A HIGH POWER DC MOTOR CONTROLLER
FOR AN ELECTRIC RACE CAR
USING POWER MOSFETS

A Thesis Presented to
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Brian A. Welchko
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1 Introduction

Throughout the ages, man has been driven to control modes of transportation to meet his needs. From our early ancestors who invented the wheel improving foot travel, to the locomotive engine speeding cross country, to the jumbo jet flying around the world, inventing faster forms of transportation has improved the quality of life. Man continually strives to control the speed of his transportation vehicle, to arrive faster, safer, and more efficiently than he has done before. Electric powered vehicles are one form of transportation continuously being changed and improved.

As the twenty-first century approaches, electric powered vehicles are beginning to gain a foothold in the transportation market. The largest potential growth area for this industry is the personal automobile. The electric car has been around since the invention of the automobile, and with modern technology, it is able to compete with the gasoline powered internal combustion engine in many ways. The best application for current electric cars is for commuters who live and work near cities. This is where air quality is low and much time is spent in heavy traffic. Electric vehicles do not use energy when stopped at a traffic light and are zero emission vehicles when generation of electricity is
not considered. Even if power generation is considered, it is much easier to control pollution at one power plant than on thousands of individual automobiles.

This thesis is an effort to control the speed of a specialized kind of electric vehicle - an electric race car. A commuter driving an electric car to work each day may seem unrelated to a race car, but many important features that consumers now expect in their own cars were first developed on the race track as an effort to be a little faster, a little more efficient, and a little safer than the competition.

1.1 An Overview of the Electric Bobcat

The Electric Bobcat is Ohio University's electric powered Formula Lightning class race car. There are about twelve universities that have Formula Lightning class electric race cars. Each year, there are several races for this class of cars organized by Electric Vehicles Technology Competitions, Inc. (EVTC). This is a stock series where all the schools have identical chassis and tires. It is the goal of the series to test different types of electric propulsion systems, hence there are tight rules on non-powertrain parts. The Electric Bobcat is shown in Figures 1.1 and 1.2 as it appeared at the Cleveland Electric Formula Classic in June, 1996. The car is powered by sealed lead acid 12 volt Optima batteries and a series wound motor manufactured by Advanced DC, Inc. It has a Webster 4-speed gearbox. The car chassis was manufactured by Stewart Racing. It has a racing weight of 2000 to 2500 pounds depending on the battery configuration being used.
1.2 Motivation for the Controller Design

At the beginning of the Electric Bobcat project at Ohio University in 1994, it quickly became apparent that the limiting factor on vehicle speed and acceleration was the motor controller. The controller in use at the time had a maximum rating of 400 amps and 120 volts. The 400 amp limit allowed for a mildly competitive top speed, but this current limit was too low for a vehicle capable of winning the races. It was determined that a controller capable of delivering 600 to 800 Amps to the motor would allow for a desired top speed in excess of 100 mph and much improved acceleration characteristics. It was also desirable to design and build a controller that was more than a black box piece of equipment. This way, the racing team could determine precisely which part of the power train contains the source of a problem without blindly placing
blame on the controller since it was a sealed, unknown unit.

1.3 Goals

The goal of this thesis is to design, build, and race a MOSFET based motor speed controller for the Electric Bobcat. It is a goal that the controller is able to be operated at a nominal 144 volt battery input voltage and have a continuous current rating of 600 amps. It should be able to intermittently handle 800 amps during periods of rapid acceleration. On the control electronics side of the controller project, the controller will have pulse width modulation (PWM) control and a changeable preset current limit. It is also a goal to have a full range from 0 to 100 percent duty cycle of the PWM unit with an adjustable delay ramp to protect the MOSFET switches from a large instantaneous increase in throttle position. Weight of the controller should be as little as possible. Air cooling must also be used for the sake of simplicity. It is also desirable to have an additional current limit option that would limit the current draw from the batteries to an adjustable limit centering around 400 amps. Lastly, the controller must be electrically isolated from the chassis of the car and have an efficiency in excess of 90 percent.

1.4 Organization of Thesis

This thesis is organized into seven chapters and two appendices. Chapter 1 is a background discussion on the Electric Bobcat and controller performance criteria. Chapter 2 covers the theory of operation of DC-DC step down chopper drives. Chapter 3 provides background and technical information on the power MOSFETs and flyback
diodes used in this controller. Chapter 4 provides theory on losses in buck chopper drives and heatsink theory and calculations. In Chapter 5, the design, design methods, and construction of the controller are presented. Chapter 6 covers the experimental test results of the controller designed in the previous chapter. Chapter 7 contains conclusions resulting from the work of this thesis, shortcomings of the design, and recommendations for improvements to the controller design. Appendix A contains data sheets for the major components of the controller and Appendix B contains photographs taken throughout the controller project.
2 DC-DC Step Down Chopper Drives

Step down choppers are used to apply a variable output dc voltage to a load from a fixed voltage dc supply. In this capacity they act as step down dc transformers with continuously variable turns ratios. This is accomplished using pulse width modulation or frequency modulation techniques on a semiconductor switch to toggle the output voltage seen by the load between the input voltage and zero. For additional information on choppers, see the references by Shepherd, Hulley, and Liang (1995) and Rashid (1993). Much of the information in this chapter is derived from the reference material.

2.1 Pulse Width Modulation

Pulse Width Modulation (PWM) is controlling the duty cycle of the output waveform of a chopper by controlling the on-time, $t_{on}$. As seen in Figure 2.2a, the duty cycle $D$ is defined by

$$D = \frac{t_{on}}{T}$$  \hspace{1cm} (2.1)

where the period $T$ is defined as
\[ T = t_{on} + t_{off}. \] (2.2)

On the other hand, frequency modulation, as defined for power electronics, would involve the use of a fixed \( t_{on} \) or \( t_{off} \) while varying the period \( T \). PWM techniques were used to control the chopper designed in this thesis.

2.2 Step Down Chopper with a Resistive Load

Figure 2.1 shows a step down chopper with a resistive load. The corresponding output voltage and current waveforms are shown in Figure 2.2. The resistive load case is the least complicated to analyze, but not very applicable to a motor load. However, it presents a simple derivation of some important features of the step down chopper. The chopper shown in Figure 2.1 is commonly referred to as a class A, one quadrant, or 'buck' chopper. It is a class A, or one quadrant chopper, because the load voltage and current are both unidirectional and positive. This type of converter has also become known as a buck chopper because the output voltage cannot be greater than the input voltage. To simplify the derivation of formulas which characterize the operation of a buck chopper, it has been assumed that the chopper is lossless.

![Figure 2.1: Buck Chopper with Resistive Load.](image-url)
In order to simplify the computation of equations which describe the behavior of buck choppers, it is helpful to define the resulting waveforms in terms of the periodic time $\omega$. Thus, a complete cycle of $t_{on} + t_{off}$ is equivalent to $2\pi$ radians. It is then possible to define the periodic time and its inverse, the chopping frequency. Restating the above gives

$$\omega (t_{on} + t_{off}) = 2\pi,$$

$$\text{periodic time} = \frac{2\pi}{\omega} = T,$$

$$\text{chopping frequency} = f_c = \frac{2\pi}{\omega}.$$
The on-period and on-time in seconds can then be given as

\[ T_{on} = D(t_{on} + t_{off}) = 2\pi D, \quad (2.6) \]

\[ \text{switch on-time} = \frac{2\pi D}{\omega} = DT. \quad (2.7) \]

With the independent variable of \( \omega t \) defined in Equations (2.3) through (2.7), the average output voltage of the buck chopper is given by

\[ V_{L_{avg}} = \frac{1}{2\pi} \int_0^{2\pi} v_L(\omega t) d\omega t = \frac{1}{2\pi} \int_0^{2\pi D} V_s d\omega t \quad (2.8) \]

\[ V_{L_{avg}} = DV_s, \quad (2.9) \]

where \( D \) is the duty cycle. Likewise, the average load current can be found, and is given by

\[ I_{L_{avg}} = \frac{1}{2\pi} \int_0^{2\pi} i_L(\omega t) d\omega t = \frac{1}{2\pi} \int_0^{2\pi D} \frac{V_s}{R} d\omega t \quad (2.10) \]

\[ I_{L_{avg}} = D \frac{V_s}{R} \quad (2.11) \]

The rms (root-mean-square) output voltage is defined in the usual way by

\[ V_{L_{rms}} = \sqrt{\frac{1}{2\pi} \int_0^{2\pi} v_L^2(\omega t) d\omega t}. \quad (2.12) \]
Substituting, the rms output voltage of a buck chopper is

$$V_{L_{ma}} = \sqrt{\frac{1}{2\pi} \int_0^{2\pi D} V_s^2 \, d\omega t}$$

(2.13)

$$V_{L_{ma}} = \sqrt{D} V_s.$$  

(2.14)

The power input to the chopper, $P$, and hence the output power since the chopper is lossless, can be calculated as

$$P = \frac{1}{2\pi} \int_0^{2\pi} v_L(\omega t) i_L(\omega t) d\omega t = \frac{1}{2\pi} \int_0^{2\pi D} \frac{V_s^2}{R} d\omega t$$

(2.15)

$$P = D \frac{V_s^2}{R}.$$  

(2.16)

Equation (2.16) shows that controlling the power to a load connected to a buck chopper is as easy as controlling the duty cycle, $D$, of the chopper. This is to be expected because all other parameters involved are fixed. Also note that the power, $P$, varies linearly with the duty cycle.
2.3 Step Down Chopper with a Series RL Load

A buck chopper with a series RL load is shown in Figure 2.3. The voltage source, \( E \), in the load represents the back emf that is present in an inductive motor load. This circuit contains a diode which is absent in the chopper configured for a resistive load. The diode is necessary due to the reactive load. When the switch is opened, the current in the load does not instantaneously drop to zero as it does in the resistive load case. The diode serves to create a continuous current path and is commonly known as a flyback diode. As can be seen in Figure 2.4a, assuming a continuous load current, the output voltage waveform is the same regardless of the load.

![Diagram of Buck Chopper with RL Load](image)

**Figure 2.3:** Buck Chopper with RL Load.

For an RL load, circuit operation can be broken down into two modes. Mode 1 occurs when the switch is closed and the flyback diode behaves as an open circuit. Mode 2 occurs when the switch is opened and the load current flows through the flyback diode. Two operating conditions are then possible; those being continuous current operation and discontinuous current operation. In discontinuous current operation, the load current
decays to zero before the switch is closed at the start of the next cycle. This operating condition is undesirable for an electric vehicle as the power output of the electric motor is intermittent. To avoid this problem, the frequency used for the PWM control of the drive should be increased. Due to the undesirability of discontinuous current operation, it is not considered in the following equations which characterize a buck chopper with a series RL load.

Figure 2.4: Waveforms for an RL Motor Load.

As in the equations developed for a resistive load, it is assumed that the chopper and flyback diode are lossless. Also for this case, the inductor is assumed to be lossless.
For mode 1, $0 \leq \omega t \leq 2\pi D$, and the corresponding loop equation is

$$V_s = L \frac{d i_L(\omega t)}{d\omega t} + R i_L(\omega t) + E.$$  \hfill (2.17)

Steady state is a condition that is reached when the waveforms are the same from cycle to cycle. This occurs several periods after the start of the first cycle. Assuming that the system is in steady state and $i_L(\omega t=0) = i_{min}$, Equation (2.17) can be solved to give

$$i_L(\omega t) = i_{min} e^{-\omega t/\tau} + \frac{V_s - E}{R} \left(1 - e^{-\omega t/\tau}\right),$$ \hfill (2.18)

where $\omega \tau = \omega L/R$ and $\tau$ is the time constant of the load. When $\omega t = 2\pi D$, the load current has reached its maximum value as seen in Figure 2.4b. Thus, $I_{max} = i_L(\omega t = 2\pi D)$ and can be found from Equation (2.18) as

$$I_{max} = i_{min} e^{-2\pi D/\tau} + \frac{V_s - E}{R} \left(1 - e^{-2\pi D/\tau}\right).$$ \hfill (2.19)

When the switch opens and the diode begins conducting, the circuit is operating in mode 2. The loop equation that describes this mode is

$$0 = R i_L(\omega t) + L \frac{d i_L(\omega t)}{d\omega t} + E.$$ \hfill (2.20)

This equation is valid for $2\pi D \leq \omega t \leq 2\pi$. Knowing that $i_L(\omega t=2\pi D) = I_{max}$ from Equation (2.19), Equation (2.20) can be solved as follows;

$$i_L(\omega t) = I_{max} e^{-\omega t - 2\pi D/\tau} \left(1 - e^{-(\omega t - 2\pi D)/\tau}\right).$$ \hfill (2.21)

Under the steady state conditions assumed, the current at the end of this mode has
decayed to \( I_{\text{min}} \). With this information, Equation (2.21) can be written as

\[
I_{\text{min}} = I_{\text{max}} e^{-2\pi (1-D)/\omega \tau} \frac{E}{R} (1 - e^{-2\pi (1-D)/\omega \tau}).
\]

Equations (2.19) and (2.22) can be solved simultaneously to yield

\[
I_{\text{max}} = \frac{V_s}{R} \left( \frac{1 - e^{-2\pi D/\omega \tau}}{1 - e^{-2\pi/\omega \tau}} \right) - \frac{E}{R}.
\]

\[
I_{\text{min}} = \frac{V_s}{R} \left( \frac{e^{2\pi D/\omega \tau} - 1}{e^{2\pi/\omega \tau} - 1} \right) - \frac{E}{R}.
\]

The average load current for the inductive load case is found by substituting Equations (2.18) and (2.21) over their respective time intervals into the basic equation

\[
I_{\text{avg}} = \frac{1}{2\pi} \int_{0}^{2\pi} i_L(\omega t) d\omega t.
\]

This long calculation can be avoided if it is considered that the average voltage across the inductor must be equal to zero over one cycle because the inductor cannot create energy. Since there is no net voltage drop across the inductor, the average current can be found using the average voltage drop across the resistance of the load. Therefore, the average load current is given by

\[
I_{\text{avg}} = \frac{V_s - E}{R}.
\]

The rms load current is calculated by substituting Equations (2.18) and (2.21) over their respective time intervals into the equation
\[ I_{L_{\text{rms}}} = \sqrt{\frac{1}{2\pi} \int_0^{2\pi} I_L^2(\omega t) d\omega t}. \] (2.26)

Due to the mathematical nature of the instantaneous current equations, calculation of the rms load current can be tedious. In practice, an approximation to this current is normally used. This approximation can be found using the first few terms of the Fourier series

\[ I_{L_{\text{rms}}} = \sqrt{I_{L_{\text{avg}}}^2 + I_{L_1}^2 + I_{L_2}^2 + \ldots}, \] (2.27)

where the harmonic currents are given by

\[ I_{L_n} = \frac{V_{L_n}}{Z_{L_n}}. \] (2.28)

\[ I_{L_n} = \frac{2V_s \sin \pi D}{\sqrt{2\pi} \sqrt{R^2 + (n\omega L)^2}}. \]

The energy delivered to the load can be found by integrating the product of load voltage and load current over a cycle. This average load power is then given by

\[ P_L = I_{L_{\text{rms}}}^2 R + EI_{L_{\text{avg}}}. \] (2.29)

The term \( E_v I_{L_{\text{avg}}} \) represents the power transferred to the motor and includes any friction and windage losses. Iron losses, including eddy current losses and those associated with magnetic properties of the motor, are not considered. The \( f_{L_{\text{rms}}} R \) term gives the copper or winding losses of the motor.

The average current in the flyback diode is found by integrating the average value of Equation (2.21) over a cycle. As seen in Figure 2.4d, the diode current is just the
load current when the switch is off. Substituting this and the value for \( I_{\text{max}} \) given in equation (2.23) into the required integral gives

\[
I_{\text{avg}} = \frac{1}{2\pi} \int_{2\pi}^{2\pi D} \frac{E}{R} \left[ 1 - e^{-(\omega t - 2\pi D)/\omega \tau} \right] \left[ \frac{V_s}{R \left( 1 - e^{-2\pi D/\omega \tau} \right)} \right] \frac{E}{R} e^{-(\omega t - 2\pi D)/\omega \tau} d\omega t
\]  

(2.30)

\[
I_{\text{avg}} = \frac{-E(1-D)}{R} + \frac{V_s \omega \tau}{2\pi R} \left( 1 - e^{-2\pi(1-D)/\omega \tau} \right) \left( \frac{1 - e^{-2\pi D/\omega \tau}}{1 - e^{-2\pi/\omega \tau}} \right).
\]  

(2.31)

On the input side of the chopper, the supply current can be found by substituting Equations (2.18) and (2.23) into the following integral

\[
I_{\text{avg}} = \frac{1}{2\pi} \int_{0}^{2\pi D} i_L(\omega t) d\omega t.
\]  

(2.32)

As shown in Figure 2.4c, the supply current is the same as the load current when the switch is on, and is zero when the switch is opened. Upon substitution, the integral becomes

\[
I_{\text{avg}} = \frac{1}{2\pi} \int_{0}^{2\pi D} \frac{V_s - E}{R} \left( 1 - e^{-\omega t/\omega \tau} \right) + \frac{V_s}{R \left( e^{2\pi/\omega \tau} - 1 \right)} \frac{E}{R} e^{-\omega t/\omega \tau} d\omega t
\]  

(2.33)

\[
I_{\text{avg}} = D \frac{V_s - E}{2\pi R} \left( e^{2\pi/\omega \tau} - e^{2\pi D/\omega \tau} \right) \left( \frac{1 - e^{-2\pi D/\omega \tau}}{e^{2\pi/\omega \tau} - 1} \right).
\]  

(2.34)

The load current waveform shown in Figure 2.4b has a large peak-to-peak ripple associated with it. This was deliberately shown to make the inductive ramping of the current waveform obvious. In practice, the chopping frequency would be increased to
keep the ripple down to a small percentage of the peak current, \( I_{\text{max}} \). As a result, the waveform would essentially be a sawtooth waveform and the inductive ramping would be nearly linear. Also, since the semiconductor switches are not ideal, the small percent change in the current calculations due to inductive ramping can be discarded without loss of significant accuracy. So by assuming that the current rises and falls linearly from \( I_{\text{min}} \) to \( I_{\text{max}} \), the simplified equations that characterize the chopper currents are

\[
I_{S_{\text{avg}}} = D \frac{V_s - E}{R} \tag{2.35}
\]

\[
I_{D_{\text{avg}}} = (1 - D) \frac{V_s - E}{R} \tag{2.36}
\]

\[
I_{L_{\text{rms}}} = \frac{1}{3} \left[ I_{\text{max}}^2 + I_{\text{max}}^2 + I_{\text{min}}^2 \right] \tag{2.37}
\]

Equations (2.35), (2.36), and (2.37), along with Equation (2.25) which gives the average load current, and Equation (2.9) which gives the average load voltage, are the most important equations derived in this chapter. This is because nearly all of the necessary and useful information about practical choppers and their applications are found with them.
3 MOSFET and Diode Background Information

The MOSFET switches used in the chopper to control the voltage to the motor of the Electric Bobcat were Motorola part number MTY55N20E. The data sheets for this part are given in Appendix A.1. This part is from the TMOS Power MOSFET line, which is the latest step in the evolutionary process that began with the small-signal MOSFET. This is a family of vertical, metal-oxide-semiconductor power field-effect transistors with matrix diffused channels. They offer a wide range of features and applications (Motorola, 1994, p. 2-1-6). The inherent advantages of Motorola's power MOSFETs include:

* Nearly infinite static input impedance featuring:
  - Voltage driven input
  - Low input power
  - Few driver circuit components

* Very fast switching times
  - No minority carriers
  - Minimal turn-off delay times
  - Large reversed biased safe operating area
  - High gain bandwidth product

* Positive temperature coefficient of on-resistance
  - Large forward biased safe operating area
  - Ease in paralleling
* Almost constant transconductance
* High $dv/dt$ immunity
* Low cost

(Motorola, 1994, p. 2-1-6).

MOSFETs were selected as the semiconductor switch of the chopper over other semiconductor devices for several reasons. Since MOSFETs require only a voltage driven input to be turned on, they will dissipate a small amount of power in the gate drive circuit which makes for a simplified gate driver circuit. They also feature the fastest switching times of large power semiconductor devices (Rashid, 1993, p. 10). This will allow for faster switching and thus reduced switching losses which contribute significantly to the power losses of the chopper. Since power MOSFETs have a positive temperature coefficient, they are easy to parallel because this temperature coefficient nearly eliminates the possibility of thermal runaway. A positive temperature coefficient means that as the device temperature increases, the on-state resistance of the device also increases. This allows for more effective current sharing between paralleled devices. As a result of paralleling many devices, the on-state resistance of the chopper is almost negligible. This, in turn, reduces conduction losses to a small amount of the total chopper power loss. They also have a high $dv/dt$ immunity which allows them to withstand the current spikes and noise associated with an inductive load, and the harsh operating environment of an electric automobile. Lastly, power MOSFETs have a low cost which could make this chopper attractive to commercial production. Overall, at the time of origin of this project, power MOSFETs proved to be the most attractive semiconductor switch readily
available. Insulated gate bipolar transistors (IGBTs) were also considered, but were
decided against because of cost, a negative temperature coefficient, and a larger on-state
resistance than paralleled power MOSFETs.

To get the most performance out of the power MOSFETs available, it is necessary
to understand their operating characteristics and limits very well. Switching
characteristics are covered in Section 3.1 of this chapter as this is an extremely important
subject when switching multiple devices in parallel and reducing switching losses. As a
result of the switching configuration used, the MOSFETs will be subjected to transients
which can damage or destroy the device. This creates the need for protection circuitry
in the form of snubbers and flyback diodes. These are discussed in Sections 3.2 and 3.3
respectively.

3.1 Gate Charge and Gate, Drain, and Source Capacitances

Traditionally it has been assumed that to switch MOSFETs, it was necessary to
charge and discharge a capacitive reactance network associated with the gate of the
MOSFET and the drive circuit. These main capacitances are $C_{gs}$ and $C_{dg}$ which are the
gate-to-source and drain-to-gate capacitances respectively. The common-source input
capacitance, $C_{iss}$, is the sum of $C_{gs}$ and $C_{dg}$ and is the dominating factor during the early
stages of switch-on and the latter stages of switch-off. The common-source reverse
transfer capacitance, $C_{rss}$, is a dominating factor to the input impedance of the device
during the latter stages of switch-on and the first stages of switch-off. $C_{rs}$ is a result of
the "Miller Effect" and is also known as the "Miller Capacitance" (Motorola, 1994, p.
2-6-1). A shortcoming to using capacitance data to determine switching times and gate drive requirements is that many of the capacitances involved are a function of the applied voltage. Therefore, a better and easier way to calculate switching times and gate drive requirements is with the use of gate charge data (Motorola, 1994, p. 2-6-6).

Gate charge data contains a wealth of information on the switching characteristics of power MOSFETs. As the name implies, these curves indicate the amount of charge that must be supplied to the gate to affect different stages of turn-on. Understanding the gate charge test circuit helps in the interpretation of the gate charge waveforms (Motorola, 1994, p. 2-6-3). A circuit to determine the gate charge waveform is given in Figure 3.1.

![Gate Charge Test Circuit](image)

**Figure 3.1:** Gate Charge Test Circuit (International Rectifier [IR], 1993, p. 1551).

The gate charge test circuit employs a constant current source to charge up the MOSFET's input capacitance. This insures that $C_{iss}$ is charged at a fixed rate. As a result of the
constant charge profile, the resulting waveform is a representation of the gate-to-source voltage, $V_{gs}$, versus gate charge, as well as $V_{gs}$ versus time. Another constant current source is used as the load. This helps to sharpen the inflection points on the waveform (Motorola, 1994, p. 2-6-4). When the switch opens in Figure 3.1, the gate of the MOSFET charges up and the MOSFET begins to conduct. The resulting waveforms are shown in Figures 3.2a and 3.2b.

![Gate-to-Source Voltage versus Gate Charge](a)

![Vgs and Id versus Gate Charge](b)

**Figure 3.2:** Basic Gate Charge Waveforms (IR, 1993, p. 1551).

Figure 3.2a shows the gate voltage, $V_{gs}$, versus time and gate charge. The waveforms shown in Figures 3.2a and 3.2b are idealized. In reality, the lines shown in the figures are not perfectly straight due to unmodeled effects. In particular, the slope of the gate charge curve in the region of Q2 cannot be horizontal as that would imply charging up an infinite capacitance. However, this idealization provides for a clear understanding and a fairly accurate portrayal of gate charge data. When the switch in the
gate drive path is initially closed at t0, $V_{gs}$ begins to rise. After a charge of Q1 is supplied to the gate, $V_{gs}$ holds at the plateau voltage, $V_{gpp}$. After supplying an additional Q2 worth of charge, $V_{gs}$ begins to rise again to its final value attained at t4. Q1 is the amount of charge that is stored between the gate and source terminals of the MOSFET, hence it is also known as the gate-to-source charge, $Q_{gs}$. Q2 is the gate-to-drain charge and is also called $Q_{gd}$. $Q_{gd}$ is associated with the charging of the Miller capacitance. The quantities $Q_{gs}$ and $Q_{gd}$ are listed as Q1 and Q2 in the figure because this is the way they are listed on all power MOSFET data sheets by Motorola. The corresponding drain-to-source voltage and drain current waveforms of the MOSFET are shown in Figure 3.2b.

As can be seen in the figure, the MOSFET is in its off-state until t1. At t1, $V_{gs}$ has charged up to its threshold voltage, $V_{gsth}$. When $V_{gs} = V_{gsth}$, the MOSFET begins to conduct. For an inductive load, this means that the current begins to rise to its final value. Thus the drain current, $I_d$, rises from zero, to its on-state value between t1 and t2. At t2, $V_{gs}$ has reached $V_{gpp}$. It is at $V_{gpp}$ during which the voltage transitions of the power MOSFET occur. From t2 to t3, Q2, or $Q_{gd}$, is supplied to the gate of the MOSFET, and $V_{ds}$ goes linearly from its off-state value of $V_s$ to its on-state value of near zero. At t3, the MOSFET is fully turned on, with the gate voltage at the plateau voltage. After t3, $V_{gs}$ continues to rise as more charge is supplied to the gate. $V_{gs}$ stops charging when the voltage being applied at the gate is equal to the gate voltage at t4 (Motorola, 1994, p. 2-6-4).

The gate charge curves are fairly invariant over the possible load, drive, and supply conditions. In general, the plateau voltage will increase slightly as the drain
current increases. The plateau voltage for most MOSFETs is around 5-6 volts and may vary in a 2-3 volts range for very light, or very heavy drain currents. Also, the amount of charge Q2 supplied in the plateau region increases slightly with the system voltage being applied to the drain (Motorola, 1994, p. 2-6-6). Unless a very accurate model of the MOSFET and the system is available, these variances are of no concern because the unmodeled effects and noise in a practical circuit are greater than the switching characteristics changing over possible operating conditions. As a result, the switching times remain independant of the load. Also, the switching losses (to be discussed in Chapter 4) become a function of the system voltage and drain current.

Even though the MOSFET is turned on completely at t3, it is desirable to charge the gate up to higher voltage than $V_{gsr}$. It is recommended by Motorola and is standard practice to charge the gate up to 10 or 12 volts. This will insure full turn-on and decrease the on-state resistance, $R_{ds(on)}$, of the MOSFET. $R_{ds(on)}$ decreases slightly as $V_{gs}$ is charged up after t3. It is possible to decrease $R_{ds(on)}$ by 10% or 15%, depending on the MOSFET, by charging $V_{gs}$ to 18 volts instead of 12 volts. This is usually undesirable due to the increased likelihood of voltage transients on the gate that could be destructive. The topic of fatal voltage transients is discussed in Section 3.2.

It is more common in application to drive the MOSFET with a voltage source and a series resistance than a constant current source. The $V_{ds}$ and $I_d$ waveforms versus time are shown in Figure 3.2b during switch-on. During switch-off, the waveforms are identical in shape, but they occur in the opposite direction. During switch-on from t2 to t3, $V_{ds}$ falls from the system voltage to its on-state value. This is known as the voltage
fall time, $t_{fv}$. Since this transition occurs while $V_{gs}$ is at the plateau voltage, the gate current, $I_g$, is also constant. Knowing the general equation which relates time, charge, and current, $t = q/i$, $t_{fv}$ can be found from

$$t_{fv} = \frac{Q_{gd}R_{eff(on)}}{V_{source} - V_{gs}}.$$  \hspace{1cm} (3.1)

In Equation 3.1, $R_{eff(on)}$ is the gate driver's effective resistance in the "on" direction and $V_{source}$ is the output voltage of the gate driver (Motorola, 1994, p. 2-6-9). Recall that $Q_{gd}$ is shown as Q2 in Figure 3.2a. In a similar fashion, the voltage rise time, $t_{vr}$, when the MOSFET is switching off, can be found. The voltage rise time is given by

$$t_{vr} = \frac{Q_{gd}R_{eff(off)}}{V_{gs} - V_{sink}},$$  \hspace{1cm} (3.2)

where $R_{eff(off)}$ is the gate driver's effective "off" resistance (Motorola, 1994, p. 2-6-9). $V_{sink}$ is the gate driver's effective sink voltage. $V_{sink}$ is usually near ground and may even be a negative voltage. It is possible to have a different output resistance for the gate driver during switch-on and switch-off. This can be accomplished with the use of two diodes to control the current path through different drive resistors. It is often desirable to control these times individually.

The current transitions for the power MOSFET occur between $t1$ and $t2$. Since it is easy to get the value of $Q_{gs}$ from the manufacturer's data sheet, it will be assumed that the current transitions take place linearly in the time from $t0$ to $t2$ instead of $t1$ to $t2$. This will provide for a reasonable estimate since the reverse recovery current of the flyback diode is not being considered and it can be a large source of error. The current
rise time, $t_{ri}$, is the time it takes the current to rise from zero to its on-state value during switch-on. This can be found using gate charge data as

$$t_{ri} = \frac{Q_{gs} R_{eff(on)}}{V_{source} - \frac{V_{gs}}{2}}.$$  \hspace{1cm} (3.3)

During switch-off, the gate charge of the MOSFET is removed through the turn-off resistor. The current fall time, $t_{rf}$, which is assumed to occur between $t_2$ and $t_0$, is given by

$$t_{rf} = \frac{Q_{gs} R_{eff(off)}}{V_{gs} - V_{sink}} \frac{1}{2}$$

$$= \frac{2Q_{gs} R_{eff(off)}}{V_{gs} - V_{sink}}.$$  \hspace{1cm} (3.4)

Power MOSFETs are usually thought of as voltage controlled devices, but Equations (3.1) through (3.4) show that they are better modeled as charge controlled devices. Also, if it is desired to switch the MOSFETs in less than a microsecond, a gate drive current on the order of several amps will be necessary, depending on the amount of charge the MOSFET requires. This makes gate driver design interesting, as they must be capable of providing large amounts of current for a very short amount of time. Now that the switching times can be calculated, the amount of energy lost during switching can be determined. The topic of power losses in a buck chopper will be covered in Chapter 4.
3.2 Protection Against Fatal Operating Conditions

Power MOSFETs are quite rugged when it comes to stressful overcurrent conditions. For example, the Motorola MTY55N20E used in this controller is rated to conduct 55 A continuous, but is capable of withstanding a 10 micro-second pulse of 165 A, or, in other terms, three times its continuous current rating. The overcurrent conditions are acceptable as long as the semiconductor junction temperature is kept below its maximum rating of 150 °C. Under normal operating conditions, this becomes a heat sinking issue. Overvoltage conditions create another problem for power MOSFETs. MOSFETs can quite easily be destroyed by overvoltage transients because this stressful condition affects the insulation layers in the MOSFET which are thin and cannot withstand exceeding their rated value for very long (Motorola, 1994, chap. 5). The devices are destroyed when the transients contain enough energy for the device to avalanche. The amount of energy that the MOSFET can absorb before going into avalanche breakdown is called the avalanche energy. The avalanche energy rating of MOSFETs decreases with an increasing junction temperature, thus increasing the susceptibility to damaging voltage transients under periods of a heavy load (Motorola, 1994, chap. 5). There are two basic overvoltage conditions which must be avoided. These are overvoltage conditions on the drain-to-source junction and overvoltage conditions on the gate-to-source junction. In contrast to most areas of engineering, in power electronics it is common to refer to these transients on signals as noise.

Exceeding the maximum rated drain-to-source voltage, $V_{(br)DS}$, of a power MOSFET is probably the most common cause of device failure (Motorola, 1994, p. 2-4-
3. In switching power supply applications such as a buck chopper, exceeding $V_{br(id)}$ is quite easy if precautions are not taken. Drain voltage transients are caused by switching high currents through load or stray inductances (Motorola, 1994, p. 2-4-3). The flyback diode is very critical here because it provides a current path for the load current, thus saving the MOSFET from absorbing the energy stored in the load inductance (Motorola, 1994, p. 2-4-3). Stray inductances pose a significant problem because they are usually unknowns during the design phase. Stray inductance is any inductance associated with the wiring and layout of the circuit and its components. Proper circuit layout is important to minimize the amount of stray inductance in the circuit.

Voltage transients encountered during switch-on are not as important as those encountered during switch-off because they are usually smaller. If the voltage transients during switch-off can be controlled, then the voltage transients during switch-on should not be of damaging magnitude.

Overshooting the system bus voltage is the type of drain-to-source voltage transient encountered during switch-off. There are several ways to protect the MOSFET from this problem. One way to reduce overshoot is to slow down the switch-off time by increasing $R_{eff(\text{off})}$. This reduces $di/dt$, which in turn lowers the voltage that will be seen across the drain of the MOSFET due to the stray inductances. This maximum drain-to-source voltage encountered during switch off can be expressed as

$$V_{ds(\text{max})} = V_s - L \frac{di}{dt}$$

(3.5)

In Equation (3.5), $L$ is the sum of the stray inductances and the package inductance.
internal to the power MOSFET (Motorola, 1994, p. 2-4-5).

Figure 3.3: Overvoltage Transient on $V_{ds}$ due to Load Inductance (IR, 1993, p. 1536).

Figure 3.3 shows a drain-to-source overvoltage transient which would result from switching an inductive load without a flyback diode. Operating a chopper on an inductive load without a flyback diode is very stressful on the chopper. Also, due to the nature of the load and chopping frequencies normally used, the chopper most likely will not even operate without a flyback diode if the duty cycle is not close to one. Figure 3.4 shows a chopper with a flyback diode in place. It still has a large overshoot during switch-off due to the stray inductance which is not clamped by the flyback diode.
A microhenry, or less, of stray inductance may seem like an insignificant amount, but it is very important to the magnitude of the voltage transients. Consider a switching time which produces a typical current transition rate of 100 amps per microsecond. If even one half of a microhenry of stray inductance is present in the circuit, the voltage associated with $\frac{L}{dt}$ of the stray inductance would be 50 volts! As a result, the voltage on the drain of the MOSFET during switch-off would overshoot the system operating voltage by 50 volts.

Another way to reduce overshoot is to place a Zener diode directly across the drain-to-source junction of each MOSFET. The Zener diode will then clamp the voltage transient at the Zener breakdown voltage. When using this technique it is important to choose the power rating of the Zener such that the clipped energy is safely dissipated
(Motorola, 1994, p. 2-4-4). A chopper utilizing a Zener diode to reduce voltage transients is shown in Figure 3.5.

![Figure 3.5: Reduction of Overvoltage Transients Using a Zener Diode (IR, 1993, p. 1536).](image)

Snubber circuits and clamp networks employing resistors and capacitors are also employed to prevent the drain voltage from exceeding $V_{brid}$ (Motorola, 1994, p. 2-4-4). A good design normally does not depend on any type of protection devices, although they are often included for an increased safety margin. A final cause of drain-to-source voltage transients needs to be considered. If the system operating voltage, $V_s$ (E in Figures 3.3 - 3.5), is not stable, the supply could produce damaging voltage transients that will destroy the circuit (Motorola, 1994, p. 2-4-4).

Voltage transients on the gate-to-source junction that exceed its breakdown rating of $V_{brigs}$ can also destroy the MOSFET. Sources of these transients can result from either the gate driver circuit or from capacitive coupling through $C_{ds}$ from transients on the
drain-to-source junction. The series resistance from the gate driver circuit helps to damp out any voltage transients on the gate. If it is possible that the gate will see transients that exceed $V_{brig}$, a Zener diode with a Zener voltage equal to or slightly less than $V_{brig}$ should be placed across the gate-to-source junction of the MOSFET (Motorola, 1994, p. 2-4-6).

Voltage transients need to be taken very seriously since they can destroy the device so rapidly. They are also predictable and repetitive so the circuit can be designed such that the transients are harmless. Special care needs to be taken when several devices are operated in parallel, such as for this thesis, because it has been the experience of the author that if a single MOSFET is destroyed, it will cause all other devices in parallel with it to fail also. This happens so rapidly that fuses are unlikely to protect the circuit.

### 3.3 Flyback Diode

The flyback diode used in the controller for this thesis was Motorola part number MURP20040CT. The data sheets for this part are given in Appendix A.2. This ultrafast recovery rectifier features a reverse recovery time of 75 nanoseconds. This part has dual diode construction which means that it is two diodes with a common cathode and separate anodes. Each "leg" of this diode is capable of conducting 100 amps and blocking 400 volts in the reverse direction. This is a larger blocking voltage than the 200 volt rating of the power MOSFETs used in this thesis. A larger blocking voltage was necessary because it was observed that the flyback diodes experience larger voltage transients than the MOSFETs.
Understanding a diode's operating characteristics is easier than understanding those of a MOSFET. Important device characteristics that need to be considered during the design phase are the reverse recovery times, and the forward voltage drop and its corresponding characteristics. The reverse recovery time is important, and should be as fast as possible. Schottky diodes feature the fastest recovery times of all diodes, but their blocking voltage is limited to about 150 volts which precluded them for use in this controller. The recovery time needs to be faster than the switching time of the drain current in order for the diode to clamp the inductive energy of the load. Like all components, the diodes are not ideal and thus have a forward voltage drop, $V_f$, when they are in a conducting mode. The forward voltage drop is a function of the junction temperature of the device. Unlike MOSFETs, diodes have a negative temperature coefficient. This means that as the devices heat up, their forward voltage drop decreases. Therefore, it is more efficient if the heat sink for the flyback diodes is sized such that the diodes are close to their maximum operating temperature. Unfortunately, this also creates the possibility of thermal runaway if paralleled diodes are not maintained at very close temperatures. This can usually be solved with proper heatsinking and by overdesigning the current carrying capacity of the flyback diode in the chopper.
4 Losses in a Buck Chopper Drive

In Chapter 2 it was assumed that all electronic components in the chopper drive were ideal. This allows for simple computations of system operation on a large scale, but does not truly model the behavior of the electronic components. In reality, none of the electronic components are ideal and the energy losses in the components are dissipated mostly as heat. To a much lesser extent, some of the losses are generated as sound which can be heard as a faint buzz if the frequency of the chopper is in the audible range. The ideal switch would be able to block an infinite reverse voltage, switch on and off instantaneously and have zero resistance when it is on. Chapter 3 discussed the requirements for switching the power MOSFETs used as the switch in this controller. As a result, in this real chopper there is a loss associated with switching the MOSFETs on and off, as well as losses when they are conducting. Also, there is a voltage drop across the flyback diode when it is conducting. Since most of this energy is dissipated as heat, heat sink theory and requirements are also discussed in this chapter.
4.1 Switching Losses

To study the power dissipation in a buck chopper or any other power electronics circuit, it is necessary to have profiles of characteristic instantaneous voltage and current waveforms. Consider Figure 4.1, which is a buck chopper configured like the one built for this controller. This is essentially the same as Figure 2.3, but with one small change. Notice that in Figure 4.1, the MOSFETs (shown here as a switch) are connected to the ground terminal of the battery input as opposed to the positive terminal connection shown in Figure 2.3. This simplifies gate drive requirements as discussed in Chapter 3.

![Buck Chopper Circuit Diagram](image)

**Figure 4.1:** Buck Chopper Circuit for Switching Characteristics (Mohan, Underland, and Robbins, 1989, p. 10).

A typical input signal is shown in Figure 4.2a while the corresponding drain-to-source voltage ($V_{ds}$) and drain current ($I_d$) waveforms are presented in Figure 4.2b. The switching transition characteristics of these waveforms are typical of those for a chopper with an inductive load (Mohan et al., 1983, p. 9). They are shown here with smoothed
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transitions. In reality, there will be some noise on both the $V_{ds}$ and $I_d$ waveforms which will cause the switching losses to be somewhat larger than what is calculated. This really is not noise by the usual definition of noise. It is really unmodeled effects that cause transients on the waveforms. In power electronics circuits, these effects are commonly referred to noise. The flyback diode in this section is assumed to be ideal since the focus is on switch characteristics. In reality, reverse recovery current can significantly affect the switch waveforms, and is one of the chief sources of noise in the circuit. For additional information on switching and conduction losses, see the reference by Mohan, Underland, and Robbins (1983). The information presented in Sections 4.1, 4.2, and 4.3 is derived from this text.

To switch the MOSFETs on, a positive voltage is applied to the gate drive circuitry. As discussed in Chapter 3, there is a turn-on delay time, $t_{d(on)}$, while the input
capacitance is being charged up to the threshold voltage. Once this threshold voltage has been reached, the drain current increases until the full load current is flowing in the switch. The time it takes for the current to flow from zero to its on-state value, is the current rise-time, and is shown as \( t_{ri} \) in Figure 4.2b. At this time the flyback diode becomes reverse biased and the voltage across the switch falls to its on-state value, \( V_{on} \). This is the voltage fall-time, \( t_{vf} \), shown in the figure. Let the total switch on-time be given as

\[
t_{s(on)} = t_{ri} + t_{vf} \quad (4.1)
\]

From Figure 4.2b it is possible to calculate the energy dissipated during switch-on. This energy is given by

\[
W_{s(on)} = \frac{1}{2} V_s I_d t_{s(on)} \quad (4.2)
\]

To switch the MOSFETs off, the positive voltage to the gate drive circuitry is reduced to ground potential. When this happens, gate charge is removed from the MOSFETs until the gate voltage has reached its plateau value. After this short turn-off delay time, \( t_{d(off)} \), \( V_{ds} \) begins to rise to the system input voltage, \( V_s \). When \( V_{ds} = V_s \), the flyback diode goes into conduction and the drain current falls from its on-state value to zero. The voltage rise-time and current fall-time are shown in Figure 4.2b as \( t_r \) and \( t_{fi} \) respectively. Although shown as equal in the graph, \( t_{d(on)} \) and \( t_{d(off)} \) are generally not equivalent. The total switch turn-off time is given as
\[ t_{s(\text{off})} = t_{f1} + t_{r1} \] \hspace{1cm} (4.3)

The corresponding energy lost during a single turn-off transition is then given by

\[ W_{s(\text{on})} = \frac{1}{2} V_s I_d t_{s(\text{on})} \] \hspace{1cm} (4.4)

Since Equations (4.2) and (4.4) give the energy lost during a single transition, the average switching power loss is found by multiplying by the switching frequency, \( f_s \).

Thus the average switch-on and average switch-off power losses are given by

\[ P_{s(\text{on})} = \frac{1}{2} V_s I_d f_s t_{s(\text{on})} \] \hspace{1cm} (4.5)
\[ P_{s(\text{off})} = \frac{1}{2} V_s I_d f_s t_{s(\text{off})} \] \hspace{1cm} (4.5)

The total power lost during switching is the sum of the power lost during switch-on and switch-off. Thus the total switching losses are given by

\[ P_s = \frac{1}{2} V_s I_d f_s (t_{s(\text{on})} + t_{s(\text{off})}) \] \hspace{1cm} (4.6)

From a power loss standpoint, Equation (4.6) clearly shows fast switching times directly decrease losses, as does using the lowest switching frequency that will provide a desirable power output.

### 4.2 Conduction Losses

Since real world electronics are not ideal, they have a small but non-negligible on-state resistance. As a result, energy is dissipated as heat while the switch is conducting. Since, in general, \( t_{on} \approx (t_{s(\text{on})} + t_{s(\text{off})}) \), the switches can be assumed to switch on and off
instantaneously for conduction loss calculations. The conduction energy lost in a given period, as shown in Figure 4.2b, is given by

\[ W_c = V_{on} I_d t_{on} \]. \hspace{1cm} (4.7)

It is important to remember that \( V_{on} \) is equivalent to the on-state drain-to-source resistance, \( R_{ds(on)} \), times the drain current. Also, the conduction power is found by multiplying Equation (4.7) by the chopping frequency. It follows that the conduction power loss is given by

\[ P_c = V_{on} I_d t_{on} f_c \]. \hspace{1cm} (4.8)

Equation (4.8) can be put into a more useful form if \( V_{on} \) is substituted as discussed, and it is noticed that \( t_{on} f_c \) is the duty cycle, \( D \), of the chopper. Thus the on-state conduction power losses are given by

\[ P_c = I_d^2 R_{ds(on)} D. \] \hspace{1cm} (4.9)

Equation (4.9) shows that it is desirable for the switch to have the smallest possible resistance when switched on.

Another important part of conduction losses are the power losses in the flyback diode when it is conducting. The flyback diode conducts when the MOSFETs are off. The conduction energy lost in a given period is given by

\[ W_D = V_F I_d t_{off}, \] \hspace{1cm} (4.10)

where \( I_d \) is the drain current, which is also the same current that flows through the flyback diode when it is conducting, and \( V_F \) is the forward voltage drop across the flyback
diode. By multiplying Equation (4.10) by the chopping frequency, the power lost in the flyback diode can be obtained. Noting that \( f_{\text{off}} = 1 - D \), the conduction power loss in the flyback diode is given by

\[
P_D = V_F I_d (1 - D).
\]  
(4.11)

### 4.3 Power Loss Results

Since Sections 4.1 and 4.2 developed the three largest components of losses in a buck chopper, it follows that the total chopper losses are the sum of Equations (4.6), (4.9), and (4.11). The total chopper power losses are given by

\[
P_{\text{loss}} = \frac{1}{2} V_s I_d f_{c} (t_{s(\text{on})} + t_{s(\text{off})}) + I_d^2 R_{ds(\text{on})} D + V_F I_d (1 - D).
\]  
(4.12)

Equation (4.12) shows that as the duty cycle increases, the conduction losses increase proportionally, while the flyback diode conduction losses decrease inversely with the duty cycle. Also, the switching losses in the chopper are mostly independent of the duty cycle. It is important to note that if the duty cycle of the chopper is one, such as when the throttle is depressed completely to the floor, the chopper is operating in its most efficient configuration. With a duty cycle of one, the chopper does not have any flyback diode losses or switching losses.

There exist a few other contributors to the power lost in a buck chopper, but they are generally on the order of a few watts. This is negligible compared to the hundreds of watts resulting from Equation (4.12). These additional sources of lost power include:
line losses in the wiring, power for the control electronics and very small leakage currents when the switch is turned off. The aforementioned noise on the voltage and current waveforms that inherently exists in a real chopper will cause the power loss given in Equation (4.12) to be a non-conservative estimate. Therefore, when designing the cooling heat sinks for the components, it is necessary to make their capacity larger than Equation (4.12) would indicate.

4.4 Heat Sink Theory

The purpose of a heat sink is to dissipate the power lost in semiconductor devices in order to maintain a safe device temperature. Most heat sinks are made of extruded finned aluminum and either remove heat by natural convection or forced air cooling. Forced air cooling allows for a smaller heat sink to be used than natural convection alone. An electrical analog to the operation of a heat sink is shown in Figure 4.3.

![Diagram of a heat sink](image)

**Figure 4.3:** Electrical Analog of a Heat sink (Thermalloy, 1990, p. 10).

In Figure 4.3, \( R_{\theta JC} \) is the thermal resistance of the semiconductor device junction to the outer case of the device. \( R_{\theta CS} \) is the thermal resistance of the case of the device
to the heat sink. This is normally comprised of the thermal resistance of a thermal grease used to attach the device to the heat sink or some type of thermal conductive pad such as mica. $R_{\theta SA}$ is the thermal resistance of the heat sink to the ambient air. As can be seen in the figure, $R_{\theta JC}$ and $R_{\theta CS}$ are attributed to the device and are connected in series. If devices are connected in parallel, these thermal resistances are also connected in parallel like their electrical analogs. Also in Figure 4.3, $Q$ is the amount of heat, in watts, entering the heat sink, and $\Delta T_{JA}$ is the temperature rise of the device junction, relative to the ambient air temperature. It is standard to give temperatures in °C and thermal resistances in °C/watt. Figure 4.3 implies that the junction temperatures of the three paralleled devices shown are equal. This is not necessarily the case. Small differences exist in $R_{\theta JC}$ in the devices due to the manufacturing process. The small difference is negligible if the paralleled devices are matched, or if they are from the same manufacturing lot number. Similar to Ohm’s law, $\Delta T_{JA}$ is given by

$$\Delta T_{\theta JA} = QR_{\theta tot},$$

(4.13)

where $R_{\theta tot}$ is the equivalent thermal resistance of the $R_{\theta JC}$'s, $R_{\theta CS}$'s, and $R_{\theta SA}$ (Thermalloy, 1990, p. 10). Equation (4.13) can be written in the following forms:

$$Q = \frac{T_J - T_A}{R_{\theta JC} + R_{\theta CS} + R_{\theta SA}}$$

(4.14)

$$Q = \frac{T_C - T_A}{R_{\theta CS} + R_{\theta SA}}$$

(4.15)
\[ Q = \frac{T_S - T_A}{R_{SSA}} \]  

(4.16)

In Equations (4.14), (4.15), and (4.16), \( T_J, T_C, T_S, \) and \( T_A \) are the temperatures of the semiconductor junction, the semiconductor case, the heat sink, and the ambient air respectively (Thermalloy, 1990, p. 10).

Forced air cooling is used when a heat sink cannot be designed to cool the device sufficiently by convection because of space or weight limitations. Forced air cooling relies on an external force, such as a fan or duct work, to increase the air flow in order to speed up the convection process. The linear velocity of the air current as it passes over the heat sink is the key factor in the amount of heat being removed. When forced air cooling is used, the thermal resistance of the heat sink is largely a function of the amount of surface area of the heat sink, and not the volume, or shape of the heat sink (Aavid, 1987, p. 6). The amount of surface area of the heat sink is the Profile Perimeter, \( PP \), and is normally given in units of (in\(^2\)/in), which is square inches of surface area per inch of length of the heat sink. Thus the Profile Perimeter becomes one of the most important design criteria when selecting a heat sink to be used with forced air cooling.

The linear velocity of air as it passes over the heat sink is normally used in the units of linear feet per minute, or LFM. Although the volume of air (cubic feet per minute, or CFM) is not a governing factor in the cooling process, most fans are rated by this volume flow and not in LFM. To find the LFM rating from the given CFM rating of the fan, it is needed to know the cross-sectional area that the air is passing through and the CFM rating of the fan. These quantities are related by
Velocity (LFM) = \frac{Volume (CFM)}{Area (ft^2)} \quad (4.17)

Most fans are nominally rated and compared at their free air delivery, or with the fan working against zero back pressure. This is rarely the case in application. As a result, it is necessary to derate the volume of air flow to 60% to 80% of the rated capacity, in the anticipation of back pressure (Aavid, 1987, p. 6).

Since the thermal resistance of a heat sink is normally given for natural convection, it is necessary to compute a new thermal resistance for the heat sink for a given air flow when forced air cooling is used. To calculate this new thermal resistance, consider the chart given in Figure 4.4.

**EXTRUSION FORCED CONVECTION THERMAL RESISTANCE TABLE**

<table>
<thead>
<tr>
<th>Air Velocity (CFM)</th>
<th>100</th>
<th>200</th>
<th>300</th>
<th>400</th>
<th>500</th>
<th>600</th>
<th>700</th>
<th>800</th>
<th>900</th>
<th>1000</th>
<th>1100</th>
<th>1200</th>
</tr>
</thead>
<tbody>
<tr>
<td>Length of Extrusion (inches)</td>
<td>0.25</td>
<td>129.63</td>
<td>120.63</td>
<td>105.84</td>
<td>91.66</td>
<td>81.98</td>
<td>74.84</td>
<td>69.28</td>
<td>64.81</td>
<td>61.11</td>
<td>57.97</td>
<td>55.27</td>
</tr>
<tr>
<td></td>
<td>0.50</td>
<td>129.63</td>
<td>91.66</td>
<td>74.84</td>
<td>64.81</td>
<td>57.97</td>
<td>52.92</td>
<td>48.96</td>
<td>45.83</td>
<td>43.21</td>
<td>40.99</td>
<td>39.08</td>
</tr>
<tr>
<td></td>
<td>1.00</td>
<td>91.66</td>
<td>64.81</td>
<td>52.92</td>
<td>48.96</td>
<td>45.83</td>
<td>43.21</td>
<td>40.99</td>
<td>39.08</td>
<td>37.42</td>
<td>35.33</td>
<td>33.68</td>
</tr>
<tr>
<td></td>
<td>2.00</td>
<td>64.81</td>
<td>45.83</td>
<td>37.42</td>
<td>34.21</td>
<td>32.44</td>
<td>30.53</td>
<td>28.97</td>
<td>27.62</td>
<td>26.44</td>
<td>25.43</td>
<td>24.64</td>
</tr>
<tr>
<td></td>
<td>5.00</td>
<td>15.94</td>
<td>15.26</td>
<td>14.68</td>
<td>14.22</td>
<td>13.86</td>
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<td>3.80</td>
<td>3.70</td>
<td>3.60</td>
<td>3.52</td>
<td>3.44</td>
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<td>2.92</td>
<td>2.80</td>
<td>2.70</td>
<td>2.60</td>
<td>2.52</td>
<td>2.44</td>
<td>2.36</td>
<td>2.30</td>
<td>2.25</td>
<td>2.20</td>
</tr>
<tr>
<td></td>
<td>12.0</td>
<td>2.05</td>
<td>1.92</td>
<td>1.80</td>
<td>1.70</td>
<td>1.60</td>
<td>1.52</td>
<td>1.44</td>
<td>1.36</td>
<td>1.30</td>
<td>1.25</td>
<td>1.20</td>
</tr>
<tr>
<td></td>
<td>13.0</td>
<td>1.05</td>
<td>0.92</td>
<td>0.80</td>
<td>0.70</td>
<td>0.60</td>
<td>0.52</td>
<td>0.44</td>
<td>0.36</td>
<td>0.30</td>
<td>0.25</td>
<td>0.20</td>
</tr>
<tr>
<td></td>
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<td>0.05</td>
<td>0.02</td>
<td>0.00</td>
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<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
</tr>
</tbody>
</table>

Figure 4.4: Chart of Performance Factors to Determine Thermal Resistances with Forced Air Cooling (Aavid, 1987, p. 7).

To calculate the thermal resistance of an extruded aluminum heat sink, find the
corresponding Performance Factor (PF) from Figure 4.4 at the intersection of the Air Velocity column and Length row. The thermal resistance of the extrusion with forced air cooling is then obtained by dividing the Performance Factor by the Profile Perimeter, PP (in²/in), exposed to the airflow. The quantity PP, is found in the technical data for each heat sink. This can also be written with an equation as

\[
R_{\text{ ESA}} = \frac{PF}{PP}. \tag{4.18}
\]

From Equation (4.18), it can be concluded that if forced air cooling is going to be used, it is desirable to have a heat sink with many thin fins in order to make PP large and the weight small (Aavid, 1987, p. 6). Also, from Figure 4.4, it can be seen that there is a diminishing return of performance when increasing the speed of the airflow, or by lengthening the heat sink extrusion.

4.5 Heat Sink Calculations

The heat sinks used in the controller for this thesis were Aavid model number 60540. The data for this heat sink is shown in Figure 4.5 as it is given in the Aavid catalog. For this project, forced air cooling was used to keep the controller at a safe operating temperature. This heat sink was chosen because it was readily available. These heat sinks were pre-cut in 7 inch lengths. This heat sink provides ample room to mount the MOSFETs flat on the underside of the heat sink as it is shown in the figure. On each heat sink, a total of 7 MOSFETs were mounted. For the flyback diodes, a total of 2 diode modules (recall each discrete diode is a dual diode) were mounted on each heat sink.
A 100 CFM fan was chosen to blow air over the heat sinks. This was the largest fan available in an approximate 5 inch square package. Since the heat sinks, are 2.25" x 5", 2 individual heat sinks were oriented on top of one another to comprise an approximate 5" x 5" x 7" long heat sink package. Between the two heat sinks is a plastic insulator to keep the heat sinks electrically isolated from one another.

Using Equation (4.17), a 100 CFM fan produces a linear air velocity across a 5" x 5" area of

\[
Velocity = \frac{100 \text{CFM}}{25 \text{in}^2} = \frac{144 \text{in}^2 \text{ft}^2}{144 \text{in}^2 \text{ft}^2} = 576 \text{ LFM}.
\]

Since there is an unknown amount of back pressure in the system, it is assumed that this airflow will be reduced by 30%. This produces a linear air velocity across the heat sinks of roughly 400 feet per minute (576x70% = 403.2). From Figure 4.4, the Performance Factor corresponding to a 7" long extrusion with 400 ft/min of airflow can be found. This Performance Factor is 17.30. The Profile Perimeter of the heat sinks used is 56
in²/in from Figure 4.5. Using Equation (4.18), the thermal resistance from the heat sink to the ambient air can be found. This is given by

\[
R_{B5A} = \frac{17.30}{56.0} \quad (4.20)
\]

=0.309 °C/watt.

It is also necessary to know \(R_{0JC}\) and \(R_{0CS}\) in order to calculate the cooling capacity of the heat sink. For the MTY55N20E power MOSFETs, \(R_{0JC}\) is given as 0.42 °C/watt from the data sheets for this part included in Appendix A.1. \(R_{0CS}\) needs to be calculated and depends on the type of thermal conductor used between the device and the heat sink.

For this controller, the conductive grease used between all power semiconductors and the heat sink material was Conducto-Lube, a product of the Cool-Amp Conducto-Lube Company in Oswego Lake, Oregon. This conductive lubricant is a mixture of pure silver and mineral oil. Its thermal and electrical properties are excellent for power semiconductors. Conducto-Lube has a conductivity of 65 btu/hr/ft²/°F/in. This conductivity can be converted to the resistivity that is normally used for heat sink calculations. This thermal resistivity, \(p\), is given by

\[
p = 4.20 \frac{\circ C \ in^2}{W \ in}. \quad (4.21)
\]

In contrast, the white silicon grease that is commonly used as a conductive layer for heat sinks has a resistivity \(p\) around 200 (Aavid, 1987, p. 4).

The thermal resistance of the semiconductor case to the heat sink is then found
from

\[ R_{\theta CS} = \rho \frac{t}{A} \]  \hspace{1cm} (4.22)

where \( t \) is the thickness of the heat sink grease and \( A \) is the area of contact with the heat sink.

Each MTY55N20E power MOSFET has a contact area with the heat sink of 0.691 in\(^2\). This was calculated from the footprint information of the part as given in Appendix A.1. It was estimated that the thickness of the Conducto-Lube used in this thesis was 0.003 inches thick. Using Equation (4.22), \( R_{\theta CS} \) of each MOSFET is then given by

\[ R_{\theta CS} = 4.20 \frac{\degree C \text{ in}^2}{W \text{ in}} \frac{0.003 \text{in}}{0.691 \text{in}^2} \]  \hspace{1cm} (4.23)

\[ = 0.0182 \ \degree C/W. \]

This value of \( R_{\theta CS} \) is nearly insignificant compared to \( R_{\theta JC} \) as a result of using Conducto-Lube. Using Equation (4.24), the amount of heat that the heat sinks can dissipate while keeping the junction temperature of the MOSFETs at or below 125 \degree C can be calculated. It will be assumed that the temperature of the ambient air is 25 \degree C. This is given as
\[
Q = \frac{T_J - T_A}{1 + \frac{R_{\theta SA}}{1 + \frac{R_{\theta JC} + R_{\theta CS}}{125 - 25}} + \frac{0.42 + 0.0182}{0.309}}
\]

\[= \frac{269 \text{ watts}}{}
\]

If it is desired to maintain the junction temperature below 100 °C, each MOSFET heat sink can dissipate about 200 watts. This works out to losses on a per MOSFET basis of about 29 to 38 watts.

In the same manner, the amount of heat that the heat sinks which hold the flyback diodes can dissipate is calculated. For the MURP20040CT diodes used in this controller, the corresponding value of \(R_{\theta JC}\) is given as 0.45 °C/watt. This is given in the data sheets in Appendix A.2. Also from Appendix A.2, the amount of surface area that is in contact with the heat sink can be calculated to be 2.24 in\(^2\). Assuming that the thickness of the Conducto-Lube is 0.003 inches, \(R_{\theta CS}\) can be calculated using Equation (4.22) as

\[
R_{\theta CS} = 4.20 \frac{\text{°C}}{\text{in}^2 \cdot \text{w}} \cdot \frac{0.003\text{in}}{2.24\text{in}^2}
\]

\[= 0.0056 \text{ °C/watt.}
\]

Using this value of \(R_{\theta CS}\) and Equation (4.24), the amount of heat that the flyback diode heat sinks can dissipate while keeping the junction temperature of the diodes below 150 °C, can be calculated as
\[ Q = \frac{T_J - T_A}{1 + \frac{R_{\text{BAS}}}{2R_{\text{BAS}}}} \]

\[ = \frac{150 - 25}{0.45 + 0.0056 + 0.309} \]

\[ = 233 \text{ watts.} \]

If it is desired to maintain the junction temperature of the flyback diodes below 125 °C, the amount of heat that the heat sinks can dissipate will be reduced to about 186 watts.

These quantities of heat that can be safely dissipated without damage to the semiconductors are a maximum value. The actual value of the losses will depend on the operating conditions of the chopper. This will be covered in the design of the chopper in Chapter 5.
5 Controller Design

Armed with the theory of buck chopper drives and the knowledge of the components used to build one, the design and construction of the controller can take place. Controller layout and design involves much time spent in the laboratory and probing and testing with an oscilloscope before a design can be considered complete. As alluded to in previous chapters, the placement of components has direct impact on the operation of controllers at the high power levels encountered in an electric vehicle. At times, this seems more like art and magic than the sound engineering principles upon which it is based. This chapter has two sections which cover the two main components of the controller; the switch section and the control and pulse width modulation section.

5.1 Switch Layout and Design

The switch section of the controller consists of the MOSFETs, gate drive circuitry, and the flyback diodes. The controller was built as modules. With the modular design, the current carrying capacity of the controller is limited only by the number of modules connected in parallel. Each module consists of 7 Motorola MTY55N20E power
MOSFETs, 2 Motorola MURP20040CT ultrafast diodes, a 100 CFM fan, and the necessary gate driver and filter capacitors.

The seven MOSFETs were attached to one heat sink and the two diodes to another. Both of the heat sinks are used as electrical conductors for the circuit. The MOSFET heat sink serves as their drain connection and the flyback diode heat sink serves as the battery plus and motor plus connection as shown in Figure 5.1. These heat sinks were stacked on top of one another with a plexiglass separator between them. Since it is necessary to have the anode of the flyback diode and the drains of the MOSFETs connected, it would be desirable to attach these terminals to a common heat sink and use the heat sink as a conductor. Unfortunately, the flyback diodes used were such that the cathode needed to be mounted to a heat sink, thus the use of separate heat sinks. As a result, the heat sinks were connected by a threaded rod through a copper tube which is capable of carrying the current flowing through the controller. This connection needs to be as short as possible because it is essentially introducing a small amount of stray inductance into the circuit which will cause overshoot on turn-off as discussed in Chapter 4.

Filter capacitors are needed to stabilize the input battery voltage. For each module, three 1600 \( \mu \)F electrolytic capacitors were used. Also, two large electrolytic 2 \( \mu \)F capacitors were connected in parallel with the 1600 \( \mu \)F capacitors since they will respond faster to higher frequency noise. The placement of the capacitors is critical to the circuit operation. The circuit needs to be designed such that the filter capacitors are as close as physically possible to the source ground connection of the MOSFETs, and the
battery plus, motor plus connection on the cathode of the flyback diodes.

An electrical schematic of two switch modules connected in parallel is given in Figure 5.1. The two switch modules that were constructed, were built with the filter capacitors on opposite sides of the heat sinks in order to minimize the distance between units and the electrical connections between them.

The schematic also contains the gate driver circuit since it is contained in a module. As can be seen in the figure, each module contains its own gate driver circuit. The gate driver circuit receives the output of the pulse width modulation circuit along with a 12 volt and ground input.

The heart of the gate driver is a push-pull pair of npn-pnp Darlington transistors. The Darlington transistors provide the high gain that is necessary to produce the large amount of short duration current to turn the paralleled MOSFETs on. The 470 $\Omega$ resistors (R3 and R4 in the Figure 5.1) produce the base current to drive the Darlington pairs from the voltage input from the PWM circuit. The Darlington gate driver configuration worked well during all tests that were conducted on the controller, but it was not considered the ideal choice. In the original design, the gate driver consisted of a pair of push-pull n-channel p-channel power MOSFETs. The MOSFETs are desirable because they do not require as much drive energy as the Darlington transistors. They are also faster and have a nearly infinite gain and can provide a very large amount of the short duration current that is needed to switch the paralleled MOSFETs. The original MOSFET gate driver proved to be very noisy. Since MOSFETs inherently have some overshoot when switching off rapidly, the output of the push-pull pair of MOSFETs was not an ideal square wave. The
Figure 5.1: Electrical Schematic of Two Paralleled Switch Modules.
overshoot in the gate driver circuit was induced onto the gates of the paralleled power MOSFETs. This, in turn, was coupled to their drain connection. This resulted in an unreasonable and dangerous amount of overshoot on the drain-to-source junction of the paralleled power MOSFETs. With the layout of the MOSFET gate driver that was used, a successful method was not found that would reduce the noise to an acceptable amount. Therefore, the MOSFET gate driver was abandoned in favor of the Darlington gate driver.

Like the battery supply of the car, the gate driver needs filter capacitors on its 12 volt supply. Each gate driver circuit has two filter capacitors; a 1000 \( \mu \text{F} \) electrolytic capacitor and a 0.47 \( \mu \text{F} \) ceramic capacitor. These capacitors are C5, C6, C7, and C8 in the schematic.

The turn-on and turn-off times of the paralleled MOSFETs were controlled individually with the use of two diodes and two 100 \( \Omega \) potentiometers in the gate drive path of each module. The potentiometers are component numbers R9-R11 and the diodes are component numbers D7-D10 on the schematic. The diodes are low voltage Schottky diodes which is the fastest diode type available. This enables the diodes to have a minimal effect on the switching and delay times. Also included in the gate drive path are the gate drive resistors labeled R1 and R2 in the schematic. These are 14 individual 27 \( \Omega \) resistors. Individual resistors are necessary to damp out any noise between paralleled MOSFETs and to even out the charging rate of the gates of each MOSFET to insure that they switch simultaneously. During turn-on, the current path is through D10 and R12 and the related 27 \( \Omega \) resistors. During turn-off, the path is through D9 and R11 along
with the 27 Ω resistors. In addition to being able to control the turn-on and turn-off times individually, the potentiometers serve the important function of matching the turn-on and turn-off waveforms of paralleled modules to one another. If the switching waveforms of paralleled modules are not well matched, they will dissipate different amounts of heat due to switching losses. This discrepancy could cause the controller to fail.

Two other resistors are located on each module. These 10 kΩ resistors are labeled as R4, R5, R7, and R8 in the schematic. These resistors are pull-down resistors that constantly discharge the gates of the paralleled MOSFETs. In the emergency case that the input to the gate drivers is removed, these resistors will turn the controller off.

Between each gate and source terminal of the paralleled MOSFETs is a 20 volt Zener diode. These are marked as D1 and D2 in Figure 5.1. They protect the gate junction from transient voltage spikes as discussed in Chapter 3.

In an effort to keep stray inductance to a minimum and construct the circuit of each module in an orderly fashion, it is necessary to create a printed circuit board. This printed circuit board houses all of the gate drive circuit components and the gate and source connections of the MTY55N20E power MOSFETs which are attached to the heat sink. This source connection creates an interesting dilemma because it carries the entire current load of the module -- around 200 amps during periods of heavy load. It is impossible for a standard printed circuit board to handle this amount of current. The solution to this problem was to attach a \( \frac{3}{8} '' \times \frac{1}{8} '' \) copper bar to the printed circuit board very close to where the source leads of the MOSFETs were attached. Cool-Amp
shown on the top part of Figure 5.3 are the gate drive resistors, Zener diodes and locations for capacitors which were never used. The Darlington gate driver is located on the bottom right and the filter capacitors are in the left center. The box on the far left is a DB-9 connector that was used to connect the modules to the PWM circuit. It is important to note that very large traces were used since they are carrying signals that have a non-trivial amount of energy associated with them. This is seen in Figure 5.2.

There is one other important connection on the printed circuit board. The vertical strip along the left side connects the 12 volt and battery input voltage grounds. This is necessary since the gate drive must be referenced to battery ground potential. This connection is made on the printed circuit board so that it is as short as possible to minimize ground loops which wreak havoc during the operation of converters.

Each module is contained in a plexiglass enclosure. The enclosure serves to keep dust and dirt away from the electronics. It is also necessary to direct the airflow from the fan so that all the air travels over the heat sinks, instead of a large portion of it escaping out the sides if the controller were not enclosed.

The completed two modules weigh 21 pounds including bus bars and attachment points for the external electrical connections. Pictures of the two completed modules connected together are shown in Figures 5.4 and 5.5.
5.2 Pulse Width Modulation and Control Circuit

The Pulse Width Modulation (PWM) unit which acts as the control circuit of the controller is built around a Motorola MC33035 integrated circuit. The MC33035 is a 24 pin IC brushless dc motor controller. This high performance motor controller can also be used as a brush dc motor controller, which is the type of motor used in the Electric Bobcat. This IC contains many built in features including: undervoltage lockout, cycle-by-cycle current limiting with a selectable time delayed latched shutdown, internal thermal shutdown, and a fault output. The Motorola data sheets on this part are given in Appendix A.3.

The MC33035 requires a power supply voltage between 10 and 30 volts. Therefore, the 12 volt auxiliary battery of the Electric Bobcat will also be adequate to power the control circuitry. This supply voltage is connected to the control circuit through a 0.5 Ω resistor and is filtered by two capacitors; a 100 μF electrolytic capacitor and a 0.01 μF ceramic capacitor. These are shown as R1, C4, and C5 in Figure 5.6,
which is an electrical schematic of the control circuitry.

The accelerator pedal in the Electric Bobcat controls the position of a 5 kΩ potentiometer. A 5 kΩ potentiometer acting as the "gas pedal" is very common throughout the electric vehicle industry. This input needs to be converted to the 1.6 volt to 4.1 volt input that the MC33035 needs to produce a zero to one hundred percent duty cycle as shown in Figure 10 in Appendix A.3. The conversion is accomplished by using the three leads of the potentiometer and two resistors along with the 6.25 volt reference output on the MC33035. This configuration acts as a voltage divider network. The two voltage divider resistors are shown as R2 and R4 in Figure 5.6. The reference output is from pin 8 of the MC33035. The two resistors that were used in conjunction with the 5 kΩ potentiometer produce a voltage input to the MC33035 between 1.43 and 4.27 volts. This range gives about a 0.2 volt safety margin on both sides of the input range to ensure that the PWM unit does not produce an output if the accelerator is not depressed, and also that the controller can go to a full-on position.

The throttle position voltage divider network is connected to the delay ramp. The delay ramp circuitry slows down acceleration to protect the controller from large instantaneous increases in throttle position. During acceleration, the output of the voltage divider network is connected to an RC network composed of R3 and C1 as shown in Figure 5.6. This network has a time constant of 1 second and will allow the current in the controller to gradually build up to help protect it from large over current conditions. During deceleration, the charge stored in the capacitor is discharged by a 100 kΩ resistor connected in series with a signal diode; R7 and D1 in Figure 5.6. This network has a
Figure 5.6: PWM Unit Electrical Schematic.

- 12V GND
  - +12V
  - Pot (top) R2 3.3k
  - Pot (wiper) R3 100k
  - Pot (bottom) R4 25k
  - + Shunt 25k
  - - Shunt 1k

Components:
- MC33835
- HEPR0170 100k
- R5 8.2k
- C2 0.01u
- C4 0.01u
- C5 0.01u

PWM Output
time constant of 0.1 seconds. Therefore, the controller turns off much more quickly than it turns on. This is not harmful to the controller because the current is clamped by the flyback diodes and its magnitude is not affected by the rapid turn off. Without the addition of the diode and 100 kΩ resistor, the controller would also turn off with a time constant of 1 second. This slow deceleration is very undesirable and potentially dangerous for an automobile, especially during braking conditions.

The PWM frequency of the MC33035 is set with a simple RC network. The components in Figure 5.4 which accomplish this are R5 and C2. This 8.2 kΩ resistor and 0.01 µF capacitor combination produces a PWM frequency of about 15.8 kHz. Any frequency can be chosen by changing these components. Figure 1 in Appendix A.3 gives a graphical display of the frequency/timing resistor response of the MC33035.

The MC33035 contains a forward/reverse selector pin. This is pin number 3. Since the brush dc motor in the Electric Bobcat is used only in a forward rotation, this pin needs to be connected to a logic high input for the PWM unit to operate in the forward mode. This is accomplished by connecting the pin to the output of 6.25 voltage reference on pin 8. Pin number 23 is the brake input selector pin. Since the motor in the car is not used for braking purposes, this pin needs to be connected to a logic low input. This is accomplished by tying the pin to ground. In order for the PWM unit to even operate, the output enable pin (pin 7) needs to be activated. This pin was connected to the reference voltage output so that the PWM unit would be capable of operating whenever the 12 volt auxiliary battery in the car is connected.

A sensor input code of 100 on pins 4, 5, and 6 will produce a drive output signal
on pin 19 of the IC. This will allow for the low side switch operation that was used for the controller. Low side switching is using a gate signal that is referenced to the common system ground. This was accomplished by putting a logic one input into pin 4, by connecting it to the output reference voltage, and by grounding pins 5 and 6 to create a logic zero condition.

An excellent feature of the MC33035 chip is its capacity for cycle-by-cycle current limiting. In cycle-by-cycle current limiting, the output of the chopper is terminated if the chopper current is too high. The chopper is turned off for the remainder of that cycle and begins the next cycle independently of the previous current limited cycle. The MC33035 accomplishes this by looking for a signal on pin 15 with respect to ground (pin 16). This pin current limits at a threshold of 100 mV but will also work with logic level signals up to a maximum of 5 volts. On the PWM unit of this controller, the current sense input goes through an RC filter to smooth out any spikes at the beginning of the cycle which would cause the unit to current limit prematurely. In an important note, Motorola recommends connecting the current sense ground input directly to the ground pin on the IC using a separate connection from that of the rest of the circuit.

In the application notes for this chip, Motorola warns that it is very important to construct any circuit with a MC33035 chip on a printed circuit board. As a result, the PWM circuit was constructed on a printed circuit board which utilized heavy copper runs to minimize radiated EMI.
6 Experimental Test Setups and Results

Once the design, layout and construction phases of creating a controller are complete, it becomes time to apply power and test the result of all this research and development. This chapter covers the laboratory test setups and methods used to gather the test results which verified controller operation. Results are given for both a resistive load test and an inductive motor load test. Also discussed in this chapter are some of the failures encountered in route to this final design and the difficulties in obtaining accurate measurements.

6.1 Test Measurement Equipment

All of the experimental waveforms shown in this thesis were obtained using a Hewlett-Packard model number 54503A, 500 MHz Digitizing Oscilloscope. The data from the oscilloscope was obtained using the ScopeLink oscilloscope data/image analysis program operating on an IBM PS/2 Model 30 computer. The program is a product of the Hewlett-Packard company. Only two of the oscilloscope's four channels were utilized in testing since the oscilloscope can only sample two channels simultaneously.
Each channel produces a data vector that contains the time axis (x axis of the oscilloscope) and the vertical output axis (y axis of the oscilloscope). This vector contains 500 data points per axis which are plotted on the screen of the oscilloscope. The ScopeLink program was used to download these data vectors to the hard drive of the computer. This data was then imported into MATLAB™, where various calculations were performed on the data, before it was subsequently plotted to prepare the plots shown in this thesis. The only alteration of the data in this process was in changing the time axis to read in microseconds as opposed to scientific notation.

For some diagnostic purposes, a Tektronix Model 2236, 100 MHz analog oscilloscope was used. This oscilloscope was most helpful to display the maximum initial overshoot encountered during turn-off. This portion of the signal was difficult to capture on the digital oscilloscope due to the way it samples data and the high frequency and short duration of the overshoot spike.

6.2 Matching Module Switching Characteristics

This controller was designed in modular form with potentiometers to adjust the switching times of the paralleled power MOSFETs. Since the modules are then connected in parallel, it is imperative that these switching times and waveform shapes be very close in nature. If the switching characteristics are poorly matched, the modules will not share the load current equally causing a temperature imbalance in the controller. This could lead to controller failure.

Figure 6.1 shows the test circuit that was used to match the switching
characteristics of the two modules. In order to connect the two loads separately to the modules for this case, it was necessary to remove the aluminum bus bar that normally would connect the drain connections of the modules together. In Figure 6.1, the two diodes and two MOSFETs represent the paralleled components that make up each module. This setup was used with a battery input voltage of 48 volts nominal. Throughout this thesis, the term nominal battery voltage is used. This refers to the voltage rating of the supply batteries. Thus for 4 batteries connected in series, the nominal voltage is 48 volts, while the actual voltage is really closer to 50 or 52 volts for fully charged batteries.

![Diagram](image)

**Figure 6.1:** Test Setup to Match Switching Characteristics.

Figure 6.2 shows the drain-to-source voltages of the two modules. Figures 6.3a and 6.3b show these signals in expanded detail during switching since these characteristics
**Figure 6.2:** Drain-to-Source Voltages to Synchronize Switching.

**Figure 6.3:** Expanded View of Drain-to-Source Voltages to Synchronize Switching.
are difficult to discern in Figure 6.2.

While the load was connected and the chopper was running, the potentiometers in the gate drive path were adjusted so that the rise and fall times of the drain-to-source voltage were about 1 microsecond. This time is reasonably fast and keeps the overshoot down to a safe level. The potentiometers were also adjusted to get the waveform profiles to match one another as closely as possible. Figure 6.3 shows that the modules have very close switching characteristics. The settings on the potentiometers that resulted in Figures 6.2 and 6.3 were 14.4 Ω and 19.5 Ω in the on direction for the left and right units respectively, and 1.0 Ω and 1.3 Ω in the off direction.
6.3 PWM Output

The output of the pulse width modulator unit is shown in Figure 6.4. This output was taken when the controller was connected up to a resistive load. The PWM output shown in the figure was recorded on the input to the gate driver circuit on a module. The signal on one module is shown, but there is no noticeable difference between modules for this signal. The output is a high quality square wave. Notice that the output is not equal zero in the off state. This does not affect circuit operation because the level is low enough to pull the darlington pair of the gate driver to its off state.

![PWM Output Waveform at Gate Driver Input](image)

**Figure 6.4:** PWM Output at Gate Driver Input.
6.4 Difficulties With Measuring Current

It is desirable to have profiles of the various current waveforms in a buck chopper to go along with the voltage waveforms. Since an oscilloscope is used to obtain the waveforms, and this device can only measure voltage, it is necessary to insert a low resistance current shunt somewhere in the relevant current path. Possible shunt locations for different currents are given in Figure 6.5.

![Figure 6.5: Potential Current Shunt Locations.](image)

In Figure 6.5, shunts S1 and S3 measure the flyback diode current, and shunts S2 and S4 measure the load current. Shunts S5 and S6 measure the drain current in the power MOSFETs. Each pair of shunts will theoretically yield the same waveforms. Shunt S1 cannot be placed in this circuit since the cathode of the flyback diodes were
used as the battery plus connection and shunt S3 cannot be used since the load is connected directly to the anode of the flyback diodes. Shunt S6 would be very desirable because it is connected to ground at one end and could directly be used to supply the current limiting circuit in the PWM unit, but this shunt cannot be used either. It is in a poor physical location for a shunt because the filter capacitors need to be as close as possible to the source connections of the MOSFETs. It was found that the inclusion of a shunt here on this controller design caused too much noise in the controller operation to be beneficial.

Shunt S5 is a possibility due to the controller layout used, and shunts S2 and S4 can be used in any layout due to the external load connections. These three shunts yield the general profile of the current that is expected when floating the oscilloscope, to move its ground to take the measurements, but the readings are corrupted by a great deal of noise. It is not possible to filter the signals without destroying the current waveforms during switching, which is the information being sought. Measuring the voltage across the shunts in a differential mode with an oscilloscope would be ideal, except that it did not work due to the sensitivity of the oscilloscope available for the readings. The reading across the shunts is on the order of tens of millivolts, while the voltage across the MOSFETs is on the order of a hundred volts. Due to this difficulty in obtaining accurate current waveform profiles, they are not included in this thesis. The current readings that are given were obtained from measuring the average voltage across respective shunts with a digital multimeter. This method will give accurate average current measurements, but yields no insight into their instantaneous nature. For this, it is necessary to rely on the
information given in Chapters 2-4.

6.5 Resistive Load Test

A high power resistive load test is a useful test for a motor controller. The resistive load, if sized properly, will produce the same load conditions in terms of drain current as a motor load, but without the ill effects of inductance. The resistive load will produce clean switching waveforms if the controller is working properly. If, under a resistive load, the waveforms are very noisy, it would be prudent to determine the cause of the noise before the controller is connected to an inductive load since an inductive load will only magnify the problem. For this test, the resistive load was eight electric heaters submerged in a tub of water. The load resistance was 1.17 ohms, and the battery voltage in use was 120 volts. The average load current for this test was 26.4 amps as a voltage measured across a current shunt with a Fluke model 77 multimeter. The starting temperature of the heat sinks was 72 °F. The time axis in Figures 6.6 through 6.9 is synchronized.

Figure 6.6 shows the resulting drain-to-source voltage waveform from this test. As can be seen in the figure, the overshoot was minimal and the waveform is about as clean as can be expected. Figures 6.7a and 6.7b show the waveform of Figure 6.6 in an expanded view during switching.
Figure 6.6: Drain-to-Source Voltage for a Resistive Load.

Figure 6.7: Expanded View of Drain-to-Source Voltages During Switching.

Figure 6.7 is slightly noisy. This is probably due to the digital oscilloscope, because on the Tektronix analog oscilloscope, the waveform is smooth. In Figure 6.7b, the oscillations on the drain during switch-off are noticeable. These oscillations decay
rapidly after the initial peak as previously discussed. The gate-to-source voltage waveform is shown in Figure 6.8. An expanded view during switching is given in Figures 6.9a and 6.9b.

**Figure 6.8:** Gate-to-Source Voltage for a Resistive Load.

**Figure 6.9:** Expanded View of Gate-to-Source Voltage During Switching.
Figure 6.8 shows the voltage waveform on the gate-to-source junction encountered during the test. Note from the figure that this waveform is quite noisy. Voltage spikes are encountered both during switch-on and switch-off. It was found that the spikes during switch-off could be eliminated by placing a small capacitor (0.01 \( \mu \)F) across each gate-to-source junction. This method was not used however, because it slowed down the switch-off times and increased the switching energy. There was no effective method found to control the spikes encountered during switch-on.

Figures 6.9a and 6.9b show the above waveform in an expanded profile during the switching transitions. From 6.9a, it can be seen that the effective sink voltage of the gate driver waveform is about 1.5 volts. This voltage is apparent at the beginning of the cycle. From Figure 6.9b, it can be seen that at the beginning of switch-off the gate is charged up to about 9.5 volts. This is the effective source voltage of the gate driver. These source and sink voltages are slightly different than the approximate 10.5 volt and 0.5 volt source and sink voltage that is applied to the gate driver from the PWM circuit, as seen in Figure 6.4. This difference is accounted for in the losses of the gate driver circuit.

Both the time delays, \( t_{d(on)} \) and \( t_{d(off)} \), that were discussed in previous chapters can be seen when comparing Figures 6.9a and 6.9b with Figures 6.7a and 6.7b. In Figure 6.9a, the drain voltage begins to fall at about -1.25 microseconds, while the gate voltage begins to rise at about -1.5 microseconds in Figure 6.9a. This small difference is the time delay-on, \( t_{d(on)} \). It is small because the threshold voltage, when the device begins to turn on, is only about 3 volts for the MTY55N20E. The time delay-off, \( t_{d(off)} \), is larger,
about 2 microseconds. In Figure 6.7b, the drain voltage begins to rise at about 17 microseconds, while, in Figure 6.9b, the gate charge has started to be removed at about 15 microseconds. This longer time is to be expected, since the gate voltage has to fall below the plateau voltage before the drain voltage can begin to fall.

Also notice in Figures 6.9a and 6.9b that the shape of the gate-to-source voltage waveforms are of the general shape that was given in the gate-to-source voltage versus time and gate charge waveform in Figure 3.2. Using Equations 3.1 and 3.2, the theoretical rise and fall times of the drain-to-source voltage can be calculated. Using a gate-to-drain charge of 128 nC as given in the data sheets for the MTY55N20E power MOSFET in Appendix A.1, and gate driver resistances $R_{\text{eff(\text{on})}}$ and $R_{\text{eff(\text{off})}}$ of 44 ohms and 28 ohms respectively, the corresponding times are calculated as

$$t_f = \frac{Q_{gd}R_{\text{eff(\text{on})}}}{V_{\text{source}} - V_{\text{gsp}}}$$

$$= \frac{128\text{nC} \times 44\Omega}{9.5\text{V} - 5\text{V}}$$

$$= 1.25 \text{ \mu S}$$

and
\[ t_{rv} = \frac{Q_{gd} R_{eff}}{V_{gs} - V_{sink}} \]

\[ = \frac{128nC \times 28\Omega}{5v - 1.5v} \]

\[ = 1.02 \ \mu S. \]

The voltage fall time of 1.25 microseconds and the voltage rise time of 1.02 microseconds calculated in Equations 6.1 and 6.2 are verified in Figures 6.7a and 6.7b. Therefore, the equations given in Chapter 3 provide a usable model for switching characteristics during the design phase. The equations which give the switching times for the current are not verified since it was not possible to obtain an accurate current waveform profile.

After this test was completed, the final steady-state temperatures of the MOSFET heat sinks were 81°F and 80 °F for the left and right modules respectively. This small imbalance shows that the modules are well matched. The flyback diode heat sinks both remained at 72 °F since the flyback diodes are not used for a resistive load.

If it is assumed that the average drain current measured was a square wave, the rms value of the load current can be found. This is given as

\[ I_{rms} = \frac{\sqrt{D} I_{avg}}{D} \]

\[ = 48.2 \ \text{amps}. \]

The rms load voltage is given by Equation (2.14). Thus, the rms load voltage is
\[ V_{\text{rms}} = \sqrt{D}V_s \]
\[ = \sqrt{0.3} \times 120 \]
\[ = 65.7 \text{ volts.} \]

The power delivered to the resistive load, as it was measured, can be calculated as

\[ P_L = V_{\text{rms}} I_{\text{rms}} \]
\[ = 65.7 \times 48.2a \]
\[ = 3.17 \text{ kW.} \]

This load power is different than the 3.69 kW obtained using Equation (2.16) which gives the load power for a resistive load. This is probably due to the accuracy of the measured current value obtained with the Fluke multimeter. Using the measured value of the heat sink temperatures at the end of the test, it is possible to obtain the power lost in the controller by using Equation (4.16) for the two heat sinks. After the temperatures are converted to centigrade, this is given by

\[ Q = \frac{2(T_S - T_A)}{R_{\theta SA}} \]
\[ = \frac{2(26.94^\circ C - 22.22^\circ C)}{0.309^\circ C/W} \]
\[ = 30.6 \text{ W.} \]

The efficiency of the controller for this test can then be calculated as
\[
\text{efficiency} = \frac{P_L}{P_L + Q} \times 100\% \\
= \frac{3.17kW}{3.17kW + 30.6W} \times 100\% \\
= 99.0\%.
\] (6.7)

An efficiency of 99% percent is excellent. The efficiency will be lower for an inductive load due to noisier operation. It will also be lower for heavier currents because of increased conduction losses from an increased junction temperature, but Equation (6.7) shows that the controller is about as efficient as can be expected.

### 6.6 Inductive Load Test

The controller was next tested with an inductive load. The motor load was a General Electric DC motor configured as a separately excited motor. Since loading the motor was difficult in the laboratory setting used, the heating elements used in the resistive load test were also used in parallel with the motor. This combination created a large load on the controller with the same effects that a motor load would produce by itself.

For this test, a battery voltage of 120 volts was used. The starting temperature of the heat sinks was 72 °F. The average load current was 150 amps and the average battery current was 34 amps. A small duty cycle was used to keep the rpm's of the motor at a safe level throughout the test. The time axis in Figures 6.10 through 6.13 is synchronized.
Figure 6.10: Drain-to-Source Voltage for an RL Load.

Figure 6.11: Expanded View of Drain-to-Source Voltages During Switching.
Figure 6.10 shows the drain-to-source waveform obtained for the inductive load test. Figure 6.11 shows the waveform of Figure 6.10 in expanded detail during switching. It is immediately apparent that the waveforms from the inductive load are noisier than those presented for a resistive load. The overshoot of the battery voltage during switch-off is about the same in Figures 6.9 and 6.10b as it is in Figures 6.6 and 6.7b for the resistive load. It is difficult to measure the true magnitude of this peak, but during switch-off, the Tektronix analog oscilloscope displays a peak that is 25% - 30% above the battery voltage. In general, an inductive load produces a slightly larger overshoot than a resistive load. This is a larger peak than the Hewlett-Packard oscilloscope, used to collect the data for these figures, shows. In the inductive load case, the peak also takes longer to decay out than for the resistive load case.

During switch-on, in Figure 6.11a, the drain-to-source junction experiences a lot of noise. This noise is not harmful because it is not of a damaging magnitude. Much effort was spent trying to filter this noise without success.

Figure 6.12 shows the gate-to-source voltage waveform of the controller. Figure 6.13 shows the same waveform in a slightly expanded view. The noise on the gate-to-source junction corresponds to the noise on the drain-to-source junction. An effort was made to reduce the noise on the gate without any success. It is the belief of the author that the source of the noise on the drain is a result of the load inductance, and this noise is then coupled to the gate junction through the gate-to-drain capacitance. The peaks of the noise exceed the gate drive voltage and extend below the gate driver’s sink voltage and ground. They will not damage the controller unless the magnitude of the spikes exceeds
Figure 6.12: Gate-to-Source Voltage for an Inductive Load.

Figure 6.13: Expanded view of the Gate-to-Source Voltage for an Inductive Load.
the ±20 volt rating of the gate-to-source junction.

At the conclusion of the test, the heat sink temperatures were 103 °F and 100 °F for the left and right MOSFET heat sinks, and 82 °F and 80 °F for the flyback diode heat sinks respectively.

A nearly identical controller to the one used for the tests in Figures 6.2 through 6.13 was also constructed. The difference was that each heat sink held 5 MOSFETs instead of 7, for a total of 10 MOSFETs. This controller was subjected to the same tests and fared just as well. It was also tested successfully with a battery input voltage of 144 volts and a load current of 165 amps on the same inductive load. During another inductive load test on this controller, the load current was maintained near 300 amps for about one minute. All indications were that the controller was capable of operating at this high current level in steady state. It was not tested for a longer duration due to the difficulty in providing a sufficient load to maintain the motor rpm's at a safe level.

6.7 Controller Failures

Several controller designs were discarded along the way to obtaining the successful controller design presented in Chapter 5 and Sections 6.1 through 6.6. The first controller design had the flyback diodes attached directly to the motor as opposed to being mounted very close to the paralleled MOSFETs. It was found that this layout had an unacceptable amount of noise associated with it.

A subsequent design corrected this problem, but retained the same gate driver design which was not constructed on a printed circuit board. The design contained 10
MOSFETs in parallel with 3 flyback diodes. This controller worked when tested at 48 volts with a series dc motor from a golf cart, but was quite noisy. When it was tested on the large inductive load, the MOSFETs sounded like firecrackers as they were being destroyed. The direct cause of this controller failure was the failure of one leg of one flyback diode. The diode had experienced voltage transients that exceeded its blocking voltage of 200 volts. When it failed, it short circuited. At the beginning of the next cycle when the MOSFETs turned on, they essentially shorted out the 120 volt battery input causing a massive amount of current to flow in the drain junction. An indirect cause of this failure was the controller's noisy operation, which induced the transients on the flyback diodes. The possibility of damaging high voltage transients was reduced on subsequent designs by replacing the flyback diodes with ones rated to 400 volts.

The next phase of designs utilized printed circuit boards to reduce noise. Several models were constructed with different layouts and different numbers of components until a controller was built with 5 MOSFETs per heat sink, but otherwise identical to the controller in Chapter 5. This controller worked very well and its tests are discussed in Section 6.6. It was tested in the Electric Bobcat on a test track with a battery input voltage of 120 volts. Unfortunately, the driver was instructed to use an improper gear sequence in the car's transmission. As a result, the driver started in 2nd gear and then shifted to 1st gear. After a short time in 1st, the driver upshifted to 4th gear. The driver believed he was upshifting from 2nd to 3rd gear. Shifting into 4th gear at such a slow speed caused the motor to draw an enormous amount of current. This massive overcurrent condition caused the controller to fail because it vaporized the source leads
on most of the MOSFETs while melting the rest. Because of this, the design was changed to have seven MOSFETs per heat sink to help make the controller more rugged to overcurrent conditions. Photographs of this destroyed controller are included in Appendix B.

During testing of the controller to obtain the results presented in this chapter, one of the 20 volt Zener diodes protecting the gate junction of a MOSFET failed to a short circuit. This did not damage the controller, but leads to the conclusion that the 1 watt Zener diodes now in use should be replaced with 5 watt components.
7 Conclusions and Recommendations

The work presented in this thesis was an effort to develop a better speed controller for the Electric Bobcat that would allow the program to achieve even more success. Chapter 1 presented the Electric Bobcat program at Ohio University and expressed the need to develop a higher power speed controller. Chapter 2 derived and presented the theory of buck chopper drives and reduced the theory to a few important equations. Buck chopper drives of step-down dc-dc converters were determined to be the most efficient form of speed control of the series dc motor used in the Electric Bobcat. Power MOSFETs were chosen as the semiconductor switch of the controller in Chapter 3. Chapter 3 also covered the switching theory and protection requirements of power MOSFETs. Additionally, flyback diodes were chosen and their characteristics were presented in this chapter. Chapter 4 covers the losses that are encountered in a buck chopper drive. These losses stem from both switching and conduction losses. Since these losses are in the form of heat, heat sink theory was presented in this chapter as well. Chapter 5 brought the theory of the preceding chapters together with the design, layout, and construction of the controller. The control circuitry of the controller was also
presented in this chapter. In Chapter 6, the experimental results of testing the controller were given. Measurement instruments and techniques were first presented. Results were then presented for both a resistive and an inductive load. Switching time calculations were made to verify some of the material in Chapter 3 and efficiency calculations were made to show that the controller is 99% efficient in certain cases.

The controller presented in this thesis worked well in the laboratory, but is not ready for the task it was designed for - to propel the Electric Bobcat to victory. For that, some enhancements need to be made. The paralleling of two modules increases the current limit of the controller, but four paralleled models would be necessary for the Electric Bobcat to run at 600-800 amps for a sustained time. The control circuitry is designed to run four modules, so this change would only involve the construction of two additional modules.

The biggest potential problem with this controller is current limiting. An acceptable method of current limiting was not found since a good way to measure the current waveforms was not found. The control circuitry is capable of implementing the desired cycle-by-cycle current limiting if it is provided with a high quality current reading. The use of Hall effect current sensors may give the desired results.

To improve the switching performance of the controller, the supply voltage to the gate driver should be increased from 12 to 15 volts. This would raise the gate driver's output voltage up to the desired 12 volts. A method to reduce the sink voltage to ground potential would also be desirable to insure that the controller fully turns off. Switching losses which account for the largest majority of the controller losses can be reduced by
replacing the PWM frequency set resistor with a potentiometer. In a race situation, the controller is operating at or near full throttle the majority of the time. If the PWM frequency was decreased, the controller would not suffer a loss in performance, and the reduction of frequency would directly translate to an increased efficiency. For public events, where efficiency is not critical, the frequency could be set back to about 15 kHz so the controller would operate silently.

Looking at the design of buck chopper circuits, they are very simple circuits consisting essentially of a semiconductor switch and a diode. Unfortunately, they are very complicated to build in a working and efficient mode for high power levels. This is because there is currently no easy and effective way to model the transient behavior of the power MOSFETs that were used as the switch for this controller. The switching behavior of these devices is alluded to in the gate charge data. This data provides an incomplete picture in a larger circuit where multiple devices are paralleled. This was seen in the noisy gate-to-source voltage waveforms in Chapter 6. Also, the voltage transients which can fatally damage the controller, change with load. This is also not easily predicted with the data available in the data sheet. In order to create an accurate model of the transient behavior of a MOSFET it is necessary to know details of a proprietary nature. Since it is unlikely that this information will be available, a model could be created from switching data measured in a controlled application in a laboratory over a range of load and gate drive conditions. This data could be used in conjunction with the SPICE model of a MOSFET to give an accurate model. Another approach could use this data to create a state space model of the MOSFET switching characteristics for
implementation either in MATLAB™ or as an electrical equivalent in SPICE. With a readily available device model, the chopper could be designed and built in much less time. In addition to the time factor, the components could be used closer to their operating limits since an accurate circuit simulation could be performed during the design phase. This would allow for a more economical use of high power components.

Overall, the controller presented in this thesis is efficient, works at the required voltage levels and current levels, is capable of a 0 - 100% duty cycle, and has a delay ramp to protect the controller. It is also air cooled, it is electrically isolated from the chassis of the car, has the capacity for current limiting, and has a modular design to change the controller's power rating to the current application.
References


Appendix A: Data Sheets
A.1: MTY55N20E

MOTOROLA
SEMICONDUCTOR TECHNICAL DATA

Designer’s™ Data Sheet
TMOS E-FET™
Power Field Effect Transistor
N–Channel Enhancement–Mode Silicon Gate

This advanced TMOS power FET is designed to withstand high energy in the avalanche and commutation modes. This new energy efficient design also offers a drain–to–source diode with fast recovery time. Designed for high voltage, high speed switching applications in power supplies, converters, PWM motor controls, and other inductive loads. The avalanche energy capability is specified to eliminate the guesswork in designs where inductive loads are switched and offer additional safety margin against unexpected voltage transients.

- Avalanche Energy Specified
- Diode is Characterized for Use in Bridge Circuits
- VDS(on) and VDS(on) Specified at Elevated Temperature

MAXIMUM RATINGS (TJ = 25°C unless otherwise noted)

<table>
<thead>
<tr>
<th>Rating</th>
<th>Symbol</th>
<th>Value</th>
<th>Unit</th>
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<tr>
<td>Drain-Source Voltage</td>
<td>VDSS</td>
<td>200</td>
<td>Vdc</td>
</tr>
<tr>
<td>Drain-Source Voltage (RDS(on) = 1 mΩ)</td>
<td>VDSR</td>
<td>200</td>
<td>Vdc</td>
</tr>
<tr>
<td>Gate-Source Voltage — Continuous</td>
<td>VGS</td>
<td>±2.0</td>
<td>Vdc</td>
</tr>
<tr>
<td>Gate-Source Voltage — Non-Repetitive (Vd ≤ 10 ns)</td>
<td>VGSN</td>
<td>±4.0</td>
<td>Vdc</td>
</tr>
<tr>
<td>Drain Current — Continuous @ TJ = 25°C</td>
<td>ID</td>
<td>55</td>
<td>Adc</td>
</tr>
<tr>
<td>Drain Current — Single Pulse (Vd ≤ 10 μs)</td>
<td>IDP</td>
<td>160</td>
<td>Apc</td>
</tr>
<tr>
<td>Total Power Dissipation Derate above 25°C</td>
<td>PD</td>
<td>300</td>
<td>Watts</td>
</tr>
<tr>
<td>Operating and Storage Temperature Range</td>
<td>Tstg</td>
<td>-65 to 150</td>
<td>°C</td>
</tr>
<tr>
<td>Single Pulse Drain–to–Source Avalanche Energy — Starting TJ = 25°C (VGS = 85 Vdc, VDSR = 15 Vdc, Peak ID = 110 Apc, L = 0.3 mH, Rg = 25 Ω)</td>
<td>EAS</td>
<td>3000</td>
<td>mJ</td>
</tr>
<tr>
<td>Thermal Resistance — Junction to Case</td>
<td>RJC</td>
<td>0.42</td>
<td>°C/W</td>
</tr>
<tr>
<td>Thermal Resistance — Junction to Ambient</td>
<td>RJA</td>
<td>40</td>
<td>°C/W</td>
</tr>
<tr>
<td>Maximum Lead Temperature for soldering purposes, 1/8” from case for 10 seconds</td>
<td>T1</td>
<td>260</td>
<td>°C</td>
</tr>
</tbody>
</table>

Designer’s Data for “Worst Case” Conditions — The Designer’s Data Sheet permits the design of most circuits orderly from the information presented. SCU Limit curves — representing boundaries on device characteristics — are given to facilitate “worst case” design.

Preferred devices are Motorola recommended choices for future use and best overall value.

REV 1

11–20 Motorola TMOS Power MOSFET Transistor Device Data
### ELECTRICAL CHARACTERISTICS (T_{J} = 25^\circ\text{C} unless otherwise noted)

<table>
<thead>
<tr>
<th>Characteristic</th>
<th>Symbol</th>
<th>Min</th>
<th>Typ</th>
<th>Max</th>
<th>Unit</th>
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<tr>
<td>3V3F CHARACTERISTICS</td>
<td></td>
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<tr>
<td>Drain–Source Breakdown Voltage</td>
<td>V_{BR}DSS</td>
<td>200</td>
<td>250</td>
<td>50</td>
<td>Vdc</td>
</tr>
<tr>
<td>Temperature Coefficient (Positive)</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td>mV/°C</td>
</tr>
<tr>
<td>Zero Gate Voltage Drain Current</td>
<td>I_{DSS}</td>
<td></td>
<td>10</td>
<td></td>
<td>μAdc</td>
</tr>
<tr>
<td>(V_{GS} = 200 Vdc, V_{DS} = 0 Vdc)</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>(V_{GS} = 200 Vdc, V_{DS} = 0 Vdc, T_{J} = 125°C)</td>
<td></td>
<td></td>
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<td></td>
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</tr>
<tr>
<td>Gate–Body Leakage Current (V_{GB} = ±20 Vdc, V_{DS} = 0)</td>
<td>I_{GSS}</td>
<td></td>
<td>100</td>
<td></td>
<td>nAdc</td>
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<tr>
<td>2N CHARACTERISTICS (1)</td>
<td></td>
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<tr>
<td>Gate Threshold Voltage</td>
<td>V_{GS(on)}</td>
<td>2</td>
<td>7</td>
<td>4</td>
<td>Vdc</td>
</tr>
<tr>
<td>Threshold Temperature Coefficient (Negative)</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td>mV/°C</td>
</tr>
<tr>
<td>Static Drain–Source On-Resistance (V_{GS} = 10 Vdc, I_{D} = 27.5 Adc)</td>
<td>R_{DS(on)}</td>
<td>0.028</td>
<td></td>
<td></td>
<td>Ωm</td>
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<tr>
<td>Drain–Source On-Voltage</td>
<td>V_{DS(on)}</td>
<td></td>
<td>1.3</td>
<td>1.6</td>
<td>Vdc</td>
</tr>
<tr>
<td>(I_{D} = 55 Adc)</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>(I_{D} = 27.5 Adc, T_{J} = 125°C)</td>
<td></td>
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<td>1.6</td>
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<tr>
<td>Forward Transconductance</td>
<td>g_{FS}</td>
<td>30</td>
<td>37</td>
<td>111</td>
<td>mhos</td>
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<td>DYNAMIC CHARACTERISTICS</td>
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<tr>
<td>Input Capacitance</td>
<td>C_{iss}</td>
<td>7200</td>
<td>10080</td>
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<td>pF</td>
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<tr>
<td>Output Capacitance</td>
<td>C_{oss}</td>
<td>1600</td>
<td>2520</td>
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<tr>
<td>Reverse Transfer Capacitance</td>
<td>C_{rss}</td>
<td>460</td>
<td>920</td>
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<tr>
<td>SWITCHING CHARACTERISTICS (2)</td>
<td></td>
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<tr>
<td>Turn-On Delay Time</td>
<td>t_{on}</td>
<td>33</td>
<td>66</td>
<td>111</td>
<td>ns</td>
</tr>
<tr>
<td>Rise Time</td>
<td></td>
<td>200</td>
<td>400</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Turn-Off Delay Time</td>
<td>t_{off}</td>
<td>150</td>
<td>300</td>
<td></td>
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<tr>
<td>Fall Time</td>
<td>t_{f}</td>
<td>170</td>
<td>340</td>
<td></td>
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</tr>
<tr>
<td>Gate Charge (See Figure 8)</td>
<td>Q_{T}</td>
<td>345</td>
<td>343</td>
<td></td>
<td>nC</td>
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<tr>
<td></td>
<td>Q_{1}</td>
<td>33</td>
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<tr>
<td></td>
<td>Q_{2}</td>
<td>128</td>
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<td></td>
<td>Q_{3}</td>
<td>72</td>
<td></td>
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<tr>
<td>SOURCE–DRAIN DIODE CHARACTERISTICS</td>
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<tr>
<td>Forward On-Voltage</td>
<td>V_{SD}</td>
<td>0.75</td>
<td>1.2</td>
<td></td>
<td>Vdc</td>
</tr>
<tr>
<td>(I_{D} = 55 Adc, V_{GS} = 0 Vdc)</td>
<td></td>
<td>1.1</td>
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<tr>
<td>(I_{G} = 55 Adc, V_{GS} = 0 Vdc, T_{J} = 125°C)</td>
<td></td>
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<tr>
<td>Reverse Recovery Time (See Figure 14)</td>
<td>t_{rr}</td>
<td>310</td>
<td></td>
<td></td>
<td>ns</td>
</tr>
<tr>
<td></td>
<td>t_{a}</td>
<td>220</td>
<td></td>
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<td></td>
<td>t_{b}</td>
<td>90</td>
<td></td>
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<tr>
<td>Reverse Recovery Stored Charge</td>
<td>Q_{rr}</td>
<td>4.6</td>
<td></td>
<td></td>
<td>μC</td>
</tr>
<tr>
<td>INTERNAL PACKAGE INDUCTANCE</td>
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<td></td>
</tr>
<tr>
<td>Internal Drift Inductance (Measured from the drain lead 0.25&quot; from package to center of die)</td>
<td>L_{D}</td>
<td>4.5</td>
<td></td>
<td></td>
<td>nH</td>
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<tr>
<td>Internal Source Inductance (Measured from the source lead 0.25&quot; from package to source bond pad)</td>
<td>L_{S}</td>
<td>13</td>
<td></td>
<td></td>
<td>nH</td>
</tr>
</tbody>
</table>

(1) Pulse Test: Pulse Width ≤ 300 μs, Duty Cycle ≤ 2%.  
(2) Switching characteristics are independent of operating junction temperature.
POWER MOSFET SWITCHING

Switching behavior is most easily modeled and predicted by recognizing that the power MOSFET is charge controlled. The lengths of various switching intervals ($\Delta t$) are determined by how fast the FET input capacitance can be charged by current from the generator.

The published capacitance data is difficult to use for calculating rise and fall times because drain–gate capacitance varies greatly with applied voltage. Accordingly, gate charge data is used. In most cases, a satisfactory estimate of average input current ($I_g (AV)$) can be made from a rudimentary analysis of the drive circuit so that

\[ I = Q_l / (G_{AV}) \]

During the rise and fall time interval when switching a resistive load, $V_{GS}$ remains virtually constant at a level known as the plateau voltage, $V_{GSP}$. Therefore, rise and fall times may be approximated by the following:

\[ t_r = Q_l x R_g / (V_{GG} - V_{GSP}) \]

\[ t_f = Q_l x R_g / V_{GSP} \]

where

$V_{GG}$ = the gate drive voltage, which varies from zero to $V_{GG}$

$R_g$ = the gate drive resistance

and $Q_l$ and $V_{GSP}$ are read from the gate charge curve.

During the turn-on and turn-off delay times, gate current is not constant. The simplest calculation uses appropriate values from the capacitance curves in a standard equation for voltage change in an RC network. The equations are:

\[ t_{(on)} = R_g \times C_{iss} \ln \left( \frac{V_{GG} - V_{GSP}}{V_{GGS}} \right) \]

\[ t_{(off)} = R_g \times C_{iss} \ln \left( \frac{V_{GGS}}{V_{GG} - V_{GSP}} \right) \]

The capacitance ($C_{iss}$) is read from the capacitance curve at a voltage corresponding to the off-state condition when calculating $t_{(on)}$ and is read at a voltage corresponding to the on-state when calculating $t_{(off)}$.

At high switching speeds, parasitic circuit elements complicate the analysis. The inductance of the MOSFET source lead, inside the package and in the circuit wiring which is common to both the drain and gate current paths, produces a voltage at the source which reduces the gate drive current. The voltage is determined by $L/\frac{dI}{dt}$, but since $dI/dt$ is a function of drain current, the mathematical solution is complex. The MOSFET output capacitance also complicates the mathematics. And finally, MOSFETs have finite internal gate resistance which effectively adds to the resistance of the driving source, but the internal resistance is difficult to measure and, consequently, is not specified.

The resistive switching time variation versus gate resistance (Figure 9) shows how typical switching performance is affected by the parasitic circuit elements. If the parasitics were not present, the slope of the curves would maintain a value of unity regardless of the switching speed. The circuit used to obtain the data is constructed to minimize common inductance in the drain and gate circuit loops and is believed readily achievable with board mounted components. Most power electronic loads are inductive; the data in the figure is taken with a resistive load, which approximates an optimally snubbed inductive load. Power MOSFETs may be safely operated into an inductive load; however, snubbing reduces switching losses.

![Figure 7: Capacitance Variation](image)
The Forward Biased Safe Operating Area curves define the maximum simultaneous drain-to-source voltage and drain current that a transistor can handle safely when it is forward biased. Curves are based upon maximum peak junction temperature and a case temperature (Tc) of 25°C. Peak repetitive pulsed power limits are determined by using the thermal response data in conjunction with the procedures discussed in AN569, "Transient Thermal Resistance—General Data and Its Use."

Switching between the off-state and the on-state may traverse any load line provided neither rated peak current (I(DM)) nor rated voltage (V(DSS)) is exceeded and the transition time (t(s)) do not exceed 10 μs. In addition the total power averaged over a complete switching cycle must not exceed (TJ(MAX) - TJ)/T(RMS).

A Power MOSFET designated E-FET can be safely used in switching circuits with unclamped inductive loads. For reliable operation, the stored energy from circuit inductance dissipated in the transistor while in avalanche must be less than the rated limit and adjusted for operating conditions differing from those specified. Although industry practice is to rate in terms of energy, avalanche energy capability is not a constant. The energy rating decreases non-linearly with an increase of peak current in avalanche and peak junction temperature.

Although many E-FETs can withstand the stress of drain-to-source avalanche at currents up to rated pulsed current (I(DM)), the energy rating is specified at rated continuous current (Ig), in accordance with industry custom. The energy rating must be derated for temperature as shown in the accompanying graph (Figure 12). Maximum energy at currents below rated continuous Ig can safely be assumed to equal the values indicated.
SAFE OPERATING AREA

Figure 11. Maximum Rated Forward Biased Safe Operating Area

Figure 12. Maximum Avalanche Energy versus Starting Junction Temperature

Figure 13. Thermal Response

Figure 14. Diode Reverse Recovery Waveform
A.2: MURP20040CT

Preliminary Data Sheet

POWERTAP II
Ultrafast
SWITCHMODE Power Rectifier

...designed for use in switching power supplies, inverters, and as free wheeling diodes. This state-of-the-art device has the following features:

- Dual Diode Construction
- Low Leakage Current
- Low Forward Voltage
- 175°C Operating Junction Temperature
- Labor Saving PowerTape Package

Mechanical Characteristics:

- Case: Epoxy, Molded with metal heatshrink base
- Weight: 80 gram (approximately)
- Finish: All External (Surface Corrosion Resistant)
- Top Terminal Torque: 25–40 lb-in max
- Base Plate Torque: See procedure given in the Package Outline Section

MAXIMUM RATINGS

<table>
<thead>
<tr>
<th>Rating</th>
<th>Symbol</th>
<th>MURP20020CT</th>
<th>MURP20040CT</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Peak Repetitive Reverse Voltage</td>
<td>VRRM</td>
<td>200</td>
<td>400</td>
<td>Volts</td>
</tr>
<tr>
<td>Working Peak Reverse Voltage</td>
<td>VRRM</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>DC Blocking Voltage</td>
<td>VBR</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Average Rectified Forward Current</td>
<td>I(FRM)</td>
<td>200 (Tc = 130°C)</td>
<td>200 (Tc = 100°C)</td>
<td>Amps</td>
</tr>
<tr>
<td>(Rated VBR) Per Device Per Leg</td>
<td></td>
<td>100 (Tc = 130°C)</td>
<td>100 (Tc = 100°C)</td>
<td></td>
</tr>
<tr>
<td>Peak Repetitive Forward Current, Per Leg</td>
<td>I(FRM)</td>
<td>200</td>
<td>200</td>
<td>Amps</td>
</tr>
<tr>
<td>(Rated VBR, Square Wave, 20 kHz, Tc = 55°C)</td>
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<tr>
<td>Non-repetitive Peak Surge Current</td>
<td>TDM</td>
<td>600</td>
<td>800</td>
<td>Amps</td>
</tr>
<tr>
<td>Per Leg (Surge applied at rated load conditions)</td>
<td></td>
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</tr>
<tr>
<td>Operating Junction Temperature</td>
<td>Tj</td>
<td>-55 to +175</td>
<td>-55 to +175</td>
<td>°C</td>
</tr>
<tr>
<td>Storage Temperature</td>
<td>Tst</td>
<td>-55 to +150</td>
<td>-55 to +150</td>
<td>°C</td>
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THERMAL CHARACTERISTICS PER LEG

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<thead>
<tr>
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<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Thermal Resistance, Junction to Case</td>
<td>RJC</td>
<td>0.45</td>
<td>0.45</td>
<td>°C/W</td>
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ELECTRICAL CHARACTERISTICS PER LEG

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<thead>
<tr>
<th>Instantaneous Forward Voltage (1)</th>
<th>Symbol</th>
<th>Tj, Tc = 125°C</th>
<th>Tj, Tc = 175°C</th>
<th>Tj, Tc = 25°C</th>
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</thead>
<tbody>
<tr>
<td>(Ig = 100 Amp, Tj = 125°C)</td>
<td></td>
<td>0.95</td>
<td>1.10</td>
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<tr>
<td>(Ig = 100 Amp, Tj = 175°C)</td>
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<td>0.85</td>
<td>1.10</td>
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<tr>
<td>(Ig = 100 Amp, Tj = 25°C)</td>
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<td>0.75</td>
<td>1.10</td>
<td>1.30</td>
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</table>

<table>
<thead>
<tr>
<th>Instantaneous Reverse Current (2)</th>
<th>Symbol</th>
<th>Tj, Tc = 125°C</th>
<th>Tj, Tc = 175°C</th>
<th>Tj, Tc = 25°C</th>
</tr>
</thead>
<tbody>
<tr>
<td>(Rated dc Voltage, Tj = 125°C)</td>
<td></td>
<td>1000</td>
<td>500</td>
<td>50</td>
</tr>
<tr>
<td>(Rated dc Voltage, Tj = 175°C)</td>
<td></td>
<td>150</td>
<td>75</td>
<td>75</td>
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<tr>
<td>Maximum Reverse Recovery Time</td>
<td>tTR</td>
<td>50</td>
<td>75</td>
<td>ns</td>
</tr>
</tbody>
</table>

(1) Pulse Test: Pulse Width = 300 μs, Duty Cycle ≤ 2.0%.

Prefered devices are Motorola recommended choices for future use and best overall value.

Rev 2

4-116 Rectifier Devices Data
PACKAGE OUTLINE DIMENSIONS AND FOOTPRINTS (continued)

MOUNTING PROCEDURE

The POWERTAP package requires special mounting considerations because of the long longitudinal axis of the copper heat sink. It is important to follow the proper tightening sequence to avoid warping the heat sink which can reduce thermal contact between the POWERTAP and heat sink.

STEP 1:
Locate the POWERTAP on the heat sink and start mounting bolts into the threads by hand (2 or 3 turns).

STEP 2:
Finger tighten the center bolt. The bolt may catch on the threads of the heat sink so it is important to make sure the face of the bolt or washer is in contact with the surface of the POWERTAP.

STEP 3:
Tighten each of the end bolts between 5 to 10 in-lb.

STEP 4:
Tighten the center bolt between 8 to 10 in-lb.

STEP 5:
Finally, tighten the end bolts between 30 to 40 in-lb.
Advance Information

**BRUSHLESS DC MOTOR CONTROLLER**

The MC33035 is a high performance second generation monolithic brushless DC motor controller containing all of the active functions required to implement a full featured open-loop, three or four phase motor control system. This device consists of a rotor position decoder for proper commutation sequencing, temperature compensated reference capable of supplying sensor power, frequency programmable sawtooth oscillator, fully accessible error amplifier, pulse width modulator comparator, three open collector top drivers, and three high current totem pole bottom drivers ideally suited for driving power MOSFETs.

Also included are protective features consisting of undervoltage lockout, cycle-by-cycle current limiting with a selectable time delayed latched shutdown mode, internal thermal shutdown, and a unique fault output that can be interfaced into microprocessor controlled systems. Typical motor control functions include open-loop speed, forward or reverse direction, run enable, and dynamic braking. The MC33035 is designed to operate with electrical sensor phasings of 60°/120° or 120°/240°, and can also efficiently control brush DC motors.

- 10 V to 30 V Operation
- Undervoltage Lockout
- 6.25 V Reference Capable of Supplying Sensor Power
- Fully Accessible Error Amplifier for Closed-Loop Servo Applications
- High Current Drivers can Control MPM303 MOSFET 3-Phase Bridge
- Cycle-By-Cycle Current Limiting
- Pin-Triggered Current Sense Reference
- Internal Thermal Shutdown
- Selectable 60°/120° or 120°/240° Sensor Phasings
- Can Efficiently Control Brush DC Motors with MPM3002 MOSFET H-Bridge

**PIN CONNECTIONS**

- Top Drive Output (A1)
- Brake (A9)
- Sensor Inputs (A4, B3, C3)
- Output Enable (B4)
- Reference Output (B5)
- Non-Inverting Input (C2, D3)
- Oscillator Non-Inverting Input (D2)
- Error Amp Inverting Input (D1)
- Error Amp Output (E2)
- PWM Input (Top View)

**ORDERING INFORMATION**

<table>
<thead>
<tr>
<th>Device</th>
<th>Operating Ambient Temperature Range</th>
<th>Package</th>
</tr>
</thead>
<tbody>
<tr>
<td>MC33035P</td>
<td>-40°C to +85°C</td>
<td>Plastic DIP</td>
</tr>
<tr>
<td>MC33035DW</td>
<td>-40°C to +85°C</td>
<td>SO-24L</td>
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# MC33035

## Maximum Ratings

<table>
<thead>
<tr>
<th>Rating</th>
<th>Symbol</th>
<th>Value</th>
<th>Unit</th>
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<tbody>
<tr>
<td>Power Supply Voltage</td>
<td>$V_{CC}$</td>
<td>40</td>
<td>V</td>
</tr>
<tr>
<td>Digital Input (Pins 3, 4, 5, 6, 22, 23)</td>
<td>$V_{IL}$</td>
<td>0.8</td>
<td>V</td>
</tr>
<tr>
<td>Oscillator Input Current (Source or Sink)</td>
<td>$I_{OSC}$</td>
<td>30</td>
<td>mA</td>
</tr>
<tr>
<td>Error Amp Input Voltage Range (Pins 11, 12, Note 1)</td>
<td>$V_{IN}$</td>
<td>-3.0 to 6.0</td>
<td>V</td>
</tr>
<tr>
<td>Error Amp Output Current (Source or Sink, Note 2)</td>
<td>$I_{OUT}$</td>
<td>10</td>
<td>mA</td>
</tr>
<tr>
<td>Current Sense Input Voltage Range</td>
<td>$V_{SENSE}$</td>
<td>-0.3 to 5.0</td>
<td>V</td>
</tr>
<tr>
<td>Fault Output Voltage</td>
<td>$V_{OE}$</td>
<td>20</td>
<td>V</td>
</tr>
<tr>
<td>Fault Output Sink Current</td>
<td>$I_{OE}$</td>
<td>20</td>
<td>mA</td>
</tr>
<tr>
<td>Top Drive Voltage (Pins 1, 2, 24)</td>
<td>$V_{CC}(top)$</td>
<td>40</td>
<td>V</td>
</tr>
<tr>
<td>Top Drive Sink Current</td>
<td>$I_{OE}(top)$</td>
<td>50</td>
<td>mA</td>
</tr>
<tr>
<td>Bottom Drive Supply Voltage (Pin 18)</td>
<td>$V_{C}$</td>
<td>30</td>
<td>V</td>
</tr>
<tr>
<td>Bottom Drive Output Current</td>
<td>$I_{O}$</td>
<td>100</td>
<td>mA</td>
</tr>
<tr>
<td>Power Dissipation and Thermal Characteristics</td>
<td>$P_{D}$</td>
<td>867</td>
<td>mW</td>
</tr>
<tr>
<td>Thermal Resistance, Junction to Air</td>
<td>$R_{JJA}$</td>
<td>75</td>
<td>°C</td>
</tr>
<tr>
<td>Operating Junction Temperature</td>
<td>$T_{J}$</td>
<td>150</td>
<td>°C</td>
</tr>
<tr>
<td>Operating Ambient Temperature Range</td>
<td>$T_{A}$</td>
<td>-40 to +85</td>
<td>°C</td>
</tr>
<tr>
<td>Storage Temperature Range</td>
<td>$T_{ST}$</td>
<td>-65 to +150</td>
<td>°C</td>
</tr>
</tbody>
</table>

## Electrical Characteristics ($V_{CC} = V_{CC} = 20$ V, $R_{F} = 4.7$ kΩ, $C_{F} = 10$ nF, $T_{A} = 25$°C unless otherwise noted)

<table>
<thead>
<tr>
<th>Characteristic</th>
<th>Symbol</th>
<th>Min</th>
<th>Typ</th>
<th>Max</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Reference Output Voltage (I$_{ref} = 1.0$ mA)</td>
<td>$V_{ref}$</td>
<td>5.9</td>
<td>6.24</td>
<td>6.5</td>
<td>V</td>
</tr>
<tr>
<td>Line Regulation ($V_{CC} = 10$ V to 30 V, I$_{ref} = 1.0$ mA)</td>
<td>$R_{line}$</td>
<td>1.5</td>
<td>30</td>
<td></td>
<td>mV</td>
</tr>
<tr>
<td>Load Regulation ($I_{ref} = 1.0$ mA to 20 mA)</td>
<td>$R_{load}$</td>
<td>16</td>
<td>30</td>
<td></td>
<td>mV</td>
</tr>
<tr>
<td>Reference Under Voltage Lockout Threshold</td>
<td>$V_{OH}$</td>
<td>4.0</td>
<td>4.5</td>
<td>5.0</td>
<td>V</td>
</tr>
</tbody>
</table>

## Error Amplifier

<table>
<thead>
<tr>
<th>Characteristic</th>
<th>Symbol</th>
<th>Min</th>
<th>Typ</th>
<th>Max</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input Offset Voltage ($T_{A} = -40$°C to +85°C)</td>
<td>$V_{IO}$</td>
<td>0.4</td>
<td>10</td>
<td></td>
<td>mV</td>
</tr>
<tr>
<td>Input Offset Current ($T_{A} = -40$°C to +85°C)</td>
<td>$I_{IO}$</td>
<td>8.0</td>
<td>500</td>
<td>nA</td>
<td></td>
</tr>
<tr>
<td>Input Bias Current ($T_{A} = -40$°C to +85°C)</td>
<td>$I_{IB}$</td>
<td>-46</td>
<td>-1000</td>
<td>nA</td>
<td></td>
</tr>
<tr>
<td>Input Common Mode Voltage Range</td>
<td>$V_{CMR}$</td>
<td>0 V to $V_{CC}$</td>
<td>V</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Open-Loop Voltage Gain ($V_{D} = 3.0$ V, $R_{F} = 15$ kΩ)</td>
<td>$A_{VOL}$</td>
<td>70</td>
<td>80</td>
<td></td>
<td>dB</td>
</tr>
<tr>
<td>Input Common Mode Rejection Ratio</td>
<td>$CMRR$</td>
<td>65</td>
<td>86</td>
<td></td>
<td>dB</td>
</tr>
<tr>
<td>Power Supply Rejection Ratio ($V_{CC} = V_{CC} = 10$ V to 30 V)</td>
<td>$PSRR$</td>
<td>65</td>
<td>105</td>
<td></td>
<td>dB</td>
</tr>
<tr>
<td>Output Voltage Swing</td>
<td>$V_{OH}$</td>
<td>4.6</td>
<td>5.3</td>
<td></td>
<td>V</td>
</tr>
<tr>
<td>Low State ($R_{F} = 15$ kΩ to Ground)</td>
<td>$V_{OL}$</td>
<td>0.6</td>
<td>1.0</td>
<td></td>
<td>V</td>
</tr>
</tbody>
</table>

## Notes:

1. The input common mode voltage or input signal voltage should not be allowed to go negative by more than 0.3 V.
2. The compliance voltage must not exceed the range of -0.3 to $V_{ref}$.
3. Maximum package power dissipation limits must be observed.
### Electrical Characteristics (continued) \( V_{CC} = V_{CC} = 20 \, V, \, R_T = 4.7 \, k \Omega, \, C_T = 10 \, nF, \, T_A = 25^\circ C \) unless otherwise noted

<table>
<thead>
<tr>
<th>Characteristic</th>
<th>Symbol</th>
<th>Min</th>
<th>Typ</th>
<th>Max</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Oscillator Section</strong></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Oscillator Frequency</td>
<td>( f_{OSC} )</td>
<td>22</td>
<td>25</td>
<td>28</td>
<td>kHz</td>
</tr>
<tr>
<td>Frequency Change with Voltage ( (V_{CC} = 10 , V , \text{to} , 30 , V) )</td>
<td>( f_{VCC} )</td>
<td>—</td>
<td>0.01</td>
<td>5.0</td>
<td>%</td>
</tr>
<tr>
<td>Sawtooth Peak Voltage</td>
<td>( V_{OSClP} )</td>
<td>—</td>
<td>4.1</td>
<td>4.5</td>
<td>V</td>
</tr>
<tr>
<td>Sawtooth Valley Voltage</td>
<td>( V_{OSClV} )</td>
<td>1.2</td>
<td>1.5</td>
<td>—</td>
<td>V</td>
</tr>
<tr>
<td><strong>Logic Inputs</strong></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Input Threshold Voltage (Pins 3, 4, 5, 6, 7, 22, 23)</td>
<td>( V_{IH} )</td>
<td>3.0</td>
<td>2.2</td>
<td>—</td>
<td>V</td>
</tr>
<tr>
<td></td>
<td>( V_{IL} )</td>
<td>1.7</td>
<td>0.8</td>
<td>—</td>
<td>—</td>
</tr>
<tr>
<td>Sensor Inputs (Pins 4, 5, B)</td>
<td>( I_{IH} )</td>
<td>—</td>
<td>−70</td>
<td>−20</td>
<td>mA</td>
</tr>
<tr>
<td>High State Input Current ( (V_{HI} = 5.0 , V) )</td>
<td>( I_{IH} )</td>
<td>−150</td>
<td>−70</td>
<td>−20</td>
<td>mA</td>
</tr>
<tr>
<td>Low State Input Current ( (V_{IL} = 0 , V) )</td>
<td>( I_{IL} )</td>
<td>−600</td>
<td>−37</td>
<td>−150</td>
<td>mA</td>
</tr>
<tr>
<td>Output Enable</td>
<td>( I_{OH} )</td>
<td>−75</td>
<td>−38</td>
<td>−10</td>
<td>mA</td>
</tr>
<tr>
<td>Current Input Current ( (V_{HI} = 5.0 , V) )</td>
<td>( I_{OL} )</td>
<td>−300</td>
<td>−175</td>
<td>−75</td>
<td>mA</td>
</tr>
<tr>
<td>Low State Input Current ( (V_{IL} = 0 , V) )</td>
<td>( I_{OL} )</td>
<td>−60</td>
<td>−29</td>
<td>−10</td>
<td>mA</td>
</tr>
<tr>
<td><strong>Current-Limit Comparator</strong></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Threshold Voltage</td>
<td>( V_{TH} )</td>
<td>101</td>
<td>115</td>
<td>mV</td>
<td></td>
</tr>
<tr>
<td>Input Common Mode Voltage Range</td>
<td>( V_{ICR} )</td>
<td>3.0</td>
<td>—</td>
<td>—</td>
<td>V</td>
</tr>
<tr>
<td>Input Rise Current</td>
<td>( I_{R} )</td>
<td>0.9</td>
<td>5.0</td>
<td>mA</td>
<td></td>
</tr>
<tr>
<td>** Outputs and Power Sections **</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Top Drive Voltage Off-State Leakage ( V_{CC} = 30 , V)</td>
<td>( V_{CC} )</td>
<td>0.6</td>
<td>100</td>
<td>—</td>
<td>V</td>
</tr>
<tr>
<td>Rise Time</td>
<td>( I_{R} )</td>
<td>107</td>
<td>300</td>
<td>ns</td>
<td></td>
</tr>
<tr>
<td>Fall Time</td>
<td>( I_{F} )</td>
<td>26</td>
<td>300</td>
<td>—</td>
<td>—</td>
</tr>
<tr>
<td>Bottom Drive Voltage</td>
<td>( V_{OH} )</td>
<td>( V_{CC} - 2.0 )</td>
<td>—</td>
<td>V</td>
<td></td>
</tr>
<tr>
<td>High State ( (V_{CC} = 20 , V, , V_{C} = 30 , V, , \text{source} = 50 , mA) )</td>
<td>( V_{OL} )</td>
<td>( V_{CC} - 1.1 )</td>
<td>1.5</td>
<td>2.0</td>
<td></td>
</tr>
<tr>
<td>Low State ( (V_{CC} = 20 , V, , V_{C} = 30 , V, , \text{sink} = 50 , mA) )</td>
<td>( t_{R} )</td>
<td>38</td>
<td>200</td>
<td>ns</td>
<td></td>
</tr>
<tr>
<td>Bottom Drive Output Switching Time ( C_L = 1000 , pf )</td>
<td>( t_{F} )</td>
<td>30</td>
<td>200</td>
<td>—</td>
<td>—</td>
</tr>
<tr>
<td>Fault Output Sink Saturation ( V_{CC} = 16 , mA)</td>
<td>( V_{CC} )</td>
<td>226</td>
<td>500</td>
<td>mV</td>
<td></td>
</tr>
<tr>
<td>Fault Output Off-State Leakage ( V_{CC} = 20 , V)</td>
<td>( t_{H} )</td>
<td>1.0</td>
<td>100</td>
<td>—</td>
<td>—</td>
</tr>
<tr>
<td>Under Voltage Lockout</td>
<td>( V_{(on)} )</td>
<td>8.2</td>
<td>8.9</td>
<td>10</td>
<td>V</td>
</tr>
<tr>
<td>Drive Output Enabled ( (V_{CC} , \text{or} , V_{C} , \text{increasing}) )</td>
<td>( V_{H} )</td>
<td>0.1</td>
<td>0.2</td>
<td>0.3</td>
<td>mA</td>
</tr>
<tr>
<td>Power Supply Current</td>
<td>( I_{CC} )</td>
<td>—</td>
<td>12</td>
<td>18</td>
<td>mA</td>
</tr>
<tr>
<td>Pin 17 ( V_{CC} = 20 , V)</td>
<td>( I_{C} )</td>
<td>14</td>
<td>20</td>
<td>—</td>
<td>mA</td>
</tr>
<tr>
<td>Pin 17 ( V_{CC} = 20 , V, , V_{C} = 30 , V)</td>
<td>( I_{C} )</td>
<td>3.5</td>
<td>6.0</td>
<td>—</td>
<td>mA</td>
</tr>
<tr>
<td>Pin 18 ( V_{CC} = 20 , V, , V_{C} = 20 , V)</td>
<td>( I_{C} )</td>
<td>5.0</td>
<td>10</td>
<td>—</td>
<td>mA</td>
</tr>
</tbody>
</table>
### PIN FUNCTION DESCRIPTION

<table>
<thead>
<tr>
<th>Pin No.</th>
<th>Function</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>1, 2, 24</td>
<td>BT, AT, CT</td>
<td>These three open collector Top Drive Outputs are designed to drive the external upper power switch transistors.</td>
</tr>
<tr>
<td>3</td>
<td>PWDREV</td>
<td>The Forward/Reverse Input is used to change the direction of motor rotation.</td>
</tr>
<tr>
<td>4, 5, 6</td>
<td>$S_A$, $S_B$, $S_C$</td>
<td>These three Sensor Inputs control the commutation sequence.</td>
</tr>
<tr>
<td>7</td>
<td>Output Enable</td>
<td>A logic high at this input causes the motor to run, while a low causes it to coast.</td>
</tr>
<tr>
<td>8</td>
<td>Reference Output</td>
<td>This output provides charging current for the oscillator timing capacitor $C_T$ and a reference for the error amplifier. It may also serve to furnish sensor power.</td>
</tr>
<tr>
<td>9</td>
<td>Current Sense</td>
<td>A 100 mV signal, with respect to pin 15, at this input terminates output switch conduction during a given oscillator cycle. This pin normally connects to the top side of the current sense resistor.</td>
</tr>
<tr>
<td></td>
<td>(Noninverting Input)</td>
<td></td>
</tr>
<tr>
<td>10</td>
<td>Oscillator</td>
<td>The Oscillator frequency is programmed by the values selected for the timing components, $R_T$ and $C_T$.</td>
</tr>
<tr>
<td>11</td>
<td>Error Amp (Noninverting Input)</td>
<td>This input is normally connected to the speed set potentiometer.</td>
</tr>
<tr>
<td>12</td>
<td>Error Amp (Inverting Input)</td>
<td>This input is normally connected to the Error Amp Output in open-loop applications.</td>
</tr>
<tr>
<td>13</td>
<td>Error Amp Output/PWM Input</td>
<td>This pin is available for compensation in closed-loop applications.</td>
</tr>
<tr>
<td>14</td>
<td>Fault Output</td>
<td>This open collector output is active low during one or more of the following conditions: Invalid Sensor Input code, Enable Input at logic 0, Current Sense Input greater than 100 mV (pin 9 with respect to pin 15), Undervoltage Lockout activation, and Thermal Shutdown.</td>
</tr>
<tr>
<td>15</td>
<td>Current Sense (Inverting Input)</td>
<td>Reference pin for internal 100 mV threshold. This pin is normally connected to the bottom side of the current sense resistor.</td>
</tr>
<tr>
<td>16</td>
<td>Ground</td>
<td>This pin supplies a ground for the control circuit and should be referenced back to the power source ground.</td>
</tr>
<tr>
<td>17</td>
<td>VCC</td>
<td>This pin is the positive supply of the control IC. The controller is functional over a minimum VCC range of 10 V to 30 V.</td>
</tr>
<tr>
<td>18</td>
<td>VC</td>
<td>The high state (VCH) of the Bottom Drive Outputs is set by the voltage applied to this pin. The controller is operational over a minimum VC range of 10 V to 30 V.</td>
</tr>
<tr>
<td>19, 20, 21</td>
<td>$C_B$, $R_B$, $A_B$</td>
<td>These three totem pole Bottom Drive Outputs are designed for direct drive of the external bottom power switch transistors.</td>
</tr>
<tr>
<td>22</td>
<td>60°/120° Select</td>
<td>The electrical state of this pin configures the control circuit operation for either 60° (high state) or 120° (low state) sensor electrical phasing inputs.</td>
</tr>
<tr>
<td>23</td>
<td>Brake Input</td>
<td>A logic low state at this input allows the motor to run, while a high state does not allow motor operation and if operating causes rapid deceleration.</td>
</tr>
</tbody>
</table>
INTRODUCTION

The MC33035 is one of a series of high performance monolithic DC brushless motor controllers produced by Motorola. It contains all of the functions required to implement a full-featured, open-loop, three or four phase motor control system. In addition, the controller can be made to operate DC brush motors. Constructed with Bipolar Analog technology, it offers a high degree of performance and ruggedness in hostile industrial environments. The MC33035 contains a rotor position decoder for proper commutation sequencing, a temperature compensated reference capable of supplying a sensor power, a frequency programmable sawtooth oscillator, a fully accessible error amplifier, a pulse width modulator comparator, three open collector top drive outputs, and three high current transistor bottom drive outputs ideally suited for driving power MOSFETs.

Included in the MC33035 are protective features consisting of undervoltage lockout, cycle by cycle current limiting with a selectable time delayed latch integral thermal shutdown, and a unique fault output that can easily be interfaced to a microprocessor controller.

Typical motor control functions include open-loop speed control, forward or reverse rotation, run enable, and dynamic braking. In addition, the MC33035 has a 80°/120° select pin which configures the rotor position decoder for either 80° or 120° sensor electrical phasing inputs.

FUNCTIONAL DESCRIPTION

A representative internal block diagram is shown in Figure 19 with various applications shown in Figures 36, 38, 42, 44, 45, and 46. A discussion of the features and function of each of the internal blocks given below is referenced to Figures 19 and 36.

Rotor Position Decoder

An internal rotor position decoder monitors the three sensor inputs (Pins 4, 5, 6) to provide the proper sequencing of the top and bottom drive outputs. The sensor inputs are designed to interface directly with open collector type Hall Effect switches or opto-isolated couplers. Internal pull-up resistors are included to minimize the required number of external components. The inputs are TTL compatible, with their thresholds typically at 2.2 volts. The MC33035 is designed to control three phase motors and operate with four of the most common conventions of sensor phasing. A 80°/120° select (Pin 22) is conveniently provided which affords the MC33035 to configure itself to control motors having either 80°, 120°, 240° or 300° electrical sensor phasing. With three sensor inputs there are eight possible input code combinations, six of which are valid rotor positions. The remaining two codes are invalid and are usually caused by an open or shorted sensor line. When an invalid input condition exists, the Fault output is activated and the drive outputs are disabled. With six valid input codes, the decoder can resolve the motor rotor position to within a window of 60 electrical degrees.

The forward/reverse input (Pin 3) is used to change the direction of motor rotation by reversing the voltage across the stator winding. When the input changes state, from high to low with a given drive output code (for example 100), the enabled top and bottom drive outputs with the same alpha designation are exchanged (A1 to A2, B1 to B2, C1 to C2). The effect sequence is reversed and the motor changes directional rotation.

Motor on/off control is accomplished by the output enable (Pin 7). When left disconnected, an internal 25 μA current source enables sequencing of the top and bottom drive outputs. When grounded, the top drive output turns off and the bottom drives are forced low, causing the motor to coast and the Fault output to activate.

Dynamic motor braking allows an additional margin of safety to be built into the final product. Braking is accomplished by placing the brake input (Pin 23) in a high state. This causes the top drive outputs to turn off and the bottom drives to turn on, shorting the motor generated back EMF. The brake input has unconditional priority over all other inputs. The internal 40 kΩ pull-up resistor simplifies interfacing with the system safety switch by insuring brake activation if open or disconnected. The commutation logic truth table is shown in Figure 20. A four input NOR gate is used to monitor the brake input and the inputs to the three top drive output transistors. Its purpose is to disable braking until the top drive outputs attain a high state. This helps to prevent simultaneous conduction of the top and bottom power switches. In half wave motor drive application, the top drive outputs are not required and are normally left disconnected. Under these conditions braking will still be accomplished since the NOR gate senses the base voltage to the top drive output transistors.

Error Amplifier

A high performance, fully compensated error amplifier with access to both inputs and output (Pins 11, 12, 13) is provided to facilitate the implementation of closed-loop motor speed control. The amplifier features a typical DC voltage gain of 80 dB, 0.8 MHz gain bandwidth, and a wide input common mode voltage range that extends from ground to VREF in most open-loop speed control applications, the amplifier is configured as a unity gain voltage follower with the non-inverting input connected to the speed set voltage source. Additional configurations are shown in Figures 31 through 35.

Oscillator

The frequency of the internal ramp oscillator is programmed by the values selected for timing components R1 and C1. Capacitor C1 is charged from the reference output (Pin 8) through resistor R1 and discharged by an internal discharge transistor. The ramp peak and valley voltages are typically 4.1 V and 1.5 V respectively. To provide a good compromise between audible noise and output switching efficiency, an oscillator frequency in the range of 20 kHz to 30 kHz is recommended. Refer to Figure 1 for component selection.
FIGURE 19 — REPRESENTATIVE BLOCK DIAGRAM

FIGURE 20 — THREE PHASE, SIX STEP COMMUTATION TRUTH TABLE

<table>
<thead>
<tr>
<th>Inputs (Note 2)</th>
<th>Outputs (Note 3)</th>
<th>Top Drives</th>
<th>Bottom Drives</th>
<th>Faile</th>
</tr>
</thead>
<tbody>
<tr>
<td>Sensor Electrical Phasing (Note 4)</td>
<td>Sg</td>
<td>Sc</td>
<td>Sr</td>
<td>Sr</td>
</tr>
<tr>
<td>1</td>
<td>0</td>
<td>0</td>
<td>1</td>
<td>0</td>
</tr>
<tr>
<td>1</td>
<td>0</td>
<td>1</td>
<td>0</td>
<td>1</td>
</tr>
<tr>
<td>1</td>
<td>0</td>
<td>0</td>
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<td>0</td>
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<tr>
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<td>0</td>
</tr>
<tr>
<td>1</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
</tbody>
</table>

NOTES:
1. 0 = Any one of six valid sensor or drive combinations.
2. 0 = Don't care.
3. The digital inputs [X = 3, 4, 5, 6, 7, 21] are all TTL compatible. The current sense input (Pin 6) has a 100 mv threshold with respect to Pin 10.
4. A logic 0 for this input is defined as < 88 mV, and a logic 1 is > 118 mv.
5. The Enable and brake inputs are open collector design and require logic level inputs in the low B state.
6. The Enable and brake inputs are open collector design and require logic level inputs in the low B state.
7. The Enable and brake inputs are open collector design and require logic level inputs in the low B state.
8. Valid 60° or 120° sensor combinations for corresponding valid top and bottom drive outputs.
9. Invalid sensor inputs with brakes = 1; all top drives off, all bottom drives off, fault low.
10. Valid 60° or 120° sensor inputs with brakes = 1; all top drives off, all bottom drives off, fault low.
11. All bottom drives off, fault low.

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Pulse Width Modulator

The use of pulse width modulation provides an energy efficient method of controlling the motor speed by varying the average voltage applied to each stator winding during the commutation sequence. As Cy discharges, the oscillator sets both latches, allowing conduction of the top and bottom drive outputs. The PWM comparator resets the upper latch, terminating the bottom drive output conduction when the positive-going ramp of C_y becomes greater than the error amplifier output. The pulse width modulator timing diagram is shown in Figure 21. Pulse width modulation for speed control appears only at the bottom drive outputs.

Current Limit

Continuous operation of a motor that is severely over-loaded results in overheating and eventual failure. This destructive condition can best be prevented with the use of cycle-by-cycle current limiting. That is, each cycle is treated as a separate event. Cycle-by-cycle current limiting is accomplished by monitoring the stator current build-up each time an output switch conducts, and upon sensing an over current condition, immediately turning off the switch and holding it off for the remaining duration of the oscillator ramp-up period. The stator current is converted to a voltage by inserting a ground referenced sense resistor Rs (Figure 36) in series with the three bottom switch transistors (Q_2, Q_3, Q_4). The voltage across the sense resistor is directly monitored by the current sense comparator inputs (Pin 9 and 16) and compared to the internal 100 mV reference. The current sense comparator inputs have an input common mode input range of approximately 3.0 volts. If the 100 mV current sense threshold is exceeded, the comparator resets the lower sense latch and terminates output switch conduction. The value for the current sense resistor is:

\[ R_s = \frac{0.1}{I_{stator(max)}} \]

The Fault output activates during an over current condition. The dual-latch PWM configuration ensures that only one single output conduction pulse occurs during any given oscillator cycle, whether terminated by the output of the error amp or the current limit comparator.

Reference

The on-chip 6.25 V regulator (Pin 8) provides charging current for the oscillator timing capacitor, a reference for the error amplifier, and can supply 20 mA of current suitable for directly powering sensors in low voltage applications. In higher voltage applications, it may become necessary to transfer the power dissipated by the regulator to the IC. This is easily accomplished with the addition of an external pass transistor as shown in Figure 22. A 8.25 V reference level was chosen to allow implementation of the simpler NPN circuit, where V_{ref} = V_{CC} exceeds the minimum voltage required by Hall Effect sensors over temperature. With proper transistor selection, and adequate heat sinking, up to one amp of load current can be obtained.

**FIGURE 22 — REFERENCE OUTPUT BUFFERS**

The NPN circuit is recommended for powering Hall or optic sensors, where the output voltage temperature coefficient is not critical. The PNP circuit is slightly more complex, but is also more robust over temperature. Neither circuit has current limiting.

**FIGURE 21 — PULSE WIDTH MODULATOR TIMING DIAGRAM**

Capacitor C_y

Error Amp Out

PWM input

Current Sense Input

Top Drive Output

Bottom Drive Output

Fault Output
Undervoltage Lockout
A triple Undervoltage Lockout has been incorporated to prevent damage to the IC and the external power switch transistors. Under low power supply conditions, it guarantees that the IC and sensors are fully functional, and that there is sufficient bottom drive output voltage. The positive power supplies to the IC (VCC) and the bottom drives (VCC) are each monitored by separate comparators that have their thresholds at 9.1 V. This level ensures sufficient gate drive necessary to attain low RDS(on) when driving standard power MOSFET devices. When directly powering the Hall sensors from the reference, improper sensor operation can result, if the reference output voltage falls below 4.5 V. A third comparator is used to detect this condition. If one or more of the comparators detects an undervoltage condition, the Fault output is activated, the top drives are turned off and the bottom drive outputs are held in a low state. Each of the comparators contain hysteresis to prevent oscillations when crossing their respective thresholds.

Fault Output
The open collector Fault output (Pin 14) was designed to provide diagnostic information in the event of a system malfunction. It has a sink current capability of 16 mA and can directly drive a light emitting diode for visual indication. Additionally, it is easily interfaced with TTL/CMOS logic for use in a microprocessor controlled system. The Fault output is active low when one or more of the following conditions occur:
1) Invalid Sensor Input code.
2) Enable Input at logic high.
3) Current Sense Input greater than 100 mV.
4) Undervoltage Lockout, activation of one or more of the comparators.
5) Thermal Shutdown, maximum junction temperature being exceeded.

This unique output can also be used to distinguish between motor start-up or sustained operation in an overloaded condition. With the addition of an RC network between the Fault output and the enable input, it is possible to create a time-delayed latched shutdown for overcurrent. The added circuitry shown in Figure 23. makes easy starting of motor systems which have high inertial loads by providing additional starting torque, while still preserving overcurrent protection. This task is accomplished by setting the current limit to a higher than nominal value for a predetermined time. During an excessively long overcurrent condition, capacitor C2 will charge causing the enable input to cross its threshold to a low state. A latch is then formed by the positive feedback loop from the Fault output to the enable input. Once set, by the current sense input, it can only be reset by shorting C2 or cycling the power supplies.

Drive Outputs
The three top drive outputs (Pins 1, 2, 24) are open collector NPN transistors capable of sinking 50 mA with a minimum breakdown of 30 volts. Interfacing into higher voltage applications is easily accomplished with the circuits shown in Figures 24 and 25.

The three totem pole bottom drive outputs (Pins 19, 20, 21) are particularly suited for direct drive of N channel MOSFETs or NPN bipolar transistors (Figures 26, 27, 28 and 29). Each output is capable of sourcing and sinking up to 100 mA. Power for the bottom drives is supplied from VCC (Pin 18). This separate supply input allows the designer added flexibility in tailoring the drive voltage, independent of VCC. A zener clamp should be connected to this input when driving power MOSFETs in systems where VCC is greater than 20 V so as to prevent rupture of the MOSFET gates.

The control circuitry ground (Pin 16) and current sense inverting input (Pin 15) must return on separate paths to the central input source ground.

Thermal Shutdown
Internal thermal shutdown circuitry is provided to protect the IC in the event the maximum junction temperature is exceeded. When activated, typically at 170°C, the IC acts as though the enable input was grounded.

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FIGURE 24 — HIGH VOLTAGE INTERFACE WITH NPN POWER TRANSISTORS

FIGURE 25 — HIGH VOLTAGE INTERFACE WITH 'N' CHANNEL POWER MOSFETS

FIGURE 26 — CURRENT WAVEFORM SPIKE SUPPRESSION

FIGURE 27 — MOSFET DRIVE PRECAUTIONS

FIGURE 28 — BIPOLAR TRANSISTOR DRIVE

FIGURE 29 — CURRENT SENSING POWER MOSFETs

Transistor Q1 is a common base stage used to level shift from VIN to the high motor voltage, VIN. The collector diode is required if VIN is present while VIN is low.

The addition of the RC filter will eliminate current-limit instability caused by the leading edge spike on the current waveform. Resistor R1 should be a low inductance type.

Series gate resistor Rg will damp any high frequency oscillations caused by the MOSFET's input capacitance and any series wiring inductance in the gate-source circuit. Return D is required if the negative current into the Bottom Drive Outputs exceeds 50 mA.

The triac output can furnish negative base current for enhanced transistor turn-off, with the addition of capacitor C.

Power Ground

Control Circuit Ground (Pin 18) and Current Sensing Input (Pin 16) must return on separate paths to the Sense Input Source Ground.

Virtually lossless current sensing can be achieved with the implementation of SENSEFET power switches.
This circuit generates $V_{\text{boost}}$ for Figure 25.

Resistor $R_1$ with capacitor $C$ sets the acceleration time constant while $R_2$ controls the deceleration. The values of $R_1$ and $R_2$ should be at least ten times greater than the speed set potentiometer to minimize time constant variations with different speed settings.

The SN74LS146 is an open collector BCD to One of Ten decoder. When connected as shown, input codes 0000 through 1011 upset the PWM in increments of approximately 10%. From 0 to 30% on-time, input codes 1010 through 1111 will produce 100% on-time or full motor speed.

The rotor position sensors can be used as a tachometer. By differentiating the positive-going edges and then integrating them over time, a voltage proportional to speed can be generated. The error amplifier compares this voltage to that of the speed set to control the PWM.

This circuit can control the speed of a cooling fan proportional to the difference between the sensor and set temperatures. The control loop is closed as the forced air cools the RTC/thermostat. For controlled heating applications, exchange the positions of $R_1$ and $R_2$. 

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MOTOROLA LINEAR/INTERFACE DEVICES
SYSTEM APPLICATIONS

Three Phase Motor Commutation

The three phase application shown in Figure 36 is a full-featured open-loop motor controller with full wave, six step drive. The upper power switch transistors are Darlington while the lower devices are power MOSFETs. Each of these devices contains an internal parasitic catch diode that is used to return the stator inductive energy back to the power supply. The outputs are capable of driving a delta or wye connected stator, and a grounded neutral wye if split supplies are used. At any given rotor position, only one top and one bottom power switch (of different transistor) is enabled. This configuration switches both ends of the stator winding from supply to ground which causes the current flow to be bidirectional or full wave. A leading edge spike is usually present on the current waveform and can cause a current-limit instability. The spike can be eliminated by adding an RC filter in series with the current sense input. Using a low inductance type resistor for $R_g$ will also aid in spike reduction. Care must be taken in the selection of the bottom power switch transistors so that the current during braking does not exceed the device rating. During braking, the peak current generated is limited only by the series resistance of the conducting bottom switch and winding.

$$I_{peak} = \frac{V_M + EMF}{R_{switch} + R_{winding}}$$

If the motor is running at maximum speed with no load, the generated back EMF can be as high as the supply voltage, and at the onset of braking the peak current may approach twice the motor stall current. Figure 37 shows the commutation waveforms over two electrical cycles. The first cycle ($0^\circ$ to $360^\circ$) depicts motor operation at full speed while the second cycle ($360^\circ$ to $720^\circ$) shows a reduced speed with about 50 percent pulse width modulation. The current waveforms reflect a constant torque load and are shown synchronous to the commutation frequency for clarity.

FIGURE 36 — THREE PHASE, SIX STEP, FULL WAVE MOTOR CONTROLLER
MC33035

FIGURE 37 — THREE PHASE, SIX STEP, FULL WAVE COMMUTATION WAVEFORMS

Rotor Electrical Position (Degrees)

Sensor Inputs
60°/120°
Select Pin Open

Code: 100 110 111 011 000 100 110 111 011 000 100

Sensor Inputs
60°/120°
Select Pin Grounded

Code: 100 110 010 011 001 101 100 110 010 011 001 101

Top Drive Outputs

B_T

C_T

Bottom Drive Outputs

A_B

B_B

Conducting Power Switch Transistors

A

B

C

Motor Drive Current

Motor Drive Current

Full Speed (No PWM)

Reduced Speed (~ 50% PWM)
Figure 3B shows a three phase, three step, half wave motor controller. This configuration is ideally suited for automotive and other low voltage applications since there is only one power switch voltage drop in series with a given stator winding. Current flow is unidirectional or half wave because only one end of each winding is switched. Continuous braking with the typical half wave arrangement presents a motor overheating problem since stator current is limited only by the winding resistance. This is due to the lack of upper power switch transistors, as in the full wave circuit, used to disconnect the windings from the supply voltage $V_M$. A unique solution is to provide braking until the motor stops and then turn off the bottom drives. This can be accomplished by using the Fault output in conjunction with the Enable input as an over current timer. Components $R_{DLV}$ and $C_{DLV}$ are selected to give the motor sufficient time to stop before latching the Enable input and the top drive AND gates low. When enabling the motor, the brake switch is closed and the PNP transistor along with resistors $R_1$ and $R_{DLV}$ are used to reset the latch by discharging $C_{DLV}$. The stator flyback voltage is clamped by a single zener and three diodes.

FIGURE 3B — THREE PHASE, THREE STEP, HALF WAVE MOTOR CONTROLLER
Three Phase Closed Loop Controller

The MC33035, by itself, is only capable of open loop motor speed control. For closed loop motor speed control, the MC33035 requires an input voltage proportional to the motor speed. Traditionally this has been accomplished by means of a tachometer to generate the motor speed feedback voltage. Figure 39 shows an application whereby an MC33035, powered from the 6.25 volt reference (Pin 8) of the MC33035, is used to generate the required feedback voltage without the need of a costly tachometer. The same Hall sensor signal used by the MC33035 for motor position decoding are utilized by the MC33035. Every positive or negative going transition of the Hall sensor signal on any of the sensor lines causes the MC33035 to produce an output pulse of defined amplitude and time duration, as determined by the external resistor R1 and capacitor C1. The output train of pulses at Pin 5 of the MC33035 is integrated by the error amplifier of the MC33035 configured as an integrator to produce a DC voltage level which is proportional to the motor speed. This speed proportional voltage establishes the PWM reference level at pin 13 of the MC33035 motor controller and closes the feedback loop. The MC33035 outputs drive an MPM3003 TMOS power MOSFET 3-phase bridge circuit capable of delivering up to 25 Amperes of surge current. High currents can be expected during conditions of start up, breaking, and change of direction of the motor.

The system shown in Figure 39 is designed for a motor having 120/240 degrees Hall sensor electrical phasing. The system can easily be modified to accommodate 60/120 degree Hall sensor electrical phasing by removing the jumper (J2) at Pin 22 of the MC33035.

FIGURE 39 — CLOSED LOOP BRUSHLESS DC MOTOR CONTROL USING THE MC33035, MPM3003, AND MC33035
Sensor Phasing Comparison

There are four conventions used to establish the relative phasing of the sensor signals in three phase motors. With six step drives, an input signal change must occur every 60 electrical degrees; however, the relative signal phasing is dependent upon the mechanical sensor placement. A comparison of the conventions in electrical degrees is shown in Figure 40. From the sensor phasing table, Figure 41, note that the order of input codes for 60° phasing is the reverse of 30°. This means the MC33035, when configured for 60° sensor electrical phasing, will equally operate a motor with either 60° or 30° sensor electrical phasing, but resulting in opposite directions of rotation. The same is true for the part when it is configured for 120° sensor electrical phasing; the motor will equally operate, but will result in opposite directions of rotation for 120° for 240° conventions.

FIGURE 40 — SENSOR PHASING COMPARISON

In this data sheet, the rotor position is always given in electrical degrees since the mechanical position is a function of the number of rotating magnetic poles. The relationship between the electrical and mechanical position is:

Electrical Degrees = Mechanical Degrees \( \left( \frac{\# \text{Rotor Poles}}{2} \right) \)

An increase in the number of magnetic poles causes more electrical revolutions for a given mechanical revolution. General purpose three phase motors typically contain a four pole rotor which yields two electrical revolutions for one mechanical.

Two and Four Phase Motor Commutation

The MC33035 is also capable of providing a four step output that can be used to drive two or four phase motors. The truth table in Figure 42 shows that by connecting sensor input Sg and Sc together, it is possible to truncate the number of drive output states from six to four. The output power switches are connected to By, Cy, Bg, and Cg. Figure 43 shows a four phase, four step, full wave motor control application. Power switch transistors Q1 through Q8 are Darlington type, each with an internal parasitic catch diode. With four step drive, only two rotor position sensors spaced at 90 electrical degrees are required. The commutation waveforms are shown in Figure 44.

Figure 45 shows a four phase, four step, half wave motor controller. It has the same features as the circuit in Figure 38, except for the deletion of speed control and braking.

FIGURE 41 — SENSOR PHASING TABLE

![Sensor Phasing Table](image)

FIGURE 42 — TWO AND FOUR PHASE, FOUR STEP, COMMUTATION TRUTH TABLE

![Truth Table](image)

*With MC33036 sensor input Sg connected to Sc
FIGURE 43 — FOUR PHASE, FOUR STEP, FULL WAVE CONTROLLER
MC33035

FIGURE 44 — FOUR PHASE, FOUR STEP, FULL WAVE MOTOR CONTROLLER

Rotor Electrical Position (Degrees)

Sensor Inputs
SA
SB
Select Pin
Open

Code
10 10 01 00 10 11 01 00

Top Drive
Outputs
BT
CT

Bottom Drive
Outputs
BB
CB

Conducting
Power Switch
Translators
A O
B O
C O
D O

Motor
Drive Current

Full Speed (No PWM)

FWD/REV = 1

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FIGURE 46 — FOUR PHASE, FOUR STEP, HALF WAVE MOTOR CONTROLLER
MC33035

Brush Motor Control

Though the MC33035 was designed to control brushless DC motors, it may also be used to control DC brush-type motors. Figure 46 shows an application of the MC33035 driving a Motorola MPM3002 MOSFET H-bridge affording minimal parts count to operate a one-tenth horsepower brush-type motor. Key to the operation is the input sensor code (100) which produces a top-left (Q1) and a bottom-right (Q4) drive when the controller’s forward/reverse pin is at logic 1; top-right (Q2), bottom-left (Q3) drive is realized when the forward/reverse pin is at logic 0. This code supports the requirements necessary for H-bridge drive accomplishing both direction and speed control.

The controller functions in a normal manner with a pulse width modulated-frequency of approximately 25 kHz. Motor speed is controlled by adjusting the voltage presented to the noninverting input of the error amplifier establishing the PWM’s slice or reference level. Cycle-by-cycle current limiting of 3.0 amperes motor current is accomplished by sensing the voltage (100 mV) across the 47 Ohm resistor to ground of the H-bridge motor current. The over current sense circuit makes it possible to reverse the direction of the motor, using the normal forward/reverse switch, on the fly and not have to completely stop before reversing.

**FIGURE 46 — H-BRIDGE BRUSH-TYPE CONTROLLER**

*Single Package MPM3002 MOSFET H-Bridge
M = 1/10th horsepower DC brush-type motor

**LAYOUT CONSIDERATIONS**

Do not attempt to construct any of the brushless motor control circuits on wire-wrap or plug-in prototype boards. High frequency printed circuit layout techniques are imperative to prevent noise. This is usually caused by excessive noise pickup imposed on the current sense or error amp inputs. The printed circuit layout should contain a ground plane with low current signal and high drive and output buffer grounds returning on separate paths back to the power supply input filter capacitor $V_{CC}$. Ceramic bypass capacitors (0.1 μF) connected close to the integrated circuit at $V_{CC}$, $V_{CC}$, and the error amp noninverting input may be required depending upon circuit layout. This provides a low impedance path for filtering any high frequency noise. All high current loops should be kept as short as possible using heavy copper runs to minimize radiated EMI.

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Appendix B: Photographs
Figure B.1: Drilling Out the Heatsinks.

Figure B.2: Tapping the Heatsinks.
Figure B.3: Measuring the Plastic Enclosure.

Figure B.4: Cutting the Plastic Enclosure.
Figure B.5: Etching Printed Circuit Boards.

Figure B.6: Front View of Modules with Tops Removed.
Figure B.7: Front View of Modules.

Figure B.8: Rear view of Modules.
Figure B.9: High Power Test Setup.

Figure B.10: Oscilloscope Test Equipment.
Figure B.11: Controller Destroyed by Overcurrent.

Figure B.12: Close-up View of Destroyed Controller.