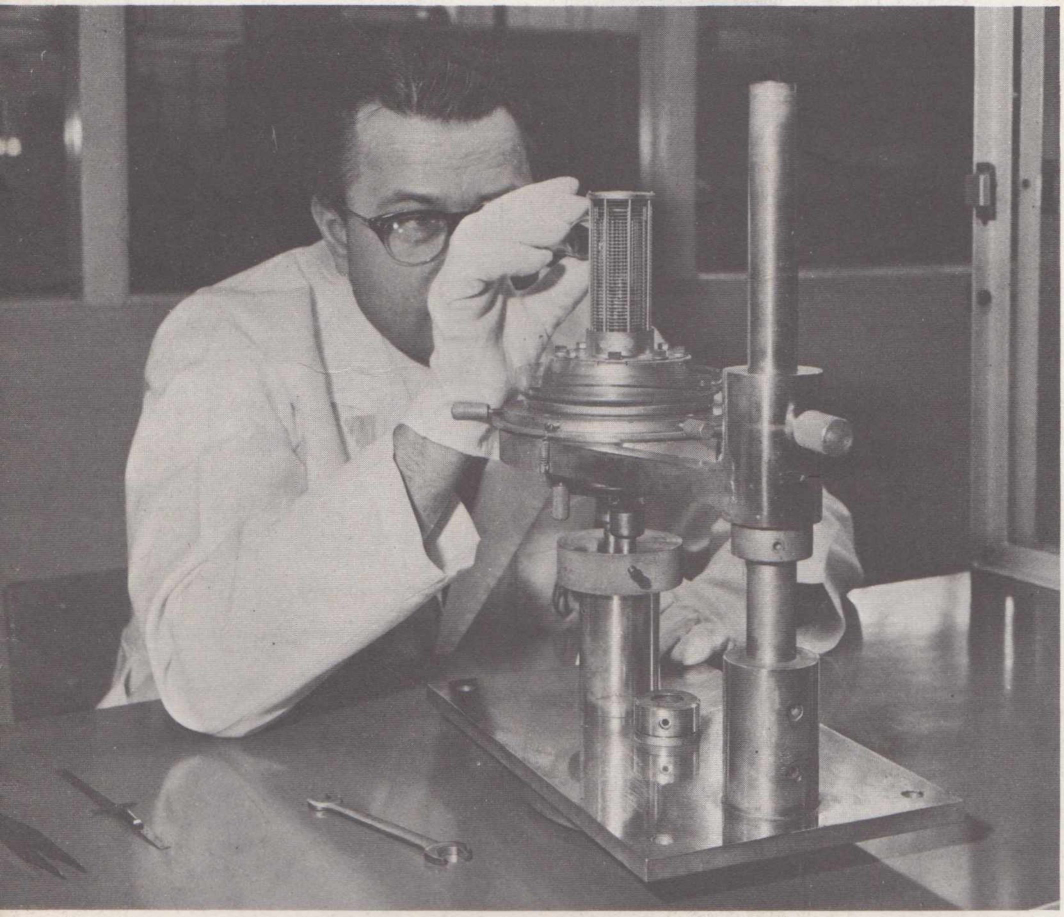
RADIOTRONICS



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COVER

Checking alignment of the GRIDS of the AWV 6166.

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A PRECISION RESISTANCE WELDING CONTROL USING INTEGRATED CIRCUITS

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Summary

A precision resistance welding control which uses solid state devices in its entirety is described. A brief discussion of the requirements of high speed resistance welding equipment as used in the electronic device manufacturing industry is included as a basis for the design approach adopted.

The practical realization of the design is demonstrated in terms of a unit incorporating a number of digital integrated circuits which achieves the multiple aims of effectiveness, reliability, and simplicity.

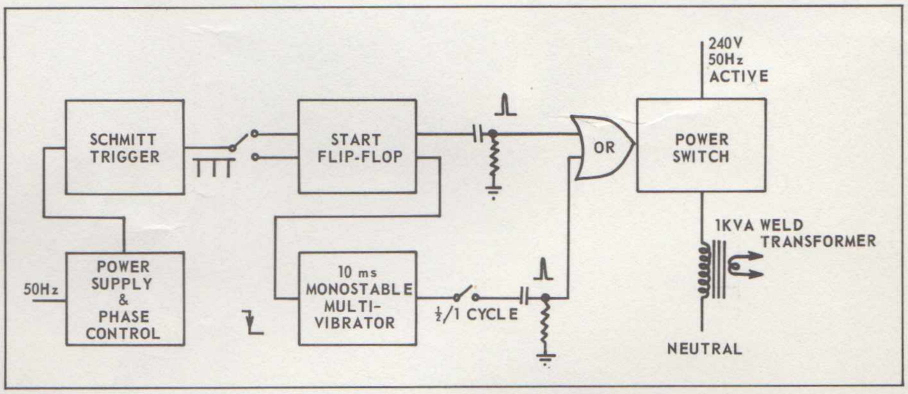

Figure 1. Block diagram of the weld control

Introduction

Resistance welding as a means of metal fabrication is commonly encountered in the manufacturing assembly processes of electronic devices. Valve structures and cathode ray tube electron guns are examples where numerous welding operations are required for each assembly, often with material thicknesses of only a few thousandths of an inch. In this type of welding operation, considerable care must be taken to avoid melting through the material or damaging adjacent areas by excessive conduction of heat.

The fundamental behaviour of the resistance welding process can be expressed by the empirical formula:

$$H = I^2 R\ T\ K \text{ watt-seconds}$$

where H = total heat generated in watt-seconds

I = current in amperes

R = resistance of work in ohms

T = duration of current flow in seconds

K = factor representing total radiation and conduction losses.

It can be seen from this expression that any desired amount of weld energy may be obtained by increasing the weld current level for a given time duration. It has been established that best results are obtained with the application of a relatively large amount of power for a correspondingly short period of time. Under these conditions, little conduction of heat from the immediate weld area takes place, and a further important advantage of reduced oxidisation results.

In practice, a weld transformer capacity of 1 KVA is common while the required time duration may range from a few milliseconds up. As the 50 Hz A.C. power mains has a half-cycle period of only 10 milliseconds, it is clear that large variations in actual heat energy will result unless the weld period is made synchronous with the supply mains.

Thus the basic requirements for a precision welding control may be established. The unit to be described fulfills the requirements of adequate power handling capacity, synchronous operation, and accuracy for short duration welds.

General Description

A block diagram of the unit is shown in figure 1. A Schmitt Trigger

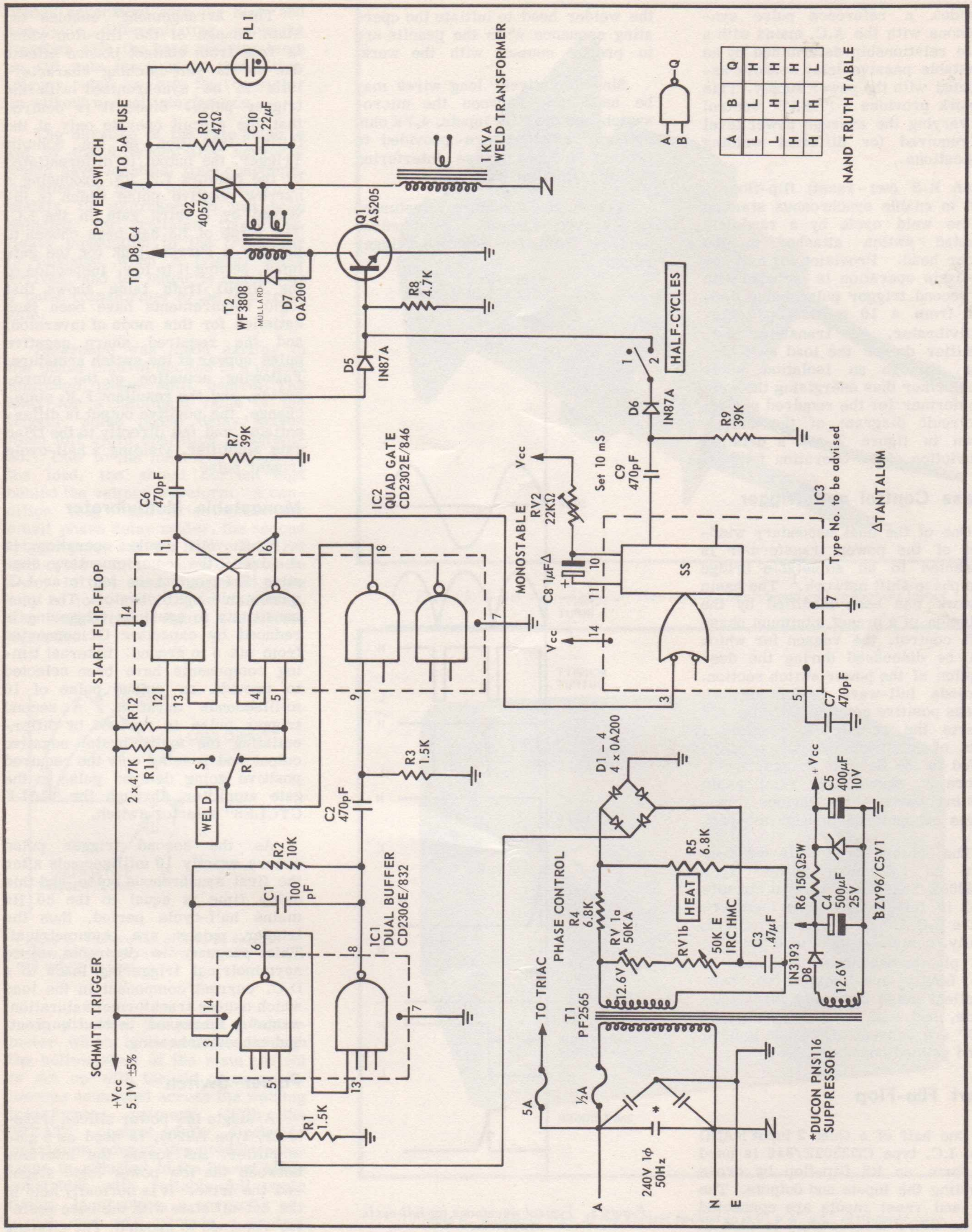

Figure 2. Circuit diagram of the Weld Control

provides a reference pulse synchronous with the A.C. mains with a phase relationship determined by an adjustable passive delay network associated with the power supply. This network provides a "HEAT" control for varying the average power level as required for different welding applications.

An R-S (set-reset) flip-flop is used to enable synchronous starting of the weld cycle by a randomly actuated switch attached to the welder head. Provision for half- or one-cycle operation is included with the second trigger pulse being derived from a 10 millisecond delay multivibrator. A transistor gate amplifier drives the load switching triac through an isolation pulse transformer thus energising the weld transformer for the required period. A circuit diagram of the unit is shown in figure 2 and a detailed description of its operation follows.

Phase Control and Trigger

One of the dual secondary windings of the power transformer is connected to an adjustable bridge type phase shift network.[1] The basic network has been modified by the inclusion of a preset minimum phase shift control, the reason for which will be discussed during the description of the power switch section. A diode full-wave bridge network passes positive going half-cycles and inverts the remainder, yielding a train of unidirectional pulses which is fed to the Schmitt Trigger input. Figure 3 shows the relationship existing between the various waveforms encountered through the unit.

The Schmitt Trigger is made up from an I.C. dual buffer[2] type CD2306E/832, the Integrated circuits used in this design being members of the R.C.A. CD2300 industrial DTL family, contained in 14 lead dual-in-line plastic packages. Members of this family are inexpensive, have excellent noise immunity characteristics, and with a lead spacing of .100" are convenient for use in hand wired printed boards.

Start Flip-Flop

One half of a Quad 2 input NAND gate I.C. type CD2302E/846 is used to form an RS flip-flop by cross coupling the inputs and outputs. The set and reset inputs are connected to an SPDT microswitch located at the welder head to initiate the operating sequence when the pencils are in proper contact with the work.

Since relatively long wires may be necessary between the microswitch and the F.F. inputs, 4.7 K ohm pull-up resistors are provided to prevent induced noise interfering with the flip-flop operation.

The switch armature is returned to a source of negative-going pulses derived from the Schmitt Trigger output.

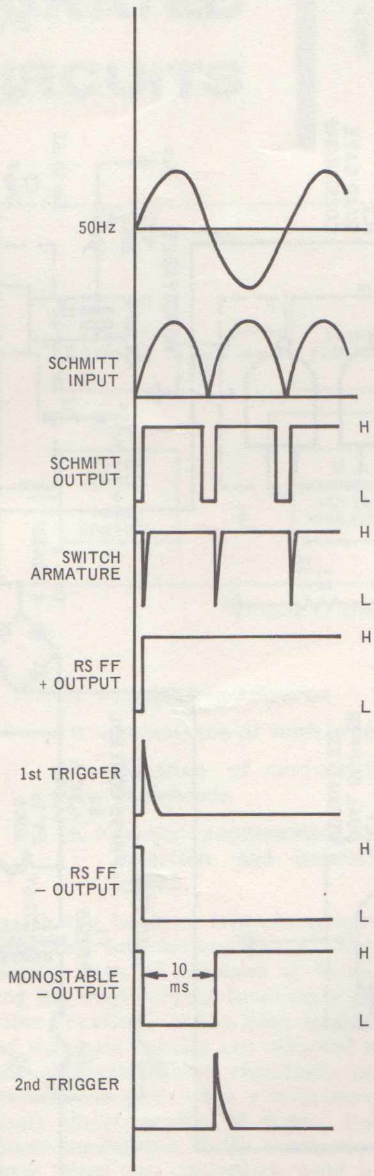

Figure 3. Typical waveforms for full-cycle operation.

This arrangement[3] enables the state change of the flip-flop which is free from contact bounce effects due to its self-latching characteristic to be synchronized with the trigger pulse. Since it is desired that the circuit operate only at the initial transition of the Schmitt Trigger, the output is differentiated by the network C2, R3, producing a narrow positive pulse which is inverted by a third gate in the I.C. The value of R3 has been chosen to provide a current sink for the gate input, setting it to low. Inspection of the NAND truth table shows that logic requirements have been thus satisfied for this mode of inversion, and the required sharp negative pulse appear at the switch armature. Following actuation of the microswitch, and the resultant F.F. state-change, the positive output is differentiated and fed directly to the triac gate amplifier, yielding a half-cycle trigger pulse.

Monostable Multivibrator

When full cycle operation is required, the complementary negative F.F. output is fed to an I.C. monostable multivibrator. The input sensitivity to spurious triggering is reduced by capacitor C7 connected from pin 5 to ground. External timing components have been selected to provide an output pulse of 10 milliseconds duration. A second trigger pulse is derived by differentiating the multivibrator negative output and passing only the required positive going delayed pulse to the gate amplifier, through the "HALF CYCLES" selector switch.

As the second trigger pulse occurs exactly 10 milliseconds after the first synchronous pulse, and this delay time is equal to the 50 Hz mains half-cycle period, thus the trigger pulses are symmetrical. This property is desirable, since asymmetrical triggering leads to a D.C. current component in the load which causes transformer saturation, wasteful increases in load current, and excessive heating.

Power Switch

A single low power silicon transistor, type AS205, is used as a gate amplifier, and forms the interface between the low power logic stages and the triac. It is normally held in the cut-off state with the base resistor returned to ground. The clipping diodes associated with the half- and one-cycle differentiating networks together with the base resistor form an OR gate function, so that either trigger pulse can turn the transistor on without mutual interference.

The duration of the gate pulse is set by the differentiating networks which have identical time constants. An effective pulse width of approximately 15 microseconds has been selected for reliable triggering. Heavy conduction in the transistor is initiated when a trigger pulse arrives, and the output is coupled by a pulse transformer to the load triac gate.

Reference was made earlier in the phase control section to a preset minimum phase shift control. The need for a minimum value of the trigger pulse phase delay arises when full cycle operation is selected. Due to the inductive nature of the load, the actual current lags behind the voltage waveform. A condition may then arise where for small phase delay angles, the second trigger pulse arrives before the first half-cycle current has decayed to zero. In this case, the second pulse may be lost, resulting in misfiring. The waveforms of figure 4 illustrate this effect.

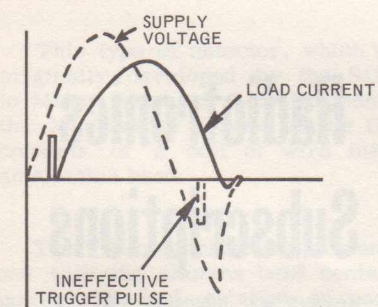

Figure 4. Illustration of loss of second trigger pulse with insufficient phase delay and inductive load.

In practice, the preset potentiometer which is accessible through the hollow shaft of the main control is set up with the aid of an oscilloscope connected across the welding transformer primary. With the "HEAT" control fully clockwise (maximum power), the preset resistance is adjusted to the lowest value consistent with reliable full cycle firing. In practice, a correctly adjusted unit will provide a control

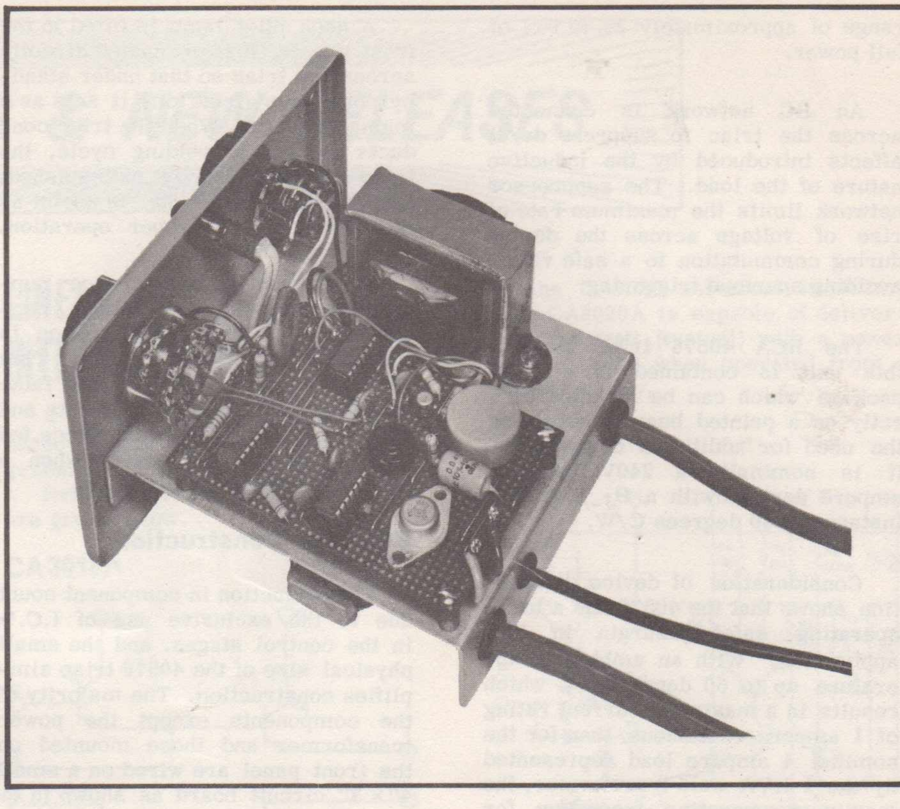

Figure 5. A rear view of the unit showing the circuit board and component layout

Figure 6. A general view of the control housed in a 6" x 4" x 6" instrument case.

range of approximately 3% to 95% of full power.

An RC network is connected across the triac to suppress dv/dt effects introduced by the inductive nature of the load. The suppressor network limits the maximum rate of rise of voltage across the device during commutation to a safe value, avoiding spurious triggering.

The RCA 40576 triac used in this unit is contained in a TO66 package which can be mounted directly on a printed board eliminating the need for additional components. It is nominally a 240V line, 15 ampere device, with a Θ_{J-A} in this instance of 40 degrees C/W.

Consideration of device dissipation shows that the 40576 has a large operating safety margin in this application. With an ambient temperature up to 60 degrees C., which results in a maximum current rating of 1 ampere continuous, then for the nominal 4 ampere load represented by the 1 KVA weld transformer, the maximum operating frequency for full-cycle welding would approximate 12 per second. Clearly this rate is far in excess of the maximum possible in an operator-limited application such as the present example.

The possibility does exist, of course, of the use of this type of equipment in automatic welding applications. In such instances specific information on permissible maximum duty cycle levels can be readily derived by calculation from the published data.

Power Supply

A simple half-wave supply is adequate for the present requirement. The nominal 5 volt supply required for the digital I.C.'s is obtained from a 5.1 volt, ± 5% zener diode which provides sufficient regulation for the comparatively light load. A half ampere fuse is connected in the power transformer primary circuit, while the load supply line incorporates a 5 ampere fuse. A standard fuse cannot interrupt destructive fault currents quickly enough to protect the triac, however it ensures that in the event of an equipment breakdown, the defective unit is isolated from the supply mains.

A neon pilot lamp is fitted to the front panel. It is connected directly across the triac so that under stand-by conditions (triac off) it acts as a mains indicator. When the triac conducts during a welding cycle, the lamp is momentarily extinguished, and the resultant flicker is useful as an indication of proper operation.

A composite interference suppressor consisting of three capacitors in a delta configuration is connected to the mains input. This device protects the unit from false triggering due to line transients and reduces the level of interference fed back to the supply mains when a weld operation is taking place.

Physical Construction

The reduction in component count due to the exclusive use of I.C.'s in the control stages, and the small physical size of the 40576 triac simplifies construction. The majority of the components except the power transformer and those mounted on the front panel are wired on a small 4" × 3" circuit board as shown in the photograph of figure 5. It will be noted that figure 5 shows the I.C.'s mounted in sockets. This was done to enable batch checking of I.C.'s in the prototype, and sockets would not normally be included in additional units. The complete equipment is housed in a small instrument case, a general view of which is shown in figure 6. The "HEAT" control is marked with an arbitrary alphabetical series enabling preferred control settings to be recorded for future use when the equipment is required to handle a variety of work.

Conclusion

A successful solid state replacement for many thyratron and capacitor discharge welding controls has been described. The economy and simplicity obtained from the use of digital integrated circuits in the greater part of the circuitry emphasise the advantages to be gained from these devices in industrial applications. The use of an improved triac device with a large power handling capacity in a small package has permitted a considerable reduction in mechanical complexity, and resulted in the construction of a unit which provides significant savings in bench space, and negligible installation costs.

Additional information regarding this application and specific data on the active devices used will be made available on request.

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3 *"SW772 Data", Stewart-Warner Microcircuits Inc. California, U.S.A. 1967.*

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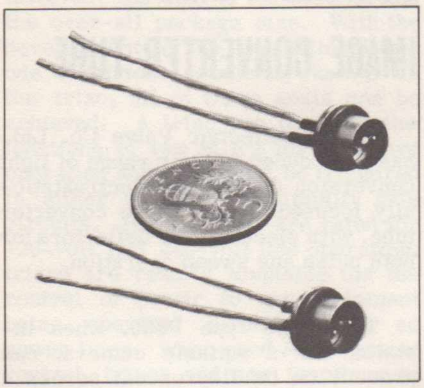
NEWS & NEW RELEASES

E.E.V. METHANE DETECTOR



English Electric Valve Co. Ltd. has produced a pinhead-sized detector (VQ1) which will continuously monitor the presence of methane in mines.

This type of detector, which was originally developed by the Safety in Mines Research Establishment of the Ministry of Power in the U.K. consists of a coil of wire inside an alumina bead.

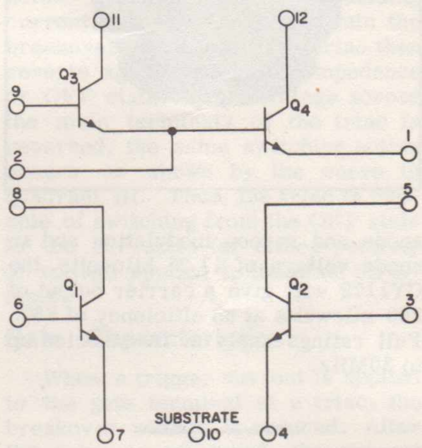
The VQ1 comprises two elements, one with the alumina bead containing a chemical catalyst to cause burning of methane on the device, and the other poisoned to prevent burning. Heat generated by the burning of methane on the one containing the catalyst causes a variation in its resistance, whereas under the same conditions the resistance of the poisoned device does not vary.

When one of each type of element is used to form two arms of a bridge circuit, the presence of methane which causes a variation of the resistance in only one of the devices gives an unbalanced condition of the bridge circuit, so providing a means of indicating the concentration present.

LINEAR INTEGRATED CIRCUITS

A further five types have been added to the range of Linear Integrated Circuits available from AWV. Brief details of the five new types are given below.

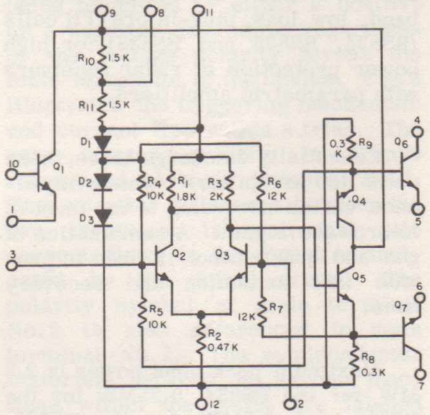
CA3018A



Schematic Diagram for CA3018A

A "Tight-spec" version of the popular CA3018 four transistor away.

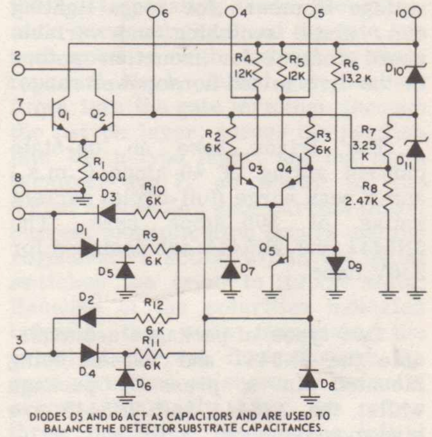
CA3020A



SCHEMATIC DIAGRAM FOR CA3020A

This type is an improved version of the CA3020 universal amplifier. The CA3020A is capable of delivering one watt (typical) with a power gain of 75dB when operated from a 12 volt supply.

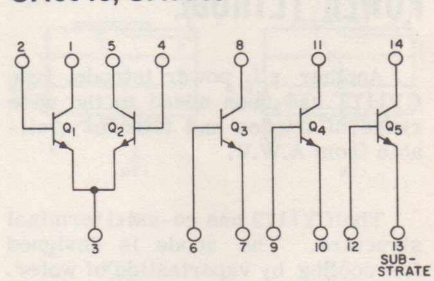
CA3044



Schematic diagram CA3044.

A second generation version of the CA3034 the circuit has been designed for use in Automatic Fine Tuning applications.

CA3045, CA3046



Schematic diagram for CA3045, CA3046

These two new types are general purpose transistors always consisting of five matched silicon N-P-N transistors on a common monolithic substrate.

The array contains three isolated transistors and one differentially connected transistor pair.

The CA3045 and CA3046 are electrically identical, with the CA3045 being a 14 lead dual-in-line ceramic package and the CA3046 a dual-in-line plastic package.

R.C.A. 40 AMP TRIACS

Four new 40 Amp Triacs have just been released by R.C.A., they are the 2N5441, 2N5442, 2N5444 and 2N5445. These devices are designed for use in residential heating control, electric range controls, high voltage dimmers for stage lighting etc., static switching and variable speed control for induction motors in the five to ten horsepower range.

The triacs have an on-state current rating of 40 amps (r.m.s.) and a peak surge (full-cycle) current rating of 300 amps peak. The 2N5441 and 2N5444 are designed for 240V lines.

Two types of packages are available the 2N5441 and 2N5442 being mounted in a press-fit package whilst the 2N5444 and 2N5445 are stud mounted.

E.E.V. 260 KW POWER TETRODE

Another r.f. power tetrode, type CY1172 has been added to the wide range of triodes and tetrodes available from A.W.V.

The CY1172 has co-axial terminal structure. The anode is designed for cooling by vaporisation of water, and contains axial passages around its circumference for the circulation of water and steam. It can dissipate 180 kilowatts.

The new tetrode is a high perveance tube, suitable for use in audio amplifiers, r.f. linear amplifiers, and class C amplifiers or oscillators. When operated as a Tyler high-efficiency circuit with

anode and screen modulation and an anode voltage of 11.25 kilovolts, the CY1172 will give a carrier output of 260 kilowatts at an efficiency of 86%. Full ratings apply at frequencies up to 30MHz.

PLUG-IN PRE-TR CELL

English Electric Valve Co. Ltd. has introduced a new range of broadband, low loss, plug-in pre-TR cells (BS834, BS836 and BS838) for high power protection of radar receivers with parametric amplifiers.

Essentially discharge tubes, these calls act as "pillars" whose impedance varies according to the level of microwave signal. A combination of gas and quartz wool is used to provide fast ionisation and recovery time.

Maximum peak input power is 2.5 MW for the BS834, 0.25MW for the BS836 and 0.5MW for the BS838.

They can be used in suitable waveguide mounts at any frequency between 2GHz and 12GHz. The handwidth and matching are determined by the design of the mount.

Dimensions of the cells are 8 in. (203.2 mm) long by 0.505 in. (12.83 mm) maximum diameter.

MULTIPLE-FRAME IMAGE CONVERTER TUBE

English Electric Valve Co. Ltd. has introduced to their range of light conversion devices an electrostatically focused triode image converter tube, with electrostatic deflectors for both pulse ans sweep operation.

This tube type P856, when installed in a suitable camera, can present on its flourescent screen a sequence of frames showing the development of a high-speed event. The shutter action is achieved by deflection of the electron beam over a slit in an aperature plate in the tube.

To nullify the blurred effect caused by the electron beam movement across the slit, a second waveform with an equivalent frequency and amplitude but out-of-phase with the first, is applied across a second pair of deflectors. A third pair of deflectors. A third pair of deflectors is used to apply a staircase voltage, triggered from the event under investigation, so enabling a sequence of images to be recorded.

Shuttering speeds of 20,000,000 per second can be obtained and, dependent on the image size, the number of frames to be recorded can be selected to be between 8 and 32. In general operation the exposure is set at 1/5 of the framing frequency and exposure times better than 10ns can be achieved.

The P856 tube has an S11 photocathode with a sensitivity (to 2854 degrees K tungsten light) of 15 uA/L min. The flat, circular face-plate of the flourescent screen provides a useful screen area of 7.5 cm by 4cm.

TRIAC POWER CONTROL APPLICATIONS

by J.V. YONUSHKA

In the control of ac power by means of semiconductor devices, emphasis has been placed upon limiting the complexity of the circuits involved, the cost of the system, and the over-all package size. With the development of the bidirectional triode thyristor, commonly known as the triac, all of these goals can be achieved. A triac can perform the functions of two SCR's for full-wave operation and can easily be triggered in either direction to simplify gate circuits. Because they are rated for 120-volt and 240-volt line operation, triacs are readily adaptable for the control of power to any equipment being operated directly from ac power lines. When used for ac power control, triacs add new functions to many designs, improve performance, and provide maximum efficiency and high reliability. This article describes triac operating characteristics and provides guidance in the use of triacs for specific applications.

Principal Voltage-Current Characteristic Diagram

Fig. 1 shows the principal voltage-current characteristic of a triac. This curve shows the current through the triac as a function of the voltage applied between main terminals Nos. 1 and 2. In quadrant I, the voltage on main terminal No. 2 is positive with respect to main terminal No. 1; in quadrant III, the voltage on main terminal No. 2 is negative with respect to main terminal No. 1. When a positive voltage is applied to main terminal No. 2, as shown by the curve in quadrant I, a point is reached, called the break-over voltage V_{BO}, at which the device switches from a high-impedance state to a low-impedance state. The current can then be increased through the triac with only a small increase in voltage across the device. The triac remains in the ON state until the current through the main terminals drops below a value, called the holding current, which cannot maintain the breakover condition. The triac then reverts again to the high-impedance or OFF state. If the voltage across the main terminals of the triac is reversed, the same switching action occurs as shown by the curve in quadrant III. Thus, the triac is capable of switching from the OFF state to the ON state for either polarity of voltage applied to the main terminals.

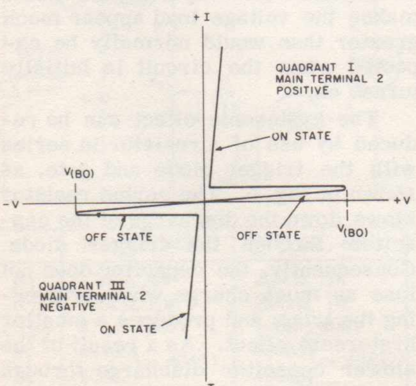


Fig. 1 – Triac principal voltage-current characteristics.

Gate Characteristics

When a trigger current is applied to the gate terminal of a triac, the breakover voltage is reduced. After the triac is triggered, the current flow through the main terminals is independent of the gate signal and the triac remains in the ON state until the principal current is reduced below the holding-current level. The triac has the unique capability of being triggered by either a positive or a negative gate signal regardless of the voltage polarity across the main terminals of the device. Fig. 2 illustrates the triggering mechanism and current flow within a triac. The gate trigger polarity is always referenced to main terminal No. 1. The potential difference between the two terminals is such that gate current flows in the direction indicated by the dotted arrow. The polarity symbol at main terminal No. 2 is also referenced to main terminal No. 1. The semiconductor materials between the various junctions within the pellet are labeled p and n to indicate the type of majority-carrier concentrations within the material.

For the various operating modes, the polarity of the voltage on main terminal No. 2 with respect to main terminal No. 1 is given by the quadrant in which the triac operates, (either I or III) and the polarity of the gate signal used to trigger the device is given by the proper symbol next to the operating quadrant. For the I (+) operating mode, therefore, main terminal No. 2 and the gate are both positive with respect to main terminal No. 1. Initial gate current flows into the gate terminal, through the p-type layer, across the junction into the n-type layer, and out main terminal No. 1, as shown by the dotted arrow. As gate current flows, current multiplication occurs and the regenerative action within the pellet switches the triac to its ON state. Because of the polarities indicated between the main terminals, the principal current flows through the pnpn structure as shown by the solid arrow. Similarly, for the other three operating modes, the initial gate current flow is shown by the dotted arrow, and principal current flow through the main terminals is shown by the solid arrow.

Because the principal current influences the gate trigger current,

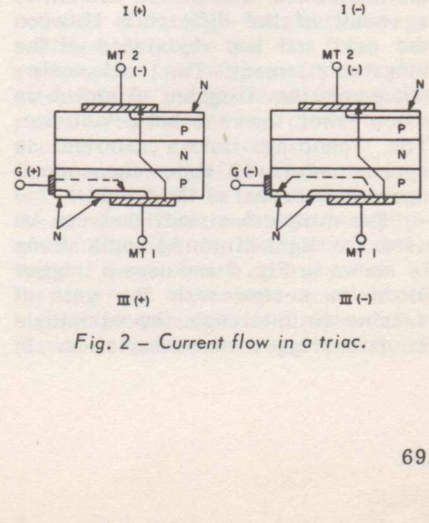


Fig. 2 – Current flow in a triac.

the magnitude of the current required to trigger the triac differs for each mode. The operating modes in which the principal current is in the same direction as the gate current required less gate trigger current, while modes in which the principal current is in opposition to the gate current require more gate trigger current.

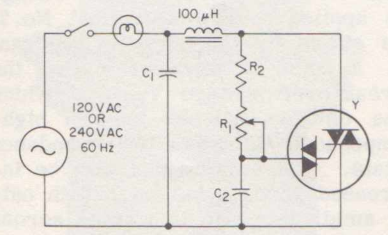
Like many other semiconductor parameters, the magnitude of the gate trigger current and voltage varies with the junction temperature. As the thermal excitation of carriers within the semiconductor increases, the increase in leakage current makes it easier for the device to be triggered by a gate signal. Therefore, the gate becomes more sensitive in all operating modes as the junction temperature increases. Conversely, if the triac is to be operated a low temperatures, sufficient gate trigger current must be provided to assure triggering of all devices at the lowest operating temperature expected in any particular application. Variations of gate trigger requirements are given in the data sheets for individual triacs.

Light Control

Because the light output of an incandescent lamp depends upon the voltage impressed upon the lamp filament, changes in the lamp voltage vary the brightness of the lamp. When ac source voltages are used, a triac can be used in series with an incandescent lamp to vary the voltage to the lamp by changing its conduction angle; i.e., the portion of each half cycle of ac line voltage in which the triac conducts to provide voltage to the lamp filament. The triac, therefore, is very attractive as a switching element in light-dimming applications.

To switch incandescent lamp loads reliably, a triac must be able to withstand the inrush current of the lamp load. The inrush current is a result of the difference between the cold and hot resistance of the tungsten filament. The cold resistance of the tungsten filament is much lower than the hot resistance. The resulting inrush current is approximately 12 times the normal operating current of the lamp.

The simplest circuit that can be used for light-dimming applications is shown in Fig. 3 and uses a trigger diode in series with the gate of a triac to minimize the variations in gate trigger characteristics. In applications where space is premium the 40431 or 40432 may be used because it combines both functions in a single package. Changes in the resistance in series with the capacitor change the conduction angle of the triac. Because of its simplicity, this circuit can be packaged in confined areas where space is at a premium.



	120VAC, 60Hz	240VAC, 60Hz
R_1	200kΩ, ½W	250kΩ, 1W
R_2	3.3kΩ, ½W	4.7kΩ, ½W
C_1	0.1µF, 200V	0.1µF, 400V
C_2	0.1µF, 100V	0.1µF, 100V
Y	40431	40432

Fig. 3 – Single-time-constant light-dimmer circuit.

The capacitor in the circuit of Fig. 3 is charged through the control potentiometer and the series resistance. The series resistance is used to protect the potentiometer by limiting the capacitor charging current when the control potentiometer is at its minimum resistance setting. This resistor may be eliminated if the potentiometer can withstand the peak charging current until the triac turns on. The trigger diode conducts when the voltage on the capacitor reaches the diode breakover voltage. The capacitor then discharges through the trigger diode to produce a current pulse of sufficient amplitude and width to trigger the triac. Because the triac can be triggered with either polarity of gate signal, the same operation occurs on the opposite half-cycle of the applied voltage. The triac, therefore, is triggered and conducts on each half-cycle of the input supply voltage.

The interaction of the RC network and the trigger diode results in a hysteresis effect when the triac is initially triggered at small conduction angles. The hysteresis effect is characterized by a difference in the control potentiometer setting when the triac is first triggered and when the circuit turns off. Fig. 4 shows the interaction between the RC network and the trigger diode to produce the hysteresis effect. The capacitor voltage and the ac line voltage are shown as solid lines. As the resistance in the circuit is decreased from its maximum value, the capacitor voltage reaches a value which fires the trigger diode. This point is designated A on the capacitor-voltage wave-shape. When the trigger diode fires, the capacitor discharges and triggers the triac at an initial conduction angle θ_1. During the forming of the gate trigger pulse, the capacitor voltage drops suddenly. The charge on the capacitor is smaller than when the trigger diode did not conduct. As a result of the different voltage conditions on the capacitor, the breakover voltage of the trigger diode is reached earlier in the next half-cycle. This point is labeled point B on the capacitor-voltage waveform. The conduction angle θ_2 corresponding to point B is greater than θ_1. All succeeding conduction angles are equal to θ_2 in magnitude. When the circuit resistance is increased by a change in the potentiometer setting the triac is still triggered, but at a smaller conduction angle. Eventually, the resistance in series with the capacitance becomes so great that the voltage on the capacitor does not reach the breakover voltage of the trigger diode. The circuit then turns off and does not turn on until the circuit resistance is again reduced to allow the trigger diode to be fired. The hysteresis effect makes the voltage load appear much greater than would normally be expected when the circuit is initially turned on.

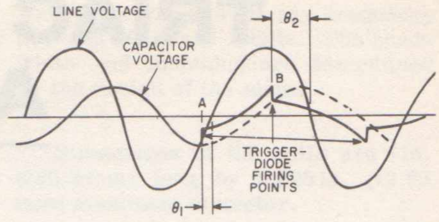
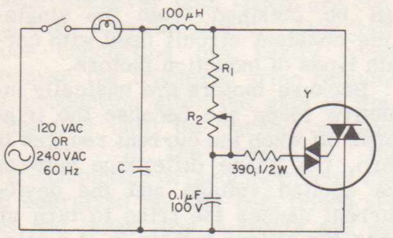


Fig. 4 – Waveforms showing interaction of control network and trigger diode.

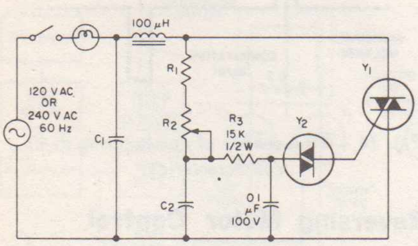
The hysteresis effect can be reduced by use of a resistor in series with the trigger diode and gate, as shown in Fig. 5. The series resistor slows down the discharge of the capacitor through the trigger diode. Consequently, the capacitor does not lose as much charge while triggering the triac, and produces a smaller hysteresis effect. As a result of the slower capacitor discharge through the trigger diode, however, the peak magnitude of the gate trigger current pulse is reduced. The size of the trigger capacitor may have to be increased to compensate for the reduction of the gate trigger current pulse.



	120VAC, 60Hz	240VAC, 60Hz
R_1	3.3kΩ, ½W	4.7kΩ, ½W
R_2	200kΩ, ½W	250kΩ, 1W
C	0.1μF, 200V	0.1μF, 400V
Y	40431	40432

Fig. 5 – Single-time-constant light-dimmer circuit with series gate resistor.

The double-time-constant circuit in Fig. 6 improves on the performance of the single-time-constant control circuit. This circuit uses an additional RC network to extend the phase angle so that the triac can be triggered at small conduction angles. The additional RC network also minimizes the hysteresis effect. Fig. 7 shows the voltage wave-forms for the ac supply and the trigger capacitor of the circuit of Fig. 6. Because of the voltage drop across R_3, the input capacitor C_2 charges to a higher voltage than the trigger capacitor C_3. When the voltage on C_3 reaches the breakover voltage of the trigger diode, the diode conducts and causes the capacitor to discharge and produce the gate current pulse to trigger the triac. After the trigger diode turns off, the charge on C_3 is partially restored by the charge from the input capacitor C_2.



	120VAC, 60Hz	240VAC, 60Hz
R_1	2.2kΩ, ½W	3.3kΩ, ½W
R_2	100kΩ, ½W	200kΩ, 1W
C_1C_2	0.1μF, 200V	0.1μF, 400V
Y_1	40485 } or 40431	40486 } or 40432
Y_2	40483 } or 40431	40483 } or 40432

Fig. 6 – Double-time-constant light-dimmer circuit.

The partial restoration of charge on C_3 results in better circuit performance with a minimum of hysteresis.

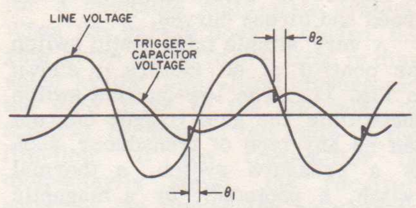
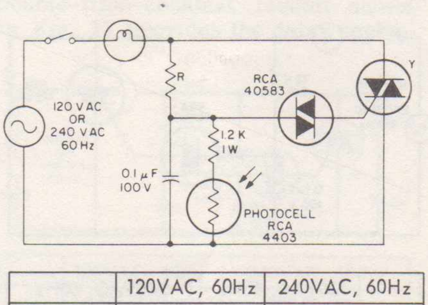


Fig. 7 – Voltage waveforms of double-time-constant control circuit.

Light-Activated Control

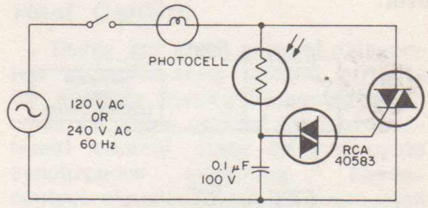
For applications requiring a light-activated circuit, such as outdoor lights or indoor night lights, the circuit shown in Fig. 8 can be employed. Although this circuit functions in the same manner as the light-dimming circuit, the photocell controls its operation. When the light impinges on the surface of the photocell, the resistance of the photocell becomes low and prevents the voltage on the trigger capacitor from increasing to the breakover voltage of the trigger diode. The circuit is then inoperative. When the light source is removed, the photocell becomes a high resistance. The voltage on the trigger capacitor then increases to the breakover voltage of the trigger diode and causes the diode to fire. The trigger pulse formed by the capacitor discharges through the trigger diode makes the triac conduct and operates the circuit. The triac continues to be triggered on each half-cycle and supplies power to the load as long as the resistance of the photocell is high. When light again impinges on the surface of the photocell and reduces its resistance, the voltage on the capacitor can no longer reach the breakover voltage of the trigger diode, and the circuit turns off.



	120VAC, 60Hz	240VAC, 60Hz
R	15kΩ, 2W	30kΩ, 3W
Photocell	4403	4453
Y	40485	40486

Fig. 8 – Light Controlled Turn-Off Circuit.

For applications requiring operation when light impinges on the surface of the photocell, the circuit of Fig. 9 is recommended. In this circuit, low resistance of the photocell allows the triac to be triggered on. When light is removed from the photocell the increased resistance of the photocell prevents the triac from being triggered and renders the circuit inoperative.



	120VAC, 60Hz	240VAC, 60Hz
Photocell	4403	4453
Y	40485	40486

Fig. 9 – Light Controlled Turn-On Circuit.

Radio Frequency Interference

The fast switching action of triacs when they turn on into resistive loads causes the current to rise to the instantaneous value determined by the load in a very short period of time. This fast switching action produces a current step which is largely composed of higher-harmonic frequencies that have an amplitude varying inversely as the frequency. In phase-control applications such as light dimming, this current step is produced on each half-cycle of the input voltage. Because the switching occurs many times a second, a noise pulse is generated into frequency-sensitive devices such as AM radios and causes annoying interference. The amplitude of the higher frequencies in the current step is of such low levels that they do not interfere with television or FM radio.

There are two basic types of radio-frequency interference (RFI) associated with the switching action of triacs. One form, radiated RFI, consists of the high-frequency energy radiated through the air from the equipment. In most cases, this radiated RFI is insignificant unless the radio is located very close to the source of the radiation.

Of more significance is conducted RFI which is carried through the power lines and affects equipment attached to the same power lines. Because the composition of the current waveshape consists of higher frequencies, a simple choke placed in series with the load slows down the current rise time and reduces the amplitude of the higher harmonics. To be effective, however, such a choke must be quite large. A more effective filter, and one that has been found adequate for most light-dimming applications is shown in Fig. 10. The LC filter provides adequate attenuation of the high-frequency harmonics and reduces the noise interference to a low level.

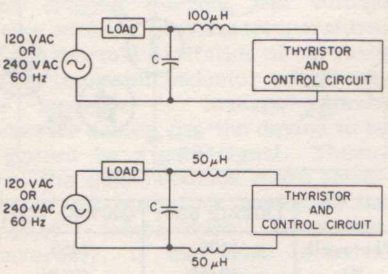
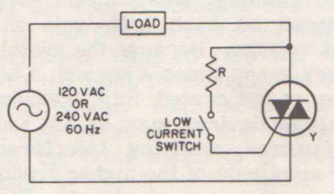


Fig. 10 – RFI-suppression networks: at 120 VAC, C = 0.1μF, 200 V; at 240 VAC, C = 0.1μF, 400 V.

Motor Control

Triacs can be used very effectively to apply power to motors and perform such functions as speed control, reversing, full power switching, or any other desired operating condition that can be obtained by a switching action. Because most motors are line-operated, the triac can be used as a direct replacement



	120VAC, 60Hz	240VAC, 60Hz
R	1kΩ, ½W	2kΩ, ½W
Y	RCA 40429	RCA 40430

Fig. 11 – Simple Triac Static Switch.

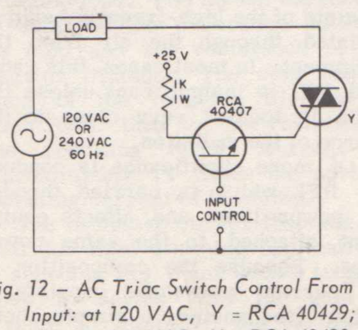


Fig. 12 – AC Triac Switch Control From DC Input: at 120 VAC, Y = RCA 40429; at 240 VAC, Y = RCA 40430.

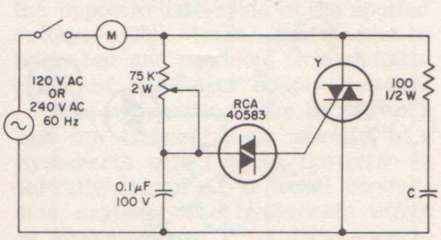
for electro-mechanical switches. In proper control circuits, triacs can change the operating characteristics of motors to obtain many different speed and torque curves.

A very simple triac static switch for control of ac motors is shown in Fig. 11. The low-current switch controlling the gate trigger current can be any type of transducer, such as a pressure switch, a thermal switch, a photocell, or a magnetic reed relay. This simple type of circuit allows the motor to be switched directly from the transducer switch without any intermediate power switch or relay.

For dc control, the circuit of Fig. 12 can be used. By use of the dc triggering modes, the triac can be directly triggered from transistor circuits by either a pulse or continuous signal.

Induction Motor Controls

Fig. 13 shows a single-time-constant circuit which can be used as a satisfactory proportional speed control for some applications and with certain types of induction motors, such as shaded pole or permanent split-capacitor motors, when the load is fixed. This type of circuit is best suited to applications which require speed control in the medium to full-power range. It is specifically useful in applications such as fans or blower-motor controls, where a small change in motor speed produces a large change in air velocity. Caution must be exercised if this type of circuit is used with induction motors because the motor may stall suddenly if the speed of the motor is reduced below the drop-out speed for the specific operating condition determined by the conduction angle of the triac. Because the single-time-constant circuit cannot provide speed control of an induction motor load from



	120VAC, 60Hz	240VAC, 60Hz
C	0.22μF, 200V	0.22μF, 400V
Y	40429	40430

Fig. 13 – Induction motor control.

maximum power to full off, but only down to some fraction of the full-power speed, the effects of hysteresis described previously are not present. Speed ratios as high as 3 : 1 can be obtained from the single-time-constant circuit used with certain types of induction motors.

Because motors are basically inductive loads and because the triac turns off when the current reduces to zero, the phase difference between the applied voltage and the device current causes the triac to turn off when the source voltage is at a value other than zero. When the triac turns off, the instantaneous value of input voltage is applied directly to the main terminals of the triac. This commutation voltage may have a rate of rise which can retrigger the triac. The commutating dv/dt can be limited to the capability of the triac by use of an RC network across the device, as shown in Fig. 13. The current and voltage waveshapes for the circuit are shown in Fig. 14 to illustrate the principal of commutating dv/dt.

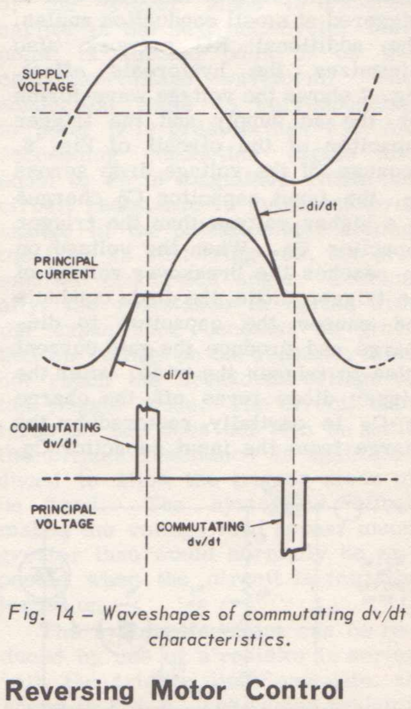
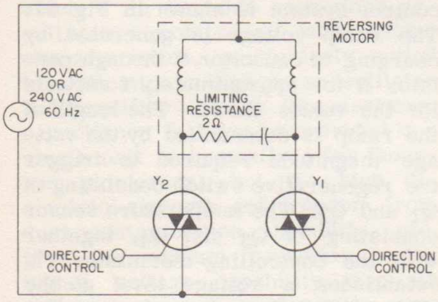


Fig. 14 – Waveshapes of commutating dv/dt characteristics.

Reversing Motor Control

In many industrial applications, it is necessary to reverse the direction of a motor, either manually or by means of an auxiliary circuit. Fig. 15 shows a circuit which uses two triacs to provide this type of reversing motor control. The reversing switch can be either a manual switch or an electronic switch used with some type of sensor to reverse the direction of the motor. A resistance is added in series with the capacitor to limit capacitor discharge current to a safe value whenever both triacs are conducting simultaneously. Simultaneous conduction can easily occur because the triggered triac remains in conduction after the gate is disconnected until the current reduces to zero. In the meantime, the non-conducting-triac gate circuit can be energized so that both triacs are ON and large loop currents are set up in the triacs by the discharge of the capacitor.

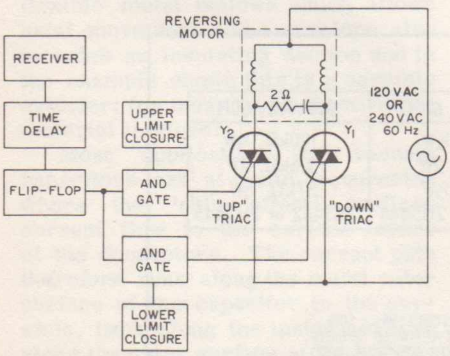


	120VAC, 60Hz	240VAC, 60Hz
Y	40485	40486
Y	40485	40486

Fig. 15 – Reversing motor control.

Electronic Garage-Door System

The triac motor-reversing circuit can be extended to electronic garage-door systems which use the principle of motor reversing for garage-door direction control. The system contains a transmitter, a receiver, and an operator to provide remote control for door opening and closing. The block diagram in Fig. 16 shows the functions required for a complete solid-state system. When the garage door is closed, the gate drive to the DOWN triac is disabled by the lower-limit closure and the gate drive to the UP triac is inactive because of the state of the flip-flop. If the transmitter is momentarily keyed, the receiver activates the time-delay monostable multivibrator so that it then changes the flip-flop state and provides continuous gate drive to the UP triac. The door then continues to travel in the UP direction until the upper-limit switch closure disables gate drive to the UP triac. A second keying of the transmitter provides the DOWN triac with gate drive and causes the door to travel in the DOWN direction until the gate drive is disabled by the lower limit closure. The time in which the monostable multivibrator is active should override normal transmitter keying for the purpose of eliminating erroneous firing. A feature of this system is that, during travel, transmitter keying provides motor reversing independent of the upper- or lower-limit closure. Additional features, such as obstacle obstructions, manual control, or time delay for overhead garage lights can be achieved very economically.

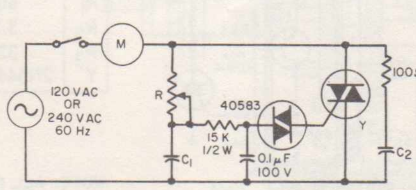


	120VAC, 60Hz	240VAC, 60Hz
Y_1	40485	40486
Y_2	40485	40486

Fig. 16 – Block diagram for remote-control solid-state garage-door system.

Universal Motor Speed Controls

In applications in which the hysteresis effect can be tolerated or which require speed control primarily in the medium to full-power range, a single-time-constant circuit such as that shown in Fig. 13 for induction motors can also be used for universal motors. However, it is usually desirable to extend the range of speed control from full-power ON to very low conduction angles. The double-time-constant circuit shown in Fig. 17 provides the delay necessary to trigger the triac at very low conduction angles with a minimum of hysteresis, and also provides practically full power to the load at the minimum-resistance position of the control potentiometer. When this type of control circuit is used, an infinite range of motor speeds can be obtained from very low to full-power speeds.

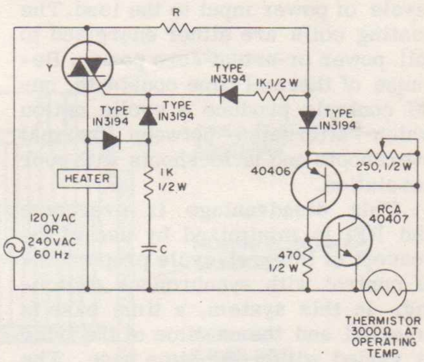


	120VAC, 60Hz	240VAC, 60 Hz
R	100kΩ, ½W	200kΩ, 1W
C_1	0.1µF, 200V	0.1µF, 400V
C_2	0.22µF, 200V	0.22µF, 400V
Y	40429	40430

Fig. 17 – Universal Motor Speed Control.

Heat Control

There are three general categories of solid-state control circuits for electric heating elements: on-off control, phase control, and proportional control using integral-cycle synchronous switching. Phase-control circuits, such as those used for light dimming are very effective and efficient for electric heat control except for the problem of RFI. In higher-power applications, the RFI is of such magnitude that suppression circuits to minimize the interference become quite bulky and expensive.



	120VAC, 60Hz	240VAC, 60Hz
R	2.2kΩ, 5W	3.9kΩ, 5W
C	0.5µF, 200V	0.5µF, 400V
Y	40575	40576

Fig. 18 – Synchronous switching on-off heat controller.

An on-off circuit for the control of resistance-heating elements is Fig. 18. The circuit also provides synchronous switching close to the beginning of the zero-voltage crossing of the input voltage to minimize RFI. The thermistor controls the operation of the two-transistor regenerative switch, which, in turn, controls the operation of the triac. When the temperature being controlled is low, the resistance of the thermistor is high and the regenerative switch is OFF. The triac is then triggered directly from the line on positive half-cycles of the input voltage. When the triac triggers and applies voltage to the load, the capacitor is charged to the peak value of the input voltage. The capacitor discharges through the triac gate to trigger the triac on the opposite half-cycle. The diode-resistor-capacitor "slaving" network triggers the triac on negative half-cycles of the ac input voltage after it is triggered on the positive half-cycle to provide integral cycles of ac power to the load.

When the temperature being controlled reaches the desired value as determined by the thermistor, the transistor regenerative switch conducts at the beginning of the positive input-voltage cycle to shunt the trigger current away from the triac gate. The triac does not conduct as long as the resistance of the thermistor is low enough to make the transistor regenerative switch turn on before the triac can be triggered.

Proportional Integral-Cycle Control

On-off controls have only two levels of power input to the load. The heating coils are either energized to full power or are at zero power. Because of thermal time constants, on-off controls produce a cyclic action which alternates between thermal overshoots and undershoots with poor resolution.

This disadvantage is overcome and RFI is minimized by use of the concept of integral-cycle proportional control with synchronous switching. In this system, a time base is selected and the on-time of the triac is varied within the time base. The ratio of the on-to-off time of the triac within this time interval depends upon the power required to the heating elements to maintain the desired temperature. Fig. 19 shows the on-off ratio of the triac. Within the time period, the on-time varies by an integral number of cycles from full ON to a single cycle of input voltage.

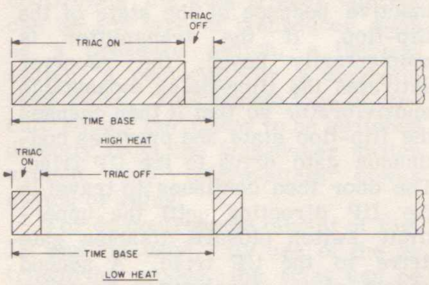


Fig. 19 – Triac duty cycle.

One method of achieving integral cycle proportional control is to use a fixed-frequency sawtooth generator signal which is summed with a dc control signal. The sawtooth generator establishes the period of time base of the system. The dc control signal is obtained from the output of the temperature-sensing network. The principal is illustrated in Fig. 20. As the sawtooth voltage increases, a level is reached which turns on power to the heating elements. As the temperature at the sensor changes, the dc level shifts accordingly and changes the length of time that the power is applied to the heating elements within the established time.

When the demand for heat is high, the dc control signal is high and little power is supplied continuously to the heating elements. When the demand for heat is completely satisfied, the dc control signal is low and no power is supplied to the heating elements. Usually a system using this principle operates continuously somewhere between full ON and full OFF to satisfy the demand for heat.

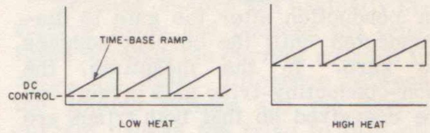
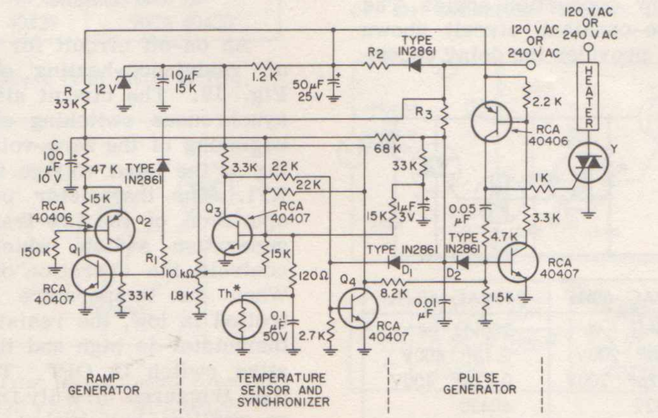


Fig. 20 – Proportional-controller waveshapes.

A proportional integral-cycle heat control system is shown in Fig. 21. The ramp voltage is generated by charging of capacitor C through resistor R for approximately 2 seconds for the values shown. The length of the ramp is determined by the voltage magnitude required to trigger the regenerative switch consisting of Q_1 and Q_2. The temperature sensor consisting of Q_3 and Q_4, together with the controlling thermistor Th, establishes a voltage level at the base of Q_3 which depends upon the resistance value of the thermistor. Q_3 and Q_4 form a bistable multivibrator. The state of the multivibrator depends upon the base bias of Q_3. When Q_3 is conducting, Q_4 is cut off. The pulse generator is energized and generates pulses to trigger the triac. The output of the pulse generator is synchronized to the line voltage on the negative half-cycle by D_2 and R_3 and on the positive half-cycle by D_1 and R_3. The pulses are, therefore, generated at the zero-voltage crossings and trigger the triacs into conduction at only these points.



	120VAC, 60Hz	240VAC, 60Hz
R_1	60kΩ, ½W	120kΩ, ½W
R_2	3.3kΩ, 2W	5.6kΩ, 2W
R_3	33kΩ, ½W	57kΩ, ½W
Y	2N5441 or 2N5444	2N5442 or 2N5445

NOTE: ALL RESISTORS ½W, ± 10% UNLESS OTHERWISE SPECIF. THERMISTOR APPROX. 3000 AT OPERATING TEMP.

Fig. 21 – Proportional integral-cycle heat controller.

●*WITH ACKNOWLEDGEMENT TO RCA*

VACUUM CAPACITORS

Vacuum capacitors, both fixed and variable, are being used today in many types of transmission and industrial equipment. The following note from the English Electric Valve Co. Pty. Ltd. describes in some detail the construction, ratings and application of their wide range of devices.

INTRODUCTION

The use of a high vacuum as dielectric in variable capacitors results in a component with unique characteristics, particularly suited to application in high power oscillator and amplifier circuits. The maximum capacitance that can be provided in a single unit is normally a few thousand picofarads, but the high dielectric strength and freedom from dust contamination, humidity, etc. give the vacuum capacitor a far higher voltage capability than air dielectric capacitors of similar size. Vacuum capacitors have frequently been used to replace existing air dielectric capacitors many times their size, with the additional benefits of greater reliability and easier resetting. The present range of EEV vacuum capacitors have maximum working voltage ratings between 3 and 30kV peak, with current ratings from 7 to 150A.

The construction of a typical EEV vacuum variable capacitor is shown in Fig. 1. Two sets of concentric cylinders form the capacitance, with one set mounted on a sliding shaft. This assembly is enclosed within a vacuum envelope, part of which is a flexible metal bellows which allows axial movement. The envelope also includes an insulating section and in the example shown this is a ceramic cylinder; for many types the envelope material is glass.

Most applications of vacuum capacitors are at radio frequencies where the 'skin effect' confines current flow to the surface layers of the conductors. The current path therefore runs along the metal outer surface of the capacitor to the ceramic, back along the inside and then along the outer surface of the bellows (See Fig. 1). The proportion of the total current which reaches the inner electrode also has to flow along both surfaces of the outer electrodes. The current path to the fixed electrodes is similar to that described above but does not include a bellows section. In order to achieve a low equivalent series resistance it is necessary to use materials of high conductivity for all parts of this current path; oxygen free, high conductivity copper is used wherever possible but a material with better elastic properties such as a bronze with a small phosphorous content is required for the bellows. The outer metal surfaces of the finished capacitor are usually silver plated to give a high conductivity skin and good electrical contact.

CHARACTERISTICS OF VACUUM VARIABLE CAPACITORS

Electrical

The method of construction employed gives a linear relationship between electrode position and capacitance over almost the full range of the capacitor, with a non-linear region at low capacitance values where the two sets of electrodes separate.

The ratio of maximum to minimum capacitance is typically of the order of fifty to one; higher ratios can be achieved, particularly in high capacitance types and some standard capacitors have maximum to minimum capacitance ratios of two hundred to one. These capacitors can give a very wide tuning range from a single unit. Both maximum and minimum capacitance can be set to specified values during manufacture if the full range of a capacitor is not required for a particular application.

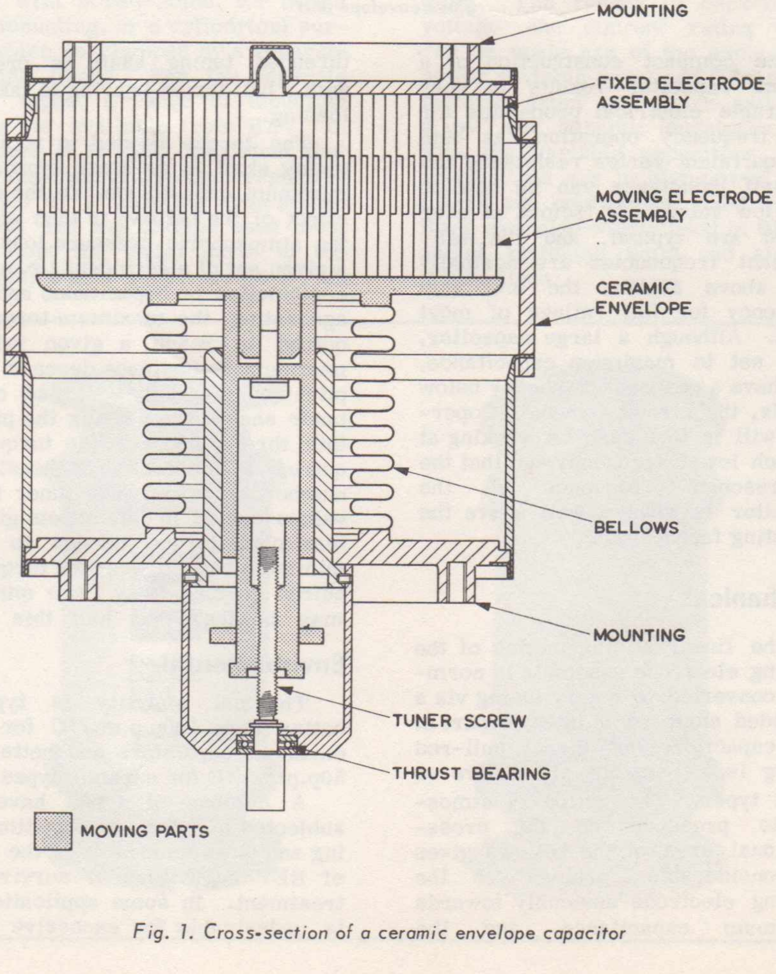
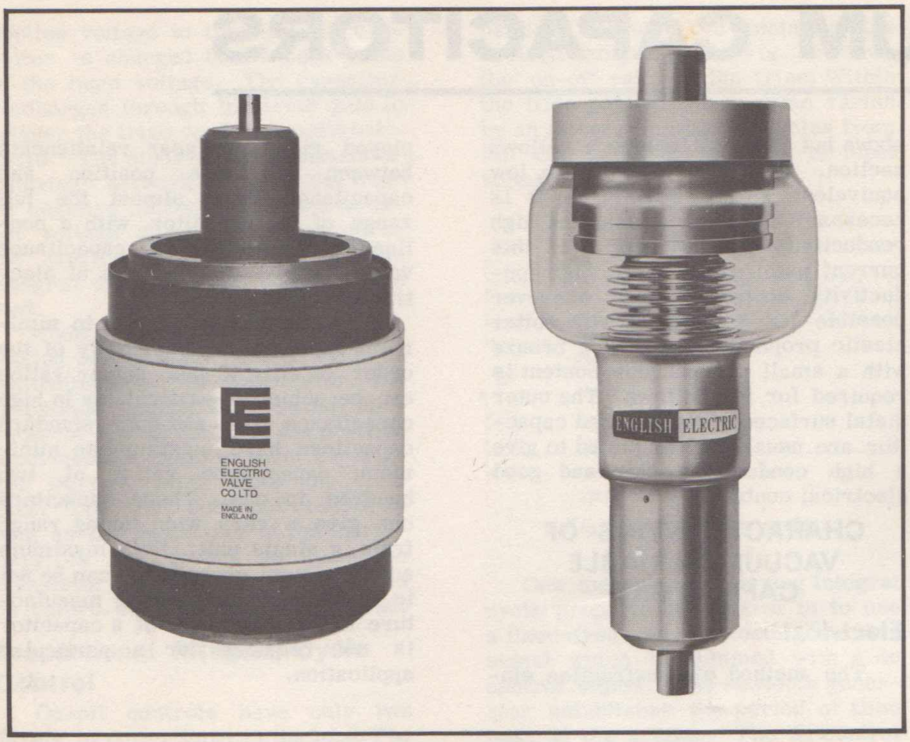


Fig. 1. Cross-section of a ceramic envelope capacitor



Typical EEV variable capacitors; a 150A metal-ceramic type and a 75A glass envelope unit

The compact construction of a vacuum capacitor results in very favourable electrical properties for high frequency operation, as both the equivalent series resistance and the self inductance can be held to very low values. Q factors of over 10,000 are typical, and the self-resonant frequencies are normally well above 30MHz, the maximum frequency for full ratings of most types. Although a large capacitor, when set to maximum capacitance, may have a resonant frequency below 30MHz, the circuit in which it operates will in that case be working at a much lower frequency, so that the self-resonant frequency of the capacitor is always well above the operating frequency.

Mechanical

The linear sliding motion of the moving electrode assembly is normally converted to rotary tuning via a threaded shaft, as an integral part of the capacitor, but direct pull-rod tuning is a possible alternative on most types. The action of atmospheric pressure on the cross-sectional area of the bellows gives a considerable 'preload' to the moving electrode assembly towards maximum capacitance, and the threaded tuning shaft is provided with a ball thrust bearing to take this loading.

The torque needed to rotate the tuning shaft is greatest approaching minimum capacitance, as the spring effect of the bellows is then added to the atmospheric pressure load. For a given set of electrodes, i.e. a given combination of capacitance and voltage rating, the maximum torque required to reach a given value of minimum capacitance depends on the pitch of the screw thread on the tuner shaft. Decreasing the pitch of this thread reduces the torque required, at the cost of increasing the number of turns. The tuner torque values quoted in data sheets give the maximum torque needed to reach minimum capacitance; the torque required to tune away from minimum may be less than half this value.

Environmental

Thermal stability is typically better than 100p.p.m./°C for glass envelope capacitors and better than 50p.p.m./°C for ceramic types.

A number of types have been subjected to extensive vibration testing and have demonstrated the ability of EEV capacitors to survive this treatment. In some applications, it is undesirable for excessive variation in capacitance to be produced by vibration; it has been found that in these capacitors such variations may be less than 1% at 10g acceleration. EEV capacitors are tested to withstand a bump test of 4000 bumps of 40g for 3.5ms and also the Mil. Std. 202 medium impact shock test of 15g for 11ms, with no subsequent change in electrical or mechanical characteristics.

At high altitudes the atmospheric pressure on the bellows is reduced, and above 30,000 feet it is insufficient to push the moving electrode assembly towards the maximum capacitance position against the spring effect of the extended bellows. For satisfactory operation under these conditions a spring can be built into the bellows to provide the required pressure; this has the effect of increasing the torque required to tune to minimum capacitance at lower altitudes. High altitude operation also raises the possibility of voltage breakdown across the outside of the insulator, owing to the loss in dielectric strength of air at low pressures.

RATINGS

The basic rating for a vacuum capacitor is the maximum peak voltage. This is determined by the spacing between the electrodes and is effectively independent of frequency. The voltage rating given in data sheets is the maximum working peak voltage which must not be exceeded in use; each EEV capacitor is tested at a voltage higher than its rated maximum to assure an adequate safety factor in service. When considering voltage rating for vacuum capacitors it is important to be sure that the true peak voltage is used; mean or r.m.s. values are only relevant if the waveform is known. It is also necessary to allow for any d.c. potential which may appear across the capacitor during operation, either by itself or in addition to an r.f. voltage, when deciding the effective peak voltage to which the capacitor will be subjected. The d.c. or d.c. plus r.f. peak voltage must not exceed 80% of the rated peak r.f. voltage. Derating under these conditions is necessary because vacuum capacitors can withstand considerably higher voltages at the peaks of an alternating waveform than with continuous d.c. Since most vacuum capacitors are used under r.f. conditions the voltage rating quoted is for peak r.f. and a derating factor of 0.8 must be used for calculating the corresponding d.c. rating.

At radio frequencies the relatively low reactance of the capacitor results in high currents which, coupled with the skin effect which increases the effective series resistance, can cause excessive power dissipation in the capacitor. For this reason it is necessary to impose a maximum rated value on the r.m.s. current, in conjunction with a limit on the frequency to which the capacitor may be operated at this current. These ratings assume natural cooling, with a local ambient temperature not exceeding 55°C. If forced-air cooling is provided it is possible to increase the current rating, or conversely if an application involves mounting a capacitor where the airflow around it is seriously obstructed, it may be necessary to reduce the rating. The manufacturer of the capacitor should be consulted for information on operation at temperatures above 55°C.

Although the ratings quoted for vacuum capacitors are all absolute limits, there is a short-term tolerance to overload. Current ratings are determined by the temperature rise of the envelope, and the mass of the vulnerable parts allows short-term current overloads to be absorbed, particularly if the current is well below maximum ratings before and after the overload. When the voltage rating is exceeded sufficiently to cause a breakdown between the electrodes, the nature of the vacuum dielectric is such that it is self-healing, provided the energy dissipated in the arc is not excessive. When the capacitor is operating in a tuned circuit, such a breakdown will normally damp the oscillations in the circuit so that the voltage falls quickly and little or no permanent damage results.

STORAGE, INSTALLATION AND MAINTENANCE

Capacitors may be stored in their original packing, in which case no special precautions are necessary. If spare capacitors are stored in open racks the glass or ceramic insulators should be shielded against dust, or cleaned before installation.

Installation of a new capacitor requires resonable care as any damage to the glass or ceramic parts may cause failure of the capacitor. Information on clamping arrangements is given in the following section. A capacitor which is installed after long storage may have a reduced voltage capability at first; this can cause internal sparking when the full operating voltage is first applied, but this effect soon disappears.

Maintenance is confined to cleaning the insulator if it becomes excessively dusty; care must be taken to avoid scratching the envelope while cleaning. The lubricants used in the moving parts of variable capacitors are effective over a wide temperature range.

MOUNTINGS

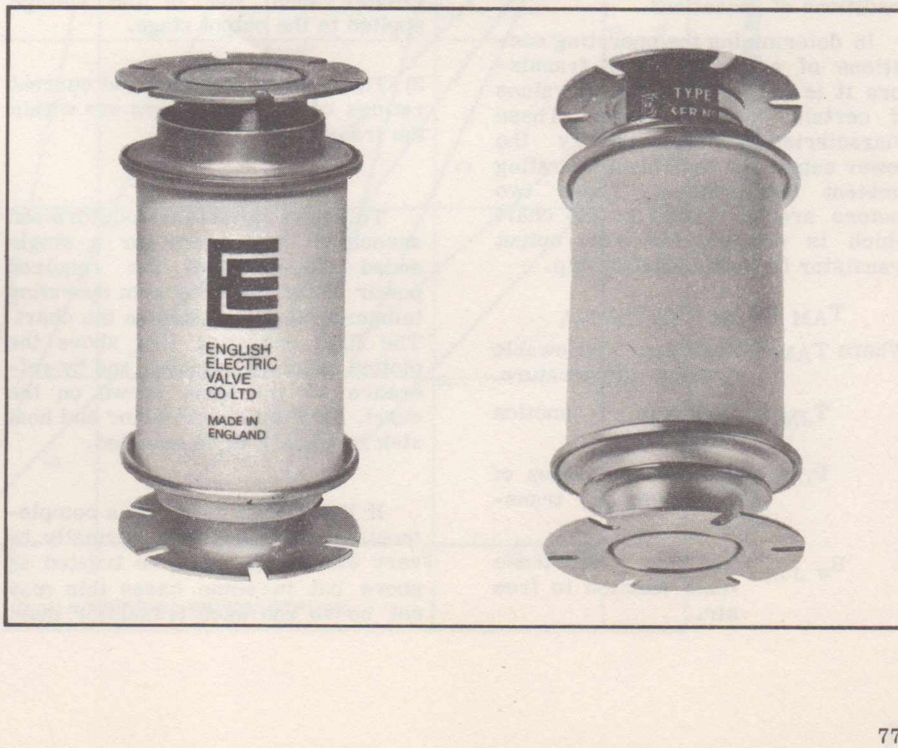
Vacuum capacitors may be mounted in any orientation; it is however necessary to observe precautions, appropriate to the high voltages present, in the placing of other components near the capacitor. The clearances essential to prevent voltage breakdown between the capacitor and nearby components may not necessarily be sufficient to ensure adequate convection cooling, particularly when a large capacitor is being operated near its maximum current rating at high frequencies.

The mounting facilities provided on the capacitor may be either a flange with screw-holes, for direct panel mounting, or a cylindrical surface which is clamped by a separate mounting flange. For the separate flange types, a range of mounting flanges is available from EEV. In all cases it is most important that only the part of the capacitor specified as a mounting area should be used and any clamps used must not distort the metal parts of the envelope, which would create a risk of seal failure. In no case should solder connections be made directly to the body of the capacitor. It is not permissible to clamp any capacitor around the glass or ceramic insulator, or to allow anything to make contact with the insulator during operation.

Variable capacitors are normally mounted by the shaft end, and it is preferable to use a flexible connection to the fixed-electrode end. The connection should be of substantial section so as to keep its losses low and improve cooling of the capacitor by conduction from both ends. Only a small amount of flexibility is needed and a copper strap is suitable. In cases where it is required to mount a capacitor at both ends, excessive strain in the envelope must be avoided.

VACUUM FIXED CAPACITORS

The advantages of the vacuum dielectric also apply to fixed capacitors, with the additional benefit that, having no bellows, the self-inductance and equivalent series resistance are so reduced that for most applications they are negligible. The ranges of capacitance, voltage and current rating which can be made are of the same order as for variable types and EEV manufactures fixed capacitors with both glass and ceramic envelopes. As with the variable capacitors, many applications are in high power amplifiers and oscillators; fixed capacitors are also widely used in filters, aerial matching units and similar applications.



TRANSISTOR DISSIPATION IN A.F. AMPLIFIERS

The operation of transistors in single ended output stages of audio amplifiers using either complementary or similar transistors, normally generates heat. This must be dissipated by the transistor case or by the heat radiating fins if the junction temperatures are to stay within manufacturer's limits. The chart given below will assist in determining the capability of a number of A.W.V. transistors under various conditions of operation.

In determining the operating conditions of a pair of output transistors it is necessary to fix the values of certain characteristics. These characteristics are usually the power output and maximum operating ambient temperature. These two factors are correlated by the chart which is derived for each output transistor from the relationship.

$$T_{AM} = T_{JM} - P_D R_{\theta\ J\text{-}A}$$

Where T_{AM} = Maximum allowable ambient temperature.

T_{JM} = Maximum junction temperature

P_D = Maximum dissipation of output stage per transistor

$R_{\theta\ J\text{-}A}$ = Thermal resistance from junction to free air.

The value of P_D is dependent upon the designed power output at the onset of clipping. This chart has been developed assuming:-

1) The dissipation per transistor is one quarter of the maximum unclipped power output.

2) The peak to peak voltage of the maximum unclipped power output is greater than 90% of the voltage applied to the output stage.

3) The operating voltage and current ratings of the transistors are within the transistor's ratings.

To select suitable transistors and associated heat sinks for a single ended output stage the required power output and maximum operating temperature are plotted on the chart. The first operating line above the plotted point is selected and by reference to the table shown on the chart, the type of transistor and heat sink required can be selected.

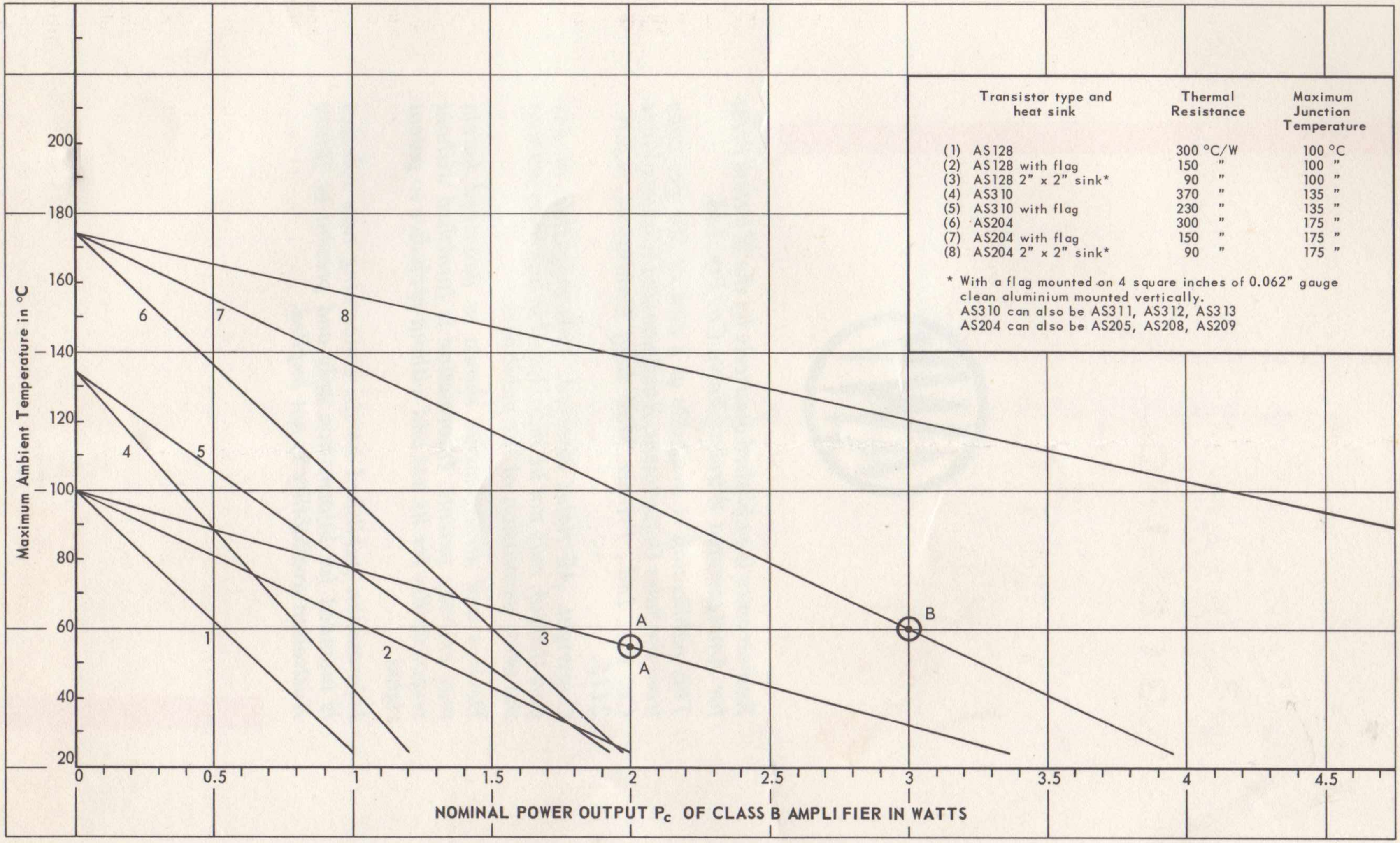
If the design calls for a complementary pair they will normally be very similar and will be treated as above but in some cases this may not be so and each transistor must be treated individually. Although they may have very different thermal characteristics, the dissipation per transistor in a given amplifier is not materially affected since this is primarily a function of the maximum unclipped power output. In such a case the transistors to be used and their associated heat sink, if required, must both lie on the side of the above determined operating point opposite the origin. The following two examples show how the chart can be used.

	Required Power Output	Required Maximum Ambient Temperature
Case 1	2 Watts	55 degrees C.
Case 2	3 Watts	60 degrees C.

Now following plotting the above characteristics on the chart (points A & B resp.) the selection would be as follows:

Case 1 — A complementary pair — One AS128 with flag and 2"×2" heat sink — One AS204 with flag.
Single ended application — Two AS204 with flags.

Case 2 — A complementary pair not possible with listed types.
Single ended application — Two AS204 with flags.



	Transistor type and heat sink	Thermal Resistance	Maximum Junction Temperature
(1)	AS128	300 °C/W	100 °C
(2)	AS128 with flag	150 "	100 "
(3)	AS128 2" x 2" sink*	90 "	100 "
(4)	AS310	370 "	135 "
(5)	AS310 with flag	230 "	135 "
(6)	AS204	300 "	175 "
(7)	AS204 with flag	150 "	175 "
(8)	AS204 2" x 2" sink*	90 "	175 "

* With a flag mounted on 4 square inches of 0.062" gauge clean aluminium mounted vertically.
AS310 can also be AS311, AS312, AS313
AS204 can also be AS205, AS208, AS209

TRANSISTOR OPERATING TEMPERATURES AND POWER OUTPUT CHARACTERISTICS.

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