

Three Piece Solution For Brushless Motor Controller Design

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Until recently, motor control designers who wished to take advantage of the brushless DC motor's unique attributes were faced with a difficult task. There were no control ICs designed to decode data coming from Hall effect sensors, let alone perform all the ancillary functions such as forward/reverse selection, overcurrent shutdown, undervoltage lockout, overtemperature shutdown, and so forth. Using discrete components to include these functions was an alternative, but discretes often consumed far too much circuit board area, especially if the control unit was to be

placed inside the motor housing.

Another problem area was the inadequate performance of the available power transistors. Power bipolars weren't favored because they can't be driven directly from a control IC, and Darlington's have on-state voltages that are sometimes too high. Power MOSFETs seemed to be the ideal choice since they are easy to drive, efficient and inexpensive. However, designers were sometimes troubled by their inability to withstand stresses common in pulse width modulate motor controllers.

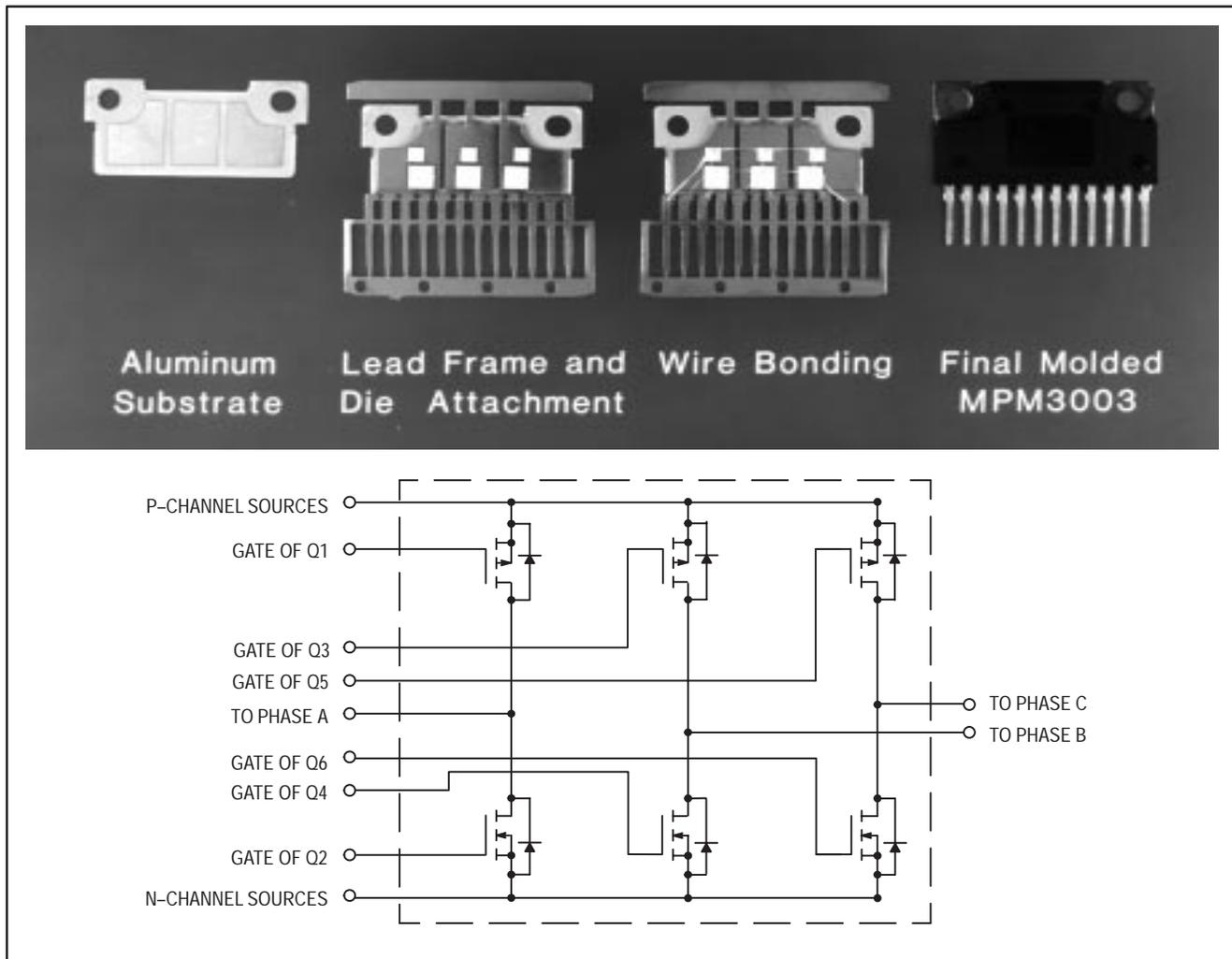


Figure 1. Construction and Diagram of the MPM3003, a Three Phase Bridge That is Ideal for Driving Small Brushless DC Motors

THE BRAINS

The system highlighted here is built with Motorola's MC33035 and a support chip, the MC33039. They are examples of new ICs that dramatically simplify the design of brushless motor control systems and reduce required circuit board area and parts count. Added benefits are shorter design times and improved system performance.

The MC33035 is a 24-pin linear IC that can operate as the control center for a brushless DC motor control system. The main duty of the MC33035 is to decode the signals from the Hall effect sensors and generate logic for electronically commutating the motor. The commutation logic is internally fed into the six output drivers consisting of three open collector NPN transistors that drive the upper legs of the bridge and three totem pole drivers that control the devices. The open collector outputs can sink 50 mA; with some additional circuitry this allows control of either N-channel MOSFETs for higher power applications or P-channel MOSFETs if a simple interface is desired. Since the three lower totem pole outputs can source and sink 100 mA, they can drive power MOSFETs directly.

FAULT MANAGEMENT

The MC33035 can detect and manage several types of faults. A common method of overcurrent detection is to tie the sources of the lower three transistors together and return them to the negative supply rail through a current sense resistor. The sense voltage, which is proportional to load current, is fed into a comparator on board the MC33035. The comparator then feeds an RS Flip-Flop, which ensures that once an overcurrent condition is detected, the output drivers turn-off the power transistors the remainder of the oscillator cycle. Without the internal flip-flop the overcurrent protection loop would rapidly cycle On and Off about the comparator's threshold, causing excessive power transistor heating.

In addition to overcurrent management, the MC33035 provides undervoltage lockouts that terminate the drive to the output transistors if any of three conditions occur. The first is insufficient voltage to operate the IC. The second occurs when there is insufficient voltage to drive the power MOSFET gates. Finally, the output drivers turn off the power transistors when the MC33035 does not sustain its on board 6.25 V reference. An invalid set of Hall effect signals or excessive temperature will cause shutdown, too.

Whenever any fault condition is present, an NPN transistor capable of sinking 16 mA pulls the Fault Output pin low. Potential uses of the Fault Output include alerting a microprocessor of a problem, lighting an LED, implementing a soft start feature to limit motor start up currents, or latching the system off at the first sign of a problem or after a fixed delay.

CONTROL FEATURES

The MC33035's circuitry contains all but one of the major elements for closed loop speed control. The only piece lacking is one that monitors motor RPM and generates a signal proportional to motor speed, a function which traditionally has been the domain of a tachometer. Once provided with a motor speed signal, the MC33035's high performance error amplifier and its internal oscillator form the last major links in the control loop.

The MC33035's on-board oscillator operates at a frequency set by an external R-C. Each cycle capacitor C_T (Figure 2) is charged from the reference output through resistor R_T and then rapidly discharged through an internal transistor. Figure 3 shows how the resulting sawtooth-shaped waveform affects output pulse width.

The MC33035 has tremendous flexibility since it works well with various Hall effect sensor spacings and the most common brushless motor windings. An MC33035 based system can easily be altered to drive motors with either delta or wye connected three phase windings with 60 or 120 degree

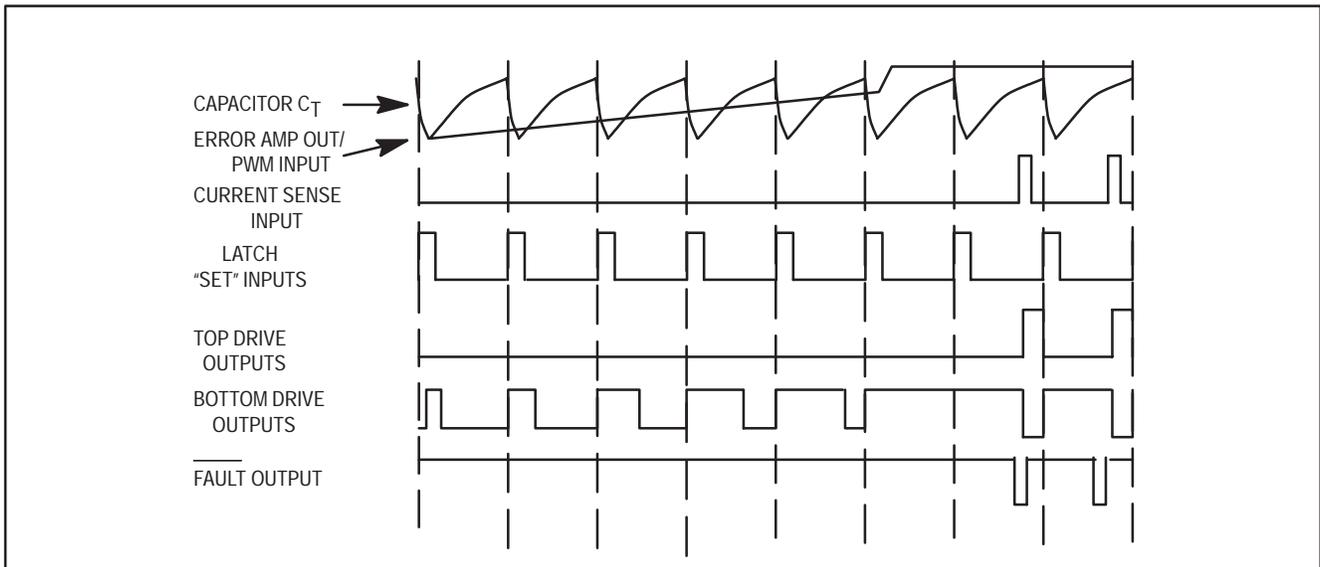


Figure 3. The MC33035's Pulse Width Modulator Timing Diagram

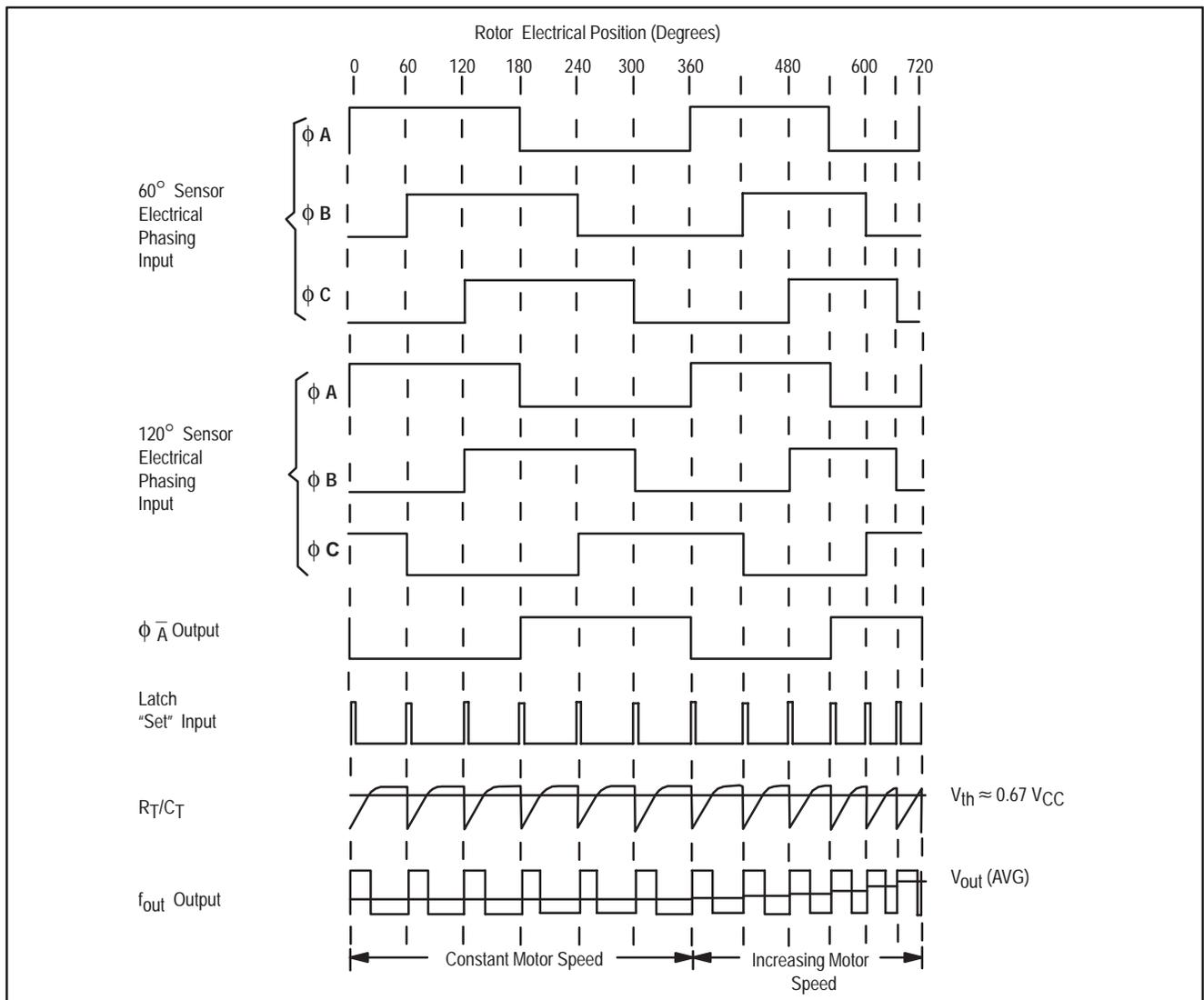


Figure 4. Timing Diagram of a Typical Three Phase, Six Step Motor Application

Hall effect sensor spacing.

A companion IC, the MC33039, is a low cost, space saving alternative to a tachometer. At each positive or negative transition of the Hall effect sensors the MC33039 generates a pulse with a fixed on time. The output signal can then be filtered to obtain a voltage proportional to motor speed. Design of an MC33035/39 based system should begin with setting the system timing, which originates in the MC33039.

ASSEMBLING THE PIECES

Figure 4 shows the MC33039 timing diagram, and Figure 5 shows its representative block diagram. Selection of timing components for the MC33039 is based on the desired maximum motor RPM. For the motor used in this application (Pittman ELCOM 5112) there are two electrical degrees for every mechanical degree since the permanent magnet on the rotor has two pairs of poles. Therefore, for every mechanical revolution each Hall effect sensor delivers two pulses and the

three sensors generate six pulses. The MC33039 generates 12, one for each rising and falling edge.

For a given maximum motor speed, the output pulse width has a maximum limit. If we assume a maximum desired speed of 5000 RPM, which is about 83 revolutions per second, the MC33039 will generate 83 x 12, or 1000 pulses per second. The 1 kHz frequency dictates that the maximum pulse width must be less than 1 ms. From Figure 6, which is taken from the MC33039's data sheet, one can determine that R1 and C1 values of 43 kΩ and 22 nF result in a pulse width of 950 μs. To set the system PWM frequency, refer to the MC33035's data sheet. There it shows that setting R2 and C2 to 5.1 kΩ and 0.01 μF gives a nominal pulse width modulation frequency of 24 kHz, just above the audible range.

Both inputs and the output of the MC33035's error amplifier are accessible to accommodate various control methods. For open loop control you can feed a reference signal proportional to desired speed into the error amplifier's non-inverting input. The error amplifier is then configured as

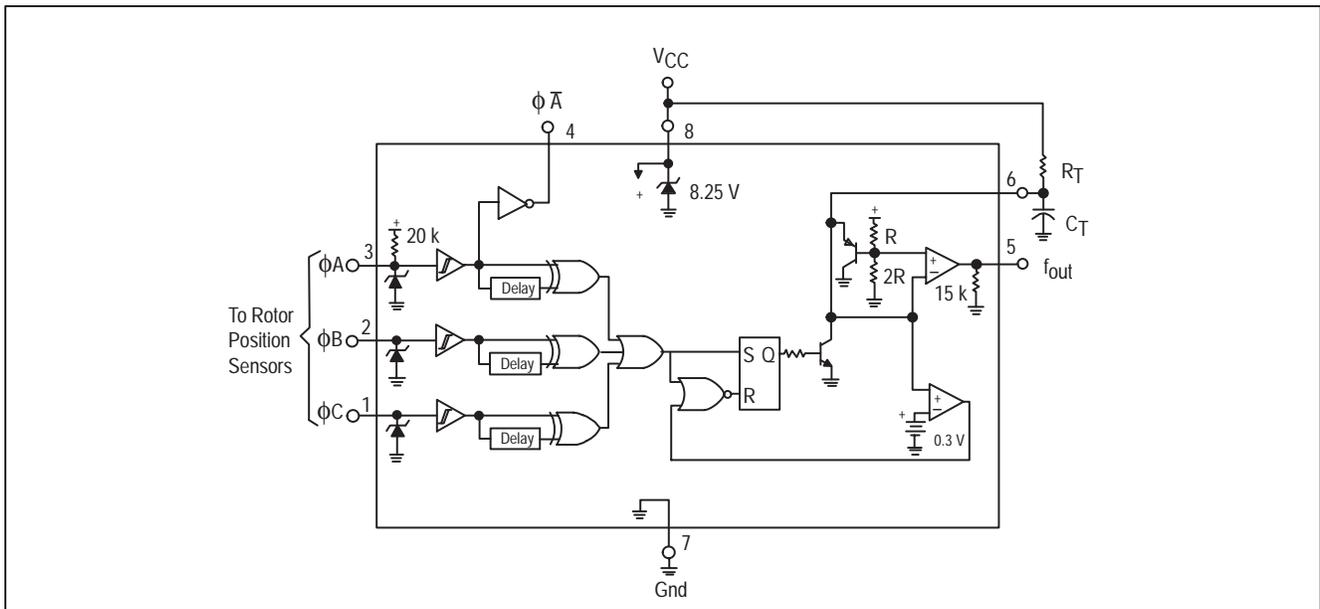


Figure 5. Representative Block Diagram of MC33039

a unity gain voltage follower by connecting its inverting input to its output. The error amplifier's output is then compared to the output of the oscillator to obtain a PWM signal proportional to desired motor speed – unless the control loop is overridden by an overcurrent or fault condition.

For closed loop control one approach is to filter the MC33039's output with a low pass filter to generate a voltage proportional to motor speed and feed the resulting signal into the inverting input of the MC33035's comparator. A signal proportional to desired motor speed drives the non-inverting input, and the ratio of the input and feedback resistors R3 and R4 control gain. In this design, low pass filtering and generating the error signal are combined by using feedback capacitor, C3.

Ideally, the integrator/error amplifier should produce a ripple free output even at low motor speeds. To do so at very low speeds reduces system response time, however. Component values must be adjusted according to the rotor's and load's inertia and friction. In this particular application the

values 1 MΩ and 0.1 μF (t = 100 ms) give good dynamic response and stability.

When motor speed falls below the desired speed, the MC33035 extends the output pulse width to the drive transistors. When motor speed is greater than the desired speed, the duty cycle decreases. However, if the input signal abruptly demands a much lower speed, the duty cycle could fall to zero and the motor would coast to desired low speed. Therefore, since the MC33035 has no provision to dynamically brake the motor and thus control rapid deceleration, it is best suited for applications which have a large frictional load or those that do not require a controlled, abrupt deceleration. The schematic of a closed loop brushless motor control system is shown in Figure 7. Shown in Figure 11 is a completed brushless motor control.

N-CHANNEL GATE DRIVE CIRCUIT

The magnitude of the system voltage effects how one might deliver power to the MC33035 and generate the gate drive supply for the N-channel power MOSFETs. Here we are only concerned with two possible supply voltages, 12 V (a range of 9 to 16 V) and 24 V (18 to 30 V). Since the MC33035 has a 40 V rating, it can be powered directly from either system supply voltage if the supply is free of large voltage transients. In addition to an electrolytic capacitor a small filter capacitor (0.1 μF) located near the IC is recommended to reduce local spiking across the DC bus.

For reduced power dissipation in the IC, the MC33035 allows driving the three lower output transistors from a supply that is independent of the MC33035's supply. Because the power transistors in this system are MOSFETs, the only required drive current is that which is needed to charge and discharge each MOSFET's gate-to-source and drain-to-gate capacitors. The lower N-channel devices require roughly 15 nC of gate charge to reach a gate-to-source voltage of 15 V. At a nominal PWM frequency of 24 kHz, the average drive

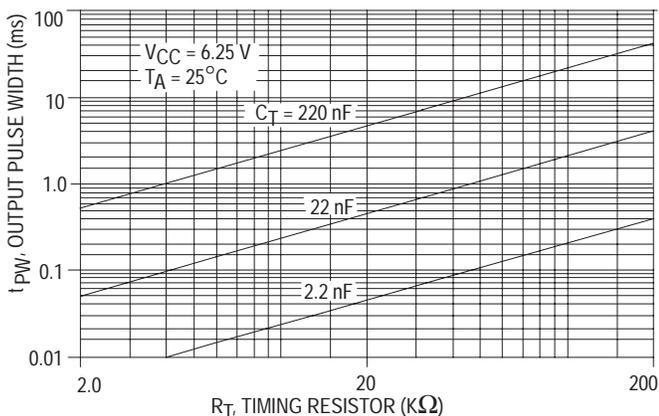


Figure 6. f_{out} Pulse Width versus Timing Resistor

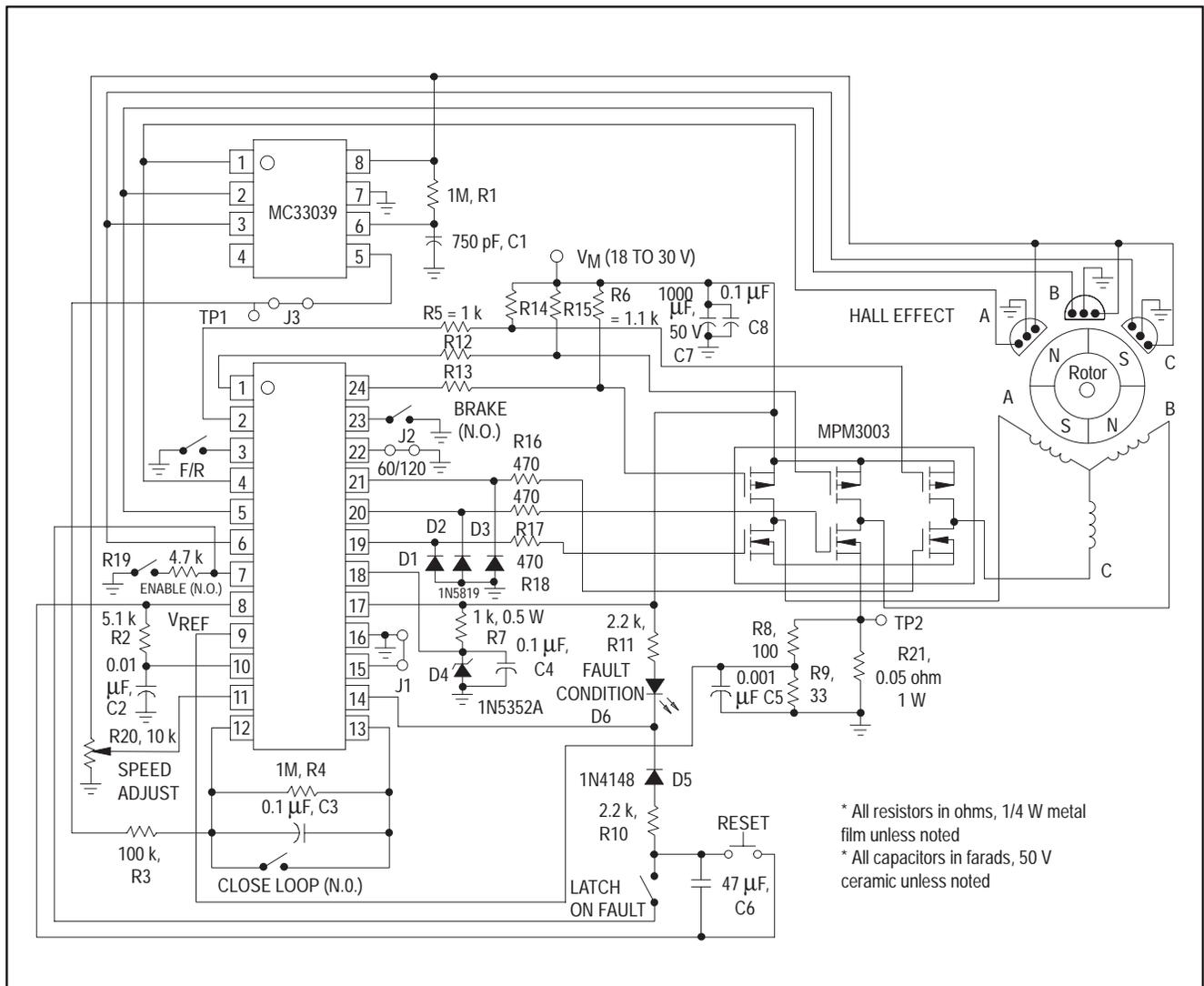


Figure 7. 24 V Brushless Motor Drive (with 120 degree sensor phasing)

current required for all three N-channel devices is only 0.36 mA. Since the charging current needed for the three P-channel MOSFETs is delivered directly from the DC bus, that charge is not included when sizing the impedance of the gate drive bus voltage.

Although the average current drawn by the MOSFETs is quite small, charging their input capacitances takes high peak currents. Therefore, the filter capacitor, C4, essentially supplies the entire turn-on current, and the capacitor is refreshed through resistor, R7. When the supply dips to its lowest specified value for a 24 V system (18 V), the dropping resistor has only three volts across it. Using a 1 kΩ resistor will provide plenty of current to keep the capacitor charged and supply at least 1 mA to the zener to ensure good regulation. At high supply voltages the resistor will see a voltage of 15 V, a current of 15 mA, and a power dissipation of about a quarter of a watt. Therefore, a 1/2 watt resistor should be adequate. That's also a good power rating for the zener. For the 12 V supply, less voltage appears across the dropping resistor, so it can be a 1/4 watt device.

Capacitor C4 should be able to deliver the entire gate charge for a single cycle without a significant dip in its voltage. For 15 nC and a maximum allowable drop in voltage of 0.5 V, the capacitor should be 0.03 μF. Therefore, a 0.1 μF capacitor should be quite adequate for charging these N-channel MOSFETs. Since a MOSFET draws essentially no drive current after it has been turned on, the output duty cycle doesn't significantly affect drive requirements. Only switching frequency and charge affect average current.

Filter capacitor C4 can be eliminated if the series resistor R7 is small enough to satisfy gate drive requirements at any instant. The cost is higher losses in the zener diode and series resistor, which is likely to require devices with greater power ratings.

The gate drive deserves two other considerations. First, the IC sees the MOSFET as a capacitor in series with a parasitic inductor. Most of that inductance is in the current sensing resistor. If switching is rapid, transients and ringing in the gate drive loop are common. Voltage spikes on the gate drives greater than 0.7 V below ground can forward bias the

substrate of the MC33035. Three Schottky diodes D1-3, one from each lower drive output to ground, are required if substrate current exceeds 50 mA.

Problems may also arise if the gate drive impedance of the three lower devices is unnecessarily low. If little or no resistance is placed between the IC and the MOSFET, the gate drive loop may cause ringing during gate voltage transitions. Such ringing is amplified by the MOSFET, the gate drive loop may cause ringing during gate voltage transitions. Such ringing is amplified by the MOSFET and may cause unacceptably high levels of noise at the drain. The solution to the problem is to reduce the circuit's Q by inserting a series gate resistor. The minimum required value depends on circuit parasitics, so it is difficult to give recommendations. However, the resistance required to keep switching losses reasonably low is usually much larger than the resistance required to eliminate oscillations. In this circuit gate drive resistors with values less than 62 ohms caused oscillations.

There is a second reason to avoid very fast turn-on of the N-channels. When a lower device turns on, the P-channel in the upper leg of that same half bridge has been conducting current in its drain-source diode. Reverse recovery charge is swept out of the diode by the lower switch. If turn on speeds are high, large reverse recovery currents and rapid swings in drain-to-source voltage will produce EMI.

P-CHANNEL GATE DRIVE CURRENT

For standard MOSFETs it is desirable, but not absolutely necessary, to have 10 volts available to drive the gate. However, as drain currents decrease, the gate voltage needed to conduct those currents also falls. Depending on the device's transfer characteristics and the desired load current, gate voltages in the 7 to 8 V range may be acceptable. In a 24 V system there is plenty of voltage available to drive the gate, so R5 and R6 are sized to ensure that the P-channel receives – 10 V of gate drive even when the supply voltage drops to 18 V.

R5, 12 and 13 govern charging of the P-channels' input capacitances and thus control turn-on speed. Similarly, R6, R14 and R15 determine turn-off speed. Lowering the value of each resistor in the divider maintains the desired – 10 V gate drive and decreases the gate drive impedance at the expense of increased current and power dissipation in the resistors.

A common pitfall in the design of the P-channel drive is to assume that since the P-channels are switching at the motor's commutation frequency (a frequency much lower than the PWM frequency) they do not need the low impedance gate drive that the N-channels require. What is often missed is that whenever the drain-to-source voltage changes (due to the greater than 20 kHz switching frequency of an N-channel in the lower legs of the bridge), the upper gate drive must charge and discharge the P-channels' gate-to-drain capacitance. If the gate drive is not sufficient, the P-channel will briefly turn on, causing shoot through currents that dramatically increase switching losses. High voltage brushless motor control systems are especially prone to gate drive problems since the excursions in V_{DS} are so large.

Avoiding shoot through currents is easy. First, slow the turn-on of the N-channels to limit impressed dv/dt 's; second, keep the P-channels' gate drive impedance low, especially in the off-state. Adding capacitance across the P-channels' gate-to-source (0.01 μ F worked well in the application) is a simple way to give the gate drive a reservoir of charge that keeps the gate off when C_{dg} demands displacement current.

When a 12 V supply is used, the full supply voltage is impressed across the gate-to-source when the MOSFET is supposed to be on. This is done by shorting the pull down resistors, R5, R12 and R13. Since there is no longer any series resistance to limit gate current, turn-on speed will be much faster than turn-off speeds.

OVERCURRENT SENSING

The MC33035 has a comparator for detecting excessive load currents. A signal from a current sensing resistor common to all the N-channel sources is fed into Pin 9. The trip

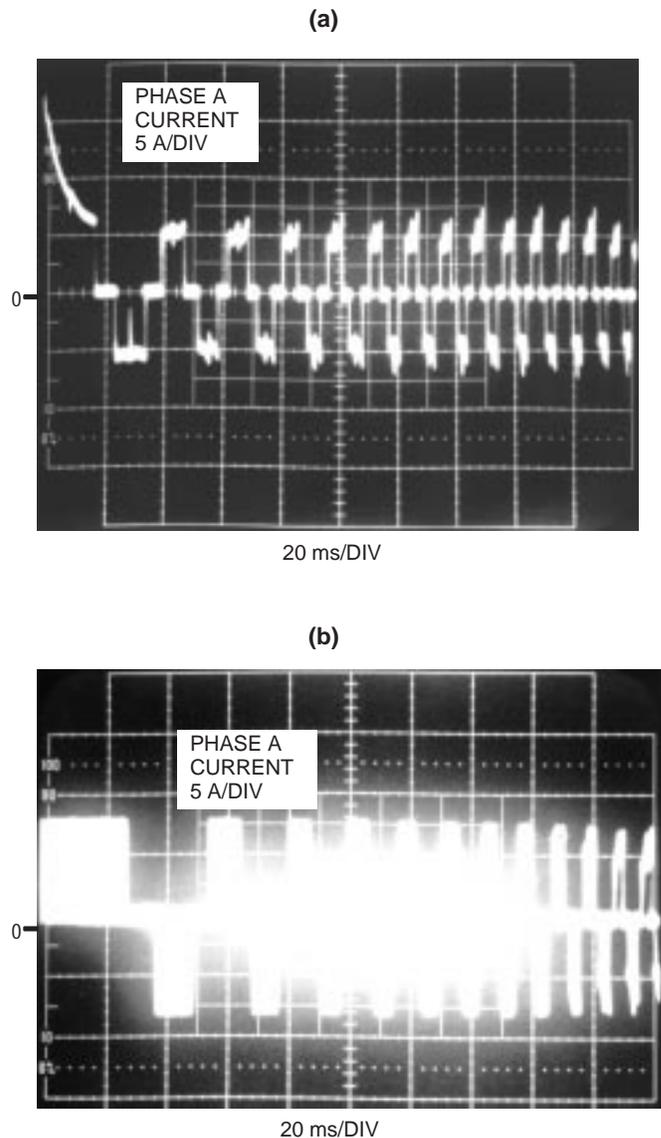


Figure 8. Motor Start Up Current Without (a) and With (b) Overcurrent Protection

threshold is internally set to 100 mV. If a greater trip voltage is desired, the MC33035 allows connection of an additional offset voltage to Pin 15. In this design a 1 W, 0.05 Ω resistor R21 is used to sense current, and the sense voltage is attenuated by a voltage divider. The values chosen for the voltage divider, 100 Ω (R8 in Figure 7) for the upper resistor and 33 Ω (R9) for the lower, set the current trip to 8 A.

Putting a small capacitor (C5) on the comparator input is a good way to keep noise or currents such as reverse recovery spikes of freewheeling diodes from tripping the overcurrent comparator. It is easy to see that the DC gain of the network is set by the resistive divider, but the time constant may not be obvious at a glance. The transfer function for the resistive divider network and the capacitor is:

$$\frac{V_o}{V_i} = \frac{R_9}{R_8 + R_9} \cdot \frac{1}{1 + \frac{C_5 R_8 R_9}{R_8 + R_9}}$$

where V_i is the voltage across the current sense resistor and V_o is the voltage appearing at the input of the comparator.

Therefore, the time constant is equal to the parallel combination of R8 and R9 times C. In this case τ is 2.4 μs (24 times 0.1 μF). Power MOSFETs are known to be able to

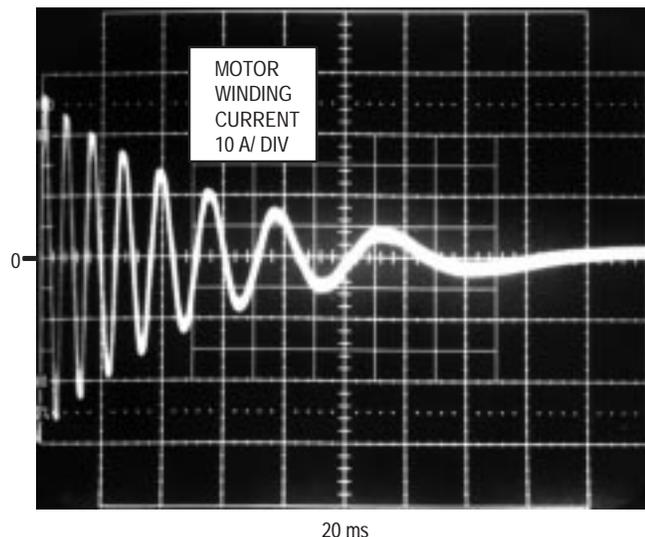


Figure 9. Application of Brake Signal May Produce Large Fault Currents

withstand high surge currents, especially if their duration is less than 10 microseconds. Therefore, a time constant of this magnitude gives adequate protection.

The benefit of over current protection is apparent in a comparison of Figures 8a and 8b. In figure 8a, where no current feedback is used, the start up current peak reaches 16 A, and in Figure 8b, current is bounded by the 8 A limit.

Another potentially stressful operating condition occurs when the motor is required to abruptly change direction of rotation. If no overcurrent sensing is used, currents are limited only by the winding resistance and the on-voltage of the power MOSFETs.

A third mode of operation that causes high currents is the brake mode. Upon application of the brake signal, all three bottom transistors are turned on, shorting the motor windings. Since current circulates between windings through the three N-channels and does not appear in the sense resistor, the MC33035 can not detect the high currents in the brake mode. Therefore, the MOSFETs must be sized to handle very large currents if the brake is used.

As Figure 9 shows, current peaks reach 35 A and last long enough to be of a reliability concern for the power transistors. Peak current is a function of the power transistor's and winding resistance and the motor's back EMF during braking. The time required for the current to decay depends on motor speed, motor winding resistance, frictional loading, and motor inertia. The photograph shows that current rings between windings until the energy stored in the spinning rotor is extinguished by dissipative elements.

FAULT INDICATOR AND OVERCURRENT LATCH

At any one of four fault conditions the Fault Output, Pin 14, is pulled low. Having an LED to indicate a fault is a handy diagnostic tool. To maintain roughly 1 mA in the LED, R11 is 2.2 kΩ in a 24 V system and 1 kΩ if the supply is 12 V.

Upon detection of a fault, it is often desirable to inhibit any further pulsing of the output transistors. This can be accomplished by tying the Fault Output to Pin 7, the Enable pin. A delay in the latch can be implemented by adding C6. The time constant of R10 and C6 fixes the delay before the system latches.

Figure 10 shows the layout of the component from the parts list on page 10.

Figure 11 is a photo of the 24 volt brushless motor control that was shown in the schematic of Figure 7.

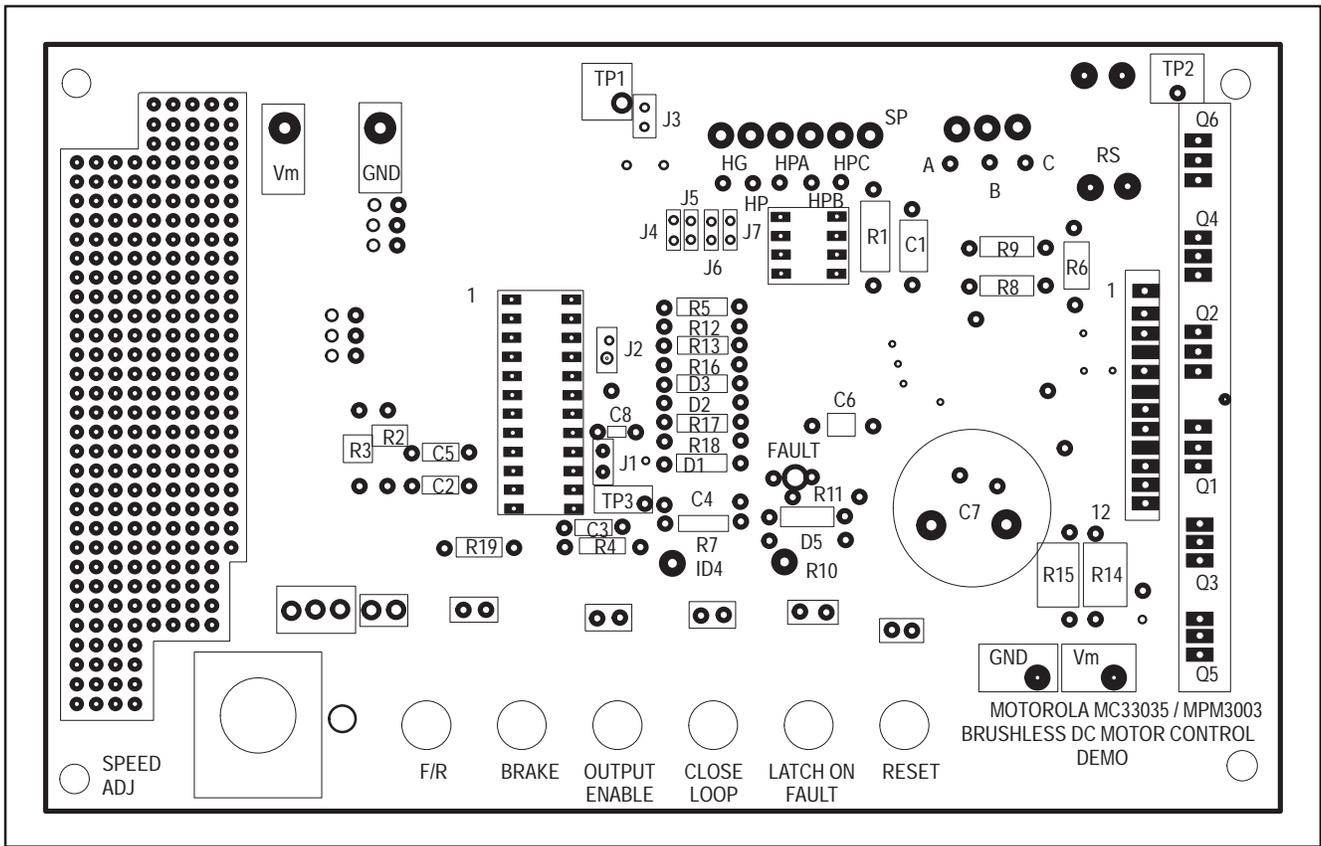


Figure 10. Printed Circuit Silk Screen of Three Phase Brushless Motor Control

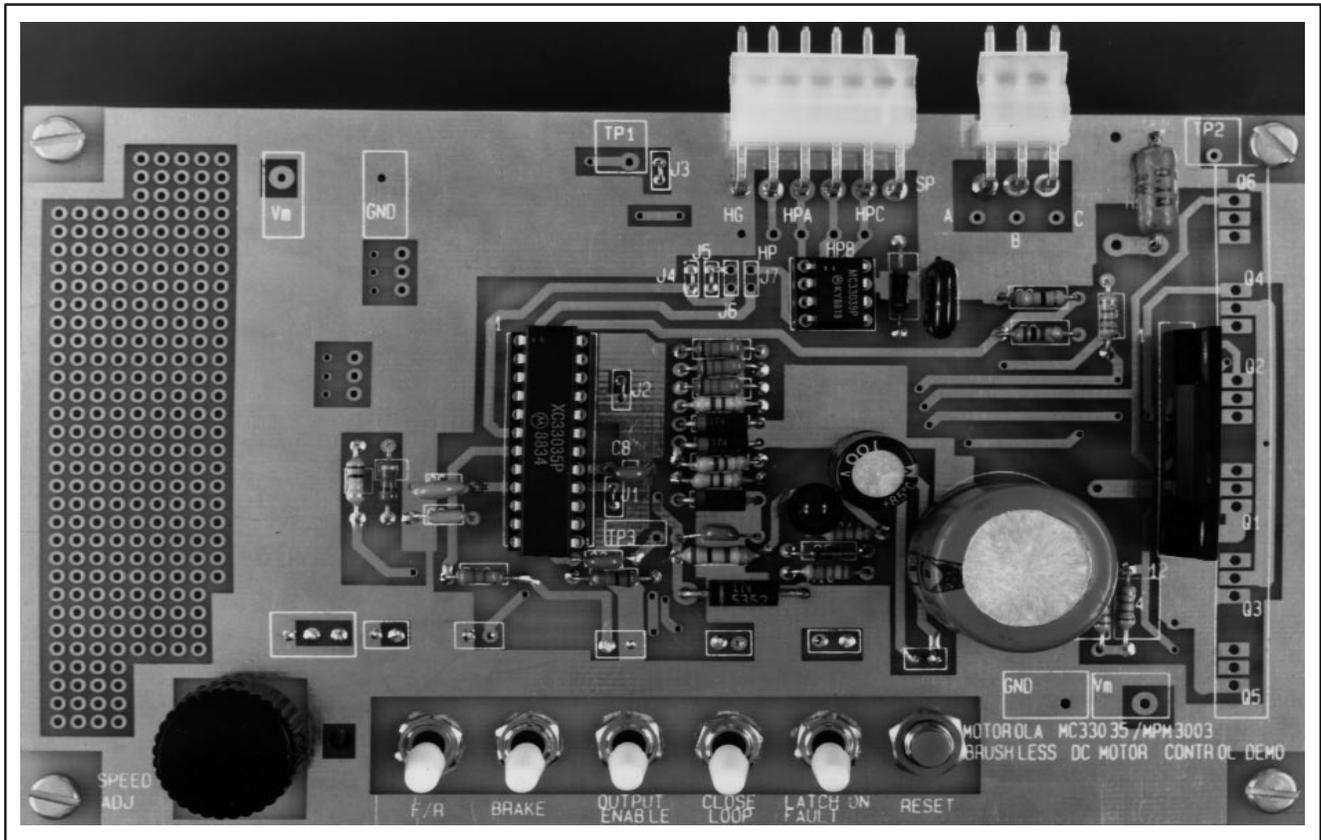


Figure 11. Brushless Motor Control Utilizing the MC33035, MC33039 and MPM3003

PARTS LIST FOR BRUSHLESS MOTOR CONTROLLER

Transistors and Integrated Circuits

MPM3003
MC33035
MC33039

Capacitors (uFd, 50 V unless otherwise noted)

C1	22 nF
C2	0.01 μ F
C3	0.1 μ F
C4	0.1 μ F
C5	0.001 μ F
C6	47 μ F (35 V tantalum or 50 V electrolytic)
C7	1000 μ F (Sprague Part #80D102P050HA5Ds or #673D108H050JJ9CS)
C8	0.1 μ F

Diodes

D1	1N5819
D2	1N5819
D3	1N5819
D4	1N5338A
D5	1N4148
D6	Red LED

5 SPDT Switches
1 Push Button Switch N.O.
3-Pin Connector for Hall Effect

Resistors (Ohms, 1/4 W, metal film unless otherwise noted)

R1	43 k Ω
R2	5.1 K
R3	100 K
R4	1 M Ω
R5	1 K
R6	1.1 K
R7	1 K, 0.5 W
R8	100 Ω
R9	33 Ω
R10	2.2K
R11	2.2 K
R12	1 K
R13	1 K
R14	1.1 K
R15	1.1 K
R16	470 Ω
R17	470 Ω
R18	470 Ω
R19	4.7 K
R20	10 K Pot.
R21	0.05 Ω , 1 W low inductance

Sensors (1 each)
6-Pin Connector for Motor (1 each)

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