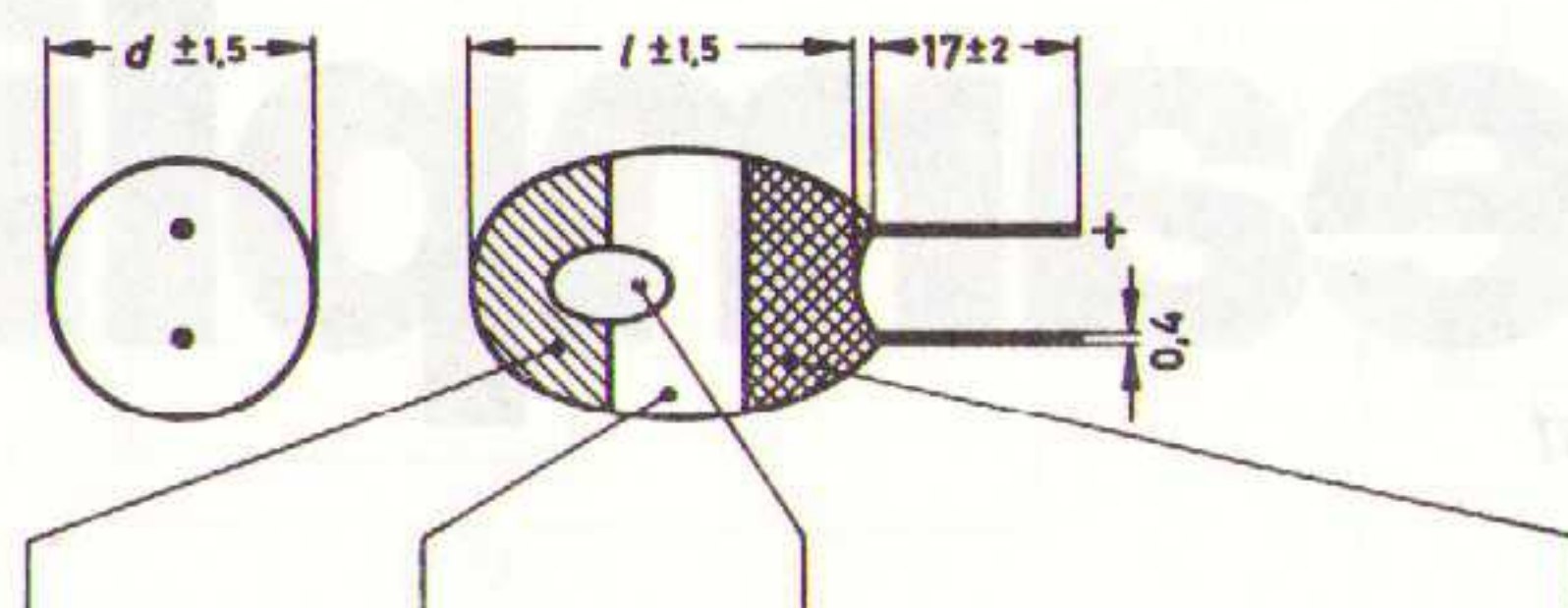
Drop-form type with rated voltages of 3/6,3/10/16/20/25/35V DC. Surge voltage = 1,15 x rated DC voltage.

Design: SF (sintered anode, solid electrolyte), polar.

Categorie according to IEC: 55 / 85 / 21,
Temperature range: -55 to +85°C.



Dimensions in mm, marking of capacity according to IEC and DIN 40820.



Capacity in μF				Rated d. c. voltage	
Colour	Dome 1. digit	Ring 2. digit	Dot multiplier marking of polarity	Colour	Voltage
black	—	0	x 1	white	3 V
brown	1	1	x 10	yellow	6,3 V
red	2	2	—	black	10 V
orange	3	3	—	green	16 V
yellow	4	4	—	blue	20 V
green	5	5	—	grey	25 V
blue	6	6	—	pink	35 V
violet	7	7	—		
grey	8	8	x 0,01		
white	9	9	x 0,1		

Marking of capacity: colour dot indicates the polarity orientation according to drawing.

Nominal capacity μF	Rated DC voltage volt	Impedance Z (Ω) at 10kHz		Dimensions d x l mm	Weight g
		limit value Ω	average value Ω		
0,1	35	310	165	3,5x6,0	0,15
0,15	35	200	110	3,5x6,0	0,15
0,22	35	150	75	3,5x6,0	0,15
0,33	35	105	51	3,5x6,0	0,15
0,47	35	68	37	3,5x6,0	0,15
0,68	35	53	27	3,5x6,0	0,15
1,0	35	34	18	4,0x6,5	0,20
1,5	25	25	13	4,0x6,5	0,20
2,2	25	17	9	4,0x6,5	0,20
3,3	25	14,5	6,5	4,0x7,0	0,25
4,7	25	10	4,5	4,5x7,0	0,30
6,8	25	6	3,0	5,5x8,5	0,40
10,0	6,3	9,5	3,0	4,0x6,5	0,20
10,0	20	5,0	2,0	5,5x8,5	0,40
22,0	16	3,5	1,0	6,0x8,5	0,45
33,0	10	3,5	0,8	6,0x8,5	0,45
47,0	6,3	3,0	0,6	6,0x8,5	0,45
100,0	3	3,0	0,5	6,0x8,5	0,45

Capacity tolerance all values: +50 to -20%.

RCA

A Simplified IC Stereo Preamplifier

by M. B. Knight

Reprinted from

EEE,

September 1969

The recent availability of quadruple-amplifier integrated circuits has provided stereo circuit designers with a new design tool. By cascading two amplifiers in each channel, the designer can reap a number of benefits.

Division of each channel into two amplifier segments permits insertion of gain control, tone control, or equalization at a signal level compatible with good signal-to-noise ratio; also it allows wide dynamic range in an amplifier with large over-all gain. In a typical system design, the signal from a tape head is inserted directly into an input and the output of that segment is applied through a frequency-response equalizing network of either the passive or the feedback type to the volume control. The resultant attenuated signal is applied to the cascaded output-amplifier segment. If tone controls that attenuate the reference signal level are desired, the output-amplifier segment is used for feedback tone control or as the driver of passive tone controls. Because this is accomplished with a single IC, each amplifier segment consumes little space and a small overall number of terminals are required.

Tape preamplifier

The advantages of a quadruple-amplifier IC

Author: M. B. Knight is an applications engineer for RCA Electronic Components, Somerville, N. J.

in stereo circuitry can be demonstrated by substituting the IC in the preamplifier of an eight-track stereo tape player. The "outboard" components and connections are shown in the schematic diagram of Fig. 1. The value of the coupling capacitors from the tape heads to the amplifier inputs may seem large from the viewpoint of coupling signal to a high-input-impedance amplifier, but this is determined by the need for the low circuit noise associated with a low-source impedance. The 50- μ F feedback capacitors and the 75- Ω feedback resistors restrict the maximum voltage gain of the input-amplifier segments to about 45 dB. The 0.22- μ F capacitors and the 200- Ω resistors provide a feedback circuit that further attenuates the higher audio frequencies to correspond approximately to the National Association of Broadcasters (NAB) standard playback frequency response.

Although the preamplifier is stable when a common supply voltage is used for the four amplifier segments with the inputs terminated, decoupling is often advisable to ease the requirements on power-supply impedance with regard to filtering, channel separation, and stability. Typical current drain is shown in Fig. 2. In this example, the supply to the input-amplifier segments is dropped to 6V; the decoupling circuit is inexpensive, the power input to the IC is small, and yet the dynamic range remains more than adequate. The total current from the 12-V supply is 20 mA.

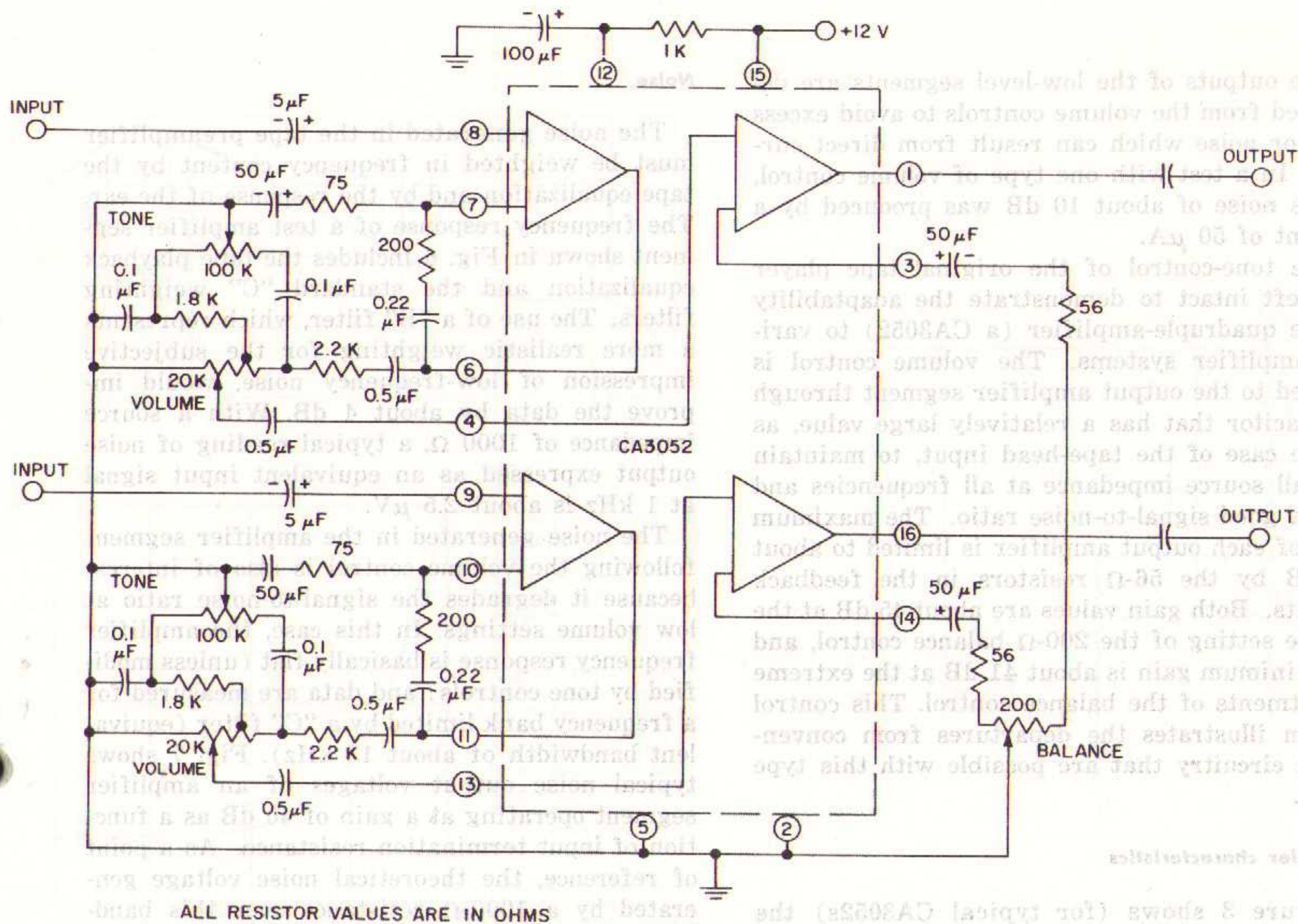


Fig. 1. Schematic diagram of complete preamplifier for eight-track stereo tape player showing discrete components used with CA3052 IC

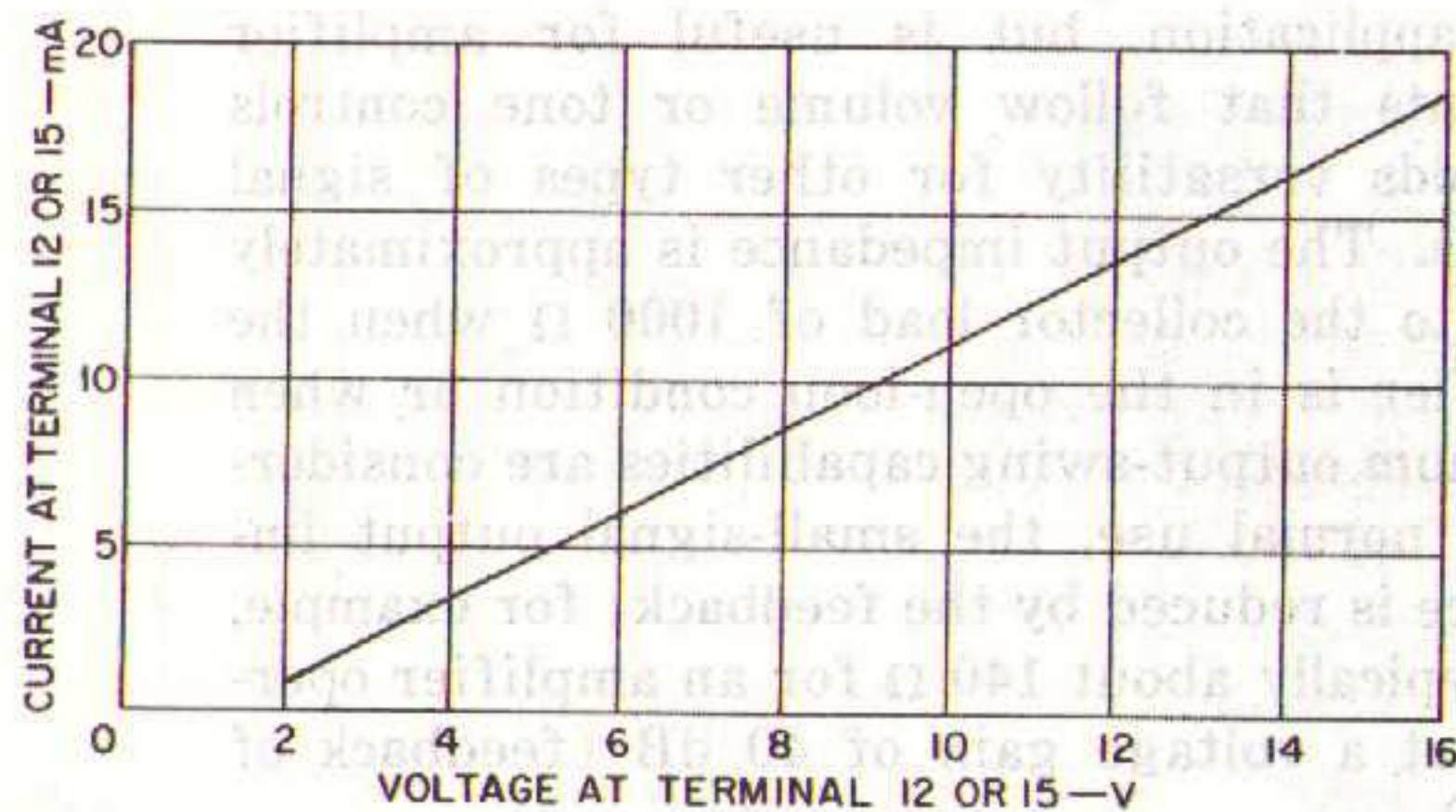


Fig. 2. Typical power-supply input currents as a function of voltages applied

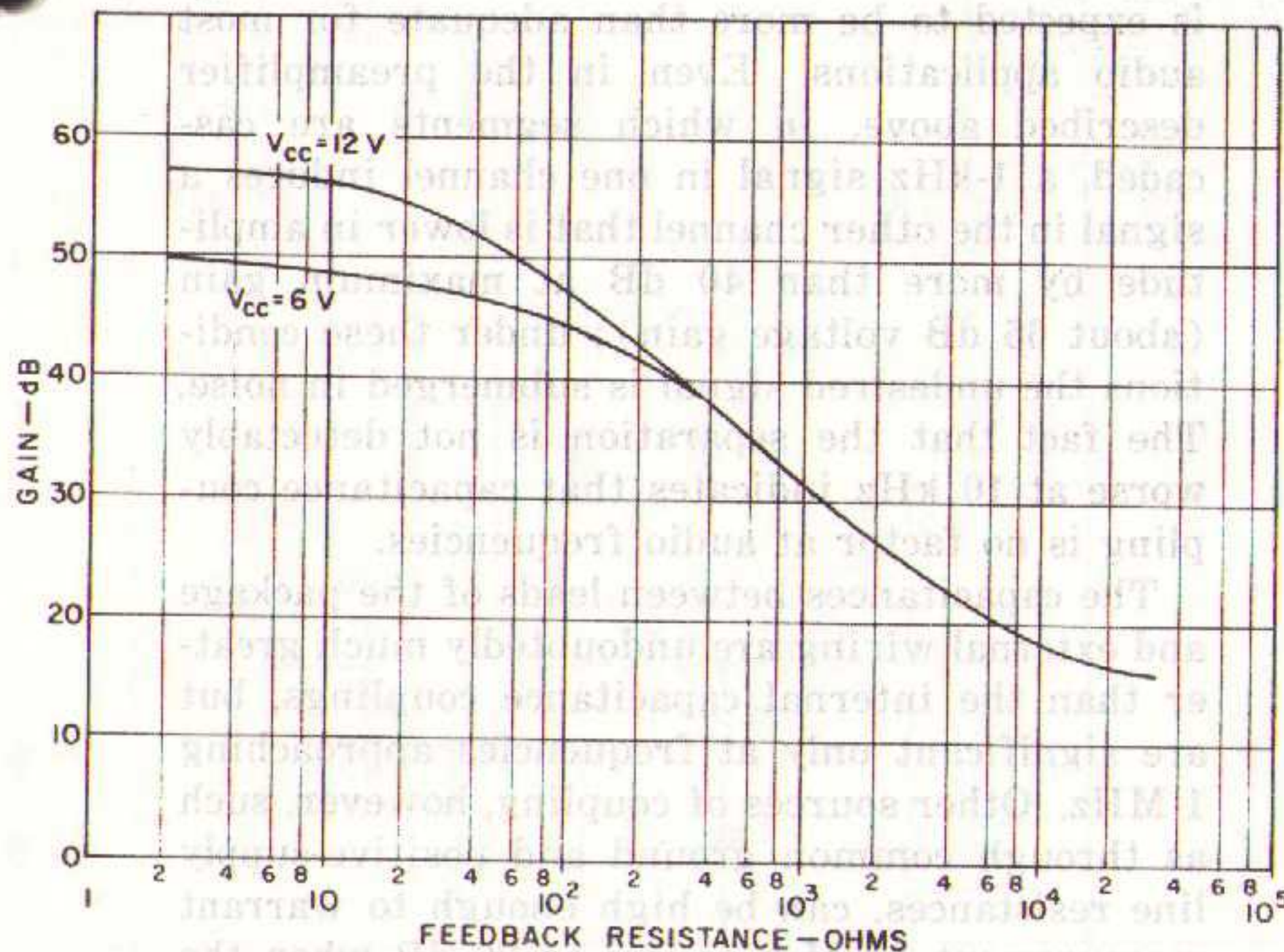


Fig. 4. Typical voltage gain of each amplifier segment as a function of feedback resistance (R_f)

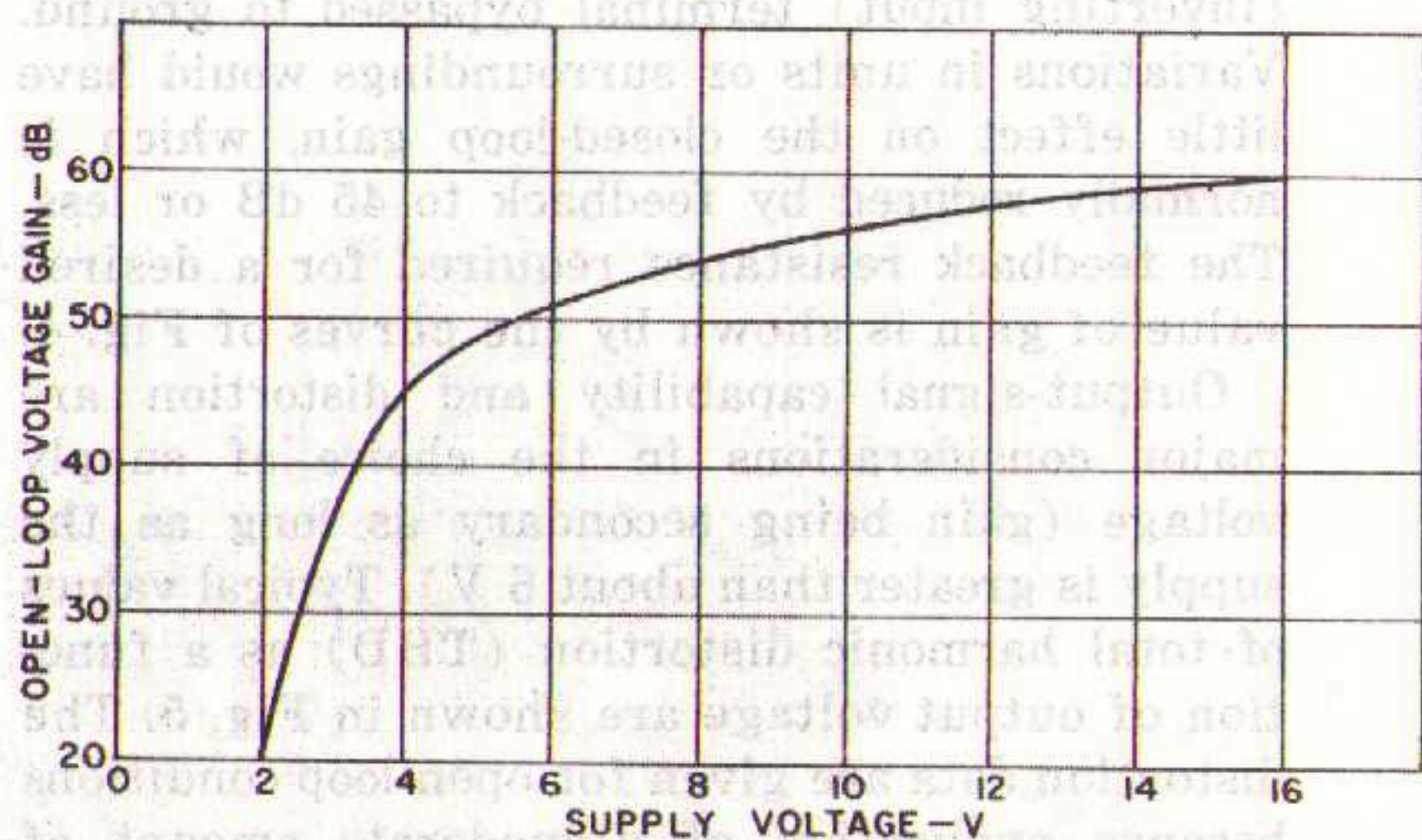


Fig. 3. Typical open-loop voltage gain (feedback resistor $R_f=0$) of each segment as a function of dc supply voltage

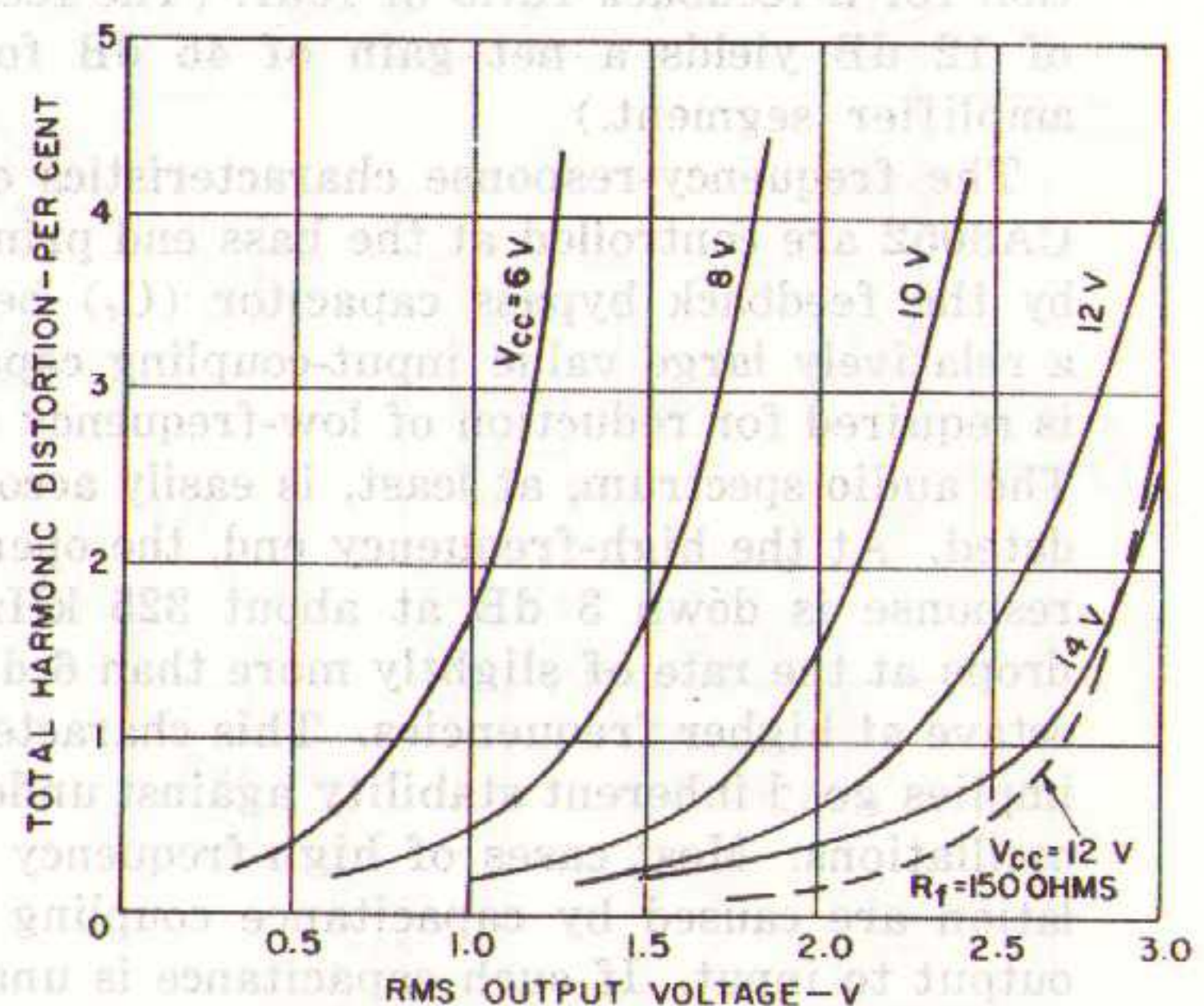


Fig. 5. Typical total harmonic distortion as a function of output voltage for each amplifier segment; open-loop (solid) and with 12 dB of feedback (dashed)

The outputs of the low-level segments are dc-isolated from the volume controls to avoid excess resistor noise which can result from direct current. In a test with one type of volume control, excess noise of about 10 dB was produced by a current of 50 μ A.

The tone-control of the original tape player was left intact to demonstrate the adaptability of the quadruple-amplifier (a CA3052) to various amplifier systems. The volume control is coupled to the output amplifier segment through a capacitor that has a relatively large value, as in the case of the tape-head input, to maintain a small source impedance at all frequencies and thus a good signal-to-noise ratio. The maximum gain of each output amplifier is limited to about 50 dB by the 56- Ω resistors in the feedback circuits. Both gain values are about 45 dB at the middle setting of the 200- Ω balance control, and the minimum gain is about 41 dB at the extreme adjustments of the balance control. This control system illustrates the departures from conventional circuitry that are possible with this type of IC.

Amplifier characteristics

Figure 3 shows (for typical CA3052s) the effect of supply voltage on the open-loop voltage gain, *i.e.*, the gain obtained with the feedback (inverting input) terminal bypassed to ground. Variations in units or surroundings would have little effect on the closed-loop gain, which is normally reduced by feedback to 45 dB or less. The feedback resistance required for a desired value of gain is shown by the curves of Fig. 4.

Output-signal capability and distortion are major considerations in the choice of supply voltage (gain being secondary as long as the supply is greater than about 6 V). Typical values of total harmonic distortion (THD) as a function of output voltage are shown in Fig. 5. The distortion data are given for open-loop conditions because application of a moderate amount of feedback reduces the THD by the feedback ratio. The dashed curve in Fig. 5 illustrates this reduction for a feedback ratio of four. (The feedback of 12 dB yields a net gain of 45 dB for the amplifier segment.)

The frequency-response characteristics of the CA3052 are controlled at the bass end primarily by the feedback bypass capacitor (C_f) because a relatively large value input-coupling capacitor is required for reduction of low-frequency noise. The audio spectrum, at least, is easily accommodated. At the high-frequency end, the open-loop response is down 3 dB at about 325 kHz and drops at the rate of slightly more than 6 dB per octave at higher frequencies. This characteristic implies good inherent stability against undesired oscillations. Most cases of high-frequency oscillation are caused by capacitance coupling from output to input. If such capacitance is unavoidable in the circuit layout, a termination impedance to reduce the high-frequency response can be used to stop oscillation.

Noise

The noise generated in the tape preamplifier must be weighted in frequency content by the tape equalization and by the response of the ear. The frequency response of a test amplifier segment shown in Fig. 6 includes the tape playback equalization and the standard "C" weighting filters. The use of a "B" filter, which represents a more realistic weighting for the subjective impression of low-frequency noise, would improve the data by about 4 dB. With a source impedance of 1000 Ω , a typical reading of noise output expressed as an equivalent input signal at 1 kHz is about 2.5 μ V.

The noise generated in the amplifier segment following the volume control is also of interest because it degrades the signal-to-noise ratio at low volume settings. In this case, the amplifier frequency response is basically flat (unless modified by tone controls) and data are measured for a frequency band limited by a "C" filter (equivalent bandwidth of about 13 kHz). Fig. 7 shows typical noise output voltages of an amplifier segment operating at a gain of 40 dB as a function of input termination resistance. As a point of reference, the theoretical noise voltage generated by a 1000- Ω resistance over this bandwidth is about 0.5 μ V (50- μ V output).

The input impedance of each amplifier segment is typically about 90,000 Ω . This value is considerably higher than needed for the tape-head application, but is useful for amplifier segments that follow volume or tone controls and adds versatility for other types of signal sources. The output impedance is approximately equal to the collector load of 1000 Ω when the amplifier is in the open-loop condition or when maximum output-swing capabilities are considered. In normal use, the small-signal output impedance is reduced by the feedback; for example, it is typically about 140 Ω for an amplifier operating at a voltage gain of 40 dB (feedback of 17 dB).

The electrical separation of amplifier segments is expected to be more than adequate for most audio applications. Even in the preamplifier described above, in which segments are cascaded, a 1-kHz signal in one channel induces a signal in the other channel that is lower in amplitude by more than 40 dB at maximum gain (about 65 dB voltage gain); under these conditions the undesired signal is submerged in noise. The fact that the separation is not detectably worse at 10 kHz indicates that capacitance coupling is no factor at audio frequencies.

The capacitances between leads of the package and external wiring are undoubtedly much greater than the internal capacitance couplings, but are significant only at frequencies approaching 1 MHz. Other sources of coupling, however, such as through common ground and positive-supply line resistances, can be high enough to warrant measurement (ratios as low as 20 dB when the over-all gain is of the order of 80 dB at low frequencies). Coupling of this degree can be

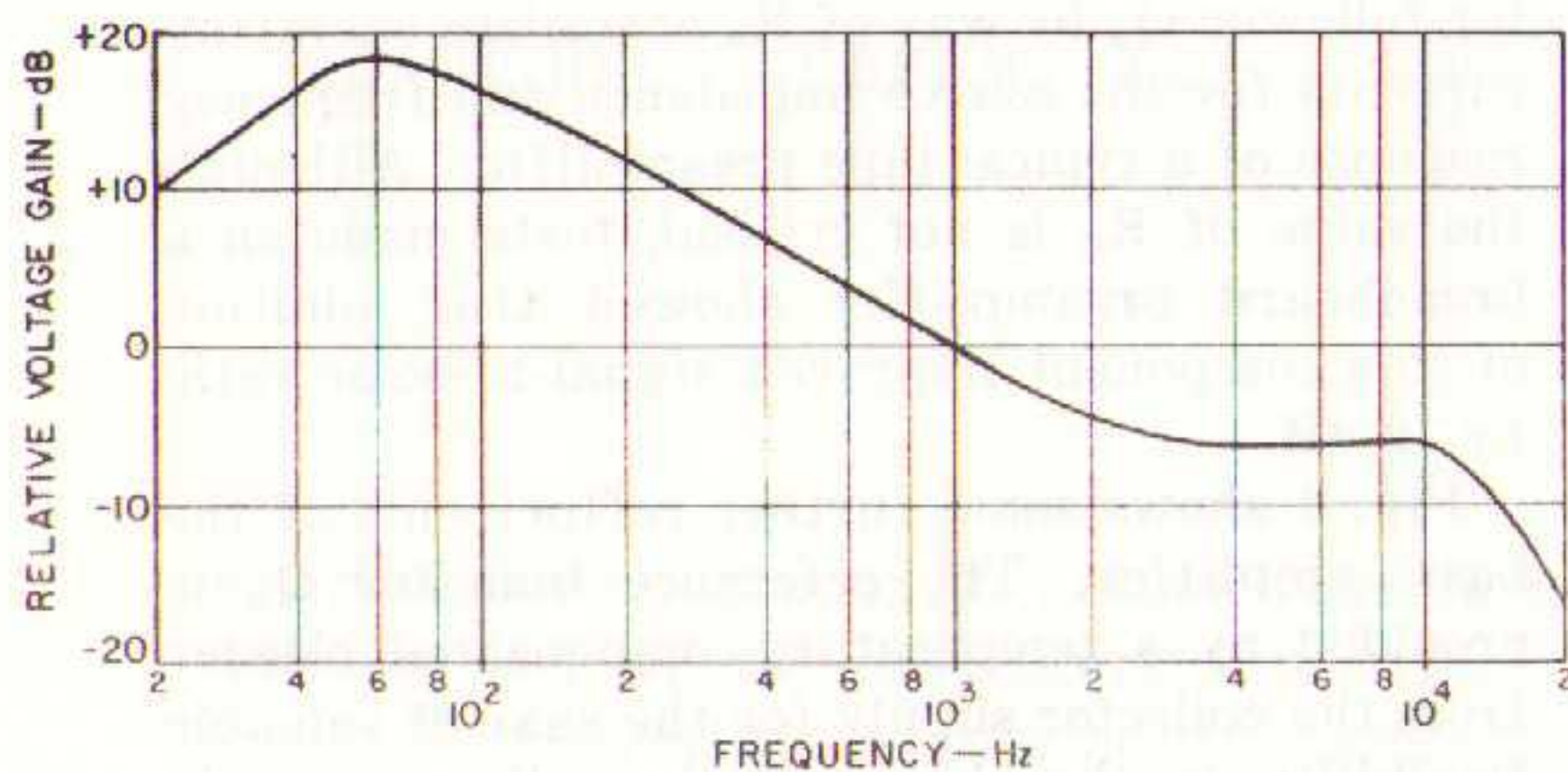
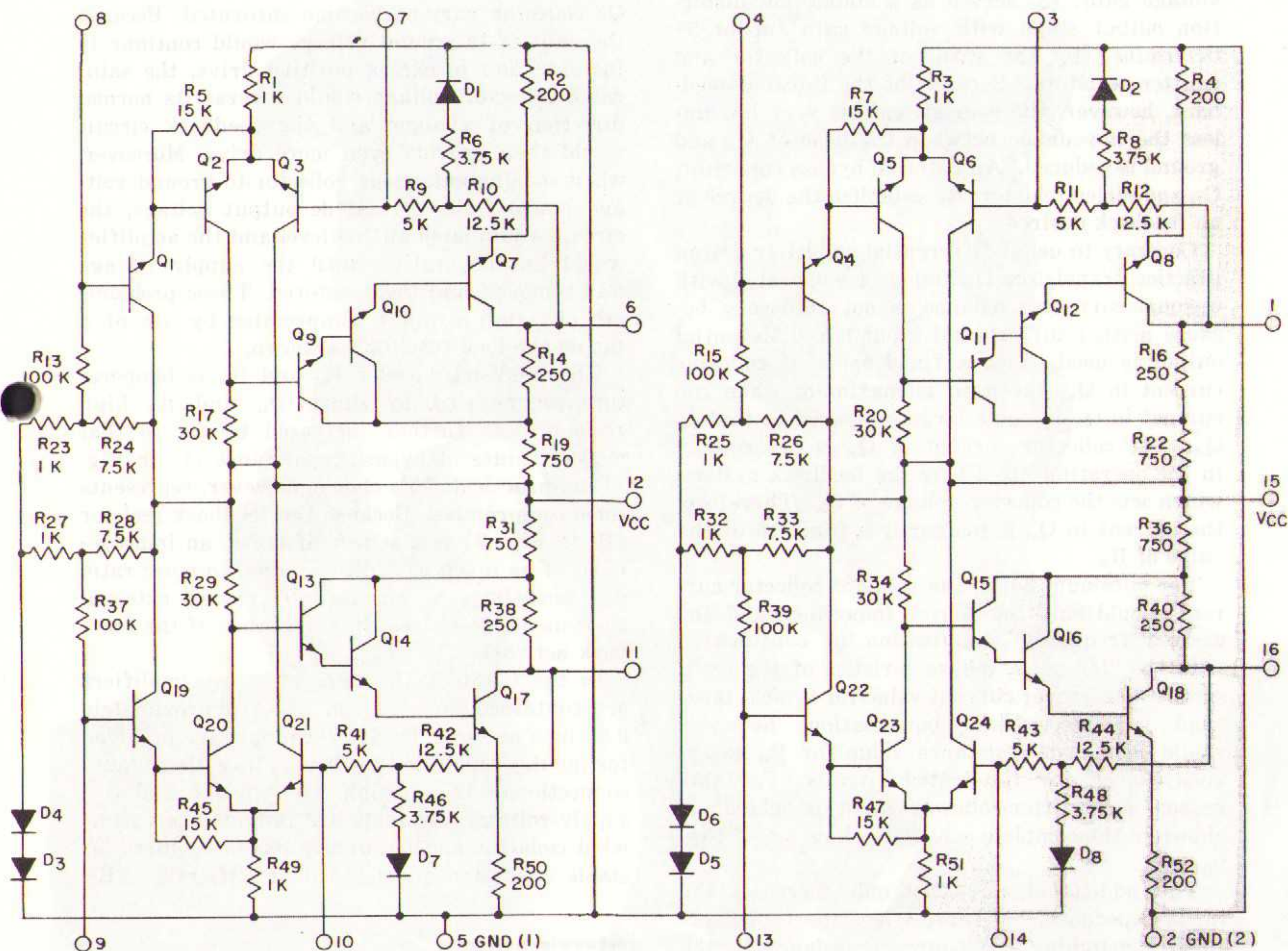
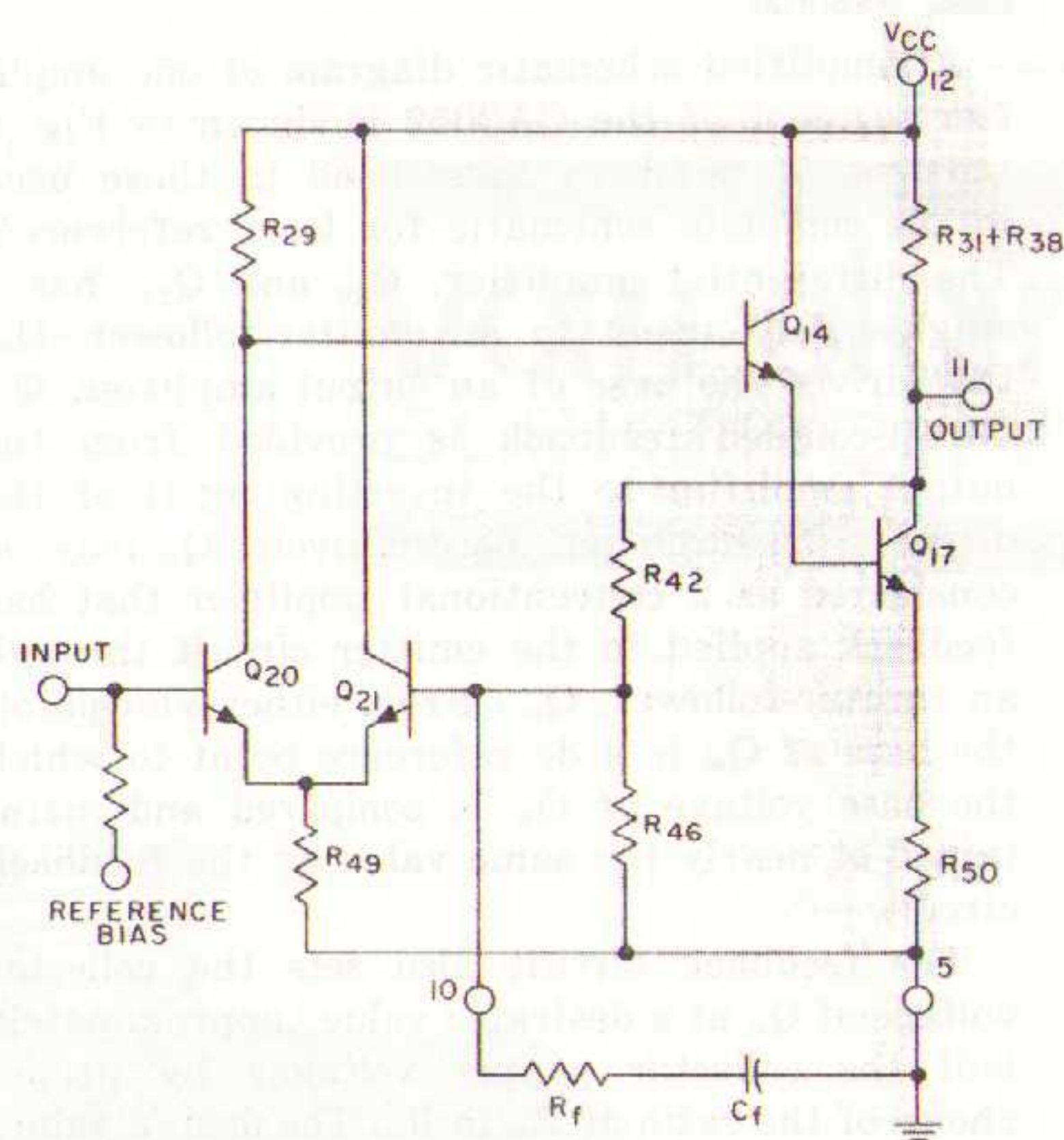
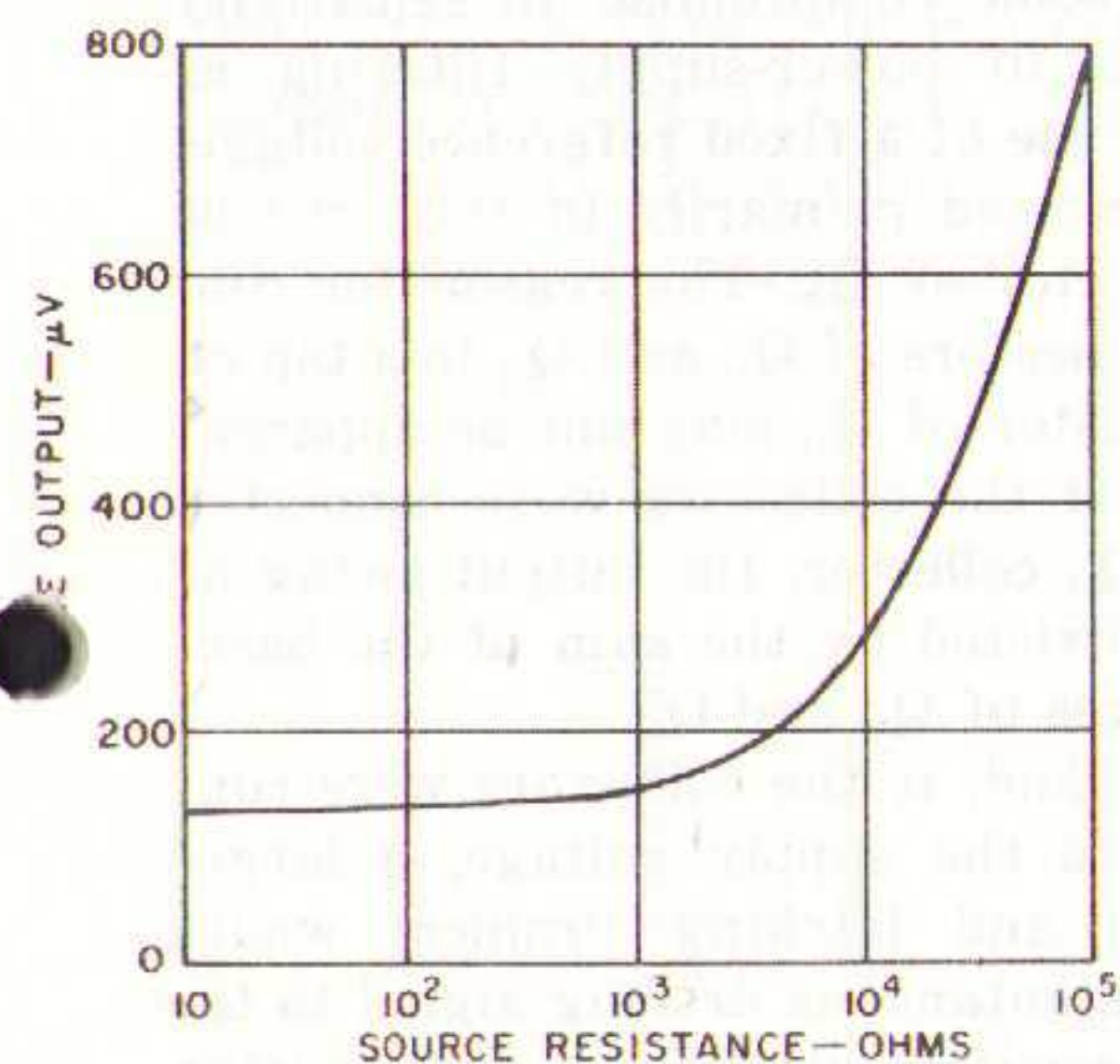


Fig. 6. (above) Frequency response of complete preamplifier equalized for tape playback and passed through "C" filter for noise measurements

Fig. 7. (left) Typical noise output of one amplifier segment operated at a voltage gain of 40 dB over the audio band (about 13 kHz) as restricted by the "C" noise-weighting filter

Fig. 8. (right) Simplified diagram of one CA3052 amplifier segment



NOTE: ALL RESISTOR VALUES ARE IN OHMS

Fig. 9. Complete schematic diagram of CA3052

prevented in external wiring of complete amplifiers by avoiding common-ground resistances. The use of two positive-voltage supply points in the CA3052 facilitates isolation of power-supply impedances.

Basic amplifier

A simplified schematic diagram of one amplifier segment of the CA3052 is shown in Fig. 8 (component numbers correspond to those used in the complete schematic for later reference). The differential amplifier, Q_{20} and Q_{21} , has a single-ended output to an emitter-follower, Q_{14} , that drives the base of an output amplifier, Q_{17} . Direct-coupled feedback is provided from the output amplifier to the inverting input of the differential amplifier. Alternatively, Q_{20} may be considered as a conventional amplifier that has feedback applied to the emitter circuit through an emitter-follower, Q_{21} . From either viewpoint, the base of Q_{20} is a dc reference point to which the base voltage of Q_{21} is compared and maintained at nearly the same value by the feedback circuit.

The feedback circuit also sets the collector voltage of Q_{17} at a desirable value (approximately half the collector supply voltage) by proper choice of the ratio of R_{42} to R_{46} . The design values are selected so that Q_{20} supplies most of the voltage gain; Q_{17} serves as a stable, low-distortion output stage with voltage gain (about 5) determined by the ratio of the collector and emitter resistors. Because of the internal feedback, however, the over-all gain is very low unless the impedance between the base of Q_{21} and ground is reduced. An external bypass capacitor, C_r , and series resistor, R_r , establish the degree of ac feedback desired.

Contrary to usual differential-amplifier design practice, transistors Q_{20} and Q_{21} are operated with unequal currents; balance is not necessary because neither differential input nor differential output is used. With a fixed value of collector current in Q_{20} , the gain is maximum when the current in Q_{21} is made large compared to that in Q_{20} . The collector current of Q_{20} is determined in the operating circuit by the feedback system, which sets the collector voltage of Q_{20} . Therefore, the current in Q_{20} is primarily a function of the value of R_{20} .

For minimum noise, the value of collector current should suit the source impedance and the desired frequency equalization in conjunction with the "1/f noise" characteristics of the transistor. The proper current value for typical tape-head impedances and equalization, however, would require a resistance value for R_{20} excessively large for integrated circuits. For this reason, an emitter-follower input is added, as shown in the complete schematic diagram of Fig. 9.

This additional stage not only increases the input impedance, but serves the important function of matching the source impedance to the differential-amplifier transistors for low noise. The bleeding of some current through the emit-

ter-follower Q_{10} by way of R_{45} optimizes operating currents for the source impedance and frequency response of a typical tape preamplifier. Although the value of R_{45} is not critical, tests made on a breadboard preamplifier showed that addition of this component improved signal-to-noise ratio by 12 dB.

Fig. 9 shows some further refinements of the basic amplifier. The reference bias for Q_{20} is provided by a temperature-compensated bleeder from the collector supply for the sake of valuable flexibility in the choice of the collector supply voltage despite some compromise in sensitivity to imperfections in power-supply filtering as compared to the use of a fixed reference voltage. Transistor Q_{13} is used primarily to shift the dc level at the collector of Q_{20} . The reason for connection of the collectors of Q_{13} and Q_{14} to a tap on the collector resistor of Q_{17} may not be apparent at first glance. If the collectors were connected directly to the Q_{17} collector, the output swing of Q_{17} would be restricted by the sum of the base-to-emitter voltages of Q_{17} and Q_{14} .

On the other hand, if the collectors were connected directly to the supply voltage, a large-signal inversion and latching problem would occur. As the instantaneous driving signal to Q_{17} swings in the positive direction, the emitter current could be forced to increase even after the Q_{17} collector current became saturated. Because the emitter-to-ground voltage would continue to increase due to excess positive drive, the saturated collector voltage would reverse its normal direction of change, and the feedback circuit would then provide even more drive. Moreover, when the instantaneous collector-to-ground voltage reached the normal dc output voltage, the circuit would latch at this level and the amplifier would be inoperative until the supply voltage was removed and then restored. These problems are obviated without compromise by use of a tap on the load resistor, as shown.

The feedback bleeder R_{42} and R_{46} is temperature-compensated by diode D_1 , and its high resistance is further increased by R_{41} so that modest values of bypass capacitance (C_r in Fig. 1) are practical. This choice, however, represents some compromise. Because the feedback resistor (R_r in Fig. 8) is a source of noise, an improvement of as much as 3 dB in signal-to-noise ratio can sometimes be obtained by use of external components to reduce the impedance of the feedback network.

In the CA3052, the four identical amplifiers are contained on a silicon pellet approximately 0.07 inch square. Twelve terminals are provided for input, output, and feedback (inverting input) connections. In addition, two ground and two supply-voltage terminals are provided to permit what isolation and decoupling may be required for stable operation of cascaded amplifiers. **EEE**

Acknowledgement

The author thanks H. W. McCord for his contributions to the low-noise design, R. G. Vano for measurements, and the many others who were responsible for layout, fabrication, and the like.

U ZET UW HALFGELEIDERS TOCH ZEKER GEEN IJSZAK OP ???

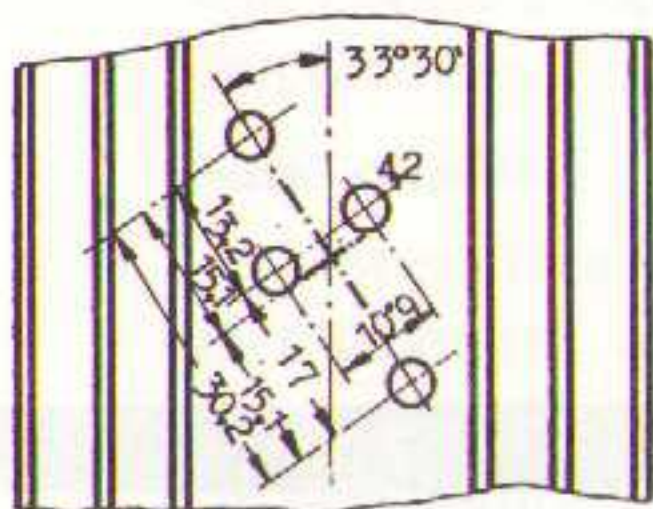
DAARVOOR HEBBEN WIJ EEN RUIM ASSORTIMENT KOELPLATEN !!!

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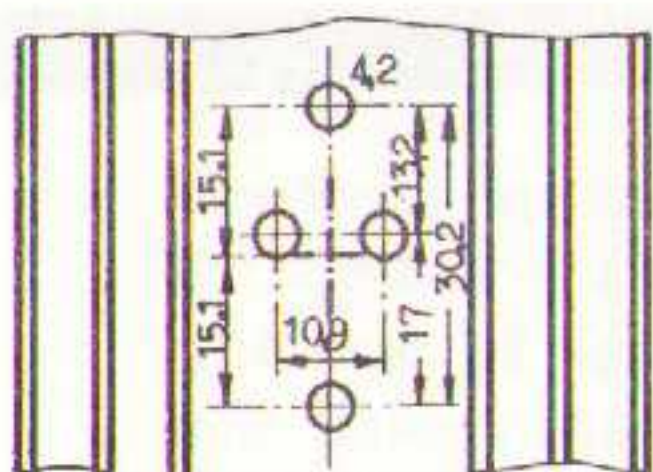
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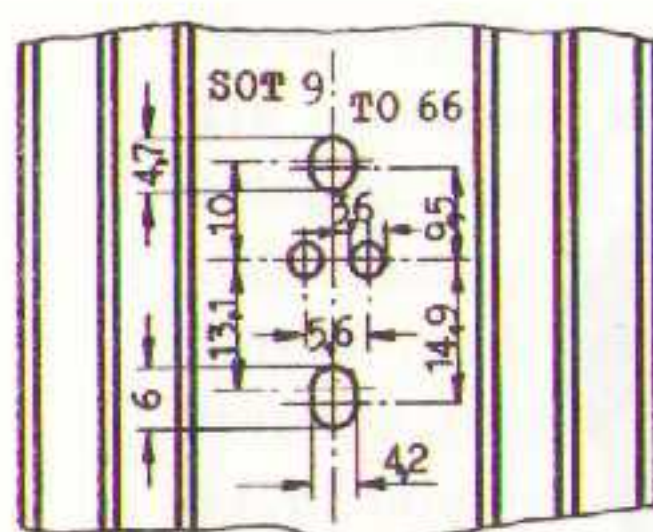
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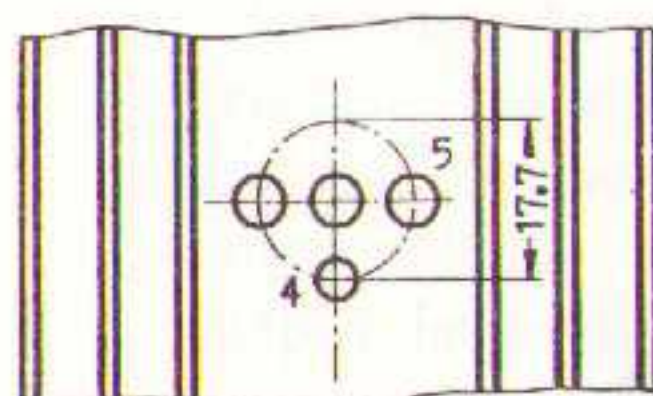
T0-3 bij
37,5 mm type



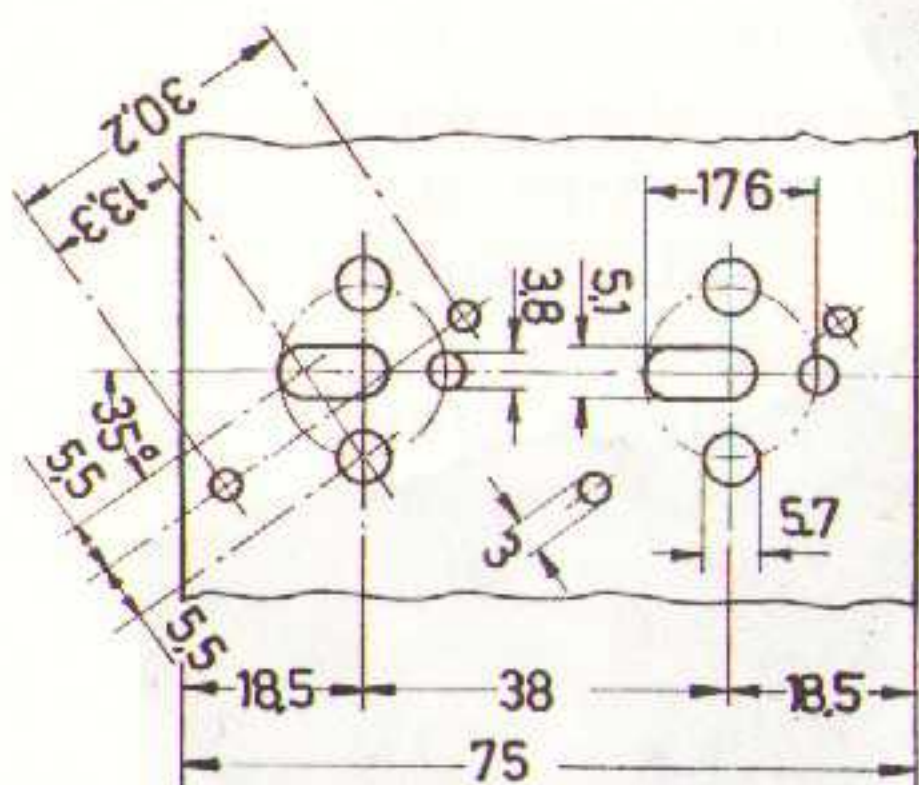
TO-3



SOT-9 + TO-66



TO-36



2 x TO-3/TO-36

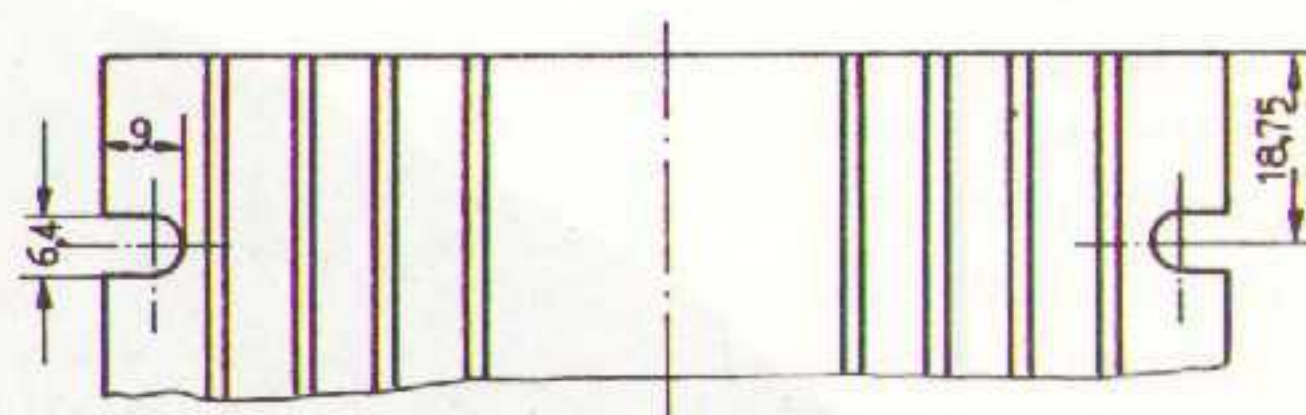
De standaard lengten van de meeste koelelementen bedragen 37,5 mm, 50 mm, 75 mm, 100 mm en 1000 mm.

Technische toelichting.

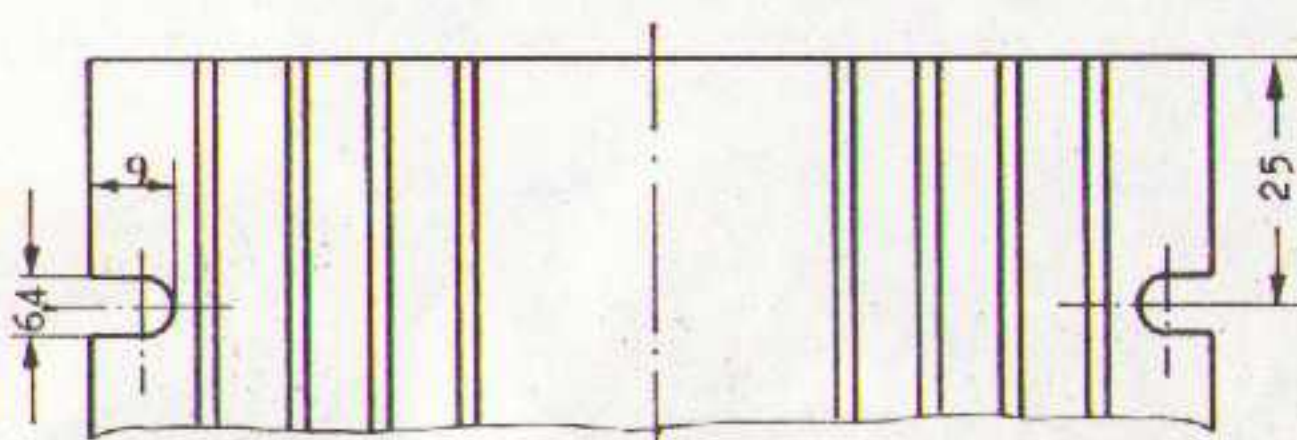
De door de halfgeleiderfabrikant gespecificeerde maximaal toelaatbare junctiontemperatuur T_j mag nimmer worden overschreden.

In iedere halfgeleider treedt bij in bedrijfstoestand een verbruiksvermogen op, dat de Tj doet toenemen. Teneinde deze temperatuur niet te overschreiden dient de warmte-afvoer door middel van een koelelement te worden vergroot. De totale warmteweerstand tussen junction, huis en omgeving wordt door de warmteweerstand van het koelelement gereduceerd. De grafieken tonen links het temperatuurverschil T tussen het montagevlak van de halfgeleider en de omgeving als functie van het vermogen PV. Uit deze grafieken kan de lengte van het koelelement bepaald worden. De grafieken zijn van toepassing voor zwart geëloxeerde uitvoering (SA) bij verticale montage; voor de blanke oppervlakken (AL) geldt een korrektiefaktor van +10 tot +15% en bij horizontale montage +15 tot +20%.

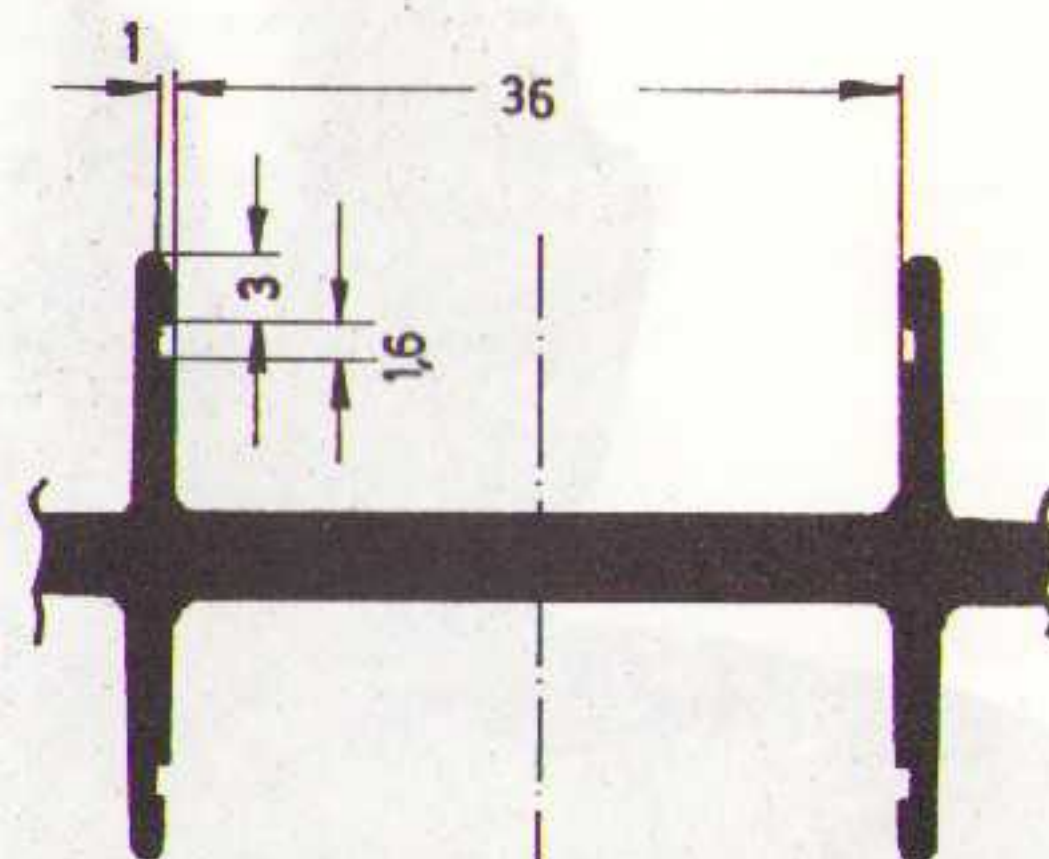
Alle typen gewalste profielen zijn van AlMgSi 05 F 22.



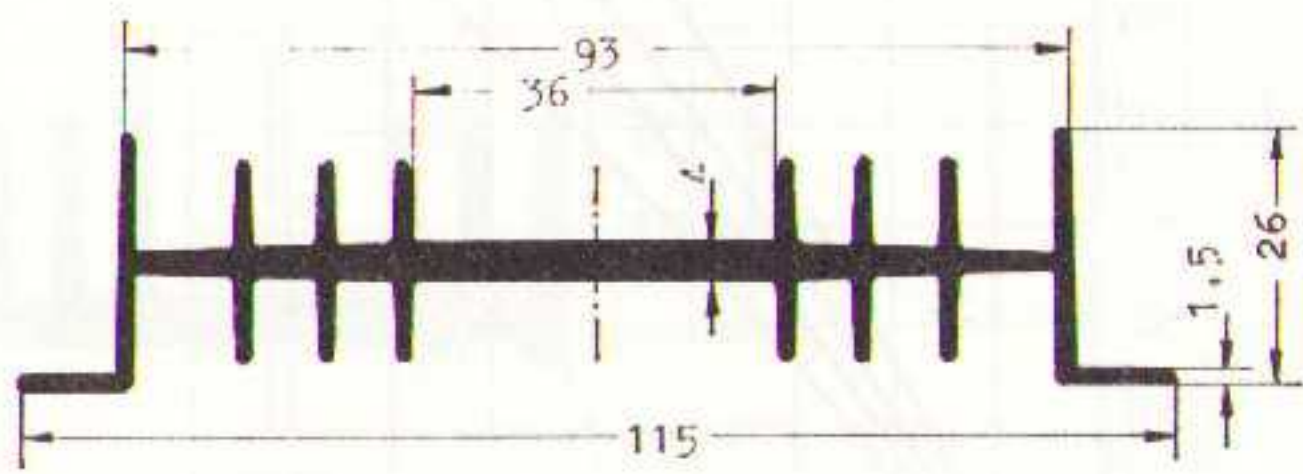
bevestigingsuitsparing, standaard



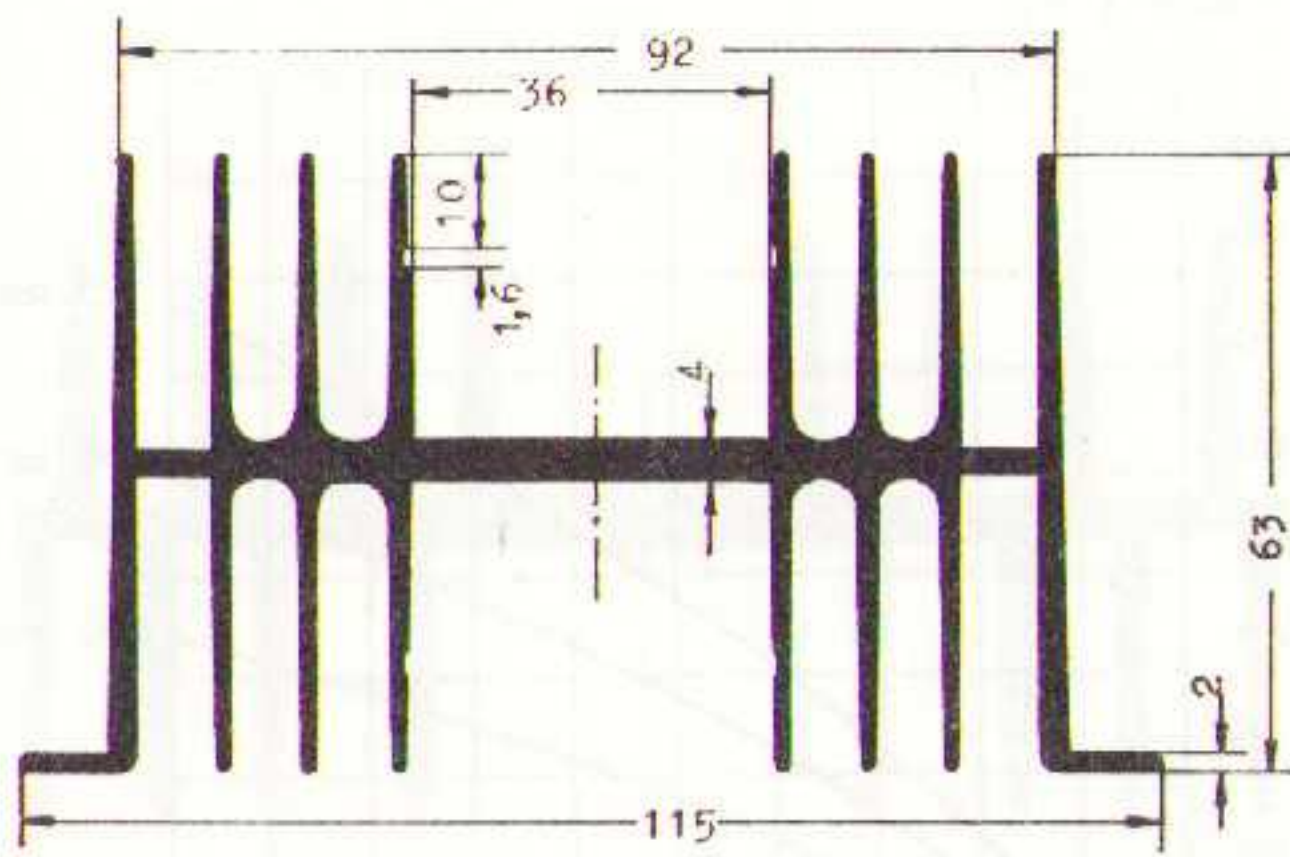
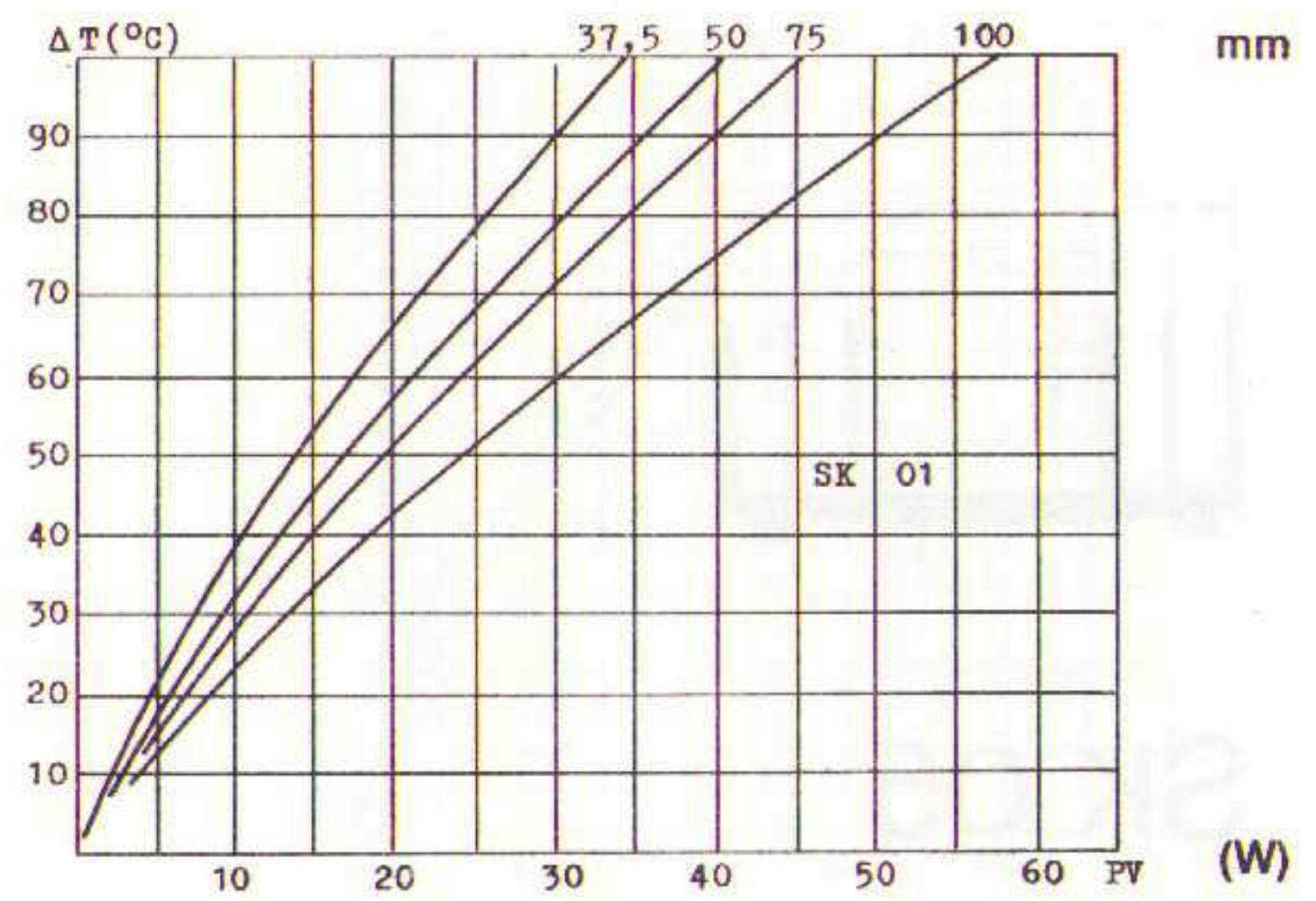
bevestigingsuitsparing, alléén bij
de 50 mm uitvoeringen.



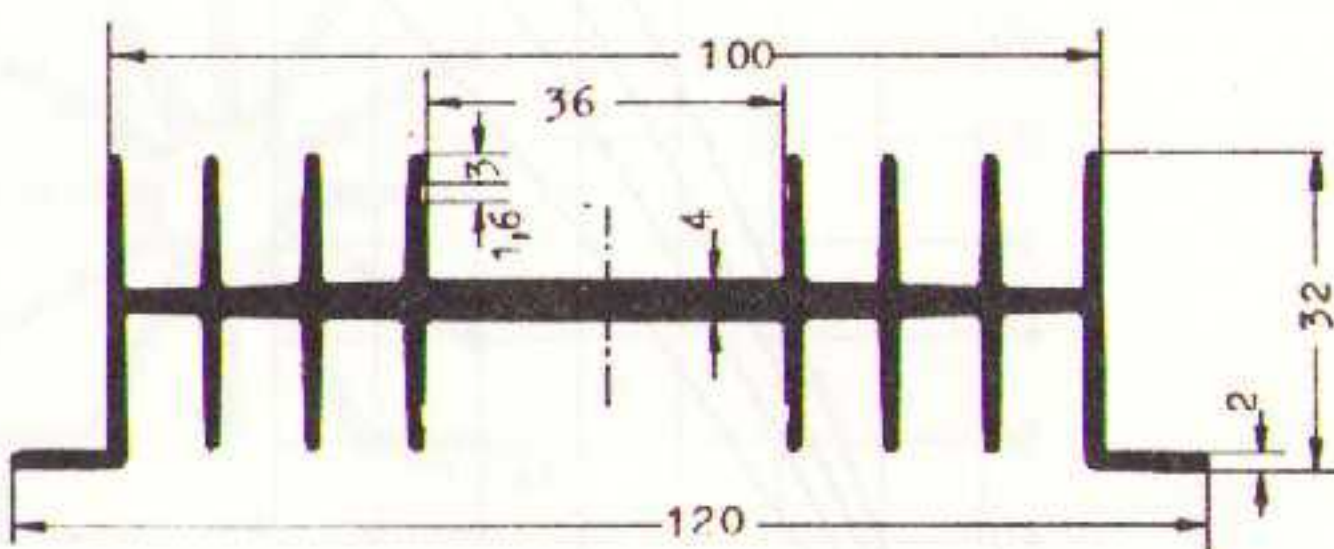
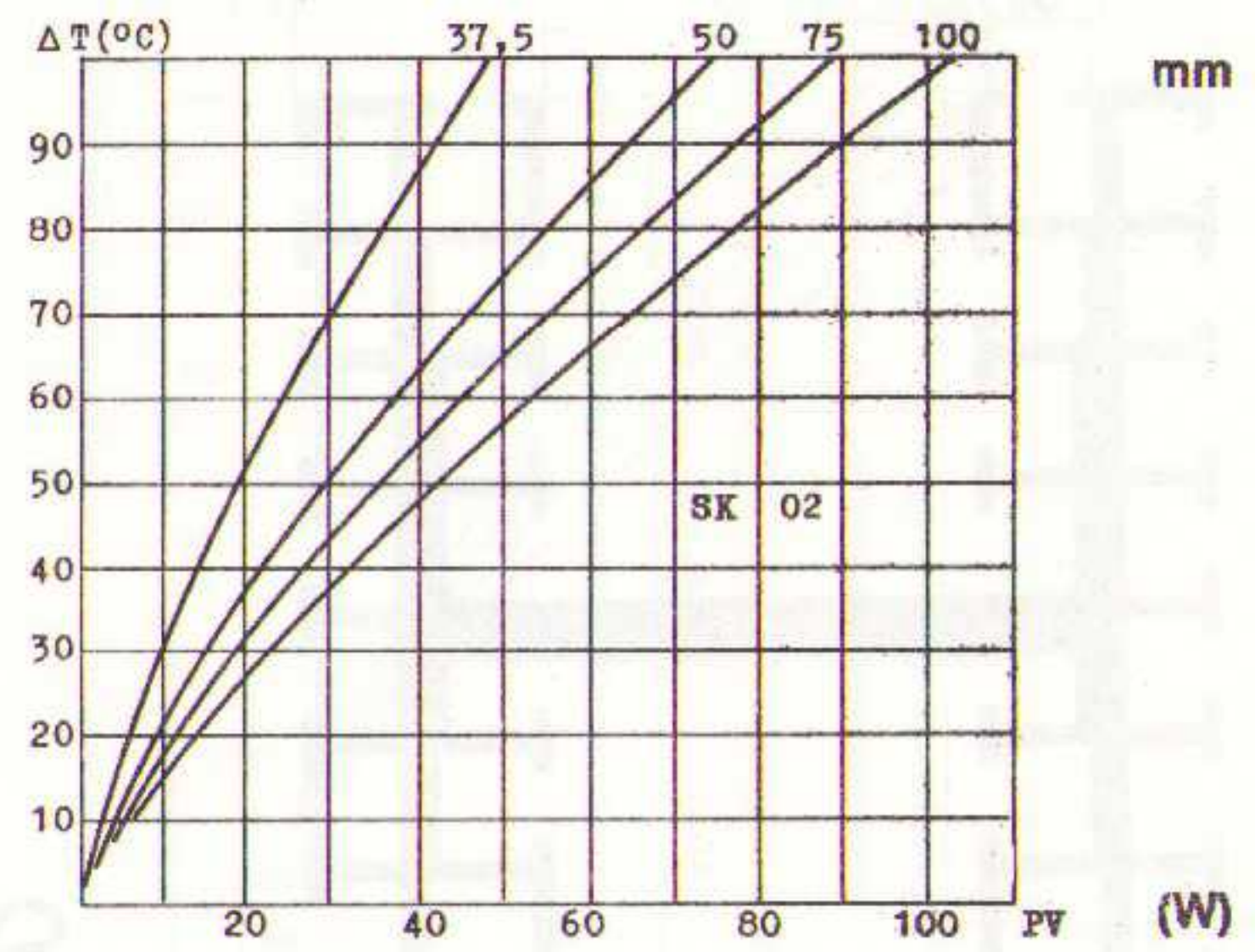
In de gleufjes kunnen bij de typen SK02, SK03, SK04, SK05, SK08 en SK 34 afdekplaatjes of gedrukte bedradingen worden geschoven.



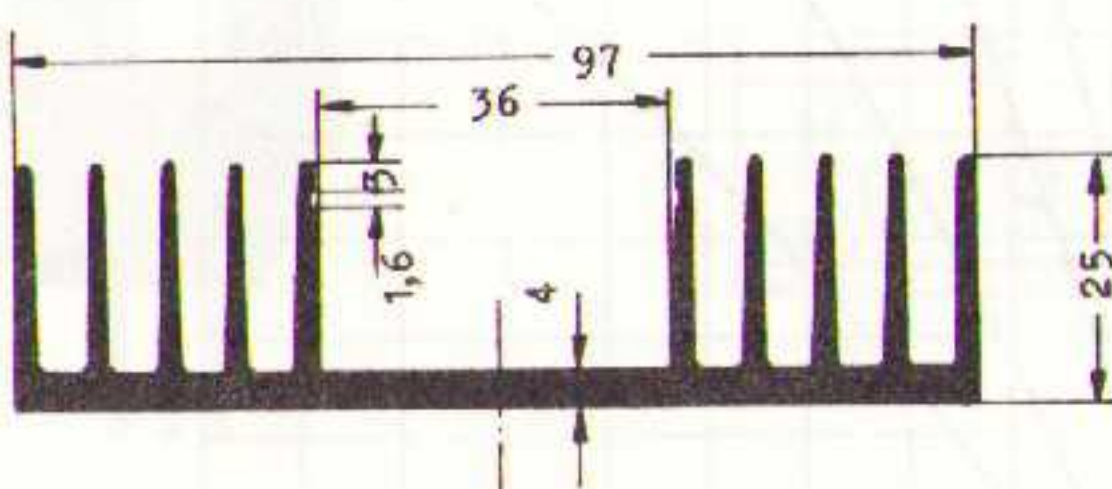
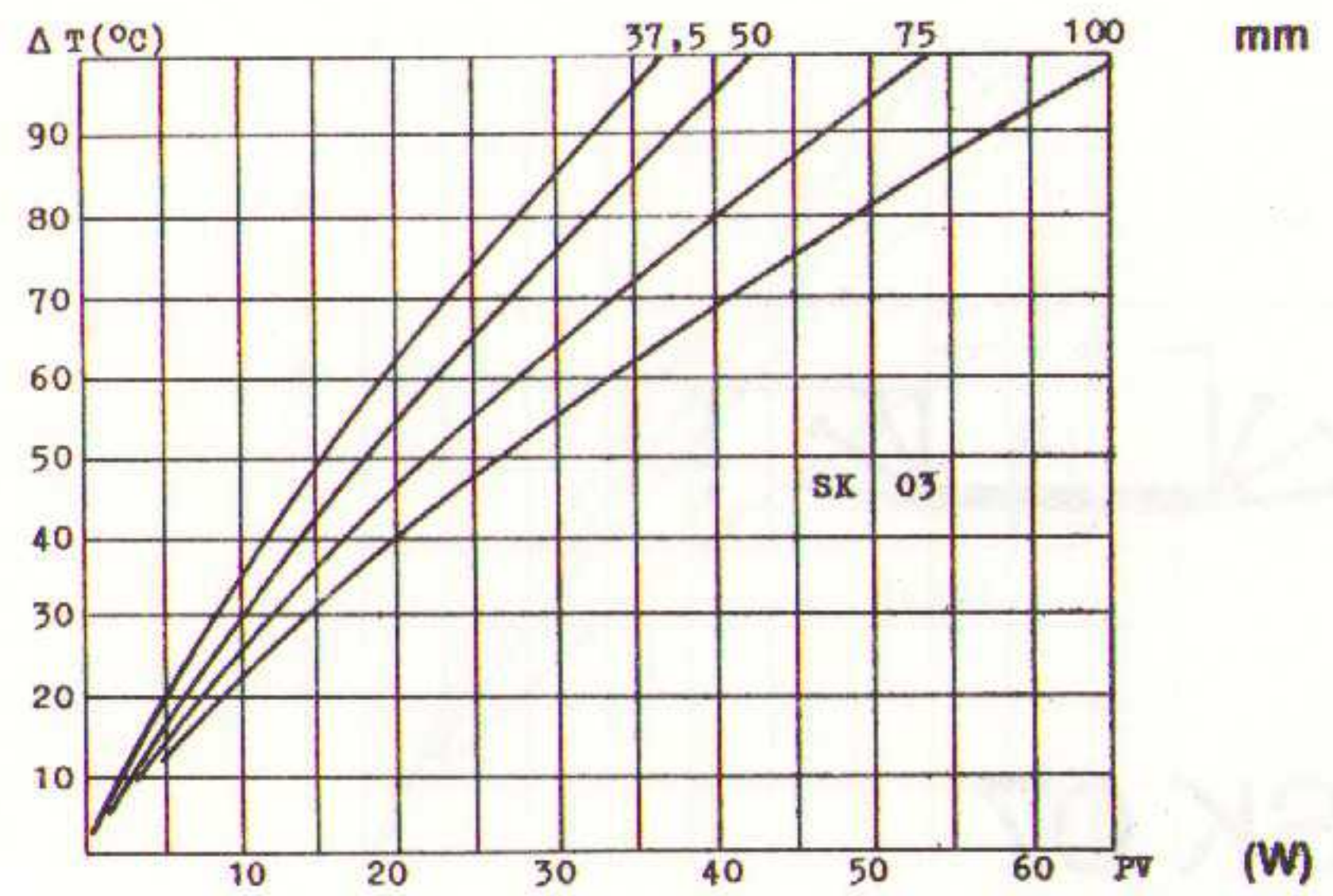
SK 01



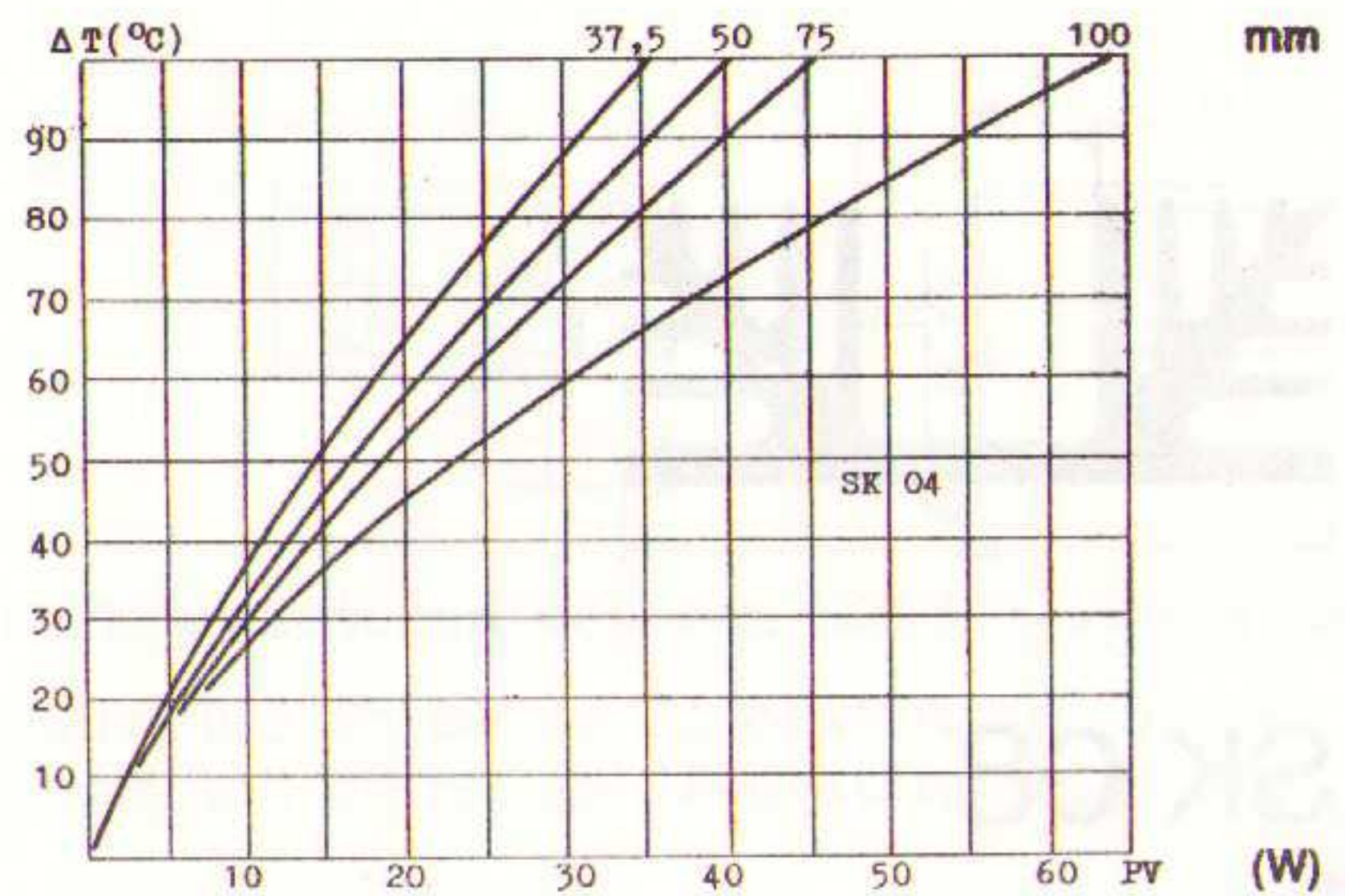
SK 02

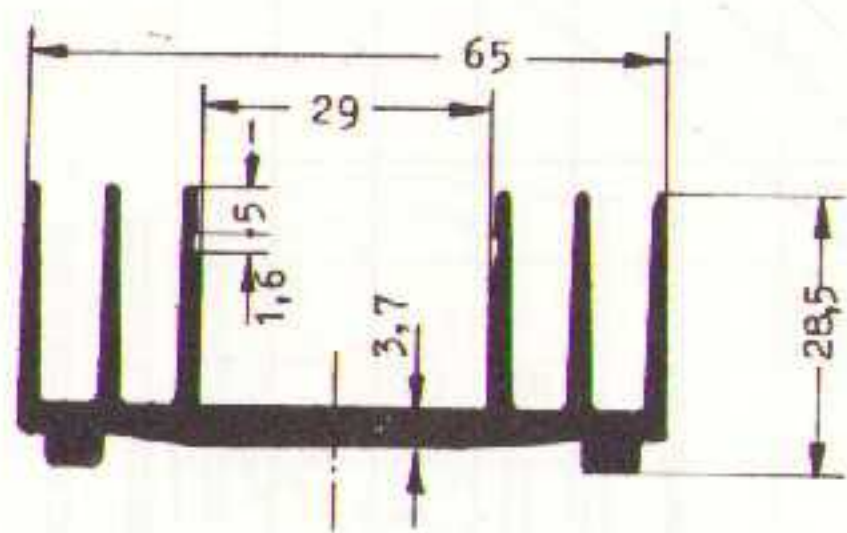


SK 03

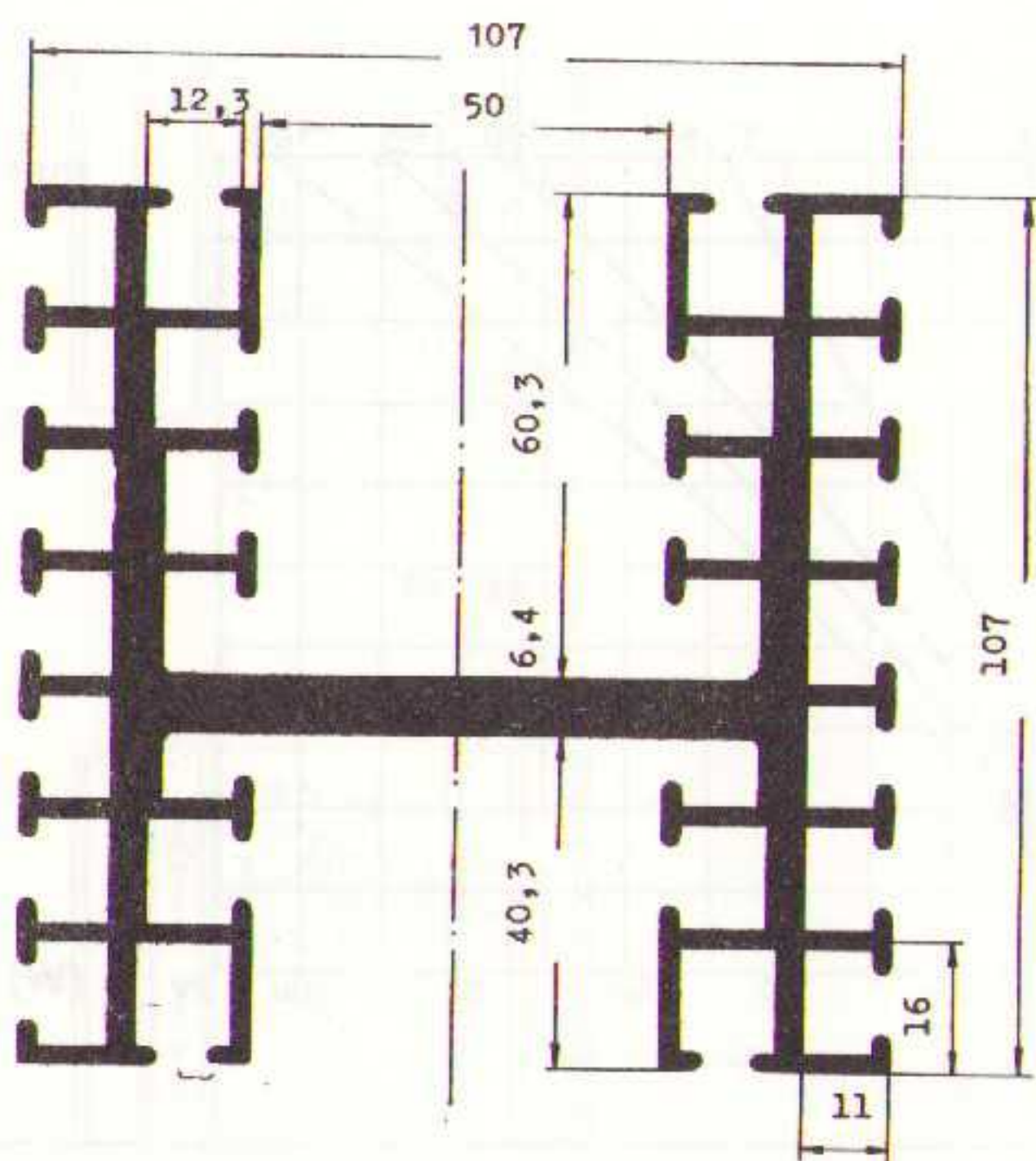
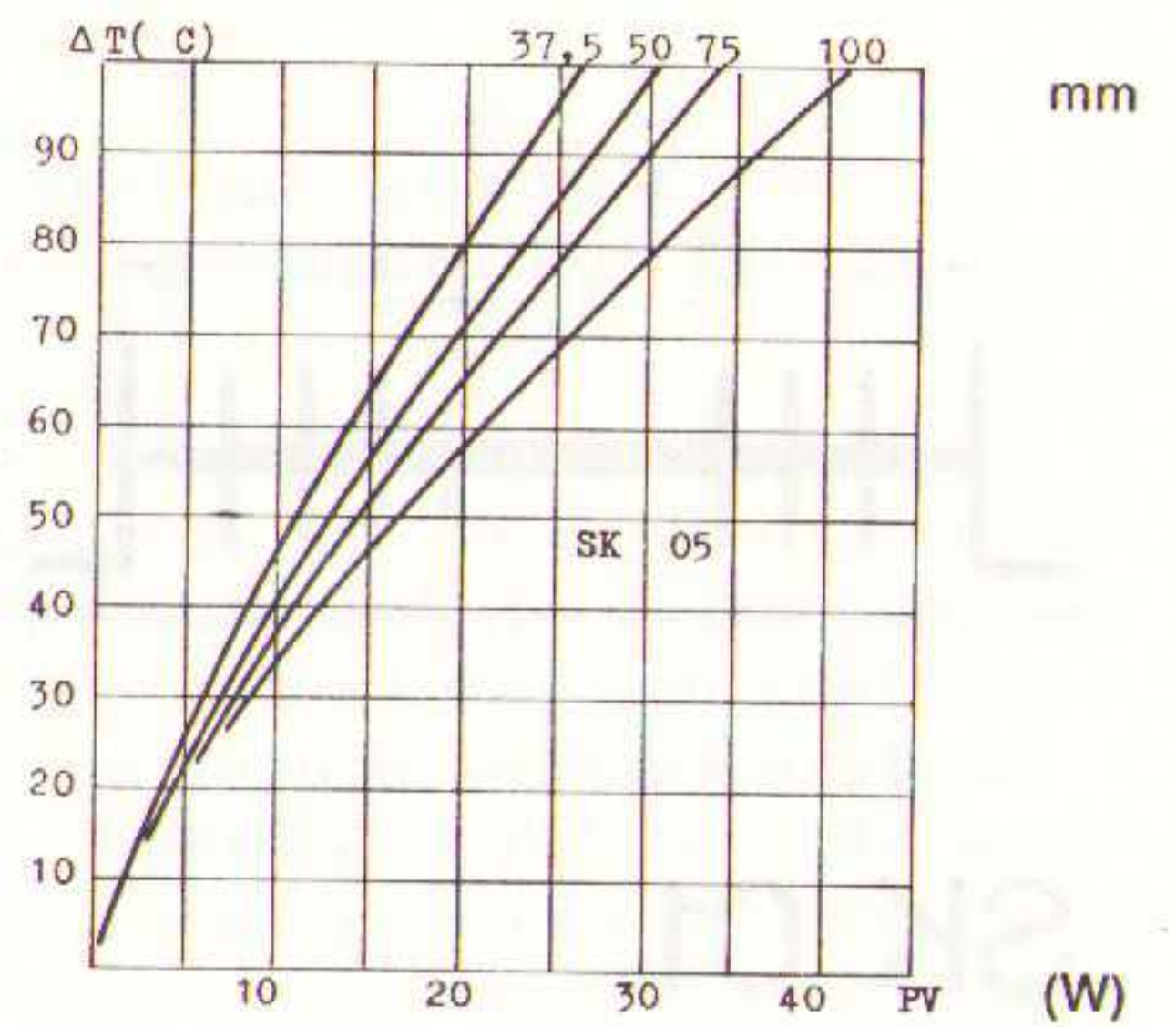


SK 04

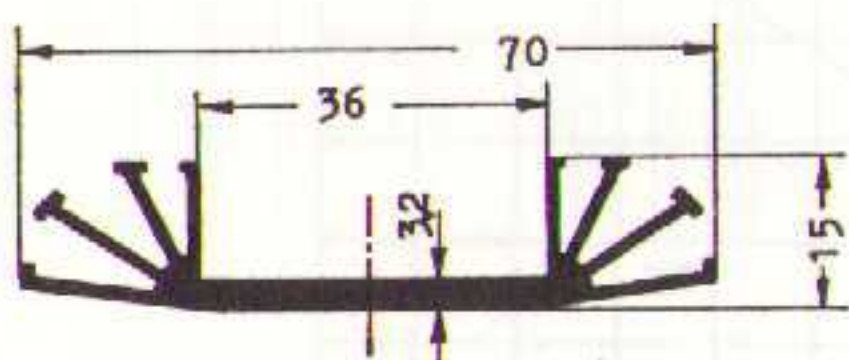
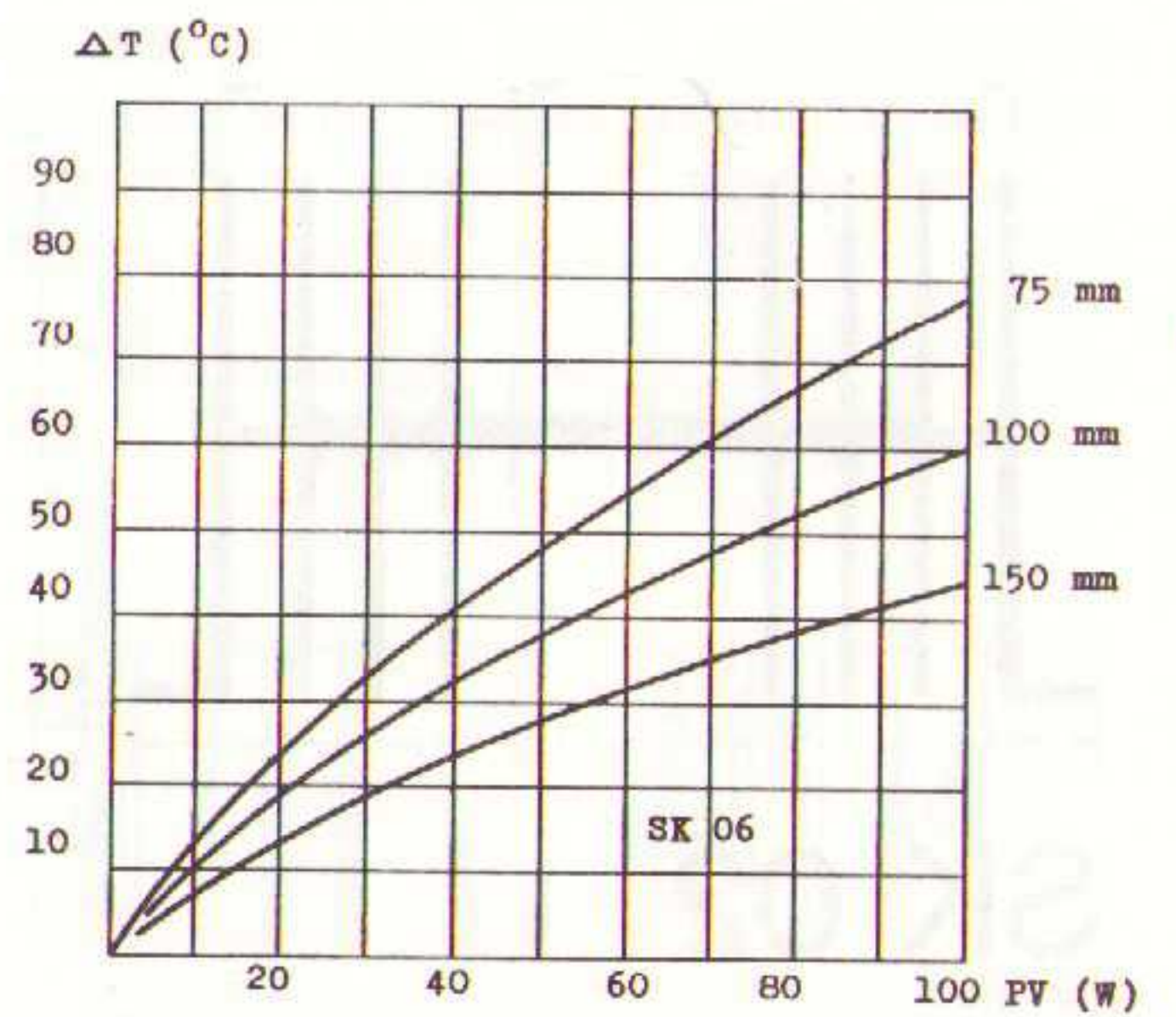




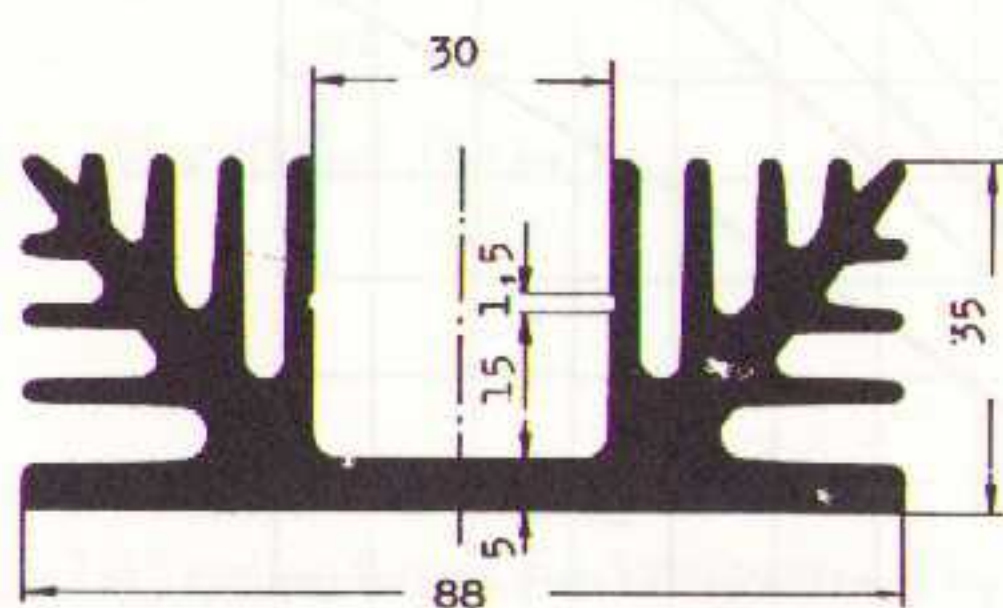
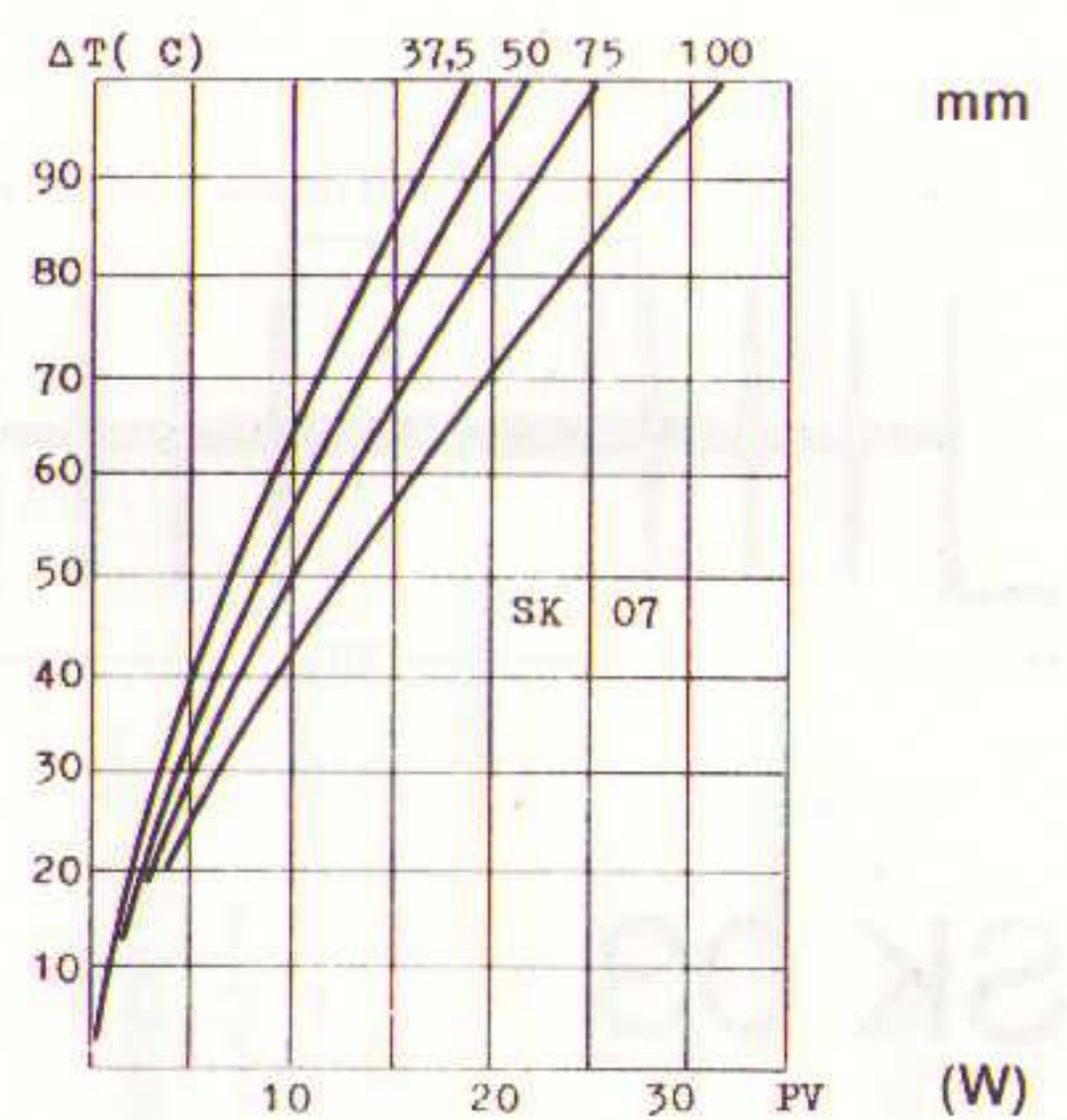
SK 05



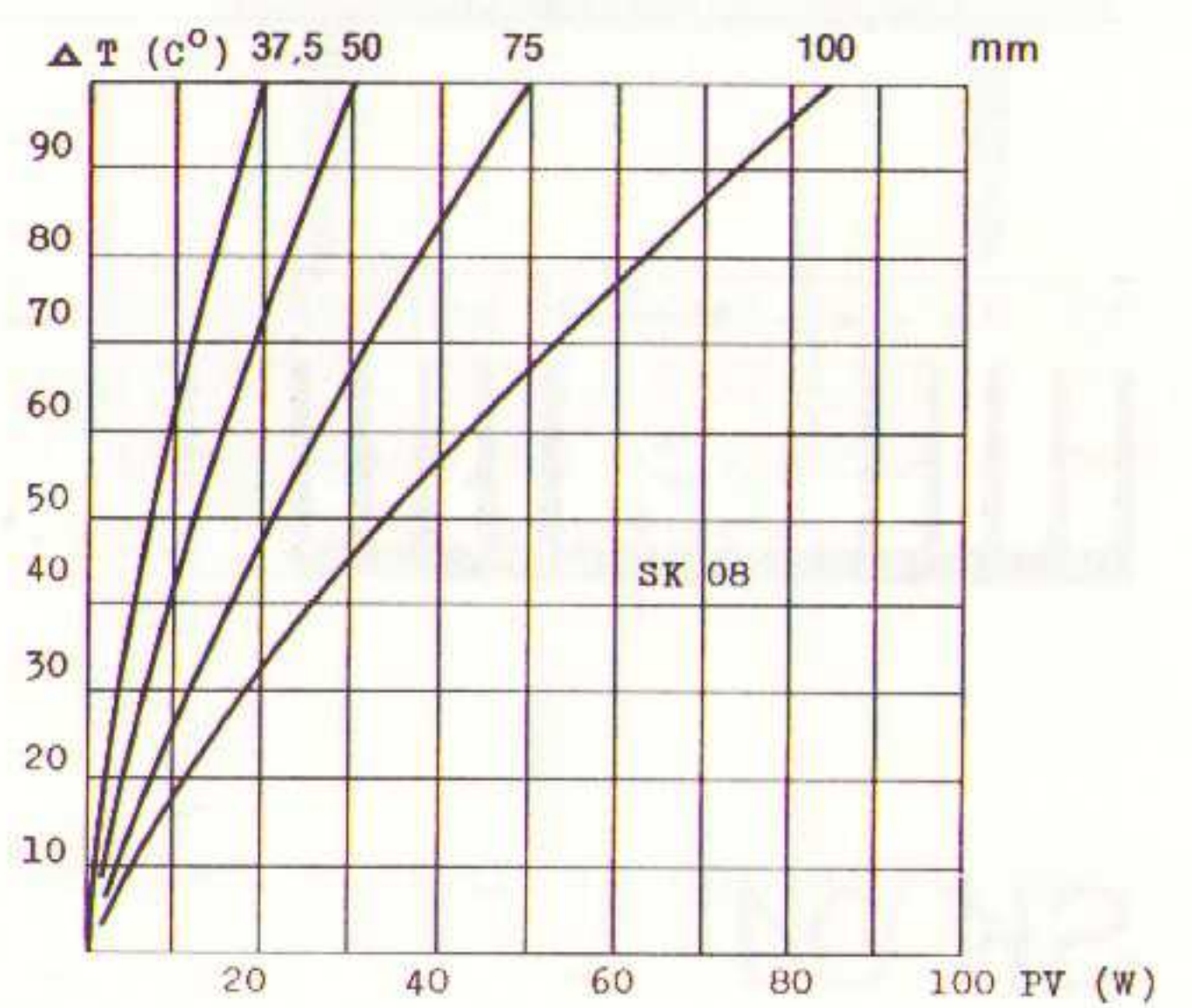
SK 06

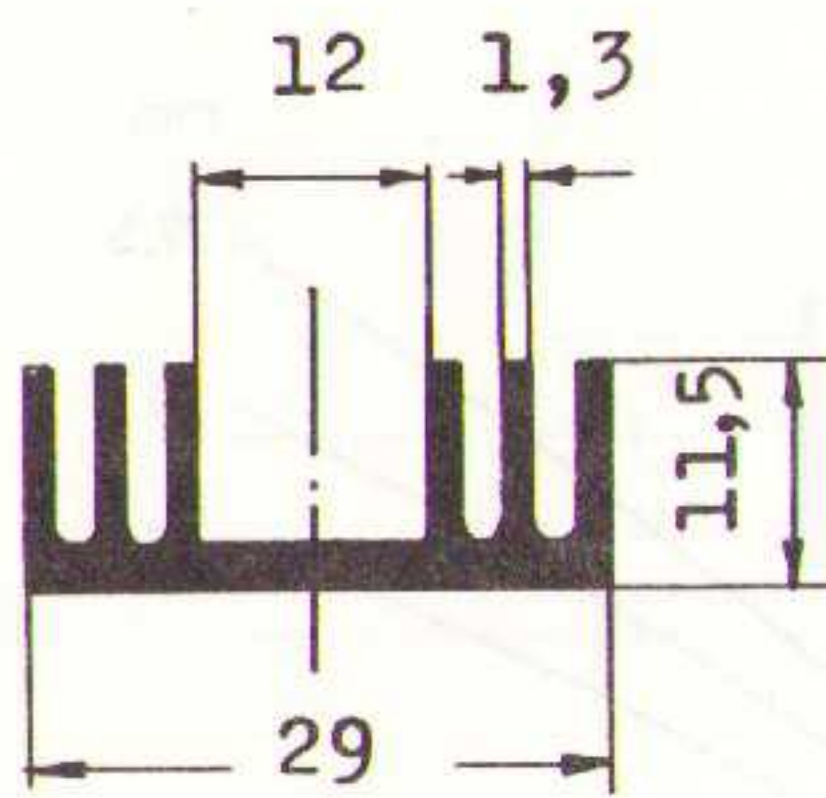


SK 07

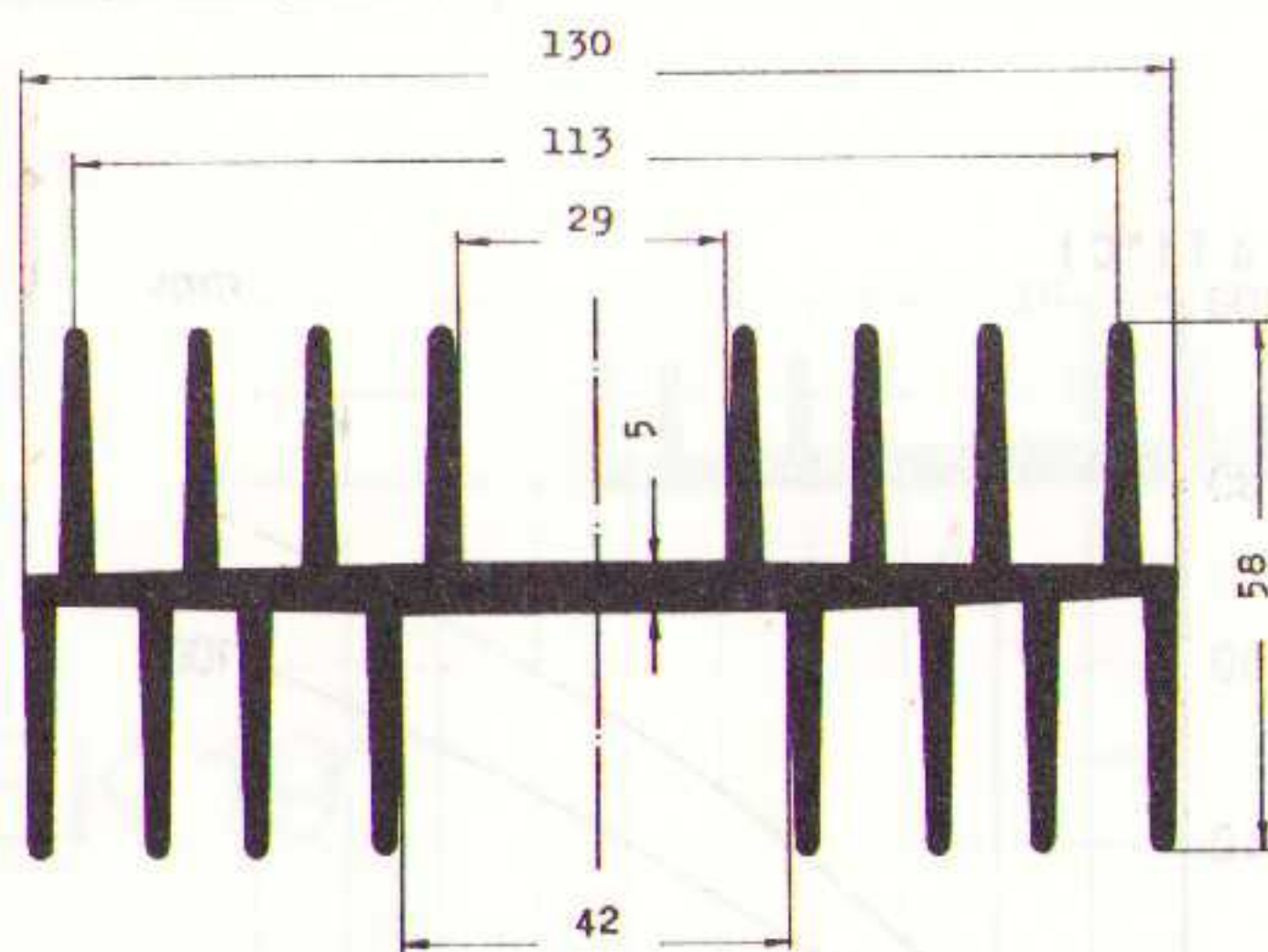
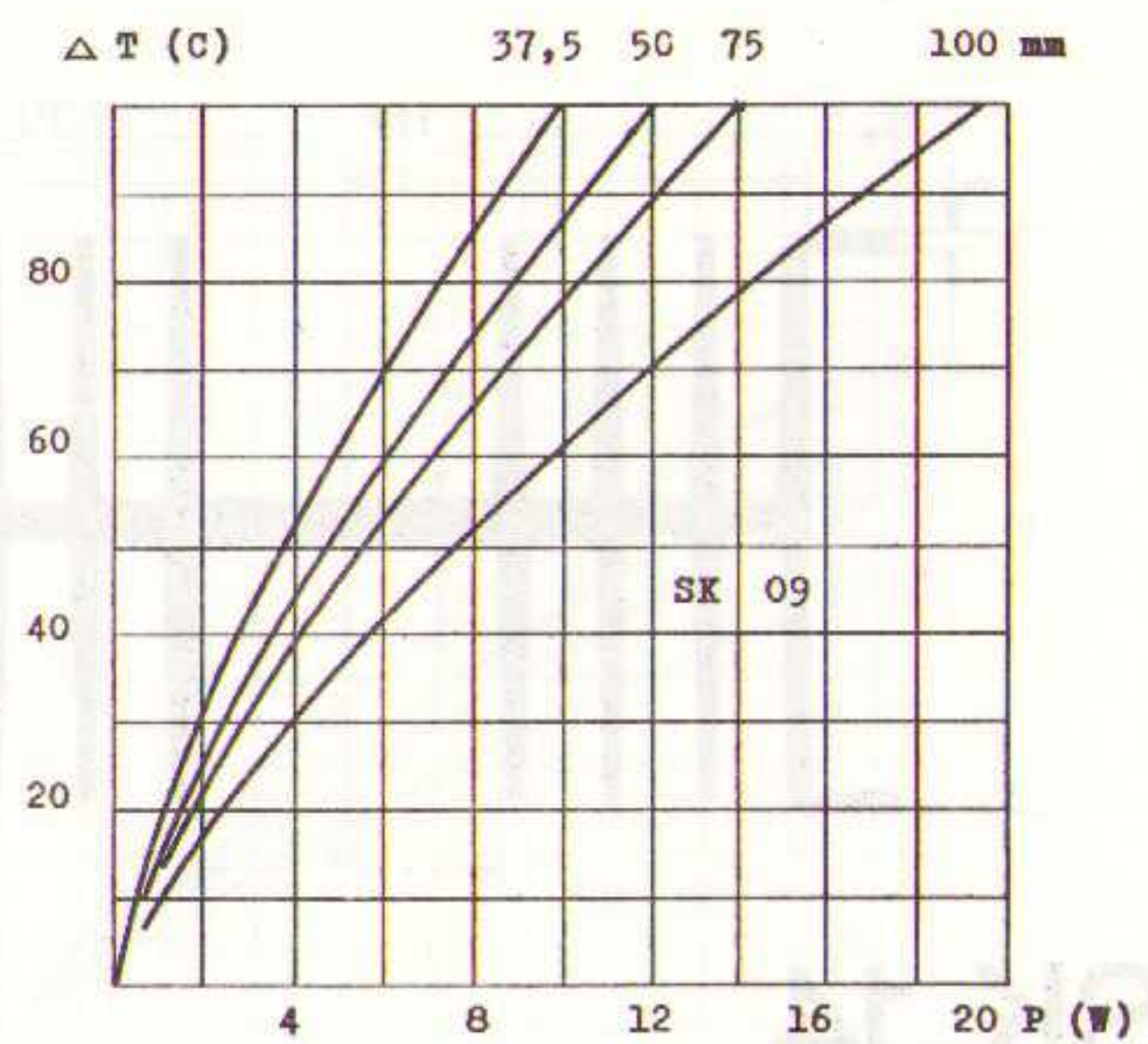
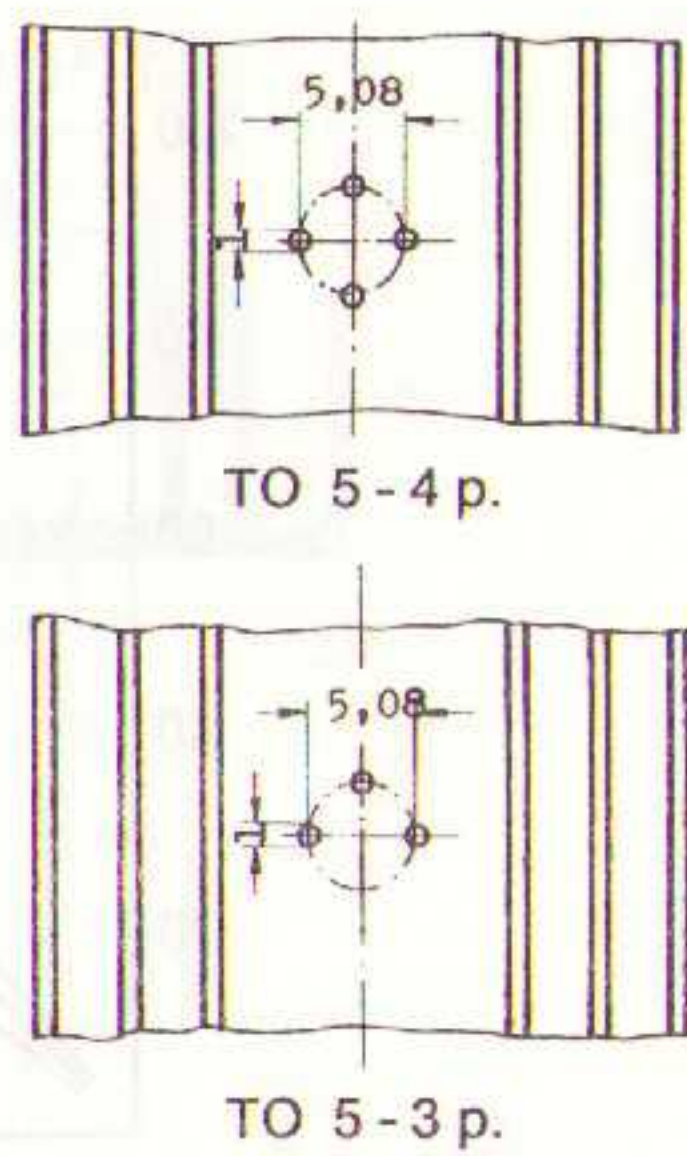


SK 08

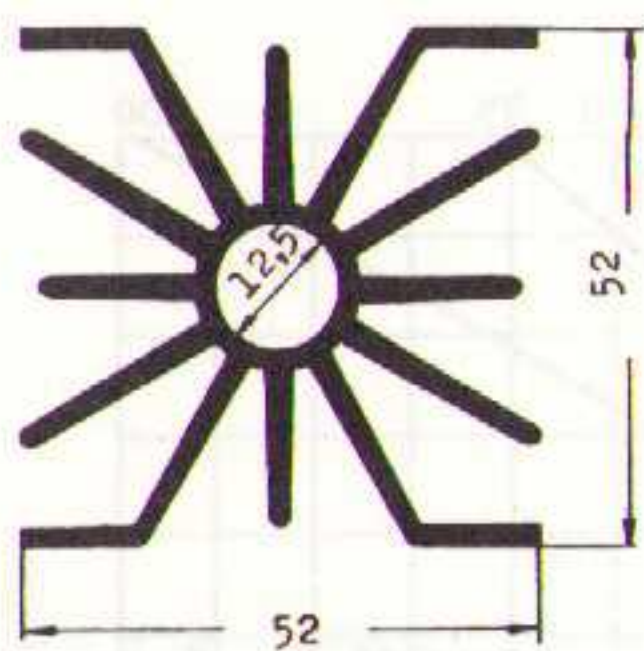
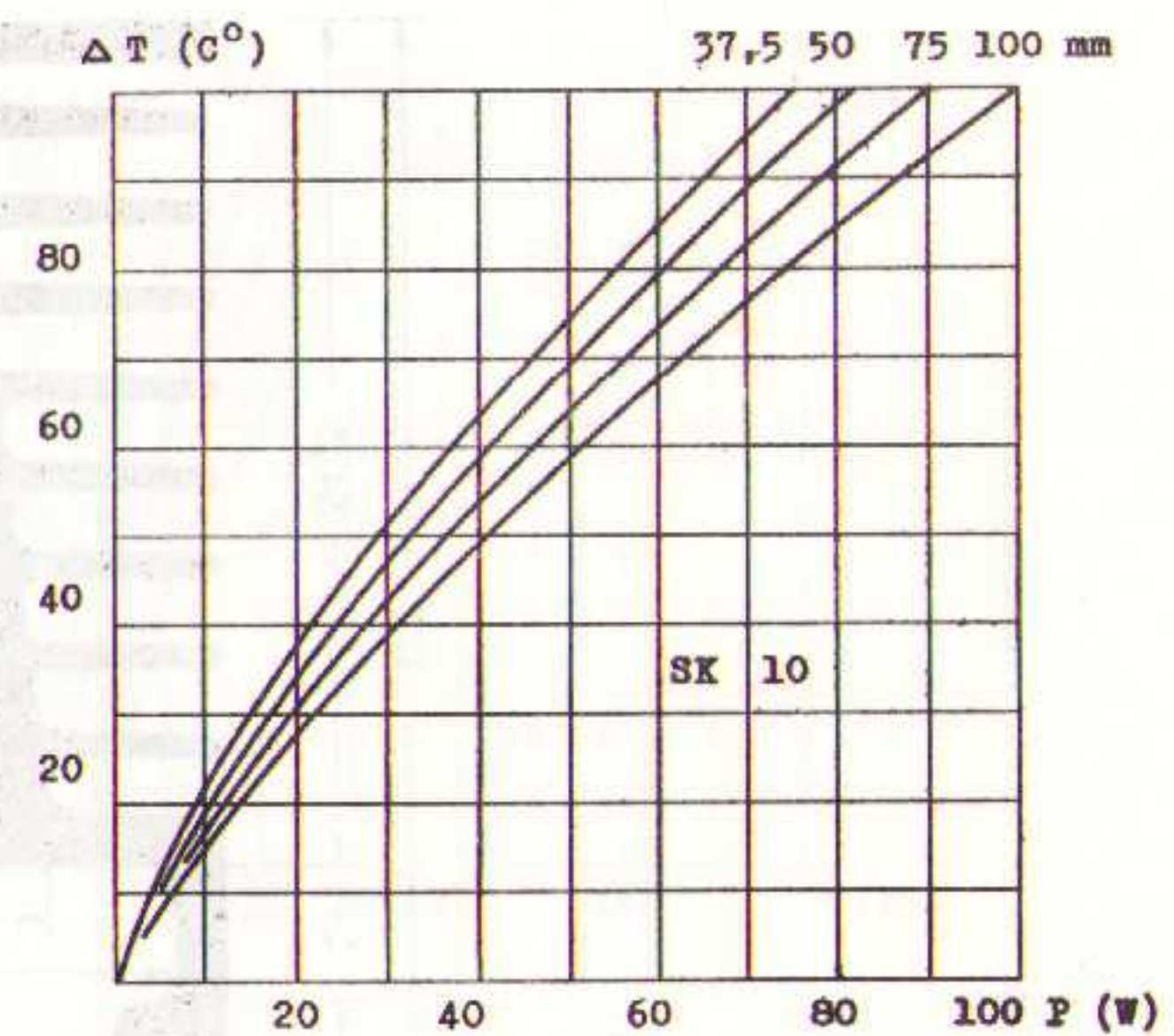




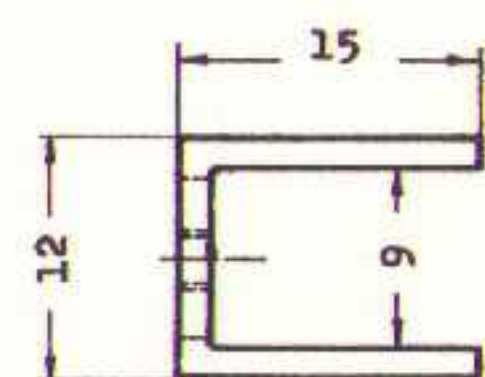
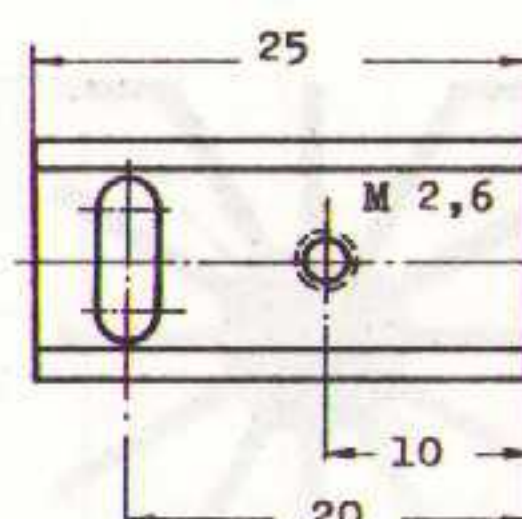
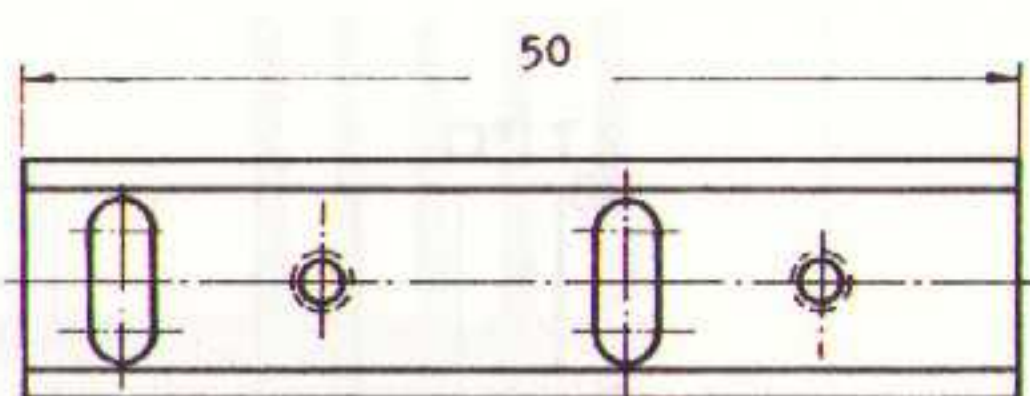
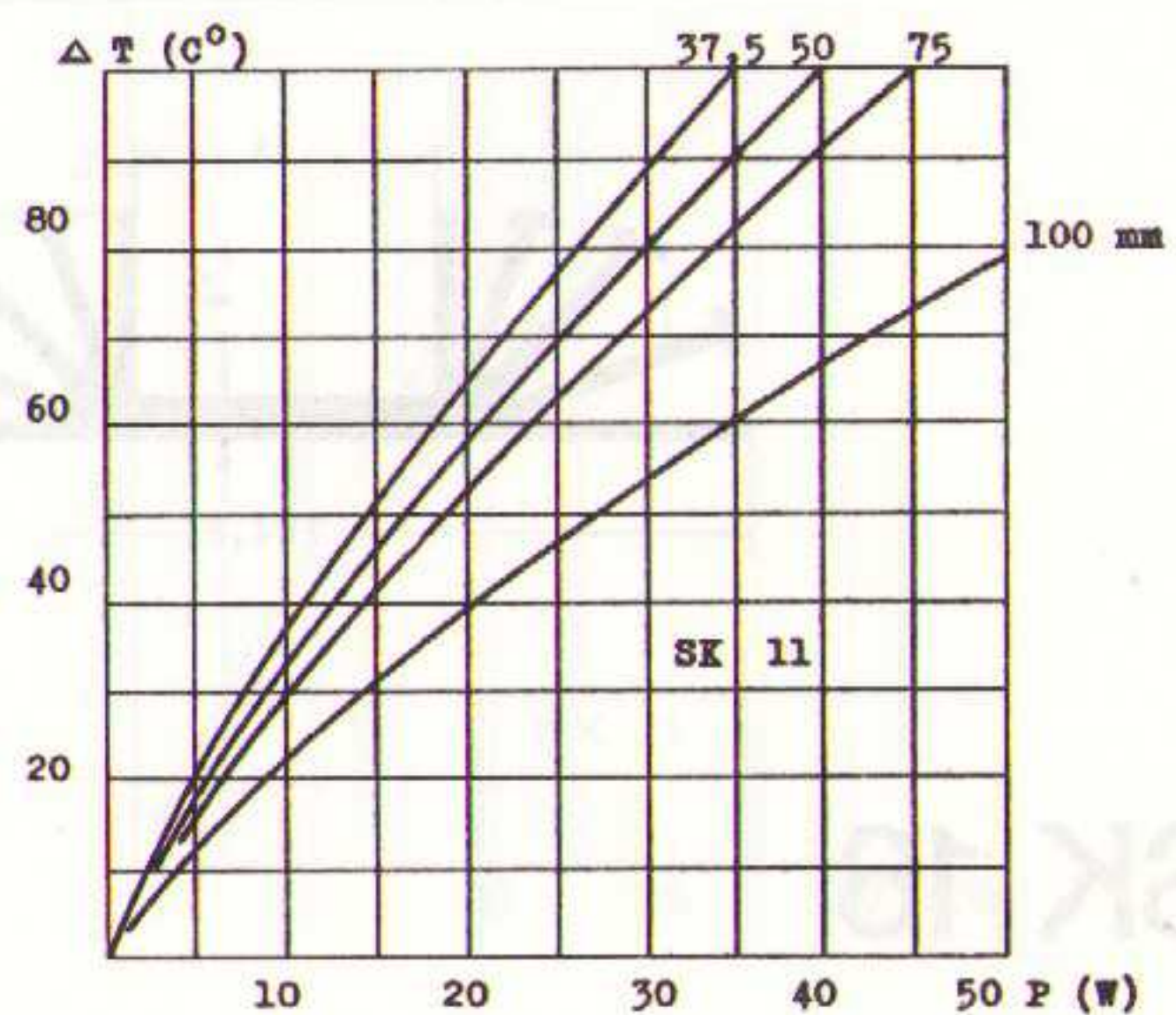
SK 09



SK 10



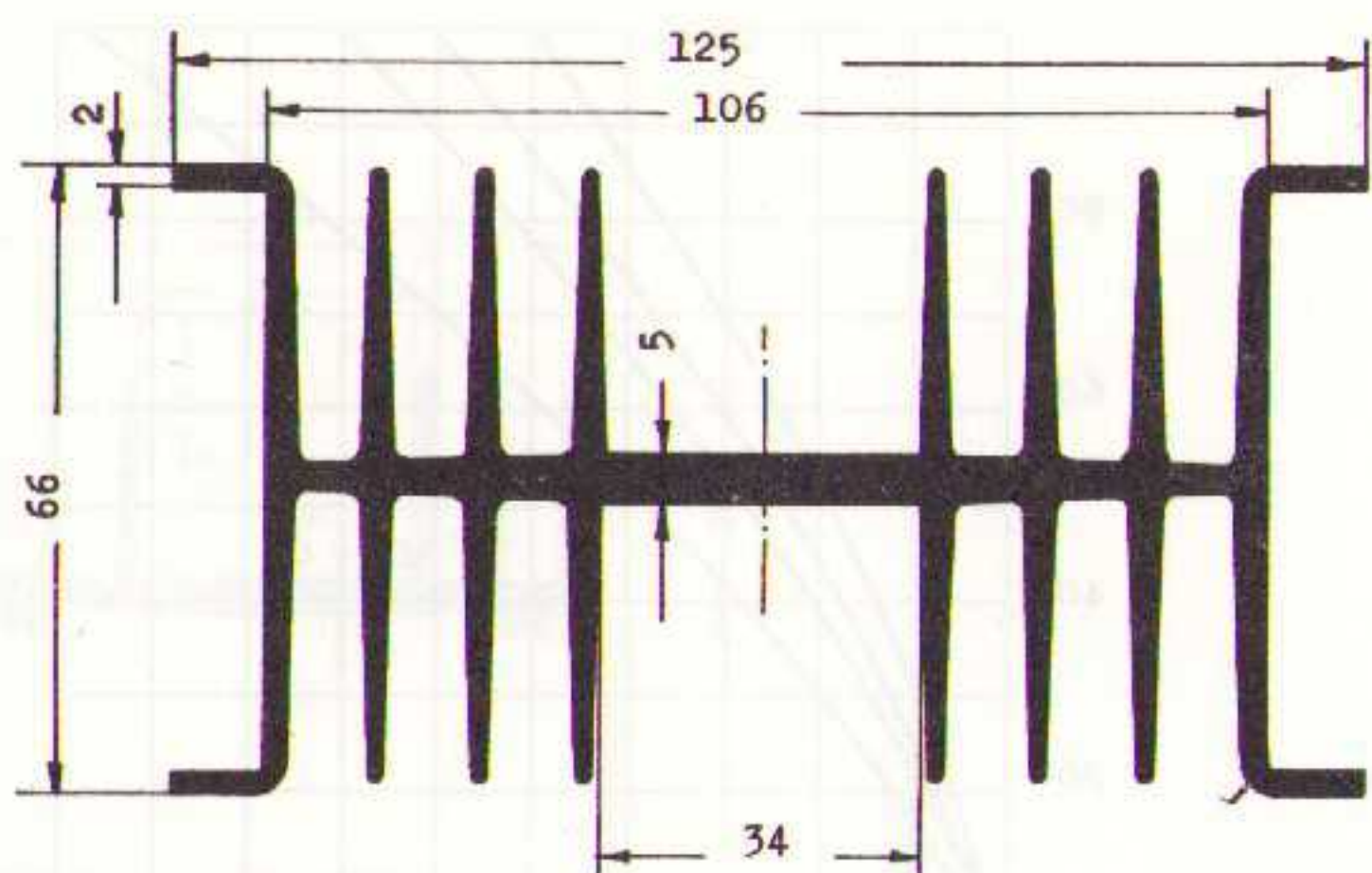
SK 11



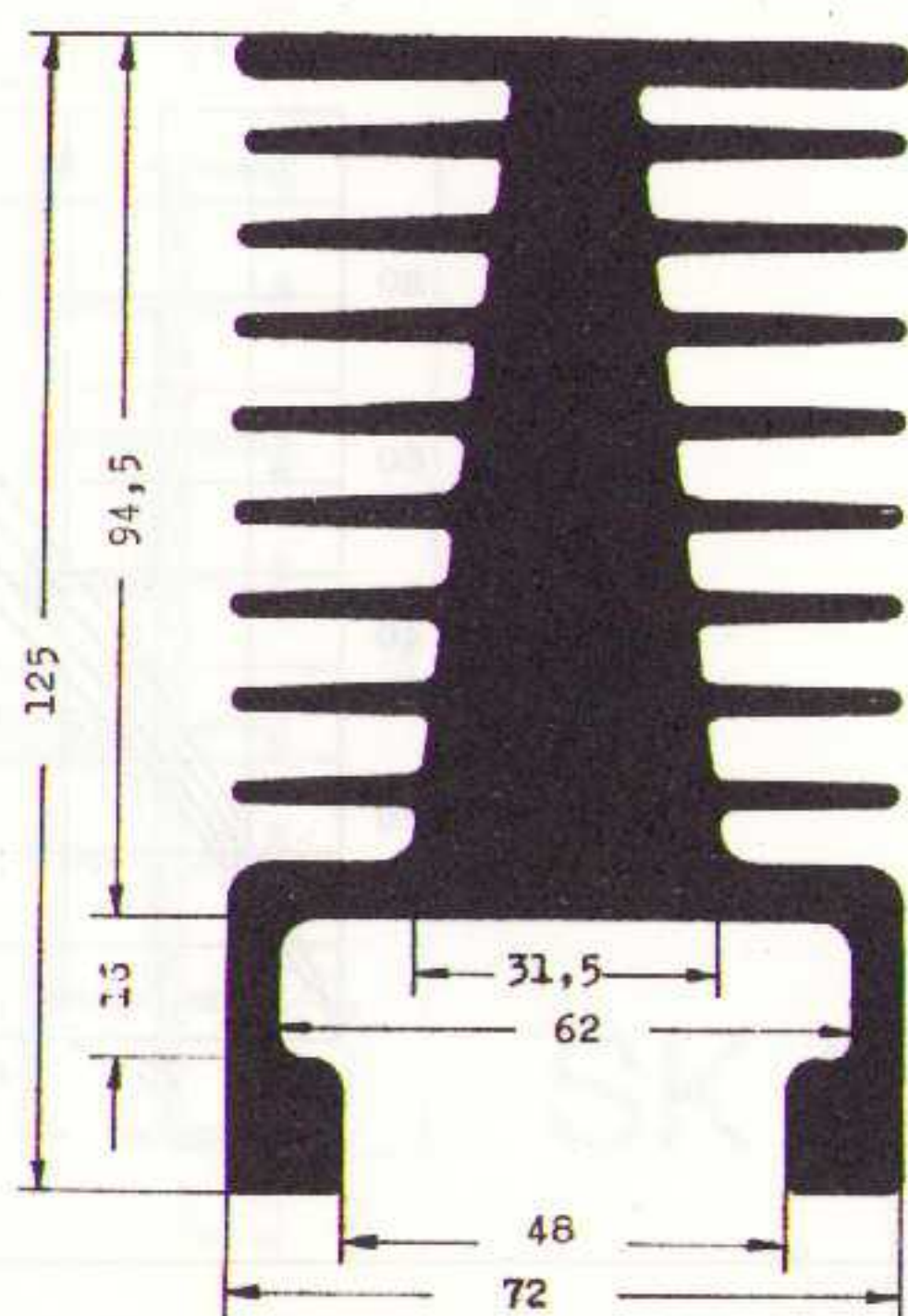
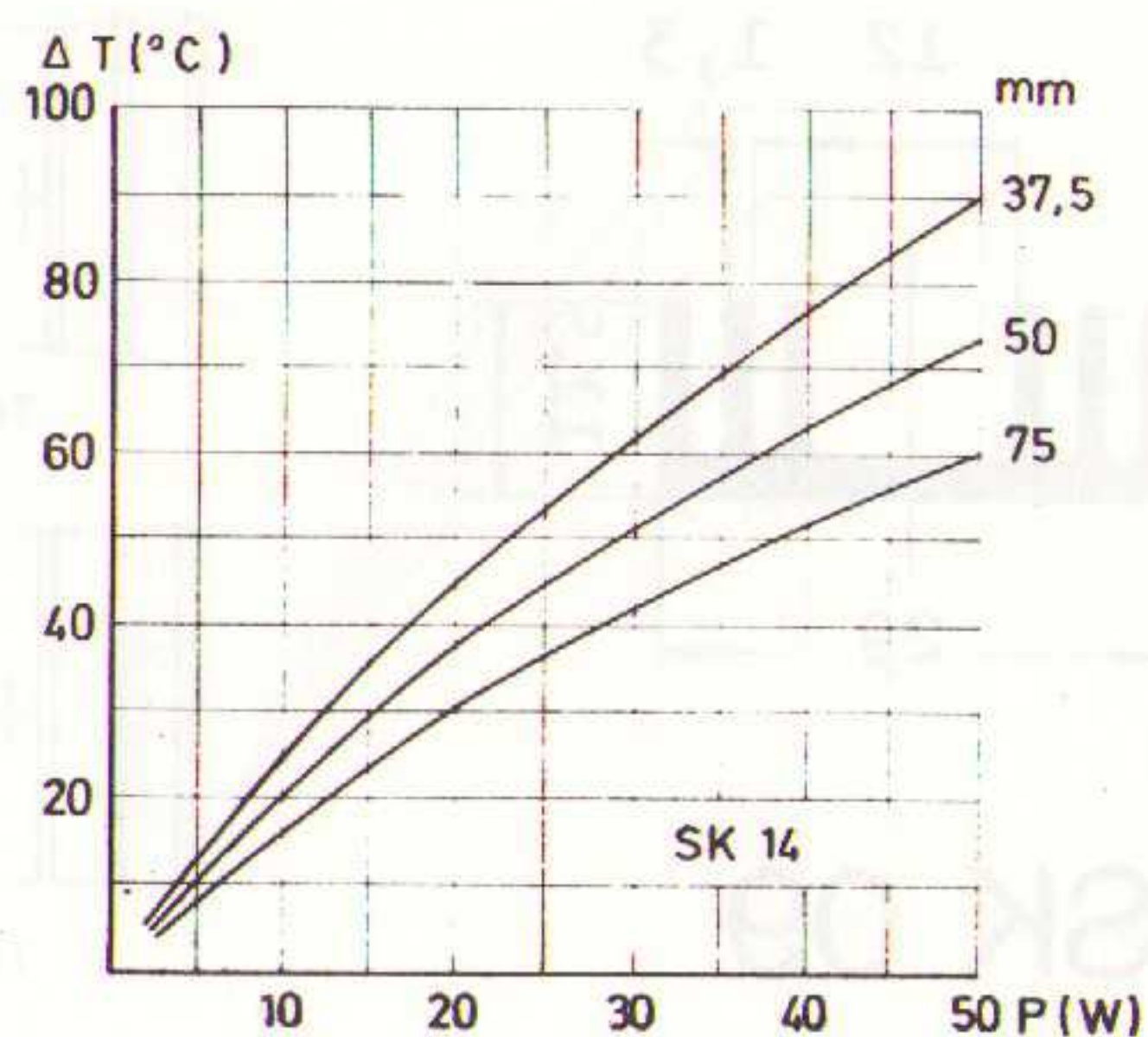
SK 12

Voor plastic transistoren met een SOT 32 behuizing.

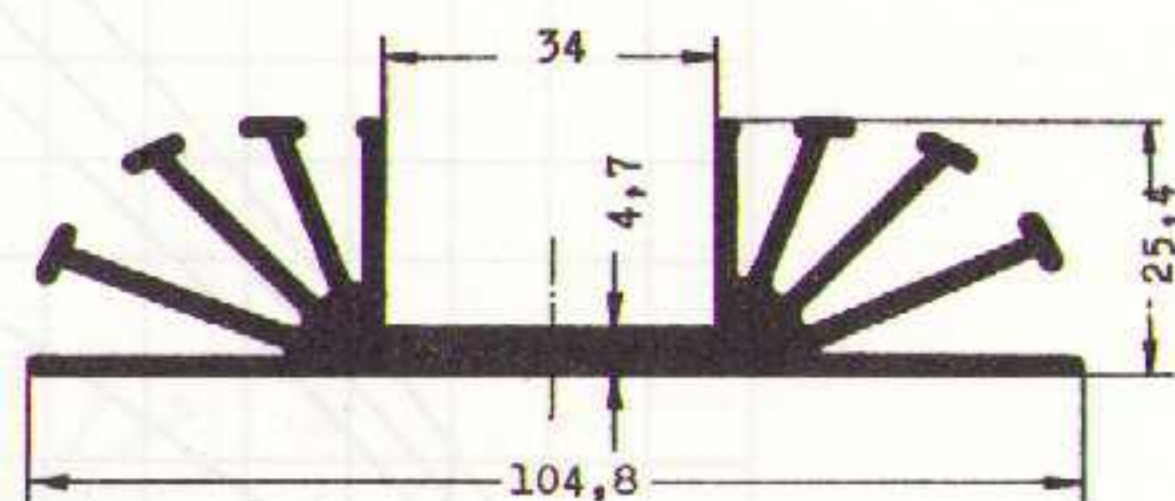
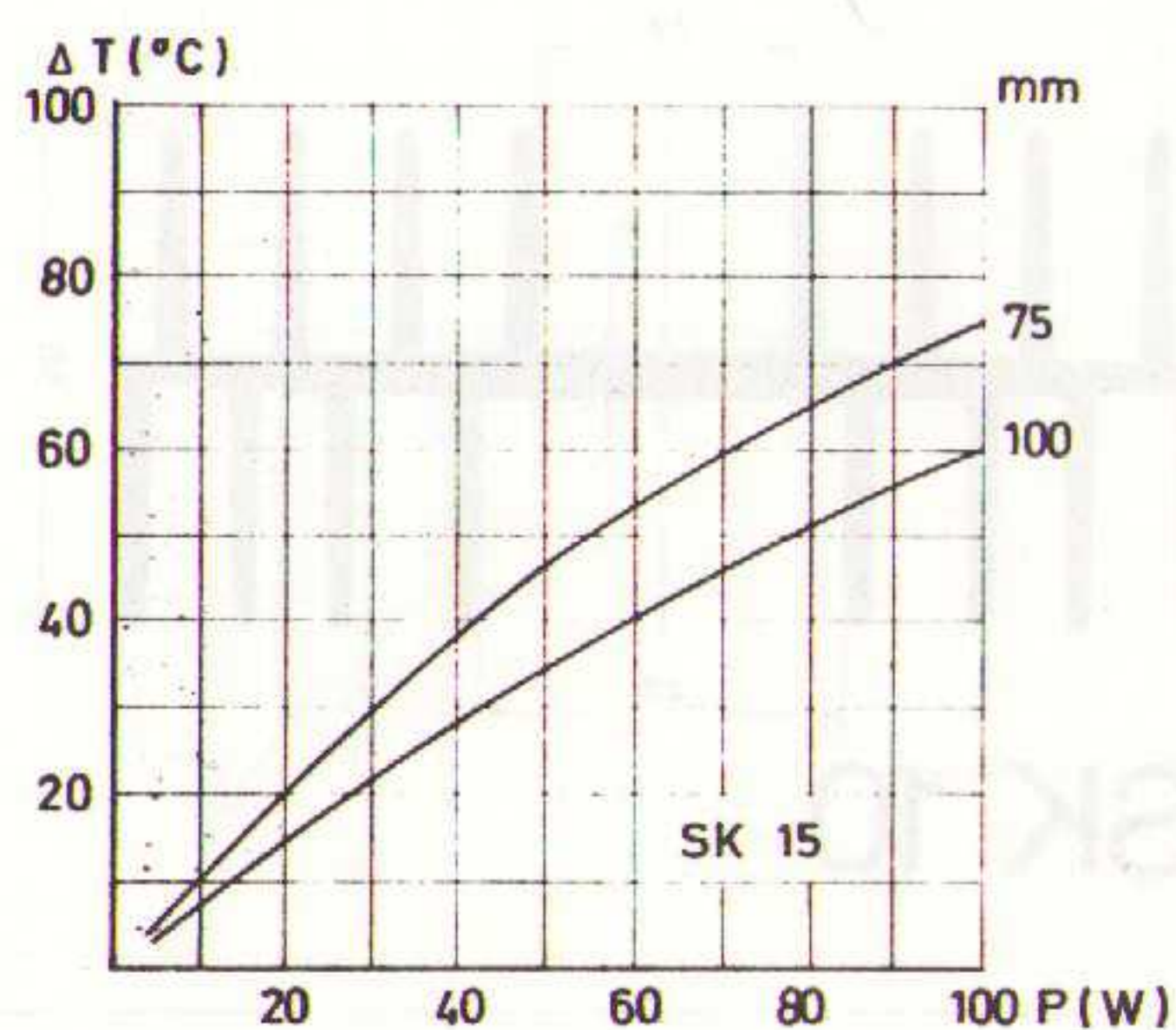
Warmteweerstand bij 25 mm en SA-coating $16,8^{\circ}\text{C/W}$
 Warmteweerstand bij 50 mm en SA-coating $8,4^{\circ}\text{C/W}$



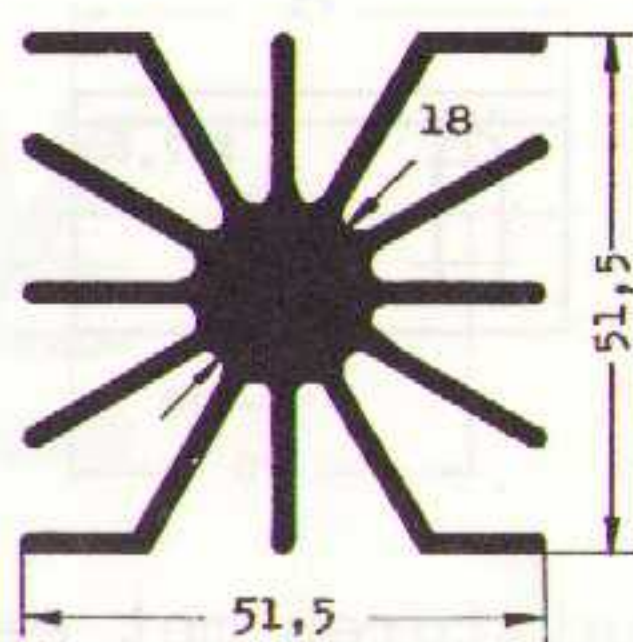
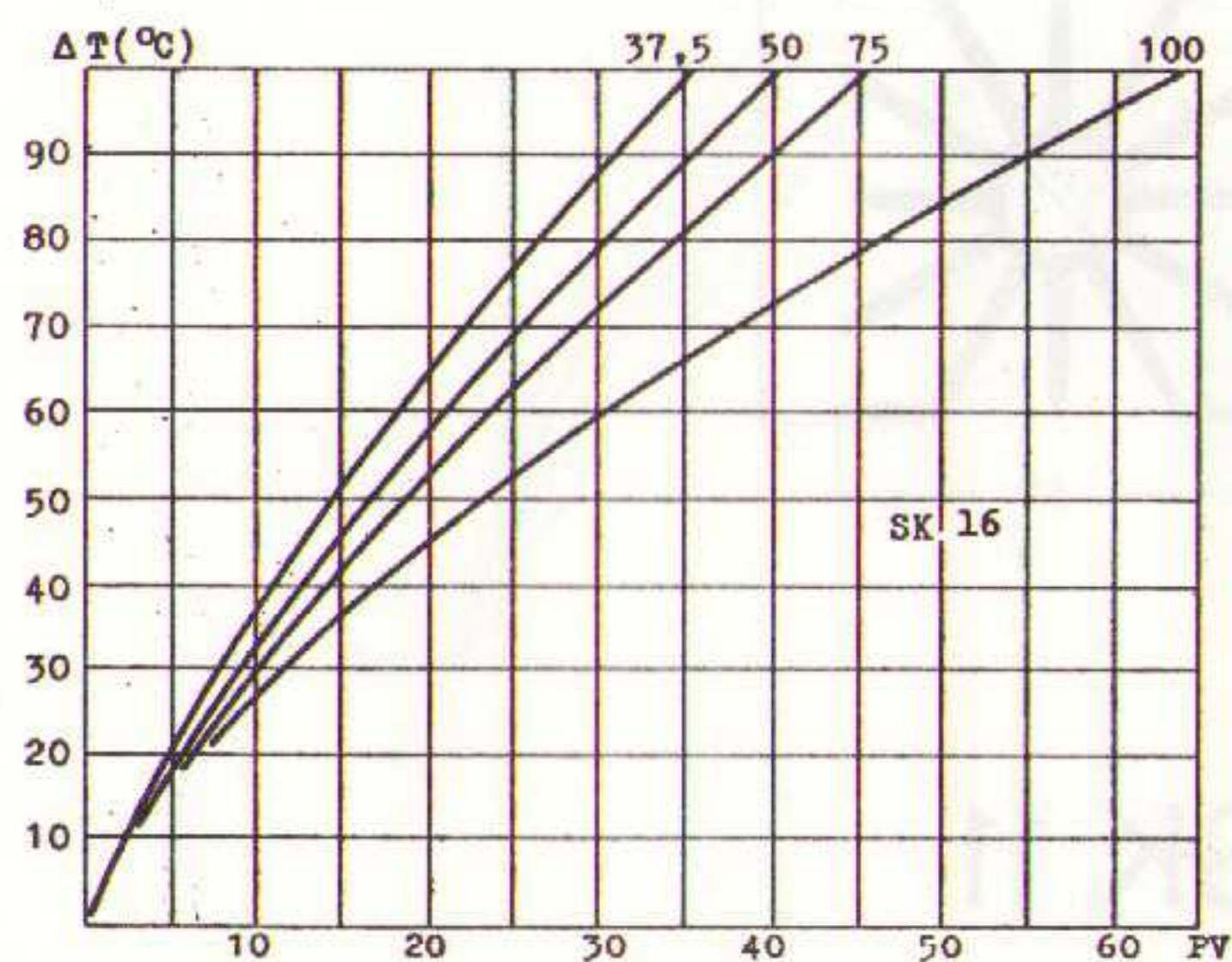
SK 14



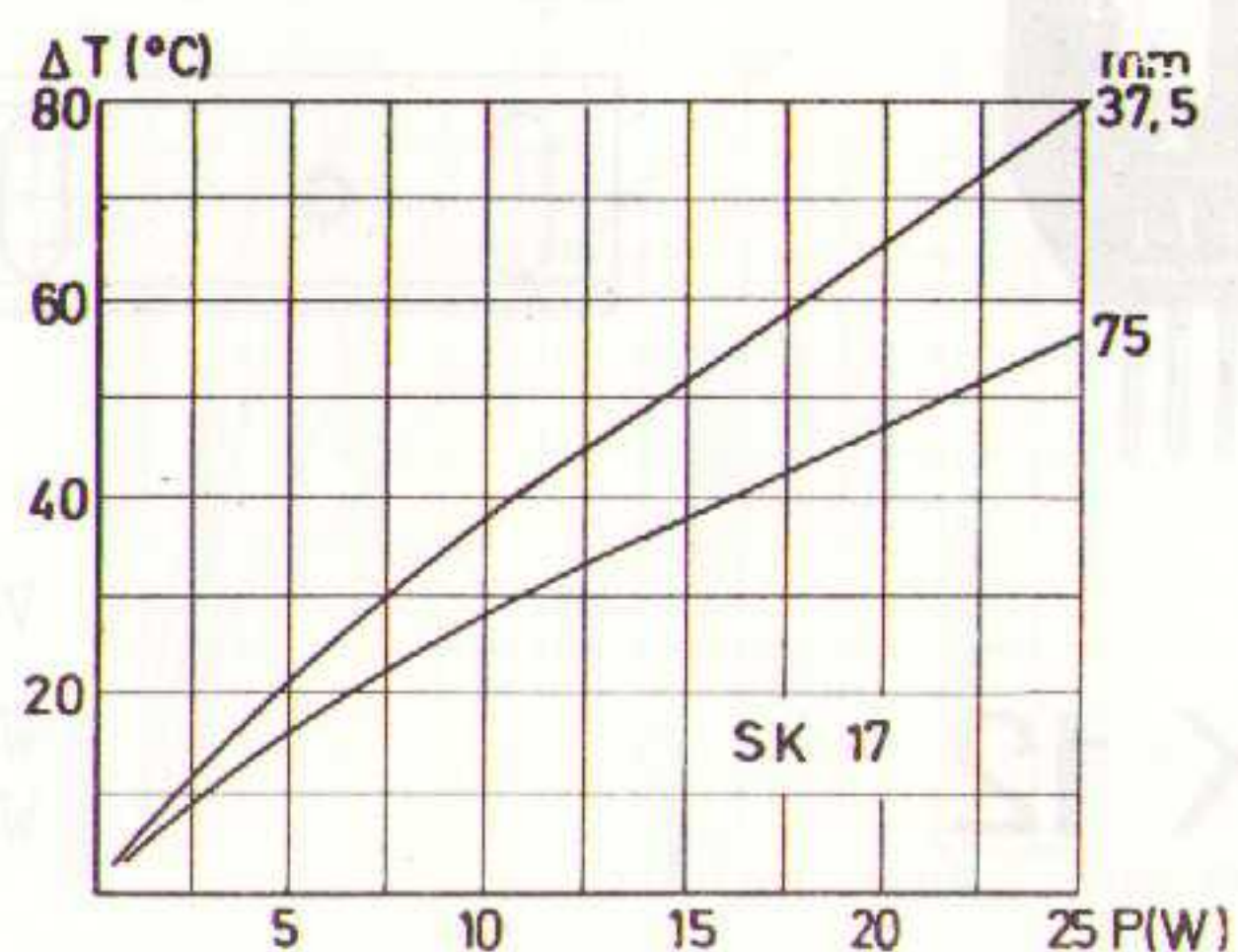
SK 15



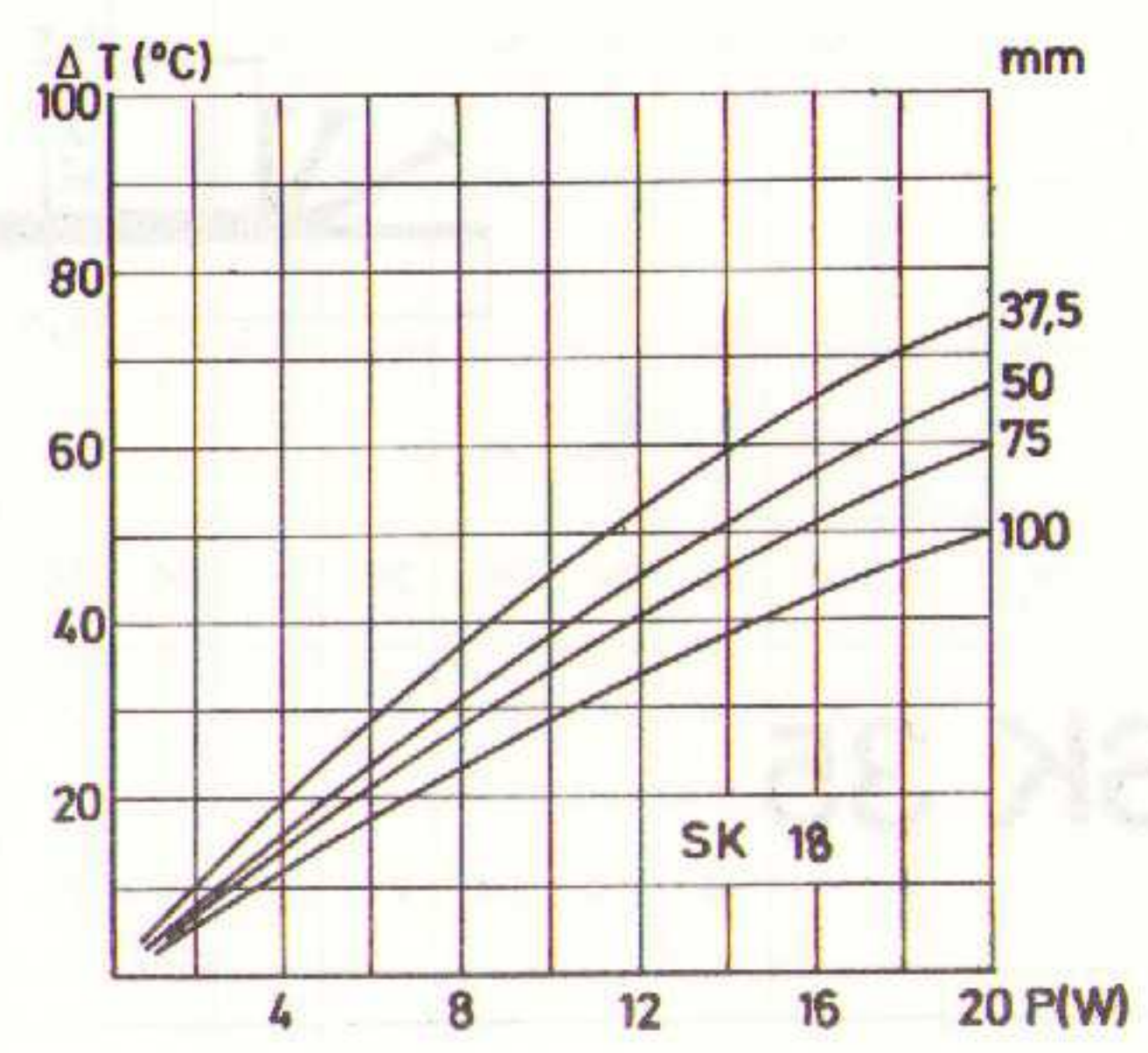
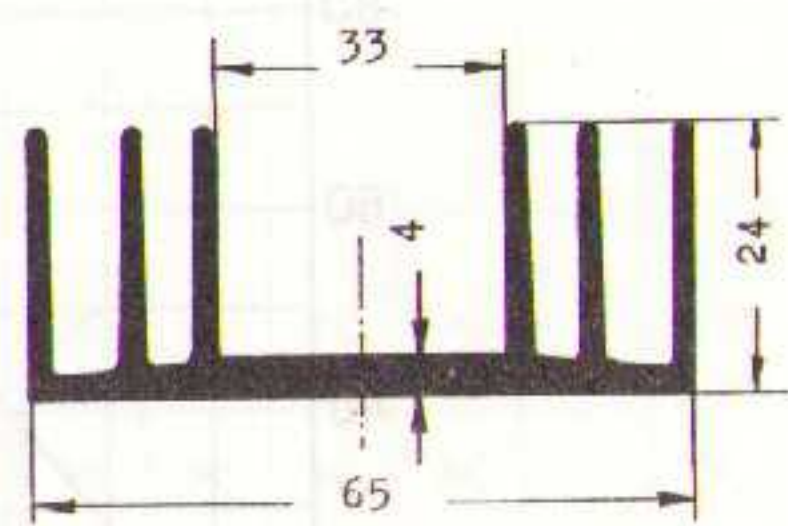
SK 16



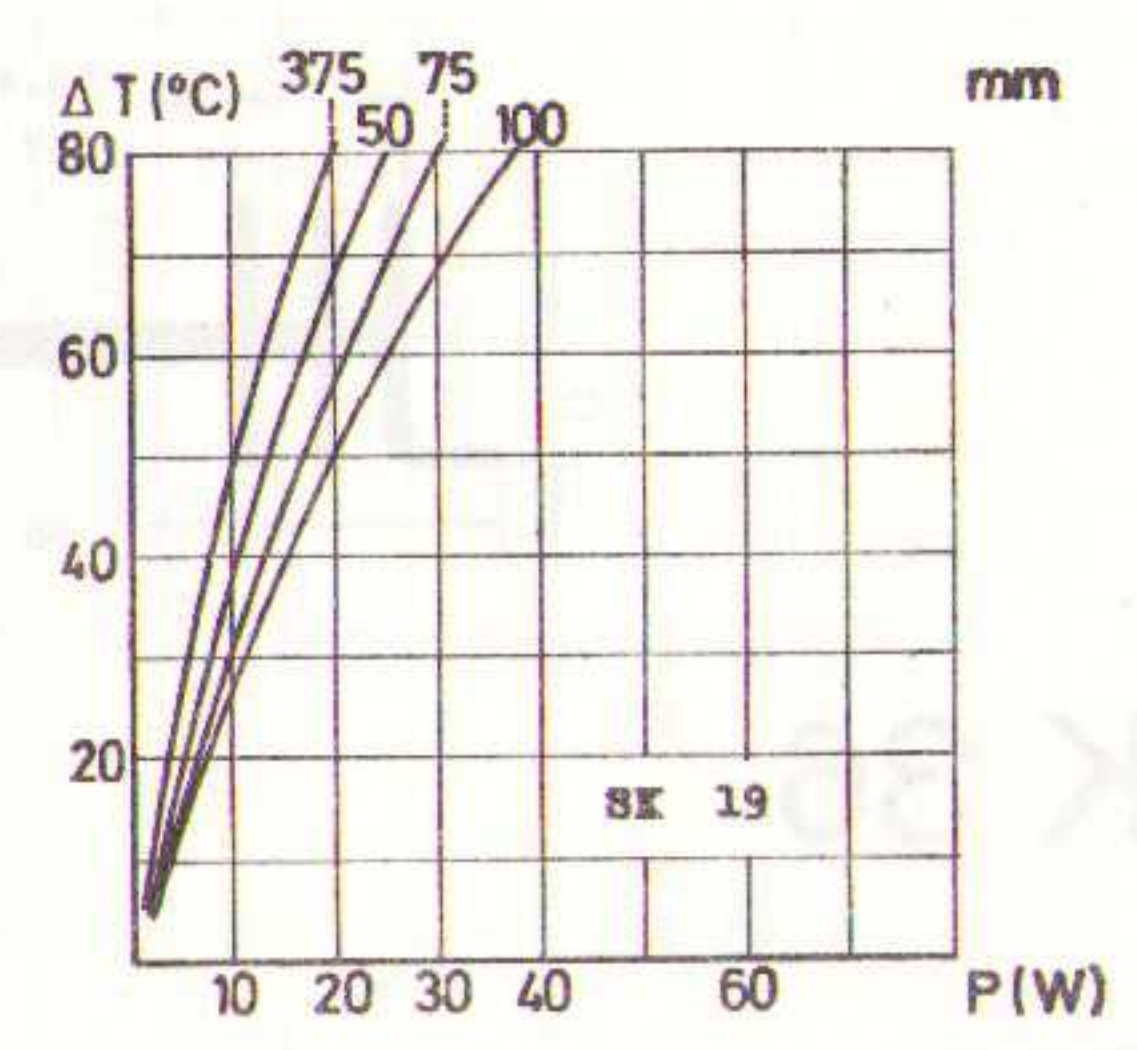
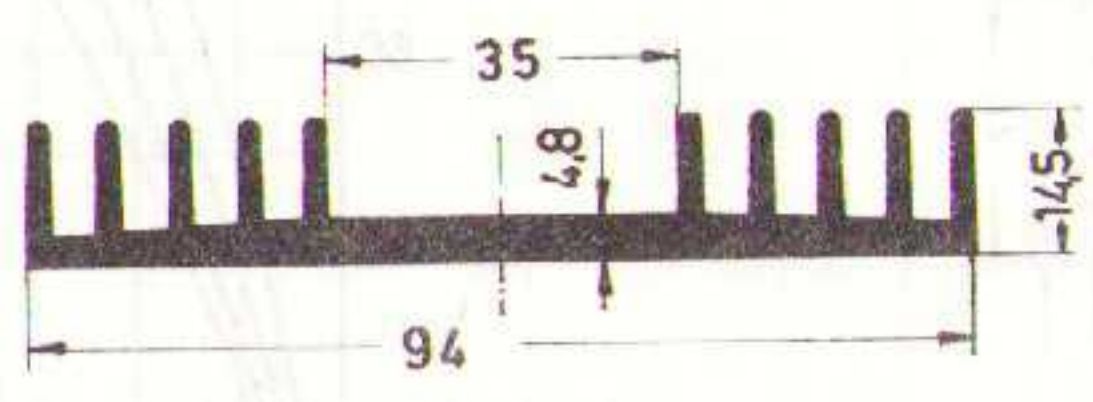
SK 17



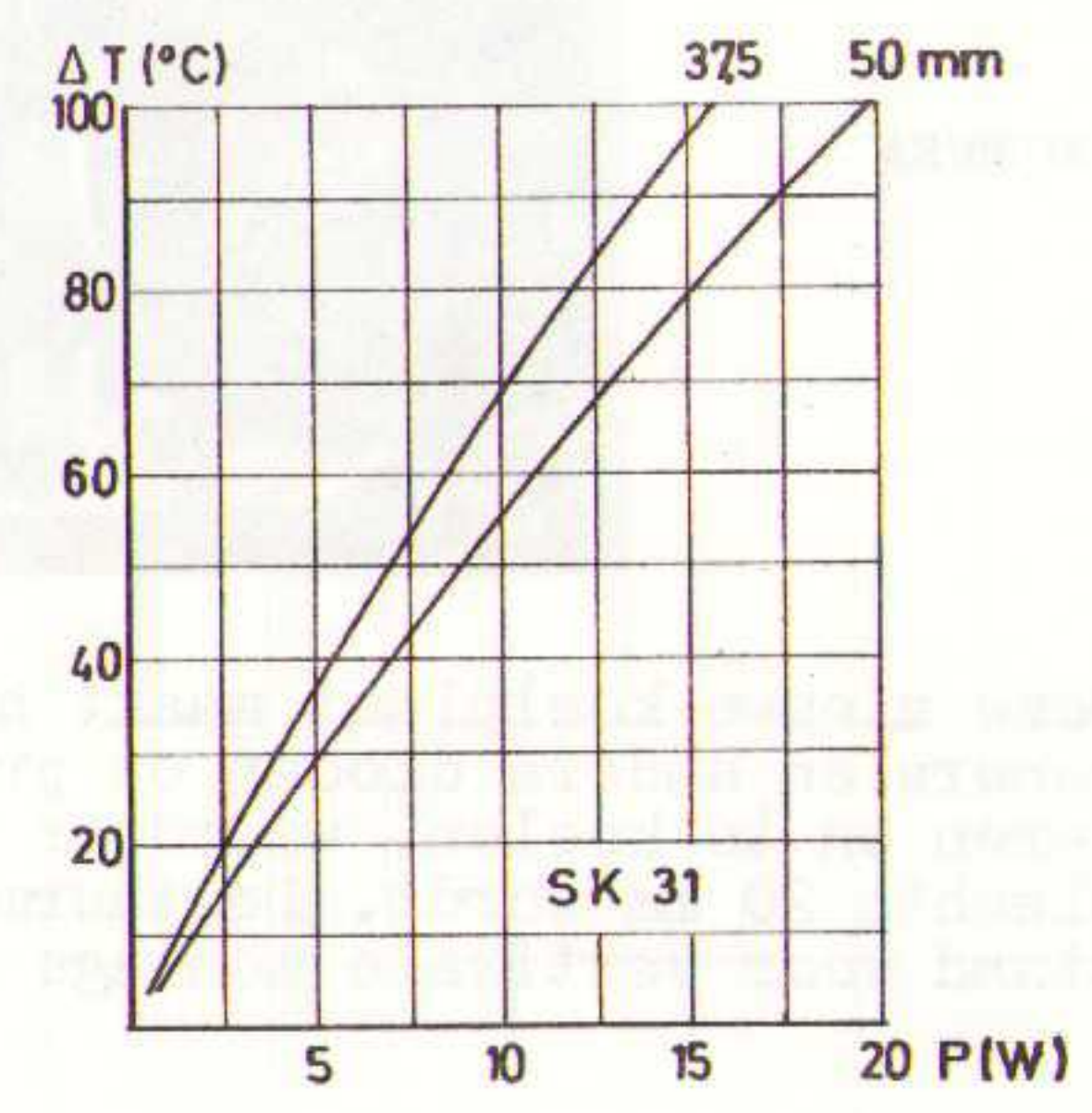
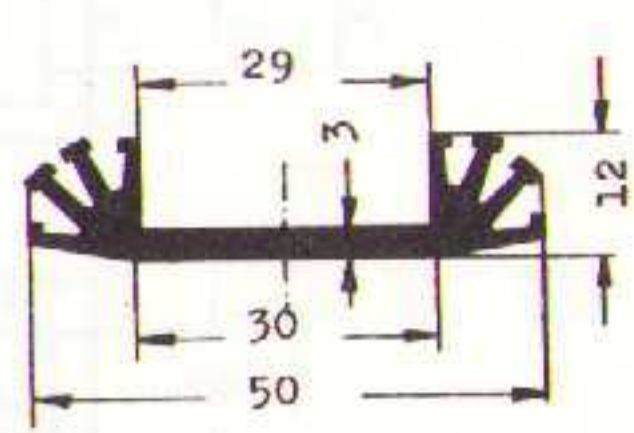
SK 18



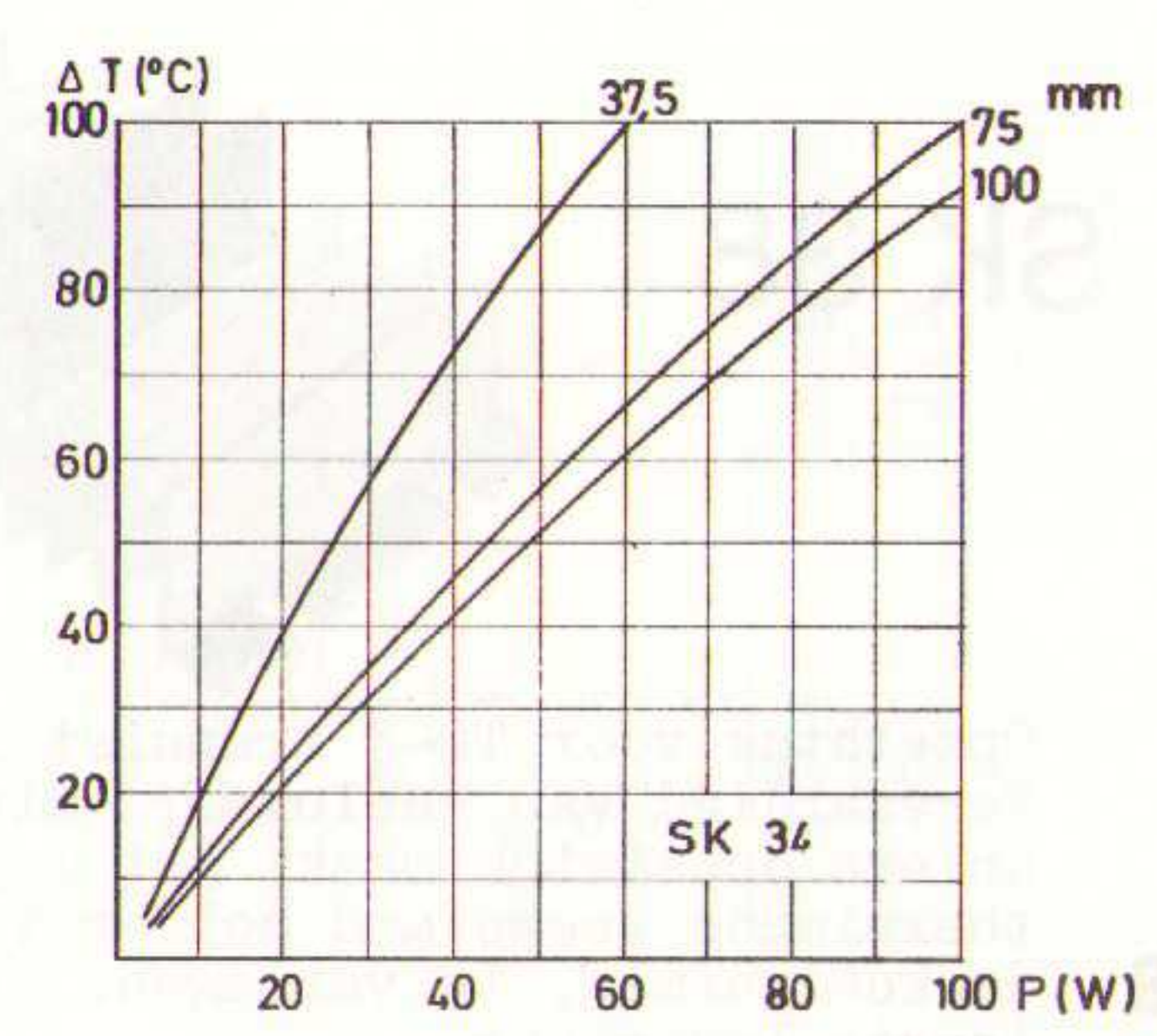
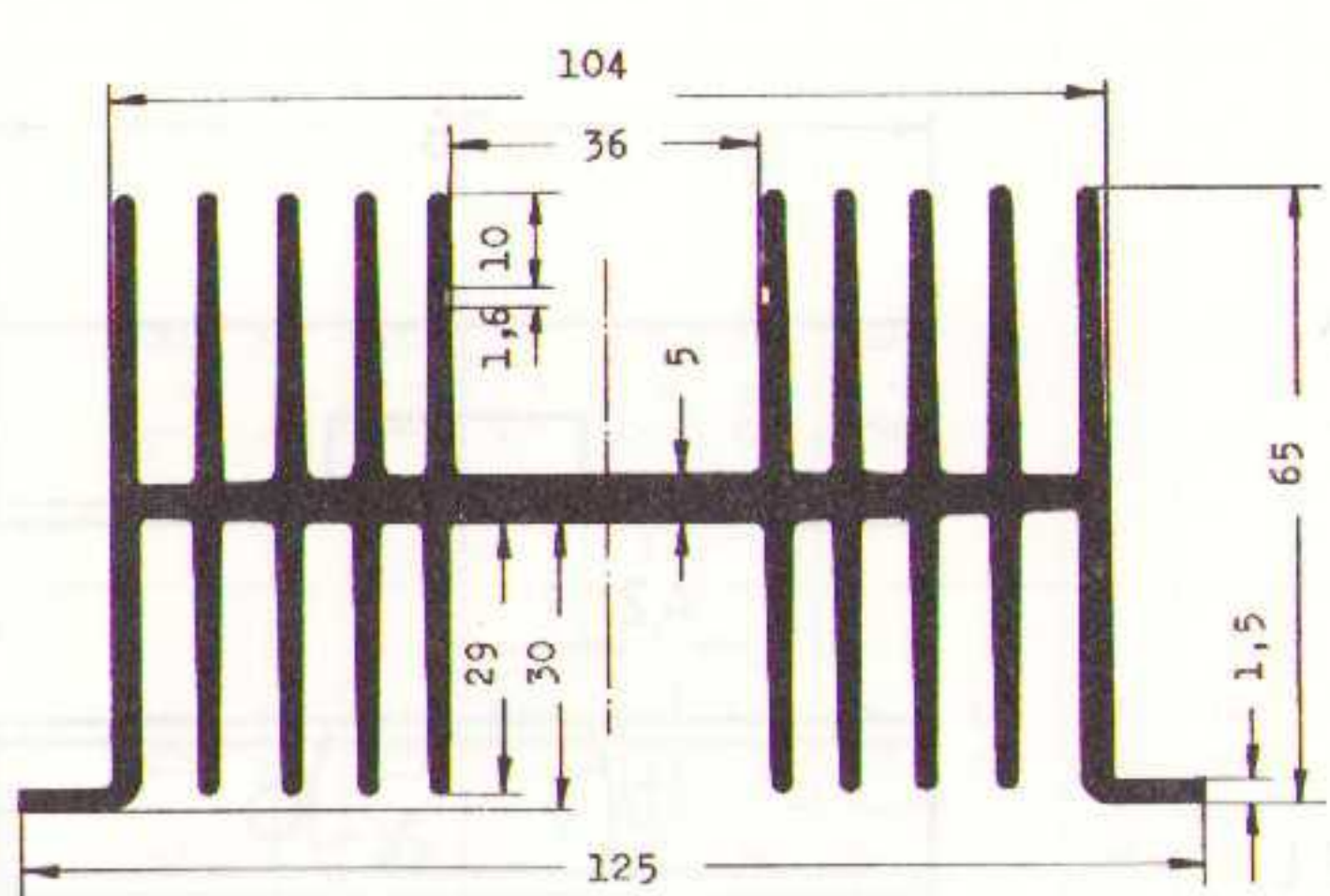
SK 19

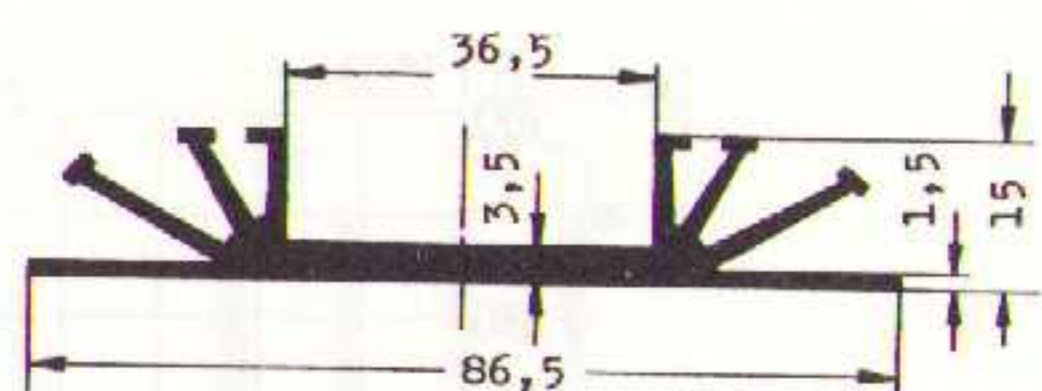


SK 31

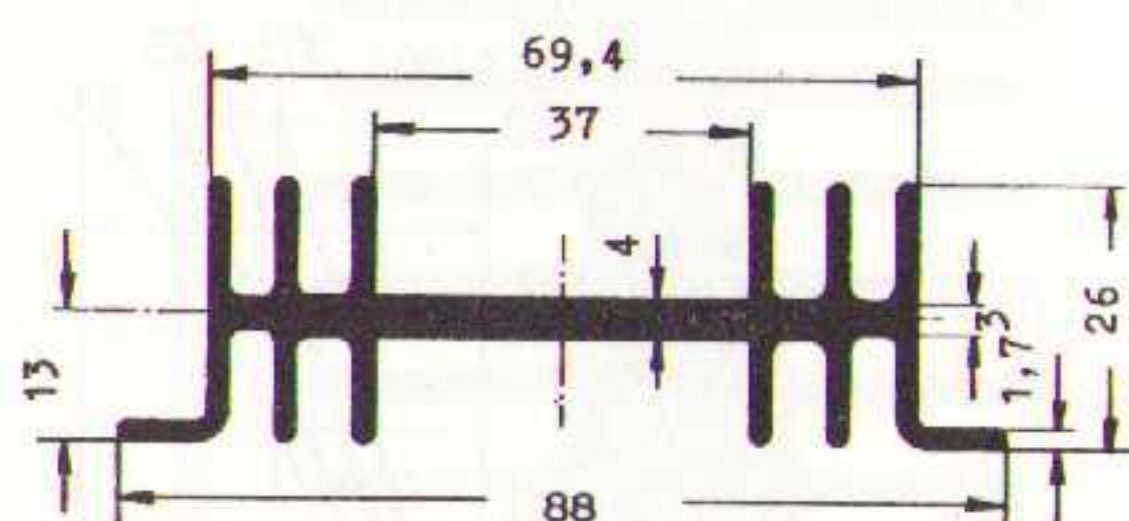
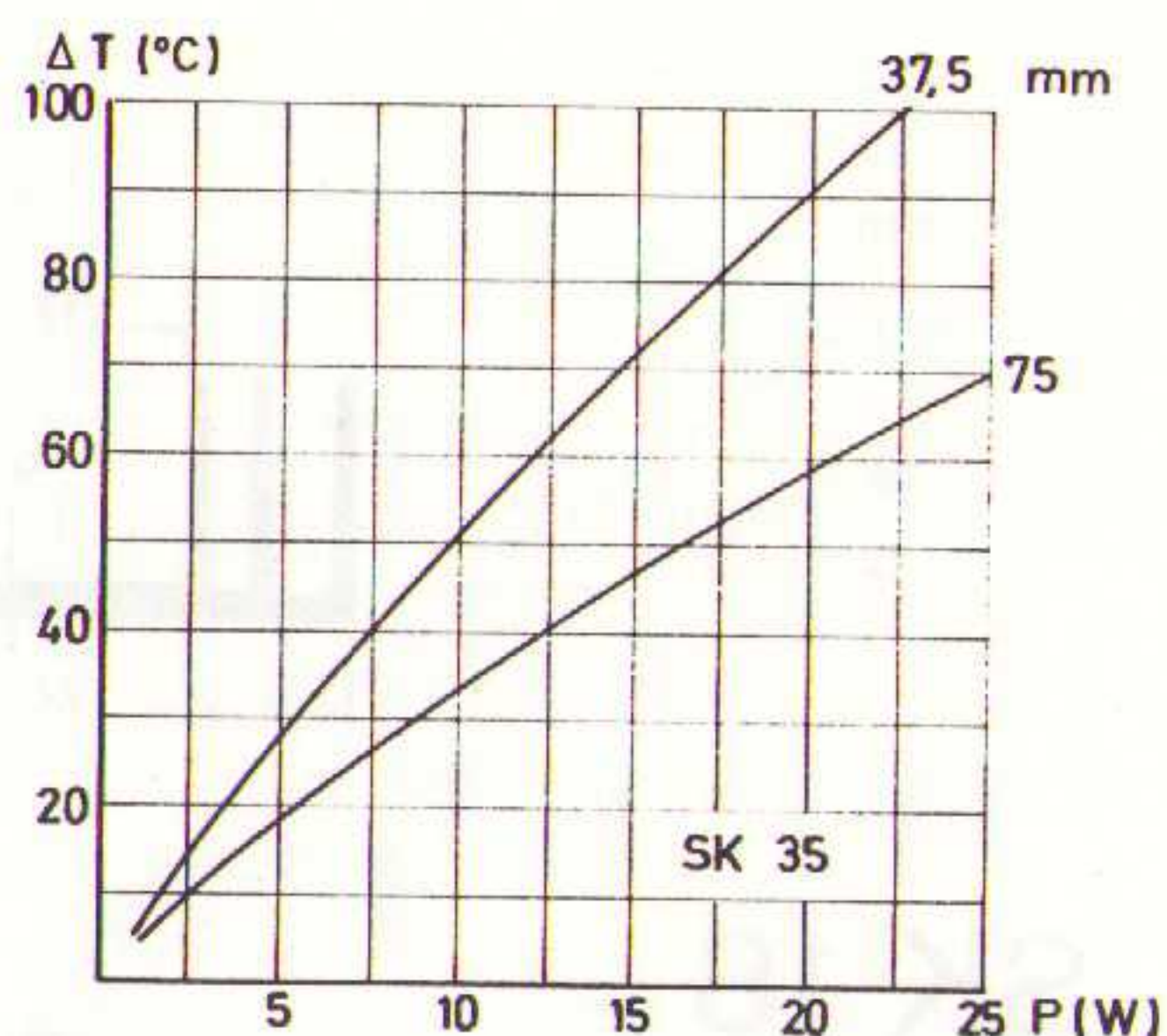


SK 34

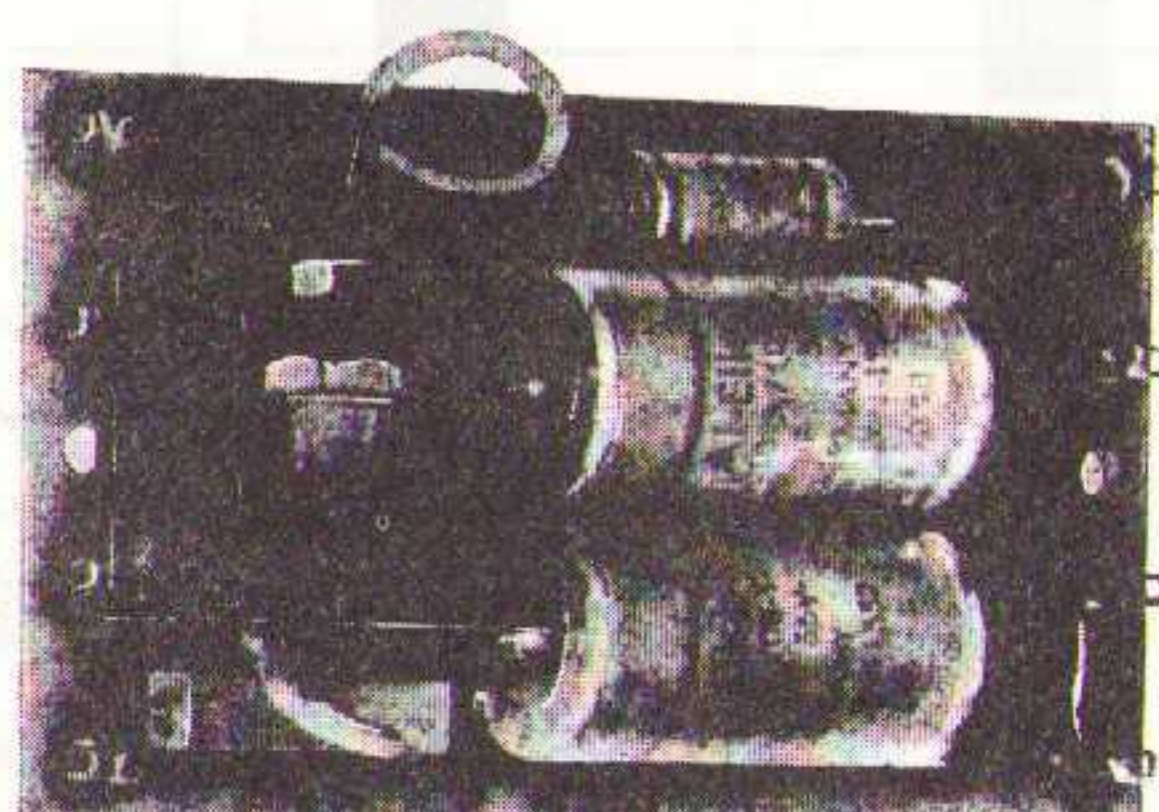
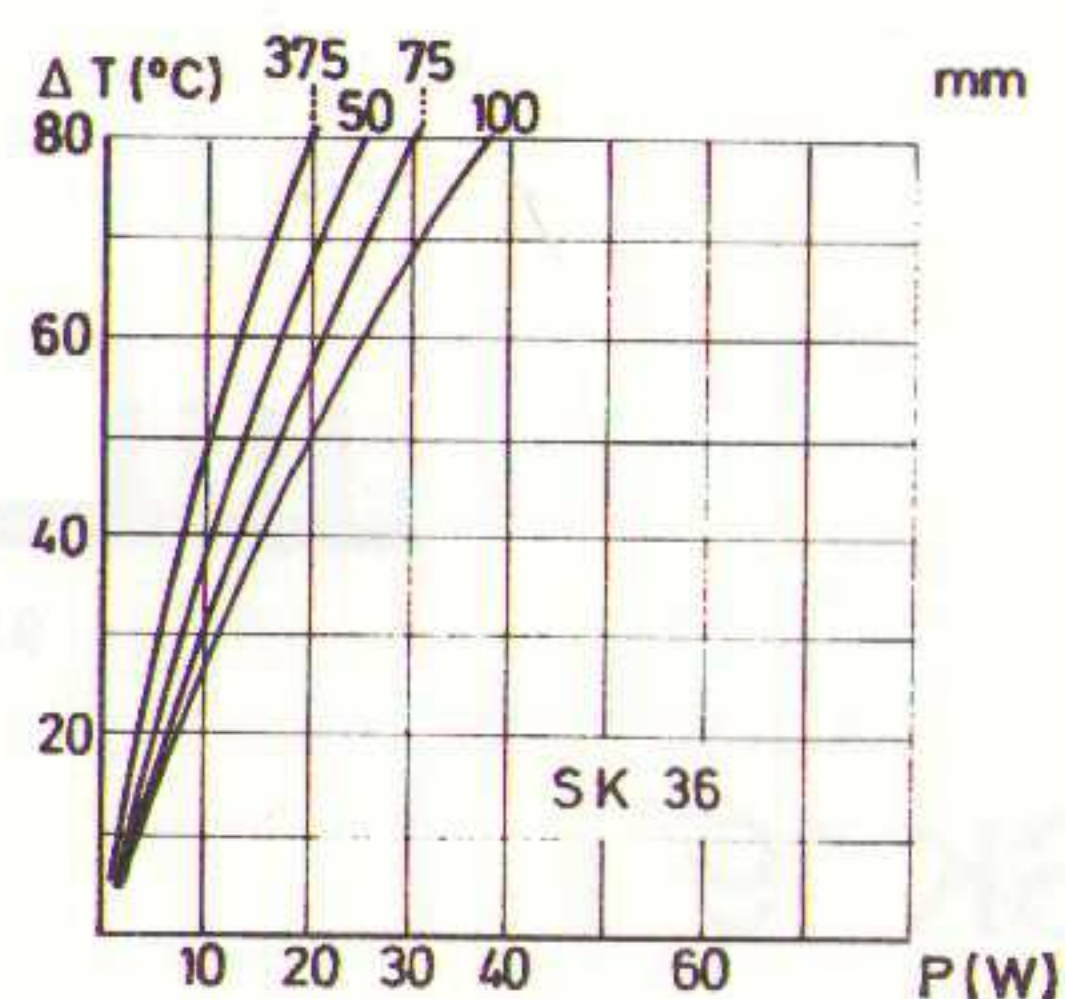




SK 35

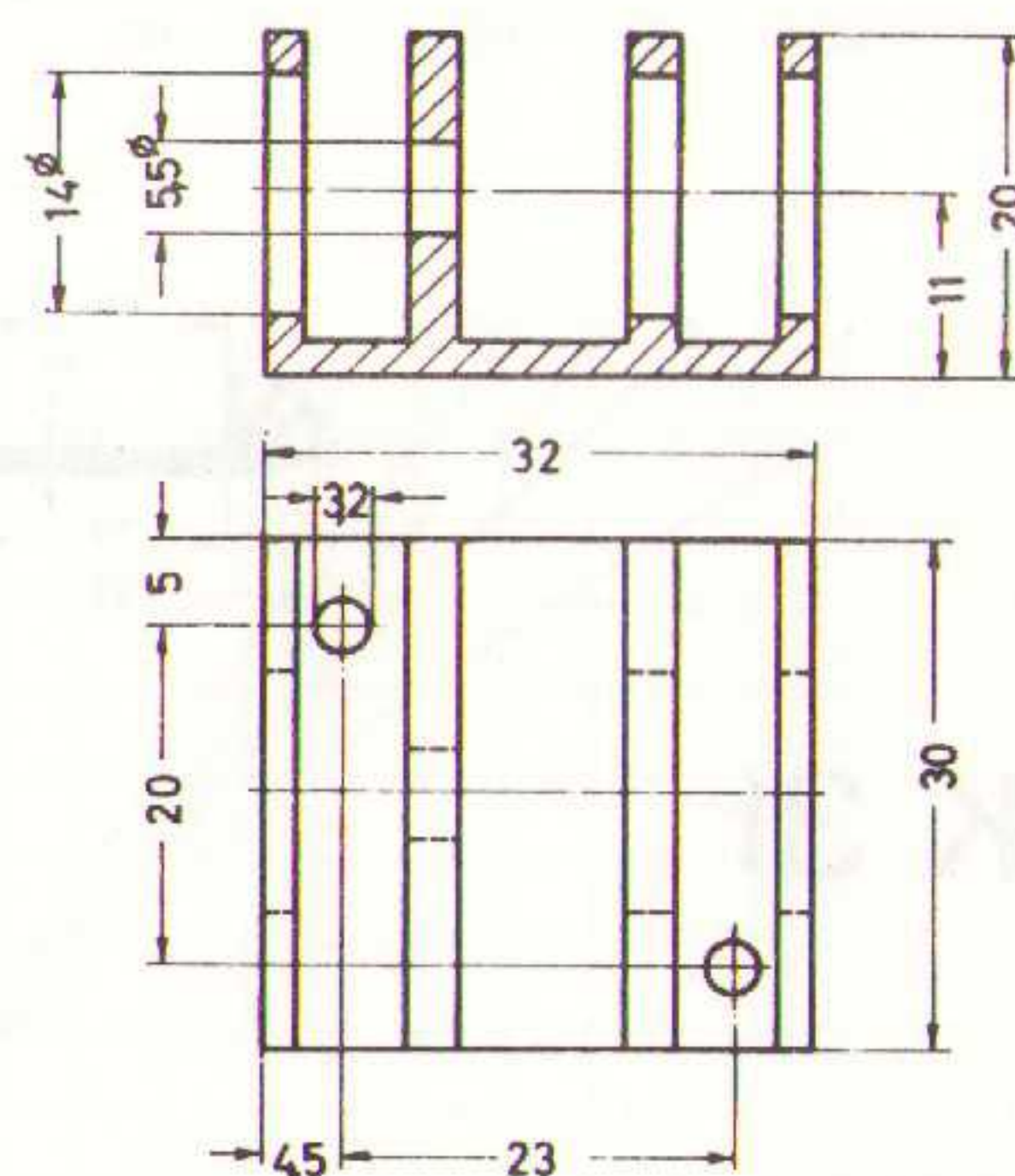


SK 36

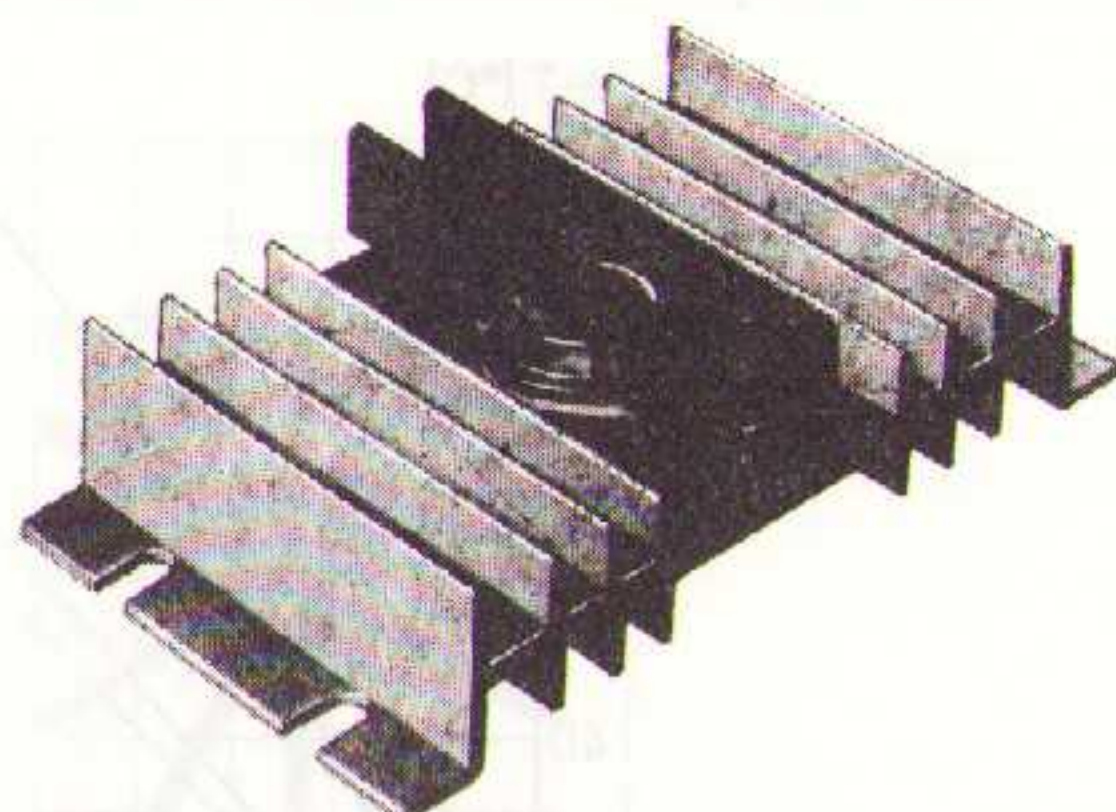


SK 37/30/SA

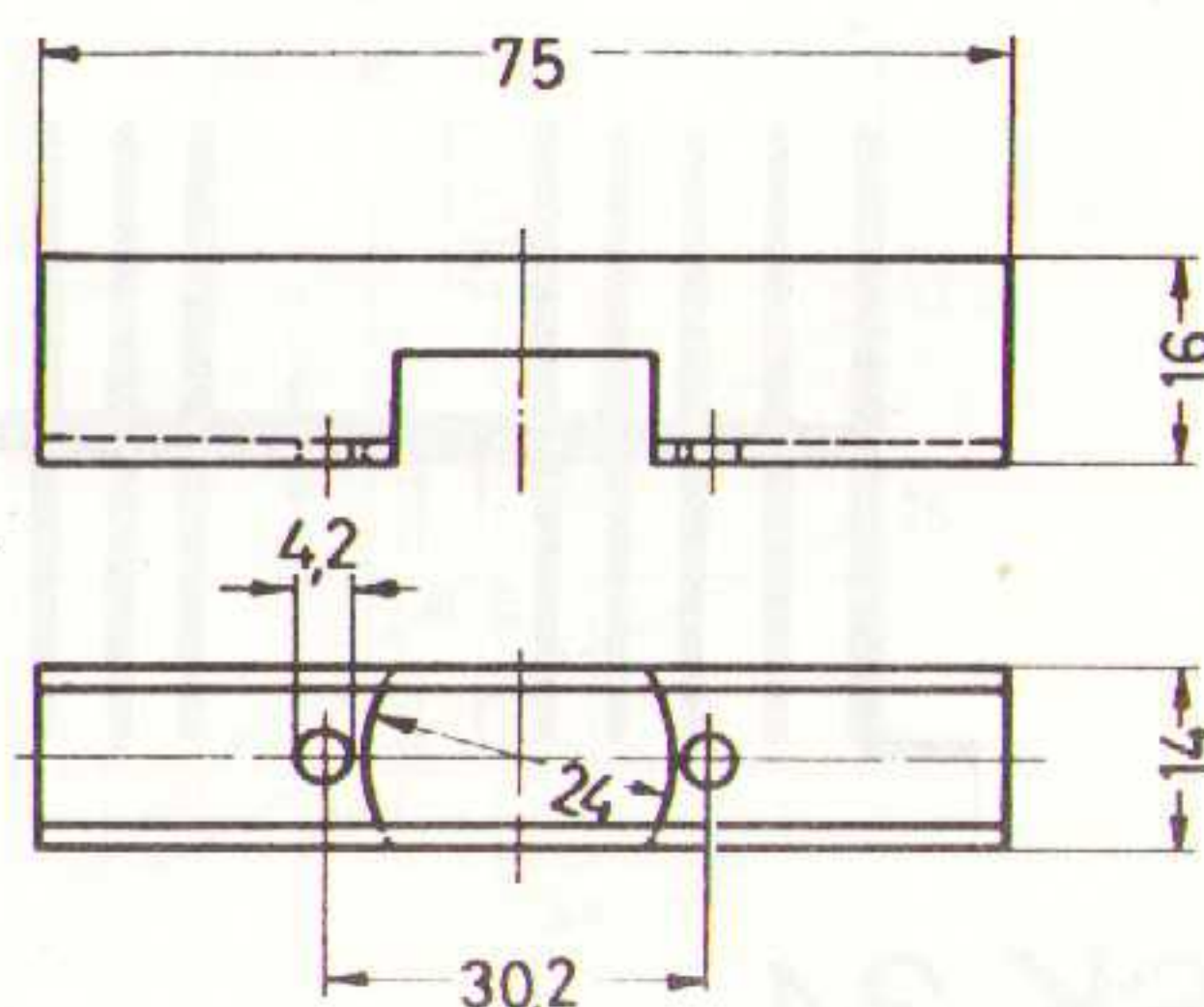
Deze nieuwe koelplaat maakt het mogelijk, zeners en andere dioden op printen te monteren en te koelen, waardoor de bouwhoogte slechts 20 mm wordt. De thermische weerstand voor verticale montage is 13,9°C/W.



SK 38



Opzetstuk voor TO-3 transistoren. Vervaardigd van geëloxeerd aluminium. Dit unieke opzetstuk maakt het mogelijk de thermische weerstand zonder vergroting van de koelruimte, te verlagen. Alleen in 75mm lengte leverbaar.

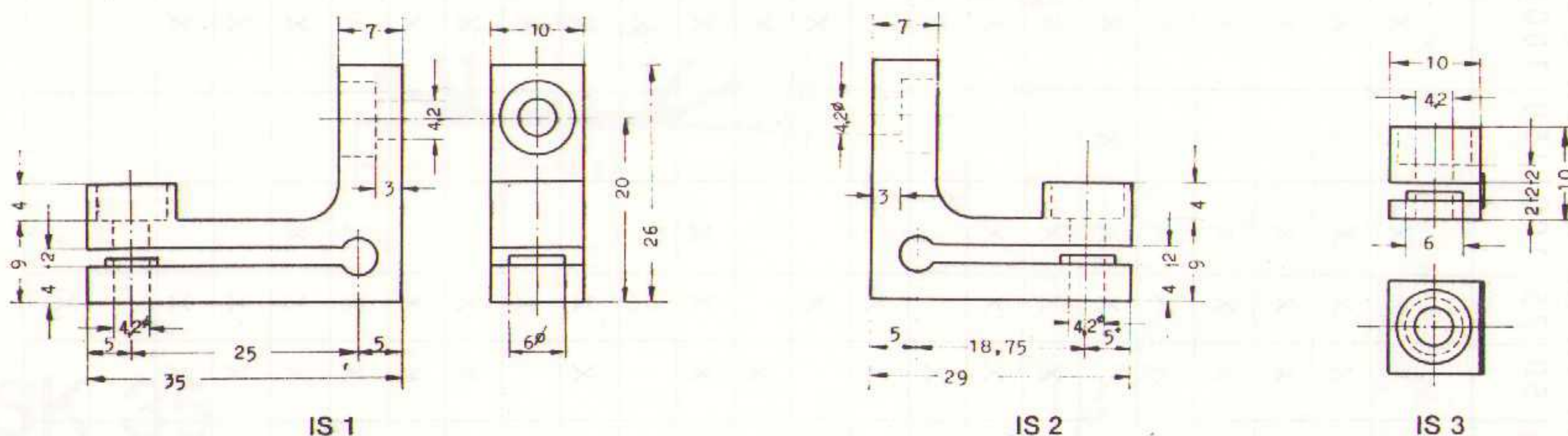


Programma-overzicht koelelementen volgens verschillende lengten en boringen.

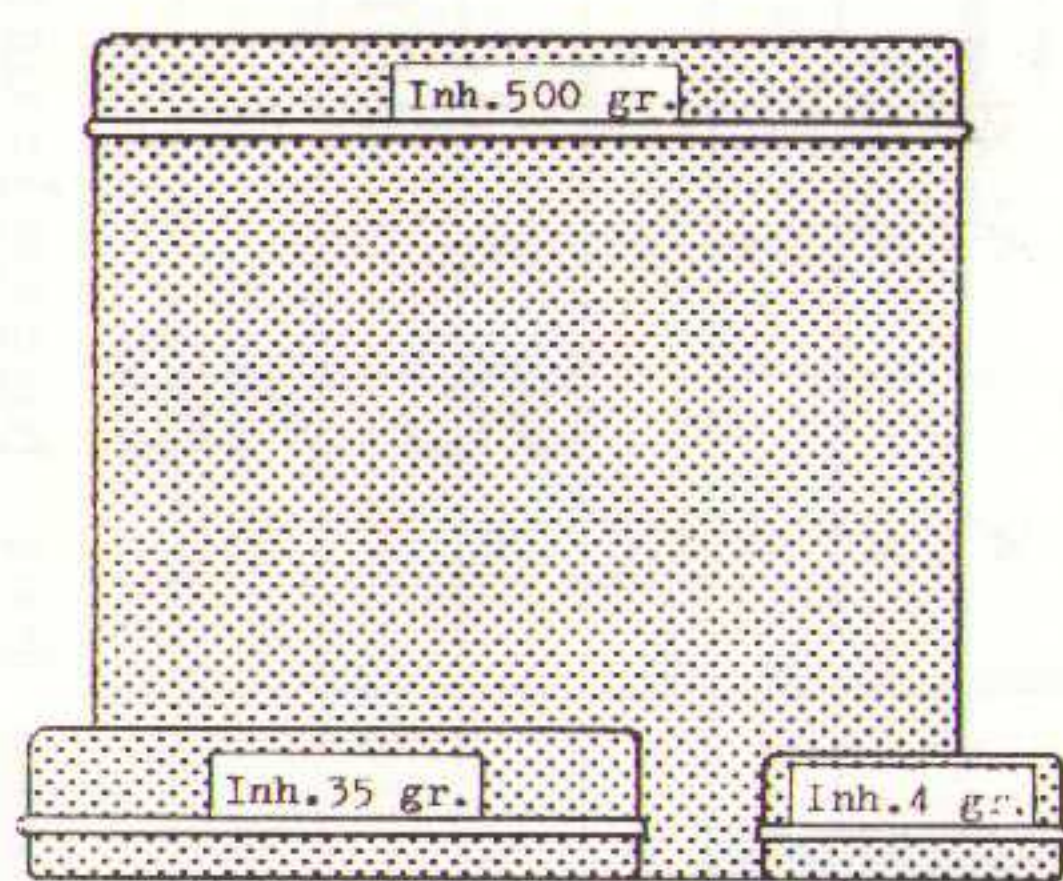
SERIE NUMMER	°C/W BIJ 100MM EN 50WATT	AL-MAT GEËTST (AL)	ZWART GEËLOXEERD (SA)	UITVOERING BORING			VERKRIJGBAAR IN DE LENGTE (MM)									
				ONGEBOORD	TO-3 SOT-9	TO-36 SPEC.	19	25	30	37½	50	75	100	150	1000	
SK 01	1,8	X	X	X	X	X				X	X	X				X
SK 02	1,1	X	X	X	X	X					X	X	X			X
SK 03	1,6	X	X	X	X	X					X	X	X			X
SK 04	1,7	X	X	X	X	X					X	X	X			X
SK 05	2,2	X	X	X	X	X					X	X	X			X
SK 06	0,7	X	X	X	X	X						X	X	X		X
SK 07	2,6	X	X	X	X	X					X	X	X			X
SK 08	1,3	X	X	X	X	X					X	X	X			X
SK 09	5,5	X	X						TO-5			X	X			X
SK 10	1,2	X	X	X	X	X						X	X			X
SK 11	1,6	X	X							X						X
SK 12	4,2	X	X							SOT32	X					X
SK 14	0,9	X	X	X	X	X						X	X	X		X
SK 15	0,7	X	X							DRAAD						X
SK 16	1,7	X	X	X	X	X						X	X	X		X
SK 17	3	X	X							DRAAD	X					X
SK 18	2,5	X	X	X	X	X						X	X	X		X
SK 19	1,9	X	X	X	X	X						X	X	X		X
SK 31	2,5	X	X	X	X	X						X	X	X	X	X
SK 34	1	X	X	X	X	X						X	X	X		X
SK 35	2	X	X	X	X	X						X	X	X		X
SK 36	1,9	X	X	X	X	X						X	X	X		X
SK 37	13,9		X							DIODE	X					
SK 38	-		X	X				X						X		

Voor het bestellen van bovenstaande koelelementen is het bestelnummer als volgt samen te stellen:
 SERIENUMMER/LENGTE/OPPERVLAKTE-BEWERKING/BORING(-EN). Voorbeeld: SK16/75/SA/2xTO-66 is een koelelement uit de SK16 serie met een zwart geëloxeerd oppervlak, een lengte van 75mm en een boring voor twee TO-66 transistoren. Voor voorraadtypen zie de met een * aangegeven typen op de prijslijst.

Montage onderdelen voor halfgeleiders en koelelementen.



De bouwelementen IS1, IS2 en IS3 zijn bedoeld voor de verticale montage (type IS1, IS2 en IS3) danwel voor de horizontale montage (type IS3) van koelelementen uit de series SK01, SK02, SK03, SK08, SK11, SK14, SK17, SK34 en SK36. Het type IS1 is alleen geschikt voor die koellichamen, met een lengte van 50 mm; de beide andere typen zijn geschikt voor gebruik bij alle andere lengten. Deze montagedelen zijn uit polypropyleen vervaardigd.

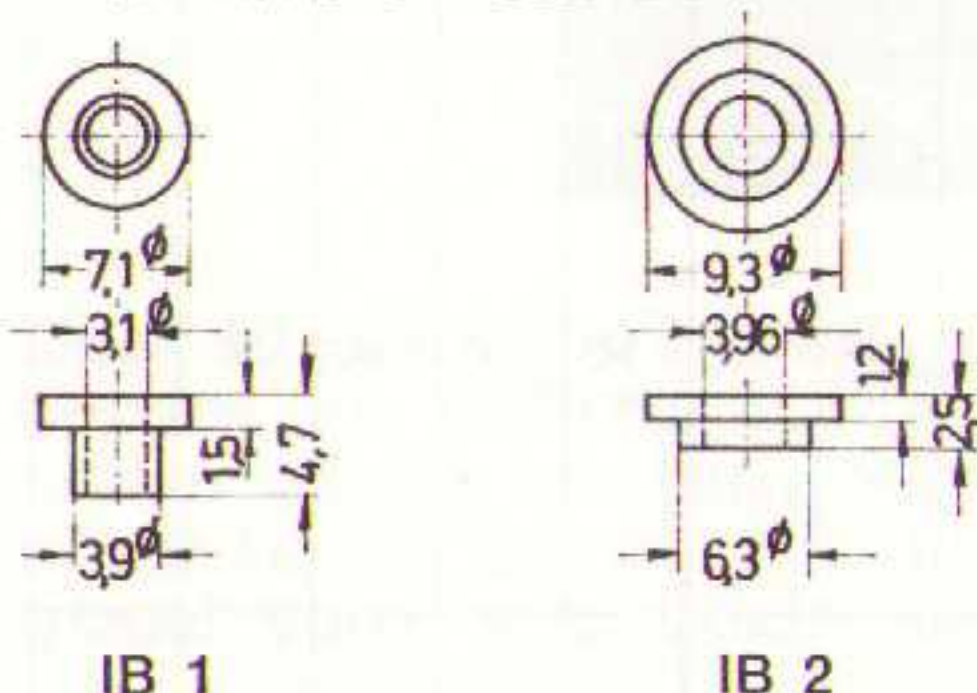


Warmtegeleidingspasta.

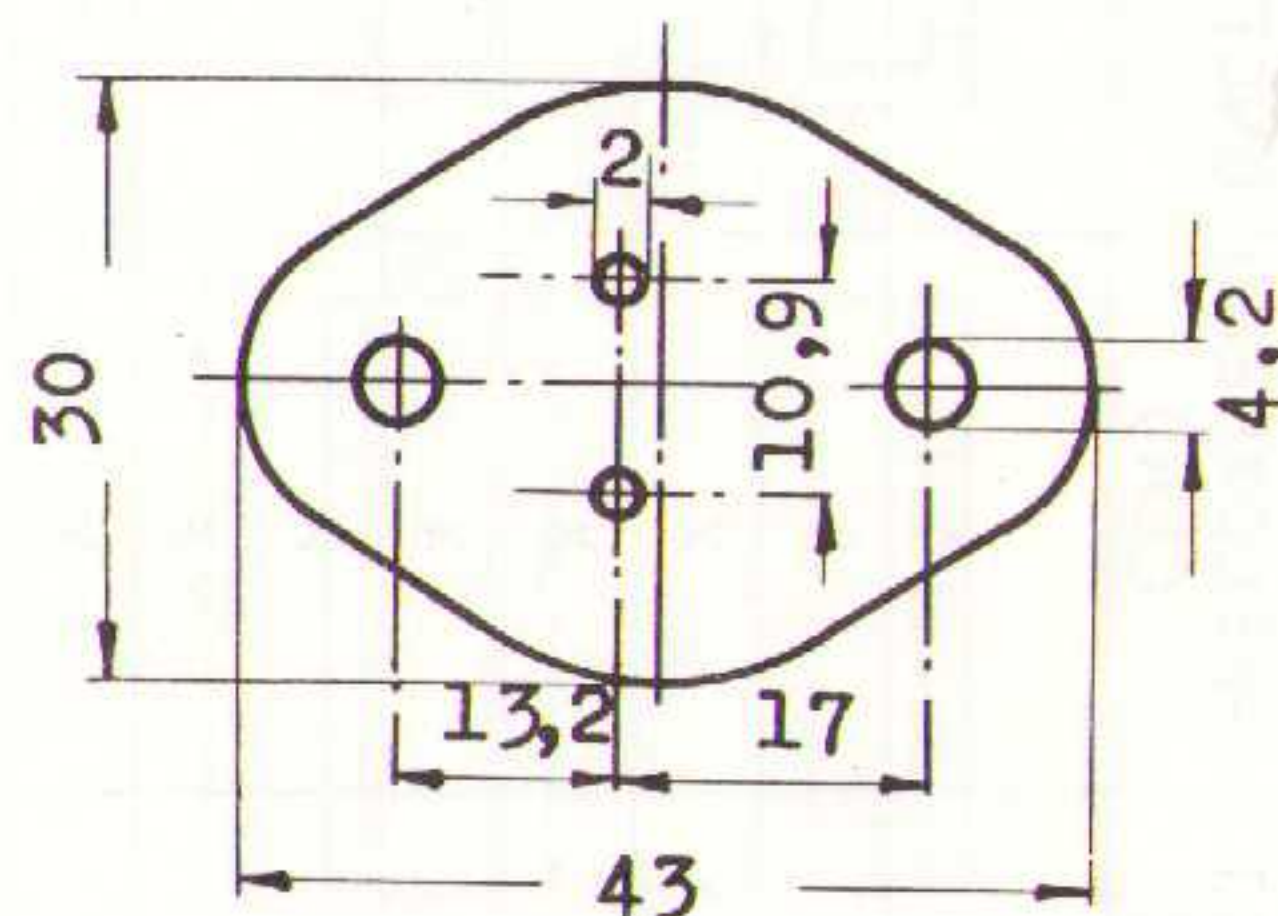
Deze warmtegeleidingspasta, ook wel siliconenpasta genoemd, verbetert de warmteoverdracht tussen uw halfgeleider en koellichaam met meer dan 50%. Leverbaar in doosjes van 4 gram, 35 gram en 500 gram. Deze pasta is ook bijzonder goed toe te passen in combinatie met de aluminium-oxyde onderlegplaatjes AOS3, AOS5 en AOS18.

Mica onderlegplaten en isolatiebusjes voor TO-3, TO-66 en SOT9 transistoren.

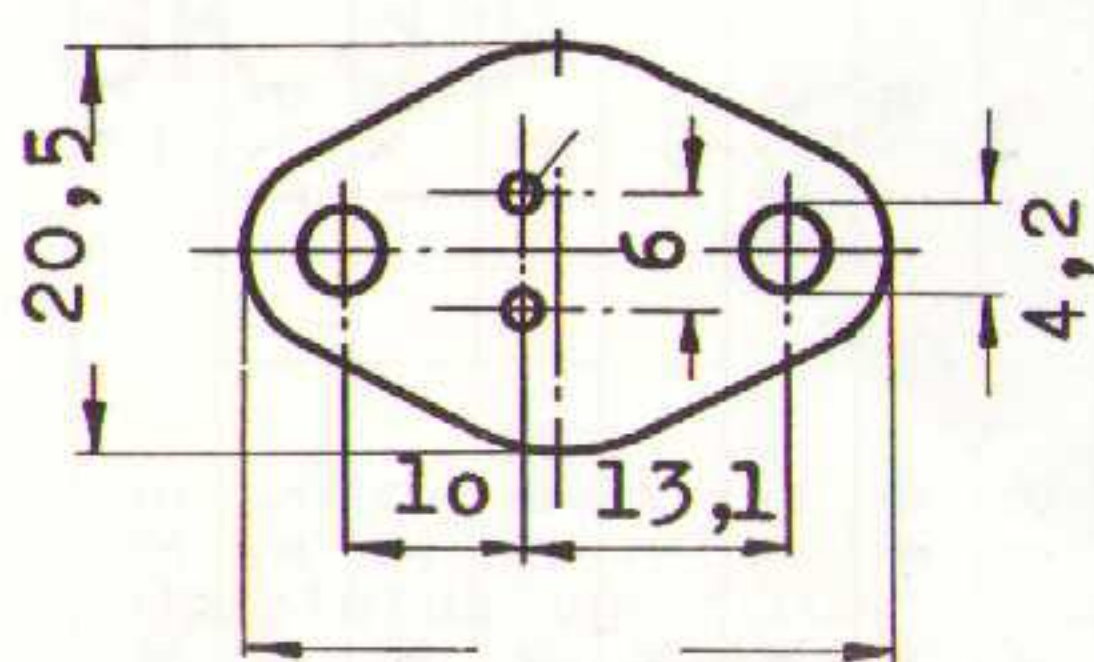
De dikte van deze plaatjes is 0,05mm.



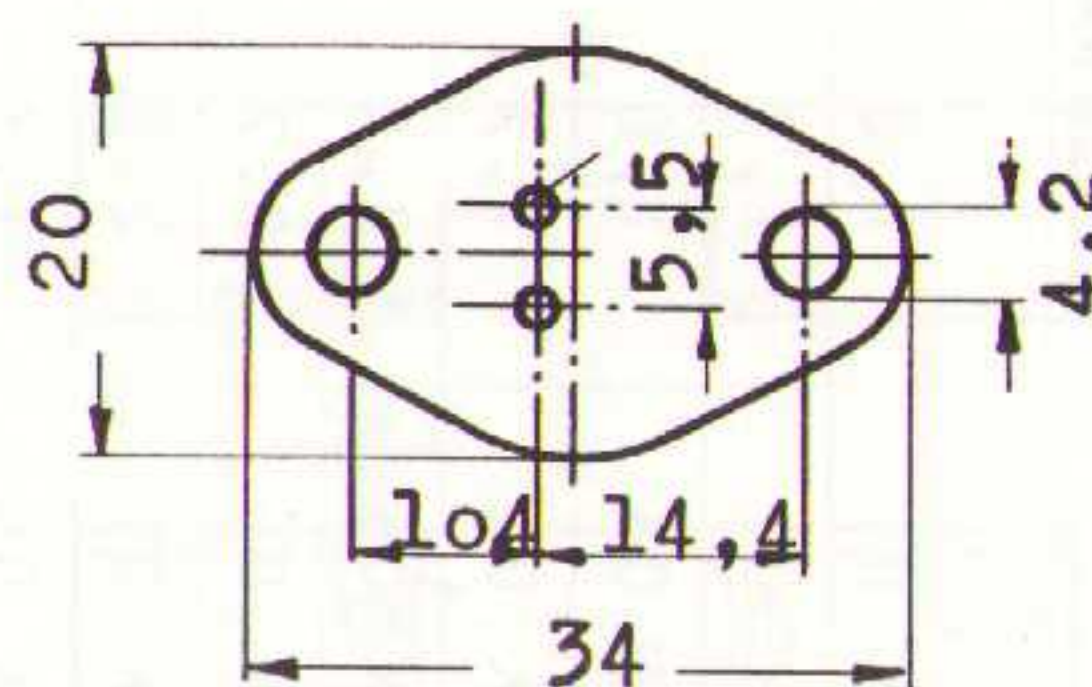
Materiaal isolatiebussen:
Glasvezelversterkte polyamide, tot 180°C vormvast



Voor TO-3 behuizing.
Typenummer: GS-3



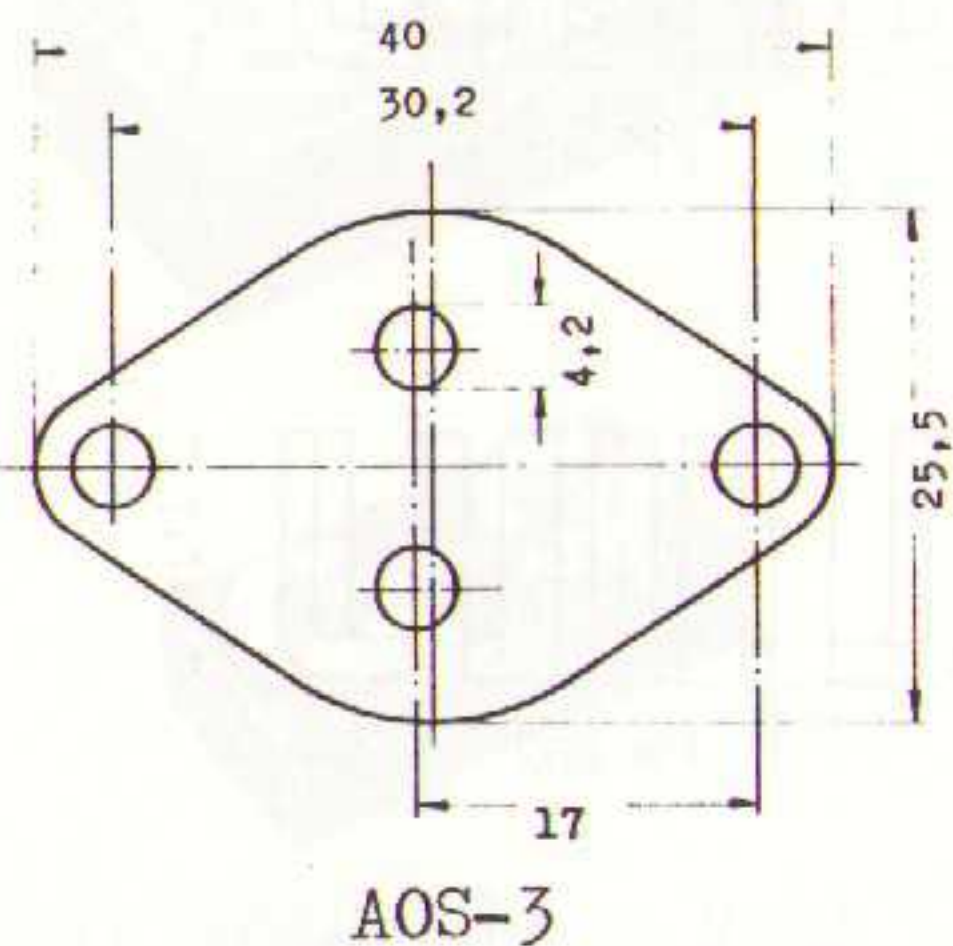
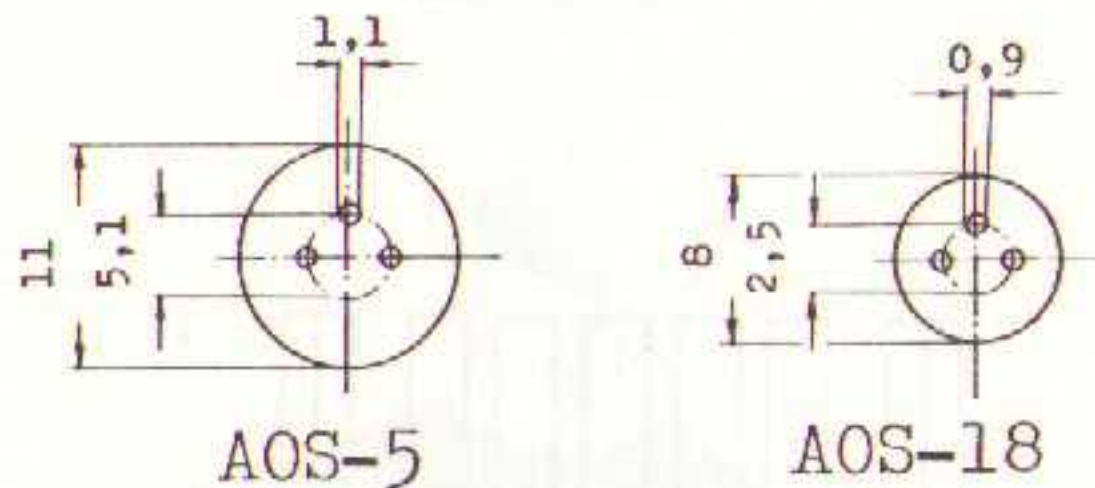
Voor SOT-9 behuizing
Typenummer: GS-9



Voor TO-66 behuizing
Typenummer: GS-66

Aluminium onderlegschijsven.

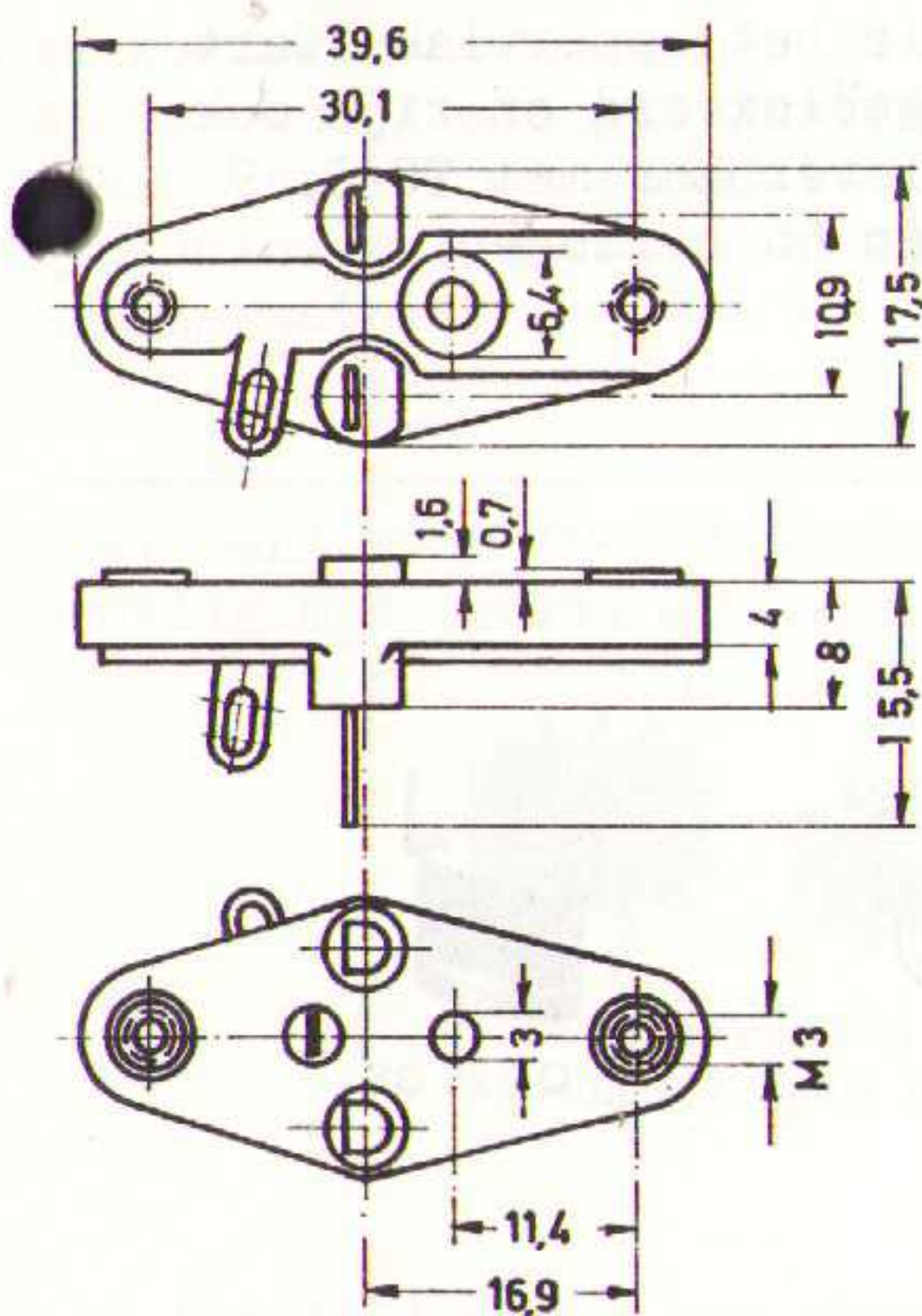
De onderlegschijsven van aluminium-oxyde-keramiek zijn warmtegeleidend en onderscheiden zich van de beryllium-oxyde typen door een lage prijs. De warmteweerstand in $^{\circ}\text{C}/\text{W}$ is meer dan een factor 8 beter dan toepassing van mica isolatieplaatjes (GS-3, GS-9 en GS-66).



Enkele technische gegevens:

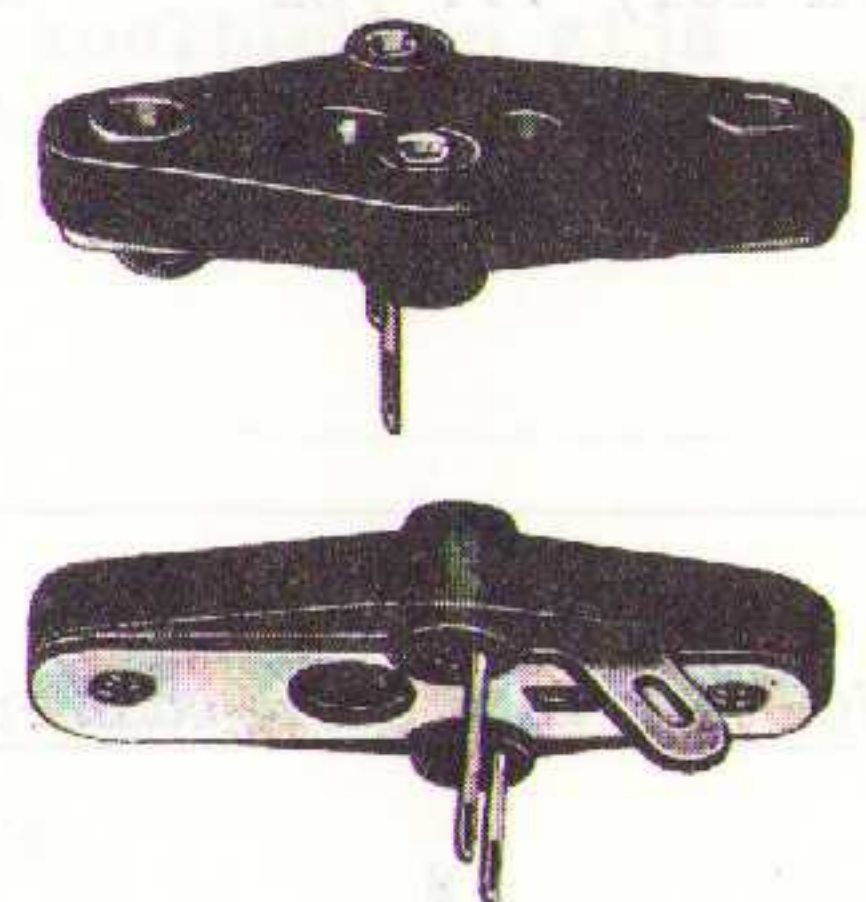
1. warmteweerstand tussen $+20^{\circ}\text{C}$ en $+100^{\circ}\text{C}$ 0,2-0,25
2. overslagspanning bij 50Hz (effectief per cm).. 200-250kV
3. isolatieweerstand bij 50Hz en $+20^{\circ}\text{C}$ $>10^{14} \Omega/\text{cm}$
4. diëlektrische constante (relatief) 9,5
5. diëlektrische dissipatiefaktor bij $+20^{\circ}\text{C}$ in 10^{-3} :

bij 50 Hz	 0,5	bij 1000 MHz	.. 0,15-0,6
bij 1 MHz	 0,1-0,2	bij 5000 MHz	.. 0,12-0,5
bij 10 MHz	 0,1-0,4		

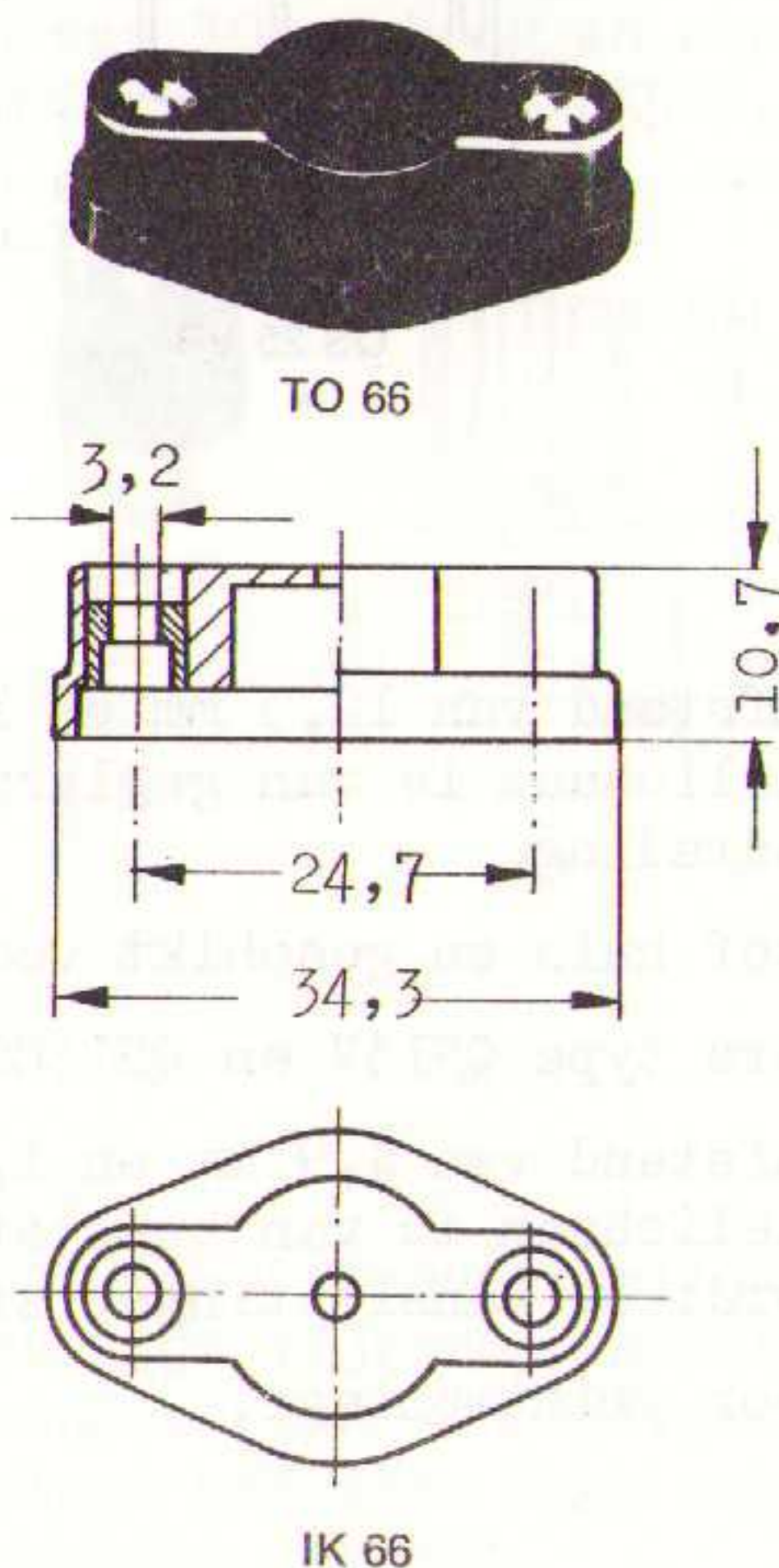
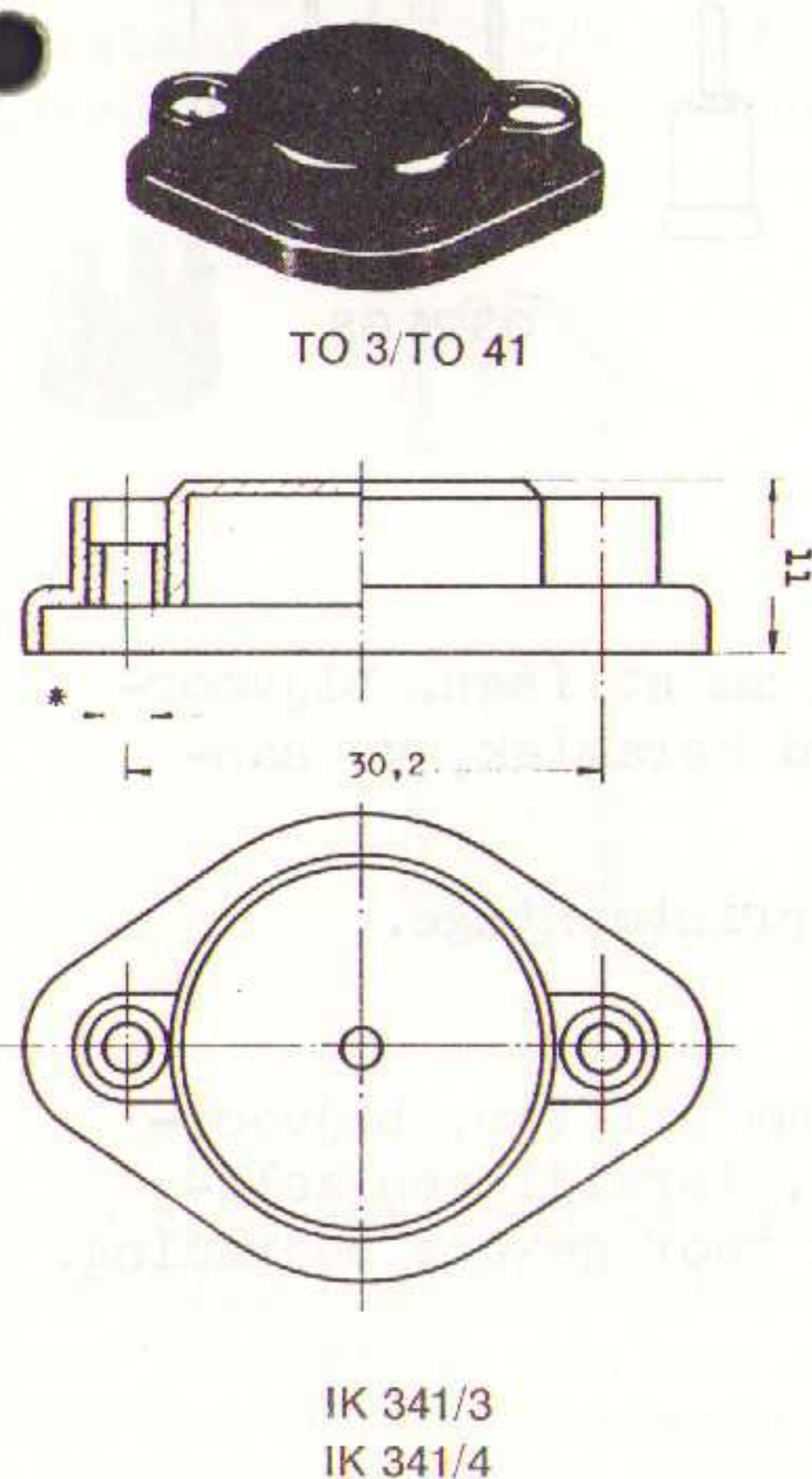


Insteeekvoet voor TO-3 transistors.

Isolatiemateriaal: noryl.
 Kontaktveren vervaardigd van fosforbrons, 3μ verzilverd en $0,5\mu$ verguld.
 Bevestigingsflens vervaardigd van messing Ms 63.
 Overgangsweerstand: 10m Ω
 Capaciteit: 1 pF



TF 3/2

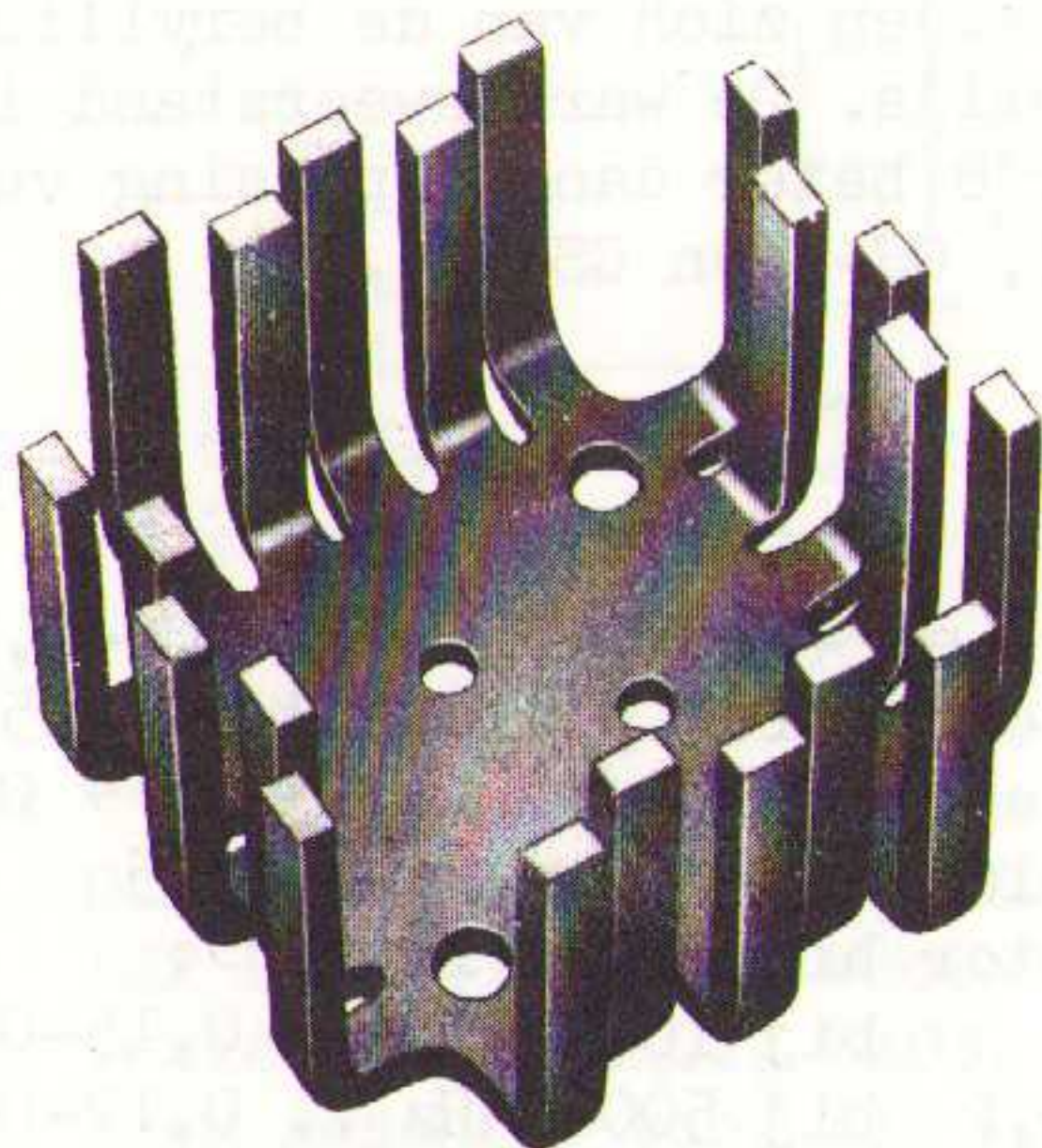


Isolatiekappen voor TO3, TO41 en TO66.

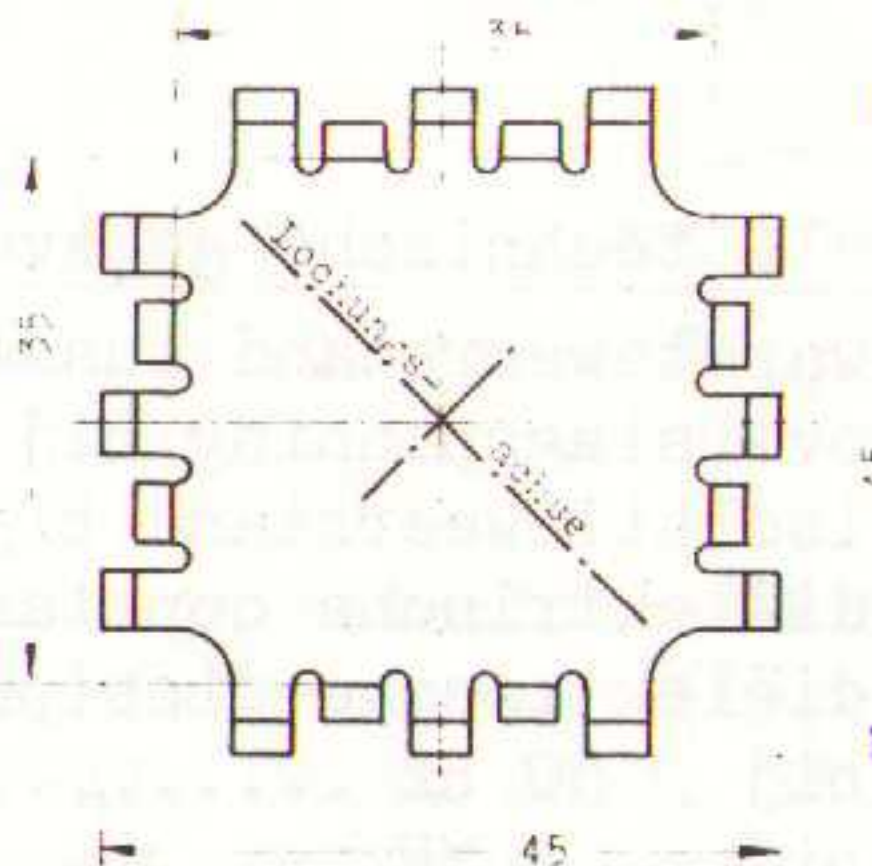
Deze kappen zijn voor het isoleren van uw transistoren ter voorkoming van een sluiting transistor - chassis of transistor - koelelement. Vervaardigd uit polyamide GV, continue bestendig tot 140°C . De geleidebussen * van de uitvoeringen IK341/3 en IK341/4 zijn van aluminium, die van de IK66 van messing.

Bij de IK341/4 worden oppervlakte veredelde boutjes volgens DIN84 medegeleverd. De doorsnede van de geleidebussen is:

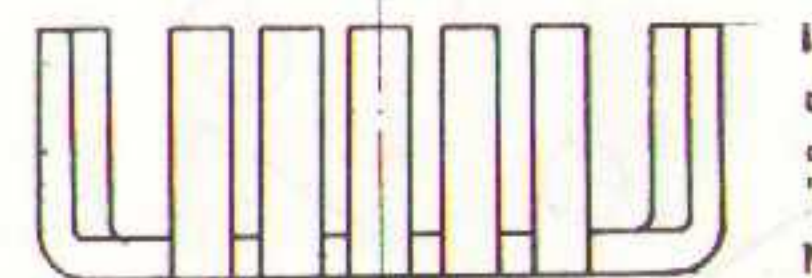
type IK 341/3	..	3,2 mm
type IK 341/4	..	4,2 mm
type IK 66	..	3,2 mm



Vingerkoelelement
matr.: aluminium.



FK 201

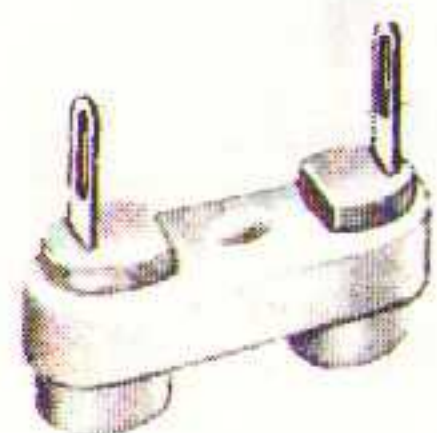


FK 202

Artikelnummer	afmetingen	thermische weerstand
FK 201/ ... /SA	45x45x25,4	6°C/W
FK 301/ ... /SA	45x45x12,5	8°C/W

Van deze koelelementen is het oppervlak zwart geëloxeerd en zijn ook leverbaar met T0-3, 9 en 66 boringen.

Voeten voor kristallen.



QS 13 V



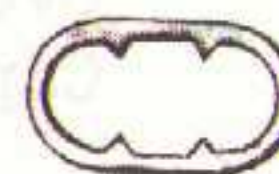
EK 1



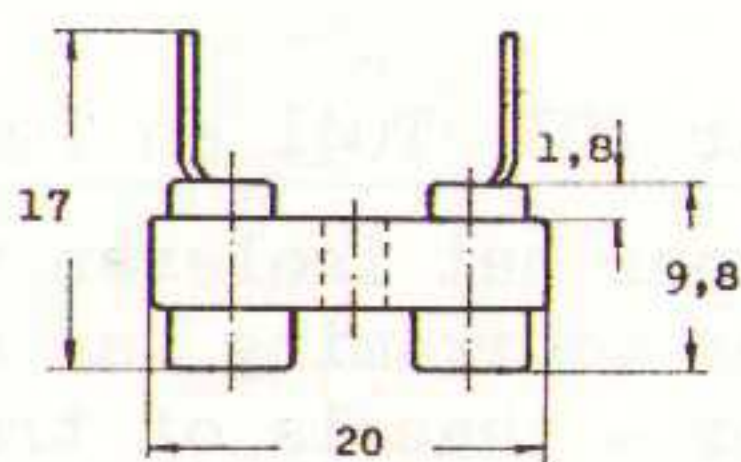
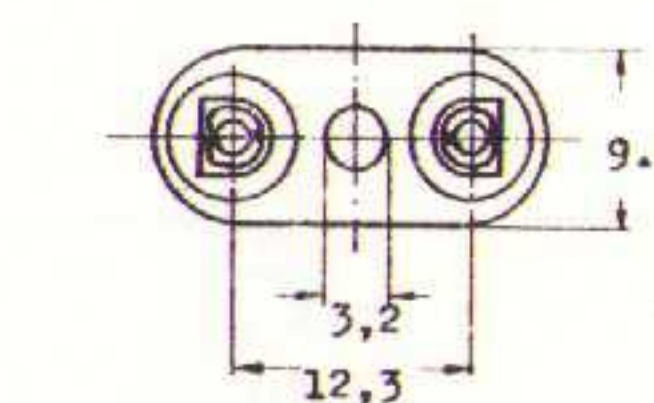
QS 13 GS



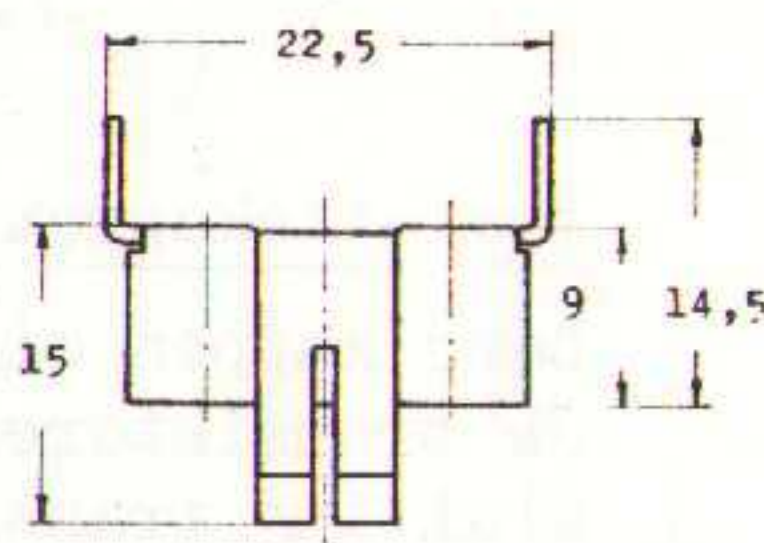
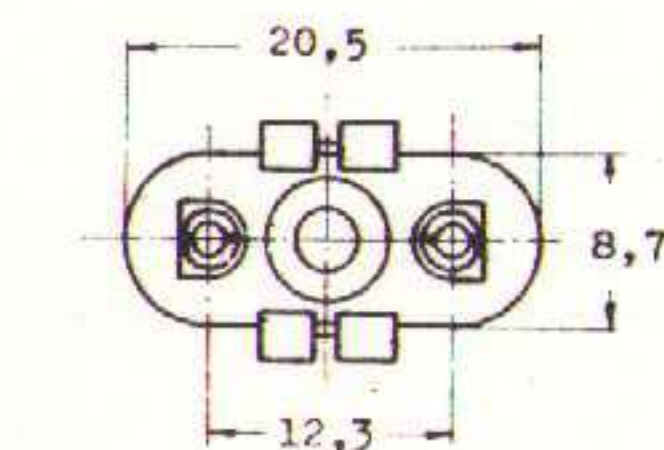
QS 25 VR



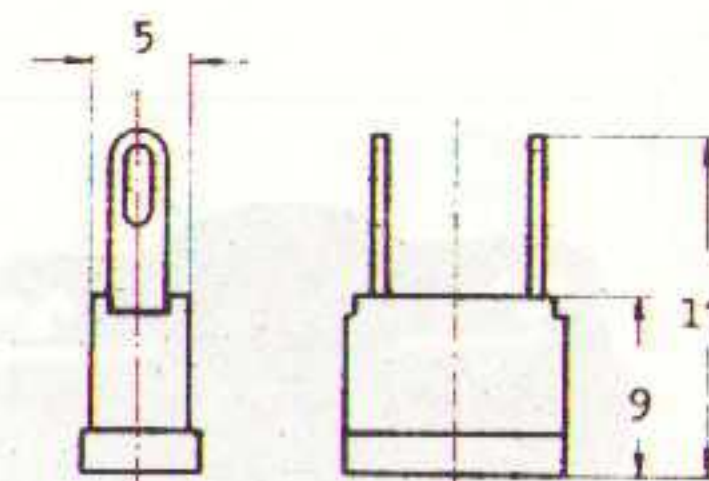
QS 25 GS



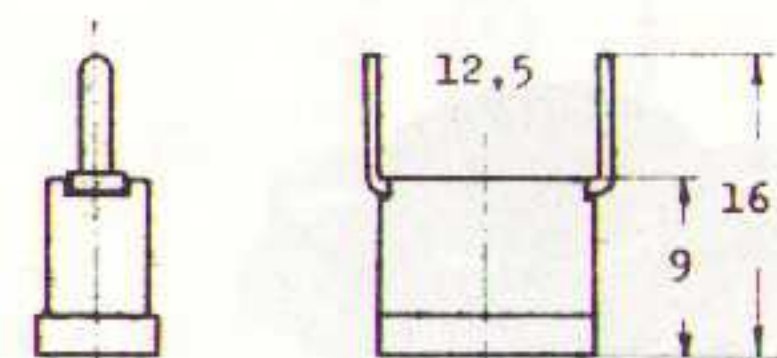
QS 13 V



QS 13 GS EK 1

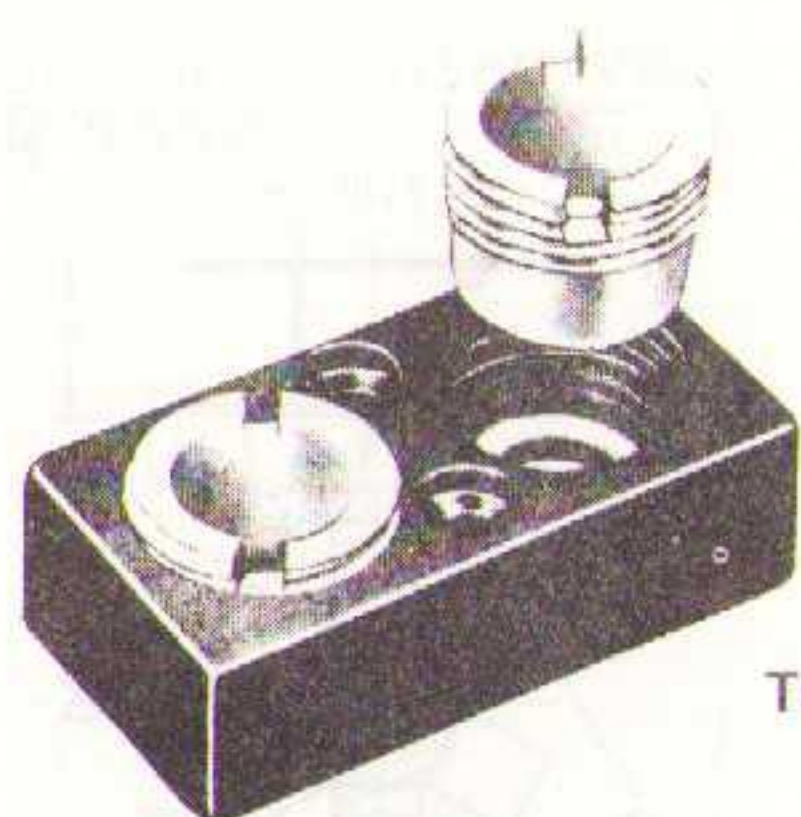


QS 25 VR



QS 25 GS

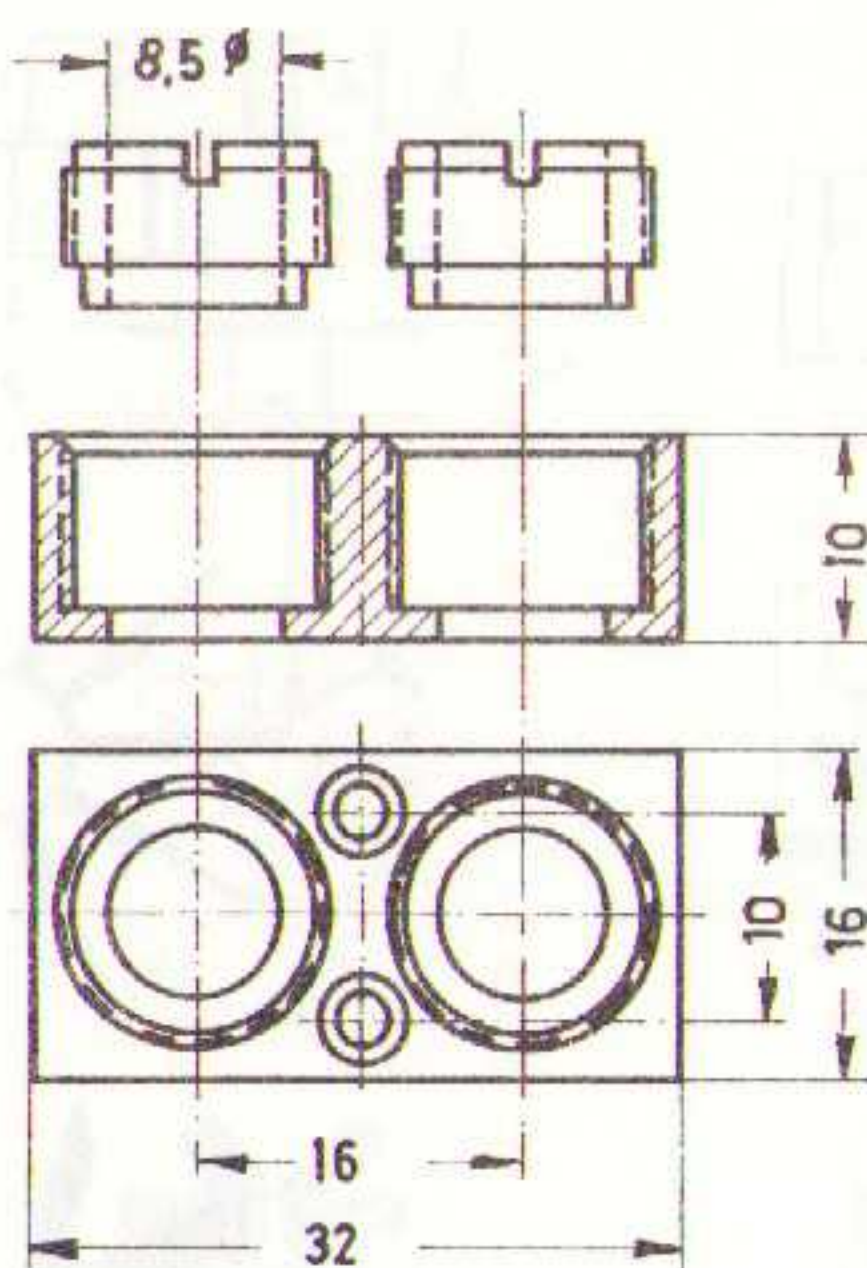
- QS 13V voor kristallen met een stiftafstand van 12,3 mm en 1,3 mm stiften, bijvoorbeeld voor type HC13U. Isolatielichaam is van geglazuurd keramiek, de aansluitingen zijn voor gewone bedrading.
- QS13GS als QS13 V, echter met kunststof huis en geschikt voor printmontage.
- EK 1 aardingsklem, passend op houders type QS13V en QS13GS.
- QS 25VR voor kristallen met een stiftafstand van 4,9 mm en 1,0 mm stiften, bijvoorbeeld voor type HC25U. Isolatielichaam is van kunststof, terwijl een zelfklemmende ring wordt bijgeleverd; de aansluitingen zijn voor gewone bedrading.
- QS25GS als QS25VR, echter geschikt voor printmontage.



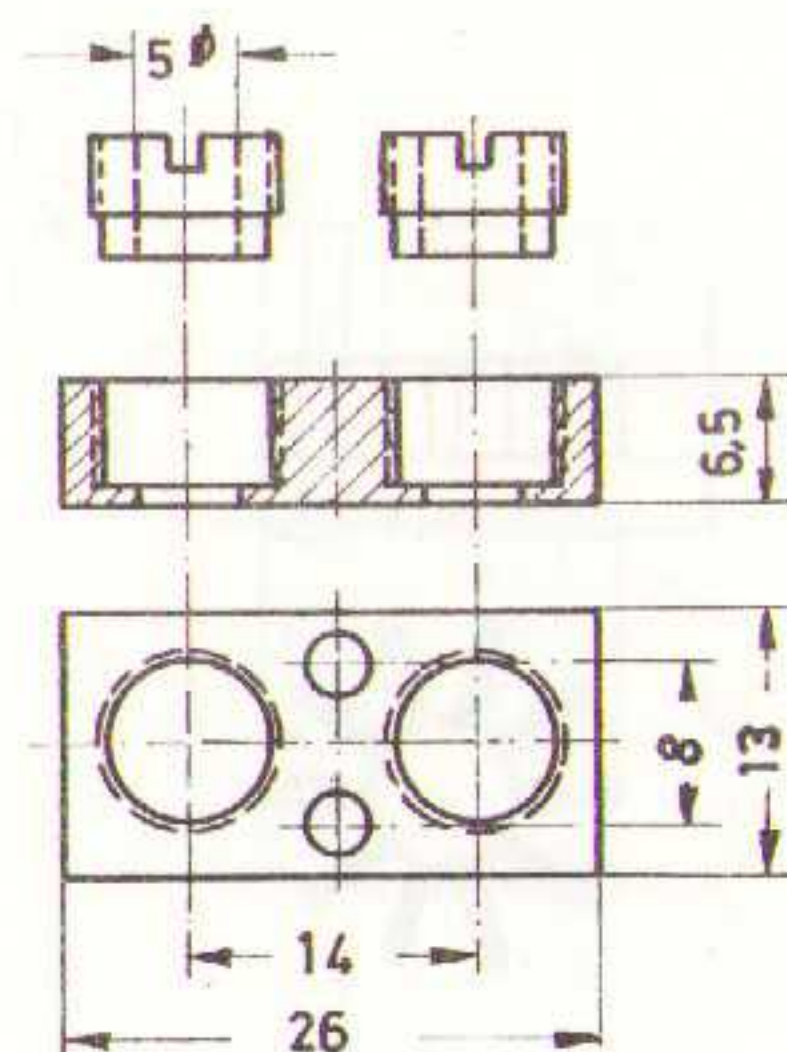
TO 5



TO 18



KK 25

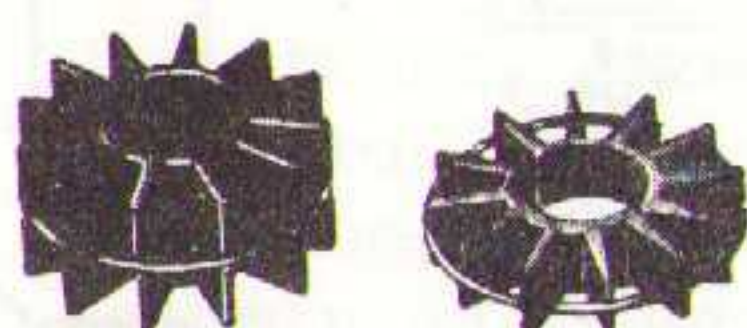


KK 218

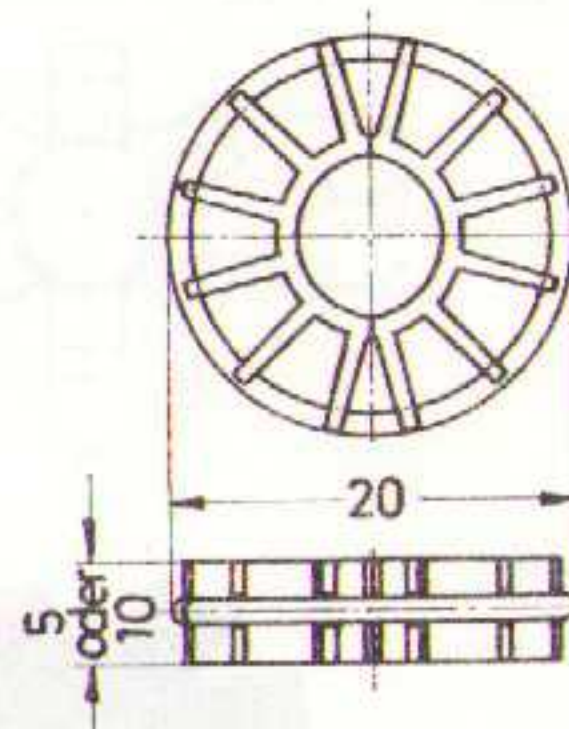
Dubbele koelblokken voor TO-5 en TO-18 transistoren.

Deze koelelementen kunnen worden gebruikt op plaatsen, waar twee (complementaire) transistoren een grote gelijkheid in temperatuur dienen te behouden. Deze koelblokken zijn geschikt voor TO-5 en TO-18 transistoren. De bevestigingsschroef is van nylon, terwijl het aluminiumblok voorzien is van een isolerende matzwarte oxyde, welke tot 500 volt doorslagvast is.

Koelelement voor TO-5 transistoren tweedelig met veerring.



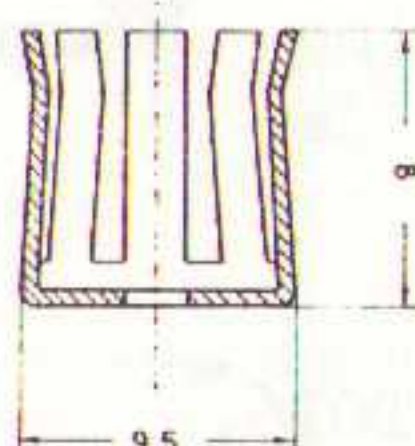
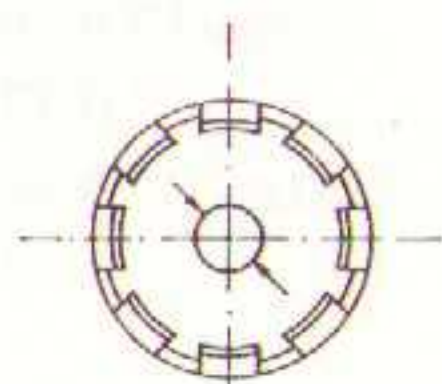
2-teilig met Federring
Aluminium schwarz eloxiert



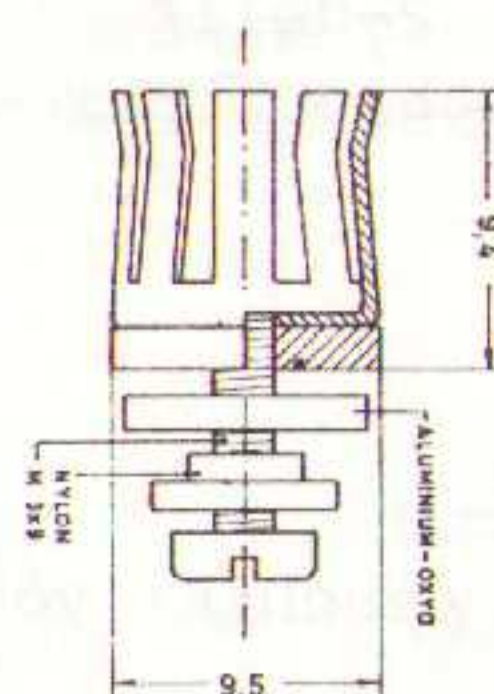
Deze tweedelige met een veerring uitgevoerde flexibele uitvoering geeft een bijzonder goed warmtecontact met Uw transistor, waardoor een goede warmteafvoer bewerkstelligd wordt (38 resp. 52°C/W). Leverbaar in twee typen (KK55 van 5 mm hoog en een warmteverstand van 52°C/W en KK510 van 10 mm hoog en een warmteweerstand van 38°C/W). Wij wijzen U ook naar de typen KK50, KK51, KK52, KK505, KK511 en KK5095.



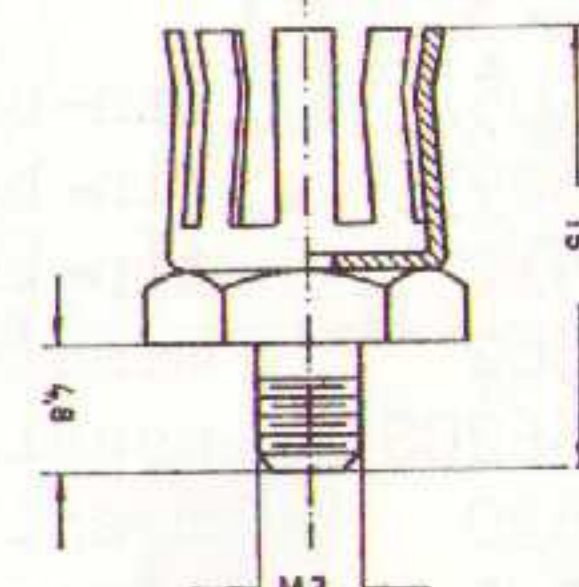
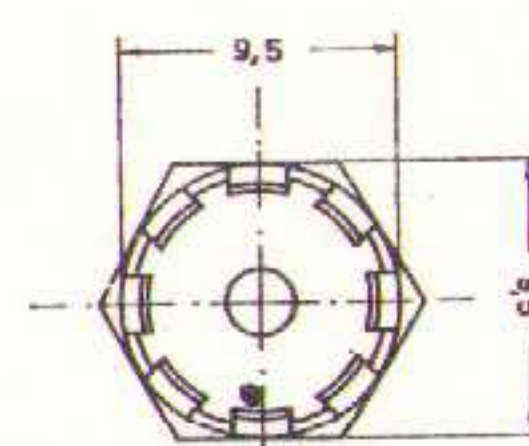
KK 50



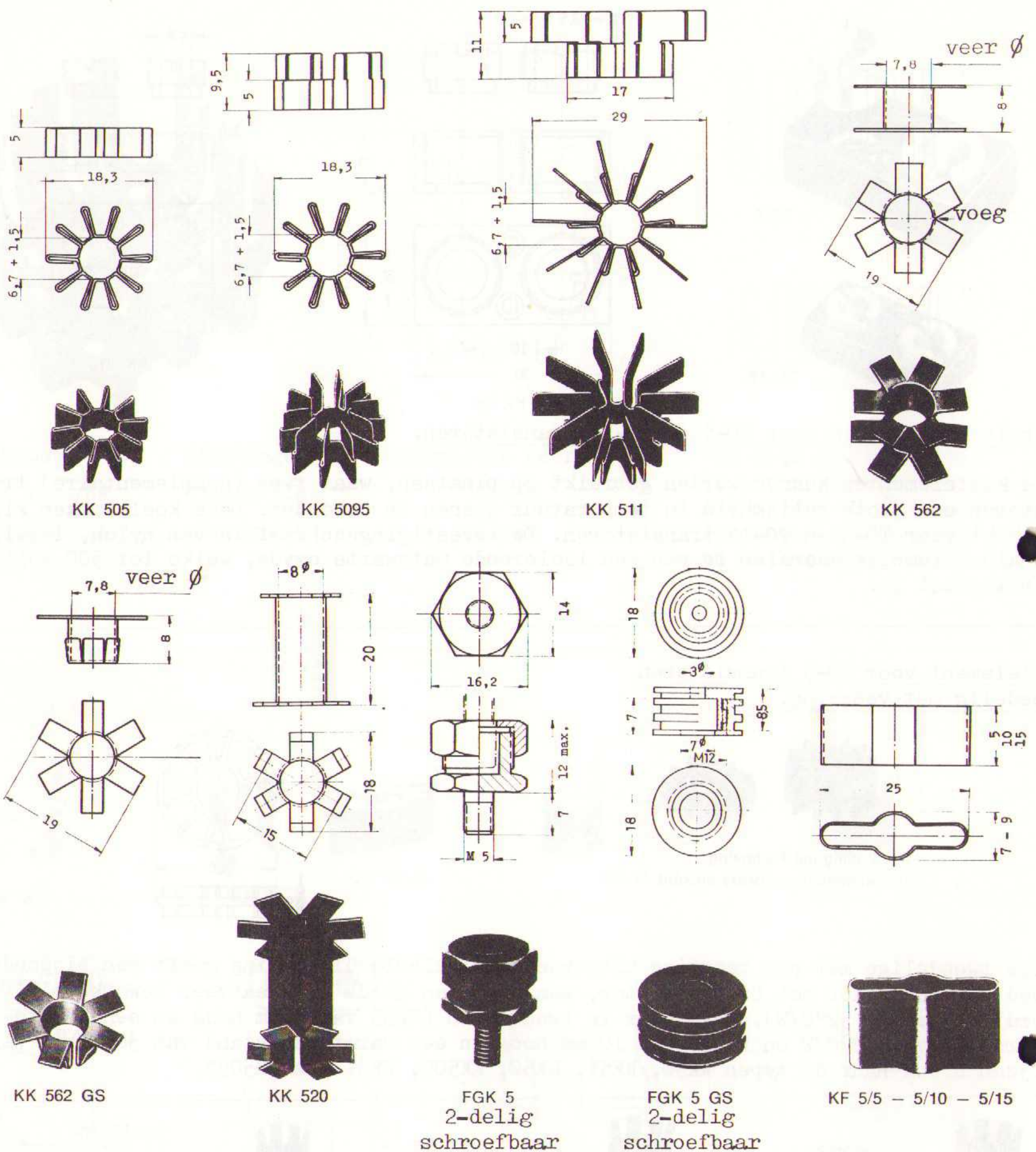
KK 51



KK 52



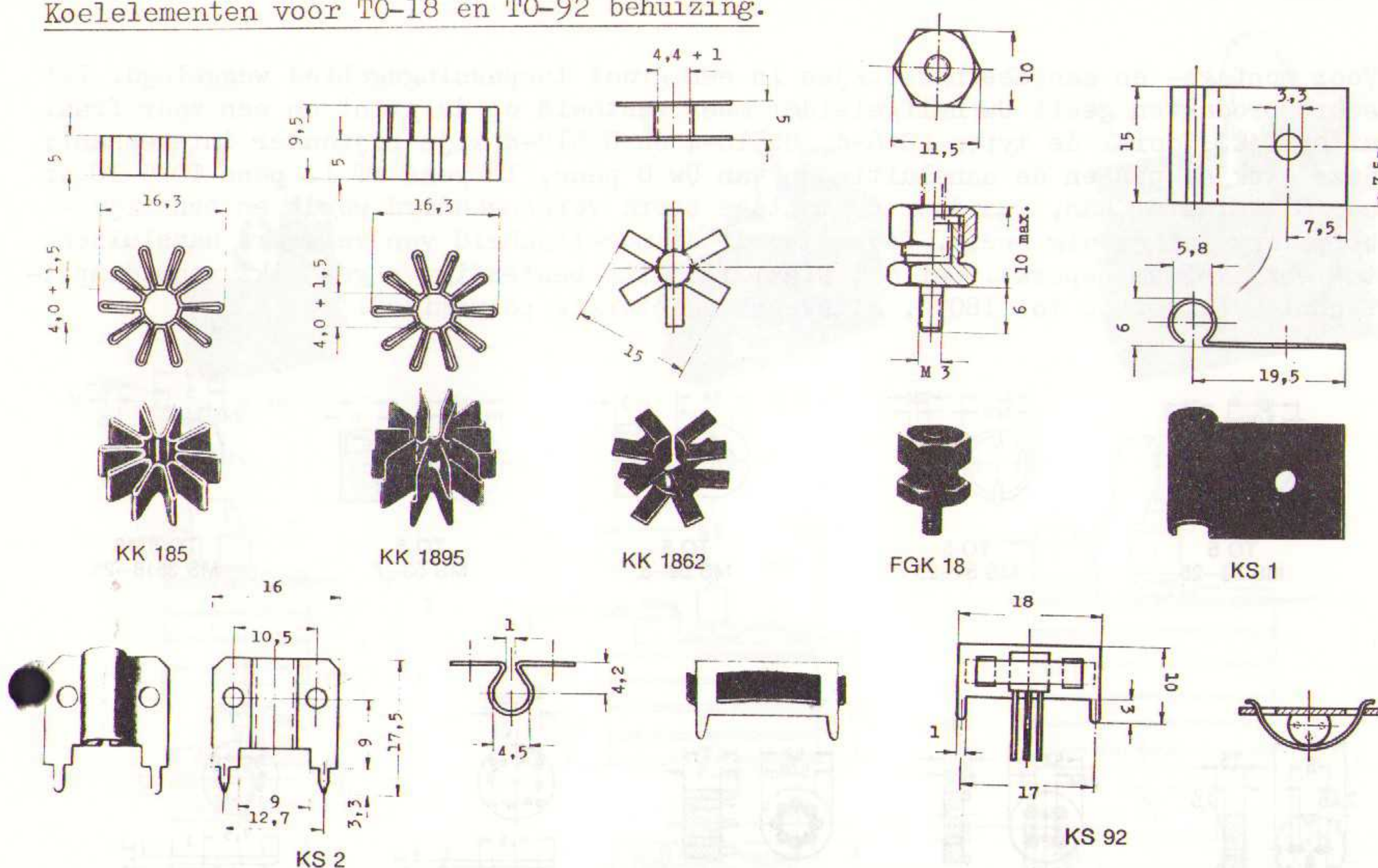
Deze opsteekkoelelementen voor TO-5 of soortgelijke behuizingen bestaan uit koperberyllium. Deze koelelementen zijn bijzonder geschikt voor hoogfrequent toepassing. Door toepassing van een aluminium-oxyde-onderlegschijs en een nylonbus bij het type KK51 kan een geïsoleerde montage met zeer lage capaciteit en een warmteweerstand van slechts 26°C/W mogelijk worden gemaakt.



Koelelementen voor TO-5 behuizingen.

typenummer	materiaal	geschikt voor Jedec behuizing	warmteweerstand
KK 505	tin-brons, gezwart	T05, T09, T011, T012, T016, T029, T033, T039, T042, T043, T045, T055.	60°C/W
KK 5095	tin-brons, gezwart		48°C/W
KK 511	tin-brons, gezwart		33°C/W
KK 562	messing MS95, gezwart		60°C/W
KK 562GS	messing MS95, gezwart		60°C/W
KK 520	messing MS95, gezwart	T05, T09, T012, T016, T033, T039, T042, T043	48°C/W
KF 5/5	koper, gezwart		57°C/W
KF 5/10	koper, gezwart		46°C/W
KF 5/15	koper, gezwart		40°C/W
FGK 5	aluminium, zwart geëloxeerd		42°C/W
FGK 5GS	aluminium, zwart geëloxeerd		42°C/W

Koelelementen voor TO-18 en TO-92 behuizing.



typenummer	materiaal	geschikt voor Jedec huis	warmteweerstand
KK 185	tin-brons, gezwart	TO-18, TO-24, TO40 en TO-44	60°C/W
KK 1895	tin-brons, gezwart	TO-18, TO-24, TO-40 en TO-44	48°C/W
KK 1862	messing, MS95, gezwart	TO-18, TO-24, TO-40 en TO-44	85°C/W
FGK 18	aluminium, zwart geëloxeerd	TO-18	65°C/W
KS 1	koper, gezwart	TO-1	85°C/W
KS 2	koper, vertind	TO-1	80°C/W
KS 92	koper, vertind; incl. staalveer	TO-92	90°C/W

Insteeckvoetjes voor transistoren en geïntegreerde schakelingen.

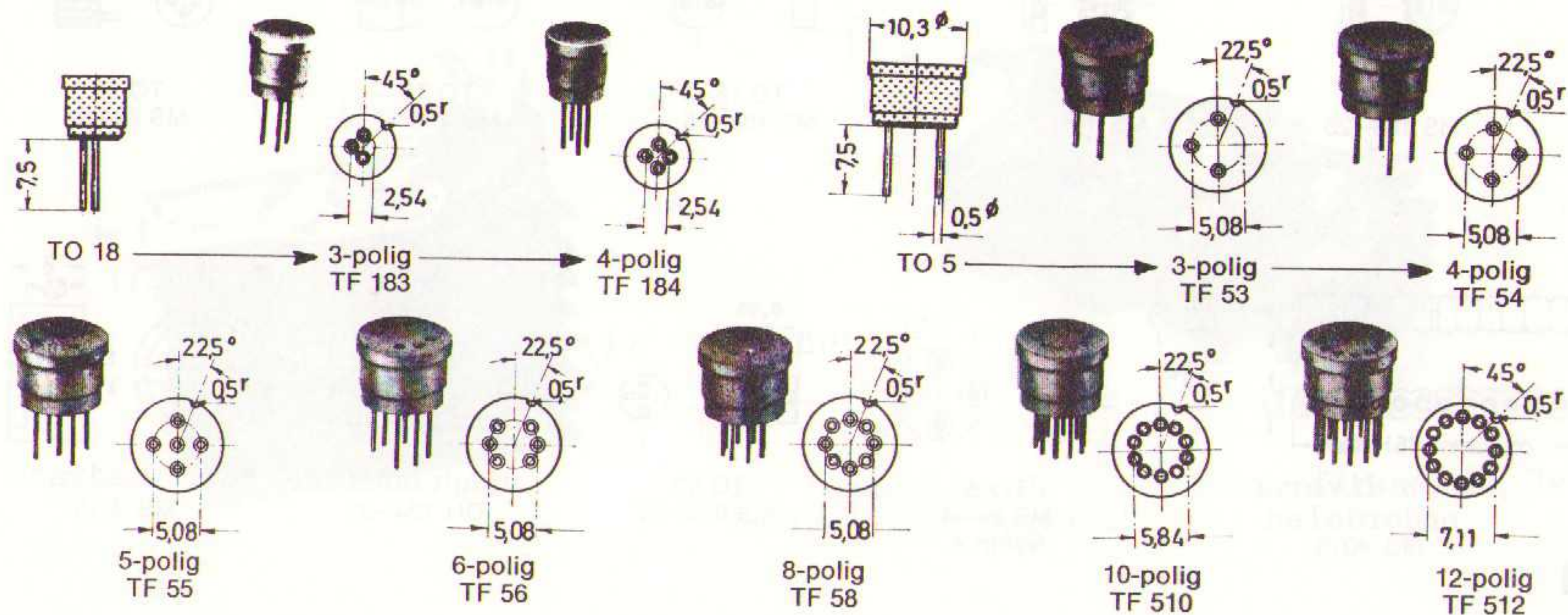
Uitvoeringen voor wire-warp techniek zijn in voorbereiding.

Isolatiemateriaal: polytetrafluoraethyleen (teflon).

Kontaktveren: gewalst fosforbrons, 3μ vernikkeld, 1μ verguld.

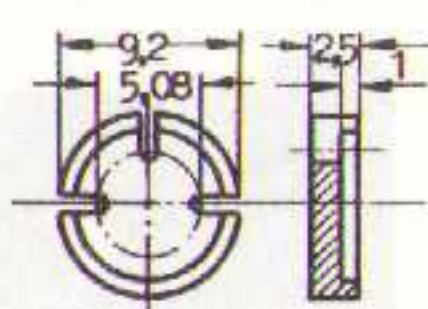
Overgangswaerstand: kleiner dan 10 mΩ.

Capaciteit: 0,4 pF.

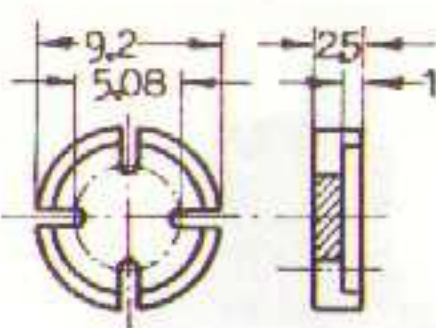


Montage- en aanpassingsstukjes voor halfgeleiders.

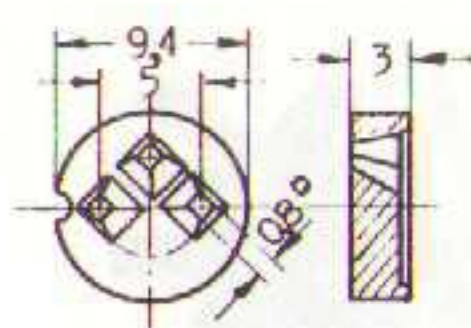
Voor montage- en aanpassingsstukjes is een groot toepassingsgebied weggelegd. Dit soort produkten geeft Uw halfgeleider meer vastheid op de print en een zeer fraai uiterlijk. Vooral de typen US58-4, US510-4 en US512-4 zijn bijzonder interessant; deze stukjes passen de aansluitingen van Uw 8 pens, 10 pens of 12 pens TO-5 IC op een 0,1" raster aan, waardoor de montage sterk vereenvoudigd wordt en een tijdsbesparing tot gevolg heeft. Tevens wordt de mogelijkheid van verkeerd aansluiten tot een minimum beperkt. Het materiaal is hitte bestendig en geschikt voor dompeltechniek. (vormvast tot 180°C, glasvezel versterkte polyamide).



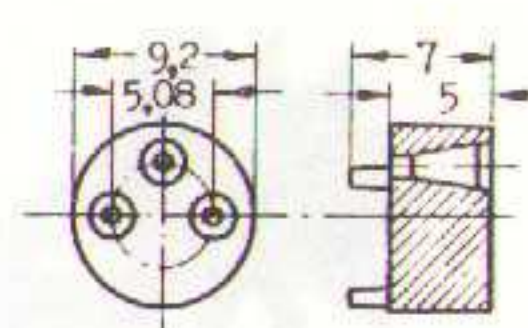
TO 5
MS 53-25



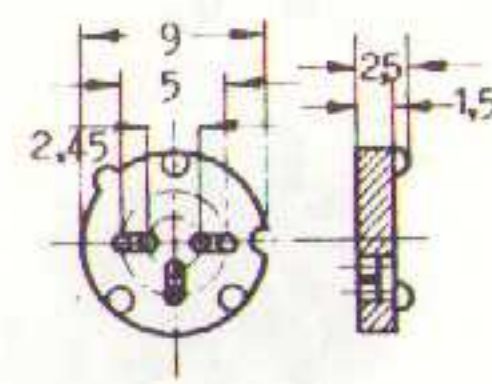
TO 5
MS 54-25



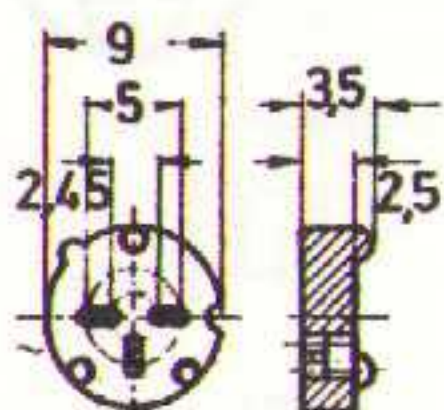
TO 5
MS 53-3



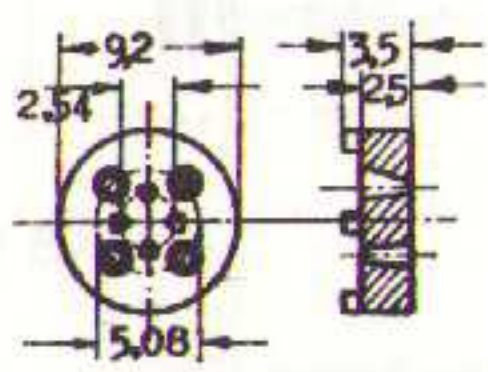
TO 5
MS 53-7



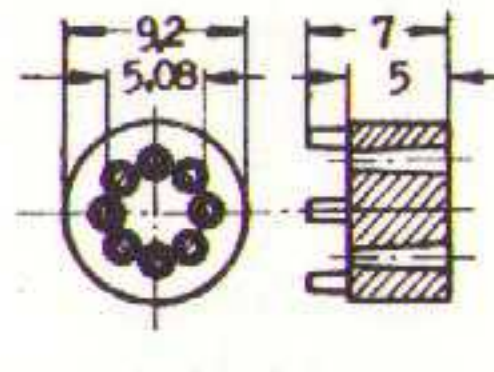
TO 5/18
MS 3518-25



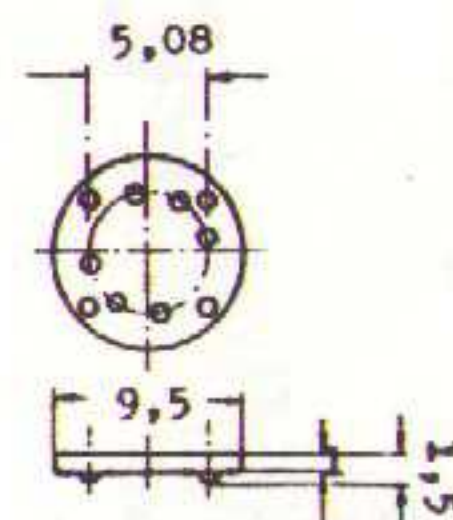
TO 5/18
MS 3518-35



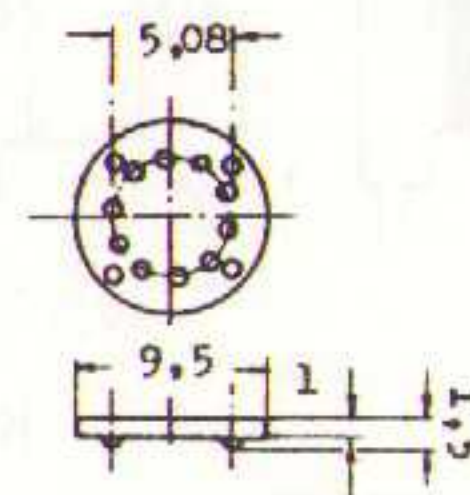
TO 5/18
MS 34-518



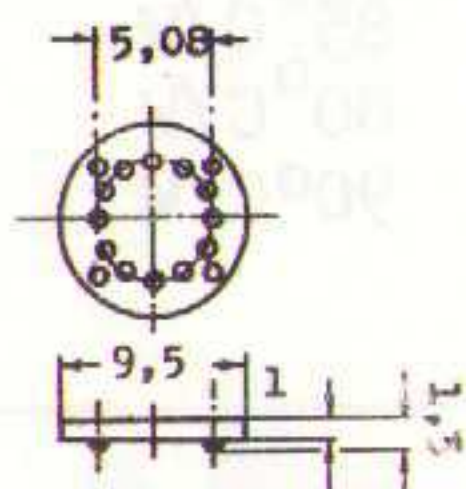
TO 5-8 p.
MS 58-7



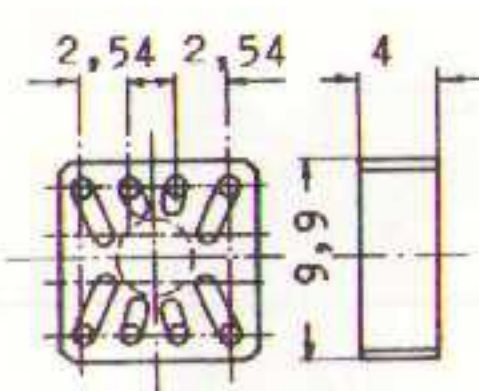
TO 5-6 p.
MS 56-15
Nylon B



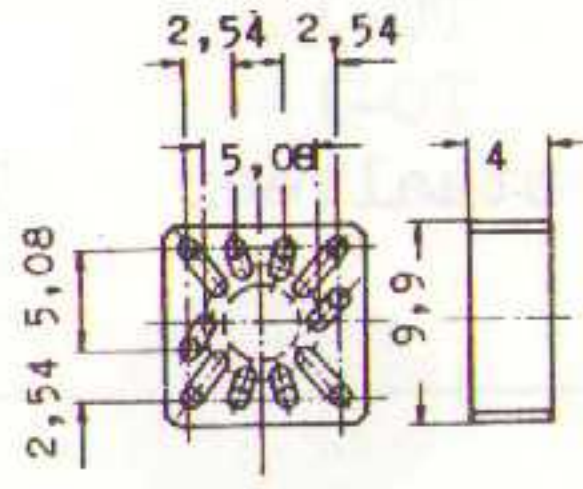
TO 5-10 p.
MS 510-15
Nylon B



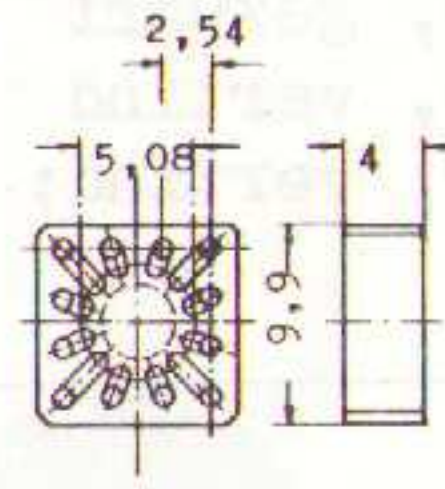
TO 5-12 p.
MS 512-15
Nylon B



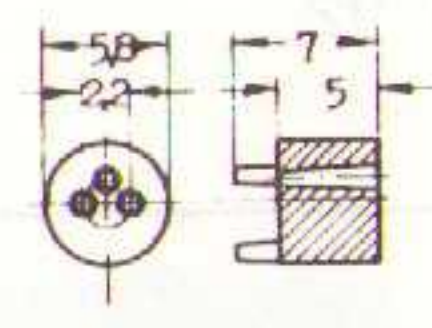
TO 5
US 58-4



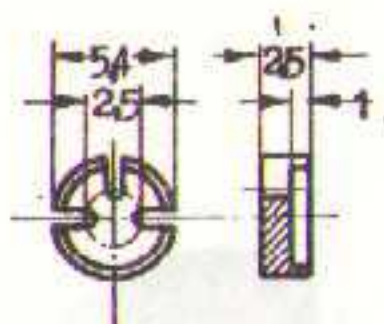
TO 5
US 510-4



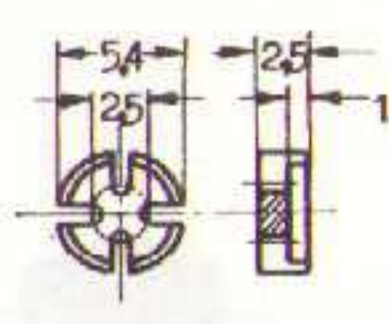
TO 5
US 512-4



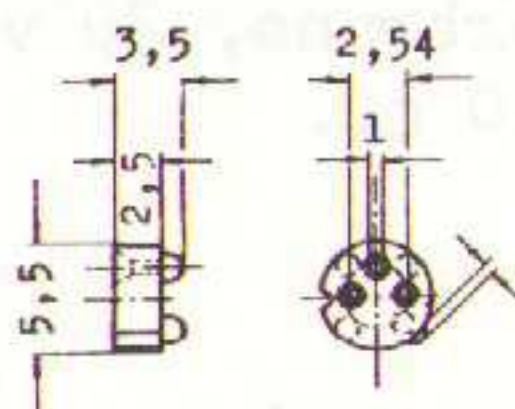
TO 1
MS 13-7



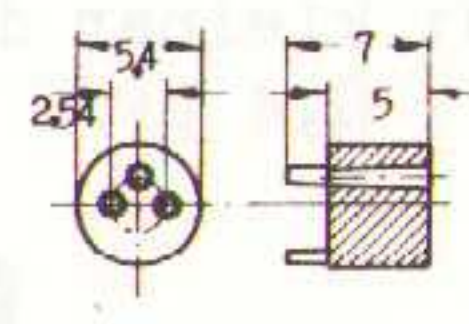
TO 18
MS 183-25



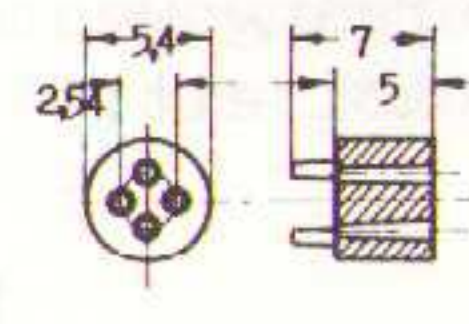
TO 18
MS 184-25



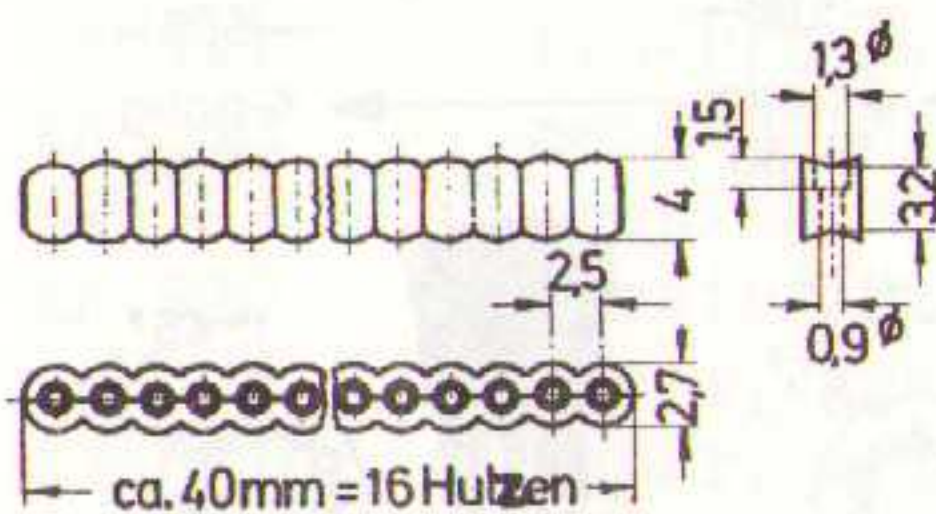
TO 18
MS 183-35



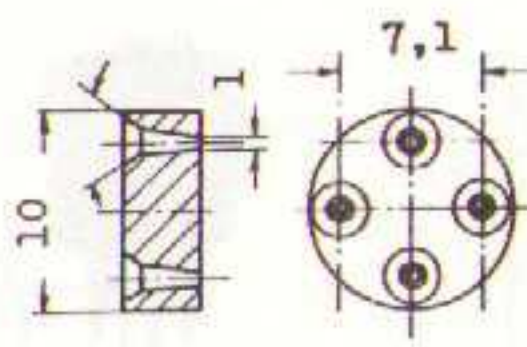
TO 18
MS 183-7



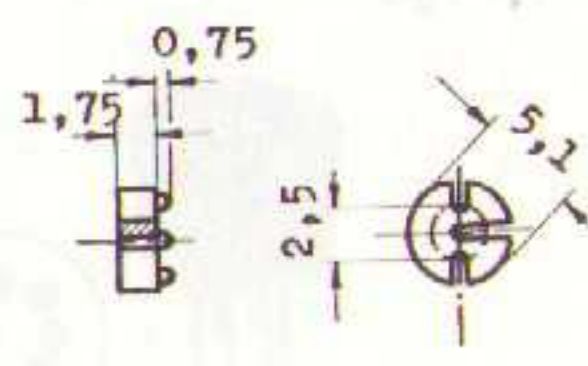
TO 18
MS 184-7



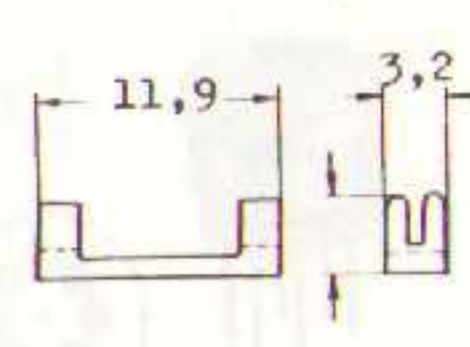
voor diverse
onderdelen
MS 4016



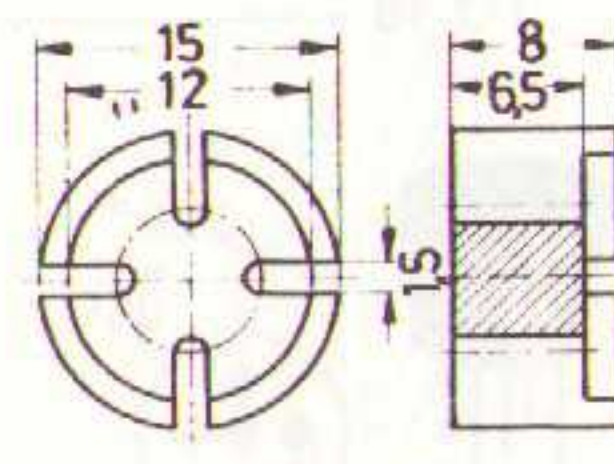
TO 8
MS 84-4
Nylon A



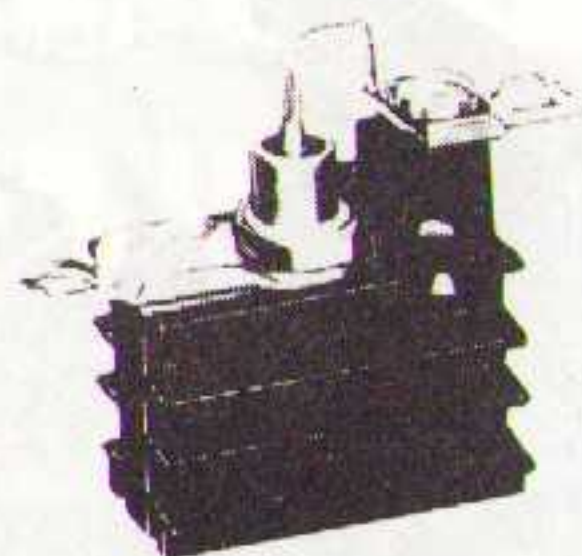
TO 92
MS 923-25



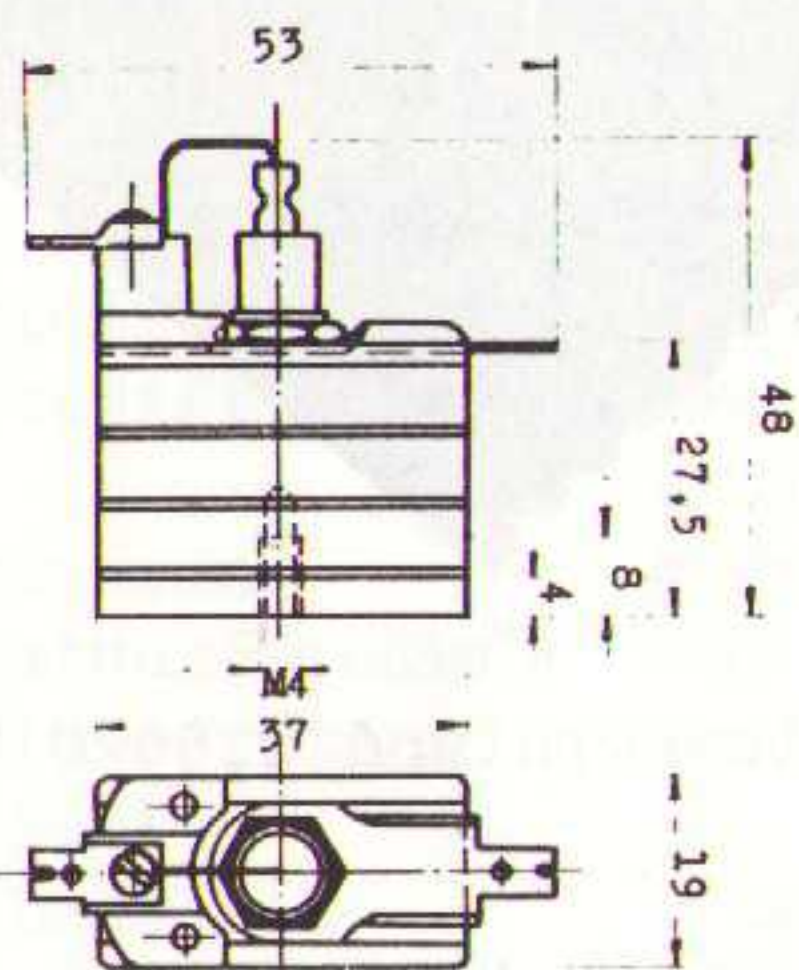
diodehouder
DH 124-32



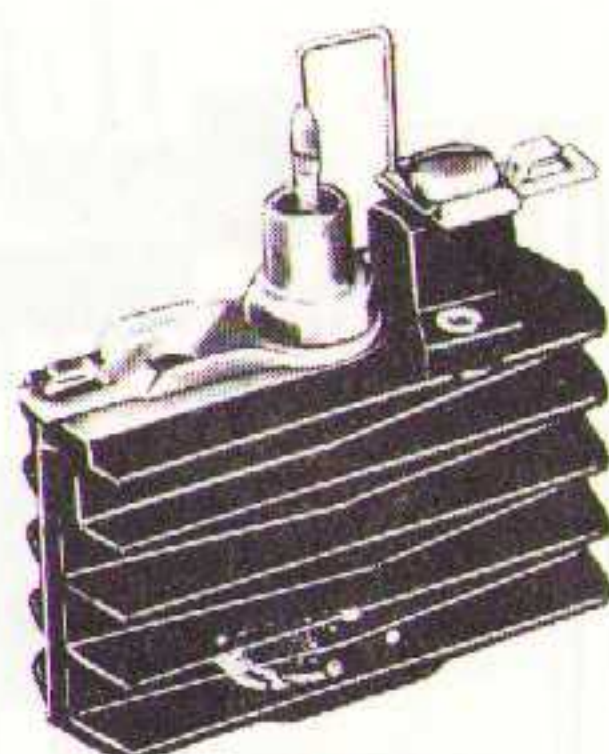
voor Quadracs
MS 4/15



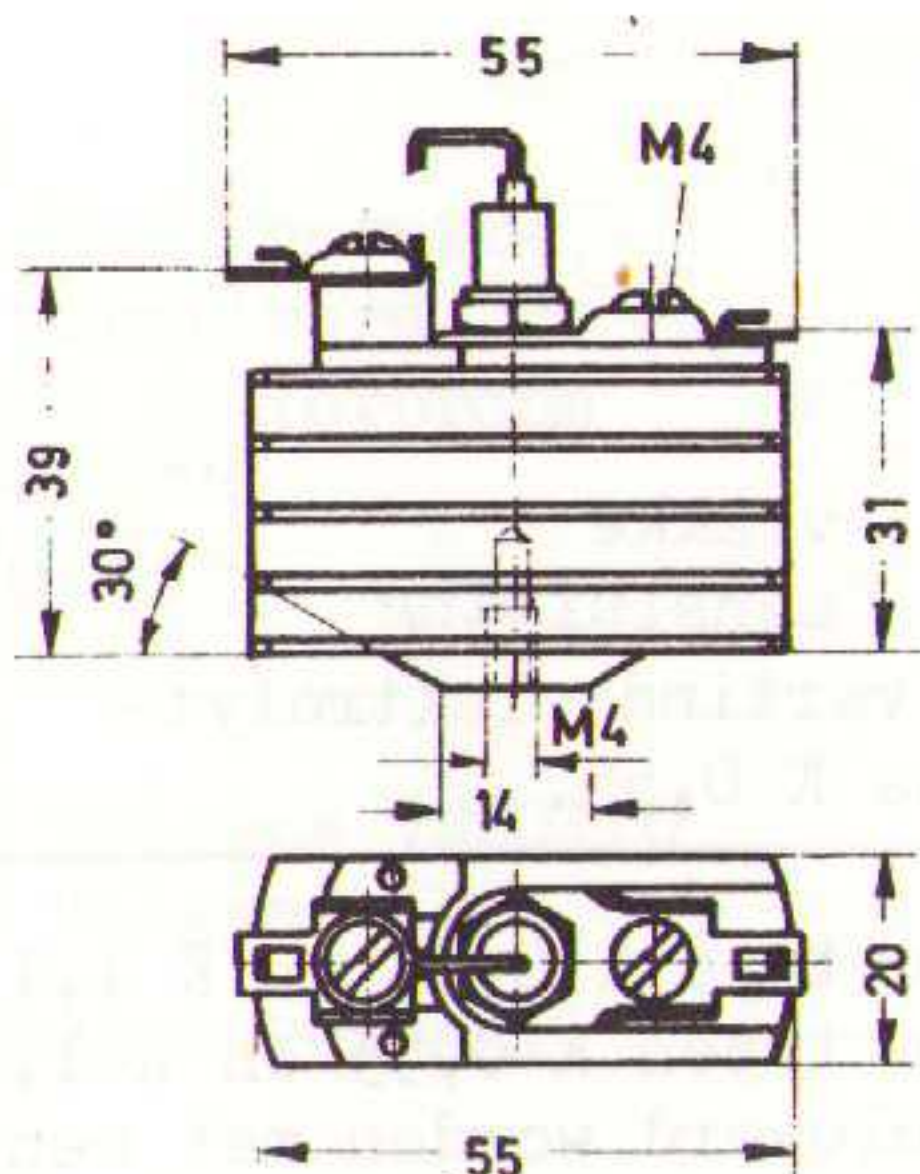
K 15



K 15



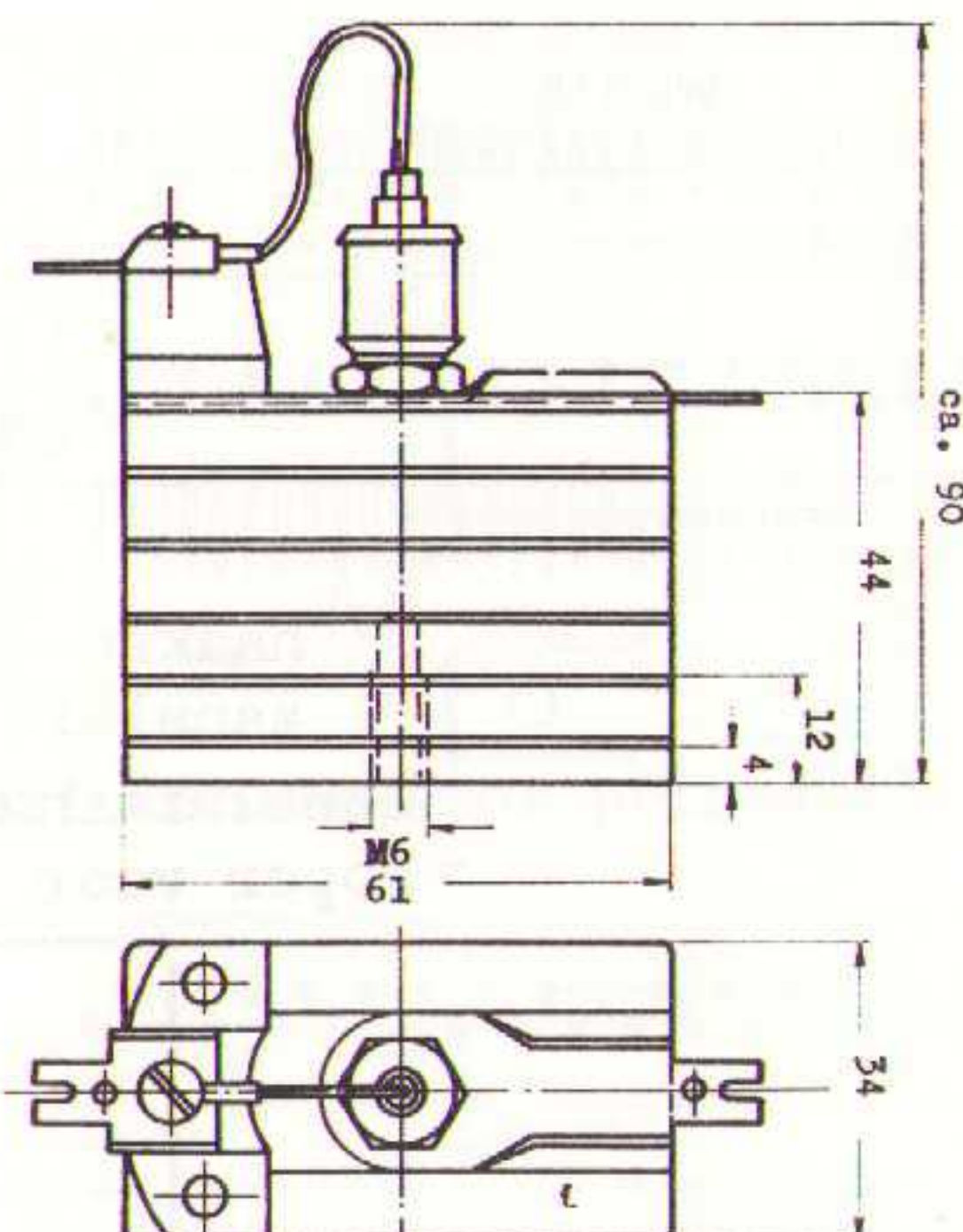
K 9



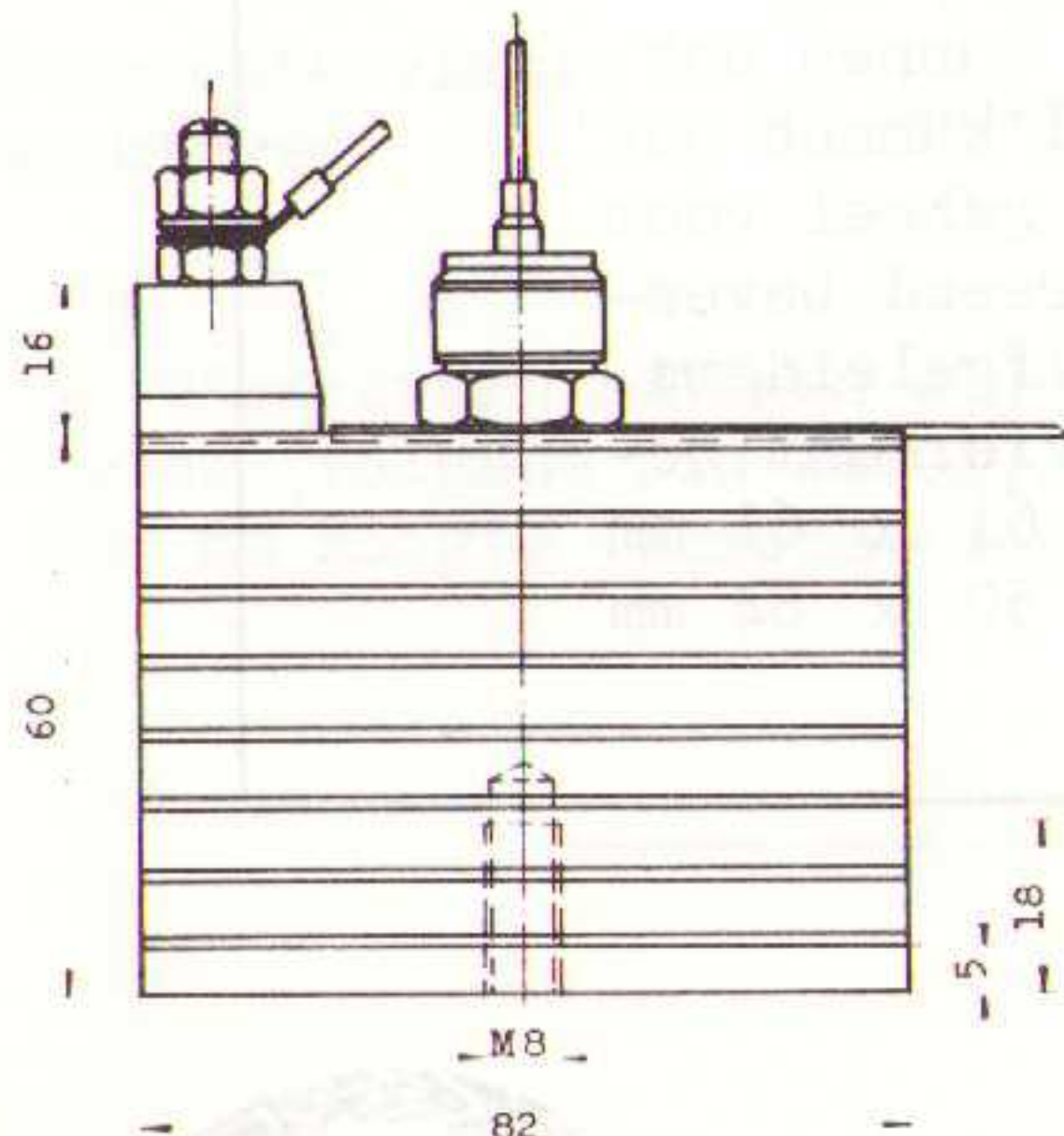
K 9



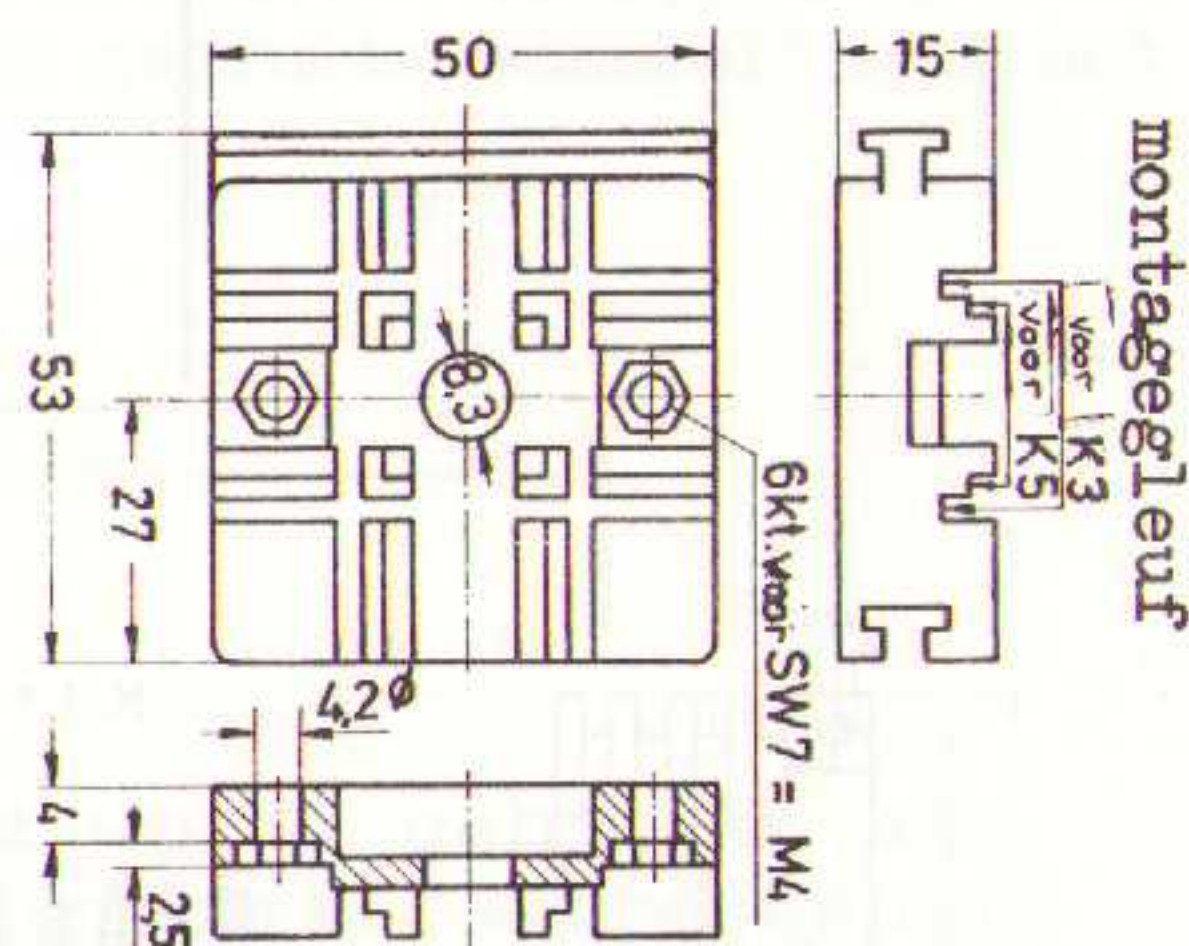
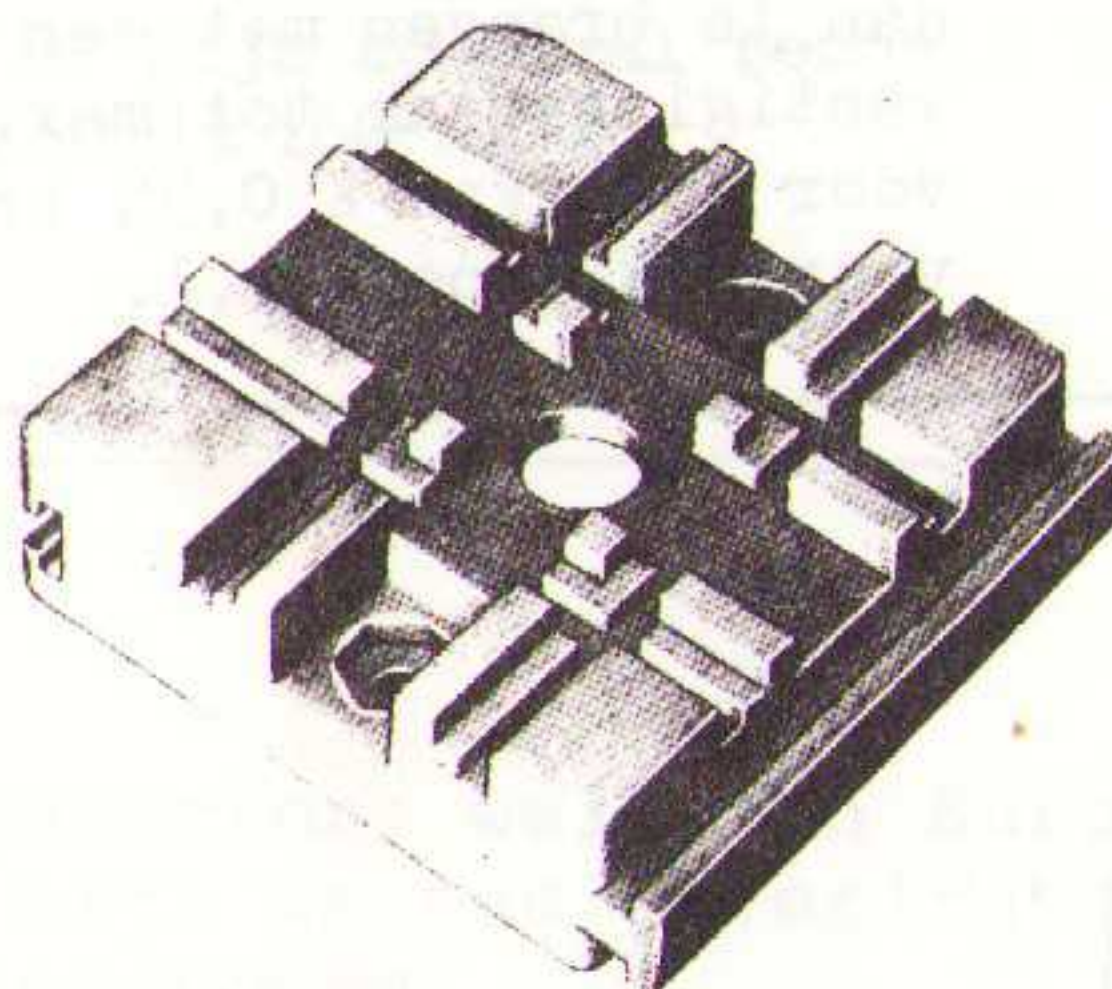
K 5



K 5

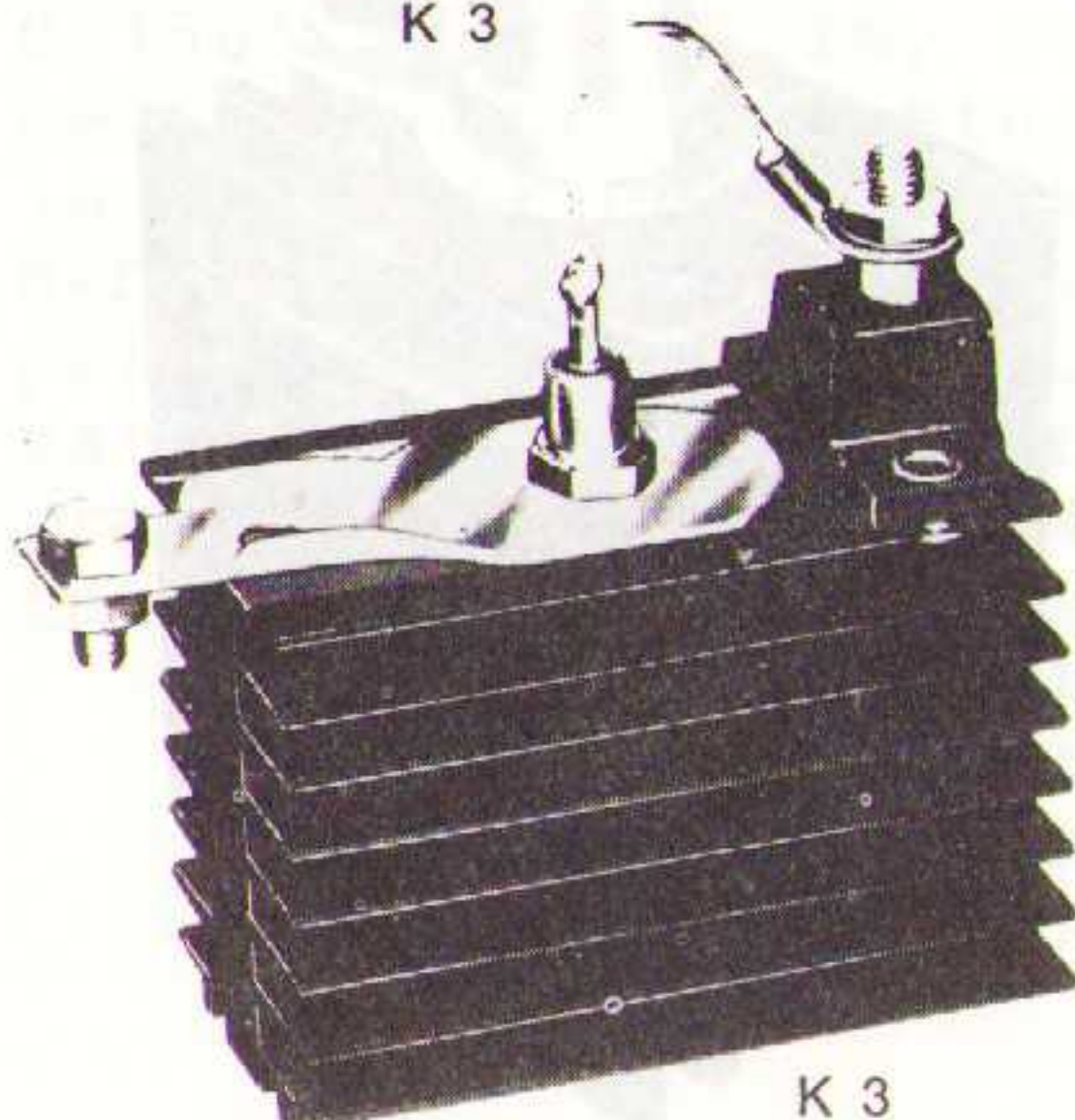


K 3

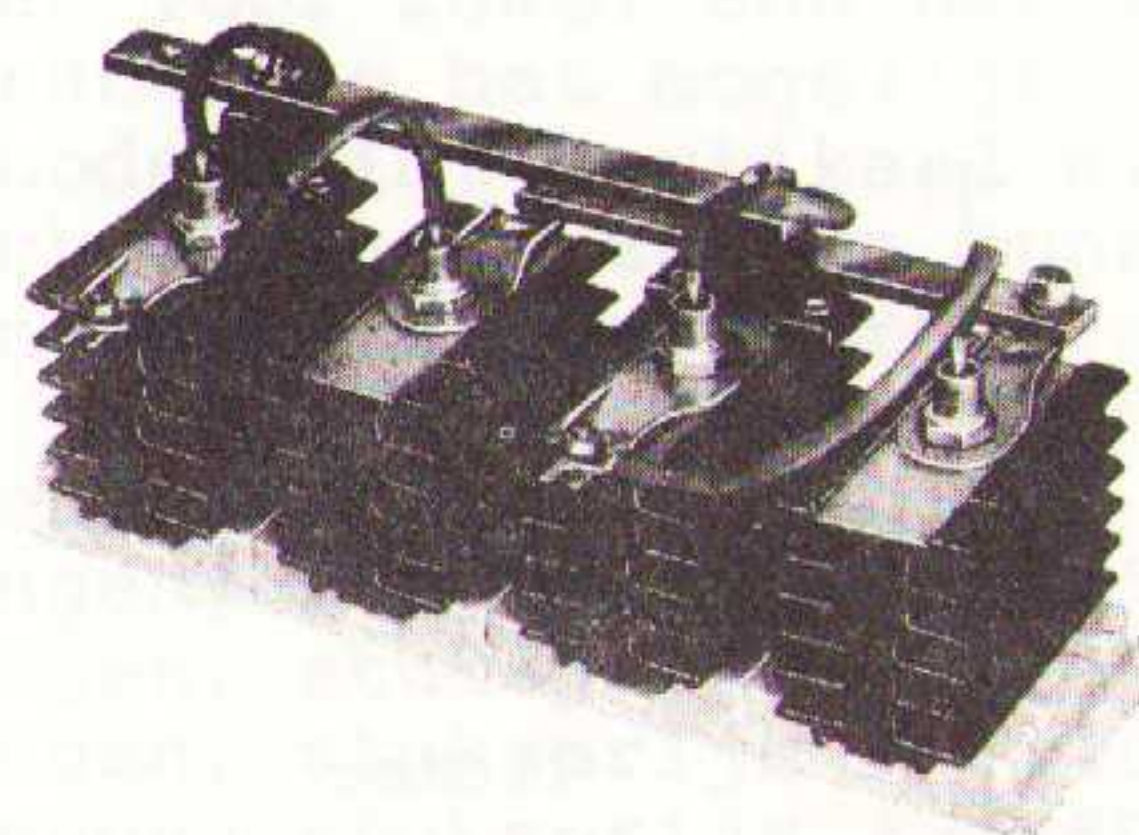


montagevoet

Montagevoet voor K3 en K5 koelelementen, type MS0 35.
Materiaal: glasvezelversterkte polyamide, langdurig
bestand tegen temperaturen van +140°C.



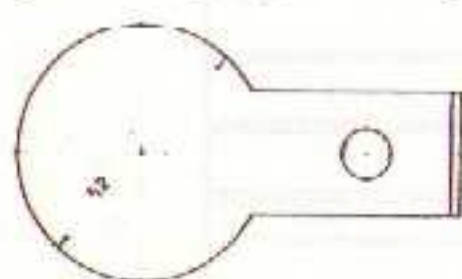
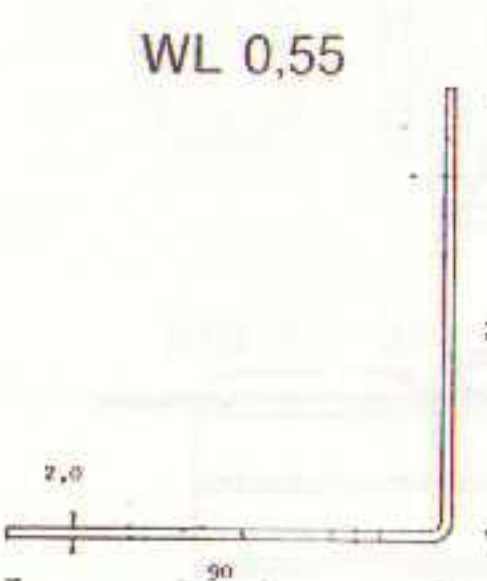
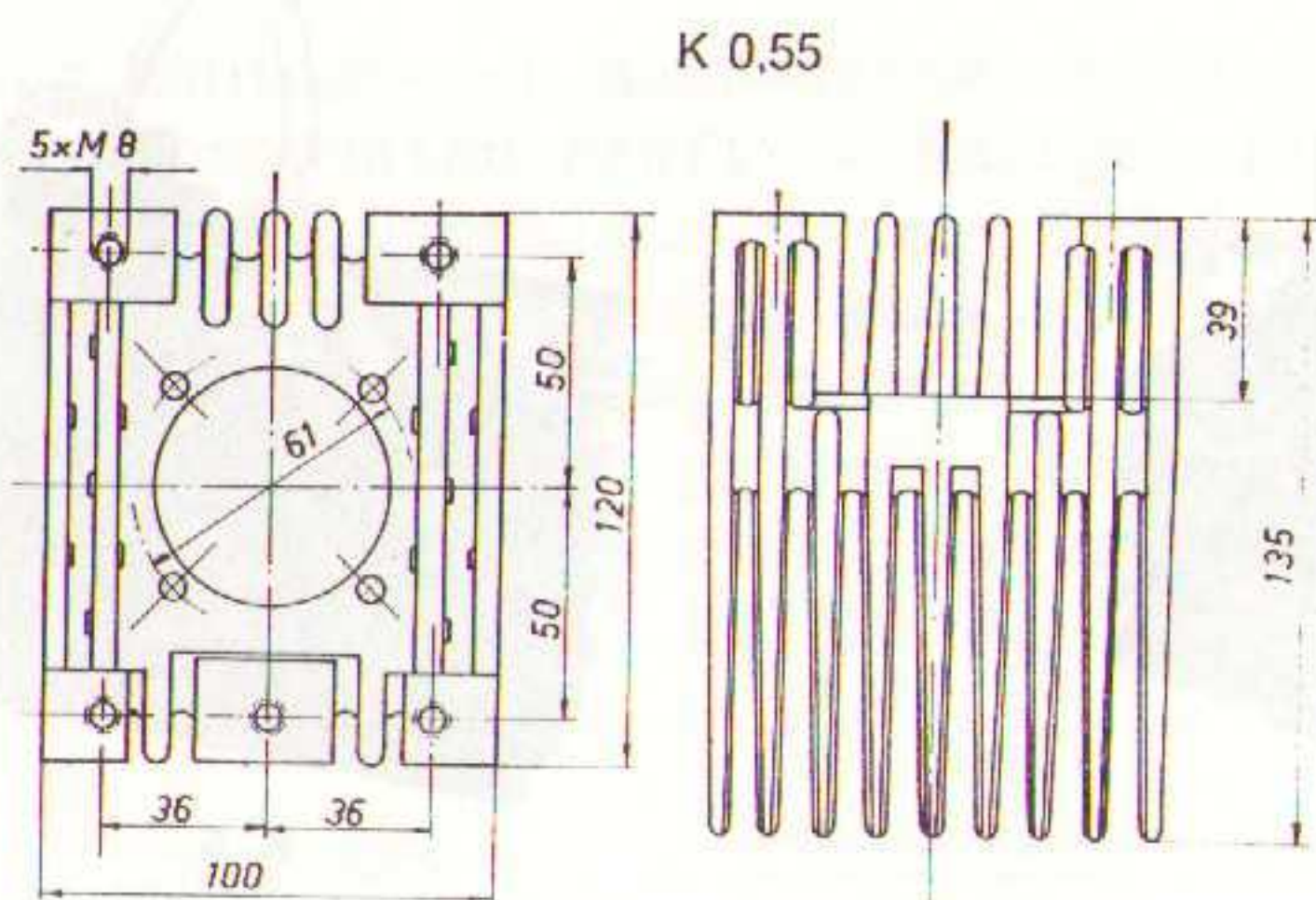
K 3



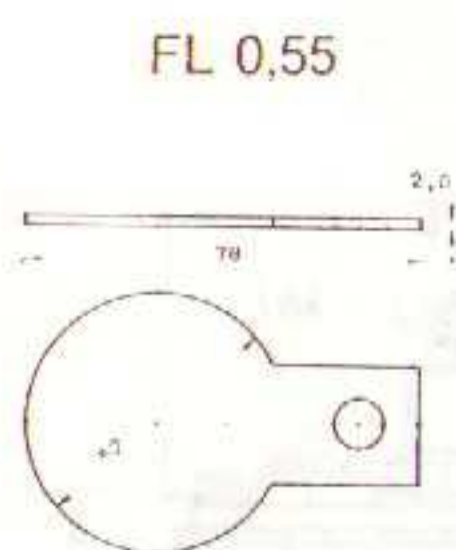
Bij bestelling de soort schroefdraad van Uw
halfgeleider opgeven!

Aluminium koelelementen volgens de duitse norm DIN 41882.

Bij bestelling de soort schroefdraad van Uw halfgeleider opgeven!

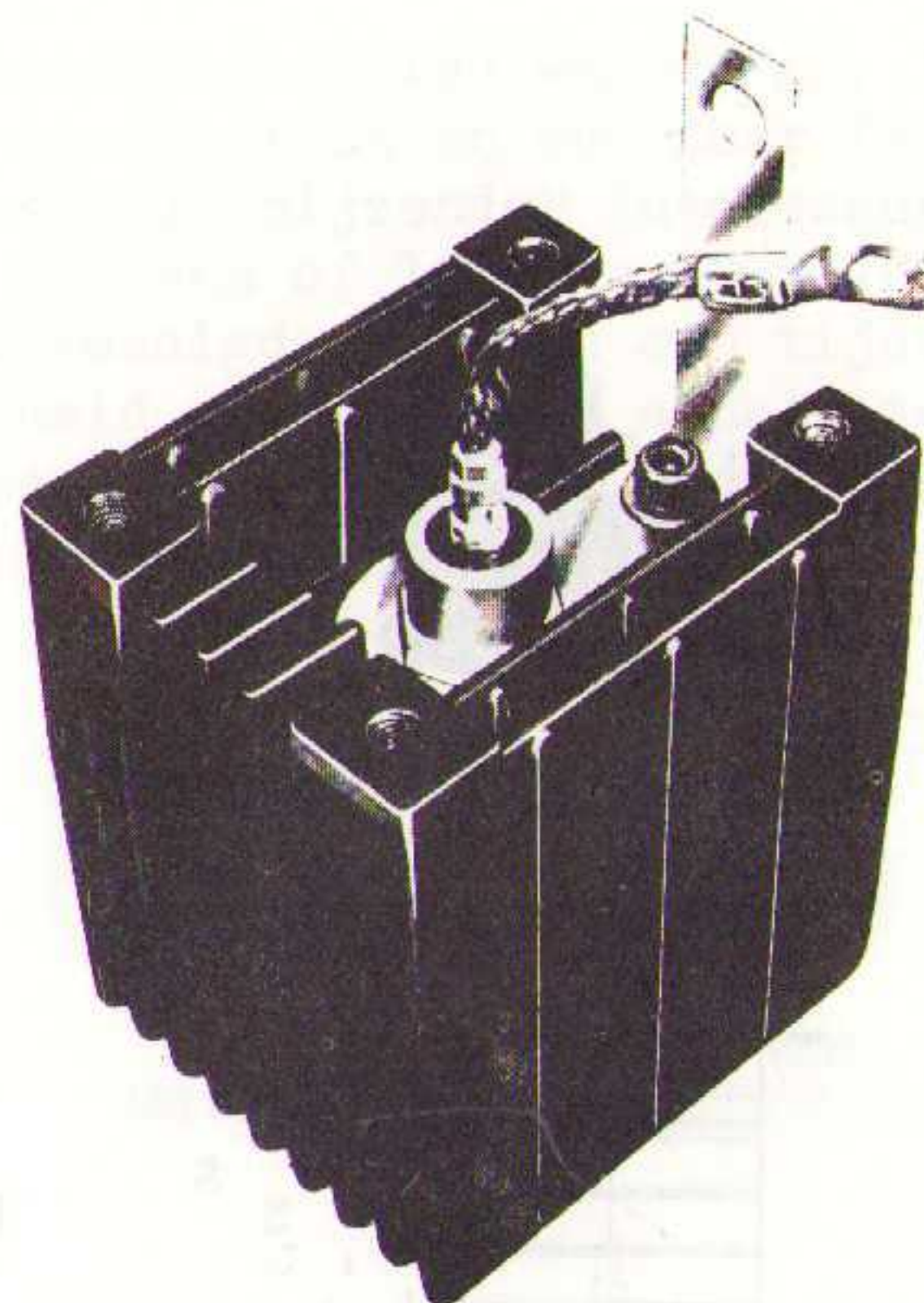


haakse
aansluiting



vlakke
aansluiting

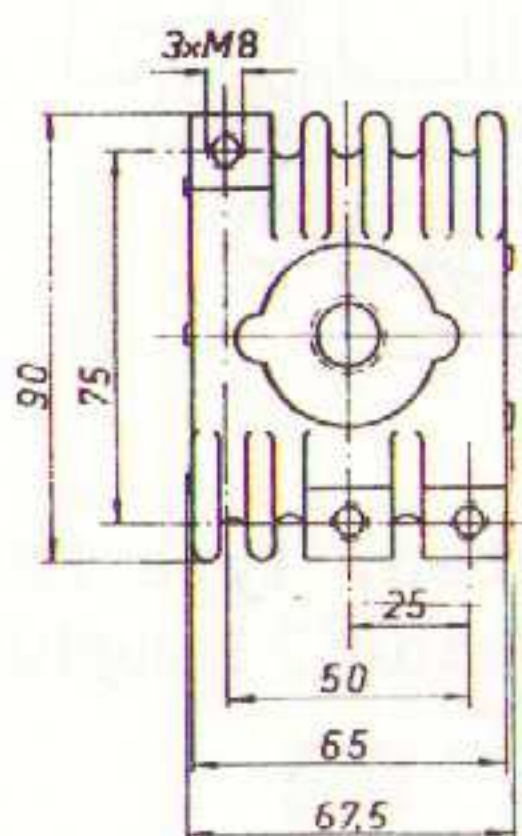
Aansluitingen uit vertind elektrolyt-
koper voor het type K 0,55.



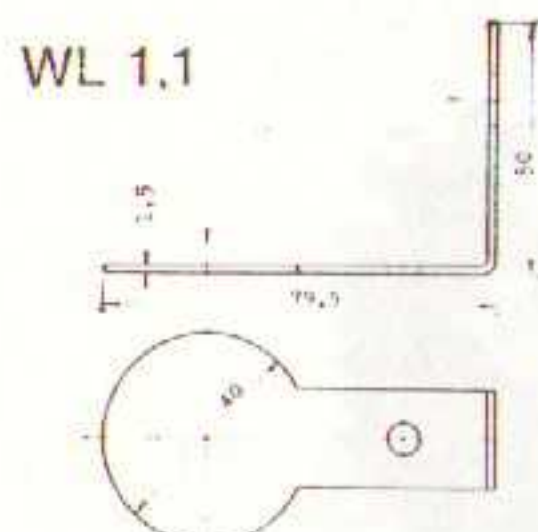
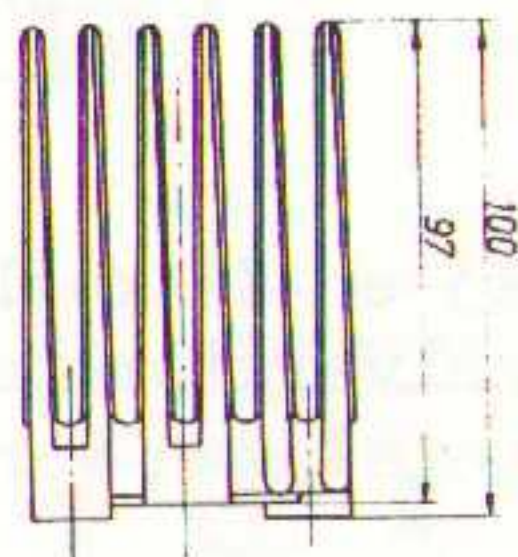
K 0,55

Warmteweerstand 0,55°C/W

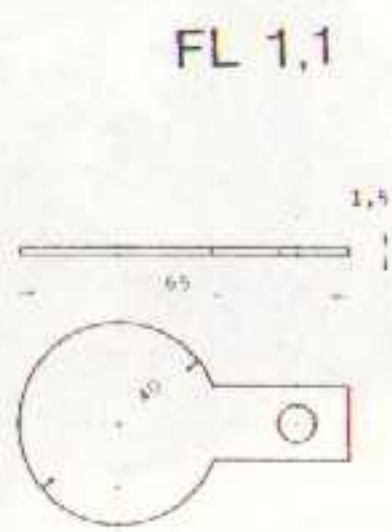
De type K 0,55 en K 1,1 kunnen ook
De typen K 0,55 en K 1,1 kunnen ook
geleverd worden met een geheel voor
de halfgeleider vlakgefreesd beves-
tigingsoppervlak om halfgeleiders
aan te brengen met een vierkant be-
vestigingsvlak tot max. 61 x 61 mm
voor het type K 0,55 en 50 x 64 mm
voor het type K 1,1.



K 1,1

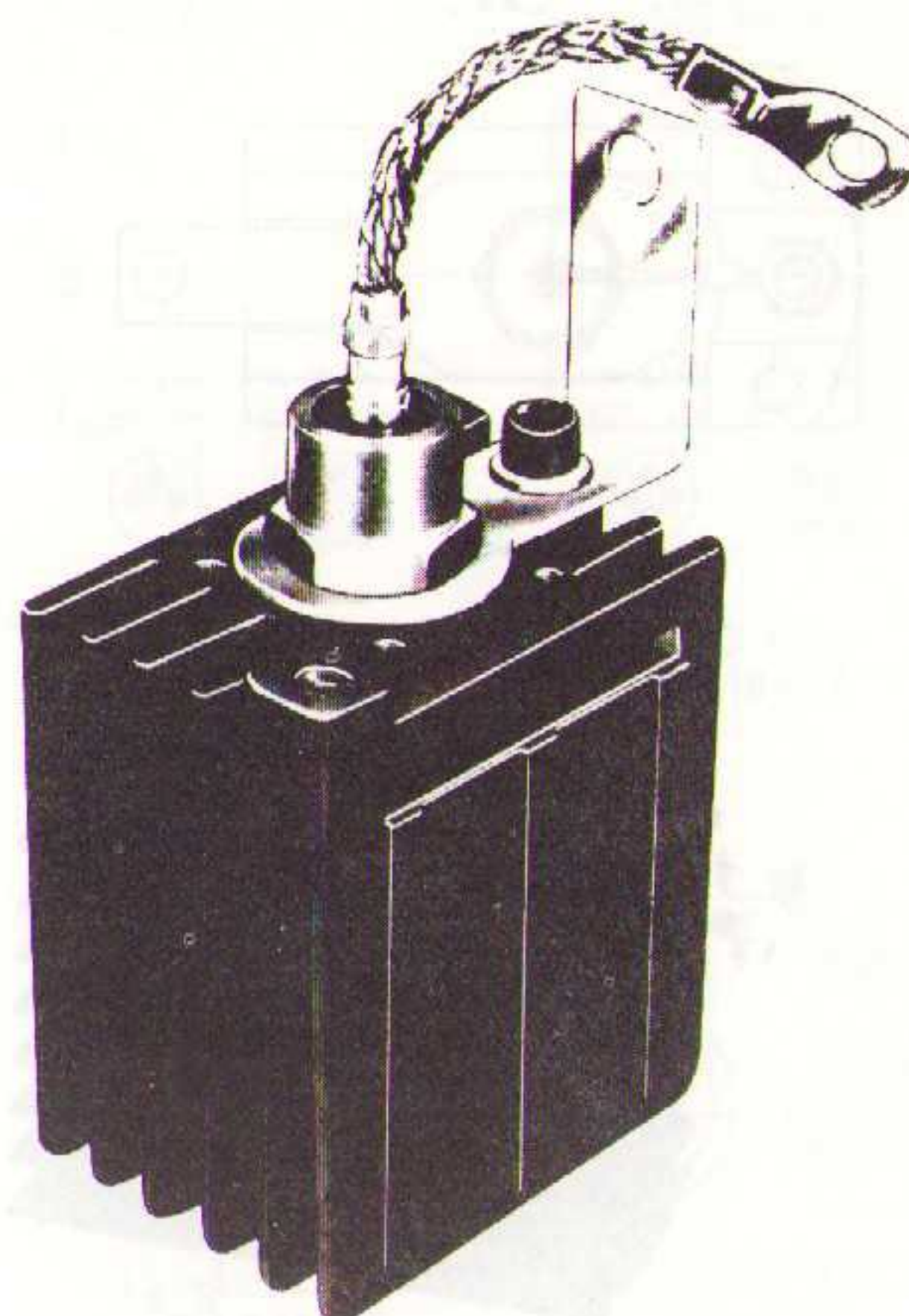


haakse
aansluiting



vlakke
aansluiting

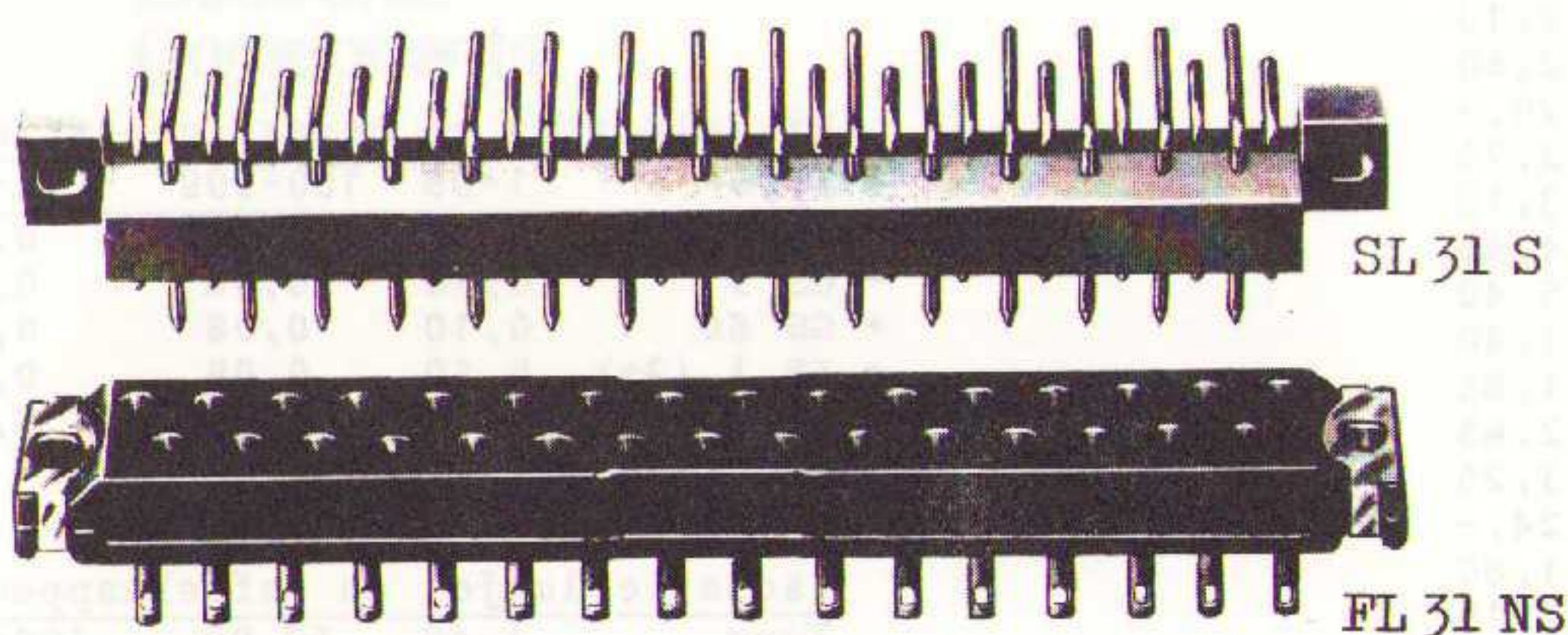
Aansluitingen uit vertind elektrolyt-
koper voor het type K 1,1.



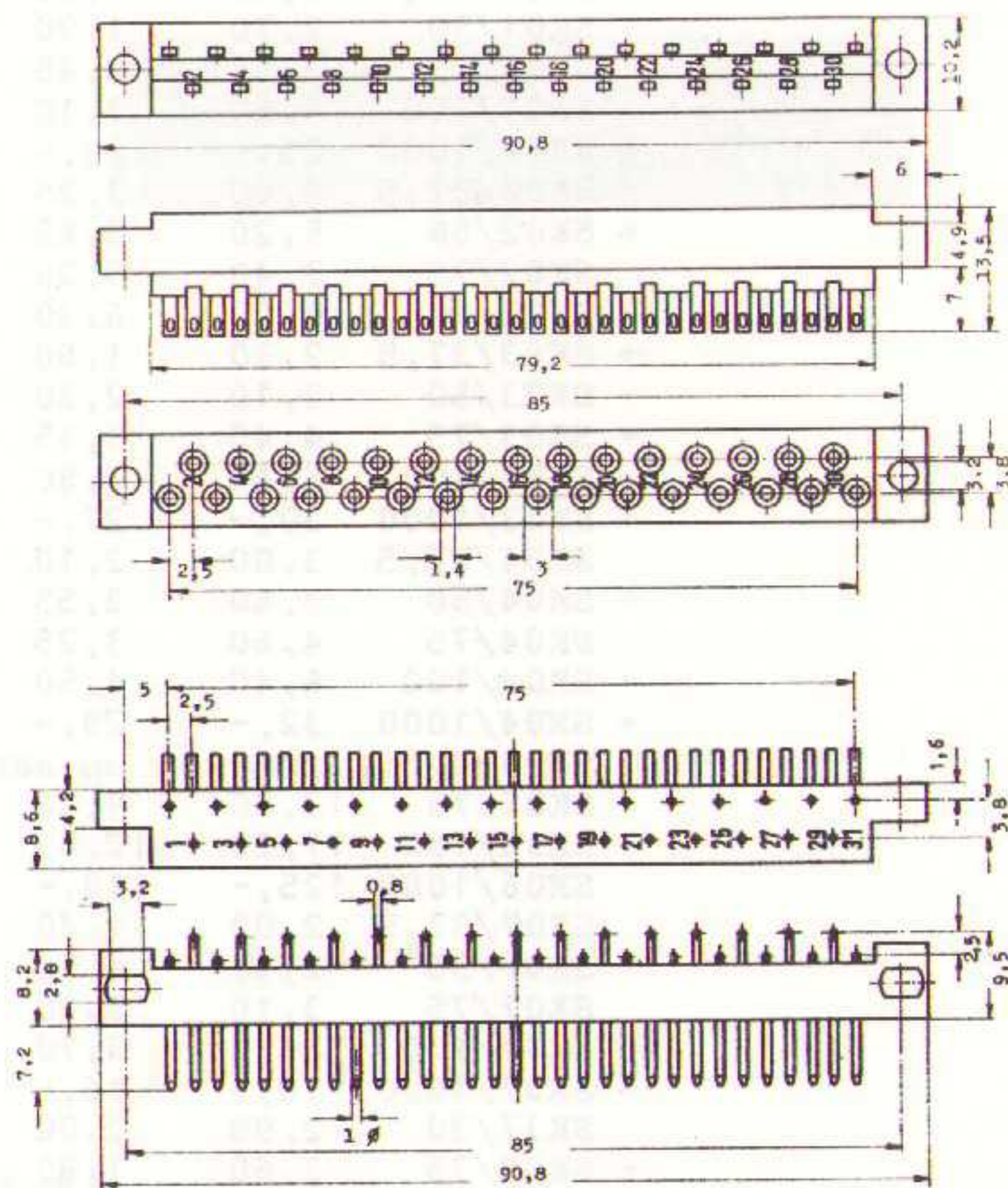
K 1,1

Warmteweerstand 1,1°C/W

31 polige stekerverbindingen voor prints naar DIN 41 617.



Aantal kontakten	31
Isolatielichaam	Makrolon
Kontaktstiften	Messing, Ms 63 vernikkeld en verzilverd
Kontaktveren	verzilverd fosforbrons
Temperatuurbereik	-55 tot +140°C
Isolati weerstand	10 ¹² Ω
Capaciteit	1 pF
Belastbaarheid	6 ampère
Werkspanning	127 V AC volgens VDE0110 350 V AC volgens IEC130-1
Andere oppervlakte bewerkingen	op aanvraag leverbaar



maattekening voor printdeel layout.

De stekerverbindingen zijn voor printplaten met een 2,5 mm raster en 1,3 mm gatdiameter (volgens DIN 40801). Ze zijn speciaal bedoeld voor gedrukte schakelingen met 100 mm hoogte (Europa-kaart 100 x 160 mm).

Naast bovenstaande connector, welke in Europa wordt gebruikt als standaard connector, zijn in ons leveringsprogramma ook verscheidene andere soorten opgenomen.

Zo is bijvoorbeeld een zeer universele reeks opgenomen van het Amerikaanse fabriek "MOLEX", een serie met een printsteek van 0,156". Deze serie, welke onder de naam "CONNECTCON" op de markt gebracht wordt, bestaat uit printdelen voor zowel een horizontale als voor een verticale print. Hierdoor is het mogelijk geworden in een horizontaal aangebrachte bodemprint vertikaal diverse printen te steken, welke standaard schakelingen zouden kunnen omvatten. Deze unieke connectors kunnen naast elkaar worden aangebracht tot maximaal 64 aansluitingen en zijn leverbaar uit voorraad magazijn Rotterdam in de navolgende aansluitingen:

- type A2145-2183-A3, 3 aansluitingen, stuksprijs f 0,85
- type A2145-2183-A4, 4 aansluitingen, stuksprijs f 1,05
- type A2145-2183-A6, 6 aansluitingen, stuksprijs f 1,40
- type A2145-2183-A8, 8 aansluitingen, stuksprijs f 1,70

TOESLAG VOOR BORING 10-5 EN 10-66 FL. 0,30

Koelelementen, ongeboord, type SA

Type	1-49	50-99	100-499
SK01/37½	2,20	1,55	1,30
SK01/50	2,70	1,90	1,60
SK01/75	3,50	2,45	2,10
SK01/100	4,40	3,10	2,60
→ SK01/1000	25,-	23,-	20,-
SK02/37,5	4,60	3,25	2,75
→ SK02/50	5,20	3,65	3,15
SK02/75	7,40	5,20	4,45
→ SK02/100	9,00	6,30	5,40
→ SK03/37,5	2,30	1,60	1,40
→ SK03/50	3,10	2,20	1,85
→ SK03/75	4,45	3,15	2,65
→ SK03/100	5,40	3,80	3,25
→ SK03/1000	30,-	27,-	24,-
SK04/37,5	3,00	2,10	1,80
→ SK04/50	3,60	2,55	2,15
SK04/75	4,60	3,25	2,75
→ SK04/100	6,40	4,50	3,85
→ SK04/1000	32,-	29,-	25,-
SK05-serie prijzen op aanvraag.			
→ SK06/75	13,00	9,10	7,80
→ SK06/100	17,20	12,05	10,35
SK06/1000	125,-	110,-	99,60
SK07/37,5	2,00	1,40	1,15
→ SK07/50	2,40	1,70	1,45
SK07/75	3,10	2,20	1,85
SK07/100	3,80	2,70	2,25
→ SK07/1000	18,-	16,-	14,-
SK37/30	2,90	2,00	1,70
→ SK38/75	2,60	1,85	1,55

→ Voorraad Rotterdam met boring voor 1 of 2 TO-3 transistoren. De 1000 mm lengten zijn ongeboord. BOVENSTAANDE PRIJZEN ZIJN EXCL. DE BOOR-TOESLAG! Van de niet vermelde typen zijn de prijs en levertijd op aanvraag beschikbaar.

Koelsterren, koelmoeren, enz.

Type	1-99	100-499	500-999
→ KK 505	1,00	0,70	0,64
→ KK 5095	1,05	0,75	0,67
→ KK 511	1,45	1,00	0,90
KK 562	0,20	0,14	0,12
KK 562GS	0,25	0,18	0,16
KK 520	0,30	0,21	0,18
→ FGK 5	1,10	0,77	0,69
→ FGK 5GS	1,20	0,84	0,76
KF 5/5	0,20	0,14	0,12
KF 5/10	0,25	0,17	0,15
KF 5/15	0,30	0,21	0,18
→ KK 25	3,50	2,95	2,65
→ KK 218	3,20	2,85	2,40
→ KK 55	1,10	0,77	0,55
→ KK 510	1,20	0,84	0,62
KK 50	1,00	0,70	0,65
→ KK 51	2,60	1,82	1,70
→ KK 52	2,20	1,55	1,45
→ KK 185	1,00	0,70	0,64
→ KK 1895	1,05	0,74	0,67
→ KK 1862	0,20	0,14	0,12
→ FGK 18	1,00	0,70	0,64
→ KS 1	0,05	0,03	0,02
KS 2	0,10	0,07	0,06
→ KS 92	0,20	0,14	0,13
→ voorraad Rotterdam.			

Vingerkoelplaten

Type	1-49	50-99	100-499
→ FK201/SA	2,00	1,40	1,20
→ FK201/3/SA	2,40	1,70	1,45
→ FK201/9-66	2,40	1,70	1,45
→ voorraad Rotterdam			

Warmtegeleidingspasta (in doosje)

Inhoud netto	1-49	50 - 99
→ 4 gram	2,25	1,00
→ 35 gram	7,60	4,75
→ 500 gram	66,00	51,75
→ voorraad Rotterdam.		

Aluminium oxyde onderlegplaten

Type	1-99	100-499	500-999
→ AOS 3	1,80	1,38	1,17
AOS 5	0,40	0,28	0,26
AOS 18	0,30	0,21	0,19
→ voorraad Rotterdam.			

Micaplaatjes en bevestigingsringen

Type	1-99	100-999	1k-10k
→ GS 3	0,15	0,10	0,08
→ GS 9	0,10	0,08	0,07
→ GS 66	0,10	0,08	0,07
→ IB 1 (2x)	0,10	0,08	0,05
→ IB 2 (2x)	0,10	0,08	0,05
→ voorraad Rotterdam			

Isolatieblokken en -afdekkappen

Type	1-49	50-99	100-999
→ IS 1	0,40	0,30	0,25
→ IS 2	0,40	0,30	0,25
→ IS 3	0,25	0,20	0,15
→ IK 341/3	0,65	0,45	0,40
→ IK 341/4	0,80	0,60	0,50
→ IK 66	0,65	0,45	0,40
→ voorraad Rotterdam			

Isolatiestukjes en vulstukjes.

Type	1-99	100-999	1000-9999
→ MS 53-25	0,15	0,10	0,06
→ MS 54-25	0,15	0,10	0,06
→ MS 53-3	0,15	0,10	0,06
→ MS 53-7	0,15	0,10	0,06
→ MS3518-25	0,15	0,10	0,06
→ MS3518-35	0,15	0,10	0,06
→ MS 34-518	0,15	0,10	0,06
→ MS 58-7	0,20	0,12	0,08
→ MS 56-15	0,20	0,12	0,08
→ MS 510-15	0,20	0,12	0,08
→ MS 512-15	0,20	0,12	0,08
→ US 58-4	0,50	0,30	0,25
→ US 510-4	0,50	0,30	0,25
→ US 512-4	0,50	0,30	0,25
→ MS 13-7	0,15	0,10	0,06
→ MS 183-25	0,15	0,10	0,06
→ MS 184-25	0,15	0,10	0,06
→ MS 183-35	0,15	0,10	0,06
→ MS 183-7	0,15	0,10	0,06
→ MS 184-7	0,15	0,10	0,06
→ MS 4016	0,40	0,25	0,20
→ MS 84-4	0,20	0,12	0,08
→ MS 923-25	0,15	0,10	0,06
→ DH 124-32	0,15	0,10	0,06
→ MS 4-15	0,50	0,30	0,25
→ voorraad Rotterdam.			

31-polige "EURO" connectorset

Type	1-24	25-99	100 up
→ FL 31 NS	5,90	4,15	3,55
→ SL 31 S	4,20	2,95	2,55
→ voorraad Rotterdam			

Voeten voor kristallen.

Type	1-99	100-499	500-999
→ QS 13V	0,40	0,28	0,26
→ EK 1	0,10	0,07	0,05
→ QS 13GS	0,45	0,31	0,28
→ QS 25GS	0,35	0,25	0,22
→ QS 25VR	0,35	0,35	0,22
→ voorraad Rotterdam.			

Insteekvoeten voor halfgeleiders

Type	1-24	25-99	100-499
→ TF 183	4,00	2,80	2,60
→ TF 184	4,60	3,22	3,00
→ TF 53	4,00	2,80	2,60
→ TF 54	4,60	3,22	3,00
→ TF 55	5,00	3,50	3,25
→ TF 56	6,00	4,20	3,90
→ TF 58	7,00	4,90	4,55
→ TF 510	8,00	5,60	5,20
→ TF 512	9,40	6,60	6,10
→ TF 3/2	2,00	1,40	1,30
→ voorraad Rotterdam			

Prijzen per 1 oktober 1971 excl. B.T.W. Prijsveranderingen zijn strikt voorbehouden.

Application of the RCA-CA3059 Zero-Voltage Switch in Thyristor Circuits

by

George J. Granieri

The RCA-CA3059 zero-voltage switch is a monolithic integrated circuit used primarily as a trigger circuit for the control of thyristors. This multistage circuit employs a diode limiter, a threshold detector, a differential amplifier, and a Darlington output driver to provide the basic switching action. The dc supply voltage for these stages is supplied by an internal zener-diode-regulated power supply that has sufficient current capability to drive the external circuit elements, such as transistors and other integrated circuits. This built-in power supply provides unique solutions to many application problems. An important feature of the CA3059 is that the trigger pulses developed by this circuit can be applied directly to the gate of a silicon controlled rectifier (SCR) or a triac. A built-in fail-safe circuit inhibits the application of these pulses to the thyristor gate circuit in the event that the external sensor for the integrated-circuit switch should be inadvertently opened or shorted.

The CA3059 is particularly suited for use in thyristor temperature-control applications. The integrated circuit may be employed as either an on-off type of controller or a proportional controller, depending upon the degree of temperature regulation required. The availability of numerous terminal connections to internal circuit points greatly increases the flexibility of the CA3059 and permits the circuit designer to exercise his creativity

to employ the integrated switch in unique ways. This Note describes the operation of the CA3059 integrated-circuit switch and discusses its operation in thyristor power-switching and control circuits.

CIRCUIT OPERATION

Fig. 1 shows a functional block diagram of the CA3059 integrated-circuit zero-voltage switch. Any triac that is driven directly from the output terminal of this circuit should be characterized for operation in the I(+) or III(+) triggering modes, i.e., with positive gate current (current flows into the gate for both polarities of the applied ac voltage). The circuit operates directly from a 60-Hz ac line voltage of 120 or 240 volts.

The limiter stage of the CA3059 clips the incoming ac line voltage to approximately plus and minus 8 volts. This signal is then applied to the zero-voltage-crossing detector, which generates an output pulse during each passage of the line voltage through zero. The limiter output is also applied to a rectifying diode and an external capacitor that comprise the dc power supply. The power supply provides approximately 6 volts as the V_{CC} supply to the other stages of the CA3059. The on/off sensing amplifier is basically a differential comparator. The triac gating circuit contains a driver for direct triac triggering. The gating circuit is enabled

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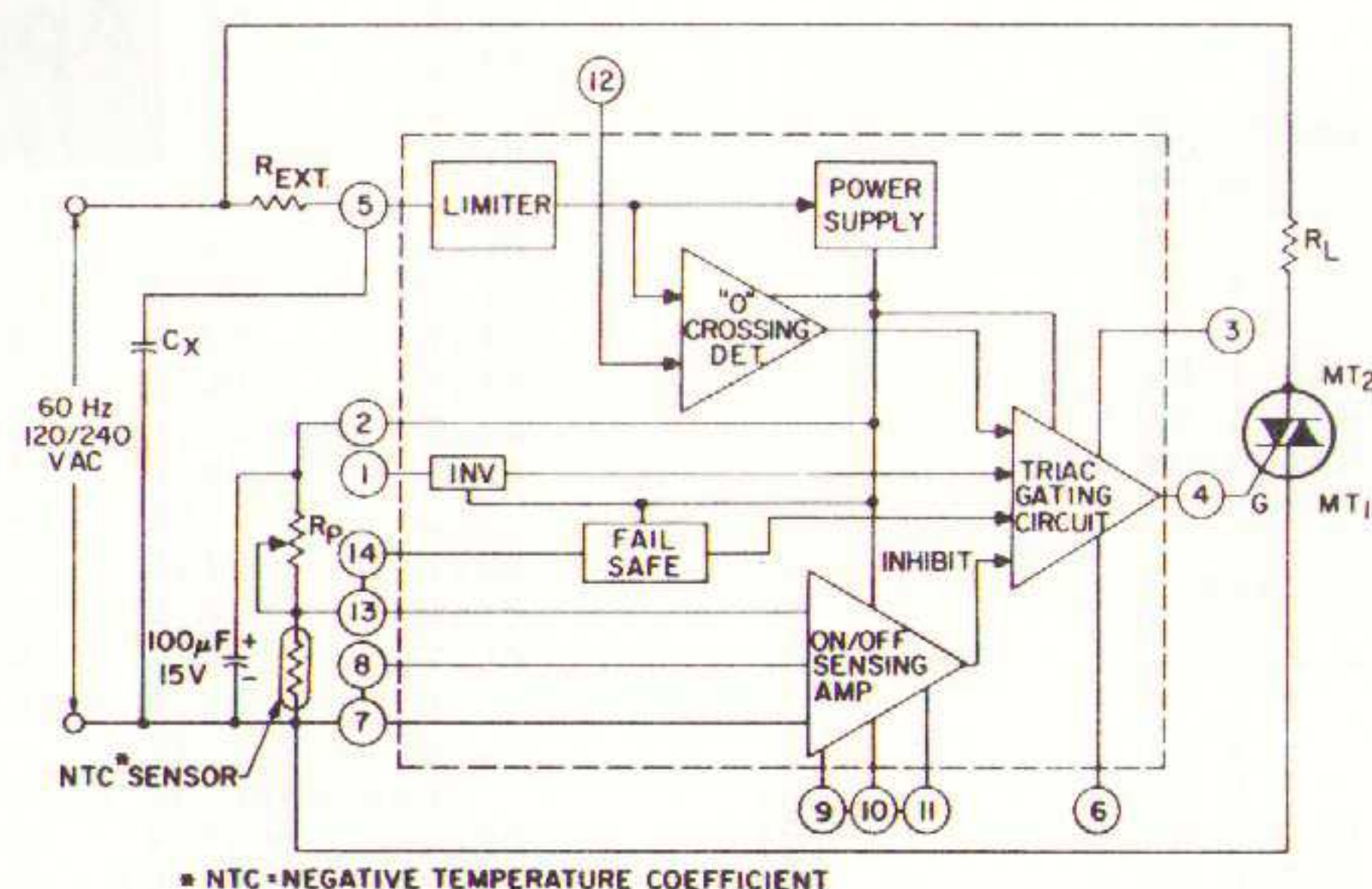


Fig. 1 - Functional block diagram of the integrated-circuit zero-voltage switch.

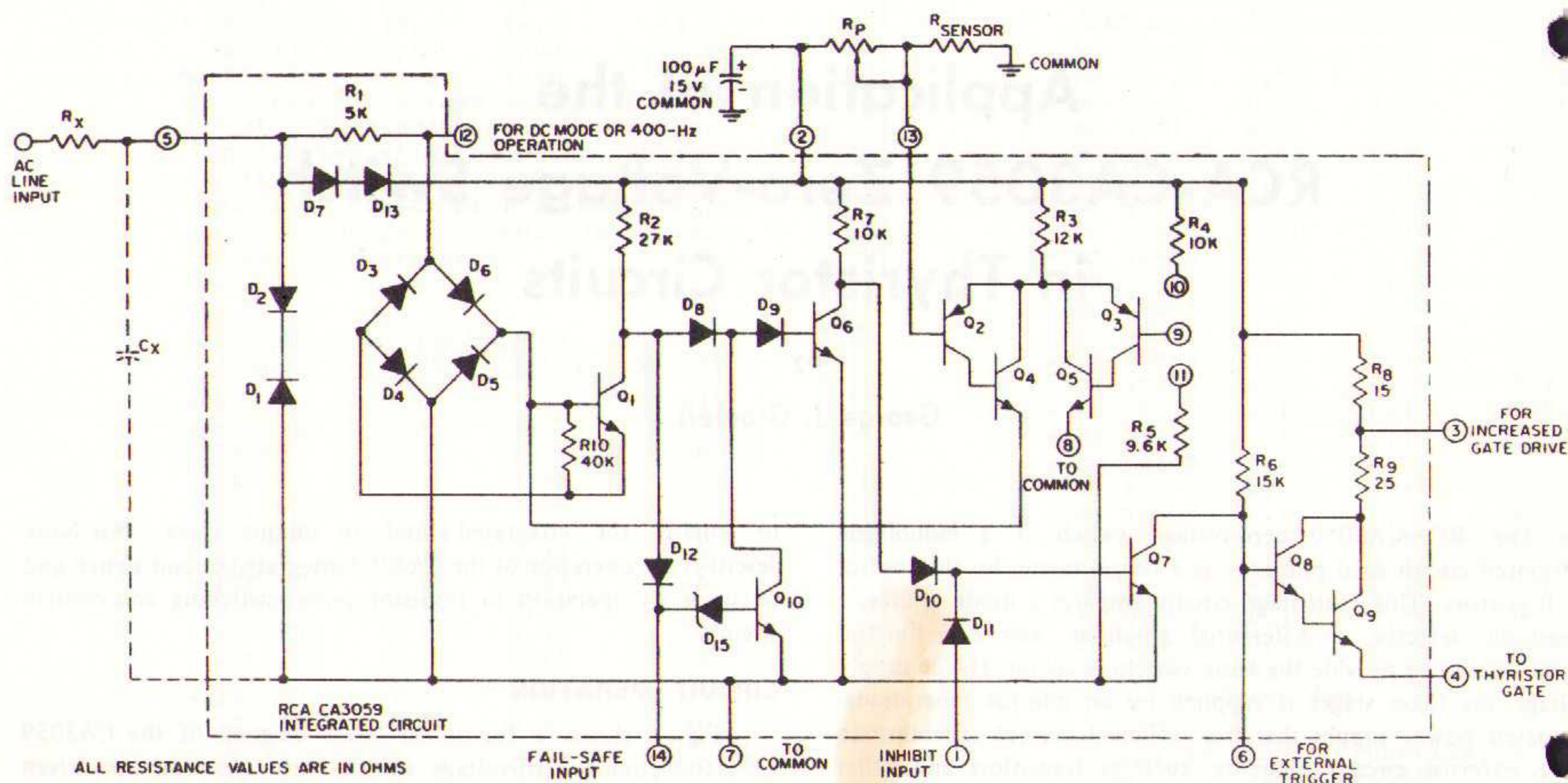


Fig. 2 - Circuit diagram for the CA3059 zero-voltage switch.

when all the inputs are at a high voltage, i.e., the line voltage must be approximately zero volts, the sensing-amplifier output must be "high," the external voltage to terminal 1 must be a logical "1," and the output of the fail-safe circuit must be "high."

Fig. 2 shows the circuit diagram of the CA3059. The zero-voltage threshold detector consists of diodes D_3 , D_4 , D_5 , and D_6 , and transistor Q_1 . The differential amplifier consists of transistor pairs Q_2 - Q_4 and Q_3 - Q_5 . Transistors Q_1 , Q_6 , Q_7 , Q_8 , and Q_9 comprise the triac gating circuit and driver stage. Diode D_{12} , zener diode D_{15} , and transistor Q_{10} constitute the fail-safe circuit. The power supply consists of diodes D_7 and D_{13} , and an external resistor and capacitor connected to terminals 5 and 2, respectively, and to ground through pin 7. If the transistor pair Q_2 - Q_4 and transistor Q_1 are turned off, an output appears at terminal 4. Transistor Q_1 is in the OFF state if the incoming line voltage is less than approximately the voltage drops across three silicon diodes

(2.1 volts) for either the positive or negative excursion of the line voltage. Transistor pair Q_2 - Q_4 is OFF if the voltage across the sensor, connected from terminals 13 to 7, exceeds the reference voltage from 9 to 7. If either of these conditions is not satisfied, pulses are not supplied to terminal 4. Fail-safe operation requires that terminal 13 be connected to 14. The addition of hysteresis and elimination of half-cycling can be obtained by a resistive voltage divider connected from 13 to 8 and from 8 to 7.

Fig. 3 shows the position and width of the pulses supplied to the gate of a thyristor with respect to the incoming ac line voltage. The CA3059 can supply sufficient gate voltage and current to trigger most RCA thyristors at ambient temperatures of 25°C. However, under worst-case conditions (i.e., at ambient-temperature extremes and maximum triggering requirements), selection of the higher-current thyristors may be necessary for particular applications. For example, the RCA-2N5444

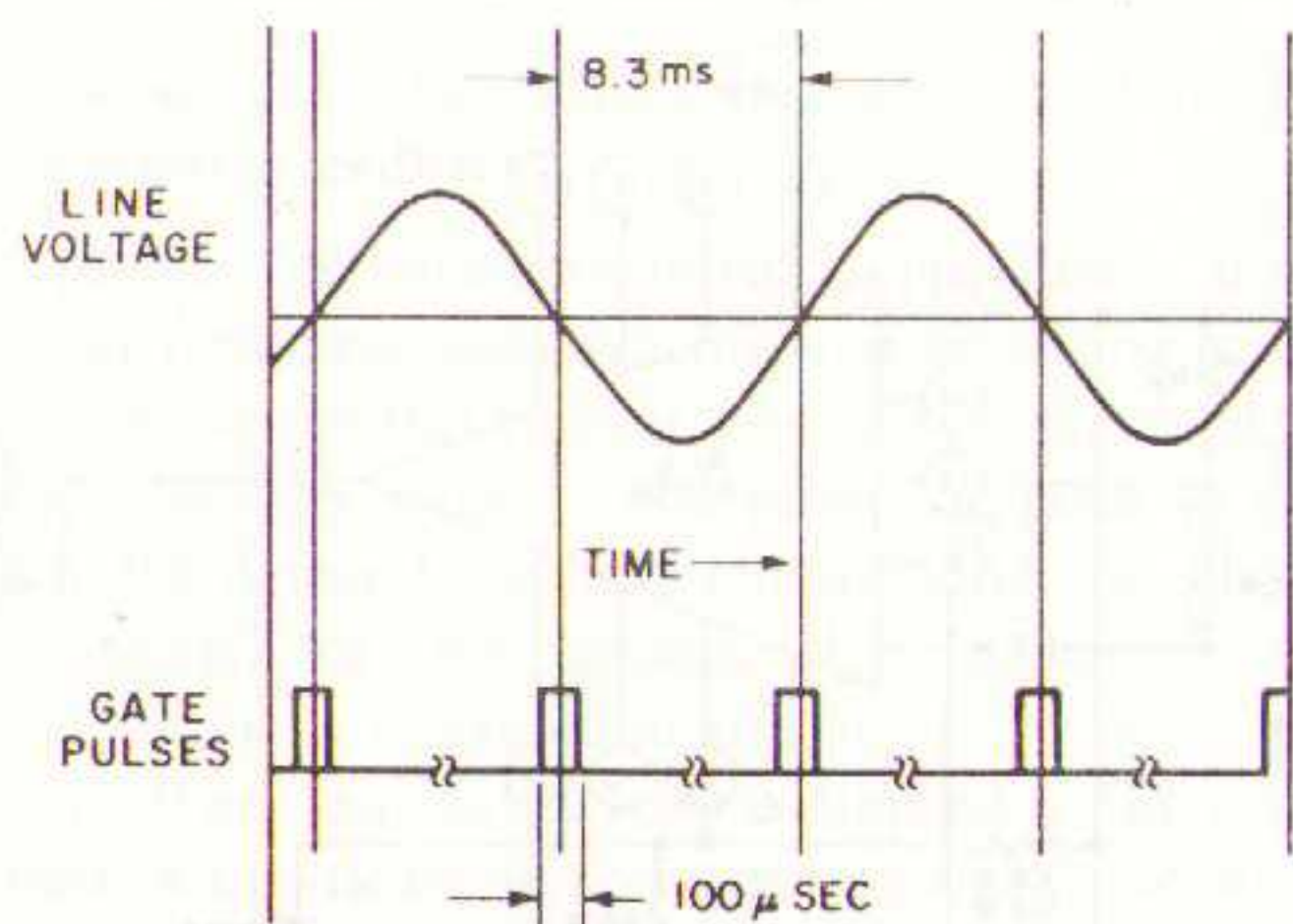


Fig.3 - Timing relationship between the output pulses of the CA3059 and the ac line voltage.

40-ampere triac has a maximum gate trigger voltage V_{gt} (max) of 2.5 volts and a maximum gate trigger current I_{gt} (max) of 80 milliamperes in the III(+) quadrant at 25°C. Because the CA3059 cannot guarantee a drive of 80 milliamperes for a V_{gt} of 2.5 volts, triac selection will be required.

EFFECT OF CA3059 ON THYRISTOR LOAD CHARACTERISTICS

The CA3059 is designed primarily to gate a thyristor that switches a resistive load. Because the output pulse supplied by the CA3059 is of short duration, the latching current* of the triac becomes a significant factor in determining whether other types of loads can be switched. (The latching-current value determines whether the triac will remain in conduction after the gate pulse is removed.) Provisions are included in the CA3059 to accommodate inductive loads and low-power loads. For example, for loads that are less than approximately 4 amperes rms or that are slightly inductive, it is possible to retard the output pulse with respect to the zero-voltage crossing by insertion of the capacitor C_x from terminal 5 to terminal 7 as shown in Fig. 1. The insertion of capacitor C_x permits switching of triac loads that have a slight inductive component and that are greater than approximately 200 watts (for operation from an ac line voltage of 120 volts rms). However, for loads less than 200 watts (for example, 70 watts), it is recommended that the user employ the RCA-40526 sensitive-gate triac with the CA3059 because of the low latching-current requirement of this triac.

For loads that have a low power factor, such as a solenoid valve, the user may operate the CA3059 in the dc mode. In this mode, terminal 12 is connected to terminal 7, and the zero-crossing detector is inhibited. Whether a "high" or "low" voltage is produced at terminal 4 is then dependent only upon the state of the differential comparator within the CA3059 integrated circuit, and not upon the zero crossing of the incoming line voltage. Of course, in this mode of operation, the CA3059 no longer operates as a zero-voltage switch. However, for many applications that involve the switching of low-current inductive loads, the amount of RFI generated can frequently be tolerated.

For switching of high-current inductive loads, which must be turned on at zero line current, the triggering technique employed in the dual-output over-under temperature controller and the transient-free switch controller described later in this Note is recommended.

* The latching current is the minimum current required to sustain conduction immediately after the thyristor is switched from the OFF to the ON state and the gate signal is removed.

FAIL-SAFE FEATURE

As shown in Figs. 1 and 2, when terminal 13 is connected to terminal 14, the fail-safe circuit of the CA3059 is operable. If the sensor should then be accidentally opened or shorted, power is removed from the load (i.e., the triac is turned OFF). The internal fail-safe circuit functions properly, however, only when the ratio of the sensor impedance at 25°C, if a thermistor is the sensor, to the impedance of the potentiometer R_p is less than 4 to 1. It is readily apparent that, if the potentiometer is adjusted for 1000 ohms and the sensor is 100,000 ohms, the zener diode D_{15} (shown in Fig. 2) would conduct because virtually all the dc power-supply voltage (from terminal 2 to terminal 7) would appear across the sensor. The CA3059 would then detect this condition as an open sensor.

For ratios greater than 4 to 1, for example 100 to 1, the circuit shown in Fig. 4 may be employed to provide fail-safe operation. In this circuit, transistor Q_1 and diode D_1 are components external to the CA3059. Transistor Q_1 detects the sensor current which maintains this transistor in saturation so that terminal 1 is effectively shorted to terminal 7 through the collector-to-emitter junction of the transistor. Transistor Q_1 provides sufficient current gain to permit operation with a sensor impedance greater than 1 megohm. If the sensor becomes open-circuited, transistor Q_1 turns OFF, and current then flows into terminal 1, the inhibit terminal of the CA3059, and results in the removal of power to the load. For the shorted-sensor condition, the external diode D_1 conducts and causes triac Y_1 to turn OFF. Diode D_2 compensates for variations in the base-to-emitter voltage of transistor Q_1 with temperature. Terminals 13 and 14 on the CA3059 should not be connected when the external fail-safe circuit shown in this illustration is employed.

HALF-CYCLING AND HYSTERESIS CHARACTERISTICS

The method by which the CA3059 senses the zero crossing of the ac power results in a half-cycling phenomenon at the control

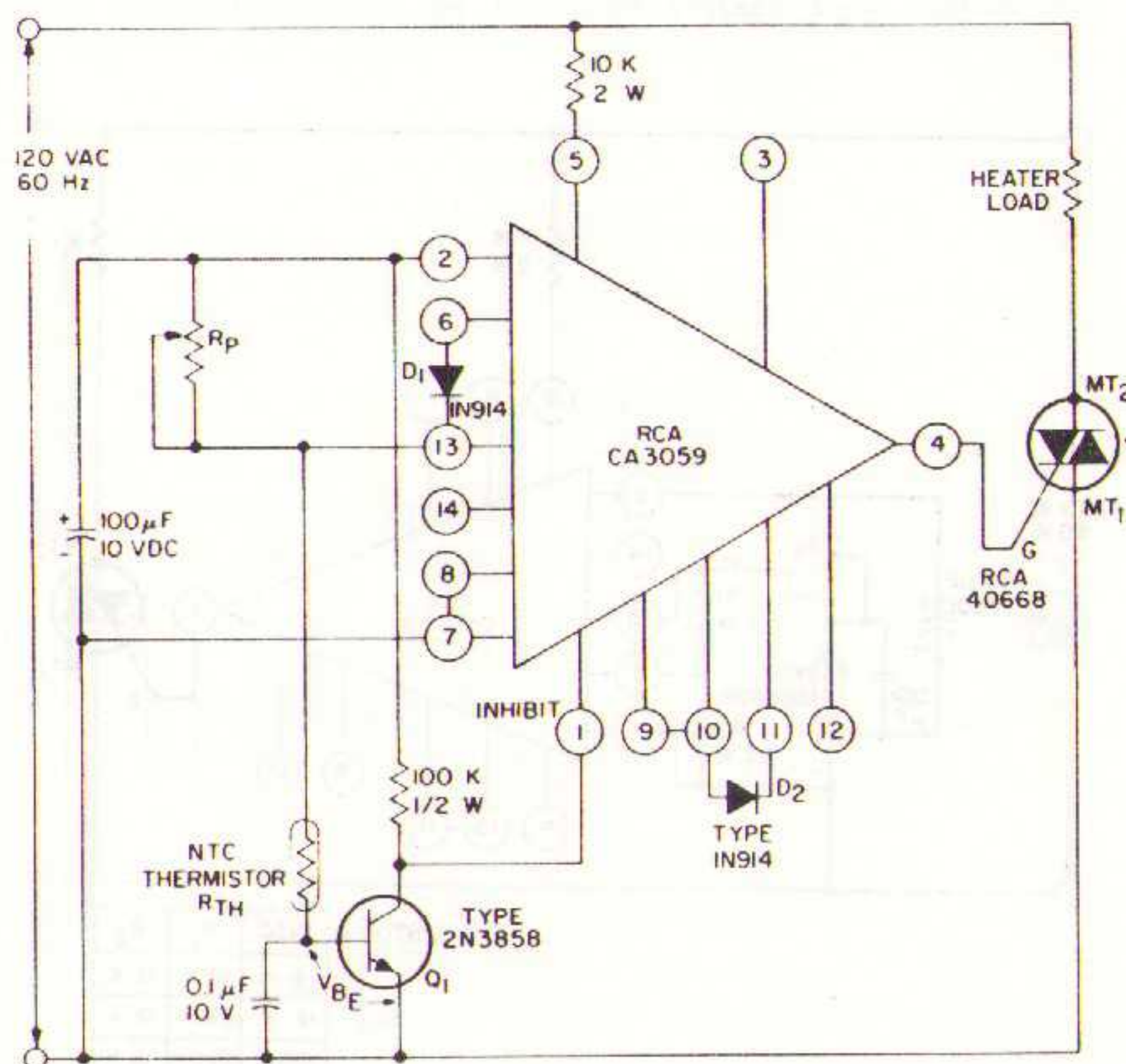


Fig.4 - CA3059 on-off controller that uses an external fail-safe circuit.

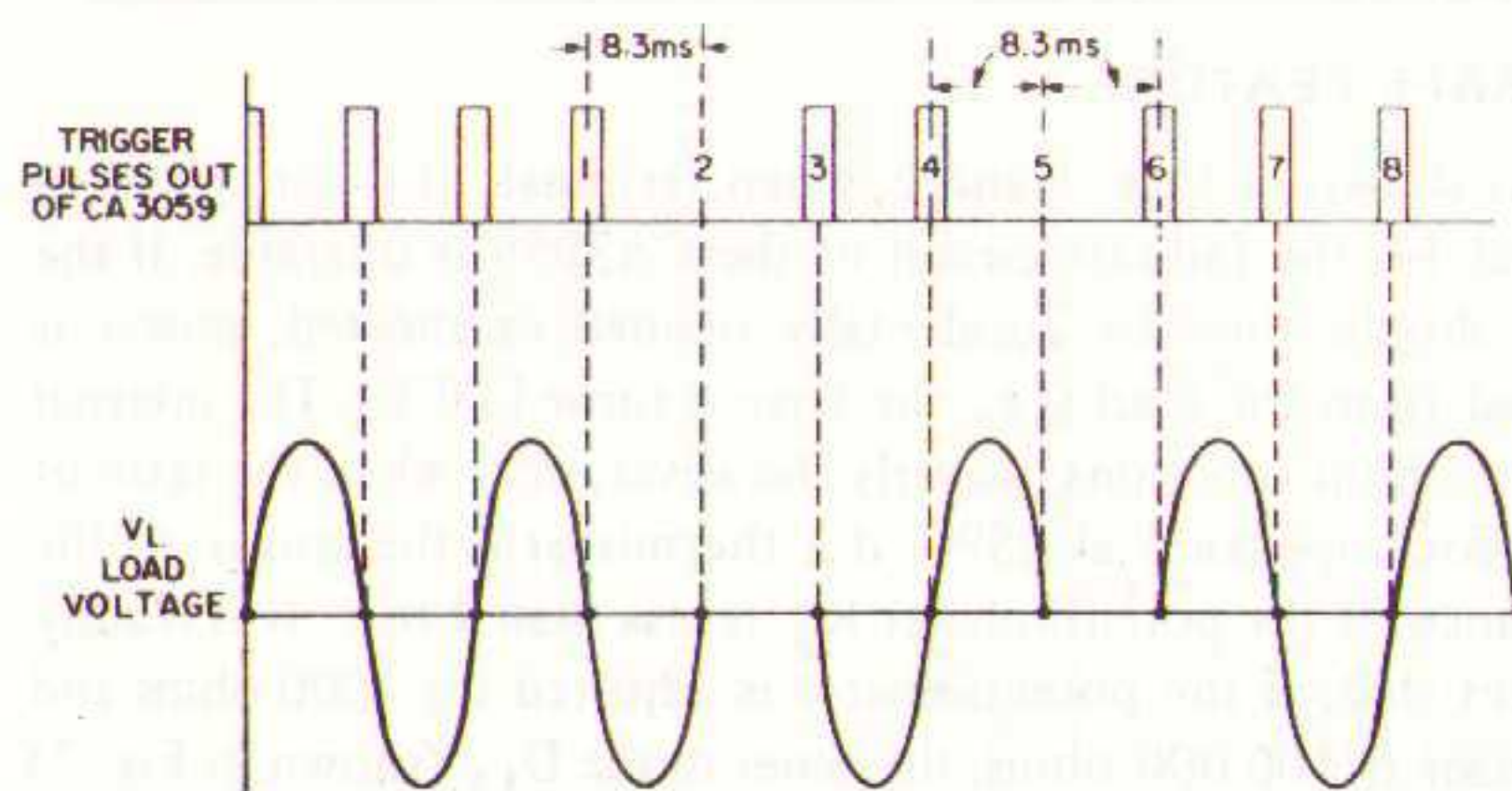


Fig.5 - Half-cycling phenomenon in the CA3059.

point. Fig. 5 illustrates this phenomenon. The CA3059 senses the zero-voltage crossing every half-cycle and an output, for example pulse No. 4, is produced to indicate the zero crossing. During the remaining 8.3 milliseconds, however, the differential amplifier in the CA3059 may change state and inhibit any further output pulses. The uncertainty region of the differential amplifier, therefore, prevents pulse No. 5 from triggering the triac during the negative excursion of the ac line voltage.

Several solutions exist for elimination of the half-cycling phenomenon. If the user can tolerate some hysteresis in the control, then positive feedback can be added around the differential amplifier. Fig. 6 illustrates this technique. The tabular data in the figure lists the recommended values of R_1 and R_2 for different sensor impedances at the control point.

If a significant amount (greater than $\pm 10\%$) of controlled hysteresis is required, then the circuit shown in Fig. 7 may be employed. In this configuration, external transistor Q_1 provides a means for addition of positive feedback to the CA3059. It should be noted that the signal developed at the collector to Q_1 could perhaps be used to provide an auxiliary time-delay function.

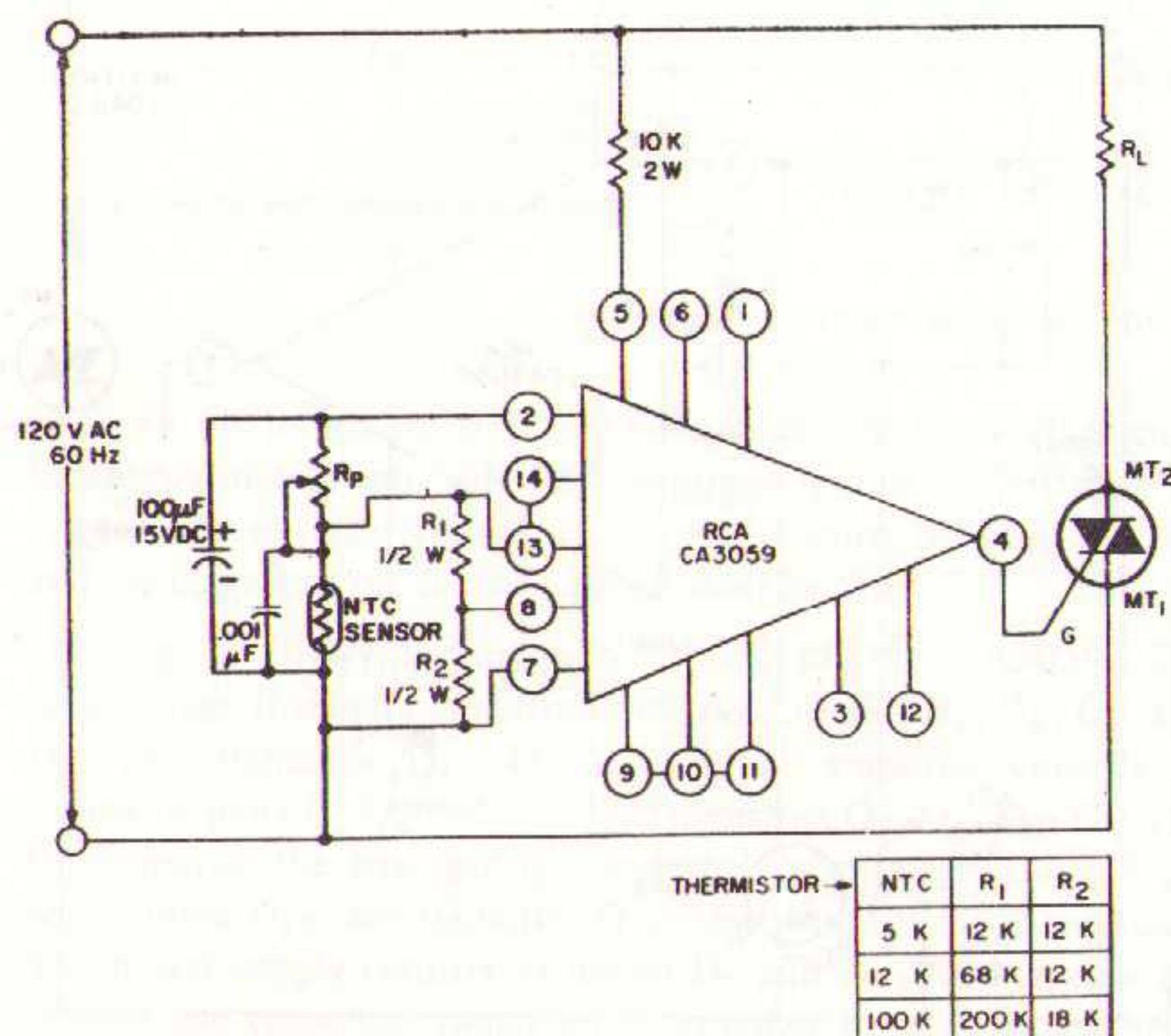


Fig.6 - CA3059 on-off controller with hysteresis.

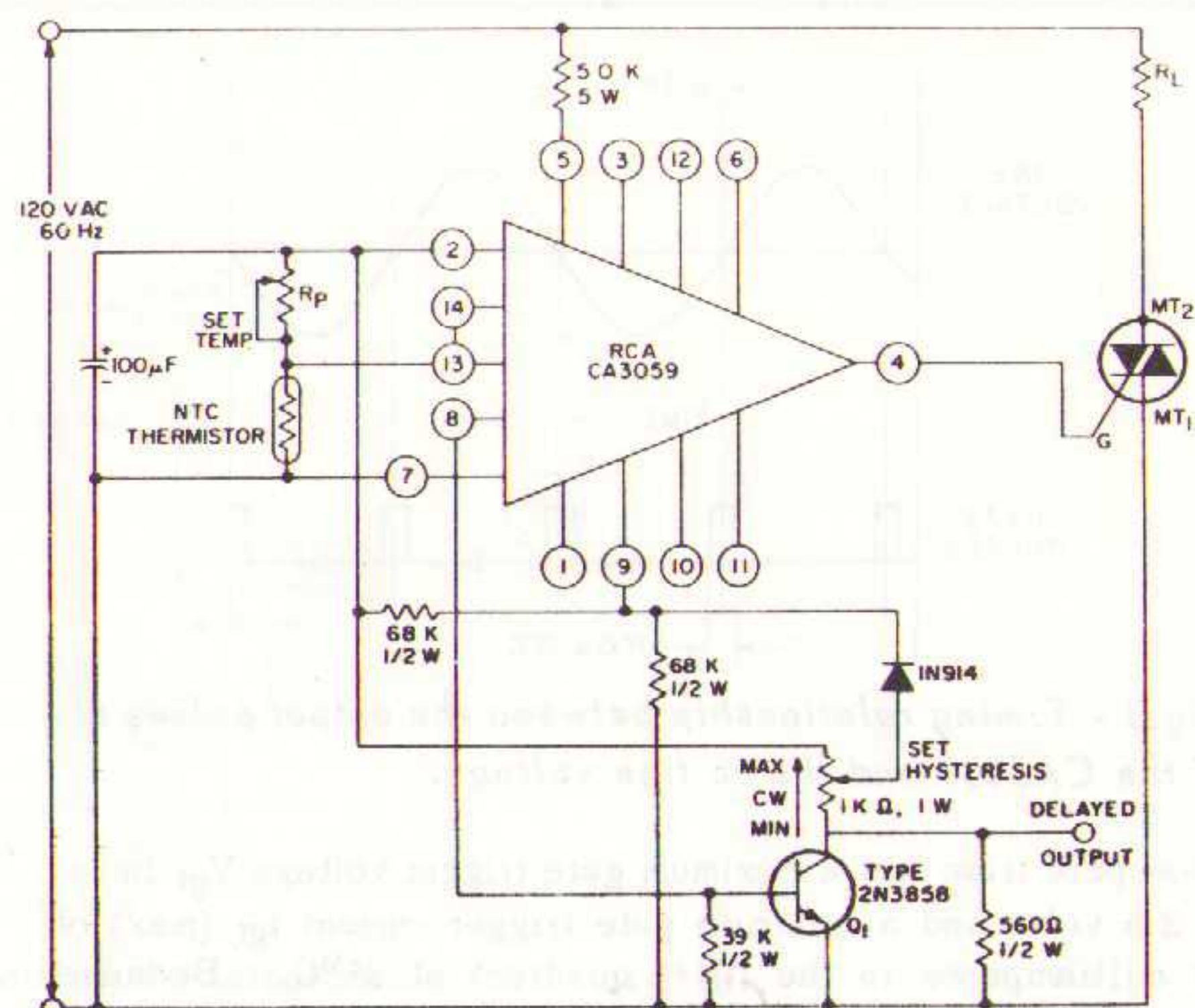


Fig.7 - CA3059 on-off controller with controlled hysteresis.

For applications which require complete elimination of half-cycling without the addition of hysteresis, the *integral-cycle* temperature controller described in a later section of this Note, which senses the zero-voltage crossing only once during the ac power cycle, can be used.

TEMPERATURE CONTROLLERS

Fig. 8 shows a triac used in an on-off temperature-controller configuration. The triac is turned on at zero voltage whenever the

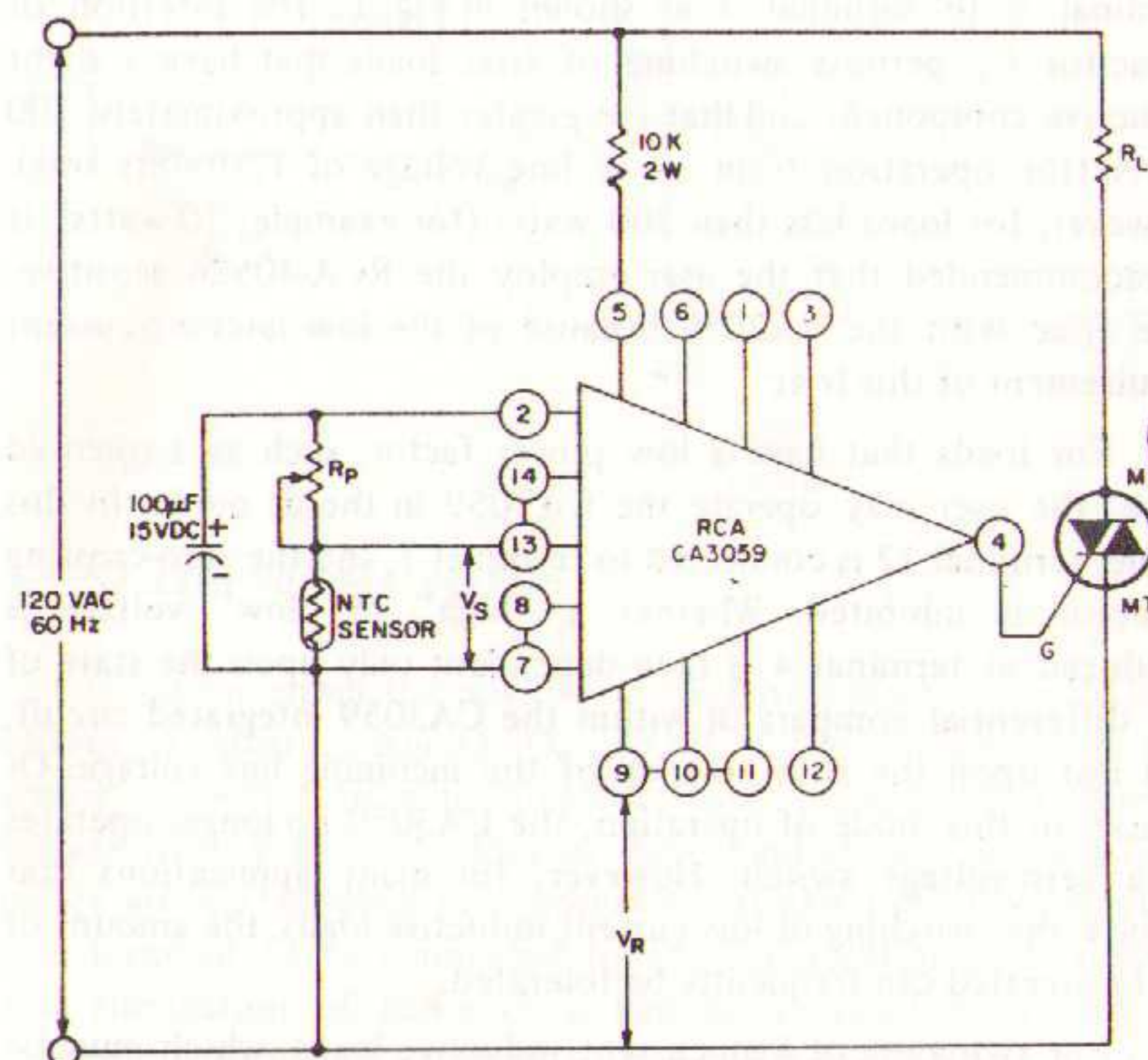


Fig.8 - CA3059 on-off temperature controller.

voltage V_s exceeds the reference voltage V_r . The transfer characteristic of this system, shown in Fig. 9(a), indicates significant thermal overshoots and undershoots, a well-known characteristic of such a system. The differential or hysteresis of this

system, however, can be further increased, if desired, by the addition of positive feedback.

For precise temperature-control applications, the proportional-control technique with synchronous switching is employed. The transfer curve for this type of controller is shown in Fig. 9(b). In this case, the duty cycle of the power supplied to the load is varied with the demand for heat required and the thermal time constant (inertia) of the system. For example, when the temperature setting is increased in an "on-off" type of controller, full power (100 per cent duty cycle) is supplied to the system. This effect results in significant temperature excursions *because there is no anticipatory circuit* to reduce the power gradually before the actual set temperature is achieved. However, in a proportional control technique, less power is supplied to the load (reduced duty cycle) as the error signal is reduced (sensed temperature approaches the set temperature).

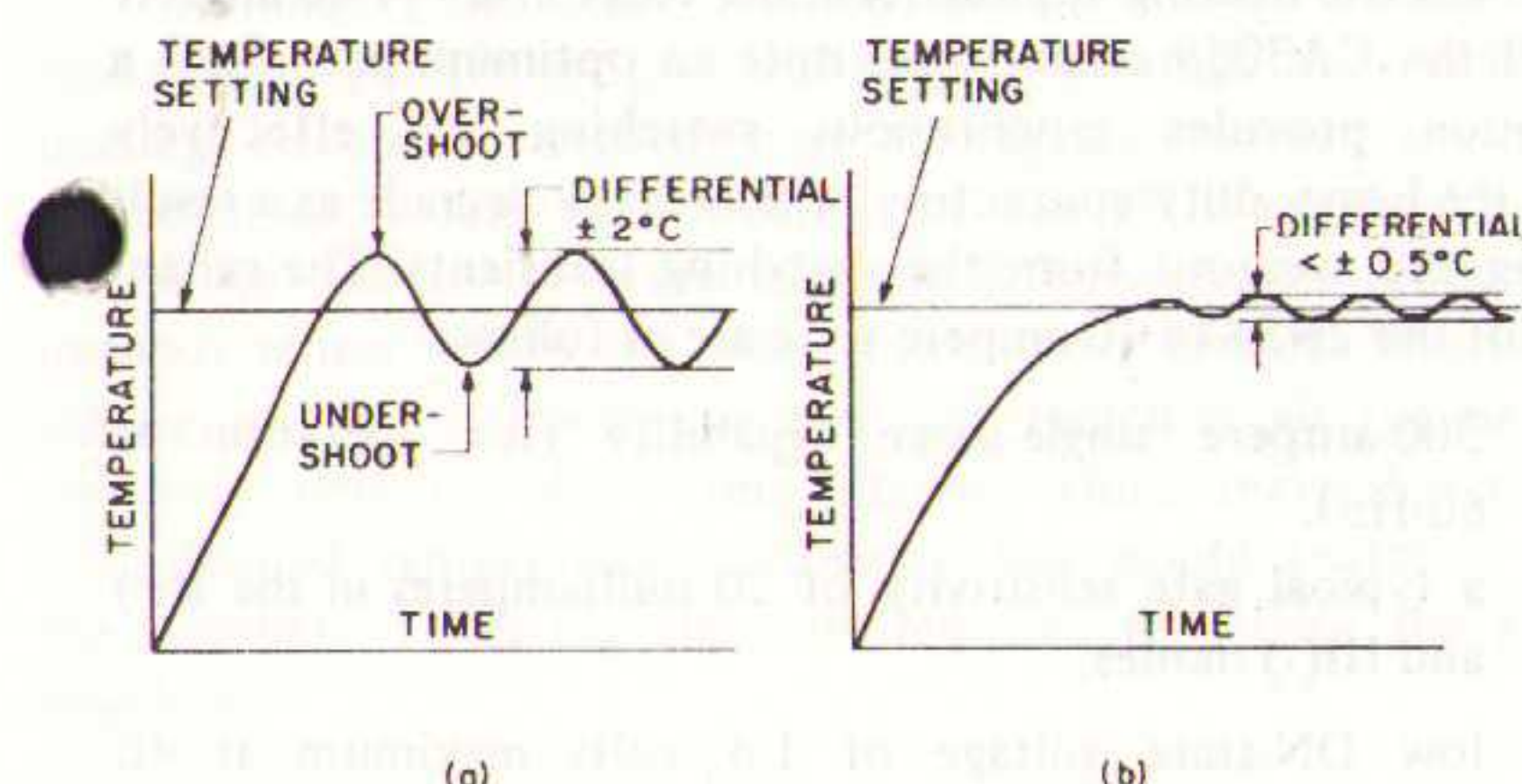


Fig.9 - Transfer characteristics of (a) on-off and (b) proportional control systems.

Before such a system is implemented, a time base is chosen so that the ON-time of the triac is varied within this time base. The ratio of the ON-to-OFF time of the triac within this time interval depends on the thermal time constant of the system and the selected temperature setting. Fig. 10 illustrates the principle of proportional control. For this operation, power is supplied to the load until the ramp voltage reaches a value greater than the dc control signal supplied to the opposite side of the differential amplifier. The triac then remains OFF for the remainder of the time-base period. As a result, power is "proportioned" to the load in a direct relation to the heat demanded by the system.

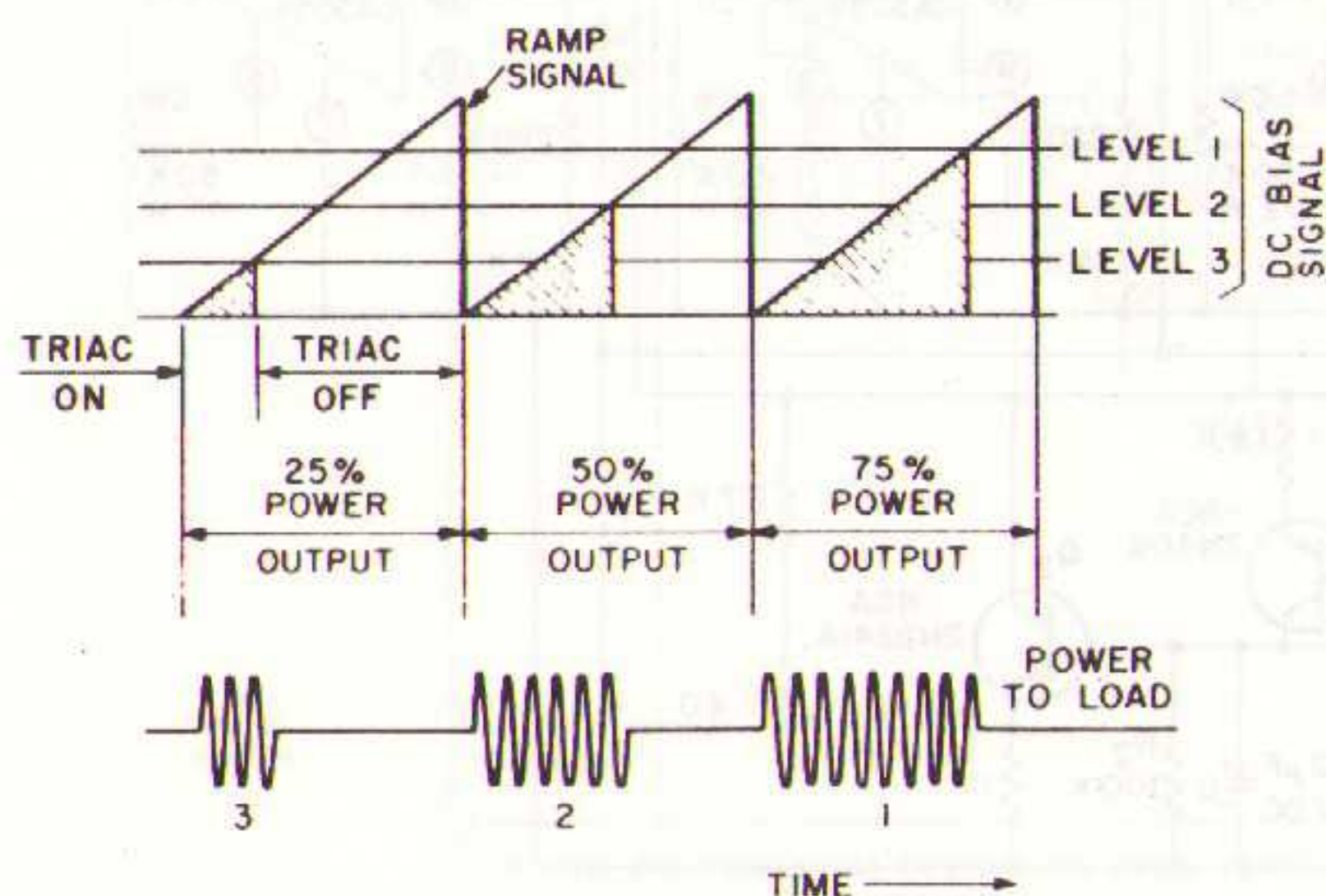


Fig.10 - Principles of proportional control.

For this application, a simple ramp generator can be realized with a minimum number of active and passive components. It is noted that a ramp having good linearity is not required for proportional operation because of the nonlinearity of the thermal system and the closed-loop type of control. In the circuit shown in Fig. 11, ramp voltage is generated when the capacitor C_1 charges through resistors R_0 and R_1 . The time base of the ramp is determined by resistors R_2 and R_3 , capacitor C_2 , and the breakover voltage of the 1N5411 diac. When the voltage across C_2 reaches approximately 32 volts, the diac switches and turns on the 2N3241A transistor. The capacitor C_1 then discharges through the collector-to-emitter junction of the transistor. This discharge time is the retrace or flyback time of the ramp. The circuit shown can generate ramp times ranging from 0.3 to 2.0 seconds through

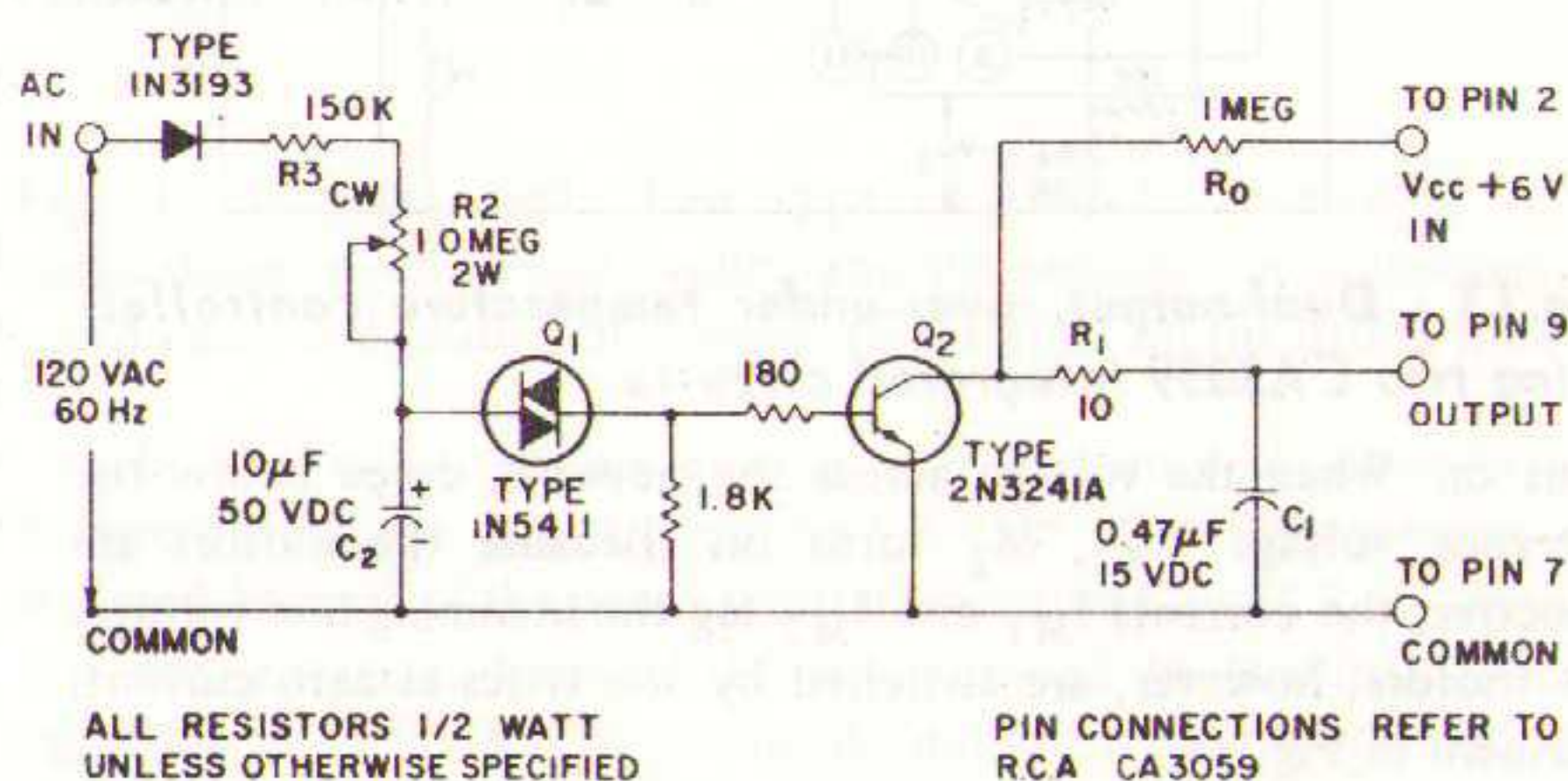


Fig.11 - Ramp generator.

adjustment of R_2 . For precise temperature regulation, the time base of the ramp should be shorter than the thermal time constant of the system, but long with the respect to the period of the 60-Hz line voltage. Fig. 12 shows a triac connected for the proportional mode.

Fig. 13 shows a dual-output temperature controller that drives two triacs. When the voltage V_s developed across the temperature-sensing network exceeds the reference voltage V_{R1} , motor M_1

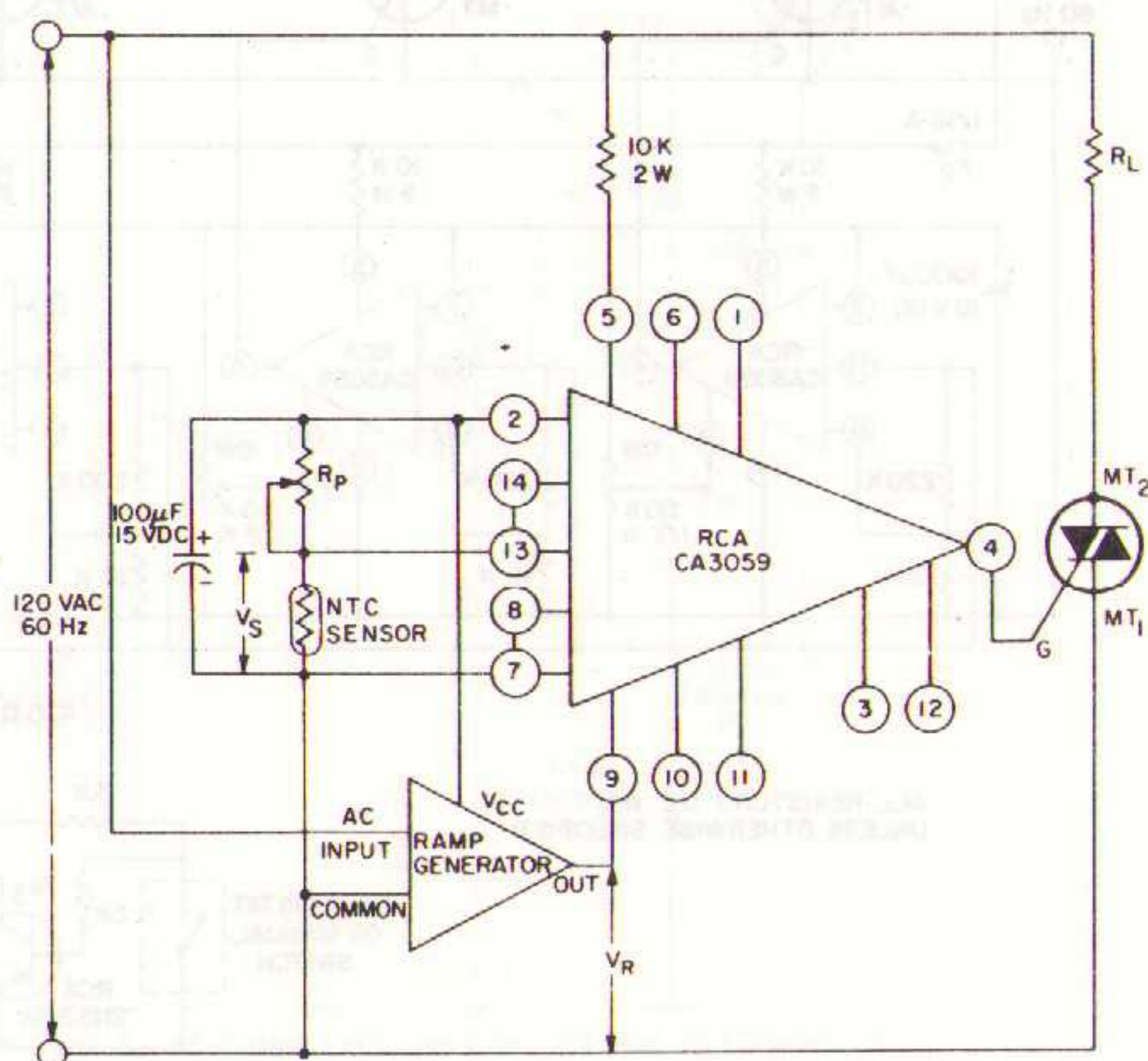


Fig.12 - CA3059 proportional temperature controller.

as heavy as 5 kilowatts are switched sequentially at zero voltage to eliminate RFI and prevent a dip in line voltage that would occur if the full 25 kilowatts were to be switched simultaneously.

Transistor Q_1 is used as a constant-current source to charge capacitor C in a linear manner. Transistor Q_2 acts as a buffer stage. When the thermostat is closed, a ramp voltage is provided at output E_o . At approximately 3-second intervals, each 5-kilowatt heating element is switched onto the power system by its respective triac. When there is no further demand for heat, the thermostat opens, and capacitor C discharges through R_1 and R_2 to cause each triac to turn OFF in the reverse heating sequence. It should be noted that some half-cycling occurs before the heating element is switched fully ON. This condition can be attributed to the inherent dissymmetry of the CA3059 and is further aggravated by the slow-rising ramp voltage applied to one of the inputs. The timing diagram in Fig. 16 shows the turn-on and turn-off sequence of the heating system being controlled.

Seemingly, the basic method shown in Fig. 15 could be modified to provide proportional control in which the number of heating elements switched into the system, under any given normal load, would be a function of the BTU's required by the system or the temperature differential between an indoor and outdoor sensor within the total system environment. That is, the closing of the thermostat would not switch in all the heating elements within a short time interval, which inevitably results in undesired temperature excursions, but would switch in only the number of heating elements required to satisfy the actual heat load.

Integral-Cycle Temperature Controller (No half-cycling)

If a temperature controller which is completely devoid of half-cycling and hysteresis is required, then the circuit shown in

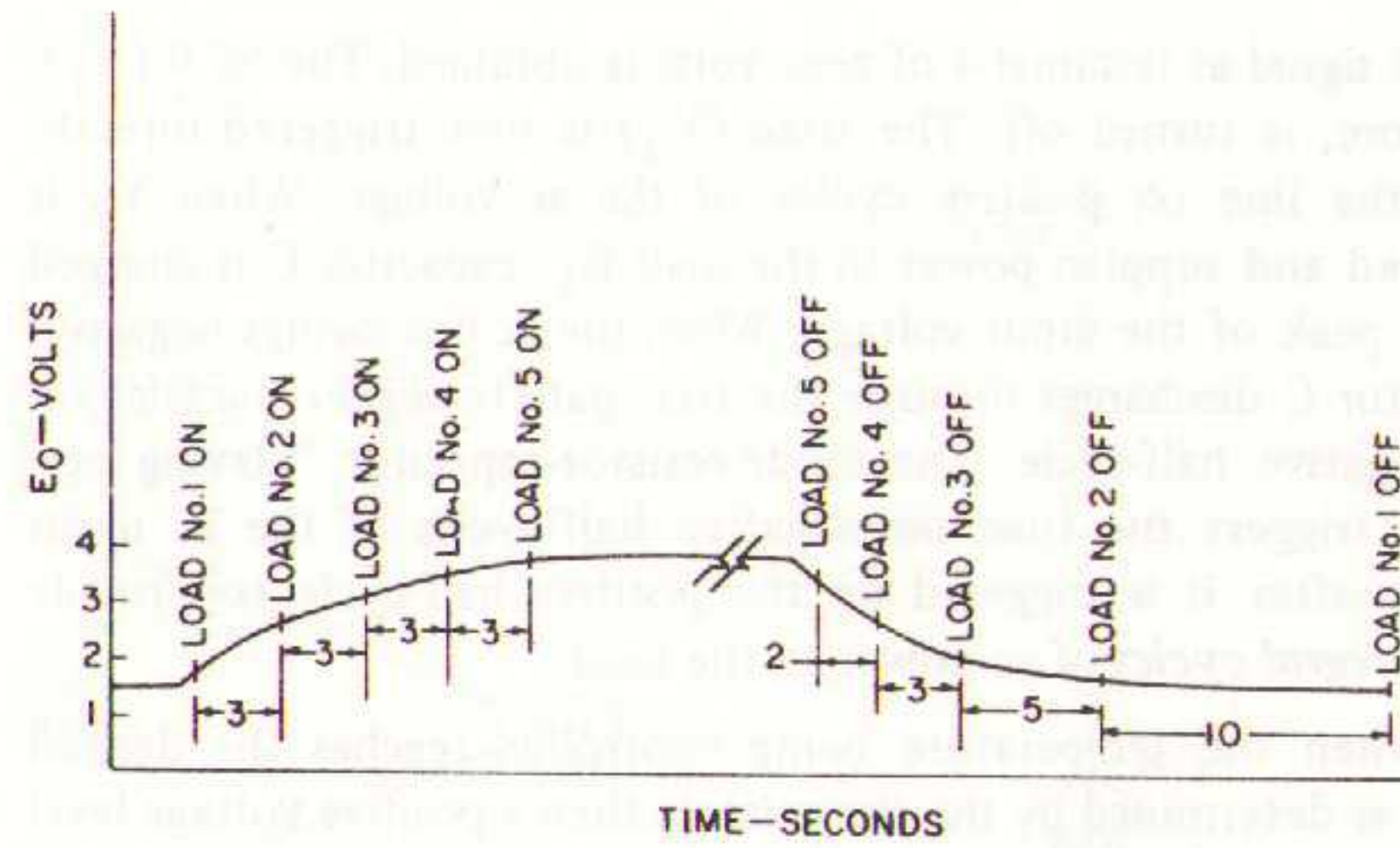


Fig. 16 - Ramp-voltage waveform for the heat-staging controller.

Fig. 17 may be used. This type of circuit is essential for applications in which half-cycling and the resultant dc component could cause overheating of a power transformer on the utility lines.

In the circuit shown in Fig. 17, the sensor is connected between terminals 7 and 9 of the CA3059. This arrangement is required because of the phase reversal introduced by SCR Y_1 . With this configuration, terminal 12 is connected to terminal 7 for operation of the CA3059 in the dc mode (however, the load is switched at zero voltage). Because the position of the sensor has been changed for this configuration, the internal fail-safe circuit cannot be used (terminals 13 and 14 are not connected).

In the integral-cycle controller, when the temperature being controlled is low, the resistance of the thermistor is high and an

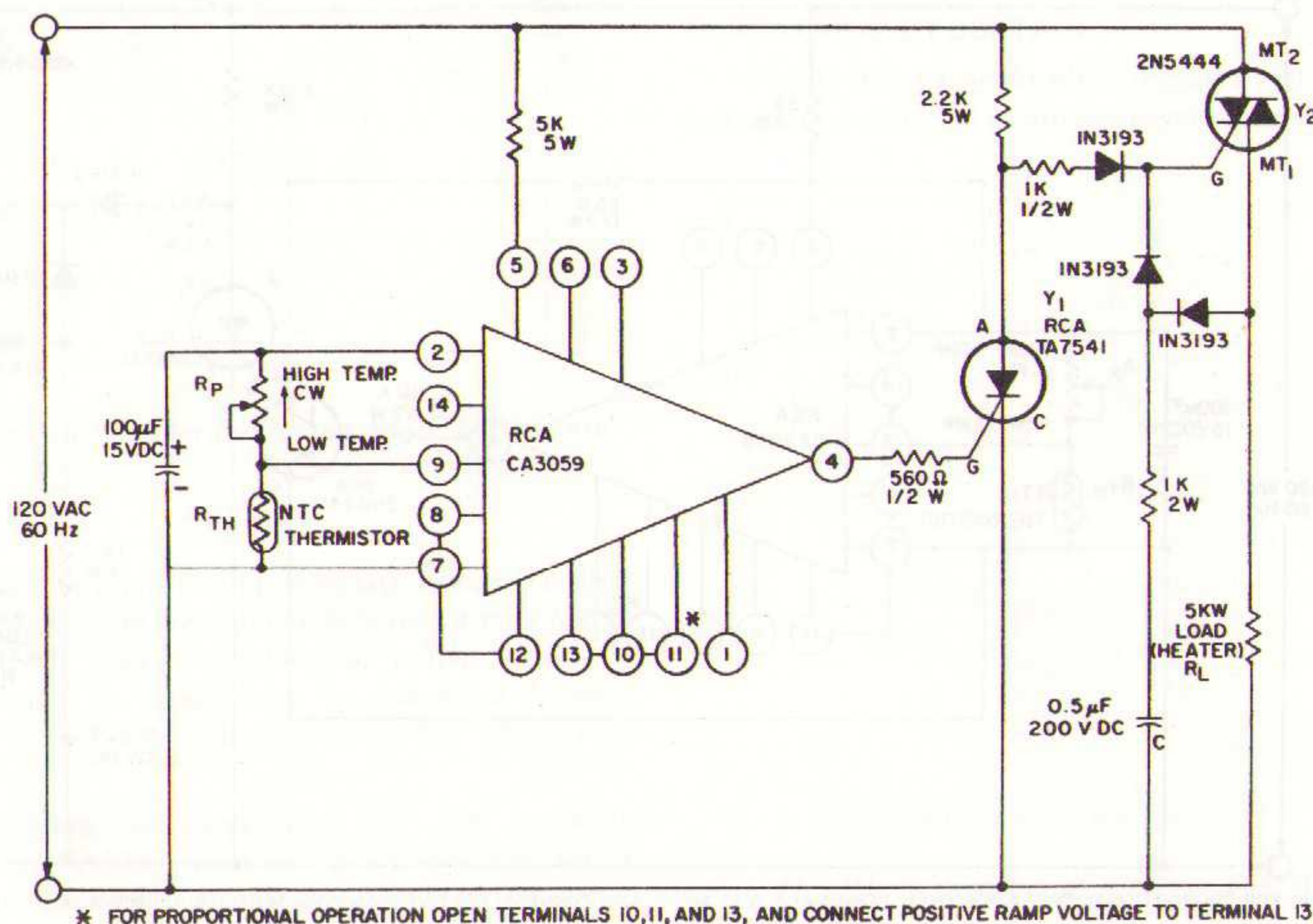


Fig. 17 - CA3059 integral-cycle temperature controller in which half-cycling effect is eliminated.

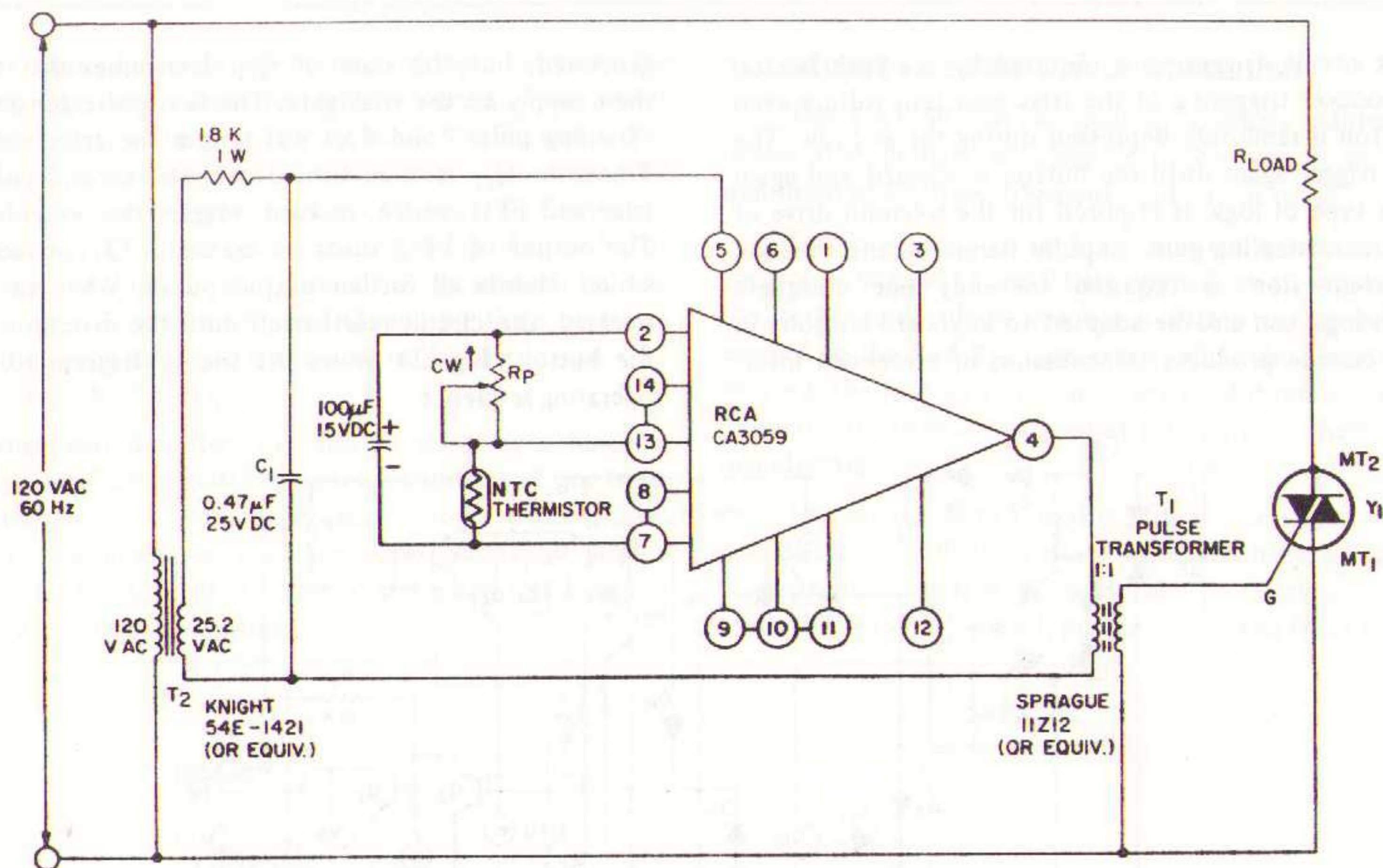


Fig. 19 - CA3059 on-off controller with an isolated sensor.

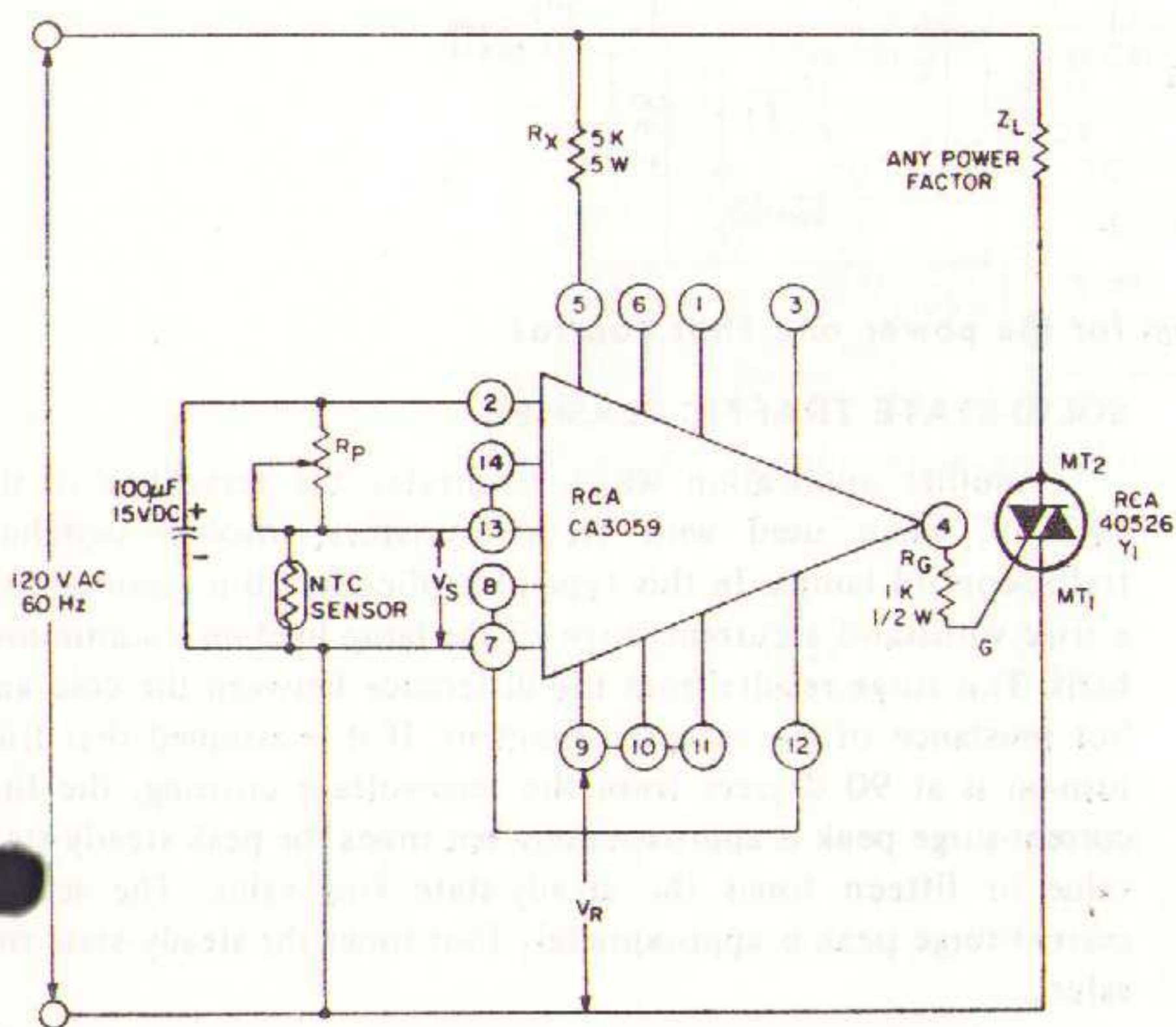


Fig. 20 - Differential comparator using the CA3059 integrated circuit.

external load current as a function of dc voltage from terminals 2 and 7. This curve is shown in the technical bulletin for the CA3059 integrated circuit. Of course, if additional output current is required, an external dc supply may be connected between terminals 2 and 7, and resistor R_X (shown in Fig. 20) may be removed.

The chart below compares some of the operating characteristics of the CA3059, when used as a comparator, with a typical high-performance commercially available integrated-circuit differential comparator.

PARAMETERS	CA3059 (typical values)	Typical Integrated- Circuit Comparator (710)
1. Sensitivity	50 mV	2mV
2. Switching speed (rise time)	$> 20 \mu s$	90 ns
3. Output drive capability	*4.5V at $\leq 4mA$	3.2V at $\leq 5.0mA$

*Refer to Figure 20; R_X equals 5000 ohms.

POWER ONE-SHOT CONTROL

Fig. 21 shows a circuit which triggers a triac for one complete half-cycle of either the positive or negative alternation of the ac line

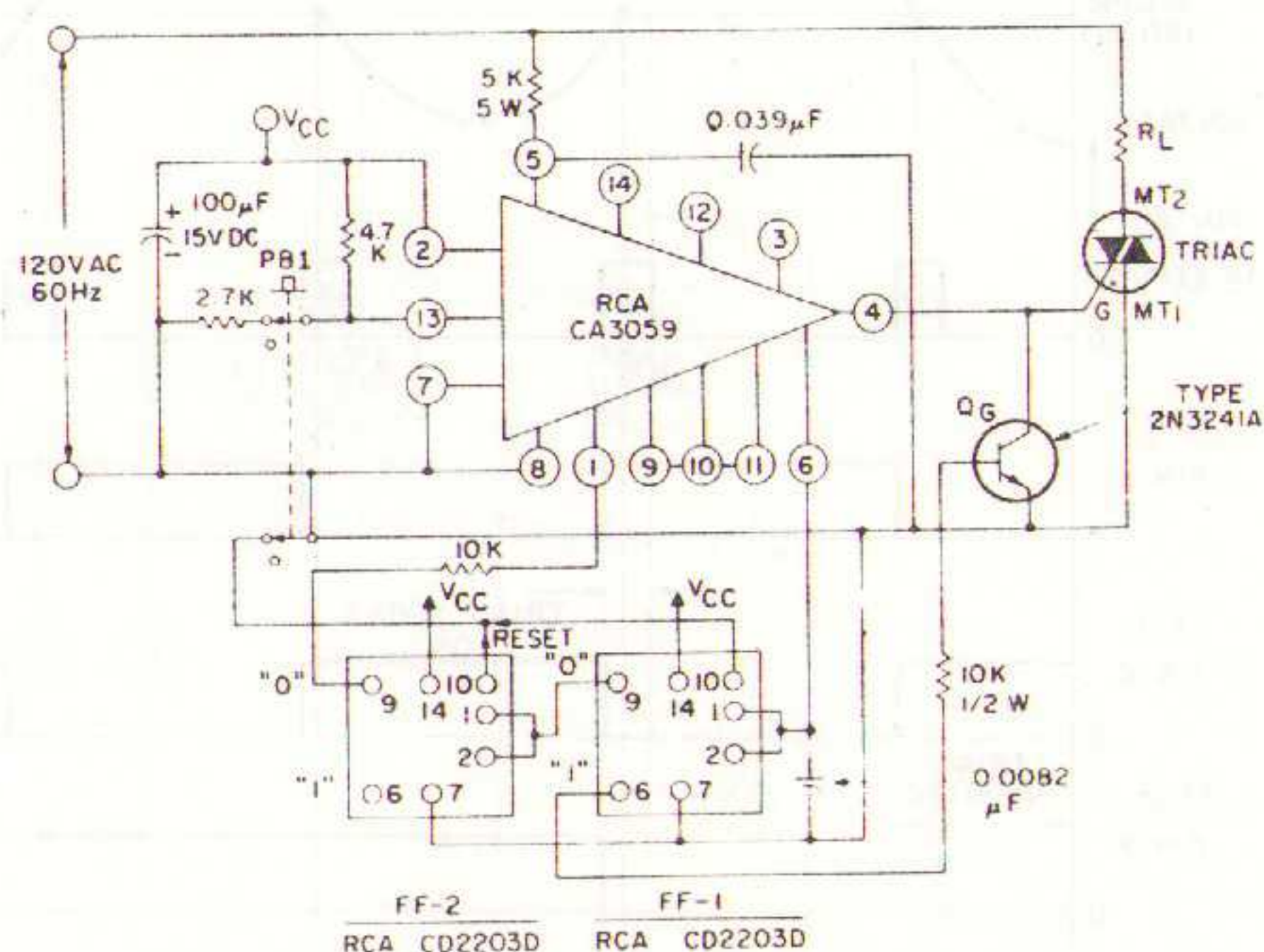
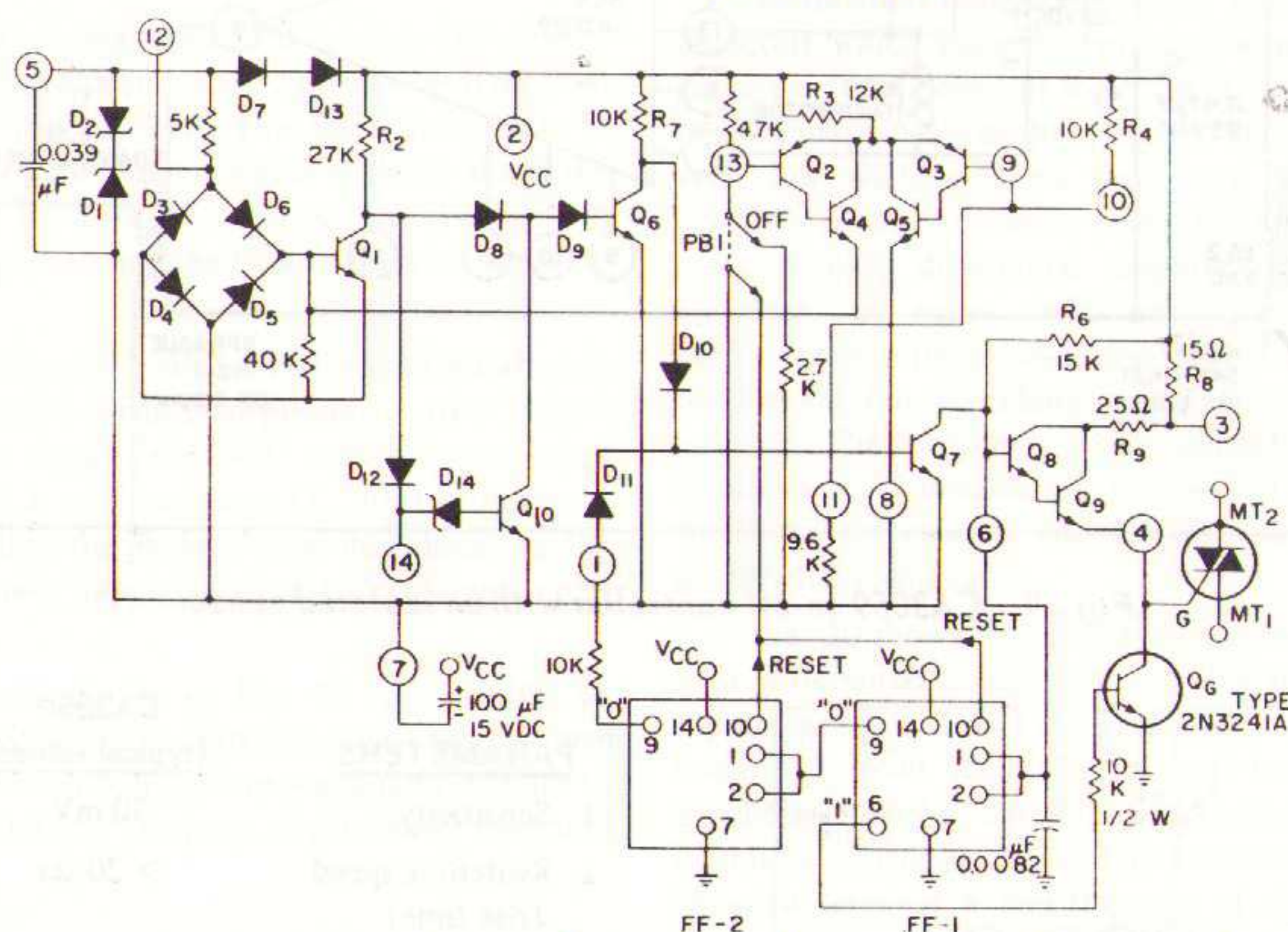
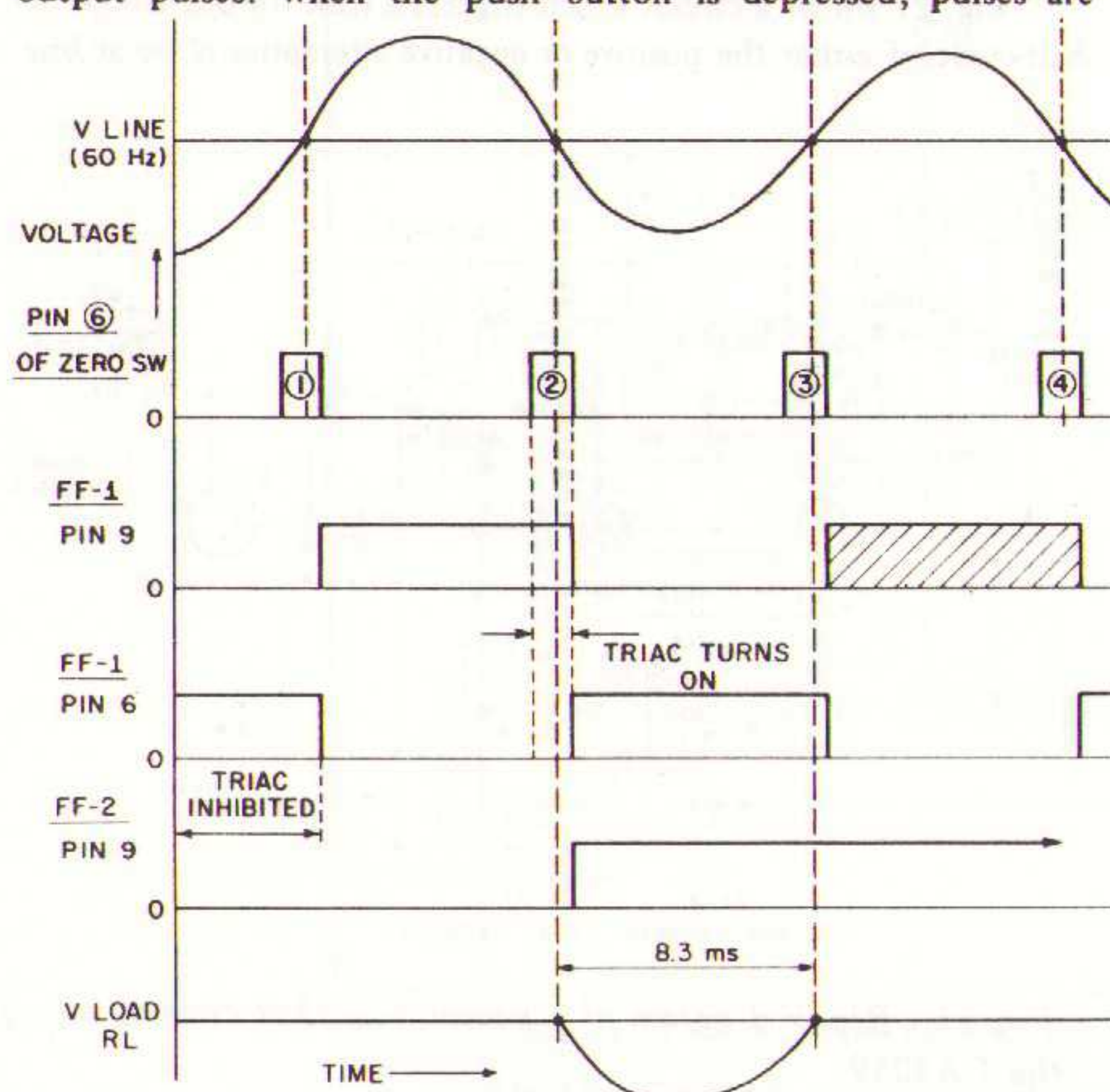


Fig. 21 - Block diagram of a power one-shot control using the CA3059.

generated, but the state of Q_G determines the requirement for their supply to the triac gate. The first pulse generated serves as a "framing pulse" and does not trigger the triac but toggles FF-1. Transistor Q_G is then turned off. The second pulse triggers the triac and FF-1 which, in turn, toggles the second flip-flop FF-2. The output of FF-2 turns on transistor Q_7 , as shown in Fig. 22, which inhibits all further output pulses. When the pushbutton is released, the circuit resets itself until the process is repeated with the button. Fig. 23 shows the timing diagram for the described operating sequence.



In the circuit of Fig. 21, before the button is depressed, both flip-flop outputs are in the "zero" state. Transistor Q_G is biased ON by the output of flip-flop FF-1. The differential comparator which is part of the CA3059 circuit is initially biased to inhibit output pulses. When the push button is depressed, pulses are



SOLID-STATE TRAFFIC FLASHER

Another application which illustrates the versatility of the CA3059, when used with RCA thyristors, involves switching traffic-control lamps. In this type of application, it is essential that a triac withstand a current surge of the lamp load on a continuous basis. This surge results from the difference between the cold and hot resistance of the tungsten filament. If it is assumed that triac turn-on is at 90 degrees from the zero-voltage crossing, the first current-surge peak is approximately ten times the peak steady-state value or fifteen times the steady-state rms value. The second current-surge peak is approximately four times the steady-state rms value.

When the triac randomly switches the lamp, the rate of current rise di/dt is limited only by the source inductance. The triac di/dt rating may be exceeded in some power systems. In many cases, exceeding the rating results in excessive current concentrations in a small area of the device which may produce a hot spot and lead to device failure. Critical applications of this nature require adequate drive to the triac gate for fast turn-on. In this case, some inductance may be required in the load circuit to reduce the initial magnitude of the load current when the triac is passing through the active region. Another method may be used which involves the switching of the triac at zero line voltage. This method involves the supply of pulses to the triac gate only during the presence of zero voltage on the ac line.

Fig. 24 shows a circuit in which the lamp loads are switched at zero line voltage. This approach reduces the initial di/dt ,

decreases the required triac surge-current ratings, increases the operating lamp life, and eliminates RFI problems. This circuit consists of two triacs, a flip-flop (FF-1), the CA3059, and a diac pulse generator. The flashing rate in this circuit is controlled by potentiometer R, which provides between 10 and 120 flashes per minute. The state of FF-1 determines the triggering of triacs Y_1 or Y_2 by the output pulses at terminal 4 generated by the zero-crossing circuit. Transistors Q_1 and Q_2 inhibit these pulses to the gates of the triacs until the triacs turn on by the logical "1" (V_{CC} high) state of the flip-flop.

TRANSIENT-FREE SWITCH CONTROLLER

The circuit shown in Fig. 25 is connected so that, after the control terminals (14 and 7) are opened, electronic logic waits until the power-line voltage reaches a zero crossing before power is applied to the load Z_L . Conversely, when the control terminals are shorted, the load current continues until it reaches a zero crossing. This circuit can switch a load at zero current whether it is resistive or inductive.

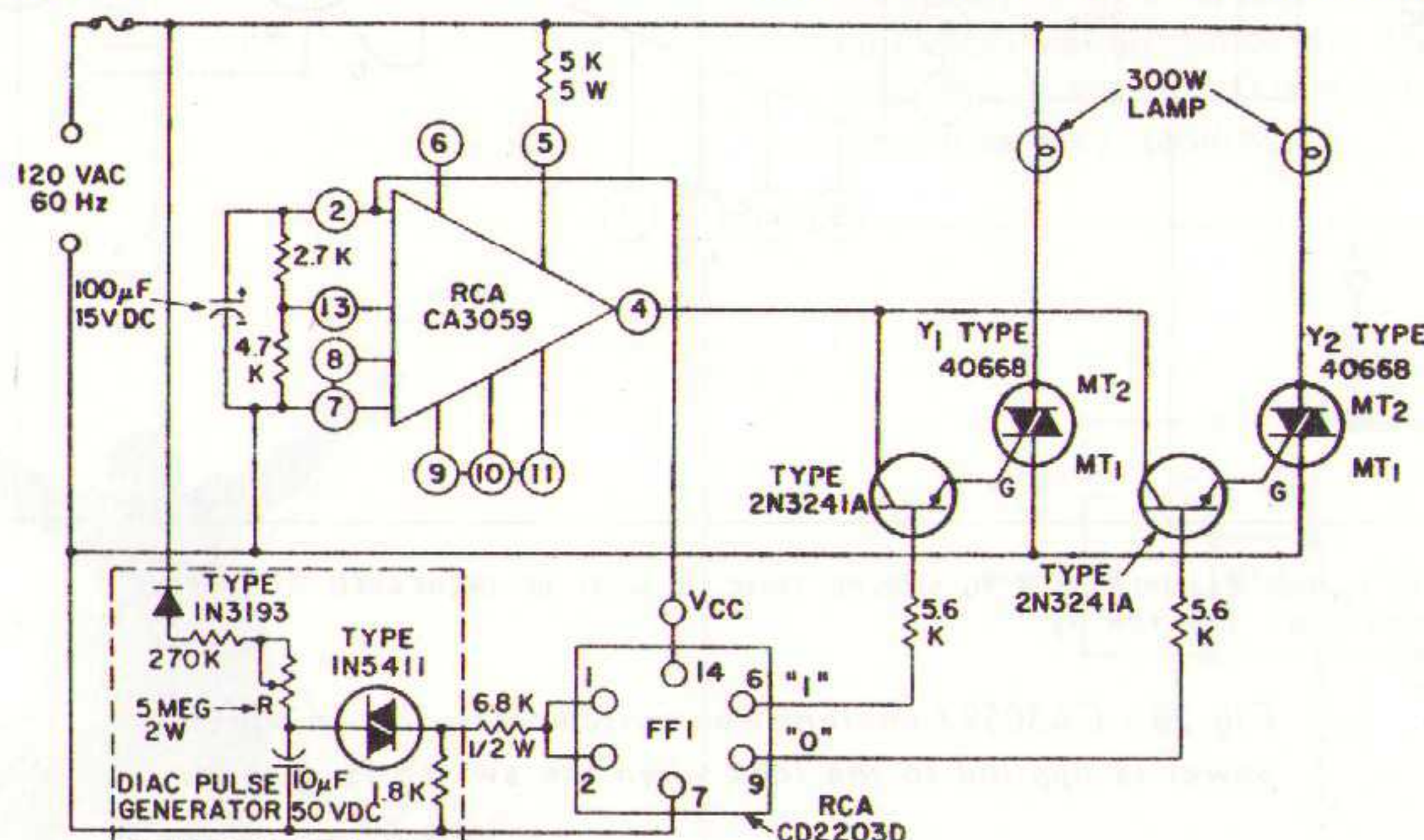


Fig.24 - Synchronous-switching traffic flasher using the CA3059.

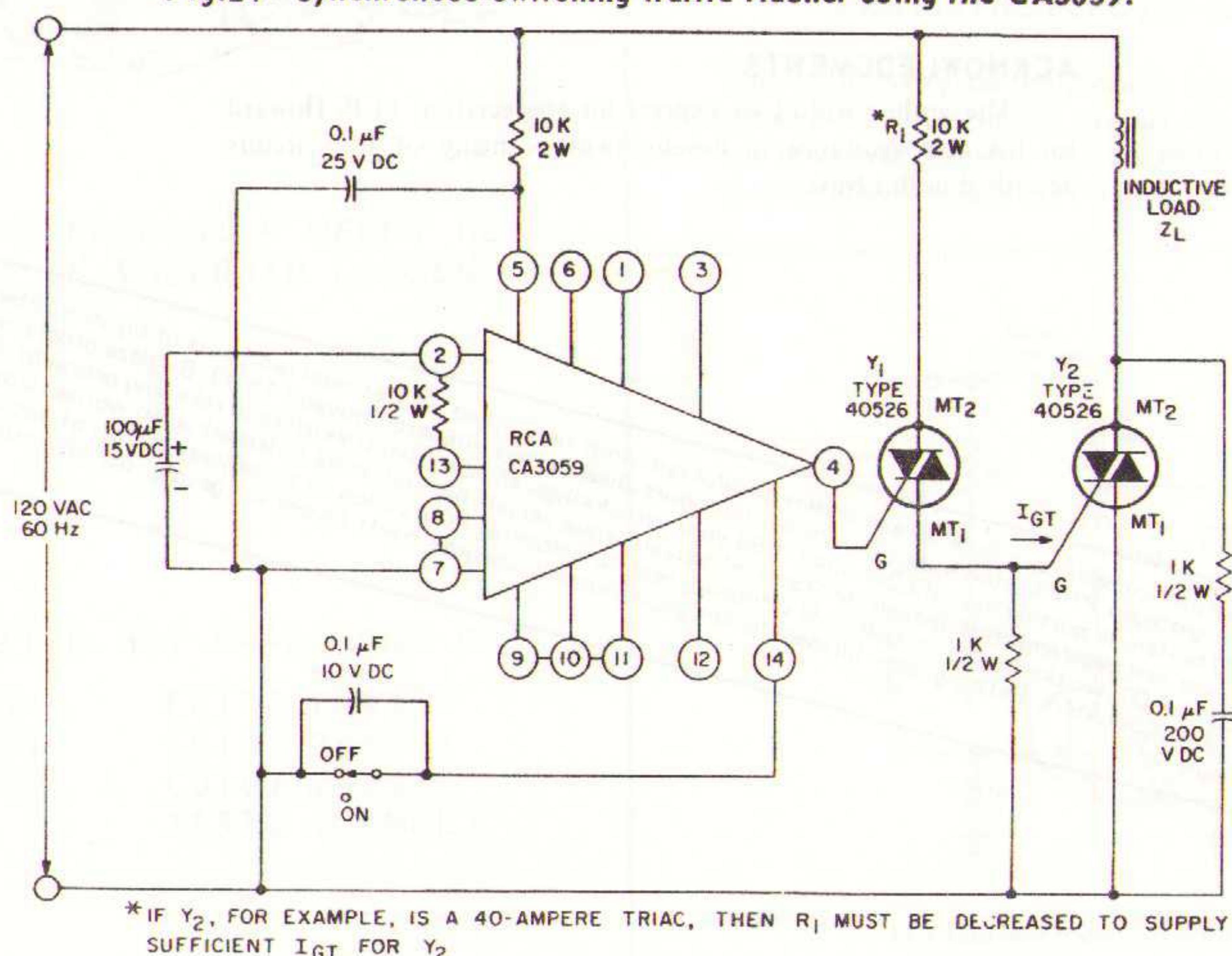
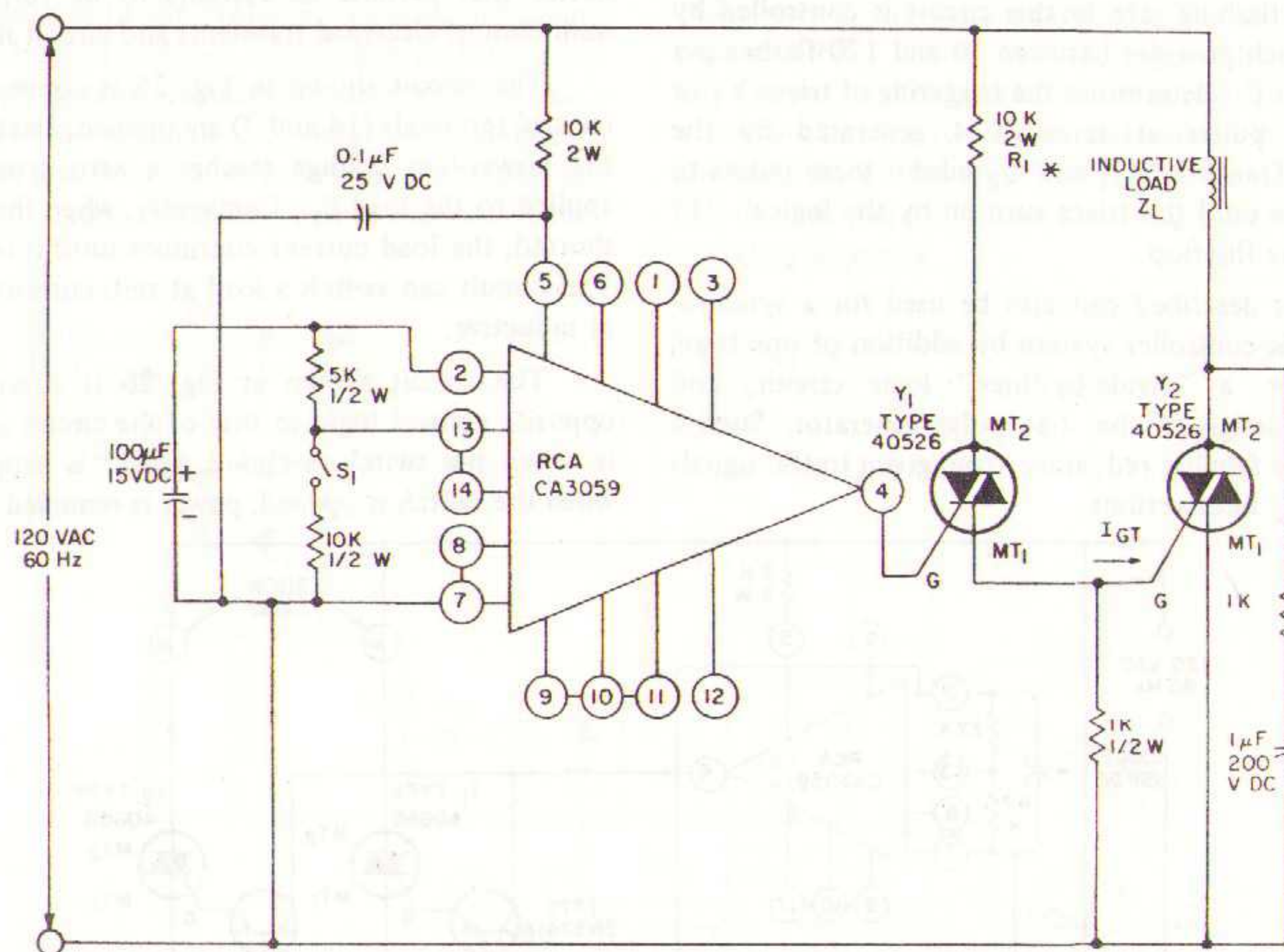


Fig.25 - CA3059 transient-free switch controller in which power is supplied to the load when the switch is open.

In both configurations, the maximum rms load current that can be switched depends on the rating of triac Y_2 . If Y_2 is an RCA-2N5444 triac, an rms current of 40 amperes can be switched.



* IF Y_2 , FOR EXAMPLE, IS A 40-AMPERE TRIAC, R_1 MUST BE DECREASED TO SUPPLY SUFFICIENT I_{GT} FOR Y_2 .

Fig.26 - CA3059 transient-free switch controller in which power is applied to the load when the switch is closed.

ACKNOWLEDGMENTS

The author wishes to express his appreciation to P. Howard for his able assistance in development of many of the circuits described in this Note.

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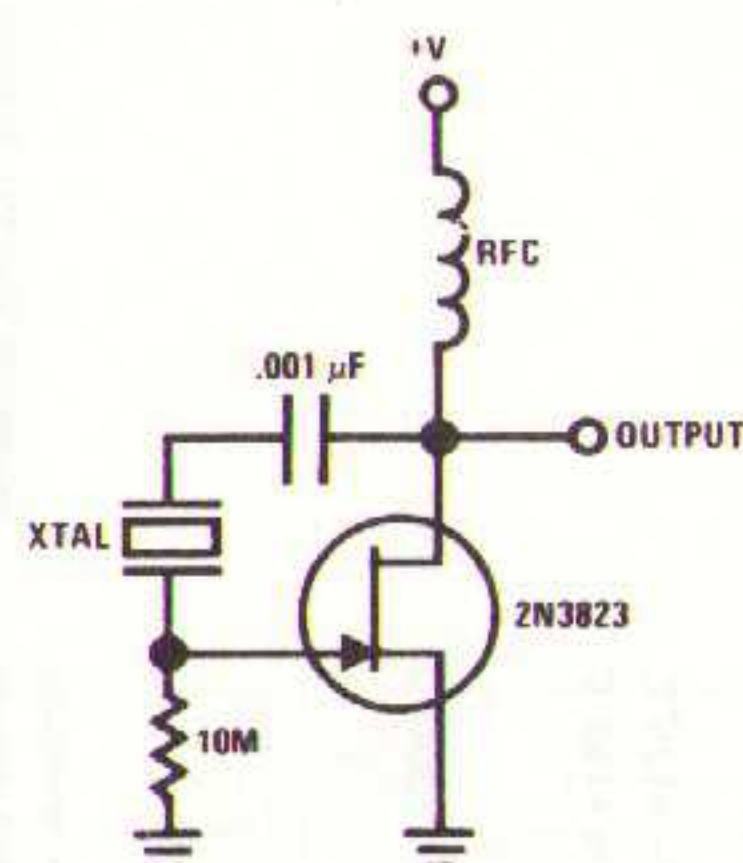
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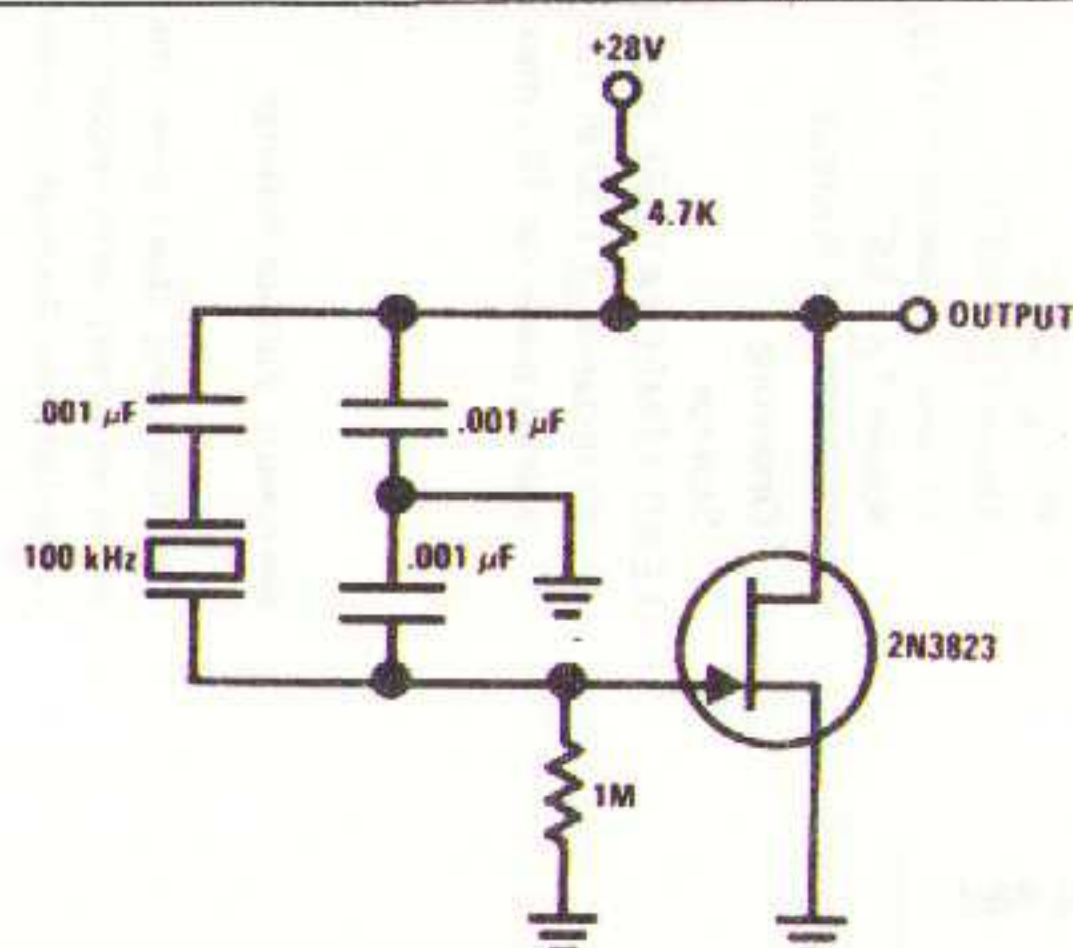
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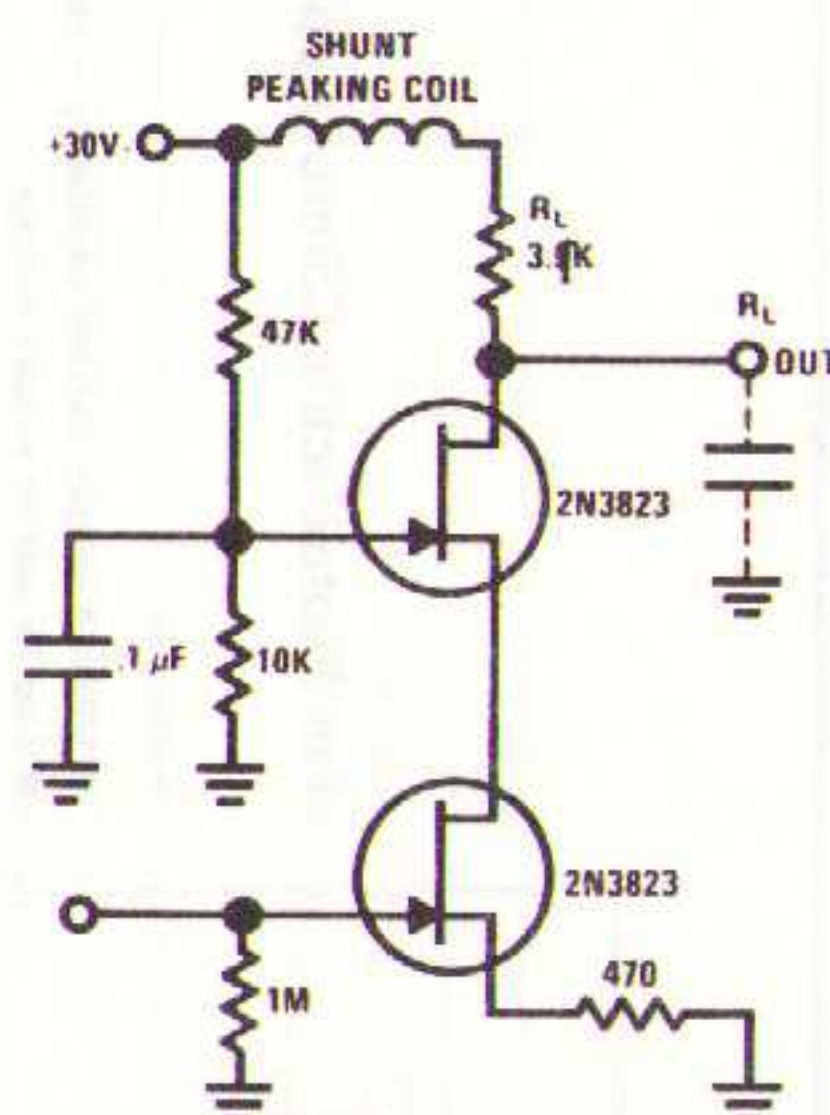
JFET Pierce Crystal Oscillator

The JFET Pierce crystal oscillator allows a wide frequency range of crystals to be used without circuit modification. Since the JFET gate does not load the crystal, good Q is maintained thus insuring good frequency stability.



Stable Low Frequency Crystal Oscillator

This Colpitts-Crystal oscillator is ideal for low frequency crystal oscillator circuits. Excellent stability is assured because the 2N3823 JFET circuit loading does not vary with temperature.



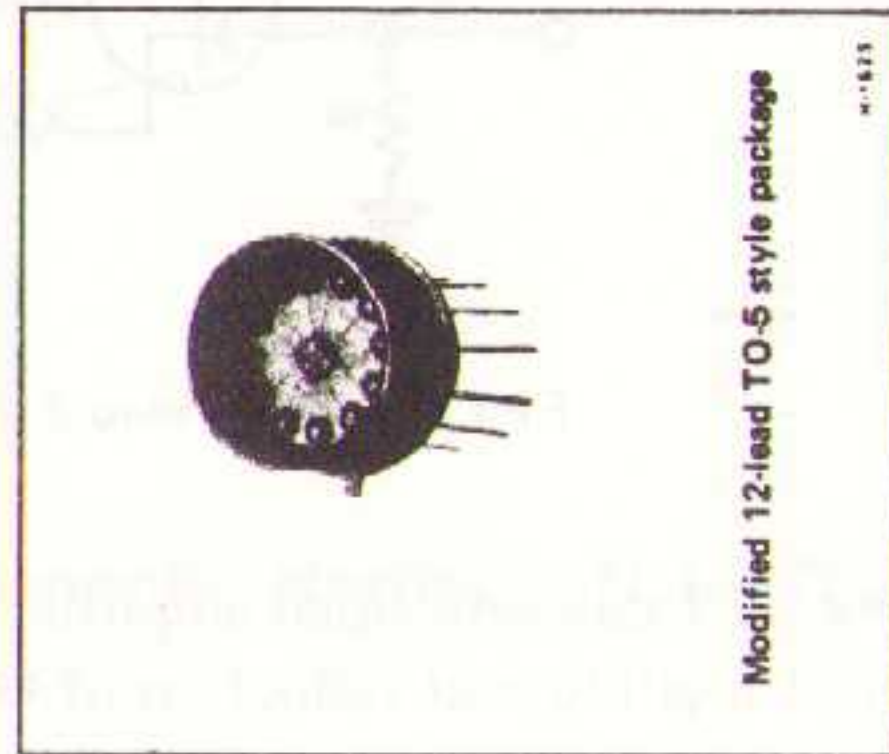
FET Cascode Video Amplifier

The FET cascode video amplifier features very low input loading and reduction of feedback to almost zero. The 2N3823 is used because of its low capacitance and high Y_{fs} . Bandwidth of this amplifier is limited by R_L and load capacitance.

Photo Detector and Power Amplifier

CA3062

File No. 421



Modified 12-lead TO-5 style package

For Photoelectric Control Applications

Features

- 100 mA output-current capability — can drive a relay or thyristor directly
- 5 to 15 volt dc supply voltage
- Compact — complete system in a TO-5 style package
- Compatible with RCA-40736R Infrared Emitter

The CA3062* is an integrated circuit consisting of a photosensitive section, an amplifier, and a pair of high-current output transistors on a single monolithic chip.

The photosensitive section consists of Darlington pairs and affords high sensitivity. The power amplifier has a differential configuration which provides complementing outputs in response to a light input — normally "ON" and normally "OFF". The separate photodetector, amplifier, and high-current switch provide flexibility of circuit arrangement. This feature plus the high current capability of the output section, can now provide the user with a complete system particularly useful in photoelectric control applications utilizing IR emitters and visible-light sources.

*Formerly developmental type TA5371B.

Applications

- Counters
- Sorting
- Level controls
- Inspection
- Intrusion alarms
- Position sensor
- Edge monitoring
- Isolators

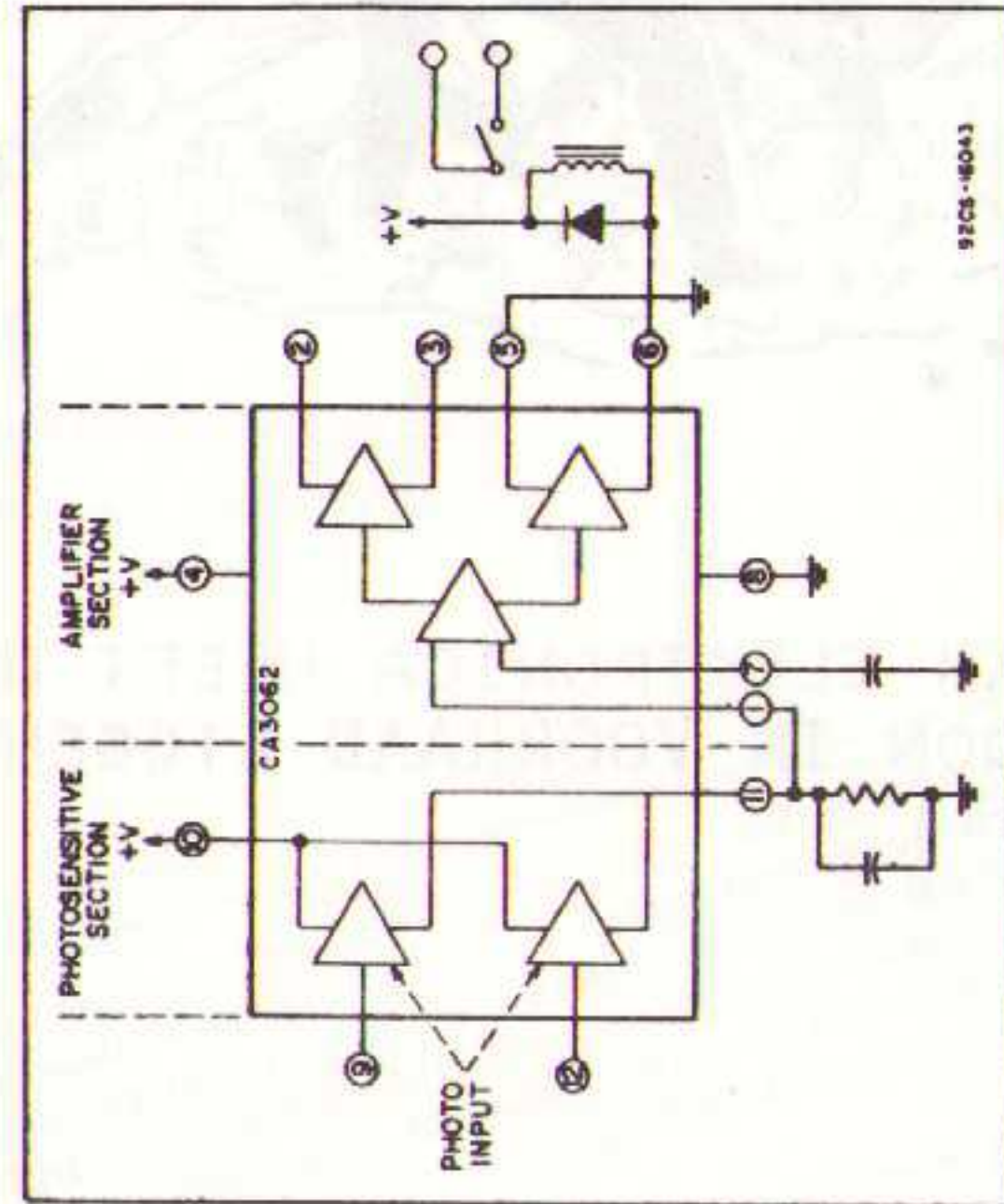


Fig. 1 - Light operated relay using CA3062.

ABSOLUTE-MAXIMUM RATINGS

DISSIPATION:
At $T_A = 25^\circ\text{C}$ 700 mW
Above $T_A = 25^\circ\text{C}$ Derate linearly 5.6 mW/ $^\circ\text{C}$
At Case Temperature ($T_C \leq 55^\circ\text{C}$) 1.5 W
Above $T_C = 55^\circ\text{C}$ Derate linearly 16 mW/ $^\circ\text{C}$

TEMPERATURE RANGE:
Operating -55°C to $+125^\circ\text{C}$
Storage -65°C to $+150^\circ\text{C}$

LEAD TEMPERATURE (During soldering):
At distance $\geq 1/32$ in. (3.17 mm) from seating plane for 10 s max $+300^\circ\text{C}$

Maximum Voltage Ratings

The following chart gives the range of voltages which can be applied to the terminals listed vertically with respect to the terminals listed horizontally. For example, the voltage range between vertical terminal 2 and horizontal terminal 3 is +15 to 0 volts.

TERM INAL No.	10	11	12	1	2	3	4	5	6	7	8
9	0	+2	.	.	.	.	.	.	.	.	.
10	9	5	+9	.	.	.	.	.	.	.	.
11	0	0	0	.	.	.	.	.	.	.	.
12	.	.	.	.	.	.	.	.	.	.	.
1	.	.	.	.	.	.	.	.	.	.	.
2	.	.	.	.	.	.	.	.	.	.	.
3	.	.	.	.	.	.	.	.	.	.	.
4	.	.	.	.	.	.	.	.	.	.	.
5	.	.	.	.	.	.	.	.	.	.	.
6	.	.	.	.	.	.	.	.	.	.	.
7	.	.	.	.	.	.	.	.	.	.	.
8	.	.	.	.	.	.	.	.	.	.	.

Reference Substrate and Case

* Voltages are not normally applied between these terminals. Voltages appearing between these terminals will be safe if the specified limits between all other terminals are not exceeded.

ELECTRICAL CHARACTERISTICS at $T_A = 25^\circ\text{C}$

CHARACTERISTICS	SYMBOLS	TEST CONDITIONS	MEASURE- MENT TERMINAL Nos.	TEST CIR- CUIT FIG.	CA3062 LIMITS	TYPICAL CHARAC- TERISTICS CURVES FIG.
STATIC CHARACTERISTICS						
Photo Darlington Section: Collector to Emitter Breakdown Voltage	$V_{(BR)CEO}$	$E = 0 \text{ lumens/ft}^2$ $I_C = 1 \text{ mA}$	10-11	—	10	V
Emitter to Base Breakdown Voltage	$V_{(BR)EBO}$	$I_E = 0.1 \text{ mA}, E = 0$	9-11 12-11	—	10	V
Dark Current	I_{Dark}	$V_{CE} = 7.5 \text{ V}, E = 0$	10	—	0.1	μA
Photo Current	I_p	$V_{CE} = 7.5 \text{ V}$ $E = 8 \text{ lumens/ft}^2$	10	3	60	μA
Wavelength of Max. Sensitivity	λ_{max}				725	Note 2 nm
Relative Angular Sensitivity				—	—	5
Area of Each Photo Transistor				—	$1.3 \times 10^{-4} \text{ cm}^2$	6
Amplifier Section Output Transistor: Collector to Emitter Breakdown Voltage	$V_{(BR)CEO}$ $V_{(BR)EC}$	$I_C = 1 \text{ mA}$	2-3 6-5	—	15	V
Emitter to Base Breakdown Voltage	$V_{(BR)EBO}$ $V_{(BR)EB}$	$I_E = 1 \text{ mA}$	3-8 6-8	—	5	V
DC Supply Current	I_{Supply}	$V_A = 7.5 \text{ V}$	4	—	5.5	10 mA
Sensitivity: Illumination, For Normal "OFF" Output For Normal "ON" Output	E_{ON} E_{OFF}	Set light input for $I_E = 70 \text{ mA}$ Set light input for $I_E = 5 \text{ mA}$	6 2	7, 15, 17	8 10	Note 1, 3 lumens per ft ²
DYNAMIC CHARACTERISTICS						
Overall Response Time: Turn-On Time	t_{on}	$E = 700 \mu\text{W/cm}^2$ at $\lambda = 930 \text{ nm}$	—	12	—	38 μs
Rise Time	t_r				—	125 μs
Turn-Off Time	t_{off}				—	43 μs
Fall Time	t_f				—	20 μs

NOTES

- (1) Tungsten filament light source at a color temperature of 2854K.
- (2) One (1) nanometer = 10 Angstrom units.
- (3) A radiant flux density of $7.5 \mu\text{W/cm}^2$ at 725 nm produces the same photocurrent as 1 lumen/ft^2 from a tungsten filament lamp at a color temperature of 2854K.

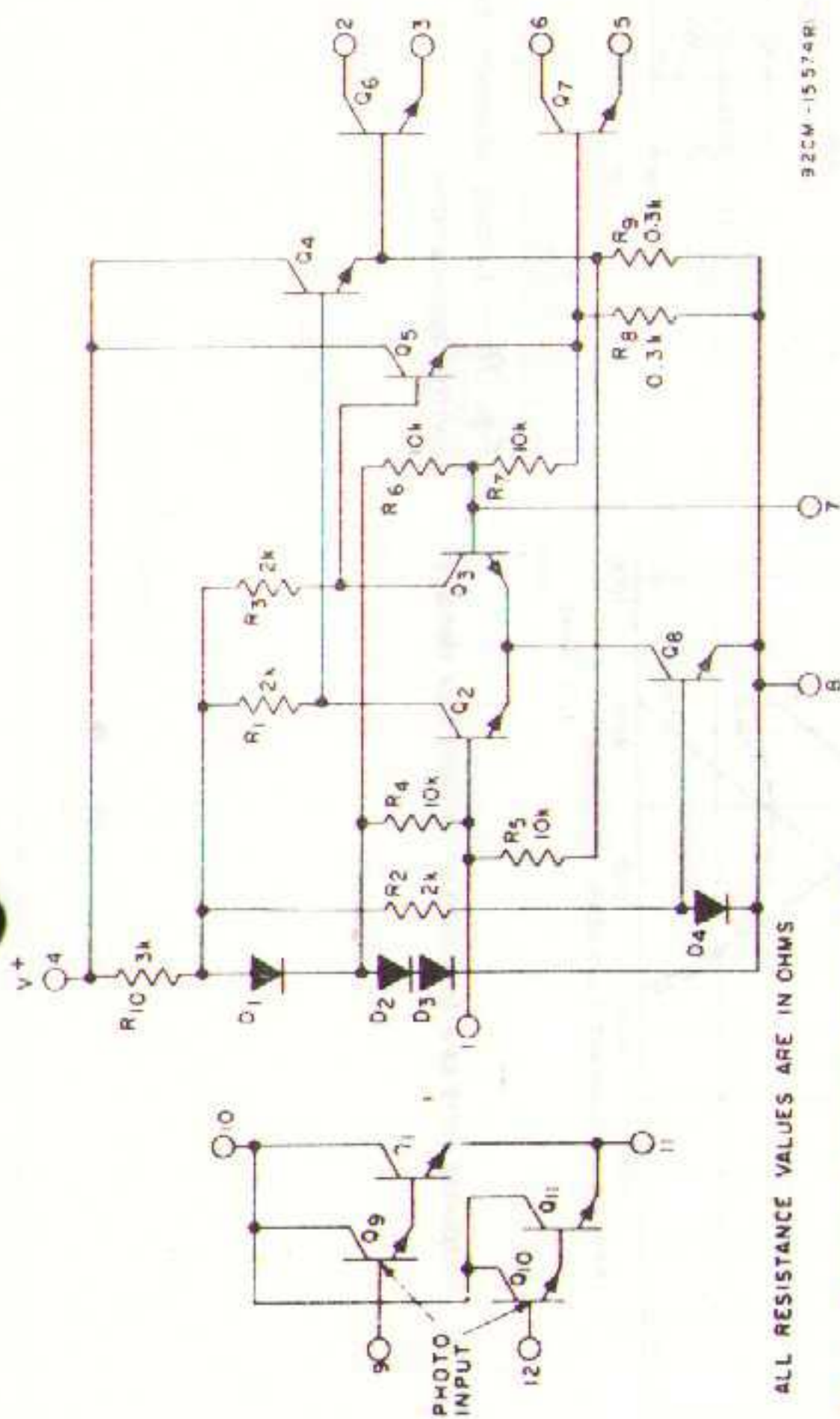


Fig. 2 - Schematic diagram of CA3062.

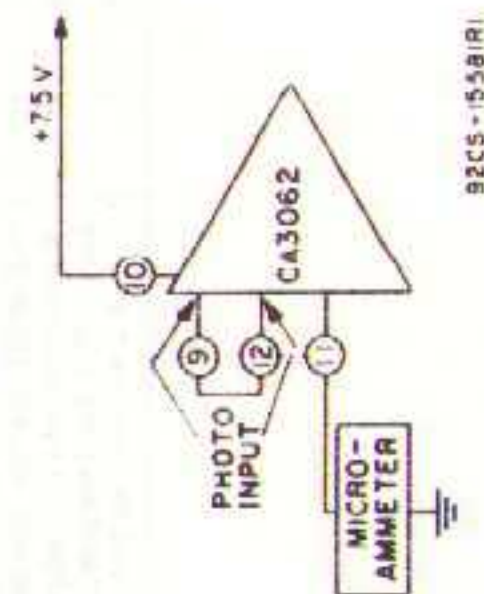


Fig. 3 - Test circuit for photocurrent and typical spectral response of photosensitive Darlington unit.

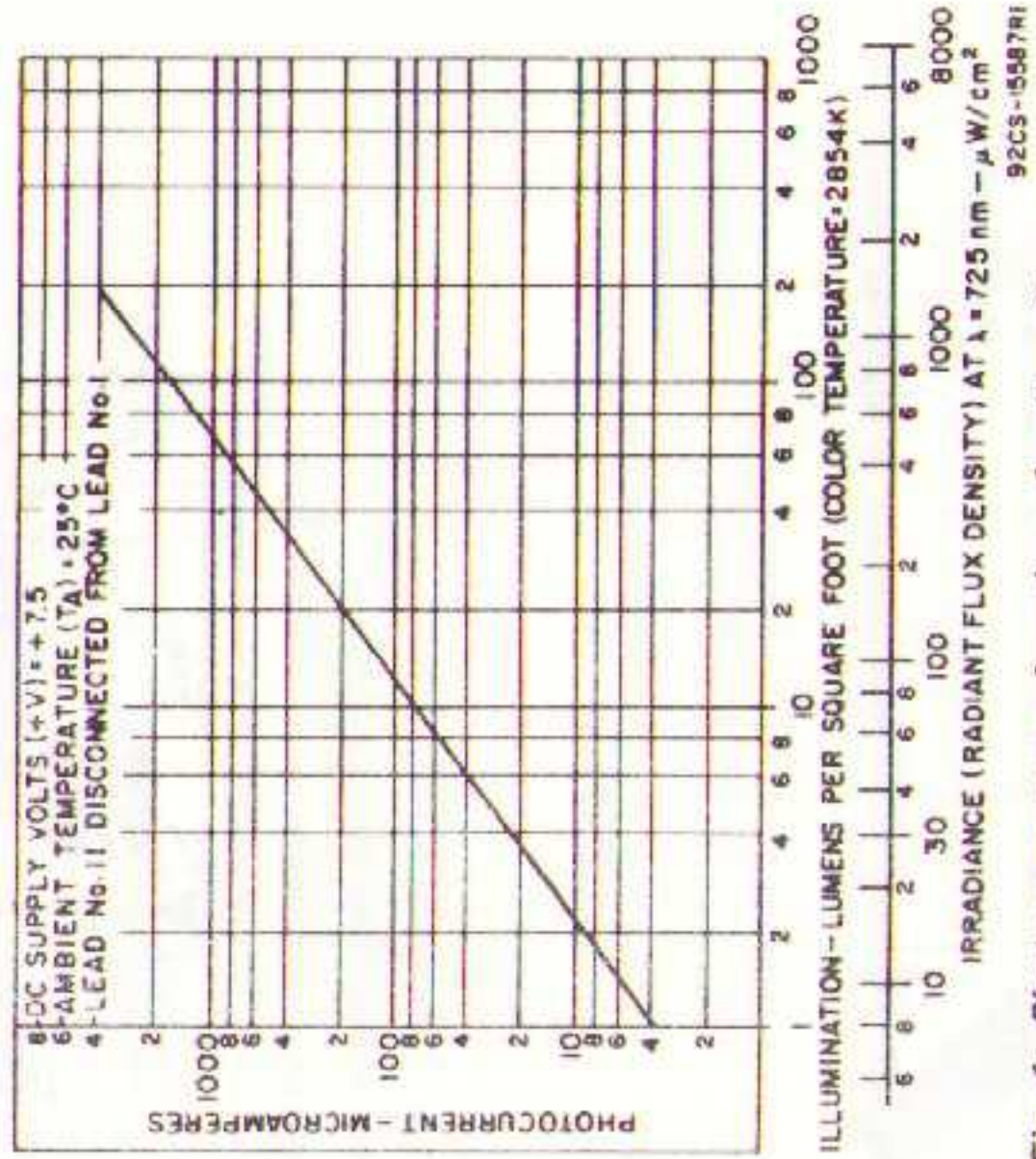


Fig. 4 - Photocurrent as a function of radiant flux.

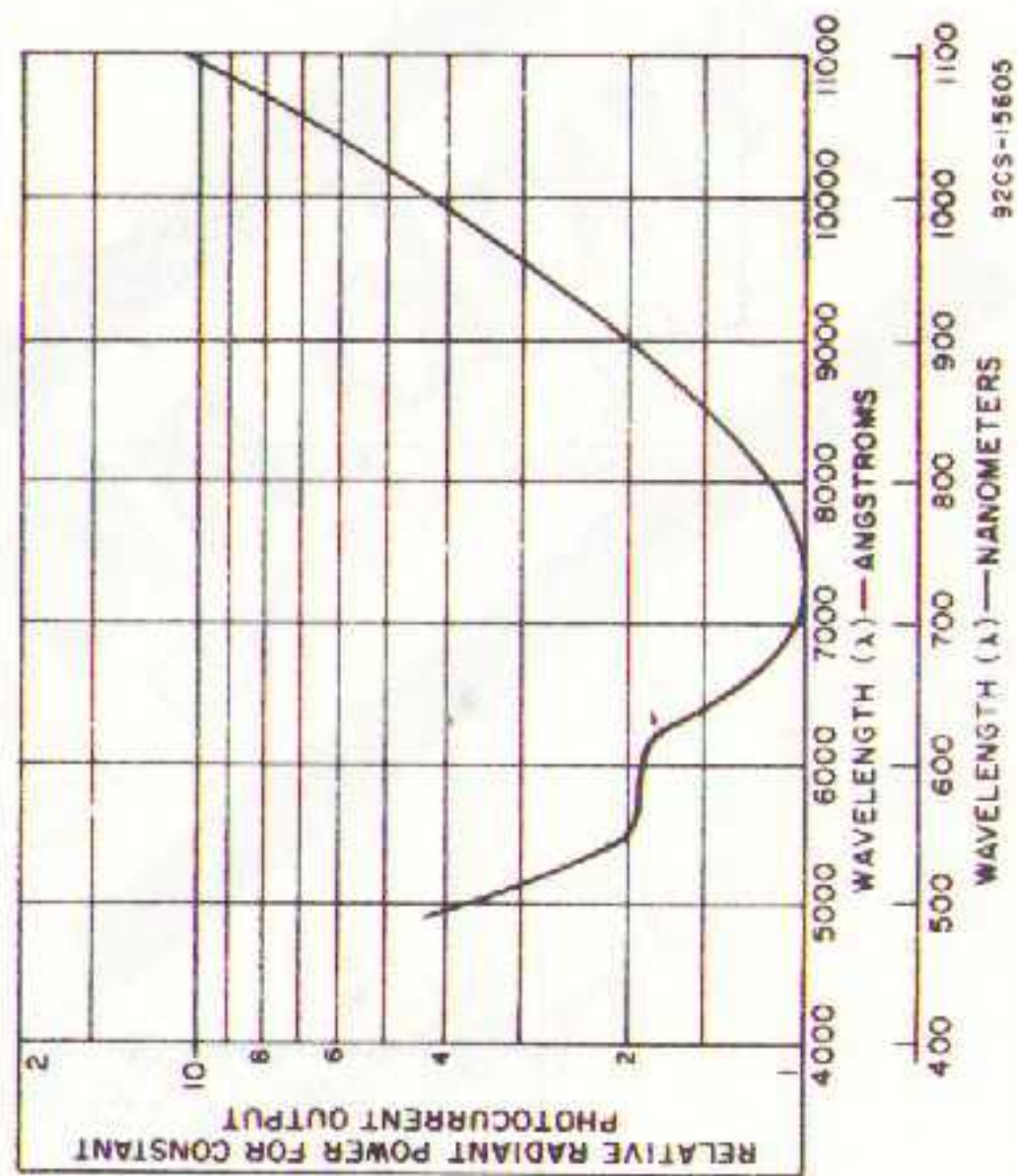


Fig. 5 - Typical spectral response of photosensitive Darlington unit.

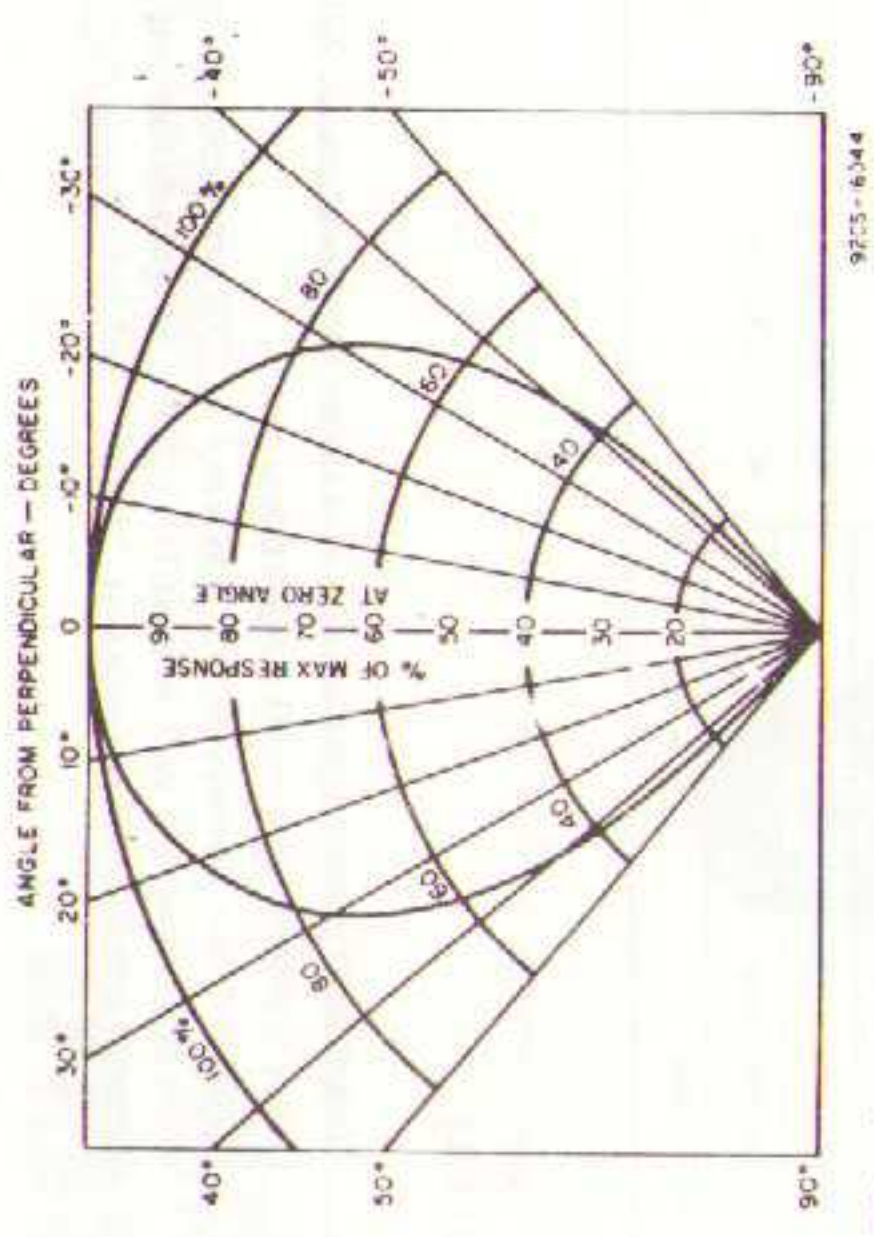


Fig. 6 - Relative angular sensitivity.

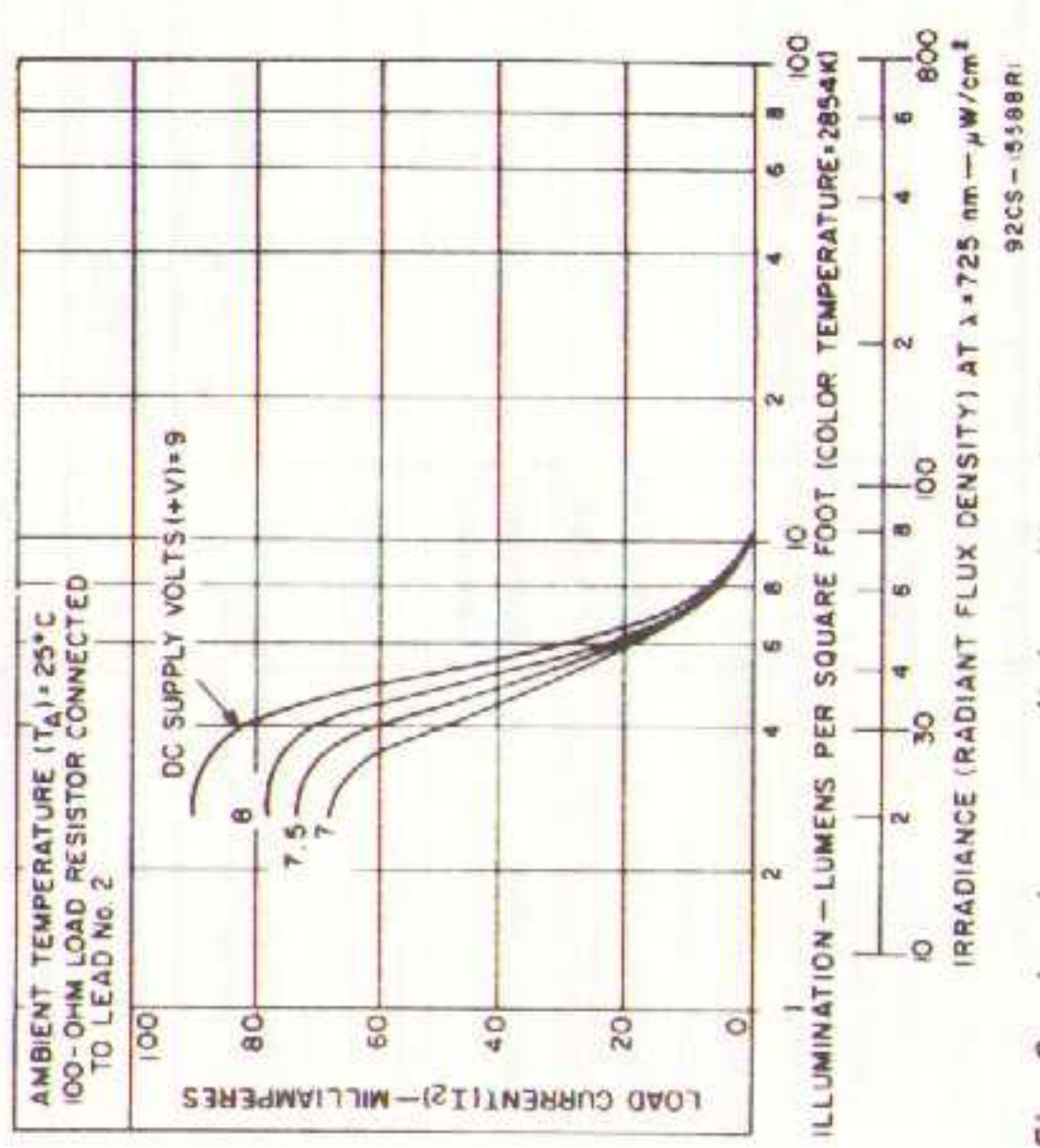


Fig. 8 - Load current (I_2) vs. illumination as a function of supply volts.

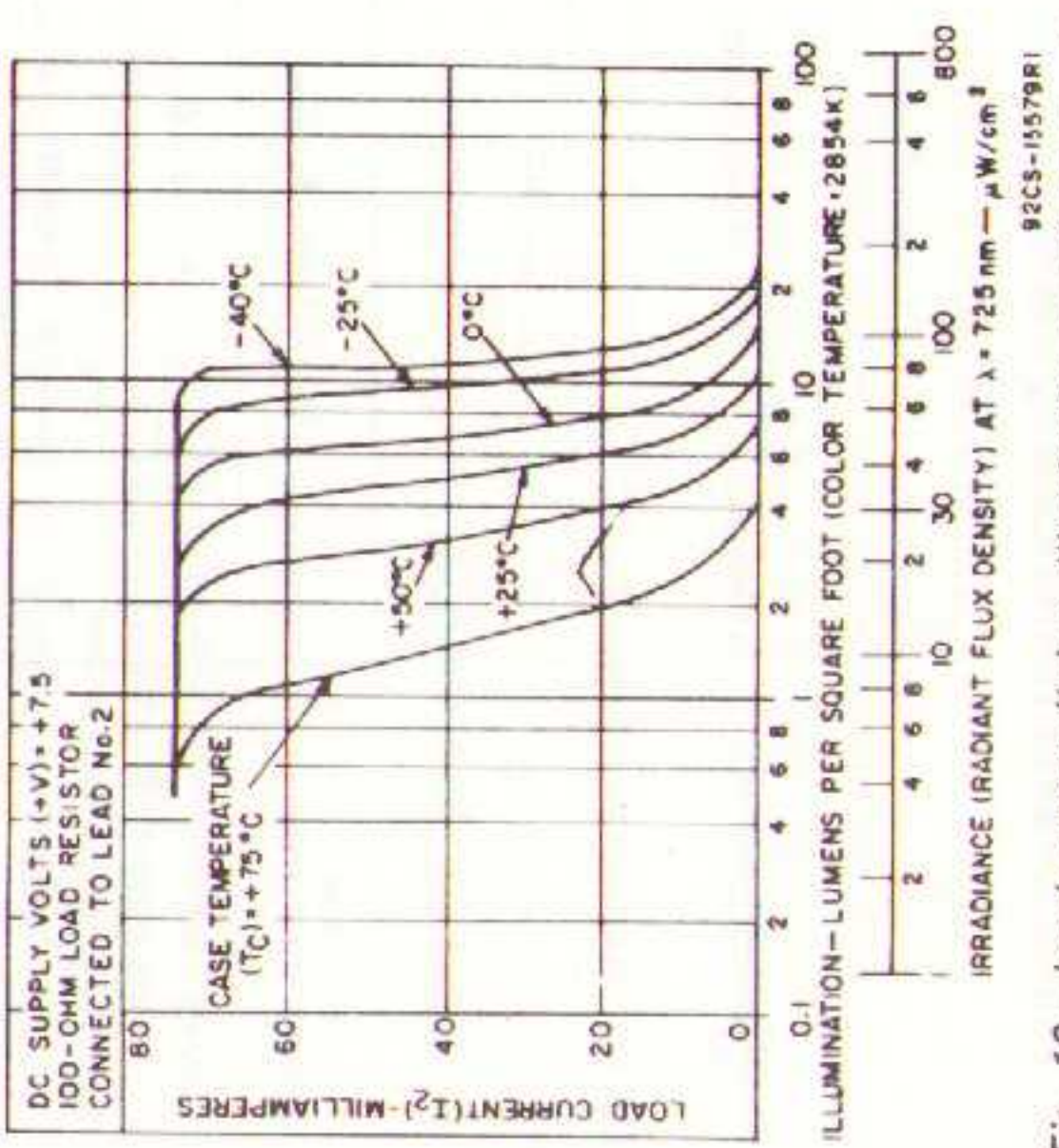


Fig. 10 - Load current (I_2) vs. illumination as a function of case temperature.

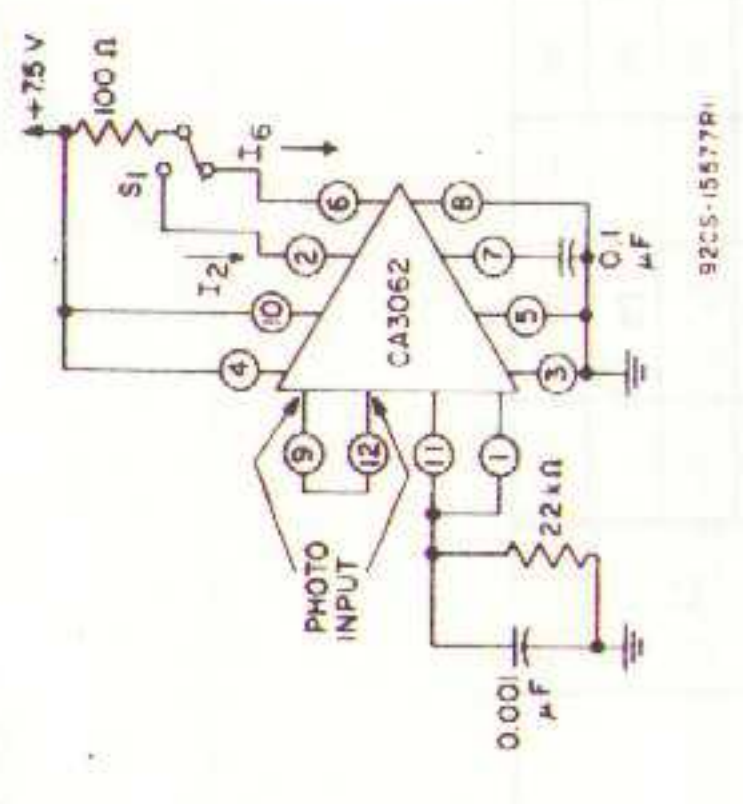


Fig. 7 - Test circuit for sensitivity and dc current measurement.

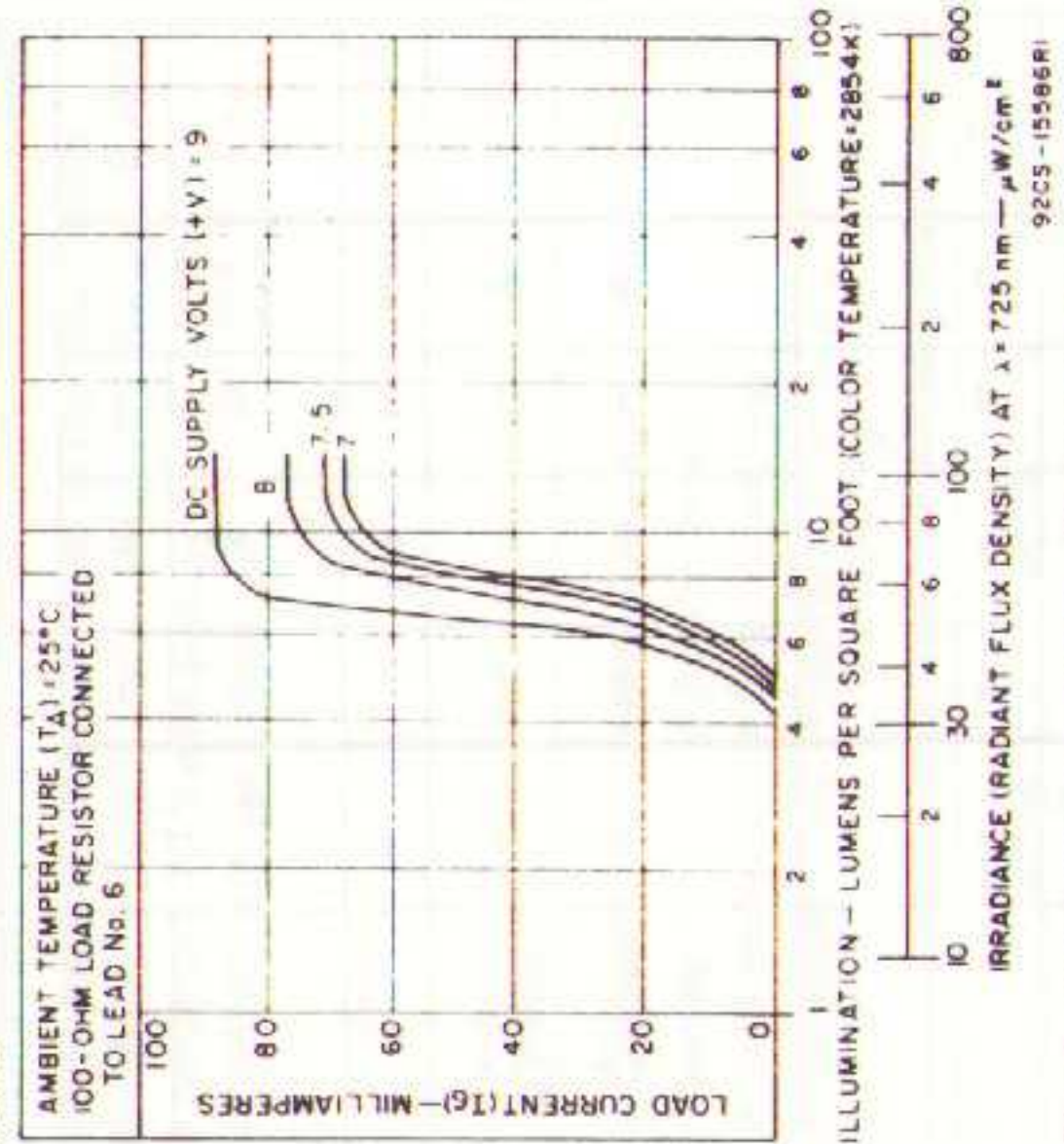


Fig. 9 - Load current (I_6) vs. illumination as a function of supply volts.

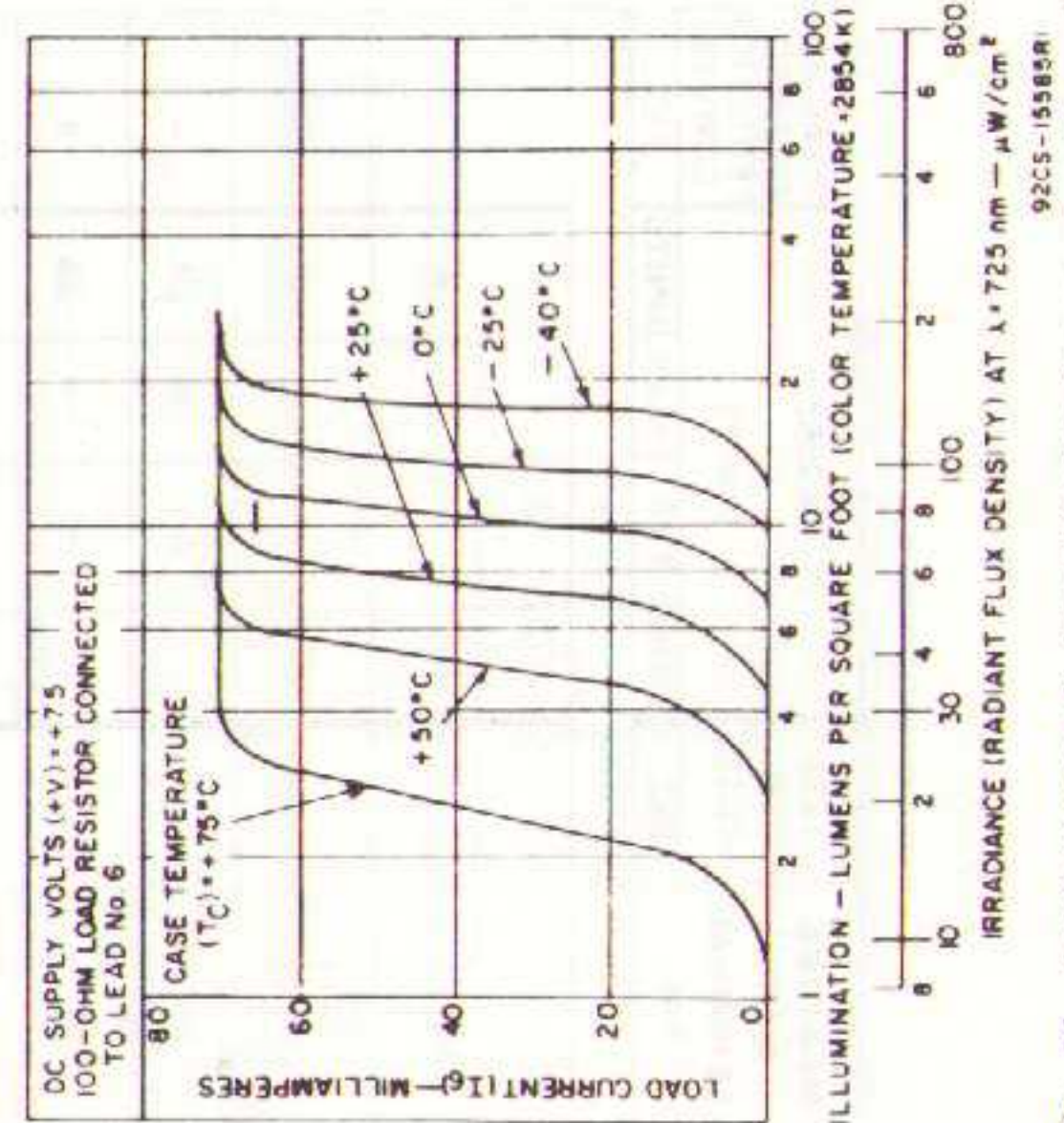


Fig. 11 - Load current (I_6) vs. illumination as a function of case temperature.

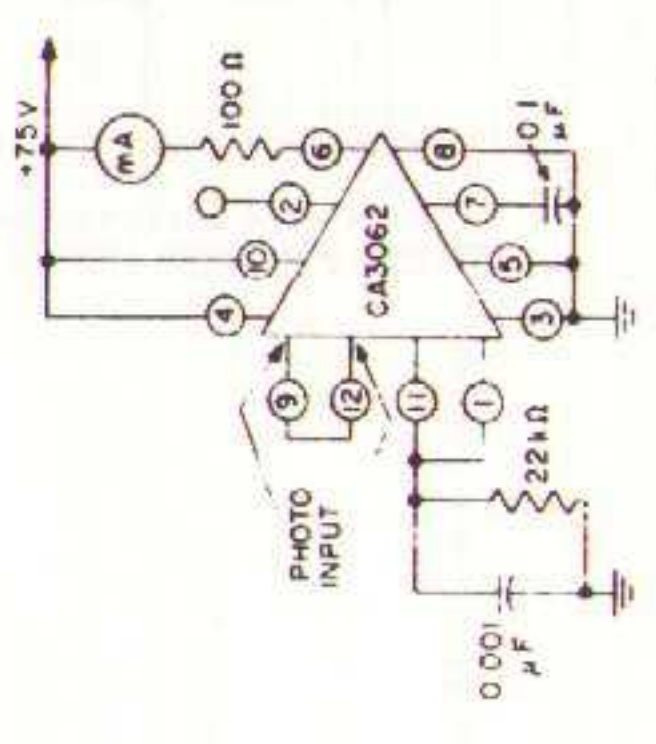


Fig. 12 - Response time test circuit.

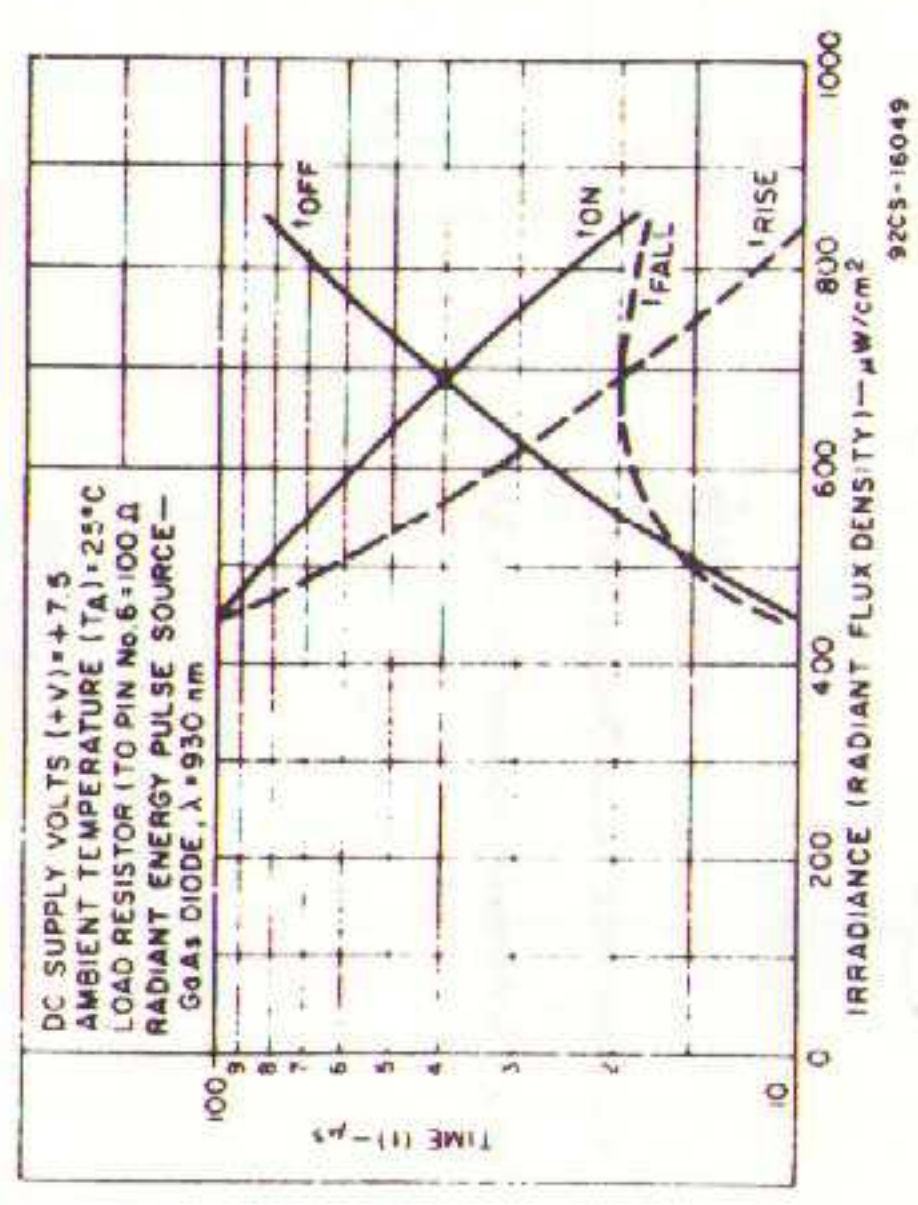


Fig. 14 - Response time as a function of radiant flux density.

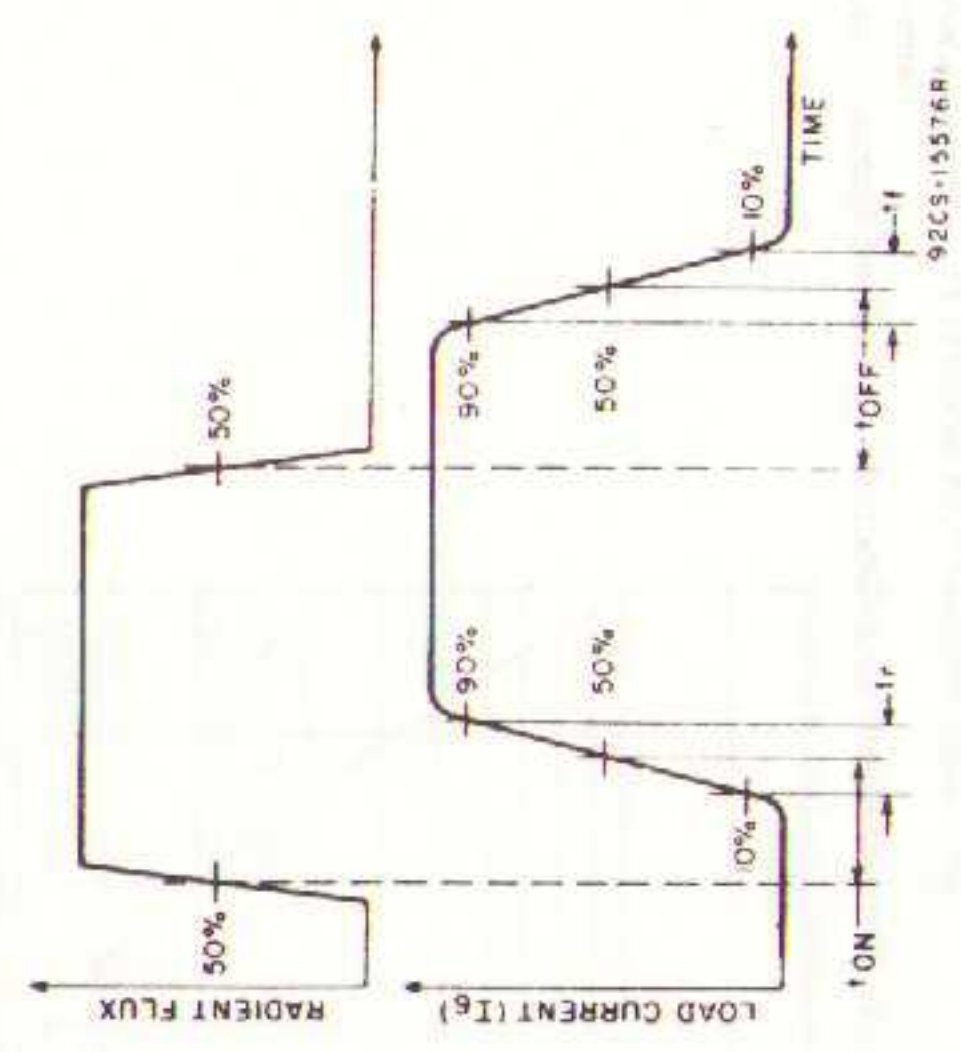


Fig. 13 - Waveforms for measurement of response time.

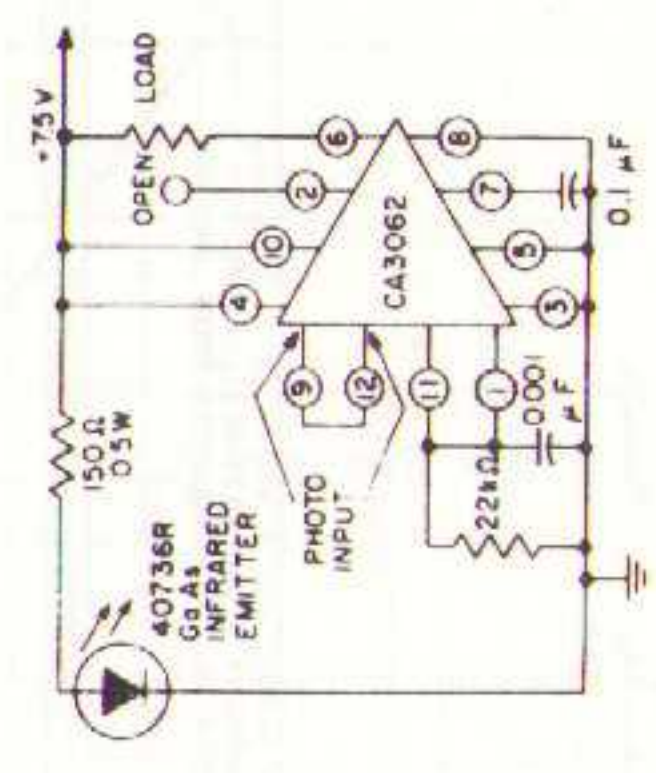


Fig. 15 - Circuit diagram for "ON-OFF" photoelectric control applications.

OPERATING CONSIDERATIONS

Switching Service

The CA3062 is primarily intended to provide "ON-OFF" output in response to a light signal. Optimum performance of this device is achieved when the output transistors are operated at values of load current sufficient to saturate the device in the "ON" state. Operation of the CA3062 at values of load current between the condition of no load current and saturation will cause substantial power to be dissipated in the silicon chip. This condition of operation is therefore not recommended because the heat rise in the silicon chip induced by the increased power dissipation causes the load current to shift in the same direction as though additional illumination were applied to the CA3062, a condition which will substantially alter the switching characteristics of the device.

The signal voltages at the input terminals (terminal No. 1 and No. 7) must not exceed 3 volts, because any increase in the signal voltage beyond the value specified will cause both output transistors to be turned "ON". In the circuit shown in Fig. 7, this condition will occur for values of illumination greater than 60 lumens/ft². This adverse operating condition can be avoided by either limiting the maximum illumination or by clamping the input so that the voltage does not exceed 3 volts.

Linear Service

The CA3062 can be connected as shown in Fig. 16 to give a linear output. The value of the load resistor should be greater

than 1000 ohms in order to limit the power dissipation and thus minimize the heating effects. Because of the many possible variations in circuit configurations, the CA3062 has not been characterized for linear service applications. A guide-line circuit for this class of service is shown in Fig. 16.

Specific inquiries for use of the CA3062 in this type of service should be addressed to your local RCA Field Technical Representative.

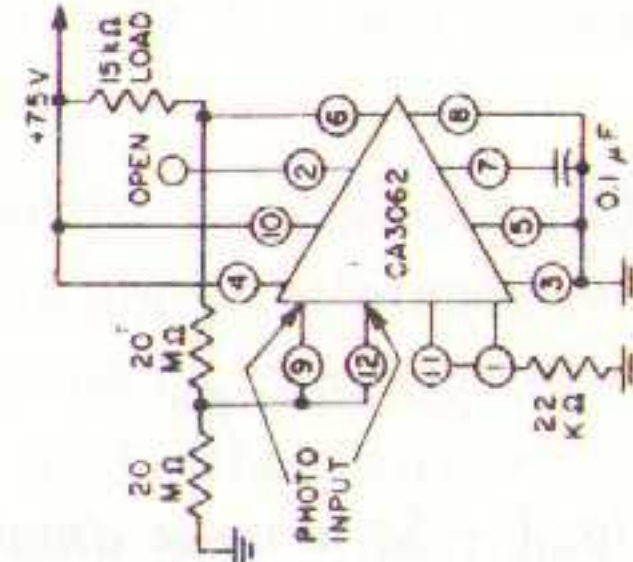
Precautions

Because of the high amplification of the CA3062, care should be taken, when wiring, to keep all lead lengths as short as possible. A recommended breadboard layout is shown in Fig. 17.

If the CA3062 is operated with an inductive load impedance, such as a relay, it is recommended that a diode be connected across the load to absorb the energy of the pulse voltages generated during switching.

Many of the graphs are shown with two sets of abscissa values for light energy input, one expressed in illumination values (lumens/sq. ft.) and the other in irradiance values ($\mu\text{W}/\text{sq. cm.}$)

Correlation between these two sets of abscissa values is accomplished by having the light source operating at the maximum sensitivity wavelength of the CA3062. See Notes on page three.



92CS-5583R1

Fig. 16 - Circuit diagram for linear output photoelectric applications.

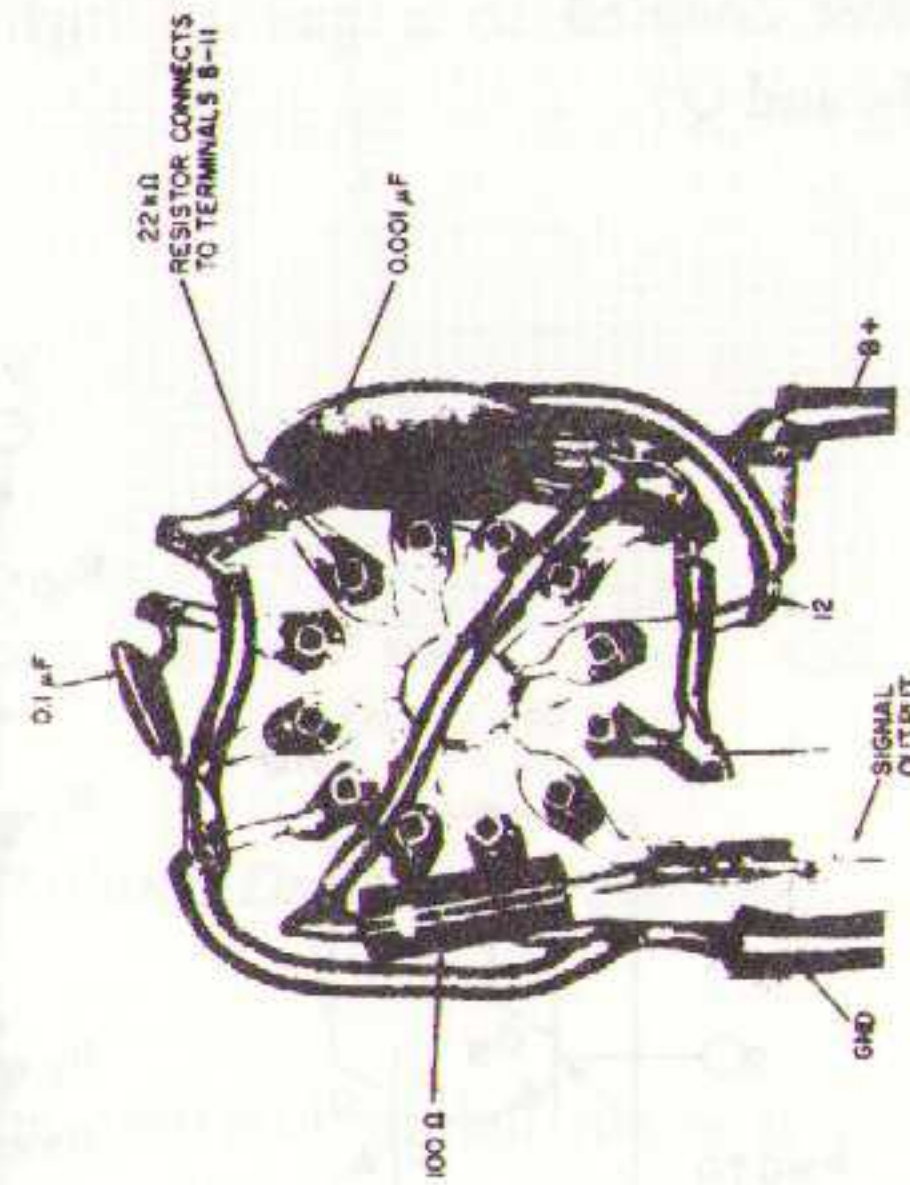
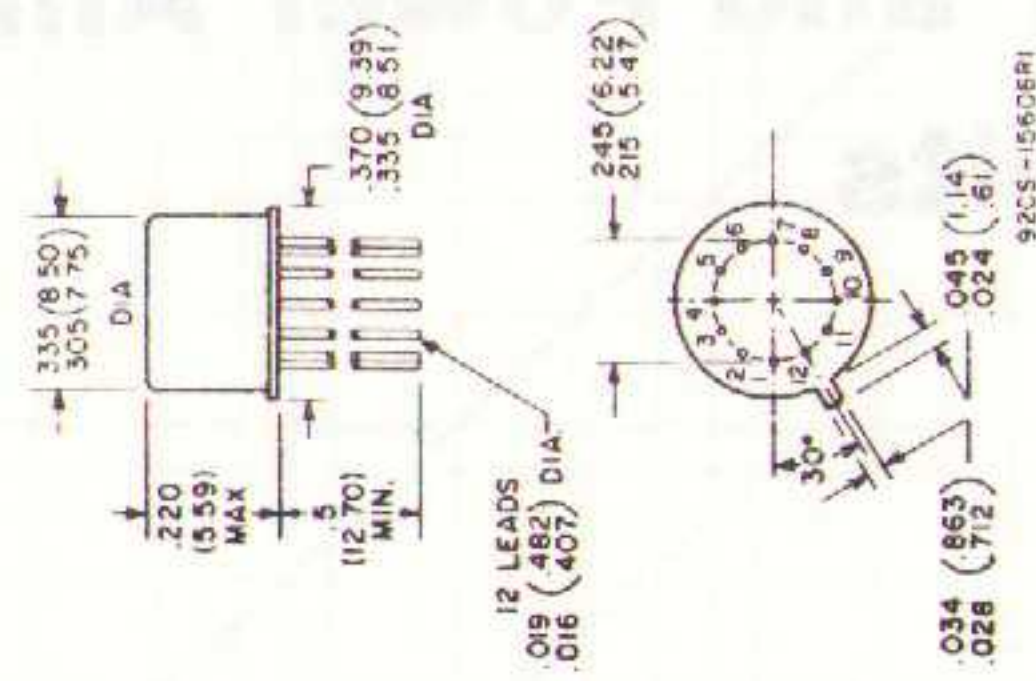


Fig. 17 - Breadboard layout of test circuit, shown in Fig. 7 for the CA3062.

DIMENSIONAL OUTLINE



Dimensions in Inches and Millimeters

Dimensions in parentheses are in millimeters and are derived from the basic inch dimensions as indicated.

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Applications of the RCA-CA3062 IC Photo-Detector and Power Amplifier in Switching Circuits

by J. D. Mazgy

The RCA CA3062 is a monolithic silicon integrated circuit consisting of a photosensitive detector and a switching amplifier with a pair of high current output transistors. This note describes how the CA3062 with only 3 resistors can provide a light activated switch which will drive a variety of practical loads; such as solenoids, relays, triacs, SCR's, etc. "Normally ON" and "Normally OFF" outputs are available simultaneously.

Circuit Description

The circuit diagram and terminal connections for the CA3062 are shown in Fig. 1. The circuit consists of a photo-Darlington pair and a differential amplifier which is emitter-follower coupled to a pair of high-current output transistors, Q₆ and Q₇.

Circuit Operation

The CA3062 is designed for operation from power supply voltages of 5 to 15 volts between terminal Nos. 4 and 8, and voltages as high as 30 volts V+ on the output transistors.

The photo-detector system consists of four silicon transistors Q₁ and Q₉, with Q₁₀ and Q₁₁ in a parallel-connected Darlington circuit. The Darlington configuration is used to provide maximum photo current from the available detector area. The area of each photo-transistor is $1.3 \times 10^{-4} \text{ cm}^2$. However, the effective photo-sensitive area is $2.6 \times 10^{-4} \text{ cm}^2$ because transistors Q₁ and Q₁₁ of each Darlington contribute a relatively small percentage of the total photo-current.

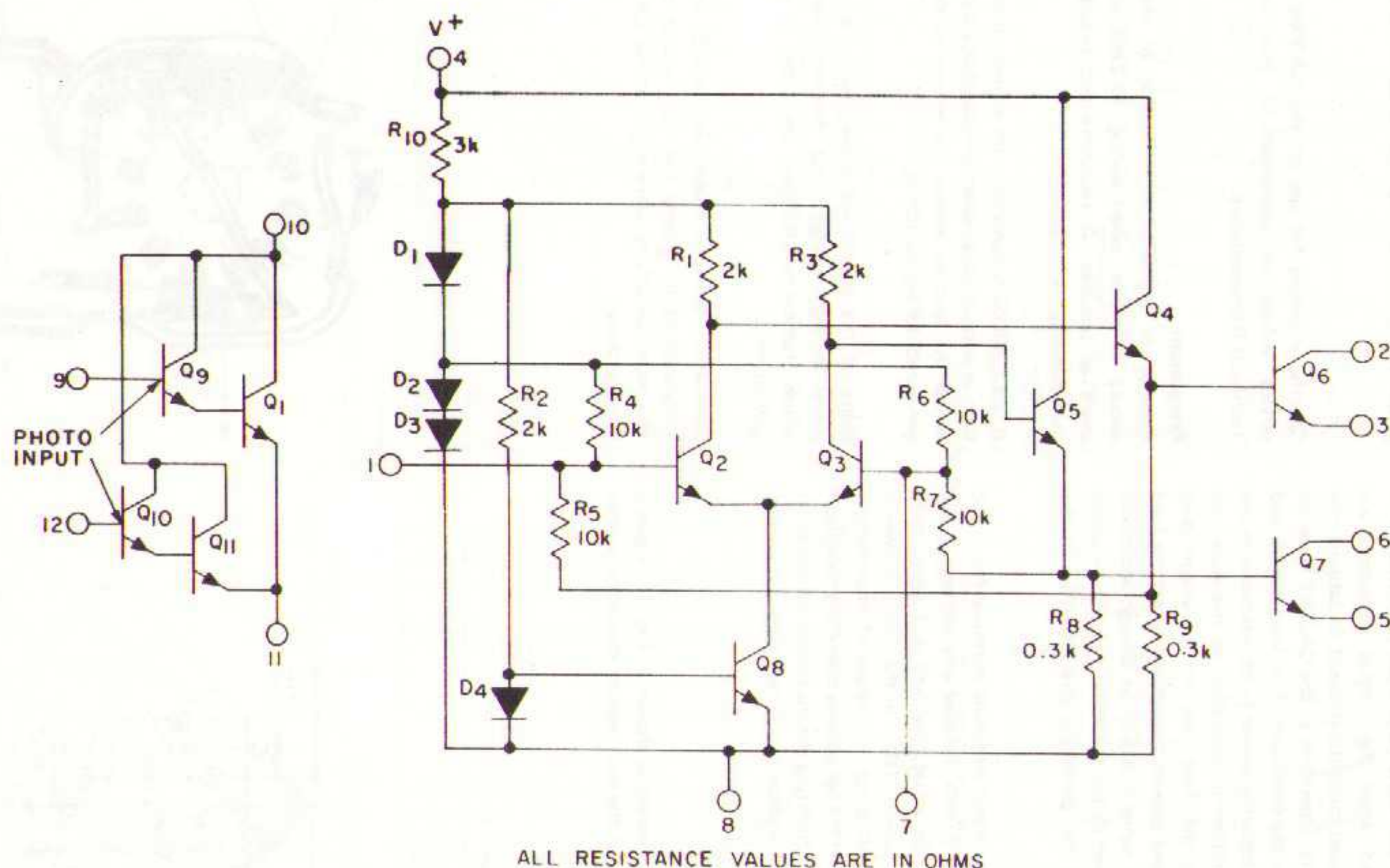


Fig. 1 — Schematic diagram of CA3062.

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Fig. 2 shows a typical curve of photo-current in the collector and base as a function of light intensity. Fig. 3a & b shows the test set up used to obtain the data for Fig. 2. A typical spectral response curve of the photo sensitive Darlington is shown in Fig. 4. Fig. 5 shows typical rise and fall times for the photo detector output.

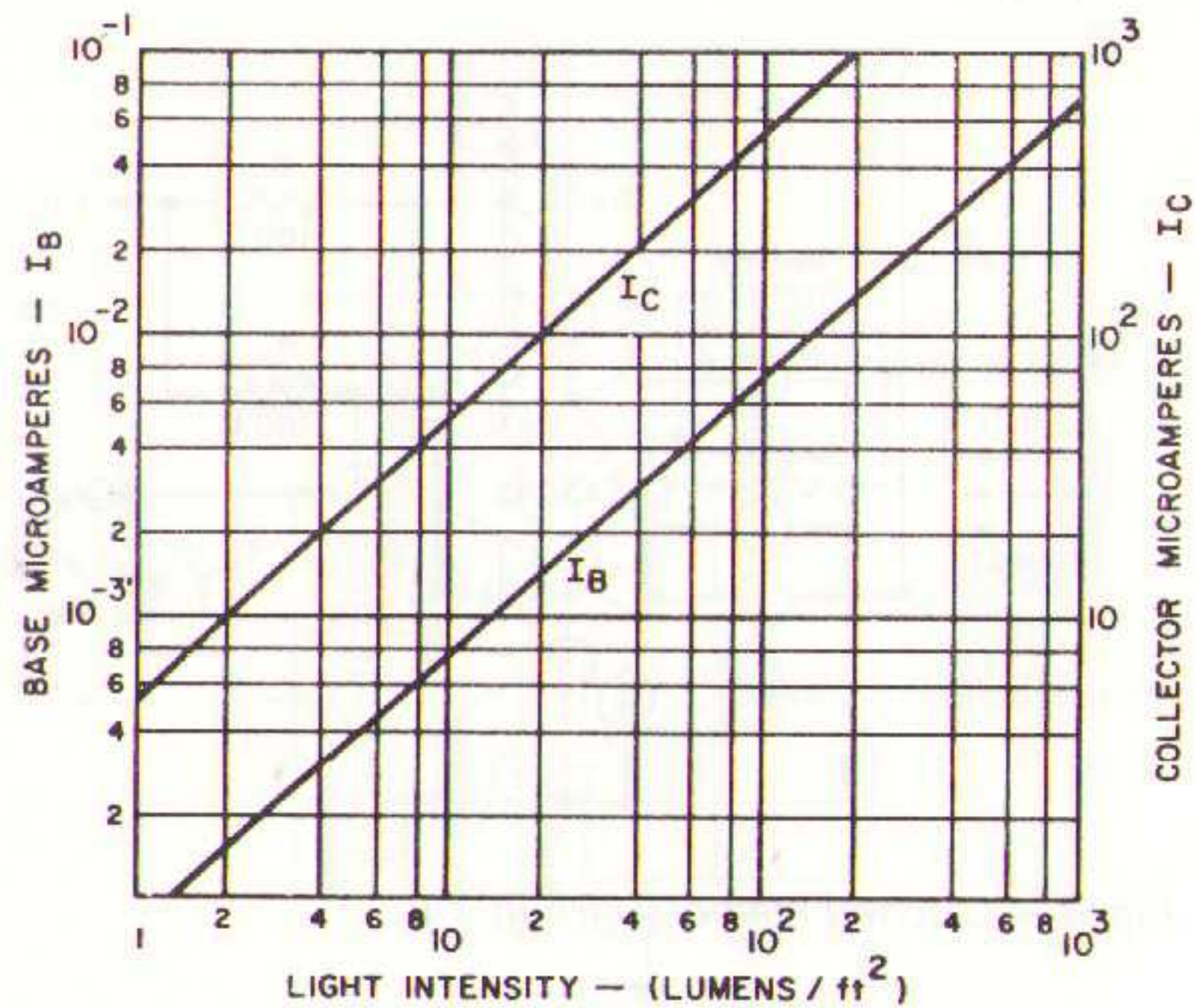


Fig. 2 – Typical I_C, I_B as a function of light intensity at 25°C.

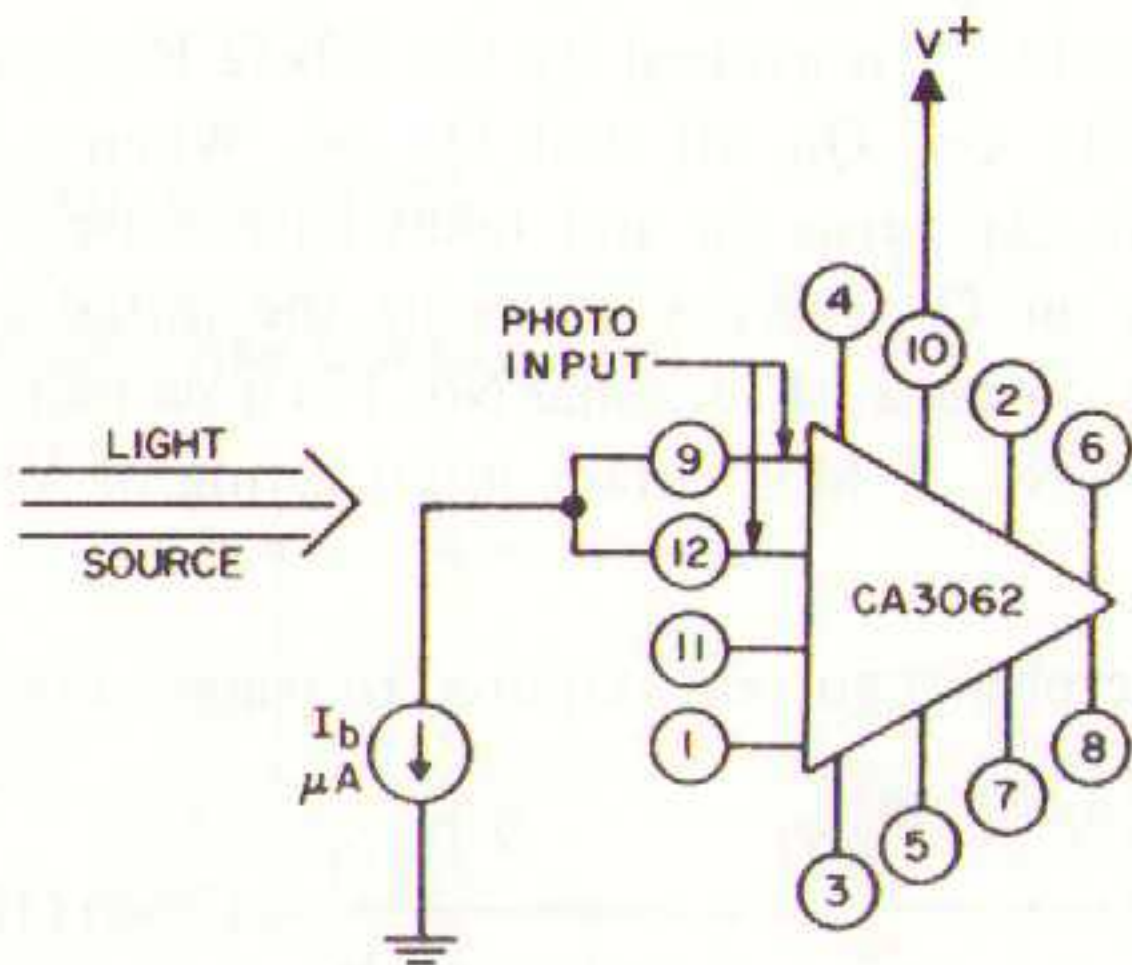


Fig. 3a – Base current test set-up.

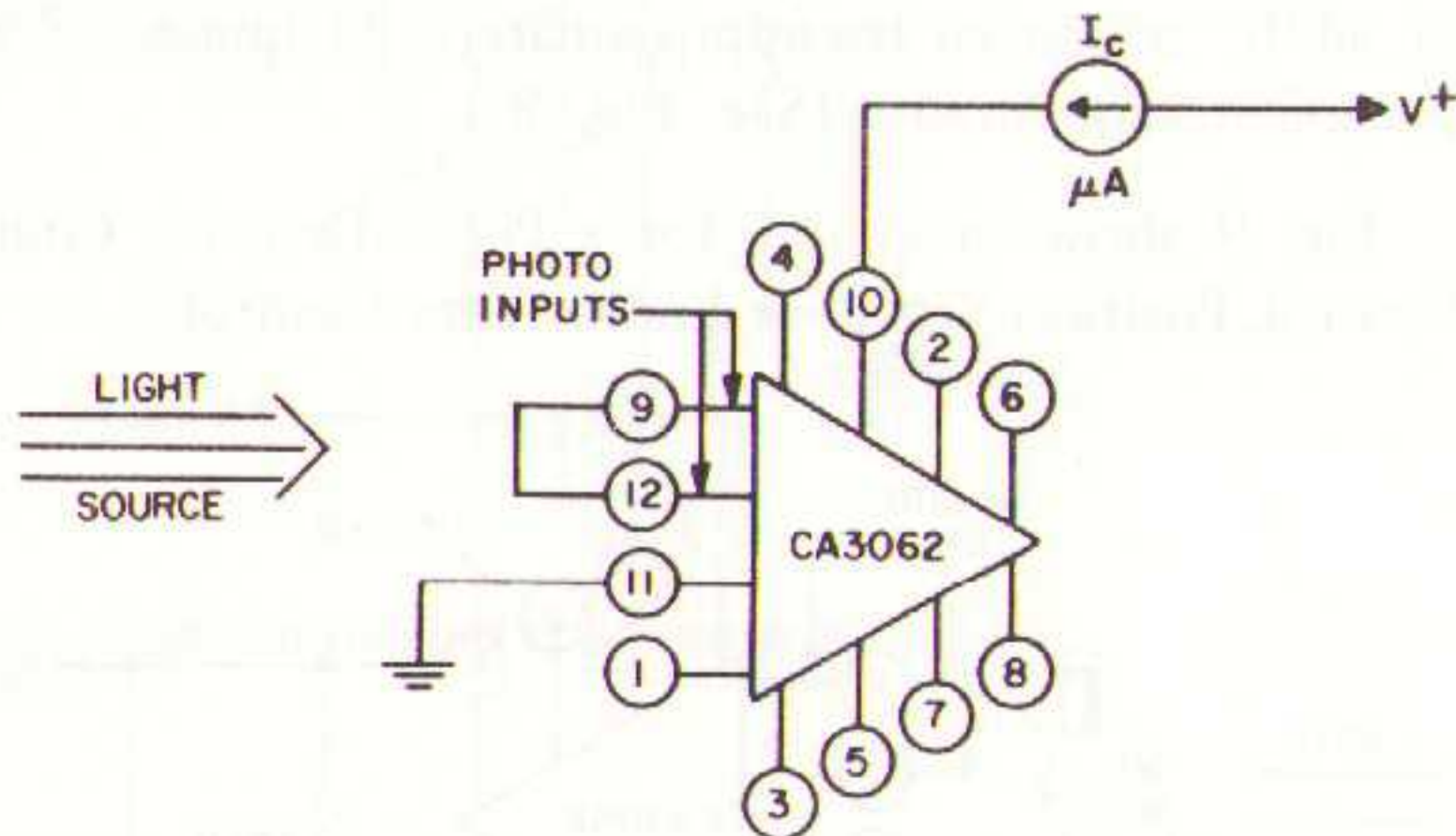


Fig. 3b – Collector current test set-up.

Fig. 3 – Test circuits.

The photo Darlington pair can be either emitter-coupled or collector-coupled to the differential amplifier consisting of Q₂ and Q₃ and constant current sink Q₈, operating with a total current of 0.7mA as shown in Figs. 6a and 6b,

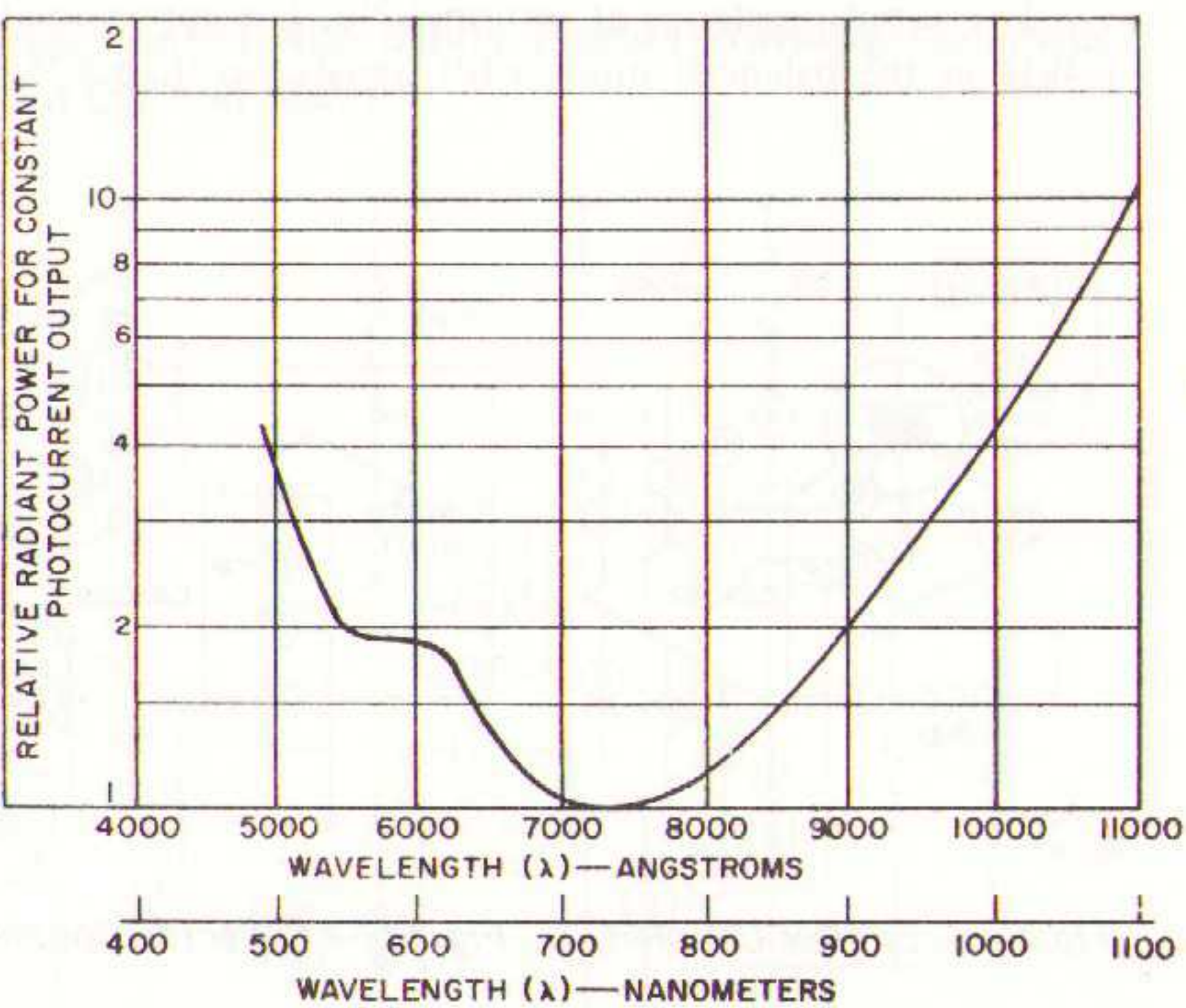


Fig. 4 – Typical spectral response of photosensitive Darlington unit.

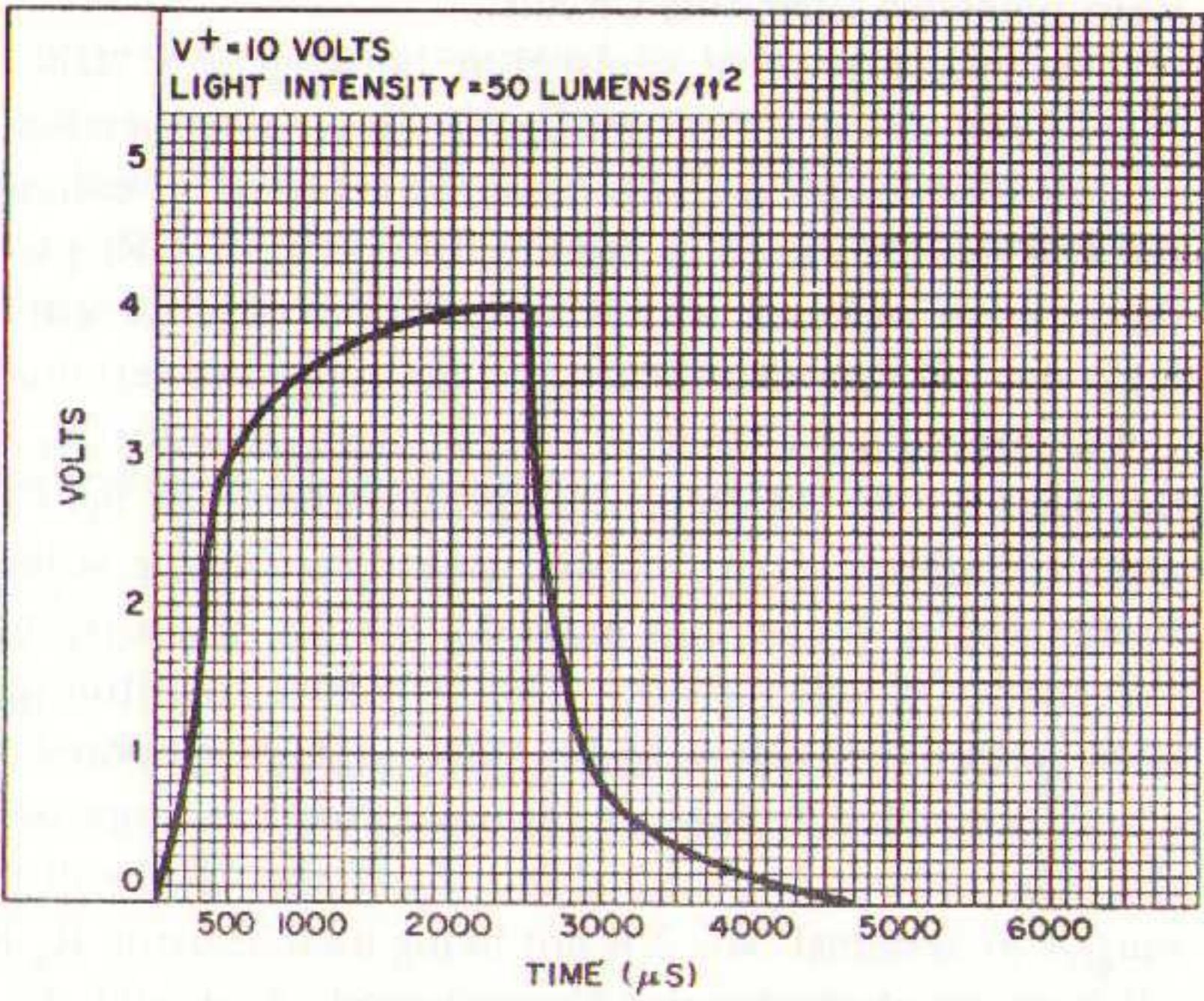


Fig. 5 – Typical photo-Darlington response.

respectively. The emitter-coupled mode is generally recommended. However, there are applications in which the collector-coupled mode is advantageous. Examples of the collector-coupled mode are shown in Figs. 8, 12 & 13.

In a balanced condition Q₂ and Q₃ are conducting 0.35mA each, which sets the dc voltage of each of the collectors of Q₂ and Q₃ at 0.7 volt below the three diode drops of D₁, D₂ and D₃.

With the dc potential on collectors Q₂ and Q₃ determined to be +1.4 volts above reference terminal No. 8, the voltage on the emitters of Q₄ and Q₅ then is 0.7 volt below the respective base, and, therefore, +0.7 volt above the reference point. Thus, the base potentials of Q₂ and Q₃ are set to +1.05 volts and +0.7 volt for Q₆ and Q₇, respectively. The emitter currents of Q₄ and Q₅ are set to approximately 2.33mA each under the balanced condition.

The input resistance at terminal No. 1 is approximately $1.4k\Omega$ in the balanced mode with signals less than $\pm 25mV$.

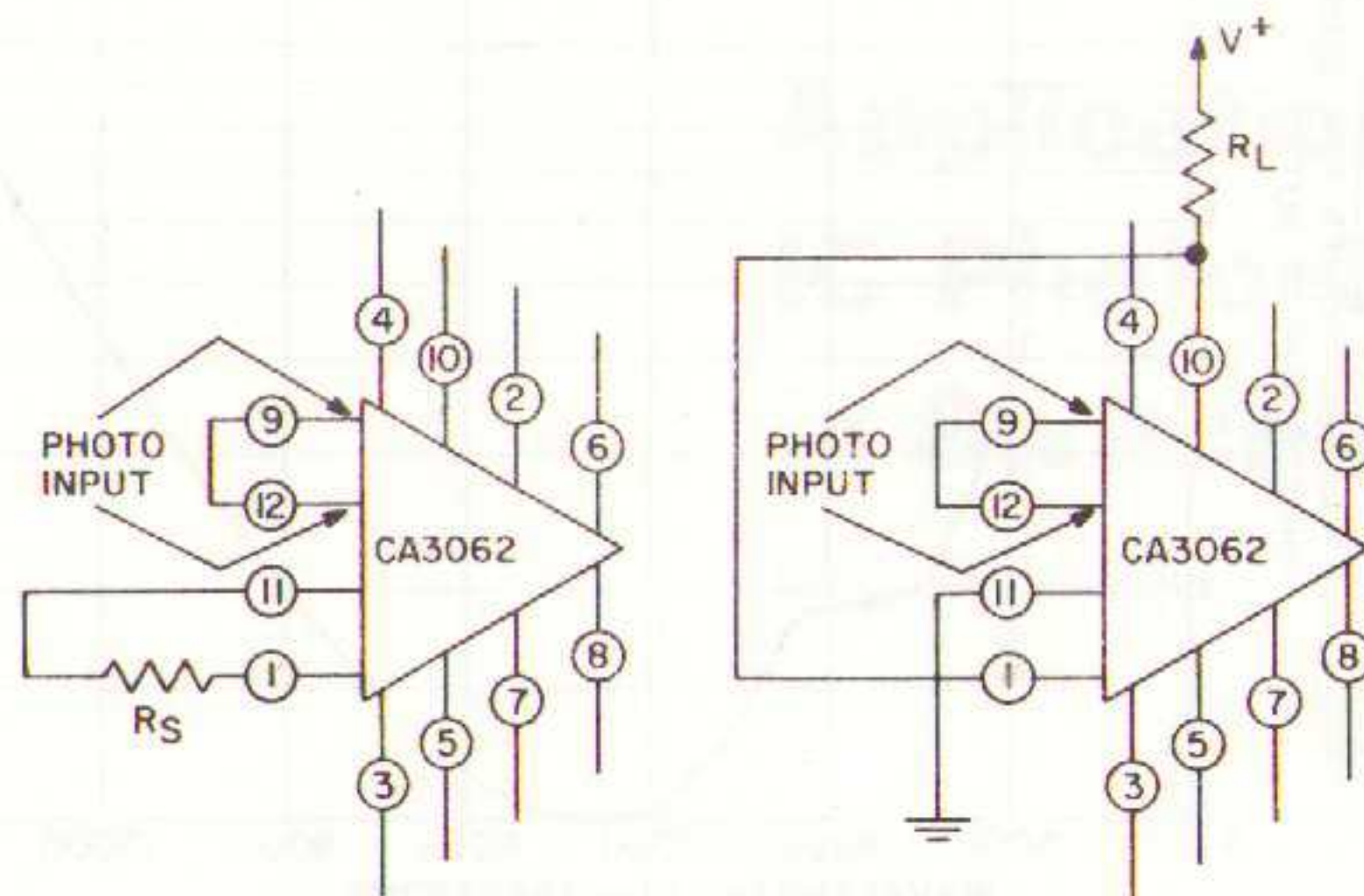


Fig. 6a - Emitter Coupled. Fig. 6b - Collector Coupled.

Fig. 6 - Methods of coupling photo-detector to amplifier portion of CA3062.

Recommended Operating Circuit:

To assure positive transition between the "ON" and "OFF" states, Schmitt trigger operation is recommended.

Schmitt trigger operation can be achieved by connecting the CA3062 as shown in the circuit of Fig. 7. R_f provides the positive feedback which causes Q_3 to conduct and holds Q_7 in cut-off. A positive going voltage applied to terminal No. 1 will result in a change of output state. Resistor R_S limits the drive to the differential amplifier when high light levels are encountered. R_S is chosen to insure that the voltage at terminal No. 1 does not exceed 1.9 volts. If this voltage is exceeded, Q_6 will turn on. This overdrive condition causes both output transistors to be "ON". (Q_6 is supposed to be cutoff when the voltage at terminal No. 1 is more positive than the voltage at terminal No. 7). If the "Normally ON" output at terminal No. 2 is not being used, resistor R_S is not required, and terminal Nos. 2 and 3 should be left unconnected. (See Fig. 7.) The magnitude of the threshold voltage and the amount of hysteresis provided are determined by the value of the feedback resistor R_f . (See appendix for calculations.)

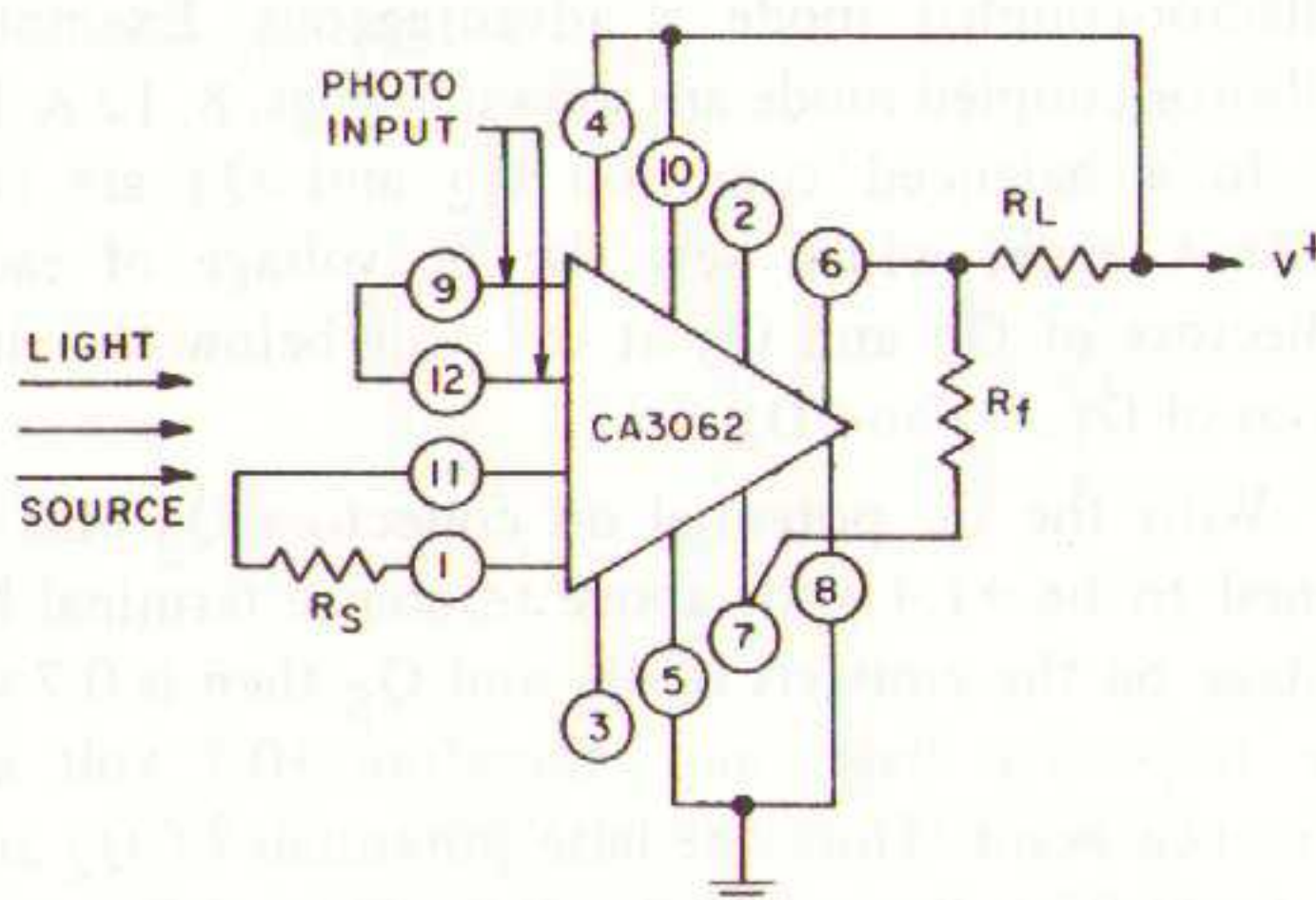


Fig. 7 - Schmitt-circuit.

Typical Applications

Latched Memory Circuit

A latched memory system can be used to stop clocks, record an intrusion, or activate light-actuated dark-room controls.

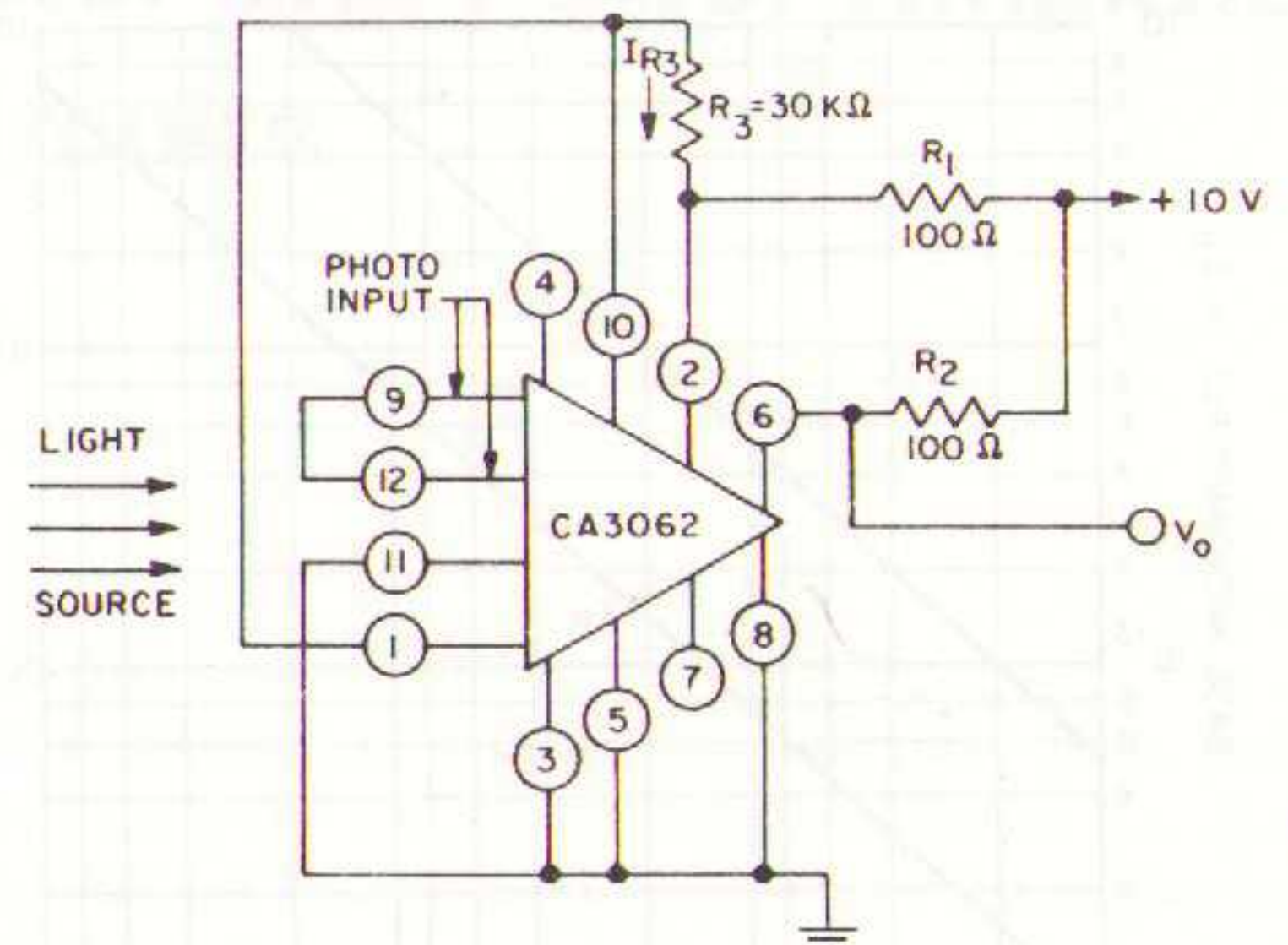


Fig. 8 - Latched memory circuit

Circuit Operation of Latched Memory

The initial conditions are: terminal No. 2 at "high"-output-voltage and terminal No. 6 at "low"-output-voltage. This condition is assured by the $30k\Omega$ Resistor (R_3), which biases Q_2 on, Q_6 off and Q_7 on. When a light pulse is received, Q_1 turns on and takes base drive away from Q_2 , turning on Q_6 thereby reversing the initial conditions. Q_6 remains on because terminal No. 1 is now more negative than terminal No. 7. Momentary interrupting of V^+ will reset the circuit.

The photo current required to trigger a typical circuit is:

$$I_{R_3} = \frac{V^+ - V_{B21}}{R_3 + R_{B2}} = \frac{9}{(35)(10^{-3})} = (256)(10^{-6}) \text{ ampere}$$

or an I_C produced by approximately 25 lumen/ft^2 at the photo-detector input. (See Fig. 3.)

Fig. 9 shows a circuit for a Photo-Detector Counting Control, Position Sensor or End-of-Tape Control.

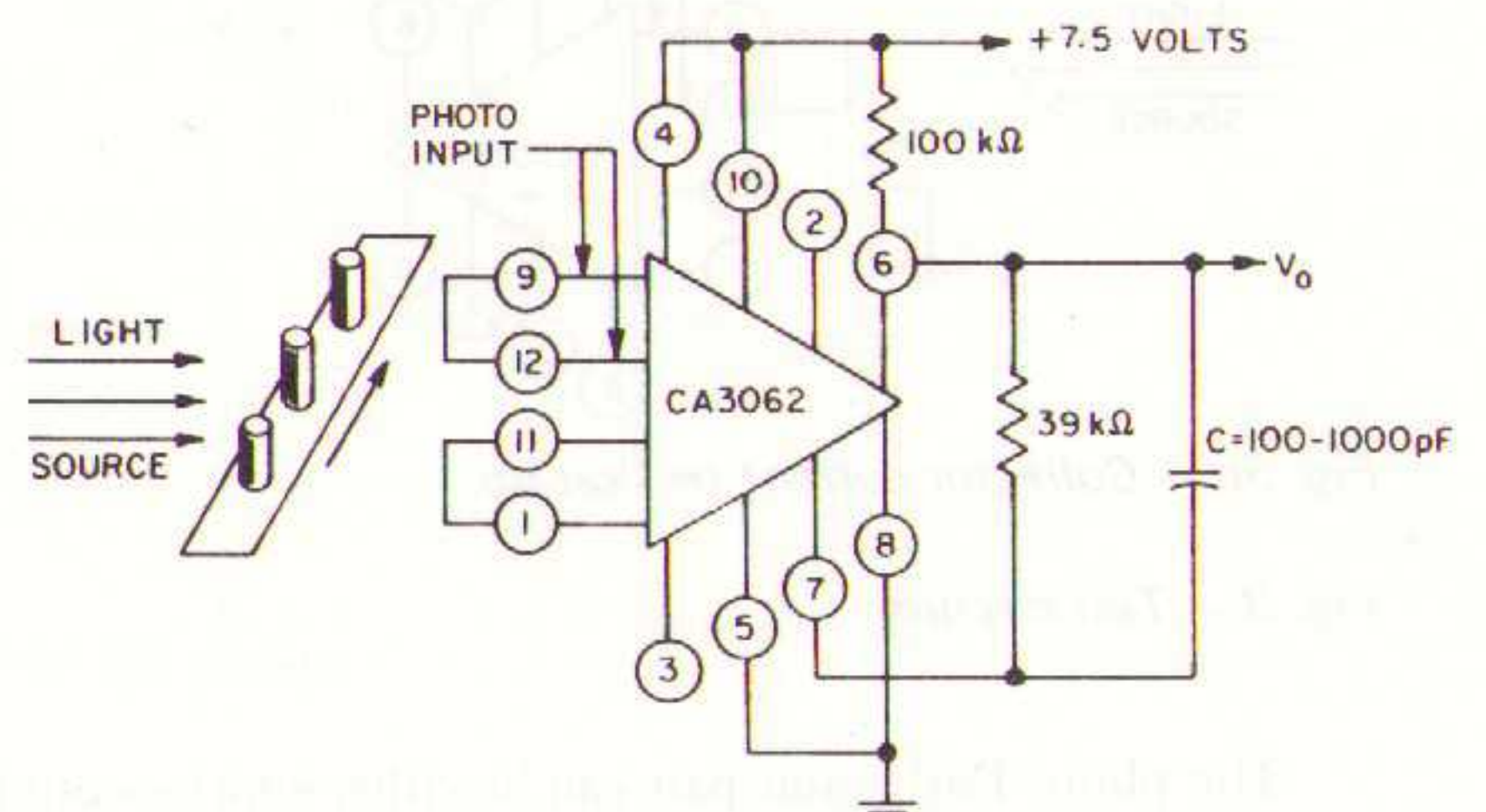


Fig. 9 - Circuit for slow speed counting level control, position sensor and end-of-tape control.

An Optically Coupled Isolator Circuit, used to transfer signals that are at substantially different voltage levels, is shown in Fig. 10. Both polarity outputs are available. Current transfer ratios of as high as 10:1 can be achieved with this circuit. The design equations for this system are the same as those presented in the appendix.

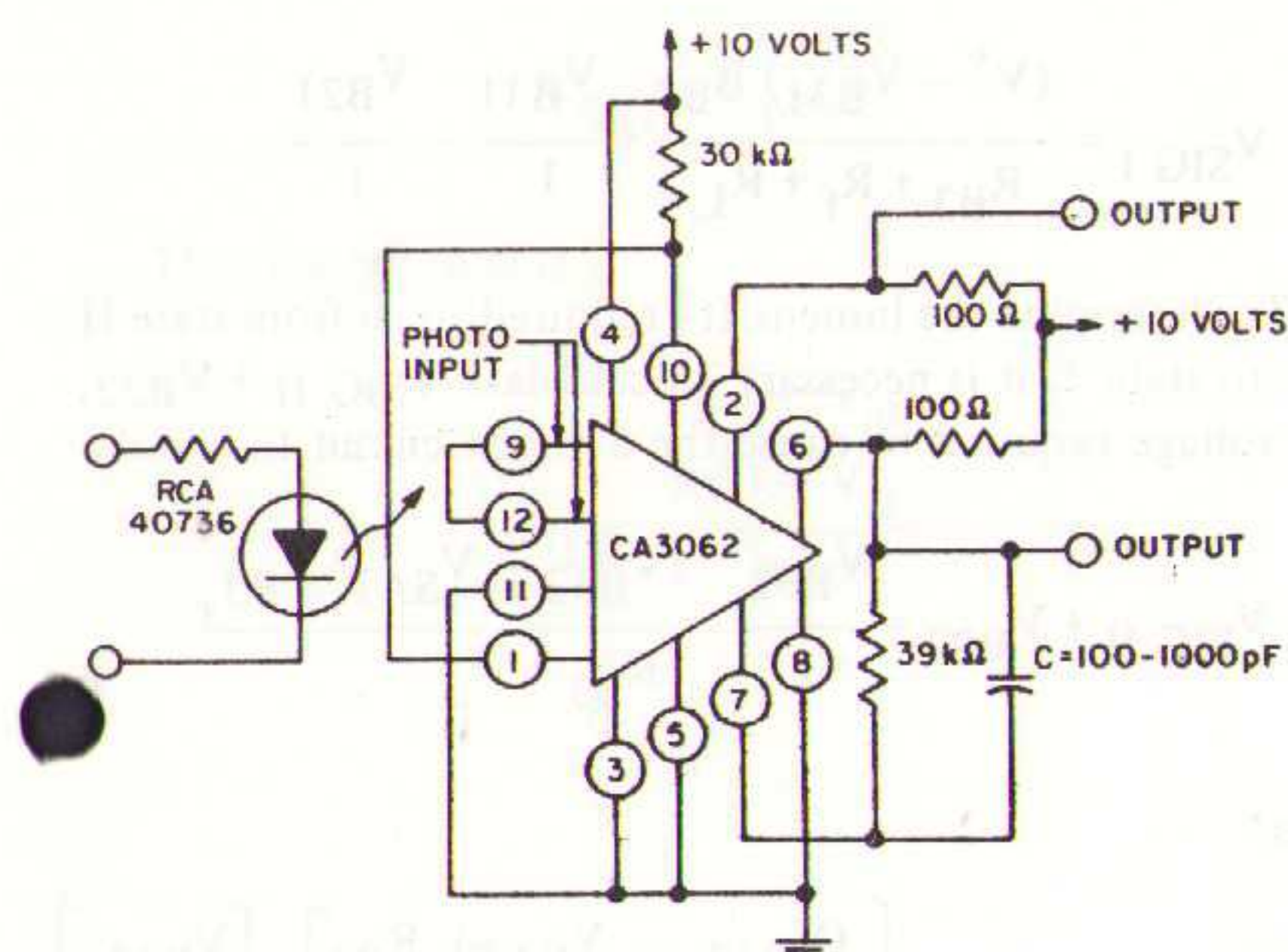


Fig. 10 – Isolator circuit.

TRIAC CONTROL SYSTEMS

A. Light Activated Triac Control

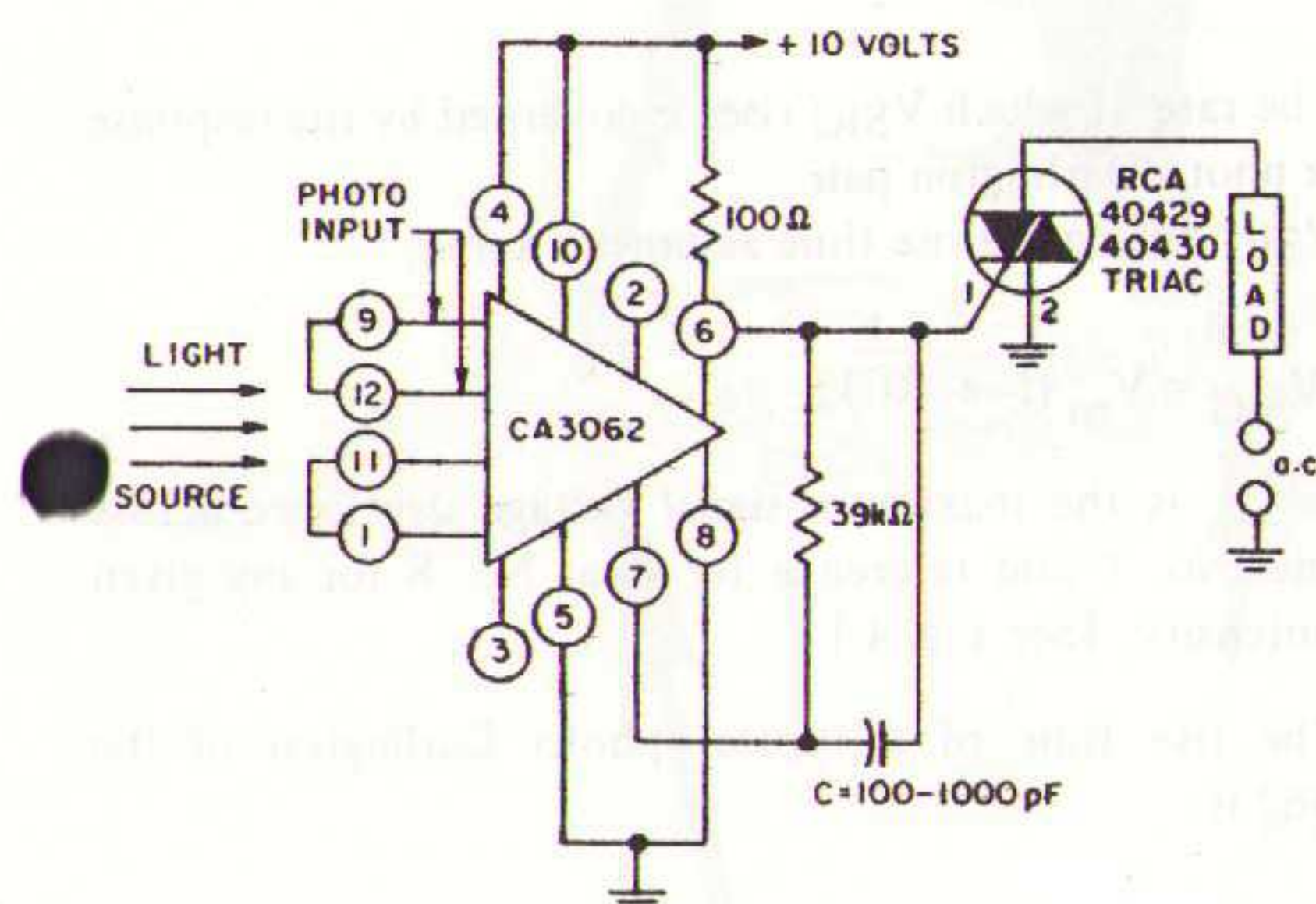


Fig. 11 – Light activated triac control.

B. Triac Control With Safety Feature Providing Automatic Shut Off And Alarm.

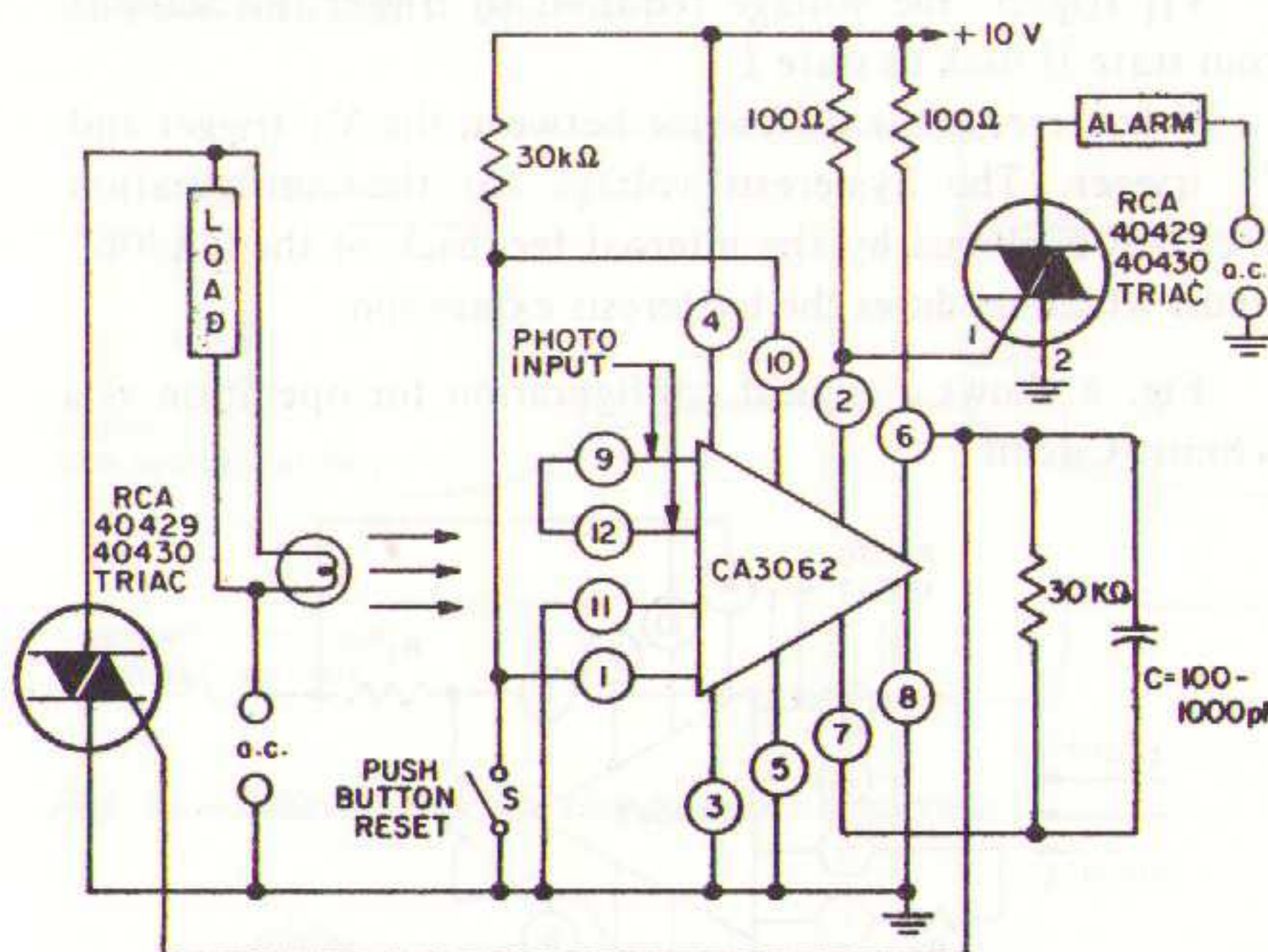


Fig. 12 – Triac automatic shut off and alarm.

In this system ac is supplied to the load as long as the light source is "on". If the light path to the CA3062 is broken, then the ac to the load and light source is opened, thereby activating the alarm circuit. The system can be reset with the push-button shown.

C. Triac Intrusion Alarm System.

If the light path is broken or the ac is interrupted, the alarm system will be activated, provided the battery is adequately charged.

The V+ acts as a charging circuit for the battery while the circuit is operating from the ac supply.

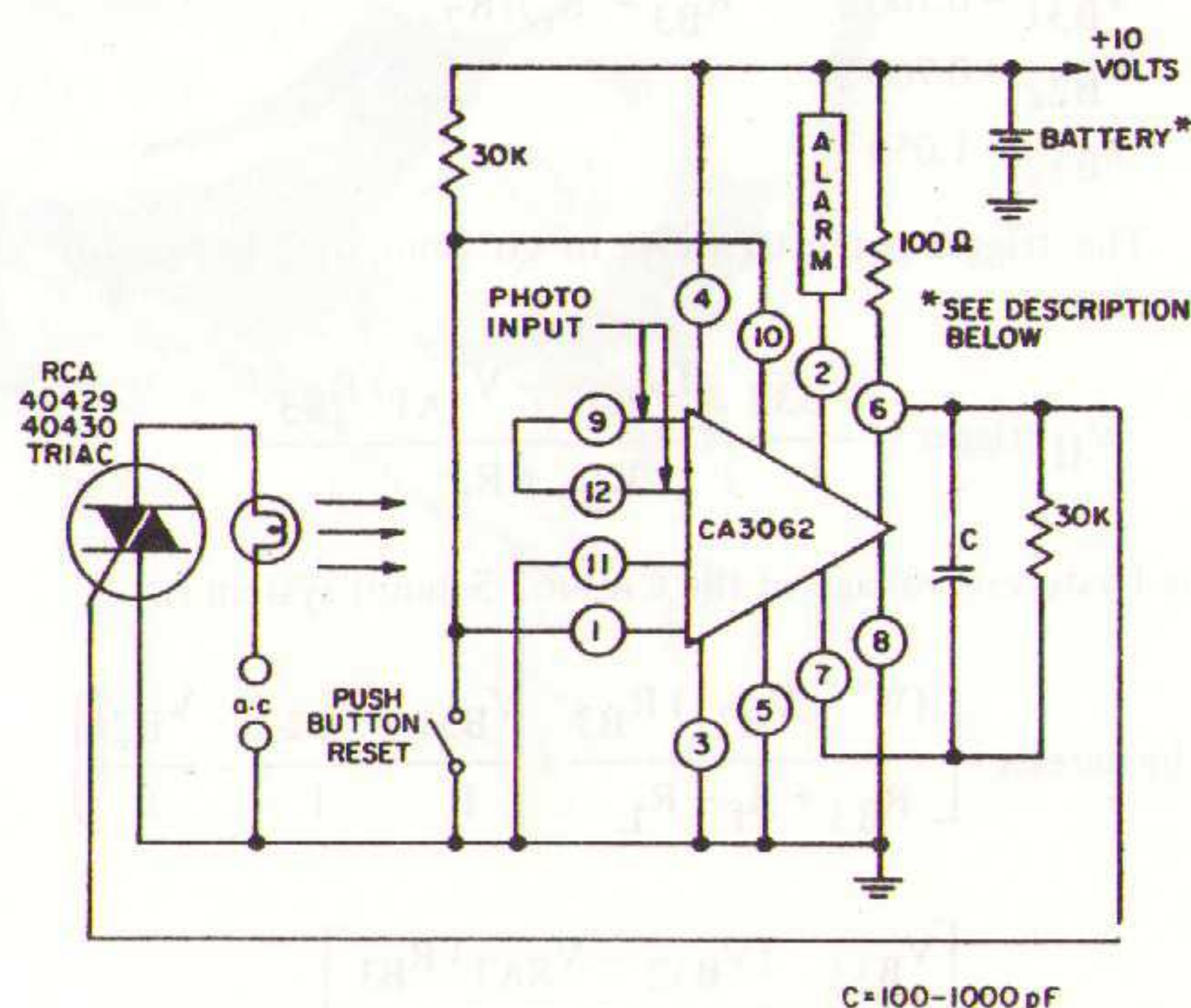


Fig. 13 – Triac intrusion alarm system.

APPENDIX

In general, Schmitt circuits have three voltages of interest.

V_I trigger: the voltage required to trigger the Schmitt from state I to state II.

V_{II} trigger: the voltage required to trigger the Schmitt from state II back to state I.

V hysteresis: the difference between the V_I trigger and V_{II} trigger. The hysteresis voltage for the configuration described is altered by the internal feedback of the CA3062 circuit which modifies the hysteresis expression.

Fig. a shows a typical configuration for operation as a Schmitt-Circuit.

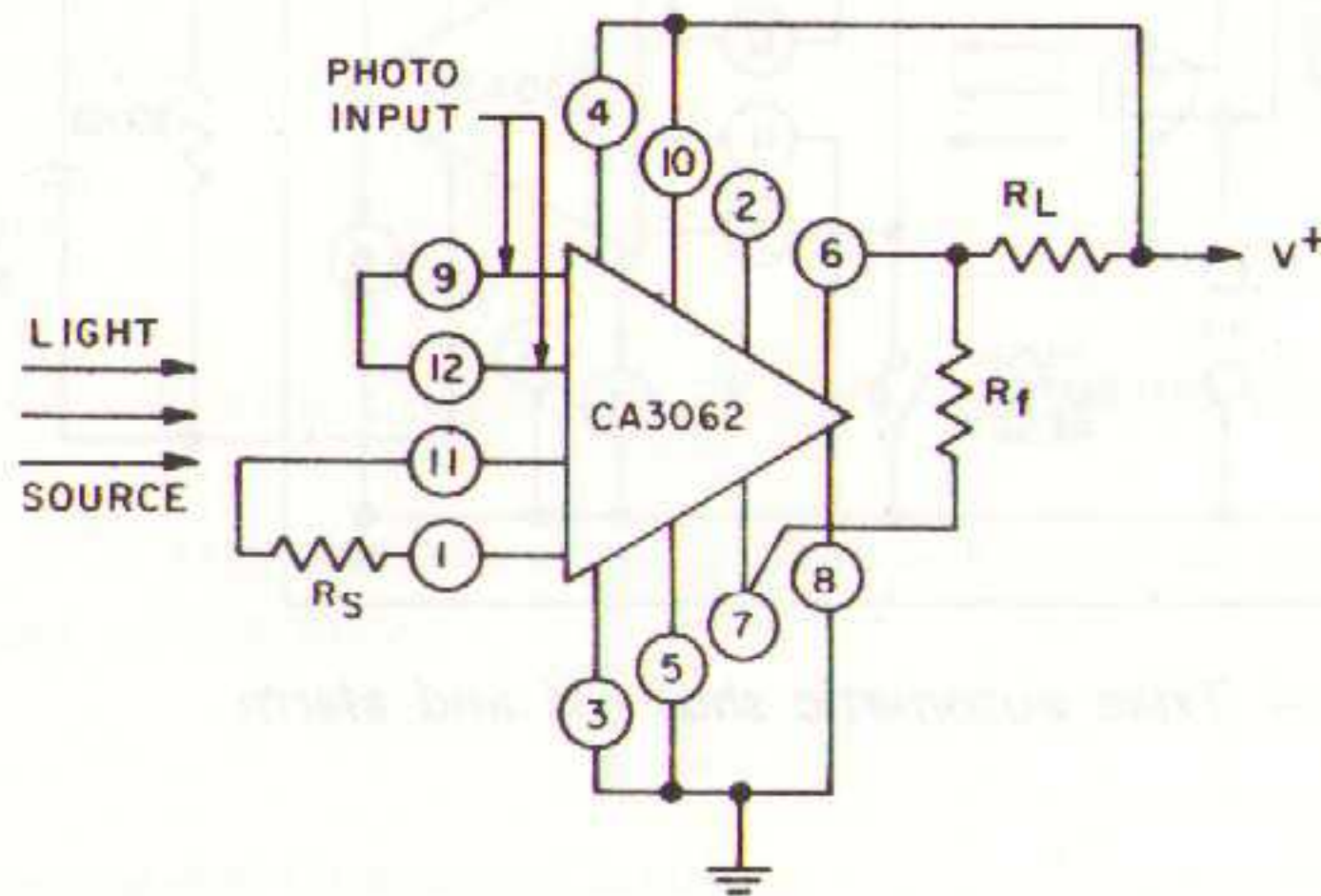


Fig. a - Schmitt-circuit

The trigger potential (V_I) to go from state I to state II is:

$$V_I \text{ trigger} = \frac{(V^+ - V_{B31}) R_{B3}}{R_{B3} + R_f + R_L} + \frac{V_{B31}}{1}$$

Where state I is defined as the case with pin 6 at a "high" output voltage, with no light into the Darlington photo-detector. State II is defined as the case with pin 6 at a "low" output voltage, with no light into the Darlington photo-detector.

$$\begin{aligned} \text{and: } V_{B21} &= 1.050 \text{ V} & R_{B2} &= R_4 // R_5 \\ V_{B31} &= 0.700 \text{ V} & R_{B3} &= R_6 // R_7 \quad (\text{Refer to Fig. 1}) \\ V_{B22} &= 0.700 \text{ V} \\ V_{B32} &= 1.050 \text{ V} \end{aligned}$$

The trigger potential (V_{II}) to go from state II back to state I is:-

$$V_{II} \text{ trigger} = \frac{V_{B32}}{1} - \frac{(V_{B32} - V_{SAT}) R_{B3}}{R_{B3} + R_f}$$

The hysteresis voltage of the CA3062 Schmitt system is:

$$\begin{aligned} V \text{ hysteresis} &= \left[\frac{(V^+ - V_{B31}) R_{B3}}{R_{B3} + R_f + R_L} + \frac{V_{B31}}{1} + \frac{V_{B22}}{1} - \frac{V_{B21}}{1} \right] \\ &\quad - \left[\frac{V_{B32}}{1} - \frac{(V_{B32} - V_{SAT}) R_{B3}}{R_{B3} + R_f} \right] \end{aligned}$$

The light intensity required to switch from state I to state II is determined from Fig. 2, which shows the number of lumens/ft² necessary to generate I_{CI} .

$$I_{CI} = \frac{V_{SIG I}}{R_{B2}} = \frac{V_{SIG I}}{5 \times 10^3}$$

where:

$$V_{SIG I} = \frac{(V^+ - V_{B31}) R_{B3}}{R_{B3} + R_f + R_L} + \frac{V_{B31}}{1} - \frac{V_{B21}}{1}$$

To determine the lumens/ft² required to go from state II back to state I, it is necessary to calculate $V_{SIG II} + V_{B22}$, (the voltage required to cause the Schmitt circuit to switch).

$$V_{SIG II} + V_{B22} = \frac{V_{B32}}{1} - \frac{(V_{B32} - V_{SAT}) R_{B3}}{R_{B3} + R_f}$$

Thus:

$$V_{SIG II} = V_{B32} - \left[\frac{(V_{B32} - V_{SAT}) R_{B3}}{R_{B3} + R_f} \right] \left[\frac{V_{B22}}{1} \right]$$

and:

$$I_{CII} = \frac{V_{SIG II}}{R_{B2}} = \frac{V_{SIG II}}{5k}$$

The rate at which V_{SIG} rises is governed by the response of the photo Darlington pair.

V_{SIG} during the rise time assumes a value:

$$V_{SIG} = V_m (1 - e^{-\frac{t}{RC}});$$

where V_m is the maximum signal voltage developed across terminal No. 1 and reference terminal No. 8 for any given light intensity. [See Fig. 4.]

The rise time of a typical photo Darlington of the CA3062 is:

$$RC \log \left[\frac{1}{1 - \frac{V_c}{V_m}} \right]$$

$$t_r = \frac{\log e}{\log e}$$

where $R=40 \times 10^6$ and $C = 8 \times 10^{-12}$ and V_C any part of V_m .

APPENDIX (cont'd.)

By setting $V_C = V_{SIG I}$ the delay time, t_d , of the pulse output of the Schmitt circuit can be determined for any given light intensity.

The rate at which the photo Darlington output decreases is:

$$V_{SIG} = V_m e^{-\frac{t}{RC}}$$

Thus the fall time is

$$t_f = \frac{RC \log \frac{V_m}{V_{ct}}}{\log e}$$

Setting $V_C = V_{SIG II}$, then t_f becomes t_d which is turn off delay time, or the time the Schmitt circuit turns off with respect to the time the light input to the photo Darlington has been turned off. See Fig. b.

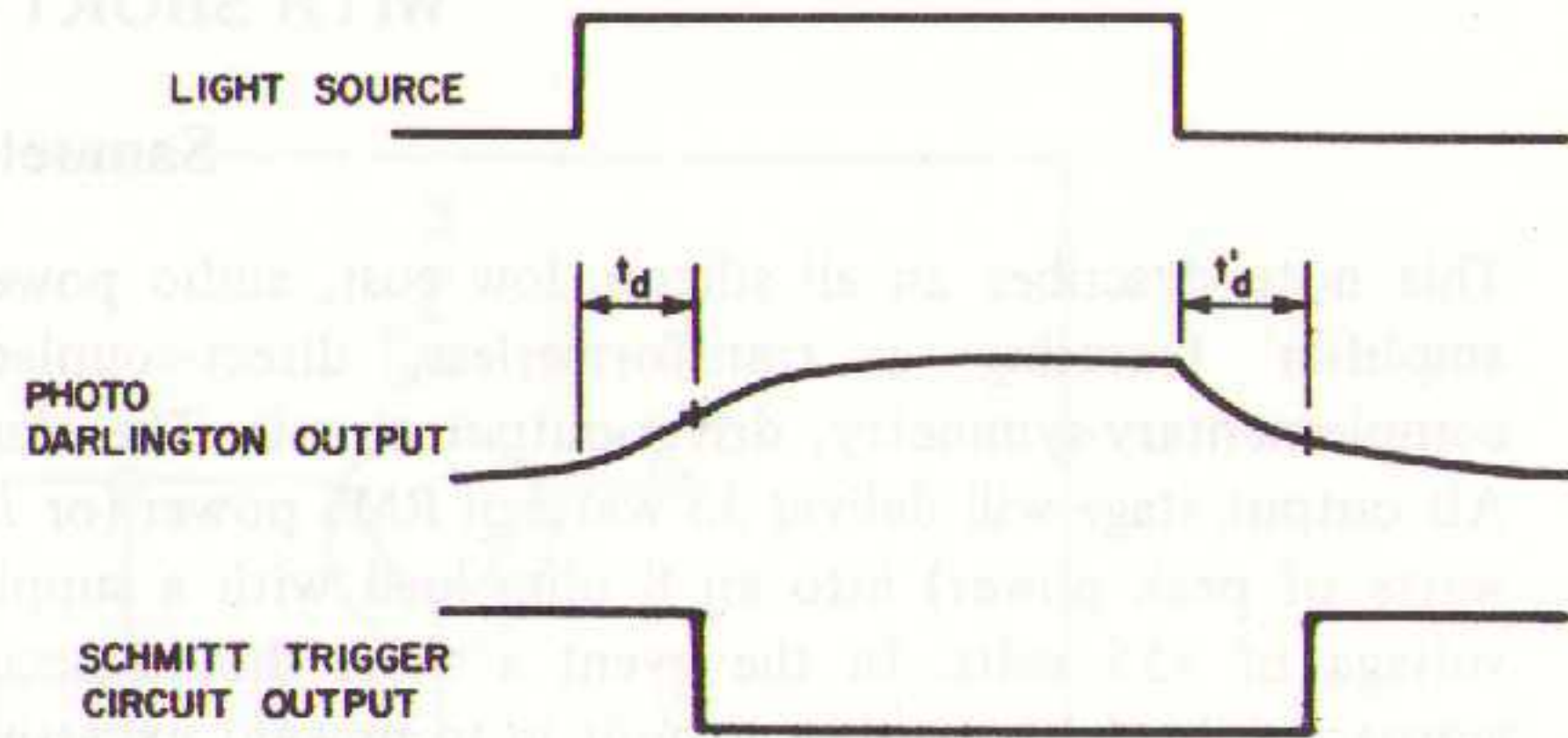
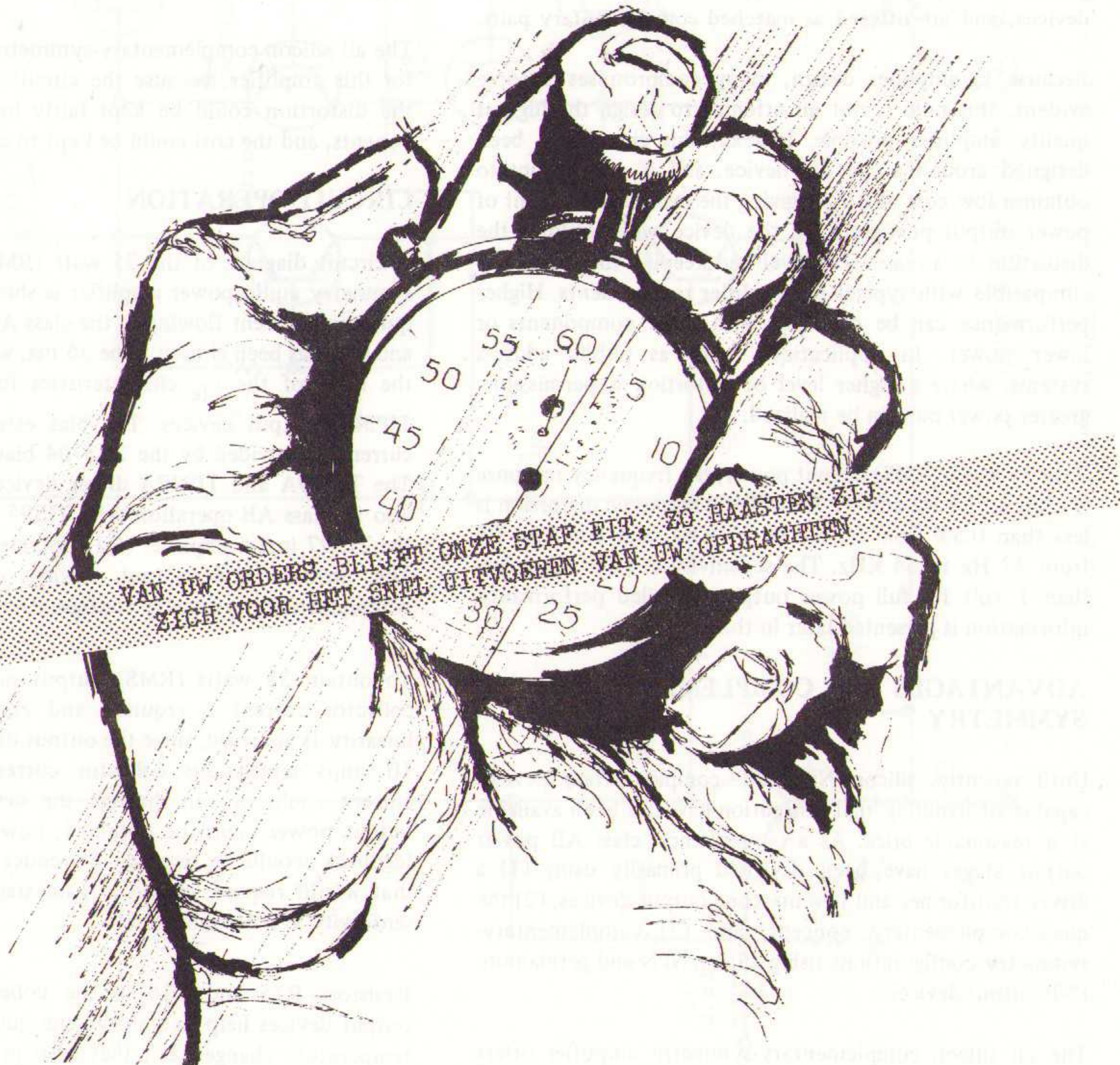


Fig. b — Waveforms for rise and fall response.



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ALL SILICON COMPLEMENTARY-SYMMETRY 35 WATT (RMS) AUDIO POWER AMPLIFIER WITH SHORT CIRCUIT PROTECTION

Samuel W. Webster, Jr.

This note describes an all silicon, low cost, audio power amplifier featuring a transformerless, direct-coupled, complementary-symmetry, driver-output circuit. The class AB output stage will deliver 35 watts of RMS power (or 70 watts of peak power) into an 8 ohm load with a supply voltage of +55 volts. In the event a short should occur across the load, protection is built in to prevent excessive current from damaging the output transistors. The amplifier has been designed around TI's new epitaxial base plastic power transistors, which are available as both NPN and PNP devices, and are offered as matched complementary pairs.

Because in amplifier design, many compromises become evident, this note is not an attempt to design the highest quality amplifier possible. Instead, this circuit has been designed around a specific device, and is an attempt to obtain a low cost amplifier giving the maximum amount of power output possible from the device, while holding the distortion to a reasonable level and keeping the sensitivity compatible with typical preamplifier requirements. Higher performance can be obtained with more components or lower power. In applications, such as public address systems, where a higher level of distortion is permissible, greater power out can be realized.

With 35 watts (RMS) output power, the frequency response is -3db at 7.4 Hz and 91 KHz. Total harmonic distortion is less than 0.5% from 60 Hz to 17 KHz, and less than 1% from 37 Hz to 54 KHz. The sensitivity at 1000 Hz is less than 1 volt for full power output. Detailed performance information is presented later in this report.

ADVANTAGES OF COMPLEMENTARY-SYMMETRY

Until recently, silicon NPN-PNP complementary devices capable of handling high dissipation have not been available at a reasonable price. As a consequence, class AB power output stages have been designed primarily using (1) a driver transformer and two matched output devices, (2) the quasi-complementary concept, and (3) complementary-symmetry configurations using silicon NPN and germanium PNP output devices.

The all silicon complementary-symmetry amplifier offers advantages over all three of the above. It does not require a costly transformer like the driver transformer circuit. It

does not have unequal input impedances that cause distortion like the quasi-complementary configuration. In the quasi-complementary, the signal sees impedances due to two base-emitter junctions on one half cycle and an impedance due to one junction on the other half cycle. The germanium-silicon complementary-symmetry concept has the disadvantage of each output device having greatly different base-emitter voltages (approximately 0.3 volts for germanium and 0.7 volts for silicon) and different temperature characteristics.

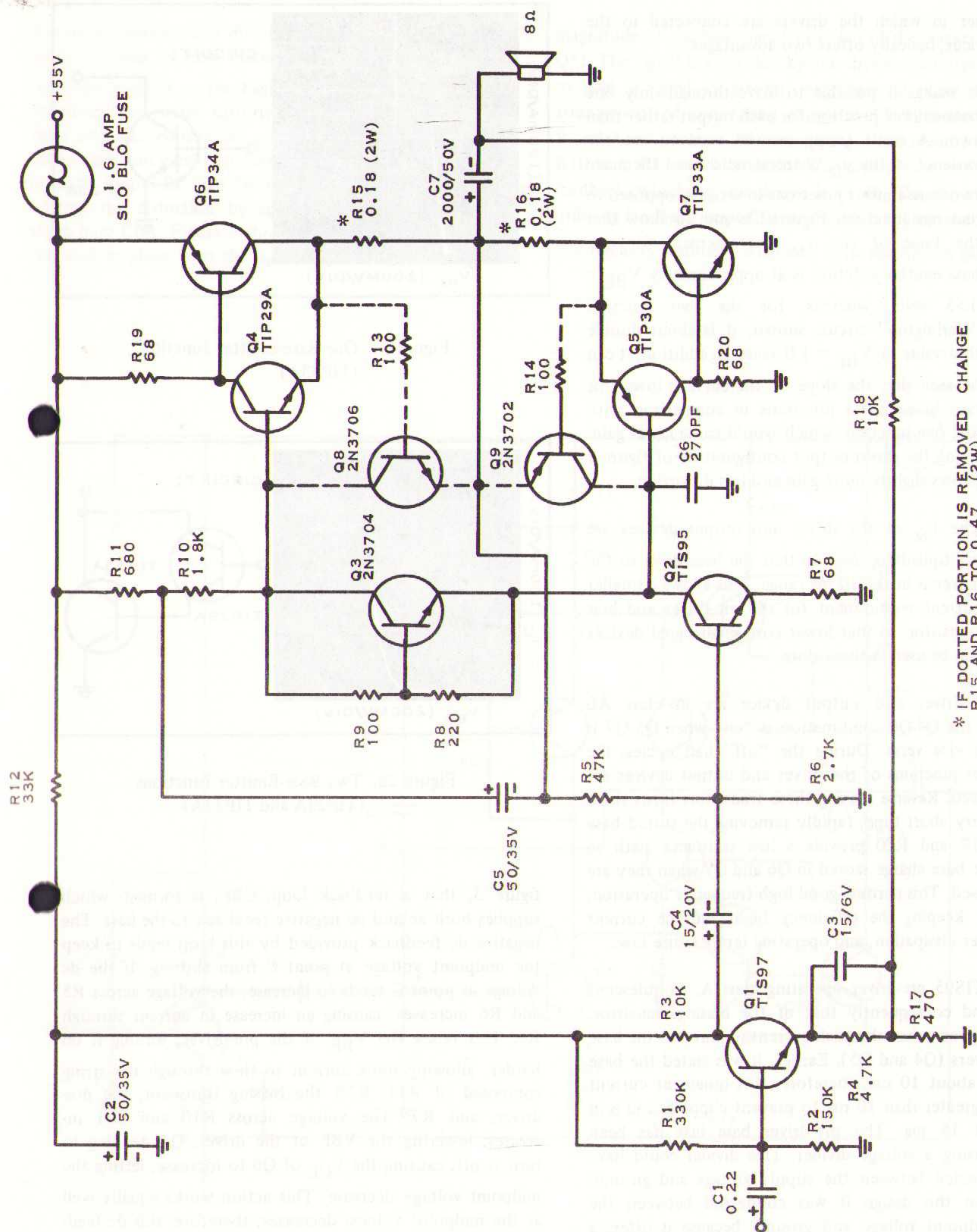
The all silicon complementary-symmetry design was chosen for this amplifier, because the circuit is relatively simple, the distortion could be kept fairly low with fewer components, and the cost could be kept to a reasonable level.

CIRCUIT OPERATION

A circuit diagram of the 35 watt (RMS) complementary-symmetry audio power amplifier is shown in figure 1. The quiescent current flowing in the class AB output stages (Q6 and Q7) has been chosen to be 36 ma, which places it above the knee of the y_{fe} characteristics for the TIP33A and TIP34A output devices. The bias establishing this idling current is provided by the 2N3704 biasing transistor (Q3). The TIP29A and TIP30A driver devices (Q4 and Q5) are also in class AB operation. The TIS95 pre-driver (Q2) and the TIS97 input amplifier (Q1) operate in Class A fashion. The drivers and output devices are used in the complementary-symmetry configuration.

To obtain 35 watts (RMS) output power, 3 amps peak collector current is required and excellent current gain linearity is achieved, since the output devices are capable of 10 amps continuous collector current. A higher peak current could be run through the devices and a greater output power could be obtained; however, more negative feedback would be required to reduce the distortion, and that would require additional gain stages to maintain the sensitivity at a desirable level.

Resistors R15 and R16 in the collector circuit of the output devices help to stabilize the quiescent current from temperature changes, but they also provide negative feedback to the output devices to help improve the h_{fe} non-linearity.



* IF DOTTED PORTION IS REMOVED CHANGE
R15 AND R16 TO 0.47 (2W).

Figure 1. 35 Watt (rms) Audio Amplifier With Short Circuit Protection

The manner in which the drivers are connected to the output devices, basically offers two advantages.

- 1) It makes it possible to drive through only one base-emitter junction for each output rather than two. A curve tracer can be used to see the variance of the y_{fe} characteristic when there are two base-emitter junctions in series as opposed to just one junction. Figures 2a and 2b show that the knee of the y_{fe} characteristic for one base-emitter junction is at approximately $V_{BE} = 0.55$ volts; whereas, for the two junction "Darlington" circuit shown, it is about double that value or $V_{BE} = 1.0$ volts. In addition, it can be seen that the slope of the curve is lower for two base-emitter junctions in comparison with just one junction, which would cause lower gain. Using the driver-output configuration of figure 1 allows slightly more gain and less distortion.
- (2) The h_{fe} of the driver and output devices are multiplied together so that the base drive to the driver is in the 10 ma range. This allows a smaller current requirement for the pre-driver and bias transistor, so that lower cost small signal devices can be used in these slots.

Since the driver and output device are in class AB operation, the Q4-Q6 combination is "on" when Q5-Q7 is "off", and vice versa. During the "off" half cycles, the base-emitter junctions of the driver and output devices are reverse biased. Reverse biasing these transistors turns them off in a very short time, rapidly removing the stored base charge. R19 and R20 provide a low resistance path to remove the base charge stored in Q6 and Q7 when they are reverse biased. This permits good high frequency operation, as well as keeping the efficiency high and the current drain, power dissipation, and operating temperature low.

With the TIS95 pre-driver operating class A, its quiescent current, and consequently that of the biasing transistor, should be greater than the peak current required to the base of the drivers (Q4 and Q5). Earlier, it was stated the base drive was about 10 ma; therefore, this quiescent current should be greater than 10 ma to prevent clipping, and is in fact about 15 ma. The pre-driver base bias has been obtained using a voltage divider. This divider could have been connected between the supply voltage and ground; however, in this design it was connected between the output midpoint voltage and ground, because it offers a couple of advantages in addition to biasing the base. By connecting to the midpoint of the output, it can be seen in

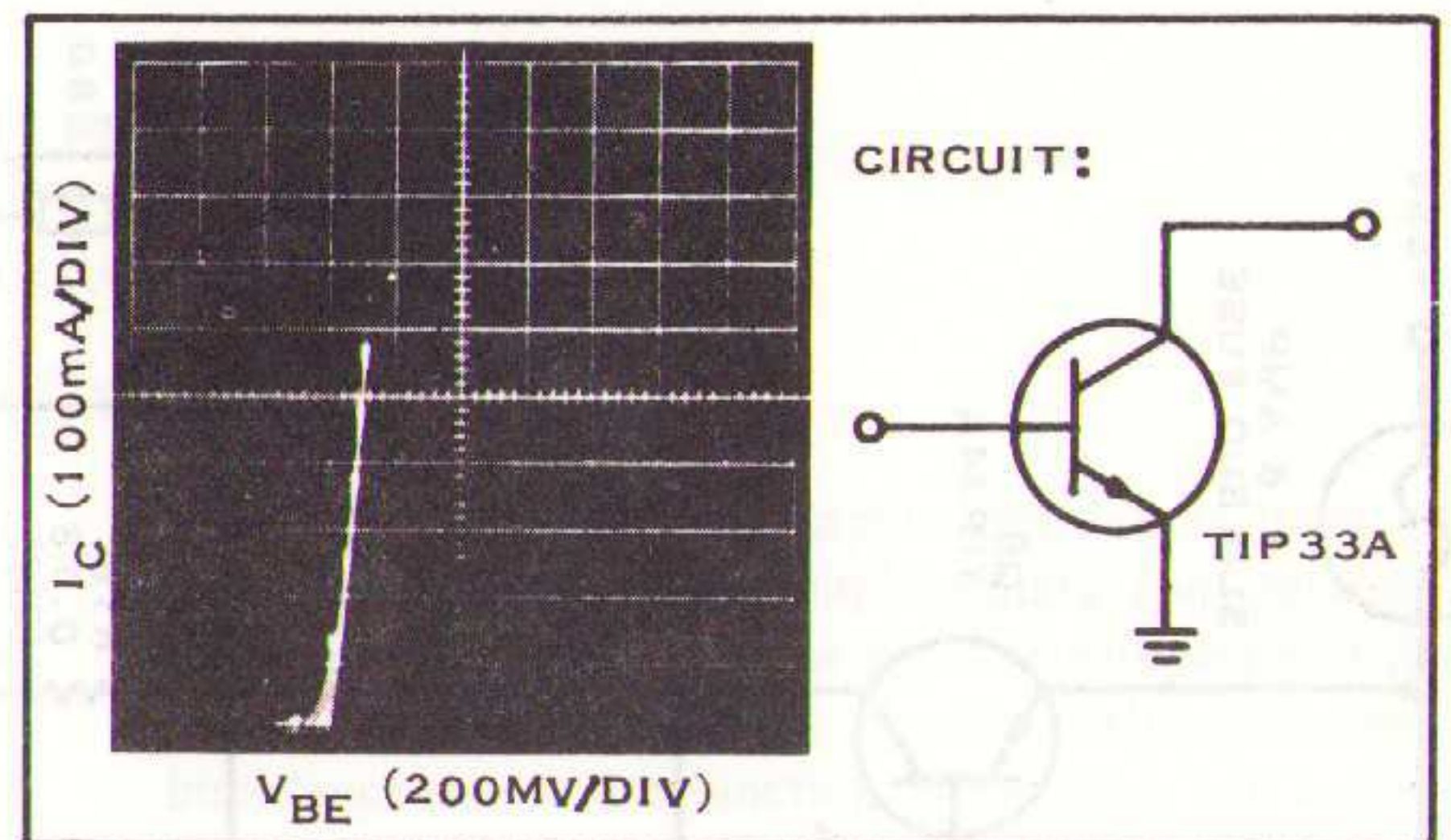


Figure 2a. One Base-Emitter Junction (TIP33A)

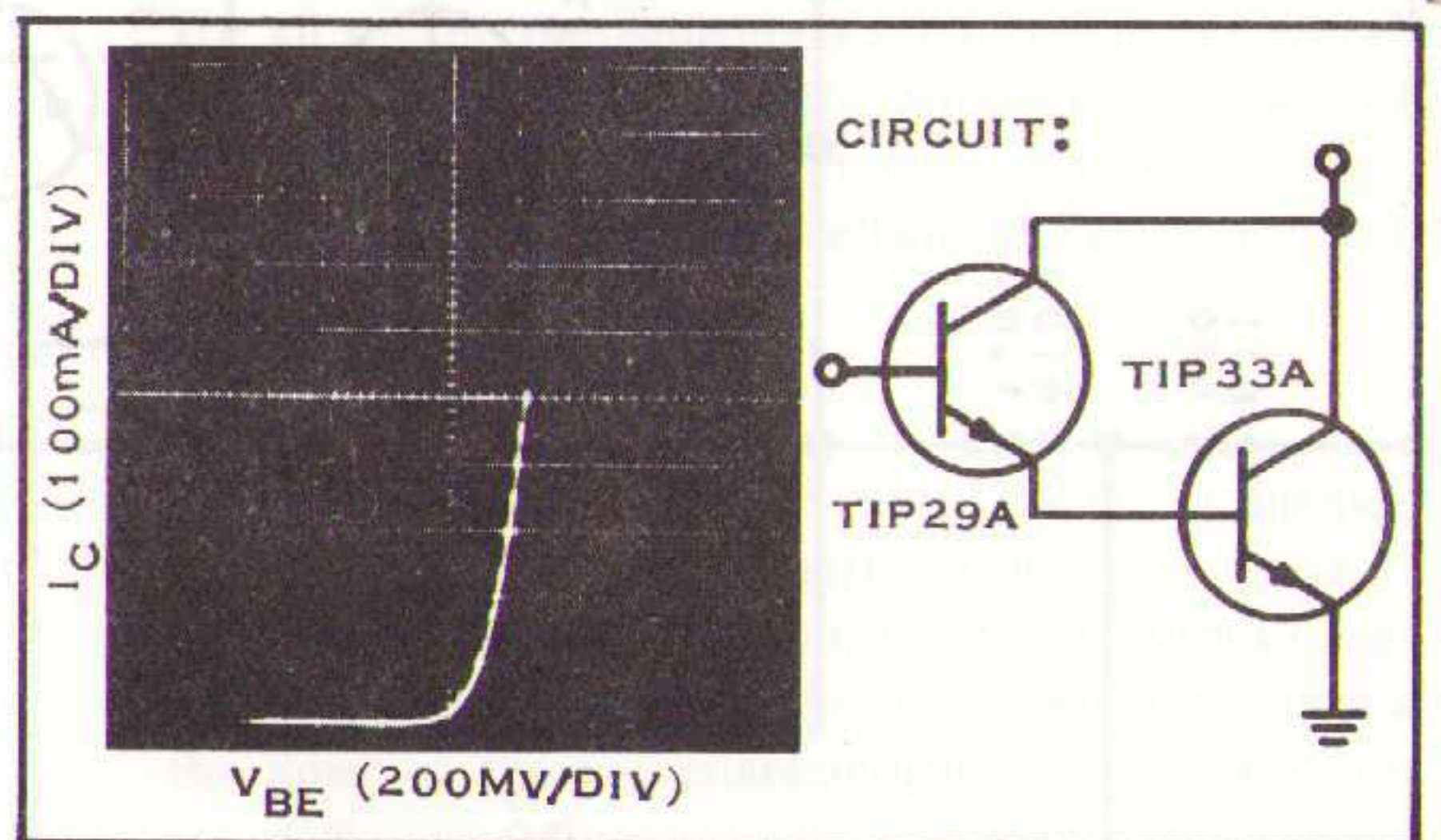


Figure 2b. Two Base-Emitter Junctions (TIP29A and TIP33A)

figure 3, that a feedback loop, CBF, is formed, which supplies both ac and dc negative feedback to the base. The negative dc feedback provided by this loop tends to keep the midpoint voltage at point C from shifting. If the dc voltage at point C tends to increase, the voltage across R5 and R6 increases, causing an increase in current through R6. This raises the V_{BE} of the pre-driver, turning it on harder, allowing more current to flow through the string composed of R11, R10, the biasing transistor, the pre-driver, and R7. The voltage across R10 and R11 increases, lowering the V_{BE} of the driver, Q4, tending to turn it off, causing the V_{CE} of Q6 to increase, letting the midpoint voltage decrease. This action works equally well as the midpoint voltage decreases; therefore, this dc feedback stabilizes the voltage at point C to the desired value of $\frac{1}{2}V_{CC}$ or approximately 27.5 volts.

The loop provides -0.5 db of negative ac feedback to the pre-driver base to help reduce the non-linear h_{FE} distortion. The negative ac feedback introduced has the effect of improving the circuit distortion characteristics, but at the same time has reduced the gain and input sensitivity. Some of the gain and sensitivity lost due to the negative feedback through loop CBF can be recovered without drastically hurting the distortion by applying positive ac feedback along loop CBA. Figure 3 shows the signal at point C being fed back in phase with the signal at point A, increasing the

magnitude of the base signal to either of the drivers (Q4 or Q5). The signal is amplified by the driver and output stage to give a signal of greater magnitude at point C. This arrangement of connecting capacitor C5 between points B and A is known as "bootstrapping." The resistor divider, R10 and R11, and capacitor C5 control the amount of feedback, which for this design is +1 db. The net effect of utilizing feedback loops CBF and CBA is to improve the h_{fe} non-linearity distortion with very little sacrifice in gain.

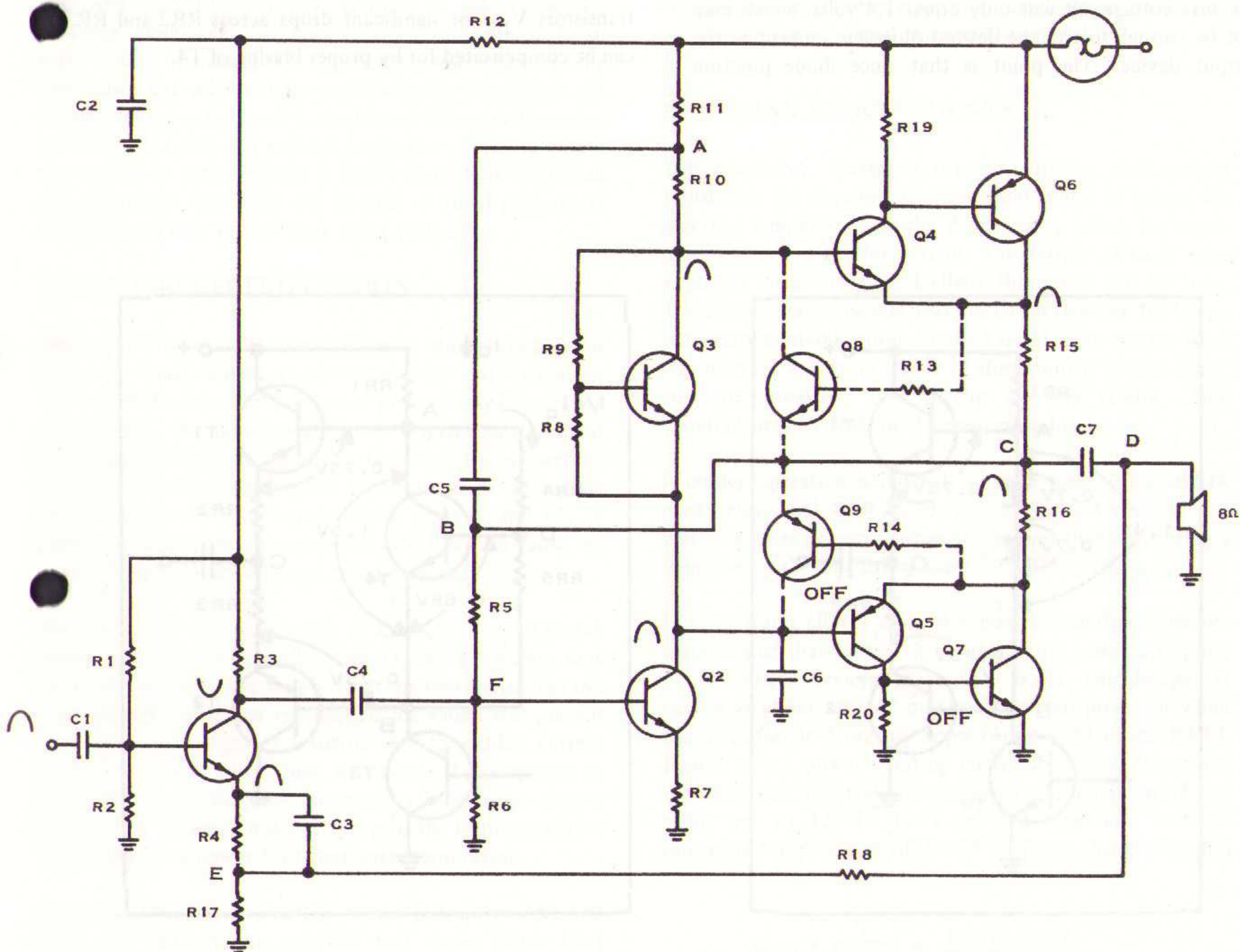
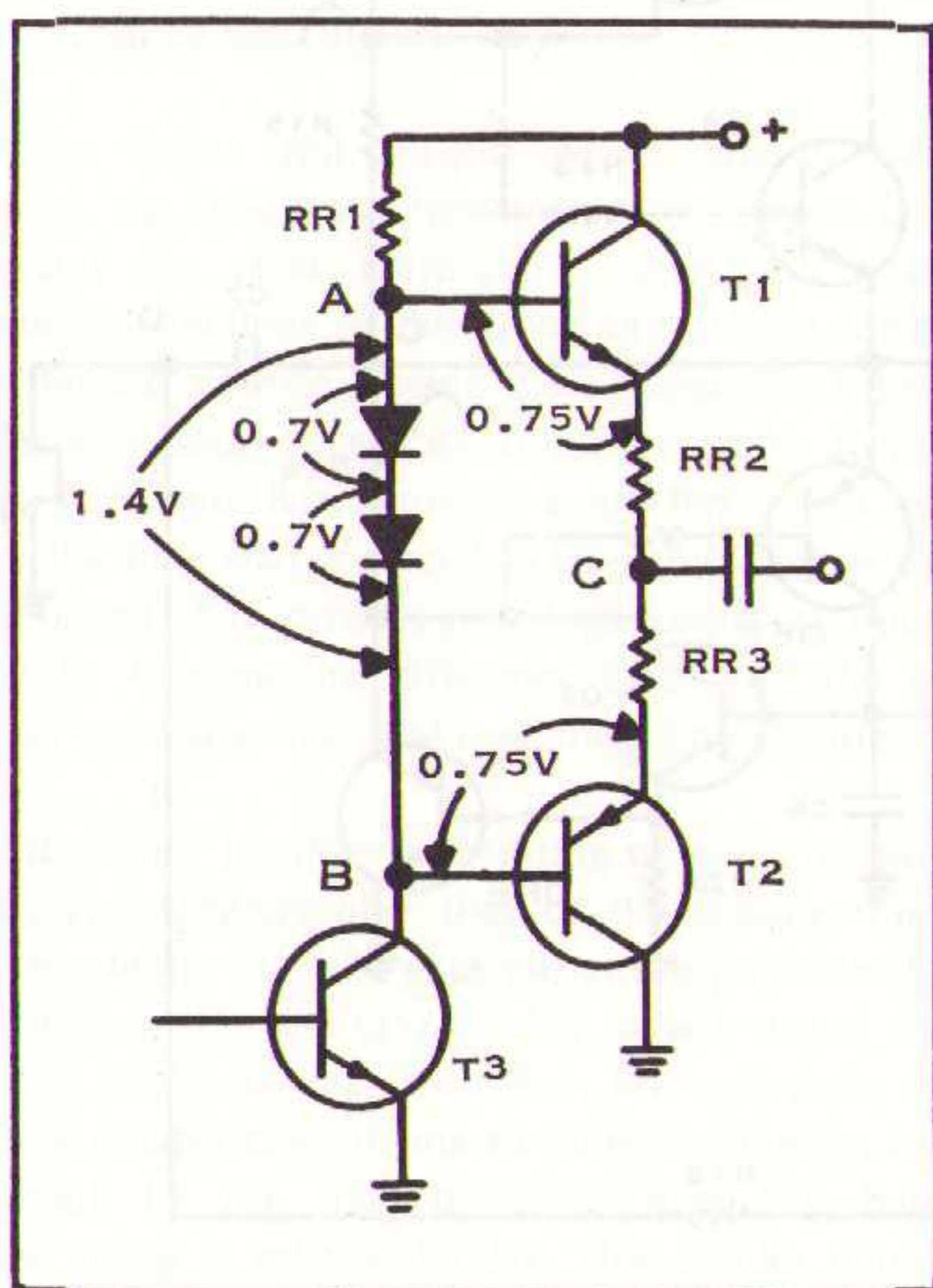


Figure 3. Phase Relationship at Different Points

voltages and transistor base-emitter voltages can vary from the typical 0.7 volts, it would become quite a task selecting diodes which would give the exact multiple of voltages to properly bias the circuit. It would not only be difficult, but very costly, and if the voltages considered negligible here were significant, it would be even more difficult. The choice for this design is, as in figure 4b, a transistor as the biasing element. The transistor can be biased to give any desired bias voltage across the output circuit. The procedure used in this design, and one which normally gives good results, is to allow approximately $\frac{1}{4}$ of the quiescent current through RR1 to flow in base bias resistors, RR4 and RR5, of T4. The voltage drop ACB is determined by summing the drops around that loop. The voltages AB and ADB should equal voltage ACB. The voltage at point D should be the voltage at point B plus the V_{BE} of T4. From this brief analysis it can be seen that variances in the transistors V_{BE} or significant drops across RR2 and RR3 can be compensated for by proper biasing of T4.



The diagram shows a differential amplifier circuit with four transistors, T_1 , T_2 , T_3 , and T_4 . Transistors T_1 and T_2 are the main output devices, while T_3 and T_4 are part of the current source and biasing network. A tail current I_Q is supplied to the circuit. The circuit includes resistors RR_1 , RR_2 , RR_3 , RR_4 , and RR_5 . Biasing voltages are indicated: $0.75V$ for the base-emitter junctions of T_1 and T_2 , $1.5V$ for the base-emitter junction of T_4 , and $0.65V$ for the base-emitter junction of T_3 . A node voltage of $1/4 I_Q$ is also shown. The output is taken from the collector of T_1 through a capacitor.

Figure 4b. Transistor Biasing

Q1 is an input amplifier and has been capacitor coupled to the pre-driver. It provides, in addition to increased amplification of the input signal, a point to which additional negative feedback can be returned to help improve the overall distortion. A signal is fed back from the output of the overall amplifier to the emitter of Q1, along path DE (figure 3), providing negative ac feedback. The feedback signal and the input signal are in phase, but in opposition, because as the input signal raises the base voltage, the feedback signal raises the emitter voltage. The larger the feedback signal, the harder Q1 is to turn on. The amount of feedback is controlled by the voltage divider R4 and R17, and the feedback resistor, R18.

Utilizing feedback loops CBF and CBA together did not create a compromise situation, since distortion was slightly improved with little, if any, sacrifice in gain. However, feedback along path DE does create a compromise situation, since increasing the feedback will improve the distortion, but will hurt the sensitivity. In this design, -14 db of negative ac feedback is used, since it lowers distortion by better than one half and still keeps the sensitivity within the desired lv level. Figures 6 and 9 show how distortion and sensitivity vary with -14 db and -5 db of negative ac feedback and with no feedback along path DE.

SHORT CIRCUIT PROTECTION

The possibility usually exists where a short could occur across the load, causing an extremely large output current to flow. If the transistor current or power dissipation rating is exceeded, the device will be destroyed. Short circuit protection has been designed into the amplifier and serves a dual purpose: (1) to prevent the current from exceeding the device ratings (mainly from sudden surges), and (2) to prevent a large continuous current which would cause the power dissipation ratings from being exceeded.

The protective circuit consists of Q8, Q9, R13, R14 (all shown with dotted lines in figure 1), and a 1.6 amp slow blow fuse in the power supply line. The two transistors and resistors make up a fast reacting circuit which will prevent damage to the output transistors due to sudden current surges. The slow blow fuse will cause the amplifier to disconnect from the power supply if a continuous current greater than 1.6 amps is drawn, keeping the transistors from dissipating more power than their maximum rating.

Transistor Q8 gives protection on the output positive half cycle and Q9 on the negative half cycle. Since their operation is the same, only the positive half cycle will be

discussed. During normal operation of the amplifier, the current through R15 is not large enough to raise the voltage across it high enough to bias Q8 on. When a short circuit across the load develops, the current through R15 rises rapidly, developing a voltage across R15 and, consequently, across the base-emitter of Q8 great enough to turn it on. A portion of the current, which normally would flow into the base of Q4, now flows as collector current in Q8, preventing Q4 and Q6 from turning on beyond a predetermined value, keeping the collector current below the maximum output device rating.

Should it be desired not to have this feature, it can be removed by eliminating Q8, Q9, R13, R14, and the slow blow fuse, and in addition, the value of R15 and R16 would have to be changed to 0.47 ohms. With these changes, the performance without short circuit protection will remain the same as if the amplifier had it.

HEAT SINK REQUIREMENTS

The pre-driver, biasing transistor, drivers, and output transistors are required to have heat sinks to insure the junction temperature of the devices stay below the maximum ratings. For this circuit, the output devices were mounted on a one piece Delbert Blinn heat sink 6" long. The biasing transistor was also placed on this sink for better temperature stabilization of the circuit. The drivers were mounted on a 4 sq. in. piece of aluminum 1/16 inch thick, and the pre-driver uses a clip on sink (Delta 203, manufactured by Wakefield Engineering Inc.).

If cooler operation is desired, a larger heat sink could be used. Heat sink size is arrived at by determining the case-to-ambient thermal impedance, and then consulting a manufacturer's table listing sinks for different θ_{C-A} .

For 35 watts (RMS) class AB power amplifier, approximately $\frac{1}{4}$ of that wattage is the maximum power dissipated by the output devices, or about 9 watts. This dissipation occurs at about 40% of maximum output power or when the amplifier is delivering approximately 21 watts (RMS). Figure 5 is a power derating curve. A look at this curve indicates that if 9 watts is dissipated, the temperature can safely go to 135°C. Let's say the maximum ambient operating temperature will be 55°C. From the relationship

$$\theta_{C-A} = \frac{T}{P_{DISS}} = \frac{135^{\circ}\text{C} - 55^{\circ}\text{C}}{9 \text{ watts}}$$

or θ_{C-A} is about 9°C per watt. Consulting a table, it will be found that the output device heat sink used is more than adequate. If it is desirable to have the temperature lower than 135°C , substitute a lower figure into the equation, and it will be found that θ_{C-A} decreases indicating a different type or a larger heat sink.

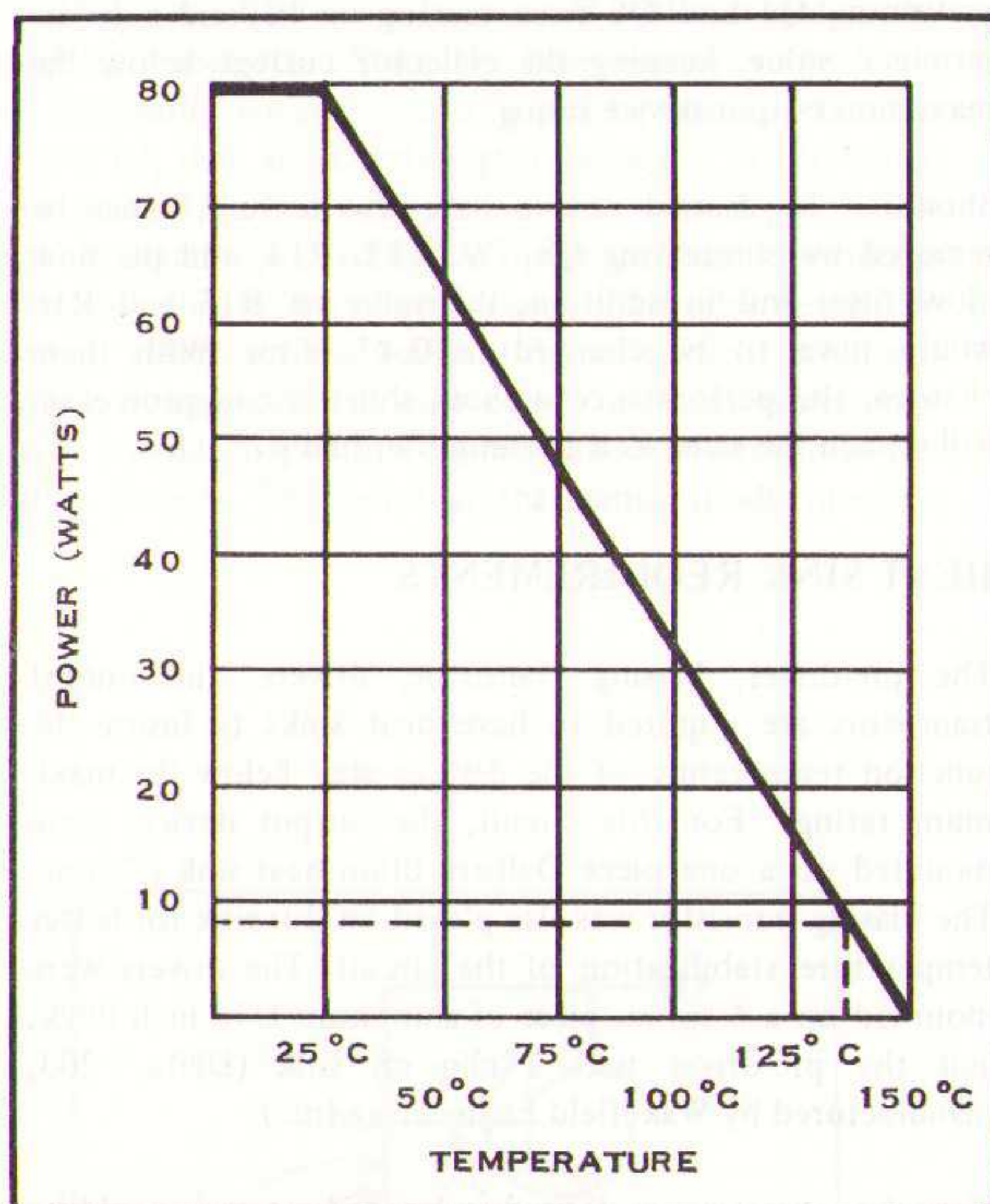


Figure 5. Dissipation Derating Curve

PERFORMANCE DATA

Output Power	35 watts (RMS) power into an 8 ohm load with a supply voltage of +55 volts.	
Sensitivity	940 mv at full rated power at 1000 Hz. (See figure 6)	
Frequency Response	7.4 Hz to 91 KHz at -3 db rolloff from rated output power (See figure 7)	
Total Harmonic Distortion	0.43% from 75 Hz to 10 KHz, and less than 1.0% from about 35 Hz to 52 KHz at rated output power (See figures 8 and 9)	
Intermodulation Distortion (60 Hz and 7 KHz mixed at a 4:1 ratio)	2.0% at rated power output 0.2% at 3.5 watts output (See figure 10)	
Power Supply Current Drain	(See figure 11)	
Square Wave Response	(See figure 12)	
Signal to Noise	Input Open:	- 98 db
	Input Shorted:	-105 db
Input Impedance	39 Kohms @ 1000 Hz	
Output Impedance	0.5 ohms @ 1000 Hz	
Damping Factor	16	
Amplifier Efficiency	70%	

Current and Voltage Data

Supply Voltage	50v	55v				60v
Ambient Temperature	25°C	25°C	35°C	45°C	55°C	25°C
Output Device Quiescent Current	11ma	36ma	82ma	120ma	161ma	65ma
Power Supply Quiescent Current Drain	31ma	58.5ma	117ma	155ma	195ma	85.5ma
Output Midpoint Voltage	24.8v	26.5v			25.0v	27.8v

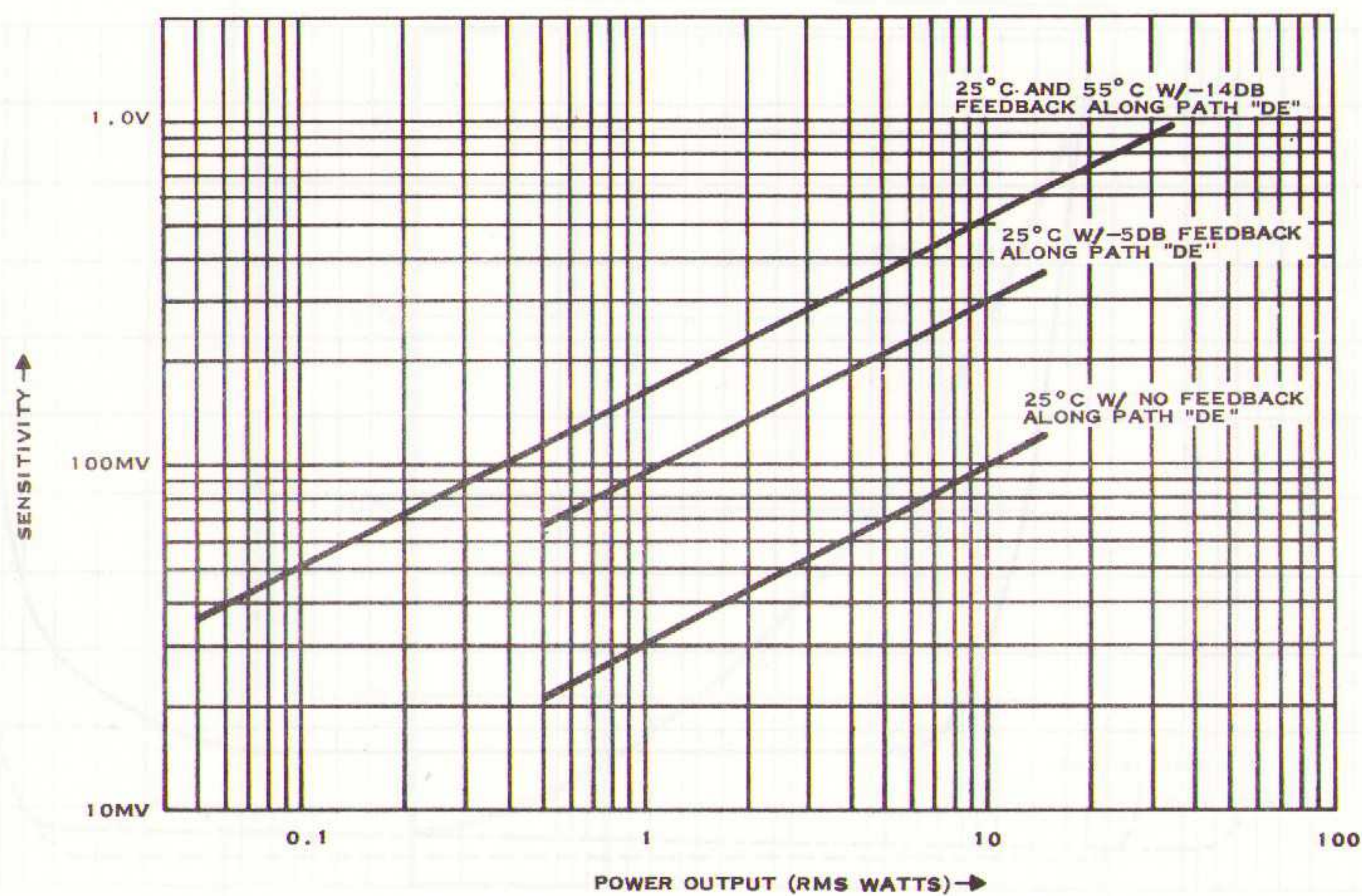


Figure 6. Sensitivity vs Power Output @ 1000 Hz

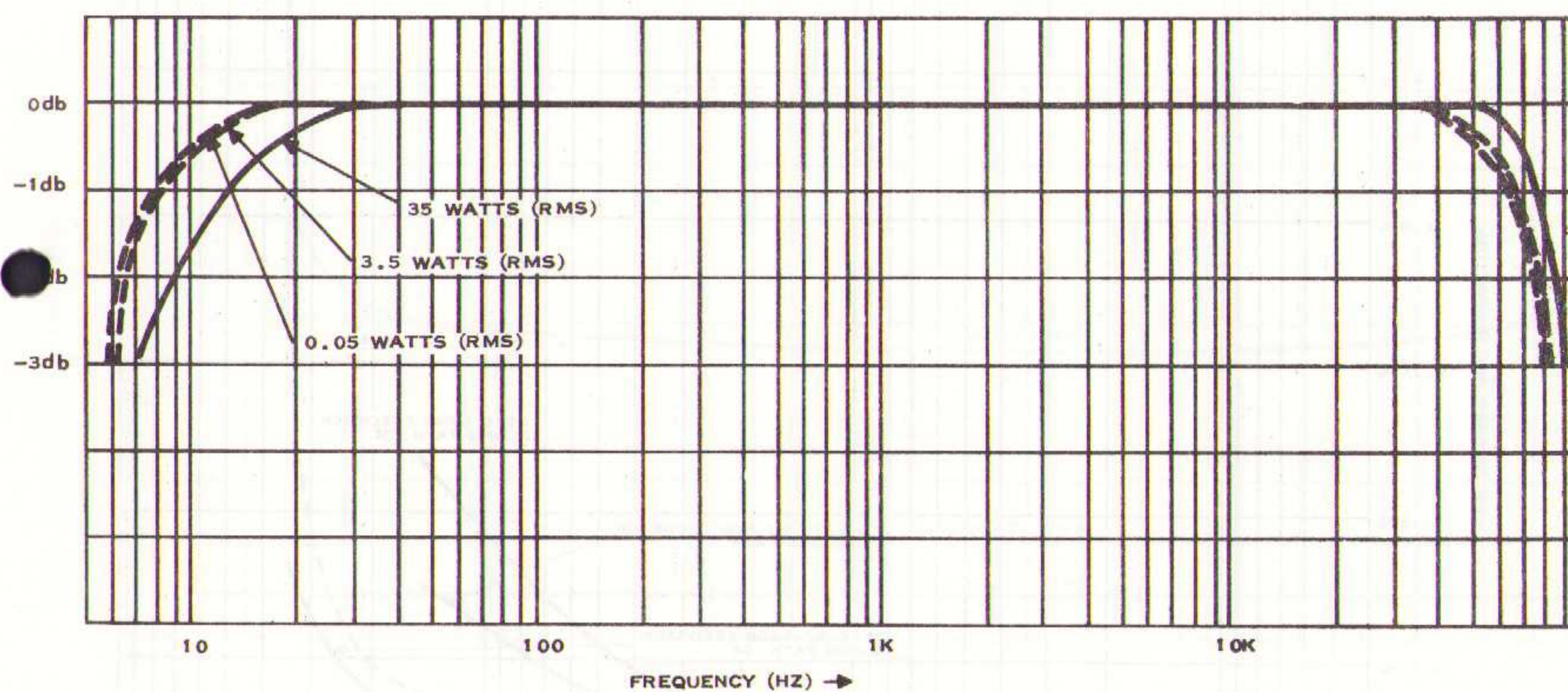


Figure 7. Frequency Response at 25°C

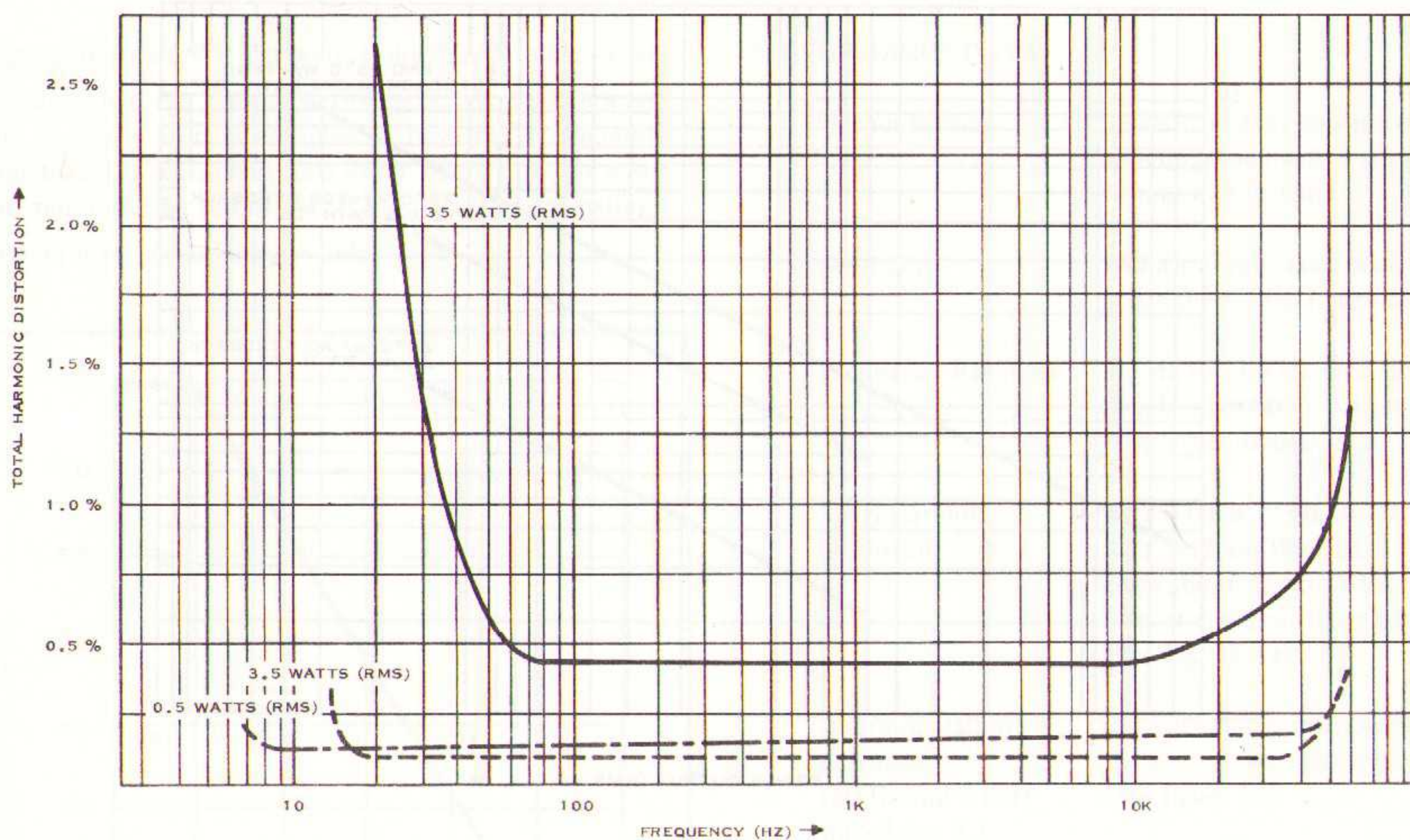


Figure 8. Total Harmonic Distortion vs Frequency at 25°C

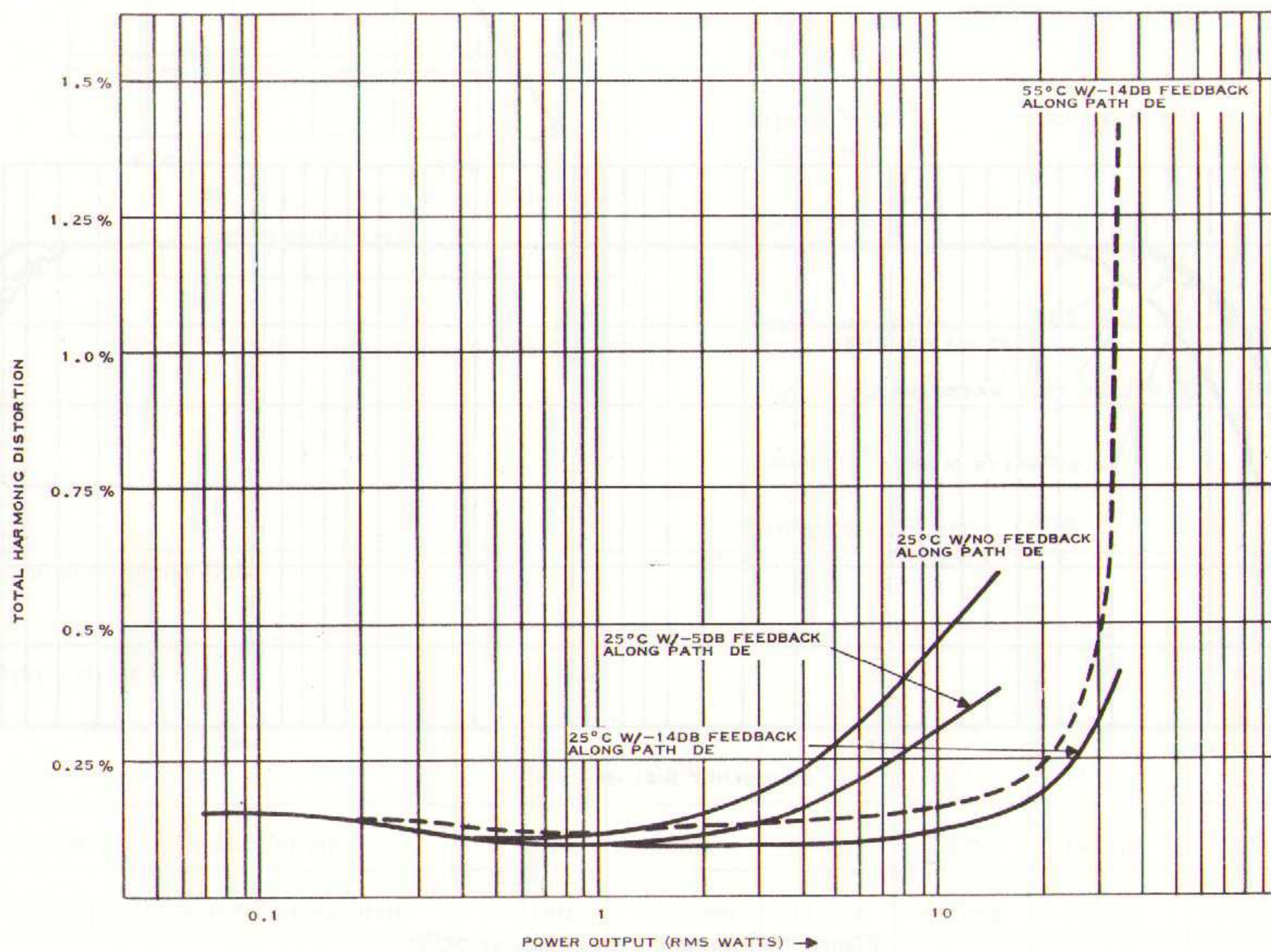


Figure 9. Total Harmonic Distortion vs Power Output at 1000 Hz

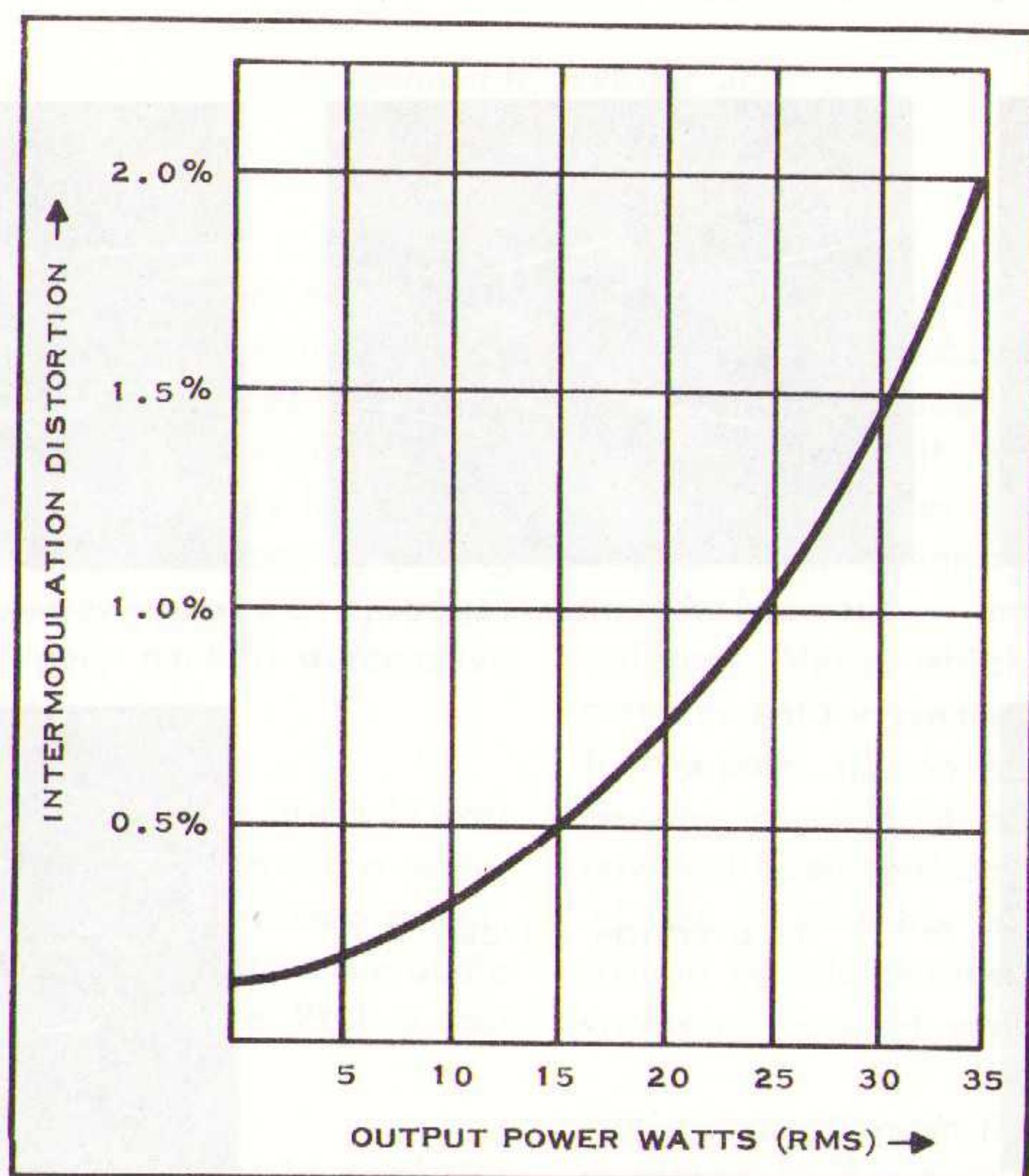


Figure 10. Intermodulation Distortion vs Power Output at 25°C

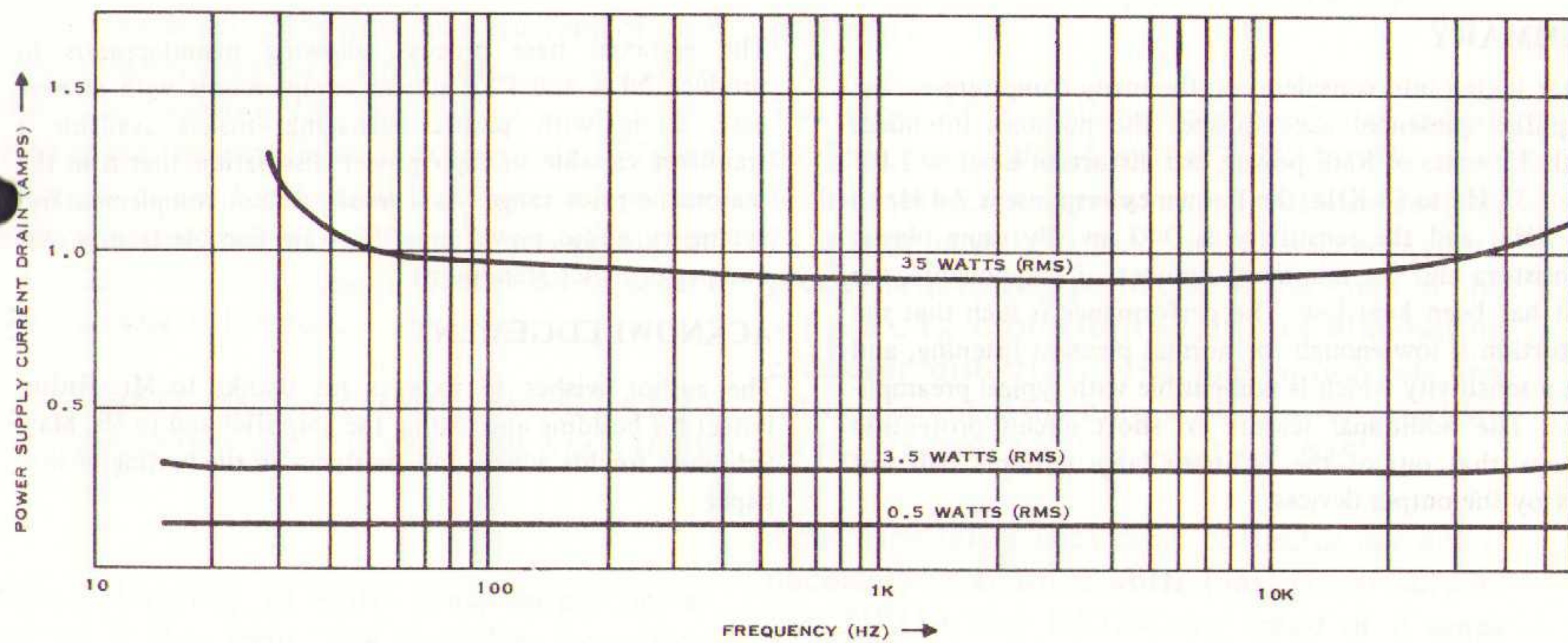


Figure 11. Power Supply Current Drain vs Frequency at 25°C

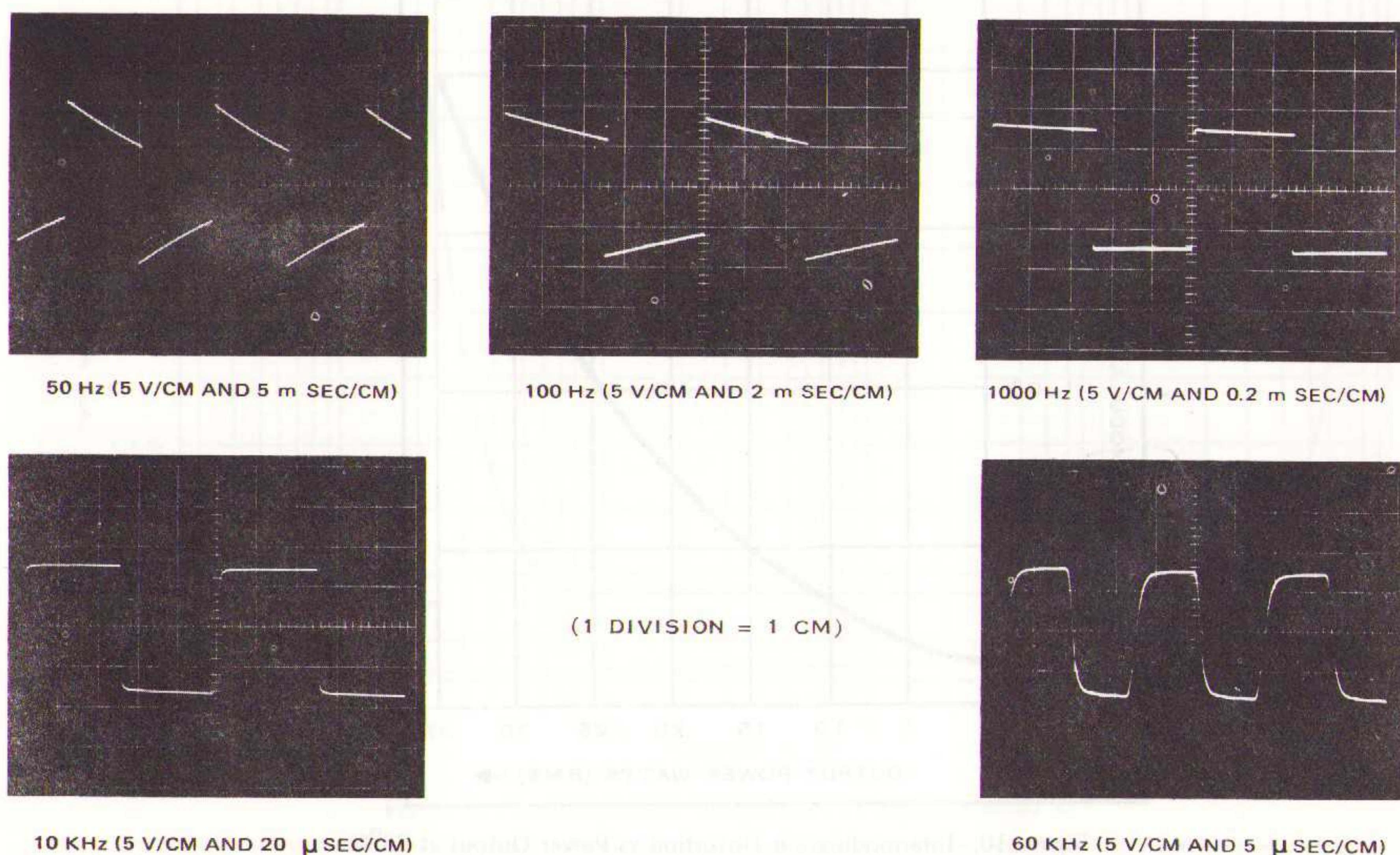


Figure 12. Output Waveforms for Square Wave Input 3.5 Watts (rms) Output

SUMMARY

After taking into consideration the many compromises, the amplifier presented accomplishes the purposes intended. With 35 watts of RMS power, the distortion is below 1.0% from 37 Hz to 54 KHz, the frequency response is 7.4 Hz to 91 KHz, and the sensitivity is 940 mv. By using plastic transistors and the minimum number of components, the cost has been kept low. The performance is such that the distortion is low enough for normal pleasant listening, and has a sensitivity which is compatible with typical preamplifiers. The additional feature of short circuit protection insures that out of the ordinary large currents will not destroy the output devices.

The epitaxial base process, allowing manufacturers to produce NPN and PNP silicon complements with relative ease, along with plastic packaging, makes available a transistor capable of high power dissipation that is in the reasonable price range. As a result, silicon complementary-symmetry audio power amplifiers are feasible from a cost and performance standpoint.

ACKNOWLEDGEMENT

The author wishes to express his thanks to Mr. Rufus Poteet for building and testing the amplifier and to Mr. Max Schreiner for his advice and assistance in the writing of this paper.

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LOW-COST SOLID-STATE AUDIO AMPLIFIERS

Samuel W. Webster, Jr.

I. INTRODUCTION

This report presents two applications of TI's new low cost, medium power, silicon epitaxial base plastic transistors. Since the devices are available as both NPN and PNP devices, and as matched complementary pairs, they lend themselves very readily to applications requiring complementary-symmetry outputs, direct coupling between stages, and to automotive usages.

The circuits described in this note are a 15 watt (rms) complementary-symmetry audio power amplifier which is direct-coupled throughout, and a 3 watt (rms) class A automotive audio power amplifier which uses a PNP output device.

Many compromises become evident in amplifier design; therefore, the circuits presented are not intended to be the highest quality possible. Instead, the designs have been made around a specific device, with the emphasis placed on obtaining a low cost amplifier giving the maximum amount of output power possible from the device, while using the minimum number of devices, and holding the distortion to a reasonable level and keeping the sensitivity compatible with typical preamplifier requirements. Higher performance can be obtained with more components or lower power. A greater power output is possible if a sacrifice in distortion or sensitivity can be tolerated.

II. 15 WATT (RMS) COMPLEMENTARY-SYMMETRY AUDIO POWER AMPLIFIER

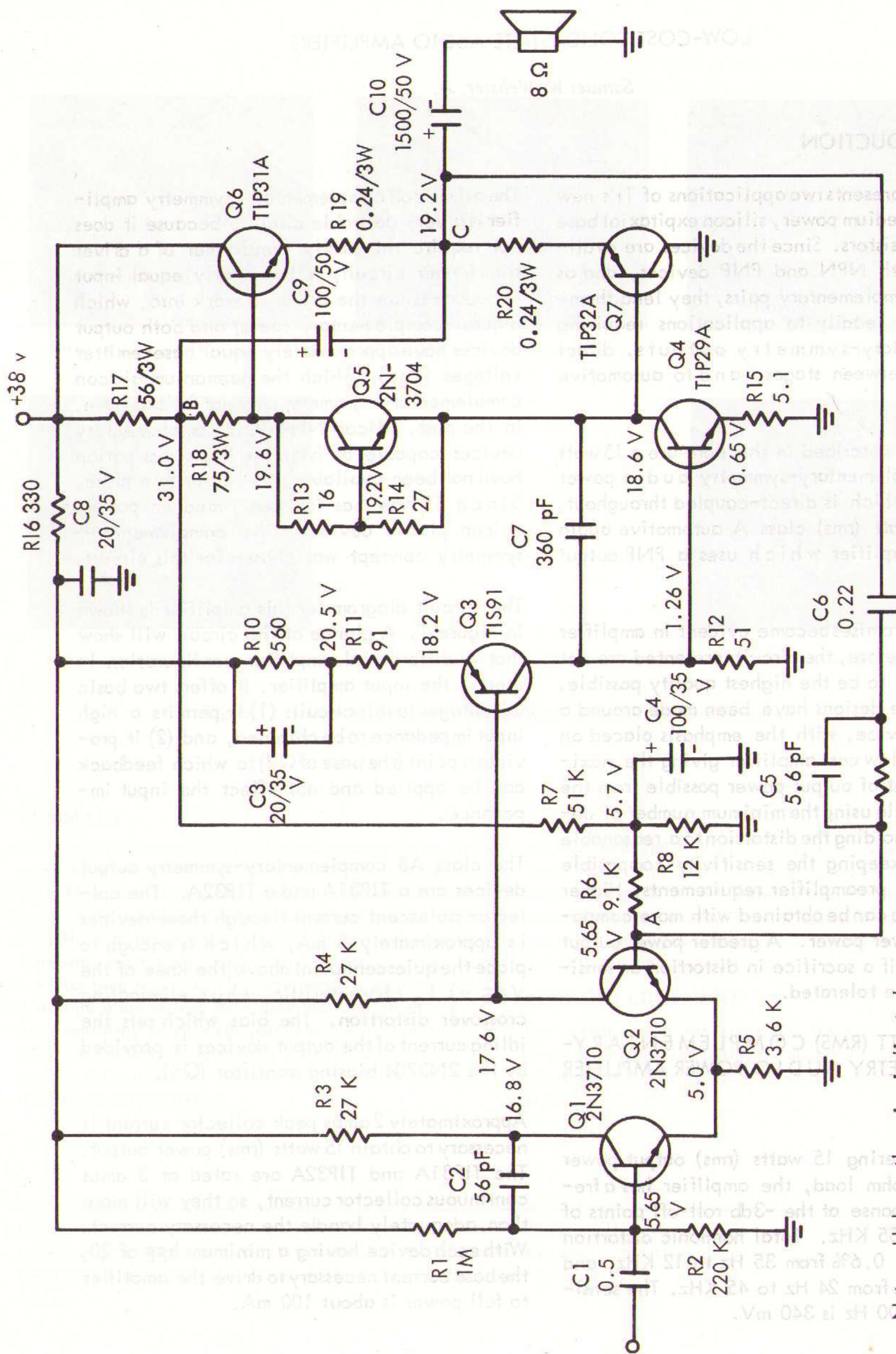
When delivering 15 watts (rms) output power into an 8 ohm load, the amplifier has a frequency response at the -3db roll-off points of 12.5 Hz to 55 KHz. Total harmonic distortion is less than 0.6% from 35 Hz to 12 KHz, and less than 1% from 24 Hz to 45 KHz. The sensitivity at 1000 Hz is 340 mV.

The all silicon complementary-symmetry amplifier is a very desirable circuit, because it does not require the costly transformer of a driver transformer circuit; it has nearly equal input impedances for the driver to work into, which a quasi-complementary lacks; and both output devices have approximately equal base-emitter voltages (V_{BE}), which the germanium-silicon complementary-symmetry concept fails to have. In the past, silicon NPN-PNP complementary devices capable of handling high dissipation have not been available at a reasonable price. Since TI now has low cost, medium power, silicon plastic devices, the complementary-symmetry concept was chosen for this circuit.

The circuit diagram for this amplifier is shown in Figure 1. A glance at the circuit will show that a differential amplifier configuration is used as the input amplifier. It offers two basic advantages to this circuit: (1) it permits a high input impedance to be obtained, and (2) it provides a point (the base of Q2) to which feedback can be applied and not affect the input impedance.

The class AB complementary-symmetry output devices are a TIP31A and a TIP32A. The collector quiescent current through these devices is approximately 5 mA, which is enough to place the quiescent point above the knee of the V_{BE} vs. I_C characteristic, thus eliminating crossover distortion. The bias which sets the idling current of the output devices is provided by the 2N3704 biasing transistor (Q5).

Approximately 2 amps peak collector current is necessary to obtain 15 watts (rms) power output. The TIP31A and TIP32A are rated at 3 amps continuous collector current, so they will more than adequately handle the necessary current. With each device having a minimum h_{FE} of 20, the base current necessary to drive the amplifier to full power is about 100 mA.



Note: Unless otherwise indicated resistor and capacitor values are in ohms and microfarads.

FIGURE 1.

15-Watt (RMS) Audio Amplifier

Resistors R19 and R20 in the output devices collector circuit provide some degenerative feedback which helps to improve the distortion due to non-linear h_{FE} .

Since the output devices are class AB, Q6 is "on" when Q7 is "off", and vice versa. The turning "off" and "on" is accomplished by the class A driver, Q4, reverse biasing and then forward biasing the output devices. Reverse biasing the output devices turns them off in a very short time, rapidly removing the stored base charge, permitting good high frequency operation, keeping the efficiency high, and giving low values of current drain, power dissipation, and operating temperature.

The driver is a TIP29A (Q4). Being in class A operation, its quiescent current should be at least equal to the base drive required to drive the output devices to full power, and preferably, it should be a little greater. With the requirement for a base drive of 100 mA, the quiescent current through the driver has been chosen to be near 128 mA in order to prevent clipping. The TIP29A is an excellent driver device for this circuit because it can handle high power dissipation and high collector current. The maximum current through the driver is around 225 mA, which could be handled by many small signal devices, but with the collector to emitter voltage swing being nearly 18 volts, the power dissipation of such devices would easily be exceeded. The TIP29A can easily handle both the current and power requirements. With a minimum h_{FE} of 20, the base drive current requirement would be about 5 mA.

The 2N3704 biasing transistor is used instead of biasing diodes, very simply, because it allows tighter control over the value of the output device collector quiescent current, and the cost of the transistor and its biasing resistors is less than two or more diodes. Silicon diode junctions and silicon transistor base-emitter junctions have a voltage of approximately 0.7 volts.

Referring to Figure 2, if V_{BE1} and V_{BE2} are both 0.7 volts and the voltage across RR2 and RR3 are negligible, and voltage across both diodes, V_{D1} and V_{D2} is 0.7 volts, then diodes could readily be used to bias the output devices, because the voltage along path AB and path ACB would be equal. However, normally, the junction voltages are not exactly 0.7 volts, and the voltage across RR2 and RR3 is not necessarily negligible. Therefore, it could become quite difficult and costly to screen for diodes which would have a combined junction voltage such that the voltage along path AB would equal that along path ACB. If the voltage across path AB is greater than required by path ACB, then T1 and T2 will be biased on too much allowing more collector quiescent current than necessary through them causing excessive dissipation. If voltage AB is less than required along path ACB, too little quiescent current will flow, causing crossover distortion. The biasing transistor used in the circuit can be biased so that its V_{CE} provides the proper bias to the output devices, in order to allow control over the output collector quiescent current.

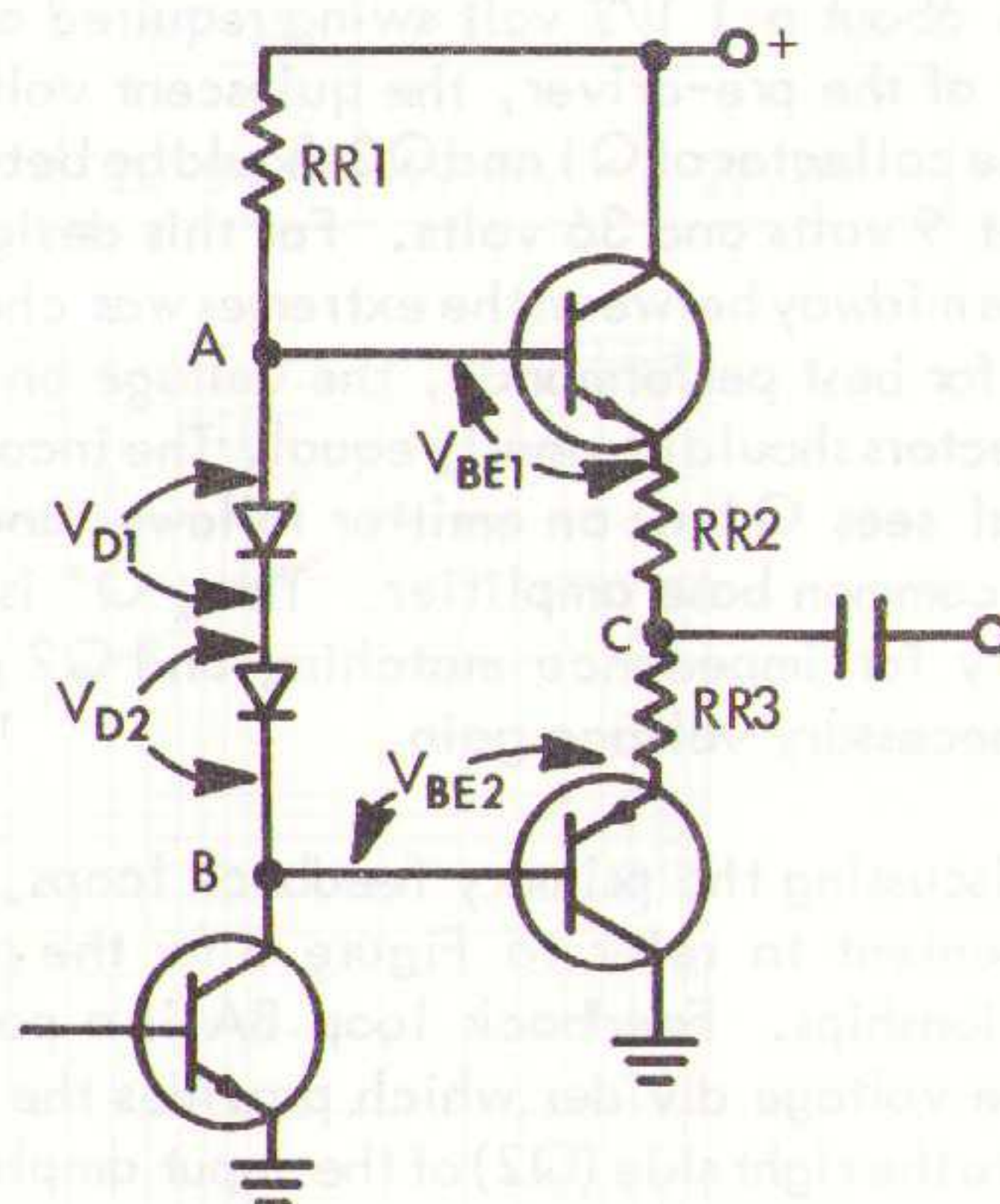


FIGURE 2. Example of Diode Biasing

ATIS91 PNP small signal device is used in the class A pre-driver slot. The collector quiescent current through this device is 25 mA and is more than adequate to provide the 5 mA base drive to the TIP29A driver. Direct coupling of the stages necessitated a PNP in this slot, to make the base voltage of the pre-driver or collector voltage of the input amplifier relatively high in order to prevent voltage clipping at this point.

The input amplifier is a differential amplifier configuration, and was used in this design for the two primary reasons mentioned earlier. The terminology "differential amplifier configuration" has been used, because the amplifier is not being used as a true differential amplifier. Certain features of the differential amplifier are being used, but the amplification is not of the differential of the voltages which appear on the two collectors. The input amplifier consists of two 2N3710's with the emitters tied together. The emitters are tied to a constant current source, which could be an active device, but in this design is R5, a 3.6 Kohm resistor.

With about a 1 1/2 volt swing required at the base of the pre-driver, the quiescent voltages on the collector of Q1 and Q2 should be between about 9 volts and 36 volts. For this design, a value midway between the extremes was chosen. And for best performance, the voltage on both collectors should be nearly equal. The incoming signal sees Q1 as an emitter follower and Q2 as a common base amplifier. Thus, Q1 is primarily for impedance matching and Q2 gives the necessary voltage gain.

In discussing the primary feedback loops, it is convenient to refer to Figure 3 for the phase relationships. Feedback loop BA is a portion of the voltage divider which provides the base bias to the right side (Q2) of the input amplifier. Capacitor C4 shorts any a-c signal to ground, so that this loop provides only d-c feedback. The purpose of this feedback is to stabilize the midpoint voltage at point C against shift tendencies due to temperature changes. It actually keeps

the 19.2-volt midpoint voltage from changing more than 0.8 volts from 25°C to 55°C. If the voltage at point C rises, that at point B will also increase; this turns Q2 on harder, thus lowering the collector voltage. By tracing this change through the circuit, it will be seen that the action lowers the midpoint voltage.

Loop CB provides about 0.5 db of positive a-c feedback through a "bootstrap" capacitor to help recover some of the gain and sensitivity lost due to the negative a-c feedback in the system without drastically affecting the distortion. Figure 3 shows that a positive signal at point C is fed back in phase with the positive signal at point B to increase the magnitude of the base signal to the output devices. Amplified by the output device, a signal of greater magnitude is obtained at point C.

36.5 db of negative a-c feedback is provided by loop CA to correct for the h_{FE} non-linearity of the circuit. By feeding back to point A, the amplifier input impedance is not reduced because of the large amount of feedback.

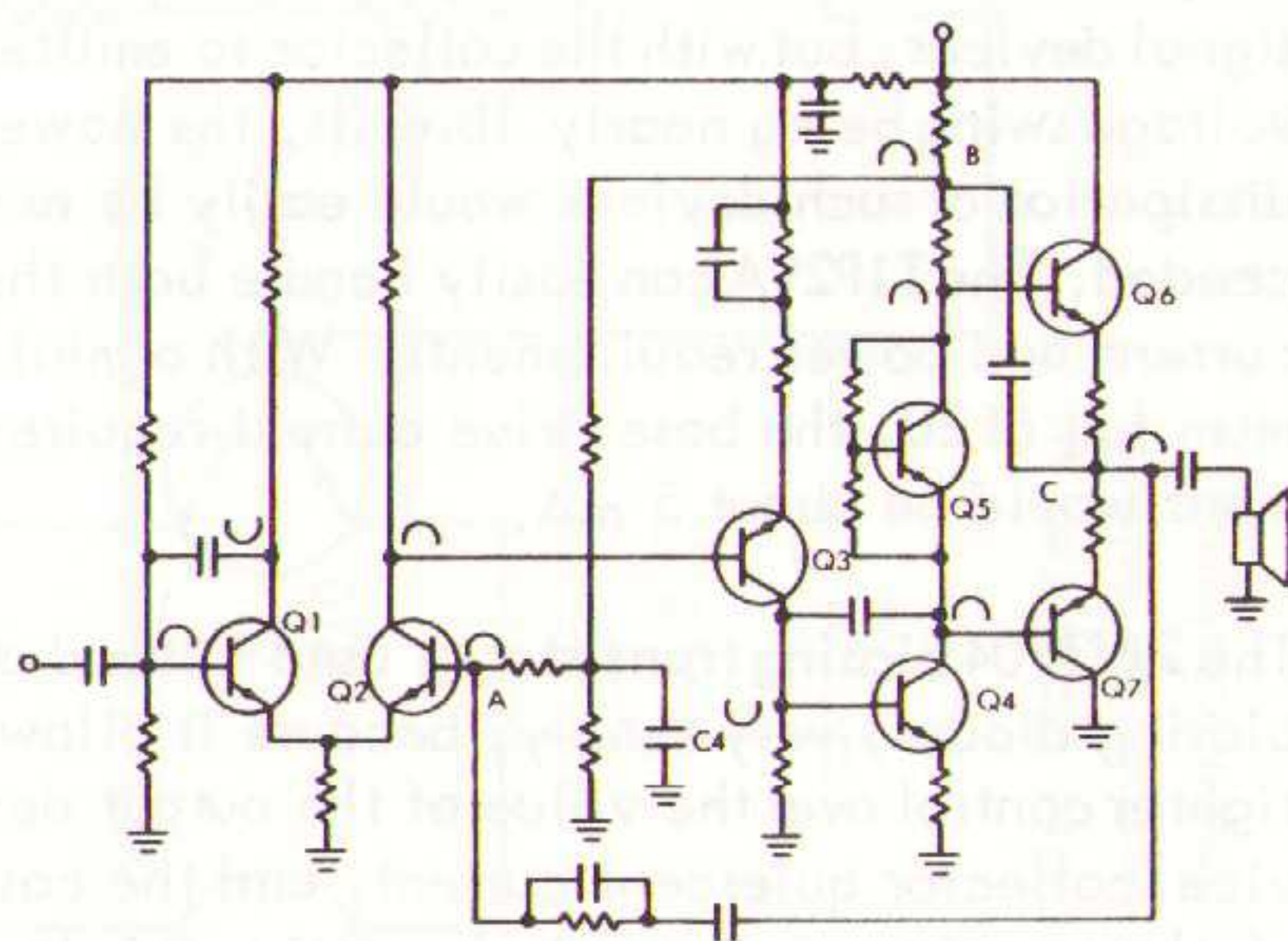


FIGURE 3. Circuit Phase Relationships for 15-Watt Amplifier

Performance Data:

Output Power	15 watts (rms) power into an 8 ohm load with a supply voltage of +38 volts.	Intermodulation Distortion (60 Hz and 7 KHz mixed at a 4:1 ratio)	(See Figure 8)
Sensitivity	340mV at full rated power at 1000 Hz. (See Figure 4)	Signal to Noise (Reference to 15 watts)	Input Open: -57 db Input Shorted: -87 db
Frequency Response	12.5 Hz to 55 KHz at -3 db roll-off from rated output power. (See Figure 5)	Input Impedance	150 Kohms
Total Harmonic Distortion	Less than 0.6% from 35 Hz to 13 KHz, and less than 1% from 24 Hz to 45 KHz at rated output power. (See Figures 6 and 7)	Damping Factor	21
		Square Wave Response	(See Figure 10)
		Power Supply Current Drain	(See Figure 9)

Current and Voltage Data:

Supply Voltage	33V	38V				43V
Ambient Temperature	25°C	25°C	35°C	45°C	55°C	25°C
Output Device Quiescent Current	2.5 mA	4.6 mA	5.4 mA	8.2 mA	11 mA	7.8 mA
Power Supply Quiescent Current Drain	124 mA	148 mA	145 mA	148 mA	152 mA	170 mA
Output Midpoint Voltage	17.6V	19.2V	20.0V	20.0V	19.8V	22.0V

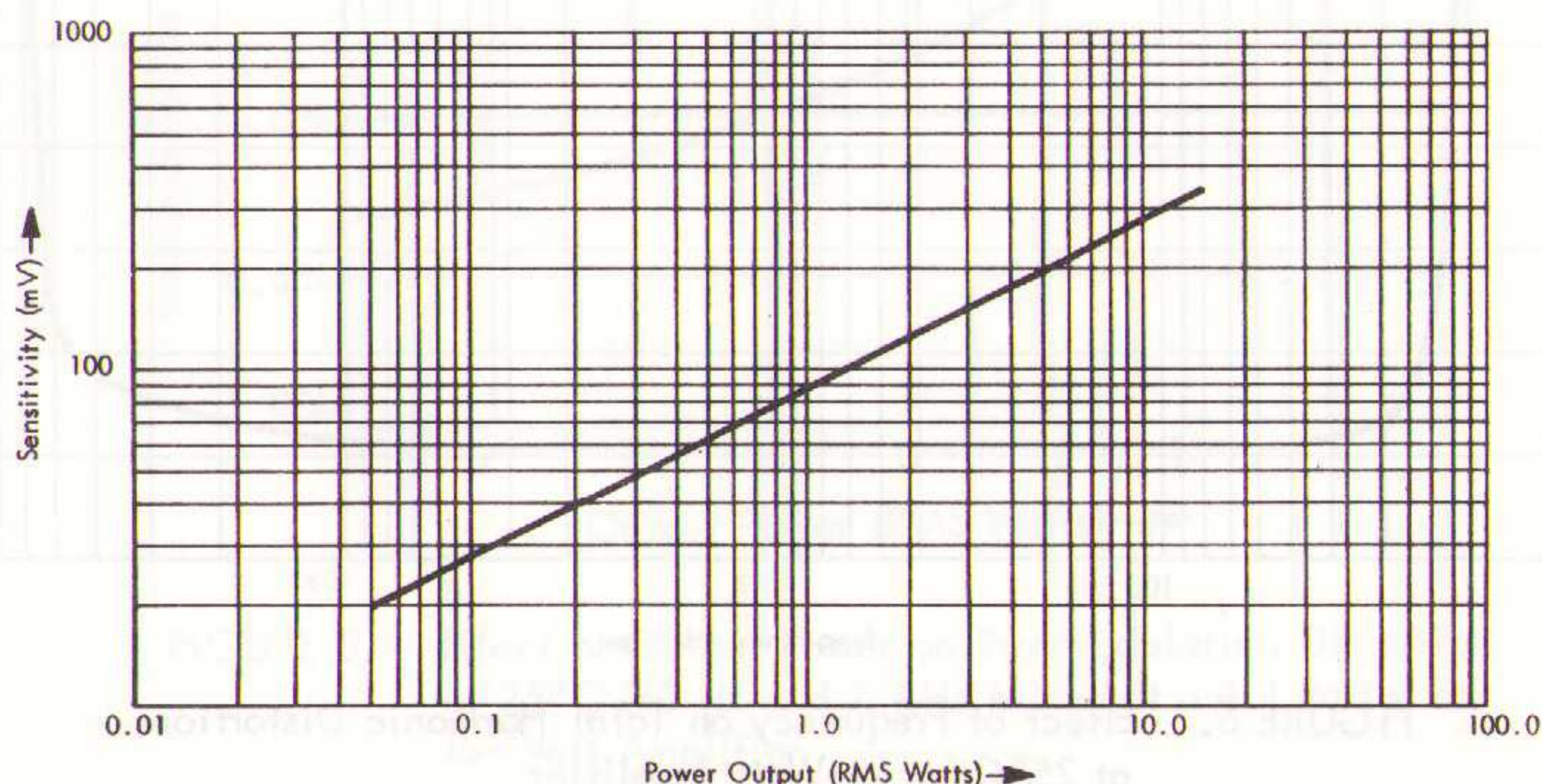


FIGURE 4. Effect of Power Output on Sensitivity at 1000 Hz and 25°C for 15-Watt Amplifier

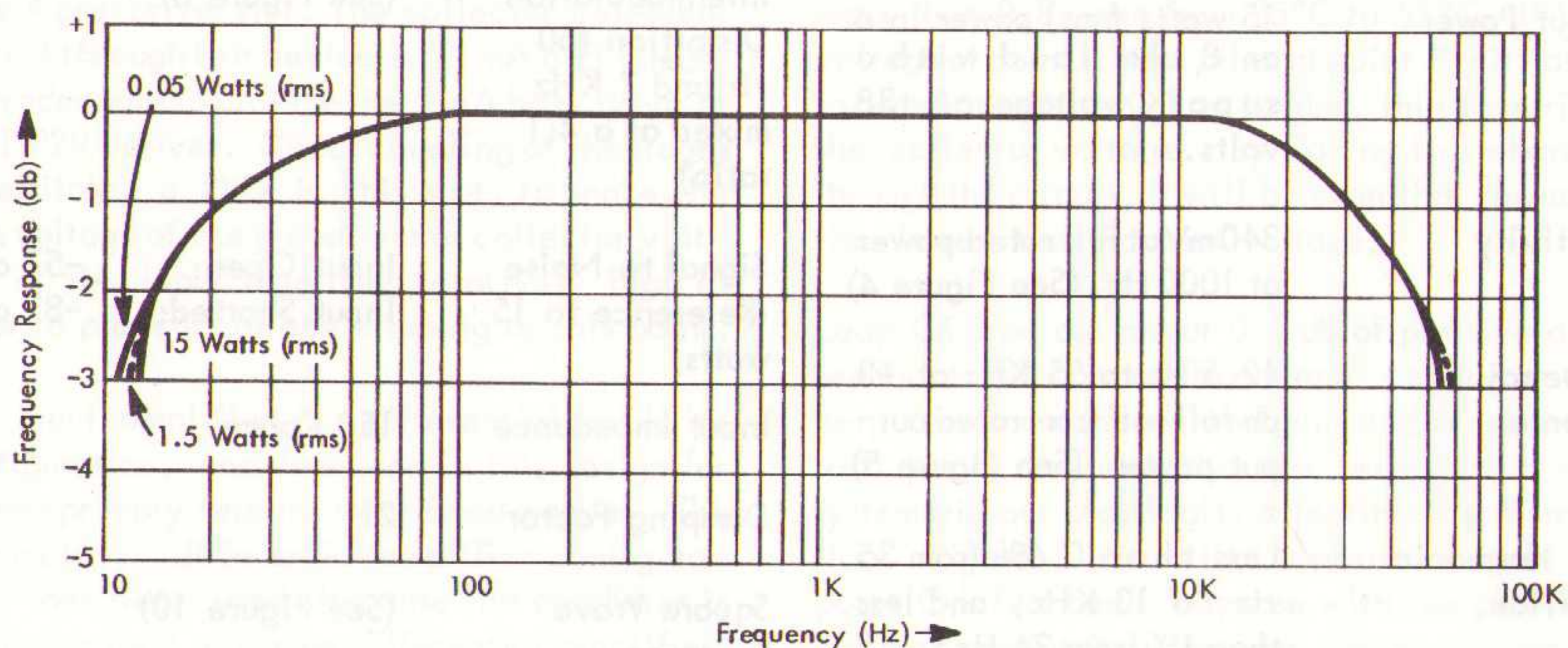


FIGURE 5. Frequency Response at 25°C for 15-Watt Amplifier

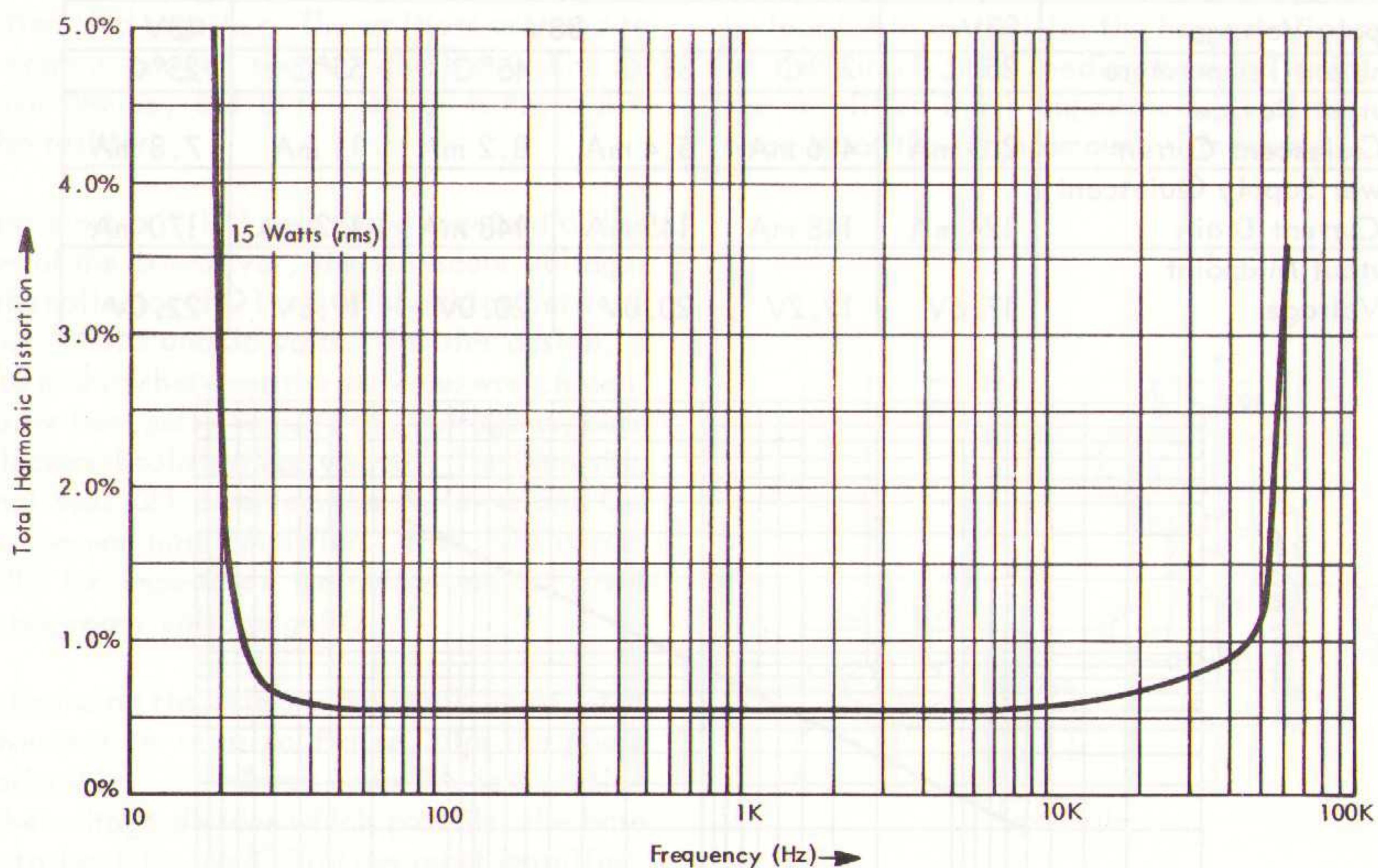


FIGURE 6. Effect of Frequency on Total Harmonic Distortion at 25°C for 15-Watt Amplifier

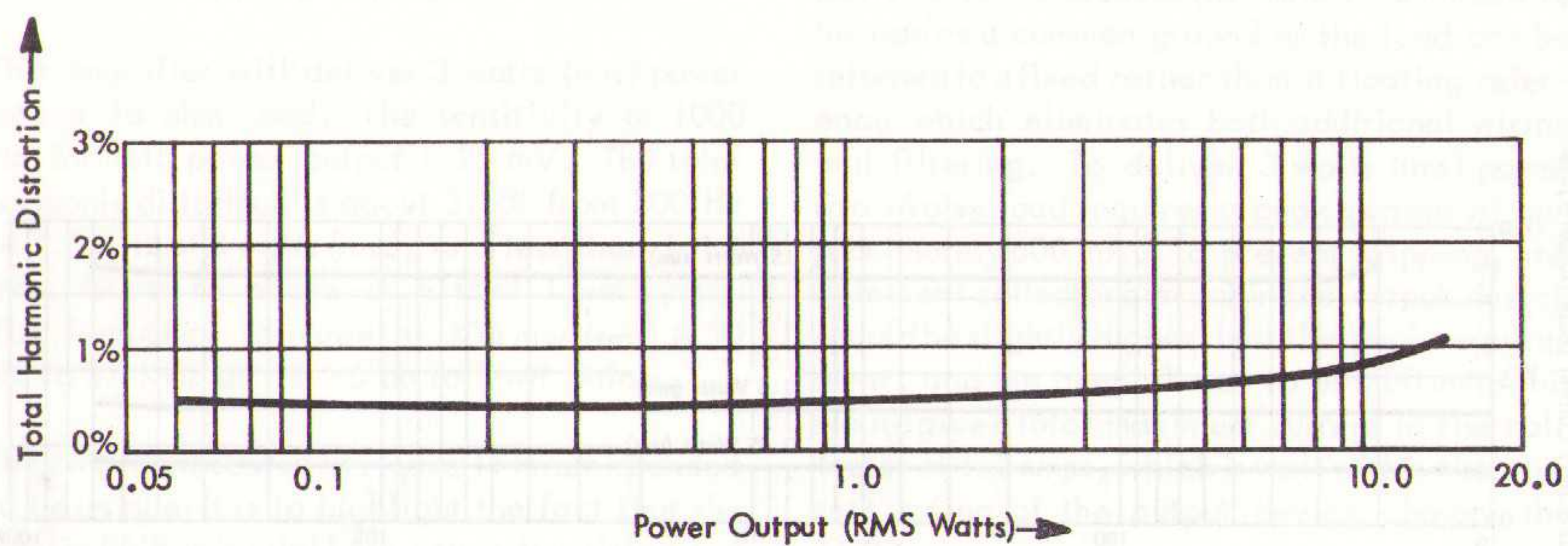


FIGURE 7. Effect of Power Output on Total Harmonic Distortion at 25°C and 1000 Hz for 15-Watt Amplifier

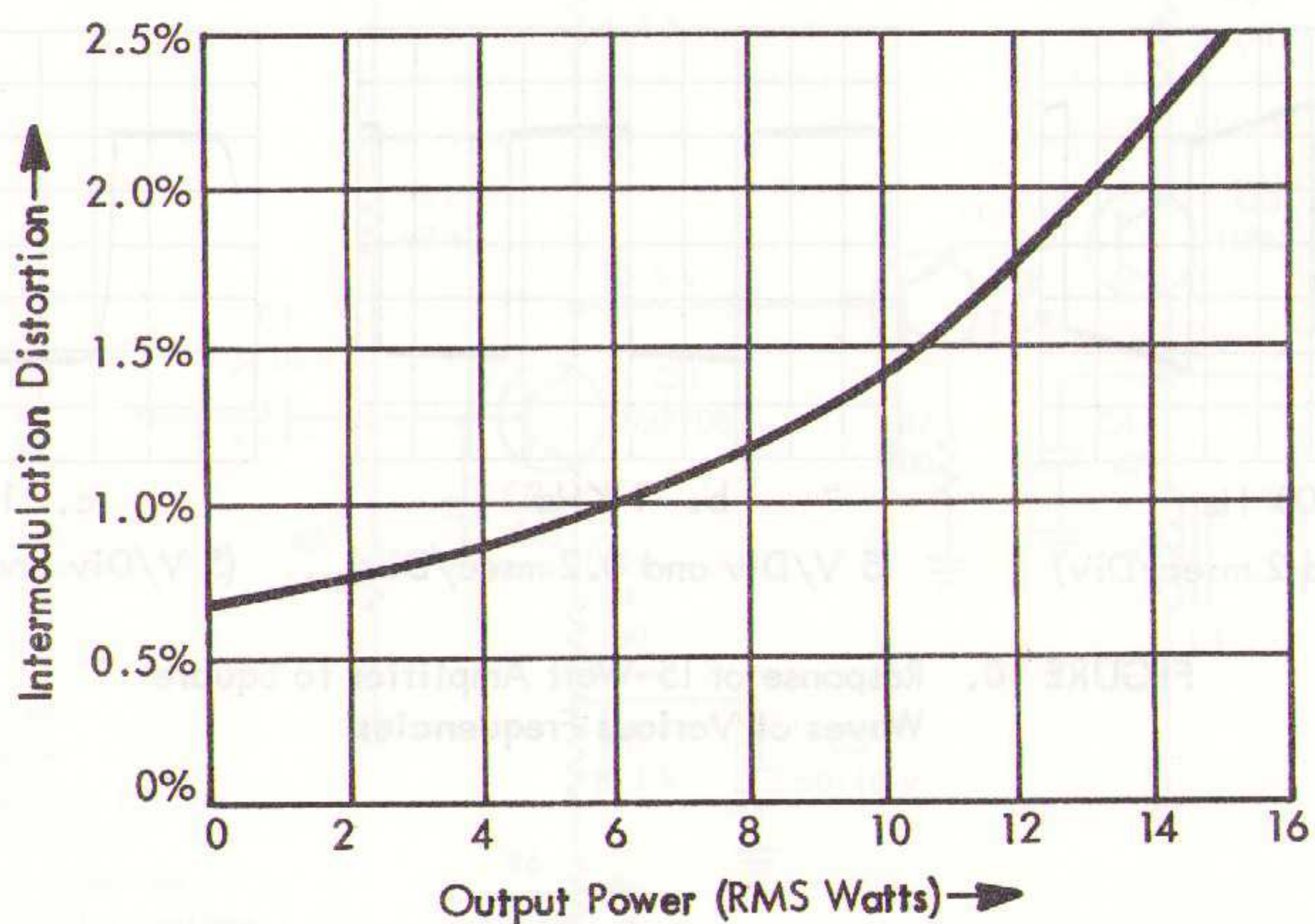


FIGURE 8. Effect of Output Power on Intermodulation Distortion at 25°C (60 Hz and 7 KHz Mixed at a 4:1 Ratio) for 15-Watt Amplifier

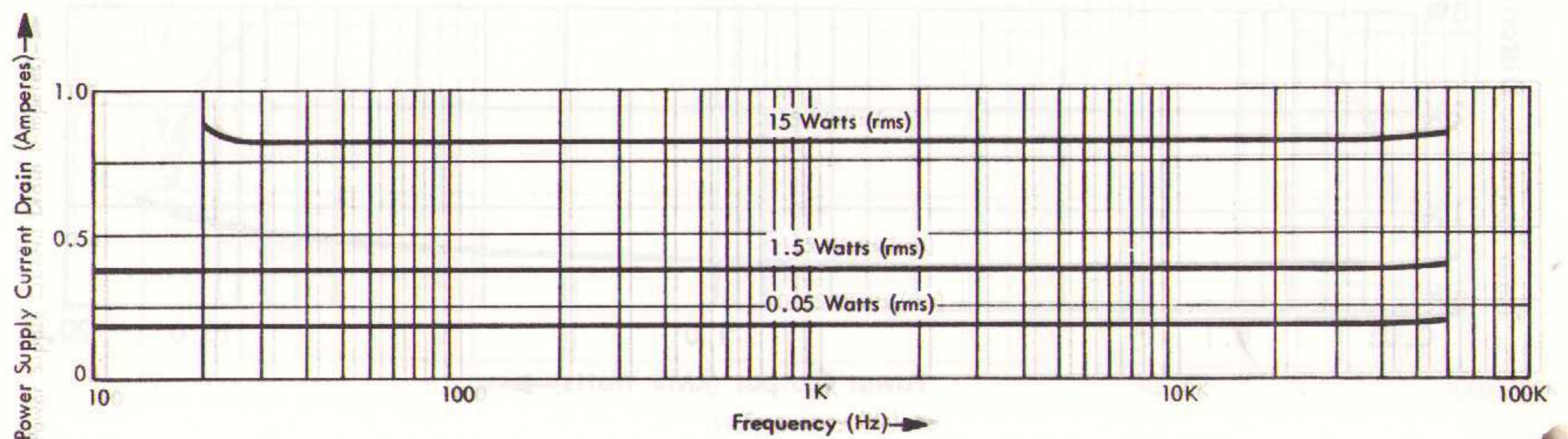
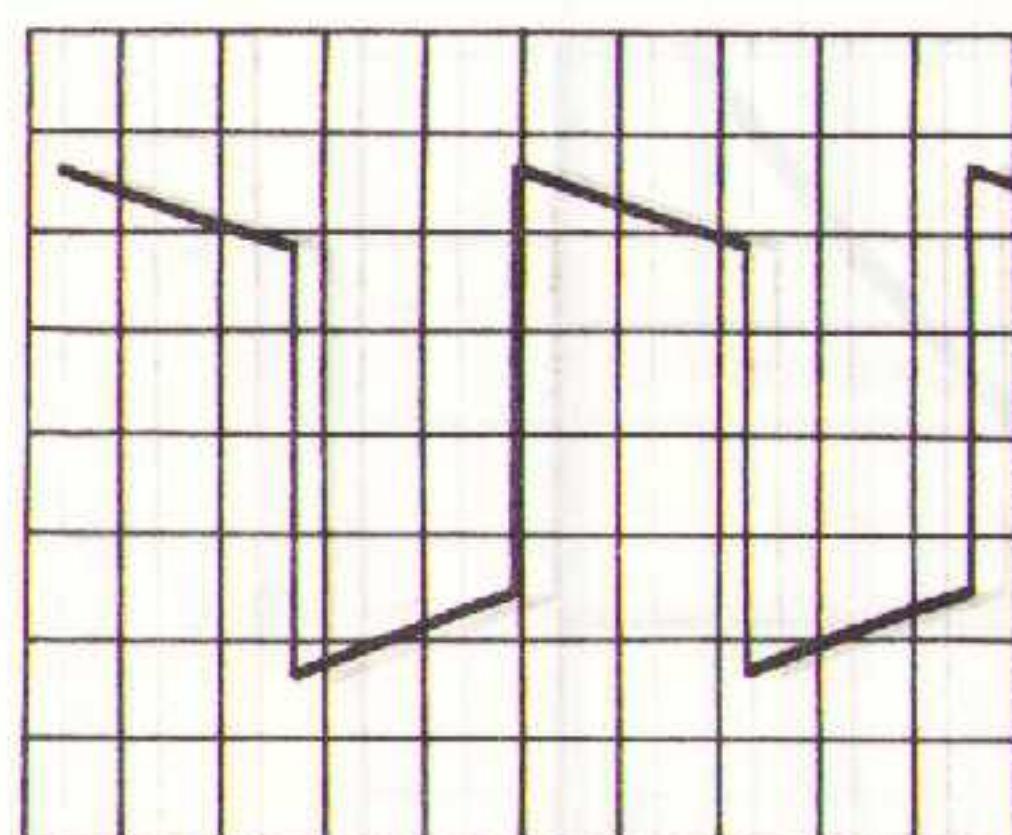
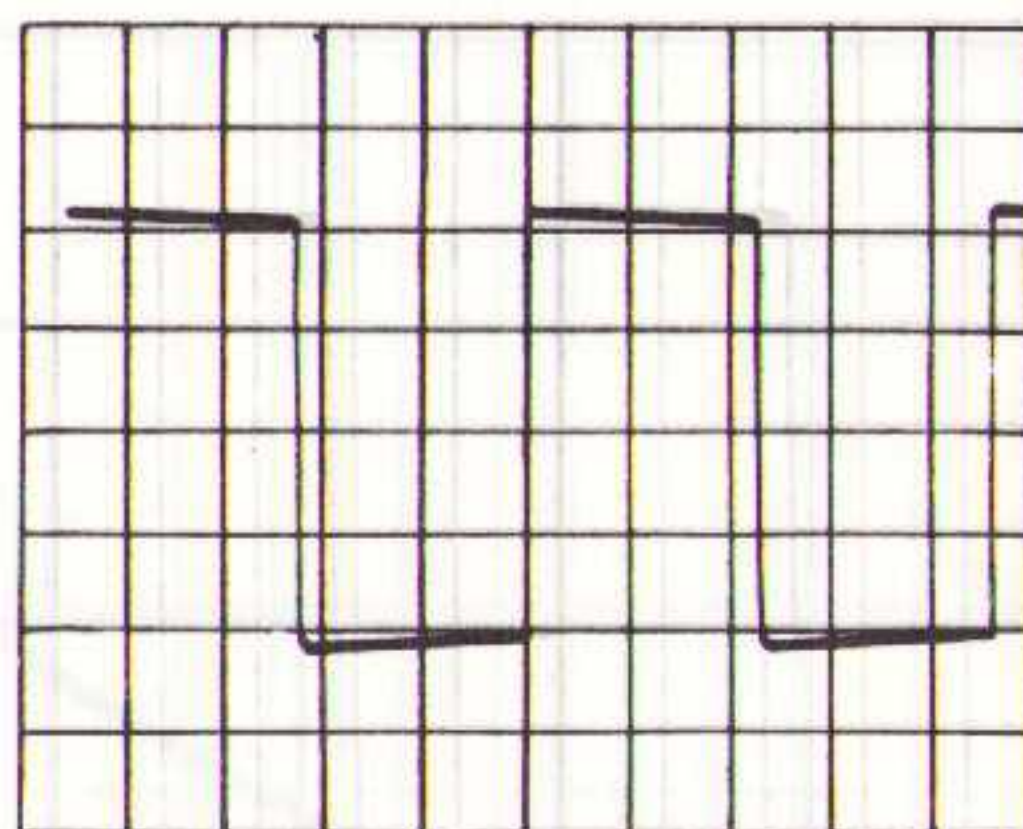


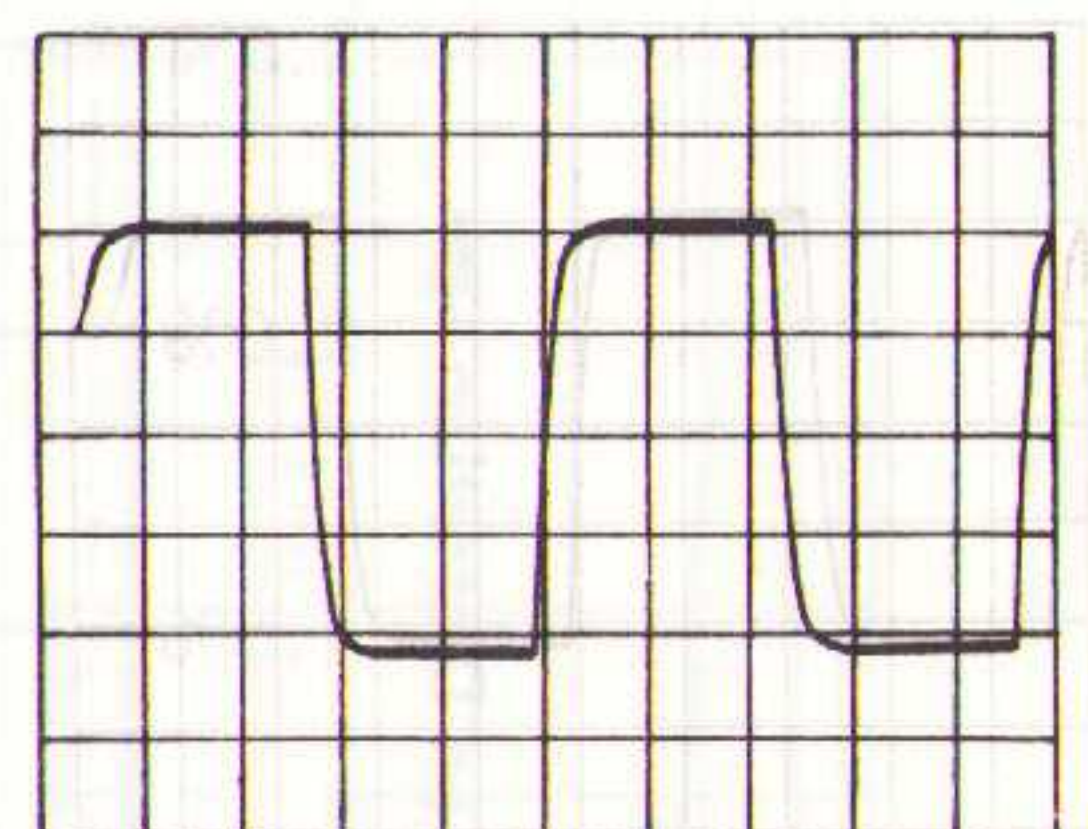
FIGURE 9. Effect of Frequency on Power Supply Current Drain at 25°C for 15-Watt Amplifier



a. 100 Hz
(5 v/Div and 2 msec/Div)



b. 1 KHz
(5 V/Div and 0.2 msec/Div)



c. 10 KHz
(5 V/Div and 20 μsec/Div)

FIGURE 10. Response of 15-Watt Amplifier to Square Waves of Various Frequencies

III. 3 WATT (RMS) CLASS A AUTOMOTIVE AUDIO POWER AMPLIFIER

This amplifier will deliver 3 watts (rms) power into a 16 ohm load. The sensitivity at 1000 Hz for full power output is 25 mV. The total harmonic distortion is about 3.5% from 200 Hz to 7.5 KHz at 3 watts (rms), and less than 1.1% from 70 Hz to above 20 KHz at 1 watt (rms). The frequency response at 300 mw (rms) is 33 Hz to 37 KHz at the -6 db roll-off points.

The circuit presented in Figure 11 is not intended to be unique; it is to highlight the fact that the plastic PNP epitaxial base power transistors are readily adaptable for amplifiers of the type used in automotive radios.

PNP transistors are normally preferred in automotive circuits because they allow the chassis to be used as a common ground so the load can be returned to a fixed rather than a floating reference which eliminates both additional wiring and filtering. To deliver 3 watts (rms) power to a 16 ohm load requires a peak current of approximately 600 mA. To prevent clipping, the quiescent collector current in the output device should be slightly higher than the peak requirement, and has been chosen to be 700 mA. This would give a total maximum current in the collector of 1.3 amps, which is well within the current rating of the output device chosen, the TIP32.

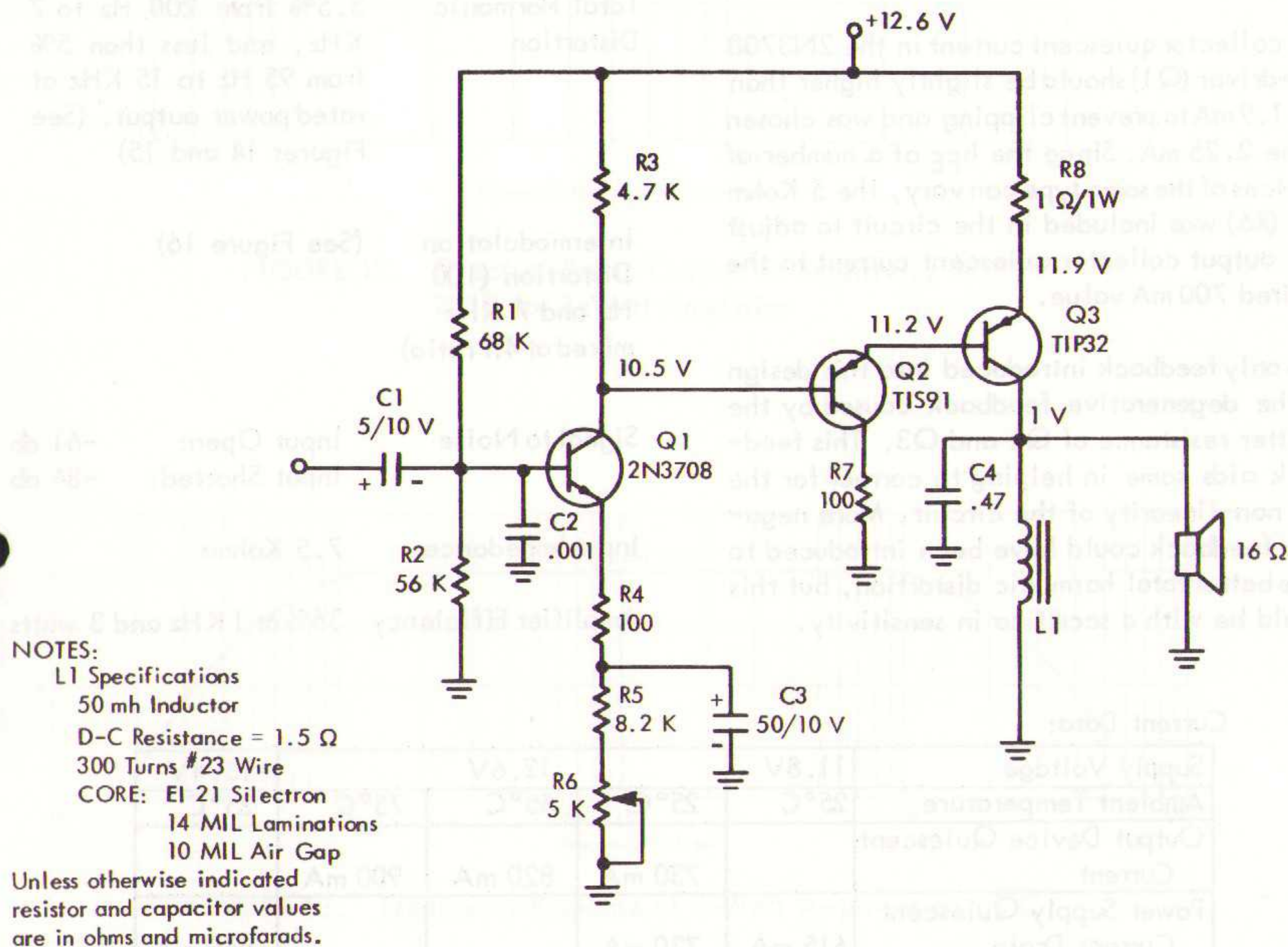


FIGURE 11. 3-Watt (RMS) Automotive Audio Amplifier

From the $\frac{V_{cep} I_{cp}}{2} = PO(rms)$ relationship, it becomes evident that V_{cep} would equal 10 volts, or a peak to peak voltage of 20 volts. The V_{CC} of 12.6 volts is an adequate supply voltage, because of the inductive load, L1. The V_{cep} of a class A amplifier with an inductive load is approximated by V_{CC} less $V_{CE(SAT)}$ and a small voltage to compensate for the V_{BE} change due to changing temperature.

A PNP, TIS91 "Silect" transistor has been chosen for the driver since it allows the base voltage and, consequently, the pre-driver collector voltage to be high. The TIP32's h_{FE} of 20 along with the h_{FE} of 160 for the TIS91 combine to reduce the current swing required at the base of the driver (Q2) to about 1.9 mA.

The collector quiescent current in the 2N3708 pre-driver (Q1) should be slightly higher than the 1.9 mA to prevent clipping and was chosen to be 2.25 mA. Since the h_{FE} of a number of devices of the same type can vary, the 5 Kohm pot (R6) was included in the circuit to adjust the output collector quiescent current to the desired 700 mA value.

The only feedback introduced into this design is the degenerative feedback caused by the emitter resistance of Q1 and Q3. This feedback aids some in helping to correct for the h_{FE} non-linearity of the circuit. More negative feedback could have been introduced to give better total harmonic distortion, but this would be with a sacrifice in sensitivity.

Performance Data:

Output Power 3 watts (rms) power into a 16 ohm load with a supply voltage of 12.6 volts.

Sensitivity 25 mv at full rated power at 1000 Hz. (See Figure 12)

Frequency Response 33 Hz to 37 KHz at -6 db roll-off from rated output power. (See Figure 13)

Total Harmonic Distortion 3.5% from 200 Hz to 7 KHz, and less than 5% from 95 Hz to 15 KHz at rated power output. (See Figures 14 and 15)

Intermodulation Distortion (100 Hz and 7 KHz mixed at 4:1 ratio) (See Figure 16)

Signal to Noise Input Open: -61 db
Input Shorted: -84 db

Input Impedance 7.5 Kohms

Amplifier Efficiency 36% at 1 KHz and 3 watts

Current Data:

Supply Voltage	11.8V	12.6V			13.4V
Ambient Temperature	25°C	25°C	55°C	75°C	25°C
Output Device Quiescent Current		730 mA	820 mA	900 mA	
Power Supply Quiescent Current Drain	615 mA	730 mA			840 mA

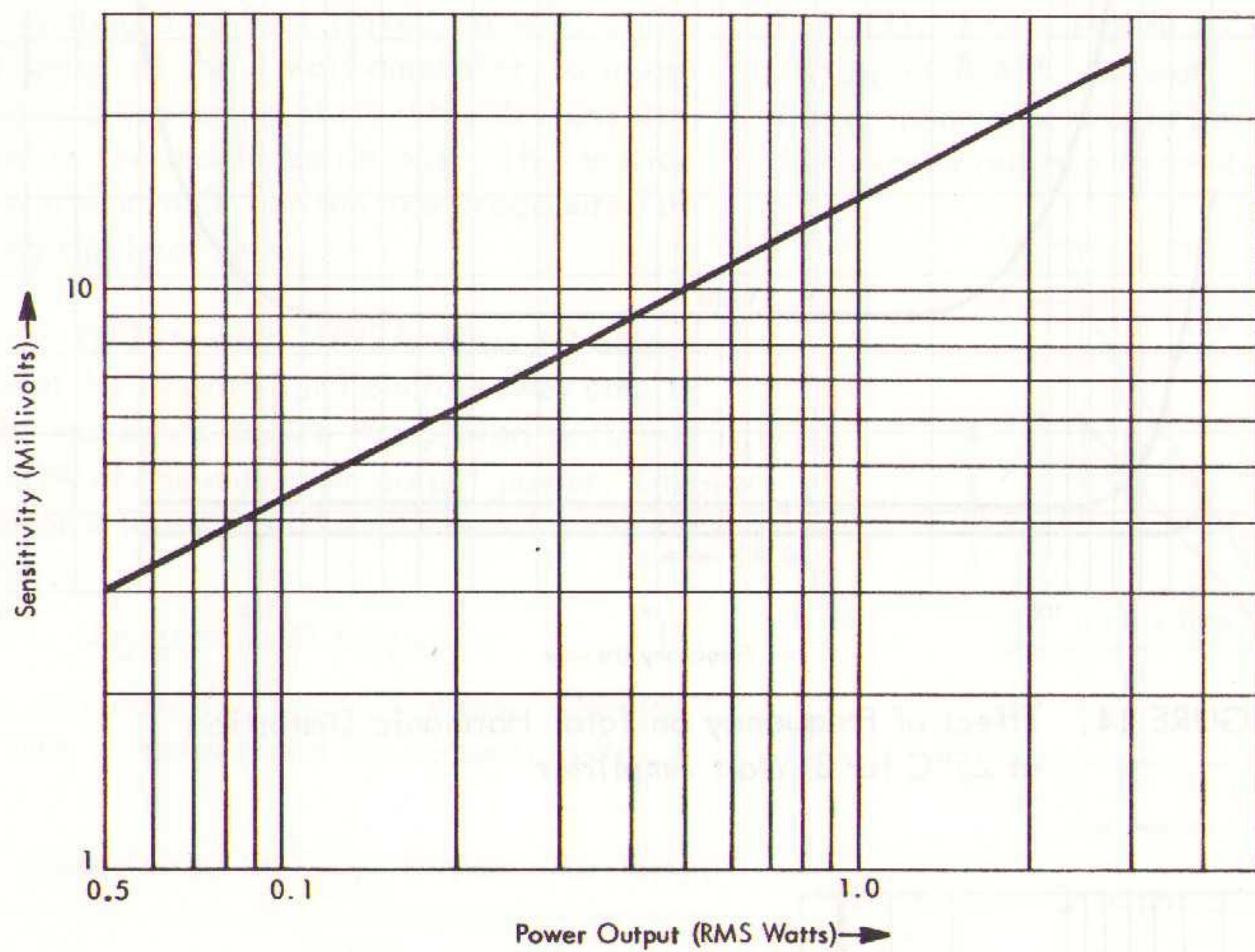


FIGURE 12. Effect of Power Output on Sensitivity at 25°C for 3-Watt Amplifier

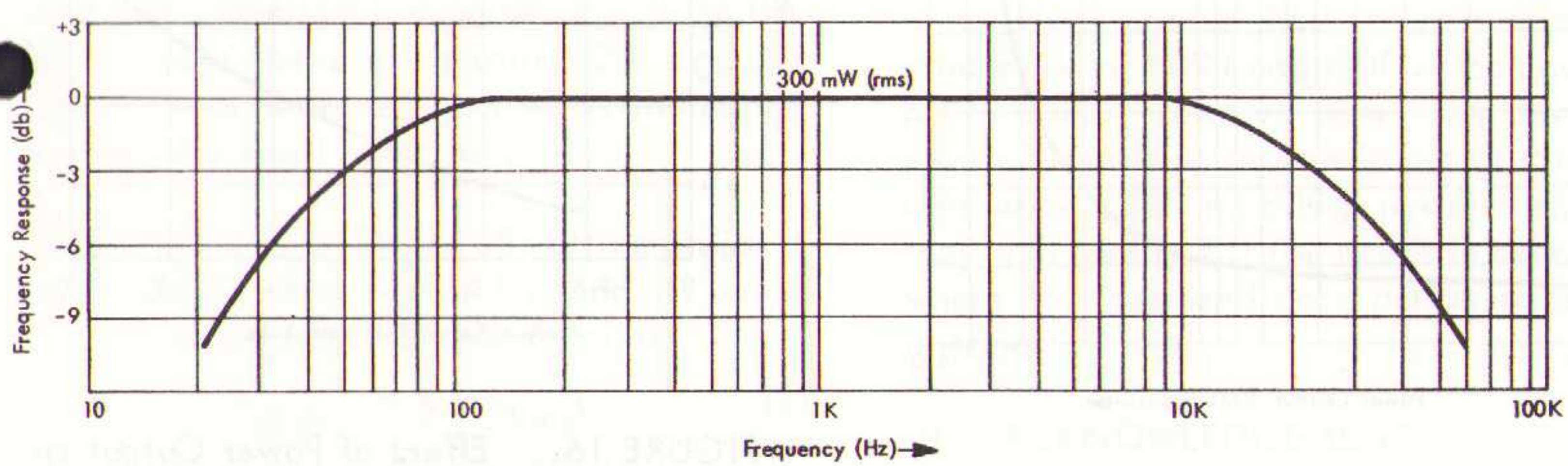


FIGURE 13. Frequency Response of 3-Watt Amplifier at 25°C

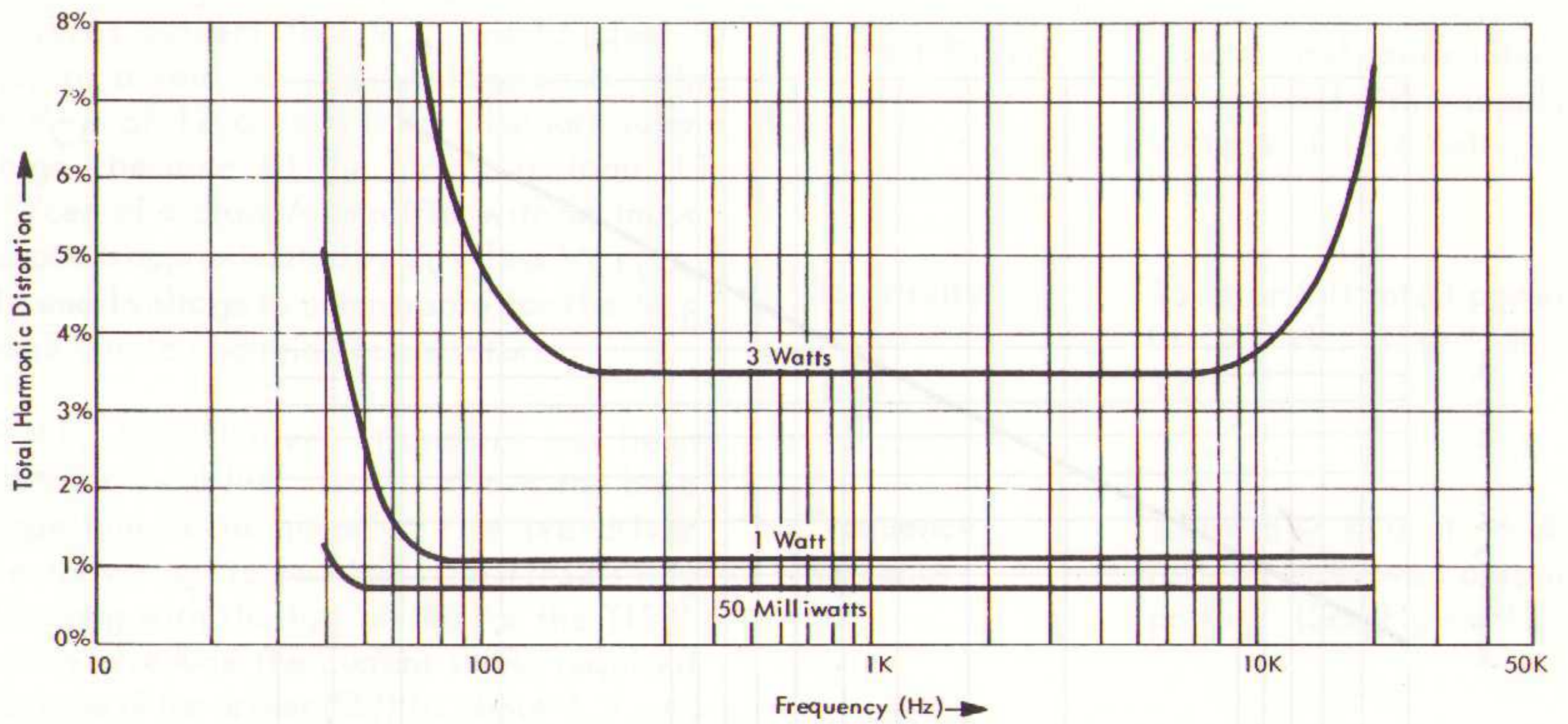


FIGURE 14. Effect of Frequency on Total Harmonic Distortion at 25°C for 3-Watt Amplifier

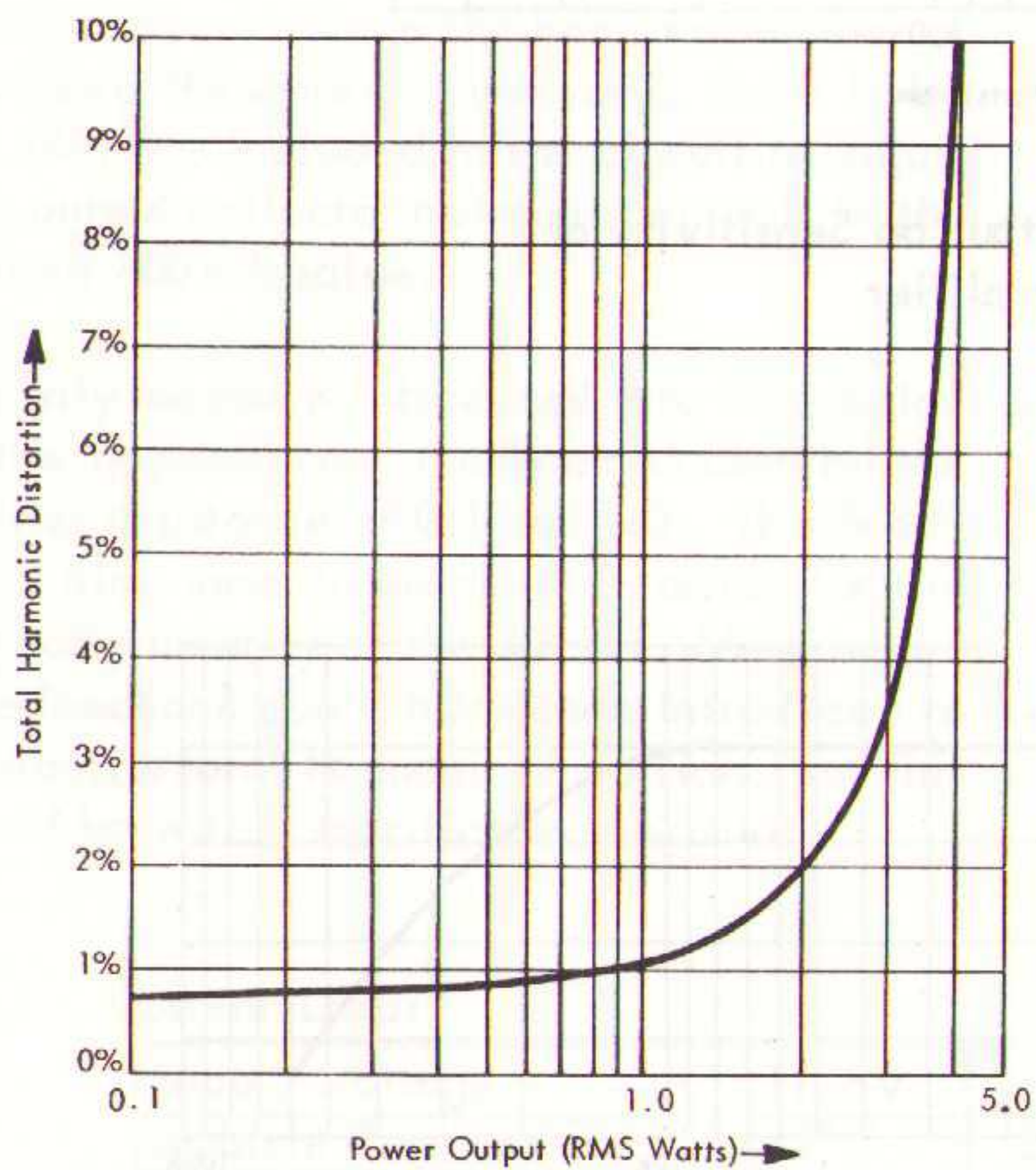


FIGURE 15. Effect of Power Output on Total Harmonic Distortion at 25°C and 1000 Hz for 3-Watt Amplifier

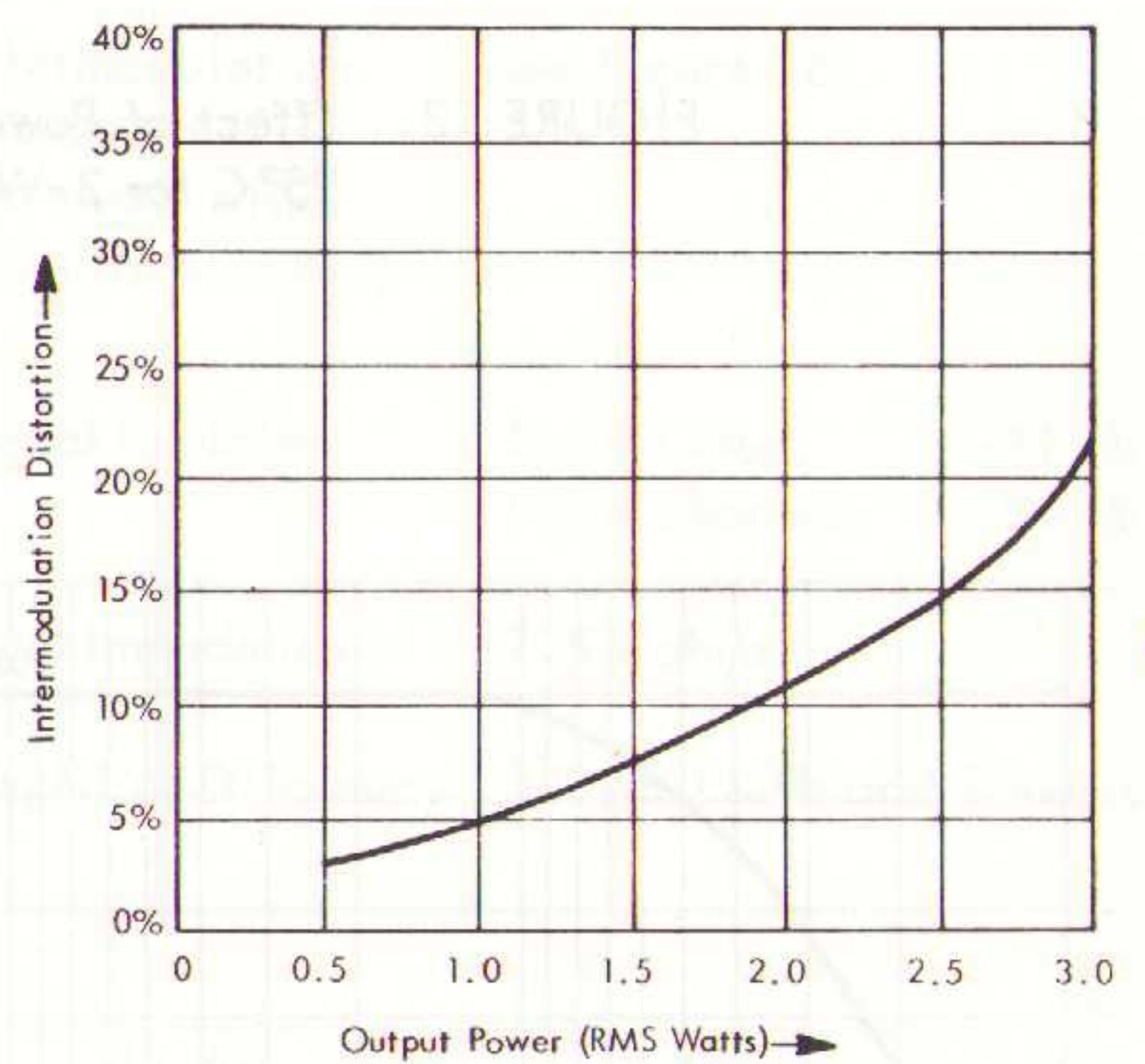


FIGURE 16. Effect of Power Output on Intermodulation Distortion at 25°C for 3-Watt Amplifier (100 Hz and 7 KHz Mixed at a 4:1 Ratio)

IV. HEAT SINK INFORMATION

The drivers and output devices of the 15 watt amplifier should be heat sinked, as well as the output device of the 3 watt amplifier, to insure that the junction temperature of the devices stay well below the maximum ratings. This section of the note will outline the procedure for choosing the heat sinks.

For the TIP31A and TIP32A class AB output devices of the 15 watt (rms) audio power amplifier, the maximum device dissipation occurs at about 40% of the maximum output power. This dissipation is found from the equation for class B operation:

$$P_{DISS} = 0.25 P_{O(rms)} \quad (1)$$

and is calculated to be 3.75 watts. The heat sink size is determined by computing θ_{C-A} from:

$$\theta_{C-A} = \frac{\text{Case Temp. at } P_{DISS} - \text{Max. Air Temp.}}{P_{DISS}} \quad (2)$$

and referring to heat sink manufacturers tables, which give heat sink size for a specific θ_{C-A} . The power derating curve of Figure 17 shows that for 3.75 watts dissipation, the device case temperature can safely go to 135°C. The amplifier has been designed to operate to at least 55°C ambient temperature. Using the three values above in equation (2), θ_{C-A} is found to be about 21.5°C per watt, indicating a relatively small heat sink.

The TIP29A driver for the 15 watt amplifier is operating in class A fashion, and its power dissipation is found from the formula:

$$P_{DISS} = 2.5 P_{O(rms)} \quad (3)$$

The rms power delivered by the driver is approximately 900 mW, therefore, the dissipation in the TIP29A is 2.3 watts. Using equation (2) and the power derating curve (Figure 17), θ_{C-A} is found to be 31.5°C per watt.

The 3 watt (rms) class A audio automotive amplifier has a power dissipation in its TIP32 output device of 7.5 watts, found from equation (3). From Figure 17 and equation (2), θ_{C-A} is 8.4°C per watt. Consequently, a larger heat sink would be required than was in the previous two examples.

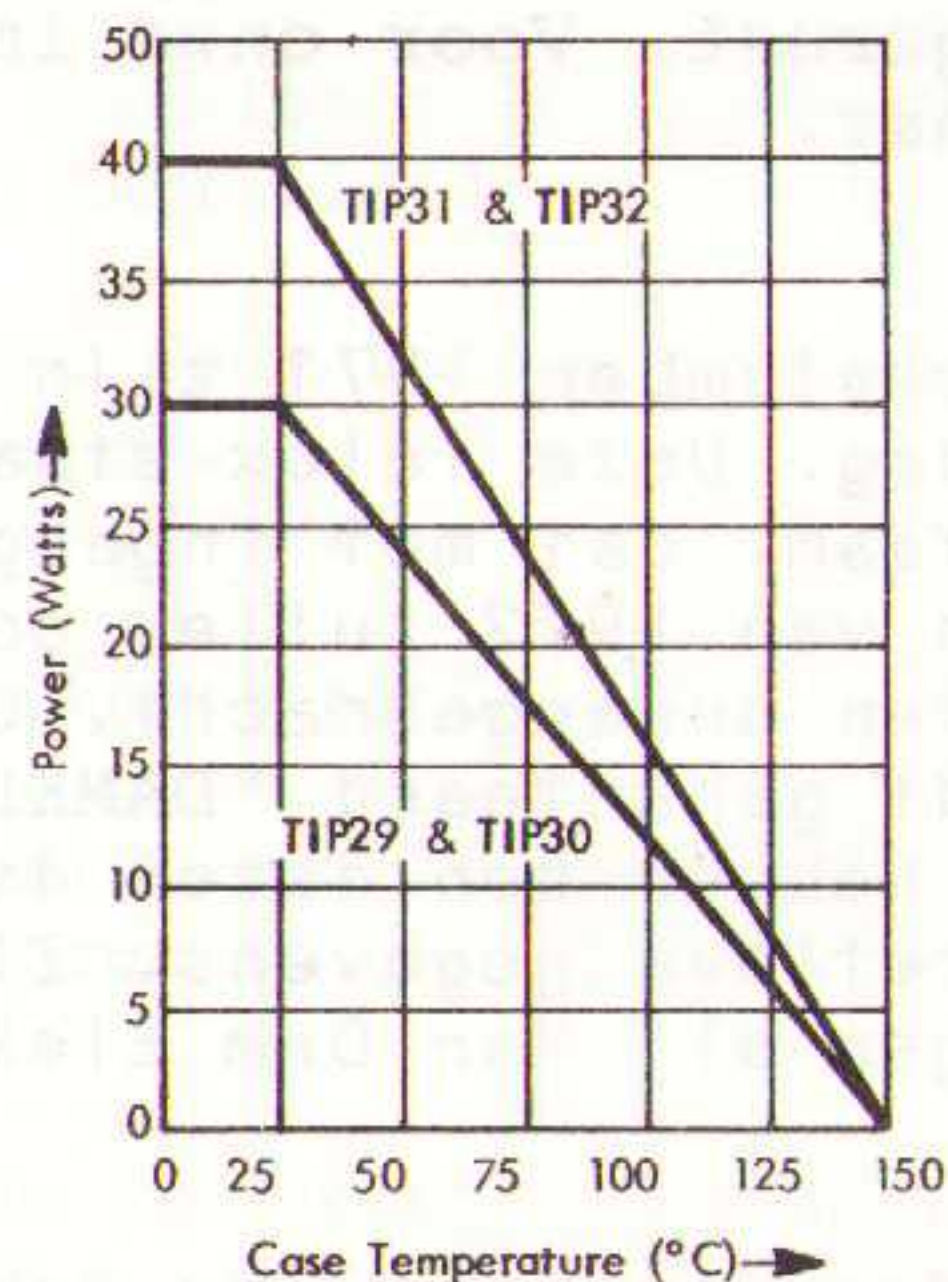


FIGURE 17. Dissipation Derating Curves

V. SUMMARY

Both amplifier designs presented in this note accomplish the purpose for which they were intended. They have demonstrated low cost amplifiers with satisfactory performance for the intended areas of usage.

Since the epitaxial base process allows the production of NPN and PNP silicon complements with relative ease, the devices in plastic packaging are available at a reasonable price. Having a high power dissipation, they will readily fill the needs in many areas where low cost and good performance are a criteria.

VI. ACKNOWLEDGEMENT

The author wishes to express his appreciation to Mr. Rufus Poteet for designing, building, and testing the 3 watt (rms) Automotive Audio Amplifier, and to Mr. Max Schreiner for his advice and assistance in the writing of this report.

In ons leveringsprogramma zijn sinds kort ook van het door ons vertegenwoordigde fabriek Bourns Security Systems Inc. industriële foto- en filmcamara's opgenomen, welke kunnen worden gebruikt voor het vastleggen van industriële processen of ongelukken, overvallen of inbraak. Er zijn verscheidene soorten in ons programma opgenomen voor de meest voorkomende installaties. Bijbehorende bedieningskasten kunnen eveneens worden geleverd, terwijl tevens uitvoeringen voorhanden zijn, welke door een ultrasonoor alarm apparaat kunnen worden bestuurd, of hiermee reeds zijn uitgerust. Voor onze industriële afnemers is een documentatiepakket beschikbaar.

Sinds 1 september 1971 zijn wij ook per telex bereikbaar gedurende 24 uur per dag. Deze telex staat in ons kantorengebouw aan de Spoorsingel te Rotterdam, dat met ingang van deze datum in gebruik is genomen. In het begin van 1972 zullen ook onze afdelingen verkoop e.d. in dit gebouw worden ondergebracht. Ons telexnummer is 25336, terwijl als roepnaam wordt gehanteerd "DAMEL" (van Dam Elektronica). Onze industriële afnemers hebben een dezer dagen een formulier ontvangen voor opgave van administratieve gegevens. Zij zullen via onze telex van de laatste ontwikkelingen bij Van Dam Elektronica op de hoogte worden gehouden.

Over de door ons voor de BeNeLux vertegenwoordigde onweer-prognose-apparatuur is een uitgebreid documentatie pakket beschikbaar met rapporten van Nederlandse instellingen en vakbladen, alsmede van buitenlandse organisaties. Dit documentatiepakket wordt U op aanvraag gratis toegezonden. Tevens zijn de prijzen van deze apparaten sinds 1 augustus 1971 verlaagd tot f 95,- voor het bouw pakket THP800K, f 150,- voor de gemonteerde uitvoering THP800, f 175,- voor de gemonteerde THP800S (THP800 met extra uitgang) en f 537,- voor de THP200 (uitvoering met dubbel meetbereik en standaard uitgangen). Een kaartje naar Postbus 3149 te Rotterdam of een telex naar nummer 25336 is voldoende om in het bezit te komen van de op dit moment beschikbare documentatie en rapporten.

NOG NET OP TIJD

Op de valreep kunnen wij U nog informeren over onze nieuwe vertegenwoordiging van het Amerikaanse fabriek FLEX KEY CORP. Deze firma houdt zich bezig met de produktie van toetsenborden voor tafelrekenmachines, computers, telefonie, enz. Het gehanteerde systeem is op het gebied van de mechanische schakeltechniek revolutionair te noemen en inmiddels volledig gepatendeerd. Elk (maak-)contact wordt gevormd door een onder druk veranderende contactweerstand: zonder druk op de toets is deze ongeveer ∞ , bij 200 gram druk $<100\Omega$ en bij 300 gram $<20\Omega$. FLEX KEY toetsenbordjes hebben een totale bouwhoogte van ± 8 mm en kunnen per toets circa 50 miljoen schakelfuncties verrichten! Het FLEX KEY systeem is direct te gebruiken met DTL, TTL en MOS systemen. De maximaal toelaatbare contactspanning is 150 volt, de toelaatbare stroom 20 mA. Het maximale schakelvermogen is 200 mW. De gemiddelde stijg- en daaltijd tussen contactmaken en 20Ω bedraagt 1-2 milliseconden! In ons voorraadprogramma zijn een geheel vlak toetsenbordje met 0-9 en een decimale punt opgenomen (f 93,50 +BTW) alsmede een geprofileerde uitvoering (f 122,00 +BTW). De typenummers zijn resp. DK1L en DK1M. Nadere documentatie komt in één van de komende documentaties.

pag. 2	VAN DAM ELEKTRONICA nu ook per telex.
pag. 3	Een IC testklem van Guest International Ltd.
pag. 4	Toepassingen van de RCA CA 3048.
pag. 12	Tantaal elco's in druppelformaat van ITT (TAG-serie).
pag. 14	Toepassingen van de RCA CA 3052 (soortgelijk type als de CA 3048).
pag. 19	Koelplaten programma Van Dam Elektronica.
pag. 21	Karakteristieken koelplaten.
pag. 27	Overzicht koelelementen van pagina 21 t/m 26.
pag. 28	Montagesteunen, siliconenpasta, mica-onderlegplaten.
pag. 29	Aluminium oxyde onderlegplaten, T0-3 voet, afdekkappen T0-3 en T0-66.
pag. 30	Vingerkoelelementen en kristalvoeten.
pag. 31	Koelblokken voor T0-5 en T0-18, koelster en opsteekkoelelementen.
pag. 32	Koelelementen voor T0-5 transistoren.
pag. 33	Koelelementen voor T0-18 transistoren en transistor-IC insteekvoeten.
pag. 34	Montage en aanpassingen voor monteren van halfgeleiders in prints.
pag. 35	Speciale koelelementen voor dioden.
pag. 37	DIN-steker (31 polig), print en chassisdeel voor Europa printkaart.
pag. 38	Prijzenoverzicht produkten van pag. 19 t/m 37.
pag. 39	Toepassingen CA3059; zie ook deel 1-3 pag. 47-62.
pag. 51	Enkele toepassingen van de 2N3823 FET.
pag. 52	Technische gegevens van de RCA fotoversterker type CA 3062.
pag. 56	Toepassingen van de RCA CA 3062.
pag. 62	Toepassing van Texas Instruments TIP transistoren in een 35 watt versterker.
pag. 73	Toepassing van Texas Instruments transistoren in een 15 watt en een 3 watt versterker.
pag. 86	Het laatste nieuws.
pag. 88	Prijzen van in deze documentatie opgenomen produkten (voor koelelementen zie pag. 38.).

BINNENKORT WORDT OOK LEVERBAAR:

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montagedelen voor een bijzonder universele opbouw van chassis: T-profielen, platte profielen, hoek-profielen, elleboogjes en verbindingshoeken. Vervaardigd van plastic (PVC) en aluminium.

++++

zijdemat uitgevoerd en geëloxeerde handgrepen voor montage op uw professionele apparatenkasten. Leverbaar in verscheidene maten en modellen. De prijzen variëren van f 2,40 tot f 3,30 excl. 14% BTW.

++

teflon soldeersteunen en -doorvoeren voor chassisgaten van $\pm 3,8$ mm ϕ . De kontaktstiften zijn van verzilverd messing. Prijzen variëren bij 1-99 stuks van f 0,40 tot f 0,55, terwijl deze prijs bij 1000 stuks daalt tot f 0,15 en f 0,17.

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van het door ons vertegenwoordigde fabrikaat Guest International Ltd. zijn binnenkort ook professionele voedingen leverbaar, welke instelbaar zijn binnen 0,5% met behulp van een 10-slagen potentiometer. In deze apparatuur wordt van silicium halfgeleiders gebruik gemaakt en is een instelbare overstroombeveiliging aanwezig. De uitgangsspanning en -stroom zijn omschakelbaar op één meter af te lezen. De line-regulation bedraagt <0,02% of 10 mV en de load-regulation <0,1% of 10 mV. Leverbaar in 4 modellen van 0-7½V, 0-15V, 0-30V en 0-60V. Prijzen van f 440,00 tot f 540,00.

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in het GUEST programma zijn naast de op pag. 3 van deze documentatie genoemde testklem voor geïntegreerde schakelingen en bovenstaande voedingen ook vele andere bijzondere produkten opgenomen. In één van de komende uitgaven van de TECHNISCHE DOCUMENTATIE zullen wij hierop nog nader terugkomen.

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PRIJZEN EXCL. 14% BTW VAN IN DEZE DOCUMENTATIE OPGENOMEN PRODUKTEN.

IC testklem, fabr. Guest International Ltd. Bij 1-24 stuks	f	25,00
Geïntegreerde schakeling type CA 3048, fabr. RCA. Bij 1-24 stuks.	f	19,45
type CA 3052, fabr. RCA. Bij 1-24 stuks.	f	16,95
type CA 3059, fabr. RCA. Bij 1-24 stuks.	f	15,55
type CA 3062, fabr. RCA. Bij 1-24 stuks.	f	23,50

Tantaal elco's, fabriikaat ITT, druppeluitvoering.

0,1uF - 35V	f 0,60	1 uF - 35V	f 0,60	10 uF - 6,3V	f 0,60
0,15uF - 35V	f 0,60	1,5uF - 25V	f 0,60	10 uF - 20V	f 0,80
0,22uF - 35V	f 0,60	2,2uF - 25V	f 0,60	22 uF - 16V	f 0,85
0,33uF - 35V	f 0,60	3,3uF - 25V	f 0,65	33 uF - 10V	f 0,85
0,47uF - 35V	f 0,60	4,7uF - 25V	f 0,70	47 uF - 6,3V	f 0,85
0,68uF - 35V	f 0,60	6,8uF - 25V	f 0,80	100uF - 3V	f 0,85

Koelelementen, zie voor prijzen pagina 38 van deze documentatie.

Transistoren, fabriikaat Texas Instruments.

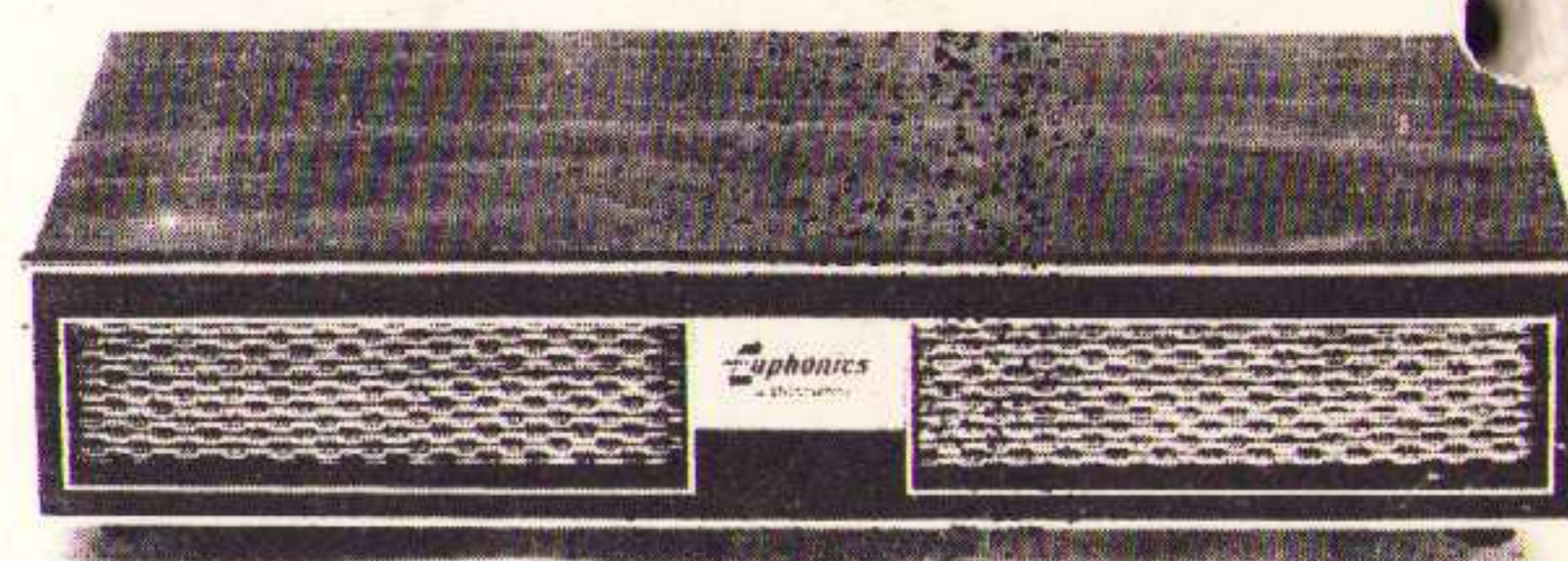
TIP 29A	f 4,00	TIP 30A	f 4,68	2N3702	f 1,04
TIP 31A	f 4,90	TIP 32A	f 5,50	2N3704	f 0,87
TIP 33A	f 7,48	TIP 34A	f 10,90	2N3710/11	f 0,87
TIP 35A	f 20,57	TIP 36A	f 27,50	TIS 97	f 1,73



Een compleet en geavanceerd programma op het gebied van alarmering en beveiliging kunnen wij u bieden voor uw woning of bedrijf op het gebied van:

- ☒ inbraak alarm
- ☒ brand alarm
- ☒ gas- en rook-alarm
- ☒ overvalalarm
- ☒ explosie-alarm
- ☒ ramp-alarm
- ☒ water-alarm
- ☒ autodiefstal alarm
- ☒ wat u nog verder wilt ...

van dam
ELEKTRONICA



Gebruik van onze systemen geeft niet alleen een gedegen alarmering, doch doet doorgaans ook Uw verzekeringspremie dalen. Vooral bij grote objecten kan de aanleg van een gecombineerde installatie bijzonder goede diensten bewijzen.

☒ gratis wordt op aanvraag een documentatiepakket toegestuurd door: **n.v. technische handelmaatschappij VAN DAM ELEKTRONICA**
Postbus 3149, Rotterdam, Holland
Telefoon: 010-240812. Telex: 25336