RADIOTRONICS

IN THIS ISSUE

COVER

Picture Tube Gun Assembly at the AWV Works at Rydalmere.

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FUNCTION DELAY TIMER

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When an application requires a time delay between the operation of the power control switch and the actual application of the power to a device, then some form of delay system must be incorporated. This type of switching system has many applications, some of which are listed below:

1) Power control of heater systems of various specialised types.

2) The operation of fans to circulate air or gas after other functions have been completed.

3) Safety device delay systems in electronic equipment.

4) The delayed control of gases by solenoid valves after other functions have been completed.

5) Incorporation in part of safety systems in domestic and industrial equipment.

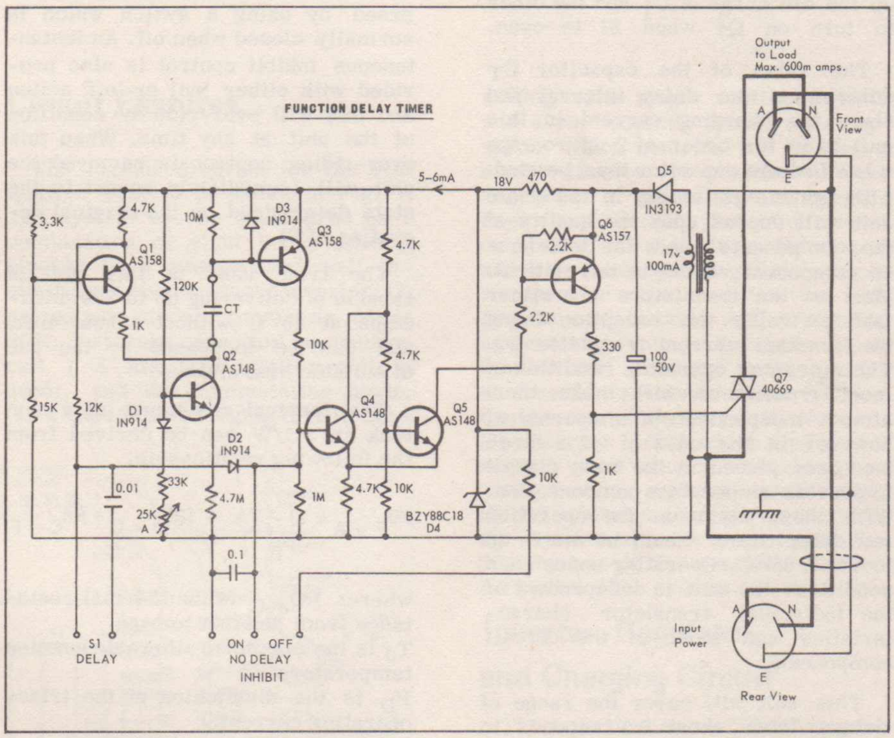
6) The coupling of several units in cascade for sequence controls.

The circuit illustrated produces a predetermined delay before applying the power to the load. In this case the design has been developed to control the power mains but can be readily applied to other devices. The delay is determined by the actual value of the capacitor C_T and the components have been adjusted so that the delay time, measured in seconds, is given by the relationship.

$$T_D = 10C_T$$

Where C_T is in microfarads.

To compensate for component tolerances a variable resistor R_T has been provided but this and the adjacent series resistor can be replaced by a single fixed resistor if required.



The circuit of the unit can be divided into three main sections, and a description of each unit is as follows:

1) THE TIME DELAY CIRCUIT CONTAINING Q2 AND Q3 WITH THE INHIBITING TRANSISTOR Q1.

The Q2 collector is a constant current source used to charge the timing capacitor C_T in a linear manner. When it is charged to the available voltage, this charging current, which also holds on the transistor Q3, ceases and Q3 turns off. The timing system is completely discharged at any instant by the turning on of Q1 which, in turn, is controlled by the external switch S1. Under these conditions the opening of S1 at any time will almost instantaneously reset the time delay to the full interval selected.

2) A SCHMITT TRIGGER CONTAINING Q4 AND Q5, WHICH CONTROLS THE THIRD SECTION.

The Schmitt trigger is controlled by the collector current of Q3, which while it exists during the charging of C_T, holds Q4 on, which in turn holds Q5 off. As soon as the collector current of Q3 falls towards a low value near zero, the Schmitt trigger reverses and the collector current of Q5 rises to its maximum value.

3) A CONTROL TRANSISTOR Q6 WHICH TURNS THE TRIAC ON AND APPLIES THE POWER MAINS TO THE LOAD.

Q6 turns on as the collector current of Q5 rises to its maximum value. This then applies a current to the gate of the triac Q7 via a series limiting resistor and turns on the triac applying the power to the load circuit. A simple power supply is also included with zener control for the first two sections which assists the suppression of unwanted transients. The small capacitors are placed to suppress other transients and two diodes are used, one to speed up the discharge of C_T and the other to turn on Q4 when S1 is open.

The value of the capacitor C_T determines the delay interval and since the charging current in this unit is in the order of 2 microamps a low leakage capacitor must be used. Although the reliability of the whole unit will depend upon the quality of the components used, the tolerance on component values is not critical. Most of the transistors are either 'on' or 'off', the exception being the constant current transistor Q2. The saturated operating condition of most transistors will make them almost independent of temperature, however in the case of Q2 a diode has been placed in the base circuit to provide temperature compensation. With these features, the operation and delay times should be stable up to about 65° C. Operating under such conditions the unit is independent of the individual transistor characteristics and most of the circuit components.

This unit will cover the range of delays from about 0.1 seconds to several tens of seconds. The delay times can be extended to several minutes by using a lower constant charging current and a darlington coupled pair in place of Q3. The value of C_T can also be increased, provided its leakage is low, to further extend the delay time.

The maximum delay time will be limited by the transistor and capacitor leakage currents, but delays of the order of 20 minutes should be possible. It should be noted that as the charging current is reduced, problems may arise from RF interference and other spurious signals.

The power supply transformer must supply about 100 milliamps when the triac Q7 is on, but under quiescent conditions drain is less than 20 milliamps. The voltage rating of the main filter capacitor should be conservative and will be dependent upon the peak voltage of the secondary of the transformer. The secondary voltage of the transformer is about 17 volts, but can be varied. Allowance should be made for about 5 to 10 milliamps to be bled through the zener diode D4 and several volts to appear across the filter resistor. The 470 ohm shown can be adjusted to suit other transformer voltages or any external power supply desired.

The delay control of the unit described is normally adjusted to turn on the power following the closing of switch S1. The effect can be reversed by using a switch which is normally closed when off. An instantaneous inhibit control is also provided with either 'on' or 'off' action and this will over-ride the condition of the unit at any time. When this over-riding control is removed the unit will immediately revert to the state determined by the original operation of S1.

The triac used in this unit is capable of delivering up to 600 milliamps at 65° C without a heat sink, and can be increased by the use of suitable heat sinks.

The thermal resistance of a heat sink in °C/W can be derived from the following relationship.

$$R\theta_{F-A} = \frac{Tj - TA}{P_D} - (R\theta_{J-C} + R\theta_{C-F})$$

where: RO_{J-C} is the thermal resistance from junction to case.
T_J is the maximum allowable junction temperature.
P_D is the dissipation at the triacs operating current.

The above characteristic values are obtainable from the triacs data sheet.

T_A is the maximum required ambient temperature. $R\theta_{C-F}$ is the thermal resistance from case to fin. The maximum values for $R\theta_{C-F}$ are

2.0° C/W for the 40669 plastic assembly.

1.0° C/W for a T066 assembly.

0.5° C/W for a T03 assembly.

The table given below indicates the maximum sizes of the heat sinks needed for higher dissipation and are based on the thermal resistance for the 40669 triac, mounted with mica washer and silicon grease.

The following explanation covers the notations associated with some of the heat sink values:

A. No heat sink is required.

B. The thickness has been increased to 10 gauge (1/8")

C. The values stated are the thermal resistance of the heat sink.

D. It is not practical to operate the triac under these conditions.

The values given in the table are the total surface areas in square inches, including both sides, of 16 gauge (1/16") natural finished vertically mounted aluminium sheet.

Conclusion

The circuit described will provide delay times from about 0.1 seconds to several tens of seconds with provision for longer delays. It is almost independent of circuit constants and temperature variations up to 65°C. Provision has been made in the circuit for over-riding control and many other inhibit or logic functions can be applied which will all enhance its versatility.

Operating Current in AMPS	Maximum Ambient Temperature TA °C				
	40	45	55	60	65
1	_A	_A	2	4	6
2	10	12	15	18	20
3	16	18	22	25	32
4	22	25	35	45	50
6	40	50	85B	2.5C	1.5C
8	100B	2.0C	1.0C	0.5C	_D

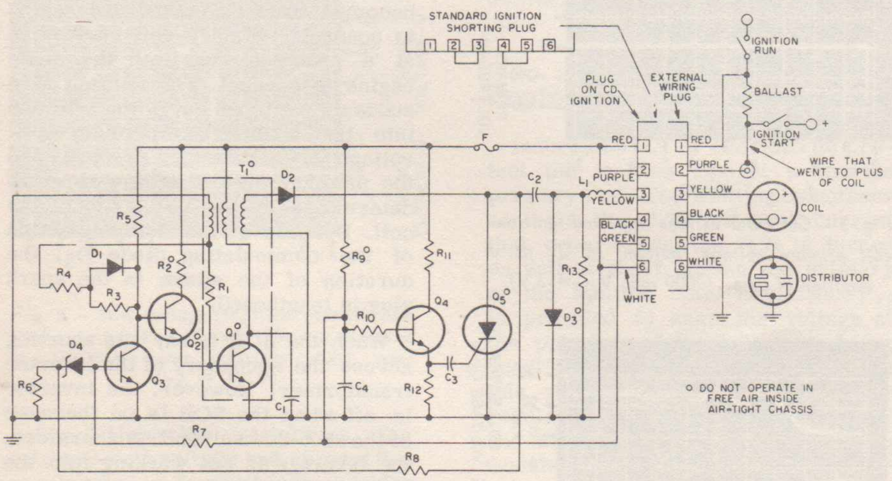
An SCR Capacitor-Discharge Ignition System with Regulated, Single-Ended Inverter

by M. S. FISHER

This Note describes a low-cost capacitor-discharge (CD) ignition system for passenger automobiles that uses a silicon controlled rectifier (SCR) in the output circuit. This system provides the advantages of reduced maintenance, low battery current drain, and full output voltage at low battery voltage (down to 4 volts) for starting under low-battery-power conditions. In addition, it provides better firing of fouled spark plugs because the high-voltage output pulse has a faster rise time.

Circuit Features

The circuit diagram for the SCR ignition system is shown in Fig. 1. The system can be considered as a combination of eight basic building blocks: (1) a single-ended, self-oscillating, swinging-choke inverter; (2) an output circuit consisting of an SCR, a storage capacitor, an ignition coil (a standard ignition coil is used), and the commutating diode; (3) a capacitor-charging circuit consisting of a single rectifying diode and a commutating diode; (4) a regulator which controls the frequency of the inverter to provide efficient regulation of the voltage on the storage capacitor; (5) a protection circuit consisting of a limiting inductance and a resistance that guards against an open or shortened ignition coil (the circuit is also self-protecting against high-voltage arcing to either ignition-coil primary terminal); (6) an inverter shutdown which disables the inverter while the points are open; (7) a trigger circuit which suppresses normal point bounce and also prevents residual voltage across the closed points from triggering the SCR; (8) a method of cummutating the SCR by use of the interplay of several parts of the system.

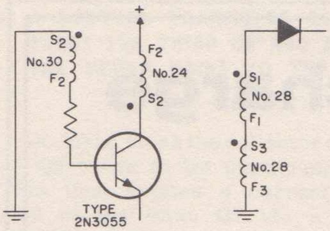


Q_1 : 2N3055	C_1 : 0.25 μF, 200 V	R_4 : 1000 ohms, 1/2 W
Q_2 : 2N3053	C_2 : 1 μF, 400 V	R_5 : 18000 ohms, 1/2 W
Q_3 : 2N3241	C_3 : 1 μF, 25 V	R_6 : 15000 ohms, 1/2 W
Q_4 : 2N3241	C_4 : 0.25 μF, 25 V	R_7 : 8200 ohms, 1/2 W
Q_5 : RCA 40657	F : 5A	R_8 : 0.39 megohm, 1/2 W
D_1 : 1N3193	L_1 : 10 μH, 100 T of No.28 wire wound on	R_9 : 220 ohms, 1 W
D_2 : 1N3195	a 2-W resistor (100 ohms or more)	R_{10} : 1000 ohms, 1/2 W
D_3 : 1N1763A	R_1 : 1000 ohms, 1/2 W	R_{11} : 68 ohms, 1/2 W
D_4 : 12 V, 1/4 W	R_2 : 35 ohms, 5 W	R_{12} : 4700 ohms, 1/2 W
	R : 22000 ohms, 1/2 W	R_{13} : 2700 ohms, 1/2 W

Fig. 1 – An SCR capacitor-discharge ignition circuit.

Inverter, Regulator, and Charging Circuit

The inverter uses a 2N3055 transistor Q_1 in the single-ended output stage, a 2N3053 transistor Q_2 in an emitter-follower driver stage, and a 2N3241 transistor Q_3 in the control stage. Regenerative feedback is provided to the bases of Q1 and Q2 from the feedback winding on transformer T_1 (details concerning transformer winding are described in Fig. 2). The high gain of the Q_1 and Q_2 combination assures oscillation and low drive-power requirements. Starting resistor R5 biases Q_1 and Q_2 into conduction to start oscillation. When Q_1 is saturated, the battery voltage across the primary winding of T_1 increases the current in the transformer. The reflected voltage in the feedback winging produces a current flow in R_4 , R1, and D_1 to hold Q_1 in saturation;



Transformer T_1 is wound as follows:
A 1/2-in. bobbin and E1 stack of grain-oriented silicon steel are used; first, 150 turns of No.28 wire are wound and labeled S1 and F1 on the winding; second, 50 turns of No.24 and No.30 wire are wound bifilar and labeled S2 and F2; third, 150 turns of No.28 wire are wound and labeled S3 and F3. All windings are wound in the same direction. A total air gap of 70 mils (35-mil spacer) is used. Connections are made as shown above.

Fig. 2 – Transformer winding detail.

simultaneously, the rectifying diode D_2 in the secondary winding is reverse-biased.

When the collector current in Q_1 reaches a level where it cannot further increase, the feedback voltage reduces, and Q_1 starts to turn off. Regenerative action of the circuit cuts off the base drive of transistors Q_1 and Q_2 and causes a "flyback" pulse of voltage at the collector of Q_1. Diode D_1 blocks the reverse voltage and limits the reverse base drive. The reverse bias current that turns off Q_1 and Q_2 is applied through resistors R_1 and R_3, respectively. The flyback pulse is stepped up in voltage on the secondary of T_1 and diode D_2 becomes forward-biased because of reversal of secondary-voltage polarity. As a result, the energy previously stored in the primary of T_1 is transferred through the secondary winding to the storage capacitor C_2, charging the capacitor. Capacitor C_1 reduces the amplitude of the leakage inductance pulse, and also the rate of voltage rise at the collector of Q_1. Charging current through capacitor C_2 is shunted out of the ignition coil by commuting diode D_3 so that no energy is transferred into the ignition coil and out into the load.

When the collector voltage of Q_1 falls below the battery voltage, Q_1 turns back on and the cycle repeats, further charging the capacitor C_2. During the time that C_2 is below a predetermined value, the voltage applied to zener diode D_4 through voltage dividers R_8 and R_6 is not sufficient to cause the zener diode to conduct. If the ignition points are also closed, resistor R_7 is returned to ground and transistor Q_3 is held off.

When the capacitor voltage reaches a level that causes zener diode D_4 to conduct, transistor Q_3 turns on and shunts the base drive current from Q_2. This effect reduces the base drive at Q_1 and causes it to pull out of saturation at a lower collector-current level which, in turn increases the frequency of oscillation. The cutback in peak primary current reduces the charging rate of C_2 to the level required to supply the losses (SCR leakage, etc.) and holds the output voltage from further rise.

Transistor Q_3 also shuts down the inverter when the ignition points are open. When these points open, current is fed from the voltage at the points through resistor R_7 to turn Q_3 on hard. This effect shorts the base of Q_2 and stops the oscillation.

Fig. 3 shows the collector voltage at Q_1 alternately saturating and then going through a "flyback" pulse of increasing amplitude. The

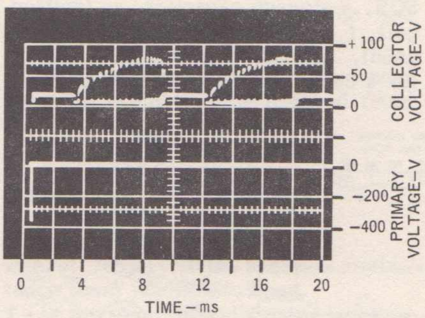


Fig. 3 – Collector voltage (top) and ignition-coil primary voltage (bottom) of Q_1 as a function of time. (2000 rpm; V_{CC}=12 V)

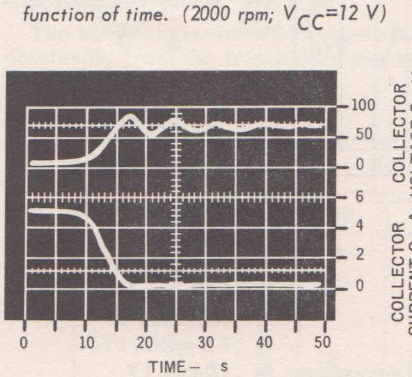


Fig. 4 – Expanded collector voltage (top) as a function of time, and collector current (bottom) as a function of time during turn-off Q_1 as the capacitor C_2 is being charged. (V_{CC}=12 V)

change in frequency as a result of regulator action can be seen. The collector voltage then drops to the supply voltage when the ignition points open and shut down the inverter. Fig. 4 shows an expanded view of the turn-off and flyback characteristic of Q_1 at a point when the capacitor is being charged.

Output Circuit

When a high-voltage pulse is required, the SCR (Q_5) is gated on. As a result, the anode voltage of the SCR drops to zero, and the voltage across the capacitor C_2 is applied to the primary of the ignition coil. (The inductance of L_1 is negligible when compared to the ignition-coil inductance in this analysis). The hot side of the ignition-coil primary (terminal 3 on the connecting plug) is driven negative to the capacitor potential. Diode D_3, in parallel with the coil, is reverse-biased at this time. The discharge of the capacitor into the primary of the coil generates a high-voltage pulse on the secondary.

The capacitor discharges sinusoidally into the primary inductance of the ignition coil, building up the primary current in the coil. When the capacitor (and coil primary) voltage reaches zero and starts to reverse, the commutating diode D_3 becomes forward-biased and begins to conduct. Primary-coil current is at a peak at the time the diode begins to conduct. The current then suddenly switches out of the SCR and into the diode. The primary-coil voltage stays clamped at zero, and the primary current decays at a rate determined by the L/R ratio of the coil. Because of the clamping action of the commutating diode D_3, the duration of the spark in the spark plug is lengthened.

When the SCR is on, it is shorting across the secondary of the inverter transformer. However, the inverter is off when the SCR is on (because of the shut-down circuitry); therefore, the inverter is not working into the short.

Figs. 5 and 6 show the SCR voltage and current as a function of time. In Fig. 5, the oscilloscope is triggered at the time the ignition points open. The anode voltage of the SCR drops to zero and its anode current builds up to the peak in a quarter-cycle. The current is then switched out of the SCR, and SCR current drops suddenly almost to zero. The small residual current is a result of the energy stored

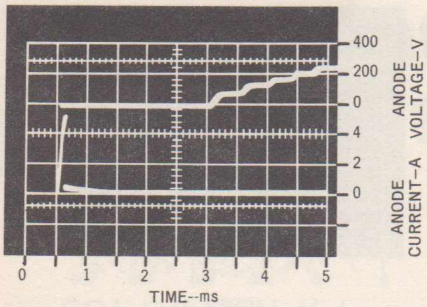


Fig. 5 – SCR voltage (top) and current (bottom) as a function of time. (2000 rpm; $V_{CC} = 12$ V)

in the inverter transformer when the inverter shuts down. This stored energy causes a current to circulate from the secondary of the transformer through the SCR. When the ignition points close and capacitor C_2 recharges, the SCR blocks the voltage on the capacitor.

In Fig. 6, the oscilloscope is triggered at the time the ignition points close. The time between the closing of the points and the initial rise of the voltage is the time for the collector current of Q_1 to build up to the switching level. The significance of this time will be explained when commutation of the SCR is discussed.

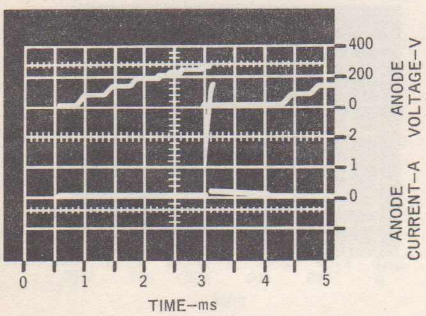


Fig. 6 – SCR voltage (top) and current (bottom) as a function of time. (Sweep triggered at ignition point closing; 4000 rpm; $V_{CC} = 12$ V)

Fig. 7 shows the waveforms for primary-coil voltage and current that result when a 7-millihenry inductor is used to simulate the primary of the ignition coil. In Fig. 7, the primary voltage moves rapidly to a peak negative value, then decays sinusoidally to zero as the current builds to a peak. The primary voltage is then clamped at zero and the current decays exponentially.

It should be noted that an actual ignition coil, operated with the secondary open, will reflect a tuned circuit into the primary. This operation causes a ringing on top of the waveforms shown in Fig. 7. The anode voltage of the SCR may actually reverse for a short time as a result of this ringing. The SCR will essentially block this reverse voltage except for a small current that will flow because of the presence of a positive gate signal. There will be some instantaneous dissipation in the reverse-blocking junction of the SCR for this short period of time which the device is capable of handling. The gate signal is kept positive through the ringing cycle so the SCR will continue to conduct when the anode voltage rings back positive. This ringing does not occur when the secondary voltage fires a plug because the ionized plug shorts the secondary winding.

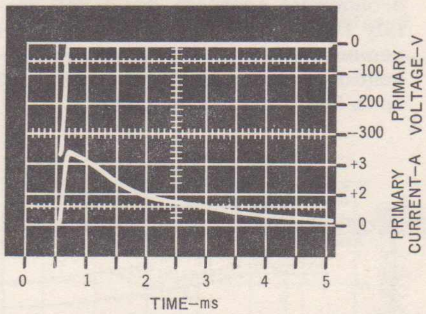


Fig. 7 – Primary voltage (top) and current (bottom) as a function of time. (7-mH dummy coil, 2000 rpm; $V_{CC} = 12$ V)

Protective Circuits

Inductance L_1 is provided to protect the ignition against a shorted primary coil. The limiting inductance controls the di/dt and peak current that occur when the SCR is turned on with a short across the primary of the ignition coil. Resistance R_{13} is provided to keep the voltage at the primary of the coil from being negative when the coil is open. If this voltage were not clamped, the regulator would not operate properly and the peak collector voltage at Q_1 would exceed the limits specified for the device.

Trigger Circuit

The functions of the triggering circuitry are as follows: (1) triggering and holding the SCR on when the ignition points open (at battery voltages down to 4 volts), (2) feeding signal back into the inverter shutdown circuitry when the ignition points are open, (3) suppressing the inverter signal that is riding on the power supply so it does not trigger the SCR, (4) preventing the residual voltage across the closed points from triggering the SCR, (5) preventing normal point bounce at point closure from false-triggering the SCR, and (6) maintaining proper operation whether or not a capacitor is present across the breaker points. Transistor Q_4 is used to perform these functions.

The trigger current for the gate of the SCR is initiated when the base voltage of Q_4 is approximately 0.6 volt above the emitter voltage. The trigger current flows from the supply through R_{11}, Q_4 and C_3 to the gate of the SCR.

When the ignition points open, capacitor C_4 (and the capacitor across the points) charges because of the current through R_3. If the points are open long enough (without bouncing), the voltage across C_4 becomes high enough to turn on Q_4. The voltage required to turn on Q_4 is the sum of the gate-cathode voltage of Q_5, the voltage across C_3, the emitter-base voltage of Q_4, and the drop across R_{10}. (R_{10} is provided so that the voltage across the open ignition points becomes high enough to feed sufficient current through R_7 to shut down the inverter.) At normal speeds, there is an average voltage level across C_3, the emitter side of the capacitor being positive with respect to the gate side of the capacitor. This voltage keeps both the gate-cathode junction of the SCR and the emitter-base junction of the transistor reverse-biased until C_4 charges high enough to turn on Q_4 Because, when the points are closed, Q_4 is off and the gate of Q_5 is reverse-biased, the desired suppression of the inverter signal and the residual point voltage is achieved.

If the points are bouncing in normal operation, they will discharge C_4 (and the distributor capacitor) almost instantly each time they close. Thus, each time the points bounce back open, these capacitors must start recharging from zero toward the triggering level. With normal bouncing, the points will not stay open long enough for the triggering level to be reached, and the SCR will not be triggered. If severe bouncing occurs at very high speed, the points can stay open long enough to cause triggering of the SCR.

Better filtering is achieved when the capacitor is used in the distributor, but satisfactory operation occurs without it. With the capacitor left in the distributor, it is possible to switch back to standard ignition by switching the plug shown in Fig. 1.

Commutating the SCR

All the parts of the system work together in such a manner as to cause the SCR current to go to zero for a sufficient length of time to cause commutation (turn-off).

As explained earlier, primary-coil current is switched out of the SCR and into the commutating diode when primary voltage drops to zero. From that time on, the diode keeps the coil current clamped out of the SCR. The SCR then carries only the small current resulting from the energy stored in the inverter transformer. When the points close again, the inverter starts back up, and, as explained previously, the rectifying diode D_2 is back-biased during the time the collector current of Q1 builds back up to the switching level. At this point, there is no current in the SCR. If current is left at zero long enough, the SCR is left in the blocking state for re-applied positive anode voltage. As shown in Fig. 6, the current is zero for about 300 microseconds before anode voltage is re-applied at 0.5 volt per microsecond. A worst-case SCR will commutate in less than 100 micro-seconds at 100°C under these operating conditions.

Performance

Fig 8 shows several performance curves for the capacitor-discharge ignition circuit. Fig. 9 shows the open-circuit output voltage as a function of time. Fig. 10 shows the out-put voltage when the secondary is loaded with a 1-megohm resistor in parallel with a 50-picofarad capacitor.

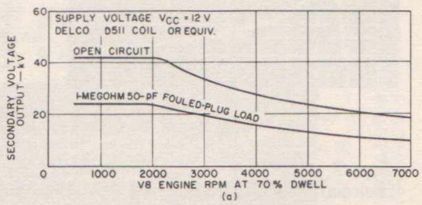
Figs 11 and 12 show secondary voltage under sparking conditions. Arc duration in Fig. 11 is 300 micro-seconds (single polarity) under wide-gap conditions. The narrower gap of Fig. 12 yielded an arc duration of 400 microseconds.

Mounting Considerations

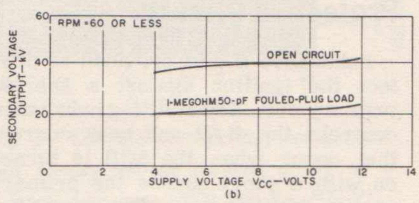
The circuit should be contained in a water-tight environment. However, heat-generating components should not be enclosed in a non-circulating atmosphere because still air has a thermal resistance 12,000 times that of copper. Generation of heat in this high thermal resistance could cause the inside ambient temperature to rise above the specified limits. All components marked "Do not operate in free air inside air-tight enclosure" should be thermally connected to a thermal conductor which makes a low thermal-resistance path to the outside environment. For example, the SCR should be mounted to an aluminum plate on a mica insulating washer. This plate should then be fastened to the inside of the chassis wall to provide the thermal path to the outside environment. The resistors and diodes which need heat sinks, as shown in Fig. 1, should be attached to the chassis with a thermally conductive epoxy.

Conclusion

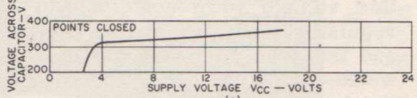
Use of this circuit provides an economical approach which will satisfy the stringent performance requi rements of ignition systems in conventional passenger cars as well as in high-performance cars.



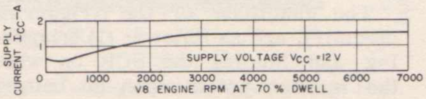
(a) Output voltage as a function of engine rpm at 12 V for both an open secondary and a fouled-plug load.



(b) Output voltage as a function of battery voltage at cranking speeds for both an open secondary and fouled-plug load.



(c) Regulation curve showing peak capacitor (and SCR) voltage as a function of battery voltage.



(d) Battery drain as a function of engine rpm at a battery voltage of 12 V.

Fig. 8 – Performance of the capacitor-discharge ignition circuit.

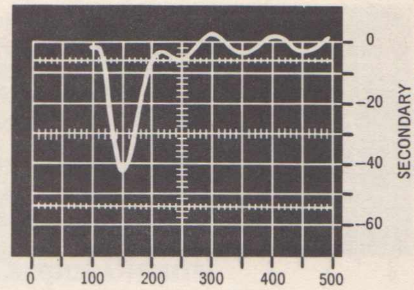


Fig. 9 – Open-circuit output voltage (Delco D511 standard ignition coil or equivalent; 2000 rpm; V_{CC} = 12V)

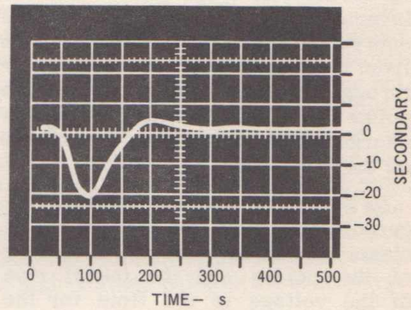


Fig. 10 – Fouled-spark-plug output voltage (Delco D511 standard ignition coil or equivalent; 50-pF load in parallel with 1 megohm; 2000 rpm; V_{CC} = 12 V)

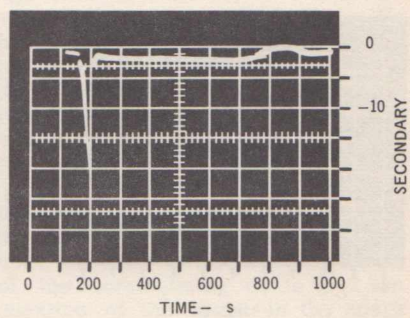


Fig. 11 – Output voltage showing spark-arc duration (Delco D511 standard ignition coil or equivalent; 2000 rpm; V_{CC} = 12 V)

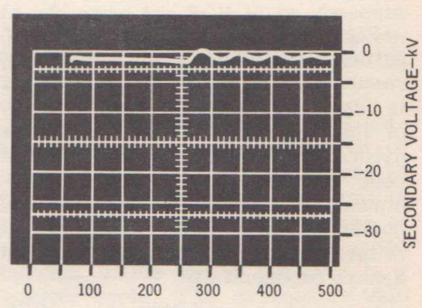


Fig. 12 – Output voltage with spark gap shortened (Delco D511 standard ignition coil or equivalent; 2000 rpm; V_{CC} = 12 V)

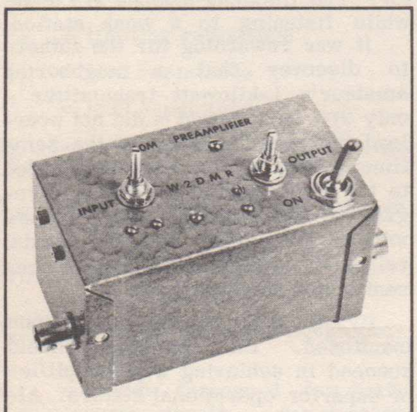
A Dual-Gate MOS-FET Preamplifier for the 10-Meter Band

BY G. E. YEWDALL
& D. W. NELSON,
RCA CAMDEN, N. J.

AUTHORS' PREFACE: Older-type receivers frequently lack the gain necessary to ferret out weaker signals on the 10-meter band. An ideal solution to this problem is provided by an inexpensive, easily constructed preamplifier which exploits outstanding performance characteristics of RCA's recently developed dual-gate metal-oxide-semiconductor (MOS) field-effect transistor. The preamplifier discussed in the article which follows boasts a gain of 26 dB without special neutralization. A noise figure of 2 dB can be appreciated when quiet conditions exist.

A dual-gate field-effect transistor, such as the RCA-3N140 used in the preamplifier described in this article, is equivalent electrically to two single-gate transistors connected in cascode and enclosed in the same package. In some respects, the resulting transistor resembles a tetrode tube; however, the main intent in using a dual-gate transistor in the preamplifier is to provide an inexpensive cascode circuit that offers maximum resistance to cross-modulation from nearby stations.

In Figure 2 are illustrated three evolutionary stages of the cascode amplifier designed to reduce cross-modulation distortion. Illustration "a" shows a tube circuit that was widely acclaimed for its superior cross-modulation reduction. The two single-gate MOS field-effect-transistor equivalent of the tube circuit is shown in illustration "b". Finally, in illustration "c", is the dual-gate MOS field-effect-transistor amplifier - or electrical equivalent of the cascode circuits - which provides the basis for the 10-meter preamplifier constructed by the authors.



Circuit Operation

Figure 3 shows the circuit schematic and parts list of the 28-30-MHz preamplifier. Figure 4 illustrates the basing diagram of the dual-gate MOS-FET transistor. Gate 1 (Lead 3) is forward-biased by R_1 and R_2 to raise its quiescent potential above ground.

Inspection of the circuit shows that the value of the source resistor, R_6, is large enough so that Gate 1 will always be negative with respect to the source. You have

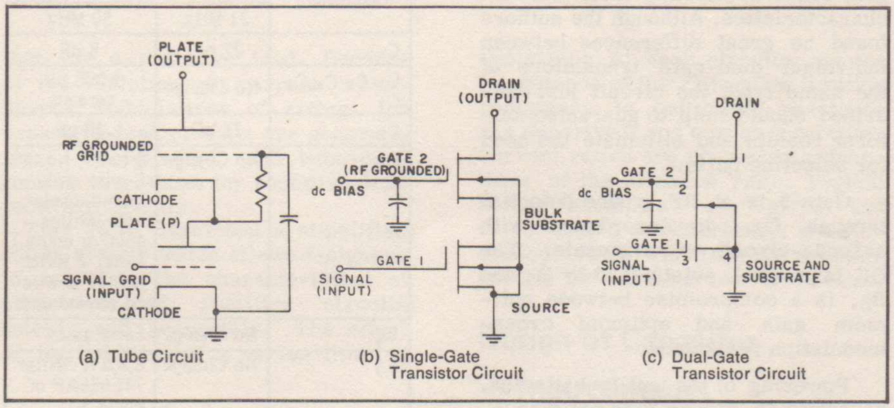
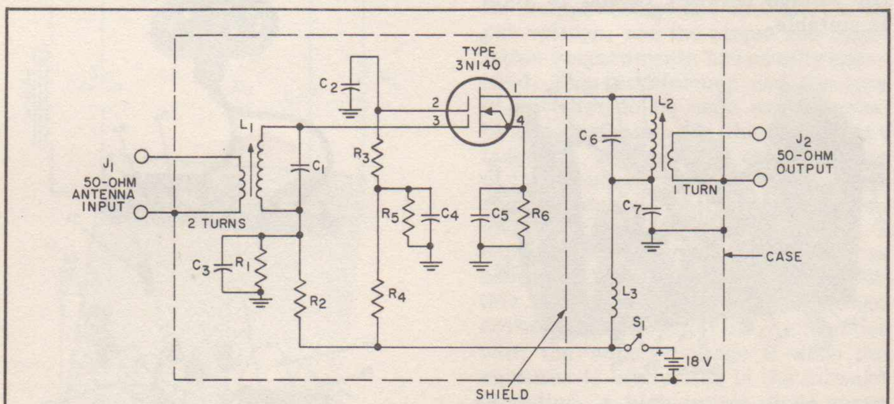


Figure.2: Evolutionary stages of a cascode amplifier designed to reduce cross-modulation distortion.



C_1 — 8 picofarads, mica or ceramic tubular
C_2, C_3, C_4, C_5, C_7 — 0.01 microfarad, ceramic
C_6 — 10 picofarads, mica or ceramic tubular
R_1 — 27,000 ohms, ¼ watt, 10%
R_2 — 150,000 ohms, ¼ watt, 10%, carbon
R_3 — 1,800 ohms, ¼ watt, 10%, carbon
R_4 — 100,000 ohms, ¼ watt, 10%, carbon
R_5 — 33,000 ohms, ¼ watt, 10%, carbon
R_6 — 270 ohms, ¼ watt, 10%, carbon
L_1, L_2 — 1.6 to 3.1 microhenries, adjustable
L_3 — 22 microhenries
Q_1 — 3N140 MOS field-effect transistor
S_1 — Toggle switch, single-pole, single-throw
J_1, J_2 — Coaxial receptacle (Amphenol BNC type UG-1094 or equiv.)
Miscellaneous — Two batteries and one case

Figure 3: Schematic diagram and parts list for 10-meter preamplifier circuit.

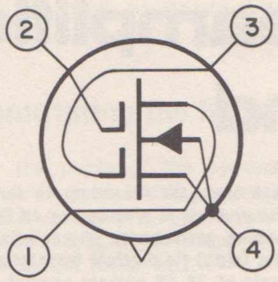


Figure 4: Base diagram of dual-gate MOS field-effect transistor.

probably recognized the resemblance of this configuration to that of an old tube circuit which was used to equalize gain differences in high-gain tubes by shifting their transfer characteristics. Although the authors found no great differences between individual dual-gate transistors of the same type, the circuit just described should help to guarantee uniform results and eliminate the need for selecting parts.

Gate 2 is at RF ground potential through C_2, in accordance with cascode-circuit requirements. The DC bias level, established by R_4 and R_5, is a compromise between optimum gain and optimum cross-modulation resistance.

Powering of the unit by batteries, as shown in Figure 3, is not mandatory. Any reasonably well-filtered DC voltage between 15 and 18 volts is suitable.

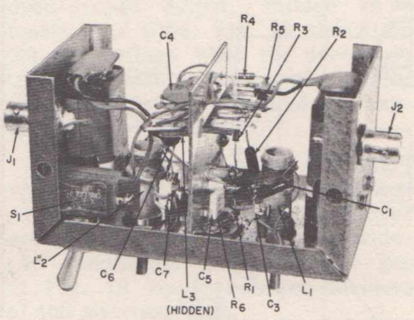


Figure 5: In this photo showing interior of 10-meter preamplifier, MOS field-effect transistor is obscured by the ceramic-standoff insulators mounted on the center partition.

Adapting the Preamplifier To Other Frequencies

The 3N140 has excellent performance characteristics up to 200 MHz. Consequently, the circuit can be used at higher frequencies with only a few changes (see Table I). For example, both tanks in the preamplifier circuit can be made to tune to 21 MHz (15 meters) by changing only C_1 and C_6 to 22 picofarads.

It must be remembered that wiring becomes critical at 50 MHz, and even more critical at 144 MHz. Bypass-capacitor leads and all leads carrying RF signals should be made as short as possible. A well-constructed circuit will show only a slight degradation of the 26-dB gain and the 2-dB noise figure at 50 MHz. At 144 MHz, the authors have achieved gains in excess of 20 dB with noise figures of 2.8 dB.

Table I – Values of Circuit Components For 21 and 50 MHz

Component	Value	
	21 MHz	50 MHz
C_1	22 pF	8 pF
C_2, C_3, C_4, C_5, C_7	No Change	1,000 pF, ceramic
C_6	22 pF	10 pF
L_1	No Change	8 turns, No. 30 E wire on ¼-inch-diameter core (Miller 4500 or equiv.) Link: 2 turns, No. 30 E wire on ground end.
L_2	No Change	Same as L
L_3	No Change	6.8μH (Miller 74F686AP or equiv.)

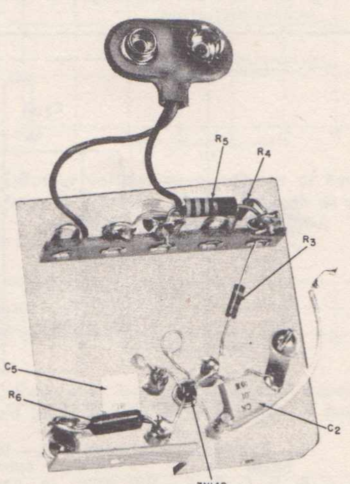


Figure 6: Detailed view of preamplifier's center partition shows method of mounting the 3N140 MOS field-effect transistor. Note that the transistor leads have been short-circuited by a piece of fine, bare wire, This wire is removed after all transistor connections have been made by merely pulling on the looped portion.

Special Handling of MOS Field-Effect Transistors

Special care must be exercised when wiring an MOS transistor into a circuit. For example, there is always a possibility that the transistor can be damaged if static electricity is discharged across the oxide layer. Such risk can be virtually eliminated, however, if all leads are shorted until the completion of all wiring. The 3N140 comes supplied with a protective ring which shorts the leads. This ring should be removed before wiring is commenced, and a fine, bare wire wrapped around the leads near the case. The shorting wire should not be removed until all soldering is completed.

Some builders may prefer to use a socket instead of soldering the transistor directly into the circuit. This practice is acceptable when used in conjunction with the rules listed below. (All transistor failures experienced by the authors have been traceable to violations of these rules. Please observe them carefully.):

- Keep transistor leads shorted until the transistor is completely connected to the circuit.
- Never insert or remove the transistor when power is on. (This rule applied to all transistors.)
- When cutting leads, grasp the leads and case simultaneously. This action will reduce the possibility of mechanical and electrical shock.

Adjustments

Preamplifier tuning is simplified because no special neutralization is needed - even at 144 MHz. Rough adjustments of the coils may be made by use of a grid-dip oscillator. The finishing touches are made while listening to a weak station.

It was rewarding for the authors to discover that a neighboring amateur's 1-kilowatt transmitter - only 200 feet distant - did not overload the preamplifier. At the same time, this word of caution is extended to the preamplifier builder with regard to his own high-power transmitter: Be certain that the coaxial relay has sufficient isolation to prevent transistor overload.

By following all the precautions mentioned, the builder should succeed in achieving a preamplifier of superior operational stature. Although small in size, the 3N140 dual-gate MOS field-effect transistor is a giant in performance.

• WITH ACKNOWLEDGEMENT TO RCA.

Application of Silicon Rectifiers to Capacitive Loads

by B. J. Roman

and J.M.S. Neilson

When rectifiers are used in capacitive - load circuits, the rectifier current waveforms may deviate considerably from their true sinusoidal shape. This deviation is most evident for the peak-to-average current ratio, which is somewhat higher than that for a resistive load. Because of the variation in current waveshapes, calculations of ratings for capactive-load circuits are generally more complicated and time-consuming than those for resistive-load rectifier circuits.

This Note describes a simplified rating system which allows designers to calculate the characteristics of capacitive-load rectifier circuits quickly and accurately. The effect of the addition of a series limiting resistance to such circuits and the importance of the ratio of the limiting resistance to capacitive reactance are described, and curves of rectifier current ratios are presented as functions of the effective ratio. Typical design examples are given, and output-ripple considerations are discussed. Table 1 defines the symbols used in the equations and calculations.

Table 1 — Definition of Symbols

E	=	sinsusoidal input voltage ($E \quad E_o \sin \omega t$)
E_o	=	peak input voltage
E_{avg}	=	average output voltage
f	=	input frequency (Hz)
ω	=	angular frequency of input ($\omega = 2\pi f$ radians per second)
t	=	time counted from beginning of cycle
R_S	=	limiting resistance
R_L	=	load resistance
C	=	load capacitance
I_o	=	absolute peak current through rectifier
I_{pk}	=	actual peak current through rectifier
I_{rms}	=	root-mean-squafe current through rectifier
I_{avg}	=	average current through rectifier
n	=	charge factor; 1 for half-wave circuit, for doubler circuit, 2 for full-wave circuit

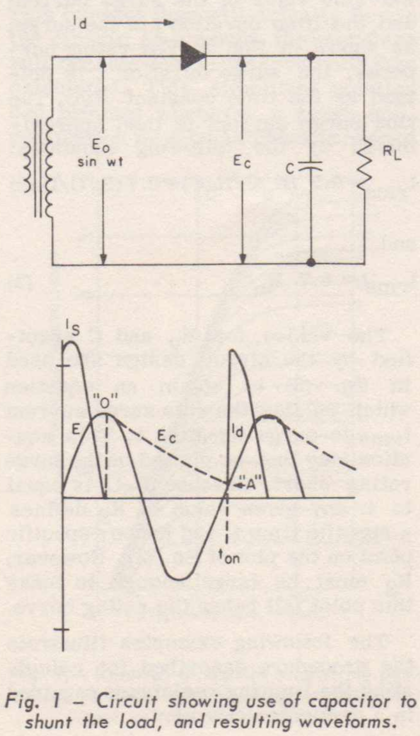


Fig. 1 – Circuit showing use of capacitor to shunt the load, and resulting waveforms.

Design of Capacitor — Input circuits

In the disign of a rectifier circuit, the output voltage and current, the input voltage, and the ripple and regulation requirements are usually specified. The transformer and the type of rectifier to be used are selected by the designer, and the load resistance is determined on the basis of the output voltage and current requirements. The ripple requirements are satisfied by use of a capacitor to shunt the load R_L, as shown in Fig 1. The waveforms for this circuit indicate that the voltage across the capacitor E_C coincides with the supply voltage E when the rectifier is conducting in the forward direction. A high initial diode surge current I_S occurs because the capacitor acts as a short circuit when power is first applied. The diode turns off at the peak of the curve (point O), and remains off until E_C is again equal to E (point A). The turn on point t_{on} is determined by the time constant R_LC, and affects the average, peak, and rms currents through the device.

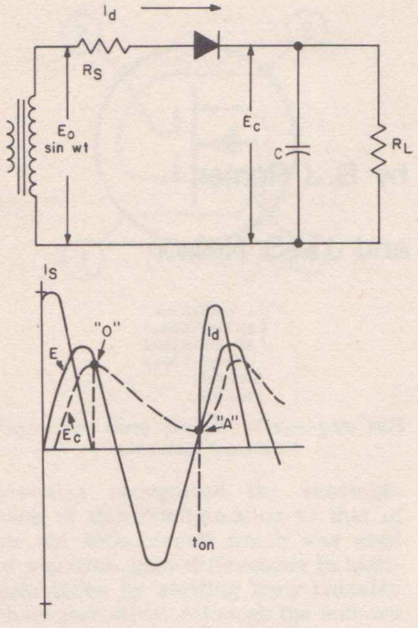


Fig. 2 – Circuit showing addition of limiting resistance and resulting waveforms.

As stated above, the low forward voltage drop of silicon rectifiers may result in a very high surge of current when the capacitive load is first energized. Although the generator or source impedance may be high enough to protect the rectifier, in some cases additional resistance must be added to the generator-rectifier-capacitor loop, as shown in Fig. 2, to keep the surge within device ratings. The waveforms in Fig. 2 show that the capacitor voltage E_C is no longer coincident with the steady state supply voltage E during any part of the cycle. The sum of the additional limiting resistance plus the source resistance is referred to as the total limiting resistance R_S. The ratio of R_S to capacitive reactance

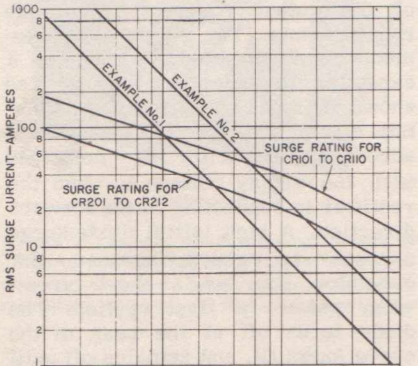


Fig. 3 – Surge-rating chart used for calculation of limiting resistance.

$1/\omega C$ is an important consideration in capacitor-input rectifier circuits; ideally, R_S should be much smaller than $1/\omega C$. The magnitude of R_S required in a particular circuit is calculated as described below.

Calculation of Limiting Resistance

The value of resistance required to protect the rectifier is calculated from the surge rating chart for the particular device used. Fig. 3 shows surge rating charts for the RCA CR100 and CR200 series of diffused junction stack rectifiers. Each point on the curves defines a surge rating by indicating the maximum time for which the device can safely carry a specific value of rms current.

With a capacitive load, maximum surge current occurs if the circuit is switched on when the input voltage is near its peak value. When the time constant R_SC of the surge loop is much smaller than the period of the input voltage, the peak current is equal to the peak input voltage E_o divided by the limiting resistance R_S, and the resulting surge I_S approximates an exponentially decaying current with the time constant $R_S C$, as follows:

$$I_S = (E_o/R_S) \exp(-1/R_SC) \quad (1)$$

Surge - current ratings for rectifiers are often given in terms of the rms value of the surge current and the time duration t of the surge, as shown in Fig. 3. For rating purposes, the surge duration t is defined as the time constant R_SC. The rms surge current is then approximated by the following equations:

$$I_{rms} = 0.7\ (E_oC/R_SC) = 0.7\,(E_o C/t) \quad (2)$$

and

$$I_{rms}t = 0.7\ E_oC \quad (3)$$

The values for E_o and C specified by the circuit design are used in Eq. (3) to obtain an equation which relates the rms surge current I_{rms} to surge duration t. This equation may then be plotted on the surge rating chart. Because R_SC is equal to t. any given value of R_S defines a specific time t, and hence a specific point on the plot of Eq. (3). However, R_S must be large enough to make this point fall below the rating curve.

The following examples illustrate the procedure described for calculating the limiting resistance required in a particular circuit.

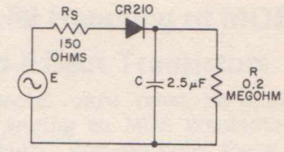


Fig. 4 – Half-wave rectifier circuit (E = 3500 V rms, E_o = 3820 V, f = 60 Hz).

<u>EXAMPLE NO.1</u>: Fig. 4 shows a half-wave rectifier circuit that has a 60-Hz frequency and a peak input voltage E_o of 4950 volts. The values of E_o and C are substituted in Eq. (3) to obtain the value of $I_{rms}t$, as follows:

$I_{rms}t$ 0.7 (4950) (2.5 x 10^{-6})

$I_{rms}t$ 0.0086

This value is then plotted on the surge-rating chart of Fig. 3 and is found to intersect the CR210 rating curve at 2.7 x 10^{-4} second. The minimum limiting resistance which affords adequate surge protection is then calculated as follows:

$$R_S C \geq 2.7 \times 10^{-4}$$

$$R_S \geq \frac{2.7 \times 10^{-4}}{2.5 \times 10^{-6}} = 108 \text{ ohms}$$

Because the value given for R_S is 150 ohms, the circuit has adequate surge-current protection for the rectifiers.

<u>EXAMPLE NO 2:</u> The doubler circuit shown in Fig. 5 has a peak input voltage of 3800 volts and a load capacitance of 10 microfarads. These values are substituted into Eq. (3), as follows:

$I_{rms}t$ (0.7) (3800) (10^{-5})

$I_{rms}t$ =0.0266

This value is then plotted on Fig. 3 and intersects the CR108 rating curve at 5.4 x 10^{-4} second. Therefore, the equation for the time constant is given by

$$R_S C \geq 5.4 \times 10^{-4}$$

$$R_S \geq \frac{5.4 \times 10^{-4}}{10^{-5}} \quad 54 \text{ ohms}$$

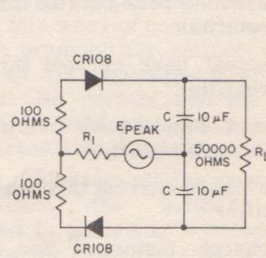


Fig. 5 – Voltage-doubler rectifier circuit (E_o= 2700 V rms, E = 3820 V, f = 60 Hz).

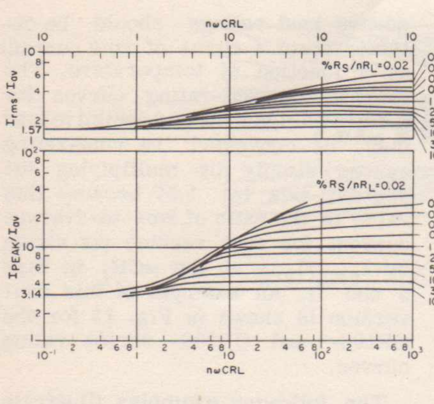


Fig. 6 – Relation of peak, average, and rms rectifier currents in capacitor-input circuits.

Calculation of Rectifier Current

The design of rectifier circuits using capacitive loads often requires the determination of rectifier current waveforms in term of average, rms, and peak currents. These waveforms are needed for calculations of circuit parameters, selection of components, and matching of circuit parameters with rectifier ratings. Actual calculation of rectifier current is a rather lengthy process. A much more direct process is to use the current relationship charts shown in Figs. 6 and 7. These curves can be readily used to find peak or rms current if the average current is known, or vice versa.

The ratios of peak-to-average current and rms-to-average current are shown in Fig. 6 as functions of the circuit constants $n\omega CR_L$ and R_S/nR_L. The quantity ωCR_L is the ratio of resistive-to-capacitive reactance in the load, and the quantity R_S/R_L is the ratio of limiting resistance to load resistance. The factor is referred to as the "charge factor" and is simply a multiplier which allows the chart to be used for various circuit configurations. It is equal to unity for half-wave circuits, 1/2 for doubler circuits, and 2 for full-wave circuits. (These values actually represent the relative quantity of charge delivered to the capacitor on each cycle).

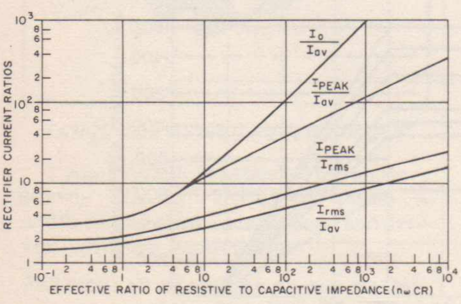


Fig. 7 – Forward-current ratios for rectifiers in capacitor-input circuits in which the limiting resistance is much less than $1/\omega C$.

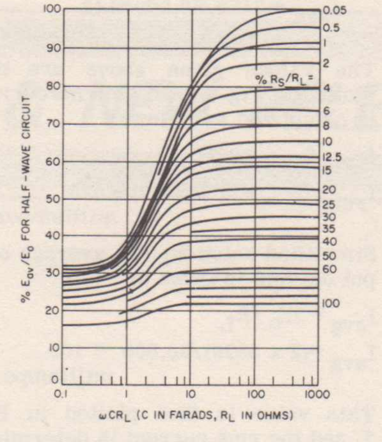


Fig. 8 – Relation of applied alternating peak voltage to direct output voltage in half-wave capacitor-input circuits.

In many silicon rectifier circuits, R_S may be completely neglected when compared with the magnitude of R_L. In such circuits, the calculation of rectifier current is even more simplified by the use of Fig. 7, which gives current ratios under the limitation that R_S/R_L approaches zero. Even if this condition is not fully satisfied, the use of Fig. 7 merely indicates a higher peak and higher rms current than will actually flow in the circuit; as a result, the rectifiers will operate more conservatively than calculated. This simplified solution can be used whenever a rough approximation or a quick check is needed on whether a rectifier will fit the application. When more exact information is needed, Fig. 6 should be used.

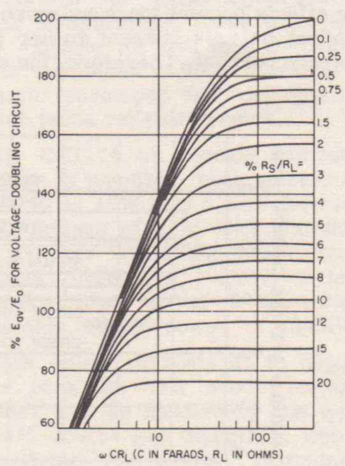


Fig. 9 – Relation of applied alternating peak voltage to direct output voltage in capacitor-input voltage doubler circuits.

Average output voltage E_{avg} is another important quantity because it can be used to find average output current. The relations between input and output voltages for half-wave, voltage-doubler, and full-wave circuits are given in Figs. 8, 9, and 10, respectively. Output ripple is shown in Fig. 11 for all three circuits. Although these curves were originally calculated for vacuum-tube rectifiers, they are equally applicable to silicon rectifier circuits.

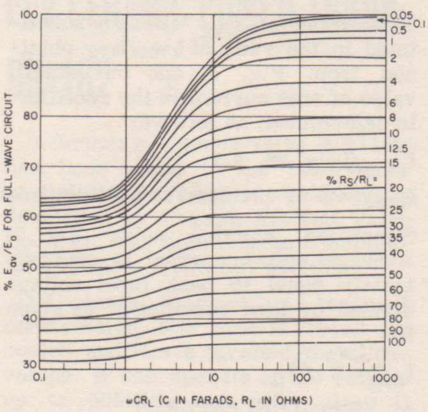


Fig. 10 – Relation of applied alternating peak voltage to direct output voltage in full-wave capacitor-input circuits.

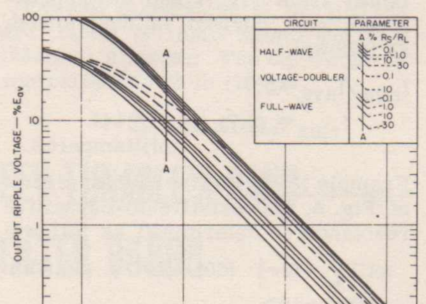


Fig. 11 – RMS ripple voltage of capacitor-input circuits.

The following examples illustrate the use of Figs. 8 through 11 in rectifier-current calculations. Both exact and approximate solutions are given for each example.

<u>Example No. 3:</u> For the half-wave circuit of Fig. 4, the resistive-to-capacitive reactance is found to be:

$$\omega CR_L = (2\pi)\ (60)\ (2.5 \times 10^{-6})\ (200{,}000)$$

$\omega CR_L = 189$

Exact solution using Fig. 6: The ratio of R_S to R_L must first be calculated as follows:

$$\% \frac{R_S}{R_L} = \frac{150 \times 100\%}{200{,}000} = 0.075\%$$

The values given above are then plotted in Fig. 8 to determine average output voltage and average output current, as follows:

$E_{avg}/E_o = 98\%$

$E_{avg} = (0.98)\ (4950) = 4850$ volts

$I_{avg} = E_{avg}/R_L$

$I_{avg} = 4850/200{,}000 = 24.2$ milliamperes

This value of I_{avg} is then substituted in the ratio of I_{rms}/I_{avg} obtained from Fig. 6, and the exact value of rms current in the rectifier is determined, as follows:

$I_{rms}/I_{avg} = 4.4$

$I_{rms} = (4.4)\ (24.2) = 107$ milliamperes

Simplified solution using Fig. 7: Average output current is approximately equal to peak input voltage divided by load resistance, as given by

$I_{avg} = E\ /R_L$

$I_{avg} = 4950/200{,}000 = 24.7$ milliamperes

This value of I_{avg} is then substituted in the ratio of I_{rms}/I_{avg} obtained from Fig. 7 and the approximate rms current is determined, as follows:

$I_{rms}/I_{avg} = 5.7$

$I_{rms} = (5.7)\ (24.7) = 141$ milliamperes

Example No 4. For the doubler circuit of Fig. 5, the resistive-to-capacitive reactance is determined as follows:

$\omega CR_L = (2\pi)\ (60)\ (10^{-5})\ (50{,}000)$

$\omega CR_L = 189$

$n\,\omega CR_L = 94$

Exact solution: The ratio of R_S to R_L is determined as follows:

$$\% \frac{R_S}{R_L} = \frac{100 \times 100\%}{50{,}000} = 0.2\%$$

This percentage is then used in conjunction with Fig. 9, and E_{avg} and I_{avg} are determined as follows:

$E_{avg}/E_o = 186\%$

$E_{avg} = (1.86)\ (3820) = 7100$ volts

$I_{avg} = E_{avg}/R_L$

$I_{avg} = 7100/50{,}000 = 142$ milliamperes

The values given above are then plotted in Fig. 6, and the rms current is calculated as follows:

$I_{rms}/I_{avg} = 3.7$

$I_{rms} = (3.7)\ (142) = 525$ milliamperes

Simplified solution: The average output current is given by

$I_{avg} = 2E_o/R_L$

$I_{avg} = (2 \times 3820)/50{,}000 = 153$ milliamperes

This value is then plotted in Fig. 7, and the rms current is determined as follows:

$I_{rms}/I_{avg} = 4.8$

$I_{rms} = (4.8)\ (153) = 734$ milliamperes

As previously noted, the simplified solution in both examples predicted a higher rms current than the actual value: about 32 per cent higher in Example No. 3 and 40 per cent higher in Example No. 4. The amount of error involved depends on both ωCR_L and R_S/R_L.

Rating Curves for RMS Current Versus Temperature

In most technical data for rectifiers, the current-verses-temperature ratings are given in terms of average current for a resistive load with 60-Hz sinusoidal input voltage. However, when the ratio of peak-to-average current becomes higher (as with capacitive loads), junction heating effects become more and more dependent on rms current rather than average current. Therefore, the capacitive-load ratings should be obtained from a curve of rms current as a function of temperature. The average current-rating curves for a sinusoidal source and resistive load may be converted to rms-rating curves simply by multiplying the current axis by 1.57 because this value is the ratio of rms-to-average current for such service (as shown by I_{rms}/I_{avg} at low ωCR_L in Figs. 6 and 7). An example of this conversion is shown in Fig. 12 for the CR100- and CR200- series rating curves.

The following examples illustrate the use of the rms current ratings.

Example No. 5: For the half-wave circuit of Fig. 4, it was found in Example No. 3 that the actual rms current in the rectifier is 107 milliamperes. The rms rating curve in Fig. 12 shows that the CR210 may carry up to 107 milliamperes at ambient temperatures up to 115 C.

Example No. 6: For the doubler circuit of Fig. 5, the actual rms current was determined to be 525 milliamperes. The rms rating curve for the CR108 in Fig. 12 shows that the circuit may be operated up to 88°C ambient temperature.

Example No. 7: If the higher values of rms current given by the simplified solution are used instead of the actual currents, the rms rating curves of Fig. 12 also give more conservative ratings because they predict a lower value for the maximum permissible ambient temperature. For example, for the half-wave circuit the exact rms current was found to be 107 milliamperes, and the approximate value was 141 milliamperes. These current values correspond to a maximum ambient temperature rating of 115°C by the exact solution and 110°C by the approximate solution.

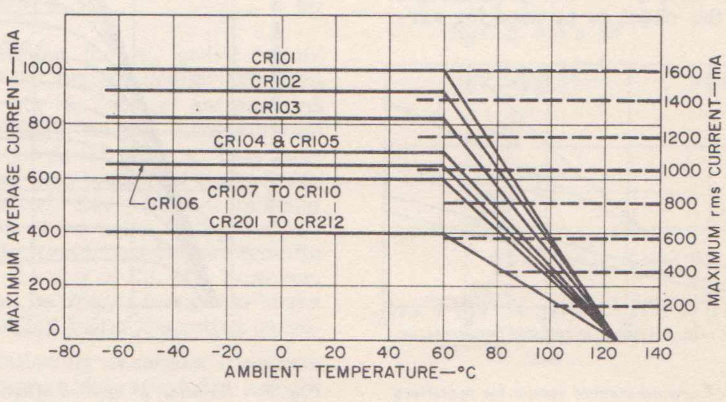


Fig. 12 – Current as a function of temperature for silicon rectifier stacks.

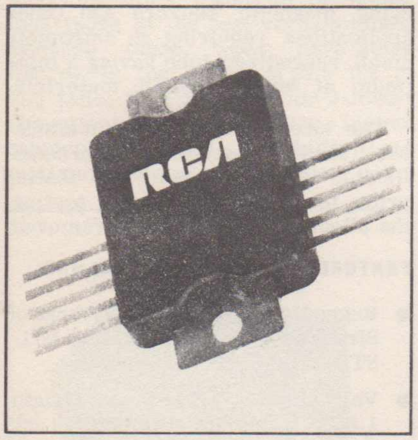
• WITH ACKNOWLEDGEMENT TO RCA

NEWS & NEW RELEASES

100W DEVELOPMENTAL HYBRID POWER MODULE

R.C.A. have recently released a developmental complete solid-state hybrid amplifier in a compact moulded-epoxy plastic pack. The device has been designated a TA7625 and is intended for audio, servo (A.C. orD.C. or pulse width modulated) amplifier.

The TA7625 employs a quasi-complementary-symmetry, class-B output circuit with built-in load-fault protection and hometaxial output transistors. This circuit may be operated from a single or split power supply (30 to 75 volts, total).



TA7625 100W HYBRID POWER MODULE

Type TA7625 is intended for audio-, servo- (A.C., D.C., or pulse - width modulated) amplifier and driven-inverter applications.

PARTICULAR FEATURES OF THE TA7625 INCLUDE:

High Power Output: Up to 100 W (RMS).

. High Output Current - 7A (peak).

. Built-in-Load-Line Limiting circuit. Protects Amplifier from Accidental Short-circuited Output Terminals. Amplifier is Stable with Resistive or Reactive Loads.

. Single or Split Power Supply (30 to 75V, Total).

. Provision for External Gain Control.

. Direct Coupling to Load.

. Class-B Output Stage. Less than 1-W Quiescent Dissipation.

. Rugged Package with Heavy Leads.

CX1174, CX1175

The first two of a new EEV 6 A range of compact deuterium - filled thyratrons, the CX1174 and CX1175, are designed for use in modulators for pulse klystrons in linear accelerators, these tubes are also suitable for use in high-power radar equipment. Both tubes have ceramic envelopes and will operate at high pulse repetition rates.

The CX1174 is a single gap tube capable of handling peak pulse currents up to 6000A at a peak forward anode voltage of up to 40kV. Suitable for average switching applications between 70MW and 100MW, it has a peak rating giving a maximum modulator output power of 120MW.

Peak pulse currents up to 6000A at a peak forward anode voltage of 80kV can be handled by the type CX1175 double gap thyratron, which is suitable for averge switching applications between 100MW and 150 MW. Peak output power rating will give a maximum modulator output of 200MW.

EEV CERAMIC POWER TRIODE BR1182

A further r.f. power valve (BR1182) has been added to the new series of industrial ceramic triodes being developed by English Electric Valve Co. Ltd., for both induction and dielectric radio frequency heating.

Of coaxial filament/grid terminal construction, this tube is forced-air cooled and has a ceramic/metal envelope. It can operate at full ratings up to 50MHz, and under class C unmodulated conditions will give an output at the valve anode of 50kW. The valve has a maximum continuous anode dissipation rating of 15kW.

Overall dimensions of the BR1182 are 14.6in (371mm) length and 10in (254mm) diameter. The net weight is approximately 35 lb (16kg).

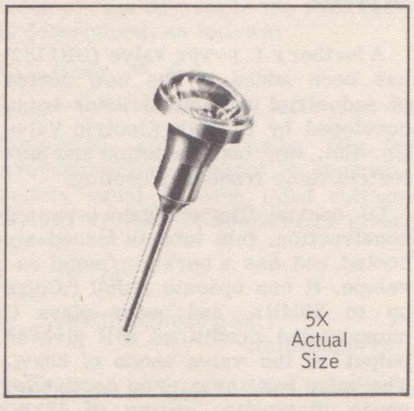
EEV TUNABLE X-BAND FILTER BS888

English Electric Valve Co. Ltd. has added a tunable X-band filter to its range of devices for radar duplexer systems.

Though the function of this type BS888 filter is similar to that of a varactor limiter stage, it is considerably cheaper and as a mechanically passive device has a virtually infinite life. The element forms a separate part of the radar equipment, making it independent of TR cell replacement.

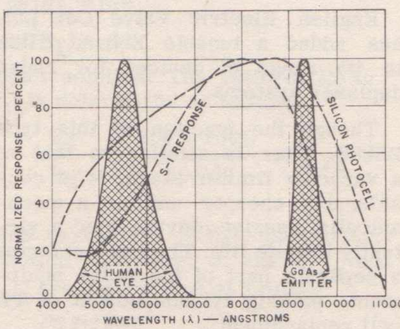
When used in conjunction with the improved EEV BS810 high Q marine radar TR cell, a spike leakage of only 0.02erg/pulse is typical, mesured at 25kW, 0.1us. Another advantage associated with the BS810 is the reduction of incident power, to a typical value of 150mW, required to initiate breakdown when the radar equipment is switched off and the TR cell is unprimed, so considerably improving the passive protection given to crystals. Frequency range of the BS888 is 9255MHz to 9565MHz. Overall length of the BS888 (including integral spacer) and BS810 mounted together is 2.11in (53.6mm).

40598A—GALLIUM-ARSENIDE INFRARED EMITTING DIODE



5X Actual Size

The R.C.A. 40598A is a high-efficiency gallium-arsenide emitting diode which radiates, when forward biased, in the near infrared region of the spectrum. The emitted radiation is collimated by means of a polished parabolic reflector which is an integral part of the diode package. The device is designed for operation in either continuous or pulse service at temperatures ranging from -73 ^oC to +75 ^oC.



Typical Response of IR Emitter, Human Eye, Silicon Photodetector, and Photomultiplier Tube (S - 1 Response)

The high radiant emission makes this device suitable for use in a wide range of optical applications.

FEATURES OF THE 40598A INCLUDE:

- 1mWmin. radiant power output at 50mA operating current.
- 9,300 ^{o}A (930nm) output (emission peak) at $T_C = 27^oC$.
- Usable with all types of silicon photodetectors.
- Focused output (15^o half-angle cone) using a unique parabolic reflector.
- Compact design for closely-spaced printed-circuit-board mounting.
- Solid-state reliability.

NEW RCA INTEGRATED CIRCUITS

The range of 45 medium-power DTL I.C.s available from R.C.A. have been increased by 12 with the addition of four new dual J-K Flip-Flops in three different package styles.

The CD2315, CD2316, CD2317, and CD 2318 are dual clocked J-K "master-slave" flip-flops, each on a single monolithic silicon chip. The CD2315 and CD2317 each consists of two CD2304-type Flip-Flops internally cross-coupled to perform the J-K function; the CD2316 and CD2318 types each consists of two CD2305-type Flip-flops also connected to perform the J-K function.

The CD2315 and CD2316 feature separate clock inputs to each flip-flop, making them suitable for ripple counter applications. The CD2315 and CD2316 have separate J and K inputs, separate Direct Set inputs, and no Direct Clear input.

The CD2317 and CD2318 feature a common clock input to each flip-flop, making them suitable for clocked counters and shift register applications. The CD2317 and CD 2318 also feature separate J and K inputs separate Direct Set inputs, and a common Direct Clear input.

The CD2315, CD2316 CD2317 and CD2318 are packaged in the R.C.A. Flat-Pack ceramic (-55^oC to $+125^oC$) package a suffix D is given to the Dual-in-line ceramic pack (-55^oC to $+125^oC$) and suffix E to the Dual-in-line plastic pack (0^oC to 75^oC).

C31016A PHOTOMULTIPLIER TUBE

The recently released RCA Developmental Type C31016A is a very short, ruggedized, 10-stage head-on type of photomultiplier tube having a bialkali photocathode of high quantum efficiency and high-stability copper-beryllium dynodes. The maximum length of this tube is only 1.85", its maximum diameter is only 1.05".

The only ferro-magnetic materials used in the construction of the tube are 14 pieces of 0.016" diameter by 0.25" long dumet in the stem.

These features make the C31016A highly suitable for space probes using magnetic sensors and other applications requiring an extremely small ruggedized tube having a minimum of ferro-magnetic materials.

The tube is supplied with a small-shell duodecal base attached to semi-flexible leads to facilitate testing prior to installation. After testing, the attached base should be removed.

FEATURES OF THE C31016A INCLUDE:

- Ruggedized Shock and Vibration Structure Designed to Meet MIL-STD-810B* Specification.
- Very Small Tube Size: Length, 1.85" Maximum Diameter, 1.05" Maximum.
- Tube Structure Having a Minimum of Ferro-Magnetic Materials.
- Bialkali Photocathode of High Quantum Efficiency: Typically 22% at 4000 ^{o}A.
- Low Dark Current: 5 x 10^{-10}A at 7 A/lm and 22^oC.
- High - Stability Copper-Beryllium Dynodes.

A Single-Gate MOS-FET Preamplifier for the 2-Meter Band

by R. M. Mendelson

The major disadvantage of bipolar transistorized RF amplifiers is their poor crossmodulation characteristics; an otherwise excellent, high-gain, low-noise amplifier can be useless in a recption area with strong local signals. Today, this handicap can be easily overcome by utilizing the superior performance qualities of the metaloxide - semiconductor (MOS) field-effect transistor. For example, two single-gate types - the 3N128 and 40467A -demonstrate (at maximum gain) crossmodulation characteristics equal to or better than those of the best vacuum tubes. Because these types have noise figures in the order of 3.5 dB at 200 MHz, they are ideal for 2-meter operation. Their small size, instant startup, excellent reliability, and minimum power requirements (12.6 volts at 5 milliamperes) are additional features which help meet the requirements of a high quality preamplifier.

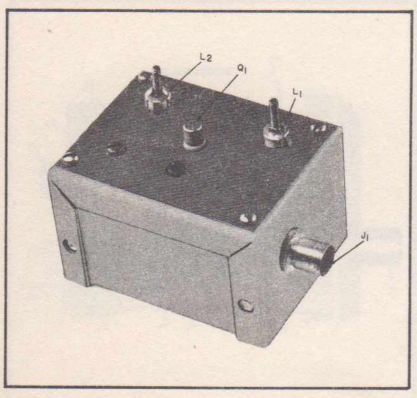
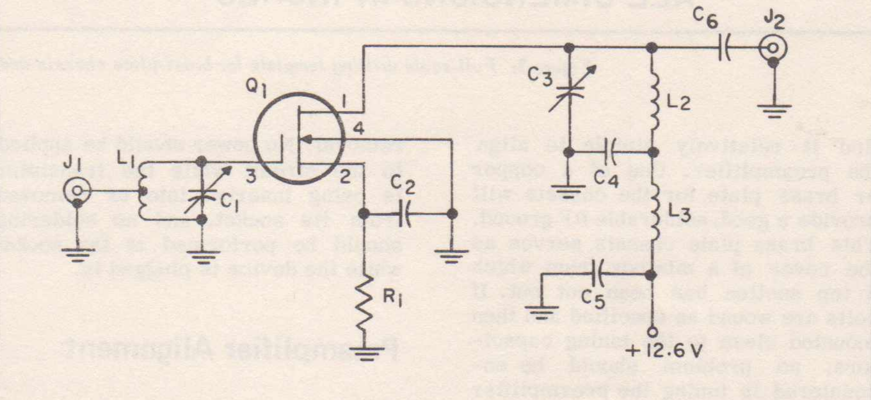


Figure 1: Exterior view of the preamplifier shows how brass-plate chassis is mounted on top of the minibox. Call-outs indicate locations of the two tuning capacitors, the MOS field-effect transistor, and the coaxial jack.



C_1, C_3 – 0.55 picifarads, tubular trimmer

C_2 – 500 picofarads, silver button

C_4 – 25 picofarads, silver button

C_5 – 1000 picofarads, ceramic disc

C_6 – 3.3 picofarads, ceramic tubular

J_1, J_2 – BNC-type coaxial jack

L_1 – 5 turns, No. 16 bare wire ¼-inch diameter, close-spaced; tap 1 turn up from bottom; do not short turns

L_2 – 3 turns, No. 16 bare wire, ¼-inch diameter, spaced to ½-inch total length

L_3 – RF choke, 1 microhenry

Q_1 – MOS field-effect transistor, 40467A or 3N128

R_1 – 270 ohms, ½ watt

Miscellaneous – One power socket; one transistor socket with four contacts; one brass plate 2.5/8 by 2 inches; one minibox 2¾ by 2.1/8 by 1.5/8 inches

Figure 2: Schematic diagram and parts list for 2-meter preamplifier circuit.

The circuit described in this article is a single-stage preamplifier that may be used ahead of an existing 2-meter converter or as the input stage of a new solid-state converter. As illustrated in the Figure - 2 schematic diagram, the circuit is straightforward and unburdensome, and use of the fullscale template (Figure 3) should expedite its completion in a few hours.

Construction

By following the illustrated layout, builders should be able to avoid any difficulty and at the same time

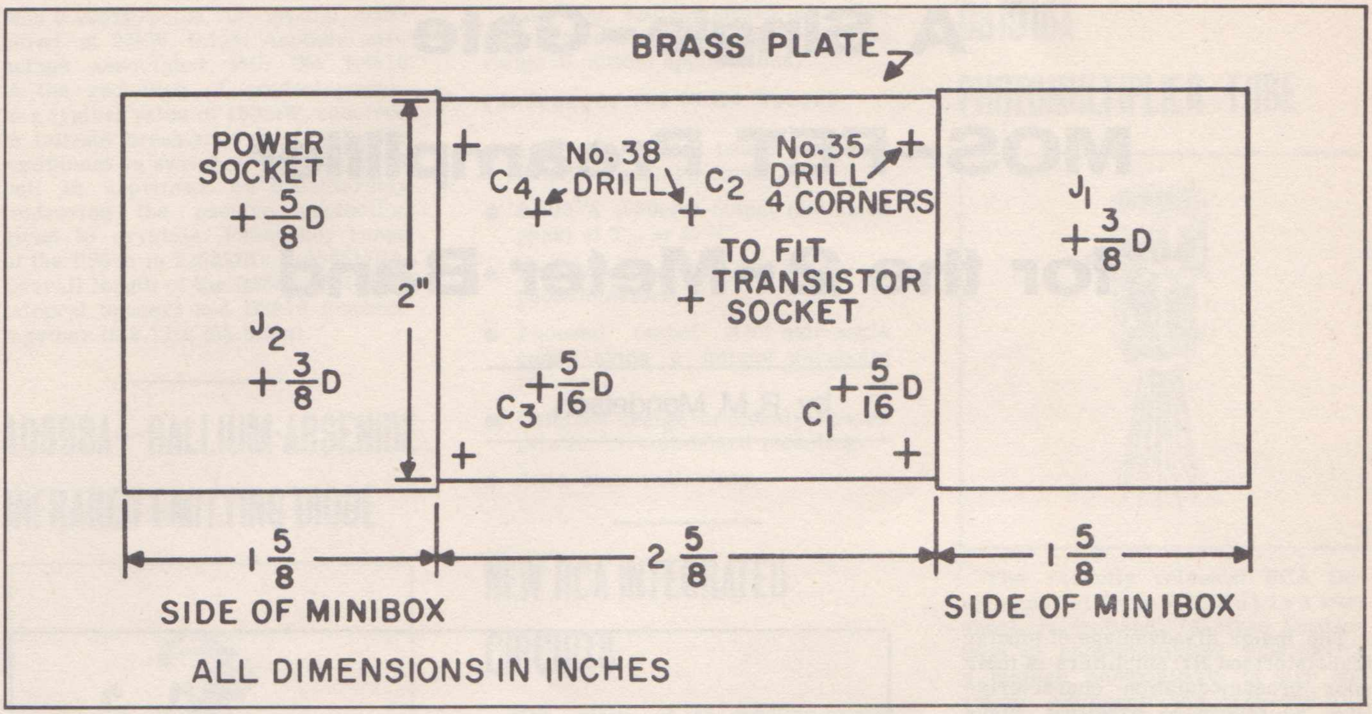


Figure 3: Full-scale drilling template for brass-plate chassis and two sides of minibox.

find it relatively simple to align the preamplifier. Use of a copper or brass plate for the chassis will provide a good, solderable RF ground. This brass plate chassis serves as the cover of a minibox from which a top section has been cut out. If coils are wound as specified and then mounted close to the tuning capacitors, no problem should be encountered in tuning the preamplifier to cover the full 4-megahertz range of the 2-meter band.

One special precaution must be observed when handling any MOS field-effect transistor; the leads must be shorted together until the device is plugged into its socket or soldered into place. Neglect of this procedure may result in permanent damage to the transistor from electrosatic discharge. After the device is in the socket, the possibility of damage by electrostatic discharge is very remote because of relatively low impenance paths between the transistor elements. MOS-FET's are factory-packaged with thin, bare protective wiring which shorts the leads. Similar-type wiring should be wrapped around the leads prior to wiring and should not be removed until soldering is completed. If it becomes necessary to remove a device from a socket, the shorting wire should be replaced prior to removal. No power should be applied to the circuit while the transistor is being inserted into or removed from its socket, and no soldering should be performed at the socket while the device is plugged in.

Preamplifier Alignment

Alignment of the MOS-FET preamplifier can be accomplished without test equipment and consists solely of two screwdriver adjustments. First, the preamplifier is connected to the antenna and 2 meter converter, and power is applied. The 12.6-volt power supply can vary one volt either way without ill effects. This power may bo obtained from the same source that feeds the solid-state converter; from the cathode of the audio-output stage of the communications receiver; from a suitable voltage divider to any positive power point in the receiver; or even from a battery.

Next, a signal near 145 MHz is tuned in and the antenna-tuned circuit is peaked for maximum signal. If no maximum can be found, the coil turns should be either squeezed closer or spread slightly apart until peaking occurs. The maximum will not bee too sharp since the circuit will pass the full 4-MHz range of the 2-meter band. A signal near 147 MHz is then tuned in, and the tuning steps repeated at the output circuit. After the preamplifier is checked for even gain across the band, the job is done.

If strong local signals have been blocking your solid-state converter, your troubles are ended. You now have a MOS-FET preamplifier that gives you an improved noise figure, generally better reception, and maintenance-free performance.

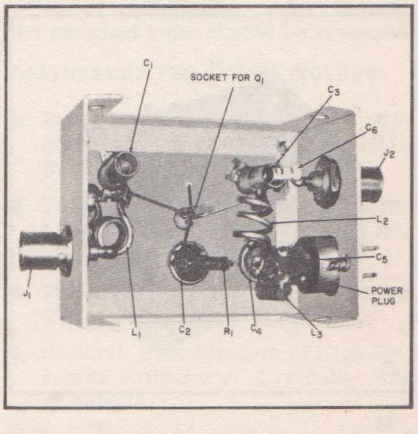


Figure 4: Interior view of assembled preamplifier showing location of all components. Note that short, straight leads are used to obtain good VHF operation.

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