SEMICONDUCTOR POWER CIRCUITS HANDBOOK

First Edition

*Compiled by the Applications Engineering Department
of Motorola Semiconductor Products Inc.*

INTRODUCTION

The Semiconductor Power Circuits Handbook was prepared to help designers make best use of such semiconductor devices as power transistors, thyristors, rectifiers, and zener diodes, in all types of applications. The book is divided into six chapters that emphasize the application areas which appear to be of greatest overall interest to prospective users of semiconductor power devices: motor speed controls, inverters, regulators, static switches, audio and servo amplifiers and miscellaneous applications.

This handbook contains a great variety of semiconductor power circuits for almost every type of power application. Even for a given application, such as dc-to-ac inversion, many different specific approaches are described: low and high voltage inputs, low and high power outputs, 60 Hz to 15 kHz switching frequencies, unregulated and regulated output with and without overload protection, SCR and transistor controlled, etc.

The majority of the circuits described in this book are unique and have not been published before. Each was designed, constructed, and evaluated for performance before it was included in this text. Each circuit performed its required task, as described, but it was not given a complete worst-case design. This means, for example, that while each transistor inverter performed well, it was not evaluated with limit transistors with minimum and maximum allowable current gain. It also means that each circuit performed well at room temperature but may need modification for best performance at high or low temperature extremes.

In the construction of these circuits, remember that good engineering practices are required for proper circuit operation. For example, power devices must have adequate cooling, and all device data sheet ratings must be observed.

While each circuit performs at the stated conditions, it is possible in most cases to change the capability of the circuit by using different semiconductor devices, such as units with higher or lower voltage or current ratings. This is apparent, for example, in the motor speed controls, where some circuits were tested with a 1/15 horsepower motor. The basic circuits will control much larger motors, and, in many cases, even with the devices given on the schematic.

In summary, each of the circuits described performs well under the stated conditions but can be modified for more demanding applications. For special assistance in the selection of semiconductor devices for OEM projects not covered in this reference manual the reader is invited to contact the Applications Engineering Department of Motorola Semiconductor Products Inc.

All resistors in this handbook are 1/2 watt, ±10% tolerance unless stated otherwise.

SEMICONDUCTOR POWER CIRCUITS

CHAPTER 1

Motor Speed Controls

1.1 Motor Speed Control

AC or DC?

Variable-speed motor drive circuits are used to provide easy and versatile control of powered drive systems. There are two basic types of speed-control systems, one for dc or universal motors and the other for ac induction motors. A dc-motor control system requires a minimum of power and control circuitry, but controls a motor that is expensive to purchase and expensive to maintain. The electronic control system for an ac motor is more complex and therefore more expensive, but the additional cost is offset by the economies of using an ac motor. An ac motor is much less expensive than an equivalent dc motor, its cost ranging from one-half the price for fractional-horsepower motors to one-fifth the price for large integral-horsepower motors. Moreover, since an ac motor has no brushes or commutators, it requires less maintenance, is more rugged and reliable, and can operate in explosive, dusty or highly humid atmospheres, and at high altitudes. In addition, ac motors can have higher maximum speeds than dc motors, and they can be liquid cooled. These advantages, when coupled with the availability of standard-rated motors from stock, make ac motors more attractive for most industrial applications than dc motors.

Speed Control for AC Motors

DESIGN CONSIDERATIONS

The speed of an ac induction motor can be controlled by varying either the amplitude or the frequency of its supply voltage, or both. The variable-voltage, fixed-frequency technique has the disadvantage that the maximum torque of the motor is porportional to the square of the applied voltage; therefore, this control is effective only over small torque ranges.

Better speed control is obtained through the use of a variable-frequency drive system, but the motor characteristics discussed below may
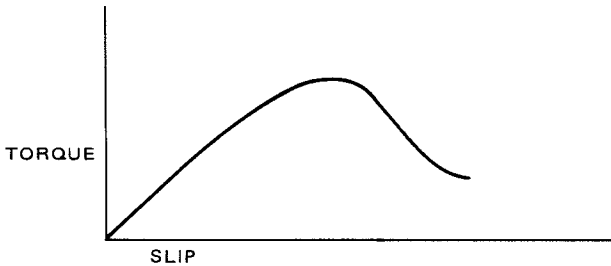
$$\text{Slip}_{\text{max torque}} = \frac{R2'}{\left[(R1)^2 + (X1 + X2')^2 \right]^{\frac{1}{2}}} \cdot$$

See Figure 1-2 For Definitions of Terms

Figure 1-1 — Slip-Torque Characteristics of Induction Motor

apply restraints that will limit the effectiveness of this approach. Generally, the best speed control is obtained through a system which varies both frequency and voltage.

No matter what type of speed control is used, a motor must be operated within its nameplate ratings. This rating gives the safe torque and speed limit for the motor as determined by the temperature limit, maximum safe rotational speed, and the saturation limit of the iron used in the motor. To keep from saturating the iron, the rms stator voltage must be proportional to the frequency of this voltage. Under this condition the maximum torque producible is independent of the frequency of the applied voltage. However, maximum torque occurs at a particular value of slip* at a particular supply frequency. The torque-versus-slip curve is shown in Figure 1-1. Peak efficiency occurs at the point of maximum torque.

It is possible to control the slip by varying the rotor resistance. The maximum torque that a motor can develop is independent of the rotor

---

*Slip is defined as

$$s = \frac{S_{sync} - S_{rotor}}{S_{sync}}$$

where S_{rotor} is the rotor speed, and S_{sync} is the synchronous speed. Synchronous speed is equal to $\frac{120\,f}{P}$, where f is the supply frequency in hertz and P is the number of poles.

resistance, as shown by the equation in Figure 1-2; however, the slip at which maximum torque is developed is a function of this resistance as shown by the equation in Figure 1-1. The effects of the rotor resistance are shown in Figures 1-3, 1-4 and 1-5. If the rotor resistance is controlled over the speed-control range of the motor, then the operating character-istics can be optimized and power requirements for a given load mini-mized.
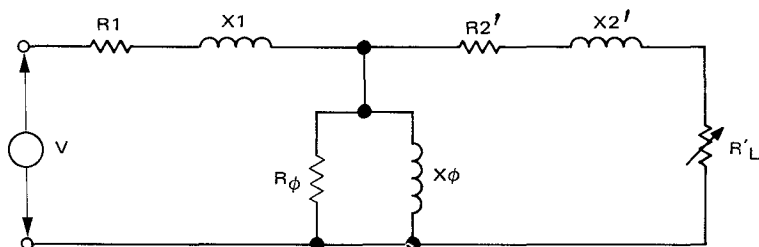
$$Torque_{max} = \frac{7.04}{S_{sync}} \cdot \frac{V^2}{2\left[R1 + \sqrt{R1^2 + (X1 + X2')^2} \right]}$$

R1 = Resistance of stator
X1 = Reactance of stator
R2' = Resistance of rotor reflected into stator
X2' = Reactance of rotor reflected into stator
RL' = Resistance of load reflected into stator
$R\phi$ = Leakage resistance
$X\phi$ = Leakage reactance
S_{sync} = Synchronous speed
V = Input voltage

Figure 1-2 — Equivalent Circuit of Induction Motor

K,K1,K2 = Constants   Sm = Speed of maximum slip
W = Speed   V = Applied voltage
W_s = Speed for zero torque

$$T = \frac{K \ V^2(1 - w/w_s)}{Sm^2 + (1 - w/w_s)^2}$$

$$T = \frac{K_1 \ V^2 \ (w_s - w)}{w_s}$$

$$T = K_2 \ V^2$$
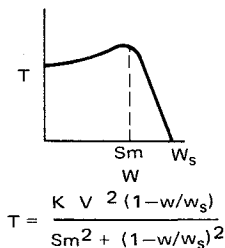
Figure 1-3

Speed-Torque
Characteristics of
Conventional
Rotor
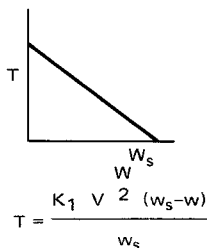
Figure 1-4

Speed-Torque
Characteristics of
High-Resistance
Rotor
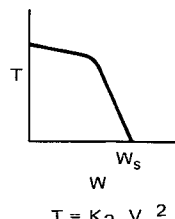
Figure 1-5

Speed Torque
Characteristics of
Constant-Torque
Rotor

It is also possible to achieve the high torque-to-current ratio typical of a dc motor by controlling the slip. Thus some of the characteristics of a dc and an ac motor can be combined. Generally, this type of control is not used as it involves the use of slip rings.

If the rotor resistance is constant, then the slip at maximum torque is inversely proportional to the applied frequency. For supply frequencies below the design value a series of speed-torque curves, as shown in Figure 1-6, can be obtained. Note that the maximum torque is constant, as it is limited by the saturation level of the iron. This level depends on the voltage magnitude at any particular frequency.
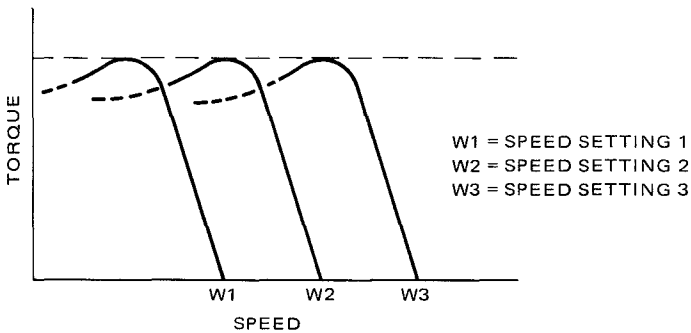
Figure 1-6 — Speed-Torque Characteristics of Induction Motors at Various Speed Settings

If the supply frequency is above that for which the motor was designed, then the horsepower rating must not be exceeded. This means the maximum torque will be reduced. Usable torque is even less at the higher speeds since friction and windage losses are increased.

A motor is capable of providing more power than listed on the name plate, but this increases the heat dissipated in the motor. The ratings which can not be exceeded without degrading the reliability are the maximum temperature limit of the winding insulation, and the maximum rotational speed of the armature.

The combination of all of these factors points to the fact that to achieve optimum performance in controlling the speed of an induction motor, a supply voltage which varies directly with the supply frequency should be used.

The equations shown in Figures 1-3, 1-4 and 1-5 hold when an induction motor is excited by a steady-state sinusoidal stator voltage.

Generally, the circuits used to control the frequency of the supply for the motor do not supply a steady-state sine-wave voltage. Therefore, the equations shown are only approximate. If it is necessary to maintain precise speed and torque control, then a speed-sensing circuit which feeds back to a variable-frequency and variable-voltage control circuit should be employed.

CIRCUIT CONFIGURATION

Either transistors or thyristors can be used as the main speed-control element for a motor. Transistors require continuous drive to remain in the "on" state and must be able to withstand high voltages and currents. When saturated, they dissipate less power than thyristors. Thyristors require a power pulse to turn on and will remain on until their anode current is reduced below the hold value. This reduction of current may occur from the natural commutation during the cycle of a sine wave, or by special circuits which bypass the current or insert high series impedance.

Two types of circuits are used to provide variable-frequency alternating voltages for motors, inverters and cycloconverters.
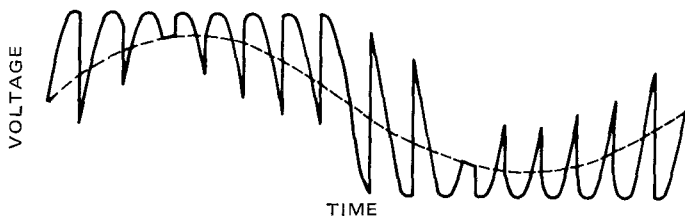
Figure 1-7a — Output Voltage of Half-Wave Cycloconverter
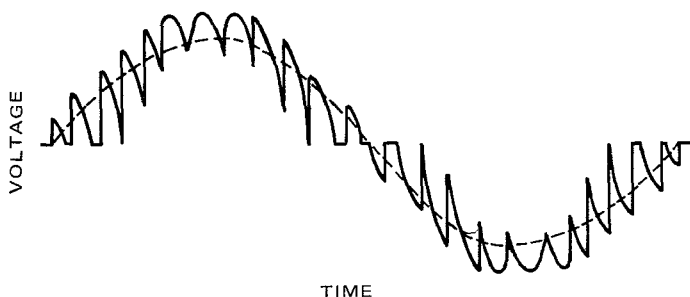
Figure 1-7b — Output Voltage of Full-Wave Cycloconverter

A cycloconverter changes the frequency of a three-phase alternating power source without the necessity of an intermediate ac-to-dc conversion stage. Cycloconverters require multiple control devices with three-phase excitation, and are limited to low output frequencies. The output of a cycloconverter has a relatively large ripple voltage, but comparatively small ripple current, which is what contributes to motor heating. The higher the output frequency, the greater is the ripple current since there are fewer supply pulses per output cycle. This limits the practical upper frequency for a cycloconverter to one sixth the supply frequency if a half-wave converter is used (i.e., three thyristors and three diodes), and one third the supply frequency if full-wave control is used (i.e., twelve thyristors) as can be seen by the output waveform for both types as shown in Figure 1-7.

An inverter is a dc-to-ac converter; it can be designed for variable-frequency output. If an inverter is to be used on alternating current, the ac must first be rectified and filtered to obtain direct current. If thyristors are used as the main controlled element in the inverter, then they must be forcibly commutated. Forced-commutation circuits can be either regenerative or nonregenerative: The regenerative inverter requires a power supply that can absorb as well as deliver power; an example is a dual phase-controlled rectifier operating off the public supply lines. Inverter circuits can supply frequencies much higher than power line frequencies and do not require as many main control devices nor control circuitry as complex as do cycloconverters. Inverters can also operate from either rectified alternating current or from a dc supply.

If operated from an ac supply, the inverter presents a higher power factor to the line than a cycloconverter and is better isolated from the line for switching spikes, due to the filtering. For these reasons, only inverter circuits are described here for variable-frequency motor speed controls.

In certain applications, cycloconverters can provide optimum characteristics for speed control. For excellent discussions on cycloconverters see references four and five in the bibliography for this chapter.

Speed Control for DC Motors

DESIGN CONSIDERATIONS

The speed of a dc motor is dependent on the applied voltage. For a series motor, the speed is directly proportional to the applied voltage. For a shunt motor, a constant voltage should be applied to the shunt field to maintain constant field flux so that the armature reaction has negligible effect. When constant voltage is applied to the shunt field, the speed is a direct function of the armature voltage and, consequently, the armature current. If the field is weak, then the armature reaction may counterbalance the voltage drop due to the brushes, windings and armature resis-

tances with the net result of a rising speed load characteristic. The performance of a dc motor operated from a rectified ac supply can be significantly different from one operated from a dc supply, as explained below.

The maximum working torque of a dc motor is governed by the design maximum temperature rise of the motor. This torque is proportional to the average armature current, whereas the copper loss is proportional to the rms value of the armature current. Therefore, for any given motor the maximum working torque is less for an ac supply than for a dc supply. The form factor (F_f), which is equal to Irms/Iavg, can be used as a measure of motor performance. It can be shown that the ratio of the dc armature copper loss to the ac armature copper loss is equal to $1/(F_f)^2$. The speed torque curves for several dc motors are shown in Figure 1-8 and 1-9.
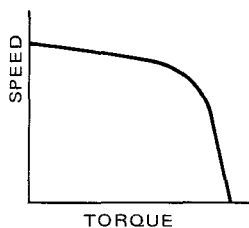
Figure 1-8 — Speed-Torque Characteristics of Shunt-Wound or Permanent-Magnet Motor
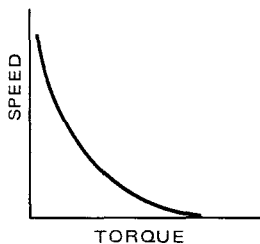
Figure 1-9 — Speed-Torque Characteristics of Series-Wound or Universal Motor

CIRCUIT CONFIGURATIONS

The speed of a shunt-wound motor can be controlled with a variable resistance in series with the field or the armature. Varying the field current for small motors provides a wide range of speeds with good speed regulation. However, if the field becomes extremely weak, a rising speed-load characteristic results. This method cannot provide control below the

design motor speed. Varying the resistance in series with the armature results in speeds less than the designed motor speed; however, this method yields poor speed regulation, especially at low speed settings. This method of control also increases power dissipation and reduces efficiency and the torque since the maximum armature current is reduced. Neither type of resistive speed control is very satisfactory.

Thyristor drive controls, on the other hand, provide continuous control through the range of speeds desired, do not have the power losses inherent in resistive circuits, and do not destroy the torque characteristics of motors.

A permanent-magnet motor has operating characteristics similar to those of a shunt-wound motor. Separate field and armature circuits are not required as in a shunt-wound motor since the field is provided by the permanent magnet.

Although a series-wound motor can be used with either dc or ac excitation, dc operation provides superior performance. A universal motor is a small series-wound motor designed to operate from either a dc or an ac supply of the same voltage. In the small motors used as universal motors, the winding inductance is not large enough to produce sufficient current through transformer action to create excessive commutation problems. Also, high resistance brushes are used to aid commutation.

The characteristics of a universal motor operated from alternating current closely approximate those obtained from a dc power source up to full load; however, above full load the ac and dc characteristics differ as
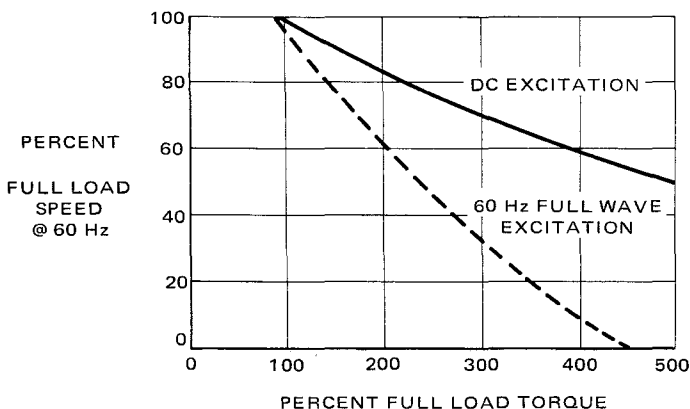
Figure 1-10 — Typical Speed-Torque Curve for Universal Motor

shown in Figure 1-10. For a series motor that was not designed as a universal motor, the speed-torque characteristic with ac rather than dc is not as good as that shown for the universal motor. At light loads, the speed for ac operation may be greater than for dc, since the effective ac field strength is smaller than that obtained on direct current. At any rate, a series motor should not be operated in a no-load condition unless precautions are taken to limit the maximum speed.
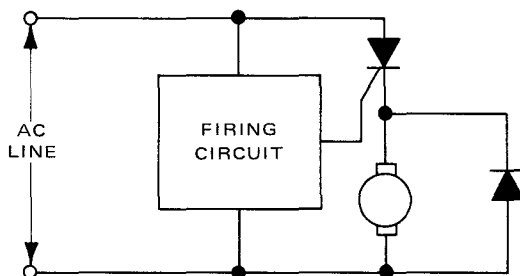
**Figure 1-11 — Typical Speed-Control Circuit
Showing Free-Wheeling Diode**

Since torque is proportional to armature current and motor voltage, a current- or voltage-sensitive circuit can be used to regulate the voltage applied to the motor. A variable-voltage power supply can consist of a circuit which provides a repetitive power pulse from either an ac or a dc source. If an LC filter is used for smoothing, a free-wheeling diode is generally included across the input of the filter to reduce commutator sparking by using the filter inductor to help maintain dc in the motor. The diode is used across the motor if such a filter is not used; this is shown in Figure 1-11. The supply thyristor shown in the figure is turned on at some phase angle to deliver the proper voltage. It will shut off at nearly 180°, if 0° is taken to be the time the voltage goes positive through zero, at high rotor speeds with low torque loads, since a large back emf is generated and the armature current is low. The free-wheeling diode will not conduct under this condition. At moderate speeds and loads, the thyristor will turn off at some angle greater than 180° when compared to the voltage angle. Here the free-wheeling diode will conduct from the time the thyristor turns off until the energy in the motor is less than that required to forward-bias the diode. At low motor speeds and high torque loads the free-wheeling diode will still be conducting when the thyristor is turned on again. The increasing current when the thyristor is on, and decaying current when the diode is conducting, can affect motor commutation to a

degree such that sparking can occur. A slight amount of sparking is not detrimental to motor performance, but should be avoided if possible since environmental conditions may not tolerate sparking.

As a rough guide, the free-wheeling diode conducts continuously when the form factor is less than 1.11, and conducts for part of the cycle when the form factor is greater than this value.

For fractional and small integral-horsepower dc motors operating from alternating voltage, a single-phase supply is generally used. For larger motors (over 3 hp), three-phase supplies are generally used. The control of a three-phase supply is more complex, but the ripple frequency is much higher, and consequently, less ripple voltage is presented to the motor. An important point to remember is that there is a maximum voltage that can be placed on the armature. This is especially important to note when phase control of the power lines is used to vary the voltage, because the peak line voltage may be placed on the armature even though the average voltage is much less.

Braking a Motor

The braking of a motor provides a method of controlling the coasting of the rotor and its load. Several methods may be used to brake a motor: mechanical, dynamic, plugging or regenerative. Mechanical braking requires the least complicated control; it also causes power to be dissipated in elements external to the motor. The disadvantages of mechanical braking are its higher initial cost and greater maintenance costs.

Dynamic braking is accomplished by placing an electrical load (generally a short) across the armature while maintaining a dc field. This may be done in an ac induction motor by replacing the ac supply with a dc supply which is removed when the machine is stopped. The motor then becomes a generator with a shorted armature, and the energy of the load is converted to electrical energy in the armature circuit and dissipated as heat.

Plugging requires that the windings be connected for reverse rotation. The large reverse-surge current resulting creates a large reverse torque. The motor must be disconnected from the power source when it has stopped or it will run in the opposite direction.

In regenerative braking, the motor must be connected to the power source during the times the back emf is greater than the source voltage, so that current is forced back into the source against the source voltage. The energy of the load is dissipated in the source and not in the motor for this method. Regenerative control may be achieved on an ac supply by using a bridge of four thyristors which are triggered in diagonal pairs at a time when the supply voltage opposes the return of current.

The method used for braking depends on the characteristics of the motor, the available power source, allowable coasting of the rotor and whether or not control is required after stopping. If the rotor is to be kept from turning after stopping then a mechanical brake must be used.

When braking power is applied to a motor, care must be taken not to exert too large a force on any element of the system. The motor itself will not be harmed as long as it does not dissipate excessive power, but elements connected to the rotor may be damaged if it is stopped too quickly. For example, gears may be stripped if a high inertial load is stopped too rapidly. Thus each part of the power train must be examined when establishing stopping requirements. If a motor's direction is to be reversed these same criteria apply and must be considered.

1.2 Inverter Driven Single Phase Induction Motor Speed Control
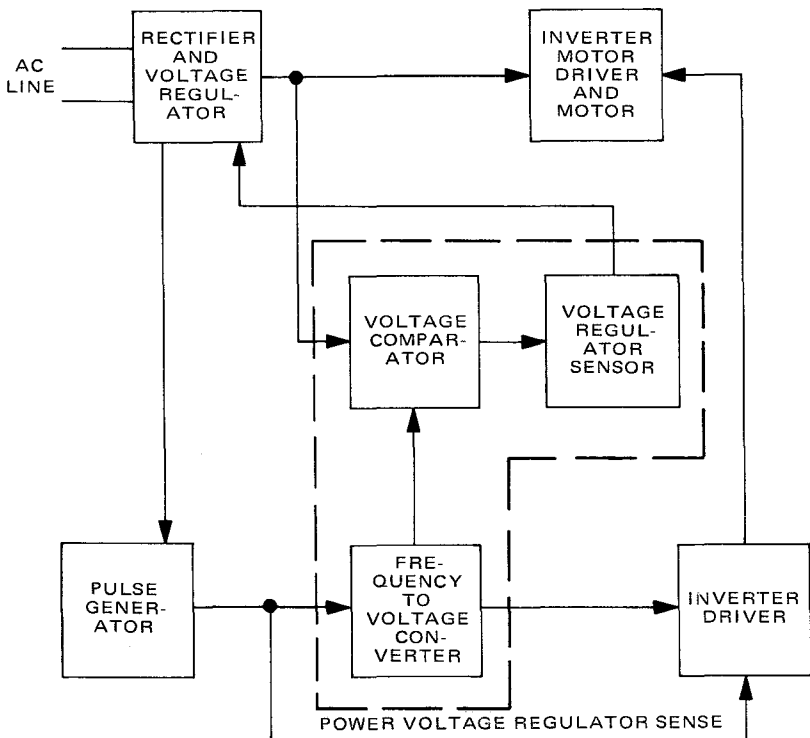
Figure 1-12 — Block Diagram of AC-Induction-Motor Speed Control

The block diagram of a single-phase, forced-self-commutated inverter which provides a variable-frequency, variable-voltage supply to a motor is shown in Figure 1-12. The circuit exhibits excellent speed control. The motor losses, due to the high frequency components of the square-wave voltage, though higher than with a sine-wave voltage, were not large enough to overheat the motor. Several speed-torque curves obtained from a 1/6 horsepower, 115 volt, 1200 r/min, 60 Hz induction motor are shown in Figure 1-14. This circuit can be used to control the speed of a single-phase induction motor with no modifications to the motor.

Inverter

The typical inverter circuit uses a transformer whose center-tapped primary windings are connected to the power supply alternately through the switching devices. This alternating voltage is transformed into the secondary which supplies the ac power to the load. A circuit such as this require a large transformer to drive a large motor. The circuit, which is shown in Figure 1-13 uses the same principle of operation but uses a center-tapped choke (L1) as an auto-transformer to supply the load. The inductance of the choke is not critical, but the windings must be able to handle the required motor current. To demonstrate that the inductance value was not critical, the choke was replaced by two resistors. The only difference in operation was that the peak-to-peak motor voltage was twice rather than four times the supply voltage, as obtained through auto-transformer operation of the choke.

An attempt was made to use the windings in the motor in place of the auto-transformer in the commutating circuit, but the only result was an overheated motor. This technique requires that the motor be modified to bring out the center point of the windings. Torque should be developed when the current is switched from one winding to the other. However, as the motor was not designed for this use, the torque developed when current increases in one winding was opposed by the torque developed in the other winding due to its current decreasing. The overheating was a result of the stator iron being saturated and allowing excessive current to pass through the windings.

The commutating capacitor (C1) must be small enough that a full charge can be placed on it at the highest frequency of operation required. The capacitor charges through the choke (L1), so the capacitor value depends on the choke inductance. Diode D1 and inductor L2 provide a reverse bias to SCR Q3 when it is being commutated off due to the firing of SCR Q4. This reverse bias must be held across the SCR long enough to ensure that it will turn off. The same operation occurs when SCR Q4 is to be commutated off except that diode D2 now conducts.

This circuit presents a square-wave voltage to the motor. This results in about 15% lower motor efficiency than that obtained with a sine-wave source, due to the high-frequency components of the waveform.

Inverter Driver and Oscillator

The inverter SCRs Q3 and Q4 are driven by the inverter driver and oscillator (See Figure 1-13). The driver is basically a flip-flop. Pulse transformers are used to couple the inverter SCRs to the driver.

The frequency of the multivibrator is controlled by a unijunction oscillator which provides pulses at a frequency set by the RC network attached to the emitter. These pulses trigger a one-shot multivibrator, which in turn triggers the inverter-driver multivibrator. The frequency of the voltage applied to the motor is one half that provided by the unijunction oscillator because of the divide-by-two action of the flip-flops.

Power Voltage Regulator

As discussed previously, for best results, the amplitude and frequency of the voltage applied to the motor must be varied proportionally; this is a function accomplished by the power voltage-regulator circuit. The basic circuit is a one-shot multivibrator whose output is filtered to obtain a dc level. This multivibrator output also provides a negative pulse necessary for triggering the multivibrator of the inverter driver. The magnitude of the filtered output voltage is a direct function of the frequency of the one-shot multivibrator since the pulse duration is fixed and the frequency variable. The filtered voltage is applied to a voltage comparator consisting of transistors Q5 and Q11. The voltage to which the comparison is made is the voltage applied to the inverter circuit. The voltage comparator does two things: It is used to set the supply voltage, and once set, it regulates this voltage. Transistor Q11 sets the reference voltage and transistor Q5 senses the supply voltage. The charging rate of capacitor C2 is controlled by the degree of conduction of Q5. The faster C2 charges, the sooner unijunction transistor Q6 fires, and thus SCR Q1 and SCR Q2 turn on sooner. This increases the supply voltage obtained from the filter composed of L3 and C3. SCR Q1 and SCR Q2 and diodes D3 and D4 form a full-wave bridge providing power to the inverter circuit. The range over which the supply voltage can be varied for a given inverter frequency range is controlled by the width of the one-shot multivibrator's pulse.

This power voltage regulator and power supply can be used as the supply for the other ac induction motor speed control circuits shown in this chapter.
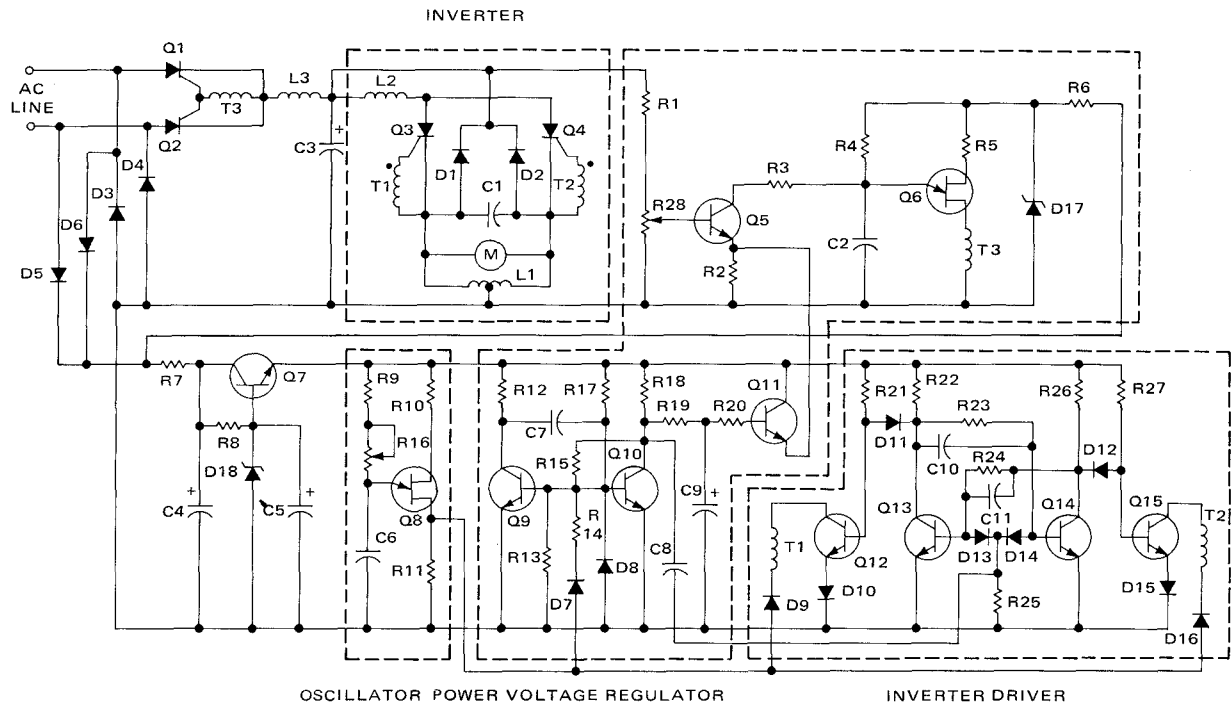
Figure 1-13 — Speed-Control Circuit for Inverter-Driven, Single-Phase, Induction Motor

INVERTER

OSCILLATOR POWER VOLTAGE REGULATOR   INVERTER DRIVER

Motor Speed Controls

Parts List for Figure 1-13

CAPACITORS
C1 — 10 µF, N.P., 300 V
C2 — 0.05 µF, 25 V
C3 — 6,000 µF, 300 V
C4 — 50 µF, 50 V
C5 — 50 µF, 25 V
C6 — 0.25 µF, 25 V
C7 — 0.25 µF, 25 V
C8 — 0.001 µF, 25 V
C9 — 50 µF, 25 V
C10 — 0.001 µF, 25 V
C11 — 0.001 µF, 25 V

DIODES
D1 — MR1125
D2 — MR1125
D3 — MR1125
D4 — MR1125
D5 — 1N4004
D6 — 1N4004
D7 — 1N4001
D8 — 1N4001
D9 — 1N4001
D10 — 1N4001
D11 — 1N4001
D12 — 1N4001
D13 — 1N4001
D14 — 1N4001
D15 — 1N4001
D16 — 1N4001
D17 — 1N4744
D18 — 1N4744

INDUCTORS
L1 — 32 mH, CT, 10 A
L2 — 1 mH, 10 A
L3 — 20 mH, 10 A

TRANSFORMERS
T1 — SPRAGUE 11 Z 13
T2 — SPRAGUE 11 Z 13
T3 — SPRAGUE 11 Z 13

RESISTORS
R1 — 30 kΩ, 1 W
R2 — 620 Ω, 1 W
R3 — 560 Ω
R4 — 43 kΩ
R5 — 330 Ω
R6 — 22 kΩ, 1 W
R7 — 3.5 kΩ, 5 W
R8 — 1 kΩ, 1 W
R9 — 20 kΩ
R10 — 2.7 kΩ
R11 — 390 Ω
R12 — 10 kΩ
R13 — 6.8 kΩ
R14 — 22 kΩ
R15 — 56 kΩ
R16 — 60 kΩ, 1 W Potentiometer
R17 — 150 kΩ
R18 — 2 kΩ
R19 — 33 kΩ
R20 — 10 kΩ
R21 — 6.8 kΩ
R22 — 10 kΩ
R23 — 36 kΩ
R24 — 36 kΩ
R25 — 51 kΩ
R26 — 10 kΩ
R27 — 6.8 kΩ
R28 — 2.5 kΩ, 2 W Potentiometer

TRANSISTORS AND SCRs
Q1 — 2N5170
Q2 — 2N5170
Q3 — 2N5170
Q4 — 2N5170
Q5 — 2N4124
Q6 — 2N4871
Q7 — 2N3019
Q8 — 2N4871
Q9 — MPS6531
Q10 — MPS6531
Q11 — 2N4124
Q12 — 2N3904
Q13 — 2N3904
Q14 — 2N3904
Q15 — 2N3904
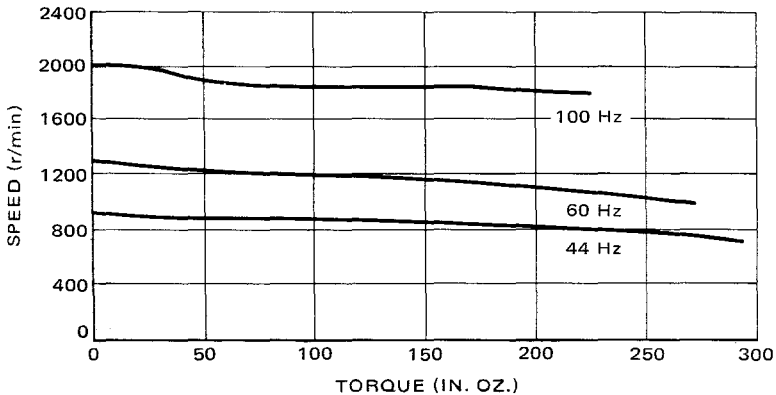
Figure 1-14 — Speed-Torque Characteristics of Induction Motor Controlled by Inverter-Driven Speed Control

1.3 Ring-Counter-Driven 3-Phase Motor Speed Control

The three-phase, forced-self-commutated inverter in Figure 1-15 is actually a ring counter which has a motor winding as the load for each stage. The successive firings of the SCRs create a moving flux field around the stator, which develops a torque on the rotor. The commutating circuit requires a dc supply. It is important to remember that the excitation current must be held constant, so the supply voltage has to be varied with the applied frequency. If the ac power lines are used as the primary source, then a voltage regulator similar to that shown for the inverter-driven, single-phase motor control of Section 1.2 can be used to provide the variable direct current for this circuit. The value of commutating capacitors C1, C2 and C3 depends upon the parameters of the particular motor used. The capacitors must be small enough to ensure that an adequate charge for commutation be stored in them at the highest frequency of interest. The charging rate depends on the motor winding inductance and the capacitance.

The time constant of the SCR firing circuit (such as R3, R4, and C4) must allow gate current to exist long enough for the anode current to exceed the holding current in this period of time. This again depends on the inductance of the motor winding. If the inductance is too large, a resistor to provide holding current can be placed in parallel with the winding.

An advantage of this circuit over other inverters is the ease with which the motor's rotation can be reversed. Interchanging any two of the three windings will reverse the rotation.

The motor's neutral connection is tied to the positive side of the supply voltage. The other end of each winding is returned to the supply through an SCR. When an SCR is on, the 1.5 kΩ resistor connected to its anode is grounded, allowing the diode connected to the other side to be forward-biased, but when the SCR is off the 1.5 kΩ resistor is at the supply potential, reverse-biasing the trigger input diode. What this does is to steer the next pulse arriving at the trigger input to the SCR at the right of the one that is on. To obtain the same speed as with an ac supply, the driving frequency must be increased by the factor equal to the number of windings used in the ring counter. In this case three windings were used so the driving frequency had to be 180 Hz to attain the same speed as normal motor operation on 60 Hz. The particular motor used in this test had two-pole, three-phase, wye-connected 220 volt windings. This meant that each winding was a 110 volt winding. Therefore, at 3600 r/min the supply voltage had to be 110 volts to deliver design torque per winding. However, the rated torque can not be achieved since only one winding is energized at
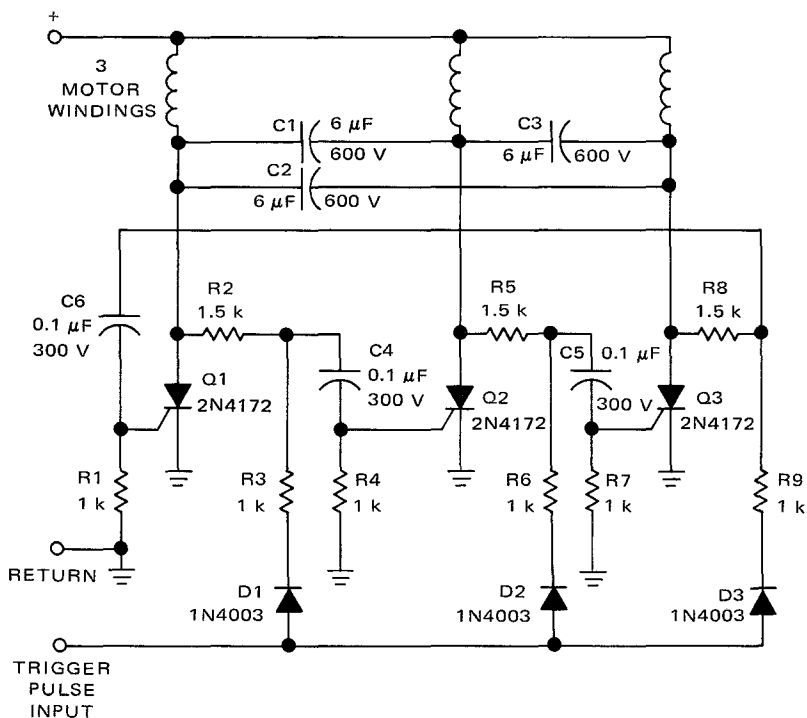
Figure 1-15 — Ring-Counter-Driven, Three-Phase Induction Motor

TORQUE (IN-LBS)
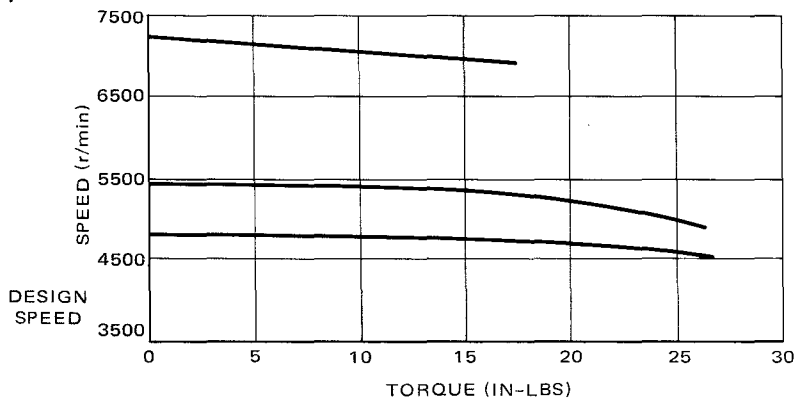
Figure 1-16 — Performance of Ring-Counter-Driven, Three-Phase Induction Motor

a time. The efficiency of the motor is about 20% less than design at rated speed due to the increased I^2R losses and core losses resulting from square wave voltage operation instead of sine wave voltage. At the higher speeds more power is required to overcome friction, windage and core losses. This particular motor was run as high as 7200 r/min; however, speeds in excess of this can be achieved. The speed-torque curve at each speed setting selected had only moderate droop. The droop increased at higher speeds as was expected, and is shown in Figure 1-16.

1.4 Series-Wound, DC-Motor Speed-Control Circuits

The speed of a series connected dc motor can be controlled by varying the average voltage applied to the motor. A variable voltage can be obtained in a number of ways, the easiest of which is phase controlling a thyristor connected in series with a motor operating from an ac supply.

The most economical circuit, in that it uses the fewest components, is the half-wave, uncompensated control shown in Figure 1-17. The control of speed is effective over a wide range, providing stable operation at low speeds with a given load. The circuit controls the average motor voltage by setting the firing point of SCR Q1. The time required for capacitor C1 to charge to the gate turn-on voltage is set by potentiometer R2. Once the SCR is on, the capacitor voltage is less than the forward voltage drop of the SCR for the rest of the half cycle. During the reverse half cycle, diode D1 blocks any current which would try to pass through the SCR gate. Capacitor C2 is then ready to begin charging when the next half cycle starts. The speed-torque curves for different speed settings of a 1/15 hp, 5,000 r/min, 115 volt motor are shown in Figure 1-18.
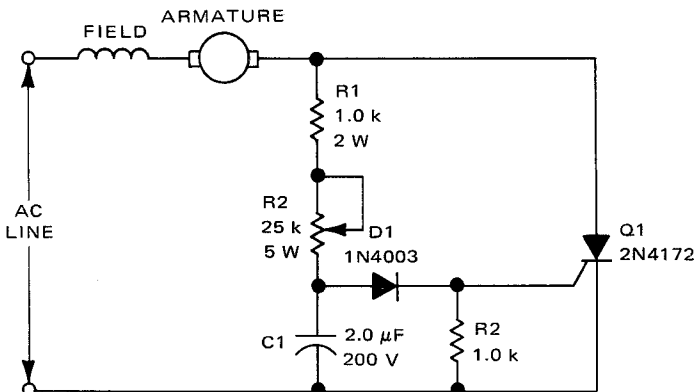
Figure 1-17 — Uncompensated Half-Wave Drive
for Series-Wound DC Motors

Better speed regulation can be obtained through the use of the half-wave compensated control shown in Figure 1-19. The circuit operates similarly to the previous one but uses the back emf of the motor to provide feedback during the time when the SCR is off. This feedback compensates for load differences, as it is inversely proportional to the load. That is, as the load is increased, the motor's speed is decreased and the back emf is decreased. When this occurs, the gate voltage exceeds the cathode voltage at an earlier phase angle, and turns the SCR on sooner.

Figure 1-18 — Speed-Torque Characteristics of Series-Wound Motor with Half-Wave Control
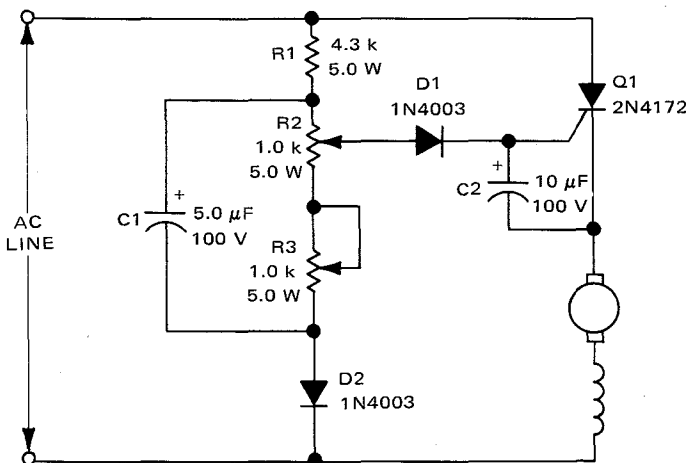
Figure 1-19 — Compensated Half-Wave Drive for Series DC Motors

This provides more drive to the motor, returning the speed to the original value. Actually, for the half-wave circuit, the speed does not remain the same, but the droop of the speed-torque curve is reduced considerably from that of the uncompensated circuit. The speed-torque curves for this circuit are shown in Figure 1-18, where they may be compared with those obtained from the uncompensated circuit. It can be seen that as the control resistance is increased, the speed is reduced, but the droop is much smaller. At slow speeds, the characteristic curves approach those of a shunt-wound motor.

An inexpensive circuit which provides full-wave control, but no compensation, uses a three-layer diode and a triac. The schematic is shown in Figure 1-20. The maximum average voltage is greater for full-wave control than for half wave; therefore, the motor can provide greater maximum torque with full-wave control. The characteristic curve is the same as that obtained for the uncompensated half-wave control circuit except that it has been translated to higher torque values. This curve is shown in Figure 1-21. The triac will turn on with gate current in either direction; therefore, three-layer diode D1 is used to set the firing point. When the voltage across capacitor C2 exceeds the breakover voltage of D1, it conducts and turns on triac Q1. The time required for the voltage to reach this point is controlled by the potentiometer. When the triac is on, the capacitors discharge to the forward voltage drop of the triac. They begin to charge towards the firing voltage again when the next half cycle from the line appears after the current passes through zero.

If greater speed control is necessary, a full-wave compensated circuit such as shown in Figure 1-22 can give characteristics very similar to those of a shunt-wound motor. The operation of this circuit is the same as that
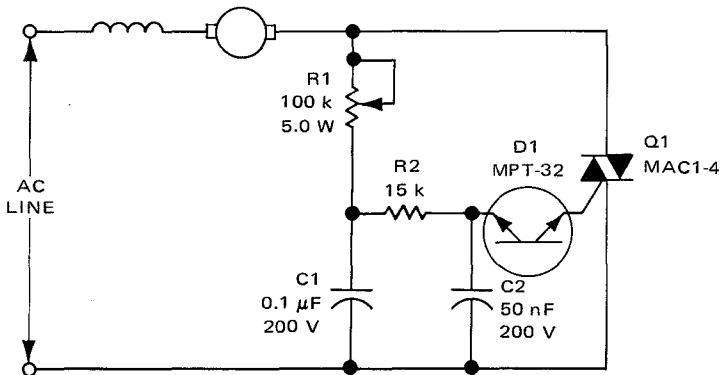
Figure 1-20 — Uncompensated Full-Wave Drive for Series-Wound DC Motors

of the half-wave circuit with compensation, with the exception that full-wave voltage is applied to the motor through the rectifier bridge, and a diode is required across the field. The function of this diode (D5) is to provide a current path for the energy stored in the inductive field, thereby enabling the SCR to turn off. An undesirable condition can occur with this circuit: At low speed settings and at no load, the SCR may not fire on every cycle due to the difference in the back emf of the motor from that at high loads and speeds. The back emf of the motor under these conditions may be greater than the gate voltage of the SCR. If this happens, the SCR does not come on and the motor slows down until the back emf is lower than the gate voltage, at which time the SCR can again be turned on.
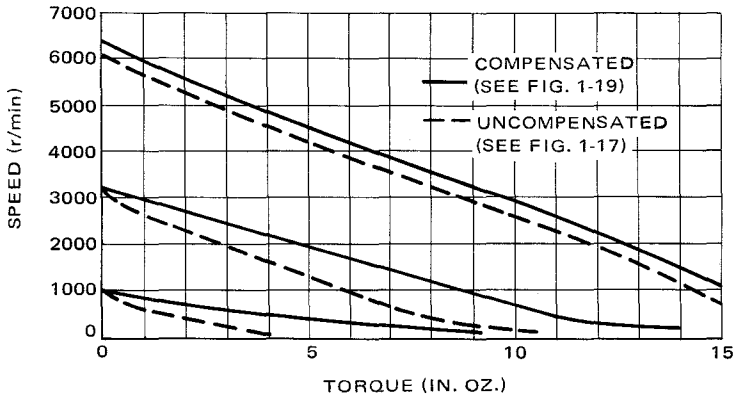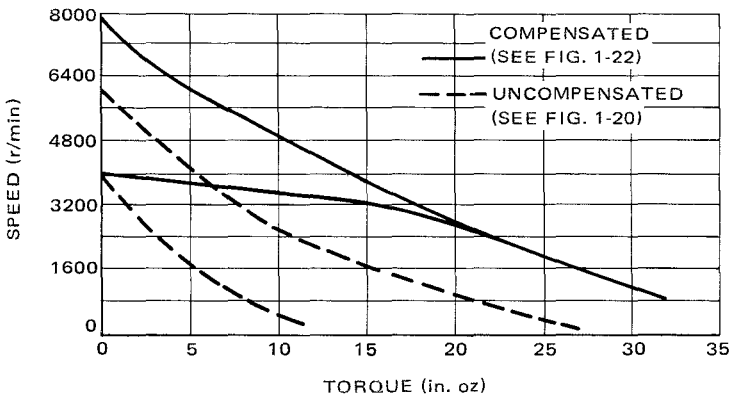

Figure 1-21 — Speed-Torque Characteristics of Series-Wound Motor with Full-Wave Control
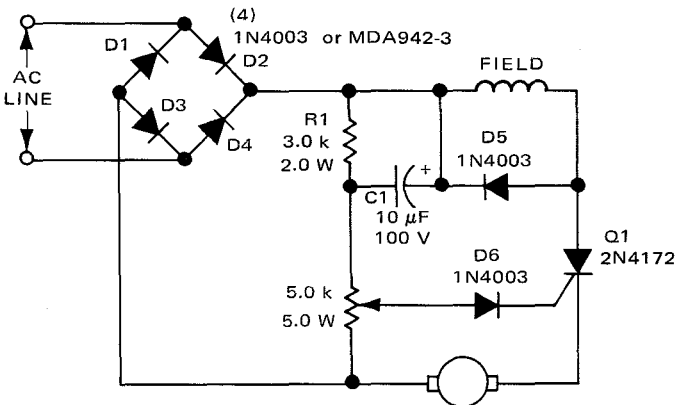
Figure 1-22 — Compensated Full-Wave Drive for Series-Wound DC Motors

This results in unstable speed control and can cause a pulsating speed. The speed-torque characteristics for this connection are shown in Figure 1-21. These curves exhibit the maximum flatness of this series. Of course, the motor characteristics as obtained when operated at rated dc voltage are the maximum that can be provided. Thus the speed-control circuit will provide effective control only up to this curve.

All of the curves shown were obtained for a 1/15 hp, 5000 r/min, 115 volt motor. This circuit is capable of controlling larger fractional
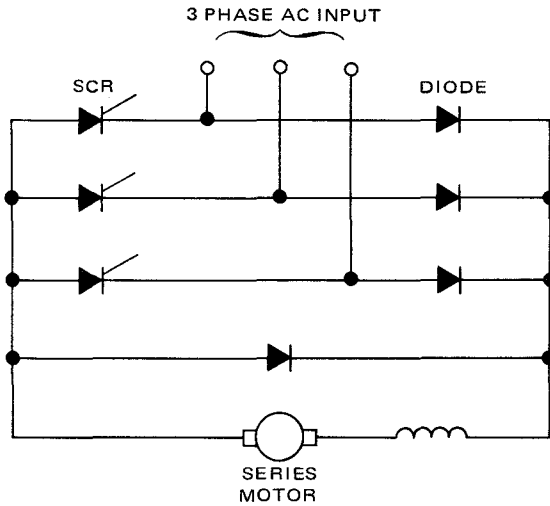
Figure 1-23 — Simplified Schematic of Half-Controlled, Full-Wave Bridge
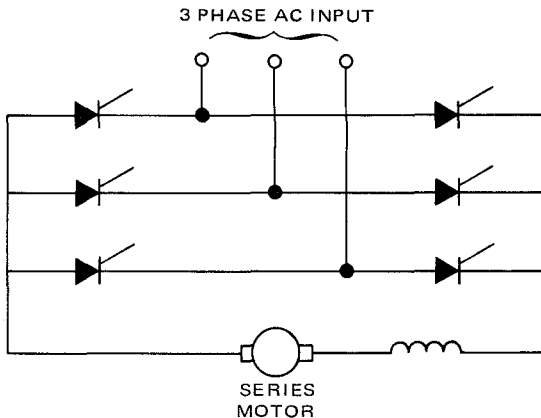
Figure 1-24 — Simplified Schematic of Fully-Controlled, Full-Wave Bridge

horsepower motors. One-quarter-horsepower motors have been controlled with the devices as used in the schematics under normal conditions. The real criterion is the average steady-state current and the stalled-rotor current under the operating conditions. These two conditions must fall within the current ratings of the SCR. In most cases, higher-current SCRs will work in this circuit. Since a fractional-horsepower motor was used, a single-phase ac supply was used as the main power source. These circuits may require some adaptation for any given motor to obtain performance similar to those shown here. For large integral-horsepower motors, it may be necessary to use a three-phase supply to obtain enough driving power. This complicates the control and rectifier circuitry. Rectifier circuits as illustrated in Figures 1-23 and 1-24 can be used on a three-phase power line. The harmonic content of the output voltage with the half-wave controlled circuit is of higher amplitude and lower frequency than that of the fully controlled circuit. The fully controlled circuit has less ripple, better transient response, less susceptibility to damage from misfirings, and is readily adaptable to regenerative braking (if two circuits are connected back to back).

1.5 Shunt-Wound DC-Motor Speed-Control Circuits

The circuit shown in Figure 1-25 provides half-wave control of a shunt-wound motor. The field is supplied through diode D1 during the half cycle that Q1 conducts. Diode D2 across the field winding is used as a free-wheeling diode.
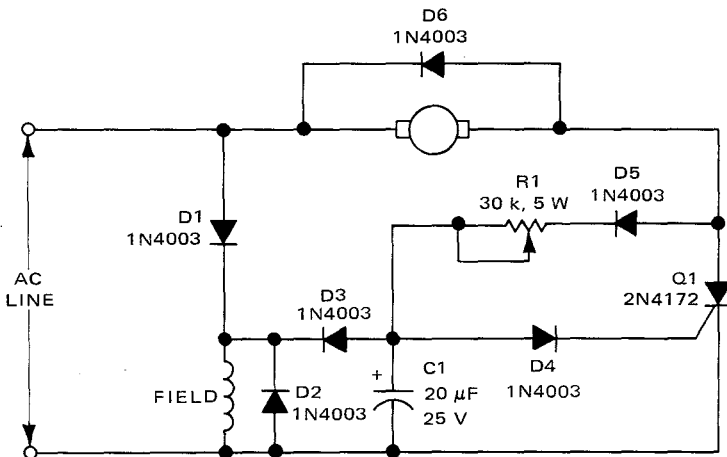
Figure 1-25 — Half-Wave Control of Shunt-Wound Motor

The average voltage applied to the armature is set by the conduction angle of SCR Q1. This angle is set by the charging rate of C1 through resistor R1. When the capacitor voltage exceeds the sum of the diode drops of D4 and the gate of SCR Q1, Q1 turns on. Q1 remains on until the current in this circuit goes to zero. Capacitor C1 is discharged through D3 every half cycle giving the same initial voltage from which the capacitor begins to charge.
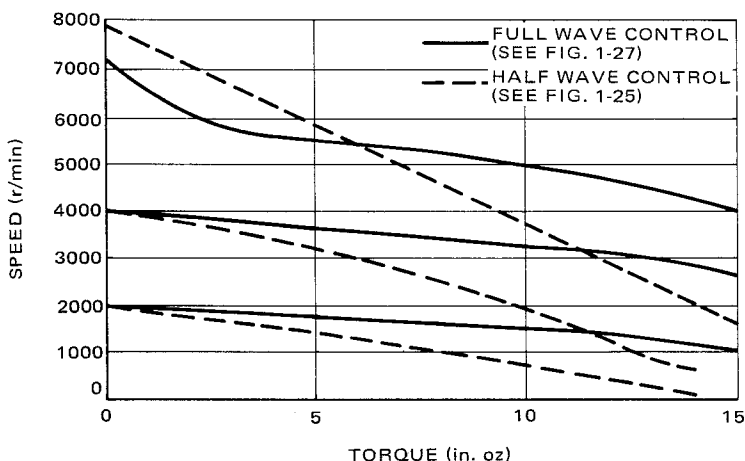
**Figure 1-26 — Speed-Torque Characteristics
of Controlled Shunt-Wound Motor**

The speed-torque curves for a 1/15 hp, 5,000 r/min, 115 V motor used with this circuit are shown in Figure 1-26. The curve for the maximum speed setting shows the droop obtained with this circuit. When the no-load speed setting is reduced, the typical shunt-wound characteristics are obtained, but with a reduction in the droop of the curve.

Improved speed-torque characteristics can be obtained through the use of full-wave control. The circuit shown in Figure 1-27 can be used to provide full-wave control of a shunt-wound motor. The operation of this circuit is similar to that of half-wave control, except that power is supplied from a full-wave bridge. Free-wheeling diode D5 connected across the armature is necessary in this circuit; without it, current would flow through SCR Q1 and the bridge rectifier, holding Q1 on. Four-layer-diode D8 was used to provide a stable firing point for Q1. The capacitor must charge to its breakover voltage before Q1 will turn on. At low speed settings and at light loads, some degree of erratic firing of Q1 was experienced. The maximum speed at which this occurred depends on the motor

and the degree of loading of that motor. Only at very low speed settings was this pulsating characteristic objectionable.

The curves obtained with this circuit and a 1/15 hp, 5,000 r/min, 115 V motor are shown in Figure 1-26. These curves show that the full-wave control provides much better speed regulation than the half-wave control circuit. These circuits may require adaptation for some motors to obtain characteristics similar to those obtained with this motor.
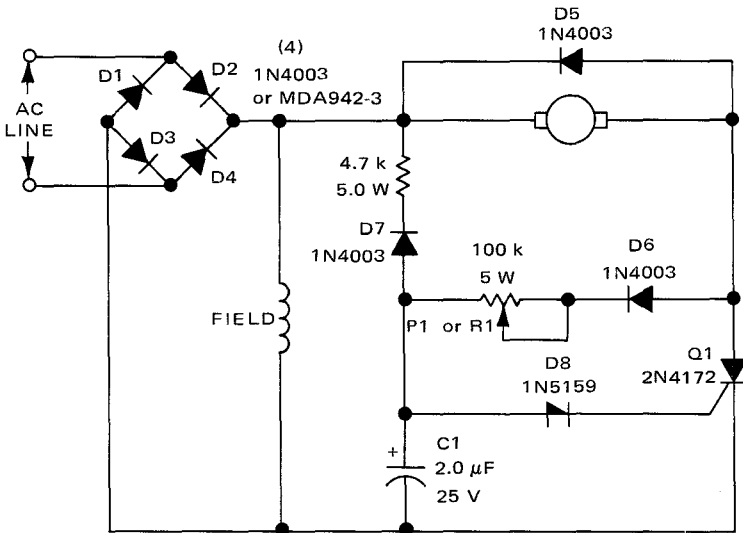
Figure 1-27 — Full-Wave Control of Shunt-Wound Motor

1.6 Direction- and Speed-Control Circuits for Series, Universal, and Shunt Motors

The circuit shown in Figure 1-28 can be used to control the speed and direction of rotation of a series-wound dc motor. Silicon controlled rectifiers Q1 through Q4 are connected in a bridge arrangement, and are triggered in diagonal pairs. Which pair is turned on is controlled by switch S1 since it connects either coupling transformer (T1 or T2) to a pulsing circuit. The current in the field can be reversed by selecting either SCRs Q2 and Q3 for conduction, or SCRs Q1 and Q4 for conduction. Since the armature current is always in the same direction, the field current reverses in relation to the armature current, thus reversing the direction of rotation of the motor.
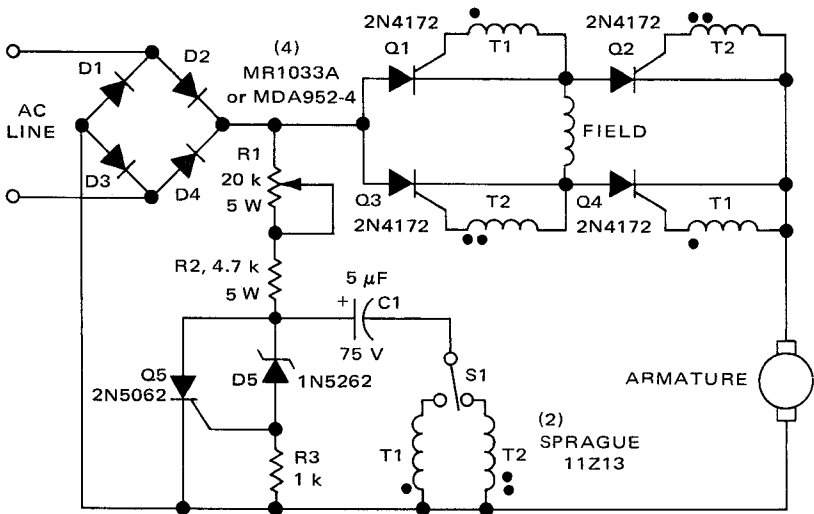
Figure 1-28 — Direction- and Speed-Control Circuit for Series-Wound or Universal Motor

A pulse circuit is used to drive the SCRs through either transformer T1 or T2. The pulse required to fire the SCR is obtained from the energy stored in capacitor C1. This capacitor charges to the voltage of zener diode D5 through potentiometer R1 and resistor R2. As the capacitor voltage exceeds the zener voltage, the zener conducts, delivering current to the gate of SCR Q5. This turns Q5 on, which discharges C1 through either T1 or T2, depending on the position of S1. This creates the desired triggering pulse. Once Q5 is on it remains on for the duration of the half cycle. This clamps the voltage across C1 to the forward voltage drop of Q5. When the supply voltage drops to zero, Q5 turns off, permitting C1 to begin charging when the supply voltage begins to increase.

The speed of the motor can be controlled by potentiometer R1. The larger the resistance in the circuit, the longer C1 requires to charge to the voltage of zener D5. This determines the conduction angle of either Q1 and Q4, or Q2 and Q3, thus setting the average motor voltage, and thereby the speed.

If a shunt-wound motor is to be used, then the circuit shown in Figure 1-29 is required. This circuit operates like the one shown in Figure 1-28; the only differences are that the field is placed across the rectified supply and the armature is placed in the SCR bridge. Thus the field current is unidirectional, but armature current is reversible, and consequently the motor's direction of rotation is reversible. Potentiometer R1 controls the speed as explained previously.
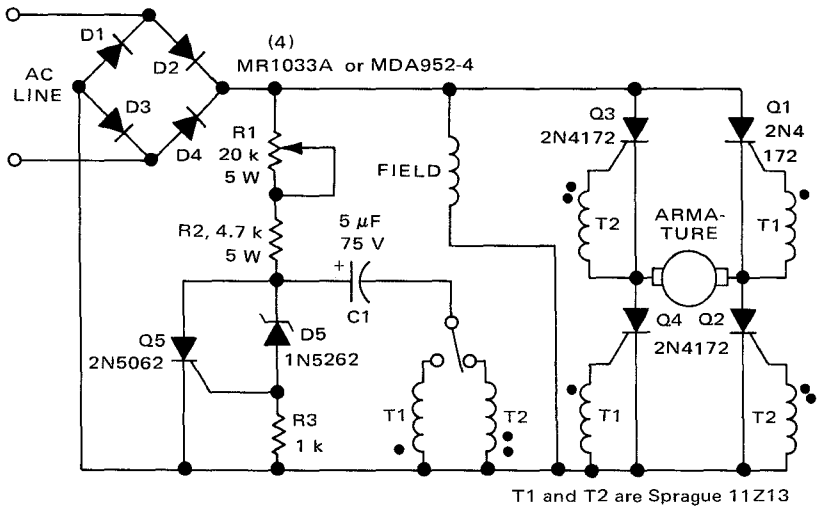
**Figure 1-29 — Direction- and Speed-Control
Circuit for Shunt-Wound Motor**

Excellent results were obtained when these circuits were used to control 1/15 hp, 115 V, 5,000 r/min motors. This circuit will control larger fractional-horsepower motors provided the motor current requirements are within the semiconductor ratings. Higher-current devices will permit control of even larger motors. The operation of the motor under worst case must not cause anode currents to exceed the ratings of the semiconductor.

1.7 Permanent-Magnet-Motor Speed Control

The circuit shown in Figure 1-30 can be used to control the speed of a permanent-magnet motor. It was tested with a 35 volt motor. A step-down transformer was used to permit circuit values that gave phase control over the full half cycle. This is not actually necessary as the SCR could have been phased back to keep the average motor voltage below its maximum rating. However, if a motor designed for a higher voltage were to be used, component values would have to be modified. If a transformer is not used, the maximum average voltage can be set by placing a resistor in the position shown by R1. The higher the value of this resistor, the lower the maximum average voltage will be.

The full-wave bridge (D1-D4) provides power to the motor and control circuits every half cycle. SCR Q1 is turned on when the charge on capacitor C1 exceeds the sum of the voltage drops in diode D5 and the
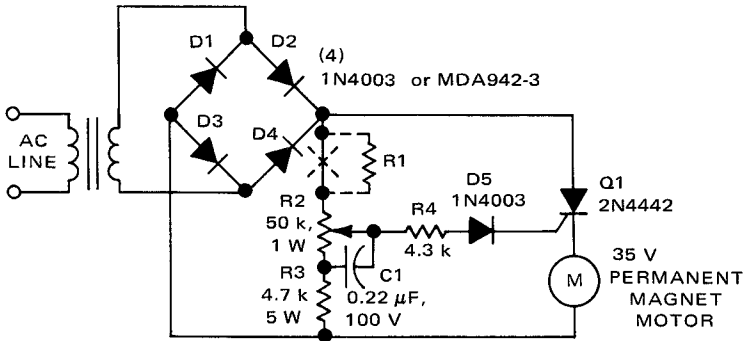
Figure 1-30 — Speed Control for a Permanent-Magnet Motor
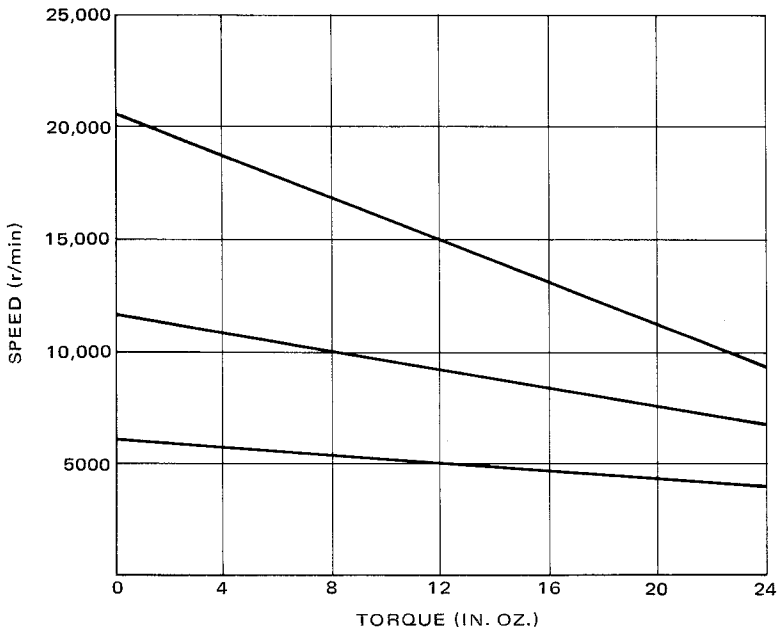
**Figure 1-31 — Speed-Torque Characteristics
of Controlled Permanent-Magnet Motor**

gate of Q1, plus the back emf of the motor. The time required to fire Q1 is set by the potentiometer, which governs the charging rate of the capacitor. As the motor load is increased, the back emf is reduced, and for any potentiometer setting, Q1 turns on sooner thereby compensating for the increased load.

1-28

This circuit demonstrated excellent speed control from 0 to 20,500 r/min as shown in Figure 1-31. When a load was placed on the motor, the control was smooth over the speed range. At light loads and slow speeds (up to 12,000 r/min), some degree of pulsation was noticed. This is caused by the back emf of the motor rising above the capacitor voltage setting, thus preventing SCR Q1 from turning on. When the motor slows down sufficiently, the SCR turns on and speeds the motor up; the process then starts over again. The amount of pulsation is small enough that it is not objectionable. When the motor is loaded, this effect disappears.

1.8 Pulse-Width-Modulated DC-Motor Speed Control

Figure 1-32 is the schematic diagram of a pulse-width-modulated dc-motor speed control. The circuit operates from a 12 volt dc source and is capable of driving motors with in-rush currents up to 20 amperes. The maximum allowable running current will obviously be less than 20 A, and will depend to a great extent on the heat sinking provided for Q5.

The modulated waveform is provided by the Schmitt trigger consisting of Q1 and Q2, phase-inversion stage Q3, and the delayed feedback, R3 and C1. The output is a variable-width, variable-frequency pulse waveform whose duty cycle and frequency are a function of the dc input. The dc input is the summation of the current through R2, which is connected to the speed-adjustment potentiometer R1, and the current through R_f, the overall feedback resistor.

The output of the modulator is fed to a Darlington-connected power-amplifier stage consisting of Q4 and Q5, which drives the dc motor. Free-wheeling diode D3 suppresses the inductive kickback of the motor. Diode D2 protects the base-emitter junction of Q5 against reverse breakdown due to voltage transients generated by the motor.

The overall feedback voltage used to maintain constant motor speed is generated by sensing the forward base-emitter voltage of Q5, causing Q6 to conduct a greater amount of current as the load is increased. The voltage at point A will then rise and Q1 will begin to conduct for a shorter period of time. This means that the duty cycle of Q2 through Q5 will increase and a larger average voltage will be applied to the motor, thus compensating for the increased load.

The curves shown in Figure 1-33 are the open-loop operating characteristics of a 12 volt, 4 ampere dc motor. Figure 1-34 shows the effect of controlling this motor with the circuit shown in Figure 1-32. As can be seen, the operating characteristics are considerably improved.
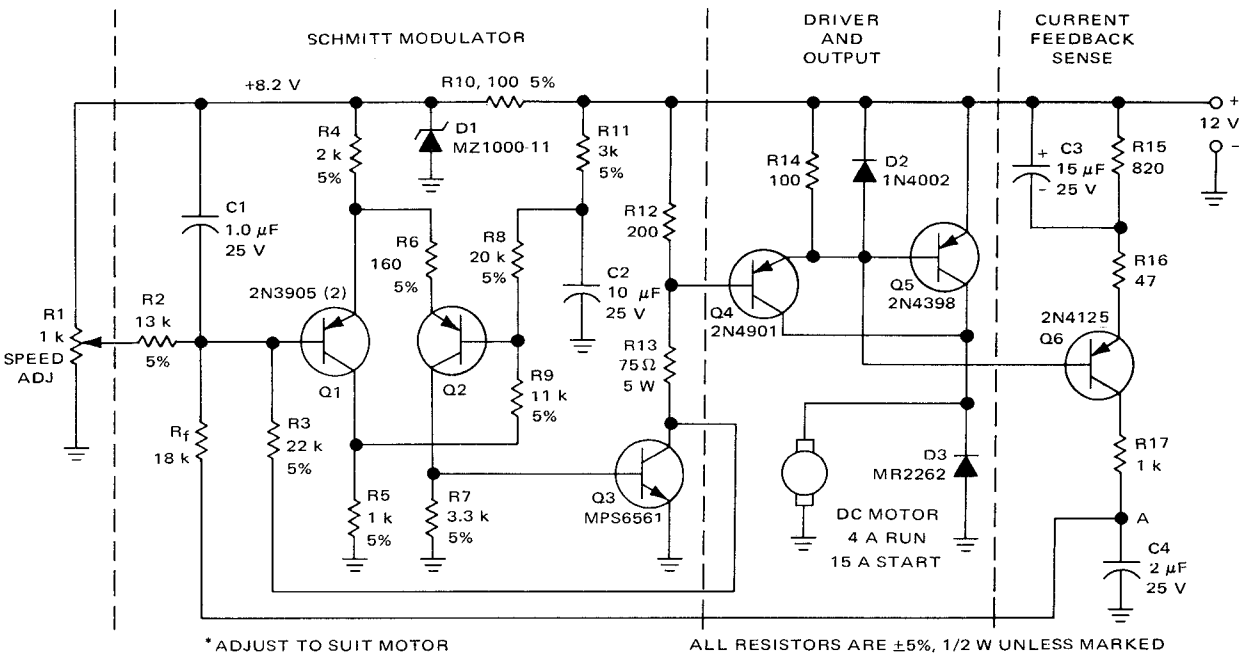
Figure 1-32 — DC Motor Speed Control with Current Feedback from V_{BE} Sensing

*ADJUST TO SUIT MOTOR

ALL RESISTORS ARE ±5%, 1/2 W UNLESS MARKED

Motor Speed Controls

The data for Figure 1-34 were taken at $-20°C$, $+25°C$ and $+65°C$, with a 12 volt supply voltage. Figure 1-35 shows the effects of varying the power-supply voltage at $25°C$. Zener diode D1 is used to provide a regulated voltage to speed-adjustment potentiometer R1, and to the two stages of the Schmitt trigger. This minimizes the effects of power-supply variations on this part of the circuit and tends to keep the duty cycle and frequency independent of the supply voltage. However, the average voltage applied to the motor is a direct function of the supply voltage. Thus some means must be provided to compensate for this, and that is the reason for
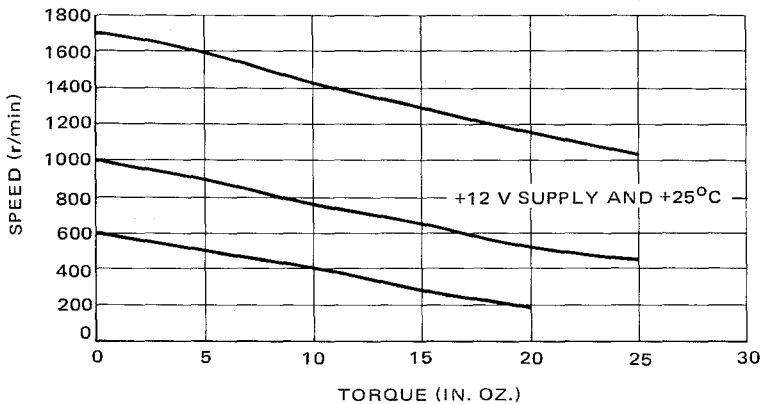
Figure 1-33 — Open-Loop Operating Characteristics of DC Motor
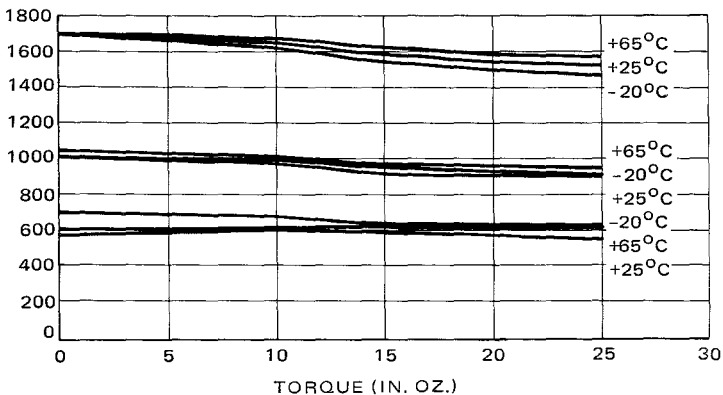
Figure 1-34 — Temperature Characteristics of Motor Controlled by Speed Control Circuit with Current-Sensitive Feedback
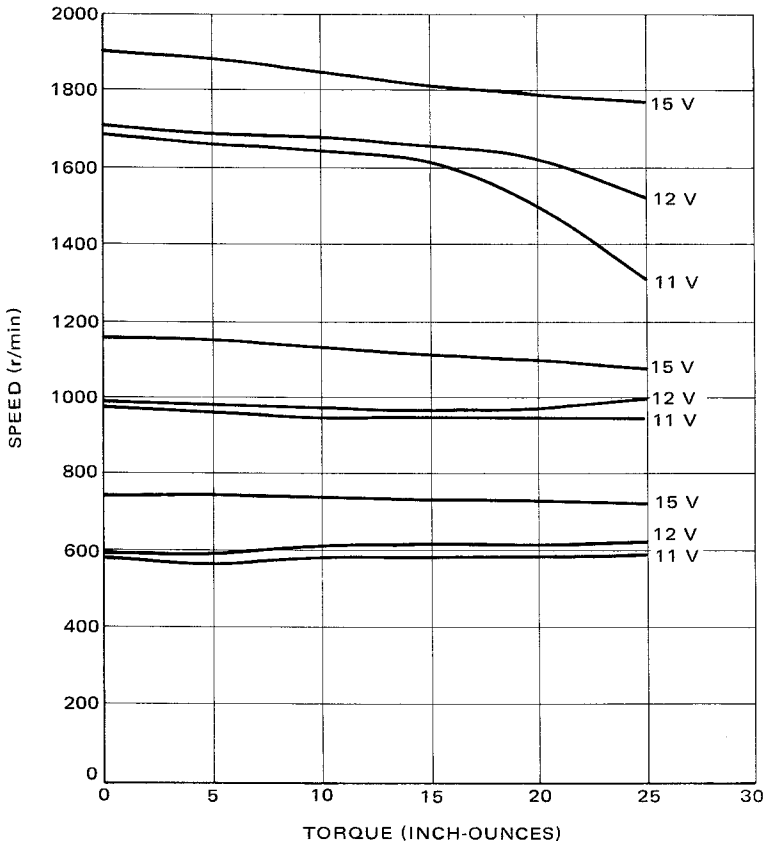
Figure 1-35 — Effect of Different Supply Voltages on Speed Control Circuit with Current-Sensitive Feedback

Parts List for Figure 1-36

CAPACITORS	RESISTORS		TRANSISTORS
C1 — 10 nF	R4 — 10 kΩ	R23 — 160 Ω	
C2 — 1.0 μF	R5 — 10 kΩ	R24 — 3.3 kΩ	
C3 — 8 μF, 15 V	R6 — 10 kΩ	R25 — 200 Ω	
C4 — 50 μF, 15 V	R7 — 10 kΩ	R26 — 75 Ω	
C5 — 1.0 μF	R8 — 2 kΩ	R27 — 100 Ω	
	R9 — 10 kΩ		
	R10 — 47Ω	TRANSISTORS	
DIODES	R11 — 10 kΩ	Q1 — MRD300	
D1 — 1N4001	R12 — 1 kΩ	Q2 — 2N4123	
D2 — 1N4001	R13 — 1 kΩ	Q3 — 2N4123	
D3 — 1N4001	R14 — 470 Ω	Q4 — 2N4221	
D4 — 1N4738	R15 — 47 Ω	Q5 — 2N3903	
D5 — 1N4001	R16 — 47 Ω	Q6 — 2N3903	
D6 — MR2262	R17 — 10 kΩ	Q7 — 2N4125	
	R18 — 22 kΩ	Q8 — 2N3905	
RESISTORS	R19 — 2 kΩ	Q9 — 2N3905	
R1 — 27 kΩ	R20 — 1 kΩ	Q10 — MPS6561	
R2 — 10 kΩ	R21 — 11 kΩ	Q11 — 2N4901	
R3 — 10 kΩ	R22 — 10 kΩ	Q12 — 2N4398	

Figure 1-36 — Regulated DC Motor Control with Feedback from Optical Pickoff

Motor Speed Controls

+12 V

LAMP

FIBER OPTICS

MOTOR ARMATURE

ALTERNATE BLACK AND WHITE STRIPES — 10 EACH

PULSE SHAPER

TACH

LEVEL SHIFTER

COMPARATOR

+8.2 V

SPEED ADJ

12 V

Q1 R3 R5 C1 D1 Q4 R7 R9 R10 D4 R15 R16

R1 D3 Q5 Q6 Q7 R11 R12 C4

R2 Q2 R4 Q3 D2 C2 R6 R8 C3 R13 R14

SCHMITT MODULATOR

DRIVER AND OUTPUT

R17 C5 R19 R22 R25 R27 D5

R23

Q8 R21 Q9 R26 Q11 Q12

R18 R20 R24 Q10 M D6

DC MOTOR
4 A RUN
15 A START

connecting the voltage divider consisting of R5, R8, R9 and R11 to the 12 volt supply. If the voltage increases, Q2 will conduct for a shorter time, reducing the duty cycle of the voltage applied to the motor, thereby keeping the average voltage seen by the motor constant.

Although this circuit can significantly improve the speed-torque characteristics of the motor, it has one major disadvantage. Resistor R15 (820 ohms in Figure 1-32) must be selected for the particular motor used and for the base-emitter drop of power transistor Q5. This can be accomplished by setting the no-load speed at one half the rated speed with rated dc applied. The motor is then loaded to full load, and R15 selected to restore the speed to the no-load speed.
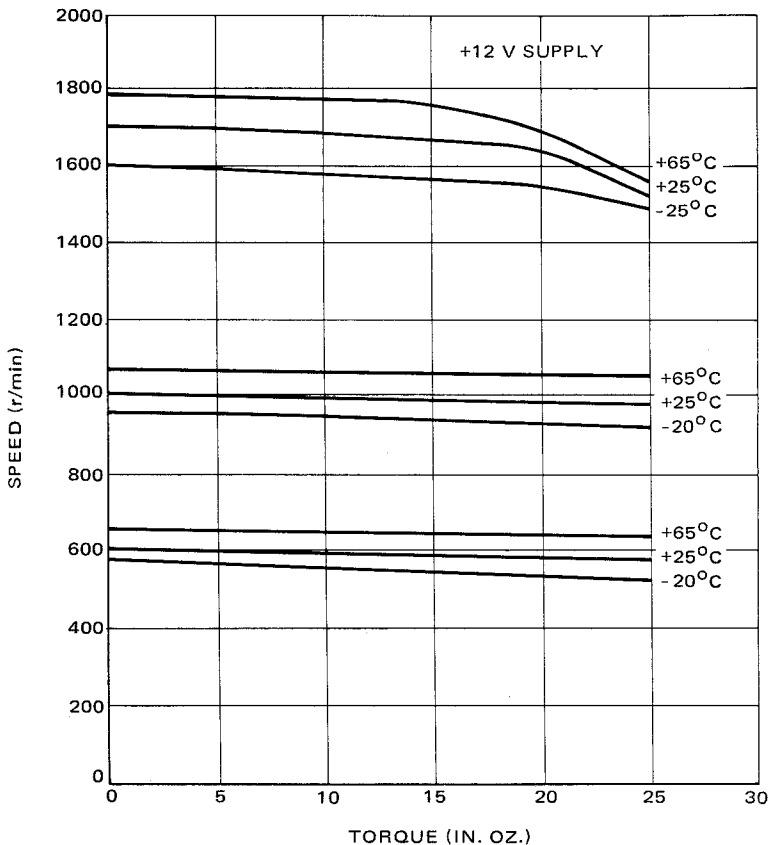
Figure 1-37 — Temperature Characteristics of Motor Controlled by Speed Control Circuit with Optical Feedback

To further improve the motor's speed-torque characteristics, the circuit shown in Figure 1-36 can be used. This circuit illustrates a dc-motor speed control using an optical pick-off to complete the closed-loop system.

The motor armature is painted with alternate black and white stripes. A total of twenty stripes was used for the circuit show. A bundle of noncoherent fiber optics is used to transmit light from a 12 volt lamp to the armature, and transmit reflected light from the armature to the input of Q1, a MRD-300 phototransistor. The light transmitted from bulb to armature to phototransistor is chopped at a frequency determined by the speed of the motor.

The output of the phototransistor is fed into a pulse-shaping circuit, and then into a tachometer circuit whose dc output is proportional to the input frequency. The output of the tachometer is fed to Q4, a junction field-effect transistor. The JFET provides a high input impedance to minimize loading on the tachometer circuit, and a reasonably low output impedance to drive the differential amplifier. In addition, it acts as a level shifter to ensure that there is suffecent output to bias the differential amplifier when the tachometer output is zero. The diode in series with the source lead of the FET is used for temperature compensation.

The differential amplifier compares the fixed-speed-adjust voltage from R13 to the output of the tachometer level shifter, and generates an output voltage that is proportional to the difference, or error, between the two. Capacitor C3 is used to prevent motor-speed overshoot when the speed-adjustment potentiometer is changed rapidly.

The T network (R15, R16 and C4) and zener diode D4 are used to reduce supply voltage variations to the pulse shaper, tachometer, level shifter and comparator circuits.

The output of the comparator is amplified and fed into a pulse-width modulator similar to the one shown in Figure 1-32.

The output of the modulator is then fed into a Darlington circuit consisting of Q11 and Q12, which drives the motor.

The feedback for this circuit is derived from a direct measurement of the armature speed of the motor. This permits the speed of the motor to be well regulated as shown in Figure 1-37. An alternate method of temperature-compensating the circuit can be accomplished by removing diode D3 and adding diodes in series with R9. Two or more diodes will probably be necessary for best temperature compensation, but one may yield satisfactory results.

Figure 1-38 shows the results of power-supply voltage variations. The effects at lower supply voltages are greater since the pulse width reaches 100% and cannot be increased further to provide additional average motor voltage required by the load.
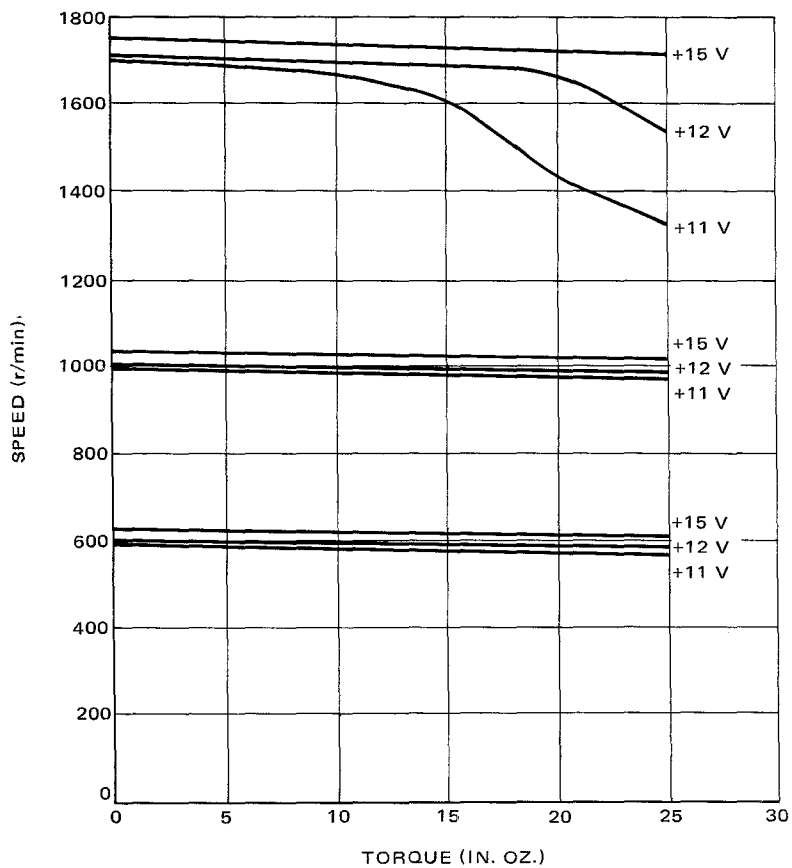
Figure 1-38 — Effect of Different Supply Voltages on Speed Control Circuit with Optical Feedback

BIBLIOGRAPHY FOR CHAPTER 1

1. Electric Machinery, Clifford C. Carr, John Wiley & Sons, Inc., New York, 1958.

2. Electro-Technology, June, 1967, Vol. 79, No. 6, "Matching DC Drives and Motors," C. J. Newell.

3. IEEE International Conference Record, Vol. 14, Part 8, 1966, "Pulse Width Modulated Inverters for AC Motor Drives," Boris Mokrytzki – Reliance Elec. & Eng. Co., Cleveland, Ohio.

4. IEEE Transactions on Industry and General Applications, IGA-2 No. 4, July/Aug., 1966, "Precise Control of Three Phase Squirrel Cage Induction Motor Using Practical Cycloconverter," Walter Slabiak and Louis J. Lawson.

5. IEEE Transactions on Industry and General Applications, IGA-3, No. 2 Mar/Apr, 1967, "Thyristor Adjustable Frequency Power Supplies for Hot Strip Mill Run-Out Tables," R. A. Hamilton and George R. Lezan.

6. IEEE Transactions on Industry and General Applications, IGA-2, No. 2, Mar/Apr, 1966, "Solid State Control for DC Motors Provides Variable Speed with Synchronous Motor Performance," E. Keith Howell.

7. Mullard Technical Communication, No. 80, March, 1966, "Thyristor Speed Control of DC Shunt Motors from a Single Phase Supply," J. Merret.

CHAPTER II

Inverters and Converters

2.1 Inverter Circuits

An inverter circuit is used to convert a dc input voltage to an ac output voltage. Subsequent rectification of the inverter ac output results in dc-to-dc conversion, or a converter circuit. In converters, output frequency and waveshape of the basic inverter output are relatively unimportant, except as they relate to efficiency of the converter and the filtering of the rectified output. When designing inverters, however, these and other characteristics must be specified before the proper circuit can be chosen.

Some inverter characteristics which may be important in various applications are listed in Table 2-1. It is the function of the inverter specifications to indicate which are important to the use intended, and to state, via tolerances, how critical each of these is expected to be. Obviously, the greater the number of controlled characteristics required and the tighter the tolerances, the more difficult and costly the design will be. In the absence of specifications requiring emphasis on other characteristics, inverters are usually designed for highest efficiency. Achievement of very precise frequency and output performance, etc. is usually at the expense of efficiency as well as circuit simplicity.

Semiconductor devices used in inverters are of two types, thyristor (generally SCR) and transistor. Some designs, of course, use both devices. Present day SCR inverters work best in an input voltage range from 50 to 600 volts dc and an input current range from 1 to 20 amperes. Transistor inverters are generally used with lower input voltages of about 1 to 100 volts and higher input currents (1 to 100 A). The typical output voltage of both types is a square wave which may be filtered to obtain a sine wave if necessary. The normal range of output power is from 1 watt to 1 kilowatt. The range of output frequencies normally encountered is 60 hertz to 100 kilohertz. The output voltage can be almost any value; it is determined by the input voltage and the output transformer turns ratio. Elaborate inverter circuits can be devised to extend the range of these characteristics. Series and parallel connections and time-sharing techniques are among those commonly used.

TABLE 2-I

COMMONLY SPECIFIED INVERTER CHARACTERISTICS

Input voltage: range and nominal

Output power

Output voltage

Output frequency accuracy

Regulation of output voltage and frequency vs. load and input voltage

Load power factor

Output waveform

Harmonic distortion of output, if sinusoidal, vs. load, power factor, input voltage

Overall efficiency vs. loading

Operating environments (Temperature, etc.)

Size and Weight

Protection required (as against shorted output, reversed polarity input, etc.)

SCR inverters can be conveniently categorized by identifying the method used to commutate or switch off the devices. Series-commutated inverters make use of a capacitor in series with the load to bring the load current to zero and turn off the appropriate SCR. When the capacitor is in parallel with the load, it is used to reverse the anode voltage on an SCR to turn the device off. This arrangement is called a parallel-commutated inverter. Finally, the inpulse-commutated inverter makes use of an auxiliary LC tank to store an impulse of energy that will be used to turn off the SCRs.

Transistor inverters may be identified by their feedback. In current-feedback inverters, a portion of the load current is fed back to drive the transistor base circuit. The voltage-feedback inverter samples a portion of the output voltage and uses it for transistor base drive. Finally, the hybrid inverter makes use of both types of feedback.

These circuits, both thyristor and transistor, have unique character-istics which make them desirable for specific applications. One circuit may have high efficiency while another may give excellent frequency regula-tion. The sections which follow will explain the theory of operation of these circuits as an aid to design. These sections will also describe the advantages and disadvantages of each circuit in an attempt to fit the circuit to a particular application.

The final section in this chapter contains a representative sample of practical inverter circuits. The circuits can be used as is or modified as desired.

2.2 Series-Commutated SCR Inverters

The basic circuit for series-commutated SCR inverters is shown in Figure 2-1. Capacitors C1 and C2 are equal and inductor L is center tapped. Resistor R is the load resistor through which a sine wave of cur-rent is to flow. The operation of the inverter starts with both SCRs off and with $E_S/2$ volts on each capacitor. When Q1 is turned on, the voltage across the upper half of L will be $E_S/2$. Current will start to flow as C1 discharges and C2 charges through the load and the upper half of L (see waveform in Figure 2-2). Current will be a maximum when the voltage across L reaches zero. With negligible load, the voltage across C1 will be zero and C2 will have charged to the battery voltage. Now as the field of inductor L collapses, C1 will be charged to $-0.5\ E_S$ and C2 to $+1.5\ E_S$ (reference battery ground). When the current drops to zero, Q2 is turned on and the voltage across the lower half of L will be $1.5\ E_S$. Autotrans-former action induces this same voltage in the upper winding and causes Q1 to be reverse biased by twice the battery voltage E_S. This insures that Q1 will turn off very quickly. Now current will again start to flow but in the reverse direction as C2 discharges and C1 charges through the load and
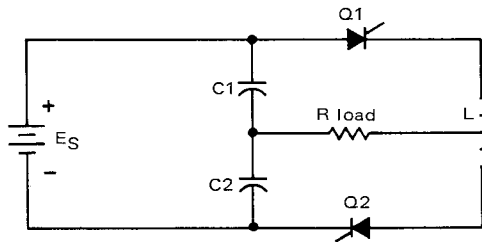
Figure 2-1 — Series-Commutated SCR Inverter

the lower half of L. This time the inductor L will charge C2 to -1.5 E_S and C1 to $+2.5$ E_S. The cycle is now complete. The average voltage across C1 and C2 will remain $E_S/2$ but the peak voltage will continue to increase until the positive alternation becomes equal to the negative alternation. The amount of load resistance will be the limiting factor for this final value of capacitor voltage.
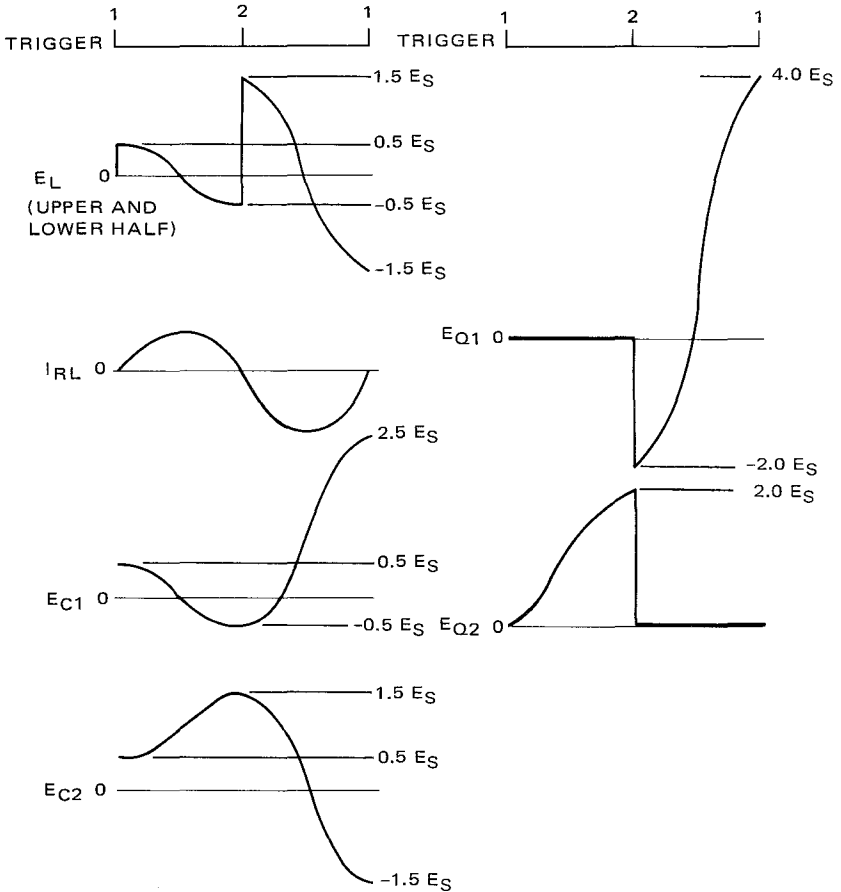
Figure 2-2 — Waveforms of Series-Commutated SCR Inverter

The peak value of capacitor voltage can be found using[*]

$$E_C = \frac{E_S}{2} \times \frac{(1 + e^{-\pi/2Q})}{(1 - e^{-\pi/2Q})} ,$$

where

$$Q = \frac{\sqrt{L'/C}}{R} \geqslant 1/2,$$

$$L' = L/4 \text{ (half of load coil)},$$

and $C = C1 + C2$ (sum of capacitors),
where the units are ohms, microfarads and microhenries.

The resonant frequency is determined by the values of L' and C:

$$f_R = \frac{1}{2\pi\sqrt{L'C}}$$

The circuit can be controlled solely by operation of the thyristor gates and operation above or below resonance is possible. The maximum frequency occurs when the circuit turn-off time has decreased to the value required by the characteristics of the SCR. The peak load voltage equals E_C/Q and becomes about 0.6 E_S with $Q = 4$. Four is the optimum value for factor Q although the circuit will operate with any value of Q greater than $1/2$. This means that the load resistance may be varied from twice the circuit impedance ($R = 2\sqrt{L'/C}$) to zero, but that a good choice for R is one-quarter of the circuit impedance ($R = 1/4\sqrt{L'/C}$). As resistance increases above this value the sine wave output becomes distorted. Critical damping occurs at the upper limit where the thyristors fail to commutate (switch off). Lower values of resistance increase the reactive voltages and decrease the load voltage. If the circuit components were rated at higher voltages the load could be short circuited.

The series inverter is the easiest way to obtain a sine wave output voltage for a resistive load. Because a series inverter is self-commutating, it is not necessary to remove the input voltage to stop the circuit. Operation can be started and stopped by controlling the trigger pulses. It can also be designed so that a short circuit in the load will not destroy any components. Output voltage regulation, however, is generally quite poor for variations in the load.

The highest frequency of operation is generally limited by thyristor turn-off time to about 10 kilohertz. It is possible to increase this upper

---

[*]Bedford, B. D. and R. G Hoft, *Principles of Inverter Circuits*. John Wiley and Sons, Inc. 1964.

frequency limit by an order of magnitude using the circuit shown in Figure 2-3. If the thyristors are triggered in the sequence indicated by their numbering, each device will have one cycle to turn off. New limitations will be present with this circuit, the most predominate of which is the rate of rise of forward voltage.
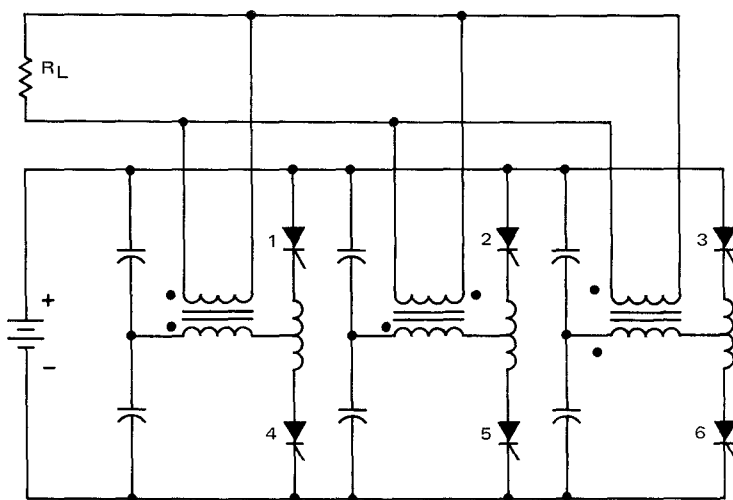
Figure 2-3 — Series-Commutated Time-Sharing SCR Inverter

2.3 Parallel-Commutated SCR Inverters

The basic circuit for parallel-commutated SCR inverters is shown in Figure 2-4, with waveforms taken from the circuit shown in Figure 2-5. SCR pair Q1 and Q2 is triggered alternately with pair Q3 and Q4 to produce a square wave across load resistor R (Figure 2-5D). The voltage across Q1 and Q2 is E_{12} and is shown positive when current flow is from Q1 through R to Q2. The voltage across R is E_R and is shown positive when the left side is positive with respect to the right. When Q1 and Q2 are turned on, the battery voltage, E_S will appear across inductor L, and current will start to flow. Current will flow from E_S through L, Q1, C, R, and Q2. At this point, there will be a constant current flow through L approximately equal to E_S/R (the voltage across R is E_S minus the sum of the voltage drops across Q1 and Q2.) During this time, the total voltage across Q1 and Q2 is approximately 1 V as shown by E_{12}. When Q3 and Q4 are turned on, C will begin to discharge. One path of discharge current will be through Q1 and Q3; another will be through Q4 and Q2. Reverse
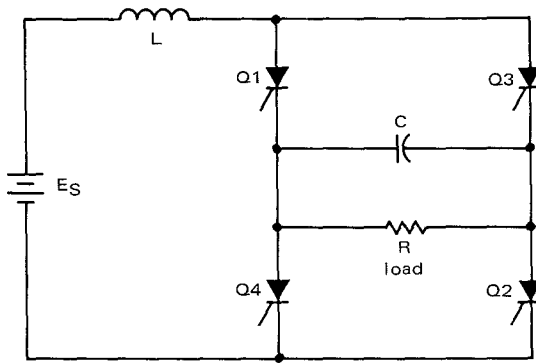
Figure 2-4 — Parallel-Commutated SCR Inverter

current will flow in Q1 and Q2 until the stored internal charge is removed and the devices cease conducting. Current through L will now be diverted from Q1, R, and Q2 to Q3, C, and Q4, and will begin to increase (see Figure 2-5B). The load voltage E_R is held positive by C and will not become negative until the current flowing through Q3 and Q4 charges C to E_S in the reverse direction (positive on the right). During the time E_R remains positive, E_{12} remains negative, and Q1 and Q2 are able to recover their forward blocking ability. So when E_R becomes negative, Q1 and Q2 will remain off, and E_{12} will go positive. Current through L reaches a maximum when E_R is approximately equal to E_S. The collapsing magnetic field will now cause a slight negative overshoot on E_R. Finally, a constant current will flow through L, and E_S will appear in reverse across R. A similar sequence of events will occur when Q1 and Q2 are triggered and E_R again becomes positive.

The design of this circuit would begin with the selection of the input voltage and load. The next step would be to choose the SCRs and a trigger circuit to meet the power and frequency requirements. Then values for the commutating capacitor and inductor should be determined: let the R-C time constant be twice the turn-off time (t_{off}). Then $R C = 2 t_{off}$ and $C = 2 t_{off}/R$. Finally let the L-R time constant be equal to the R-C time constant to obtain critical damping. Then $L/R = CR$ and $L = CR^2$.

If the inductance is reduced by a factor of three, the circuit will be underdamped.* The load voltage shown as E_{RU} in Figure 2-5E will result. Smaller inductance would cause more ringing and the circuit would turn itself off following commutation.

---

*Mapham, Neville "An SCR Inverter with Good Regulation and Sine Wave Output," *IEEE Transactions* IGA-3 No. 2 Mar/Apr 67.
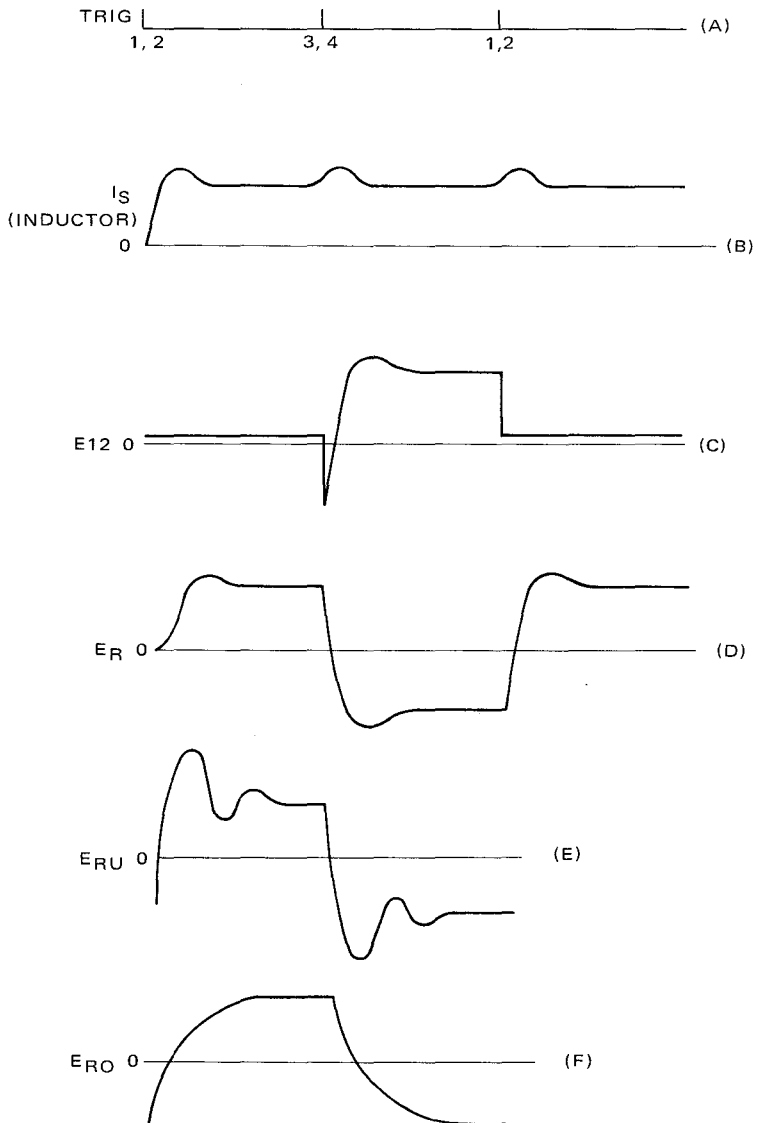
Figure 2-5 — Waveforms of Parallel-Commutated SCR Inverter

An overdamped load voltage is shown in Figure 2-5F as E_{RO}. The overdamped load voltage occurs when the inductance is increased by a factor of four.* Therefore, the actual value of the inductance should not be greater than 4L nor less than L/3. If the load is to be fixed, L should be chosen for critical damping. If the load is to be variable choose C for minimum load (R_{min}) and L for overdamping ($L = 4\ CR^2_{min}$). The load resistance can then be increased by a factor of 3.4 before the circuit will ring off. Any decrease in resistance would tend to shorten the turn-off time and commutation would then become unreliable.

It is possible to obtain any output voltage by replacing load resistor R with an output transformer. If the primary of this transformer has a center tap, SCRs Q1 and Q3 can be removed. The circuit would then appear as shown in Figure 2-6. In this case both C' and R' must be reflected across one-half of the transformer primary and the values to be used are $C = 4C'$ and $R = \left(\dfrac{n_1}{n_2}\right)^2 R'$.
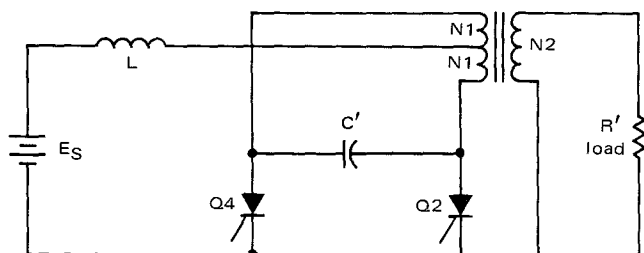
**Figure 2-6 — Parallel-Commutated SCR Inverter
with Center-Tapped Transformer Load**

This approach to a parallel inverter is about the simplest way to obtain a square wave from an SCR inverter. Even the triggering is simplified since both SCRs have a common ground. However, the output waveform will always have an exponential leading edge and, the waveform is also affected, as we have shown, by variations in load. Because the turn-off time can represent a considerable portion of each alternation, the operating frequency will normally be less than 2 kilohertz for most available SCRs.

*See Reference Four.

2.4 Impulse-Commutated SCR Inverters

The basic circuit for impulse-commutated SCR inverters is shown in Figure 2-7. Battery voltages E1 and E2 are equal. Either thyristor Q3 or Q1 must be triggered first to store turn-off energy in capacitor C1. Assume Q3 is triggered first. Current will flow from battery E1 through inductor L1, capacitor C1, and load resistor R1. When this current ceases to flow, C1 will be charged to E1 (positive on the side connected to L1) and Q3 will turn off. Q4 can be turned on next and current will then flow from E2 through R1 and Q4. At this point the battery side of R1 will be positive, and C1 will remain charged as before. Using this condition, we can now refer to the waveforms in Figure 2-8A. A trigger pulse is applied to Q1. The capacitor current is initially zero. E_R, the voltage on resistor R1, is positive and the capacitor voltage, E_C, is negative. Both are approximately equal to E1. The current flowing through R1 (I_L) is approximately equal to E1/R1. C1 will now discharge a sine wave pulse of current through L1, Q1, E2 and R1. As I_C increases toward the value of load current I_L (shown dotted), the current through Q4 will decrease to zero at t1. When I_C exceeds I_L, E_R will increase until diode D2 is forward biased. The amount of I_C that exceeds I_L will therefore flow through D2. During this time, Q4 is reversed biased by D2 and will turn off. When I_C is maximum, E_C is zero and energy is now stored in the magnetic field of L1. As this field collapses, I_C will decrease below I_L and D2 will become reverse biased at t2. Now charging current will flow from E2 through R1, C1, L1, and Q1, and E_R will decrease. When C1 becomes charged at t3 to E2, E_R will be zero, and Q1 will turn off. The triggering of Q2 and Q3 will create a similar sequence of events for the negative alternation of load voltage (E_R).
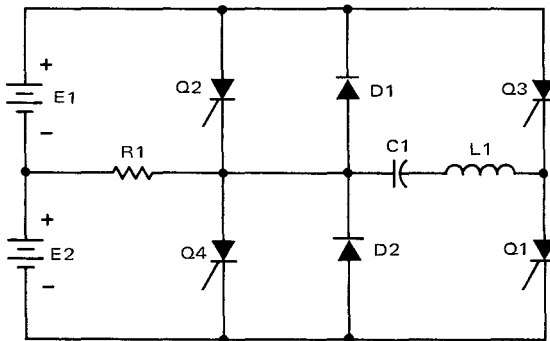
Figure 2-7 — Impulse-Commutated SCR Inverter

An alternate triggering sequence is shown in Figure 2-8B. Q1 has just been triggered; E_C is negative and Q4 is on. During the time D2 is forward biased, I_C exceeds I_L, and Q4 is recovering its blocking capability. At the point where I_C decreases below I_L, the turn off of Q4 is completed and Q2 may be triggered. The load voltage E_R will reverse instantaneously at this point. Since Q2 was triggered before I_C reached zero, Q1 will still be on. This means that charging current for C1 will now flow from both E2 and E1 through Q2, C1, L1, and Q1. The net result is that the final voltage on C1 will be about double that of the previous operation.
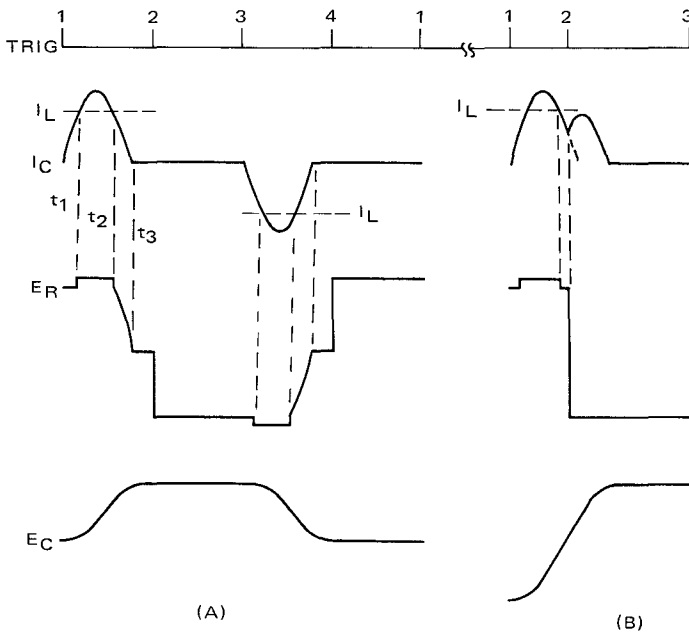
Figure 2-8 — Waveforms of Impulse-Commutated SCR Inverter

In the design of this circuit, the supply voltage and load are selected first. The center-tapped supply may be obtained by connecting two capacitors in series across a single dc supply. Any value of load voltage may be obtained by using an output transformer. After the thyristors and diodes have been selected, a triggering circuit can be designed. A variable-time ring counter is often a good choice. Finally, values for L1 and C1 must be determined. Let I_C be 1.5 I_L, and the pulse width ($\pi \sqrt{LC}$) be at least double the turn-off time (2 t_o) because this pulse shape requires the least

energy.* Since I_C is determined by E_C and the characteristic impedance (Z_0), we obtain

$$1. \quad I_C = \frac{E_c}{\sqrt{L/C}} = 1.5\, I_L \ \text{ or } \ Z_0 = \sqrt{L/C} = \frac{E_c}{1.5\, I_L},$$

where E_c is either equal to E1 or 2E1 depending on the mode of operation. The remaining condition is

$$2. \quad \pi\sqrt{LC} = 2\, t_0 \ \text{ or } \ f = \frac{1}{2\pi\sqrt{LC}} = \frac{1}{(4\, t_0)},$$

where f is the resonant frequency. Knowing f and Z_0, values for L1 and C1 may now be obtained directly from a reactance chart.

In the mode of operation shown in Figure 2-8A, turn off can occur at any time. This means that the output voltage can be pulse-width modulated. In the mode of 2-8B, a near-perfect square wave and fast switching and no overshoots is obtained. Very little power is lost in the commutating circuit so the circuit is very efficient. The circuit can be operated with no load and with a reactive load. The limiting value of load current (I_L) is that chosen for use in equation (1). Commutation becomes unreliable for load currents which exceed this value. It is possible to operate this circuit with any frequency up to about 12 kHz. This limit is imposed by a typical turn-off-time requirement of 20 μs. The one disadvantage of this circuit is that it requires four thyristors and a fairly complex triggering circuit.

2.5 Current-Feedback Transistor Inverters

The basic current-feedback, two-transformer transistor inverter is shown in Figure 2-9. Possible modifications include the use of NPN transistors (with proper power supply polarity) and/or common collectors, with feedback windings in the emitters. In the circuit of Figure 2-9, feedback transformer T1 functions as a current transformer to provide base drive current proportional to collector current. If Q1 is assumed to be conducting, $I_{B1} \approx (N_C/N_B)\, I_{C1}$. This flow of base current causes a voltage drop in the base circuit of Q1. In the absence of other series voltage drops, the voltage developed will be the V_{BE} of the conducting transistor and voltage

---

*Bedford, B. D. and R. G. Hoff, *Principle of Inverter Circuits*. John Wiley and Sons, Inc. 1964.
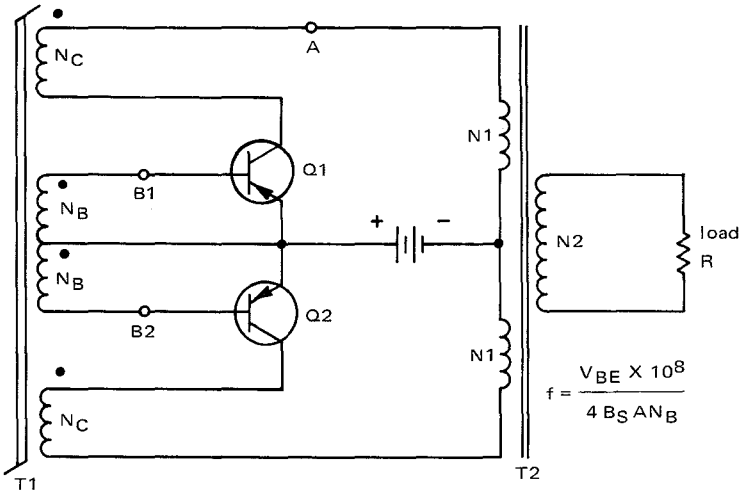
Figure 2-9 — Current-Feedback Two-Transformer Transistor Inverter

across N_B will be clamped at V_{BE}. This voltage across N_B determines how long it will take T1 to saturate and establishes the operating frequency of the inverter according to equation (1), where $V = V_{BE}$, $N = N_B$, and $\beta_s A$ is the flux capacity of the core of T1:

$$f = \frac{V \times 10^8}{4\,\beta_s\, A\, N} \; .$$

β_s is the maximum flux density in gauss

A is core area in cm^2.

(1)

When T1 saturates, the conducting transistor Q1 loses base drive, and the inverter switches to the other transistor.

Collector current I_C results in $H_1 = \dfrac{I_C N_C}{\ell}$ where ℓ is core magnetic path length. Prior to core saturation, only a small portion of this applied magnetic-field intensity is required for core magnetization, and current induced in base winding N_B will be approximately $(N_C/N_B)\,I_C$. As the core saturates, increasing current required for core magnetization detracts from the available base drive, causing the conducting transistor to turn off, and the other transistor to switch on regeneratively.

Because V_{BE} is relatively insensitive to inverter input voltage and load current, the frequency of the basic current feedback inverter (Figure 2-9) is correspondingly independent of input and load. However, the following precautions are necessary: The duration of the half period associated with conduction of each transistor is proportional to the V_{BE} of that respective transistor. To avoid half-period asymmetry which would result in net direct current in transformer T2, it is recommended that the two transistors be matched for V_{BE}. Further, since V_{BE} is typically one volt or less, considerable deviation from calculated frequency can be caused by even tenths of a volt of base-circuit I-R drops such as may be caused by appreciable series resistance in winding N_B. Empirical adjustment of the number of turns (N_B) is generally easier than taking winding resistance into account in the original calculation.

Current-feedback inverters are easily adapted to stabilization or control of inverter frequency by the addition of controlled series voltage drops in the base circuits. Several possibilities are shown in Figure 2-10. In each case, frequency is proportional to the sum of V_{BE} and the added series voltage drop.

In Figure 2-10A, a zener diode in series with V_{BE} reduces the effect on frequency of variations in V_{BE} due to changing load or temperature. If V_{BE} is small compared to zener voltage V_Z, good frequency accuracy is possible.

A diode in series with the drive circuit is used in Figure 2-10B to compensate for the negative temperature coefficient of β_s, and provide almost constant frequency over a wide temperature range. A resistor in series with the base of the transistor may be used to obtain a series voltage that will provide a frequency proportional to $V_{BE} + I_B R_B$. With a variable resistor, as shown in Figure 2-10C, frequencies over a range exceeding ten to one have been observed. Also, frequency variation or precise frequency control may be obtained by using a voltage-regulating transistor in the feedback circuit, as shown in Figure 2-10D.

In adding series voltage to the base circuit, care must be taken not to exceed BV_{EBO} of the nonconducting transistor, or a protective external diode must be placed in the base circuit. In the configurations shown in Figure 2-10, which have series elements common to both base circuits, reverse emitter-base voltage will be twice the induced series voltage plus V_{BE} of the conducting device. Common-collector circuits or other configurations having independent series elements in each base circuit will result in emitter-base reverse bias equal to the series voltage plus V_{BE}. Note, however, that when independent series elements are used, the voltages

Figure 2-10 — Stabilized Current-Feedback Inverters

developed across each should be equal, so that the two half cycles are of equal duration.

Some means must generally be provided to assure starting of current feedback inverters. Frequently, introduction of a slight circuit imbalance is adequate to initiate regenerative conduction of one transistor and begin oscillation. Such imbalance may consist of a resistor from point A to point B1 in Figure 2-9, or a capacitor from A to B2. For severe starting requirements, more sophisticated starting means, such as the circuit of Figure 2-11, may be used.
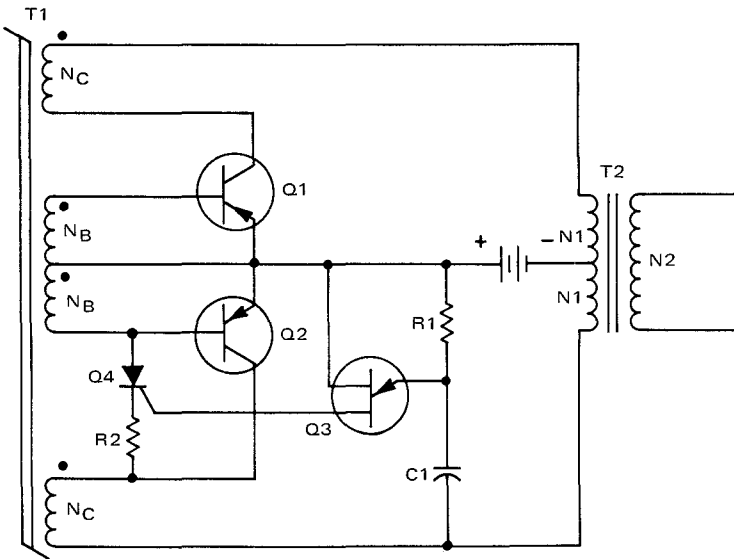
Figure 2-11 — Current-Feedback Inverter with Starting Circuit

In the circuit of Figure 2-11, inverter starting is accomplished by turning on SCR Q4, causing current to flow through the emitter-base diode of Q2, and thus turning Q2 on. When Q2 turns on, the collector-base junction of Q2 becomes forward biased and Q4 is commutated off. Q4 is triggered by the discharge of C1 through unijunction transistor Q3 when the voltage on C1 becomes adequate to fire the unijunction. Resistor R2 limits both gate and anode current of Q4. If the time constant of R1 and C1 is long compared to the inverter half period, the starting circuit will not fire after the inverter starts, and very little loss of efficiency will result from use of the starting circuit.

Simplifications in the starting circuit of Figure 2-11 may be adequate for many applications. For example, if Q4 were eliminated, and base one of the unijunction were connected to the collector of Q2, discharge of C1 would be through the Q2 feedback winding N_C. Q2 would be biased on by current-feedback action of transformer T1. For NPN circuits, unijunction base one may be connected directly to the base of an inverter transistor, and discharge of C1 would provide starting base current to the transistor.

Note that because high output current results in proportionately high base current, current-feedback inverters start well into a full load, or even a somewhat-capacitive load. On the other hand, starting into inductive loads is difficult. This is the converse of voltage-feedback inverters, which start well into light or inductive loads, but with difficulty into full or capacitive load. Low input voltage does not detract from starting performance of current-feedback inverters.

For applications involving varying loads or a range of input supply voltages, current-feedback inverters are more efficient than voltage-feedback inverter because of the proportionality between base current and output current in current-feedback inverters. Conventional voltage-feedback inverters must be designed to provide adequate base drive at full load and minimum input voltage. As loading is reduced, voltage feedback and its associated losses remain constant in magnitude and represent an increasing share of input power. As the input voltage is increased from its minimum value, drive losses increase, again detracting from efficiency. The combination of light load and high input voltage may result in very inefficient operation. A further inefficiency of voltage-feedback inverters is that base resistors or other dissipative means must usually be provided to control base current.

2.6 Voltage Feedback Transistor Inverters

One-Transformer-Inverter Operation

Operational theory common to most transistor inverters may be illustrated by considering the basic over-driven, push-pull, transformer-coupled transistor oscillator circuit of Figure 2-12A, and the transformer B-H curve of Figure 2-13. Assume that transistor Q1 is nonconducting, Q2 conducting, and the transformer saturated at point J on the B-H curve. When Q1 starts to conduct, the voltage developed across the primary windings (N1) induces voltage in the feedback windings (N3) which drives Q1 into saturation rapidly and turns Q2 off. When this transition is completed, constant voltage, $V_P = V_{CC} - V_{CE(sat)}$, is applied to N1. Since
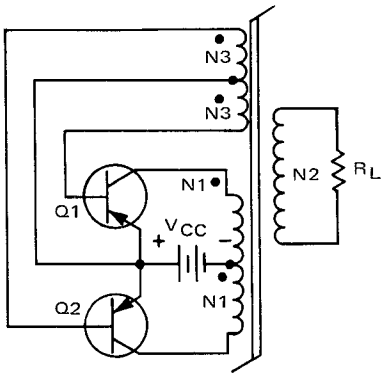
(a) COMMON-EMITTER CIRCUIT
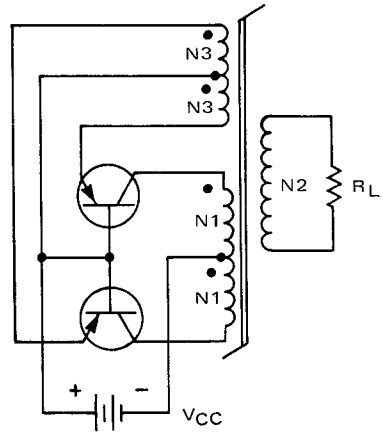
(b) COMMON-BASE CIRCUIT
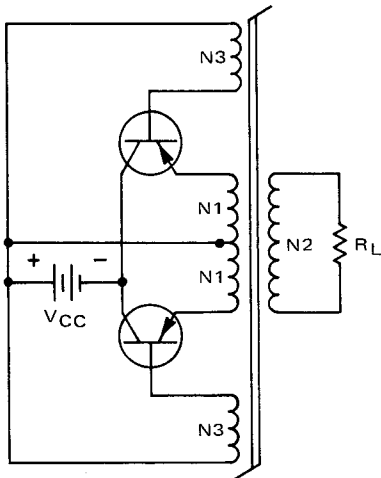
(c) COMMON-COLLECTOR
CIRCUIT (N3>N1)
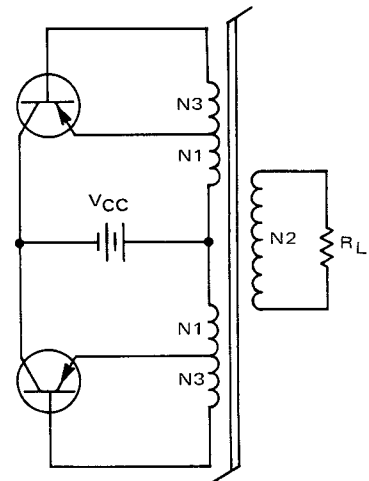
(d) AUTOTRANSFORMER
COMMON-COLLECTOR
CIRCUIT

Figure 2-12 — Basic One-Transformer Inverter Circuits

FLUX DENSITY β (Gauss)

L   K

B_r

MAGNETIC FLUX H (oersteds)

M   J
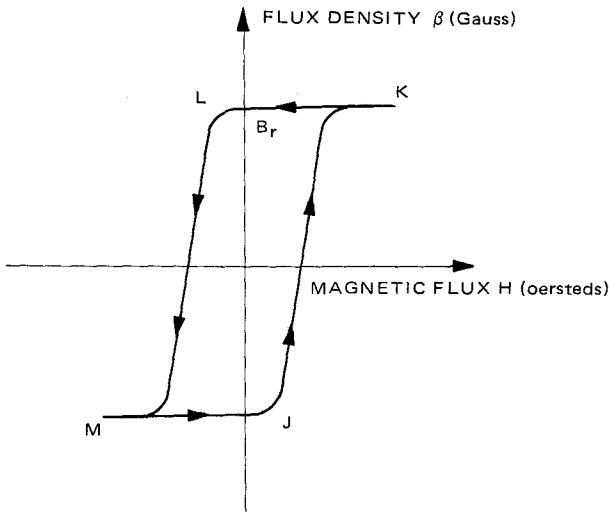
Figure 2-13 — Transformer B-H Curve

$\dfrac{d\phi}{dt} = \dfrac{V_P}{N1}$, flux ϕ must increase in the transformer core at a constant rate, causing flux density $B = \phi/A$ to increase from point J toward point K on the B-H curve. As long as the core remains nonsaturated, magnetization current i_m ($= H/N1$) is small, but as saturation (point K) is approached, high magnetization current i_m is required to keep $\dfrac{d\phi}{dt}$ constant. When reflected load current, plus this sharply increasing magnetization current, exceeds the collector current which Q1 can supply (with the drive available), Q1 begins to come out of saturation causing V_P to decrease. The feedback voltage (V_{FB}) and collector current (I_C) decrease regeneratively, turning Q1 off and ending the half cycle.

As flux in the transformer core collapses from point K to point B_r, voltage which biases transistor Q2 into conduction and initiates the next half cycle is induced in the winding. The operation is similar to the first half cycle except that supply voltage (less $V_{CE(sat)}$) is applied to the other half of the primary, causing a reversal of polarity in the induced output voltage. Q2 conducts until the core is driven into negative saturation at point M on the B-H curve. As flux collapses from M to J, the full cycle is completed.

Typical voltage and current waveforms for a one-transformer inverter are shown in Figure 2-14. It can be seen from the waveforms of
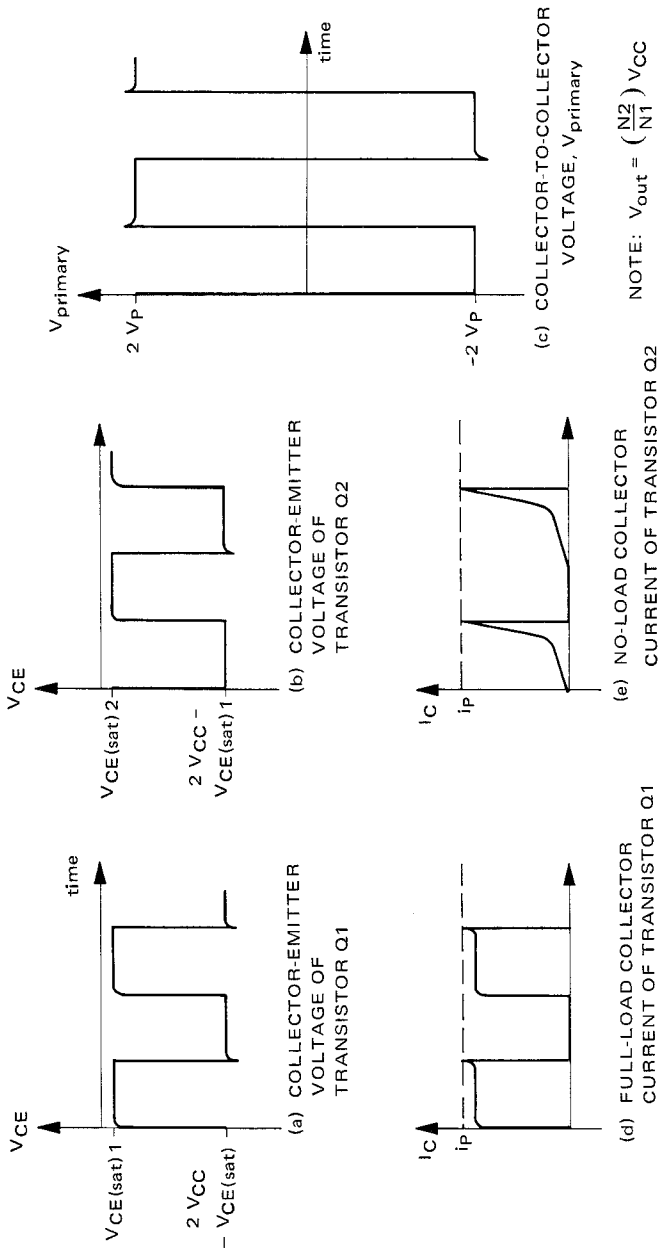
**Figure 2-14 — Typical Voltage and Current Waveforms
for One-Transistor Inverter**

collector-to-emitter voltage that, in the off condition, each device is subjected to a voltage approximately equal to twice the supply voltage plus any induced voltage that may occur in the circuit due to leakage inductance, etc. Also significant is the fact that the same maximum collector current i_p is required for switching action whether this current is primarily reflected load current, as in Figure 2-14D, or totally magnetization current, as in 2-14E. This will obviously limit efficiency at low output loads.

Operating frequency of the inverter is determined by the voltage V_p and by the saturation characteristics of the transformer core according to the relationship, $f = \dfrac{V_P \times 10^8}{4\beta_s A N_1}$ Hz. β_s is saturated flux density in gauss, A is cross-sectional area of the core in cm^2, and N_1 is the number of turns on one half of the primary.

Two-Transformer Inverters

At high load currents, high frequency, and high output power, it becomes increasingly difficult to design and build a one-transformer inverter in which the transformer fulfills the dual role of frequency control and efficient transformation of output voltage satisfactorily. For this reason, the two-transformer inverter designs of Figure 2-15 are advantageous in many applications. Operation of two-transformer inverters is similar to that of one-transformer inverters except that in a two-transformer circuit (Figure 2-15), only the small feedback transformer T2 need be saturated. Since the magnetization current of T2 is small, high current levels due to transformer saturation currents are reduced significantly compared to one-transformer design, as is device stress due to these transformer saturation current levels. (Compare Figure 2-16 with Figures 2-14D and 2-14E.) Furthermore, use of a conventional output transformer with normal core material permits lower transformer costs as well as higher efficiency.

Another major advantage of two-transformer inverter designs (circuits similar to Figure 2-15) is that the inverter frequency is determined by V_{FB}. This voltage can be regulated to provide a constant frequency or can be changed with a variable resistor for R_{FB} to provide variable frequency output.

Bridge Inverters and Series-Connected Inverters

Inverters such as the bridge circuits of Figure 2-17, and the series-connected inverter of Figure 2-19, are useful when input voltage exceeds transistor voltage capabilities. The bridge arrangements apply the supply voltage across each transistor, instead of applying approximately twice the
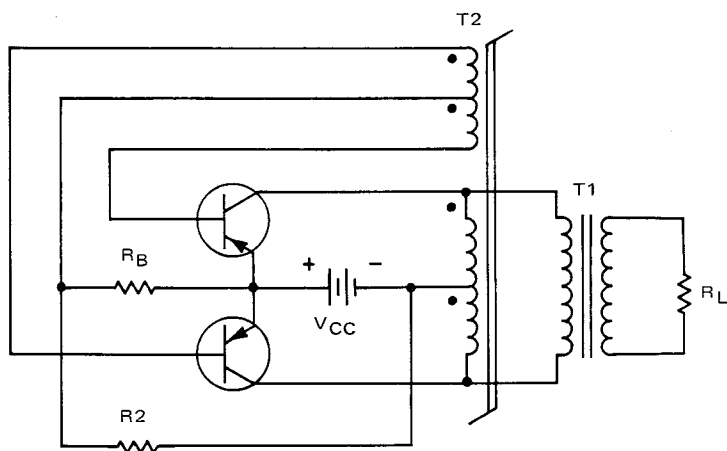
(a) TWO-TRANSFORMER INVERTER WITH
SIMPLE OUTPUT TRANSFORMER T1
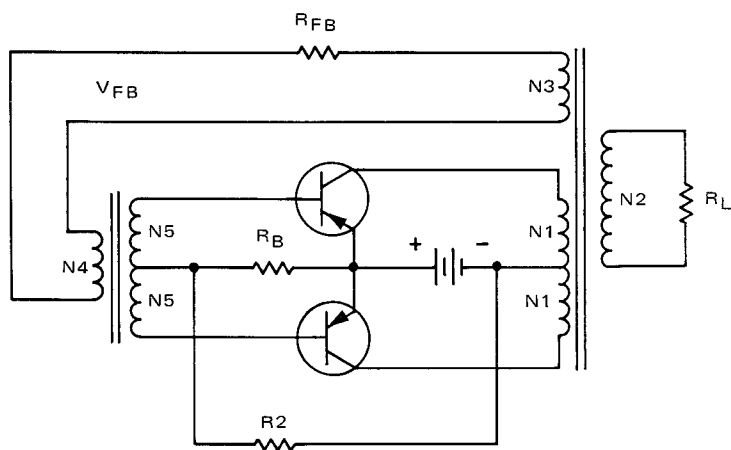
(b) TWO-TRANSFORMER INVERTER HAVING
EASILY REGULATED V_{FB}

Figure 2-15 — Typical Two-Transformer Inverters

(a) FULL LOAD
COLLECTOR CURRENT

(b) NO LOAD COLLECTOR CURRENT
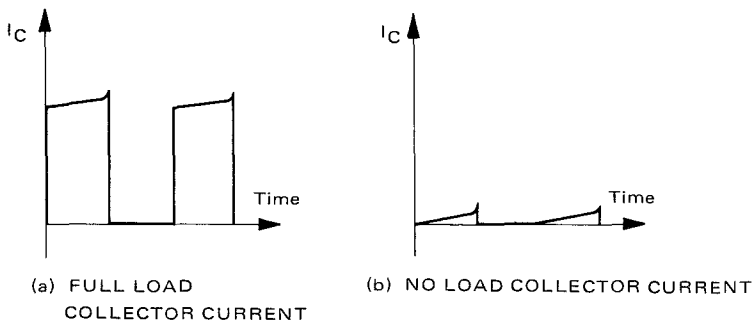
Figure 2-16 — Collector-Current Waveforms for Two-Transformer Inverter

supply voltage as in simple push-pull circuits. This reduction in voltage to which each device is subjected is true of all circuits of Figure 2-17, but in the half-bridge circuit of Figure 2-17C, transistor current must double to maintain the original output power.

In considering bridge inverters, the designer should be aware of a problem tolerable in simple inverters but of major significance in bridge connections. This has to do with the current-voltage excursions of the devices as the circuit switches. If the previously nonconducting side of the circuit turns on before the other side is completely off, both high voltage and high current may be imposed on the transistors, and their safe areas may be exceeded. Additionally, high transients may be generated. The problem may be somewhat alleviated by reducing device "on" drive, by device protection against transients, or by compensating base-drive networks which retard turn on of the nonconducting device. One possibility is to use a driven bridge having the input waveforms of Figure 2-18.

In the circuit of Figure 2-19, n simple inverters connected in series divide the supply voltage equally so that each device is required to withstand $\dfrac{2\,V_{CC}}{n}$. The magnetic circuit requires that the voltage divide equally among the series stages.

Each of the basic approaches summarized above has a unique combination of advantages and disadvantages which should be considered in light of the design requirements.

Inverter Starting Circuits

In general, the basic circuits of Figure 2-12 and their derivatives (including basic two-transformer inverters) will not oscillate readily unless some means is provided to begin oscillation. This is especially true at full load and low temperature, the most severe starting condition for resistive
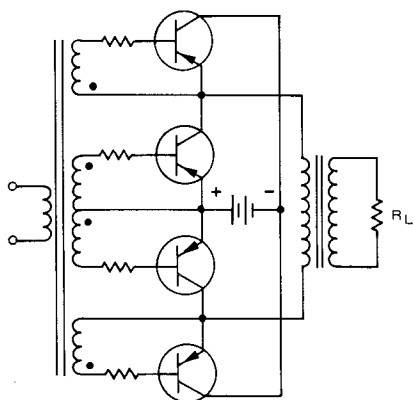
(a) DRIVEN BRIDGE-CONNECTED
OUTPUT STAGE
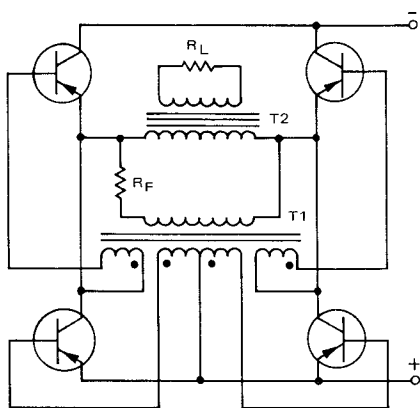
(b) SELF-OSCILLATING TWO-TRANSFORMER
FULL-BRIDGE INVERTER
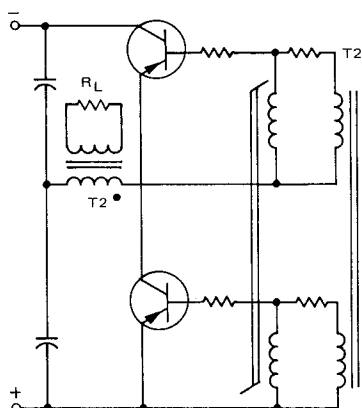
(c) HALF-BRIDGE INVERTER WITH
SATURABLE-BASE INDUCTORS

Figure 2-17 — Typical Bridge Inverter Circuits
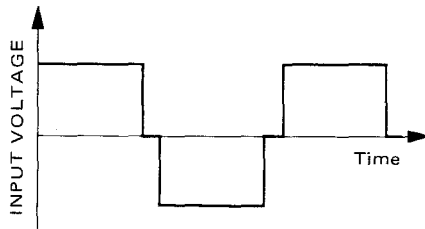
**Figure 2-18 — Typical Magnetic-Amplifier-Output Waveform Used
as Bridge-Inverter Input Prevents Simultaneous Conduction
of Both Inverter-Circuit Halves During Conduction**
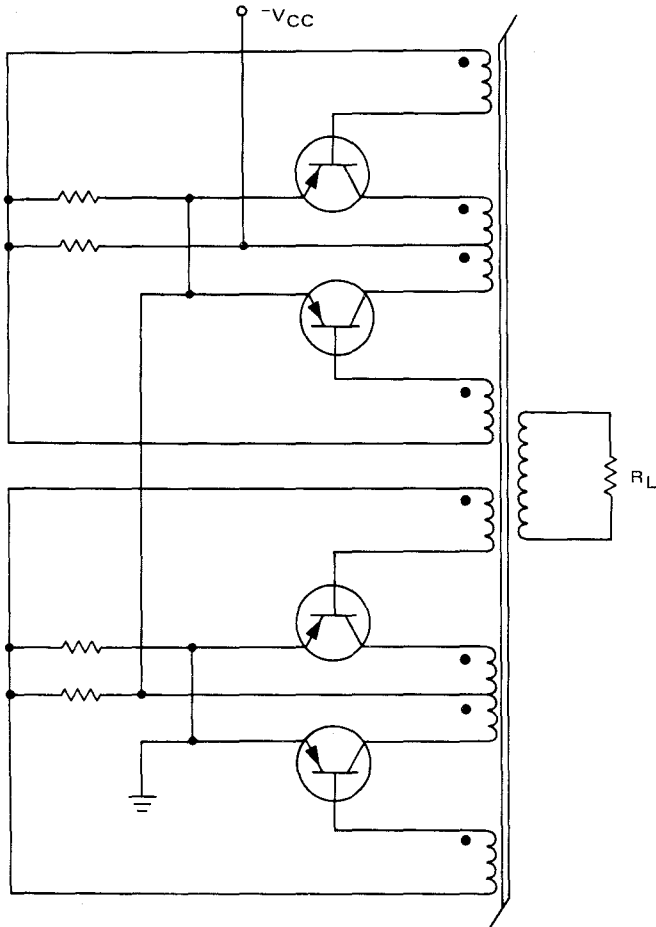
Figure 2-19 — Series-Connected Inverter for High DC Input Voltages

loads. The discussion of basic inverter operation assumed that one of the transistors was conducting, which will start the oscillation. The function of the starting circuit is to ensure this condition.

A simple, commonly used starting circuit is shown schematically in Figure 2-20A. In this circuit R1 and R2 form a simple voltage divider to bias the transistors to conduction before oscillation starts. A good rule of thumb for the base starting bias developed by this circuit is to use 0.3 volts for germanium transistors and 0.5 volts for silicon. This voltage, V_B, is equal to $\dfrac{R1\ V_{CC}}{R1 + R2}$. Since R1 occurs in the feedback circuit in series with the base of each circuit half, R1 must not exceed R_B, which is equal to $\dfrac{V_{FB} - V_{EB}}{I_B}$. If R1 is set equal to R_B, then R1 and R2 are uniquely determined for any given starting bias. The resistance of R2 is uniquely determined for any given starting bias. The value of R2 may be adjusted if starting is not satisfactory. The advantage of this straightforward starting technique is that only resistors need to be added to the circuit, but it has the disadvantage of additional power dissipation, which may become excessive in high power circuits.

An improved, but somewhat more costly, starting circuit is the diode self-starting circuit shown in Figure 2-20B. This circuit dissipates less power than its resistive counterpart and is less temperature dependent. Operation is similar to resistive starting, but when power is first applied the bases of the transistors are driven negative by full supply voltage and oscillation starts rapidly.

Capacitive filters, starting motors or incandescent lamps may temporarily present extremely high loads to inverters during the starting period. Starting with such loads is often simplified somewhat by using a driven inverter, and this approach may be preferable to the circuit complications needed to assure self-oscillation.

Conclusion

Transistor voltage-feedback inverters provide many advantages in conversion of direct current to alternating current. The basic two-transistor, one-transformer design has excellent efficiency and performance, but is inferior to a two-transistor inverter at power outputs in excess of 100 watts or if precise frequency is required. Design requirements are dictated by inverter specifications, and modifications to the basic circuit probably will be necessary to achieve precisely controlled frequency or output voltage.
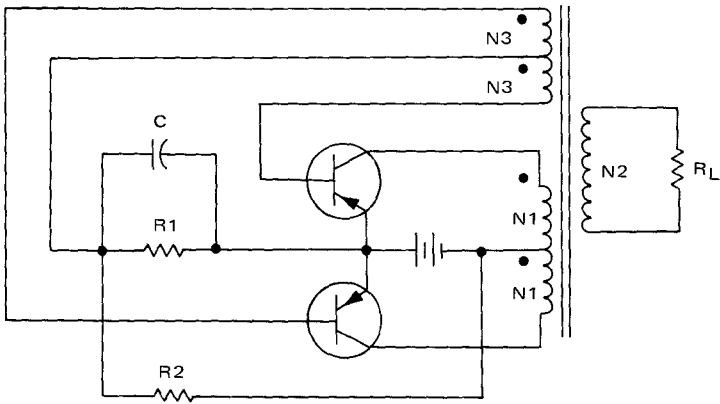
(a) SIMPLE RESISTIVE SELF–STARTING CIRCUIT WITH
SPEED-UP CAPACITOR ACROSS R1
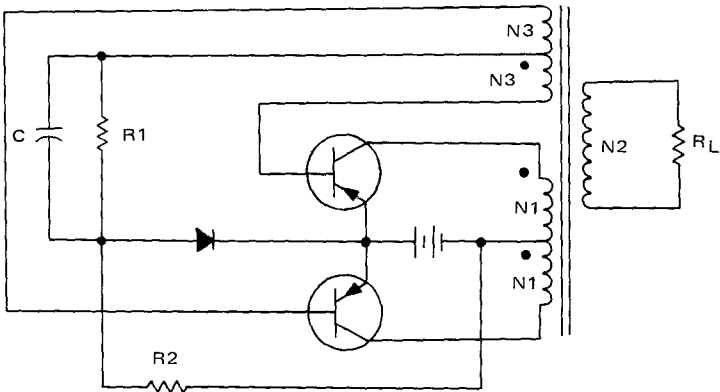
(b) DIODE SELF-STARTING CIRCUIT

Figure 2-20 — Starting Circuits for Transistor Inverters

2.7 Hybrid Feedback Inverters

It has been noted that current-feedback inverters have some advantages over voltage-feedback inverters. The frequency of a current-feedback inverter is almost independent of supply voltage and load, and its efficiency is nearly constant over a wide range of load because its base current is proportional to load current. This constant forced gain also enhances the overload capability of current-feedback inverters. Current-feedback inverters operate well with low input voltage. They start well into full loads, but are difficult to start into light or inductive loads, and shut down with open output. They can be operated into center-tapped resistive loads as well as into transformers.

Voltage-feedback inverters start well into light or inductive loads, but are difficult to start into full or capacitive loads, and shut down automatically if their output is shorted. The frequency of voltage-feedback inverters is determined by feedback voltage and therefore is usually proportional to input voltage. Because base drive is constant with respect to load, voltage-feedback inverters are comparatively inefficient at low load. Base drive increases with increasing input voltage, and decreases efficiency if input voltage ranges widely.

By combining current feedback and voltage feedback in a single inverter, it is possible to obtain many of the advantages of each type of feedback. For example, the hybrid inverter of Figure 2-21A has voltage feedback added to the basic current-feedback inverter. Although the polarity of N4 could be arranged to obtain differential feedback, additive feedback is assumed in the following discussion. The relative contributions of N4 and N_C to total drive may be varied, depending upon the characteristics desired.

The combination or hybrid feedback results in excellent starting over a full range of loads, including reactive loads. Frequency stability with changing input voltage and changing load is essentially that of the current-feedback inverter. As previously discussed for current-feedback inverters, a resistor, rectifier, zener diode or other series element may be placed in the base circuits for additional control of inverter frequency. Low-load efficiency and overload capability are improved over the voltage-feedback inverter since base current is still somewhat proportional to load current. Hybrid inverters operate well over an extremely wide range of input voltage. Either common-emitter or common-collector NPN or PNP transistors can be used in hybrid-feedback configurations.

The modification of hybrid feedback shown in Figure 2-21B, was found to be useful with inductive and light loads. Because of capacitor C_{FB} in the voltage feedback loop, the voltage feedback is maintained only
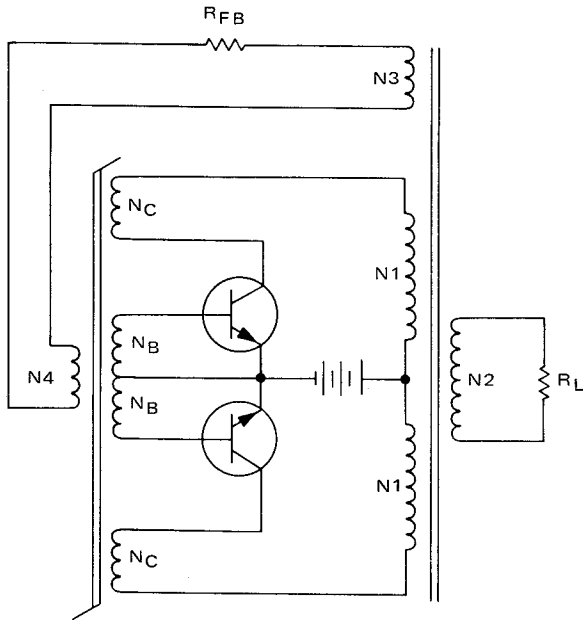
(a) CURRENT FEEDBACK + VOLTAGE FEEDBACK
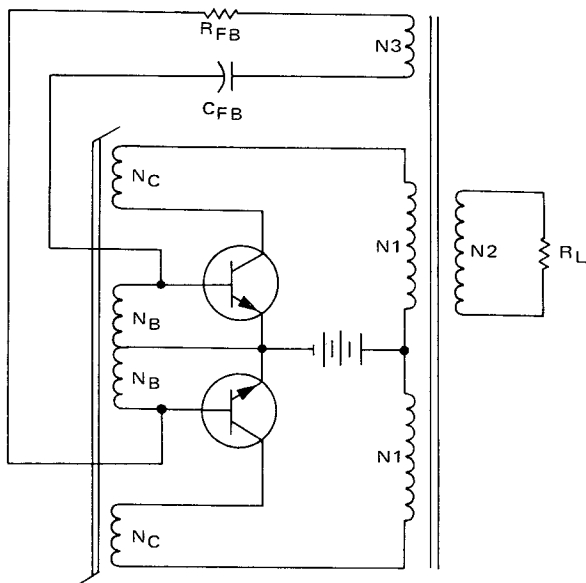
(b) CURRENT FEEDBACK + "BOOSTER" FEEDBACK

Figure 21 — Hybrid-Feedback Two-Transformer Inverter

during the first portion of each half cycle. Each time the inverter switches, a pulse of current from the voltage feedback "booster" circuit consisting of R_{FB}, C_{FB} and N3 is injected at the base of the transistor being turned on. This supplemental current adds to the base current from the current-feedback transformer to assist transistor switching. If the load is inductive, it also serves to maintain bias until collector current is adequate to provide the feedback needed to sustain conduction. At low loads (low current in N_C), the booster circuit is adequate to provide oscillation. At full load, regular current-feedback operation predominates and the booster circuit has little effect on the inverter operation.

2.8 SCR Inverter with Well-Regulated, Sine-Wave Output 165 Vdc - 120 Vac, 800 W, 400 Hz

The 800 watt, 400 hertz series-commutated SCR inverter shown in Figure 2-22 is designed to operate with an input of 165 volts. The output voltage is 125 volts rms and the waveform is a sine wave. This circuit is probably the best sine-wave inverter known today. The output voltage is extremely well regulated for variations in load, the frequency is independent of normal load and input voltage variations, and the circuit itself can be designed to operate over a wide range of frequencies and output power levels.

The 800 watt limit imposed on this circuit is actually a device limitation. The circuit was tested at 1000 watts for efficiency, and the load was then reduced to 3 ohms before the thyristors failed to commutate. The 3 ohm load represents almost 5 kilowatts of output power. The circuit will also operate at the other extreme of no load. The actual operating range of output power is 0 to 800 watts with an input of 165 volts dc.

Efficiency and output voltage are plotted as a function of output power in Figure 2-23. Efficiency increases with power from 5% at 30 watts to 60% at 800 watts. This shows that the best efficiency is obtained with full load. The output voltage decreases slightly as output power is increased. The voltage dropped from 127.5 V at no load to 125.0 V at full load. This means that output voltage will decrease by only 0.3 volts for every 100 watt increase in load. Frequency was not plotted because it does not vary with load. The frequency remains 400 hertz over the full range of output power. The waveform itself is also quite good over this range; there is no perceptible distortion of the sine wave as viewed on an oscilloscope.

At full load, the input current is about 9 A. If the trigger circuit were externally driven, the inverter circuit would operate with an input voltage of 25 volts dc. The zener will start to lose regulation in this circuit
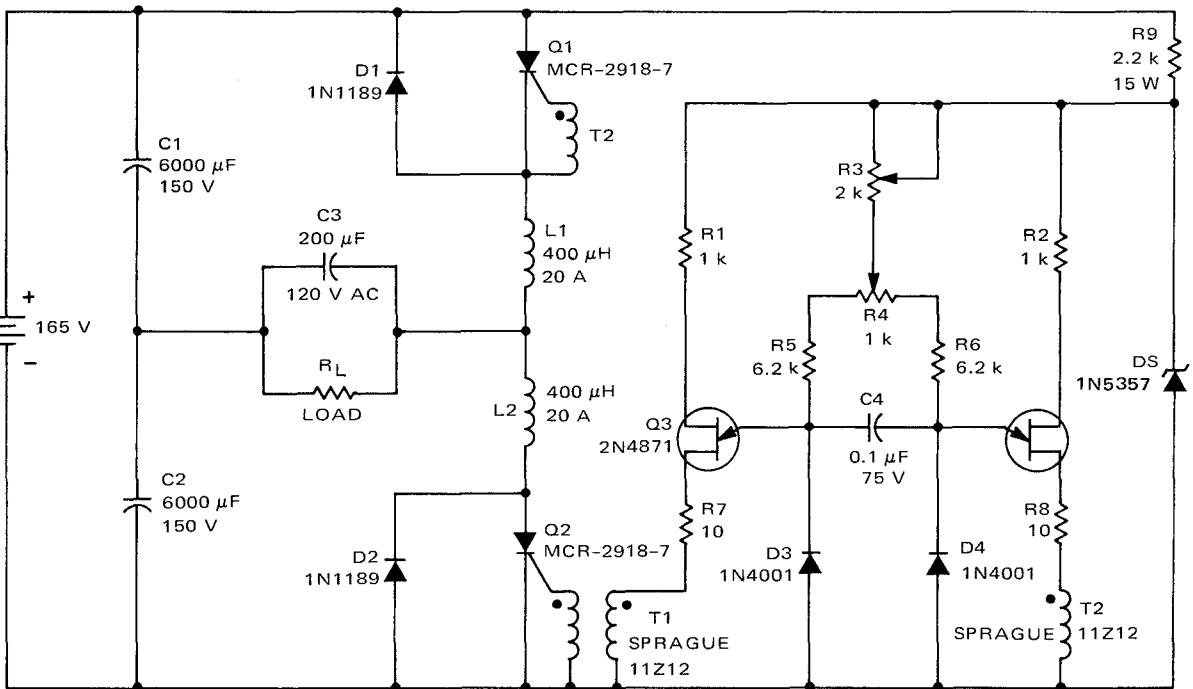
Figure 2-22 — Sine-Wave SCR Inverter with Well-Regulated Output. 165 Vdc-120 Vac, 800 W, 400 Hz

Inverters and Converters
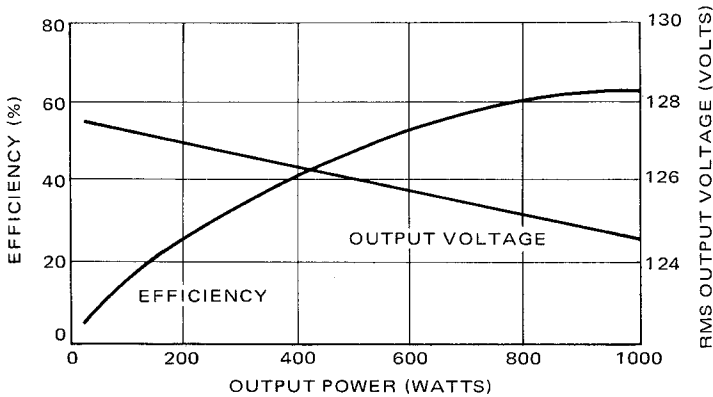
Figure 2-23 — Performance Curves for Series-Commutated SCR Inverter in Figure 2-22

at about 75 volts input. Between 75 and 165 volts input, the output frequency remains 400 hertz. Output power and output voltage, however, will vary with the input. The peak output voltage remains approximately equal to the input voltage. This means that the rms output is about 0.7 times the input voltage.

This circuit does differ from the basic series inverter: Diodes D1 and D2 are connected across the thyristors Q1 and Q2. Inductors L1 and L2 do not have a common core and the resonant frequency is higher than the operating frequency. The circuit is discussed in more detail in Reference 6.

2.9 Basic Sine-Wave Inverter
28 Vdc - 110 Vac, 40 W, 400 Hz

The 40 watt, 400 hertz series-commutated SCR sine-wave inverter shown in Figure 2-24 is designed for an input of 28 volts dc. The output voltage is 110 volts rms. The circuit is quite simple; it uses a capacitor and inductor on the output to produce the sine-wave output. It may be operated above or below resonance without any commutation problems. The circuit is driven by a pair of unijunction transistors, and the frequency regulation is excellent for variations in both input voltage and load. The circuit can be operated indefinitely into either an open- or short-circuited load. The output voltage and power are affected by variations in circuit Q and will change with variations in load resistance.

Figure 2-25 shows the effect of variations in load resistance on frequency, efficiency, power, and output voltage. The load resistance was

increased from 50 to 500 ohms with a 28 volt dc input. Frequency increases slightly with resistance but the overall regulation is better than 5%. The output voltage varies considerably with load; it increases from about 30 to 130 volts rms. If it is desirable to keep output voltage within a certain range, the load resistance must be limited accordingly. The maximum transfer of power occurs with a load resistance of 290 ohms and the output power at this point is approximately 42 watts. Variations in resistance above and below this power cause the power to decrease. Efficiency increases with increasing load resistance from about 20% to 60%. At 250 ohms, the efficiency is slightly better than 40%.

In Figure 2-26, output voltage and frequency are plotted as a function of input voltage, with a constant load of 250 ohms. As the input voltage is increased from 15 to 30 volts, the output voltage increases linearly from about 50 to 110 volts. Doubling the input voltage also causes the output voltage to double. The output frequency decreases from 415 to 395 hertz over this same range of input voltage. Frequency regulation is again better than 5%.

The nominal input voltage is 28 volts dc. Variations of $\pm 10\%$ are the allowable limit. Input current will vary with load from 2 A at 500 ohms to 6 A with a short circuit (0 Ω load resistance). The operating range of load
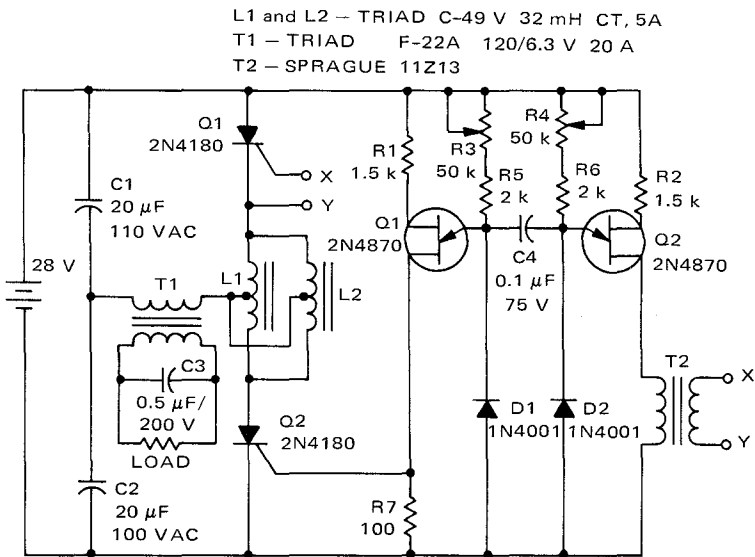
Figure 2-24 — Basic Sine-Wave SCR Inverter.
28 Vdc to 110 Vac, 40 W, 400 Hz
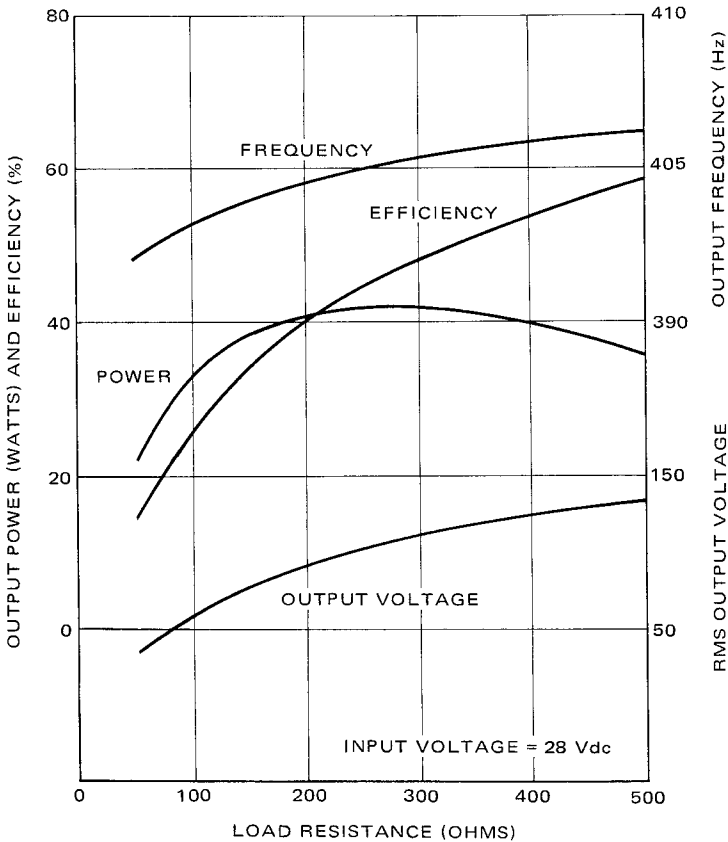
**Figure 2-25 — Performance of SCR Sine-Wave Inverter
in Figure 2-24 for Varying Load**

resistance should be from 200 to 300 ohms in order to obtain ±10% output voltage regulation.

2.10 Driven Inverter with Stable Frequency Output
24 Vdc - 45 Vac, 120 W, 400 Hz

A simple means to approximately double a voltage is to use the parallel-commutated SCR inverter shown in Figure 2-27. The nominal input voltage for this circuit is 24 Vdc, but input voltage may be varied from 10 to 28 Vdc. Output power is typically 120 watts but can be increased to approximately 150 watts with 28 Vdc input. The output voltage is a 45 V, 400 Hz square wave. This voltage is independent of load
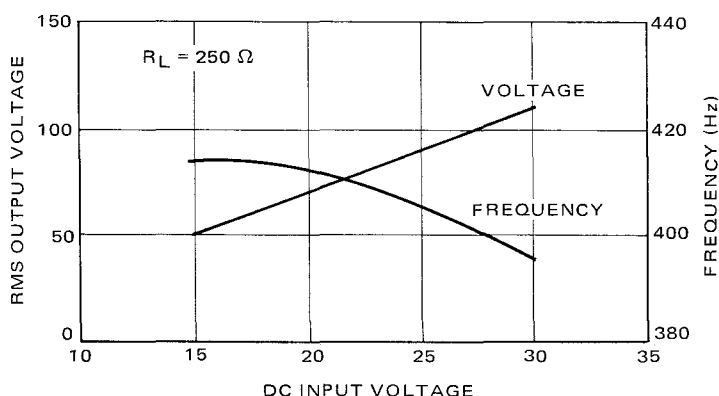
**Figure 2-26 — Performance of SCR Sine-Wave Inverter
in Figure 2-24 for Varying Input**

current but does vary with the input voltage. A capacitive load is required to commutate the thyristors properly. Load power factor in this circuit is about 0.9.

Load current does affect the output voltage waveform of this circuit. The leading and lagging edges rise and fall exponentially. As load current decreases, the rise and fall times increase. A certain amount of rise and fall time is necessary for commutation. The minimum load for this circuit is ten ohms, resulting in a maximum load current of 2.4 A.

The magnitude of the input voltage does not affect the output-voltage rise and fall times but will, as stated before, affect output voltage magnitude.

Since this is a driven inverter, the output frequency is not affected by changes in the load. Frequency is also very stable for variations in input voltage. Frequency stability is maintained by the voltage drops of diode D1 and the base-emitter junctions of Q1 or Q2, which are relatively independent of current variations.

The overall efficiency of this circuit is approximately 70%, resulting in power losses low enough that this inverter can be used in ambient temperatures from $0°C$ to $+60°C$ if heat sinks with thermal resistance of less than $3°C$ per watt are used for Q1, Q2, Q5, and Q6.

The circuit is actually an inverter-driven inverter. The drive circuit is a current-feedback inverter with a UJT starting network. Additional windings on the small timing transformer drive a pair of sensitive-gate SCRs. These SCRs drive the main parallel-commutated inverter, which supplies power to the load.
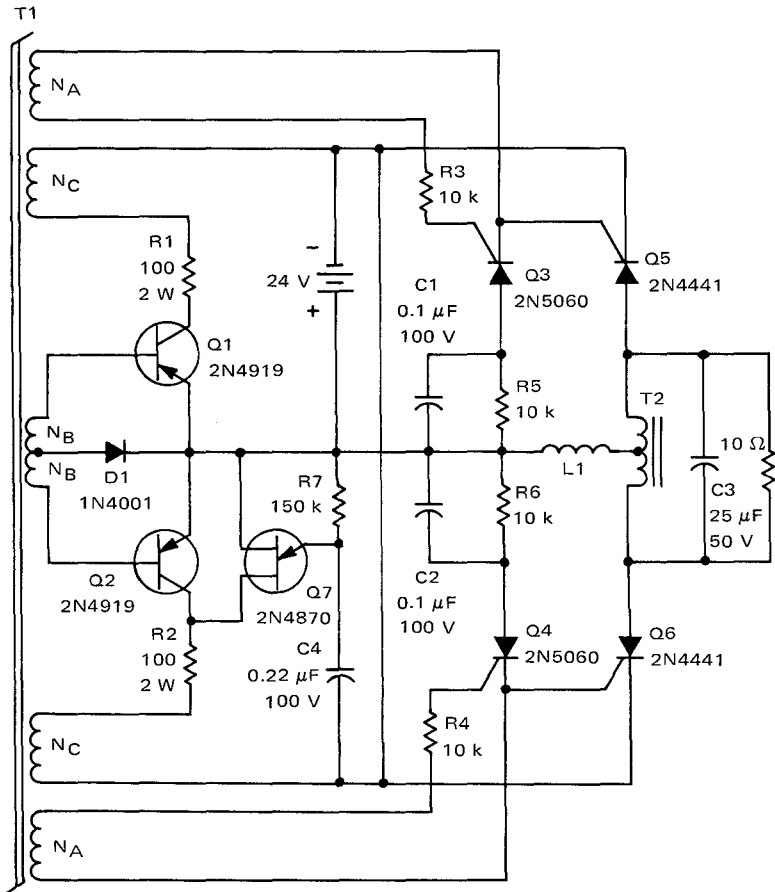
TRANSFORMERS
T1 — ARNOLD CORE 6T-5340-D1
        $N_A + N_B$ 83 TURNS EACH #30 WIRE
        N_C 8 TURNS #26 WIRE
T2 — PHOENIX TRANSFORMER PX2677
        350 V, 1 kVA, 1 kHz
    INDUCTOR

INDUCTOR
L1 — PHOENIX TRANSFORMER PX1710 2 mH, 30 A

Figure 2-27 — Driven SCR Inverter.
24 Vdc-45 Vac, 120 W, 400 Hz

The value of inductor L1 was chosen to provide a damped output waveform with no overshoot. If the waveform is not critical, this value may be reduced to as low as 80 μH. This not only reduces the weight of the reactive components but will improve efficiency as well. This circuit is subject to damage from short circuits or overloads. In either case, both thyristors Q5 and Q6 will fail to commutate, permitting sufficient current to be drawn from the line to destroy them unless suitable protection is provided.

2.11 Sine-Wave Inverter with Stable Output Frequency 48 Vdc - 120 Vac, 100 W, 60 Hz

A 100 watt, 60 hertz parallel-commutated SCR inverter requiring a 48 volt dc input is shown in Figure 2-28. It was designed by the Forest Electric Company and is commercially available as the Felco Model W-1694. The output voltage is a 120 volt rms sine wave. The output frequency is extremely stable for variations in load and input voltage, since the drive circuit is a frequency-controlled inverter. The drive circuit and the main inverter both use the same input voltage, but the drive-circuit input is regulated by the combination of resistor R5 and zener diode D5. The 62 V zener diodes, D3 and D4, have been placed across the input to suppress voltage transients.

The performance curves shown in Figure 2-29 were taken with a fixed load of 92 watts (with 48 volts dc input). Variations in efficiency, frequency, and output voltage are plotted as a function of input voltage. The frequency decreases by 0.6 hertz when the input voltage is increased by 10 volts. In other words, a 10% change in input voltage causes the frequency to vary by only 0.5%. The efficiency remains at about 60% over this range of input voltage. However, the output voltage does increase in almost direct porportion to the input. That is, a ±10% change in input voltage will cause the output voltage to vary by approximately ±10%.

Additional performance information is shown in Figure 2-30 for a smaller load (79 watts at 48 V input). The frequency and output voltage regulation are identical to Figure 2-29 but the efficiency shows a noticeable decrease as the input voltage is increased. It drops from 59% at 43 volts to 56% at 53 volts.

An interesting comparison can be made between Figures 2-29 and 2-30. The frequency at 92 watts (Figure 2-29) is identical to the frequency at 79 watts (Figure 2-30), with 48 V input. The frequency at these points is 60.3 hertz and is not dependent on the load. The output voltage, however, does increase from 120 to 123 volts as the load is decreased. This
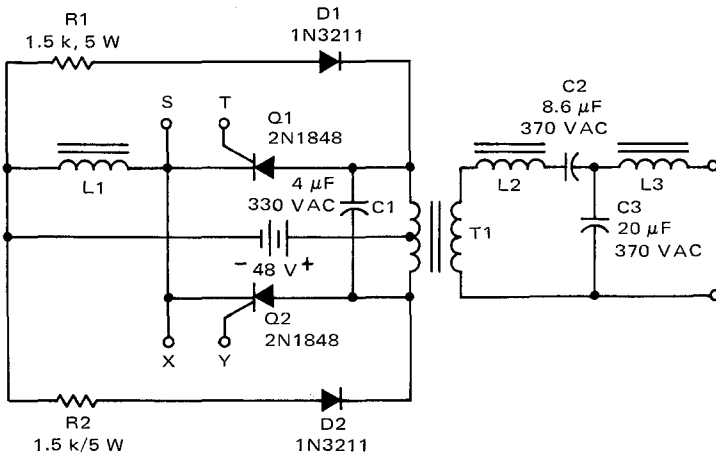
INDUCTORS

L1 — FOREST ELECTRIC W-1708
L2 — FOREST ELECTRIC W-1706
L3 — FOREST ELECTRIC W-1707

TRANSFORMERS

T1,T2 — FOREST ELECTRIC W-1705
T2 — FOREST ELECTRIC W-1704

Figure 2-28A — SCR Inverter with Stable Output Voltage and Frequency. 28 Vdc – 120 Vac, 100W, 60 Hz. Drive Circuit in Figure 2-28B. Felco Model W-1694 Inverter
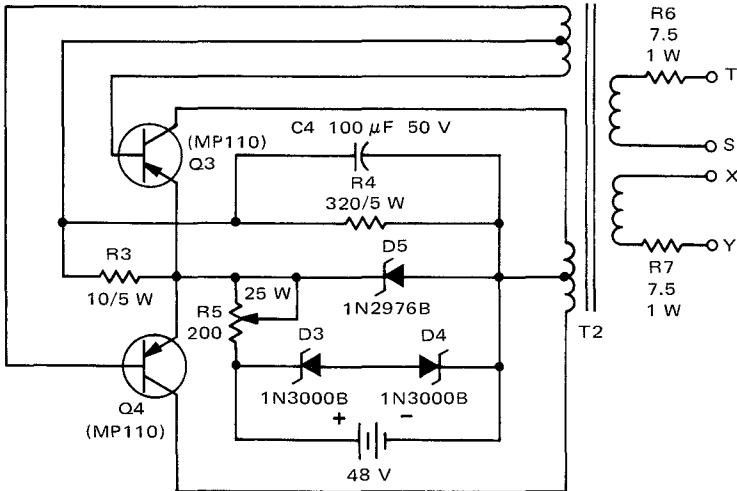
INDUCTORS

L1 — FOREST ELECTRIC W-1708
L2 — FOREST ELECTRIC W-1706
L3 — FOREST ELECTRIC W-1707

TRANSFORMERS

T1, T2— FOREST ELECTRIC W-1705
T2 — FOREST ELECTRIC W-1704

Figure 2-28B — Inverter Drive Circuit for Inverter in Figure 2-28A

means that a 5 watt increase in load would cause the output voltage to decrease by nearly 1 volt. Efficiency also varies with load. At 79 watts the efficiency is 57.5%, but this increases to 60.3% at 92 watts.

The output power may be increased to 150 VA with lagging power factor as low at 0.84. The total harmonic distortion will be about 6%. The permissible input voltage variation is ±10%, as shown on the curves. The maximum input current will be approximately 6 A. Best frequency regulation is obtained by setting resistor R5 to its maximum resistance of 200 Ω. The setting can be reduced to about 100 Ω if the frequency changes when the input voltage is decreased to its minimum value.
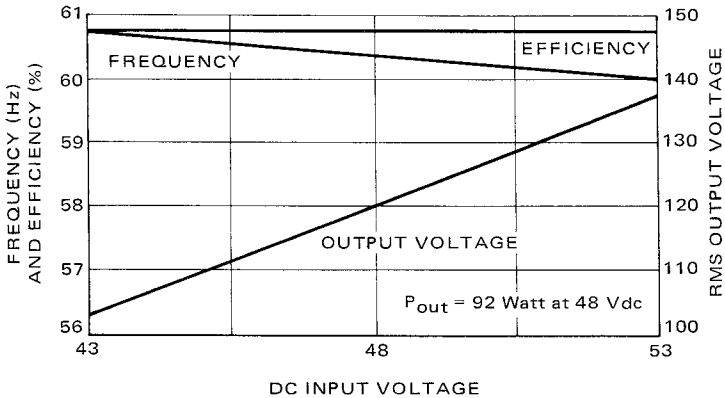
**Figure 2-29 — Performance Curves for Inverter
of Figure 2-28 at P_o = 92 W**
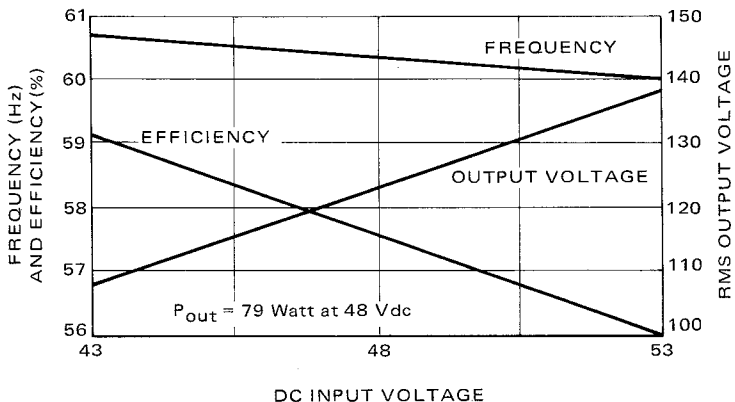
**Figure 2-30 — Performance Curves for Inverter
of Figure 2-28 at P_o = 79 W**

The design of the sine-wave output filter is covered in "A Filter for SCR Commutation and Harmonic Attenuation in High Power Inverters," in *Communications and Electronics*, May, 1963, by Richard R. Ott.

2.12 SCR Inverter with Stable Output Voltage and Frequency 28 Vdc - 12 Vac, 60 W, 465 Hz

The 60 watt, 465 hertz impulse-commutated SCR inverter shown in Figure 2-31 requires 28 Vdc input. The output square-wave voltage and frequency are well regulated for variations in load. This particular output frequency can be used to drive a 400 Hz sine-wave transformer. Efficiency and output power are relatively low compared to SCR inverters operating from higher voltages because of the SCR forward voltage drop. The 2N4441-44 thyristor series will handle from 50 to 600 volts. If the remaining components were rated at 600 volts, a 400 Vdc supply could be used. This would increase the output power capability to nearly 2 kilowatts.

Variations in efficiency, frequency, and output voltage are shown in Figure 2-32. Output voltage decreases as the load increases. The no-load-to-full-load regulation is better than 10%. In other words, the output voltage will decrease by about 0.1 volt for an increase in output power of 6 watts.

Frequency regulation is better than 2%. The frequency tends to increase with load. The increase is approximately 1 hertz for each 10 watt increase in output power. Potentiometers R13 and R14 are used to make adjustments to the output waveform. Resistor R14 determines the balance between positive and negative alternations. The tap is therefore normally set midway for equally timed alternations. Resistor R13 is used to adjust the frequency; it was set for maximum resistance for this test. Frequency can be increased by decreasing the resistance.

The efficiency of this circuit also increases with load. At 2 watts the efficiency is about 10%. This increases to 65% at full load (60 watts). Most of the power lost in this circuit does not depend on load. This includes power in the trigger circuits and in the commutating elements, capacitor C3 and inductor L1. Because of this fact, the best efficiencies can be realized by operating at full load.

The maximum input current is 4 A. Input voltage can be reduced to 20 volts and the circuit will operate with no load. The maximum load is approximately 75 watts. The load resistance at this point is about 1.5 ohms. The circuit will not commutate (switch) with lower values of load resistance.

The commutating inductor, L1, is larger than necessary because a smaller inductor was not available. The design value for L1 is about 20 μH.
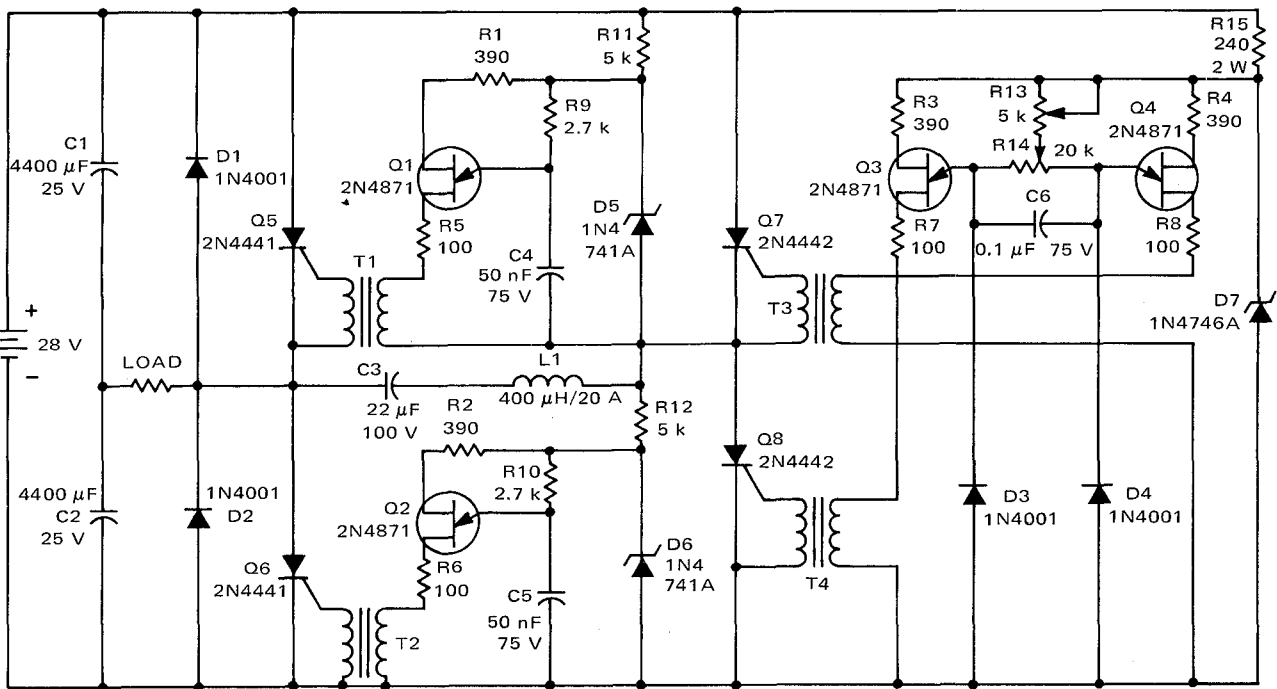
Figure 2-31 — SCR Inverter with Stable Output Voltage and Frequency. 28 Vdc-12 Vac, 60 W, 465 Hz

T1 — T4 — SPRAGUE 11Z12

Lower-resistance loads could be commutated if this value had been available. Output power could have been increased to approximately 200 watts. There would also be a disadvantage in this case. Timing of the trigger circuits involving transistors Q1 and Q2 would be more critical. Instability here could cause either early or late firing. With full load, it is possible that both of these cases would prevent commutation. When the load is not commutated from Q5 to Q6, both thyristors remain on. The result is that they would then draw excessive current and destroy themselves.
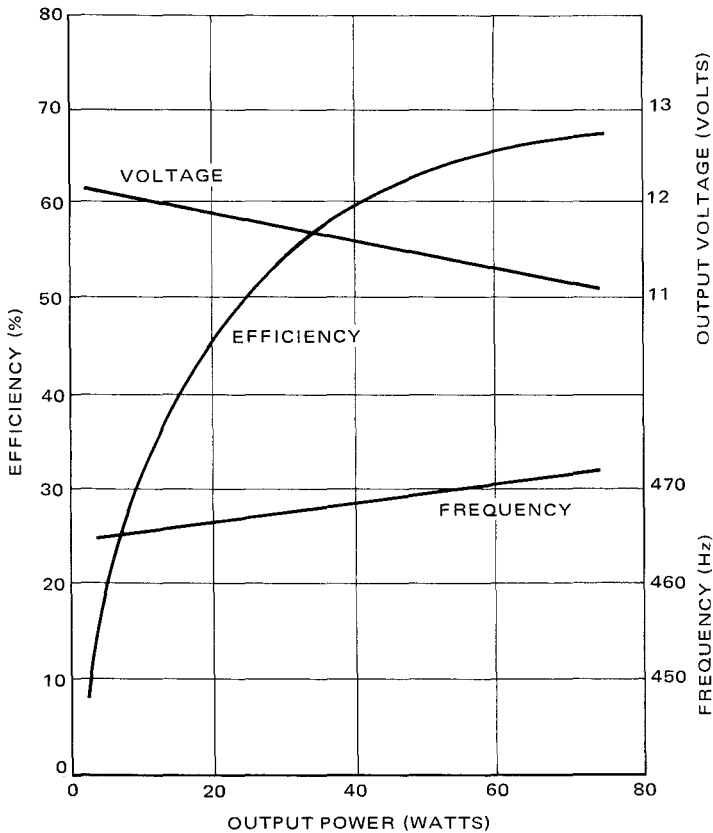
Figure 2-32 — Performance Curves for Inverter of Figure 2-31

2.13 Transistor Inverter with Stable Output Frequency
12 Vdc - 110 Vac, 40 W, 400 Hz

A 40 watt, 400 Hz current-feedback inverter designed for 12 volt input is shown in Figure 2-33. The output voltage is a 110 volt square wave whose frequency is extremely stable and almost unaffected by changes in input voltage and output power. In Figure 2-34 the percent change in frequency is plotted as a function of input voltage for various loads. It can be seen that if the input voltage is varied ±25% at full load (40 watts), the frequency will change by less than ±2%. When the load was reduced to 6 watts, the nominal output frequency dropped to 390 hertz. However, at this reduced load, there was virtually no variation in frequency as the input voltage was varied by ±25%.

Input voltage is nominally 12 Vdc. It is possible to operate the circuit with input voltages as low as 6 Vdc and as high as 20 Vdc. The main function of capacitor C2 is to suppress switching transients caused by leakage inductances. It will also suppress any power-supply spikes at the input. There is no provision for filtering input ripple voltage or for regulating the input voltage. It is therefore desirable to keep input-voltage changes to minimum to obtain a clean, well-regulated output voltage waveform.
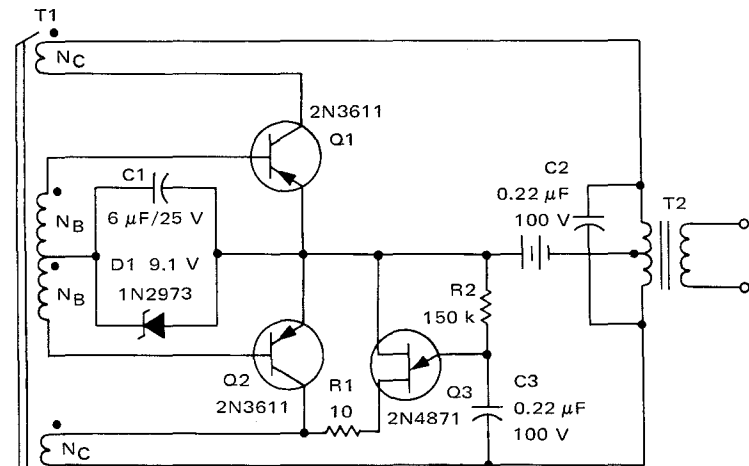
T1 — ARNOLD CORE 4T — 4179 D1
N_B 200 TURNS #20 WIRE
N_C 20 TURNS THREE #16 WIRES IN PARALLEL
T2 — TRANSFORMER 110-24 VCT FELCO VJ-1
(PRODUCT OF FOREST ELECTRIC CO.)

Figure 2-33 — Transistor Inverter with Stable Output Frequency.
12 Vdc-110 Vac, 40 W, 400 Hz
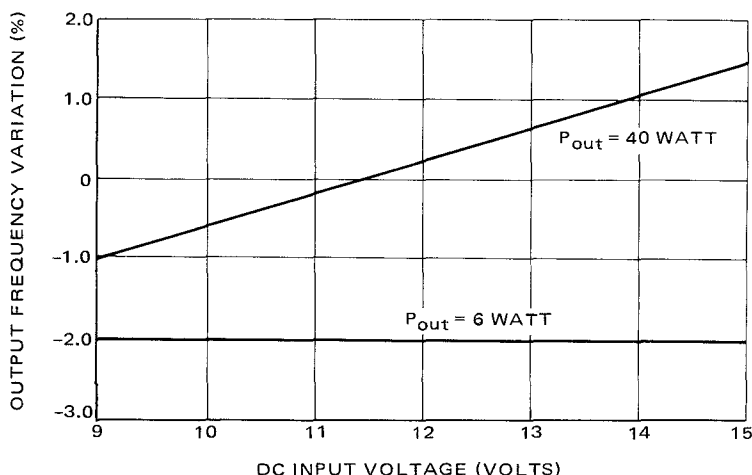
**Figure 2-34 — Output Frequency versus Input Voltage
for Inverter of Figure 2-33**

Nominal input current is 4 A but can vary from 240 mA to 7 A depending on the input voltage and the output load. This means that if the input voltage is 12 V, the permissible range of output current will be from 25 mA to 760 mA. Capacitive loads are permissible if the load surge current is less than 1.5 A.

Circuit efficiency is approximately 80%. Power is purposely sacrificed in the zener diode to obtain frequency stability. Decreasing the zener voltage would improve efficiency.

A unijunction transistor, Q3, is used to start the inverter. Its output voltage pulse causes current to flow into the lower collector winding of transformer T1. Transformer action then creates the base current which turns transistor Q2 on.

In a low-voltage inverter such as this, the $V_{CE(sat)}$ of the transistors used should be low for good efficiency and output voltage regulation. Because the zener voltage adds to the reverse base-emitter voltage drop, a high V_{BE} rating is necessary. The 2N3611 germanium power transistor was chosen mainly because it fulfilled these requirements. In addition, it met the following specifications imposed by the circuit: V_{CES} over 28 V, h_{fe} greater than 10 with I_C - 4 A, V_{BE} over 10 V, and $V_{CE(sat)}$ less than 1 V with I_C - 4 A. The $V_{BE(sat)}$ must be small in relation to the zener voltage; it must also change very little as input voltage and load current vary to obtain good frequency stability. For changes in collector current

from 3 to 5 A, the change in $V_{BE(sat)}$ is only 0.1 V or 1% of the zener voltage.

The ambient temperature range of the circuit is from $0°C$ to $70°C$ without a heat sink and from $0°C$ to $100°C$ with a heat sink on transistors Q1 and Q2. The thermal resistance of the heat sink should be no more than $3°C$ per watt.

2.14 Low Voltage Input High-Current Inverter 2 Vdc - 35 Vac, 80 W, 1 kHz

The 80 watt, 1 kHz current-feedback inverter shown in Figure 2-35 requires an input of 2 volts dc. The output voltage increases only 30% when the load is reduced from full to one-fourth load. Figure 2-36 shows the effects of load changes on both output voltage and efficiency.

Efficiency also varies with load but remains quite high even at full load, as illustrated in Figure 2-36. Maximum efficiency of 87% occurs with a 30 watt load. As the load is increased to 80 watts, efficiency drops slightly to about 70%.
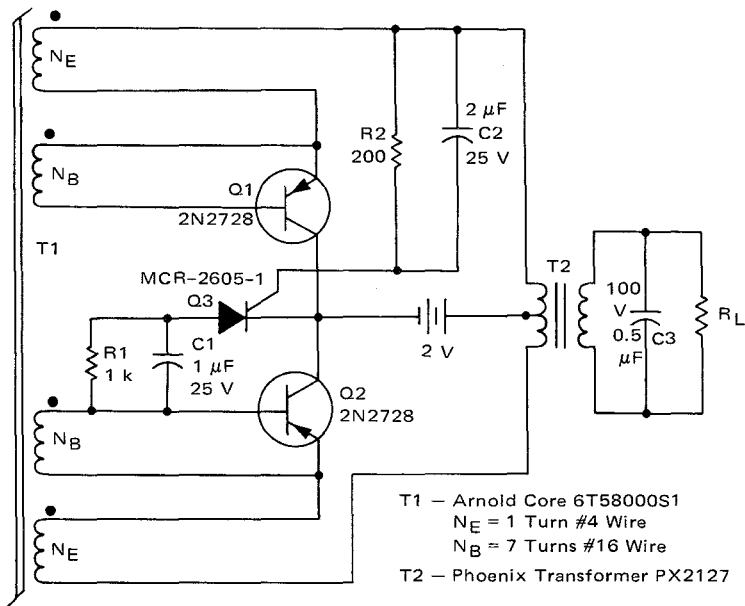
Figure 2-35 — Low-Voltage Input, High-Current Inverter. 2 Vdc-35 Vac, 80 W, 1 kHz
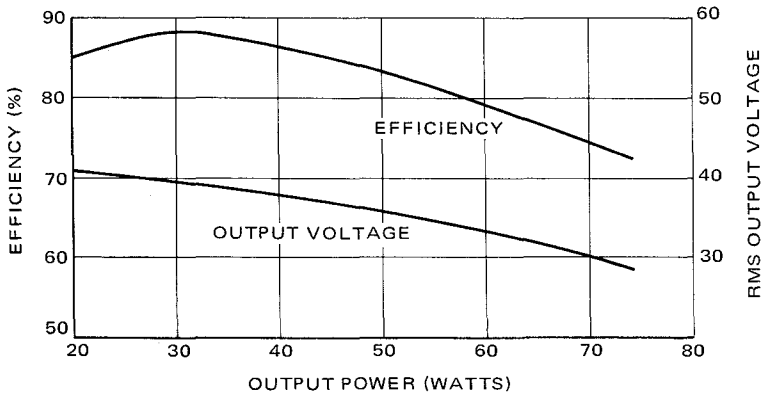
Figure 2-36 — Performance of Inverter in Figure 2-35

The allowable range of input current is 10 to 50 amperes. This means that with an input voltage of 2 V, the output load can be varied from 20 to 80 watts. It is also possible to vary the input voltage from 1.5 to 2.5 V. The frequency is slightly dependent on the load since the resulting change in collector current will cause the forward base-emitter voltage drop to change. However, as input current is increased from 10 to 50 A, the frequency will increase only 20%. In a more practical case, if the load is doubled, the frequency will change by less than 10% (100 Hz.)

The starting circuit for this inverter consists of R1, C1, R2, C2 and Q3. When supply voltage is first applied, both gate and anode of the thyristor are positive with respect to the cathode. This turns the thyristor on and provides base current to start transistor Q2. Saturation of Q2 commutates Q3. After starting of the inverter, the thyristor cannot turn on. When Q2 is on, the anode is reversed biased, and when Q1 conducts, gate voltage is too low to fire Q3 since $V_{CE(sat)}$ is less than 0.1 volt at $I_C = 50$ A.

The operating temperature range of this circuit is 0 to 75°C when Q1 and Q2 are mounted on a heat sink with thermal resistance of less than 3.0°C per watt. Only one heat sink is needed because the cases (collectors) of the two transistors are common electrically. The heat sink must be capable of cooling both transistors.

2.15 High Power Inverter
28 Vdc - 110 Vac, 2 kW, 580 Hz

The 2 kilowatt, 580 hertz current-feedback inverter shown in Figure 2-37 requires an input of 28 volts dc. Its output voltage is a square wave
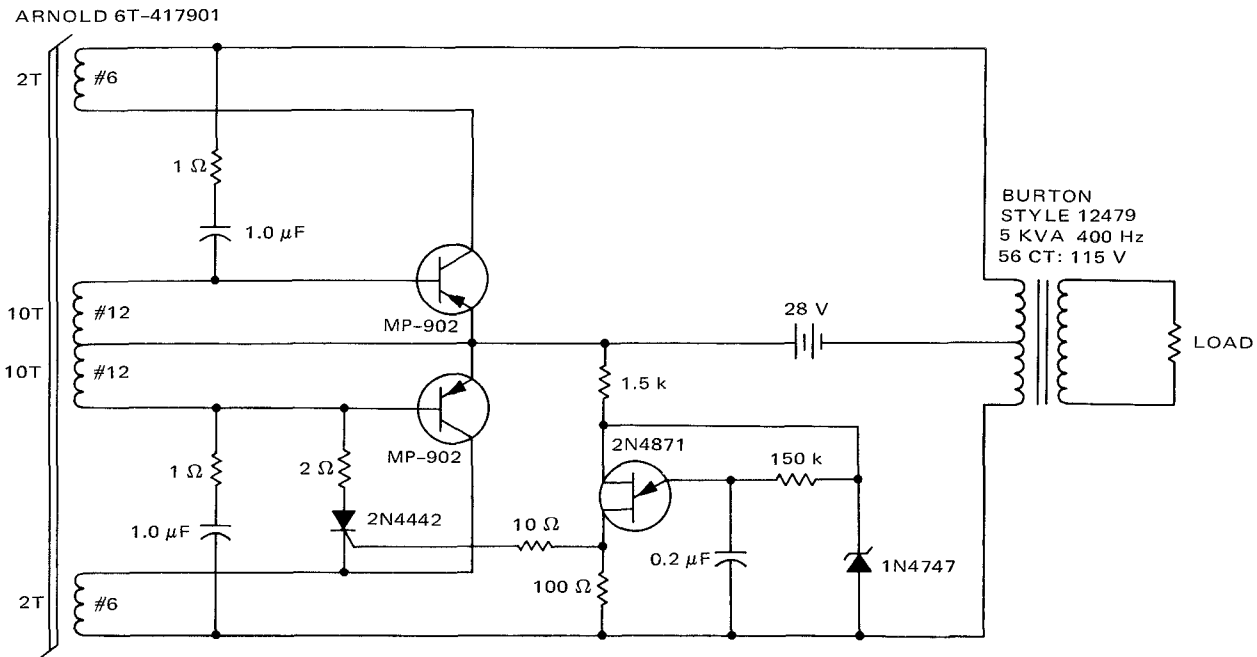
Figure 2-37 — High Power Inverter. 28 Vdc-110 Vac, 2 kW, 580 Hz

2T | #6

1 Ω

1.0 µF

10T | #12

MP-902

10T | #12

1 Ω

2 Ω

MP-902

1.0 µF

2N4442

10 Ω

2T | #6

100 Ω

1.5 k

2N4871

0.2 µF

150 k

1N4747

28 V

BURTON
STYLE 12479
5 KVA 400 Hz
56 CT: 115 V

LOAD

Inverters and Converters

whose peak magnitude is approximately 100 volts. The output voltage regulation and efficiency are excellent for variations in load. The circuit was designed basically for high power and high efficiency. No attempt has been made to stabilize frequency. If it becomes desirable to operate this inverter at 400 hertz, Figure 2-38 can be used to determine the appropriate limit for output power. For instance, if the load is allowed to vary from 500 to 1000 watts, the frequency will increase only 70 hertz (from 360 to 430 Hz). The output voltage over this range decreases only 3 volts (from 113 to 110 V). Another advantage is gained by operating between 500 and 1000 watts: efficiency is greater than 90% in this range.

With the input voltage held constant at 28 volts dc, the variations in output voltage, frequency, and efficiency are plotted as functions of output power in Figure 2-38. Output power was varied from 125 watts to full load of 2050 watts. The output voltage decreased from 114 to 98 volts which means that the regulation is better than 20%. It takes more than a 100 W increase in load power to decrease the output voltage by 1 volt. Frequency, on the other hand, jumped from 260 to 580 hertz which means that it more than doubled as the output power was increased from very low load to full load. The increase in frequency is approximately 20 hertz for each 100 W increase in load. The efficiency of this circuit remains high over the entire range of output power. At 125 W output, efficiency is 77%. It increases to 93% at 500 watts and then decreases again to 82% at full load.

The performance curves in Figure 2-39 were obtained with a con-
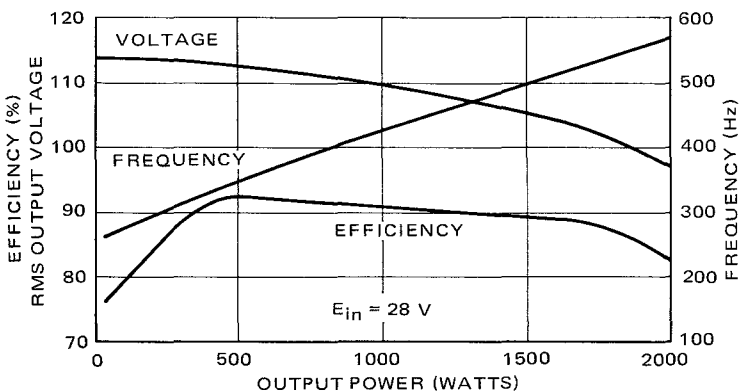
Figure 2-38 — Performance of High-Power Inverter of Figure 2-37 for Varying Output Power

stant load of about 5 ohms, which represent 2 kilowatts or full load with a 28 volt dc input. The variations of output voltage, frequency, and efficiency are plotted as a function of input voltage. The input voltage was varied from 5 to 32 volts dc, a factor of about six. As could be expected, the output voltage did nearly the same; it increased by a factor of six from 18 to 115 volts. The increase in frequency was not as drastic; it only increased from 340 to 610 hertz which means that it did not even double. Efficiency again remains quite high as input voltage is varied; it remains constant at 82% from 15 to 28 volts but decreases rapidly beyond these limits.

A special feature of this circuit is the ac network connected to each base. This is a "booster" type of voltage feedback which provides additional base drive during switching. The nominal input voltage is 28 volts dc but the circuit can be operated with voltages from 24 to 32 V. Starting is unreliable at lower voltages. The maximum input current is 100 A.

In general, the power transistors used in this circuit must have low $V_{CE(sat)}$ and adequate current gain at $I_C = 100$ A. The MP902 germanium transistor has an extremely low $V_{CE(sat)}$ of 0.5 V but the gain at $I_C = 150$ A is only 5. The bases were driven with a low forced gain, 5 instead of 10, to insure good operation at $I_C = 100$ A. The most critical requirement was for a clamped inductive safe operating area of 70 V with 100 A current. The test point which insures that this specification is met is at 55 V and 150 A.

The load resistance can be varied from 100 to 5 ohms. Higher resis-
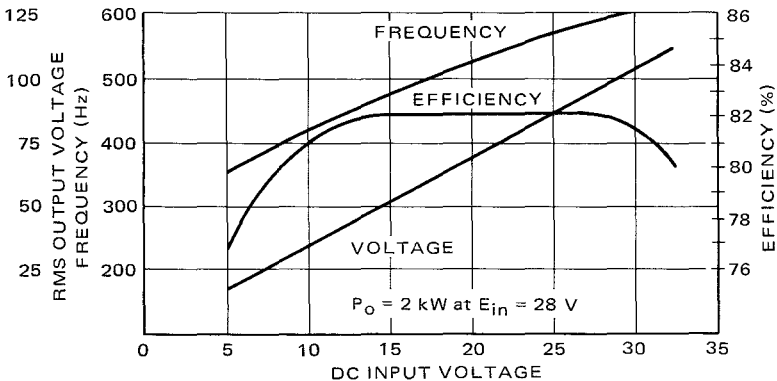
**Figure 2-39 — Performance of Inverter of Figure 2-38
for Varying Input Voltage**

tances tend to make starting difficult. Lower resistances can cause the power transistors to pull out of saturation, which can destroy them. The power transistors were mounted side by side on a silicone-greased heat sink (Wakefield MN-2131-6.0 without insert). Without forced air cooling, the circuit can be operated at ambient temperatures up to 55°C at full load.

2.16 High Efficiency Inverter
120 Vdc - 240 Vac, 1 kW, 1200 Hz

The nominal input voltage of the current-feedback inverter shown in Figure 2-40 is 120 Vdc, but it can be operated with an input voltage greater than 20 V but less than 160 V. The output voltage is a 1200 Hz, 240 V square wave (with 120 V input.) Output current is about 4 A, but the transistors can handle up to 5 A in the load.

The efficiency of this circuit is excellent. Figure 2-41 shows that efficiency is 92% at 1000 watt output. Frequency stability is also very good. In Figure 2-42, it is shown that a 6:1 change in input voltage will

T1 — ARNOLD CORE 6T 4168S1
46T OF #23 FOR EACH BASE WINDING
5T OF 3 #18 FOR EACH COLLECTOR WINDING
T2 — PHOENIX TRANSFORMER PX-2677
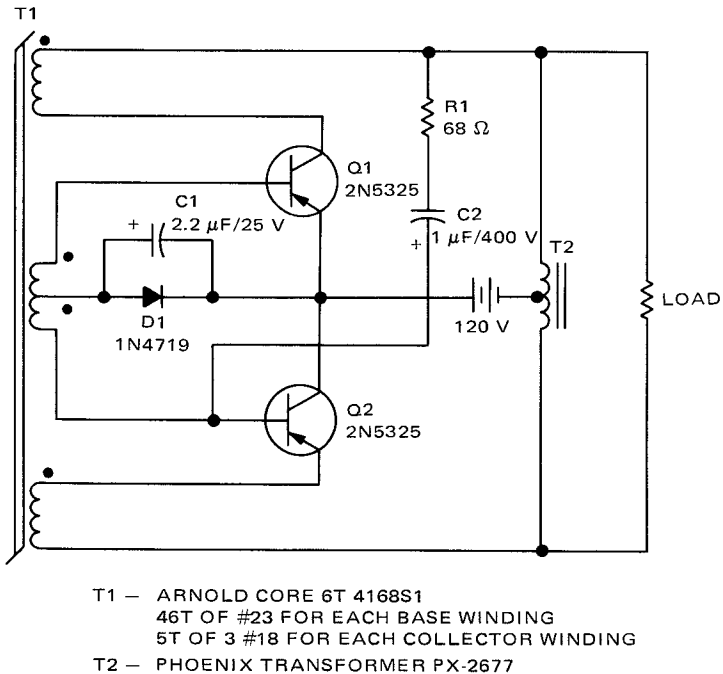
Figure 2-40 — High-Efficiency Inverter. 120 Vdc-240 Vac, 1 kW, 1200 Hz

shift the frequency by only 20%. Variation in load impedance produces similar results. Voltage regulation is less than 2% when the load is varied from no load to full load. However, the output voltage will change in direct proportion to the input voltage unless a preregulator stage is used.

Even though this circuit is self-excited, it will fail safe if either input voltage or load is removed. That is, one transistor will not remain on and burn out. However, there is no current limiting and the transistors can be damaged by a short circuit in the load. There are no starting limitations for the circuit. It can be started into a full load or a capacitive input load up to the surge rating capability of the transistor.
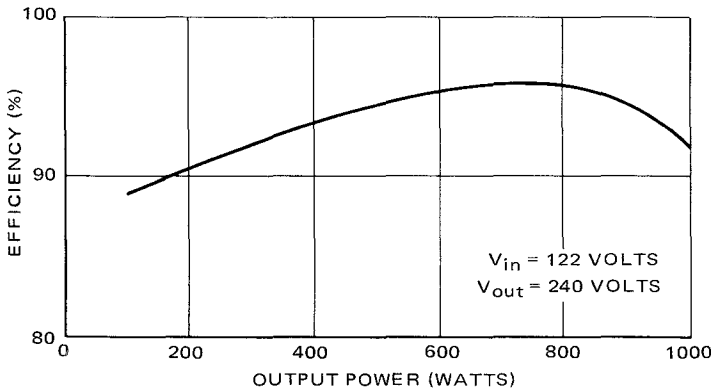
Figure 2-41 — Efficiency versus Output Power for High-Efficiency Inverter of Figure 2-40
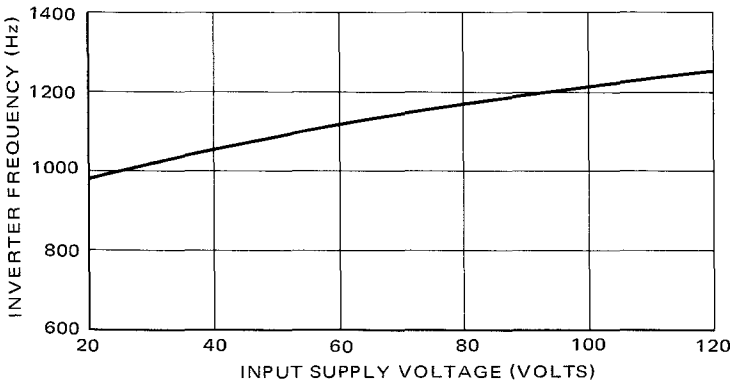
Figure 2-42 — Frequency versus Input Voltage for High-Efficiency Inverter of Figure 2-40

If the transistors are mounted on a heat sink with a thermal resistance of less than 3°C per watt, the ambient temperature range will be from 0 to 80°C.

2.17 One-Transformer, Car-Battery-to-AC-Line-Voltage Inverter 13 Vdc - 115 Vac, 180 W, 60 Hz

The schematic of a 180 watt, 60 hertz voltage-feedback inverter designed for 13 volt dc input is shown in Figure 2-43. The output of this inverter is a 113 volt square wave; its magnitude varies with load and input voltage. Figure 2-44 shows that as load is increased, the voltage will decrease. The voltage regulation from no load to full load (180 watts) is less than 30%. This means that it takes a 6 watt increase in load to decrease the output voltage by 1 volt. A corresponding curve of output voltage versus input voltage is shown in Figure 2-45. As input voltage increases, the output also increases. However, a change in input voltage of ±11.5% (±1.5 V) will only cause the output voltage to vary by ±7.5%. This is a ratio of about 2 to 3.

Output frequency is even more stable than the output voltage. In Figure 2-44, it can be seen that frequency decreases linearly as the load is increased. The change in frequency from no load to full load is slightly less than 12%. This means that it takes a 30 watt increase in load to decrease the frequency by approximately 1 hertz. Frequency also varies with input

T1 — Forest Electric VJ–18
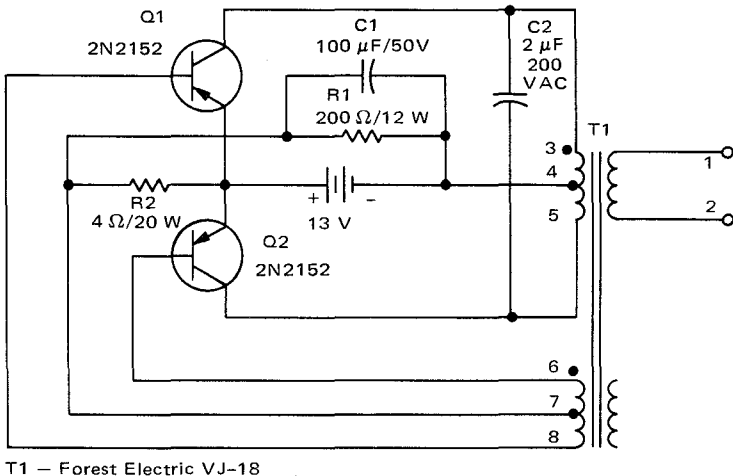
Figure 2-43 — Car Battery-to-Line-Voltage Inverter. 13 Vdc-115 Vac, 180 W, 60 Hz. Felco Model TR-181

voltage as shown in Figure 2-45. Here a change in input voltage of ±11.5% produces about the same change in frequency (±12%).

The curve of efficiency versus output power is shown in Figure 2-44. Efficiency increases from about 60% at 40 watts to 75% at full load. This is a good figure for a one-transformer inverter.

The operating range of input voltage is from 12 to 14 volts. Input current can vary from 2 to 20 A. This means that it is possible to operate a load of 200 watts with 14 volt input as shown in Figure 2-45. It is also possible to operate with no load. Loads in excess of 200 watts and short circuits can destroy the transistors and should be avoided.
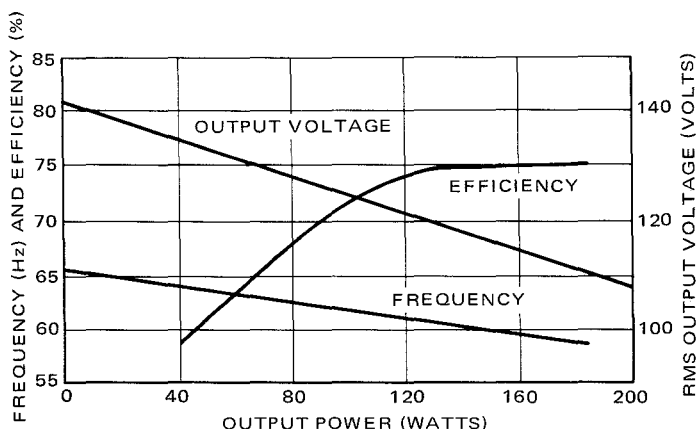
Figure 2-44 — Performance of Inverter in Figure 2-43
as a Function of Output Power
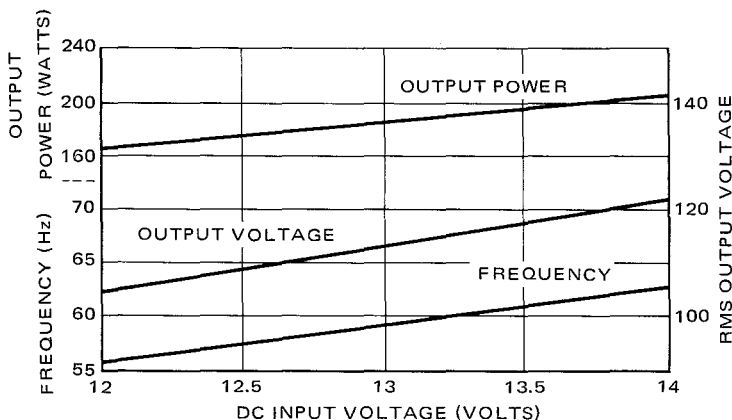
Figure 2-45 — Performance of Inverter in Figure 2-43
as a Function of Input Voltage

The 2N2152 germanium power transistor was chosen for this inverter because of its low cost and low $V_{CE(sat)}$. Specifically, the transistor has a $V_{CE(sat)}$ of 0.3 V at an I_C of 25 A and costs less than \$3.00. Other specifications necessary were h_{fe} greater than 10 at $I_C = 25$ A, V_{CES} of 45 V, and V_{EB} greater than 8 V because of the drop across R2 in the base-emitter circuit.

The circuit can be operated in ambient temperatures from 0 to 60°C. The transistors should be mounted on heat sinks with thermal resistances of about 3°C per watt.

2.18 Efficient, High Frequency Inverter
28 Vdc - 52 Vac, 100 W, 15 kHz

The 100 watt, 15 kHz voltage-feedback inverter shown in Figure 2-46 requires 28 volt dc input. The output voltage is a square wave with a magnitude of 52 volts rms.

T1 — ARNOLD CORE GT-5502-D500
N1 26 TURNS OF AWG #14 WIRE
N2 52 TURNS OF AWG #14 WIRE
N3 3 TURNS OF AWG #22 WIRE
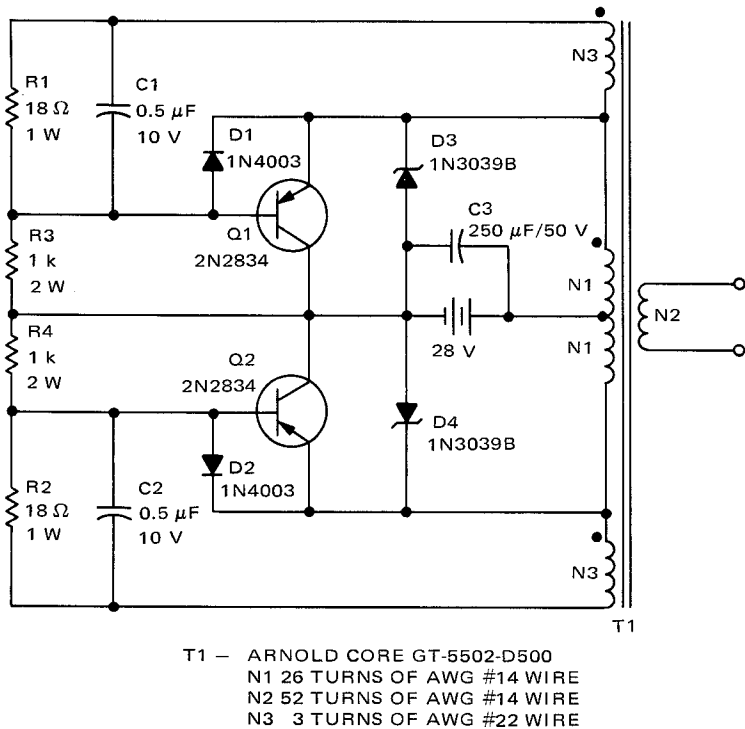
Figure 2-46 — Efficient, High-Frequency Inverter.
28 Vdc-52 Vac, 100 W, 15 kHz
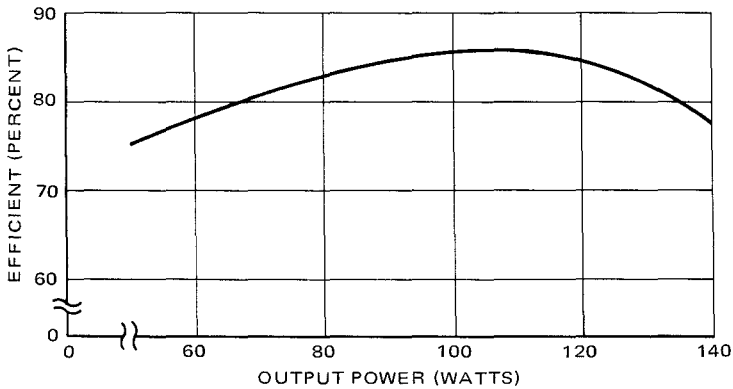
Figure 2-47 — Efficiency versus Output Power for Inverter of Figure 2-46

The frequency stability of this circuit is excellent. Frequency is dependent only on input voltage and not load. Any variation in output power will not affect the frequency. However, a 10% increase in input voltage will cause a 10% change in frequency. In other words, if the input voltage increases by 1 volt, the output frequency will increase by approximately 500 hertz.

The efficiency, as shown in Figure 2-47, is quite high. It reaches a maximum of 85% at approximately 100 watts. Even at 50 watts of output power, the efficiency drops only slightly to about 75%. This would be higher if there were a way to reduce the base drive as the output power decreases.

The output voltage, like frequency, is well regulated for changes in load. Figure 2-48 shows that the output voltage remains constant as the load is increased from 50 to 100 watts. The transistors are not fully saturated above this point and the output voltage starts to decrease at higher power levels. The output voltage, however, is affected by changes in input voltage. The output transformer almost doubles the input voltage. The output voltage can be increased or decreased by adding or subtracting turns from secondary winding N2.

The output power is nominally 100 watts but the inverter can be operated at any level from no load to this value. The output transformer will overheat at higher power levels, but the transistors will handle 140 watts of output power. The fast switching speed of the transistors permits efficient switching at 15 kHz. The range of output power could be increased to this value if the tape-wound core in transformer T2 were re-

placed by a ferrite core. Diodes D1 and D2 were used to protect the base-emitter junctions of Q1 and Q2 from reverse voltage.

Input voltage variations of ±10% will not affect the output significantly. At full load (100 watts) the input current will be 4 A.

Both transistors can be mounted on a single heat sink with the collectors uninsulated. The thermal resistance of the sink should be 3°C per watt or less. The ambient temperature may then be allowed to vary from 0 to 60°C.

Since this is a high-frequency inverter, it is excellent for applications which require minimum size and weight.
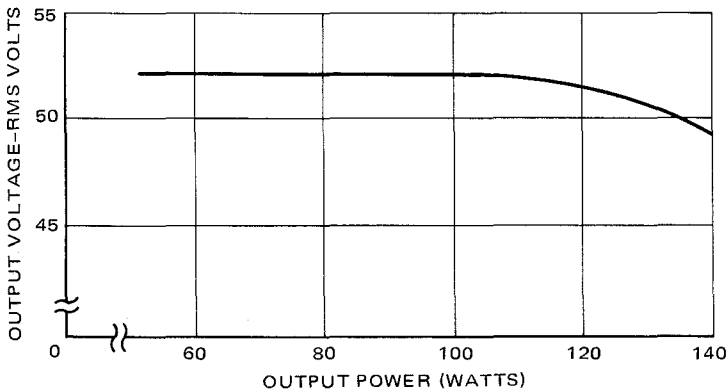
Figure 2-48 — Output Voltage versus Output Power for Inverter of Figure 2-46

2.19 Efficient Inverter with Good Output Regulation 12 Vdc - 115 Vac, 400 W, 400 Hz

The 400 watt, 400 hertz hybrid-feedback inverter in Figure 2-49 furnishes a 115 volt square wave output with a 12 V input. Figure 2-50 shows the output voltage variation with load. As load is reduced to 50 watts, the voltage increases by 10%. Voltage regulation is therefore quite good. The curves were taken with the input voltage constant at 12.6 volts; input voltage changes will affect the output voltage. An increase in input voltage will cause the output to increase.

The efficiency of this circuit is high; it remains greater than 85% for the entire range of output power. In Figure 2-50 the efficiency is shown reaching 89% with a 200 watt load.

The nominal input voltage is 12.6 volts dc, but the input voltage may be allowed to vary from 10 to 14 volts. The output load may be varied from no load to a maximum of 400 watts. As the load is varied over this range, the input current will increase from 0.75 to 40 A. Loads that exceed 400 watts, and output short circuits, can damage power transistors Q1 and Q2. This inverter will operate into both resistive and inductive loads. It has been used with fluorescent and incandescent lights, and a 400 hertz, 90 watt motor.
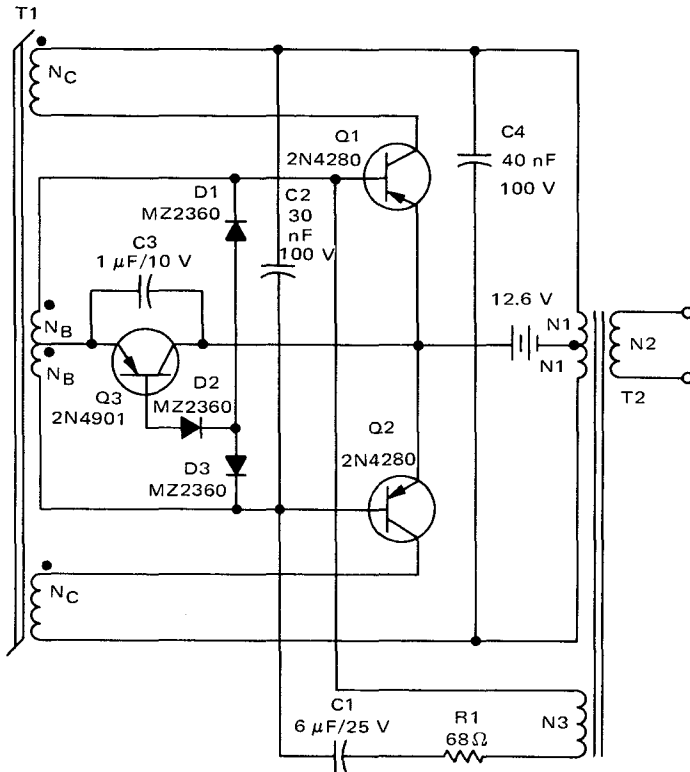
TRANSFORMERS

T1 — ARNOLD CORE 6T 7699D1
N_B 54 TURNS #20 WIRE
N_C 3 TURNS TWO #12 WIRES IN PARALLEL

T2 — ARNOLD CORE 3T 6464-L4
N1 12 TURNS TWO #12 WIRES IN PARALLEL
N2 120 TURNS #16 WIRE
N3 7 TURNS #20 WIRE

Figure 2-49 — Hybrid-Feedback Inverter with Good Output Regulation. 12 Vdc-115 Vac, 400 W, 400 Hz
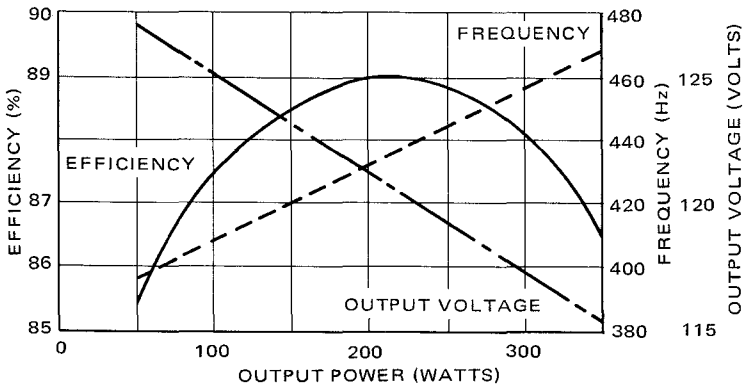
Figure 2-50 — Performance of Inverter of Figure 2-49

The output frequency is slightly dependent on load and input voltage. Figure 2-50 shows an output frequency of 470 hertz at 350 watts; it decreases to 400 hertz at 50 watts. For a 7-to-1 reduction in load, the frequency decreases by only 15%. The base-emitter voltage, which would normally determine frequency, decreases from 0.63 to 0.40 volts over this range of output power. However, the voltage across winding N_B only decreases from 1.8 to 1.6 volts because of frequency-stabilizing transistor Q3.

The circuit was originally operated without feedback-winding N3. Efficiency dropped off at light loads and it would not start without a load. However, the output frequency regulation remained the same and the range of variation was lowered by about 30 hertz. That is, the frequency at 350 watts was 435 hertz, and it dropped to 378 hertz at 50 watts.

The output frequency can be decreased in another way if desired. It turns out that each additional turn on winding N_B will lower the frequency by 10 hertz. In order to obtain sufficient current gain at 400 watts, winding N_B must be limited to 62 turns. This means that up to eight turns may be added. This would decrease the frequency by 80 hertz. With a 400 watt load, output frequency would then decrease from 480 to an even 400 hertz.

2.20 Line-Operated, High Frequency Inverter
120 Vac - 120 Vac, 200 W, 15 kHz

Figure 2-51 shows a 200 watt, 15 kHz, line-operated hybrid-feedback inverter. It is designed for an input voltage of 120 V at 60 Hz. In order to keep size to a minimum, there is very little filtering of the recti-
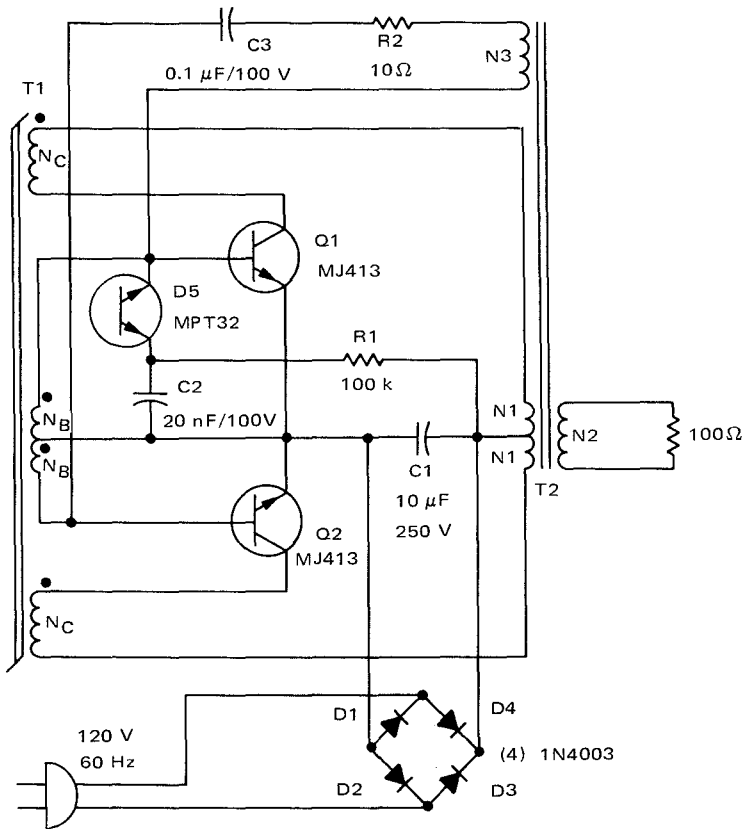
TRANSFORMERS
T1 — CORE — MAGNETICS INC. #80623-50-080
   N_B 15 TURNS OF #26 AWG WIRE
   N_C 3 TURNS OF #22 AWG WIRE
T2 — CORE — ARNOLD 6T-5800-D1
   N1 100 TURNS OF #22 AWG WIRES (3 EA.)
   N2 104 TURNS OF #19 AWG WIRES (3 EA.)
   N3   7 TURNS OF #26 AWG WIRE

Note: N1 and N2 are Trifilar Wound in Parallel.

**Figure 2-51 — Line-Operated, High-Frequency Inverter.
120 Vac-120 Vac, 200 W, 15 kHz**

fied input. This means that the voltage across C1 varies from 10 to 150 and back to 10 volts during each alternation of the input voltage. This makes the output voltage an amplitude-modulated 15 kHz square wave. Because the additional turns on the output winding compensate for the internal voltage drops, the output voltage magnitude remains 120 V rms. The output voltage is unaffected by changes in load but will vary with input voltage.

The output frequency is dependent on both input voltage and output load as shown in Figure 2-52. With a 100 ohm load, a dc voltage was connected across C1. As the voltage varied from 120 to 70 V, the frequency changed by less than 1%. Below 70 V, the frequency begins to increase as shown. With a 100 Vdc input and variable load, similar changes in frequency occur. That is, with an increase in load from 40 to 80 watts, the frequency change was less than 1%. Above 80 watts, the frequency again increases as shown.

Output power is nominally 150 watts, but the circuit may be operated with loads between 2 watts and 200 watts. At the nominal load, efficiency is 88%.

This inverter was developed for an application which required isolation from the line with minimum size and weight. Better frequency stability can be obtained with increased input filtering. The voltage across C1 should ideally be kept above 70 V. It can also be seen that variations in load will affect frequency unless the output power is kept below 80 watts.

The hybrid feedback used in this circuit is useful with inductive
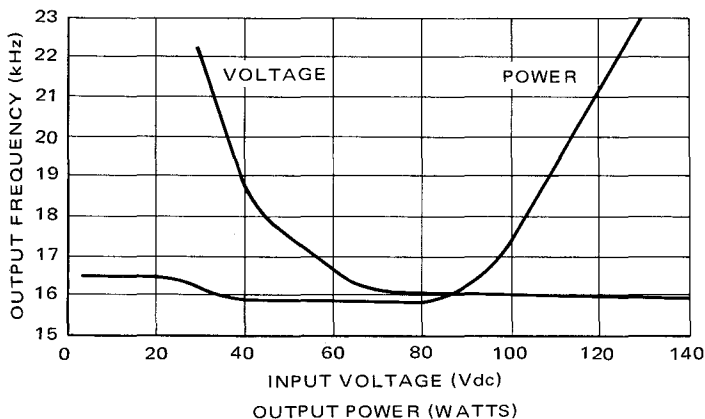
Figure 2-52 — Output Frequency versus Input Voltage and Output Power for Inverter of Figure 2-51

loads, and with light loads. Because of capacitor C3 in the voltage feed-back loop, the voltage feedback is maintained only during the first portion of each half cycle. Each time the inverter switches, a pulse of current from the voltage feedback or "booster" circuit consisting of R2, C3, and N3 is injected at the base of the transistor being turned on. This supplemental current adds to the base current from the current feedback transformer to assist transistor switching. If the load is inductive, it also serves to maintain bias until collector current is adequate to provide the feedback needed to sustain conduction. At low loads (low current in N_C), the booster circuit is adequate to provide oscillation. At full load, regular current-feedback operation predominates and the booster circuit has little effect on the inverter operation.

2.21 Multivibrator-Driven Inverter
12 Vdc - 20 Vac, 50 W, 1 kHz

The 50 watt, 1 kHz, driven inverter shown in Figure 2-53 requires a 12 volt dc input. The output voltage is a square wave with an rms magni-tude of 20 volts. The inverter is driven by a free-running multivibrator which provides excellent frequency regulation for variations in both input voltage and load.

In driven inverters, switching of the output power transistors is accomplished by multivibrator drive rather than by feedback from the output transformer. Multivibrator driven transistor inverters are useful for precision systems requiring carefully controlled frequency, waveform, etc.,

T1 — Phoenix Transformer PX2677
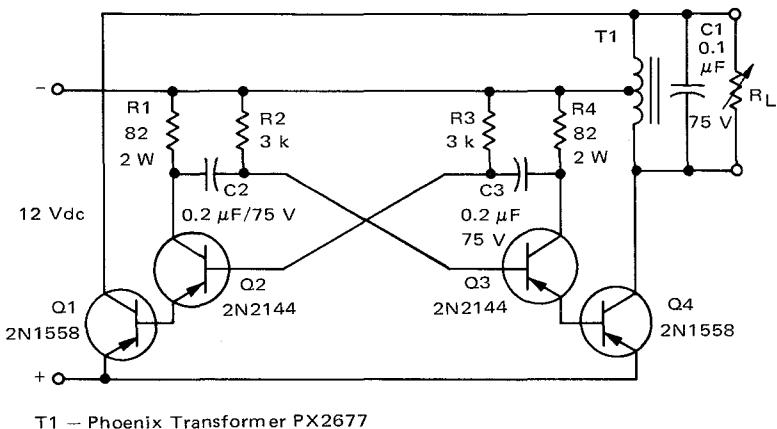
Figure 2-53 — Multivibrator-Driven Inverter.
12 Vdc- 20 Vac, 50 W, 1 kHz

and for load independent systems. Load-independent systems are especially attractive for reactive loads, and when transient or starting conditions impose loads which would cause self-oscillating inverters to shut down or operate abnormally. Power requirements of the multivibrator are largely offset by lack of output transformer saturation and elimination of the feedback drive. The multivibrator-driven inverter is not inherently less efficient than a self-oscillating inverter, but use of the driven power stage as a linear amplifier rather than as a saturated switch will result in high dissipation in the transistors and low system efficiency.

Figure 2-54 shows the effect of variations in load on output voltage, efficiency, and frequency. Output power is varied from about 10 to 70 watts. Frequency increases with load from 1030 to 1080 hertz. Frequency regulation is better than 5% and frequency increases by only 1 hertz as the load is increased by 1 watt. Output voltage decreases with load by almost 4 volts. This is just less than 20% regulation and means that the output voltage will decrease about 1 volt for a 20 watt increase in load. Efficiency is about 36% at 10 watts and 64% at 50 watts. This drops again to about 56% as the load is further increased to 70 watts.

Figure 2-55 shows the effect of variations in input voltage on output frequency. As the input voltage ranges from 10 to 14 volts, frequency decreases from 1090 to 1040 hertz. Regulation here is again better than 5% and frequency increases approximately 10 hertz per volt. Below 10 volts, the increase in frequency is more pronounced but the circuit is not normally operated in this range.

Figure 2-56 shows the effect of variations in input voltage on the output voltage magnitude. As the input is increased from 10 to 14 volts, or
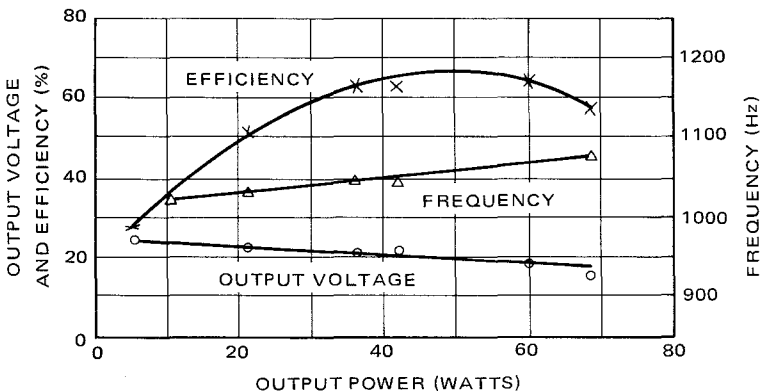
Figure 2-54 — Performance of the Inverter in Figure 2-53 as a Function of Output Power
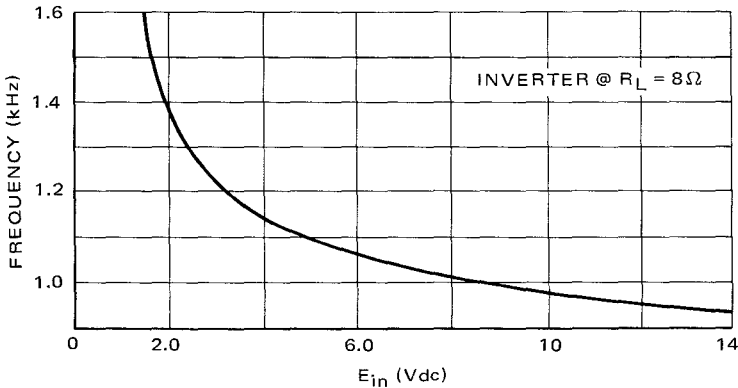
**Figure 2-55 — Frequency versus Input Voltage
for the Inverter of Figure 2-53**
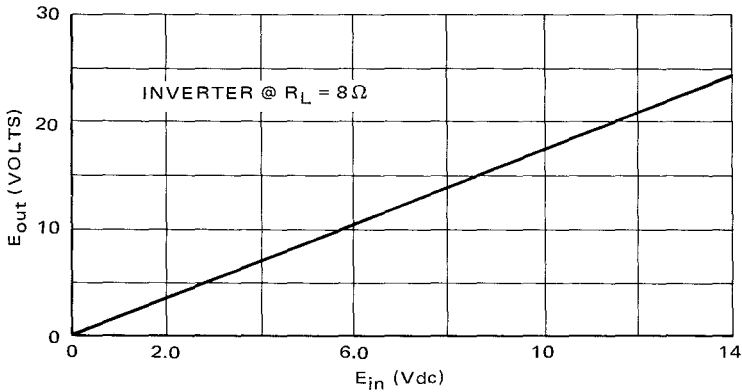
**Figure 2-56 — Output Voltage versus Input Voltage
for the Inverter of Figure 2-53**

about 40%, the output does nearly the same. That is, the output voltage increases from 17 to 24 volts rms, just less than 40%.

Output power may be varied from 10 to 70 watts. The maximum input current will be 10 A dc. Input voltage is normally 12 volts with an operating tolerance of ±2 volts dc. The circuit will operate with input voltages as low as 2 volts dc, but the frequency regulation will be poor (see Figure 2-55). The nominal load resistance is 8 ohms and this corresponds to 50 watts of output power. With 4 ohms in the load, the output power will be about 70 watts. This represents the minimum amount of load resistance. If load is increased beyond this point, transistor Q1 and Q4 come out of saturation and may be destroyed.

BIBLIOGRAPHY FOR CHAPTER 2

1. Bedford, B. D., and R. G. Hoft, *Principles of Inverter Circuits*, John Wiley & Son, Inc., 1964.

2. Glorioso, Robert "Converter Cuts Start-Up Power, Offers Good Regulation" *Electronics*, Feb. 6, 1967.

3. Jensen, James Lee, "An Improved Square-Wave Oscillator Circuit." *I.R.E. Transactions on Circuit Theory*. Ct 4:3 pages 276-279. September 1957.

4. Kusko, A. and B. Szpakowski, "Load Ranges of Series SCR Inverters," *Electro-Technology*, April 1965.

5. Magnetics, Inc. *Design Manual Featuring Tape Wound Cores*. Buttler, Pa.: 1962 pages 46-50.

6. Mapham, Neville "An SCR Inverter with Good Regulation and Sine Wave Output," *IEEE Transactions* IGA-3 No. 2 Mar/Apr 67.

7. Murphy, R. H. and K. P. P. Nambiar, "A Design Basis for SCR Parallel Inverters," *Proceedings of the IEEE Paper 3642E*, Sept. 1961.

8. Ott, Richard R., "A Filter for SCR Commutation and Harmonic Attenuation in High Power Inverters," *Communications and Electronics* May 63.

9. Palmer, Merrill, "The ABC's of DC to AC Inverters." Application Note AN-222. Phoenix, Arizona: Motorola Semiconductor Products, Inc. 1966.

10. Palmer, L. Merrill. "Current Feedback Inverter for Low Input Voltage Applications." *Electronic Communication* Vol. 1 No. 5, (Sept.-Oct. 1966) page 5.

11. Saladin, Herb, "Low-Cost Power Inverter Circuits Using Off-the-Shelf Components" Motorola AN-134.

12. Royer, G. H., "A Switching Transistor DC to AC Converter Having an Output Frequency Proportional to the DC Input Voltage." *A.I.E.E. Transactions*, pt. I (*Communications and Electronics*) vol. 74, pages 322-324, July 1955.

13. Takesuye, Jack, "A 15KC Power Inverter," Motorola AN-199.

14. Takesuye, Jack, "A Low Voltage High Current Converter." Application Note AN-169. Phoenix, Arizona: Motorola Semiconductor Products, Inc. 1965.

15. Takesuye, Jack, and Howard Weber, "Silicon Power Transistors Provide New Solutions to Voltage Control Problems" Motorola AN-163.

16. Thompson, R., "Designing Series SCR Inverters," *Electronic Design*, July 5, 1963.

17. Thompson, R., "Team Saturating Chockes with SCRs," *Electronic Design*, Nov. 8, 1966.

18. Tibbetts, L. M., "SCR Bridge Inverter Eliminates Transformers," *Electronics*, September 5, 1966.

19. Weber, Howard, "Low Voltage Inverter Features High Frequency Operation With High Efficiency," Motorola AN-174.

20. Weisz, T., "Disregard Load Impedance in SCR Inverter Design," *Electronic Design*, May 24, 1967.

21. Wilson, T. B. and E. T. Moore, "Inverters for Use with Very Low Input Voltages." *IEEE Transactions on Communications and Electronics* 63-1450. 83:73 pages 424-248. July 1964. Also available as NASA Technical Brief B65-10178. "DC to AC Converter Operates Efficiently at Low Input Voltages." Greenbelt, Maryland: Goddard Space Flight Center.

22. Wilson, T., "Voltage Controls Dual-Pulse SCR Trigger," *Electronics*, Nov. 2, 1964.

CHAPTER 3

Regulated Power Supplies

3.1 Regulated Power Supplies

The power supplies considered in this chapter are electronically regulated supplies designed for ac and dc inputs and outputs. Included are series-pass dc-output voltage regulators, switching-mode dc-output voltage regulators, dc-output current regulators and ac rms voltage regulators. The purpose of a dc regulator is to maintain a constant ripple-free dc output for changes in input voltage, load impedance, and temperature. The device used as the regulator control can be either in series with the load or in shunt with the load. Shunt regulators are not considered here since their low efficiency limits their practical usage to regulators with low output voltages and fairly constant loads. If the regulating element is to maintain continuous control of the output voltage, then the device must be a transistor. If the regulator is to operate in the switching mode and the primary power source is dc then a transistor is also the best choice for the control element. If the primary power source is ac then either a thyristor or a transistor (a rectifier must precede the transistor) can be used as the switching element.

Series Pass Regulators

The ideal voltage-regulated power supply would have a zero output impedance so that the output voltage would remain constant for any load current requirements. A zero output impedance cannot be achieved although supplies with output impedance levels on the order of milliohms can be constructed. The properties of semiconductor devices place a limit on the maximum current and voltage that can be supplied to a load; these two parameters are dependent on each other and upon the particular circuit used as a regulator. Two simple series-pass voltage regulators are shown in Figures 3-1 and 3-2. Transistor Q1 in each of these circuits is required to buffer the difference between the unregulated input voltage and the required output voltage, thus the collector-emitter voltage rating of the device used for Q1 determines the maximum input and output voltage. The current delivered to the load must pass through Q1. Therefore
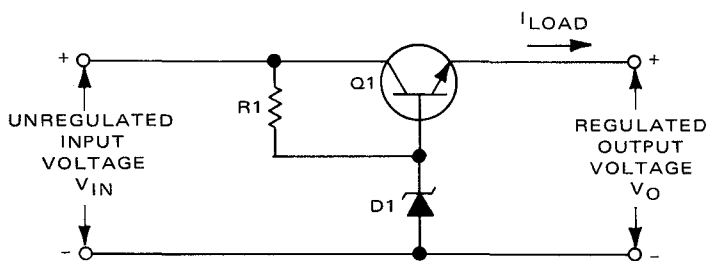
Figure 3-1 — Simple Series-Pass, Fixed-Voltage Regulator
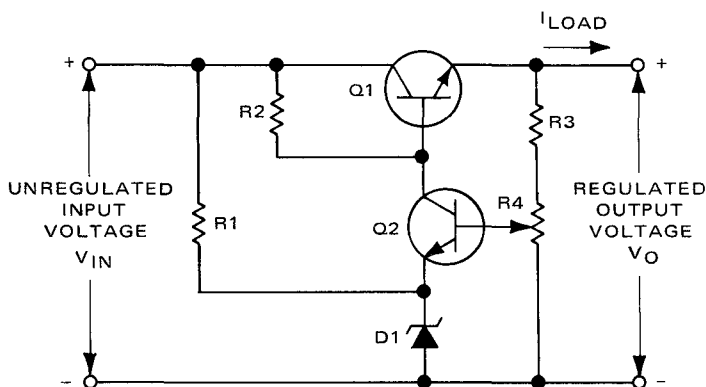
Figure 3-2 — Simple Series-Pass, Variable-Voltage Regulator

the average power dissipated in Q1 is $(V_{in} - V_{out})$ times I_{LOAD}. If high input voltages and low output voltages and/or large load currents are required, large amounts of power can be dissipated in Q1. Obviously under these conditions, the regulator efficiency will be low. The power rating for the particular heat sinking used, therefore, limits regulating capacity. For limited load variations a resistor may be placed in shunt with Q1, thus shifting some power from Q1 to the resistor. For any input supply voltage and output load combinations, the safe operating area of Q1 must not be exceeded. (See Motorola Applications Note AN415.) The output voltage of the circuit shown in Figure 3-1 is the zener voltage minus the base-emitter voltage of Q1. R1 must provide enough current to the base of Q1 and to D1, to keep the voltage of D1 above its breakover point at all times. If the voltage versus temperature characteristic of D1 is equal and opposite to that of the base-emitter junction of Q1, the output voltage will remain constant with temperature. If a variable output voltage is required, then a

circuit similar to that shown in Figure 3-2 can be used. The lowest output voltage attainable is the breakdown voltage of D1 plus the base-emitter voltage of Q2, if R3 is small compared to R4. The highest output voltage is limited by the lowest excursion of the unregulated input plus enough voltage to provide adequate drive to Q1 through R2. R1 provides bias current to D1; having its voltage compared to that at the potentiometer arm through Q2. If the load increases, thereby decreasing the regulator output, Q2 will conduct less and Q1 will conduct more, thus restoring the voltage to the original level.

If the requirements of the regulator are greater than one transistor can handle, then the power section of the regulator may be as shown in Figure 3-3 or 3-4.

Series operation of several transistors can be used to decrease the power dissipation in each unit or to regulate voltages greater than one transistor can withstand. Since the transistors all carry the same current, the voltage must be divided equally to equalize individual power dissipations and to prevent one device from failing due to operation outside its safe area. Resistors connected between the bases of each transistor can be used to equalize the voltages, but the power dissipation in the resistors can be excessive for high output current and for wide fluctuations in output voltage such as expected in series regulator service. Transistors which can withstand 700 V are presently available. If voltages above this are required, then series transistor operation can be used.

The controlled device for the circuit shown in Figure 3-3 is Q2. The maximum voltage across Q2 is the breakdown voltage of zener diode D1 minus the base-emitter voltage of Q1, thus the maximum power dissipated in Q2 is easily controlled. Q1 and shunt resistor R2 must now withstand the remainder of the total series pass voltage. The higher this voltage becomes, the greater is the share of the total current which passes through R2. Therefore, both the power and the safe operating area of the transistors have been reduced. Only two transistors are shown, but more can be placed in series if required.

The circuit shown in Figure 3-4 will deliver load currents greater than that possible with a single device, since the series-pass transistors are placed in parallel. The major problem with this connection is having the current divide equally between Q1 and Q2 in order to equalize power dissipations. To achieve this, either the transistors must be matched or feedback must be used to compensate for individual characteristics. Resistors R1 and R2 provide such feedback at the expense of increased power loss in the resistor; also, more control signal is required. However, a bonus of this connection is that the thermal stability of the regulator is improved. The maximum output-to-input voltage is limited by the collector-

emitter voltage rating of the transistors used. Again, only two transistors are shown, but more can be paralleled if required.

Basic limitations of series-pass regulators are maximum current, voltage and power dissipation ratings, and safe operating area of the series control transistor. The design of the particular circuit and the components used control the degree of regulation, output impedance, temperature coefficient and other parameters mentioned.
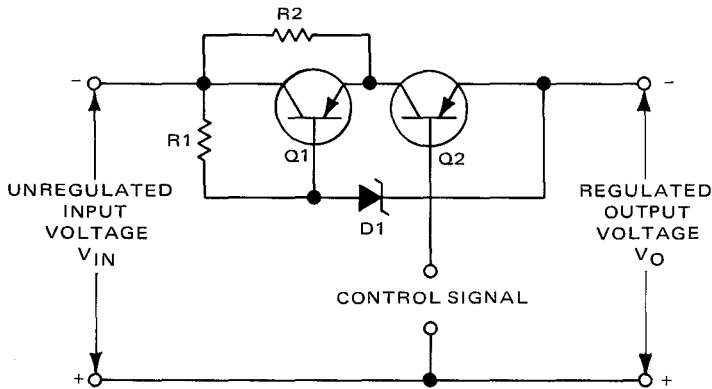
Figure 3-3 — Typical Series-Connected Transistors
with Power-Dissipation Resistor
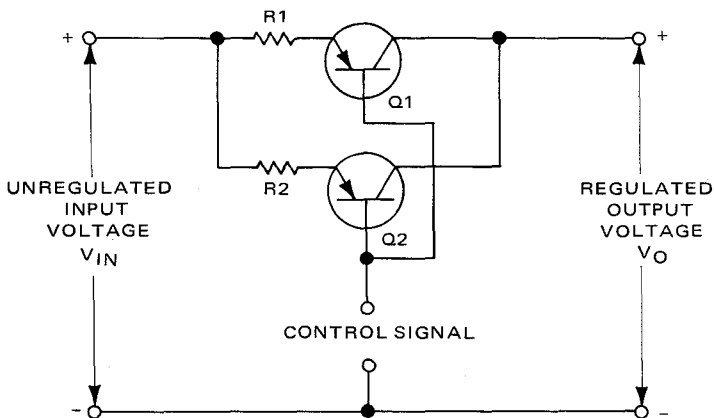
Figure 3-4 — Typical Parallel-Connected Transistors Used in a Regulator

Series-pass regulators can provide extremely good regulation, fast response time, low output impedance and low ripple. Their performance is almost independent of input frequency fluctuations and they have excellent dynamic response by virtue of transient-free output. They also can provide variable output voltage and are readily adaptable to remote voltage sensing, remote programming and current limiting or current regulation.

Switching Regulators

Switching regulators have the distinct advantage over series-pass regulators (especially at high power outputs) of low power dissipation in the main control device. Since the series control element is either saturated or off, it has a low forward voltage drop when conducting load current, and only leakage current flowing when it is not conducting. Thus, transistor power dissipation is small, and this results in high regulating efficiencies. Either transistors or thyristors can be used as the controlled element if appropriate drive circuits are employed. The basic factors limiting the regulator voltage and current are the voltage and current specifications of the controlled power element. The drive requirements and operating characteristics of transistors naturally make them ideal for application in dc input and dc output regulators while the natural application of thyristors is in ac input regulators. Transistor circuits have the advantages of simple circuits, no commutation problems, no dv/dt nor di/dt problems, low saturation voltages, and fast switching speeds with consequent high frequency operation. Thyristor circuits are capable of handling high voltages (to 1500 volts) and high currents (to 400 amperes), require low average driving power at low frequencies and have no safe operating area problems. Switching regulators must be followed by adequate filtering and/or regulation to keep output ripple low. For this reason, switching regulators provide excellent, efficient preregulation to reduce the voltage across and the average power dissipation of series pass regulators.

The important switching transistor parameters are switching speed, pulse safe operating area, high-temperature collector-leakage current, dc gain, $V_{CE(sat)}$ and $V_{BE(sat)}$. The predominate power dissipated in the transistor is that which occurs during the switching time. This power loss is a function of the frequency since the higher the frequency, the higher the proportion of the total period the switching interval becomes. The load line during the transition time determines the safe operating area required of the transistor.

If the circuit elements will force the transistor to exceed its safe operating area, then protection must be provided for reliable operation. The transistor can be protected by connecting a capacitor or zener diode between the collector and emitter terminals of the transistor. If a zener

diode is used, it should have a breakdown voltage greater than the supply voltage but less than the collector-emitter sustaining voltage. If a capacitor is used, its capacitance must not be too high as it may cause excessive charging currents and slow the transistor switching response. Another protective method is to use a free-wheeling diode as shown in Figure 3-5 by D1. For some cases a capacitor can be used in place of the diode, but again it is subject to the conditions above. The voltage requirement resulting from full input voltage being sustained by the switching transistor limits present regulators to about 400 volts at 4 amperes. A typical transistor switching regulator circuit is shown in Figure 3-5. The circuit is connected so that the Schmitt trigger turns Q2 on when the voltage at the arm of R3 falls below the Schmitt triggering voltage. This turns Q1 on and increases the output voltage, and thus the voltage at the arm of R3. When this voltage exceeds the Schmitt triggering voltage, then Q2 and thus Q1 are turned off. While Q1 is off, free-wheeling diode D1 conducts so that current in inductor L1 is preserved. This increases the effectiveness of the filter and prevents excess voltage on Q1. The frequency of operation is a function of the filtering capability of L1 and C1 and the load. The output ripple is a function of the feedback factor from R3 and R2 to the triggering level of the Schmitt trigger.
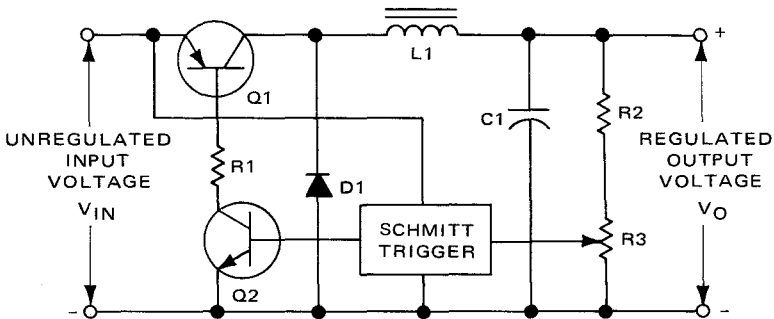
Figure 3-5 — DC-Input Switching Voltage Regulator

Silicon controlled rectifiers are useful as the main control element in phase-controlled regulators. Regulation is achieved through phase control by adjusting the portion of the half sine wave during which conduction occurs. Because this is a fraction of the 60 Hz sine wave, extensive filtering is required to achieve low output ripple. Phase control is highly efficient since the SCRs are either on or off and have a low power dissipation in either condition. Also, low average power is required to drive the SCRs.

SCR regulators are most useful for high current supplies in which the high current would impose severe requirements upon transistors, and for phase-controlled preregulators for series-pass voltage regulators. The SCR turns off every time the current through it goes to zero and is held there for a few microseconds. Therefore a circuit which is sensitive to the line voltage must be used to turn on the SCRs every half cycle (for full-wave control). The function of this circuit is to compute the required conduction angle for each half cycle of input voltage so the output voltage remains constant for any combination of changes in input voltage, output voltage and output current. Since this firing control circuit functions each half cycle, it provides correction for sudden changes in line voltage or load current by altering the timing of the trigger pulse which occurs during the next half cycle of supply voltage. It is desirable to operate the SCR into an inductor to limit the inrush current when the SCR is turned on, since this improves reliability and suppresses electromagnetic interference (EMI). Generally the filter will reduce the EMI on the load; however, a high frequency capacitor across the regulator output may be desirable to bypass any high frequency switching effects on the SCR.

A typical SCR regulating circuit with ac input and dc output is shown in Figure 3-6. SCRs Q3 and Q4 are connected in a bridge to deliver energy to the filter each half cycle of input voltage. Inductor L1 limits the input surge current when either SCR is turned on. A portion of the output voltage (as determined by R4 and R5) is compared to the zener voltage of D6; the difference controls the conduction of Q1, which sets the charging rate of C1 and thus the firing point of Q2 each half cycle. T1 couples the firing pulse from Q2 to the SCRs. D3 and D4 in conjunction with D1 and D2 form a full wave bridge which synchronizes the operation of the firing control circuit to the line. Which SCR comes on is determined by the polarity of the line voltage.

Phase-controlled operation of SCRs is not limited to dc output regulators; it can be used to regulate the rms value of an ac output voltage. A problem is encountered in the method of detecting the rms voltage delivered to the load. This value is difficult to measure due to the irregular waveshapes generally encountered. An irregular waveshape forces some form of integrator to be used to obtain a usable feedback signal. This restricts the response time of the regulator to several cycles of input voltage, limiting its usefulness in compensating for over-voltage and increased load current requirements. Since the line voltage has a uniform waveshape, its rms value is easily detectable and can be used to provide a control signal to regulate the output voltage. However, this does not provide any feedback thereby restricting its usefulness to a narrow range of input voltages.
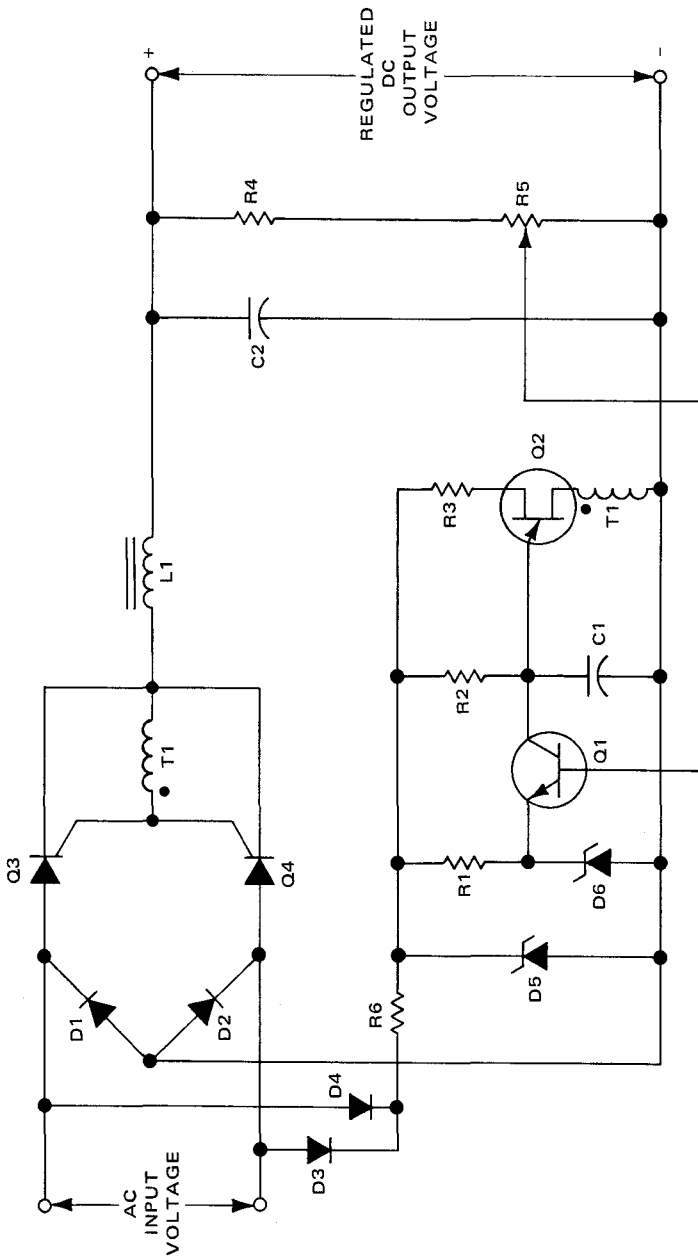
Figure 3-6 — Phase-Controlled DC Voltage Regulator

Current Regulators

The dual of a *voltage* regulated power supply is one that regulates the output *current.* The ideal current source would have an infinite internal impedance and if open-circuited, would produce an infinite voltage. Practically, the devices used as the regulator elements limit the maximum current and voltage that can be delivered to the load. In many instances current regulators are coupled with voltage regulators to limit the maximum current to a load and thus protect the regulator components and the load. Current is regulated indirectly by regulating the voltage developed by the load current across a resistor in series with the load.

For constant output current, the voltage across the resistor must be constant. If the power dissipated in the resistor is sufficient to change the resistance, then current is changed also. The sensitivity of the circuit depends on the gain of the circuits and the size of the resistor. Thus the working voltage can be increased for higher sensitivity at the cost of increased power dissipation, or the gain of the comparison amplifier can be increased with a result of a poorer signal-to-noise ratio. A compromise must be made for any given regulation range.

The circuit shown in Figure 3-7 demonstrates a method of current limiting. Q1, R1, and D1 form a simple voltage regulator. The output current is sensed by R2, which develops a voltage which controls Q2. As the output current increases, Q2 turns on harder and robs base current from Q1. The level to which the output current is limited is equal to the forward base-emitter voltage of Q2 divided by the value of R2.
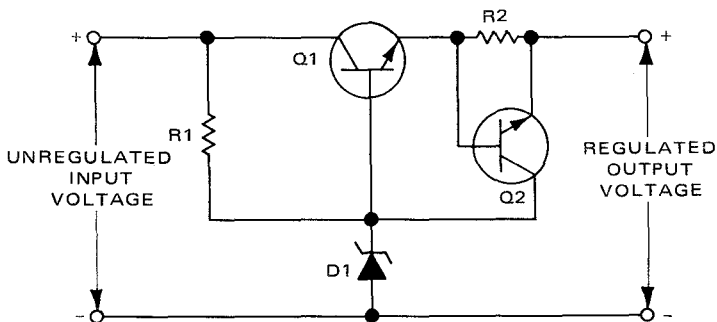
Figure 3-7 — Current-Limited Voltage Regulator

Figure 3-8 shows the circuit of a simple current regulator. Zener diode D1 and resistor R2 set a voltage across R1 equal to the zener voltage minus the base-emitter voltage of Q1. The level at which the output current is regulated is the voltage across R1 divided by the value of R1. Obviously the input voltage must be greater than the zener voltage for regulation to occur.

For low level current regulation, a current-regulating diode can be used as shown in Figure 3-9. This device will maintain a fairly constant load current for an input voltage variation from the knee to the maximum power point. This device is the dual of the zener diode in that it is a regulator for current rather than voltage.
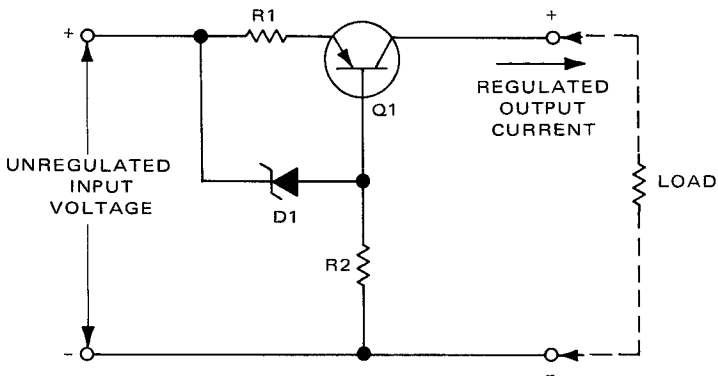
Figure 3-8 — Simple Current Regulator
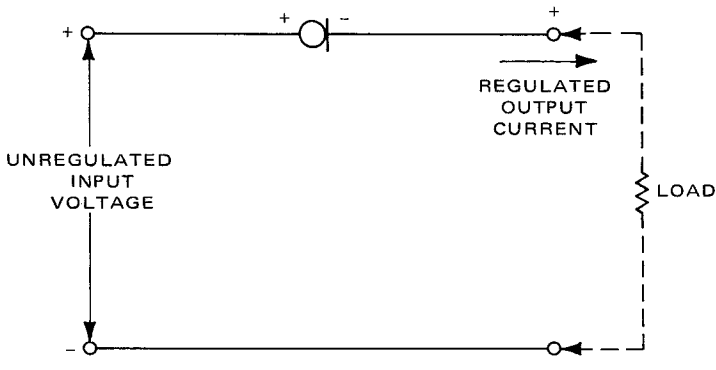
Figure 3-9 — Current Regulator Using Current-Limiting Diode

3.2 11-32 V Output Regulated Power Supply With Overcurrent Protection

The circuit shown in Figure 3-10 is a regulated power supply with overcurrent protection and an output of 11-32 Vdc. The maximum output current with the specified circuit components is 700 mA. The diode bridge of D1 through D4 provides full-wave rectification of the 25 volt secondary of T1. Capacitor C1 is used to reduce the ripple of the rectified current thereby holding a positive voltage on the collectors of Q1 and Q2. Resistor R1 and capacitor C2 further reduce the ripple in order to maintain a pure dc voltage as a collector supply for Q3 and Q4, and for driving the base of Q1. Series-pass transistor Q2 regulates the output voltage. The resistive divider, R4, R5, and R6, is a sensing network. The voltage at the arm of potentiometer R5 is applied to the base of Q4, which compares the voltage to that of zener diode D5. The difference between the two voltages determines the degree of conduction of Q4. If the output voltage increases, then the base voltage of Q4 increases, turning it on more. This reduces the base current of Q1, which in turn reduces the conduction of Q2, thereby lowering the output voltage. If the output voltage drops, Q4 begins to turn off, which turns Q1 and Q2 on, thereby increasing the output voltage. In essence, the circuit is a feedback amplifier which tries to maintain the output voltage at a constant level independent of load condition.

The output voltage is determined by the potentiometer setting according to the following formula:

$$V_{out} = \frac{(V_{D5} + V_{BE} Q4) (R4 + R5 + R6)}{R6 + R5 \text{ (Setting)}}$$

If R5 is set at the low end (A), the output is maximum and is given by

$$V_{out(max)} = \frac{(10.7) (24.2k\Omega)}{8.2k\Omega} = 31.7 \text{ volts.}$$

The minimum output voltage occurs when the arm of R5 is set to the high side (B), or

$$V_{out(min)} = \frac{(10.7) (24.2k\Omega)}{23.2k\Omega} = 11.2 \text{ V.}$$

The regulation is a function of the potentiometer setting since this determines the amount of feedback. The closer the base of Q4 is to the positive output the more feedback and the better the regulation will be.

Regulated Power Supplies
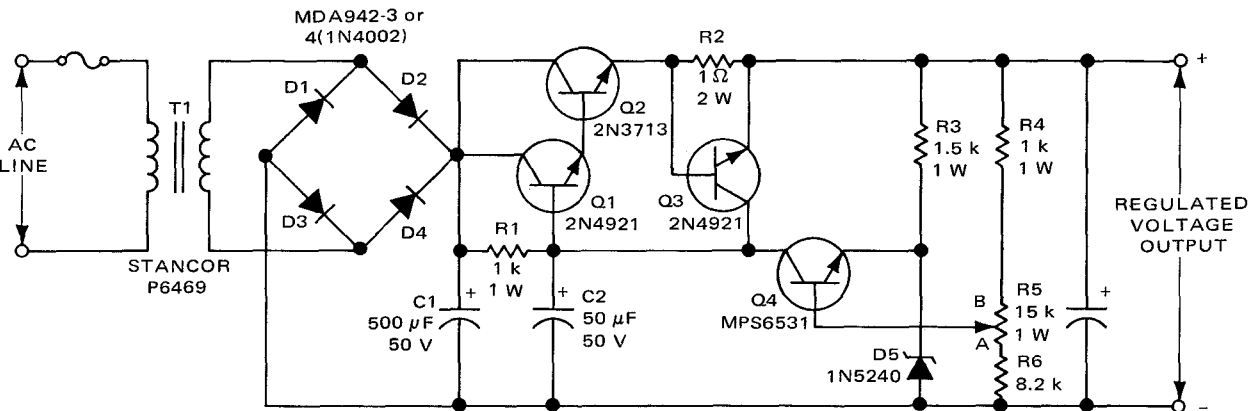
Figure 3-10 — Variable 11-32 Vdc Voltage-Regulated Power Supply with Overcurrent Protection

AC LINE

T1

STANCOR P6469

MDA942-3 or 4(1N4002)

D1 D2
D3 D4

R1
1 k
1 W

C1
500 μF
50 V

C2
50 μF
50 V

Q1
2N4921

Q2
2N3713

R2
1 Ω
2 W

Q3
2N4921

Q4
MPS6531

D5
1N5240

R3
1.5 k
1 W

R4
1 k
1 W

R5
15 k
1 W

R6
8.2 k

B
A

REGULATED VOLTAGE OUTPUT

+
−

The overcurrent protection is provided by R2 and Q3. Since R2 is in series with the output, the voltage across it is proportional to the output current. This voltage is used to drive Q3 so that the larger the output current becomes, the more Q3 turns on. When Q3 comes on, base drive is removed from Q1 which turns off power regulator Q2 thus limiting the output current. The values were chosen to limit the maximum current to approximately 700 mA. The curves shown in Figure 3-11 show the effectiveness of the current limiter.

The load regulation obtained for several output voltages is shown in Figure 3-12. The overcurrent circuit actually degrades the performance of
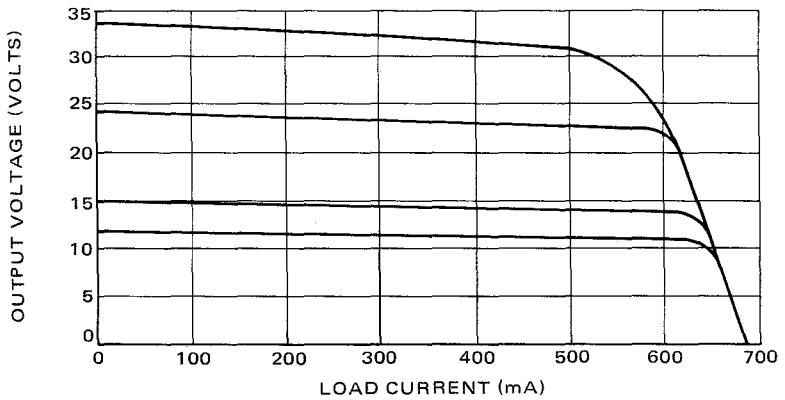
Figure 3-11 — Voltage Output versus Load Current
for Regulator of Figure 3-10
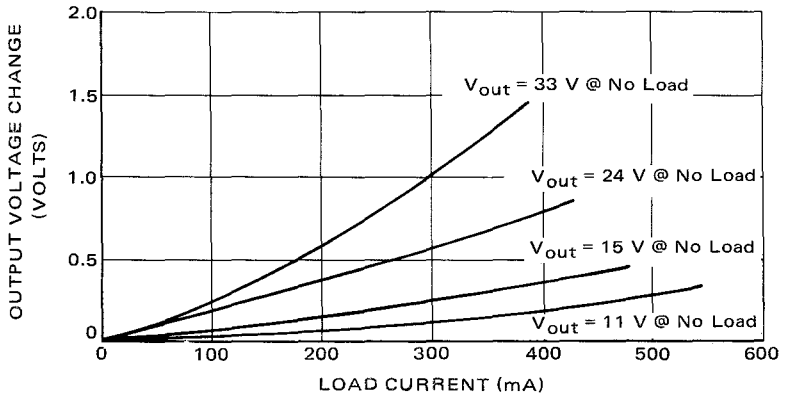
Figure 3-12 — Regulation versus Load Current
for Regulator of Figure 3-10

the regulator circuit, particularly at the higher current levels, since it decreases the base drive of Q1.

The power stage of Q1 and Q2 was selected to have a high gain at -55°C in order to maximixe the control the voltage regulator and current regulator exerts on the power stage. At maximum output voltage and at low temperatures, Q1 requires a base current of about 1 mA. R1 was selected to deliver approximately 2.5 mA. This sets the minimum voltage regulator current in Q4 at 1.5 mA, when the current regulator is not working. When the current regulator, Q3, comes on, it must handle this current. For the opposite condition at high temperatures and low output voltages, R1 provides about 25 mA maximum current. The drive requirement for the power stage is not critical here, but Q3 and Q4 must be able to handle this current for effective control.

The maximum temperature at which the circuit will operate is set by the heat sinking of Q1 and Q2. If these transistors are mounted directly to a heat sink and a good silicone grease is used, then the worst-case temperature rise above the temperature of the heat sink for Q2 is 25°C, while that for Q1 is about 20°C. Since the maximum operating temperature of Q1 is 150°C, this means the heat sink temperature must be kept below 130°C.

3.3 24 Vdc Switching Voltage Regulator

The blocking-oscillator voltage regulator shown in Figure 3-13 produces an output of 24 volts ±1% for currents from 100 mA to 2 amperes. The power switching transistor(Q1) is controlled by the feedback winding of T1 and transistors Q2 and Q4. Initially Q4 turns on, which turns on Q2 and this turns on Q1. This supplies energy to the output pi filter consisting of C3, C4, and L1. The output voltage is fed back to one side of the differential voltage comparator consisting of Q3 and Q4. This voltage is compared to that of zener diode D1, the difference controlling the conduction of Q4, and thus of Q2 and Q1, thereby completing the feedback loop. Q2 was selected to be a high speed switch to work in conjunction with the high-gain voltage-comparator transistors to minimize loading the supply and thus maximize the efficiency. Reverse bias is provided to Q1 by C1 and R2. This helps turn Q1 off once it has started to go off due to T1 becoming saturated and thus reversing the base drive voltage. The fall time of the collector current for Q1 is approximately 1/2 microsecond. Diode D3 is a free-wheeling diode which allows current to continue through the load when Q1 is off. D3 must be a fast-recovery diode to minimize the reverse current through it, which occurs while it is turning off when Q1 is again turned on. This increases the efficiency of the
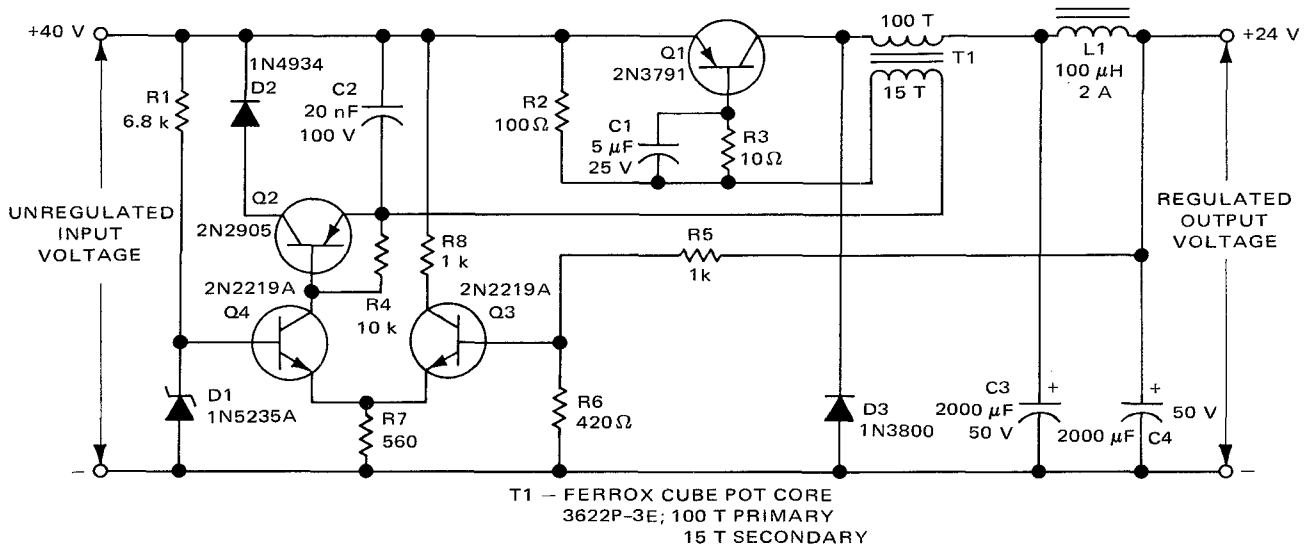
Figure 3-13 — 24 Vdc Switching Voltage Regulator

Regulated Power Supplies

+40 V

R1
6.8 k

1N4934
D2

C2
20 nF
100 V

R2
100 Ω

Q1
2N3791

100 T

T1

15 T

L1
100 μH
2 A

+24 V

UNREGULATED
INPUT
VOLTAGE

C1
5 μF
25 V

R3
10 Ω

REGULATED
OUTPUT
VOLTAGE

Q2
2N2905

R8
1 k

R5
1k

2N2219A
Q4

R4
10 k

2N2219A
Q3

D1
1N5235A

R7
560

R6
420 Ω

D3
1N3800

C3
2000 μF
50 V

2000 μF C4

50 V

T1 — FERROX CUBE POT CORE
3622P-3E; 100 T PRIMARY
15 T SECONDARY

circuit since load current is maintained by C3 and L1 in addition to C4, when Q1 is both on and off. The maximum voltage that Q1 must withstand is the supply voltage plus the forward voltage drop of D3. As the output voltage is decreased, the feedback voltage available from T1 is increased; this can cause the peak current through Q1 to increase for a given supply voltage. This increased drive to Q1 can be reduced by allowing Q2 to come out of saturation where the high drive conditions exist. If changes are made in the circuit values, then these conditions as well as the load-line stress on all transistors must be considered for reliable operation.

The maximum frequency of the blocking oscillator is limited by the transformer used; it is 6 kHz for this circuit at the higher input voltages. The peak collector current of Q1 is 5 to 7 A depending on the input voltage and transistor gain. The safe operating area of Q1 must be able to withstand this stress. For high efficiency the switching time and the saturation voltage should be minimized. A 2N3791 was selected for Q1, since its switching time is less than 500 μs, $V_{CE(sat)}$ is about 1 V, and it is fairly inexpensive.

The performance curves of Figure 3-14 show that the 24 volt output is regulated within ±1% for a load current range of 100 mA to 2 A and for supply voltage changes of ±10%. The efficiency as shown in Figure 3-15 is approximately 85% for currents in excess of 1 ampere at the nominal input voltage (40 volts) and for a 10% supply voltage change at 2 amperes of load current. Additional gain in the voltage sensor may improve the
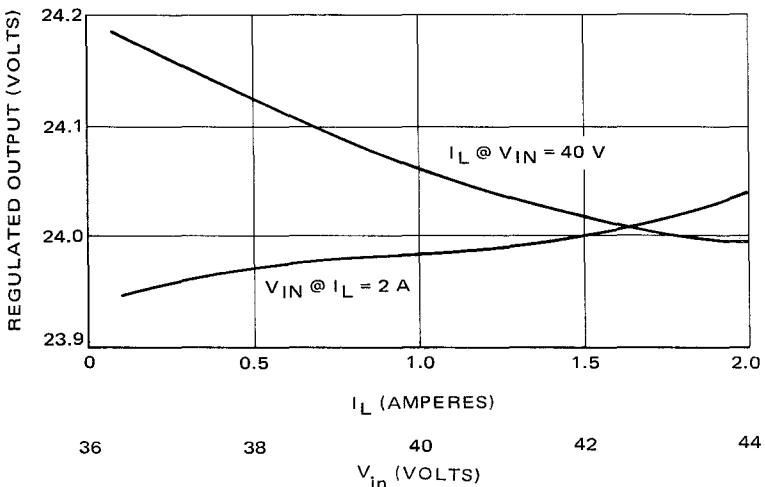
Figure 3-14 — Output Voltage Regulation for Switching Regulator of Figure 3-13
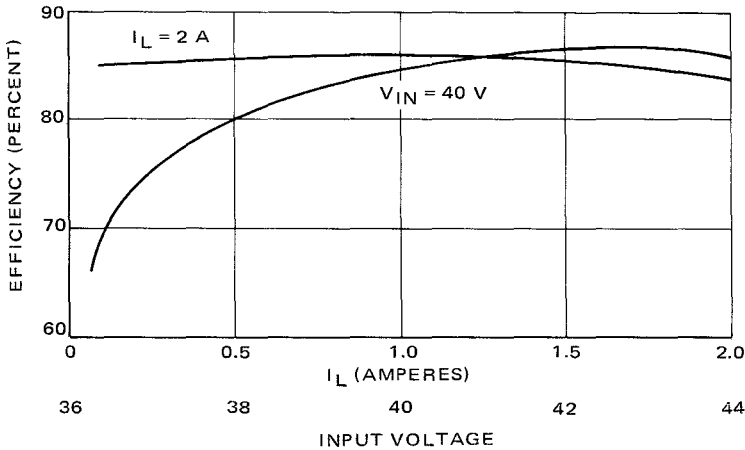
Figure 3-15 — Efficiency for Switching Voltage
Regulator of Figure 3-13

regulation, and a lower dc resistance in the transformer should increase the efficiency. If any changes are made, enough current must be supplied to the base of Q1 to start the circuit oscillating.

3.4 20 Vdc Switching Voltage Regulator With Inverted Output

The inverted-output switching regulator shown in Figure 3-16 provides a 20 volt output that is regulated to within ±1% over a load current range of 50 mA to 1 amperes. The basic circuit of the power section is a blocking oscillator formed by Q1 and T1; its drive is obtained by feedback from the output. Initially Q1 is turned on from current through Q3, R4, the base-emitter junction of Q2, T1, and R1. As collector current is drawn through the primary of T1, more base drive is provided by the secondary through Q2, thus turning Q1 on more and driving it into saturation. When T1 saturates, its secondary voltage reverses and removes base drive from Q1; this begins to turn Q1 off. The energy stored in C1 helps to turn Q1 off. While Q1 is off, the output pi filter is charged through D1 from the energy stored in T1. For this reason the voltage at the filter output is greater than that at the positive input terminal. Thus the output voltage is inverted from the input. Resistors R1, R2 and capacitor C1 provide reverse bias to Q1 while Q1 is off. This is necessary in order to overcome the forward bias current from Q3 through R4, the emitter-base of Q2 and the feedback winding of T1. During the time the collector

3-17
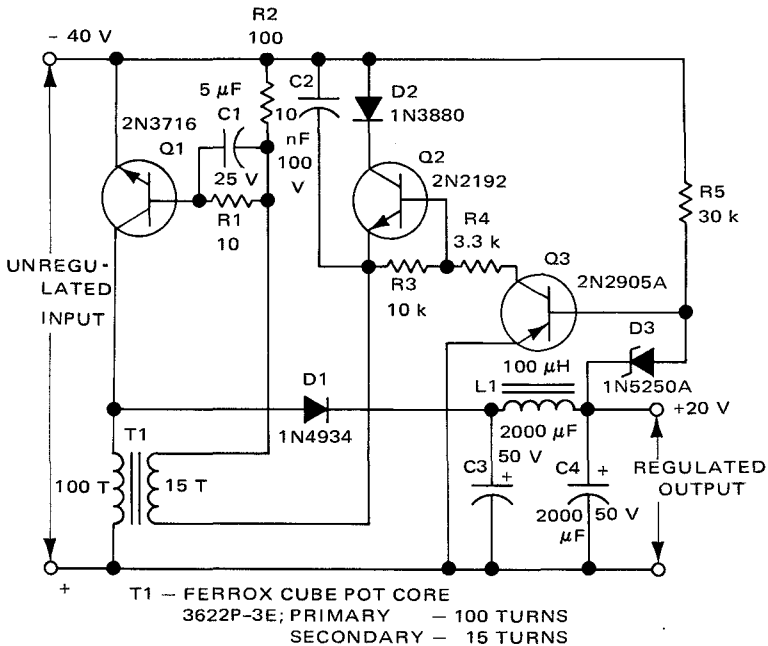
Figure 3-16 — Switching Voltage Regulator with Inverted 20 Vdc Output

current of Q1 is falling (approximately 0.5 microsecond), capacitor C2 provides a low impedance path for turn-off bias from the feedback winding.

The minimum transistor off time depends upon the decay time of the current from the transformer primary to the output filter. The current must decay to a low level before the feedback circuit will turn Q1 on.

For a given input and output voltage, and peak transistor current, the primary inductance of T1 determines the maximum frequency of the blocking oscillator. For a given filter, the higher the frequency, the lower the output ripple voltage will be. Therefore, high frequency of operation and low inductance is desirable. However, high frequencies result in high transistor power dissipation and high core losses in transformers. These factors must be considered in the design. Transformer design set the operating frequency of this circuit at 6 kHz with a peak collector current in Q1 of 5 to 7 amperes depending on transistor gain and input voltage levels. The collector-to-emitter voltage of Q1 is the sum of the supply voltage and the output voltage, so Q1 must be capable of withstanding this stress level.

In addition to this, for maximum efficiency Q1 must have a low

saturation voltage and fast switching times. A 2N3716 is used as it is inexpensive and satisfies these requirements. A 2N2192 was chosen for Q2 since it is a high gain, high current switch; this is necessary since it must handle the base current of Q1, which could be about 500 mA. To minimize loading of the supply and maximize efficiency, Q3 should be a high
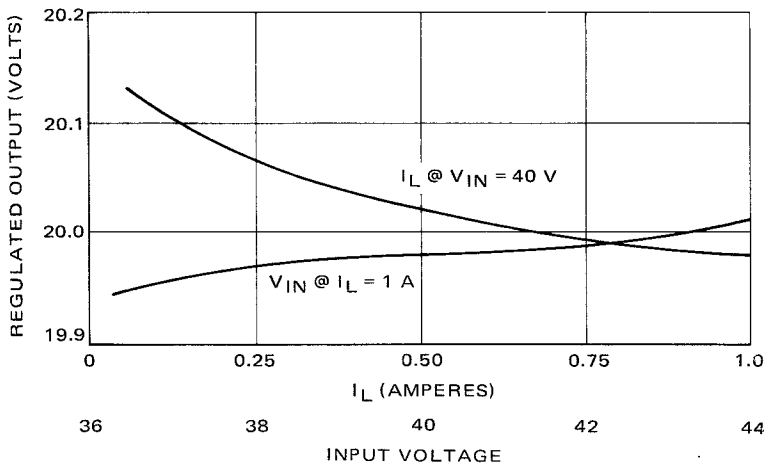
Figure 3-17 — Output Voltage Regulation for Inverted Output
Switching Regulator of Figure 3-16
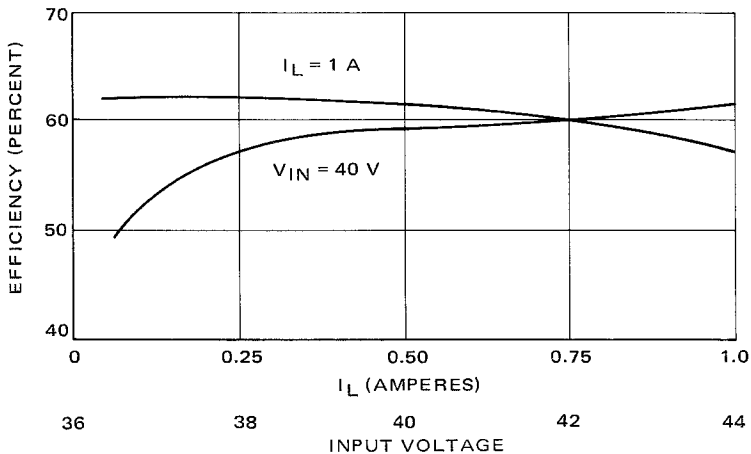
Figure 3-18 — Efficiency for Inverted Output Switching
Regulator of Figure 3-16

gain, high speed switch. This requirement is satisfied through the use of a 2N2905A.

The performance curves of Figure 3-17 show the regulation is ±1% for a ±10% input voltage change at output currents up to 1 ampere. The efficiency of the circuit versus both input voltage and load current is shown in Figure 3-18. Additional gain in the voltage sensing circuit will improve the regulation. However, if any changes are made, enough current must be supplied to the base of Q1 to start the circuit oscillating. The efficiency may be improved by using a transformer with a lower dc resistance.

3.5 Pulse-Width-Modulated 38 Vdc, 5 A Switching Pre-Regulator

The pulse-width-modulated 38 Vdc, 5 ampere switching voltage regulator whose circuit is shown in Figure 3-19 utilizes the voltage sensing property of a Schmitt trigger to control the output voltage of the regulator. The Schmitt trigger is formed by transistors Q3 and Q4; Q4 is the voltage sensor and Q3 is used to provide drive for the Darlington series switch, Q1 and Q2. To start the circuit oscillating, zener diode D4 receives enough current to bias it above its knee through resistor R10. Diode D5 is used to block the load from robbing current from D4. After oscillation begins, D5 becomes forward biased and current is supplied through R9. The resistive divider, R11, R14, and R12, is used to set the dc voltage level at the output. Since the switching regulator is basically an ac circuit in that Q4 responds to changes in the output voltage, C3 is used to provde an ac bypass around the divider so that the output voltage ripple is fed to the base of Q4 with little attenuation. The level at which Q4 is turned on is governed by the difference between the voltage at the arm of R14 and the voltage across R5. The voltage on R5 is set by the zener voltage of D4, the divider consisting of R8, R7, R6, the base-emitter junction of Q3, and the small drop across R4. When the output voltage increases and the upper trigger level of the Schmitt is exceeded, Q4 turns on, which removes base drive from Q3, turning it, Q2 and Q1 off, thus removing output drive. These devices remain in this condition until the output voltage decreases and the base voltage of Q4 falls below the lower trigger level, which turns Q4 off, permitting Q3 to turn on. This provides base drive to Q2 and Q1 through R3, thus turning Q1 on. Since Q1 cannot be saturated, because of the Darlington connection, resistor R13 is used to reduce the voltage across Q1 and thereby its power dissipation. This minimizes the requirements of the heat sink for Q1. While Q1 is on, energy is stored in L1 and C2, and the output voltage is increased. When Q1 is off, free-wheeling

Figure 3-19 — Pulse-Width-Modulated Switching Preregulator

Regulated Power Supplies
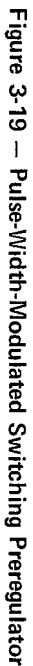
diode D3 conducts, permitting load current to continue through L1. This current is provided by the energy stored in L1 and C2. Use of the free-wheeling diode accomplishes two things: (1) it increases the efficiency of the regulator by providing a path for L1 to discharge its energy into the load and (2) it helps to keep Q1 in its safe operating area since voltage spikes on the collector are eliminated. There is still the possibility of exceeding safe operating area, however, since D3 is on when Q1 comes on. This means that Q1 will be subjected to a momentary short circuit until D3 turns off. The use of a fast-recovery diode for D3 minimizes this
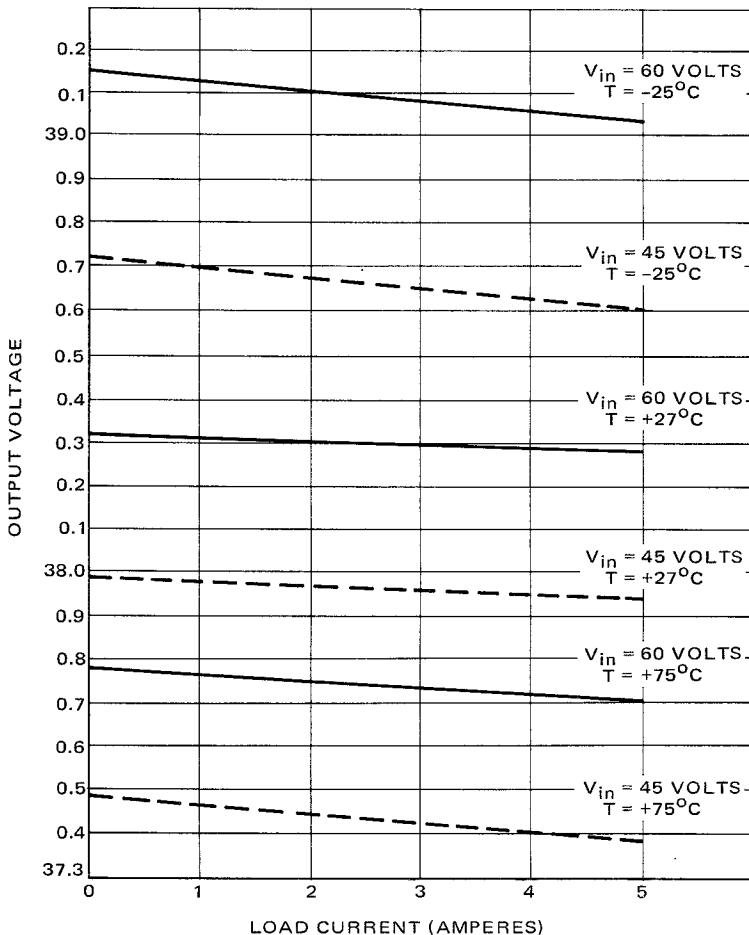
Figure 3-20 — Voltage Regulation of Preregulator of Figure 3-19

problem since turn-off times for the device chosen is less than 200 nanoseconds.

The frequency at which the circuit operates is governed by the filtering capability of L1 and C2 and the hysteresis of the Schmitt trigger. It is found that the frequency is also dependent on the resistive and inductive parameters of the capacitor used for C2. Because of this, some trimming may be required to achieve the desired operating frequency. Once the frequency is set it will remain almost constant from no load to full load. The electrolytic capacitor used for C2 in the test circuit gave an operating frequency of approximately 16.5 kHz.

The circuit was designed to deliver 5 amperes at 38 volts. The regulating characteristics are shown in Figure 3-20; from no load to full load the output voltage dropped about 150 millivolts. The output ripple voltage remained almost constant over this range at approximately 270 mV peak to peak, and the frequency varied from about 19 kHz with 60 V input to about 10 kHz with 45 V input.

3.6 Current Regulators

The circuits shown in Figure 3-21 through 3-24 can be used to regulate the current in a load connected as shown on the schematics. The circuits of Figures 3-21 through 3-23 are three-terminal current regulators and the only difference between these three circuits is the biasing arrangement for the control transistor. Figure 3-24 shows a two-terminal current regulator.

For the circuit shown in Figure 3-21, diodes D1 and D2 set the base potential of Q1. The base-emitter voltage of Q1 is approximately equal to the voltage drop across D1, so the voltage across R2 is the voltage drop of diode D2. The current level at which the regulator will regulate, therefore, is the voltage drop of D2 (0.7 volt) divided by the value of R2. R2 is 10 ohms in this circuit for a regulated current level of 70 mA. The regulation is somewhat dependent on the value of R1 since this controls the biasing current level in D1 and D2. If R1 is too large the diode is working on the knee of the curve and the circuit operation will be very dependent upon the supply voltage. If R1 is too small, excessive power is dissipated in the circuit elements. The value shown for R1 works adequately as can be seen by the results shown in Figure 3-25 and 3-26.

For the circuit shown in Figure 3-22, the voltage drop of D1 compensates for the base-emitter voltage of Q1; thus the current regulated is equal to the breakdown voltage of zener diode D2 divided by the value of R2. To obtain the 70 mA desired, a 5.6 volt zener is selected and R2 is
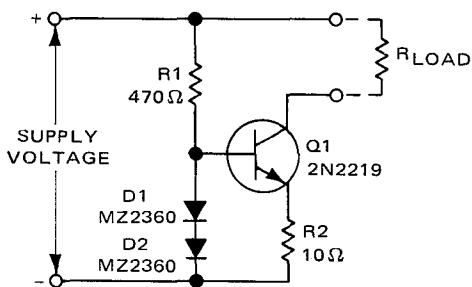
Figure 3-21 — 70 mAdc Current Regulator
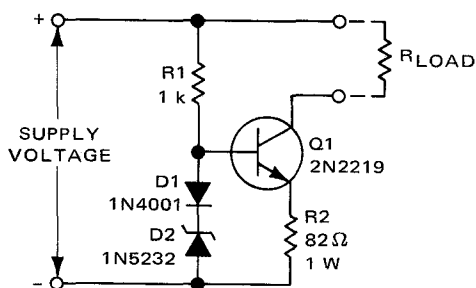
Figure 3-22 — 70 mAdc Current Regulator
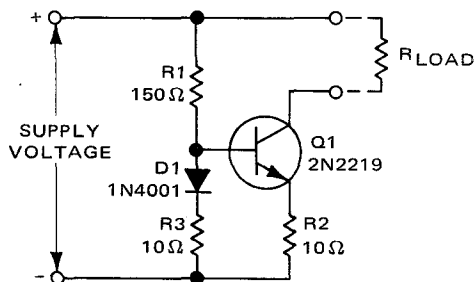
Figure 3-23 — 70 mAdc Current Regulator
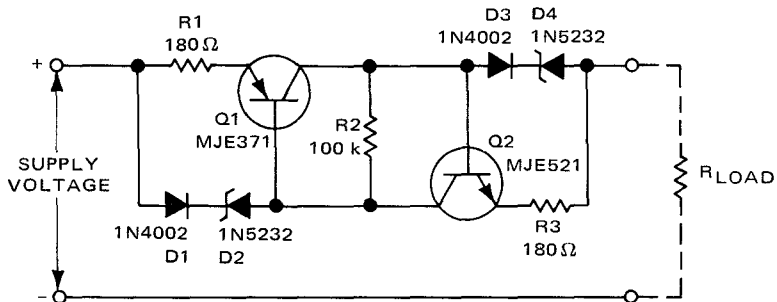
Figure 3-24 — Two-Terminal 70 mAdc Current Regulator
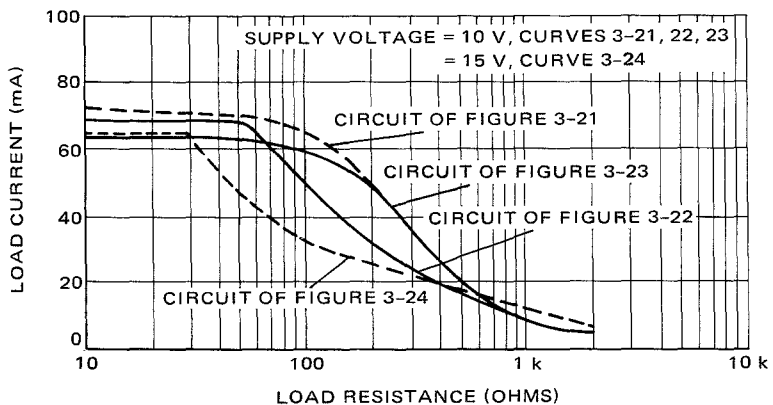
Figure 3-25 — Current-Regulator Load Current versus Load Resistance
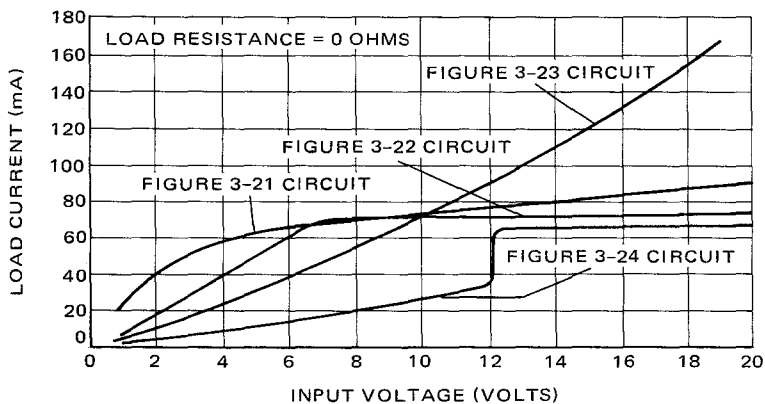
Figure 3-26 — Current-Regulator Load Current versus Input Voltage

set at 82 ohms. The value of R1 for this circuit is not as critical as it was for the circuit in Figure 3-21, since the zener breakdown voltage of D2 is less sensitive to current changes than the forward voltage of the diodes in Figure 3-21, and the current through Q1 is established primarily by the voltage across D2. The 5.6 volt zener diode (D2) was selected since its temperature coefficient is approximately zero. The variation of the base-emitter voltage of Q1 with temperature is compensated by the temperature variation of D1 so that the circuit should exhibit good regulation with temperature. The curves of Figures 3-25 and 3-26 show the results obtained with this circuit for variations in load and input voltage. Figure 3-26 shows that the output current is regulated if the supply voltage is above the knee of the combined curves of D1 and D2.

The circuit shown in Figure 3-23 provides very good current regulation if the supply voltage remains constant and the load is varied. Diode D1 provides compensation for the base-emitter voltage of Q1 and the voltage across R2 equals that across R3. The level at which the current will be regulated is therefore, the voltage across R2 divided by the resistance of R2. If R2 is equal to R3, and R1 is large compared to R2 and R3, the current through R1 is equal to the regulated output current. The load and input-voltage regulation obtained with this circuit is shown in Figures 3-25 and 3-26. This circuit regulates the output current up to the point where the load resistance becomes too large, thereby forcing Q1 into saturation. This prevents further regulation of the load current.

If a three-terminal current regulator cannot be used, then the two terminal regulator shown in Figure 3-24 may be useful. For this circuit to regulate, the supply voltage must be greater than the sum of the voltage drops of D1, D2, D3, and D4. For the components shown, this voltage is 12.5 volts. The level at which the current will be regulated is this voltage divided by the resistance of R1 or R3 since R1 is equal to R3. The curve shown in Figure 3-25 shows the regulation for a supply voltage of 15 volts. The regulation can be increased to higher load resistances if the supply voltage is raised. The maximum load resistance at which regulation can be maintained is shown by the following equation.

$$R_{max} = \frac{V_{supply} - (V_{D1} + V_{D2} + V_{D3} + V_{D4})}{I_{Regulated}}$$

The curve shown in Figure 3-26 demonstrates that the supply voltage must be above the sum of the diode voltage drops for effective regulation. As the regulated current level is increased, the forward-voltage drops of D1 and D3 increase slightly. This circuit has been operated at a regulated current level of 700 mA; for this, the supply voltage must be above 13.5 volts.

3.7 100 V rms Voltage Regulator

The circuit shown in Figure 3-27 will regulate the rms output voltage across the load (which is a projection lamp) to 100 volts ± 2% for an input voltage between 105 and 250 volts ac. This is accomplished by sensing the light output of L1 indirectly and applying this feedback signal to the firing circuit (Q1 and Q2) which controls the conduction angle of triac Q3. The load is a 150 watt projection lamp which has a reflector mirror included inside the glass envelope. If the light output of the lamp were sensed directly by the photocell, it would respond to the 60-Hz variation of the supply voltage unless additional filter components were used. Therefore, another approach was used to generate the feedback signal. The reflector inside the lamp's envelope glows red due to the heat of the filament. Since the reflector has a relatively large mass it cannot respond to the supply frequency, and its light output provides a form of integration. This light is then used as a feedback signal. To eliminate 60 Hz modulation of the photocell, it is mounted at one end of a black tube with the other end of the tube directed at the back side of the reflector in the lamp. The lamp is energized through triac Q3, whose conduction angle is set by the firing circuit for unijunction transistor Q2. This circuit is synchronized with the line through the full-wave bridge rectifier. The voltage to the circuit is limited by zener diode D5. Phase control of the supply voltage is set by the charging rate of capacitor C1. Q2 will fire when the voltage on C1

L1 — 150 WATT PROJECTION LAMP WITH
BUILT-IN REFLECTOR MIRROR
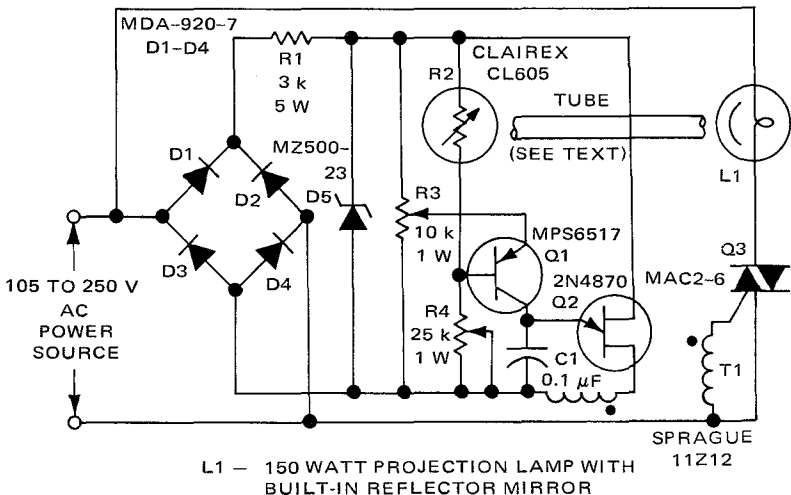
Figure 3-27 — 100 Vrms Voltage Regulator

reaches approximately 0.65 times the zener voltage. The charging rate of C1 is set by the conduction of Q1, which is controlled by the resistance of photocell R2. Potentiometers R3 and R4 are used to set the lamp voltage to 100 volts when the line voltage is 105 volts and 250 volts, respectively. This assures that the lamp voltage will be within the desired tolerance over the operating range of input voltage. Some interaction will occur between R3 and R4 and the adjustment of each potentiometer may have to be made several times. Since this is an rms voltage regulator, a true rms meter must be used to adjust the load voltage.

If the excursion of the supply voltage is not large, then a circuit without feedback such as shown in Figure 3-28 may be used. This circuit regulates the input voltage to the step-up transformer T2 at 90 volts, thereby maintaining the load voltage at 120 volts ±2%. Q2 and Q3 are both pulsed at the same time by Q1 through T1, but the only one that turns on is the one that is forward biased by the supply voltage. The circuit which controls the phase of Q1 is synchronized to the line voltage through full-wave bridge D1 to D4. The voltage to the charging circuit of capacitor C1 is regulated by zener diode D5. The charging circuit is a ramp-and-pedestal combination and the voltage on C1 is as shown in Figure 3-29. R2 and R5 have low values compared to R6 so that C1 can
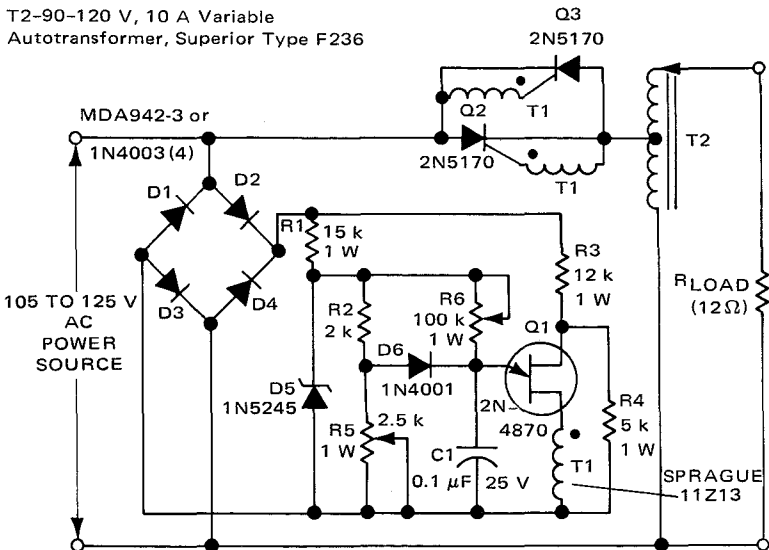
Figure 3-28 — 105 to 125 Vac Voltage Regulator

charge quickly to the level set by R5. Diode D6 then becomes reverse biased, and C1 continues to charge through R6, but at a slower rate, until the breakover voltage of Q1 is reached. Since the interbase voltage of Q1 is proportional to the line voltage, the value to which C1 charges is variable. When the supply voltage is low, R5 is adjusted to provide 120 volts to the load. When the supply voltage is high, R6 is adjusted to provide 120 volts to the load. R5 is set so the firing point of Q1 is close to the voltage at which D6 becomes reverse biased. R6 then adjusts the ramp to delay the firing point as the line voltage is increased. R5 and R6 interact, so the adjustment will have to be made several times. As in the previous circuit, a true rms meter must be used for this adjustment.

The operation of this circuit depends upon the waveshape of the supply voltage. Therefore, the ac power source must provide a low-distortion sine wave to assure good rms voltage regulation at the output. Since there is no feedback in this circuit the regulation at the load depends upon the regulation of transformer T2. As the load is changed some adjustment of R5 and R6 may be required to maintain voltage regulation at the load. This circuit was used to regulate a 10 ampere load and provided a load voltage change of only 2 volts for a line voltage change from 105 to 125 volts using an adjustable transformer for T2.
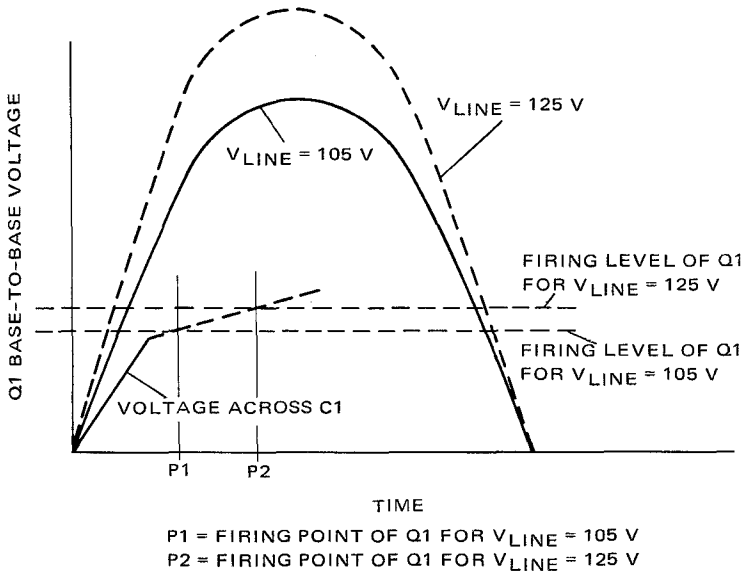
Figure 3-29 — Firing-Point Change of Q1 for Line-Voltage Change

3.8 Short-Circuit Protection for Series-Pass Regulator

Power supplies with series-pass regulators can be adequately protected against damage due to short-circuited loads through the simple addition of a few inexpensive components. The circuit shown in Figure 3-30 not only protects the regulator against a shorted output but incorporates an automatic reset circuit that restores normal operation when the short is removed.

The series-pass portion of the circuit is conventional in operation and is designed to present very low impedance to the load.

The part of the circuit which provides protection against short circuits is shaded in Figure 3-30. It operates as follows: When a short circuit occurs across the output the whole supply voltage appears across D2, R2, and the emitter-base junction of Q2. Since the input supply voltage is greater than the breakdown voltage (V_Z) of zener diode D2, the diode breaks down and supplies base drive to Q2. As Q2 approaches saturation, it robs the series-pass transistor (Q1) of base drive and cuts it off. Capacitor C1, which is charged up to the output voltage before a short appears across the load, aids in turning on transistor Q2 by discharging through the base of Q2 when the output is shorted. Although capacitor C1 discharges to a very low voltage when the regulator is in the short circuited mode, transistor Q2 remains saturated due to the base drive provided through D2 and R2.

The automatic reset incorporated in this circuit operates in the following manner: When the short is removed, the voltage across the output begins to rise due to the current supplied through Q2. Since the
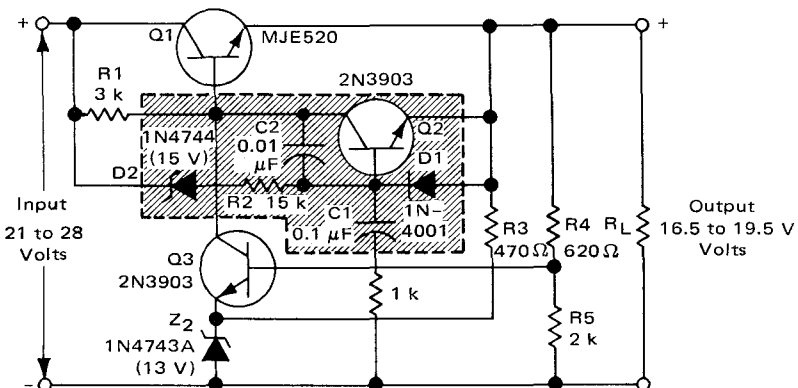
Figure 3-30 — Circuit Protection for Series-Pass Voltage Regulator

voltage differential across the series-pass transistor (Q1) determines the base drive of Q2, and this voltage has decreased due to the rise of voltage across the output, Q2 begins to turn off, allowing Q1 to turn on.

When the voltage across the output exceeds the voltage across C1, this capacitor begins to charge through D1. The voltage thus dropped across D1 back-biases the emitter-base diode of Q2, turning this transistor off. When Q2 turns off, normal base current is supplied to the base of Q1 and the circuit resumes normal operation. Capacitor C2 enhances the ac stability of Q2 and may not be necessary in all circuits.

Under normal series-pass operating conditions, the protective circuit is biased off and does not affect the circuit performance in any way.

The protective circuit is limited in speed only by the storage and fall time of the series-pass transistor (Q1), and operates in approximately one millisecond.

BIBLIOGRAPHY FOR CHAPTER 3

1. *DC Power Supply Handbook* – Hewlett Packard Application Note Number 90.

2. *Power Supply Handbook* – Kepco, Inc., Flushing, N.Y. 11352.

3. QS Series Power Supplies, *Technical Manual,* Behlman-Invar Electronics Corp., Santa Monica, Calif.

4. "Avoiding Second Breakdown", Motorola Application Note Number AN-415.

5. Wechsler, Reuben, "Designing SCR Circuits for High Inrush Loads", Motorola Application Note Number AN-170.

6. Wechsler, Reuben, "Reducing (di/dt)-Effect Failures in Silicon Controlled Rectifiers", Motorola Application Note Number AN-173.

CHAPTER 4

Static Swiching

4.1 Solid State Static Relays

There is no exact, universal electronic replacement for the mechanical relay. Semiconductors can be used to duplicate the function of a relay but must be tailored to fit a particular application.

Either transistors or thyristors can be used as switches to isolate a load from, or connect it to, a power source. The control of either device can be completely electronic, in which case the circuit is called a static switch, or through the use of a switch with small current capability, which forms a static contactor. A transistor requires continuous drive power during the time it is conducting. A thyristor requires only a pulse of power to turn it on but it must be commutated off either forceably or by natural commutation of an alternating voltage source. Either type of device can be used on ac or dc, but the circuit used must be designed for the power source used.

AC Power Source

Semiconductor devices which are basically dc devices, such as transistors and SCRs, require special circuits when they are used on ac; these special circuits are not required for devices such as triacs, which will pass current in both directions. Figures 4-1 through 4-3 show circuits which use a transistor, a silicon controlled rectifier, and a triac as the switching element. The circuit of Figure 4-1 maintains ac to the load and at the same time provides dc to the transistor, which is the controlled element, through the use of a bridge rectifier. The transistor must be able to withstand the peak voltage of the power source when it is off and it must be capable of handling the peak load current when it is on. It also must have adequate safe operating area for the switching load line. Since the transistor requires continuous drive power to keep it on, it will begin to turn off immediately upon removal of the driving signal. This could create large voltage or current spikes in the circuit if reactive loads are used. If inductive loads are to be switched, then a circuit which will protect the transistor from over-voltage must be included. If high voltages or large currents
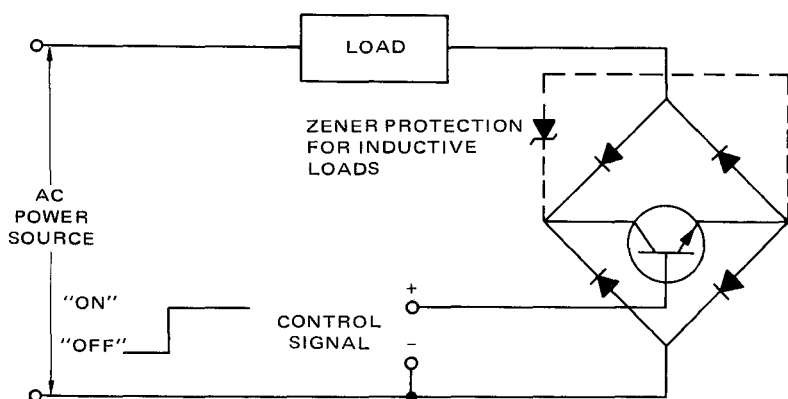
Figure 4-1 — AC Solid-State Static Relay Using
Transistor Control Element
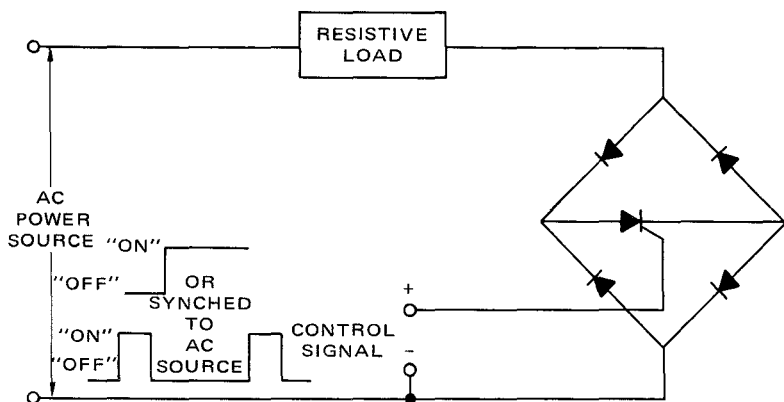
Figure 4-2 — AC Solid-State Static Relay Using SCR Control Element
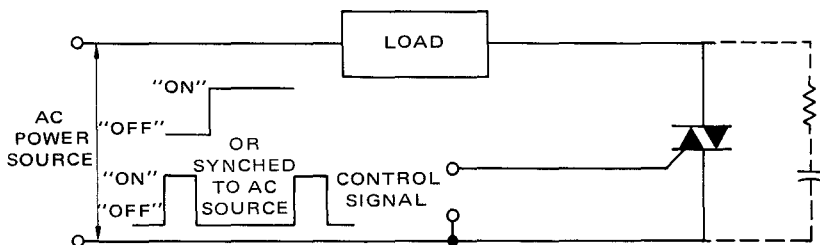
Figure 4-3 — AC Solid-State Static Relay Using Triac Control Element

are to be switched, the transistors that can be used are usually expensive and the driving power required can get quite high. For these reasons the circuit in Figure 4-1 is not recommended for switching high voltages and/ or high currents.

For resistive loads, a circuit which performs much like that in Figure 4-1 is shown in Figure 4-2. In this case the transistor has been replaced by an SCR. The SCR does not require large drive power; it will turn on as soon as the gate drive is provided and will turn off each half cycle when the current goes to zero. Therefore, to keep the SCR on, either continuous gate drive must be provided or a gate driving power pulse must be delivered each half cycle. The turn-off time of the SCR limits the maximum frequency of present SCR switching circuits to approximately 30 kHz. If the supply frequency is around 60 Hz or less, a triac can be used as the control element as shown in Figure 4-3. It will conduct (or control) current in both directions. Unlike the SCR, the triac can be triggered by either positive or negative gate drive, but more power is required for negative drive than for positive drive. Since the triac also turns off every time the current goes to zero, the drive requirements for a triac used on ac are the same as described for the SCR. The series R-C network across the triac is required for inductive loads to prevent the rate of voltage rise (dv/dt) developed when the triac turns off from immediately returning it to the on state.

DC Power Source

Either thyristors or transistors can be used as the controlled switching element with a dc supply if the circuits are adapted to satisfy each device's particular characteristics. Transistors play an important role here since the continuous drive required to keep them on is readily available and easily obtained. Special circuit adaptation is required for thyristors since these devices remain on once turned on. Thus, another circuit which will force the current in the device to drop to zero is required to turn the thyristor off.

Figure 4-4 shows a typical connection for using a transistor as a switch. The transistor must be able to withstand the supply voltage when off and be able to handle the load current when on. Also, it must have a safe operating area adequate to handle the switching load line. The amount of drive power required is an inverse function of the gain of the device and a direct function of the load current. Therefore, the drive requirements are dependent on the circuit parameters. If the load is inductive then the transistor must be protected from voltage spikes created by the switching

action. One method for achieving this is to place a diode across the load as shown in Figure 4-4. If high voltage supplies are to be used or large currents are to be switched this circuit is not recommended since an expensive transistor and large driving power are required.

The thyristor circuit shown in Figure 4-5 uses silicon controlled rectifiers as the controlled switching elements. A power pulse applied to the gate of SCR Q1 will turn the device on, connecting the "cold terminal" of the load and the commutating capacitor to the negative side of the supply. The capacitor then charges to the supply voltage minus the SCR drop. Q1 will stay on until a power pulse is applied to the gate of Q2. This turns Q2 on and applies a reverse voltage to Q1 via capacitor C1. The capacitor then discharges through Q1 forcing the current to drop below the holding current, turning Q1 off and disconnecting the load from the power source. Once Q2 comes on it will remain on until the control signal pulse is again applied to the gate of Q1. At this time the reverse of the
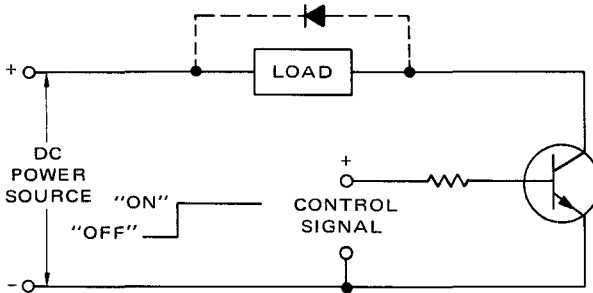
Figure 4-4 — DC Solid-State Static Switch Using
Transistor Control Element
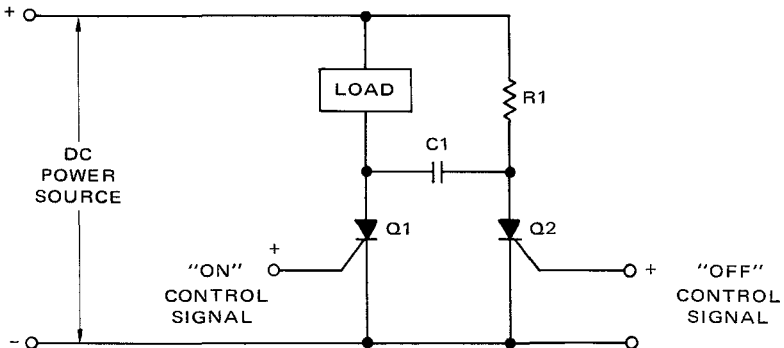
Figure 4-5 — DC Solid-State Static Switch Using SCRs
as the Control Element

commutating action just described occurs and Q2 is turned off while Q1 is turned on. If the duty cycle of the control signal is low, Q2 can be smaller than Q1 since Q2 must be able to withstand a nonrepetitive current surge equal to the peak commutation current plus the current through R1, rather than the continuous load current. This circuit gives excellent results when switching with high voltage supplies and/or with large load currents.

4.2 Triac Static Motor-Starting Switch for 1/2 hp, 115 Vac Motor

Single-phase induction motors require a starting winding which is used only until the rotor obtains a speed of approximately 75% of full-load speed. The circuit shown in Figure 4-6 can be used to control the starting winding of such a motor; essentially a triac replaces the centrifugal switch normally used for this purpose.
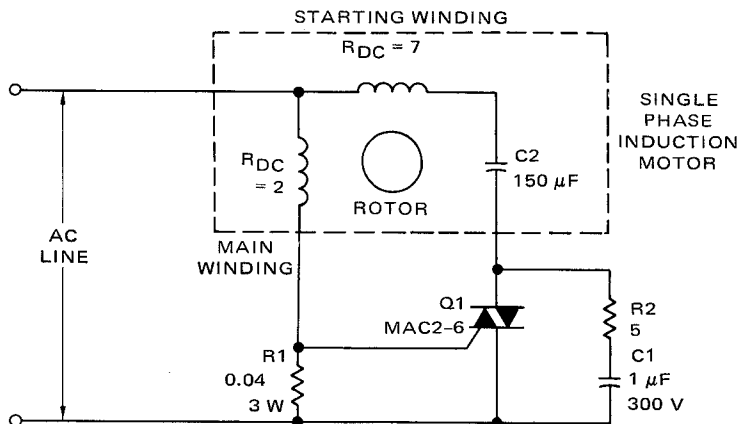
Figure 4-6 — Triac Static Motor-Starting Switch for
½ Hp 115 Vac Single-Phase Induction Motor

Since the inrush current during starting in the main winding of the motor is several times the running current, this inrush current can be used as the source of control for the starting winding. The voltage developed across resistor R1 is used to gate on triac Q1 when the voltage exceeds its threshold level. The resistance must be large enough to develop sufficient voltage to turn Q1 on, but small enough that the normal peak running current will not develop enough voltage to turn it on. Some selection of this value may be necessary to accommodate any triac that may be used; a slide-wire resistor may be used to select the proper trigger level.

When the current drops to zero in the triac, the device turns off and the voltage across it will increase rapidly to the value of the line voltage since there is phase shift due to the inductive load. If the rate of rise of this voltage (dv/dt) is too great, the triac will again turn on. Therefore, a circuit must be provided to assure the dv/dt is low enough. The network consisting of R2 and C1 is used to limit the dv/dt across the triac to within its capability. This rating is lowest at high temperatures, so the network values should be chosen for proper operation at the highest temperature the circuit will encounter.

The values shown on the schematic were used with a 1/2 hp, 115 volt motor. The peak starting current of this motor was 40 amperes while the peak running current was 8 A. The value of the resistor was chosen so that the triac would not turn on when current was less than 12 A. This occurred for this motor after 12 cycles. It should be possible to use the triac shown for integral-horsepower motors since the triac conducts for only a few cycles. Of course, maximum current ratings must be observed. This circuit as shown has been operated at temperatures as high as 80° C. The motor-starting capacitor limits the maximum ambient temperature to +65°C and the duty cycle of the motor of a maximum of 60 starts of one second duration per hour. The triac will perform satisfactorily under these conditions.

4.3 Triac Prevents Relay Control Arcing

A common problem in contacts switching high current is arcing which causes erosion of the contacts. A solution to this problem is illustrated in Figure 4-7. This circuit can be used to prevent relay contact arcing for loads up to 50 amperes. There is some delay between the time a relay coil is energized and the time the contacts close. There is also a delay between the time the coil is de-energized and the time the contacts open. For the relay used with this circuit both times were about 15 ms. The triac across the relay contacts will turn on as soon as sufficient gate current is present to fire it. This occurs after switch S1 is closed but before the relay contacts close. When the contacts close, the load current passes through them, rather than through the triac, even though the triac is receiving gate current. If S1 should be closed during the negative half cycle of the ac line, the triac will not turn on immediately but will wait until the voltage begins to go positive, at which time diode D1 conducts providing gate current through R1. The maximum time that could elapse before the triac turns on is 8-1/3 ms for a 60 Hz supply. This is adequate to assure that the triac will be on before the relay contact closes. During the positive half cycle, capacitor C1 is charged through D1 and R2. This stores energy in the
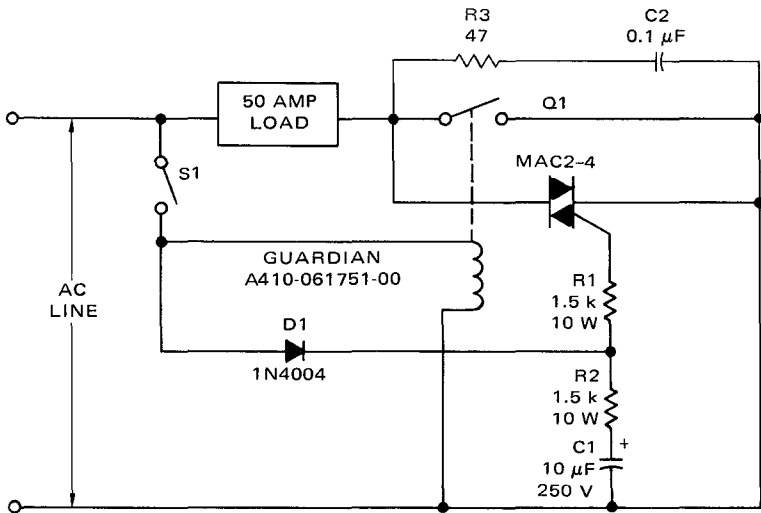
Figure 4-7 — Triac Prevents Relay Contact Arcing

capacitor so that it can be used to keep the triac on after switch S1 has been opened. The (R1 + R2)C1 time constant was set such that sufficient gate current would be present at the time of relay dropout after the opening of S1, to assure that the triac would still be on. For the relay used, this time was 15 ms. The triac therefore limits the maximum voltage across the relay contacts upon dropout to its voltage drop of about 1 volt. The triac will conduct until its gate current falls below the threshold level after which it will turn off when the anode current goes to zero. The triac will conduct for several cycles after the relay contacts open.

This circuit not only reduces contact bounce and arcing but also reduces the physical size of the relay. Since the relay is not required to interrupt the load current, its rating is only based on two factors: the first is the rms rating of the current-carrying metal, and the second is the contact area. This means that many well-designed 5 ampere relays can be used in a 50 ampere load circuit. Because the size of the relay has been reduced, so will the noise on closing. Another advantage of this circuit is that lifetime of the relay will be increased with no contact burning, welding, etc.

The R-C circuit shown across the contact and triac is to reduce dv/dt if any other switching element is used in the line. Typical values are 47 ohms and 0.1 μF.

4.4 AC Static Switches and Static Contactors

AC static switches and static contactors with essentially zero-point switching on all but the first cycle are illustrated by the circuits in Figures 4-8 through 4-11.

The circuit shown in Figure 4-8 can be used to switch resistive loads on an ac supply at frequencies below 30 kHz with electrical control signals. SCR Q1 is used to prevent gate current in Q2 when Q2 is reverse-biased by the supply voltage, which could occur the first time Q2 is turned
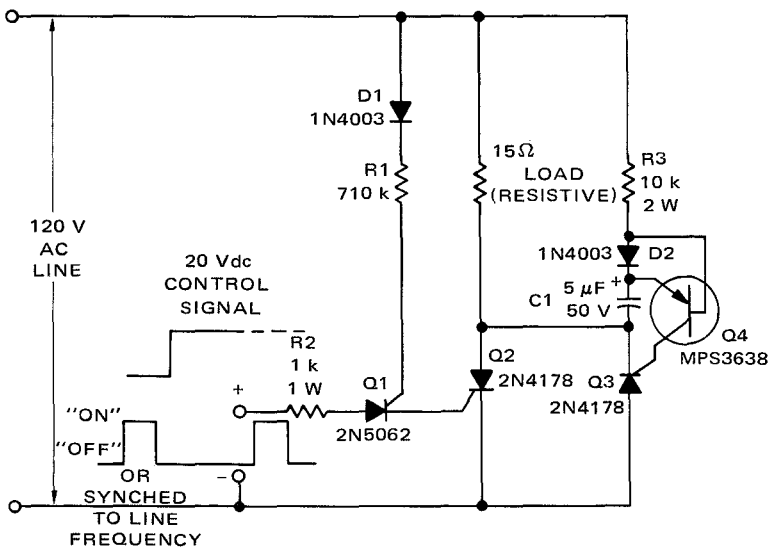
Figure 4-8 — SCR AC Static Switch
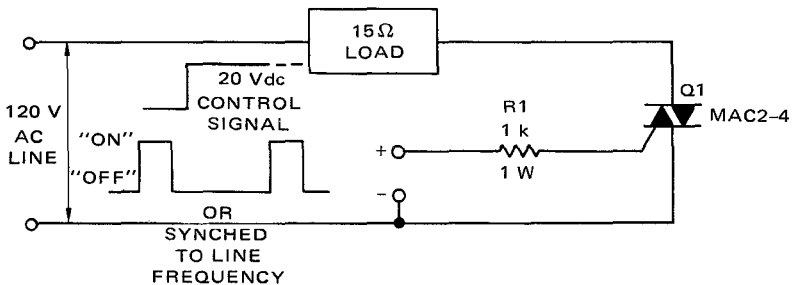
Figure 4-9 — Triac AC Static Switch
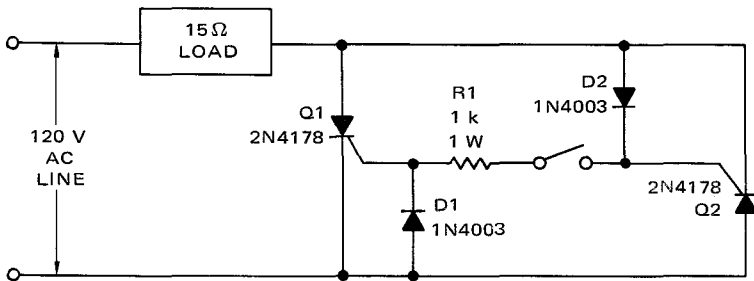
Figure 4-10 — SCR AC Static Contactor
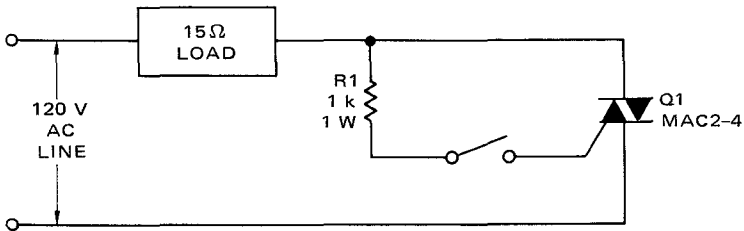
Figure 4-11 — Triac AC Static Contactor

on. Q1 will turn on during the first positive half cycle following the application of the control signal to R2. Q1 provides gate current to Q2, thereby turning Q2 on. The device used for Q1 is a sensitive gate SCR. This allows the value of R1 to be large, which limits the amount of current in the gate of Q2 to a value below turn on when no voltage is applied to R2. Diode D1 prevents reverse breakdown of the gates of Q1 and Q2. During the half cycle that Q2 is on, capacitor C1 charges through D2 and R3 to the peak line voltage. When D2 becomes reverse-biased by the decaying line voltage, C1 discharges through the base-emitter junction of Q4, R3 and the load, thus turning Q4 on and providing another discharge path through Q4 and the gate of Q3. The gate current of Q3 must be above its threshold level when the line voltage reverses for Q3 to turn on. The values shown were chosen to provide 50 mA of gate current when the anode-cathode voltage of Q3 reaches 10 volts. This circuit will work at other supply frequencies below approximately 30 kHz. This maximum frequency is due primarily to the recovery time of the SCRs.

This circuit will not switch large reactive loads since Q2 turns off when its current goes to zero, and by the time that occurs with a current which lags the voltage, the gate current of Q3 has decayed to a value less

than its threshold level thereby preventing the device from turning on.

A circuit that can be electronically controlled to switch resistive or reactive ac loads statically is shown in Figure 4-9. The triac will turn on when the control signal is applied to R1. It will remain on until the control signal is removed, after which it will turn off when the current goes to zero. As mentioned in section 4.1, present triacs are limited to a maximum supply frequency of about 60 Hz by commutation dv/dt.

If it is not necessary to provide complete static switching, then a small switch can be used as in the static contactor shown in Figure 4-10, in which the SCRs remove the load from the line when the switch is open. When the switch is closed, diode D2 provides gate current to Q1 during the positive half cycle, turning Q1 on, and D1 supplies current to Q2 during the negative half cycle, turning Q2 on. Resistor R1 limits the gate current immediately after closing the switch until one of the SCRs comes on. The maximum load current capability is limited only by the ratings of the devices used.

In Figure 4-11, the main control element is the triac. Gate current is provided when the switch is closed, turning the device on. The supply voltage must be greater than the gate threshold voltage each half cycle for the triac to turn on. If the load is reactive, the current will either lead or lag the voltage. For this condition, the triac will turn off each half cycle when the voltage is greater than zero. If the phase shift is great enough that the magnitude of the voltage is larger than the trigger level of the triac at turn off, then the triac will turn on immediately. Should the load be resistive, so that there is no phase shift, there will be a dead band until the supply voltage exceeds the triggering level. When the control signal is obtained from a 120 volt rms supply, the dead band is negligible. After the switch is opened, the triac turns off when the current goes to zero.

4.5 DC Static Switches and Static Contactors

The circuits shown in Figure 4-12 through 4-15 are dc switches and static contactors. Static contactors use small switches with low current-carrying capacity to switch heavy loads; typical contactors are shown in Figures 4-12 and 4-13. The switches shown can be remote from the load. In Figure 4-12, a transistor is used as the controlled device. The circuit is off when the switch is open. When the switch is closed, resistor R1 provides base current to the transistor holding it on. Resistor R2 is used to hold the base at the potential of the emitter when the switch is open. If there is an inductance in the load, diode D1 is required to protect the transistor from voltage spikes which occur during the switching time. A circuit that does not require continuous drive is shown in Figure 4-13. The

switch required is spring-loaded SPDT switch with a center off position. When S1 is thrown to the on position, resistor R1 provides gate current to SCR Q1, connecting the load to the supply. The switch is returned to the center off position since gate drive is not required once the SCR is on. While Q1 is on, capacitor C1 charges to the supply voltage through resistor R2. When it is desired to turn the circuit off, S1 is pushed to the off position which provides gate current to Q2 through R1 and R3. As before, S1 is returned to the center off position. Q2 is smaller than Q1 since it is used only to commutate Q1 off and its average power dissipation is small. Since Q2 is a smaller device than Q1, its gate current requirement is less
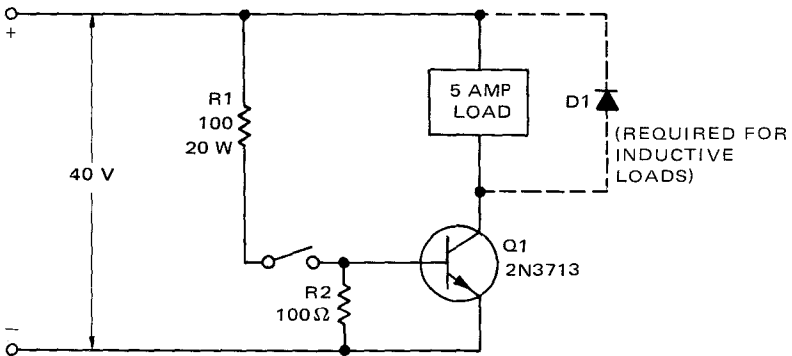
Figure 4-12 — Transistor DC Static Contactor
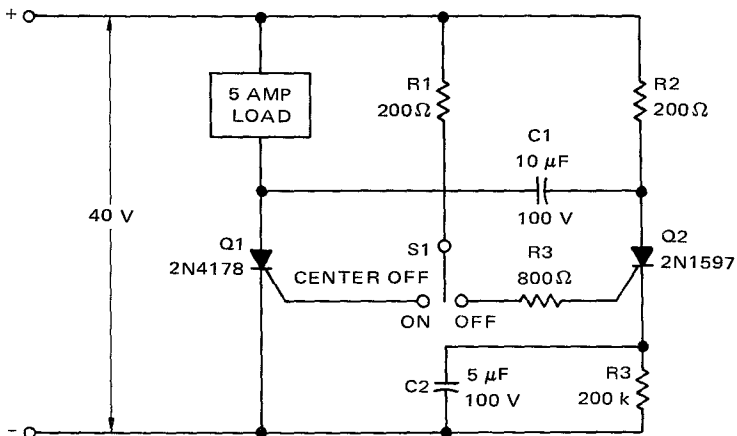
Figure 4-13 — Thyristor DC Static Contactor

than that of Q2; therefore, resistor R3 is required to limit the gate current of Q2 below its maximum value. When Q2 first turns on, capacitor C2 has no voltage across it, permitting the energy stored in C1 to turn Q1 off. C2 then charges to the supply voltage, at which time Q2 turns off since its current falls below the holding level. Following this, R3 discharges C2, returning C2 to its initial uncharged state. R3 must be large enough that it
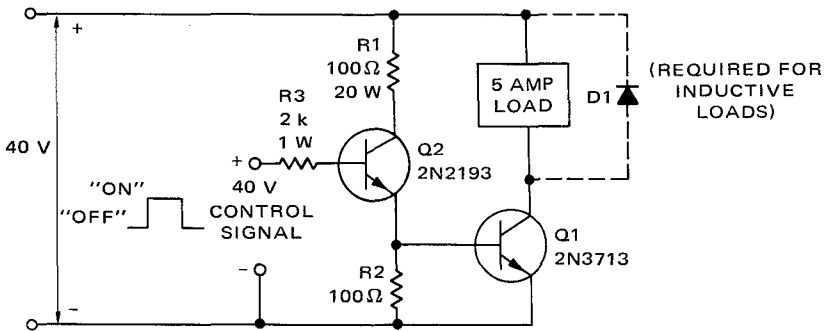
Figure 4-14 — Transistor DC Static Switch
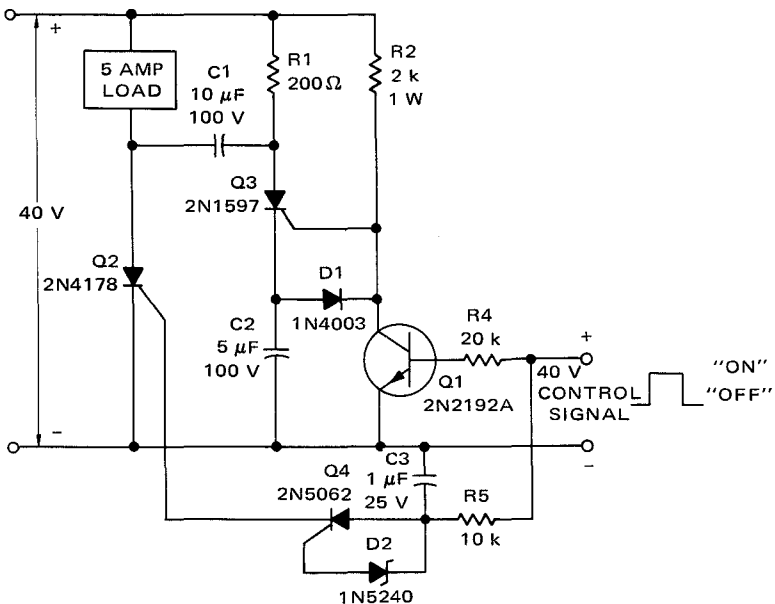
Figure 4-15 — Thyristor DC Static Switch

will not draw enough current to hold Q2 on after it has been turned on. The C2-R3 network assures that this circuit will draw no current when off.

These basic circuits can be used to provide static switching if the switches are replaced by semiconductors as shown in Figures 4-14 and 4-15. In Figure 4-14, transistor Q2 is used to control power transistor Q1. When the control signal is applied to R3, transistors Q1 and Q2 turn on, connecting the load to the power supply. When the control signal is removed both devices turn off. If a voltage other than that shown is to be used as the control signal, then R3 should be changed to keep the base current of Q2 at 2 mA. Figure 4-15 shows the schematic of a thyristor static switch. The basic circuit is the same as that shown in Figure 4-13 except that the components associated with Q1 and Q3 replace the switch. When the control signal is applied, transistor Q1 turns on, clamping the gate of Q3 to ground, thus preventing it from turning on. Diode D1 is used to remove any charge on C2 through itself and Q1, thus protecting the gate of Q3. At the same time C3 begins charging. When the voltage across C3 reaches the breakdown voltage of zener diode D2, D2 conducts and turns on Q4, which in turn fires Q2. This circuit is required to prevent false triggering of Q2 and a possible "latch up" condition in which both Q2 and Q3 are on. When the control signal is removed, Q1 and Q4 turn off. This removes gate drive from Q2 and provides gate drive to Q3, which then commutates Q2 off. Q3 turns off when C2 becomes charged, thus removing the circuit from the supply.

4.6 Overvoltage and Overcurrent Protective Circuit with Automatic Reset

The circuit shown in Figure 4-16 can be used to protect a resistive load from both excessive voltage and excessive current. It can be used with 20 A, 115 Vac power supplies. There are three basic sections to the circuit, the power driver, the overcurrent detector and the overvoltage detector. Each one will be analyzed separately.

Power Driver

When line "A" goes positive with respect to line "B", C2 charges through R6 and R7 until the breakover voltage (about 28 volts) of three-layer diode D8 is reached. D8 then turns on, and a pulse of current is supplied to the gate of SCR Q4, causing Q4 to conduct. At this time all the line voltage, minus the forward voltage drop of Q4 (about 1 volt), is across R_L. This voltage also appears across D9, R8, R9 and C3. Capacitor C3 will charge until the line voltage falls below the voltage on C3. At this
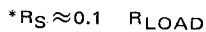
Figure 4-16 — AC Overvoltage and Overcurrent Protection with Automatic Reset for 20 A 115 Vac Power Supply

OVER-VOLTAGE DETECTOR

OVER-CURRENT DETECTOR

POWER DRIVER

LINE "A"

120 V AC 60 Hz

LINE "B"

D1 1N4004

200 V 0.25 μF C1

D2 1N4004

R11 5 k

D3 MPT-28

R1 10 k 1 W

MJ-431 Q1

MJ-423 Q2

100 R4

D4 1N4004

R12 200

A

R_S

R2 620

D5 1N4004

D6 1N4004

R3 1 k 10 W

D7 1N4004

Q3 2N5062

R5 1 k

R6 15 k 1 W

RL

D8 MPT-28

C2 0.25 μF 200 V

R7 51

D9 1N4004

R9 3.3 k 1 W

C3 400 V 0.5 μF

Q4 2N5169

R10 220 k

I1 NE-51

2.2 k R8 1 W

Q5 2N5169

*$R_S \approx 0.1$ R_{LOAD}

time, D9 becomes reverse-biased, and C3 discharges through R8 and the gate of Q5. The values of R8, R9, and C3 are adjusted so that as the line voltage goes through zero there is still ample charge on C3 to provide sufficient gate current to turn on Q5. Q5 is therefore slave-fired from Q4, since Q4 must be turned on before Q5 can be turned on.

Overcurrent Detector

This circuit offers a unique approach to overcurrent protection in that it does not require a current sensing element in series with the load. One way to keep the load from being connected across the line in the event of a short circuit or some other excessive current load is to prevent diode D8 from firing Q4. In order to do this, the overcurrent detector must work before the voltage across C2 and R7 reaches the breakover voltage of D8. When line "A" goes positive with respect to line "B", Q1 and Q2 are biased on by R3, thus allowing the load current to flow through the sense resistor R_S. Assuming that R_L is much greater than R_S, R_L will act as a current source and the voltage developed across R_S is sufficient to forward-bias D5 and the gate of Q3, then Q3 will turn on and clamp the voltage across C2 and R7 to approximately 1 volt. This prevents D8 from turning on, and also turns off Q1 and Q2 since base drive is removed from Q2 by D7. The only current now in the load is the leakage currents of Q1, Q2, Q4, Q5, and the "on" current of the fault indicator, neon lamp I1.

Component tolerance considerations force the overcurrent circuit to work before the line voltage reaches 20 volts. The available circuit gain limits the collector current of Q1 to 3.5 amperes which sets the maximum peak load current at

$$I_p = \frac{(V_p)(I_{C\ max})}{V\ sense\ (max)} = \frac{165 \times 3.5}{20} = 28.8\ A\ .$$

The maximum rms load current is therefore 20 A. Consideration of the voltage levels required for circuit operation allows the following equation to be used in determining the minimum value of R_S:

$$R_{S(min)} = (R_L)\left(\frac{V_{RS}}{V_{RL}}\right),$$

where V_{RS} = the voltage across R_S,

and V_{RL} = the voltage across R_L.

V_{RS} must be large enough to assure that D5 and the gate of Q3 can be forward-biased. Two volts will satisfy this requirement. Therefore,

$$R_{S(min)} = R_L\left(\frac{2}{20}\right),$$

or $\quad R_{S(min)} = \left(\frac{1}{10}\right)R_L .$

In order to assure that R_L will appear as a current source to R_S, however, the *maximum* value of R_S should be less than or equal to $1/10\ R_L$; therefore, the proper value for R_S in this case is $1/10\ R_L$.

The voltage divider consisting of R12 and R2 permits a selection of the "trip" point of overcurrent protection. This divider must be set so that the voltage at point A is adequate to fire Q3 before the collector current of Q1 reaches 3.5 A. Diode D4 protects Q1 and Q2 from the reverse line voltage. Resistor R4 compensates for leakage current (I_{CBO}) in Q1. D5 and D6 form an OR gate permitting the overcurrent and the overvoltage circuits to both work into Q3. The maximum voltage across Q3 is the breakover voltage of D8, therefore it need only withstand this voltage. R7 is used to limit the discharge current of C2 through Q3.

The maximum forward surge current (I_{FM} surge) of Q4 and Q5 is 240 A. This is the maximum current this circuit will withstand. If a short in the load should occur at peak line voltage (168 volts), then this circuit will withstand the resulting current if the total source impedance R_G in the 115 V line is equal to or greater than 0.7 ohms:

$$R_{G(min)} = \frac{V_{LINE\,(max)}}{I_{FM\,(surge)}} = \frac{168}{240} = 0.7\ \text{ohm}.$$

Q4 and Q5 will never have to handle more than one-half cycle of overload current each before the overcurrent detector reacts. A 5 μH choke in series with the load will protect the SCRs from destruction due to excessive di/dt if a fault occurs at this worst-case condition. When the cause of the excessive current is corrected, the overcurrent circuit no longer fires Q3 and the load is automatically connected to the line through Q4 and Q5.

Overvoltage Detector

This circuit is a peak voltage detector. The detector circuit detects

any excessive voltage on the negative half cycle (line "B" positive with respect to line "A") and prevents Q4 and Q5 from turning on as long as the overvoltage exists. This is accomplished as follows: capacitor C1 charges through D1 and D2 to a voltage level determined by R11 and R1. D2 prevents C1 from discharging back through R11. With normal line voltages, R11 is set so that C1 charges to a value less than the breakover voltage of three-layer diode D3 plus the sum of the forward voltage drops of D6 and the gate of Q3. If the line voltage goes above normal, then this level is exceeded and Q3 breaks over, discharging C1 through D6, R11, and R1, and into the gate of Q3.

The point in the cycle at which D3 will fire, depends on the setting of R11. During the negative half cycle (line B positive, line A negative), the voltage across D3 is equal to the voltage on C1, which is determined by the setting of R11, minus the line voltage from B to A (and the diode drops of D6 and the gate of Q3). Thus, if R11 is set so that C1 charges to 40 volts, and D3 fires at 30 volts, then D3 will fire when the line voltage equals approximately 10 volts. Once the line voltage goes positive (A with respect to B), the voltage on D3 equals the voltage on C1. Therefore, the overvoltage circuit will trigger Q3 only during the negative half cycle. The time constant of R11 plus R1 and C1 must be long enough to provide adequate gate current to fire Q3 when the line voltage goes positive.

The overvoltage circuit detects an overvoltage each cycle in which an overvoltage is present. When the condition is corrected, D3 no longer breaks over and the load is again connected to the line through Q4 and Q5.

4.7 DC Overvoltage Current Control

The circuit shown in Figure 4-17 is a dc electronic circuit breaker designed to protect a load against transient voltage or current overloads. It can protect circuits which cannot be protected by slower-acting mechanical circuit breakers, as it will remove power to the load in 4 to 10 μs as opposed to 8 to 10 ms for a mechanical circuit breaker. This overvoltage, overcurrent control can be used to safeguard the output stages of transistor transmitters, receivers, amplifiers, or any other circuit where transient overloads longer than 10 μs in duration may cause damage.

When power is applied to the circuit, the control circuitry is activated and, under normal conditions, connects the load to the power source. The supply voltage to base two of unijunction transistors Q1 and Q2 in the control circuit is regulated by zener diode D1. Q2, the overcurrent control transistor, senses the load current since the voltage at its emitter is a direct function of the load current. This is so since R8 is
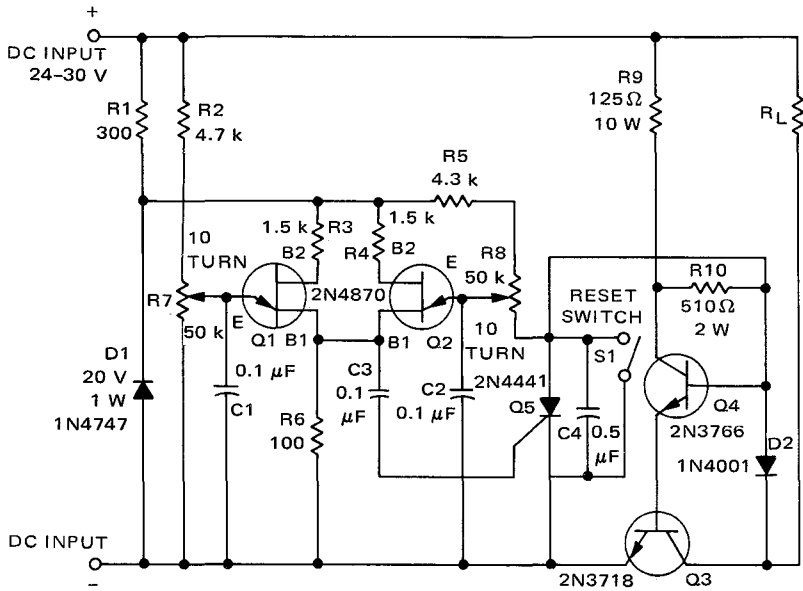
Figure 4-17 — Electronic Circuit Breaker
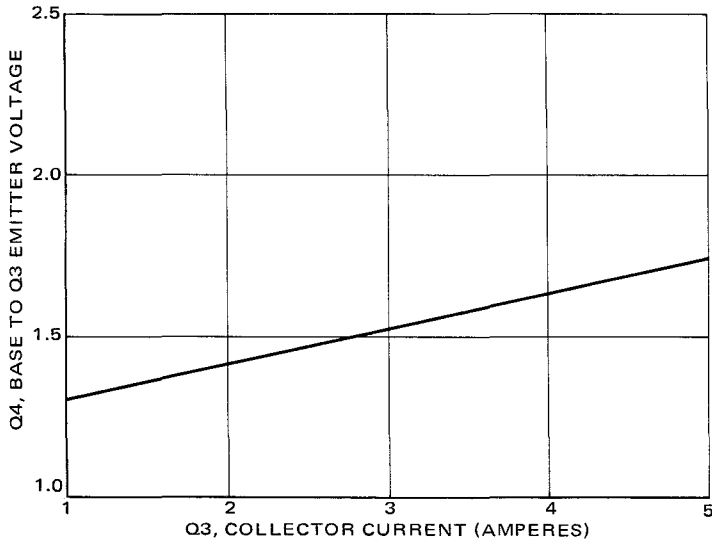
Figure 4-18 — Base-Ground Voltage of Q4 versus Collector Current of Q3 (30 V Supply)

connected to negative common through the base-emitter junctions of Q4 and Q3, and in parallel with these, D2. Q4, which is initially biased on by R10, biases Q3 on. Q3 is driven hard enough that it is in saturation and is capable of handling the current through D2 and through the load. The $V_{CE(sat)}$ of Q3 is less than 1 V for load currents as high as the rated current of 5 amperes. D2, when forward biased, maintains an almost constant voltage between the base of Q4 and the collector of Q3. After Q3 is turned on, any current increase through it due to the load will raise its collector-emitter voltage. D2 will follow this increase and raise the voltage at the base of Q4. The increase in voltage between the base of Q4 and the negative supply terminal, versus current through Q3, is quite linear from 1 to 5 A (see Figure 4-18), and is used as a current-to-voltage converter and as the sense point for Q2. The voltage at the arm of R8 is set just below the firing level of Q2 at the maximum load current desired. If the current rises above this level, Q2 fires and discharges C2 through the emitter into base one and R6. The rapid discharge of C2 creates a pulse that is transmitted through C3 to the gate of SCR Q5 and turns it on. When Q5 is on, Q4 is switched off which shuts off Q3 because the sum of the forward diode voltage drops of the emitter-base junctions of Q4 and Q3 is more than the forward voltage drop of Q5. If the cause of the overload is removed, the circuit can be reset by closing the reset switch. This drops Q5 out of conduction and the circuit is reset for normal operation when the switch is released.

The overvoltage control works in a similar manner. Q1 receives its sense voltage through potentiometer R7 and resistor R2, which are in series and are connected across the input supply. The sense voltage is set slightly below the voltage which will fire Q1 so that any increase in the supply voltage will trigger Q1. This discharges C1 into R6 and the gate of Q5. Q3 is then turned off as previously described.

4.8 Optical Logic Drivers for Power Devices

It is possible to provide logic control of power devices with optical signals if circuits such as those shown in Figures 4-19 through 4-22 are used to convert the light signals into electrical signals. The truth table for these circuits show outputs for positive logic.

For the circuit shown in Figure 4-19, a "one" output is obtained at all times except when both Q1 and Q2 are exposed to bright light. Approximately 220-foot candles to Q1 and Q2 will drive Q3 into saturation. Resistor R2 provides a path for leakage currents so that Q3 does not conduct until an adequate light level is presented to Q1 and Q2. The positive output will provide almost 2 mA to a load.
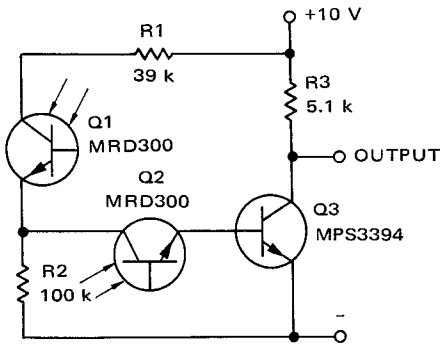
Figure 4-19 — Optical Logic Driver
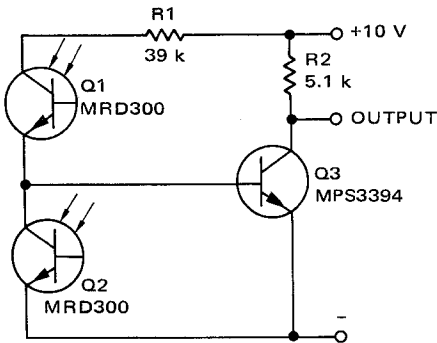
TRUTH TABLE	
Q1, Q2	OUTPUT = 0
Q1, $\overline{Q}$2	OUTPUT = 1
$\overline{Q}$1, Q2	OUTPUT = 1
$\overline{Q}$1, $\overline{Q}$2	OUTPUT = 1

Figure 4-20 — Optical Logic Driver

TRUTH TABLE	
Q1, Q2	OUTPUT = 0
Q1, $\overline{Q}$2	OUTPUT = 1
$\overline{Q}$1, $\overline{Q}$2	OUTPUT = 1
$\overline{Q}$1, Q2	OUTPUT = 1

Figure 4-21 — Optical Logic Driver

TRUTH TABLE	
Q1, Q2	OUTPUT = 1
Q1, $\overline{Q}$2	OUTPUT = 0
$\overline{Q}$1, $\overline{Q}$2	OUTPUT = 0
$\overline{Q}$1, Q2	OUTPUT = 0
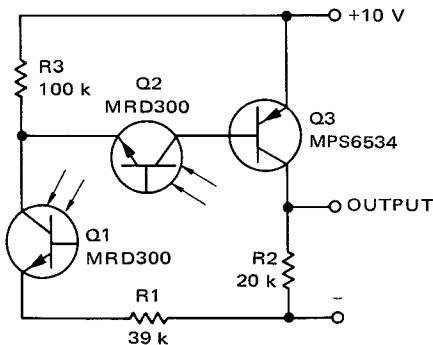
TRUTH TABLE	
$\overline{Q}1$, Q2	OUTPUT = 1
Q1, Q2	OUTPUT = 0
$\overline{Q}1$, $\overline{Q}2$	OUTPUT = 0
Q1, $\overline{Q}2$	OUTPUT = 0
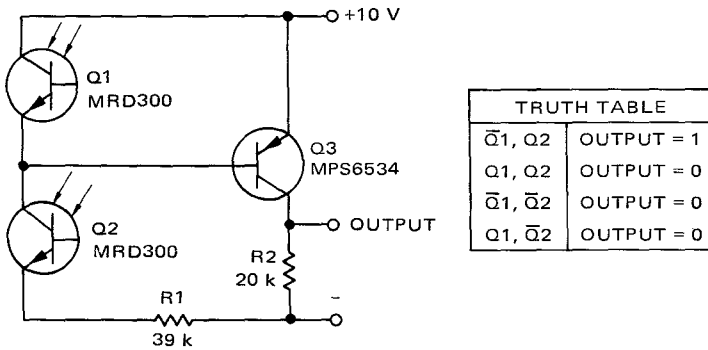
Figure 4-22 — Optical Logic Driver

The circuit shown in Figure 4-20 provides a positive output voltage at all times except when Q1 is on and Q2 is off. For this case, the base drive to Q3 is provided through R1 and Q1. As in the previous circuit, normal room lighting does not turn Q3 on, but about 220 foot-candles from a flashlight shining on Q1 will allow Q3 to saturate.

The maximum current this circuit can provide a load is 2 mA. The inverse of the output of the two previous circuits can be obtained through the use of an additional inverting stage or by the circuits shown in Figure 4-21 and 4-22. If a current sink is required, an inverter should be used, but if a current source is required, the circuits in Figures 4-21 and 4-22 should be used. The output of the circuit shown in Figure 4-21 is zero at all times except when both Q1 and Q2 are on. As in the previous circuits, about 220 foot-candles of illumination on Q1 and Q2 is enough to saturate Q3 so that it can provide at least 10 mA to a load. The value of R2 is high so it does not require any current that can be supplied to the load. A current in R2 of 1/2 mA gives satisfactory results. Resistor R3 provides a path for leakage currents so that Q3 does not conduct until Q1 and Q2 are illuminated.

The circuit in Figure 4-22 provides a zero output for all conditions except when Q1 is off and Q2 is on. For this condition, base current is provided to Q3 through Q2 and R1. Here also, 220 foot-candles is enough to allow Q3 to saturate and drive a 10 mA load. For this circuit, also, the value of R2 is high so it does not rob current from the load. Q1 must be turned on hard enough to shunt the base-emitter junction of Q3 to keep Q3 from coming on when Q2 is on. This is achieved with the illumination intensity mentioned previously.

All four of these circuits can be used as the sensing part of a two-input static switch. These outputs then become the control signal for the power switching circuits shown earlier in this chapter.

CHAPTER 5
Servo and Audio Power Amplifiers

5.1 Introduction to Servo and Audio Power Amplifiers

The circuits in this chapter are low-distortion power amplifiers for servo and audio applications. All are economical to build, and are efficient for minimum power consumption. The input to each amplifier is assumed to be conditioned to perform the function desired and the level raised to greater than 1 volt rms. For audio amplifiers this would be accomplished in a pre-amplifier which is capable of providing the desired frequency equalization. In servo systems, the pre-amplifier generally performs the required control function rather than amplifying the signal since these systems work at high level.

To satisfy these requirements and to keep the design as simple as possible, the output stage of the amplifiers should operate in class AB. There are many possible approaches to the design of class-AB amplifiers: transformer-coupled, capacitor-coupled, direct-coupled, and various combinations of the three. All have their limitations. Transformer-coupled amplifiers can be made relatively simple, but wideband operation puts fairly stringent requirements on the transformer design. This makes them expensive to use in wideband audio amplifiers but ideally suited to use in narrowband servo systems. Feedback in transformer-coupled amplifiers can lead to ac instability.

Capacitor-coupled amplifiers offer fewer problems with ac stability than transformer-coupled amplifiers, but their design is generally more complicated since more components are required for biasing and more stages are usually needed to compensate for the power loss resulting from impedance mismatch between stages.

Direct-coupled stages have relatively good ac stability, but can cause some difficulty in the area of dc stability, resulting in circuit complexity. Amplifiers using a combination of coupling methods will enjoy the benefits and suffer the limitations of all three methods.

Bias

The bias voltage applied to the base-emitter junction of a transistor amplifier determines its collector current and sets the amplifier stage in class A, B, or C operation. A class-A transistor amplifier conducts $360°$ (full) of a sine wave. In class-B operation the transistor has approximately zero bias (near cutoff) and allows $180°$ conduction. In class-C operation the transistor is biased beyond cutoff and allows less than $180°$ conduction. A transistor which is biased just above cutoff is operating in class AB and allows small-signal class-A operation and large-signal class-B operation. These classes are illustrated in Figure 5-1. Class C amplifiers are excellent for narrowband, high-Q circuits encountered in rf work, but they are almost never used in servo and audio amplifiers because of their high distortion, and they will not be discussed further here.

The load for a class-A amplifier such as shown in Figure 5-2 is generally transformer-coupled to the amplifier since this provides dc isola-
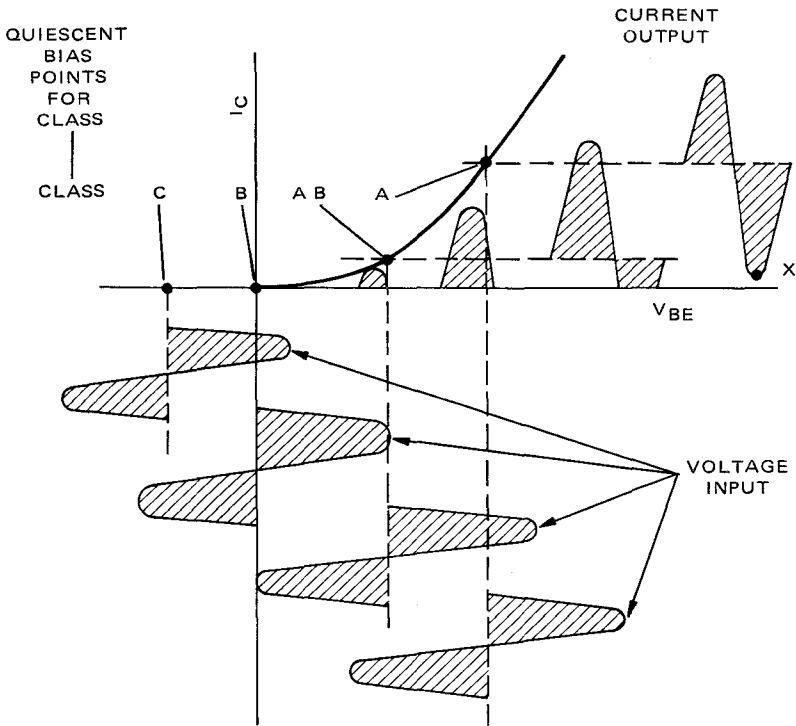
Figure 5-1 — Waveshapes of Classes of Operation
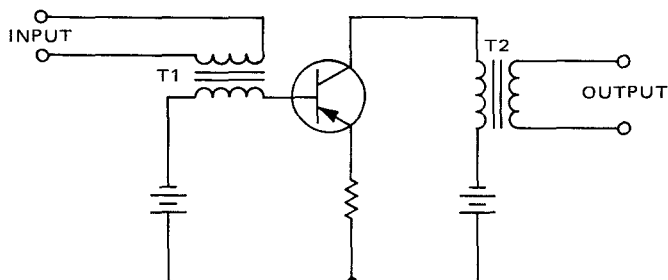
Figure 5-2 — Typical Class-A Amplifier
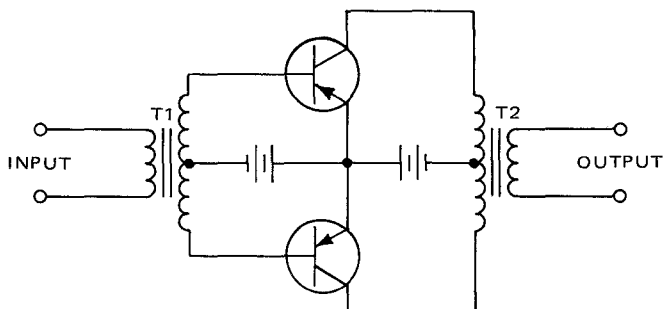
Figure 5-3 — Typical Class-A Push-Pull Amplifier

tion and, through the proper turns ratio, the optimum collector load impedance. The efficiency of the amplifier is defined as the output power at the fundamental frequency divided by the dc power supplied to the collector circuit. The maximum efficiency of class-A operation is obtained when the negative peak of the output signal approaches zero (point X in Figure 5-1) and the positive peak is close to twice the supply voltage. Theoretically the maximum efficiency is 50%, but practical amplifiers operate with less than 35% efficiency. As can be seen from Figure 5-1, the output voltage excursion swings over a nonlinear collector characteristic. The input signal must be adequate to drive the base over this range. Since the device is operated over a nonlinear range, distortion is produced; to minimize the distortion, the power output must be reduced from maximum. Careful selection of the quiescent operating point and the ac load line will give an adequate compromise between maximum power output and minimum distortion. A push-pull class-A amplifier (See Figure 5-3)

may be used to give more power output per transistor for a given distortion. For this type of operation there is no dc saturation of the output transformer core and no current of signal frequency through the power source, but special driving and supply circuits must be used. The push-pull class-A amplifier is arranged so that identical transistors are driven 180° out of phase and the outputs combined through the use of a center-tapped output transformer. If transformer drive (T1) is not used, then R-C coupling can be employed through the use of a phase-inversion stage. For relatively low power amplifiers, class-A operation may be sufficient. However when the power level exceeds about 2 watts the power supply and operating costs can be reduced and the efficiency improved by going to class-B or class-AB operation.

A class-AB amplifier is similar to the push-pull class-A amplifier shown in Figure 5-3, but it is biased to a value between cut-off and that required for class-A operation. Class-AB amplifiers are used where higher power is required from a given set of transistors than is obtainable from class-A operation. The instantaneous collector current of each transistor exists for more than 180°, but becomes zero for a small portion of each cycle as can be seen from Figure 5-1. This results in considerable distortion in the individual collector current, but the mutual coupling in the output transformer greatly reduces the effect of the interrupted collector current. Since the quiescent collector current is lower than for class A, a higher supply voltage may be used to obtain the same quiescent collector dissipation. The average value of collector current rises as a base signal is applied, and this increases the power supplied by the collector power supply. Generally, this raises the collector dissipation. The output power is increased over that of class-A operation using the same devices, due to the higher collector voltage and the extended current excursion into the nonlinear region of the collector characteristics. However, for a given voltage and a given load, class A will deliver more output power. Class-AB amplifiers are used where higher output power is required than can be obtained using the same devices in class A, and lower distortion is required than can be obtained with (as will be shown) class B. The typical efficiency that can be obtained for a class-AB amplifier is about 55%.

Class-B push-pull amplifiers are used where higher power is required than can be achieved with the class-AB amplifier using the same devices and power supply. This higher output is achieved at the expense of an increase in the driving power and the distortion. However, the maximum efficiency is increased. The theoretical limit to the efficiency is 78-1/2%, but most practical amplifiers operate around 60% to 65% efficiency. A class-B amplifier is shown in Figure 5-4. For class-B operation the base bias on each transistor is set so that the individual collector currents are nearly
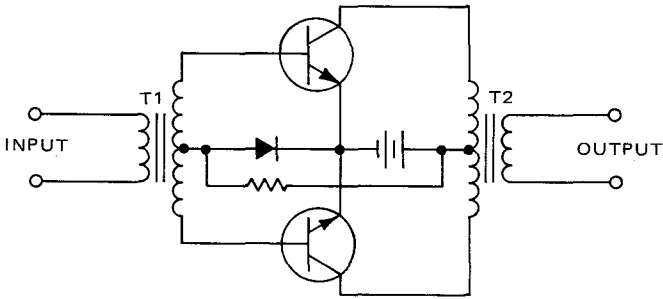
Figure 5-4 — Typical Class-B Push-Pull Amplifier

zero at quiescence as can be seen in Figure 5-1. The drive signal then causes collector current in either transistor only when the base-emitter junction of that device is forward biased. Since the driving signals at the bases are 180° out of phase, only one device is on at any given time. Thus the collector current is proportional to the driving voltage, and the power dissipated in the transistors is small with small signals, as opposed to the large dissipation for class-A amplifiers with low input. A relatively large driving power is required for the transistors when they are driven to achieve the maximum power output. This requires large collector currents and the power supply used must have good voltage regulation. If this is not so, the bias potential would vary with the input signal, leading to severe distortion. The frequency response for the class-B circuit shown in Figure 5-4 depends upon the transformers used, as does the response in the class-A and class-AB amplifiers of Figures 5-2 and 5-3.

Low-frequency rolloff is caused by the primary inductance, whereas high-frequency rolloff is caused by the leakage inductance and winding capacitance. In relatively narrowband servo systems, the use of transformers can greatly ease design problems. However, for audio amplifiers they create problems if wideband operation is desired. The advent of matched complementary transistors has made class-B operation very attractive since these transistors lend themselves readily to direct coupling, and, in addition, a phase inversion is not needed in the driver. This eliminates the need for transformers. One of the basic difficulties encountered with class-B amplifiers is the elimination of crossover distortion. One way to overcome this is to bias for class AB, but this results in higher quiescent power dissipation. With some transistors, it is sometimes useful to drive from a high-impedance, or current, source. This can give low crossover distortion with a minimum of quiescent current.

Feedback

The operating characteristics of a system can be improved through the use of negative feedback. Proper use of negative feedback can increase the bandwidth of an amplifier, improve its gain stability, reduce the noise generated in the stage, lower its output impedance and improve its linearity, which reduces intermodulation and harmonic distortion. Negative feedback must be used with care since it is possible to create regeneration at the band edges even though midband gain is stable. This is caused by excessive phase shift through both the amplifier and the feedback path. It generally occurs where the frequency roll-off characteristics are greatest. There is no simple way to relate the bandwidth of an amplifier with feedback to that without feedback. Therefore each system must be analyzed through the use of the feedback gain equation:

$$A_f = \frac{A}{1 - A\beta} \, ,$$

where, A_f is the gain with feedback,

A is the gain without feedback (load connected to the output), and

β is the feedback factor.

In the same manner, the distortion of a stage can be reduced with negative feedback as shown by the equation.

$$D_f = \frac{DA_f}{A} = \frac{D}{1 - A\beta}$$

where, D_f is the output distortion with feedback,

D is the output distortion without feedback, and

A, β and A_f are as defined above

The noise generated in the amplifier stage is reduced according to the following equation,

$$No = Ni \left(\frac{An}{1 - A\beta} \right) ,$$

where,　No is the output noise amplitude,

Ni is the noise amplitude at point of introduction,

An is the gain without feedback from point of introduction of noise to amplifier output, and

A and β are as defined above.

The output impedance of an amplifier is reduced with negative voltage feedback. If negative current feedback is used the output impedance is increased. The equation for the output impedance of an amplifier with negative voltage feedback is

$$Z_f = \frac{Z}{1 - \left(A + \dfrac{AZ}{R_L}\right)\beta}$$

where,　Z_f is the output impedance with feedback,

Z is the output impedance without feedback,

R_L is the load resistance, and

A and β are as defined above.

Another important parameter of an amplifier is its transient response. The transient response is measured by the degree of damping of the amplifier. The system can be either under-damped, critically damped or over-damped. These terms express whether the output waveform, with a step-function input, will overshoot the final output value, rise to the final value in a minimum time with little overshoot or require a long time to acquire the final value. An analytical determination of the transient response is quite difficult because of the varied nature of the amplifier's transfer function. Therefore this characteristic is generally measured by subjecting the amplifier to a step-function input voltage (square wave) and observing the degree to which the output follows the input.

Care must be taken to assure stability if a transformer is within the feedback loop, because of the phase shifts involved. Consider the low-frequency equivalent circuit of a transformer as shown in Figure 5-5, where E_g is the source voltage; R_g, the source resistance; L_p, the transformer primary inductance; and R'_L, the reflected load resistance. For this circuit, the input-output relationship is
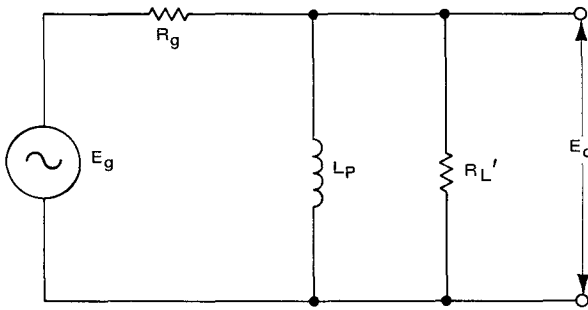
Figure 5-5 — Low-Frequency Equivalent Circuit of Transformer Coupling

$$\frac{E_O}{E_g} = \frac{R'_L \,(j\omega L_p)}{R_g R'_L + j\omega L_p \,(R_g + R'_L)} \, , \qquad (1A)$$

and the phase shift is

$$\theta \;=\; 90° - \text{ARCTAN}\frac{\omega L_p \,(R_g + R'_L)}{R_g R'_L} \qquad (2A)$$

One obvious way to minimize phase shift is to make R_g small. Thus, it is desirable to drive from a low source impedance. Low impedance means an emitter-follower or a common-emitter stage, with heavy degenerative feedback. Neither offers high voltage gain, consequently, another stage would usually have to be added.

Making R_g very small compared to R'_L reduces equation (2A) to

$$\theta \;=\; 90° - \text{ARCTAN}\frac{\omega L_p}{R_g} \qquad (3A)$$

which indicates that, for minimum phase shift, ωL_p should be large with respect to R_g.

If it is not practical or economical to drive from a low-impedance source, the situation changes somewhat. Considering the case where the transformer is in a transistor collector circuit, the source resistance R_g will generally be much larger than R'_L. Now equation (2A) becomes

$$\theta \;=\; 90° - \text{ARCTAN}\frac{\omega L_p}{R'_L} \, . \qquad (4A)$$

Increasing the ωL_p-to-$R'L$ ratio will reduce the phase shift. If the load is constant (or nearly so), for example, when the feedback resistor is connected to a feedback winding, it is only necessary to make the transformer primary reactance large compared to the reflected load resistance to minimize phase shift. When the load is a transistor base-emitter junction, however, this may not be so easy. The designer does not always have complete limit curves on h_{ie} (input impedance), in which case he would have to make an educated guess or greatly over-design the transformer.

As a final note some factors pertaining to feedback should be mentioned. Negative feedback reduces the overall gain of an amplifier, so the open-loop gain must be high enough that the closed-loop gain will provide the desired results. If this is done, then the $A\beta$ product can be made large, with the result that the amplifier closed-loop gain is virtually independent of the gains of the devices used in the amplifier. This reduction in open-loop gain requires that the input voltage be increased to obtain the same output as obtained from open-loop operation, or in other words, the input must be a high-level signal. These few disadvantages of feedback are far outweighed by the increase in performance obtainable through the use of feedback.

Stability

It is important to maintain a stable operating point that remains unchanged by transistor and temperature variations. An immediately noticeable effect of a substantial change in the quiescent operating point is an increase in distortion. This results from clipping of the peak of the output waveform. The more the operating point shifts, the more the waveform is clipped. If the operating point is not adequately stabilized, thermal runaway may result and this can be fatal to the transistors. The transistor parameters that can be affected by temperature are the base-emitter voltage, collector leakage current and current gain. For stability, the effects of changes in these parameters must be minimized. An effective way to achieve this is to place a small resistor in series with the emitter of each output transistor.

5.2 115 Vac, 7.5 W Transformer-Coupled Servo Amplifier

The transformer-coupled servo amplifiers shown in Figures 5-6 and 5-7 are designed to drive ac motors drawing 7.5 watts. Because excellent impedance matching is achieved with transformers, less power is lost between stages than with other coupling configurations. Thus only three transistors are required to provide a stable voltage gain of 100.

The common-emitter first, or driver, stage for Figure 5-6 contains

Figure 5-6 — 115 Vrms Transformer-Coupled 7.5 Watt Servo Amplifier
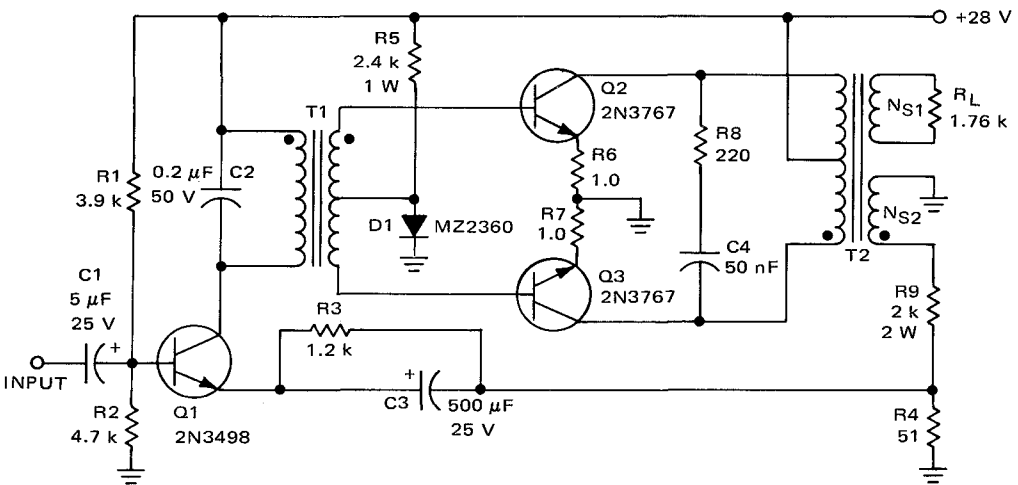
All resistors ±5%
Q2 and Q3 mounted on heat sinks with case-to-ambient thermal resistance $\leq 3.5^\circ$C/watt (each transistor)

Figure 5-7 — 115 Vrms 7.5 W Modified
Transformer-Coupled Servo Amplifier

Servo and Audio Power Amplifiers
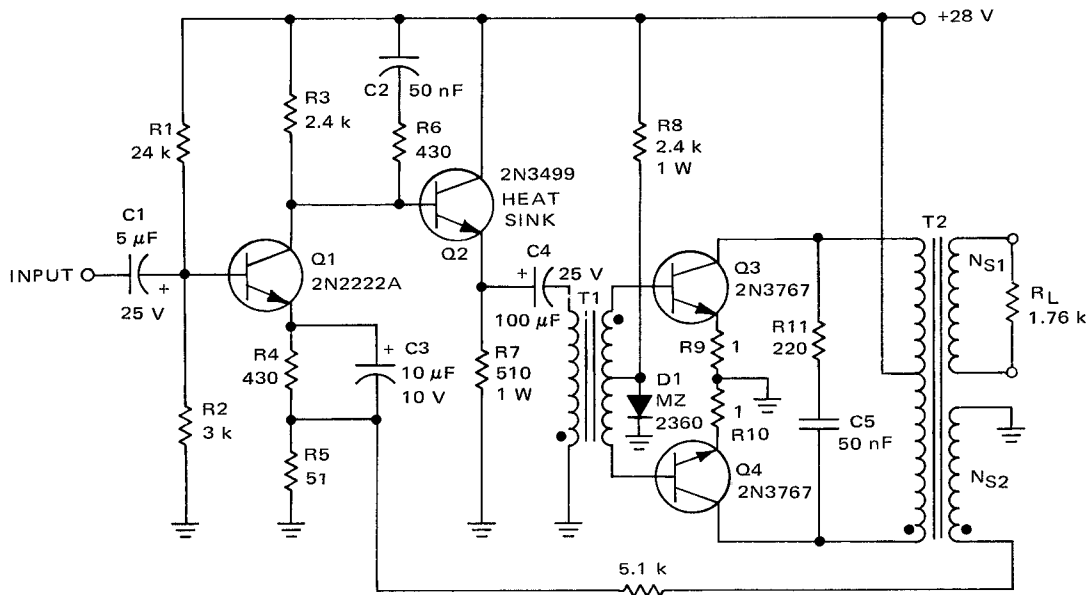
T1, T2 — See Figure 5-8
All resistors ±5%
Q3 and Q4 mounted on heat sinks with case-to-ambient thermal resistance ≤ 3.5°C/watt (each transistor)
Q2 mounted on heat sink with case-to-ambient thermal resistance ≤ 25°C/watt

the driver transformer in its collector circuit. Normally it is undesirable to drive from a transistor collector when the transformer is driving a non-linear load such as a transistor base-emitter junction, but in this amplifier, this problem has been overcome. In order to minimize phase shift when driving from a high-impedance source, the primary shunt inductance of the transformer must be much larger than the reflected load impedance. In the case of a transistor base-emitter load, the highest value of the junction impedance must be considered. Fortunately, a servo amplifier generally works at a nearly constant frequency (or over a fairly narrow band). Consequently, in the amplifier circuit shown in Figure 5-6, a capacitor (C2) was used to tune the driver transformer primary. This minimizes the phase shift. This technique may be used as long as the low-frequency phase shift of the transformer does not cause frequency instability. If wideband operation of the servo amplifier were necessary, if the amplifier tended to be unstable when the feedback loop was closed, or if the amplifier phase characteristics created problems in the total servo loop operation, a common-collector (emitter-follower) stage could be added to provide low-source-impedance drive for the transformer (Figure 5-7). This would give low phase shift across a fairly wide band (less than 10 degrees from 50 Hz to 5 kHz).

Again referring to Figure 5-6, it can be seen that the output stage is operating class B push-pull. Bias voltage for the two output transistors, Q2 and Q3, is provided by the drop across diode D1. A diode is used for biasing, since it will tend to compensate for variations in the base-emitter voltage of the transistor with temperature, thus increasing thermal stability. It is also important, for good thermal stability, that the secondary winding of the driver transformer have low dc resistance. In this circuit, dc secondary-winding resistance is about 58 ohms total (29 ohms per half). Emitter resistors R6 and R7 have been added to provide additional thermal stability.

As can be seen, feedback for the amplifier is derived from a separate feedback winding rather than from across the load or motor winding. One of the problems associated with transformer-coupled amplifiers is the proper use of feedback. It is rather difficult to carry any reasonable amount of feedback around two transformers. Also when the motor time constant is added, the effective use of feedback becomes even more difficult. One approach would be to add a network to compensate for the motor time constant. Since the time constant can vary quite widely from motor to motor, this would require an individually tailored network for each motor. A more practical solution is the use of a separate feedback winding. This eliminates the effect of the time constant of the motor on

feedback loop stability. The difficulty with this approach is arriving at the proper turns ratio between the primary and feedback windings. For minimum phase shift − important with regard to frequency stability − it is desirable to keep the primary-to-feedback winding ratio as small as possible. This results in a lower reflected load impedance, and therefore, less phase shift. On the other hand, if this ratio becomes too low, then a large voltage will be developed across the feedback winding, leading to excessive power dissipation in the feedback resistor. A compromise between the two approaches is necessary.

The voltage at the feedback winding is carried to the input through resistor R9. The voltage gain of the amplifier is given approximately by

$$A_V \approx \left(\frac{R_4 + R_9}{R_4} \right) \left(\frac{N_{S1}}{N_{S2}} \right)$$

Capacitor C4 and resistor R8 across the primary winding of transformer T2 prevent high frequency instability. This instability is partially caused by the high-frequency characteristics of T2, and consequently,

N_P = 1000 turns No. 34 AWG
N_S = 500 turns No. 36 AWG, center tapped
Wind bifilar on magnetic metals EI-75 square stack core SL-14 material, 5 x 5 interleaved

(A) − DRIVER TRANSFORMER (T1)

N_P = 512 turns No. 24 AWG, center tapped
N_{S1} = 1070 turns No. 28 AWG
N_{S2} = 400 turns No. 28 AWG

Wind bifilar on Arnold AA-81 Silectron Core

(B) − OUTPUT TRANSFORMER (T2)
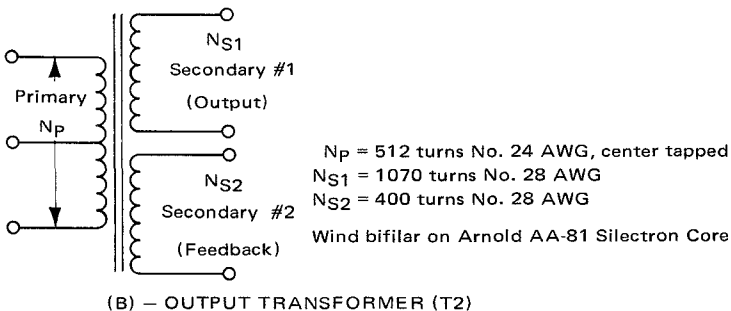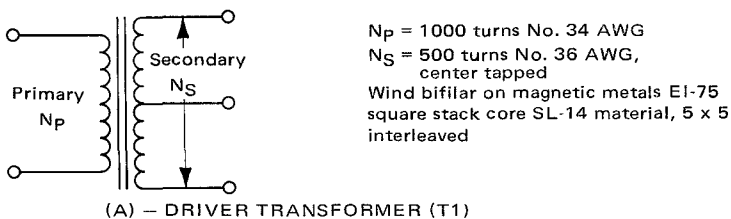
Figure 5-8 − Transformer Construction

depending upon the construction of T2, R8 and C4 may or may not be necessary.

Although data was taken only on the circuit illustrated in Figure 5-6, the other circuit (Figure 5-7) should perform similarly with one exception. This circuit (Figure 5-7) should have better phase-shift characteristics over a wider frequency range. Of course, it contains one more transistor, and is less efficient owing to the shunt drive of transformer T1.

The operating temperature range of this transformer-coupled servo amplifier is $-55°C$ to $+125°C$. If operation to only $100°C$ is needed, the case-to-ambient thermal resistance of the heat sinks for the output transistors Q2 and Q3 can be increased from $3.5°C$/watt (maximum) to $13.5°C$/watt (maximum).

The voltage gain of the amplifier at $25°C$ is 40 dB ± 1 dB, and the gain variation over the operating temperature range of $-55°C$ to $+125°C$ is within ± 2 dB. The power gain of the amplifier is a minimum of 37 dB. Maximum output is 115 volts rms into a 1.75 kΩ load, for an output of 7.5 watts.

Input impedance of the amplifier is 2 kΩ and the output impedance is approximately 200 Ω. Harmonic distortion is less than 5% up to the maximum output voltage of 115 volts rms.

5.3 20 V rms, 10 Watt Complementary Output AC Servo Amplifier

The complementary servo amplifier shown in Figure 5-9 can be used for both ac and dc loads, though several components must be changed for dc operation, as will be explained later. A disadvantage of complementary-amplifier design is that for output voltages of 115 volts rms, output transistors with breakdown voltages (BV_{CEO}) approaching 400 volts would be required. The input and driver transistors would need 200 volt breakdown voltages. The primary advantage of this type design is the elimination of transformers and the use of direct coupling throughout, thus permitting usage with both ac and dc loads. This circuit is designed for applications requiring 20 volts rms, thus it does not require relatively expensive, high-voltage transistors.

For the circuit shown in Figure 5-9, transistors Q1 and Q2 form a differential-amplifier input stage. Resistor R2 and diodes D1 and D2 in the collector circuit of Q1 act as a clamp so that a low voltage transistor can be used for Q1. In a similar manner, the collector of Q2 is clamped by the base-emitter junction of Q3 and diode D3. again allowing the use of a low-voltage transistor. Q3 is connected in a common-base configuration to give a high source-impedance drive for Q4 and Q5, which helps minimize

Figure 5-9 — 20 Vrms 10 W Complementary Servo Amplifier

Servo and Audio Power Amplifiers
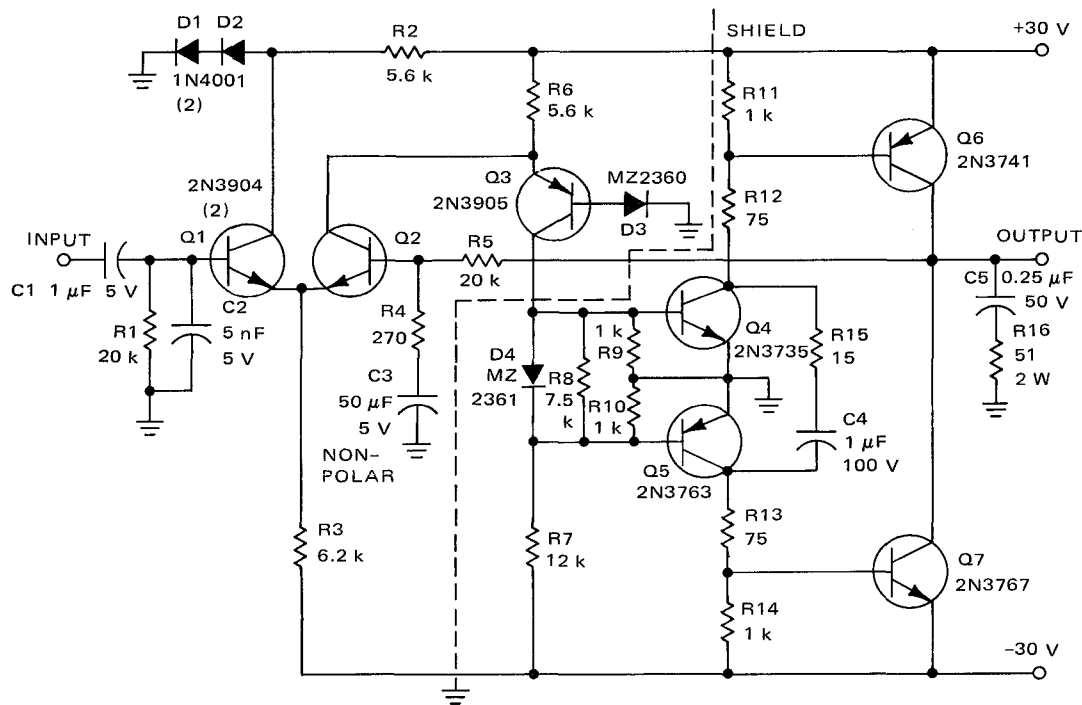
All resistors ±5%

Q6 and Q7 mounted on heat sinks with case-to-ambient thermal resistance ≤ 3.5°C/watt (each transistor)

Q4 and Q5 mounted on heat sinks with case-to-ambient thermal resistance ≤ 18°C/watt (each transistor)

crossover distortion. In addition, diode D4, a dual forward-reference diode, provides enough voltage to bias Q4 and Q5 into conduction. For low crossover distortion (low threshold or dead band for dc applications) the quiescent current through Q6 and Q7 should be between 3 and 5 mA. This value can be adjusted by R8, which shunts D4. For good thermal stability, diode D4 should be mounted directly on the heat sinks of Q4 and Q5, as close to the transistors as possible.

Although diode D3 helps clamp the collector voltage of Q3, it also serves a more useful purpose. Without D3, the base-emitter voltage drop of Q4 would tend to forward-bias the collector-base junction of Q3. Q3 would then operate in the saturation region, resulting in very nonlinear amplification. Diode D3 maintains the collector-base junction of Q3 in a reverse-biased condition.

Transistors Q4 and Q5 form a complementary driver pair which is direct coupled to the complementary output pair, Q6 and Q7. Through the use of positive and negative power supplies, the quiescent dc output voltage is approximately zero, so that the load can be direct coupled to the amplifier output.

Negative feedback is provided via resistor R5. R5 and resistor R4 set the ac voltage gain of the amplifier, which is given (approximately) by

$$A_V \approx \frac{R_4 + R_5}{R_4} \approx \frac{R_5}{R_4} \ .$$

As was mentioned, the complementary amplifier is capable of driving both ac and dc motors. The circuit in Figure 5-9 is for ac operation only. For dc, capacitors C1 and C3 must be shorted. The amplifier is then direct coupled from input to output. One point should be made, however; for ac operation, when R4 is returned to ground through C3, the dc closed-loop gain is approximately unity. This provides exceptionally good stability of the dc voltage at the output. With resistor R4 connected directly to ground, the dc stability will not be as good, since the closed-loop dc gain will now be equal to the ac gain. In addition, if the input is returned to ground through a low-impedance source, thereby effectively bypassing R1, there will be an offset voltage at the output due to the current flow through R5. (Ordinarily, in the ac amplifier, the offset voltage contributed by R5 is balanced out by the voltage drop across R1.) Increasing R4 and/or decreasing R5 will reduce the initial offset and increase the dc stability of the output for temperature changes. Of course, the gain of the amplifier will be reduced. Matching the two input transistors for current gain and base-emitter voltage drop, and matching the

current-gain products of the Q4-Q6 and Q5-Q7 pairs will help reduce the offset voltage and improve the dc stability of the output.

As can be seen, most of the transistors are working into rather low impedances (usually base-emitter junctions). Because of this, the amplifier has a rather wide open-loop bandwidth, and exhibits a strong tendency to oscillate. Several networks may be added to stabilize the amplifier, namely, C2, C4 and R15, C5 and R16, but the most effective way to eliminate oscillation is careful construction. All leads should be kept as short as possible. The input stages should be isolated from the output. It may even be necessary to use a shield as shown in the figure.

The operating temperature range of the amplifier is $-55°C$ to $+100°C$. The upper temperature is limited by plastic transistors Q2 and Q3, which have an upper junction temperature limit of $+135°C$. They were selected on the basis of economy. If $+125°C$ operation is required, Q1 and Q2 can be replaced by 2N3947's, and Q3 by a 2N3250. In addition, the case-to-ambient thermal resistance of the heat sinks for Q6 and Q7 must be decreased from a maximum of $3.5°C$/watt to $2.0°C$/watt.

The amplifier will drive 20 volts rms into a 40 Ω load, giving an output of 10 watts. The voltage gain of the amplifier at $25°C$ is 37 dB ± 1 dB. The gain variation over the $-55°C$ to $+100°C$ ambient temperature range is less than ±0.5 dB. Minimum power gain is 60 dB.

Input impedance of the amplifier is 15 kΩ and the output impedance is less than 1 Ω. Harmonic distortion is less than 5% at all levels up to 20 volt rms, and is typically less than 1% at $25°C$.

One final note: The voltage at the base of Q2 should not exceed approximately 1.2 volts positive with respect to ground. If this voltage is exceeded, the collector-base junction of Q2 will be forward-biased and Q2 will saturate. This sets a maximum value for the feedback ratio, namely,

$$\frac{R_4}{R_4 + R_5} \leqslant \frac{1.2}{V_{OUT(peak)}} \ .$$

5.4 115 Vrms, 7.5 W Complementary-Output AC Servo Amplifier

The servo-amplifier circuit shown in Figure 5-10 uses high-voltage transistors to deliver a high output voltage without using an output transformer. This reduces phase shift and its accompanying problems and permits the circuit to be operated from the 117 V ac line without a power transformer. The direct-coupled driver consisting of Q1 and Q2 is used because of its simplicity. It is decoupled from the high voltage supply by

R7 and C2. The feedback is brought through R8 to the emitter of Q2 for stability. If the feedback were returned to the base of Q1, R5 would have to be bypassed to provide adequate open-loop gain, and this increased gain, in conjunction with the phase shift across coupling transformer T1, could cause oscillation. The feedback around T1 reduces the effect of its nonlinearities, and sets the amplifier gain as determined by the following equation,

$$A \approx \frac{R3}{R4} \frac{R8}{R5} .$$

The RC network across the primary of the transformer is used to minimize phase shift at 400 Hz. This narrows the bandwidth, and care must be exercised that additional phase shift at higher frequencies does not cause instability.

Capacitors C4 and C5 in parallel with diodes D1 and D2 provide a drive path for the base currents of Q3 and Q4 when the diodes turn off. For example, refer to Figure 5-10, and note the drive loop for the base drive current indicated for Q3. The base drive current flows through D1 in the reverse direction. This is possible initially because the forward diode current supplied by R9 is greater than the base current. As transistor Q3 is further turned on, however, the voltage at point P_L goes positive and the voltage across R9 will decrease. Therefore, the current supplied by R9 will decrease until the forward current through D1 is less than the base current. When this happens, the diode will turn off and the base current must be driven through some other path. When D1 turns off, the impedance of the drive loop would change from approximately 200 ohms to approximately 50 kΩ if capacitor C4 were removed. This would require a very large drive signal to supply sufficient base current to Q3. The capacitors have there-fore been added to maintain a low-impedance drive path for Q3 and Q4. The addition of the capacitors has virtually no effect on the quiescent current.

The output stage is biased to prevent crossover distortion. Diodes were used to bias transistors Q3 and Q4 because the diode voltage drop tends to track the base-emitter drop, which simplifies the problem of thermal stability in the output stage. Due to the differences in the diodes and transistors, a value cannot be given for R9 and R10. They should be picked to produce a quiescent current of approximately 4 mA in Q3 and Q4, and will be approximately 20 kΩ to 50 kΩ. The quiescent currents I_{Q3} and I_{Q4} should be matched to minimize the dc offset voltage at the output. If I_{Q3} is not equal to I_{Q4}, the difference in the two currents will flow through R_L, causing a quiescent dc output voltage.
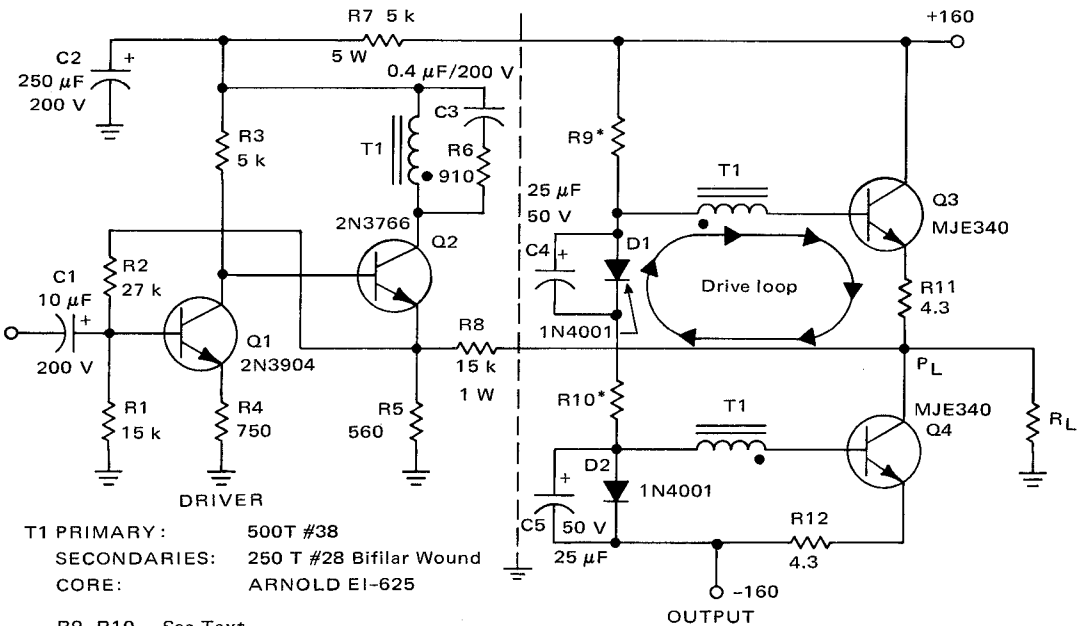
Figure 5-10 — 115 Vrms 7.5 W Complementary-Output Servo Amplifier

T1 PRIMARY: 500T #38
SECONDARIES: 250 T #28 Bifilar Wound
CORE: ARNOLD EI-625

R9, R10 — See Text
All Resistors ±5%

The frequency response is primarily limited by the transformer. The phase shift introduced by the transformer is critical because it limits the amount of feedback that can be used.

If a large amount of feedback is used and the transformer phase shift approaches 90°, there will be a strong tendency for the circuit to oscillate. If a wide bandwidth is required, three methods are available to reduce the phase shift of the transformer: (1) Increase the primary inductance of the transformer, (2) decrease the load reflected from the secondary, (3) decrease the impedance of the driving source.

Each biasing diode of the output stage should be mounted on the heat sink close to its associated output transistor so its voltage will track the transistor base-emitter voltage for thermal stability. The amplifier is designed to operate at 100°C if a heat sink of 3.5°C/watt is provided for each transistor. This heat sinking is adequate if R9 and R10 are adjusted so that quiescent current does not exceed 15 mA at 100°C.

The gain variation as a function of temperature is given in Figure 5-11. At −55°C the gain has dropped 1 dB, but from −20°C to +200°C the gain is within 0.25 dB. As can be seen from the plot of output voltage vs. input voltage in Figure 5-12, the amplifier has good linearity up to
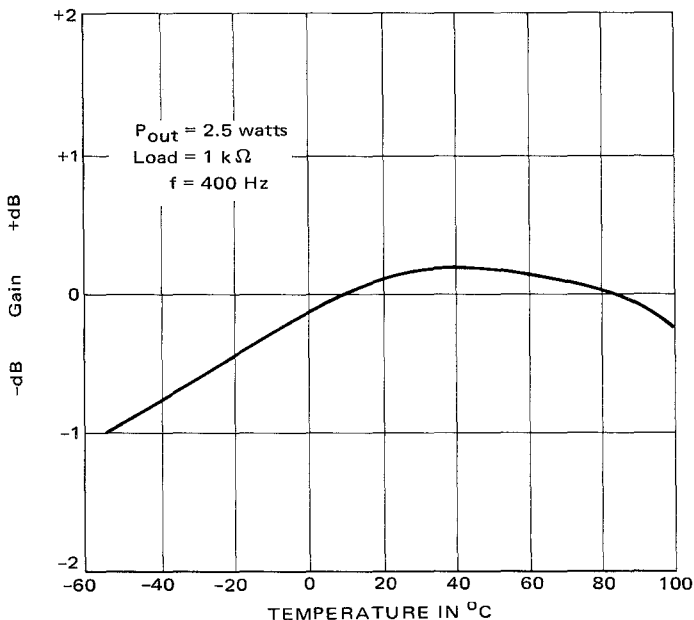
Figure 5-11 — Voltage Gain as a Function of Temperature
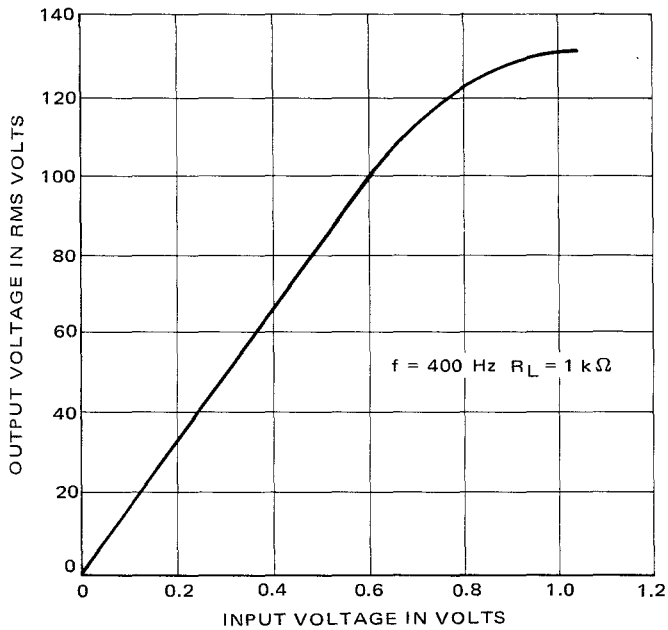
Figure 5-12 — Output Voltage versus Input Voltage
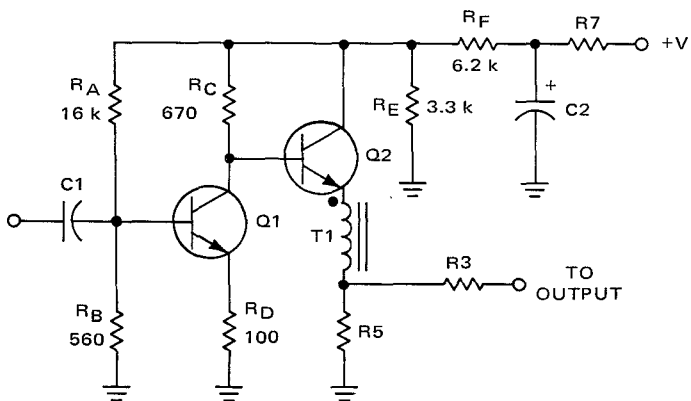
All Resistors ±5%.

**Figure 5-13 — Emitter-Follower Driver Stage for Circuit
Shown in Figure 5-10**

110 volts rms output. When driving a 7.5 watt, 115 volt rms servo motor, the output voltage changed approximately 2 volts from no load to stall conditions. The overall efficiency at 400 Hz with 90 volt rms output is 54.4%.

The frequency response of this amplifier is +0, −1 dB from 95 Hz to 2 kHz. The response is limited primarily by the secondary interwinding capacitance of T1 which causes the gain of the output stage to roll off at high frequencies. To overcome the capacitive effects, a low driving-source impedance such as the emitter follower shown in Figure 5-13 can be used. This driving circuit extends the 1 dB frequency response to 30 Hz and 12.5 kHz. The other operating parameters for the circuit using the emitter-follower driver are the same.

5.5 28 V, 28 Watt DC Servo Amplifier

The schematic shown in Figure 5-14 is that of a four-stage, direct-coupled servo amplifier capable of driving a 1 ampere load. The input stage is a differential amplifier formed by transistors Q1 and Q2. One input to this stage is the amplifier input and is applied to the base of Q1 while the other input to the stage is obtained from the amplifier output, as will be explained. The input resistors (R1 and R2) are used to give the same driving impedance to Q1 as R7 and R8 present to Q2. The emitters are connected to the negative supply through a constant-current source comprised of Q3, D1, R6 and R5. This constant-current source provides stable operation of the first differential-amplifier stage. The differential output voltage of the first stage is applied to the second stage (Q4 and Q5) which is a differential-to-single-ended converter. High-frequency roll-off is provided by the series combination of C1 and R10 between the collectors of Q4 and Q5. This roll-off is used to increase the stability margin of the stage.

The drive for the driver stage (Q6 and Q7) is obtained from the collector circuit of Q5. This connection provides a relatively high source impedance to Q6 and Q7, the power drivers. A practical way of achieving a low threshold or dead band area at a minimum of quiescent power dissipation, is to drive from a high-impedance source such as this. Diode D2 provides enough voltage to bias Q6 and Q7 into conduction. This complementary pair is directly coupled to the output pair, Q8 and Q9. The driver and the output stage are complementary devices since this keeps the design simple. There is no need for an additional stage of phase inversion and the two stages are readily adapted to direct coupling. Two complementary stages are used since this achieves maximum utilization of devices.

The use of positive and negative power supplies sets the quiescent output voltage within a few millivolts of zero. Negative feedback is pro-
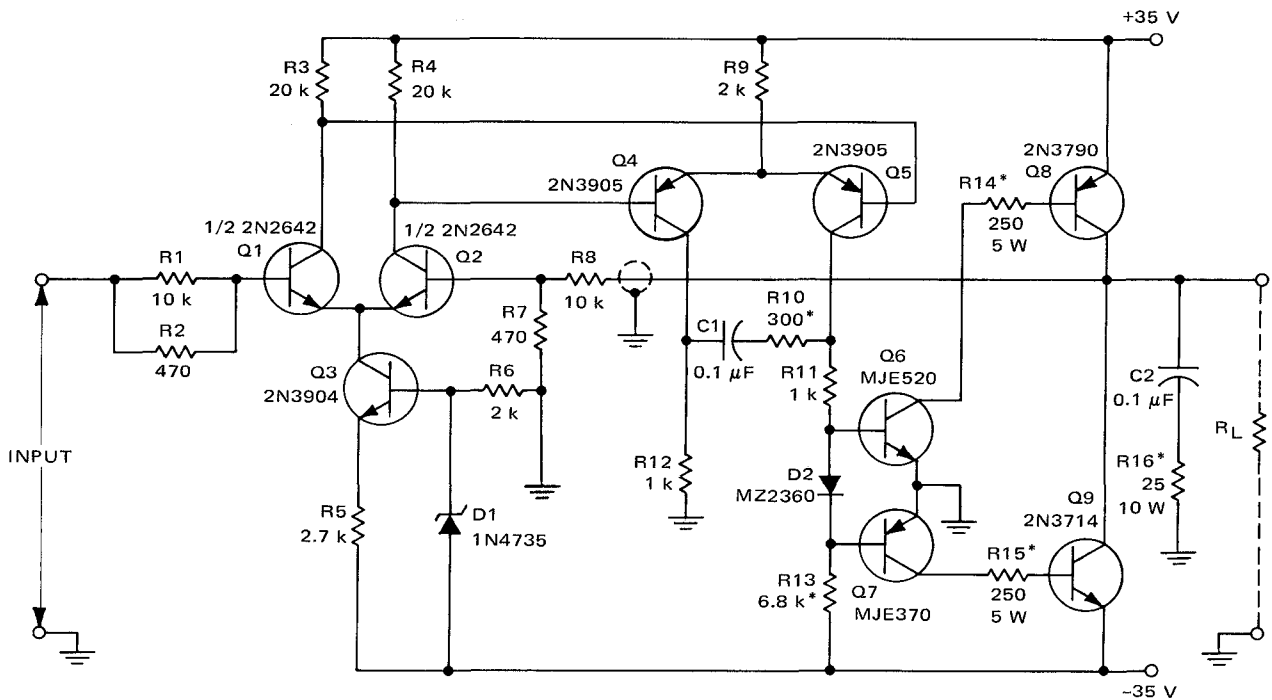
Figure 5-14 — 28 V 28 W DC Servo Amplifier

vided via resistor R8. The combination of R7 and R8 sets the voltage gain of amplifier. This is approximately given by

$$A_V \approx \frac{R7 + R8}{R7} \approx \frac{R8}{R7} \quad .$$

As a precaution against oscillation, resistor R8 should be mounted as close to Q2 as possible and the feedback from the output should be a shielded lead with the shield grounded at the end by R8 only.

Matching the two input transistors for current gain and base-emitter voltage drop, and matching the current-gain products of the Q6-Q8 and the Q7-Q9 pairs will help reduce the offset voltage and improve the dc stability of the output.

The gain of the amplifier is approximately 22 as shown by the gain characteristics plotted in Figure 5-15. The temperature characteristic of the offset voltage is shown in Figure 5-16. The change in offset is 26 mV for a temperature change from $-40°C$ to $+75°C$. The operating temperature range of the circuit is $-40°C$ to $+75°C$ with Q8 and Q9 mounted on a heat sink yielding a junction to ambient thermal resistance of $5°C$/Watt.
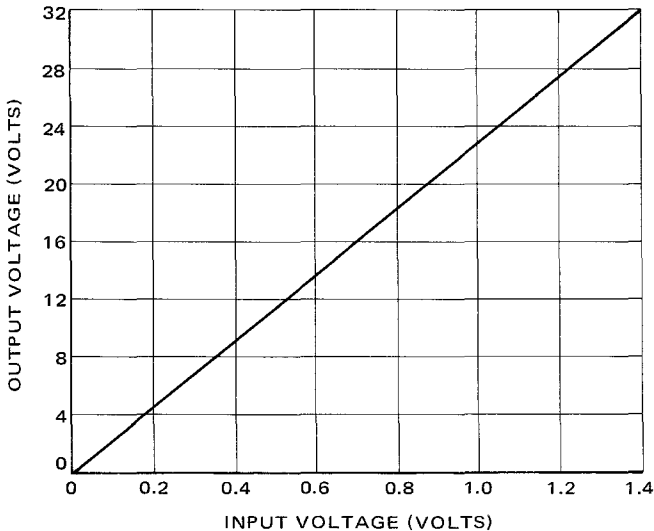
Figure 5-15 — DC Servo Amplifier Gain Characteristics
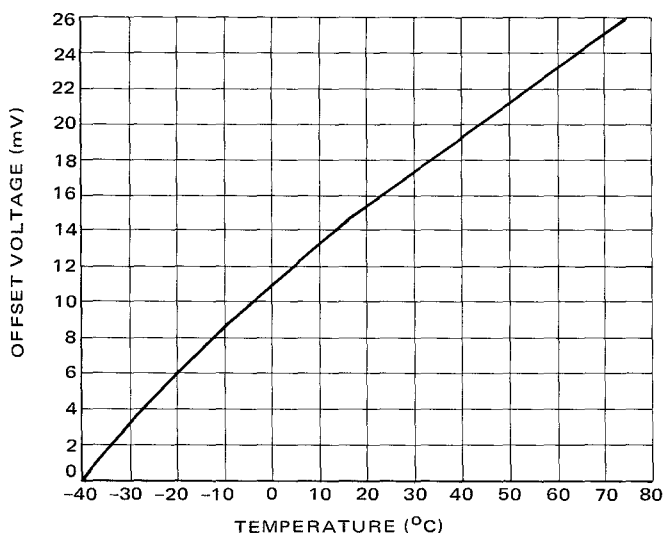
Figure 5-16 — DC Servo Amplifier Offset Voltage versus Temperature

5.6 Pulse-Width-Modulated DC Servo Amplifier

Figure 5-17 is a schematic of a pulse-width-modulated direct-coupled servo amplifier. The heart of the operation of this circuit is the input stage formed by Q1 and Q2. These two transistors and their associated resistors form a Schmitt trigger whose input voltage is that across capacitor C1. This capacitor is used with R1, R2 and R3 to integrate the voltages applied to these resistors. A reference voltage is applied to R2, the input voltage is applied to R1 and, as will be explained, the output voltage from the stage following the Schmitt is applied to R3. R2 is used to set the duty cycle of the output waveform at 50% when there is no input.

The output of the Schmitt drives Q3, which is used as a signal splitter to drive three stages. First of all, a feedback voltage is delivered to the input via D6, R7 and R3. The positive and negative excursions of the voltage are limited to the sum of D1, D2, D3, and D4. The net result of the integrator and the Schmitt trigger is a pulse-width-modulated signal at the output of Q3. The two other outputs of Q3 are the sources for the output driver stages. The output drivers and the output stages were split so the load voltage could easily be driven from the positive supply voltage. The signal splitting is accomplished by isolating diode D7. When Q3 is turned on, D7 is forward-biased and the voltage at the junction of R20 and R21 is

Figure 5-17 — Pulse-Width-Modulated DC Servo Amplifier
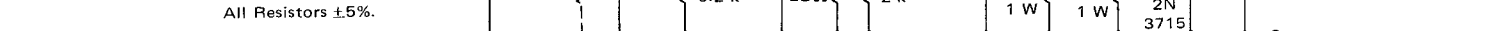
pulled to the diode drop plus the saturation voltage of Q3, below the positive supply. This removes base drive from Q4 and turns it off. At the same time base drive is applied to Q6 and turns it on. When Q3 is turned off, drive is removed from Q6, turning it off, and diode D7 is reverse-biased, which allows Q4 to turn on. Thus the drivers for the negative output, Q4 and Q5, are on when the drivers for the positive output, Q6 and Q7, are off, and vice versa. Capacitors C3 and C4 are used to provide a slight delay in the turn on of Q4 and Q6, respectively, so that Q8 and Q9 turn off before Q10 and Q11 turn on, and Q10 and Q11 turn off before Q8 and Q9 turn on. This assures that the output transistors will not be on at the same time, thus preventing excessive current and probable damage to them. When Q4 turns on, base drive is removed from Q5, turning it off. This turns the negative output stage (Q10 and Q11) off. Likewise, when Q6 is turned on, Q7 is turned off, which turns off the positive output stage, Q8 and Q9. Diodes D9 and D10 across the two output transistors are used to keep these devices in their safe operating area by limiting voltage transients; therefore they must be fast-recovery rectifiers. This amplifier is capable of driving a 5 ampere load at temperatures as high as 75°C if the output devices are mounted on a heat sink with a thermal resistance less than 10°C/watt. The operating frequency is approximately 3 kHz.

One final note: If the hysteresis of the Schmitt trigger is reduced, some temperature compensation may be required for a stable operating frequency. This can be provided by adding D5 and R13, and removing the ground at the junction of R9 and R12. The value of R13 should be chosen such that the temperature characteristic of D5 compensates for the changes due to temperature effects in the firing levels of the Schmitt trigger.

5.7 1 Watt Integrated-Circuit Audio Amplifier

The only active device in the 1 watt audio amplifier shown in Figure 5-18 is an integrated-circuit power amplifier. The device is designed to amplify signals as high as 300 kHz and to deliver one watt to a directly or capacitively coupled load. Three distinct voltage gains can be obtained from the device by simply changing several pin connections as shown in Table 5-1. The typical voltage gain ratios available are 10, 18 and 36 as measured with a 16 ohm load and a 16 volt supply. The power bandwidth for an output with less than 5% total harmonic distortion is lowest at a voltage gain of 36 and is typically 210 kHz.

When this device is used, care must be exercised to lay out the circuit properly. Due to the large bandwidth of the amplifier, coupling
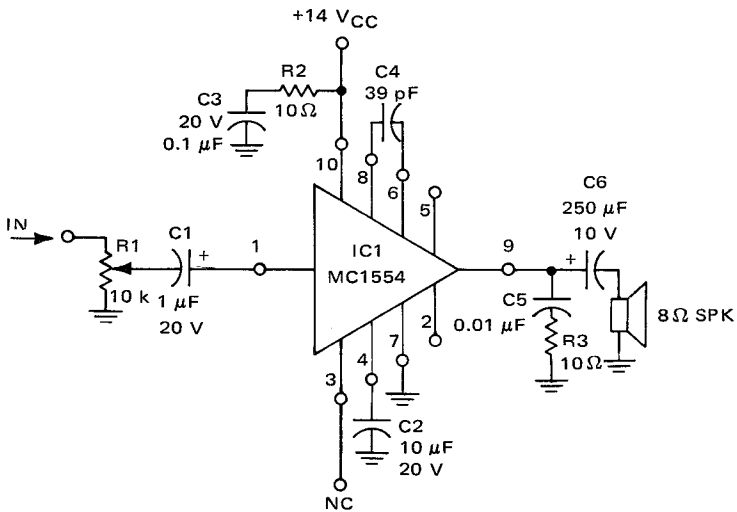
Figure 5-18 — 1 W Integrated Circuit Audio Amplifier

TABLE 5-1

VOLTAGE GAIN	PIN CONNECTIONS
V/V	
10	PIN 2 AND 4 OPEN
	PIN 5 A-C GROUND
18	PIN 2 AND 5 OPEN
	PIN 4 AC GROUND
36	PIN 2 CONNECTED TO PIN 5
	PIN 4 AC GROUND

must be avoided between the output and input leads. This can be minimized by either (a) the use of short leads which are well isolated, (b) narrow-banding the overall amplifier by placing a capacitor from pin one to ground to form a low-pass filter in combination with the source impedance, or (c) use of a shielded input cable. In applications which require upper band-edge control, the input low-pass filter is recommended. Also to avoid oscillations, an RC stabilizing network (R3 and C5) must be placed from the output (pin nine) to ground, with short leads, to cancel the effects of lead inductance to the load. Inductance of the power supply leads can also cause instability, thereby creating the necessity for R2 and C3. These components must be connected directly from pin ten to ground with leads as short as possible.

The total harmonic distortion from 20 Hz to 20 kHz for 1 watt into a 16 ohm load, with a voltage gain of 10, is typically 0.4%. The frequency response is shown by the curves of Figure 5-19, and the total harmonic distortion by the curves of Figure 5-20. The component values shown give a total harmonic distortion at 1 kHz of less than 1% at 1 watt output power into an 8 ohm speaker. The speaker is capacitively coupled to the
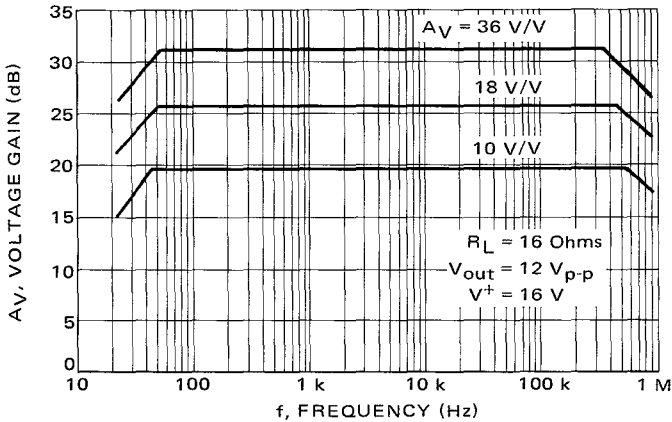
Figure 5-19 — Frequency Response of IC Audio Amplifier
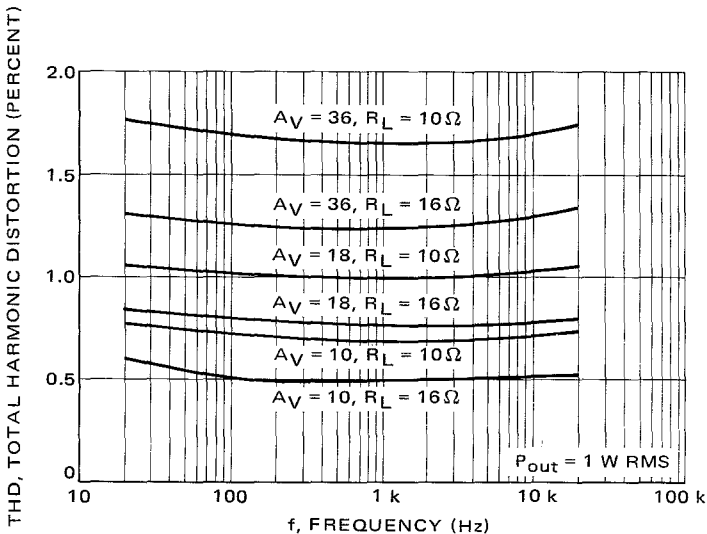
Figure 5-20 — Total Harmonic Distortion of IC Audio Amplifier

amplifier to eliminate quiescent dc in the speaker since a single-voltage power supply is used to drive the amplifier. The low harmonic distortion, low output impedance, excellent gain-temperature stability and selectable voltage gains makes this an excellent low-power audio amplifier.

5.8 4 Watt Wideband Amplifier

The amplifier whose circuit is shown in Figure 5-21 has a 100 kHz bandwidth and a minimum power output of 2 watts. The use of direct coupling and complementary transistors permits wide bandwidth and simplicity in design since there is no need for a phase-inversion stage. This also minimizes power requirements. Two complementary pairs are used, one as the driver and the other as the output stage.

The input stage, which consists of transistors Q1 and Q2, is a differential amplifier. The collector of Q2 drives the emitter of Q3, a common-base amplifier. Diode D1 in the base of Q3 overcomes the base-emitter drop of transistor Q4, thereby maintaining a reverse bias on the collector-base junction of Q3. This assures operation of Q3 in the linear region. Without D1, Q3 would operate in the saturation region resulting in very nonlinear amplification.

One of the basic difficulties encountered with class-B amplifiers is the elimination of crossover distortion. Generally the amplifier is biased

Figure 5-21 — 4 W Wide-Band Amplifier

somewhere between class A and class B (class AB), but this results in reduced efficiency due to the amount of quiescent power required. A more practical way is to drive from a high-impedance, or current, source. This gives low crossover distortion with a minimum of quiescent current. This amplifier offers a compromise between distortion and quiescent current. Dual forward diode D2 provides enough voltage to bias Q4 and Q5 just into conduction. In order to keep the quiescent current of Q6 and Q7 at a reasonable value (3-10 mA) from the standpoint of crossover distortion and efficiency, the voltage drop of D2 at 0.5 mA diode current should be between 1.05 and 1.15 volts. Diode D2 provides an added bene-fit since its variation over temperature changes will approximately match the base-emitter voltage changes of Q4 and Q5. The source impedance for Q4 and Q5 is effectively the value of R5, or 20 kΩ, since the output impedance of Q3, the common-base amplifier, is several megohms.

Transistors Q4 and Q5 form the complementary driver circuit and they are coupled directly to Q6 and Q7, the complementary output pair. Since the amplifier operates from positive and negative power supplies, and the quiescent output level is approximately 0 volts dc, the load can be directly coupled to the amplifier output.

Negative feedback is carried from the output by feedback resistor R3. This reduces the amount of distortion at the output, including that caused by crossover, resulting in a very clean output waveform. The ac gain of the amplifier is determined by resistor R6 which is decoupled from ground for dc by capacitor C2. The large amount of dc feedback gives a closed-loop dc gain of approximately one and results in exceptional dc stability of the output. This is an important consideration when driving speaker loads since any dc current in the speaker will cause an offset of the speaker cone, and may result in distorted output.

In order to assure a low dc offset at the amplifier output, and to keep distortion low, the beta product of the Q4-Q6 and Q5-Q7 pairs should be matched. The degree of matching required will depend upon the limits set for dc offset and distortion; i.e., the closer the beta-product match, the lower the offset and the distortion.

It should be noted at this point that certain precautions should be taken when building this amplifier. Because of the high gain-bandwidth product of the transistors used, there is a definite tendency for the ampli-fier to oscillate. To help minimize this tendency, all leads should be kept as short as possible. Also, an attempt should be made to isolate the input stages from the output with a shield. Capacitors C3 and C4 will also generally be necessary to provide sufficient ac stability to prevent oscillation.

The response of the amplifier is indicated in Figure 5-22. This shows

that the amplifier is capable of delivering 4 watts of output within +0, −1 dB from 45 Hz to 100 kHz. Harmonic distortion from 35 Hz to 20 kHz (the limit of the test equipment used) at an output of 4 watts is less than 1 per cent.
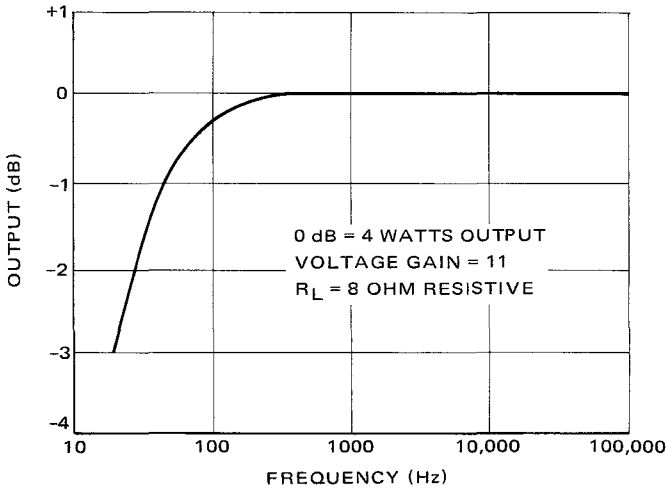
Figure 5-22 — Frequency Response Curve
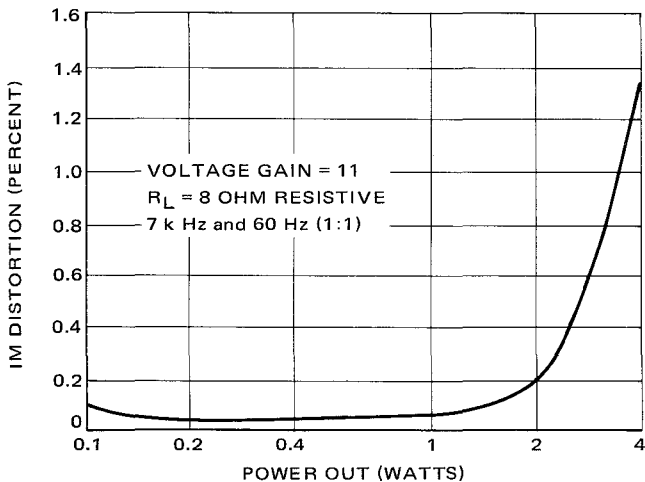
Figure 5-23 — IM Distortion versus Power Output
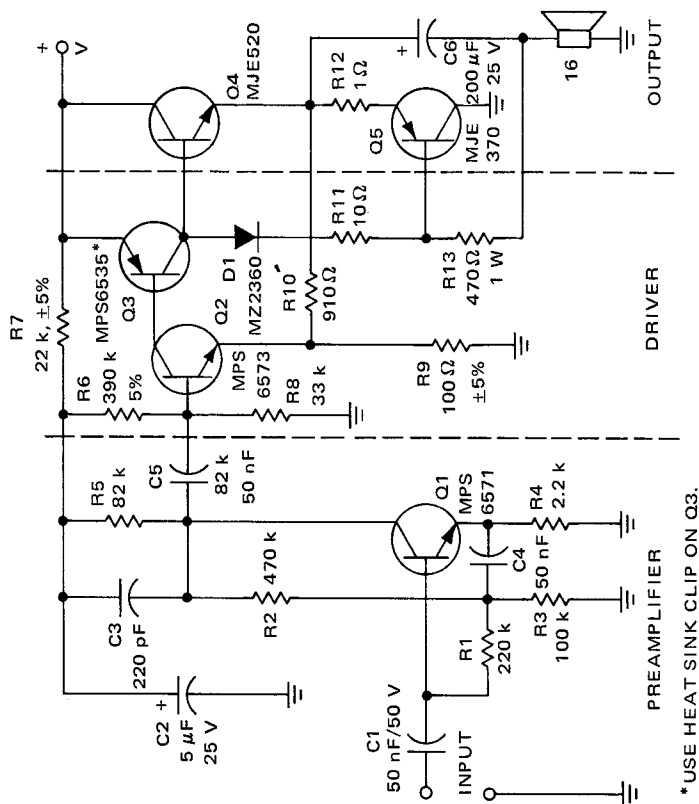
**Figure 5-24 — 5 and 10 W Audio Power Amplifiers. Values Shown
are for 5 W. For 10 W, Change Components
to Those Indicated Below**

C2	2 μF/30 V	Q5	MJE371
C6	200 μF/40 V	R7	51 k, ±5%
Q2	MPS6575	R10	1 k, ±5%
Q3	MPS6533	R13	680 Ω, 1 W
Q4	MJE521		

the amplifier, the selection and adjustment of parts values for R6, R7 and
R8 is straightforward.

Q2 acts as the common-emitter driver and, in practice, provides the
total open-loop voltage gain of the amplifier since the actual voltage gain
of Q1 is near one, and the output devices are emitter followers (also with
gains near unity). The dissipation of Q2 is inversely proportional to the
rated h_{FE} of the output transistors at the peak load current seen. Thus a
compromise is necessary between dissipation allowances for Q2 and the

h_{FE} specifications on Q4 and Q5. Bias for the complementary-symmetry output transistors is provided through D1 and its series resistor, R11. Because of the base-to-emitter voltage characteristics of the larger output devices, and the dc bias current passing through Q3 and its collector load network, it is not possible to use two silicon diodes for bias although this would be better. Therefore a diode-resistor combination is used, and is selected for threshold bias for the output devices. The dc collector load resistor for Q3 is bootstrapped by the output coupling capacitor so as to provide full ac current drive to the base of the PNP output transistor in peak signal excursions. This arrangement results in a small direct-current component through the loudspeaker voice coil, but its magnitude (about 2 or 3% of peak load current) is low enough that little or no displacement of the voice coil is experienced.

Q4 and Q5 are the complementary-symmetry emitter-follower output amplifiers. The feature of this complementary connection is that the drive signal is inherently split by the NPN-PNP combination to provide a push-pull output. The emitter-follower operation is obviously the easiest connection to use, and it is also the transistor configuration least affected by normal device parameter variations. The emitter-degenerative resistor required for thermal stability may be put in the emitter leg of the PNP transistor alone rather than being split between the output transistors since the effective dc path is identical, and the expected imbalance in ac performance is hardly measurable. The output coupling capacitor is the primary limitation on low-frequency response and may be changed as desired. The voltage rating for this component should be between half and full supply voltage since the normal dc bias (half supply) will see slow fluctuations of several volts due to the charge and discharge of stored energy when full output is being delivered at frequencies below about 100 Hz.

No specific load-fault protection has been incorporated in the circuit since the intended application is a sealed system. A shorted output will induce destruction of the circuit shown because the loss of feedback signal results in severe overdriving of driver and output stages, and an attempt to drive many times the normal peak load current into the short. The peak power will destroy both the output and driver stages. Fault protection can be incorporated at some increased expense by using a current-limiting scheme similar to that discussed in the 20, 35, and 60 watt amplifiers described in sections 5.11 and 5.12.

The operating characteristics of this amplifier are shown in Figures 5-25, 5-26, and 5-27. The input required for full rated output is 0.1 Vrms into 250 kΩ. The 16 ohm speaker was driven to 5 and 10 watts of output power for the circuit components given in the parts lists.
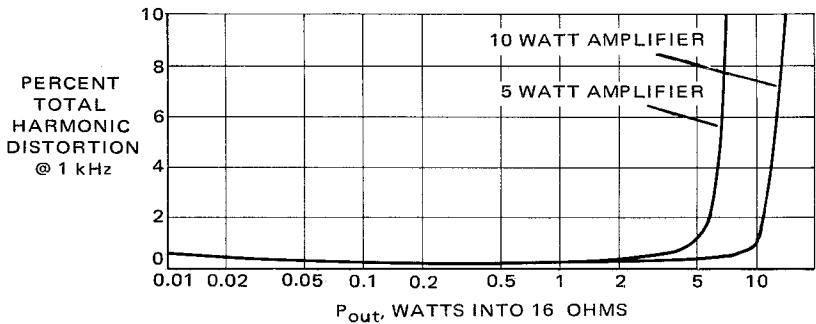
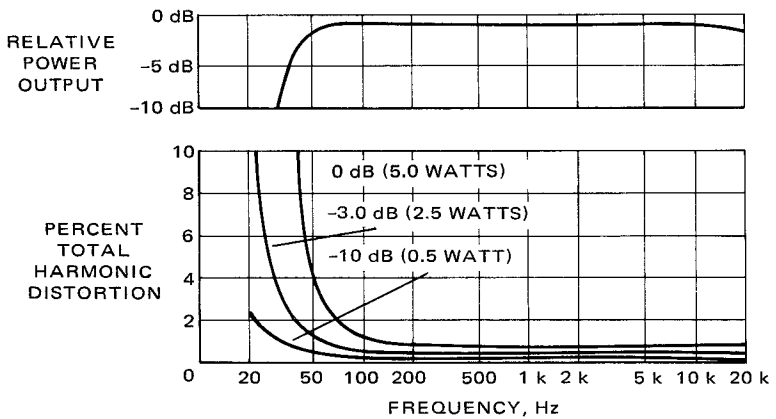**Figure 5-25 — Harmonic Distortion versus Power Output**

Figure 5-26 — 5 W-Amplifier Bandwidth and Distortion Characteristics
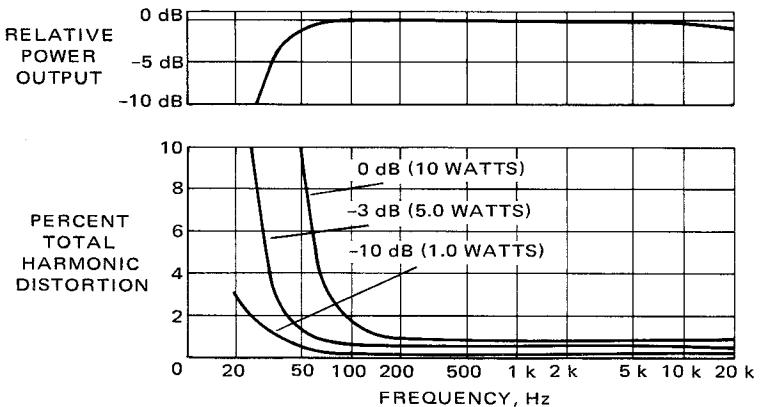
Figure 5-27 — 10 W Amplifier Bandwidth and Distortion Characteristics

5.10 High-Performance 10 Watt Audio Amplifier for 8Ω Load

The direct-coupled, class-B, complementary audio amplifier shown in Figure 5-28 provides 10 watts of music power output into an 8 ohm load with less than 1% distortion from 20 Hz to 20 kHz.

The input stage is a differential amplifier consisting of transistors Q1 and Q2. Biasing for this stage is achieved by resistor R2 (10 kΩ) and the feedback resistor R5 (10 kΩ). The collector of Q2 drives the emitter of Q3, which is connected as a common-base amplifier. The common-base configuration is used to provide high-source-impedance drive for the driver stage, which consists of Q5 and Q6. This high-impedance drive results in lower crossover distortion than low-impedance drive could provide. A common-base stage (Q4) is also connected to the base of Q6 to maintain the properties of the high-source-impedance drive. The base-emitter drop of Q5 sets the collector voltage of Q3 at approximately 0.6 volts. If the base of Q3 were grounded, the collector-base junction would be forward

*RBB — AS REQUIRED TO GIVE 15-30 MILLIAMPS QUIESCENT CURRENT IN Q7 AND Q8.

Figure 5-28 — High-Performance 10-Watt Amplifier

biased, resulting in very nonlinear amplification. Diode D1 offsets the base-emitter drop of Q5. Diode D2 in the base circuit of Q4 accomplishes the same purpose.

The emitter-base junctions of the three germanium transistors (Q9, Q10, and Q11) connected between the bases of Q5 and Q6 provide just enough forward bias to establish about 15 mA of quiescent current in the output transistors, Q7 and Q8. Resistors can be added in the emitters of Q5 and Q6, but there is a significant loss in amplifier gain before they have any appreciable effect on thermal stability. Better results can be obtained by using the transistor base-emitter junctions as shown in Figure 5-28. These three junctions track the base-emitter junctions of Q5 and Q6 very closely over a wide temperature range. Resistor R_{BB} is used to establish the quiescent currents in Q7 and Q8 between 15 and 30 mA so that crossover distortion is minimized and the upper operating temperature is not limited. Generally 500 to 1000 ohms will be adequate.

The complementary driver transistors, Q5 and Q6, are direct coupled to the complementary output pair, Q7 and Q8. Since the amplifier uses positive and negative power supplies, the quiescent output voltage is approximately 0 volts. This allows the output to be coupled directly to the load.

Negative feedback for the amplifier is provided by R5. Approximately 35 dB of feedback is used, resulting in very low distortion. The closed-loop gain of the amplifier is given by

$$Av \approx \frac{R_4 + R_5}{R_4} = 22.3$$

The gain can be varied by adjusting R_4.

Although the closed-loop ac gain of the amplifier is about 22, the closed-loop dc gain is approximately unity. The large amount of dc feedback results in exceptionally good dc stability of the output voltage.

Resistors R9 and R10 are used as current limiters for Q5 and Q6. Under normal operation, their effect is negligible. Capacitor C3 and the network consisting of C5 and R8 provide ac stability for the amplifier. It may also be necessary to add capacitor C4 from collector to base of Q3 to prevent spurious oscillations in the 50 to 70 MHz region. Obviously, any oscillation at this frequency could not be heard. In fact, it would encounter much difficulty in getting through the driver and output transistors. It can, however, contribute to distortion in the audible frequency region, and therefore is undesirable. Capacitor C6 was added to improve the transient
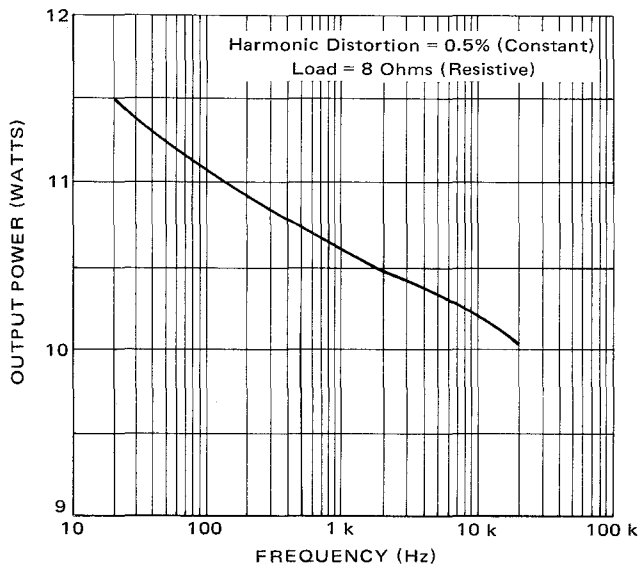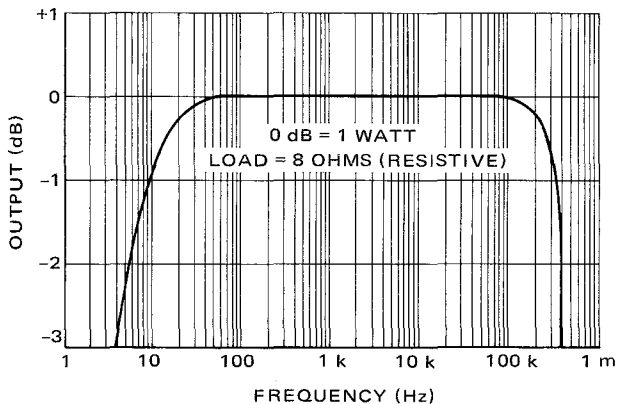
Figure 5-29 — Output Power versus Frequency
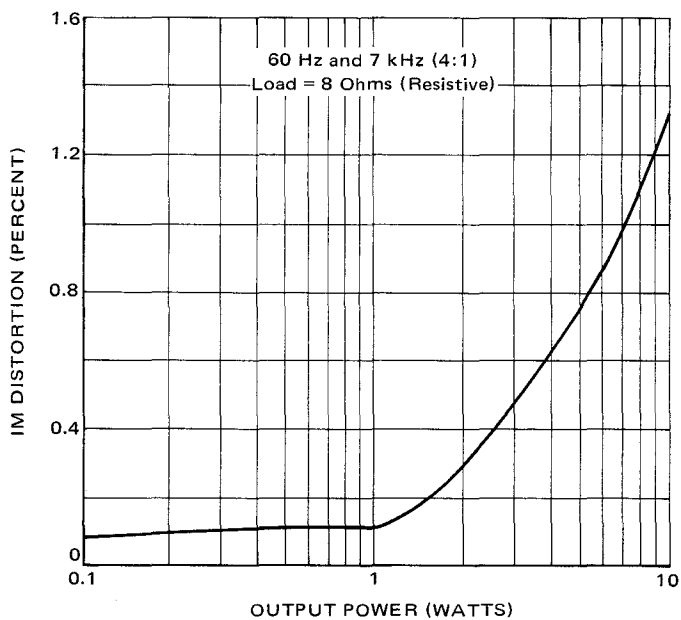

Figure 5-30 — Frequency Response

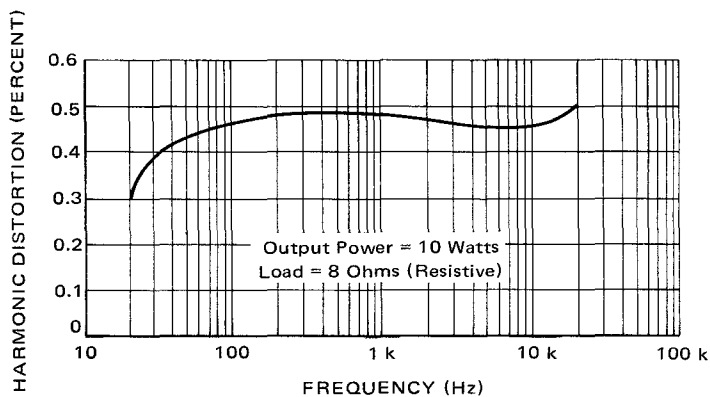**Figure 5-31 — Intermodulation Distortion**

Figure 5-32 — Harmonic Distortion

response characteristics of the amplifier. Resistor R1 was also added for this purpose.

No special heat sinking is necessary for transistors Q1 through Q6. For 100°C operation, the maximum case-to-ambient (θ_{C-A}) thermal resistance of Q7 and Q8 must be 10°C/watt per transistor.

As can be seen from Figure 5-29, the amplifier provides more than 10 watts at 0.5% total harmonic distortion (THD) from 20 Hz to 20 kHz. The frequency response at the 1 watt level (shown in Figure 5-30) is +0, −0.3 dB from 20 Hz to 20 kHz. It was down 1 dB at 9 Hz and 330 kHz. The −3 dB points were 4 Hz and 400 kHz. Phase shift was 16 degrees at 20 Hz and 4 degrees at 20 kHz.

Intermodulation distortion (IM) using 60 Hz and 7 kHz mixed 4:1 is shown in Figure 5-31. It is less than 0.5% at all levels up to 3 watts and increases to only 1.3% at 10 watts of output. Figures 5-32 and 5-33 show the harmonic distortion. Figure 5-32 is a plot of harmonic distortion versus frequency at a power output of 10 watts. It is less than 0.5% from 20 Hz to 20 kHz. Harmonic distortion versus output power at 1 kHz is shown in Figure 5-33.

Input impedance of the amplifier is approximately 10 kΩ, and is almost constant from 20 Hz to 20 kHz. The output impedance is 0.5 Ω at 1 kHz, 0.5 Ω at 20 Hz, and 0.25 Ω at 20 kHz.
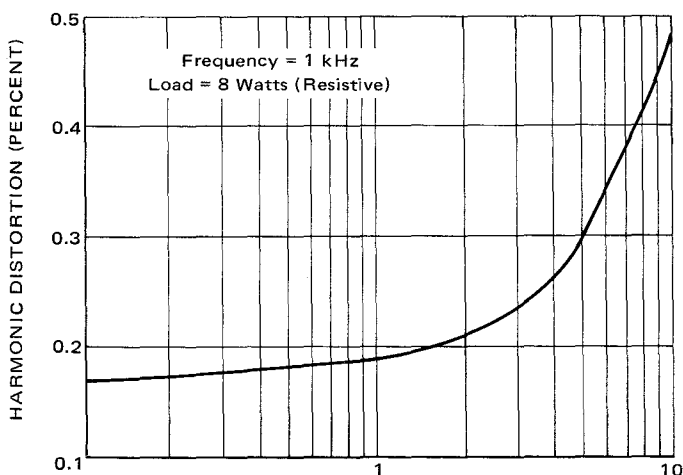
Figure 5-33 — Harmonic Distortion versus Output Power

5.11 20 Watt Audio Power Amplifier

The amplifier shown in Figure 5-34 will deliver 20 watts of music power to an 8 ohm speaker with a total harmonic distortion at rated output of less than 0.5% at 1 kHz. It has a typical 3 dB bandwidth of 40 Hz to 100 kHz. The input required for rated output is 0.3 volts rms into 100 kΩ.

The input transistor Q1 serves as a dc emitter follower through R6 to the output, and this makes the entire amplifier a dc emitter follower as
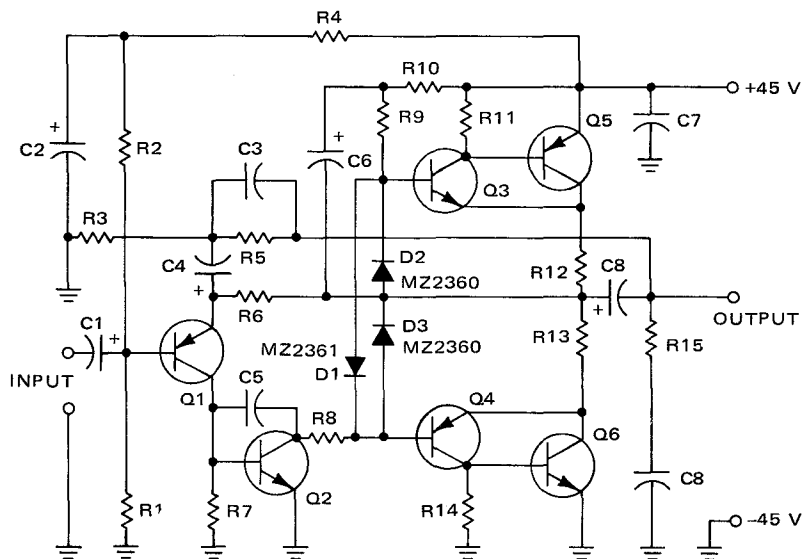
PARTS LIST:

C1	— 1 μF, 15 V	R1	— 150 k, 1/2 w, 5%
C2	— 1 μF, 50 V	R2	— 560 k, 1/2 w, 5%
C3	— 50 pF, 50 V	R3	— 100 Ω, 1/2 w, 10%
C4	— 5 μF, 15 V	R4	— 560 k, 1/2 w, 5%
C5	— 50 pF, 50 V	R5	— 5.6 k, 1/2 w, 10%
C6	— 50 μF, 15 V	R6	— 33 k, 1/2 w, 10%
C7, C8	— 0.1 μF, 50 V	R7	— 1.2 k, 1/2 w, 10%
Q1	— MPS6517	R8	— 100 Ω, 1/2 w, 10%
Q2	— MPS6575	R9	— 2.2 k, 1/2 w, 10%
Q3	— MPS6530 USE CLIP-ON	R10	— 2.2 k, 1/2 w, 10%
Q4	— MPS6533 HEAT SINK	R11	— 220 Ω, 1/2 w, 10%
Q5	— MJ-4931	R12	— 0.43 Ω, 1/2 w, 5%
Q6	— MJ-4831	R13	— 0.47 Ω, 1/2 w, 5%
		R14	— 220 Ω, 1/2 w, 10%
		R15	— 10 Ω, 1/2 w, 10%

Figure 5-34 — 20 W Audio Amplifier

well. Thus the bias voltages inside the amplifier are very stable; the input bias circuit of R1, R2, and R4 form a voltage divider which sets the desired voltage at the output driving point at 1/2 the supply voltage, to insure symmetrical signal clipping. Because of bypass capacitor C4 in the emitter of Q1, however, the amplifier has very high ac gain. Ac feedback from the output is applied to emitter resistor R3 so that the closed-loop ac gain of the amplifier is determined by the ratio of resistors R3 and R5. In the suggested circuit design given, the closed-loop gain is 50. Some ac feedback is applied through R6 so that the parallel resistance of R6 and R5 form the total effective feedback resistor, although the contribution of R6 is small.

Q2 functions as a large-signal class-A driver. Since the output circuit is effectively an emitter follower, Q2 must swing the entire load voltage. As it operates with a voltage gain of some 60 dB or better, it also provides a convenient point for setting the dominant high-frequency pole in the amplifier's open-loop Bode response. A small capacitor from collector to base, reflected back to the base driving point as a high Miller capacitance, is sufficient to locate the pole conveniently far away from the basic turn-over frequencies of the input and output circuits. Thus the amplifier has a dominant -6 dB/octave rolloff characteristic above about 50 kHz.

The collector load of Q2 consists primarily of the output load (the 8 ohm speaker) multiplied by the product of the h_{FE}'s of the transistors in the complementary Darlington circuit. Because of the bootstrapping by capacitor C6, the effective impedance presented to Q2 by R9 and R10 is much higher than the reflected speaker load. The main purpose of the bootstrapped output is to provide the drive to the upper Darlington amplifier (Q3 and Q5) on peak signal excursions.

Q3 and Q4 appear as emitter followers between the load and the driver. Q5 and Q6, the complements of Q3 and Q4, operate effectively with their drivers as Darlington pairs. One may analyze the output circuit as an amplifier with two cascaded common-emitter transistors (the NPN into the PNP, and vice versa) with total feedback for unity overall gain. Thus a basically linear, local amplifier is created: the output voltage equals the input voltage. The point to this complementary Darlington connection is to have the base-emitter junctions, which are biased through D1, control the response of the output circuit to the exclusion of the power transistor characteristics. Degenerative resistors R12 and R13 are located in the effective emitter of the complementary Darlington amplifier, and provide feedback for bias control purposes; this results in excellent thermal stability. Since the drivers are biased from a dual forward voltage reference diode, the chosen idling bias for the output circuit is fully compensated

for ambient temperature variations. It needs no adjusting potentiometers and is largely insensitive to variations in supply voltage.

Diodes D2 and D3 provide fault protection. D1 and D3 effectively appear in series between the base of Q3 and the output on the positive half cycle of output signal, and act to limit shorted-output fault current to a value slightly above the normal peak load current. Similarly, D2 appears in series with D1 for the negative half cycle of output signal.

The low-frequency response is limited by the output coupling capacitor. A larger value will extend the low-frequency limit. Any high-frequency roll off should be set at the input since any attempt to place the roll off inside the feedback loop will increase the distortion and decrease the àc-stability margin of the amplifier.

Figures 5-35 and 5-36 show the harmonic and intermodulation-distortion characteristics of the amplifier.
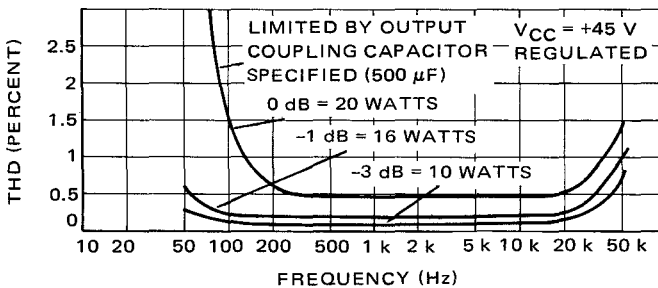

Figure 5-35 — Distortion versus Frequency Response
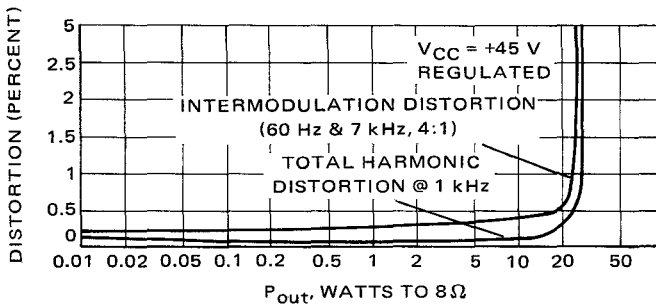

Figure 5-36 — Linearity and Clipping Characteristics

5.12 35 and 60 Watt Audio Power Amplifiers

The circuit shown in Figure 5-37 will provide either 35 or 60 watts of music power to an 8 ohm speaker load, depending on the circuit com-
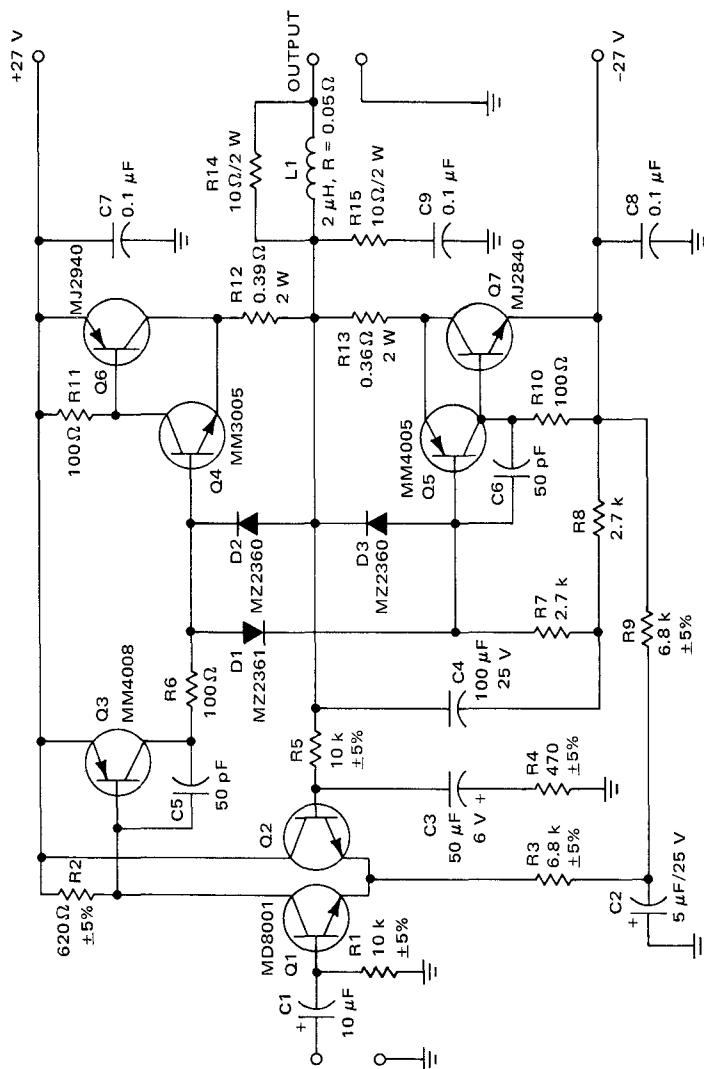
**Figure 5-37 — 35 and 60 Watt Audio Power Amplifier.
Values Above are for 35-W Amplifier. For 60 W,
Change Components Listed Below.**

D2, D3	—	MR2361	R3, R9	—	7.5 k ±5%
Q1, Q2	—	MD8002	R7, R8	—	3.9 k
Q3	—	MM4009	R12	—	0.47 Ω, 2 W
Q4	—	MM3006	R13	—	0.43 Ω, 2 W
Q5	—	MM4006	+V	—	+36 V
Q6	—	MJ2941	−V	—	−36 V
Q7	—	MJ2841			

Figure 5-38 — Intermodulation Distortion (60 Hz & 7 kHz Mixed 4: 1) versus Power Output

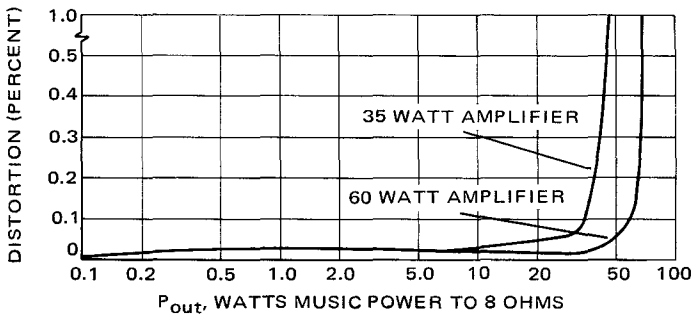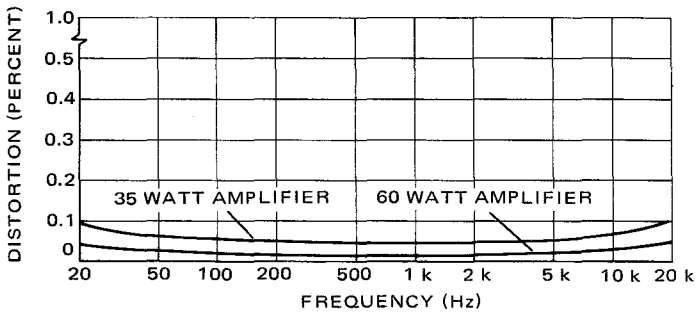

Figure 5-39 — Total Harmonic Distortion versus Frequency @ Rated Output to 8 Ohms

ponents used. The 35 watt amplifier provides less than 0.1% total harmonic distortion and less than 0.2% intermodulation distortion at full power output. It has a 3-dB bandwidth of 10 Hz to 100 kHz. The intermodulation distortion for the 60 watt amplifier is typically less than 0.15% with its other performance specifications the same as the 35 watt amplifier. Both amplifiers require an input of 1.0 volt rms into 10 kΩ for full power output. Figures 5-38 and 5-39 give the intermodulation and harmonic distortion.

Q1 and Q2 form a differential amplifier which is used because the inherent temperature-compensated balance of both base legs and the controlled symmetry of voltages in those legs provide excellent thermal stability. Q1 is biased through a resistor in its base leg to ground; Q2, biased through a similar resistor, will cause the output to be at equivalent ground because of the symmetry of the two halves of the local circuit. The error voltage appearing as an offset at the output consists of the differential base-emitter voltage between Q1 and Q2 and the differential voltage estab-

lished by the base currents flowing through the bias resistors (R1 and R5). When Q1 and Q2 are supplied as a dual transistor, typical offsets of less than 10 mV are realized at the output.

Q1 also serves as a common-emitter amplifier to the input signal, and uses the common-base input impedance of Q2 as its effective emitter load. Constant-current emitter bias for Q1 and Q2 is supplied through the common-mode path (R3 and R9), and decoupling is provided in this path for optimum rejection of power-supply ripple.

Q3 functions as a common-emitter driver. Since the output circuit (as will be explained) is only an emitter follower, Q3 must swing the full load voltage plus the losses in saturation and transconductance of the output devices. This stage commonly operates with a voltage gain of 60 dB or better, and thus provides a convenient point for setting the single dominant high-frequency pole in the amplifier's open-loop Bode response. A small capacitor (C5) from collector to base, as multiplied by the Miller effect, appears back at the base driving point to set this pole conveniently far away from the basic turnover frequencies of the input and output circuits, so that the amplifier as a whole demonstrates a basic 6 dB/octave roll off above about 50 kHz. In this way, a generous amount of feedback can be applied to the circuit to optimize linearity as desired while still maintaining a comfortable stability margin.

The dc bias path for Q3 proceeds through a small collector resistor R6 (used to limit the power dissipation in Q3 under load-fault conditions), through biasing diode D1 and dc collector load resistor (R7 and R8). This latter resistor is split and bootstrapped through C4 so that full alternating current drive may be available to the lower half of the output circuit during peak signal excursion. Because of this bootstrapping, the effective ac load seen by Q3 is the reflected loudspeaker load impedance (i.e., load R_L multiplied by the product of the h_{FE}'s of the complementary Darlington output devices). R5, the dc base resistor for Q2, also serves as the loop feedback resistor by providing the feedback signal to the base of Q2. The ratio of R5 to R4 approximately determines the closed-loop voltage gain of the amplifier. Capacitor C3 in series with R4 is chosen to provide the desired low-frequency roll off and force the amplifier to demonstrate unity dc voltage gain for optimum bias stability.

The output circuit is a full complementary-symmetry Darlington emitter follower. The emitter-follower connection is established by Q4 and Q5, and is chosen because it provides the best basic linearity among the three possible transistor configurations, and is the least critical for normal variations in device parameters. The availability of complementary-symmetry silicon power transistors has allowed the practical development

of the output circuit given, which is an improvement over the quasi-complementary connection with its many bias problems. The simplicity and advantage of a full complementary symmetry amplifier is seen in the way Q4 and Q5, biased with dual diode D1, completely control the output circuit. Transistors Q4 and Q6, and transistors Q5 and Q7 each form a local unity-gain dc operational amplifier. Complete control of the output is thus given to the drive voltage appearing at the bases of Q4 and Q5. Since bias is established by diode reference, excellent compensation and control through variations in line voltage and operating temperature are assured, and this represents a significant improvement over the bias schemes used with the quasi-complementary circuit.

Diodes D2 and D3, in conjunction with D1, provide load-fault protection. D1 and D3 will effectively appear in series between the base of Q4 and the output on the positive half cycle of output signal, and will act to limit the peak fault current to a chosen value (through R12 and R13) slightly above the normal peak-load current by limiting the voltage that can be developed between Q4's base and the output. Similarly, D2 in conjunction with D1, provides fault protection for the negative half cycle by operating between the output and the base of Q5. The peak fault current is kept to a value within the safe-area capability of the output devices (provided enough heat sinking is available) so that overload interruption may be accomplished by thermal cutouts or slow-blow fuses.

Some additional compensation networks appear at the output. Among these is a standard RC pad across the output, which inserts a stabilizing pole and zero into the amplifier's Bode response, and a choke in series with the load. The latter is a low-resistance rf choke used to provide a buffer between the amplifier and a capacitive load. A capacitor across the output of many amplifiers will upset stability and cause oscillation because it provides an uncompensated pole in the Bode response, usually at just the wrong place. A small inductance (about 2 μH) is used to buffer out this effect.

It is also advisable to provide power-supply bypass capacitors as close to the circuit as possible. These capacitors are used to provide very-high-frequency, return-path shorts to the amplifier. The high load currents flowing through long power-supply leads (consisting of series R and L) generate enough of a ground loop at very high frequencies to cause the amplifier to oscillate. Disc-ceramic bypass capacitors are chosen as needed to eliminate this source of parasitic oscillation.

5.13 50 Watt Audio Power Amplifier

Figure 5-40 is the circuit diagram of a 50 watt (music power) audio
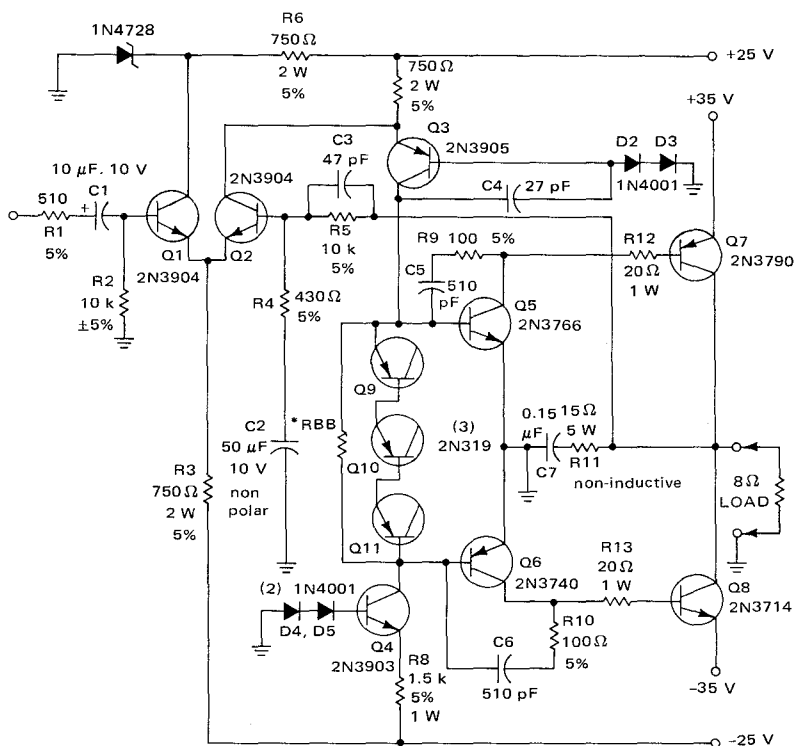
*RBB — AS REQUIRED TO GIVE 20-40 mA QUIESCENT CURRENT IN Q7 AND Q8.

Figure 5-40 — 50 Watt Amplifier

power amplifier. Except for a few minor changes, it is the same circuit used for the 10-watt high performance audio amplifier discussed in section 5.10. A zener diode (D1) has been added in the collector of Q1. This is used to limit the voltage applied to Q1, thus allowing the use of a low-cost plastic-encapsulated transistor. The zener is also a low-cost, 1 watt Surmetic device. Two diodes are used in the base circuits of Q3 and Q4, instead of the one used in the 10 watt amplifier, because of the larger dynamic voltage swings of the base-emitter junctions of Q5 and Q6 as compared to those in the 10 watt amplifier. Also, it was found that a single capacitor from collector to collector of Q5 and Q6 was not adequate for good transient response, so a 510 pF capacitor in series with a 100 ohm resistor was added from each collector to base (C5 and R9, and C6 and R10).
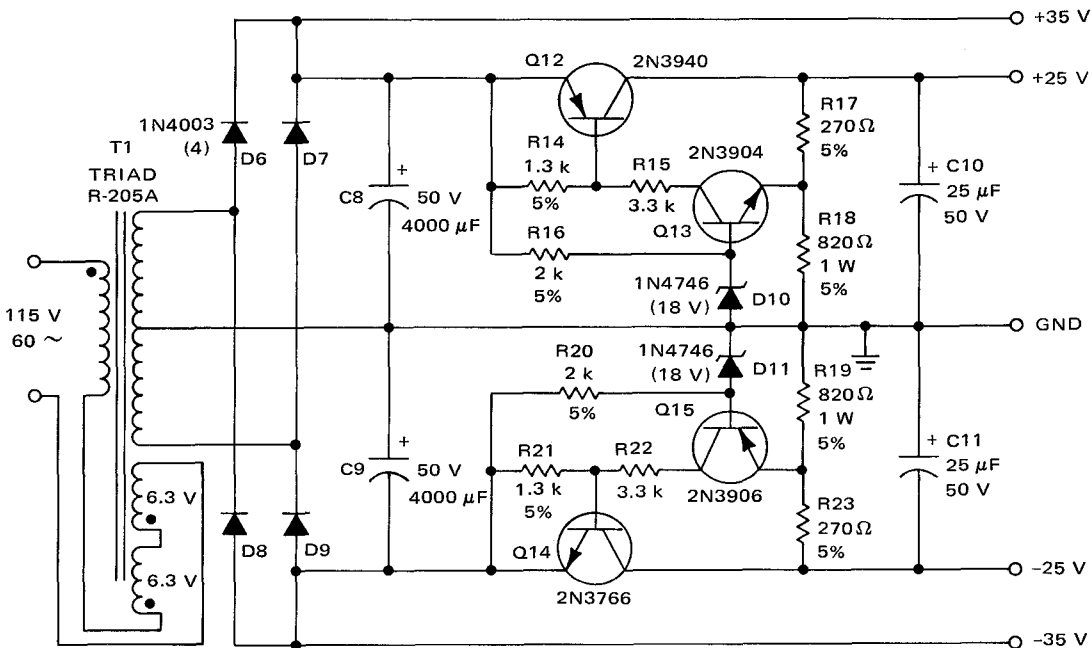
Figure 5-41 — 50 Watt Amplifier Power Supply

No heat sinks are required for transistors Q1 through Q4. The maximum case-to-ambient thermal resistance (θ_{CA}) of transistors Q5 and Q6 is 8°C/watt (per transistor) for 100°C operation. For Q7 and Q8 the maximum case-to-ambient thermal resistance is 2.5°C/watt (per transistor). In addition, for good tracking, transistors Q5 and Q6 should be mounted on one heat sink along with the three germanium transistors Q9, Q10, and Q11.

The power supply used for testing the 50 watt amplifier is shown in Figure 5-41. Filtered, but unregulated, voltage is supplied to the two output transistors. The zener-regulated supplies provide ±25 volts to the input stages. The power transformer used in this supply has a secondary voltage that is too high. This is the reason for using the two 6.3 volt windings as bucking voltages. The filtered voltage that results is ±38.5 volts at 0 watts out of the amplifier and ±34.5 volts at 50 watts output. If a power transformer were to be designed for the amplifier, better performance and lower distortion would be realized if the no-load output voltage were 39 volts, dropping to no less than 36 volts at 50 watts output. Higher voltages, though reducing distortion even further, could lead to breakdown in the output transistors. Also, power dissipation in the output transistors (and the drivers as well) would increase, thereby reducing the maximum operating temperature of the amplifier.

The ideal supply to use would be one that is completely regulated at approximately 36 volts. If ±36 volt regulated supplies are available, the
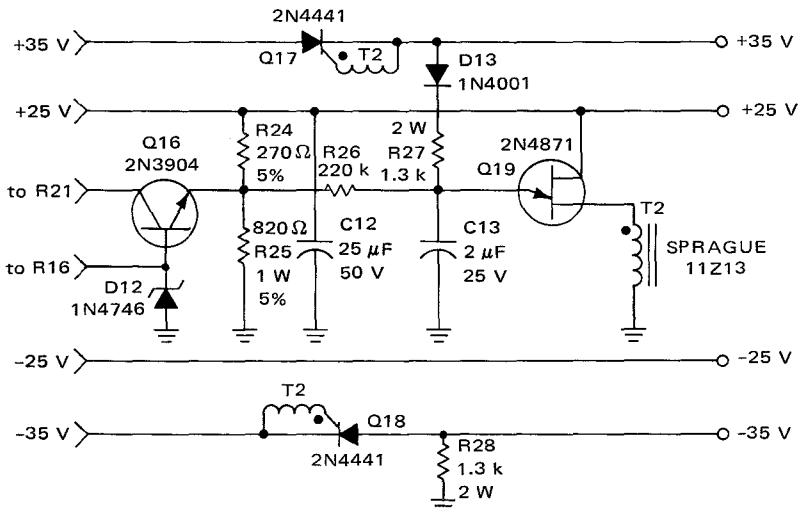

Figure 5-42 — 50 W Amplifier Delayed Switching Circuit

input and output stages can be operated from the same supplies by changing R3 and R7 to 1.3 kΩ, R6 to 1.1 kΩ, and R8 to 2.7 kΩ.

Two problems related to the power supplies arise in this amplifier. The first is that both the positive and negative supplies should come on at the same time. If power is applied to only one side, the driver-output pair associated with that side will conduct very high currents, possibly damaging one or both transistors due to high power dissipation.

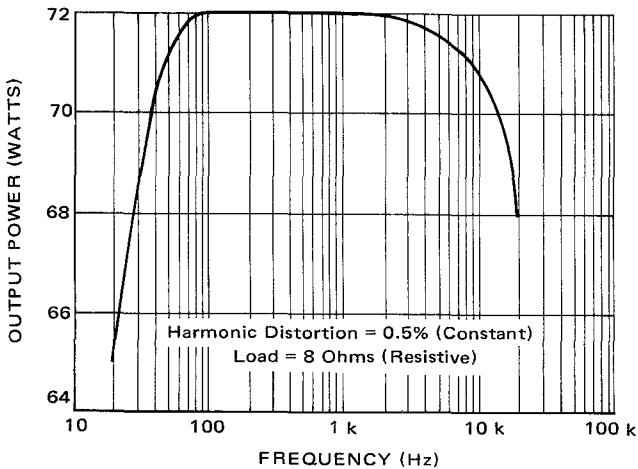The second problem arises primarily because of capacitors C1 and

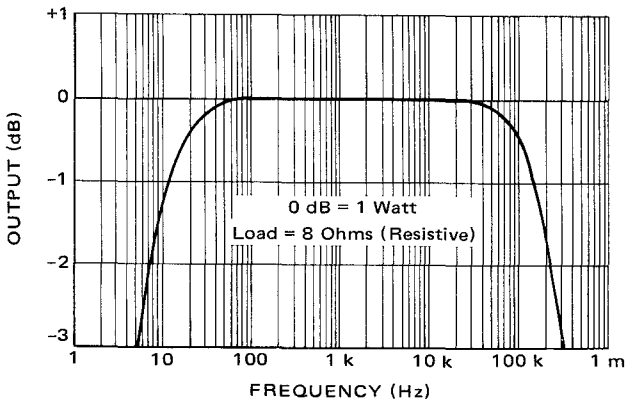**Figure 5-43 — 50-W-Amplifier Output Power versus Frequency**

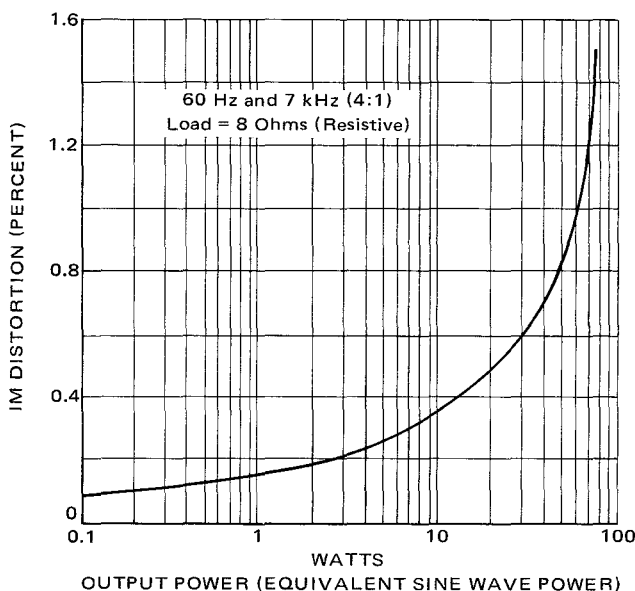**Figure 5-44 — 50-Watt-Amplifier Frequency Response**

Figure 5-45 — 50-Watt-Amplifier Intermodulation Distortion

C2 in the amplifier. When power is applied, the charging of these capacitors causes some high-magnitude transients in the amplifier output. To eliminate or minimize this, the voltages to the input stages should be applied a few hundred milliseconds before the voltages to the output stage. This allows the capacitors to become fully charged. One way to do this would be with two switches or a two-position rotary switch. It could also be done electronically. One possible circuit is shown in Figure 5-42. The RC timing network (R26 and C13) is connected at the junction of the voltage divider, at the output of the positive regulated supply, thus assuring that the timing sequence won't begin until this supply is up to voltage. When the capacitor C13 has charged to the breakover voltage of the unijunction transistor, Q19, the unijunction conducts and discharges C13 through T2. This results in a pulse being applied, via T2, to the gates of SCRs Q17 and Q18, turning them on. The result is the required delay before voltage is applied to the output stage.

Output power capability of the amplifier at a constant total harmonic distortion of 0.5% is shown in Figure 5-43: the amplifier can provide 65 watts at 20 Hz, 72 watts at 1 kHz, and 68 watts at 20 kHz, at 0.5% THD.

Figure 5-44 is a plot of the frequency response at 1 watt output. It is almost flat from 100 Hz to 20 kHz, being down only 0.35 dB at 20 Hz.
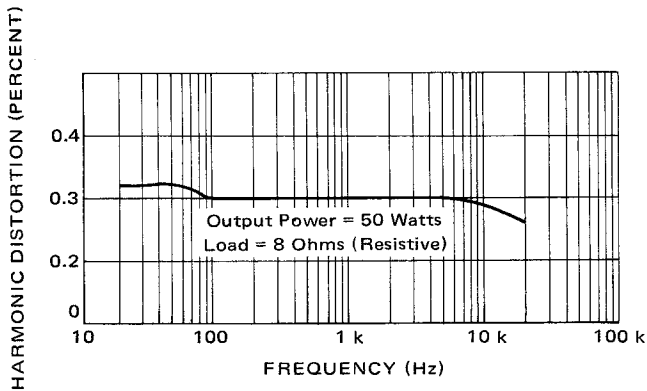
Figure 5-46 — 50-Watt-Amplifier Harmonic Distortion
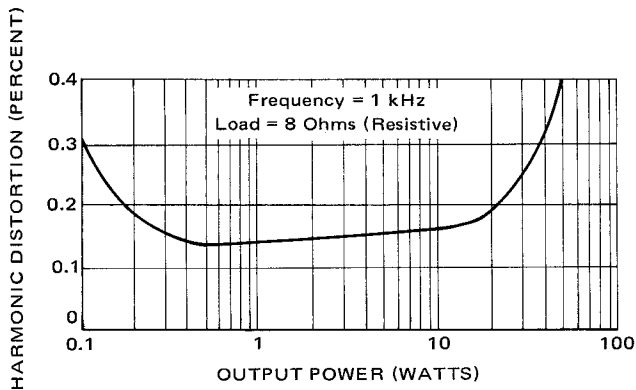
Figure 5-47 — 50-W-Amplifier Harmonic Distortion versus Output Power

The −1 dB points are at 11 Hz and 150 kHz, and −3 dB response is reached at 5 Hz and 320 kHz. Phase shift is 18 degrees at 20 Hz and 6 degrees at 20 kHz.

The results of intermodulation-distortion tests can be seen in Figure 5-45. IM is about 0.1% at 0.1 watt output, rising steadily to 0.5% at 20 watts output, and reaching 0.85% at 50 watts output. The output capability at 1.5% IM is approximately 65 watts.

Harmonic distortion is shown in Figures 5-46 and 5-47, Figure 5-46 being a plot of harmonic distortion versus frequency at 50 watts output and Figure 5-47 a plot of harmonic distortion versus power output at 1 kHz.

Input impedance of the 50 watt amplifier is approximately 10 kΩ

from 20 Hz to 20 kHz. Output impedance is less than 0.1 ohm from 20 Hz to 20 kHz.

As a final note it should be mentioned that this amplifier can be used as a dc amplifier with excellent stability. Minor circuit modifications are required to achieve this. The resulting circuit is shown as the 25 volt, 28 watt, dc servo amplifier of Figure 5-14.

BIBLIOGRAPHY FOR CHAPTER 5

1. Freyling, Nick, "A 4-Watt Wide-Band Solid-State Amplifier," Motorola Application Note AN-209.

2. Freyling, Nick, "Complementary Solid-State Audio Amplifiers," Motorola Application Note AN-230.

3. Freyling, Nick, "High-Performance All Solid State Servo Amplifiers," Motorola Application Note AN-225.

4. Landee, R. W.; Davis, D. C.; Albrecht, A. P., *Electronic Designers Handbook*, McGraw Hill, New York, 1957.

5. Long, E. L., "Low-Cost High-Voltage Servo Amplifiers," Motorola Application Note AN-241.

6. Terman, Frederick, *Electronic and Radio Engineering*, 4th Edition, McGraw-Hill, New York, 1955.

CHAPTER VI

Thyristor and Transistor Switches

6.1 Thyristor Zero-Point Switches

Zero-point switches are highly desirable in many applications because they do not generate electro-magnetic interference (EMI). A zero-point switch controls sine-wave power in such a way that either complete cycles or half cycles of the power-supply voltage are applied to the load as shown in Figure 6-1. This type of switching is primarily used to control power to resistive loads such as heaters. It can also be used for controlling the speed of motors if the duty cycle is modulated by having short bursts of power applied to the load and the load characteristic is primarily inertial rather than frictional. Modulation can be on a random basis with an on-off control, or on a proportioning basis with the proper type of proportioning control.

In order for zero-point switching to be effective, it must be true zero-point switching. If an SCR is turned on with an anode voltage as low as 10 volts and a load of just a few hundred watts, sufficient EMI will result to nullify the advantages of going to zero-point switching in the first place. The thyristor to be turned on must receive gate drive exactly at the
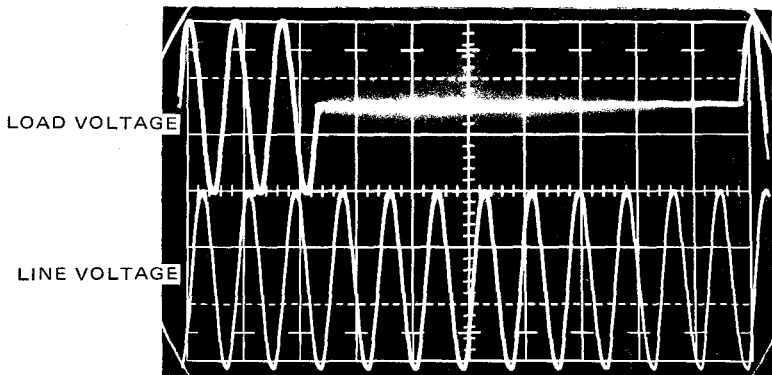
LOAD VOLTAGE

LINE VOLTAGE

Figure 6-1 — Load Voltage and Line Voltage for 25% Duty Cycle

zero crossing of the applied voltage. Because of this exact-timing require-
ment, pulse-type thyristor triggering is usually impracticable, since even
small timing drifts will result in off-zero switching, or possibly no switch-
ing at all.

The most successful method of zero-point thyristor control is there-
fore, to have the gate signal applied before the zero crossing, and have it
remain somewhat past the zero crossing. As soon as the zero-crossing
occurs, anode voltage will be supplied and the thyristor will come on. This
is effectively accomplished by using a capacitor to derive a $90°$-leading
gate signal from the power line source. However, only one thyristor can be
controlled from this phase-shifted signal, and a slaving circuit is necessary
to control the other SCR to get full-wave power control. These basic ideas
are illustrated in Figure 6-2. The slaving circuit fires only on the half cycle
after the firing of the master SCR. This guarantees that only complete
cycles of power will be applied to the load. The gate signal to the master
SCR receives all the control; a convenient control method is to shunt the
gate signal to ground whenever the SCRs are supposed to remain off. The
gate shunt can be a low-power transistor, which can be controlled by
bridge sensing circuits, manually controlled potentiometers, or various
other techniques.

A basic SCR slaving circuit such as shown in Figure 6-2 is very
effective and trouble free. However, it can dissipate considerable power.
This must be taken into account in designing the circuit and its packaging.
Slaving circuits which dissipate less power can be devised but they usually
require active devices which add considerably to circuit cost.

In the case of triacs, a slaving circuit is also usually required to
furnish the gate signal for the negative half cycle. However triacs can use
slave circuits requiring less power than do SCRs, as shown in Figure 6-3.
Other considerations being equal, the ease of slaving will sometimes make
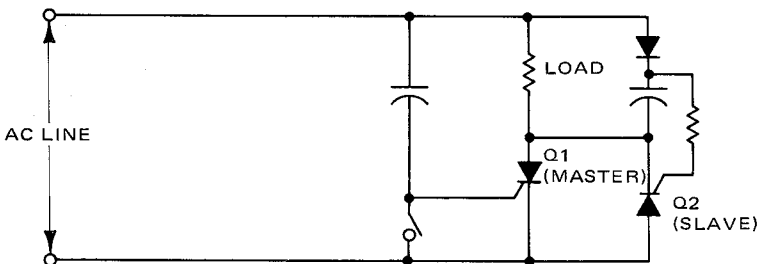the triac circuit more desirable than the SCR circuit.

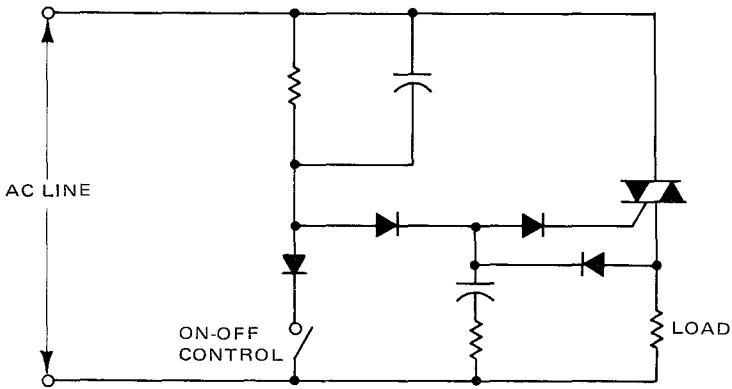**Figure 6-2 — Slave and Master SCRs for Zero-Point Switching**

Figure 6-3 — Triac Zero-Point Switch

Besides slaving-circuit power dissipation, there is another considera-
tion which should be carefully checked when using high-power zero-point
switching. Since this is on-off switching, it abruptly applies the full load to
the power line every time the circuit turns on. This may cause a temporary
drop in voltage which can lead to erratic operation of other electrical
equipment on the line (light dimming, TV picture shrinkage, etc.). For this
reason, loads with high cycling rates should not be powered from the same
supply lines as lights and other voltage sensitive devices. On the other
hand, if the load-cycling rate is slow, say once per half minute, the loading
flicker may not be objectionable on lighting circuits.

A note of caution is in order here. Neither of the full-wave, zero-
point-switching controls illustrated in Figures 6-2 and 6-3 should be used
as half-wave controls by removing the slave SCR: When the slave SCR in
Figure 6-2 is removed, the master SCR has positive gate current flowing
over approximately 1/4 of a cycle while the SCR itself is in the reverse-
blocking state. This occurs during the negative half cycle of the line
voltage. When this condition exists, Q1 will have a high leakage current
with full voltage applied and will therefore be dissipating high power.
This will cause excessive heating of the SCR and may lead to its failure.
If it is desirable to use such a circuit as a half-wave control, then some
means of clamping the gate signal during the negative half cycle must be
devised to inhibit gate current while the SCR is reverse-blocking.

Practical Zero-Point Switches

The zero-point switches shown in Figures 6-4 and 6-5 are used to
insure that the control SCR turns on at the start of each positive alterna-

tion. If the SCR were turned on later in the alternation, the turn-on voltage and current spikes could cause electro-magnetic interference (EMI). In Figure 6-4, a pulse is generated before the zero crossing and provides a small amount of gate current when line voltage starts to go positive. This circuit is primarily for sensitive-gate SCRs. Less-sensitive SCRs, with their higher gate currents, require smaller values for R1 and R2, and the result can be high power dissipation in these resistors. The circuit of Figure 6-5 uses a capacitor, C2, to provide a low-impedance path around resistors R1 and R2, and can be used with less-sensitive, higher-current SCRs. This circuit actually oscillates near the zero-crossing point and provides a series of pulses to assure zero-point switching. The basic circuit is that shown in Figure 6-4. Operation begins when switch S1 is closed. If the positive alternation is present, nothing will happen since diode D1 is reverse-biased. When the negative alternation begins, capacitor C1 will charge through resistor R2 toward the limit of voltage set by the voltage divider consisting of resistors R1 and R2. As the negative alternation reaches its peak, C1 will have charged to about 40 volts. Line voltage will decrease, but C1 cannot discharge because diode D2 will be reverse-biased. It can be seen that C1 and three-layer diode D4 are effectively in series with the line. When the line drops to −10 volts, C1 will still be 40 volts positive with respect to the gate of Q1. At this time D4 will see about 30 volts and will break back about 10 volts. This allows C1 to discharge through D3, D4, the gate, R2, and R1. This discharge current will continue to flow as the line voltage crosses zero and will insure that Q1 turns on at the start of the positive alternation. Diode D3 prevents reverse gate current and resistor R3 prevents false triggering.

The circuit in Figure 6-5 operates in a similar manner up to the point where C1 starts to discharge into the gate. The discharge path will now be from C1 through D3, D4, R3, the gate, and capacitor C2. C2 will eventually charge from the pulse at the cathode of Q1. This will reduce the voltage across D4 and it will turn off, and again revert to its blocking state. Now C2 will discharge through R1 and R2 until the voltage on D4 again becomes sufficient to cause it to break back. This repetitive exchange of charge from C1 to C2 causes a series of gate-current pulses to flow as the line voltage crosses zero. This means that Q1 will again be turned on at the start of each positive alternation, as desired. Resistor R3 has been added to limit the peak gate current.

Additional zero-point-switching circuits are shown in the temperature-control section of this chapter (6.2). A slave circuit which can be added to the circuits just discussed to provide full-wave power is shown in section 6.3.
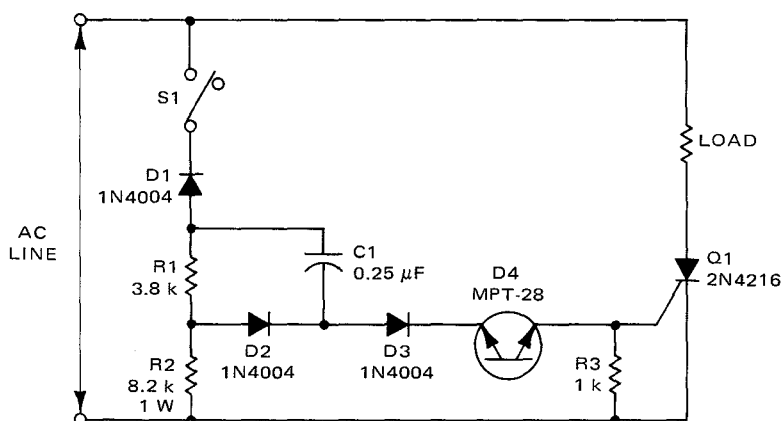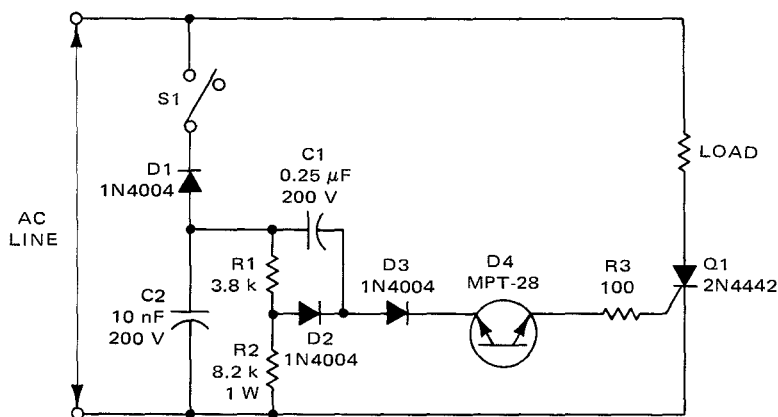
Figure 6-4 — Sensitive-Gate Switch

Figure 6-5 — Zero-Point Switch

6.2 Zero-Point Temperature-Control Circuits

117 Vac, 3 kW Temperature-Control Circuit

The circuit shown in Figure 6-6 is a temperature controller. It can directly drive any heating element which consumes 3 kW of power or less. The temperature-sensing element is a thermistor which controls base drive to a transistor. Switching of the load is done at the zero point of the ac input voltage to prevent electro-magnetic interference (EMI). When the thermistor is chosen to have a resistance of 5 kΩ at the desired operating temperature, temperature regulation will be better than 1%.

Figure 6-6 — 3 kW Temperature-Control Circuit

C1
0.5 µF
200 V

R3
100 k
100

D3
1N4003

R5
4.3 k
2 W

R6
4.3 k
2 W

R7
2.2 k
1 W

3 kW
HEATING
ELEMENT

117 V
LINE

R2
1 k

R1
5 k

MPS
6512
Q1

C2
0.5 µF
200 V

R_T*

Q2
MCR2935-5

7.5 V

D1
1N
4737

THERMISTOR
5 k NOMINAL

R4
1 k

D2
1N4933

Q3
MCR2935-5
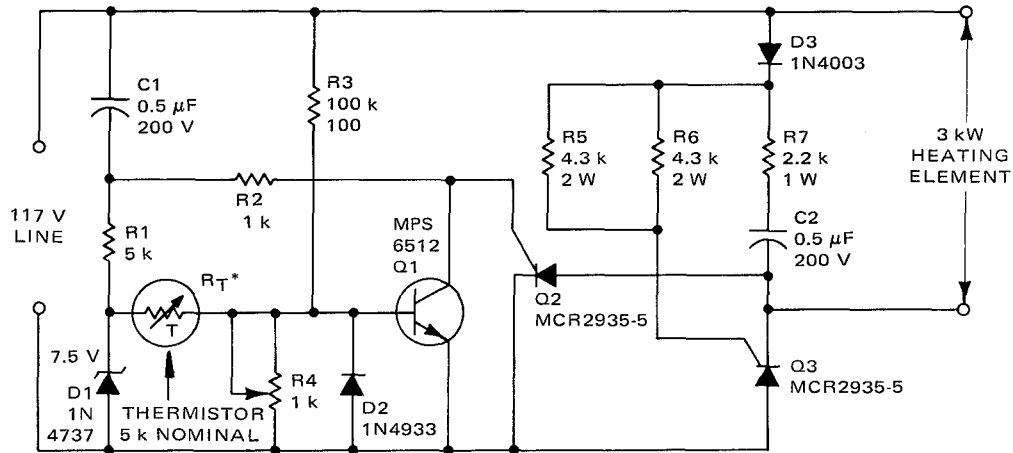
*Thermistor, 5 k nominal. Mount in proximity of heated area.

Operation of the circuit is as follows: capacitor C1, which is in series with resistor R1, causes current in zener diode D1 to lead the line voltage. In other words, the reference voltage will appear across the zener while the line voltage is changing from the negative to the positive alternation. The reference is divided by 10 through thermistor R_T, and potentiometer R4, to the transistor base. The nominal potentiometer setting will be 500 Ω. If the thermistor is cold (high resistance), the base voltage will be low and the transistor will not turn on. Current through resistor R2, which is also leading the line voltage, will then flow into the gate of SCR Q2, and Q2 will fire at the zero crossing of the line voltage. When the thermistor becomes heated by the load, its resistance will decrease. Now base drive during the zero-point interval will increase and the transistor will remain on. In this case, no current can flow into the gate of Q2 and the heater will be turned off at the desired temperature. Resistor R3 prevents Q2 from being gated on beyond the zero-crossing point by turning on transistor Q1 and shunting the gate drive.
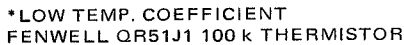
The circuit can actually be operated over a range of temperature by manually adjusting the 1 kΩ potentiometer. Tests performed with a 5 kΩ, 25°C rod thermistor indicate that the circuit would regulate up to 60°C. Thermistor resistance changed from 5 kΩ to 1.4 kΩ over this range.

The RC circuit connected across the load is used to sense the stored energy during the positive voltage alternation. If Q2 has been triggered on, this energy will then be used to gate Q3 on during the following negative alternation. Capacitor C2 will begin discharging through the gate of Q3 while the line voltage is still positive. Thus, Q3 will also be turned at the zero point of the line voltage. Since Q3 is slaved to Q2, the power to the load will be applied in full cycles.

230 Vac, 4 kW Temperature-Control Circuit

Figure 6-7 shows a modulated SCR zero-point-switching circuit designed to control 4 kW heater loads operating from 230 V power lines. Circuit operation will be explained by splitting the circuit into two halves. The right half consists of the SCRs and their triggering circuits. The phase-shift network (C3, R8) supplies a 90° shifted sine wave gate signal to the master SCR (Q4). Each time Q4 applies a positive half cycle of power to the load, it energizes the slave circuit which will then fire Q5 and apply the following negative half cycle of power to the load. Hence, load power is completely controlled on a full-wave basis by controlling the master SCR. When no load power is desired, the gate signal (Ig) is shunted to ground through a small SCR (Q3). This switch is commutated by the gate signal and turns off each negative half cycle assuring control on a cycle-to-cycle

Figure 6-7 — 4 kW Temperature-Control Circuit

Thyristor and Transistor Switches

SLAVE CIRCUIT

230 V
60 Hz

D1 1N4005
R1 12 k 5 W
R2 470
R3 330 k ±5%
R6* 3 k ±5%
R4* 20 k ±5%
R5* 82 k ±5%
2N4870
Q1
C1 40 µF 50 V
C2 40 µF 30 V
Q2 2N3905
D2 MZ2361 TEMP.
R7 10 k
RT
C3 0.47 µF 400 V
R8 220
I_g
Q3 2N5060
R9 4.7 k
D3 MZ2361
D4 MZ2361
R_L LOAD
Q4 2N5166 I_{gt} MAX 30 mA
D6 1N4005
R10 10
R11 24 k ±5% 2 W
D5 1N 4005
C4 0.68 µF 400 V
R12 10 k ±5% 2 W
Q5 2N5166 I_{gt} MAX 12 mA

*LOW TEMP. COEFFICIENT
FENWELL QR51J1 100 k THERMISTOR

basis. Diodes D3 and D4 perform dual functions. They provide dc restoration for capacitor C3, and they also provide a dc offset which is more easily controlled by switch Q3 than if this switch were placed directly between the gate of Q4 and ground.

Modulation is accomplished with a circuit network which controls the master SCR (Q4) through switch Q3. The line frequency is divided into twelve-cycle groups, and from one to all 12 cycles from each group can be applied to the load, thus allowing the load power to be modulated in 8% steps from 0% to 100% duty cycle. The number of cycles per group can be changed by changing C2.

Modulation is achieved in the following manner: First the main on-off control for this circuit is supplied by a bridge circuit consisting of R3, R5, R_T, R6, R7, and D2. The detector for the bridge is Q2. As R_T decreases, Q2 turns on, turning on Q3, shunting the gate signal to ground and removing power from the load. Now as the temperature drops, Q3 does not come on and (if modulation were omitted) full-wave power would be applied to the load on a continuous basis. However, the modulation is applied to proportion the load power in response to small changes in R_T. The modulation is achieved by superimposing a sawtooth voltage on one arm of the bridge through R3 and R5. This sawtooth voltage is generated by the unijunction-transistor relaxation oscillator consisting of R2, R4, Q1, and C2. The sawtooth wave modulates the bridge voltage so that over a portion of the ten-cycle group the bridge voltage will be above the null point, and over the other portion it will be below the null point.
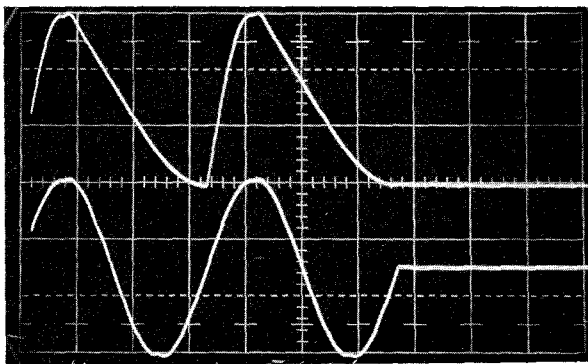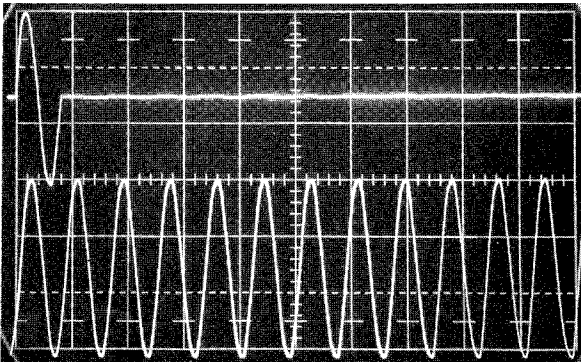

UPPER VERTICAL 100 V/cm
LOWER VERTICAL 200 V/cm
HORIZONTAL 5 ms/cm

Figure 6-8A — Capacitor Voltage and Load Voltage versus Time
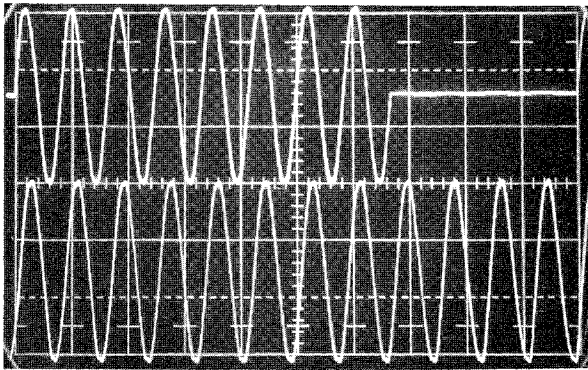
UPPER VERTICAL 100 V/cm
LOWER VERTICAL 200 V/cm
HORIZONTAL 5 ms/cm

Figure 6-8B — Load Voltage and Line Voltage for 65% Duty Cycle

UPPER VERTICAL 100 V/cm
LOWER VERTICAL 200 V/cm
HORIZONTAL 5 ms/cm

Figure 6-8C — Load Voltage and Line Voltage for 8% Duty Cycle

This action divides each ten-cycle group into an on and off portion -- the proportioning depending upon the amount R_T has varied from the nominal value. This circuit provides excellent control of a resistance heater as it will tend to stabilize and apply the correct amount of power on a continuous basis at a steady-state duty cycle depending on the load requirements. The temperature is therefore controlled over a very narrow range and no EMI is generated.

Since this circuit can switch the load power on and off at 6 Hz, it may cause undesirable flickering of incandescent lamps which are supplied
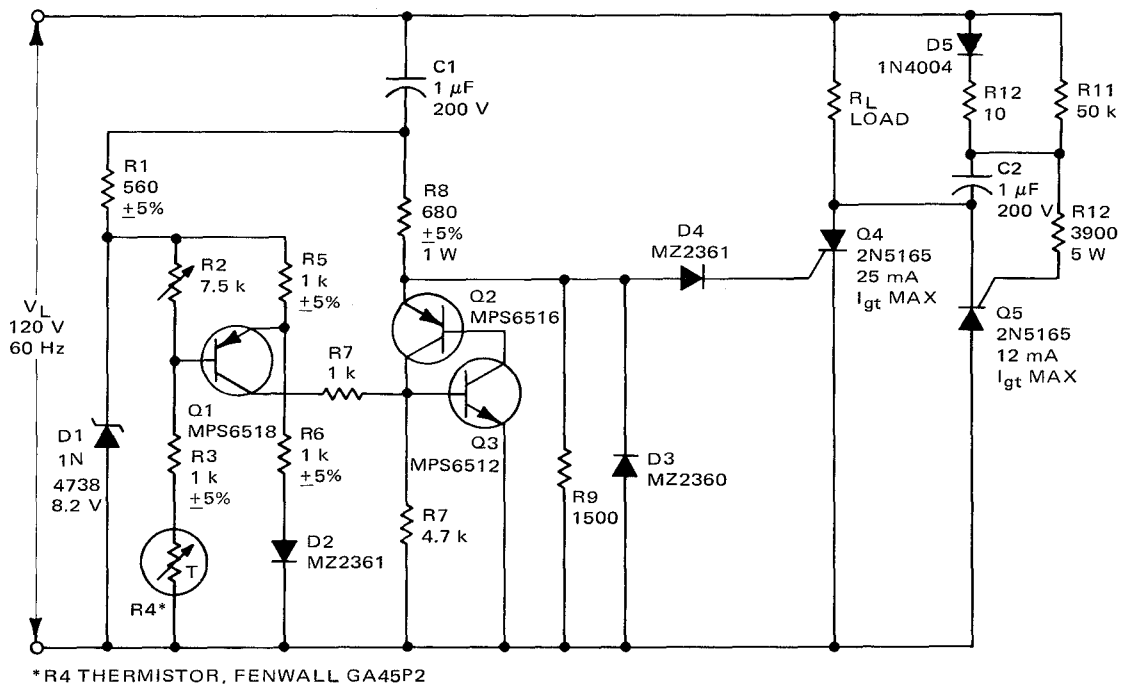
Figure 6-9 — 2 kW, 120 V Zero-Point Temperature-Control Circuit

*R4 THERMISTOR, FENWALL GA45P2

from the same power line as the heater load. Therefore, its primary application would be in an industrial situation where separate transformers supply banks of heaters.

Figure 6-8A shows the voltage waveform on slave-circuit capacitor C4. Figures 6-8B and 6-8C illustrate the load voltage for the different duty cycles (65% and 8%).

120 Vac, 2 kW Temperature-Control Circuit

Figure 6-9 shows a zero-point temperature-control switching circuit capable of controlling 2 kW watts from a 120 V supply. The circuit philosophy is basically the same as the temperature-control circuit shown in Figure 6-7, except that the proportioning feature has been removed. In many applications this type of circuit can be used on lighting circuits to control small heating loads such as portable heaters and electrical appliances. Here the cycling rate is controlled by temperature, so it will overshoot slightly because of the time lag between the heat source, the surface being controlled, and the thermistor (R4). By tailoring the amount of overshoot to the device being controlled, the cycling rate will be determined by the temperature overshoot and can be made slow enough – say once per minute – that it will not cause objectionable light blinking or TV picture size changes.

6.3 110 Vac SCR Slaving Circuit

When half-wave control of an ac line has been realized, it is often desirable to extend this to full-wave control. A slaving circuit is used to accomplish this. The slave circuit must sense when the master SCR has been gated and fire a slave SCR for the following alternation.

The SCR slaving circuit shown in Figure 6-10 provides a single power pulse to the gate of SCR Q3 each time Q2 turns on, thus turning Q3 on for the half cycle following the one during which Q2 was on. Q3 is therefore turned on only when Q2 is turned on, and the load can be controlled by a signal connected to the gate of Q2 as shown in the schematic. The control signal can be either dc or a power pulse. If the pulse is synchronized with the line, as shown in section 6-22, this circuit will make an excellent zero-point switch. During the time that Q2 is on, capacitor C1 is charged through R1, D1, and Q2. While C1 is being charged, D1 reverse-biases the base-emitter junction of Q1, thereby holding it off. The charging time constant, R1-C1, is set long enough that C1 charges for practically the entire half cycle. The charging rate of C1 follows a S-shaped curve, charging slowly at first, then faster as the supply voltage peaks, and finally slowly again as the supply voltage decreases. When the supply voltage falls
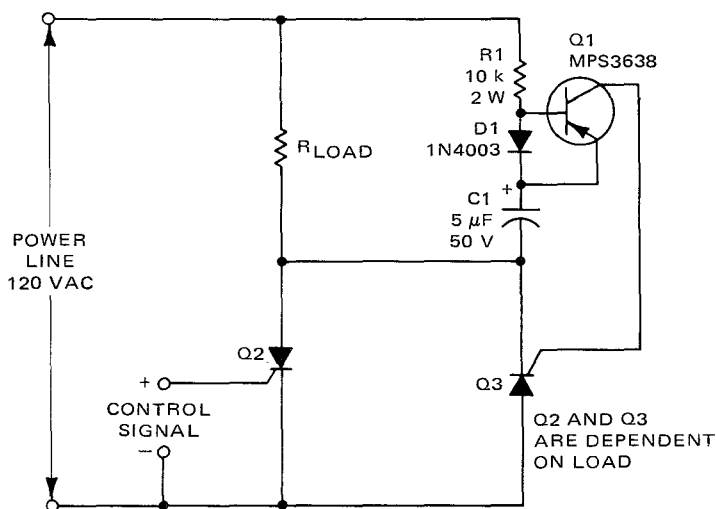
Figure 6-10 — 110 Vac SCR Slaving Circuit

below the voltage across C1, diode D1 becomes reverse-biased, and the base-emitter of Q1 becomes forward-biased. For the values shown, this occurs approximately 6° before the end of the half-cycle conduction of Q2. The base current is derived from the energy stored in C1 and is returned to C1 through R1 and the load. This turns Q1 on, and discharges C1 through the gate of Q3. As the voltage across C1 decreases, the base drive of Q1 decreases and limits the collector current somewhat. The current pulse must last until the line voltage reaches a magnitude such that holding current will exist in Q3. The values shown will deliver a current pulse which peaks at 100 mA and has a magnitude greater than 50 mA when the anode-cathode voltage of Q3 reaches +10 volts. This circuit completely discharges C1 during the half cycle that Q3 is on. This eliminates the possibilities of Q3 being slaved for additional half cycles after the drive is removed from Q2. The peak current and the current duration are controlled by the value of R1. Therefore, the proper value of R1 can be chosen for any given family of SCRs. The particular SCR used must be capable of handling the maximum current requirements of the load to be driven.

6.4 Light Dimmers

800 W Soft-Start Light Dimmer

A circuit capable of controlling incandescent lamp loads up to 800 watts from a 120 V, 60 Hz line is shown in Figure 6-11. Lamp failures
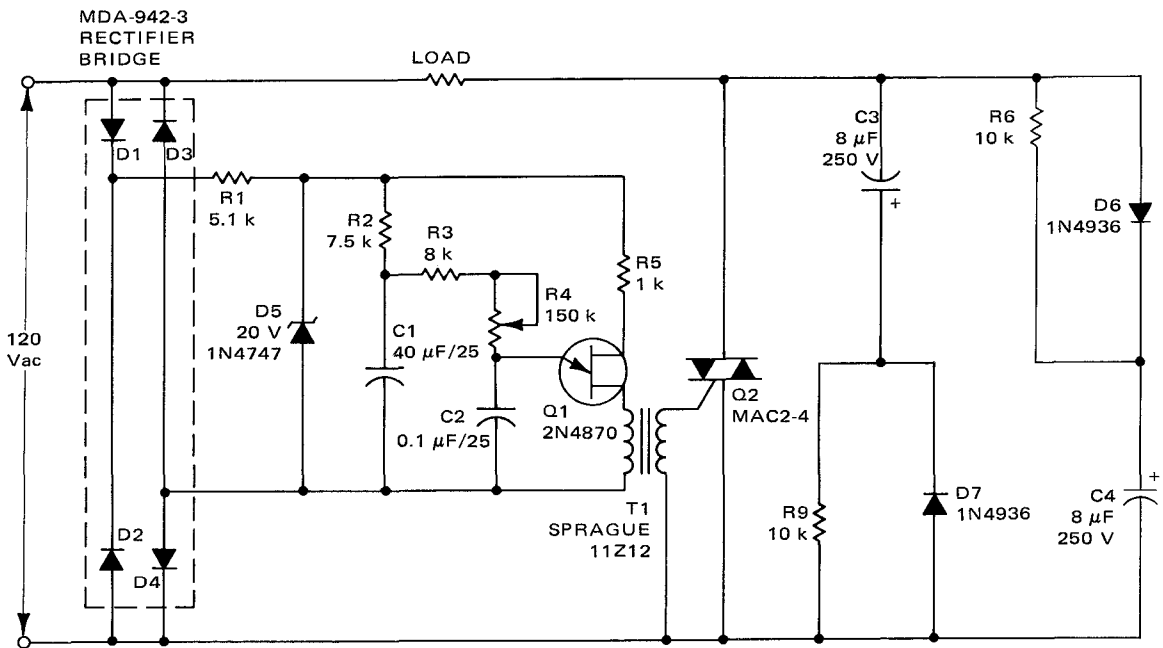
Figure 6-11 — 800 W Soft-Start Light Dimmer

Thyristor and Transistor Switches

normally caused by high inrush currents are eliminated by the soft-start feature. Accidental turn on, which could nullify this advantage, is prevented by a special dv/dt network.

Operation of this circuit begins when voltage is applied to the diode bridge consisting of D1 through D4. The bridge rectifies the input and applies a dc voltage to resistor R1 and zener diode D1. The zener provides a constant voltage of 20 volts to unijunction transistor Q1, except at the end of each alternation when the line voltage drops to zero. Initially, the voltage across capacitor C1 is zero, and capacitor C2 cannot charge to trigger Q1. C1 will begin to charge, but because the voltage is low, C2 will have adequate voltage to trigger Q1 only near the end of the half cycle. Although the lamp resistance is low at this time, the voltage applied to the lamp is low and the inrush current is small. Then the voltage on C1 rises, allowing C2 to trigger Q1 earlier and earlier in the cycle. At the same time the lamp is being heated by the slowly increasing applied voltage and by the time the peak voltage applied to the lamp has reached its maximum
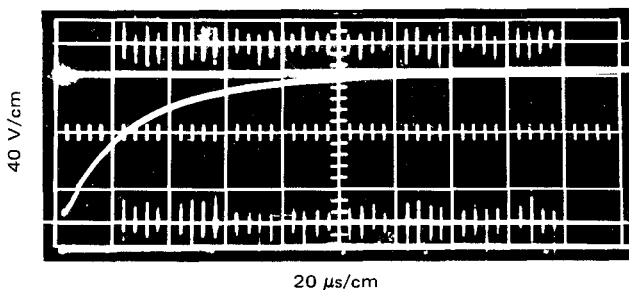

20 μs/cm

Figure 6-12 — Voltage Rise Across Triac into One 500 W Light Bulb, Using dv/dt Network. 20 μ/div, Horizontal. 40 V/div, Vertical.
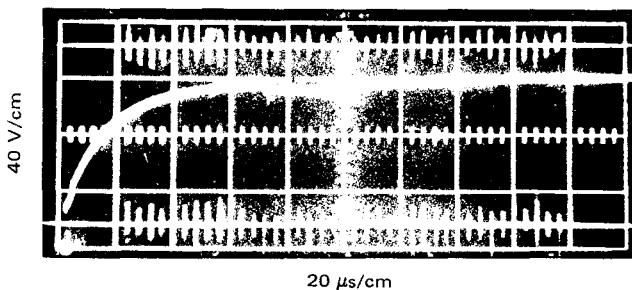

20 μs/cm

Figure 6-13 — Voltage Rise Across Triac into Two 500 W Light Bulbs, Using dv/dt Network. 20 μs/div, Horizontal. 40 V/div, Vertical.

value, the bulb has been heated sufficiently that the peak inrush current is kept to a reasonable value. Resistor R4 controls the charging rate of C2 and provides the means to dim the lamp. Power to the load can be adjusted manually by varying the resistance of R4.

The dv/dt network is used to prevent the line voltage from triggering triac Q2 before the light has warmed up. This would occur at the instant power is first applied to the circuit if the instantaneous line voltage were sufficient to cause triggering. Capacitor C3 will delay a negative rise in voltage by charging through the load and diode D7. Capacitor C4 does the same for a positive rise. Resistors R6 and R7 are used to discharge capacitors C3 and C4.

A test used to evaluate this network is as follows: A 120 volt dc source was used to charge C4 through a cold 500 watt bulb. The voltage rise shown across the triac in Figure 6-12 is 6 volts per microsecond. A 500 watt load should therefore be the maximum allowable load for an ambient temperature of 65°C. The same test was also confirmed with a 1000 watt load. The rate of rise for this case, as shown in Figure 6-13, is 10 volts per microsecond and the device did not turn on. This is considered to be a safe rate of rise when the ambient temperature is 25°C. In other words, this network will protect the triac from loads as high as 1000 watts at room temperature.

800 W Light Dimmer

A light dimmer circuit which operates from a 120 volt ac source and can control 800 watt incandescent bulbs is shown in Figure 6-14. Output power is varied by controlling the phase of conduction of triac Q1. All of the circuits in Table 6-I have this capability, but the single triac circuit
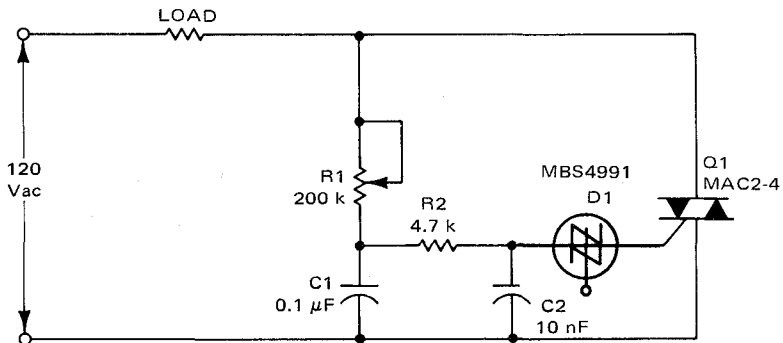

Figure 6-14 — Simple 800 W Triac Light Dimmer

shown as circuit two in Table 6-I is the simplest by far and is the one chosen for this particular application. The control circuit for this triac must function as shown in Figure 6-15. That is, it must create a delay between the time voltage is applied to the circuit (shown dotted) and the time it is applied to the load. The triac is triggered after this delay and conducts current through the load for the remaining part of each alternation. This circuit can control the conduction angle from 0 to about 170° and provides better than 97% of full-power control.

The operation of this circuit can best be understood by referring to the waveforms of Figure 6-16 which are taken from the circuit in Figure 6-14. Figure 6-16A shows the voltage across capacitor C1, and Figure
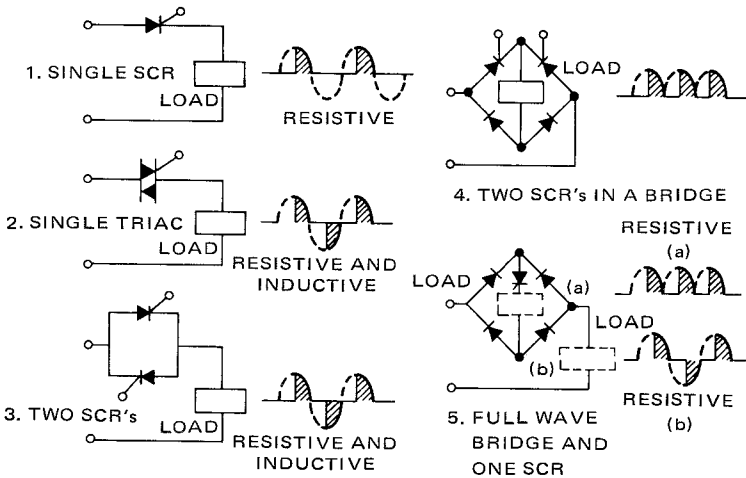
SOME COMMON PHASE-CONTROLLED CIRCUITS
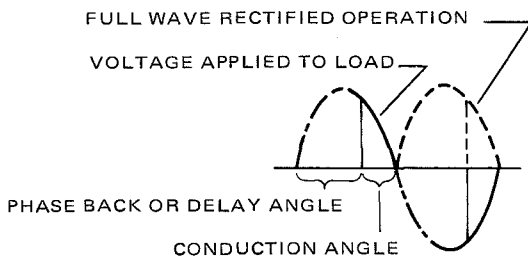

TABLE 6-I

Figure 6-15 — Sine Wave Showing Principles of Phase Control

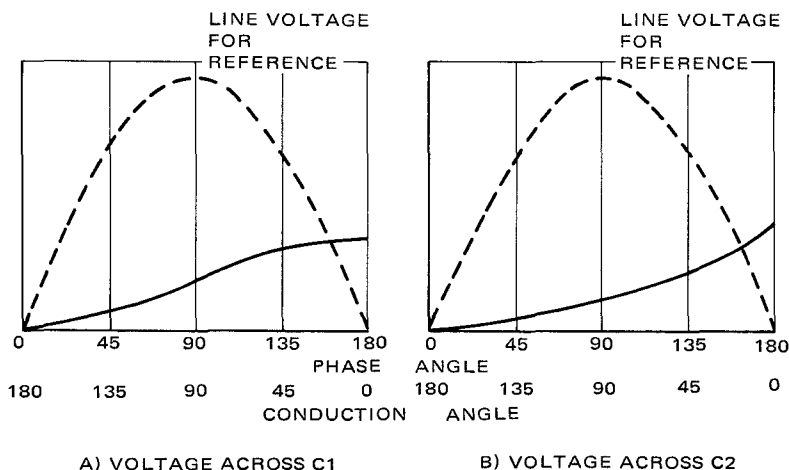A) VOLTAGE ACROSS C1   B) VOLTAGE ACROSS C2

**Figure 6-16 — Comparison of Trigger Voltage Waveforms
in Single-Section and Two-Section Phase-Shift Circuits.
Circuit Shown in Figure 6-14.**

6-16B shows the voltage across capacitor C2. Each begins to charge as
shown when the positive alternation of line voltage is applied to the cir-
cuit. It is affected by the decrease in line voltage for phase angles greater
than 90°. However, capacitor C2 is better isolated, and continues to charge
at a fairly linear rate even beyond 90°. In order to trigger the triac, this
voltage must exceed the breakover voltage of three-layer diode D1 (20
volts) sometime during this alternation. This is accomplished by reducing
the in-circuit resistance of potentiometer R1 so that the capacitors will
charge faster. When the diode breaks down, the capacitors discharge into
the gate of triac Q1 and turn it on. These same voltages will appear during
the negative alternation and trigger the triac by pulling current out of the
gate.

The two-section phase-shift circuit allows reliable and stable trigger-
ing at low conduction angles. The 180° phase shift of this circuit (when
unloaded) gives diode D1 an almost-linear ramp from which to trigger,
rather than the flat top of a sine wave which occurs as the phase ap-
proaches 180° with a single-section circuit (see Figure 6-16). Slight power-
supply voltage variations which might cause 5° to 10° jitter at low conduc-
tion angles in single-section circuits, cause less than 1° variation with the
two-section phase shift circuit.

The starting current in this circuit could cause a problem. The cold
resistance of the filament in a 500 watt bulb is about 2 Ω. A 120 volt line

20 ms/cm

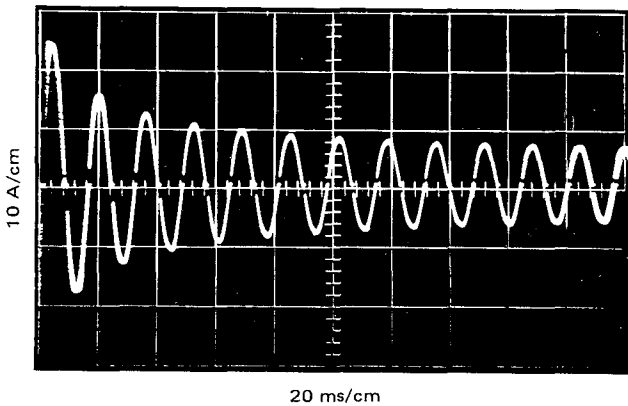Figure 6-17A — Turn-On Surge Current at 180° Conduction Angle into a 500 W Bulb. 10 A/div, Vertical. 20 ms/div, Horizontal.
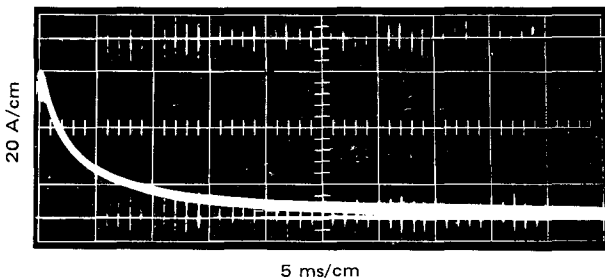

5 ms/cm

Figure 6-17B — Turn-On Time for a 500 W Bulb from 120 Vdc. 20 A/div, Vertical. 5 ms/div, Horizontal.
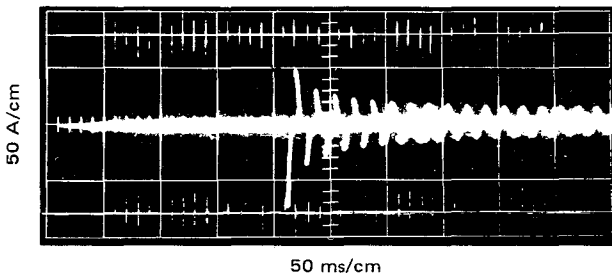

50 ms/cm

Figure 6-17C — Surge Current into a Cold 500 W Light Bulb Using a Triac Light Dimmer with a 60° Conduction Angle. 50 A/div, Vertical. 50 ms/div, Horizontal.

could therefore produce peak-surge currents of 85 amperes. The starting current waveform for a 180° conduction angle is shown in Figure 6-17A. The peak surge current is less than 30 A and Figure 6-17B shows why. With 120 volts dc applied, the peak current reaches 60 A but decreases to less than 30 A in 5 milliseconds. This means that the filament resistance can double in just a few milliseconds. In Figure 6-17A, the filament actually has 4 milliseconds to warm up before the peak voltage is reached, but this is enough time for the resistance to increase and limit the current. A more severe test is shown in Figure 6-17C. Here the triac is triggered with a 60° conduction angle. The 150 volts applied at this time into the cold filament produces a 75 A current surge. Again, the filament resistance increases rapidly so that this surge was reduced by 60% within two cycles. Based on this information and the I^2t rating of the triac, surge currents from a 1000 watt lamp load at 90° conduction would be the maximum that this device can withstand.

Tests with two 500 watt bulbs in parallel revealed some interesting data as shown in Figure 6-18. When the bulbs had been operating for a while so that resistance stabilized, the steady-state resistance increased from 1 Ω starting resistance to about 12 Ω, and varied as a function of the conduction angle. For instance, there is more peak current at 70° conduction than at 90°, even though the voltage is less at 70°. The probable explanation is that the filament has more time to cool when operating at
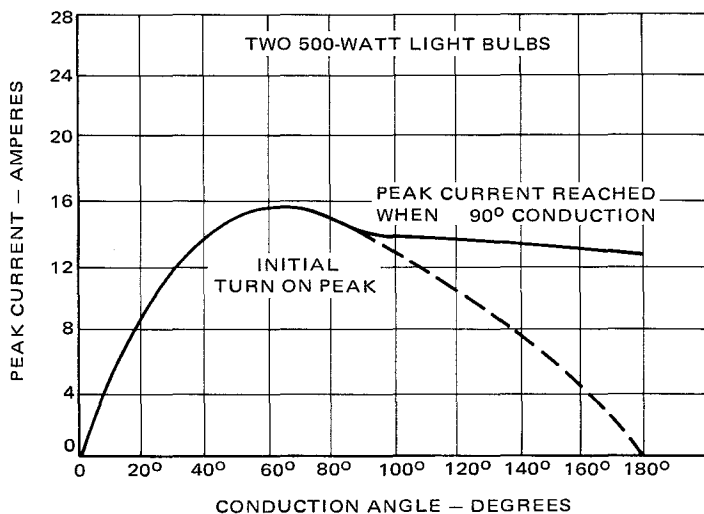

Figure 6-18 — Repetitive (60 Hz) Peak Currents of Triac Light Dimmer versus Conduction Angles
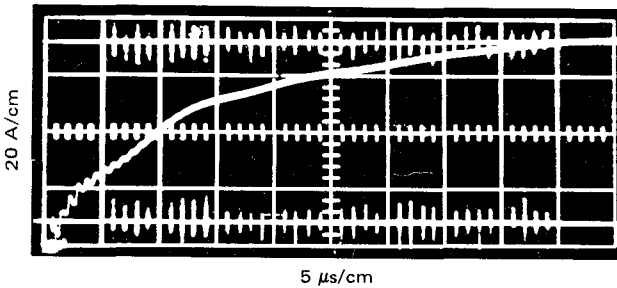
5 μs/cm

Figure 6-19A — Current Rise into a 500 W Bulb into a 60° Conduction Angle. 20 A/div, Vertical. 5 μs/div, Horizontal.

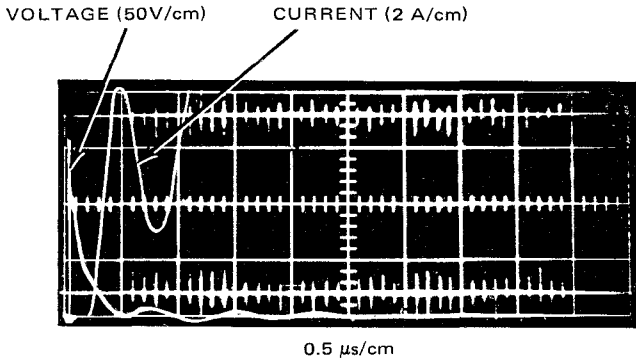VOLTAGE (50V/cm)   CURRENT (2 A/cm)

0.5 μs/cm

Figure 6-19B — Turn-On Current and Voltage Across Triac with 500 W Bulb and a 60° Conduction Angle (Single Sweep). 2 A/div Current Trace, Vertical. 50 V/div Voltage Trace, Vertical. 0.5 μs/div, Horizontal.

lower conduction angles and its average resistance will therefore be lower. And, because beyond 90° the filament had less time to cool, average resistance increased and the peak current fell slightly as shown. This variation actually has very little effect on the normal operation of the triac.

Another possible problem could be caused by the power dissipated while the triac is turning on. Figure 6-19A shows the current increasing to 75 A for the previous case (Figure 6-17C) of a cold filament operating at a 60° conduction angle. The current increases to 60% of peak current in about 10 μs. Since the only resistance in the circuit is the 2 Ω filament, it determined that there must be about 20 μH of circuit inductance. This explains the turn on waveforms in Figure 6-19B. The triac turns on in less than 0.5 μs as shown by the voltage waveform. The current surge however, is delayed for 0.3 μs, and therefore no appreciable amount of power is

INITIAL CURRENT (2A/cm)
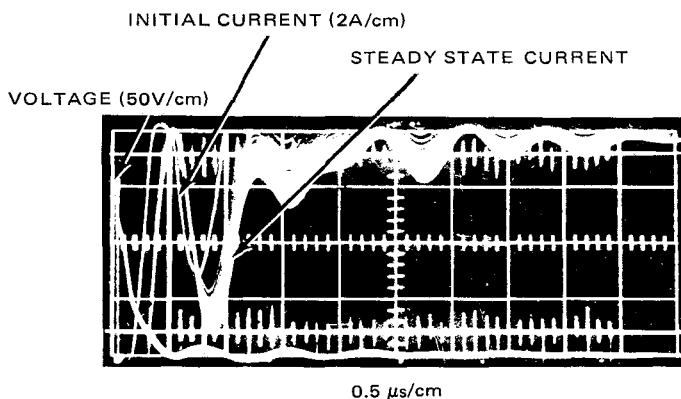
STEADY STATE CURRENT

VOLTAGE (50V/cm)

0.5 µs/cm

Figure 6-19C — Turn-On Current and Voltage Across Triac with a 500 W Bulb and a 60° Conduction Angle (Repetitive Sweep). 2 A/div Current Trace, Vertical. 50 V/div Voltage Trace, Vertical. 0.5 µs/div, Horizontal.

generated. Figure 6-19C shows the initial surge plus the steady-state turn-on currents. This further illustrates the fact that the circuit inductance is sufficient to limit the power dissipated during turn on.

There is one other possible mode of failure for the triac. That is, if the power switch were closed at the peak of line voltage, the rise of voltage would trigger the triac on. This will not destroy it since the inrush current is limited, even at this time, by the circuit inductance.

6.5 Light-Operated High-Voltage Series Switch

The circuit shown in Figure 6-20 is a high-voltage switch composed of a series string of SCRs which are triggered simultaneously from a single light source. Triggering is accomplished with phototransistors (MRD-300) driven by a xenon flash tube through a fiber-optic bundle.

Across each SCR is a 1.5 MΩ resistor in series with a 51 kΩ resistor. These resistors form not only the voltage equalization network for the series-connected SCRs, but also voltage dividers for each SCR, with 20 volts developed across each 0.1 µF capacitor and 51 kΩ resistor. This voltage provides collector bias for each phototransistor and a source of gate current for each SCR.

When the xenon tube is flashed the light is fed into the fiber-optic bundle where it is split into ten outputs of approximately equal amplitude and fed simultaneously to each phototransistor. As each phototransistor
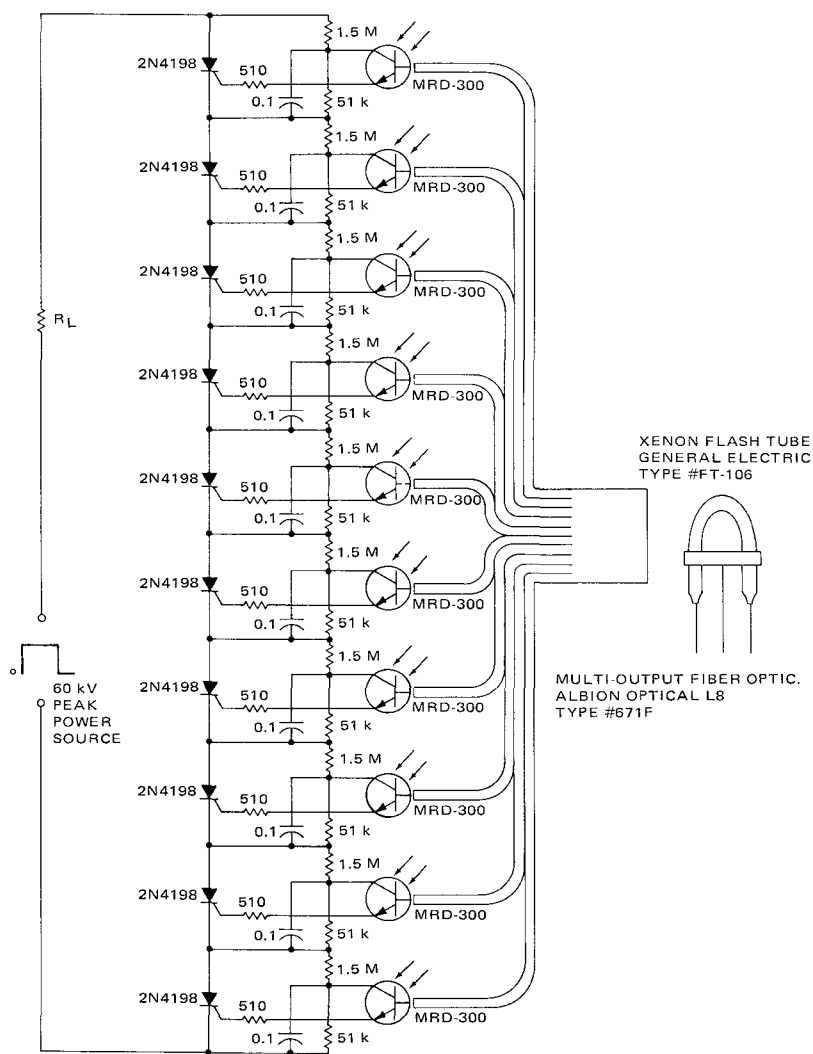
Figure 6-20 — Light-Operated 6 kV Series Switch

conducts, it discharges the 0.1 μF capacitor into the gate of its corresponding SCR via the 510 Ω resistor, turning all of the SCRs on at once.

Triggering series-connected SCRs by this method offers several advantages over other types of triggering. The use of light via phototransistors results in simultaneous firing of the SCRs, thus eliminating the inductive delays that would result if conventional wiring were used to transport a gate signal from a central source. Also, the use of fiber optics eliminates the need for special trigger transformers that can withstand the high voltages that would exist between windings.

The turn-on time of the circuit shown was measured to be about 300 ns at 1 A anode current; the maximum capability of the circuit is a 100 A pulse for 4 ms with a 6 kV input voltage. The SCRs must have matched rise times so that the slowest units will not be gated on by anode breakover due to earlier turn on of the faster devices.

This circuit would be most useful in high-voltage crowbar circuits or high-voltage pulse-forming networks. Also, with minor modifications it could be adapted to certain proportional control systems.

6.6 Time Delay Circuits

Long Duration FET Timer

The circuit shown in Figure 6-21 is used to obtain a time delay of up to 10 hours. Basically, it is a unijunction-transistor sawtooth oscillator. Output pulses can be obtained from either base one or base two of Q3. The conventional charging resistor for capacitor C_E has been replaced by a current source consisting of resistors R1 through R3 and transistor Q1. As
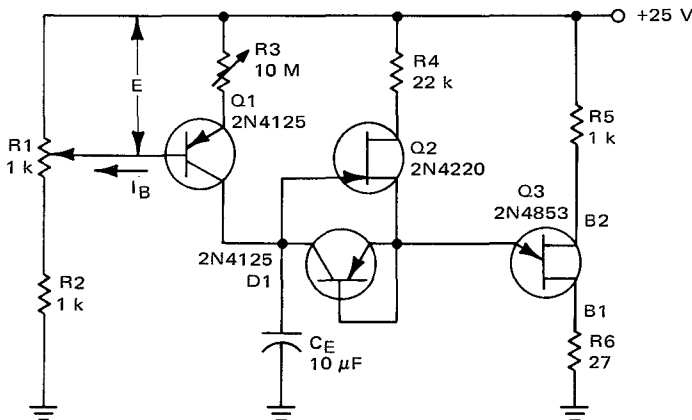

Figure 6-21 — Long-Duration Time Delay

is the case in most long-delay timers using UJTs, the charging current from Q1 will generally be inadequate to supply the peak point current required to trigger Q3. Field-effect transistor Q2 has been placed in parallel with R3 and Q1 to provide the peak point current. Diode D1 has been added to provide a low-resistance path to discharge C_E once its voltage has become sufficient to trigger Q3. Otherwise, the discharge current would pass through the gate-source junction of Q2 and cause damage to the FET.

The time required for C_E to charge is determined by the formula,

$$T = \frac{C_E V_T}{I_C} ,$$

where

T = the time in seconds,

C_E = the capacitor value in μF,

V_T = the trigger voltage (about 20 V), and

I_C = the charging current (variable) in μs

The maximum time is obtained when the current source is zero and only leakages are considered, although this is not recommended as a mode of operation. Figure 6-22 shows that worst case leakages at 25°C could total 100nA. The maximum delay time would then be 1/2 hour. In order to remove the dependence of V_T on leakage current, it would be advisable to operate the current source an order of magnitude higher. A certain
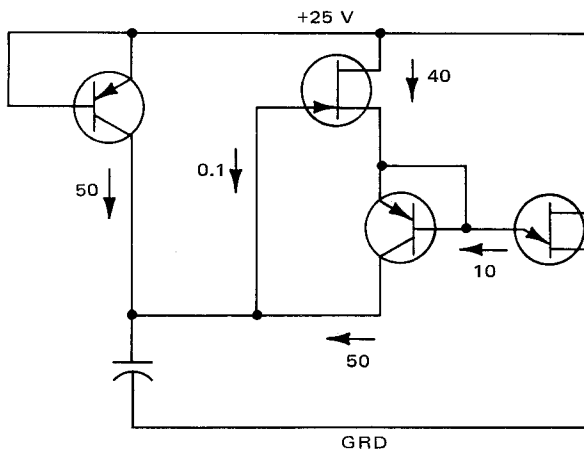

Figure 6-22 — Worst-Case Leakage Currents in Nanoamperes

amount of stability and repeatability would occur when $I_C = 1\ \mu A$, but time delay would be reduced to about 3 minutes maximum. Fortunately, the typical values of leakage currents are less by a factor of 50 (only 2 nA) as shown on the data sheet for the 2N2217, a device similar to the 2N4125. This means that leakage currents alone would create a 30 hour time delay. The insulation resistance of C_E could limit this time if R_C is less than $V_T/I_C = 20/2n = 10\ G\Omega$. It turns out however, that the minimum insulation resistance of C_E at $25°C$ is 20 GΩ and we were able to operate in practice with up to a 10 hour time delay. Again, for stable operation, it would be advisable to operate the current source with at least 20 nA (10×2 nA) for a maximum delay of 3 hours.

The current source is actually voltage sensitive since it converts a constant voltage reference to a constant current reference. It can be seen that $I_C = V3/R3$, if we ignore leakages. Both V3 and R3 are variable. R3 would normally be set at its maximum value to make V3 larger and less sensitive to voltage changes in the circuit. And $V3 = E - V_{BE}$ where E is the manually variable voltage across the upper half of R1 and V_{BE} is a relatively constant base-emitter voltage drop. If we consider the typical maximum-delay case of 3 hours, I_C will equal 20 nA. If the time delay is to change by less than 10% for voltage variations, then the change in I_C must be less than 2 nA. The voltage E, which would be 0.8 volt, cannot change by more than 20 mV. This means that voltage regulation of the supply must be better than $\dfrac{20\ mV}{800\ mV}$ or 2.5%. Regulation should, of course, be specified to match the accuracy of time that is desired.

Finally, it is desirable to analyze the effects of temperature, or more specifically, the effects of an increase in temperature from $25°C$ to $125°C$, on the 3 hour delay. If C_E is a 100 V polycarbonate capacitor, its insulation resistance will increase to about 2 GΩ. The limiting value of R_C now will be 20 V/20 nA or 1 GΩ so we are safe by a factor of 2. V_{BE} of Q1 will decrease by 0.2 volts with this temperature rise. This doubles I_C and decreases delay time by 50%. The standoff ratio (η) of Q3 will decrease by 7% with this temperature change. This decreases V_T and delay by 7%. The leakage current of Q1 and D1 will increase to 20 nA (a factor of 10). This again doubles I_C and decreases time by 50%. Multiplying these factors (.50 $\times$ .50 $\times$ .93) gives about .23, indicating that the delay time will decrease from 3 hours to 3/4 of an hour. The basic increase of I_C was 40 nA due to V_{BE} and leakage currents. If it is desirable for delay time to change by less than 10%, I_C should start at 400 nA. This is equivalent to saying that time delays should be kept at or below 10 minutes for operation over this range of temperature.

Figure 6-23 — A Sequential UJT Timer Circuit

Sequential UJT Timer

The circuit shown in Figure 6-23 is a timer in which a pulse is formed in sequence across the load resistor of each stage. In this circuit there are a total of three stages. More or fewer stages may be used if desired. The output pulse from each stage will follow that of the previous stage by an interval which is adjustable in length. A pulse at the first SCR, Q1, will start the various time intervals. SCR Q10 is used only to turn off the last stage SCR, Q7. If a timing circuit were added between Q10 and Q1, the sequence would run continuously.

Initially, all capacitors will have zero charge. A pulse at the gate of Q1 will turn it on and allow load current to flow through load resistor R1, and charging current to flow through R2 into C1. Current will also flow through the emitter-base junction of Q2 and through the 10 kΩ base resistor. This base current saturates Q2 and allows C2 to start charging through the 5 MΩ potentiometer and the 10 kΩ timing resistor. When the voltage across C2 becomes sufficient to trigger Q3, a pulse which turns on SCR Q4 is formed at base one of Q3. This places the negative voltage on C1 across Q1, causing it to turn off. Stages two and three operate in the same way. However, after SCR Q10 is pulsed on, it will turn itself off because anode current through the 10 kΩ resistor is less than holding current.

It is possible to drive relays or other types of loads directly with these SCRs. The minimum value of load resistance is determined by the turn-off time requirements and is 50 Ω. Holding current requirements set the maximum value at 1 kΩ. If it is desirable to obtain output pulses from the gate of each SCR, then R1 through R3 should each be 100 Ω. It is possible to set the delay time of each stage between 10 ms and 5 s with the 5 MΩ potentiometer. This circuit would be most useful in control operations where different cycles require different time durations.

6.7 Pulse Generators

Pulse Generator with Power Amplifier

The circuit shown in Figure 6-24 is a pulse generator which provides independent control of both "on" pulse width and "off" pulse width. Two outputs of the generator are provided; they are inverted from each other and are capable of driving 50 Ω loads. The pulse generator outputs are used to drive two power stages, which are capable of furnishing 6 A to a 5 Ω load.

The basic circuit of the generator is an astable multivibrator formed by Q5 and Q6. The timing elements for the "on" period are C4, R20, and

Figure 6-24 — Pulse Generator and Power Amplifier

POWER STAGE

PULSE GENERATOR

POWER STAGE

Thyristor and Transistor Switches

6-29
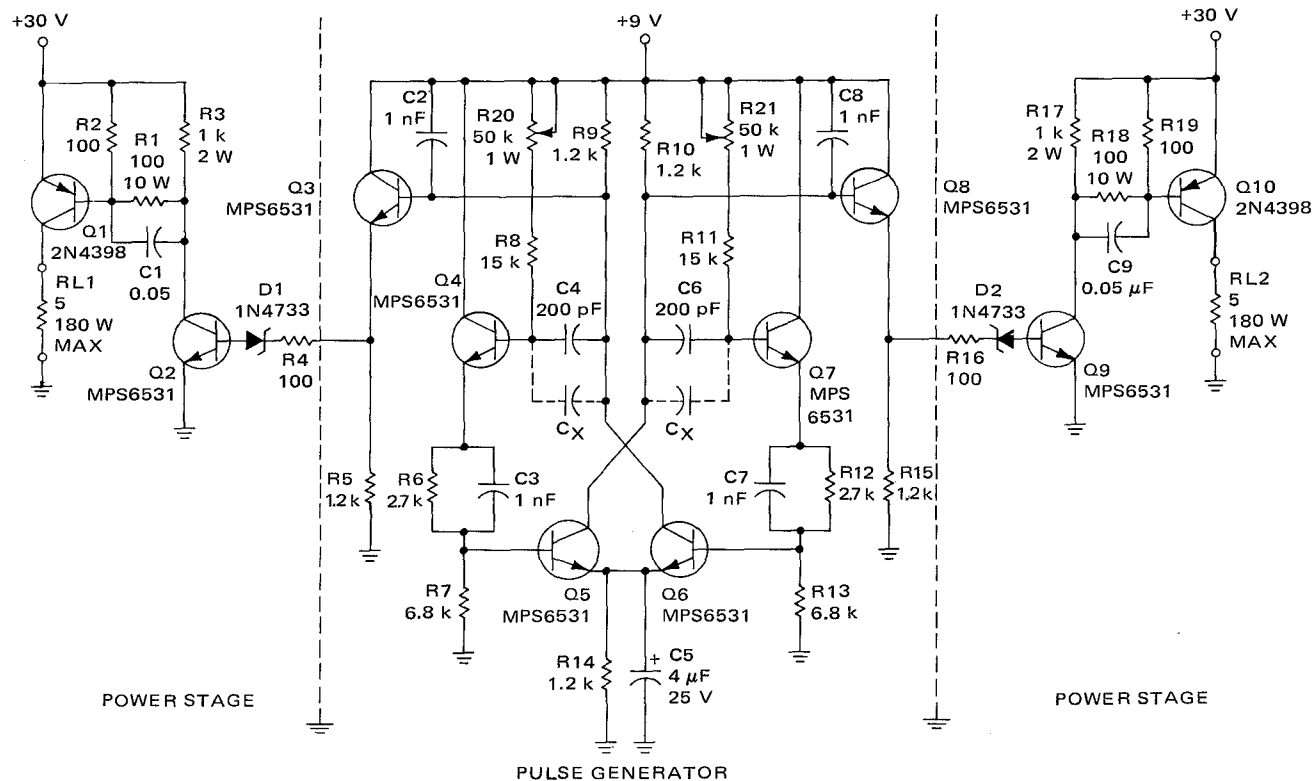
R8, and for the "off" period are C6, R21, and R11. Emitter followers (Q4 and Q7) provide isolation and a current gain for driving Q5 and Q6. This arrangement permits a wide variation in the value of capacitors C4 and C6. Therefore, the timing of the multivibrator's periods can be made with single potentiometers and various values of capacitors. Table 6-II is a summary of the periods obtained for various values of C_X. The adjustment from minimum to maximum of each period is made with R10 and R21. The values of C_X shown in the table provide some overlapping of period timing so the minimum time setting of a range where C_X is the larger of two adjacent values is less than the maximum time setting of the range where C_X is the smaller of the two adjacent values.

PULSE WAVEFORM AT LOAD
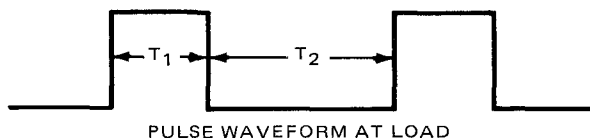
CAPACITANCE	T1 (MIN)	T1 (MAX)	T2 (MIN)	T2 (MAX)
0 pF	5.6 μs	23 μs	5.9 μs	27 μs
750 pF	20.5 μs	96 μs	23.5 μs	118 μs
3.9 nF	86 μs	430 μs	94 μs	490 μs
0.017 μf	420 μs	1.9 ms	460 μs	2.4 ms
0.10 μf	1.8 ms	9.2 ms	2.3 ms	11.6 ms
0.50 μf	8.6 ms	36.5 ms	9.4 ms	45.5 ms
1.5 μf	30 ms	140 ms	34 ms	176 ms

Table 6-II. Value of C_X versus Pulse Period

The maximum pulse duration given in Table 6-II is 176 ms. The delay time is not limited to this value and can be increased by increasing the value of capacitor C_X.

The output for each side of the multivibrator is an emitter follower which provides a low output impedance. Capacitors C2 and C8 are required to prevent the emitter followers from oscillating when a low-impedance load is driven. The rise and fall times of the output pulse are affected by the value of C2 and C8, so these values may be optimized for any particular load used. A load impedance as low as 50 Ω may be driven with the values shown.

The power amplifier stages are duplicates of each other, so only one

will be described. The output will drive 6 A into a 5 Ω load with a rise time of 100 ns and a fall time of about 2 μs. Zener diode D1 blocks the quiescent output voltage of Q3 (approximately 4 volts) from turning on Q2. When the output of Q3 increases to approximately 8 volts, the zener conducts and turns on Q2. Since the zener has no turn on time, Q2 comes on as fast as its intrinsic properties permit. Capacitor C1 is discharged when Q2 comes on, so Q2 receives a large emitter-base current which turns it on fast. The rise time of the turn-on pulse is typically 100 ns. While Q1 and Q2 are on, C1 charges to the supply voltage. When Q2 is turned off, the voltage on C1 reverse-biases the base-emitter junction of Q1 and helps to turn it off. At a current of 6 amperes, Q1 will turn off with no base capacitor (C1) in about 5 ms. With a capacitor, this time is decreased in proportion to capacitance used. The 0.05 μF capacitor used brings the turn-off time to less than 2 ms. Faster turn-off times can be obtained by using larger values of capacitance for C1. However, the larger C1 is made, the larger the turn-on current becomes. Therefore, to prevent destruction of either Q1 or Q2, or both, a small resistor should be placed in series with C1. The value of the resistor is dependent upon the size of C1, the maximum rated collector current of Q2, and the maximum rated base current of Q1.

10 ns, 1 A Pulse Generator

Figure 6-25 is the circuit diagram of a pulse generator which is unique in that its output pulse can be positive, negative, or both (Figure 6-26). The positive and negative output levels can be varied from 0 to 10 volts maximum, thus giving a maximum output of 20 volts peak to peak. The generator has a maximum output capability of one ampere, both positive and negative. The output will drive a resistive load with a current greater than 10 mA, and rise and fall times of about 10 ns.

The operation of the circuit is as follows: The relaxation oscillator uses a unijunction transistor (Q1) in which R1 and R2 charge C1 until the emitter voltage reaches V_p, then C1 discharges through R3, forming a positive voltage spike. Consider the monostable multivibrator in its normal state; Q2 is cut off, Q3 is biased on through R6 and R7, and capacitor C3 is charged to approximately 12 volts (positive on the end tied to the collector of Q2). When the relaxation oscillator fires, the positive spike from Q1 is coupled through C2, turning on Q2. The voltage across C3 will then reverse-bias the base of Q3, turning it off. Q3 will remain off until C3 charges through R6 and R7 to the voltage across R5 plus the V_{BE} of Q3. With switch S1 in position one, the predriver transistors Q4 and Q5 are normally off. When Q3 turns off, current flows through R8 and R10,

Figure 6-25 — Pulse Generator

Thyristor and Transistor Switches
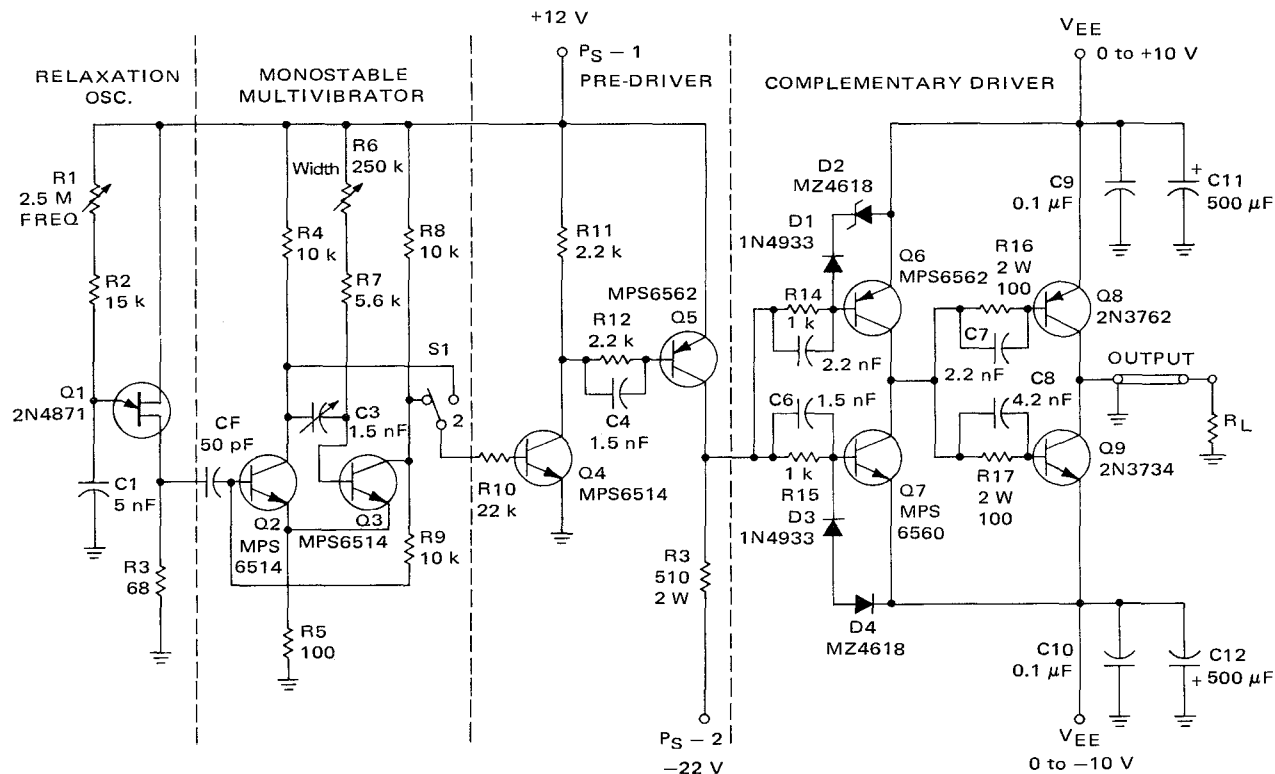
turning on Q4 and Q5. A speed-up capacitor is not used across R10 because it will load the monostable multivibrator.

Consider Q5 normally off. Then Q6 of the complementary driver will be biased on through R13 and R14, which in turn will bias Q9 on through R17. The output at this time will be negative with respect to ground. When Q5 turns on, Q7 and Q8 will go on, and at the same time Q6 and Q9 will turn off, which moves the output through ground to a positive level. The zener diode and rectifier combination, D1 through D4, are used together so that I_{B2} for Q6 and Q7 is cut off about 2 volts below BV_{EBO}, which is 4 volts.

By varying resistors R1 and R6, the frequency and pulse width, respectively, can be changed. The frequency band of operation can be shifted by changing C1. The pulse-width band can be shifted by changing C3. The maximum frequency of this circuit is limited by the unijunction
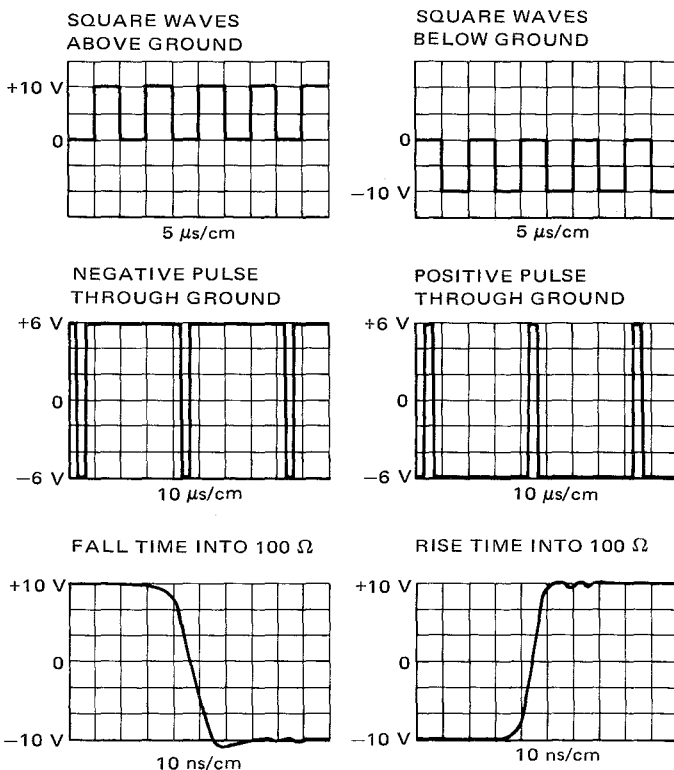
SQUARE WAVES
ABOVE GROUND

SQUARE WAVES
BELOW GROUND

5 μs/cm

5 μs/cm

NEGATIVE PULSE
THROUGH GROUND

POSITIVE PULSE
THROUGH GROUND

10 μs/cm

10 μs/cm

FALL TIME INTO 100 Ω

RISE TIME INTO 100 Ω

10 ns/cm

10 ns/cm

Figure 6-26 — Pulse Generator Output Waveforms into 100 Ω

oscillator to about 1 MHz. Switch S1 inverts the output pulse. There is about a 30% overlap in duty cycle between positions one and two with the values shown for C1 and C3. Good high frequency wiring techniques must be used to minimize ringing. The power supplies for the complementary driver must be very stiff, and the supply busses must be carefully bypassed at the transistors. The output level can be set by varying the positive and negative V_{EE} power supplies. For two-power-supply operation, $+V_{EE}$ and $-V_{EE}$ can be obtained from PS-1 and PS-2 through series-pass regulators.

6.8 Light Flashers

Simple 12 V, 1 W Light Flasher

The simple, inexpensive light flasher shown in Figure 6-27 has many advantages of more complex and expensive flasher circuits, such as flash-rate control, light-duration control, and high efficiency.

The circuit uses only two active devices: Q1, a unijunction transistor, and Q2, an NPN silicon switching transistor. Flash rate may be varied from 6 to 120 flashes per minute and light on duration from 40 ms to 0.5 seconds.

The circuit functions as follows: The firing (flash) rate of Q1 is determined by R1, R2 and C1, R1 being the flash-rate control. When the charge on C1 exceeds the firing level of Q1, it discharges through the emitter into base one of Q1, and the rate of discharge from Q1 to Q2 is
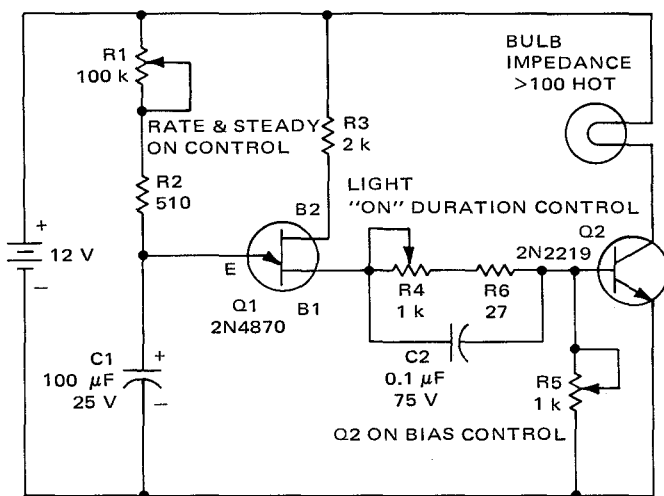

Figure 6-27 — 12 Volt 1 Watt Light Flasher

determined by R4 and C2. R5 sets the on-bias threshold of Q2 and also serves as part of the on-duration control composed of R4, R5 and C2. The high efficiency comes from a low off-state current of less than 2 mA when compared to the on-state saturation current of Q2 of approximately 100 mA.

High-Stability Flasher with Variable Flash Rate

The reliable, high-power flasher shown in Figure 6-28 is excellent for a boat, aircraft, or emergency vehicle. It is a dependable circuit with built-in protection against troublesome transients, and provision for changing the on-time duty cycle of the lamp.

Basically, the circuit is similar to many other SCR flasher circuits which make use of the familiar flip-flop. However, there are two important differences. In this circuit provision is made to guarantee that both SCRs can never come on at the same time, thus preventing latch-up of the circuit. Also, an adjustment is provided to allow the on time of the lamp to be varied from 50% duty cycle to 10% duty cycle. In the basic flip-flop, SCRs Q2 and Q3 are alternately turned on by unijunction transistor Q1 and commutated off by capacitor C1.

With only the basic circuit, a voltage transient, a momentary load short circuit, or some other disturbance could cause both SCRs to turn on at the same time. When this occurs, the circuit ceases to operate and it
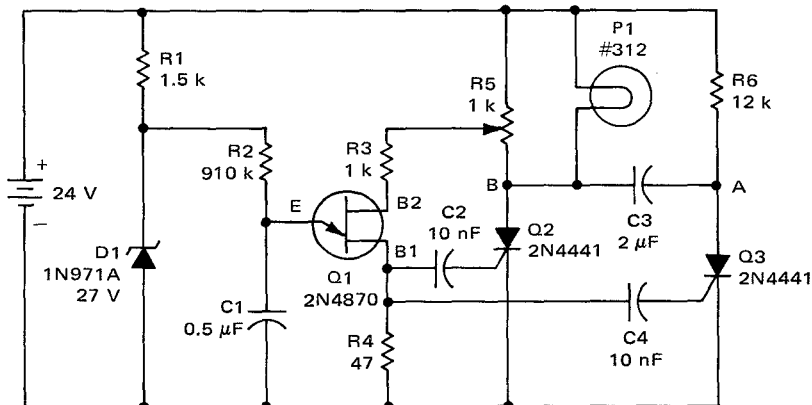

Figure 6-28 — High-Stability Flasher

cannot resume operation until the supply voltage is interrupted momentarily. In this circuit, however, resistor R6 is made large enough that the current which it supplies to Q3 is always less than the holding current. Therefore, Q3 can never remain in conduction. This improvement also eliminates the need for an elaborate starting circuit. Being able to vary the on-time duty cycle of the lamp improves the efficiency of this flasher considerably. A 30% duty cycle not only has been proven to be much easier to see than a 50% duty cycle but also realizes a 40% savings in overall power dissipated.

The circuit operates as follows: The unijunction transistor (Q1) operates as a stable relaxation oscillator which produces trigger pulses for both SCRs. The rate of oscillation varies between flashes in the following manner: When the lamp (P1) is off, the voltage at base two of Q1 is approximately equal to the supply voltage. This means that C2 must charge to the supply voltage times the intrinsic standoff ratio (η) of Q1 before Q1 will fire. When the lamp (P1) is on, however, and the voltage on the wiper of R5 is, let us say, one-half the supply voltage, then capacitor (C1) must only charge to one-half the supply voltage times the η of Q1 before Q1 fires. Since the capacitor is being charged from the full supply voltage, this time will be much shorter, therefore the lamp will stay on a much shorter period of time than it will stay off.

At the start of operation, both SCRs attempt to turn on. Q3 cannot turn on because its anode current is limited by R6 and it never reaches the holding current level. With Q2 in conduction, point B is reduced to a voltage level which is above ground by an amount equal to the forward drop of Q2. C3 charges through R6 with point A rising to the supply voltage. When the next trigger pulse occurs, Q3 can turn on since it is supplied with anode current by C3. With Q3 in conduction, Q2 is effectively reverse-biased by C3, which causes it to turn off. The load current is supported by C3 for a few microseconds during the turn off interval of Q2. Q3 continues to remain in conduction as C3 charges through the lamp with point B rising to the supply voltage. When the sum of the charging current of C3 and the current through R6 drops to a value below the holding current of Q3, Q3 drops out of conduction. C3 then discharges through R6 and the lamp at a low current level. When the next trigger pulse arrives, the cycle is repeated.

Note that when Q3 discharges C3 to turn off Q2, Q2 is also being supplied with a turn-on trigger pulse. The trigger pulse must be kept short in duration in comparison with the turn-off pulse supplied by C3, to insure that Q2 will turn off. If for some reason Q2 should fail to turn off as intended, C3 will recharge with point A positive with respect to point B, and thus turn-off pulses will be supplied until it does turn off. Likewise, if

one of the SCRs should fail to turn on, turn-on pulses will continue to be applied until it does turn on. Since the firing voltage of the unijunction transistor is relatively insensitive to changes in base voltage, the repetition rate of the trigger pulses are not appreciably affected by voltage changes.

The resistor (R3) in series with base two reduces the effects of temperature changes, and thus the repetition rate is relatively insensitive to temperature changes.

If severe transients are encountered, it is recommended that resistor R1 and zener diode D1 be added to protect the unijunction transistor.

6.9 Relay Drivers

Electronic control of either ac or dc relays is possible through the use of a transistor placed in series with the coil of the relay. This must be done through a bridge when an ac relay is used. Figure 6-29 is the schematic of a dc relay that can be controlled by an electronic signal which can supply a current of 1/2 mA. The relay coil is the load for transistor Q1. When a positive control voltage is applied to R1, Q1 receives base drive and saturates, thus connecting the relay coil to the supply voltage. R1 must allow enough base current to saturate Q1. For the components shown, the 1/2 mA of base current will assure this since the relay requires 5 mA and the transistor has a minimum gain of 55. When the control voltage drops to zero, the transistor turns off, de-energizing the relay. Since the relay coil is inductive, a voltage spike could occur at the collector of Q1; to prevent damage, a protective circuit for the transistor should be included. The diode (D1) across the relay coil does this; it clamps the collector voltage to the supply voltage by providing a path for the current in the relay coil when the transistor turns off.

A modification of this relay circuit is shown in Figure 6-30. In this case, the control of the relay is provided by light. When sufficient light is directed at Q1, it turns on. This drives Q2 which energizes the relay coil as
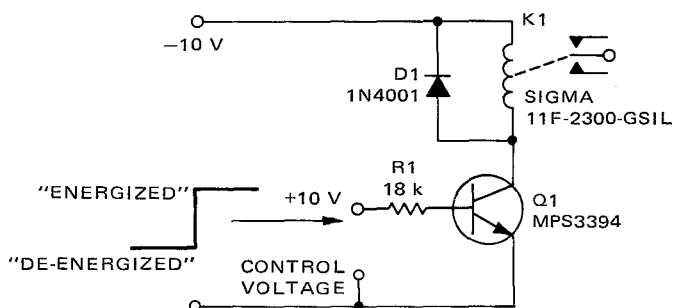

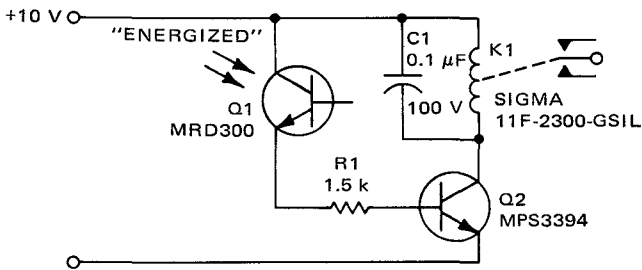**Figure 6-29 — Electronic Control of a DC Relay**
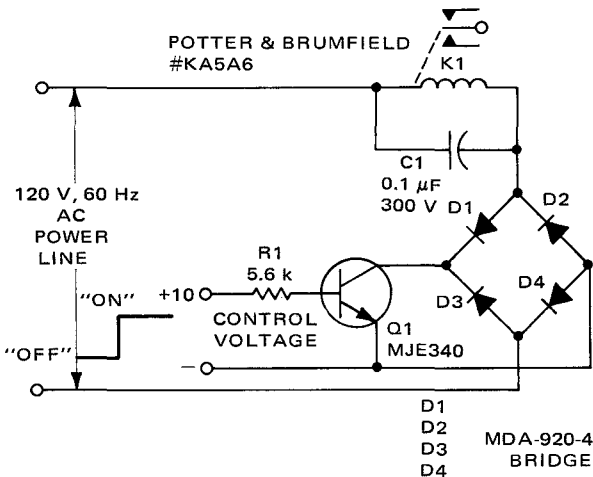
Figure 6-30 — Light-Operated Relay

Figure 6-31 — Electronic Control of an AC Relay

in the previous circuit. A light magnitude of 220 foot-candles was enough to drive relay driver Q2 to saturation. When light is removed from Q1, base drive is removed from Q2, and Q2 turns off. In this circuit, Q2 turns off slightly more slowly than in the previous circuit, therefore a small capacitor (C1) across the relay is adequate to limit the maximum voltage spike to below the breakdown level of Q2.

The circuit shown in Figure 6-31 can be used to control an ac-operated relay. The bridge consisting of D1, D2, D3 and D4 provides dc to the transistor while the relay sees an ac voltage. When a dc control voltage is applied to R1, Q1 saturates and energizes the relay coil. As before, adequate base current must be provided to saturate Q1. A disadvantage of this circuit is that the control signal must be isolated from the power line. The prime advantage is, of course, that an ac relay can be controlled by a

single transistor. A forced gain of 10 guarantees that Q1 will be saturated, therefore the base current of 1.6 mA will drive a relay coil requiring 16 mA. In this circuit, protection against voltage spikes must also be provided for the transistor when it is turned off. Capacitor C1 across the relay coil provides such protection.

6.10 Power Supply Monitors

The circuit shown in Figure 6-32 provides a visual indication of momentary interruptions of a power source. The circuit operates as follows: the push-button switch momentarily applies power to the gate of SCR Q1, which will turn it on and clamp the voltage across the lamp to the forward voltage drop of the SCR. This voltage (about 1 volt) is not enough to light the lamp. As long as power is present the SCR will be on and the lamp will be off. If there is a momentary interruption of the power for a time which is greater than the recovery time of the SCR (no greater than 50 μs at 240 mA), then the SCR will turn off and remain off. When power service is resumed, the lamp will light indicating that the service had been interrupted. Resistor R2 clamps the gate of the sensitive-gate SCR and prevents spurious signals from firing the SCR. The only way to extinguish the lamp when power is applied is to close the start switch, thus the lamp will glow until noticed. Then the circuit can be reset manually. This circuit will only indicate momentary power failures.

Should an indication of either momentary or continuous power failure be required then the circuit shown in Figure 6-33 can be used.
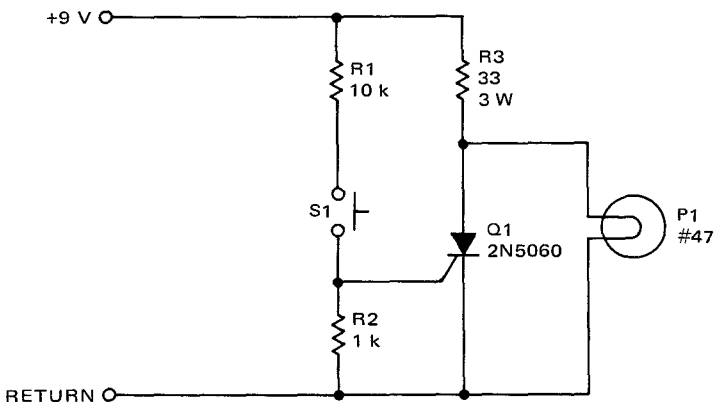

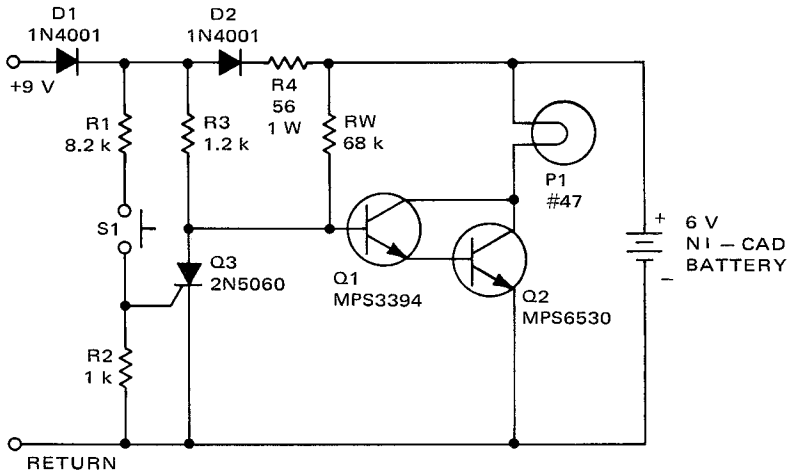**Figure 6-32 — Interrupted-Power Indicator**

Figure 6-33 — Power-Failure Indicator

When the push-button switch is closed, SCR Q3 turns on and holds the base voltage of Q1 below the value required to turn it on. Resistor R3 is used to provide holding current to SCR Q3. The SCR will stay on as long as power is applied to the circuit. If the power fails either as an open or a short, Q3 will turn off since diode D2 prevents the battery from providing holding current through R3, and the current through R5 is less than the holding current. If the power source fails as a short, then diode D1 prevents the current through R5 from bypassing the base of Q1 through resistor R3 and the shorted supply. This guarantees that Q1 and Q2 will turn on, thus lighting the lamp with current supplied by the battery. If the power should come back on, the SCR will remain off and the lamp will remain lit. This circuit will also indicate momentary power failures if the duration of the failure is greater than the recovery time of the SCR. This time is approximately 5 μs when the current is about 7 mA. When the circuit is reset, the supply will trickle-charge the battery, thus keeping it at full charge.

6.11 Battery-Charger Control Circuit

The foolproof battery-charger control circuit shown in Figure 6-34 protects a battery being charged from overcharging or reverse charging. It will also protect itself and/or a separate charging supply from short-circuit damage.

The power transformer and full-wave bridge rectifier can supply approximately 16 A of charging current to the battery.
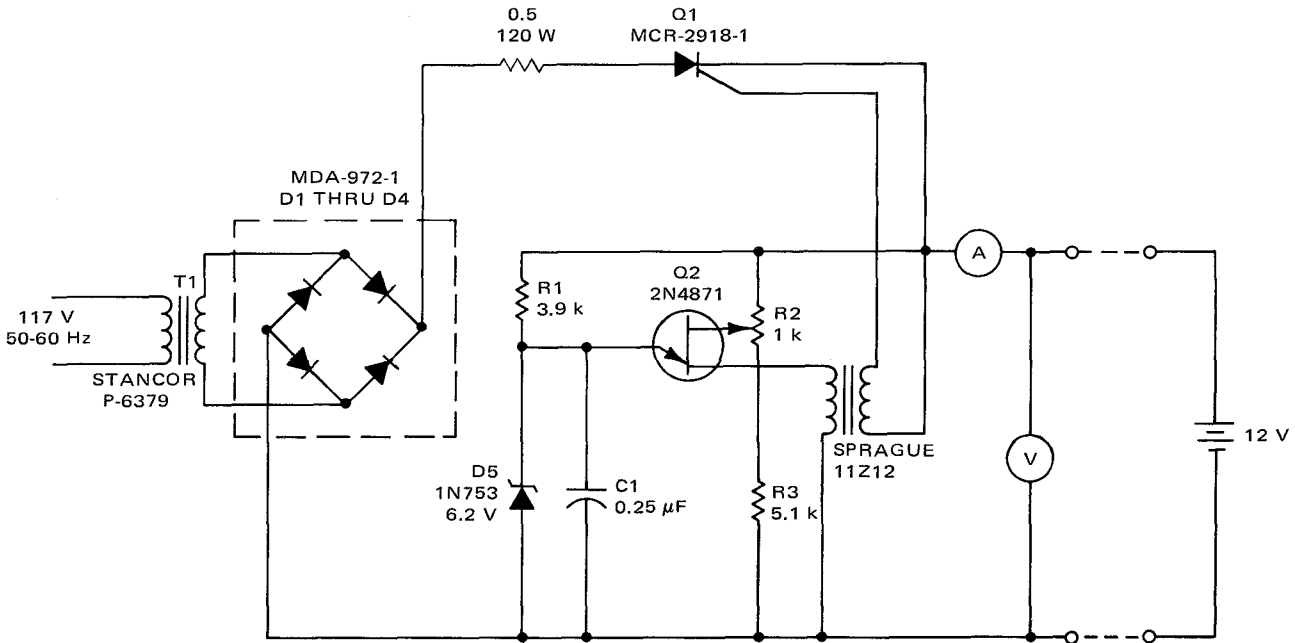
Figure 6-34 — 12 Volt Battery-Charger Control

The unijunction transistor (Q2), R1, R2, R3 and C1 form a relaxation oscillator which is used to trigger SCR Q1 through transformer T2. Power for operation of the relaxation oscillator is obtained from the output which is connected to the battery. The interbase voltage, V_{B2B1}, of the unijunction is therefore determined by the battery voltage. Because the firing voltage of the unijunction is a function of the interbase voltage, as the battery charges and its terminal voltage increases, the firing voltage also increases. The zener diode (D5) limits the voltage to which the emitter of the unijunction can rise. When the required firing voltage of the unijunction as determined by the battery voltage exceeds the breakdown voltage of D5, the unijunction can no longer oscillate. It therefore cannot trigger SCR Q1 and the charging ceases. This voltage cutoff point is controlled by the setting of R2.

The unijunction cannot oscillate unless a positive voltage less than the cutoff setting is present at the output terminals. Therefore, the SCR cannot conduct under conditions of a short circuit, an open circuit, or a reversed polarity connection to the battery.

The trigger transformer specified functions satisfactorily, but its design is not critical. The basic requirements are that its series impedance be low, and that it must be capable of passing a pulse with a width of a few microseconds.

6.12 Remote Strobeflash Slave Adapter

At times when using an electronic strobe flash it is desirable to use a remote, or "slave" flash synchronized with the master. The circuit in Figure 6-35 provides the drive needed to trigger a slave unit, and eliminates the necessity for synchronizing wires between the two flash units.
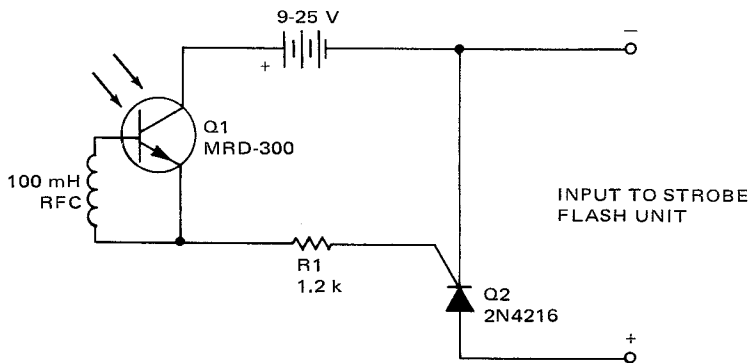

Figure 6-35 — Strobeflash Slave Adapter

The MRD-300 phototransistor used in this circuit is cut off in a V_{CER} mode due to the relatively low dc resistance of rf choke L1 even under high ambient light conditions. When a fast-rising pulse of light strikes the base region of this device, however, L1 acts as a very high impedance to the ramp and the transistor is biased into conduction by the incoming pulse of light.

When the MRD-300 conducts, a signal is applied to the gate of SCR Q2. This triggers Q2, which acts as a solid-state relay and turns on the attached strobeflash unit.

In tests this unit was unaffected by ambient light conditions. It fired up to approximately 20 feet from strobelight flashes using only the lens of the MRD-300 phototransistor for light pickup. This distance could be greatly increased if a lens or parabolic reflector were used to concentrate light into the phototransistor.

BIBLIOGRAPHY FOR CHAPTER 6

1. Bergersen, Thor, "Unijunction Transistor Timers and Oscillators," Motorola AN-294.

2. Brookmire, James L; "Zero-Voltage Switching for Proportional Temperature Control," *EEE*, April 67.

3. Buchanan, John K.; "A Modulated SCR Zero-Point Switching Circuit" Motorola AN-242.

4. Wechsler, Reuben, "A Unique Battery Charger Control Circuit" Motorola

5. Wechsler, Reuben, Designing SCR Circuits for High Inrush Loads" Motorola AN-170.

6. Wechsler, Reuben, "Reducing (di/dt)-Effect Failures in SCR's" Motorola AN-173.

7. Zinder, David A., "SCR Power Control Fundamentals," Motorola AN-240.

8. Zinder, David A., "Thyristor Trigger Circuits for Power-Control Applications," Motorola AN-227.