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Application Note

SCR CONTROLLER FOR A SERIES FIELD DC MOTOR

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A dc chopper using SCRs was built and installed in a golf cart. This note gives general design formulas and uses these to design a chopper for a 2 hp 36 V series wound motor. The note also discusses motor characteristics, SCRs versus transistors, and simple control methods.

MOTOROLA Semiconductor Products Inc.

SCR CONTROLLER FOR A SERIES FIELD DC MOTOR

INTRODUCTION

This application note describes in detail an SCR chopper for a series field dc motor, beginning with a short discussion of motor characteristics and various simple control approaches. Chopper methods and the chopper's integration with auxiliary switching functions are described. The pros and cons of transistors versus SCRs are discussed. Finally an SCR chopper is described in detail, and an actual design shown for a controller using a 36 V, 2 hp motor in a golf cart.

MOTOR CHARACTERISTICS

Series field dc motors are used in traction applications such as fork-lift trucks and rapid transit where high starting torque is required. The series field motor has high current flowing through it at zero speed, since the motor back-emf is zero also. This high current causes high motor flux, since it flows through the field winding as well as the rotor. The high motor flux causes high torque. As the speed picks up, the current and torque reduce until equilibrium is reached, where the torque output balances vehicle losses, friction and windage, etc. Figures 1a and 1b show typical series traction motor characteristics, and Figure 1c shows the circuit schematic.

MOTOR CONTROL

If more than OFF-ON control is needed, the voltage to the motor must be variable. At low speeds, measurements in a golf cart have shown that vehicle speed is roughly proportional to applied voltage. See Figure 2, which also shows motor current.

Battery Switching

The curves in Figure 2 were obtained by moving the connection from the 36 V battery to the golf cart along the battery bank in 6 V steps. Some speed control systems use four equal battery banks in series and parallel to vary motor voltage and therefore speed. This technique loads the batteries equally, and is the most efficient.

The main disadvantage of battery switching is jerky vehicle behavior as the vehicle accelerates through the speed range when the accelerator is depressed. Also in-rush currents can be high, and contactor life limited. The contactor bank is quite expensive.

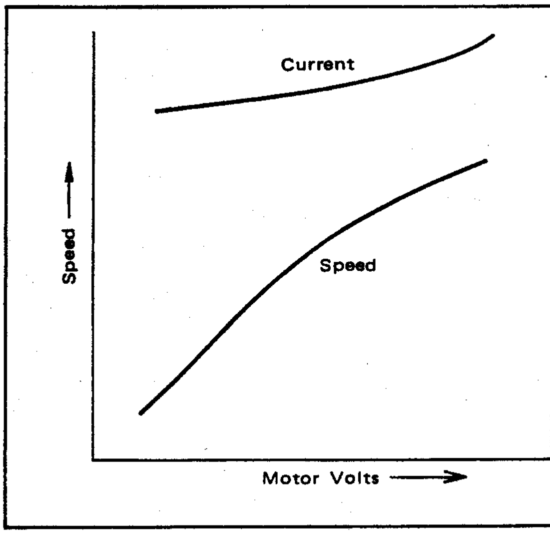


FIGURE 2 – Vehicle speed and Motor Current versus Applied Voltage

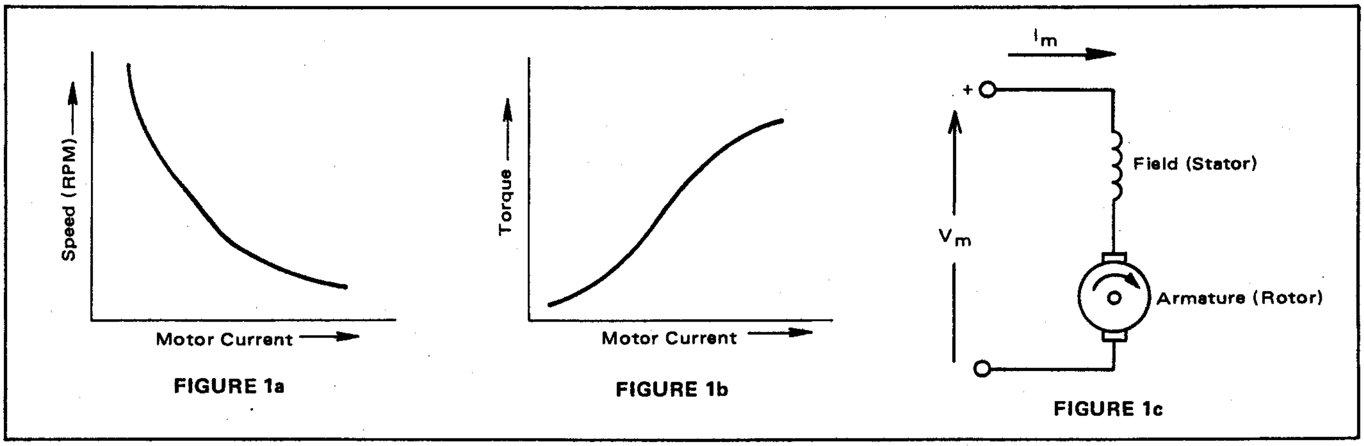


FIGURE 1 – Series Motor Characteristics and Schematic

Resistor Control

Where jerky speed control is of limited importance, for instance in an outdoors application such as a golf cart, then resistor switching, either by contactors or switches, may be used. Figure 3 shows a typical circuit. R1, R2, and R3 are equal power resistors, and SW1 is a 5-position switch ganged to the accelerator pedal. As the accelerator pedal is depressed, the SW1 wiper moves counter-clockwise from the OFF position, progressively reducing the series resistance, and allowing the vehicle to accelerate.

Such a control is economical in parts cost, but expensive in battery charge if there are many stop-start or intermediate speed cycles in the vehicle's usage.

Thus if smooth control together with maximum battery utilization is required, another type of speed controller must be used.

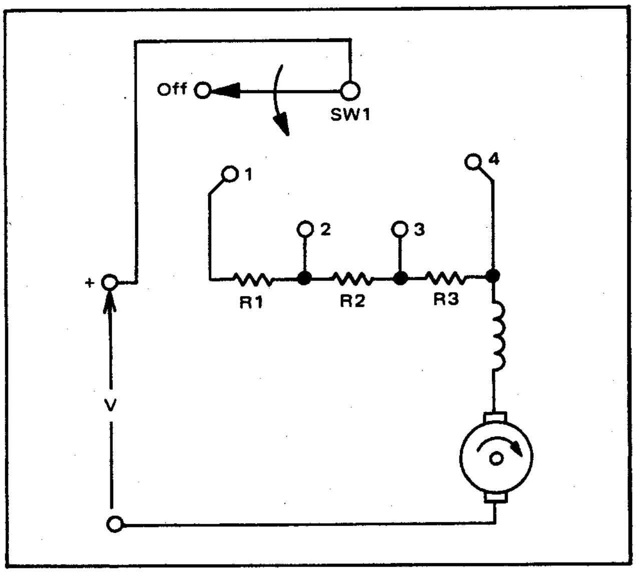


FIGURE 3 – Resistor Control

DC Chopper Controller

The dc chopper controller uses a solid-state switch, usually a silicon controlled rectifier (occasionally a transistor) to "chop up" the battery voltage and apply it to the motor. Figure 4 shows the solid-state switch in series with the motor. The rectifier D1 is the well-known "freewheel rectifier" and it carries the motor current when the switch opens, thus maintaining nearly constant motor current.

Figure 5a shows the current flow from the supply, through the motor and the solid-state switch (represented by a contact) and back to the supply. The freewheel rectifier is reverse biased, and shown in phantom, as it is essentially out of circuit. The current, I1 in Figures 5a and 5c, ramps up at a rate dependent upon speed, voltage, motor, etc.

Figure 5b shows the switch open, and the motor current flowing through the freewheel rectifier. This is shown as I2, ramping down as the inductively stored energy produces work at the motor shaft. For this period the supply is out of the circuit.

The motor current is the average of I1 and I2. The average supply current is the average of I1 only, or about half the motor current at a 50% duty cycle. These ratios vary with the ON-to-OFF ratio, or the duty cycle of chopper.

There are two commonly used methods of varying the duty cycle of the chopper, and thus the speed of the vehicle.

1. Fixed Pulse Width, Variable Repetition Rate Chopper – This is perhaps the most common type of chopper circuit in use. The battery or supply voltage is applied to

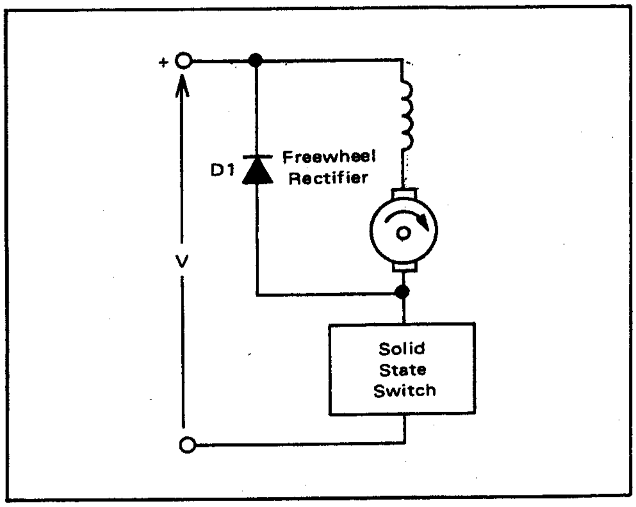


FIGURE 4 – Basic Chopper Circuit

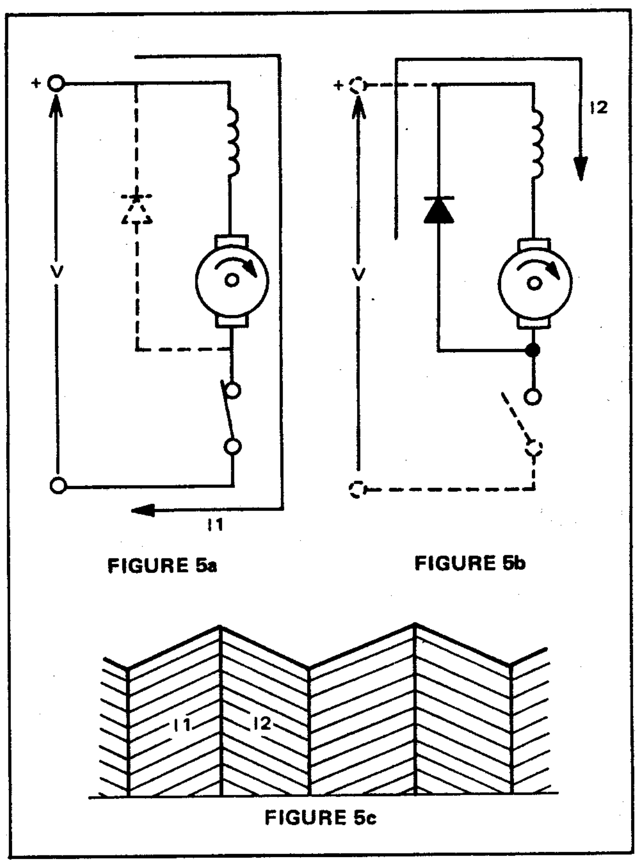


FIGURE 5 – General Chopper Waveforms

the motor in fixed width pulses. The pulse repetition frequency is then varied to increase or decrease the average voltage applied to the motor. Figures 6a and 6b show chopper voltage and current waveforms for low speed and high speed settings respectively. A voltage control range of 20 to 80% is usually claimed for this system.

See Figure 6a and 6b on next page

2. **Pulse Width Modulated Chopper** – This type of chopper runs at a nominally fixed frequency, and varies the duty cycle at that frequency. A broader range of control is available, typically 10 to 90% and possibly 5 to 95%. Figures 7a and 7b show the pulse width modulated chopper at low and high average voltage respectively. Motor current is also shown.

A pwm SCR chopper is described in detail on the following pages.

See Figure 7a & b on page after Fig 6

MAXIMUM SPEED CONNECTIONS

It is usually difficult to run at 100% duty cycle with the chopper, because of circuit constraints. Since there is no need for control (the vehicle is running at full speed) the solid-state switch would be a liability as it does have some voltage drop across it. Most controllers, therefore, have a shorting switch across the solid-state switch, shown in Figure 8 as SW1. SW1 is normally pulled in by a microswitch actuated by "flatting" the accelerator pedal, although it is sometimes delayed to avoid wheel spin. Shorting out the SCR means that the dissipation in the device is reduced to zero. This, combined with thermal feedback from the SCR heatsink itself, enables the designer to minimize SCR heatsink ratings.

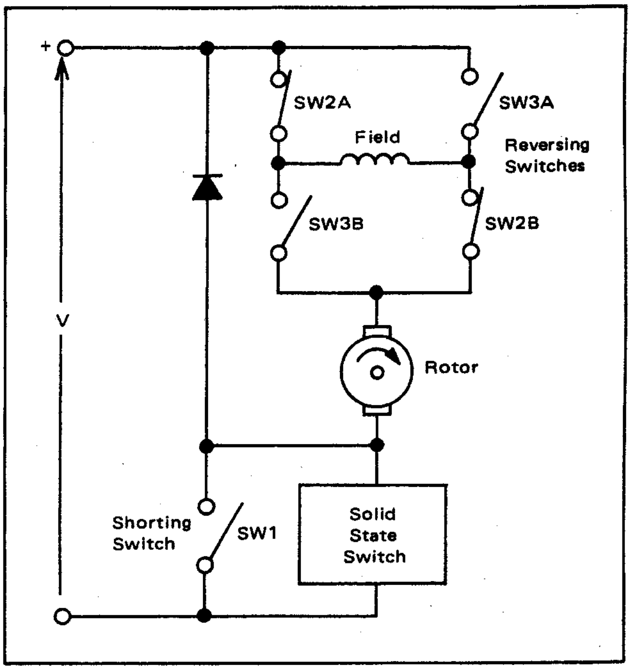


FIGURE 8 – Maximum Speed and Reversing Connections

REVERSE CONNECTIONS

Figure 8 shows the normal connections for reversing. SW2 and SW3 are ganged together such that if the contacts of SW2 are closed, then the contacts of SW3 are open and vice-versa. Change-over reverses the direction of current in the field winding, and the motor then reverses its direction of rotation. An alternative would be to reverse connections to the rotor rather than the field.

PLUGGING

As soon as a reversing switch is installed, plugging becomes a design consideration. If the vehicle is moving forward, and the controls are thrown in reverse, the condition arises where the motor is driven as a generator by the momentum of the vehicle. The "generator" voltage is series additive with the supply as shown in Figure 9 and large currents can flow, resulting in a high braking force on the vehicle.

Operating the chopper in this mode can be a problem as the chopper must be designed for these high currents. It is possible to sense the plugging current with a relay coil by sensing the current in an additional rectifier used as the "plugging rectifier". This relay would either disable the chopper or operate the chopper at a controlled rate, producing controlled vehicle braking. Controlled braking system.

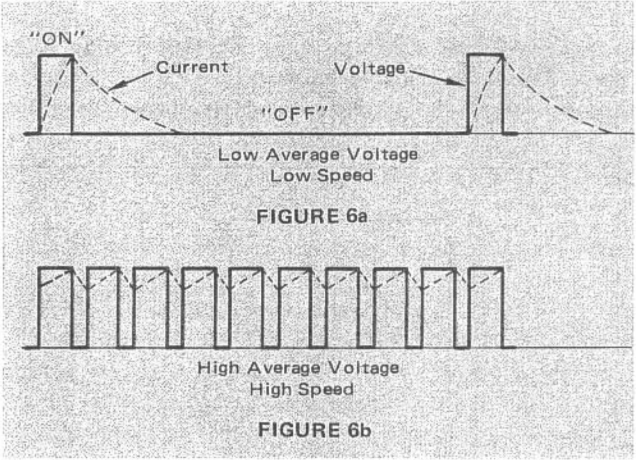


FIGURE 6 — Fixed Pulse Width Chopper Waveforms

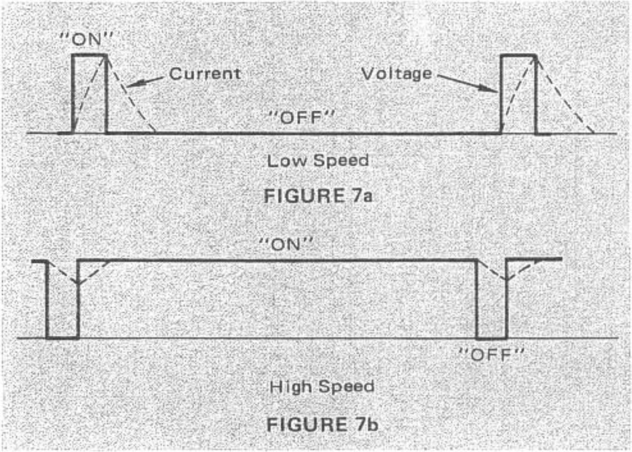


FIGURE 7 — Pulse Width Modulated Chopper Waveforms

is necessary, as the braking effort available in this mode may be sufficient to damage the mechanical transmission.

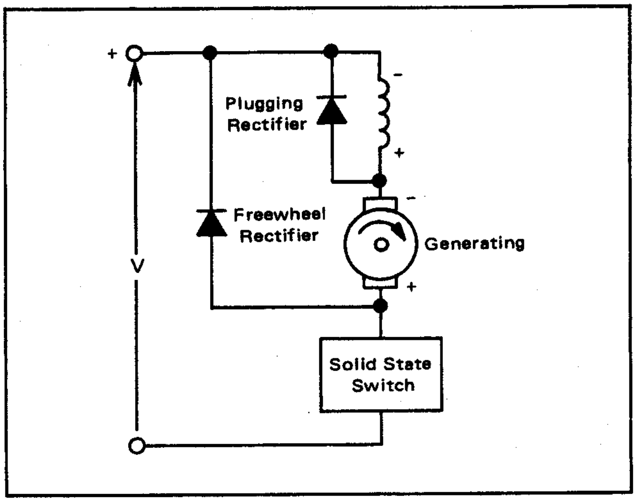


FIGURE 9 – Motor Plugging

CHOPPER CURRENT RATING

Returning the Figure 1a, starting or locked rotor currents for the series field motor can be very high. At locked rotor, the current is limited only by the motor resistance and circuit (battery, wires, switches, etc.) impedance. For a 36 V, 2 hp motor, running current is approximately 60 A. Motor resistance is approximately 60 mΩ, and with low circuit impedance and a healthy set of batteries, locked rotor currents of 500 A are possible.

While 500 A will produce the maximum locked rotor torque available, design requirements may be less stringent. A rule of thumb seems to be a factor of 5, leading to a maximum required average motor current of 300 A. This would be sufficient to drive away from stop up a moderate slope, for instance, and is known as the breakaway torque current.

Fortunately, when power is applied, the current takes time to build up due to the inductance of the motor. For the motor considered here the inductance is about 300 μH. This gives a motor time constant, L/R, of 5 ms, or current rise times of about 70 A/ms, close to what was observed. Time is available to sense motor current and turn the solid-state switch OFF, thus reducing switch current rating, as shown in Figure 10.

See Figure 10 on the next page

CHOPPER FREQUENCY

Every time the chopper "chops," either from ON-to-OFF or vice versa, some losses inevitably occur. These losses show up in the various circuit elements to be described later such as inductors, commutating circuitry, etc.

Normal Running

To minimize these losses, the chopper frequency should be kept as low as possible. This also helps to maintain as low as possible the frequency of the audio noise generated by the chopper and motor and reduces the possibilities of irritation to the user. However, to maintain a continuous current in the motor at half voltage, or 50% duty cycle in the pulse width modulated system, some minimum frequency is required. When the frequency is too low, as shown in Figure 11a, the motor current in the freewheel (rectifier) mode falls to zero. Then in order to maintain the average motor current, the peaks are higher than in Figure 11b, where the frequency is higher. Operation in the lower frequency mode, (Figure 11a) where the motor current is allowed to fall to zero, results in torque pulsations and high rotor commutator brush wear. Some discontinuity of current is inevitable at lower speeds and duty cycles. The compromise frequency chosen for the 2 hp machine was 125 Hz.

See Figure 11 on page after Fig 10

HIGH TORQUE RUNNING

In a high torque situation, such as moving up a slope or pulling away from a stopped position, the motor speed will drop. This in turn reduces the back emf of the motor

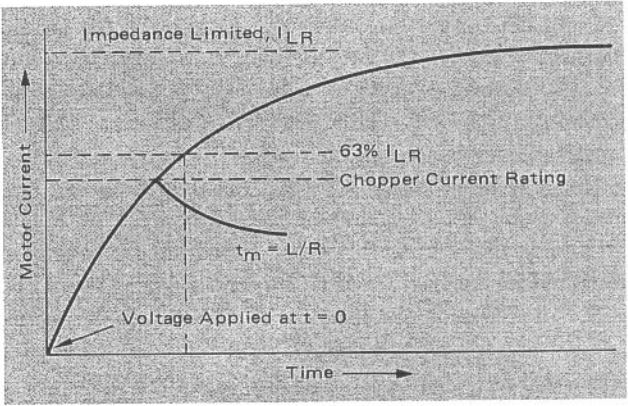


FIGURE 10 – Locked Rotor Current

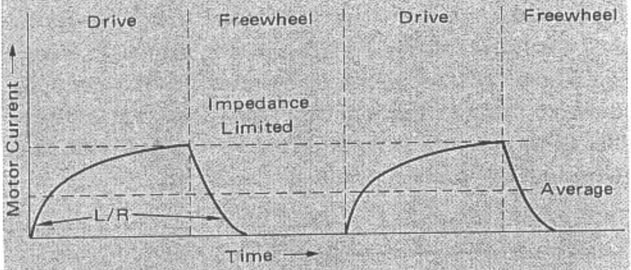


FIGURE 11a

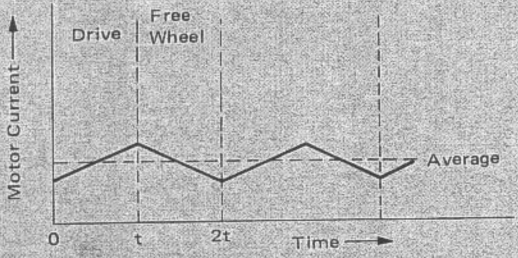


FIGURE 11b

FIGURE 11 — Variation of Chopper Frequency

resulting in a lower net voltage appearing across the motor inductor-resistor combination. Rates-of-rise of motor current therefore increase. If the chopper frequency is maintained at the normal running rate, the current will rise to too high a level for the chopper to handle unless the ON time of the chopper is reduced. However, doing only this would lower the average motor current and give low output torque. The solution is to reduce the ON time of the chopper and at the same time reduce the OFF time. This permits high average motor current and yet limits the peak current that the chopper must handle. In Figure 11 this would result in increased current and decreased time per division, with the waveforms remaining approximately the same. Frequency and current may go as high as 5 times normal running levels.

THE SOLID-STATE SWITCH

The choice of solid-state switch is transistors, either one device or several in parallel, and the Silicon Controlled Rectifier (SCR). The majority of systems in use today use the SCR.

THE TRANSISTOR SWITCH

Consider the system discussed so far, the 36 V, 2 hp motor drive. For breakaway torque requirements to be met, 300 A average is required. Single transistors nominally rated at 300 A are available, though more commonly parallel transistors have been used. When parallel transistors are used, steps must be taken to ensure that the collector current shares, either by matching devices or by emitter resistor ballasting, as indicated in Figure 12. These resistors are typically rated to drop a voltage equivalent to the base-emitter voltage of the transistor used (V_{BE}), approximately 0.7 V for Silicon transistors and about 0.4 V for Germanium transistors.

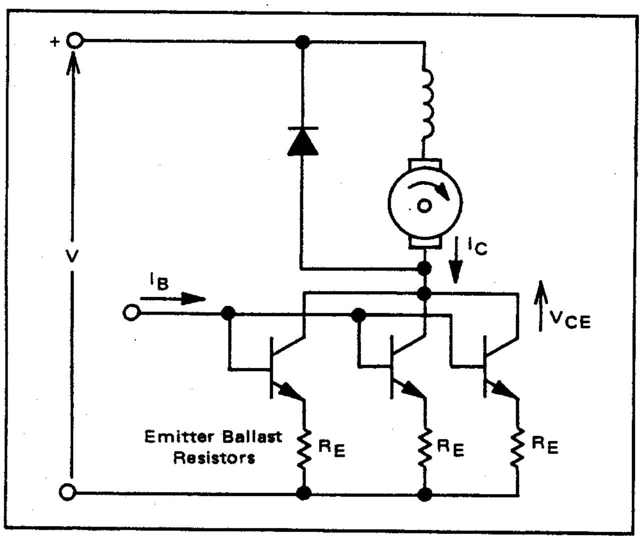


FIGURE 12 – Transistor Solid State Switch

The base drive requirements are usually quite high. If a gain of 10 for the transistors is assumed, then 30 A must be controlled at the base. The emitter resistors will also help force sharing of this current.

As an alternative to Silicon transistors, Silicon Darlington transistors are available. These have higher gain but lower speed and higher saturation voltage. Germanium transistors are also available and have lower saturation voltage but slower speed.

Speed is mentioned as this is a switching system, and an inductive switching system as well. The switching characteristics of the transistor are very important. Figure 13 shows a full ON-OFF cycle, though not necessarily to scale. At interval t0 to t1, the device is off. For t1 to t2 the collector-emitter voltage (V_{CE}) is falling and the collector current is rising as the on-going transistor picks up the current from the freewheel rectifier. The spike of current during this interval is caused by the stored charge being recovered from the freewheel rectifier as it becomes reverse biased. During interval t2 to t3 the transistor is ON. At time t3 the transistor starts to turn off. However the inductive current of the motor load will continue to flow through the transistor until the voltage reaches the supply voltage plus one diode drop. Then the freewheel rectifier will pick up the motor current.

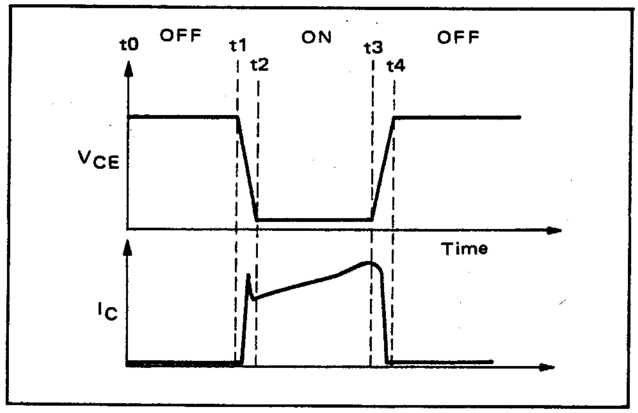


FIGURE 13 – Transistor Switch Waveforms

Intervals t1 to t2 and t3 to t4 are highly dissipative, and for slow devices may limit the maximum chopping frequency. Load line shaping and the use of a fast recovery rectifier for the freewheel rectifier may help. The inductive turn-off means that the safe operating area characteristic of the transistor must be considered.

Transistors may be used in parallel. However, if they switch at different times, because of variation in storage, rise and fall times from device to device, then the fastest transistor will turn on into the rectifier recovery current, and the slowest device will switch off all the current as the bank turns off. These two devices therefore will be more highly stressed than the others.

The advantage of the transistor is that it can be controlled from its base, unlike the SCR which responds only to an ON signal at its gate. However, the problems described above have resulted in most systems using the SCR. The remainder of this note describes an SCR design.

THE SILICON CONTROLLED RECTIFIER SWITCH

The SCR has a latching characteristic and pulse firing (gating) may be used. This means that the SCR's anode current must be reduced to zero, either by turning off a switch in series with it, or, more commonly, reverse biasing the anode for a time sufficient for it to turn off. This usually results in one or two lower current SCRs, a power capacitor, and a power inductor in the commutating circuit, the whole comprising the solid-state switch. However, paralleling of SCR's is usually unnecessary. Figure 14 shows a well known chopper circuit, the Jones Chopper. This circuit will be explained in detail, design data given, and a control scheme described.

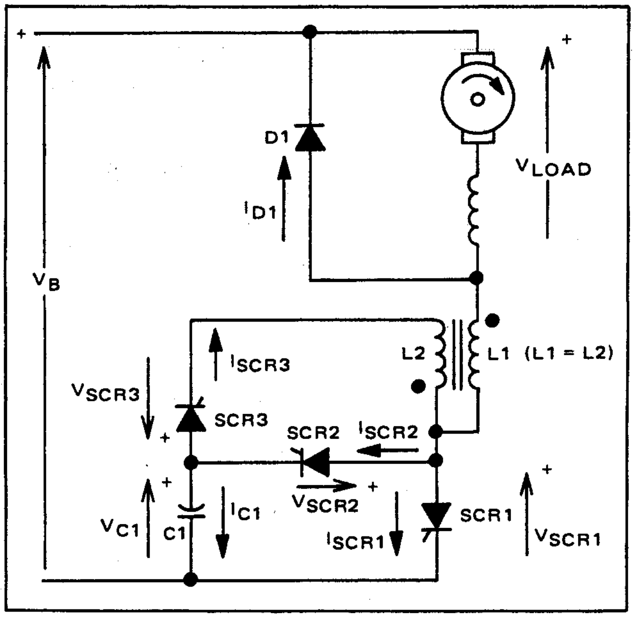


FIGURE 14 – Jones Chopper Power Circuit Diagram

THE JONES CHOPPER

The main advantage of the Jones Chopper over other types is that it allows the use of higher voltage, lower microfarad commutating capacitors. This is done by trapping energy in the commutating inductor (Figure 14) L1, and forcing it into the commutating capacitor C1, rather than simply charging the commutating capacitor to the supply voltage V_B. This of course means that the semiconductors and the capacitor are rated at a multiple of V_B, but as most vehicle battery banks are 110 V or less, this requirement can be met with ease.

Circuit Operation

Figure 15, Chopper Circuit Waveforms, is used with Figure 14 to explain circuit operation. The switching intervals of the waveforms, t1 to t4, and t5 to t8 have been expanded for clarity.

Time t0

All SCRs are off. Capacitor C1 is charged positively to some multiple of V_B. D1 is in conduction and its current I_{D1} (and therefore the motor current) is ramping down.

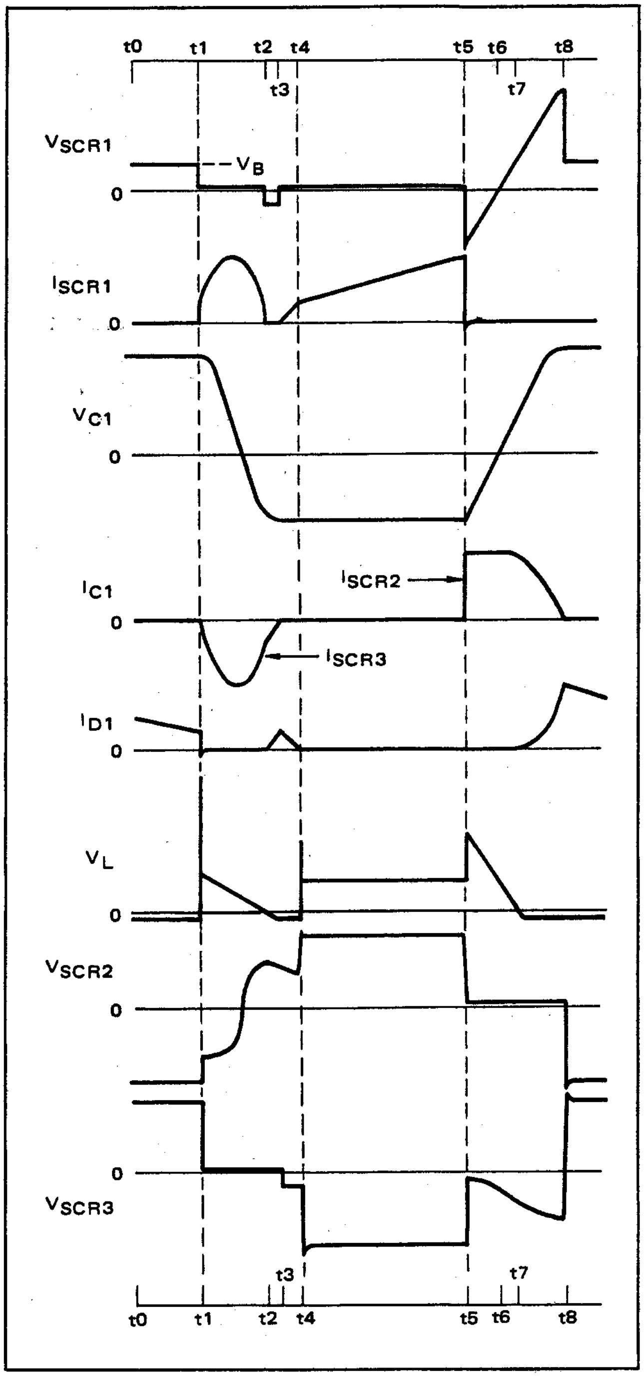


FIGURE 15 – Chopper Circuit Waveforms

Time t1

SCR1 and SCR3 are gated on.

Interval t1 to t2

C1 discharges resonantly through SCR3, L2 and SCR1. This discharge current does not flow through L1 and back to the battery because of the transformer action of L2 coupled to L1. SCR1 also picks up the motor current from D1, reverse biasing D1. The reverse recovery current of D1 causes a voltage spike across it from circuit inductance. This must be snubbed out (see Figure 17) and can be reduced by using a fast recovery rectifier.

As V_{C1} swings negative the reverse bias on D1 decreases until after a time equal to $\pi[(L2)\ (C2)]^{1/2}$ at ....

Time t2

D1 becomes forward biased again, causing the change of slope in the $I_{C1}(I_{SCR3})$ waveform.

Interval t2 to t3

Due to the voltage developed across L1, SCR1 becomes reverse biased in this interval and must have its gate signal inhibited through this period, and then regated at time t3. I_{SCR3} decreases until at zero current the current tries to reverse resonantly when at ....

Time t3

SCR3 blocks, leaving C1 charged negatively with 50% to 90% of its previous positive voltage, the rest of the energy having been dissipated in the circuit. SCR1 is re-gated and the current that had built up in L1 and D1 now decreases for the ....

Interval t3 to t4

and flows through SCR1.

Time t4

I_{D1} comes to zero and again causes a reverse recovery spike on V_L.

Interval t4 to t5

SCR1 is now conducting motor current, which is ramping up. The interval t3 to t5 must be long enough to allow SCR3 to commutate. C1 is charged negatively in preparation for commutation of SCR1 at ....

Time t5

when SCR2 is gated on. The negative voltage on C1 is applied across SCR1 which blocks after its recovery current has ceased. The motor current, nominally constant, starts to flow in SCR2 and C1. The rate-of-rise (di/dt) of this current in SCR2 is limited only by circuit stray inductance and by any designed-in saturable inductor.

Interval t5 to t6

For this period, SCR1 is reverse biased and the motor current is charging C1 positively. This interval must be long enough to allow SCR1 to reform its forward blocking junction (commutate) since at ....

Time t6

C1 has charged through zero volts and SCR1 becomes forward biased.

Interval t6 to t7

Motor current continues to flow through L1 and SCR2 charging C1 positively until at ....

Time t7

V_{C1} reaches V_B, forward biasing D1, the freewheel rectifier. D1 begins to pick up the motor current and $I_{C1}(I_{SCR2})$ starts to reduce.

Interval t7 to t8

Here the energy ($1/2LI^2$) stored in the inductor L1 is being forced into C1 and is charging it positively ($1/2CV^2$). I_{C1} is decreasing and I_{D1} increasing until at ....

Time t8

D1 has picked up all the motor current. SCR2 blocks and all SCRs are now off.

Interval t8 to t0 to t1

Motor current is ramping down. The period t8 to t1 must be long enough to allow SCR2 to fully commutate off, and at ....

Time t1

the cycle begins again.

JONES CHOPPER DESIGN PROCEDURE

If a complete system is being designed from scratch, then the designer may have all the variables of motor size, battery voltage, breakaway torque, etc., at his command. Battery and motor constraints coupled with the type of usage expected will dictate the design numbers of the main parameters. These are:

The battery voltage – V_B

The maximum average rotor current – I_m

The motor time constant – t_m

The battery voltage will usually be between 24 V and 110 V. For the chosen motor, I_m is the current that will supply the torque for the breakaway load condition, and t_m is the L/R time constant for that motor under locked rotor conditions.

Component Voltage Rating

The SCRs and the commutating capacitor C1 will all have some maximum economical voltage rating, such as QV_B for the most constraining component, where Q is the circuit voltage multiplier factor. The commutating capacitor however does not retain all of its energy when it is resonantly "rung around", and will be negatively charged at the instant of commutation to KQV_B, where K lies between 0.5 and 0.9.

Commutating Capacitor

The commutating capacitor C1 must divert the nominally constant motor current long enough so that before C1 is charged to zero volts, the SCR is off. From the charge formula CV = it,

$$C1 = \frac{I_m T_x}{KQV_B}$$

A safety factor for t_q of 100% is recommended here, therefore $T_x = 2\ t_q$, where t_q is the turn-off time of SCR1.

Commutating Inductor

The energy stored in the capacitor when it is charged positively has been transfered from the commutating inductor L1, and from $1/2CV^2 = 1/2LI^2$,

$$L1 = \frac{(C1)\,(QV_B)^2}{(I_m)^2}$$

Chopper Frequency Selection

Maximum

The maximum frequency is used in high torque situations. The locked rotor time constant of the motor, t_m gives us an approximate rate-of-rise and rate-of-fall of motor current, $\frac{0.63 I_{LR}}{t_m}$ A/s. (See Figure 10.) Then allowing the motor current to vary (ripple) between 120% and 80% of I_m (see Figure 11b), the chopper ON or OFF period T, may be calculated from:

$$\frac{(0.2)I_m}{T} = \frac{0.63 I_{LR}}{t_m}$$

The <u>maximum</u> chopper frequency f_{max} will then be:

$$f_{max} = 1/2T$$

Minimum

To minimize switching losses, the chopper frequency may be lowered for normal operating when average currents are lower. Also the current rise and fall rates are lower because the machine is rotating and generating a back emf. These combine to allow a minimum frequency three to five times lower than the maximum frequency and still maintain continuous motor current above a 30% duty cycle (See Figure 11).

Power Semiconductor Selection

The designer will have many trade-offs to make. The trade-off of voltage rating has already been mentioned, in regard to all semiconductors and also the commutating capacitor. The size of the commutating capacitor is also directly influenced by the rated turn-off time, t_q, of the main SCR. The faster the SCR, the smaller the capacitor.

Another area for trade-off is the current rating of the devices. Here, breakaway torque rating, the type and duration of usage, and the size of heatsinking allowed are intimately involved with the selection of the SCRs and the rectifier. Well chosen thermal feedback to the control circuit from some temperature sensitive component such as the main SCR's heatsink can also allow economies to be made.

In general, the selection of the various devices to fulfill the current requirements is not simple. It involves choosing SCRs of an adequate rms current rating, then the dissipation can be worked out using the forward voltage drop versus current plot usually available (See Figure 16). This dissipation is checked against the device ratings, and a heatsink selected. One or two iterations with various devices before building and testing may be required, and a check using currents measured in a prototype is needed. With all devices, the worst case power dissipation and current will be at locked rotor at a chopper frequency of f_{max}, assuming a shorting switch is used for maximum speed.

See Figure 16 on Next Page

SCR1

The current can be broken into two parts:

The rms current from the semisinusoid (Figure 15, interval t1 to t2) has amplitude $= QV_B/[(L2)/(C1)]^{1/2} = I_a$ Amp peak and duration $= \pi[(L2)\ (C1)]^{1/2} = T_a$ seconds and which occurs once a cycle, every 2T seconds, is

$$I_a\left[\left(\frac{T_a}{2(2T)}\right)\right]^{1/2} = I_b$$

The rms current from the ramp section of interval t2 to t5:

Start amplitude $= 0.8\ I_m = I_c$ Amp
End amplitude $= 1.2\ I_m = I_d$ Amp
Duration $= (T - T_a) = T_b$ seconds

every 2 T seconds is

$$\left[\frac{T_b}{3(2T)}\,(I_c^2 + I_d^2 + I_c I_d)\right]^{1/2} = I_e$$

<u>Total rms current for SCR1</u>

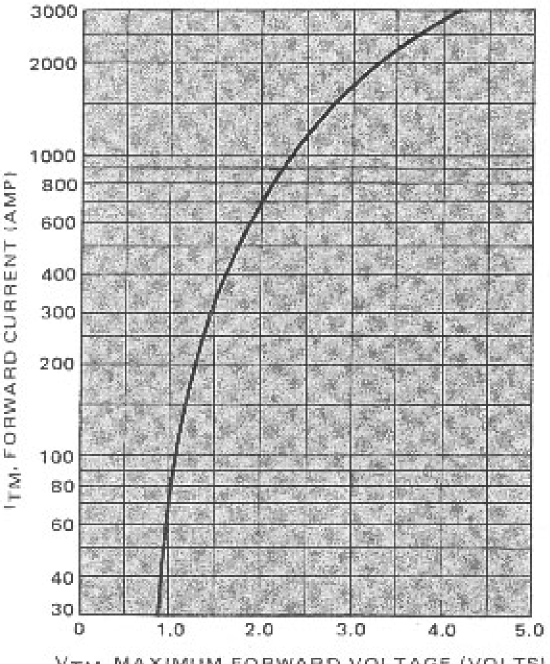


FIGURE 16 — Forward Voltage Drop versus Current (MCR380)

$$= \left[I_b^2 + I_e^2\right]^{1/2} = I_f$$

I_f is also the current in L1

$$\text{SCR1 dv/dt} = \frac{I_m}{C1} \text{ V/S}$$

$$\text{SCR1 di/dt} = \frac{I_a}{[(L2)(C1)]^{1/2}} \text{ A/s}$$

SCR 2

Again the current can be broken into two parts: the constant current section of interval t5 to t7:

amplitude = I_d for

$$\text{duration} = \frac{(C1)(KQ+1)V_B}{I_d} = T_C \text{ s}$$

occuring every 2 T seconds

$$\text{thus the rms value is } I_d \left[\frac{T_c}{2T}\right]^{1/2} = I_g \text{ Amp}$$

The inductor-to-capacitor energy transfer section (Interval t7 to t8) has an initial value of I_d and tails of to zero in $\frac{\pi}{2}[(L1)(C1)]^{1/2} = T_d$ seconds, and occurs every 2 T seconds. Its rms value is:

$$I_d \left[\frac{T_d}{3(2T)}\right]^{1/2} = I_h \text{ Amp}$$

The rms current in SCR2 is $(I_g^2 + I_h^2)^{1/2} = I_j$ Amp

SCR2 dv/dt = $\frac{(KQ-1)V_B}{[(L1)(C1)]^{1/2}}$ V/s in the gross. However there is a transient arising from D1's current cessation which must be snubbed out (see Figure 17).

SCR2 di/dt = $\frac{KQV_B}{L_s}$ A/s where L_s is the inductance of the SCR1, SCR2 and C1 circuit loop. The stray inductance, $\simeq$ 1 μH, may be increased with a saturable inductor, L3 in Figure 17. SCR2 circuit turn-off time is set by the maximum duty cycle allowed by the control design.

SCR3

The rms current through SCR3 = I_b Amp (Interval t1 to t2). SCR3 dv/dt = $QV_B(\frac{R_s}{L2})$ V/s where R_s is the snubbing resistor value in the L2, SCR3, SCR2 loop (R6 in Figure 17). This occurs when SCR2 blocks, and the positive voltage of C1 is suddenly applied across SCR3 and L2.

$$\text{SCR3 di/dt} = \frac{Ia}{[(L2)(C1)]^{1/2}}$$

SCR3 circuit turn-off time is set by the minimum duty cycle allowed by the control design.

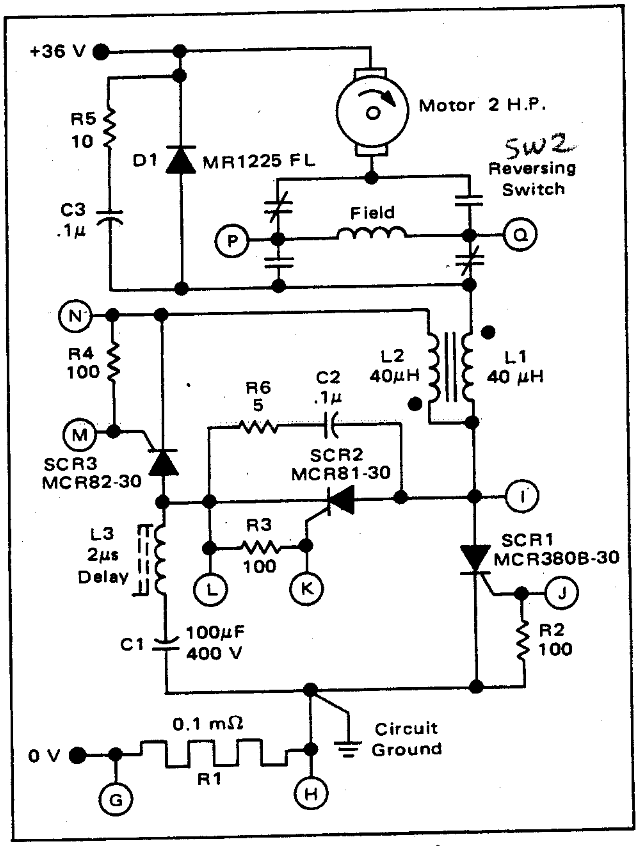


FIGURE 17 – Chopper Design

D1

The rms current in D1 = I_e Amp. A fast recovery device may be used here, to reduce reverse recovery current-induced voltage spikes across D1 itself.

C1

The rms current in C1 = $(I_b^2 + I_j^2)^{1/2} = I_k$ Amp.

L1 and L2

The rms current in L1 = $(I_e^2 + I_j^2)^{1/2} = I_1$ Amp. The rms current in L2 = I_b Amp. The peak current in the coupled inductor is either I_a or I_d.

The core must not be allowed to saturate during the interval t1 to t2. This volt-second integral is

$$\left[QV_B \times \frac{2}{\pi}\right]\frac{(\pi\sqrt{LC})}{2} \text{ Vs}$$

The core is reset during interval t5 to t8. Then using the familiar equations:

$$Es \leqslant 2B_{max}A\,N \text{ Volt-seconds and}$$

$$L = \frac{N^2A}{lg}(4\pi \times 10^{-7}) \text{ Henries.}$$

(This assumes that the air-gap is the dominant contributor to the overall inductance.)

Where, in S.I. units,

E_s = volt-seconds applied

B_{max} = Maximum allowable flux density for core, Tesla

A = area of core, square meters
N = number of turns
lg = length of air gap, meters
L = inductance, Henries

These formulas, and the winding current density, will allow rough definition of the commutating inductor.

DI/DT Reactor L3

Depending on circuit conditions and the SCRs chosen, this may not be required. However its use is advisable as it protects SCR2 from on-going di/dt failure, and also SCR1 from off-going di/dt failure.

$$\text{The delay time } t_D = \frac{N\,A(\Delta B)}{E}$$

and the magnetizing current $I_{mag} = \frac{H_s\, l_m}{N}$ which must be much less than the load current.

Where, in S.I. units,

t_D is the delay time, seconds
N is the number of turns
A is the core area, square meters
ΔB the flux swing, Tesla
E is the circuit voltage, Volts
I_{mag} is the magnetizing current, Amp
H_s is magnetizing force for saturation, ampere-turns per meter
l_m is the mean length of the core, meters

Placing this inductor in series with C1 will ensure the inductor is reset every cycle, as L3, Figure 17.

JONES CHOPPER DESIGN EXAMPLE

The motor and battery available had the following characteristics:

Battery voltage, V_B = 36 V
Maximum average motor current required, I_m = 300 A
Motor time constant, t_m = 5 ms
Locked rotor current, I_{LR} = 500 A

Component Voltage Rating

High frequency SCRs are available in voltage ratings to 600 V, as are capacitors. A standard peak-to-peak voltage rating for a commutating type capacitor is 400 V, giving a Q of $200/36 \simeq 6$. Assume a worst-case value of K = 0.5, and the most negative voltage will be $KQV_B \simeq 110$ V and the most positive will be $QV_B \simeq 215$ V.

These voltages are summed to give a capacitor voltage rating, 215 + 110 = 325 V, or 400 V, the nearest standard rating, with which we started. All power semiconductors must be rated at least 300 V, the next voltage grade above 215 V. For experimental purposes a higher rating is recommended, say 400 V.

Commutating Capacitor Value

Assume that SCRs with t_q = 15 μs are available with the required current rating. Then $T_x = 2\,t_q$ = 30 μs and

$$C1 = \frac{I_m T_x}{KQV_B} = \frac{300 \times 30}{110}\ \mu F = 80\ \mu F.$$

Commutating Inductor

$$L1 = \frac{(C1)\,(QV_B)^2}{I_m^2} = \frac{80(215)^2}{(300)^2}\ \mu H \simeq 40\ \mu H$$

Assume L2 = L1

Frequency Selection

$$T = \frac{(0.2) I_m t_m}{0.63 I_{LR}} = \frac{(0.2)\,(300)5}{(0.63)\,(500)}\text{ms} \simeq 1\text{ ms}$$

$$f_{max} = \frac{1}{2T} = 500\text{ Hz}$$

$$f_{min} = f_{max}/4 = 125\text{ Hz}$$

Power Semiconductor Selection

As mentioned previously, these calculations enable preliminary selection of SCRs. The dissipation and heat sink calculations and trade offs must still be made.

SCR1

$$I_a = QV_B/[(L2)/(C1)]^{1/2} = 6(36)/[40/80]^{1/2}\text{ A pk}$$

giving $I_a \simeq 300$ A pk

$$T_a = \pi[(L2)\,(C1)]^{1/2} = \pi[40.80]^{1/2}\ \mu s$$

giving $T_a \simeq 180$ μs

$$I_b = I_a \left(\frac{T_a}{2(2T)}\right)^{1/2} = 300 \left(\frac{180}{(2)2000}\right)^{1/2}\quad \text{A rms}$$

giving $I_b \simeq 64$ A rms

$$I_c = 0.8\,I_m = (0.8)300 = 240\text{ A}$$

$$I_d = 1.2\,I_m = (1.2)300 = 360\text{ A}$$

$$T_b = (T - T_a) = (1000 - 180) = 820\ \mu s$$

$$I_e = [\frac{T_b}{3(2T)}(I_c^2 + I_d^2 + I_c I_d)]^{1/2}$$

$$I_e = [\frac{820}{(3)\,(2)\,1000}(240^2 + 360^2 + (240)\,(360)]^{1/2}\text{ A rms}$$

giving $I_e \simeq 200$ A rms

Total rms current for SCR1 = $[I_b^2 + I_e^2]^{1/2} = I_f$

$$I_f = [64^2 + 200^2]^{1/2} = 210\text{ A rms}$$

Try MCR235B-30 or MCR380B-30, depending on dissipation calculations.

$$\text{SCR1 } dv/dt = \frac{I_m}{C1} = \frac{300}{8}\ V/\mu s \simeq 4\ V/\mu s$$

$$\text{SCR1 } di/dt = \frac{I_a}{[(L2)\,(C1)]^{1/2}} = \frac{300}{[40.80]^{1/2}} A/\mu s \simeq 5\ A/\mu s$$

SCR2

$I_d = 360$ A

$$T_c = \frac{(C1)(KQ+1)V_B}{I_d} = \frac{80(0\cdot5(6)+1)36}{360}\,\mu s$$

giving $T_c \simeq 32\ \mu s$

$$I_g = I_d\left(\frac{T_c}{2T}\right)^{1/2} = 360\left[\frac{32}{2(1000)}\right]^{1/2} = 45 \text{ A rms}$$

$$T_d = \frac{\pi}{2}[(L1)(C1)]^{1/2} = \frac{\pi}{2}[40.80]^{1/2}\,\mu s$$

giving $T_d \simeq 90\ \mu s$

$$I_h = I_d\left[\frac{T_d}{3(2T)}\right]^{1/2} = 360\left[\frac{90}{3.2.1000}\right]^{1/2}$$

$= 44$ A rms

SCR2 rms current $= I_j = (I_g^2 + I_h^2)^{1/2} \simeq 63$ A rms
Try MCR81 - 30

$$\text{SCR2 dv/dt} = \frac{(KQ-1)V_B}{[(L1)(C1)]^{1/2}}$$

$$= \frac{(0\cdot5(6)-1)36}{[40.80]^{1/2}}\ V/\mu s \simeq 1\ V/\mu s$$

but transiently may be worse.

$$\text{SCR2 di/dt} = \frac{KQV_B}{L_s} = \frac{(0\cdot5)(6)\,36}{1}\ A/\mu s$$

$\simeq 110$ A/ μs, which is rather high, so consider a saturable inductor later.

SCR3

SCR3 rms current $= I_b = 64$ A rms
Try MCR82 - 30, an isolated stud 80 A part.

$$\text{SCR3 dv/dt} = QV_B\left(\frac{R_s}{L2}\right) = \frac{6.36}{40}\cdot R_s\ V/\mu s$$

$\simeq 5\cdot5.R_s\ V/\mu s$

Let $R_s = 10\Omega$, then SCR3 dv/dt = 55 V/μs

Choose C_s, the snubbing capacitor in series with R_s to accomodate the voltage rise-time, and optimize the two by trial-and-error.

$$\text{SCR3 di/dt} = \frac{I_a}{[(L2)(C1)]^{1/2}} \simeq 5\ A/\mu s$$

D1

D1 rms current $= I_e = 200$ A rms
Try MR1225FL, or fast recovery equivalent.

C1

C1 rms current $= I_k = (I_b^2 + I_j^2)^{1/2}$
$= (64^2 + 63^2)^{1/2} = 90$ A rms

There is now enough data to make a preliminary selection of C1. (80 μF, 400 V, 90 A). Cornell Dubilier and General Electric are manufacturers of such capacitors.

L1 and L2

L1 rms current $= I_1 = (I_e^2 + I_j^2)^{1/2}$
$= (200^2 + 63^2)^{1/2} \simeq 210$ A rms
L2 rms current $= I_b = 64$ A rms
The peak current, used in some design methods, is $I_d = 360$ A.

$$\text{L2 volt-second rating} = \left[QV_B \times \frac{2}{\pi}\right]\frac{(\pi\sqrt{LC})}{2}$$

$$= \left[\frac{6(36)2}{\pi}\right]\left[\frac{\pi((40)80)^{1/2}}{2}\right]\ V\cdot\mu s = 12\cdot4\ \text{Vms}$$

Using available components;
Core - Arnold Engineering AH-33, Area 11×10^{-4} M^2
Turns - 4 turns for L1, 4 turns for L2, #4 AWG welding cable
Gap - $5\cdot5 \times 10^{-4}$ M total
B_{max} - 1.6 T

$$L = \frac{N^2A(4\pi \times 10^{-7})}{l_g} = \frac{4^2(11\times10^{-4})(4\pi\times10^{-7})10^6}{5\cdot5\times10^{-4}}\ \mu H$$

$= 40\ \mu H$

Volt-seconds available $= 2B_{max}AN =$
$2(1\cdot6)(11 \times 10^{-4})\,4$ Vs
$= 14$ Vms

The No. 4 AWG welding cable used was not optimum for current density. A foil wound design would be preferable as this would give much better packing factors.

DI/DT Reactor (L3)

For a Ferroxcube No. 400T750-3C5
Area $= 1\cdot81 \times 10^{-4}$ M^2
$\Delta B = 2\,B_{max} = 0.7$ T
$l_m = 0\cdot127$ M
$H_s = 79.6$ AT/M
Circuit Voltage $= KQV_B = (0\cdot5)(6)(36) = 110$ V

$$\text{For 4 turns, } t_D = \frac{NA(\Delta B)}{E} = \frac{4(1\cdot81\times10^{-4})(0.7)10^6}{110}\ \mu s$$

$= 4.6\ \mu s$. (A delay of 2 μs was observed).

$$I_S = \frac{H_s l_m}{N} = \frac{(79\cdot6)(0\cdot127)}{4}\ A = 2.5\ A,$$ which is much less than circuit current.

Final Power Circuit

Figure 17 shows the final circuit used. It was found necessary to encapuslate the commutating inductor in a silicon rubber (Dow Corning No. 184) to cut down on lamination noise. This was very successful.

SCR1 and SCR2 were mounted on the same heatsink. The use of an isolated stud part for SCR3 allowed it to be mounted on that heatsink also. L1, L2 and L3 were designed as described above.

The control shunt was made of a strip of aluminum.

The three 100 Ω resistors from gate-to-cathode on each SCR (R2, R3, R4) are to improve noise immunity and dv/dt performance.

Points G through N, P and Q are connection points to the control circuitry.

CHOPPER CONTROL CIRCUIT

There are many ways of building a control circuit; designs may be based on programmable unijunction transistors (2N6027), timers (MC1455), or monostables. This design is based on the MC14528 complementary MOS Dual Monostable Multivibrator.

As mentioned in the Section on Chopper Frequency, a variable frequency approach is required; a relatively low frequency to minimize drive losses, increasing in frequency, and at the same time changing duty cycle. This limits the peak current, but at the same time permits high average motor currents for high torque situations. Figure 18 shows the basic schematic. Monostables M1 and M2 are cross-coupled so that when one has finished its cycle, its $\overline{Q}$ output triggers the other. The relative times of each are cross-coupled also via potentiometer R9, which is connected so that increasing the period of one decreases the period of the other. Point (A) from M1 leads to interface circuitry (described later) which fires SCR1 and SCR3, thus M1 is shown as the "ON" monostable, and these SCRs are fired when point (A) goes high. Similarly Point (B) fires SCR2. Point (E) leads to circuitry which ensures that SCR2 is fired first when power is connected, precharging the commutation capacitor.

Points (C) and (D) are connected to a voltage source whose value is proportional to the motor current. As the motor current increases so does this voltage, and more current is supplied to the timing capacitors C4 and C5 thus shortening the monostable periods. Diode D2 causes the "OFF" period to be less affected by this voltage, tending to limit motor current, that is, the "ON" period to be reduced more than the "OFF" period.

Current Sensing

To simplify ground and sensing connections, the current sensing shunt R1 (Figure 17) is placed in the battery return line. This current is the battery current which is a square wave. Figure 19 shows the amplifier used to convert this current signal to a voltage signal.

Operational amplifier OA1 is configured as a peak detecting high gain stage. R11-R14 place OA1 in its active region and allow it to sense the battery current signal at circuit ground. C6 is required to snub the spikes appearing across the shunt due to its inherent self-inductance. As the motor, and therefore the battery, current increase, Point (G) and the inverting input of OA1 go more negative. This is peak detected and the voltage

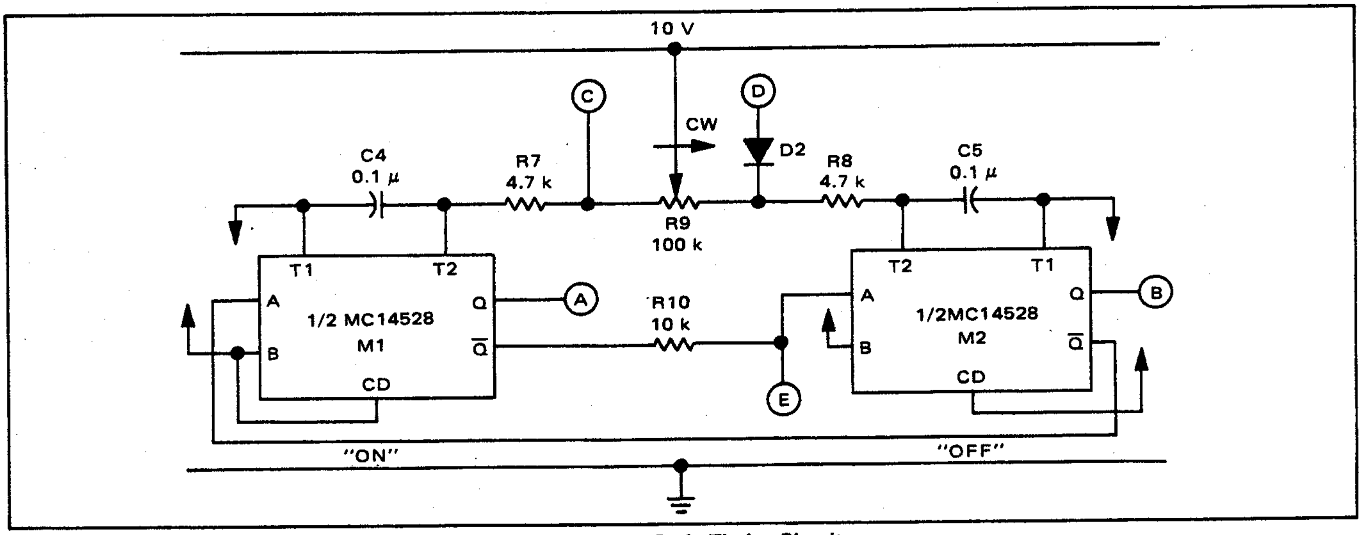


FIGURE 18 – Basic Timing Circuit

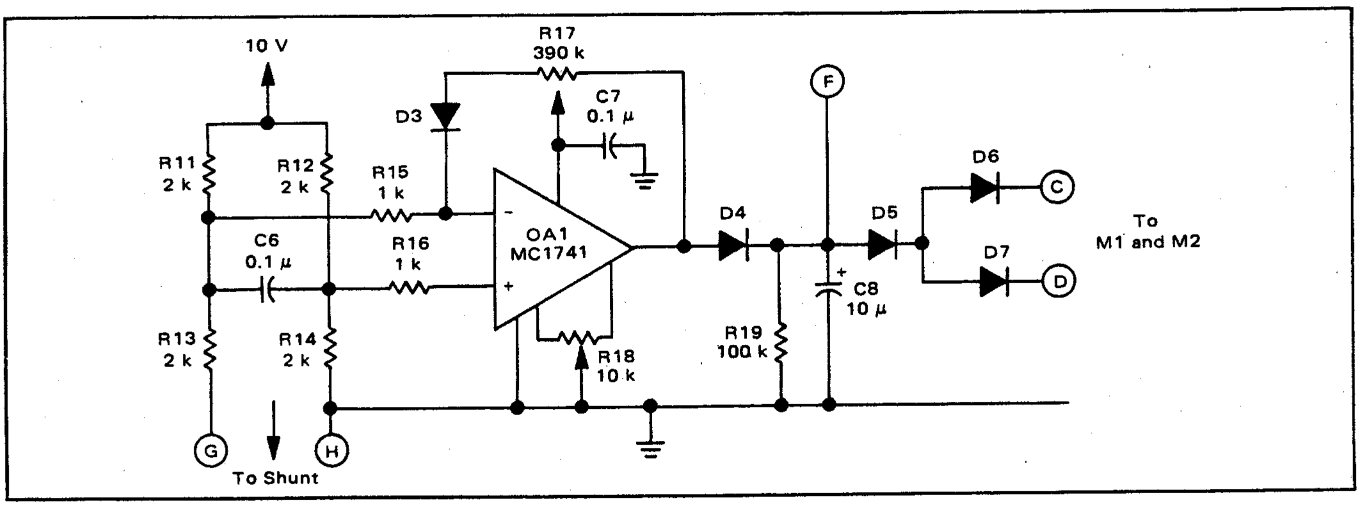


FIGURE 19 – Current to Voltage Converter

on C8 increases, increasing the current feed to M1 and M2. Point F is connected to the start circuitry, and is pulled high at turn-on, causing the timing circuits to run with short cycle times.

Gate Firing Circuitry

SCR1 can be gated directly as its cathode is common with the control circuit. SCR2 and SCR3 must be gated through some isolating device, in this case pulse transformers. Figure 20 shows the circuits used, with connections from the timing monostables and to the SCRs.

The input square waves from the monostables, at points A and B are differentiated and all the SCRs are gated with an approximately 70 μs pulse.

Point O is pulled to ground at start-up to ensure that SCR2 is fired first. Point I is tied to the anode of SCR1 and causes the gate-drive to that SCR to become disabled when its anode is reverse biased (interval t2 to t3, Figure 15). Similarly the gate pulses to SCR2 and SCR3 must be shorter than the anode current pulses to avoid gating them when the anodes are reverse biased.

Start/Stop Circuitry

To avoid relying on the transformer action of L1 and L2 to charge the commutating capacitor C1, a difficult operation to quantify, it is necessary to precharge C1 positively by firing SCR2 first. Then when SCR1 and SCR3 are fired, this voltage is reversed and is available for commutation of SCR1.

C1 can only be charged to V_B in this precharge operation which limits the current that can be commutated in the ensuing cycle. This does not present any problem if the control is set to low speed, but could create misfiring if the battery is reconnected when the control is set to a high speed. The start circuitry shown in Figure 21 is one solution.

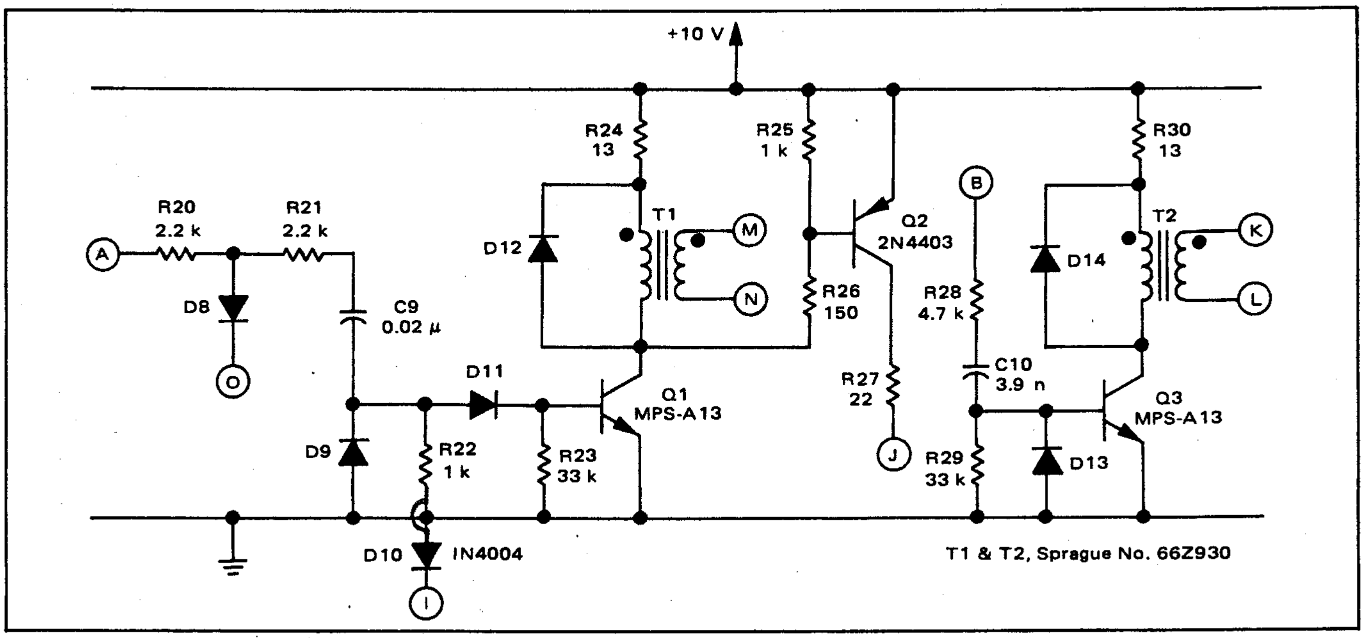
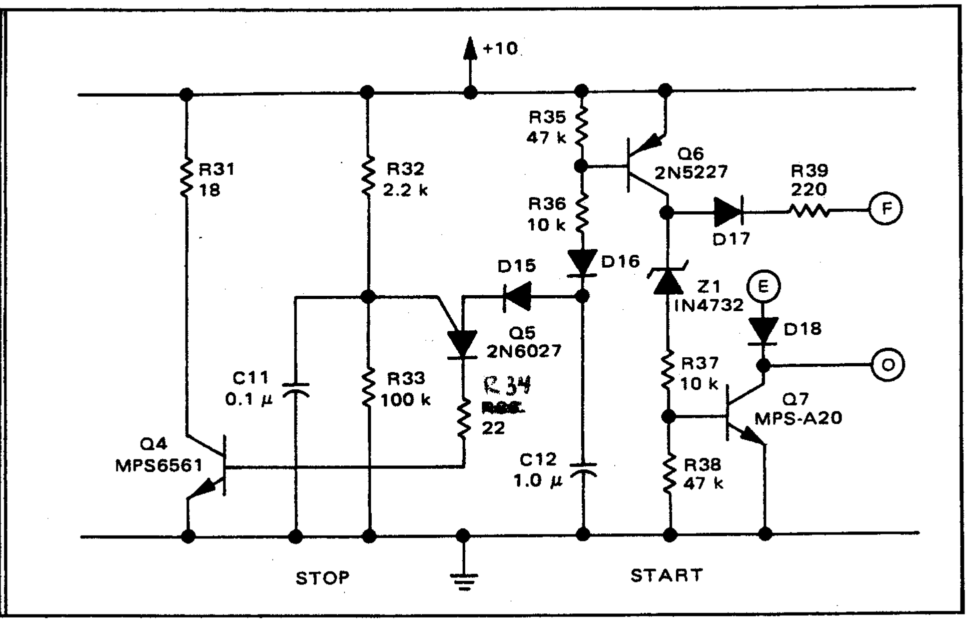


FIGURE 20 – Gate Firing Circuitry

FIGURE 21 – Start/Stop Circuit



On connection of the supply, C12 charges through the base of Q6, whose collector charges C8, Figure 19 via Point (F). Q6 being ON also turns on Q7, pulling down points (O) and (E). As Point (O) is low, the gate drives to SCR1 and SCR3 are disabled. Point (E) stays low, then as C12 charges, goes high, triggering M2 which in turn gates SCR2.

When power is removed from the circuit, the 10 V supply starts to drop, firing programmable unijunction transistor Q5 which discharges C12, resetting it for the next power reconnect. Q4 is also turned on at this time, pulling down the 10 V supply rapidly, reducing the possibility of momentary power disconnects causing uncertainty.

Power Supplies and Reversing Interlock

The gating circuits require fairly high currents which occur in pulses, causing noise on the power supply. Two separate zenered power supplies are used here, as shown in Figure 22, one to feed the gate circuits, the other to feed the low level supplies.

Also shown in Figure 22 is the reversing switch interlock. The approach taken here is to disable and restart the entire control circuit when the reversing switch is operated.

Figure 23 shows the complete schematic and interconnections.

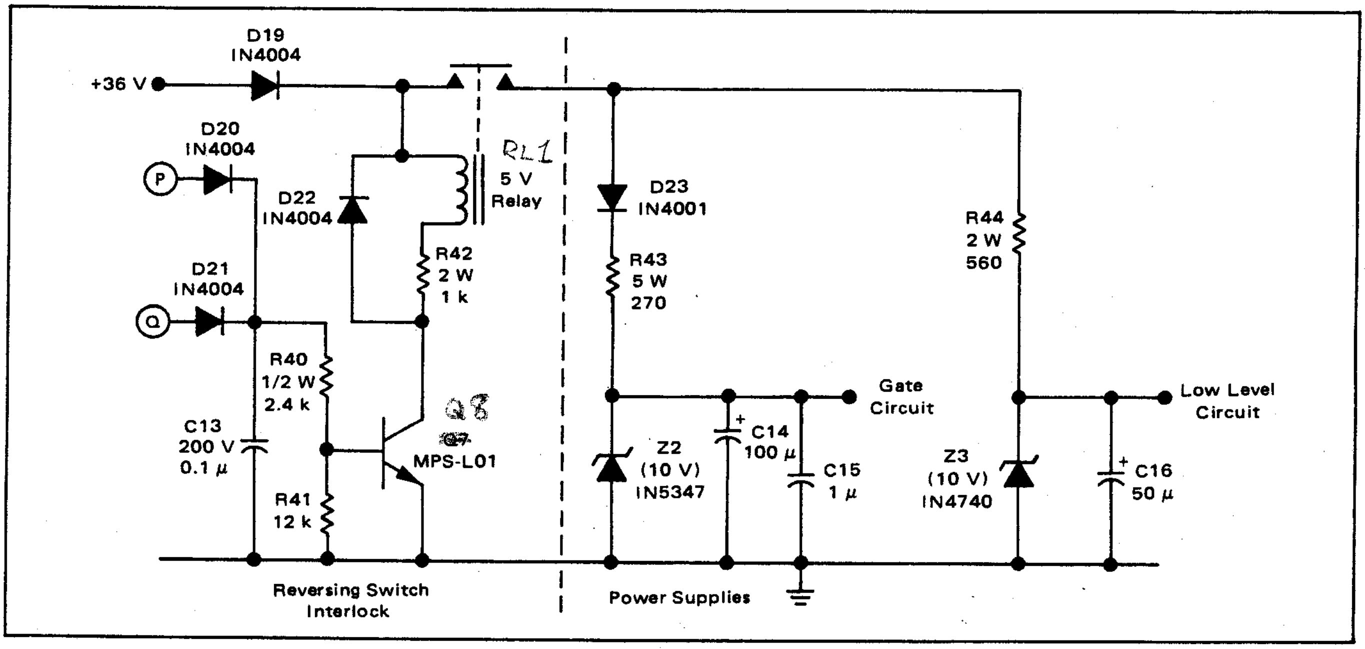


FIGURE 22 – Power Supplies and Reversing Switch Interlock

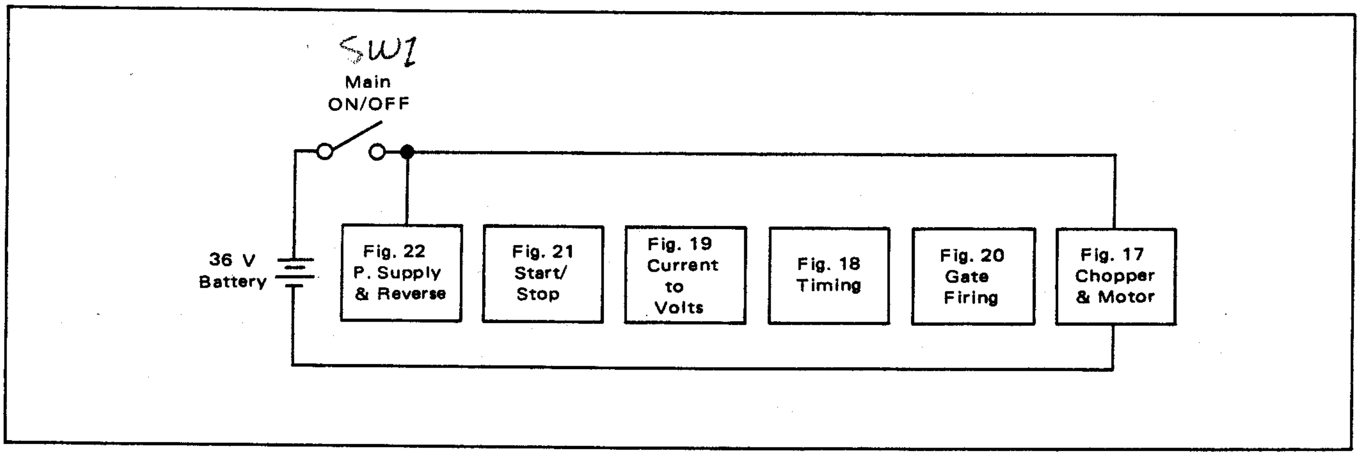


FIGURE 23 – Complete Motor Control Schematic