

**BRUSHLESS DC MOTOR SPEED CONTROL CIRCUIT
DESIGN**

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PREFACE

This study has focused to Closed Loop Speed Control Application of a BLDC Motor. I have tried to design a control circuit based upon MC33035 microcontroller produced by the MOTOROLA and gave the theory of operation regarding the application.

I am sincerely grateful to my instructor Asst. Prof. Dr. Özgür ÜSTÜN who made many contributions to this Project. I have learned many things about the PCB design and Power Electronics from him.

I would also like to thank my department manager MSc. Civil Engineer Ali Levent Kuzum for his helps and sensibility in my difficult thesis process. His goodwill was very motivating in my stressfull days.

My valuable thanks go to my family for their limitless tolerance. Their approach was very helpful and valuable to me.

Lastly, I like to thank my friend Can Gökçe for introducing and providing me the Altium Designer Software which was very useful in the process of designing the PCB circuit.

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ABBREVIATIONS

BLDC	: Brushless DC Motor
EMF	: Electromotor Force
EPROM	: Erasable Programmable read-only memory
PWM	: Pulse Wide Modulation
PI	: Proportional Integral
PID	: Proportional Integral Derivative
MOSFET	: Metal-Oxide-Semiconductor Field-Effect Transistor
LED	: Light-Emitting Diode
AMP	: Amplificator
IC	: Integrated Circuit
PCB	: Printed Circuit Board
GND	:Ground
PWR	:Power
FWD	:Forward
REV	:Reverse

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SYMBOL LIST

$e(k)$: Input error in step k

$w(k)$: Desired value in step k

$m(k)$ = Measured value in step k

$u(k)$ = Controller output in step k

$u_p(k)$ = Proportional output portion in step k

$u_i(k)$ = Integral output portion in step k

$u_i(k-1)$ = Integral output portion in step k - 1

T_i = Integral time constant

T = Sampling time

K_c = Controller gain

τ_{ind} = Torque induced on the rotor

B_R = Rotor Magnetic Field

B_S = Stator Magnetic Field

$U_{DCB} = DC$

U_{DCB} = DC bus voltage

C_T = Timing capacitor

R_T = Timing Resistor

i_a = A phase current

i_b = B phase current

i_c = C phase current

R_a = A phase stator winding resistance

R_b = B phase stator winding resistance

R_c = C phase stator winding resistance

L_a = A phase stator winding inductance

L_b = B phase stator winding inductance

L_c = C phase stator winding inductance

E_a = Induced voltage on the a phase stator winding

E_b = Induced voltage on the b phase stator winding

E_c = Induced voltage on the c phase stator winding

ω = Angular velocity

M = Torque

M_m = Motor torque

M_y = Load torque

k = Motor constant

k_a = Motor constant for A phase

k_b = Motor constant for B phase

k_c = Motor constant for C phase

θ = Rotor position angle

J = Moment of Inertia

FIRÇASIZ DOĞRU AKIM MOTORU HIZ KONTROL DEVRESİ TASARIMI

ÖZET

Günümüzde fırçasız doğru akım motorları gün geçtikçe önem kazanmakta ve yaygınlaşmaktadır. Fırçalı doğru akım motorlarıyla karşılaştırıldığında fırçasız doğru akım motorlarında komütasyon işlemi mekanik olarak fırçalarla değil elektronik olarak yapılmaktadır ve bu kullanılabilirliklerini arttırır. Fırçasız doğru akım motorlarında rotorun manyetik yapısı tarafından üretilen manyetik alan motorun verimliliğini arttırır bu yüzde fırçasız doğru akım motorlarının çok geniş bir kullanım alanı vardır. Bu tez çalışması 3 fazlı bir fırçasız doğru akım motorunun kapalı çevrim hız kontrolünü yapan bir sürücü devrenin yapılması ve devrenin çalışma ilkesinin açıklanmasını hedef almıştır. Yakın zamana kadar fırçasız doğru akım motorlarının avantajlarından yararlanmak isteyen motor üreticileri önemli bir sorunla karşılaşmışlardır. Bu sorun Hall sensörlerden gelen digital sinyalleri çözümleyecek ve bunun yanında bir motorun sorunsuz çalışması için gerekli bazı fonksiyonları yerine getirecek bir tümdevrenin yokluğuydu. Bu fonksiyonları farklı komponentleri kullanarak gerçekleştirmek alternatif bir çözüm gibi görünse de devrenin alanın büyümesi ve maliyetin artması tüm bu işlemleri tek bir tasarımda çözecek bir entegreyi gerekli kılmıştır. MOTOROLA firmasının üretmiş olduğu MC33035 tümdevresi bütün bu ihtiyaçlara cevap verebilecek, istenilen fonksiyonları sağlayabilecek bir entegredir. İçerisinde ihtiva ettiği decoder yapısıyla hall sensörlerden gelen sinyalleri çözümleyerek motora komütasyon verebilmektedir. Ancak kapalı çevrim hız kontrolünü gerçekleştirememektedir. Bu sorun da yine MOTOROLA firmasının ürettiği MC33039 elektronik takometreyle çözümlenebilir. Devrenin uygulama aşaması devre şemasının ALTIUM DESIGNER bilgisayar programıyla oluşturulması ve yine aynı programla PCB şemasının oluşturulmasıyla başlamıştır. PCB devre basıldıktan sonra devre komponentleri PCB devreye lehimlenmek suretiyle yerleştirilmiştir. Daha sonra bu devre özel bir kontrolör yardımıyla minyatür bir fırçasız DA motorunun hız denetiminde kullanılmıştır. Deneysel sonuçlar alınmış ve karşılaştırmalar yapılmıştır.

BRUSHLESS DC MOTOR SPEED CONTROL CIRCUIT DESIGN

SUMMARY

BLDC motors are very popular in a wide array of applications. Compared to a DC motor, the BLDC motor uses an electric commutator, replacing the mechanical commutator and making it more reliable than the DC motor. In BLDC motors, rotor magnets generate the rotor's magnetic flux, allowing BLDC motors to achieve higher efficiency. Therefore, BLDC motors may be used in high-end white goods (refrigerators, washing machines, dishwashers, etc.), high-end pumps, fans, and other appliances that require high reliability and efficiency.

This thesis describes the design of a 3-phase brushless DC (BLDC) motor drive based on MOTOROLA's MC33035 microprocessor. 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. 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. 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, also. But it has not an ability to perform closed loop speed control. In this point the application design solves the problem using a Closed Loop Brushless Motor Adapter MOTOROLA MC33039. The MC33039 is a high performance closed-loop speed control adapter specifically designed for use in brushless DC motor control systems. Implementation will allow precise speed regulation without the need for a magnetic or optical tachometer.

The application part of the thesis consist of the designing the circuit with Altium Designer software and obtaining the PCB format. After that this circuit has been used for the speed control of a miniature BLDC motor by means of a special controller. Has been obtained experimental results and some comparisons has been performed.

1. INTRODUCTION

This project merges the theory, design, construction and testing of a two quadrant adjustable DC speed driver for a brushless DC motor. The proposed drive system will consist of two ICs MC33035 (for decoding and control purposes) and MC33039 (speed signal determining IC) produced by the MOTOROLA and a power module MSK3003 produced by the MS Kennedy.

1.1 Brushless DC (BLDC) Motor Drives

Nowadays, the speed control of a dc motor is accomplished by terminal voltage control. Most of the modern servomotors are brushless type motors (brushless ac or brushless dc).

A conventional DC motor can operate in four different quadrants by changing the polarity of voltage and direction of current. These four modes are: forward motoring (positive voltage and current), forward regeneration (positive voltage, negative current), reverse motoring (negative voltage and current) and reverse regeneration (negative voltage and positive current). The term regeneration (also known as regenerative braking) means operating the motor as a generator. This brakes the motor by converting its mechanical energy into electrical energy and sending it back to the batteries. As mentioned above, it is designed two quadrant operation in this application. In brushless dc motors, this can be accomplished by PWM control of terminal voltage and inserting the direction information to motor drive decoder. [1]

A Brushless DC motor driver is more complicated than brushed DC motor driver. Because the motor cannot commutate the windings, so the control circuit and software must control the current flow correctly to keep the motor turning smoothly. There are two basic types of Brushless DC motors; sensor and sensorless. It is critical to know the position of the rotor to energize the correct winding of the motor therefore some method of detecting the motor position is required. A sensor motor directly reports the controller by Hall Effect sensors. Driving a sensor motor requires

a look-up table. Hall sensors send logic signals to IC, then IC electronically commutates the motor. A sensorless motor requires that the induced voltage in the un-driven winding be sensed and used to determine the current speed of the motor. Then, the next commutation pattern can be determined by a time delay from the previous pattern.

Sensorless motors are simpler to build due to the lack of the sensors, but they are more complicated to drive. A sensorless motor performs very well in applications that do not require the motor to start and stop. A sensor motor would be a better choice in applications that must periodically stop the motor. However, the improvements in Hall Effect sensor technology allow the higher temperatures and small volumes. [2]

In designed system, a special IC (MC33039) is gathering the data from sensors and giving the actual speed information as a frequency output. A low pass filter can be used to obtain the analog speed information signal.

2. BLDC MOTORS

A brushless DC motor (BLDC) is an AC synchronous electric motor that looks very similar to a DC motor. Sometimes the difference of BLDC motors is explained as being an electronically controlled commutation system, instead of a mechanical commutation but this is misleading, because as physically the two motors are completely different.

2.1 General Characteristics

Brushless Direct Current (BLDC) motors are one of the motor types rapidly gaining popularity. BLDC motors are used in industries such as Appliances, Automotive, Aerospace, Consumer, Medical, Industrial Automation Equipment and Instrumentation. As the name implies, BLDC motors do not use brushes for commutation; instead, they are electronically commutated. BLDC motors have many advantages over brushed DC motors and induction motors. Some of these are:

- Better speed versus torque characteristics
- High dynamic response
- High efficiency
- Long operating life
- Noiseless operation
- Higher speed ranges

In addition, the ratio of torque delivered to the size of the motor is higher, making it useful in applications where space and weight are important.

2.2 Construction

BLDC motors are typically of synchronous motor as mentioned before. This means that the magnetic fields generated by the stator and the rotor rotate at the same frequency.

There are 2-phase and 3-phase BLDC motor configurations. 3-phase motors are the most popular and widely used.

2.2.1 Stator Structure

Traditionally, BLDC stator resembles stator of induction motor, however the windings are distributed in a different way. Most BLDC motors have three stator windings connected in star form. There are two types of stator windings variants: trapezoidal and sinusoidal motors. This difference comes from the basis of the interconnection of the coils in the stator windings and these two windings form give the different types of back Electromotive Force (EMF). As their names indicate, the trapezoidal motor gives a back EMF in trapezoidal form and the sinusoidal motor give back EMF in sinusoidal, as shown in Figure 2.1 and Figure 2.2. In addition to the back EMF, in different types of motor the phase current also has trapezoidal and sinusoidal variations. This makes the torque output of a sinusoidal motor smoother than that of a trapezoidal motor. However, this comes with an extra cost, cause the sinusoidal motors have extra winding interconnections because of the coils distribution on the stator periphery.

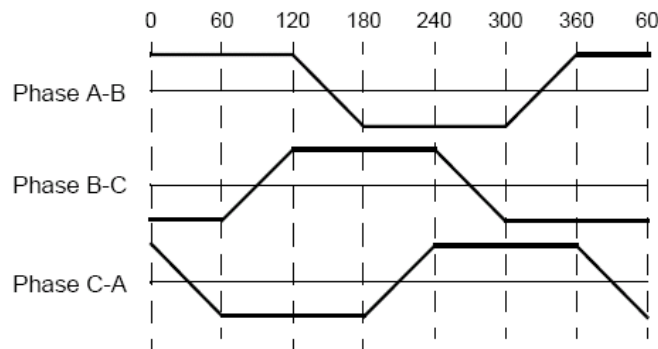


Figure 2.1: Trapezoidal Back EMF

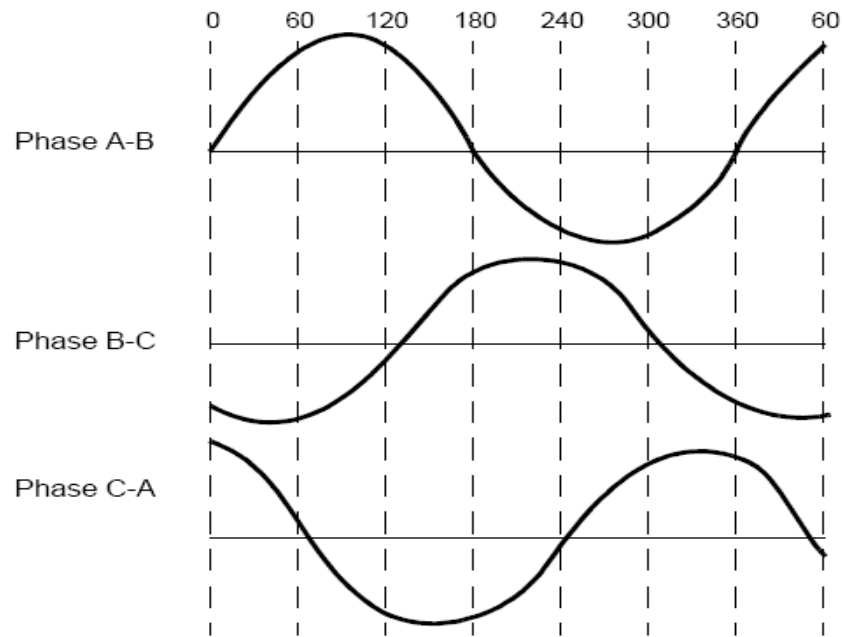


Figure 2.2: Sinusoidal Back EMF

2.2.2 Rotor Structure

The rotor is made of permanent magnet and can vary from two poles to eight poles. Ferrite magnets are traditionally used to make permanent magnets. The ferrite magnets are less expensive but they have the disadvantage of low flux density for a given volume. In contrast, the alloy material has high magnetic density per volume and enables the rotor to compress further for the same torque. Figure 2.3 shows cross sections of different arrangements of magnets in a rotor.

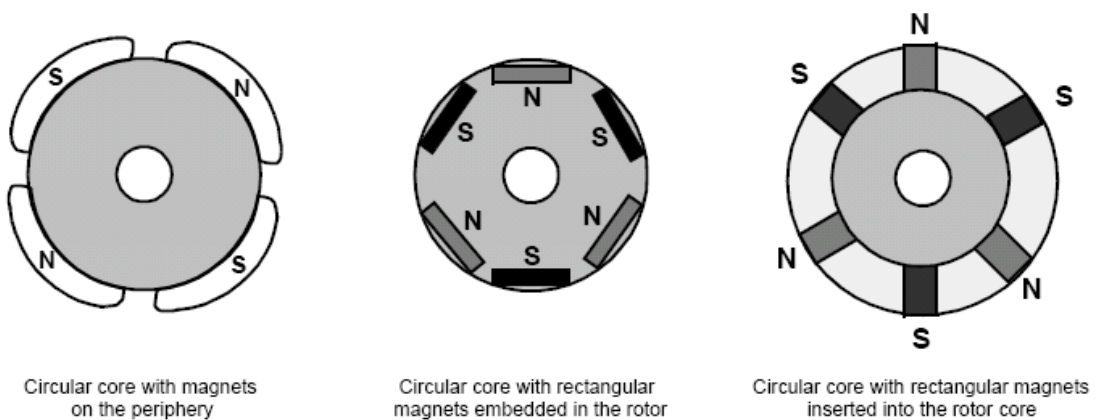


Figure 2.3: Roto Magnet Cross Section

2.2.3 Hall Sensors

Unlike a brushed DC motor, the commutation of a BLDC motor is controlled electronically. To rotate the BLDC motor, the stator windings should be energized in a sequence. It is important to know the rotor position in order to understand which winding will be energized. Rotor position is sensed using Hall effect sensors embedded into the stator. Most BLDC motors have three Hall sensors embedded into the stator on the non-driving end of the motor. Whenever the rotor magnetic poles pass near the Hall sensors, they give a high or low signal. Based on the combination of these three Hall sensor signals, the exact sequence of commutation can be determined.

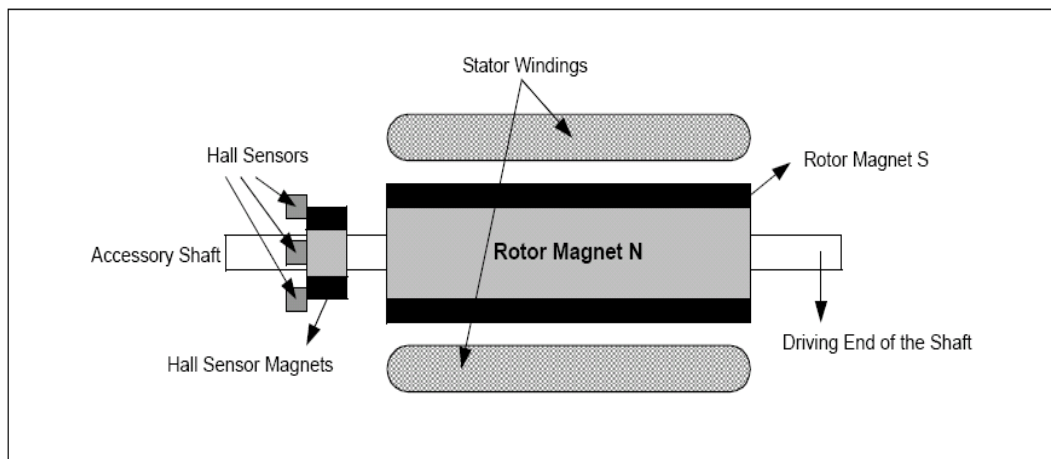


Figure 2.4: BLDC Motor Transfer Section

Figure 2.4 shows a transverse section of a BLDC motor with a rotor that has a N and S permanent magnets. Hall sensors are embedded into the stationary part of the motor. Embedding the Hall sensors into the stator is a complex process because any misalignment in these Hall sensors, with respect to the rotor

magnets, will generate an error in determination of the rotor position. The Hall sensors are normally mounted on a PC board and fixed to the enclosure cap on the non-driving end. This enables users to adjust the complete assembly of Hall sensors, to align with the rotor magnets, in order to achieve the best performance. The Hall sensors can be mounted at 60° or 120° shifted each others.

2.3 Mathematical Model of Brushless DC Motor

In this section, a mathematical model of brushless motor is improved. This mathematical model is independent of pole number, winding form, rotor shape and electronic switches. In the model, it is considered that the motor works in unsaturated area and electronic switches are ideal. System model is 4th order and variables are three phase currents and motor speed. As depending on the rotor position, differential equations are obtained for every rotor position.

To improve the mathematical model, for every rotor position, commutation and conduction differential equations are obtained. As defining the commutation and conduction equivalent circuit, differential equations for the mathematical model of the system are derived.

In the Table 2.1, commutation intervals are given according to the rotor position for a one electrical rotation.

Table 2.1: Commutation Intervals Depending on Rotor Position for One Electrical Rotation

Rotor Position	Pair of Switch on Conduction	Equation Number
$\frac{\pi}{6} < \theta_e < \frac{\pi}{2}$	S1-S6	(2.1)
$\frac{\pi}{2} < \theta_e < \frac{5\pi}{6}$	S1-S2	(2.2)
$\frac{5\pi}{6} < \theta_e < \frac{7\pi}{6}$	S3-S2	(2.3)

$\frac{7\pi}{6} < \theta_e < \frac{3\pi}{2}$	S3-S4	(2.4)
$\frac{3\pi}{2} < \theta_e < \frac{11\pi}{6}$	S5-S4	(2.5)
$\frac{11\pi}{6} < \theta_e < \frac{\pi}{6}$	S5-S6	(2.6)

Here the passing from S5-S6 conduction step to S1-S6 step will be analysed as improving the mathematical model of the system. Figure 2.5 and 2.6 show the conduction (red arrow) and commutation (blue arrow) times respectively when the rotor is positioned on $\frac{\pi}{6} < \theta_e < \frac{\pi}{2}$ interval. As is seen in the figure, after finishing the conduction of the C phase, A phase conduction is started. The commutation circuit in this rotor position corresponds to the turning off C phase current equivalent circuit. This equivalent circuit is valid until the C phase current becomes zero.

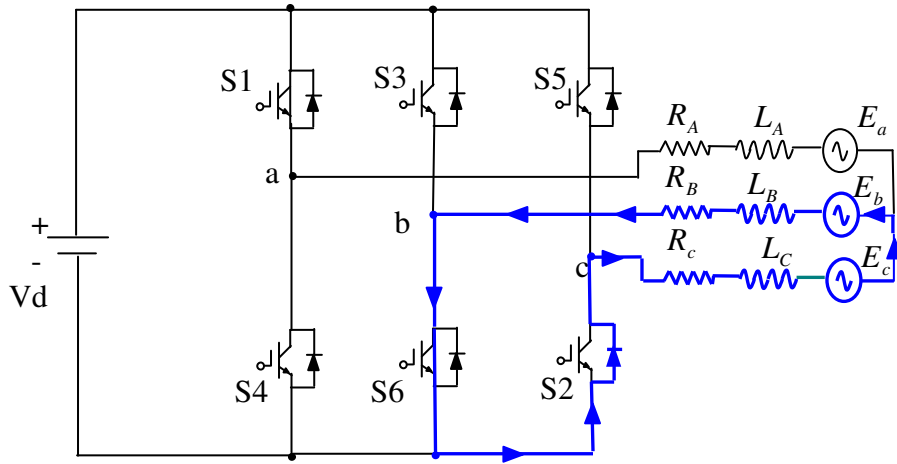


Figure 2.5: Conduction equivalent circuit for the $\frac{\pi}{6} < \theta_e < \frac{\pi}{2}$ interval

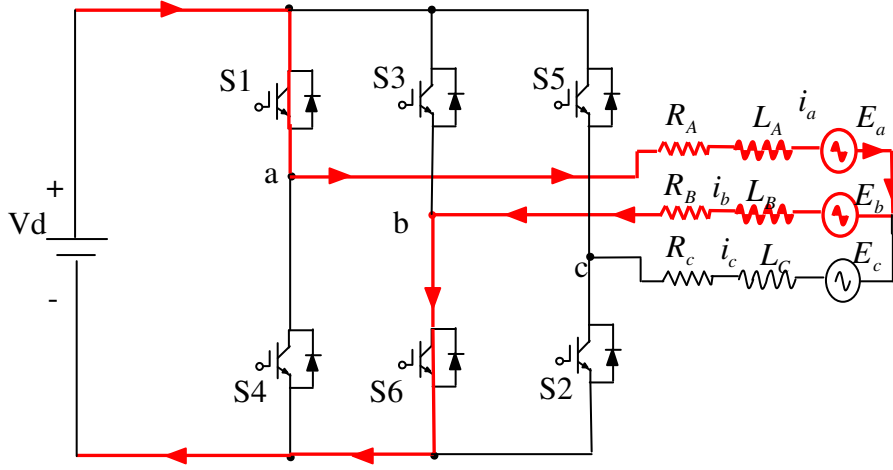


Figure 2.6: Conduction equivalent circuit for the $\frac{\pi}{6} < \theta_e < \frac{\pi}{2}$ interval

The required differential equations are obtained by the help of conduction and commutation equations and star node point equation. According to the star node point equation, sum of the three phase currents is zero [12].

$$i_a + i_b + i_c = 0, R = R_A = R_B = R_C, L = L_A = L_B = L_C \quad (2.7)$$

$$0 = Ri_c + (L + M) \frac{di_c}{dt} + E_c - E_b - Ri_b - (L + M) \frac{di_b}{dt} \quad (2.8)$$

$$V_d = Ri_a + (L + M) \frac{di_a}{dt} + E_a - E_b - Ri_b - (L + M) \frac{di_b}{dt} \quad (2.9)$$

From the 2.7 equation, replacing the C phase current on the 2.8 equation

$$\frac{di_a}{dt} = -2 \frac{di_b}{dt} - \frac{1}{(L + M)} [2.Ri_b + Ri_a - E_c + E_b] \quad (2.10)$$

is obtained. Derivative expression of the A phase current is replaced on 2.9 equation,

$$\frac{di_b}{dt} = \frac{1}{3(L + M)} [-V_d - 3Ri_b + E_a + E_c - 2E_b] \quad (2.11)$$

is obtained. This equation consists of the B phase current derivative expression, is replaced on 2.10 equation

$$\frac{di_a}{dt} = \frac{1}{3(L+M)} [2V_d - 3.Ri_a + E_b + E_c - 2E_a] \quad (2.12)$$

is obtained. From the star node point equation,

$$\frac{di_a}{dt} + \frac{di_b}{dt} + \frac{di_c}{dt} = 0 \quad (2.13)$$

$$\frac{di_c}{dt} = \frac{1}{3(L+M)} [-V_d - 3.Ri_c + E_a + E_b - 2E_c] \quad (2.14)$$

C phase current derivative equation is obtained.

After the C phase current becomes zero, the Figure 2.5 circuit will be valid. While conduction of S1-S6 switches, C phase current are zero thus star node point equation is given by the 2.15 equation.

$$i_a + i_b = 0 \quad (2.15)$$

As considering the Fig. 2.6 circuit, if the A phase current is wrote as depending on the B phase current,

$$\frac{di_b}{dt} = \frac{1}{2(L+M)} [-V_d - 2.Ri_b + E_a - E_b] \quad (2.16)$$

$$\frac{di_a}{dt} = \frac{1}{2(L+M)} [V_d - 2.Ri_a + E_b - E_a] \quad (2.17)$$

equations are obtained. 3.16 and 3.17 equations are defined as the phase current equations of brushless DC motor system, for the $\frac{\pi}{6} < \theta_e < \frac{\pi}{2}$ electrical rotor position interval.

To complete the mathematical model of the system in this step, it is needed to write the differential equations for mechanical side. Mechanical differential equations are same for all the conduction steps and given as follow,

$$M_m - M_y = J \frac{d\omega}{dt} + B\omega \quad (2.18)$$

$$\frac{d\omega}{dt} = \frac{1}{(J_m + J_y)} [M_m - M_y - (B_m + B_y)\omega] \quad (2.19)$$

Instant torque value of brushless DC Motor is given by 2.20 equation.

$$M(t) = k_a i_a + k_b i_b + k_c i_c \quad (2.20)$$

Electrical and mechanical equations for the other conduction steps can be obtained by using the same method.

For the $\frac{\pi}{2} < \theta_e < \frac{5\pi}{6}$ rotor position interval, the commutation equations:

$$\frac{di_a}{dt} = \frac{1}{3(L+M)} [V_d - 3Ri_a + \omega(k_b + k_c - 2k_a)] \quad (2.21)$$

$$\frac{di_b}{dt} = \frac{1}{3(L+M)} [V_d - 3Ri_b + \omega(k_a + k_c - 2k_b)] \quad (2.22)$$

$$\frac{di_c}{dt} = \frac{1}{3(L+M)} [-2V_d - 3Ri_c + \omega(k_a + k_b - 2k_c)] \quad (2.23)$$

$$\frac{d\omega}{dt} = \frac{1}{J} [k_a i_a + k_b i_b + k_c i_c - M_y] \quad (2.24)$$

For the $\frac{\pi}{2} < \theta_e < \frac{5\pi}{6}$ rotor position interval, the conduction equations:

$$\frac{di_a}{dt} = \frac{1}{2(L+M)} [V_d - 2Ri_a + \omega(k_c - k_a)] \quad (2.25)$$

$$i_b = 0 \quad (2.26)$$

$$\frac{di_c}{dt} = \frac{1}{2(L+M)} [-V_d - 2Ri_c + \omega(k_a - k_c)] \quad (2.27)$$

$$\frac{d\omega}{dt} = \frac{1}{J} [k_a i_a + k_c i_c - M_y] \quad (2.28)$$

For the $\frac{5\pi}{6} < \theta_e < \frac{7\pi}{6}$ rotor position interval, the commutation equations:

$$\frac{di_a}{dt} = \frac{1}{3(L+M)} [-V_d - 3Ri_a + \omega(k_b + k_c - 2k_a)] \quad (2.29)$$

$$\frac{di_b}{dt} = \frac{1}{3(L+M)} [2V_d - 3Ri_b + \omega(k_a + k_c - 2k_b)] \quad (2.30)$$

$$\frac{di_c}{dt} = \frac{1}{3(L+M)} [-V_d - 3Ri_c + \omega(k_a + k_b - 2k_c)] \quad (2.31)$$

$$\frac{d\omega}{dt} = \frac{1}{J} [k_a i_a + k_b i_b + k_c i_c - M_y] \quad (2.32)$$

For $\frac{5\pi}{6} < \theta_e < \frac{7\pi}{6}$ rotor position interval, the conduction equations:

$$i_a = 0 \quad (2.33)$$

$$\frac{di_b}{dt} = \frac{1}{2(L+M)} [V_d - 2Ri_b + \omega(k_c - k_b)] \quad (2.34)$$

$$\frac{di_c}{dt} = \frac{1}{2(L+M)} [-V_d - 2Ri_c + \omega(k_b - k_c)] \quad (2.35)$$

$$\frac{d\omega}{dt} = \frac{1}{J} [k_b i_b + k_c i_c - M_y] \quad (2.36)$$

For $\frac{5\pi}{6} < \theta_e < \frac{3\pi}{2}$ rotor position interval, the commutation equations:

$$\frac{di_a}{dt} = \frac{1}{3(L+M)} [-2V_d - 3Ri_a + \omega(k_b + k_c - 2k_a)] \quad (2.37)$$

$$\frac{di_b}{dt} = \frac{1}{3(L+M)} [V_d - 3Ri_b + \omega(k_a + k_c - 2k_b)] \quad (2.38)$$

$$\frac{di_c}{dt} = \frac{1}{3(L+M)} [V_d - 3Ri_c + \omega(k_a + k_b - 2k_c)] \quad (2.39)$$

$$\frac{d\omega}{dt} = \frac{1}{J} [k_a i_a + k_b i_b + k_c i_c - M_y] \quad (2.40)$$

For $\frac{5\pi}{6} < \theta_e < \frac{3\pi}{2}$ rotor position interval, the conduction equations:

$$\frac{di_a}{dt} = \frac{1}{2(L+M)} [-V_d - 2Ri_a + \omega(k_b - k_a)] \quad (2.41)$$

$$\frac{di_b}{dt} = \frac{1}{2(L+M)} [V_d - 2Ri_b + \omega(k_a - k_b)] \quad (2.42)$$

$$i_c = 0 \quad (2.43)$$

$$\frac{d\omega}{dt} = \frac{1}{J} [k_a i_a + k_b i_b - M_y] \quad (2.44)$$

For $\frac{3\pi}{2} < \theta_e < \frac{11\pi}{6}$ rotor position interval, the commutation equations:

$$\frac{di_a}{dt} = \frac{1}{3(L+M)} [-V_d - 3Ri_a + \omega(k_b + k_c - 2k_a)] \quad (2.45)$$

$$\frac{di_b}{dt} = \frac{1}{3(L+M)} [-V_d - 3Ri_b + \omega(k_a + k_c - 2k_b)] \quad (2.46)$$

$$\frac{di_c}{dt} = \frac{1}{3(L+M)} [2V_d - 3Ri_c + \omega(k_a + k_b - 2k_c)] \quad (2.47)$$

$$\frac{d\omega}{dt} = \frac{1}{J} [k_a i_a + k_b i_b + k_c i_c - M_y] \quad (2.48)$$

For $\frac{3\pi}{2} < \theta_e < \frac{11\pi}{6}$ rotor position interval, the conduction equations:

$$\frac{di_a}{dt} = \frac{1}{2(L+M)} [-V_d - 2Ri_a + \omega(k_c - k_a)] \quad (2.49)$$

$$i_b = 0 \quad (2.50)$$

$$\frac{di_c}{dt} = \frac{1}{2(L+M)} [-V_d - 2Ri_c + \omega(k_b - k_c)] \quad (2.51)$$

$$\frac{d\omega}{dt} = \frac{1}{J} [k_a i_a + k_c i_c - M_y] \quad (2.52)$$

For $\frac{11\pi}{6} < \theta_e < \frac{\pi}{6}$ rotor position interval, the commutation equations:

$$\frac{di_a}{dt} = \frac{1}{3(L+M)} [V_d - 3Ri_a + \omega(k_b + k_c - 2k_a)] \quad (2.53)$$

$$\frac{di_b}{dt} = \frac{1}{3(L+M)} [-2V_d - 3Ri_b + \omega(k_a + k_c - 2k_b)] \quad (2.54)$$

$$\frac{di_c}{dt} = \frac{1}{3(L+M)} [V_d - 3Ri_c + \omega(k_a + k_b - 2k_c)] \quad (2.55)$$

$$\frac{d\omega}{dt} = \frac{1}{J} [k_a i_a + k_b i_b + k_c i_c - M_y] \quad (2.56)$$

For $\frac{11\pi}{6} < \theta_e < \frac{\pi}{6}$ rotor position interval, the conduction equations:

$$i_a = 0 \quad (2.57)$$

$$\frac{di_b}{dt} = \frac{1}{2(L+M)} [-V_d - 2Ri_b + \omega(k_c - k_b)] \quad (2.58)$$

$$\frac{di_c}{dt} = \frac{1}{2(L+M)} [-V_d - 2Ri_c + \omega(k_b - k_c)] \quad (2.59)$$

$$\frac{d\omega}{dt} = \frac{1}{J} [k_b i_b + k_c i_c - M_y] \quad (2.60)$$

To complete the mathematical model of the system, it is needed to express induced voltage on every phase for every conduction step. If star point is considered reference point, the instant value of the induced voltage is the function of motor constant, angular rotor velocity and rotor position. Motor constant depends on rotor position in brushless dc motor. In this concept, induced voltage on the phase windings for the unit velocity is improved in accordance with general mathematical model expression.

Motor constant change for A phase:

$$0 < \theta_e < \alpha \quad k_a = \frac{k\theta_e}{\alpha} \quad (2.61)$$

$$0 < \theta_e < (\alpha + \beta) \quad k_a = k \quad (2.62)$$

$$(\alpha + \beta) < \theta_e < (\pi + \alpha) \quad k_a = \frac{-k(\theta_e - \pi)}{\alpha} \quad (2.63)$$

$$(\pi + \alpha) < \theta_e < (\pi + \alpha + \beta) \quad k_a = -k \quad (2.64)$$

$$(\pi + \alpha + \beta) < \theta_e < 2\pi \quad k_a = \frac{k(\theta_e - 2\pi)}{\alpha} \quad (2.65)$$

Motor constant change for B phase:

$$0 < \theta_e < \frac{2\pi}{3} - \alpha \quad k_b = -k \quad (2.66)$$

$$\frac{2\pi}{3} - \alpha < \theta_e < \frac{5\pi}{3} - \alpha - \beta \quad k_b = \frac{k(\theta_e - \frac{2\pi}{3})}{\alpha} \quad (2.67)$$

$$\frac{5\pi}{3} - \alpha - \beta < \theta_e < \frac{5\pi}{3} - \alpha \quad k_b = k \quad (2.68)$$

$$\frac{5\pi}{3} - \alpha < \theta_e < \frac{8\pi}{3} - \alpha - \beta \quad k_b = \frac{-k(\theta_e - \frac{5\pi}{3})}{\alpha} \quad (2.69)$$

$$\frac{8\pi}{3} - \alpha - \beta < \theta_e < 2\pi \quad k_b = -k \quad (2.70)$$

Motor constant change for C phase:

$$0 < \theta_e < \frac{\pi}{3} - \alpha \quad k_c = k \quad (2.71)$$

$$\frac{\pi}{3} - \alpha < \theta_e < \frac{4\pi}{3} - \alpha - \beta \quad k_c = \frac{-k(\theta_e - \frac{\pi}{3})}{\alpha} \quad (2.72)$$

$$\frac{4\pi}{3} - \alpha - \beta < \theta_e < \frac{4\pi}{3} - \alpha \quad k_c = -k \quad (2.73)$$

$$\frac{4\pi}{3} - \alpha < \theta_e < \frac{7\pi}{3} - \alpha - \beta \quad k_c = \frac{k(\theta_e - \frac{4\pi}{3})}{\alpha} \quad (2.74)$$

$$\frac{7\pi}{3} - \alpha - \beta < \theta_e < 2\pi \quad k_c = k \quad (2.75)$$

2.4 Torque Equation of Brushless DC Motor

To explain the torque generation of brushless DC motors, it is necessary to understand the characteristics of movement voltages induced on stator windings [12].

Magnetic Flux induced on just one coil is given by:

$$\psi = (\pi r l) B_m, (-\pi/2 \leq \theta \leq \pi/2) \quad (2.76)$$

The Movement Voltage induced on the coil:

$$e = \frac{d\psi}{dt} = \frac{d\psi}{d\theta} \cdot \frac{d\theta}{dt} = \frac{\pi r l B_m \omega_r}{\pi/2} = 2B_m l r \omega_r \quad (2.77)$$

$$\frac{d\theta_r}{dt} = \omega_r, \quad \theta_e = p\theta_r \quad (2.78)$$

are obtained. Total EMF induced on stator windings given by the following equation:

$$E = 2NB_m l r \omega_r \quad (2.79)$$

and the Torque generated by a brushless DC motor is given analytical as follow,

$$M_{total} = \frac{E_a i_a + E_b i_b + E_c i_c}{\omega_r} \quad (2.80)$$

$$M_{total} - M_y = J \frac{d\omega_r}{dt} + B\omega_r \quad (2.81)$$

2.5 Fundamental of BLDC Motor Operation

To simplify the explanation of how to operate a three-phase BLDC motor, we may consider that BLDC motor has only three coils. To make the motor rotate, for each commutation sequence one of the three windings energized by positive power (current enters into the winding), the second winding is negative (current exits the winding) and the third one is non-energized condition. Torque is produced because of the interaction between the magnetic field generated by the stator coils and the permanent magnets of the rotor. The magnetic field attracts and rejects the permanent magnets of the rotor. In order to keep the motor running, the magnetic field produced by the stator windings sequence should change thus the rotor rotates to catch up with the stator magnetic field. By changing the current flow in the coils, the polarity of the magnetic fields change at the right moment and the motor rotates. Ideally, the peak torque occurs when the angle of these two fields are at 90° .

3. BLDC MOTOR CONTROL

3.1 Torque-Speed Quadrants Concept

DC motor controls can be classified by the quadrants of operation referring to the torque versus speed plot. In this respect, there are four quadrants control. (See the Fig 3.1)

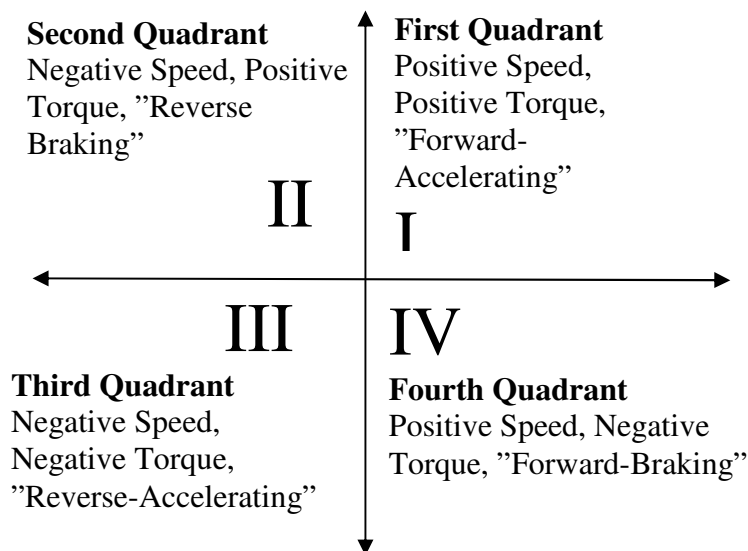


Figure 3.1: Torque/Speed Quadrant of Operation.

3.1.1 One-Quadrant Control

Single - quadrant controls only operate in the first quadrant with positive speed and positive torque. A single quadrant drive usually consists of a single transistor and a single clamp diode. This type of control can only move the motor in one direction and cannot generate any braking forces. [3]

3.1.2 Two-Quadrant Control

The most widely used control method for BLDC motors is the 2 Quadrant Speed (Voltage) control. In the two-quadrant control you can not generate any braking forces. The motor can only operate in quadrant I (forward accelerating) and quadrant III (reverse accelerating). In order to reverse the directions of the rotor, the motor must coast down to zero before reversing directions. If the average applied voltage is less than the back EMF of the motor, the motor current will decrease to zero and the motor will coast. If there is no friction, the motor may spin forever. Many loads, such as fans or pumps, are mostly frictional. Two quadrant control can be easily used in these systems. In a six steps drive system, to implement 2 quadrant speed control it is sufficient to implement PWM only the **bottom** power switches in the power inverter (See Fig. 3.2). In this case, a 0-100% PWM duty cycle adjusts the average voltage applied to the motor and creates a controlled minimum to maximum Speed range. As the average applied voltage increases, motor current increases to accelerate the motor. As motor speed increases its back EMF voltage increases proportionally, but opposes to the applied voltage. In 2-Quadrant Voltage control applications, the **top** power switches are opened and closed at the commutation frequency which is proportional to $(n \times P)/60$, where n is motor speed in rpm and P is the number of pole pairs. As an example, a 6000min^{-1} , 16 pole (8 pole-pair) motor will have a commutation frequency of $(6000 \times 8)/60 = 800$ Hz maximum. The **bottom** switches must operate at the PWM switching frequency (typically 20kHz). Since both switching losses and power switch gate drive requirements increase with switching frequency, not having to PWM on the **top** switches results in Higher Operating Efficiency. The **top** gate drive circuitry is also simpler than the **bottom** gate drive circuitry.

2-Quadrant operating BLDC motor can be reversed by reversing the Electronic Commutator switching sequence. However, this cannot be done quickly because the motor current is not directly controlled (the motor voltage is controlled). In a 2-Quadrant control configuration, the motor must coast to decrease speed. The actions like controlled deceleration (dynamic braking), hard reversal rotation as required by a typical servo application, can be achieved by operating the BLDC motor with a 4-Quadrant BLDC controller.

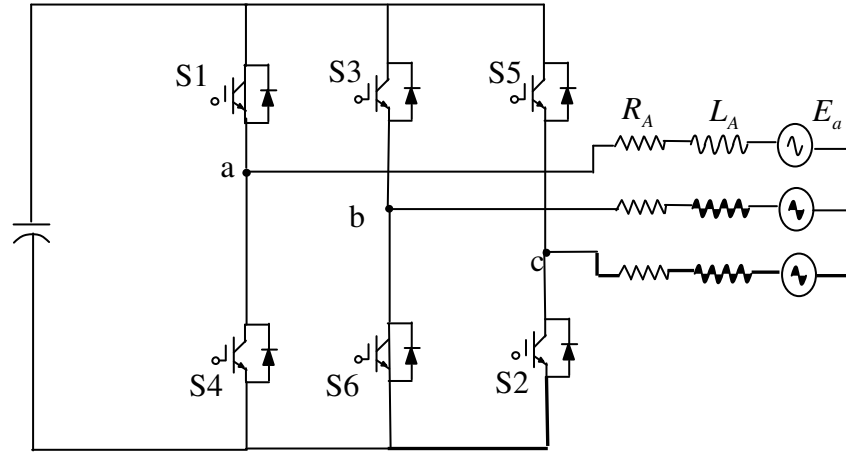


Figure 3.2: Six-Step Drive System for BLDC Motor.

3.1.3 Four-Quadrant Control

4-Quadrant Squarewave BLDC controller is operated with an internal current (Torque) control loop. In a 4-Quadrant Power Inverter, both the **top** and **bottom** power switches are simultaneously Pulse-Width-Modulated. During the PWM OFF cycle, the current in the power inverter freewheel backwards-through the anti-parallel connected power diodes, and a the DC bus capacitor. The current flows in the same shunt, but in the opposite direction to the current during the PWM ON cycle. Thus, a continuous feedback signal proportional to current (Torque) is obtained by sensing the shunt voltage with a current sense amplifier that also detects the absolute value of the current signal. This signal is then subtracted from an external current reference (Torque or current command signal) and the resultant current loop error signal is amplified and used to control the PWM modulator. Hence, the motor current (Torque) is controlled directly.

The Power losses in the 4-Quadrant control are higher than in the 2-Quadrant since all power switches are Pulse - Width-Modulated as discussed above. In 4-Quadrant controllers, a Dynamic Braking resistor and transistor (usually connected in series across the DC bus) is used to absorb the kinetic energy released by the motor during rapid deceleration and hard reversing. The released kinetic energy appears as a reverse current flowing out from the Power Inverter into the DC bus capacitor, through the flyback diodes connected in anti-parallel with each Inverter power switch. This reverse (braking) current charges the bus capacitor, increasing the average DC bus voltage. A Dynamic Braking control circuit must sense this excess

DC bus voltage and properly switch the Dynamic Braking transistor across the DC bus in order to bound the DC bus voltage to safe operating levels. When the braking transistor Turns-ON, power is dissipated in the braking resistor proportional to braking resistance times the square of the RMS braking current. The excess kinetic energy is converted to heat dissipated by the braking resistor while, the DC bus is simultaneously maintained at a safe DC operating level [4].

3.2 Closed Loop Speed Control Theory

Commutation provide the proper rotor rotation of the BLDC motor, while the motor speed only depends on the amplitude of the applied voltage. The amplitude of the applied voltage is adjusted using the PWM technique.

The required speed is controlled by a speed controller, which is performed proportional-integral (PI) controller. To generate a voltage proportional to desired speed, the difference between the actual and required speeds is fed to input of the PI controller and set the duty cycle of the PWM pulses (See Fig. 3.3.)

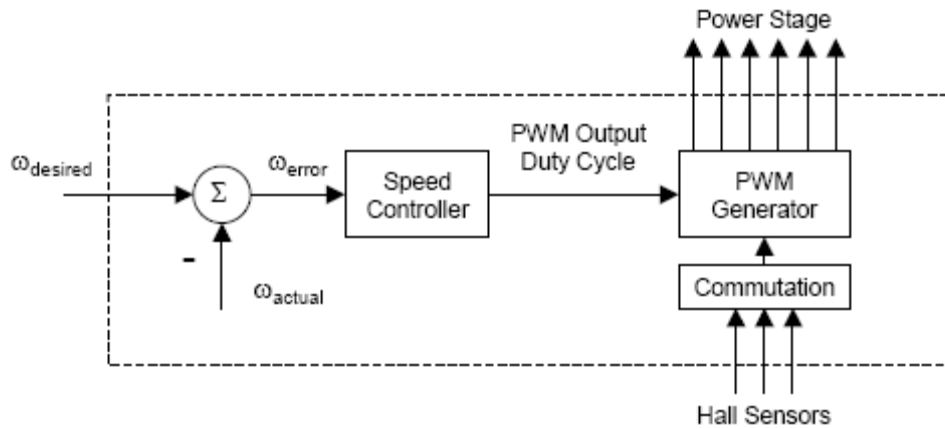


Figure 3.3: Speed Controller

The speed controller calculates the PI algorithm given in the equation below:

$$u(t) = K_c [e(t) + \frac{1}{T_I} \int_0^t e(t) dt] \quad (3.1)$$

After transforming the equation into a discrete time domain using an integral approximation with the Backward Euler method, it is obtained the following equations for the numerical PI controller calculation:

$$\begin{aligned}
 u(k) &= u_p(k) + u_i(k) \\
 u_p(k) &= K_c . e(k) \\
 u_i(k) &= u_i(k-1) + K_c \frac{T}{T_i} . e(k)
 \end{aligned}
 \tag{3.2}$$

Where:

$e(k)$ = Input error in step k

$w(k)$ = Desired value in step k

$m(k)$ = Measured value in step k

$u(k)$ = Controller output in step k

$u_p(k)$ = Proportional output portion in step k

$u_i(k)$ = Integral output portion in step k

$u_i(k-1)$ = Integral output portion in step k - 1

T_i = Integral time constant

T = Sampling time

K_c = Controller gain

3.3 Digital Control

The BLDC motor is driven by rectangular voltage pulses according to the given rotor position (see Figure 3.4). Rotor flux generated by the rotor magnet intersect with the generated stator flux thus created a torque that defines the speed of the motor. Rotor flux is generated by a rotor magnet and defines the torque and thus the speed of the motor as mentioned before. The voltage pulses must be properly applied to the phases of the three-phase winding system so that the angle between the stator flux and the rotor flux is kept as close to 90° as possible, to get the maximum generated torque. Therefore, the motor requires electronic control for proper operation.

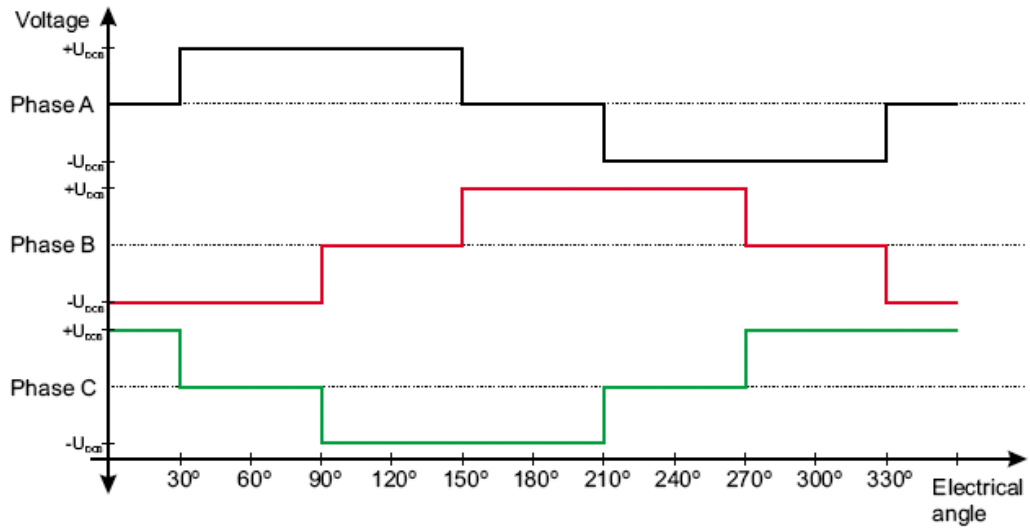


Figure 3.4: Voltage Strokes Applied to the 3-Phase BLDC Motor.

For the 3-phase BLDC motors, a standard 3-phase power stage is used (see Figure 3.5). The power stage consist of six power transistors. In both modes, the 3-phase power stage energizes two motor phases simultaneously. The third phase is nonenergised. Thus, it is obtained six possible voltage vectors that are applied to the BLDC motor using a pulse width modulation (PWM) technique.

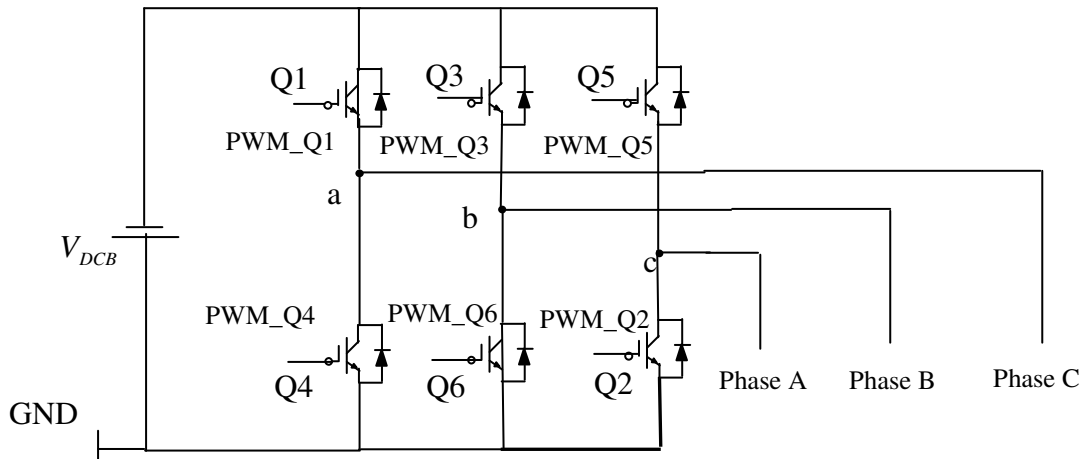


Figure 3.5: 3-Phase BLDC Power Stage.

4. CLOSED LOOP SPEED CONTROL DRIVER

4.1 Driver Construction

The main difficulty for control BLDC motors is to decode data coming from Hall effect sensors and perform some important functions like forward/reverse selection, overcurrent shutdown, undervoltage lockout, overtemperature shutdown. It is possible to use discrete components to perform these functions but it means too much circuit board area, especially if the control unit is to be placed inside the motor.

Another problem is insufficient performance of the existing power transistors. Power bipolars can not be favored because they can not be driven directly from a control IC and here power MOSFETs may be the best choice since they are easy to drive, efficient and cheaper.

As explained in the following sections with the details, three main devices underlies our control circuit. MC33035 is the brain IC of the circuit and control all the operation. MC33039 is the closed-loop speed control IC (electronic tachometer) give the speed information of the rotor and lastly MSK3003 is a three phase bridge inverter electronically commutate the motor and make the control easier.

4.1.1 MC 33035 IC

The MC33035 is a brushless DC motor controller IC can perform all of the active functions mentioned above. This IC has a rotor position decoder to provide proper commutation, sawtooth oscillator, three open collector top drivers and three totem pole bottom drivers suited for driving power MOSFETs (See Fig. 4.1 and Fig. 4.2). MC33035 has the following features:

- 10 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 External 3–Phase MOSFET Bridge
- Cycle–By–Cycle Current Limiting
- Pinned–Out Current Sense Reference
- Internal Thermal Shutdown
- Selectable 60°/300° or 120°/240° Sensor Phasings
- Can Efficiently Control Brush DC Motors with External MOSFET H–Bridge

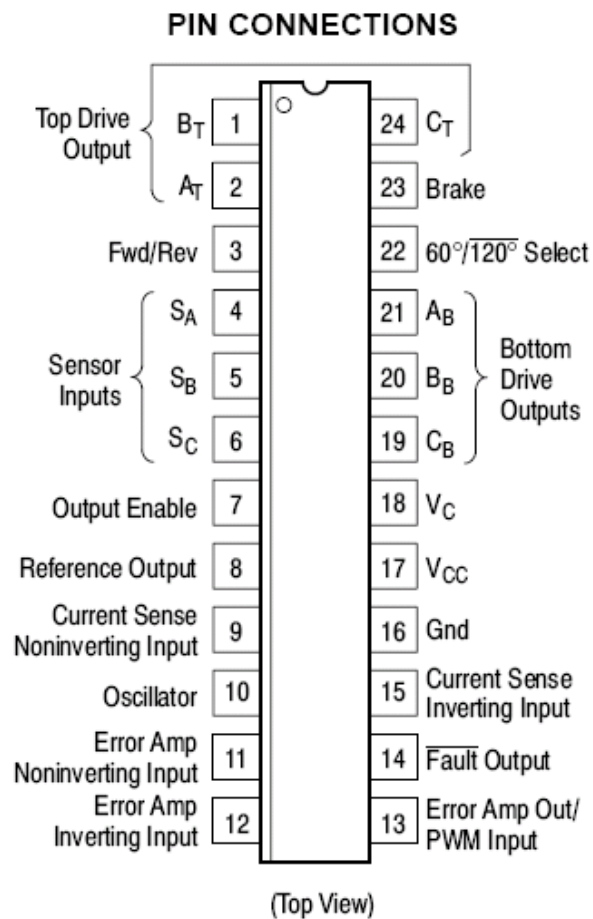


Figure 4.1: MC33035 Pin Connections [6]

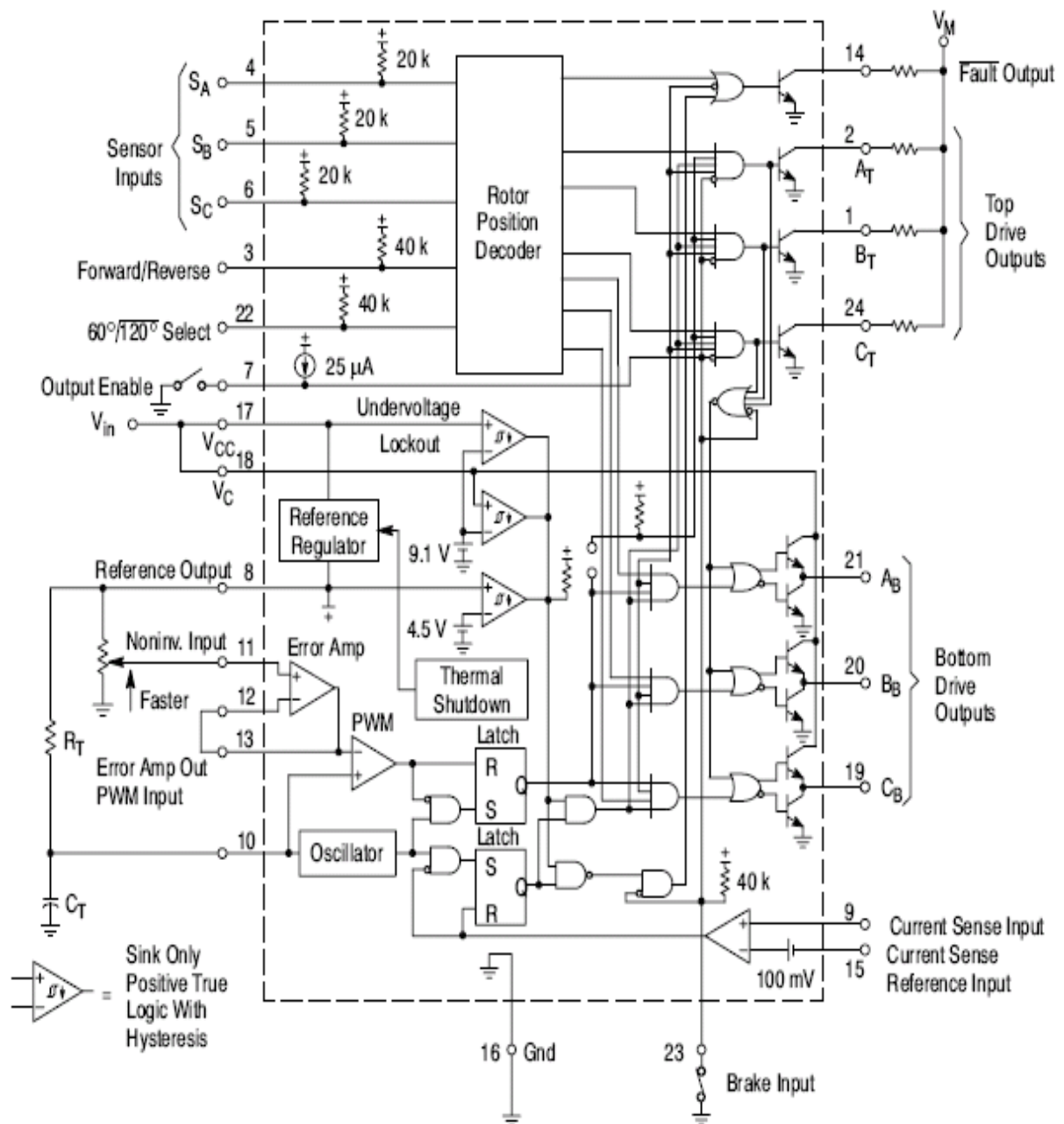


Figure 4.2: MC33035 Representative Block Diagram [6]

Table 4.1: MC33035 Pin Descriptions

PIN	SYMBOL	DESCRIPTION
1,2,24	B_T, A_T, C_T	These three open collector Top drive outputs are designed to drive the external upper power switch transistors.
3	Fwd/Rev	The Forward/Reverse Input is used to change the direction of motor rotation.
4, 5, 6	S_A, S_B, S_C	These three Sensor Inputs control the commutation sequence.
7	Output Enable	A logic high at this input causes the motor to run, while a low causes it to coast.
8	Reference Output	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.
9	Current Sense Noninverting input	A 100mV 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.
10	Oscillator	The Oscillator frequency is programmed by the values selected for the timing components, R_T, C_T .
11	Error Amp Noninverting Input	This input is normally connected to the speed set potentiometer.
12	Error Amp Inverting Input	This input is normally connected to the Error Amp Output in open loop applications.
13	Error Amp Out/PWM Input	This pin is available for compensation in closed loop applications.
14	$\overline{Fault}$ Output	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.
15	Current Sense Inverting Input	Reference pin for internal 100 mV threshold. This pin is normally connected to the bottom side of the current sense resistor.
16	Gnd	This pin supplies a ground for the control circuit and should be referenced back to the power source ground.
17	V_{CC}	This pin is the positive supply of the control IC. The controller is functional over a minimum VCC range of 10 to 30 V.
18	V_C	The high state (VOH) 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 to 30 V.
19, 20, 21	C_B, B_B, A_B	These three totem pole Bottom Drive Outputs are designed for direct drive of the external bottom power switch transistors.
22	60° 120° Select	The electrical state of this pin configures the control circuit operation for either 60° (high state) or 120° (low state) sensor electrical phasing inputs.
23	Brake	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.

4.1.1.1 Rotor Position Decoder

The main duty of the rotor position decoder is to decode signals coming from Hall Effect Sensors and to provide proper sequencing for the top and bottom drive outputs. Here the inputs are TTL (Transistor-Transistor Logic) compatible, with their thresholds typically at 2.2 V. That means 0.0 – 0.8 V correspond to logic 0 and 2.2-5V correspond to logic 1. Details of rotor position decoder will be discussed in commutation process.

4.1.1.2 Error Amplifier

An important structure is a high performance internal error amplifier is designed as unity gain voltage follower that is possible to access to both inputs and output (Pins 11, 12, 13). This structure enables open and closed loop speed control. In the following figure, error amplifier output is connected to the PWM input.

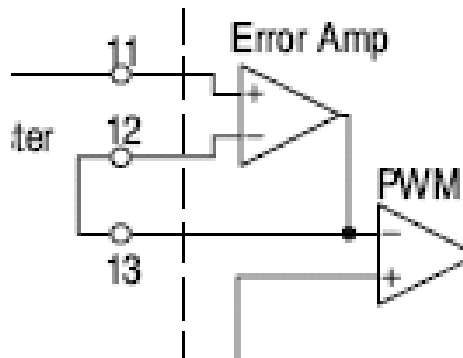


Figure 4.3: Error Amplifier [5]

4.1.1.3 Oscillator

Duty of the oscillator is to set the both R-S flip flop and thus control the conduction of the top and bottom drive outputs. The frequency of the oscillator is set by the timing components $R_T(R_2)$ and $C_T(C_2)$ (See Fig. 4.8). Capacitor C_T is charged from the MC33035 Reference Output (Pin 8) through resistor R_T and discharged by an internal transistor.

4.1.1.4 Pulse Width Modulator

It's a good and energy efficient method to control the speed of the motor as varying pulse widths of the applied voltage to each stator windings during the commutation. Here as C_T (C_2) discharges, the sawtooth oscillator adjusts both latches and control the top and bottom drive outputs. When positive rising voltage of C_T becomes higher than the error amplifier output, the PWM comparator cut off the bottom drive output transmission as resetting the upper latch. Pulse width modulation is performed only at the bottom drive outputs. The pulse width modulator timing diagram is shown in Figure 4.4.

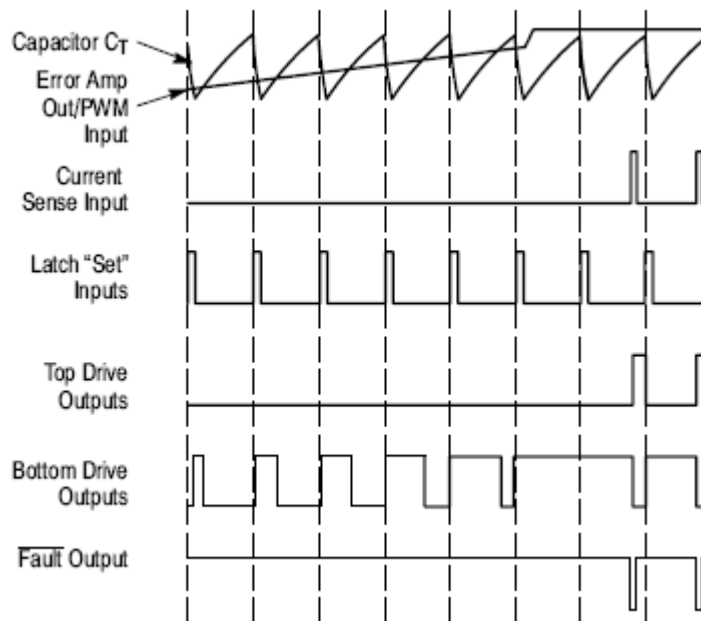


Figure 4.4: Pulse Width Modulator Timing Diagram [5]

4.1.2 MC33039 ElectronicTachometer

The MC33039 is an electronic tachometer can perform closed-loop speed control of a brushless DC motor as cooperating with MC33035. This part consists of three input buffers, three digital edge detectors, a programmable monostable and an internal shunt regulator. This device can be used in many closed-loop speed control applications. Refer to Figure 4.5 for the block diagram.

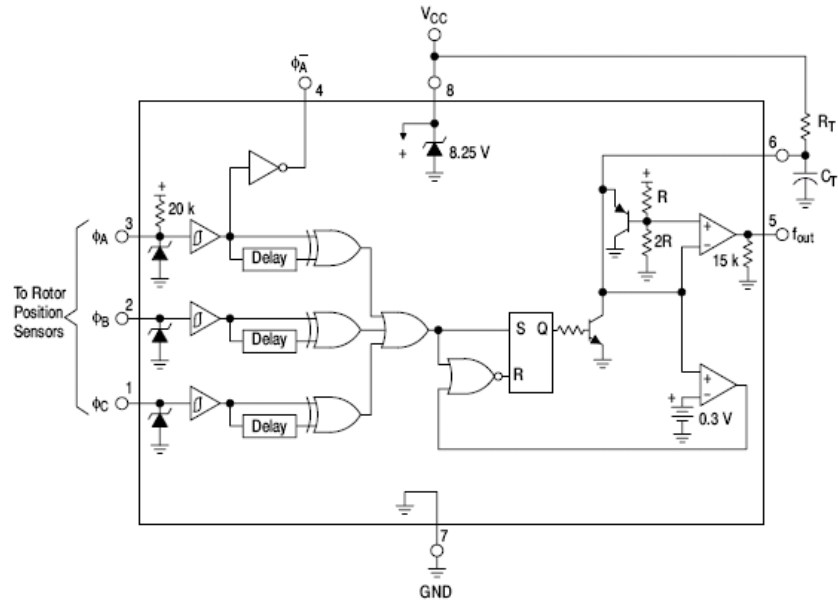


Figure 4.5: MC33039 Block Diagram [7]

4.1.3 MSK3003 Power Module

The MSK3003 is a three phase bridge power module consisting of P-Channel MOSFETs for the top transistors and N-Channel MOSFETs for the bottom transistors. The MSK3003 can be used directly with many brushless motor drive IC's without additional circuits. Refer to Figure 4.6 for the MSK3003 circuit schematic.

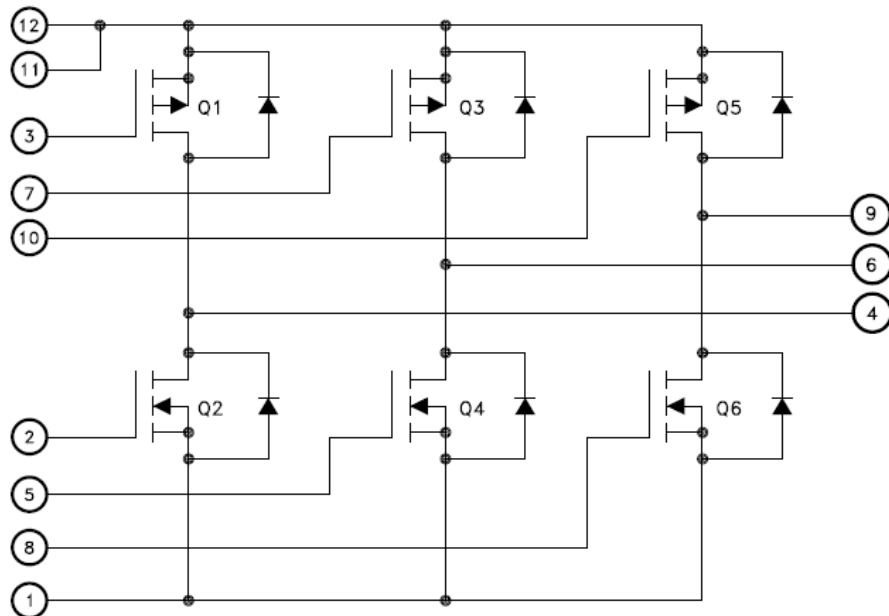


Figure 4.6: MSK3003 Circuit Scheme

4.2 Assemblying the Circuit

4.2.1 Timing Components

The brushless DC motor used in this project has one pair of pole on its permanent magnet and there is one electrical degree for every mechanical degree. Each Hall effect sensor generates one pulse and the three sensors generate three pulses for every mechanical revolution. MC33039 generates one pulse for every rising and falling edge and totally generates 6 pulses.

In Fig 4.8 R_1 and C_1 are the MC33039 timing components and the values of C_1 and R_1 set the f_{out} pulse width which takes the maximum value for a given maximum speed. Figure 4.7 shows the MC33039 timing diagram. The timing components are selected according to the desired maximum motor RPM.

R_2 and C_2 on Figure 4.8 are the timing components for the MC33035. Capacitor C_2 (C_T) is charged from the Reference Output (Pin 8) through resistor R_2 (R_T) and discharged by an internal transistor as mentioned before in chapter 4.1.1.3. The values of the timing components set the frequency of the internal ramp oscillator.

4.2.2 Drive Circuits

MC33035 has six output drivers. Three top drive outputs open collector NPN transistors (Pins 1, 2, 24) drive the P-Channel MOSFETs. Three totem pole bottom drive outputs (Pins 19, 20, 21) can drive directly N-Channel MOSFETs. Bottom drive outputs are supplied from V_C (Pin 18) as being independent source of V_{CC} . While V_{CC} is greater than 20V, MOSFETs gates might damage, therefore a zener diode must be connected to Pin 18.

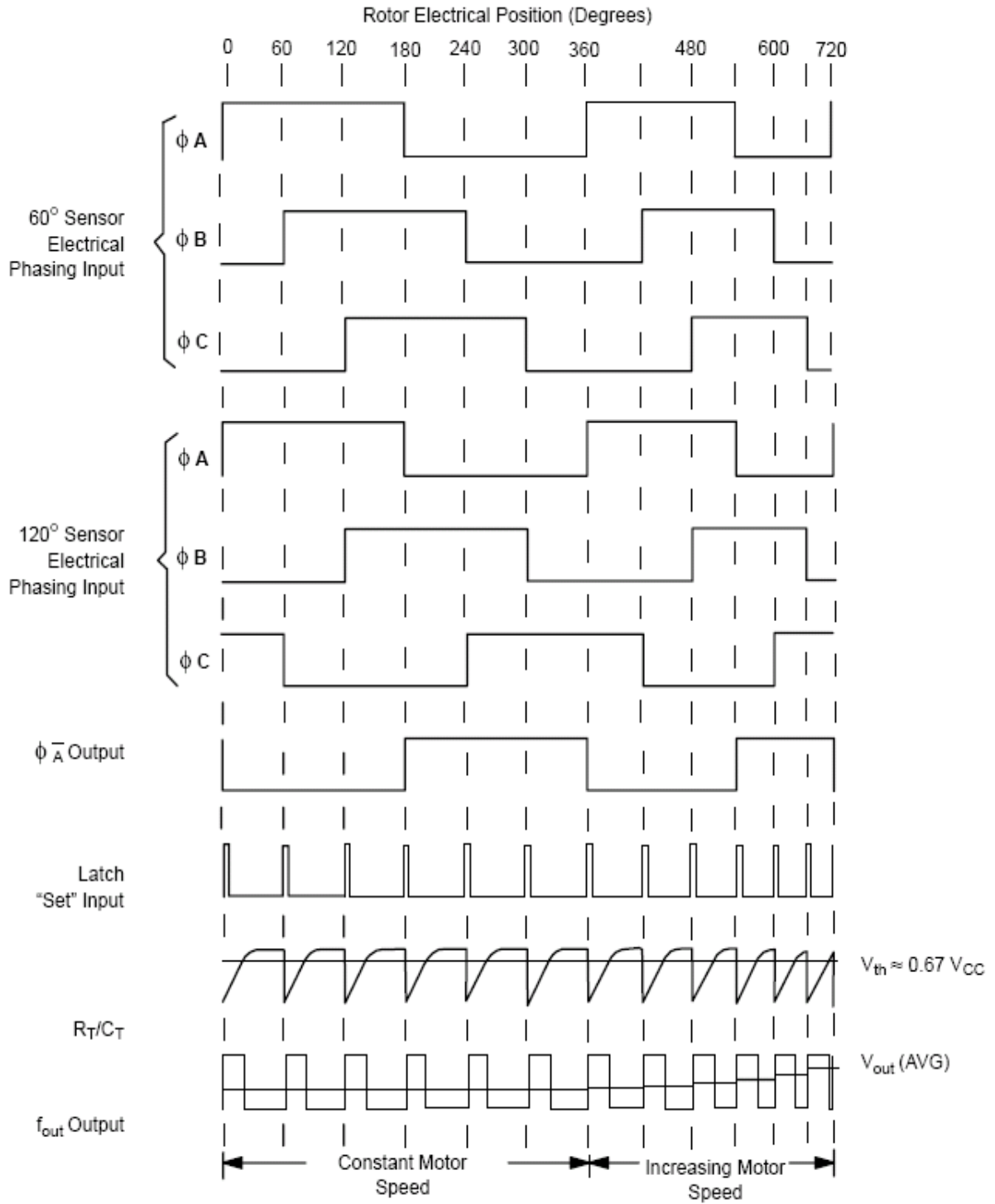


Figure 4.7: Timing Diagram of A Typical Three Phase, Six Step Motor Application [6]

4.2.2.1 N-Channel Gate Drive Circuit

If that considered our system supplied from 24V (18 to 30V), MC33035 can be powered directly from system voltage since the IC has a 40 V rating. Here with the electrolytic capacitor (C_8), a small filter capacitor ($C_7 - 0.1\mu\text{F}$) is placed close the IC to minimize local spiking across the DC bus.

To minimize the power losses in the IC, three lower output transistors are driven with a separate supply (V_C -Pin18) from the MC33035. The required current to drive the MOSFETs is just the current to charge and discharge the gate-to-source and drain-to-gate capacitors of each MOSFET. C_4 filter capacitor supplies the turn-on current while refreshed through resistor R_7 because the MOSFETs draw very small average current and high current values are required to charge their input capacitances. Dropping resistor (R_7) gets 3 V when the main supply takes the lowest value 18 V in the 24 V system. For charging the capacitor it's a good selection to use a 1 k Ω resistor, it will also supply at least 1 mA current to the zener to provide good regulation. At high supply voltages the resistor will see a voltage of 15 V, a current of 15 mA, and it means to a power dissipation about 0.25 W. Therefore, a 0.5 W resistor will be a good choice. That's also a good power rating for the zener. Also three Schottky diodes D_1 , D_2 , D_3 are placed between the Gates of the N-Channel MOSFETs and the ground to prevent the rupture if the substrate current exceeds 50mA.

If the gate drive impedance of the three lower devices is low it may be the problem that gate drive loop may cause ringing during gate voltage transitions. Such a ringing is amplified by the MOSFETs and may occurs high levels of noise at the drain. It may be the solution to insert the series resistance to gate drive as reducing the circuit's Q. Any gate drive resistor value lower than 62 Ω may occur oscillations in this circuit. (See Fig.4.8)

4.2.2.2 P-Channel Gate Drive Circuit

At least 7 - 8 V may be acceptable on gates for standard MOSFETs. **R5** and **R6** are selected to provide that the P channel gate drives take 10 V value even if the supply voltage gets lowest value (18 V). **R5**, **12** and **13** control turn-on speed as providing keep charged of the P-channel input capacitances. Similarly, **R6**, **R14** and **R15** control turn-off speed. Since P-channels MOSFETs work at the same frequency with

the motor commutation frequency (lower than the PWM frequency), it's not required the low impedance for P-Channel gate drives.

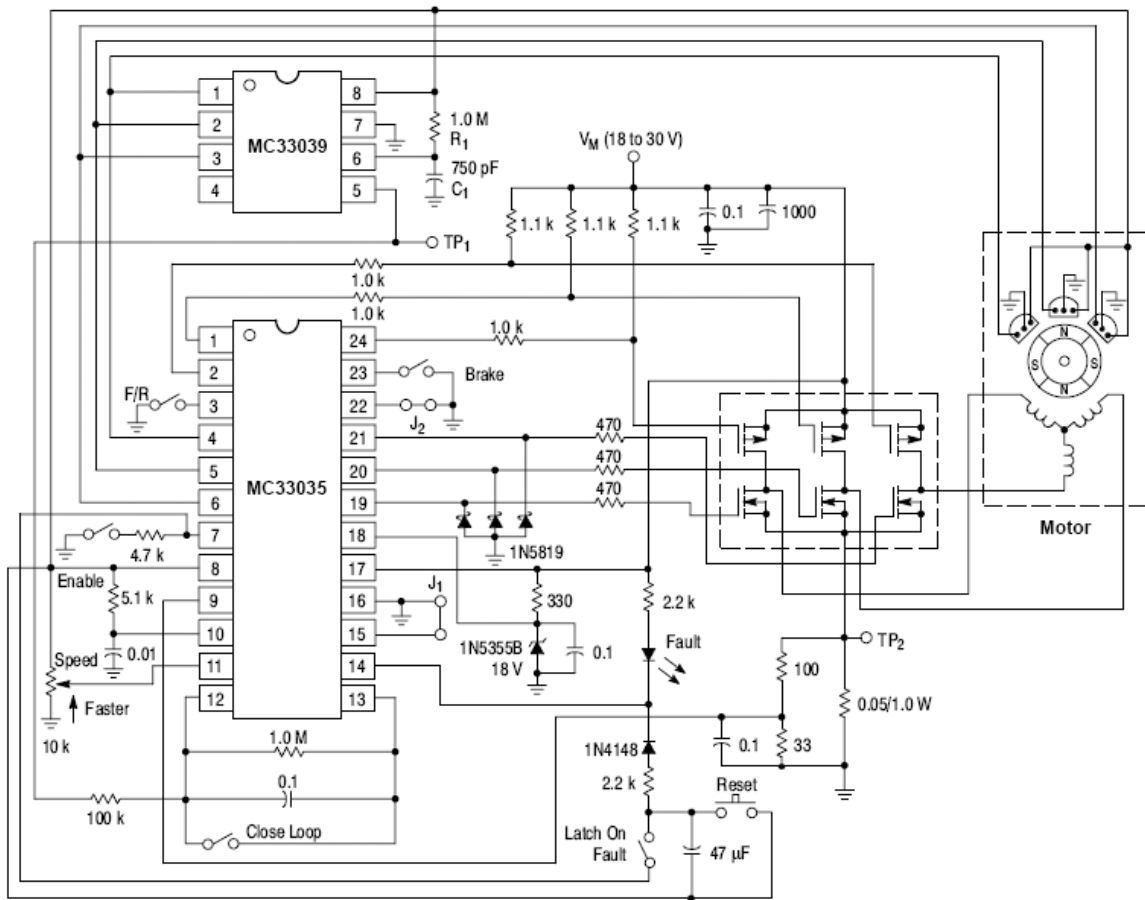


Figure 4.8: Closed Loop Brushless DC Motor Control Using The MC33035, MC33039, MSK3003 [6]

4.3 Control Features

The MC33035 is not capable of closed loop speed control. The IC can not monitor the motor speed and generates a signal proportional to the motor RPM, generally performed by a tachometer. If the motor speed signal is provided, MC33035 can manage closed loop speed control application.

4.3.1 Open Loop Speed Control

It's not required to know motor speed data to perform open loop control. It is enough to give a signal proportional to desired motor speed into the error amplifier's non-inverting input (Pin 11). Then output of the error amplifier is compared to the output

of the oscillator to obtain a PWM signal proportional to desired motor speed until the control loop is terminated by an overcurrent or fault condition.

4.3.2 Closed Loop Speed Control

For closed loop motor speed control, the MC33035 requires an input voltage proportional to the motor speed. This input voltage is generated by the MC33039. Figure 4.8 shows the application that 6.25 V reference from the MC33035 (Pin 8) is supplied to the MC33039 used to generate the feedback voltage proportional to the motor speed without need a tachometer. The both MC33035 and MC33039 use the same Hall sensor signals. Although MC33035 use them to decode rotor position, MC33039 use Hall sensor signals to detect the speed of the rotor.

With every rising and falling edge of the Hall sensor signals, MC33039 generates an output pulse which its amplitude and time duration are setted by the values of the resistor R_1 and the capacitor C_1 . These output pulses released from MC33039 (Pin 5) are integrated by the error amplifier of the MC33035 and generated a DC voltage level proportional to motor speed.

After generating a signal proportional to motor speed, this signal set the PWM reference level at Pin 13 of the MC33035 and closes the last major link of the feedback loop and the signal is fed into the inverting input (Pin 12) of the MC33035's comparator. Here the MC33039's output is low pass filtered by the R_4 , C_3 . The signal proportional to desired motor speed drives the non-inverting input (Pin 11) and the ratio of the input and feedback resistors R_3 and R_4 control the gain. Here feedback capacitor C_3 combines the low pass filtering and generating the error signal.

MC33035 expands the output pulse width to the drive transistors if the motor speed becomes lower than the desired speed, inversly the duty cycle decreases if the motor speed is greater than the desired speed. If the desired speed is so much lower than the motor speed, the duty cycle fall to zero and the motor would coast to desired speed.

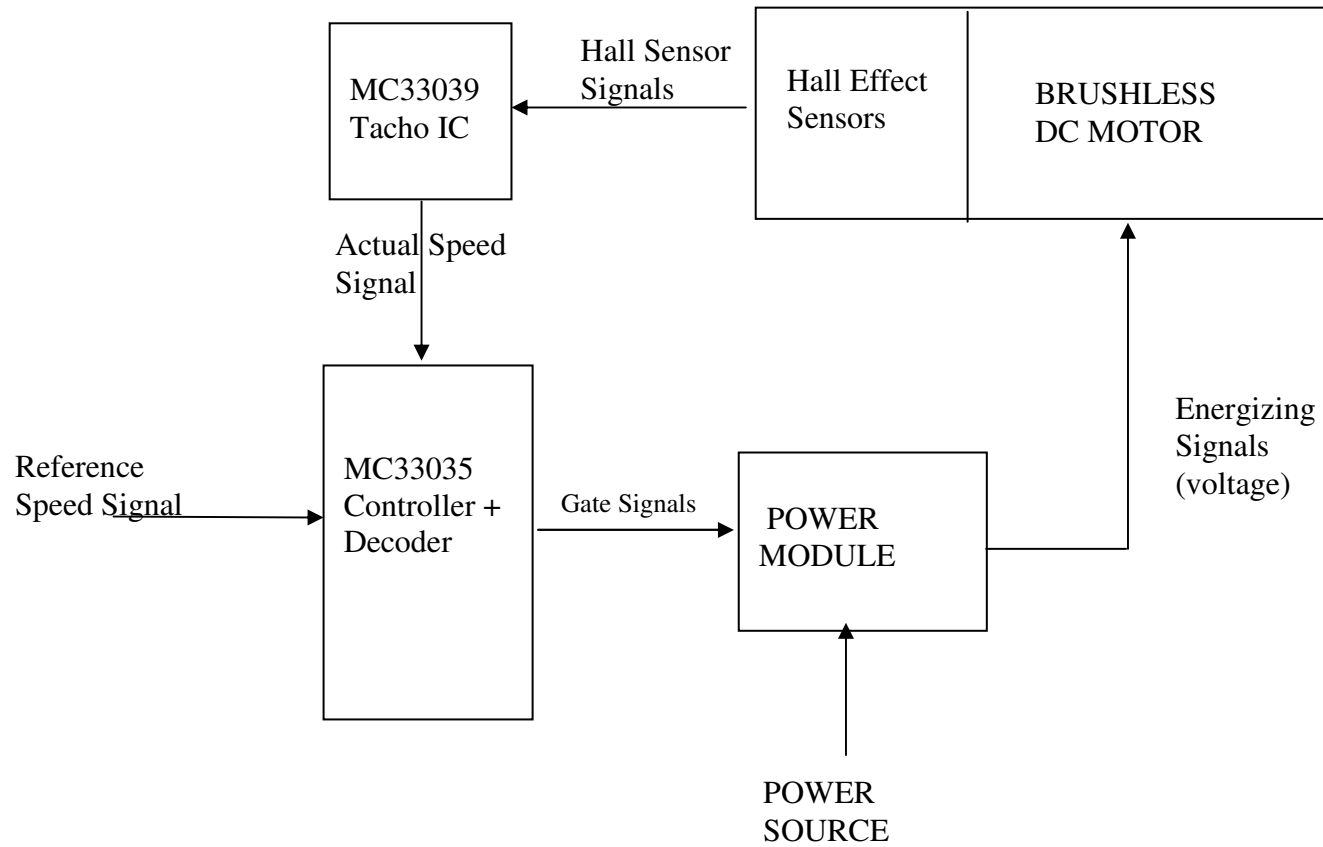


Figure 4.9: The Functional Block Diagram of the System

4.4 Commutation

4.4.1 Rotor Position Decoder

Corresponding to three sensor inputs, eight possible input code combinations are available and two of them are invalid input codes, leaving six valid input codes. The decoder can define the rotor position using the six valid input codes. The direction of motor rotation can be changed by the Forward/Reverse input (Pin 3) as giving a reverse voltage to the stator winding. The commutation sequence is reversed when the input changes from high to low with a given sensor input code; the active top and bottom drive outputs are exchanged (AT to AB, BT to BB, CT to CB). Consequently, the motor changes directional rotation. The Output Enable pin (Pin 7) controls the on/off state of the motor. A 25 mA current source provides sequencing of the top and bottom drive outputs when the pin is left disconnected. When grounded, the motor coasts as turning off the top drive outputs and forcing low the bottom drives, and the Fault output activates. Braking is performed by setting the Brake Input (Pin 23) in a high state. Thus, the top drive outputs turn off and the bottom drives turn on, shorting the motor windings and generating back EMF. The brake input has unconditional priority over all other inputs.

4.4.2 Commutation Process

The commutation decoder of the MC33035 receives signals from the position sensors regarding the position of the rotor and translates them into the switching signals as supplied to the firing circuit, which consists of the electronic switches. In a six-step drive system, two switches are activated simultaneously to energise two of the stator phases, and a rotating field is established in the air-gap by the interaction of the currents on the two coils.

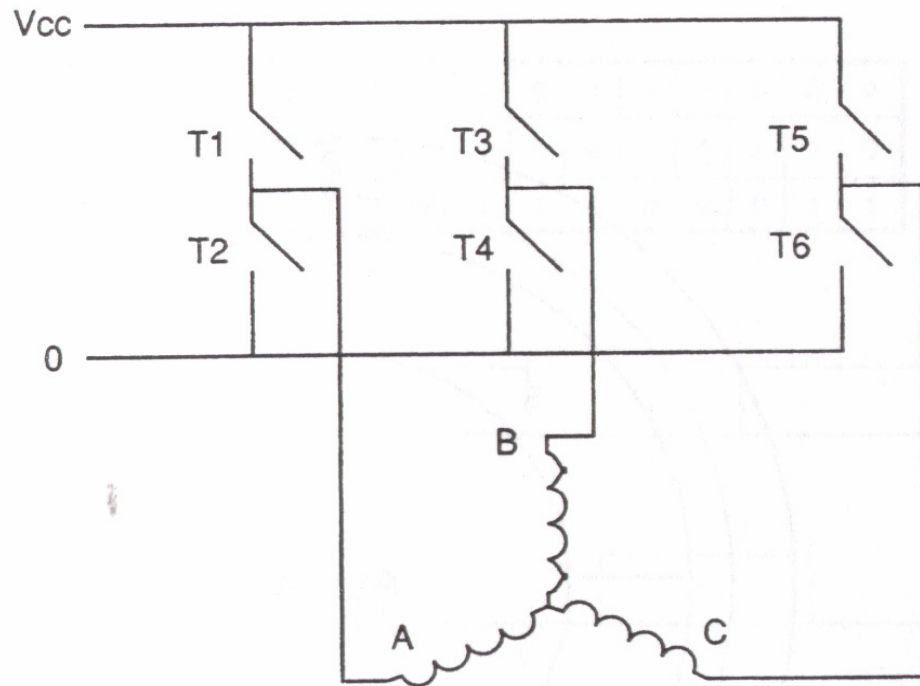


Figure 4.10: An Illustration of the Principle Commutation Circuit of A Brushless D.C. motor. [11]

The general principle of a three-phase brushless dc motor drive is illustrated in Fig 5.10.

On the Fig 4.10, switches T_5 and T_2 , T_3 and T_2 , T_3 and T_6 , etc. are activated sequentially to move the field around the air-gap in the fwd (clockwise) direction. The switching sequence and the direction of the resulting air-gap field are shown in Table 4.2. For reverse (anticlockwise) rotation, the switching sequence is reversed.

In general, a 3-to 6 line decoder is needed for a system with three position sensors and a six-switch inverter bridge. The three inputs line is connected to the output of the sensors and the six outputs are connected to the switches firing circuits. The sensors may be mounted with 30, 60 or 120 mechanical degrees spacing. The decoder design discussed here for the case of 60 and 120 degrees spacing.

Table 4.2: Switching Sequence and Resulting Air-Gap Field Direction [11]

Coils Energised	A & B	A & C	B & C	B & A	C & A	C & B
Direction of Magnetic Field						
Switches Activated	T5&T2	T3&T2	T3&T6	T1&T6	T1&T4	T5&T4
Shaft Angle	0-60	60-120	120-180	180-240	240-300	300-360

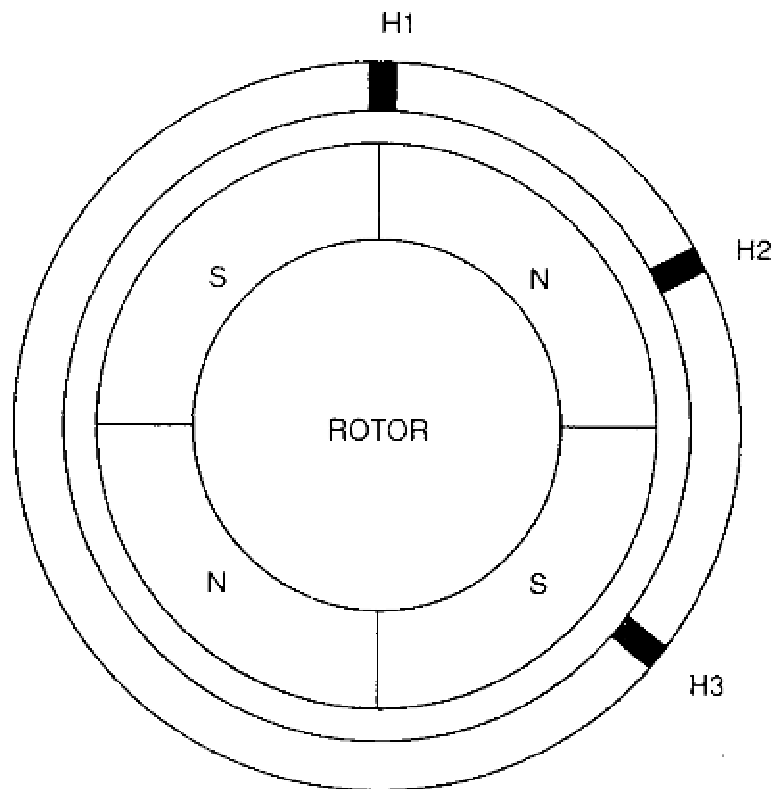


Figure 4.11: Four - Poles Permanent Magnet Rotor and Three Sensors Spacing at 60 Degrees [11]

Schematic of the four poles permanent magnet rotor and the position sensors is shown in Fig 5.11. Each sensor generate high (logic 1) pulse when it is passed by a South pole and low (logic 0) when it is passed by a North pole. For the initial position shown in Fig 4.11, the first sensor (H_1) operates at 0° the second (H_2) operates after 60° and the third (H_3) releases after 30° of rotation in the clockwise (forward) direction. Table 4.3 is the comutation truth table and shows the sensor ouput code for forward and reverse rotation. There are six different commutation logic combinations per one electrical revolution and the logic codes for both directions are the same, but the code sequence is reversed.

The decoder truth table is readily derived with the aid of Table 4.2, Fig.4.11 and 4.12. In order to enable rotation to both directions, a code corresponding to the desired direction of rotation (F/R select) is added as an input to the commutation decoder. Here codes 1 and 0 are chosen for forward (clockwise) and backward rotation respectively, but this is arbitrary. Other inputs to the commutation decoder are sensors output codes, Output Enable code (Pin 7), brake input (Pin 23), Curren Sense input and 60° 120° sensor phase selection. The commutation decoder truth table is given with the Table 4.3.

The Boolean functions of the semiconductor switches are as follows:

$$\begin{aligned} T1 &= \overline{\overline{F.H1.H2}}(\overline{R.H1.H2}) \\ T2 &= \overline{\overline{F.H1.H2}}(\overline{R.H1.H2}) \\ T3 &= \overline{\overline{F.H2.H3}}(\overline{R.H2.H3}) \end{aligned} \tag{4.1}$$

$$\begin{aligned} T4 &= \overline{\overline{F.H2.H3}}(\overline{R.H2.H3}) \\ T5 &= \overline{\overline{F.H3.H1}}(\overline{R.H3.H1}) \\ T6 &= \overline{\overline{F.H3.H1}}(\overline{R.H3.H1}) \end{aligned} \tag{4.2}$$

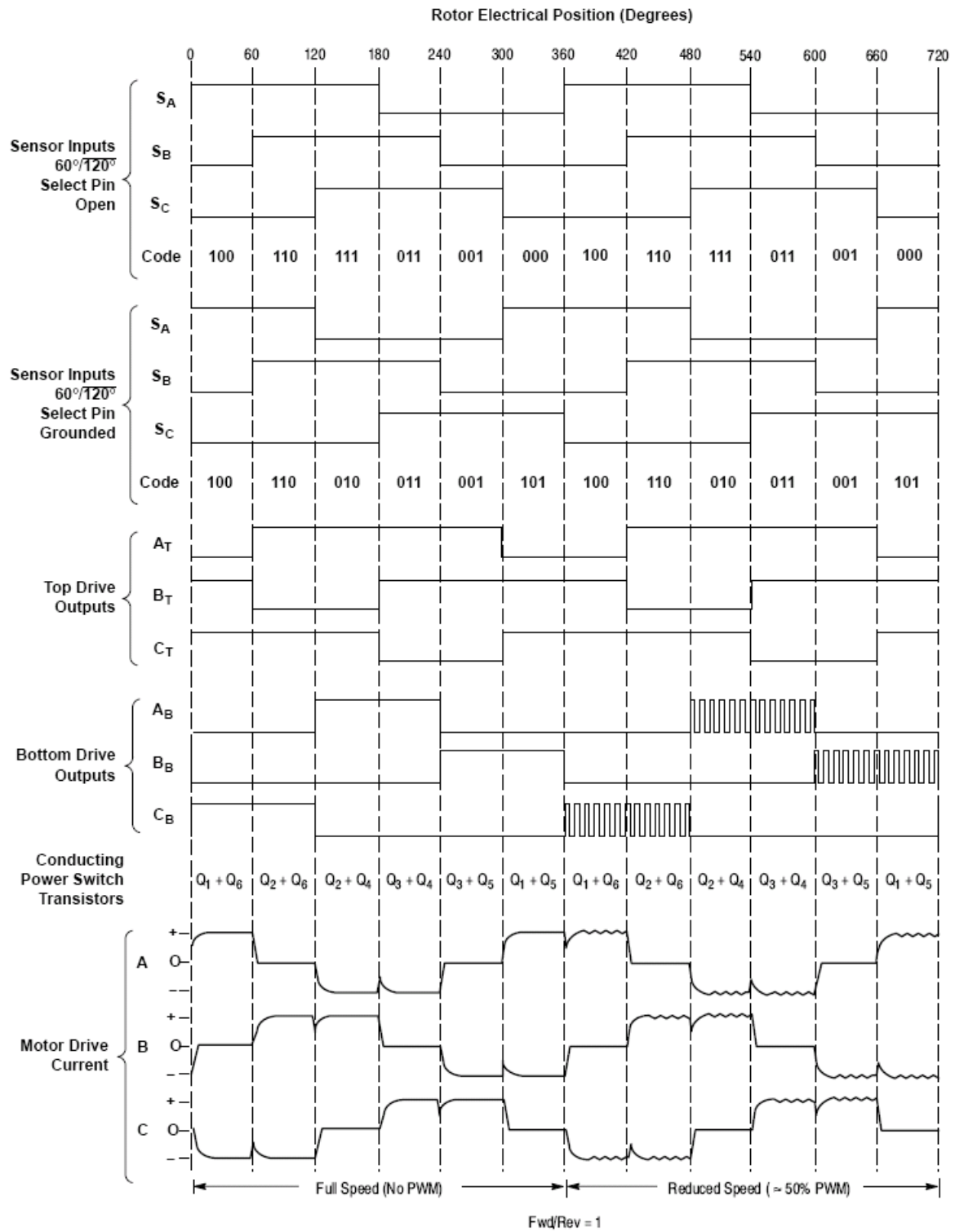


Figure 4.12: Three Phase, Six Step, Full Wave Commutation Waveforms [6]

Table 4.3: Three Phase Six Step Commutation Truth Table

Sensor Electrical Phasing										Top Drives			Bottom Drives			Fault	Active Switches
60°			120°							A_T	B_T	C_T	A_B	B_B	C_B		
S_A	S_B	S_C	S_A	S_B	S_C	F/R	Enable	Brake	Current Sense								
1	0	0	1	0	0	1	1	0	0	0	1	1	0	0	1	1	T5, T2
1	1	0	1	1	0	1	1	0	0	1	0	1	0	0	1	1	T3, T2
1	1	1	0	1	0	1	1	0	0	1	0	1	1	0	0	1	T3, T6
0	1	1	0	1	1	1	1	0	0	1	1	0	1	0	0	1	T1, T6
0	0	1	0	0	1	1	1	0	0	1	1	0	0	1	0	1	T1, T4
0	0	0	1	0	1	1	1	0	0	0	1	1	0	1	0	1	T5, T4
1	0	0	1	0	0	0	1	0	0	1	1	0	1	0	0	1	T1, T6
1	1	0	1	1	0	0	1	0	0	1	1	0	0	1	0	1	T1, T4
1	1	1	0	1	0	0	1	0	0	0	1	1	0	1	0	1	T5, T4

0	1	1	0	1	1	0	1	0	0	0	1	1	0	0	1	1	T5, T2
0	0	1	0	0	1	0	1	0	0	1	0	1	0	0	1	1	T3, T2
0	0	0	1	0	1	0	1	0	0	1	0	1	1	0	0	1	T3, T6
1	0	1	1	1	1	X	X	0	X	1	1	1	0	0	0	0	—
0	1	0	0	0	0	X	X	0	X	1	1	1	0	0	0	0	—
1	0	1	1	1	1	X	X	1	X	1	1	1	1	1	1	0	T2,T4,T6
0	1	0	0	0	0	X	X	1	X	1	1	1	1	1	1	0	T2,T4,T6
V	V	V	V	V	V	X	1	1	X	1	1	1	1	1	1	1	T2,T4,T6
V	V	V	V	V	V	X	0	1	X	1	1	1	1	1	1	0	T2,T4,T6
V	V	V	V	V	V	X	0	0	X	1	1	1	0	0	0	0	—
V	V	V	V	V	V	X	1	0	1	1	1	1	0	0	0	0	—

4.5 Fault Management

The duty of the open collector Fault Output (Pin 14) is alerting the IC in the event of the system malfunction. It pulls the Fault Output Pin “low” as supplying maximum 16 mA current when the system has any fault condition and can directly drive a LED for visual indication thus alert the IC for a problem. Here R₁₁ (See Fig. 4.8) takes the 2.2 k values in a 24 V system to supply 1 mA to the LED. Upon fault detection, it is suitable to terminate any further pulsing to the output transistors. This is possible by connecting the Fault Output to the Enable pin (Pin 7). Thus the motor start up current is limited or latched the system off. It is possible to create a time – delayed latched shutdown for overcurrent condition as inserting of a RC (R₁₀ and C₆, see Fig. 4.8) network between the Fault Output (Pin 14) and the enable input (Pin 7). Here the latch can be delayed by inserting C₆ (47μF). The delay is set by the time constant of R₁₀ and C₆ before the system latches. The Fault Output is active low when one or more of the following conditions occur:

- 1) Invalid Sensor Input code
- 2) Output Enable at logic [0]
- 3) Current Sense Input greater than 100 mV
- 4) Undervoltage Lockout
- 5) Thermal Shutdown

4.5.1 Over Current Detection

An over-current condition may occur while continuous operation of the motor that is severely results in overheating and eventual failure. If the pulse width is increased abruptly to quickly accelerate the motor, very high currents may flow. This will cause an undesirable jerk on the motor and the mechanical system. Worse, it could exceed the current rating of the power devices.

4.5.1.1 Overcurrent Sensing

The excessive load currents are detected by the comparator of MC33035. A signal related with all the N-channel sources is fed into a comparator from the current sensing resistor (R₂₁). Here the trip threshold is 100 mV. The comparator then feeds

the RS Flip–Flop and if an overcurrent condition is detected, the output drivers turn – off the power transistors the remainder of the oscillator cycle. In here, R_{21} is the current sensing resistor 0.05Ω and has 1W power rating and the sense voltage is reduced by a voltage divider. The voltage divider values are, 100Ω (R_8 in Figure 4.8) for the upper resistor and 33Ω (R_9) for the lower one that set the trip current to 8A. To preserve the overcurrent comparator from noise or currents such as reverse recovery spikes of freewheeling diodes replacing a small capacitor (C_5). 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 calculated as follow:

In the s domain analysis a capacitance C is replaced by an admittance sC , or equivalently an impedance $1/sC$,

$$T(s) = \frac{V_o}{V_i} = \frac{R_9 // Z_{C_5}}{R_8 + (R_9 // Z_{C_5})} = \frac{\frac{R_9}{R_9 \cdot s \cdot C_5 + 1}}{R_8 + \frac{R_9}{R_9 \cdot s \cdot C_5 + 1}} = \frac{R_9}{R_8 \cdot R_9 \cdot s \cdot C_5 + R_8 + R_9} \quad (4.3)$$

$$T(t) = L^{-1}\{T(s)\} = L^{-1} \left\{ \frac{R_9}{R_8 \cdot R_9 \cdot s \cdot C_5 + R_8 + R_9} \right\} = \frac{R_9}{C_5 \cdot R_8 \cdot R_9} \cdot e^{-\frac{R_8 + R_9}{C_5 \cdot R_8 \cdot R_9} \cdot t} \quad (4.4)$$

where V_i is the voltage across the current sense resistor and V_o is the voltage appearing at the input of the comparator. In this case τ is $2.4 \mu s$.

4.5.1.2 Current Limiting

Limiting the rate of acceleration is a fairly simple task for an IC. However, a simple rate limit may not protect the MOSFETs under an overcurrent condition. This destructive condition can be best prevented with the use of cycle–by–cycle current limiting. Cycle–by–cycle current limiting is accomplished by monitoring the stator current. While the current is building–up an output switch conducts each time, and upon sensing an over current condition, immediately turned off the switch and holded it off for the remaining duration of the oscillator ramp–up period. The stator current is converted to a voltage by inserting a grounded sense resistor R_s (R_{21}) (See Figure 4.8) in series with the three bottom switch transistors (Q_4 , Q_5 , Q_6). The sense voltage,

which is proportional to load current, is fed into a comparator (Pin 9) on board the MC33035. Then this voltage is monitored by the current sense input (Pin 9 and Pin 15) and compared to the internal 100 mV reference. If the 100 mV current sense threshold is exceeded, the comparator resets the lower sense latch (RS Flip-Flop) and the output drivers turn-off the power transistors on the remainder of the oscillator cycle. The Fault output activates during an over current condition. The value for the current sense resistor is:

$$R_s = \frac{0.1V}{I_{STATOR(MAX)}} \quad (4.4)$$

4.5.2 Undervoltage Lockout

The MC33035 provides undervoltage lockouts as terminating the conduction of the drive output transistors if any of three conditions occur. The first one is inadequate voltage to operate the IC. The second one is inadequate voltage to drive the power MOSFET gates. The third one is the condition that MC33035 can not continue its on-board 6.25 V reference. If one or more of the comparators detects an undervoltage condition mentioned above, the Fault Output is activated, the top drives are turned off and the bottom drive outputs are held in a low state.

4.5.3 Thermal Shutdown

If the maximum operation temperature is exceeded, typically at $170^{\circ}C$, the Internal thermal shutdown feature is activated. Then the IC acts as though the Output Enable was grounded.

4.6 Braking

Brake mode is the different mode operation that causes high currents. Upon application of the brake signal, all three bottom transistors are turned on, shorted the motor windings. Since the current circulates between the windings through the three N channels does not appear in the sense resistor and the MC33035 can not detect the high currents in the brake mode. To prevent the rupture, the MOSFETs must be designed to resist very large currents if the brake is used. Motor speed, motor winding resistance, frictional loading and motor inertia effects the time for the current to decay.

5. PRINTED CIRCUIT BOARD (PCB) DESIGN

For preparing the PCB circuit design in this thesis, Altium Designer software is used.

5.1 Creating the PCB Project on Altium Designer

A project in Altium Designer consists of links to all documents and setups related to the design. Once the project is compiled, design verification, synchronization and comparison can take place. Any changes to the original schematics or PCB, for example, are updated in the project when compiled.

5.2 Creating and Drawing the Schematic Document - Circuit

After opening the schematic document, the project name is given to the new schematic document then it is automatically added (linked) to the project. The schematic sheet is now listed under Source Documents beneath the project name in the Projects tab. As the name “thesis” is assigned to the project and schematic document (See Figure 5.1).

5.3 Locating the Component and Loading the Libraries

After creating the schematic document, it is ready to draw the circuit. To manage the thousands of schematic symbols included with Altium Designer, the Schematic Editor provides powerful library search features.

After all the components of the circuit is located to the schematic document following figure is obtained.

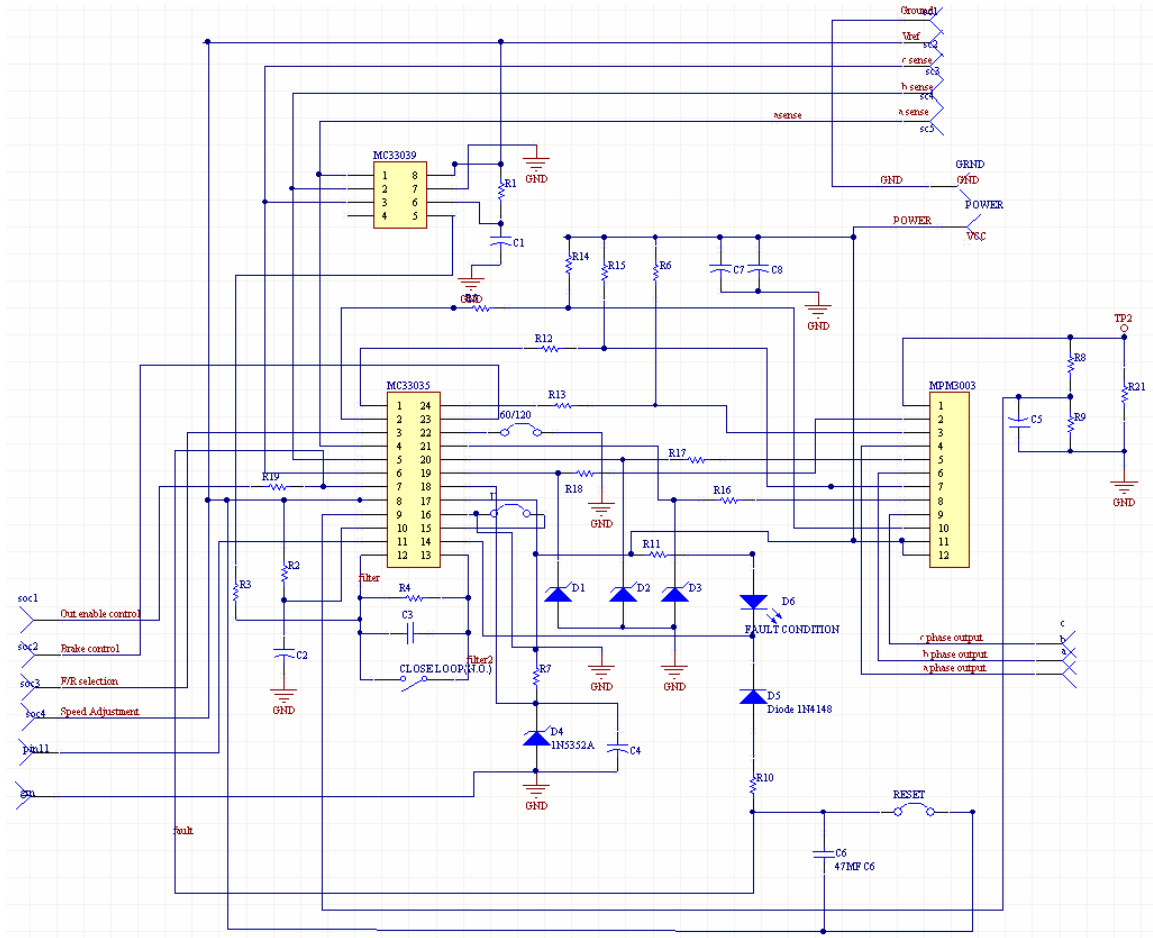


Figure 5.1: Schematic Form of the Closed Loop Speed Control Circuit

5.4 Creating a New PCB and Component Layout

Components layout is very important. Inconvenient component localizations result incompleted routine process. After components are placed on the board, Altium Designer automatically route the board. If some components are placed at irrelevant positions, auto routing action can not be completed or completed with very much contentions and disconnected routes. In this context, after has been tried layout combinations plenty of times, the best layout plan is found as is seen in the Figure 5.3.

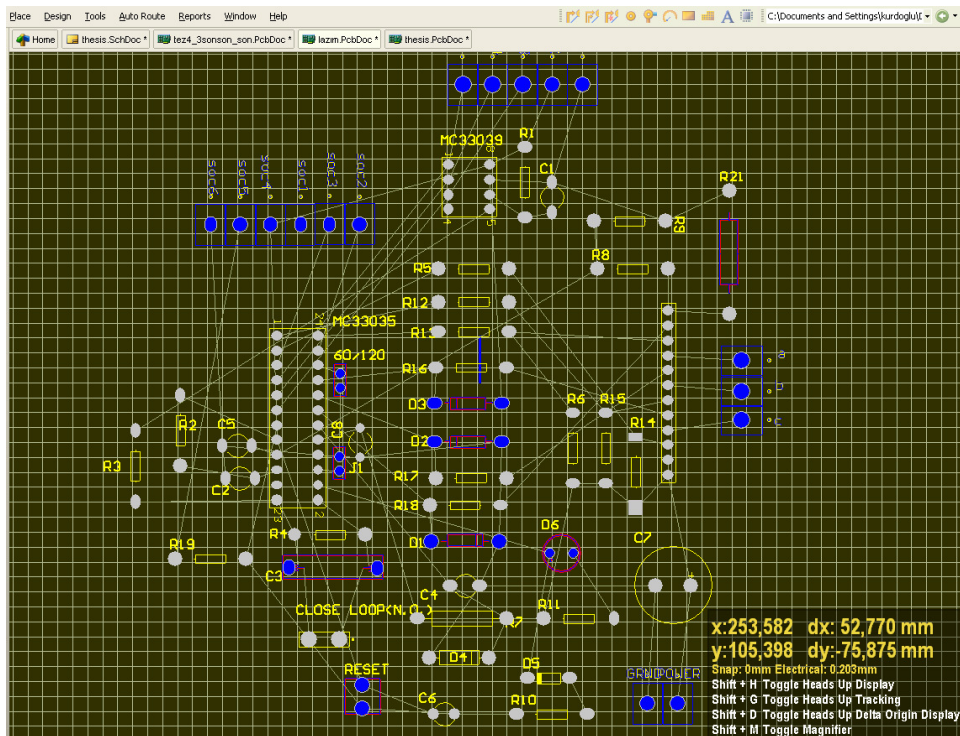


Figure 5.2: The Best Layout Plan Discovered After Many Times of Trials.

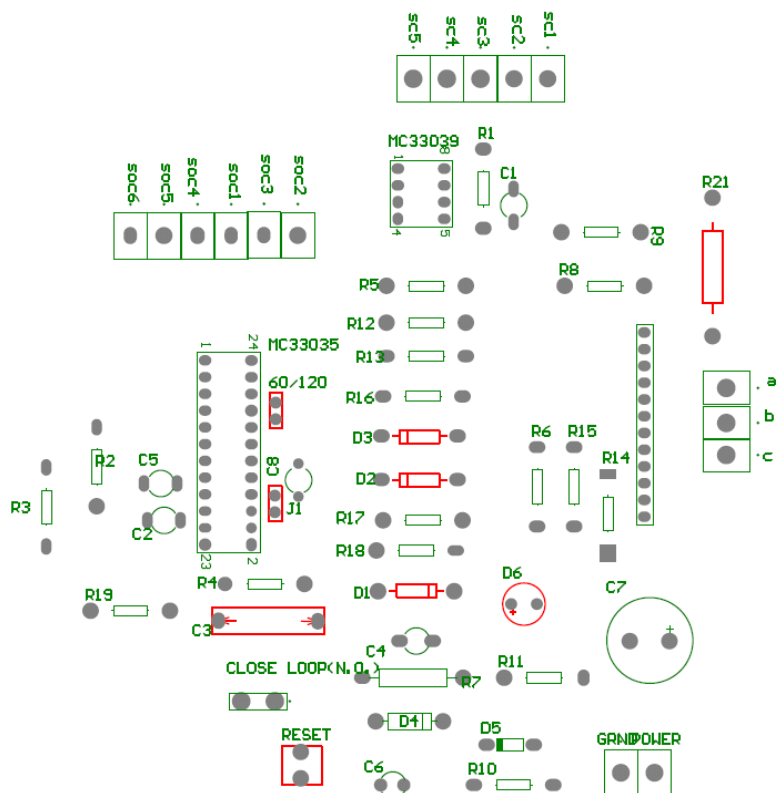


Figure 5.3: The Best Layout Plan Output

After auto-routing and adjusted all of the layers by manual, the following PCB format is obtained (See Figure 5.6).

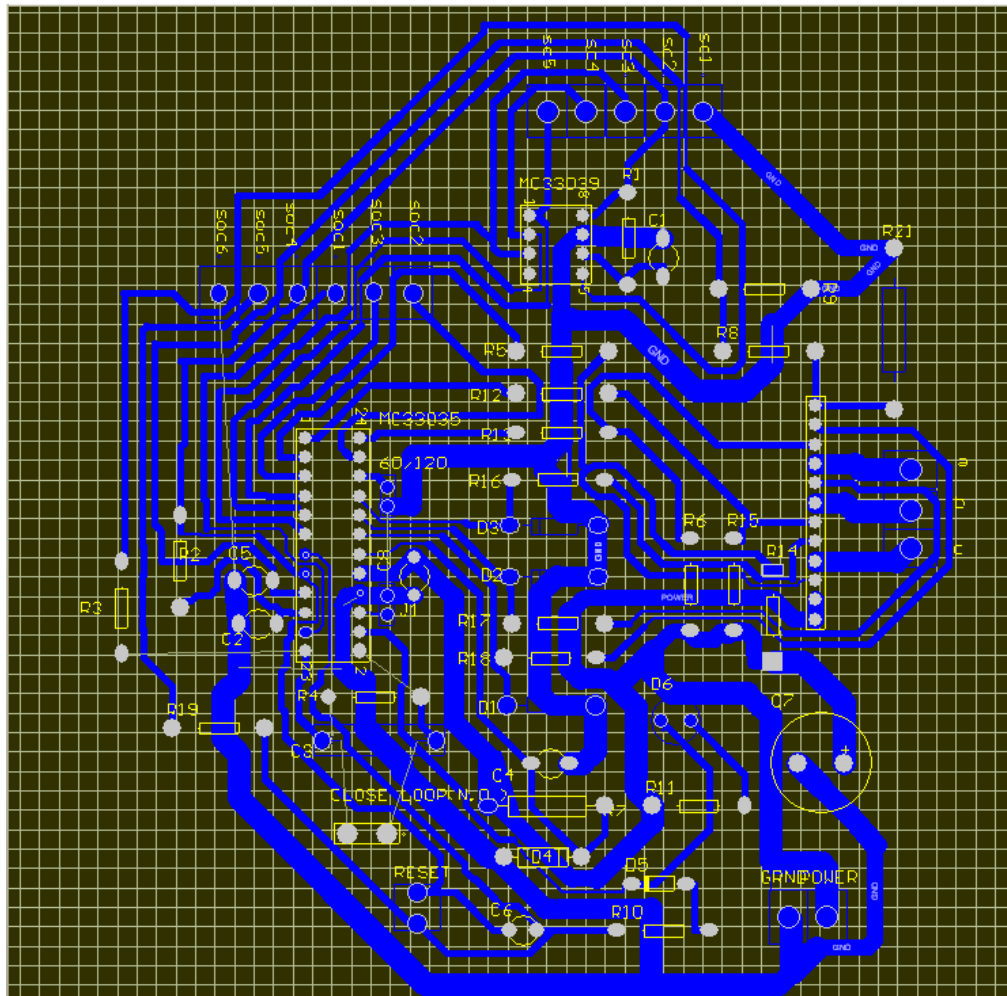


Figure 5.4: PCB Format with the Layers Connections

Although the automatic routing is performed several times, failed to complete 5 connections. Cause not possible to use the top layers, there is only way to complete the disconnections that using the jumpers. Five jumpers are connected as is seen in the Figure 5.7.

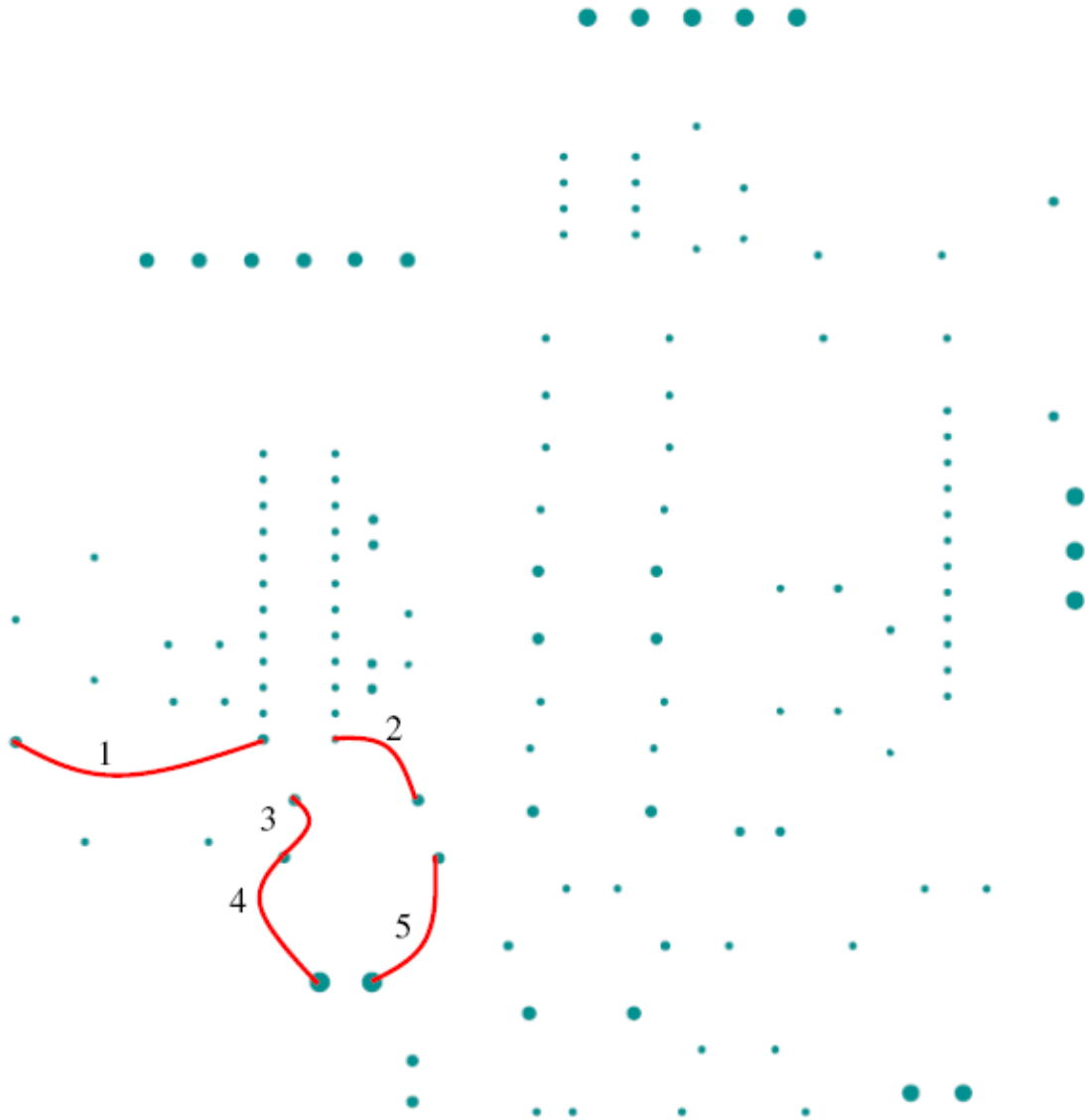
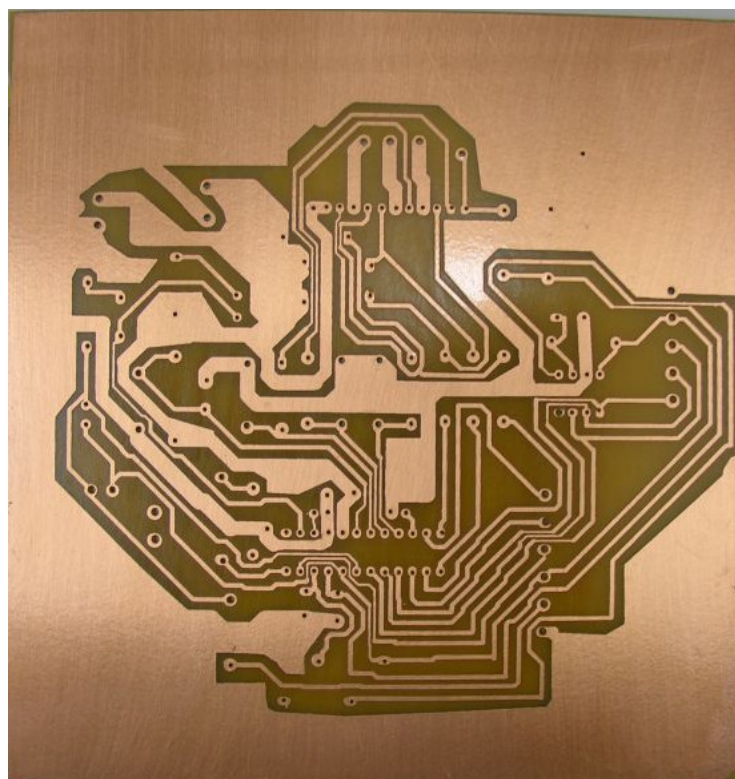
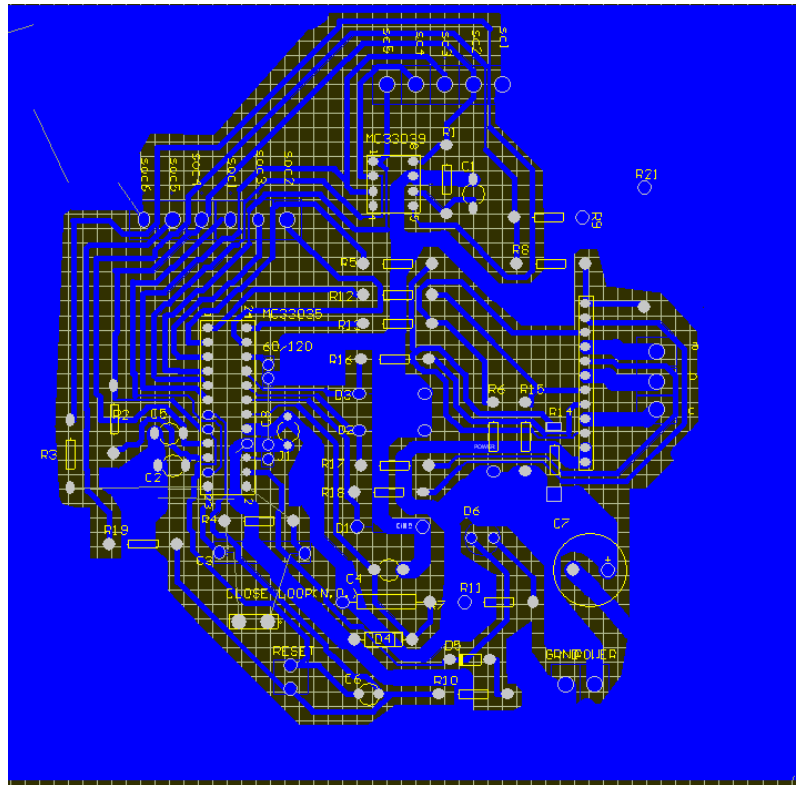


Figure 5.5: Five jumping connections

Next action is to fill the spaces on board regarding the ground and the power routes. Then the PCB Circuit had the final form as is seen in the Figure 5.8 and 5.9.



6. EXPERIMENTAL WORK

The designed circuit is used for driving a miniature BLDC motor. First, the circuit is operated in open-loop control mode which allows a certain PWM value in any loading condition. The voltage and current wave forms are in below figures.

The Motor parameters : 24V, 22W, 10.000 rpm.

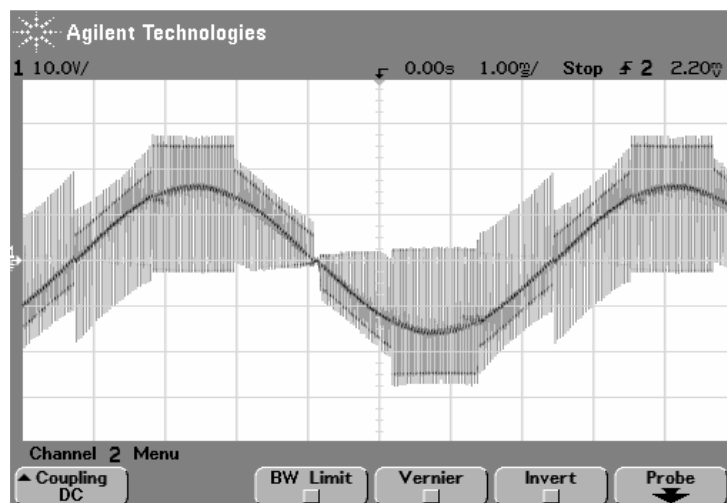


Figure 6.1: Voltage Waveform in 50% PWM

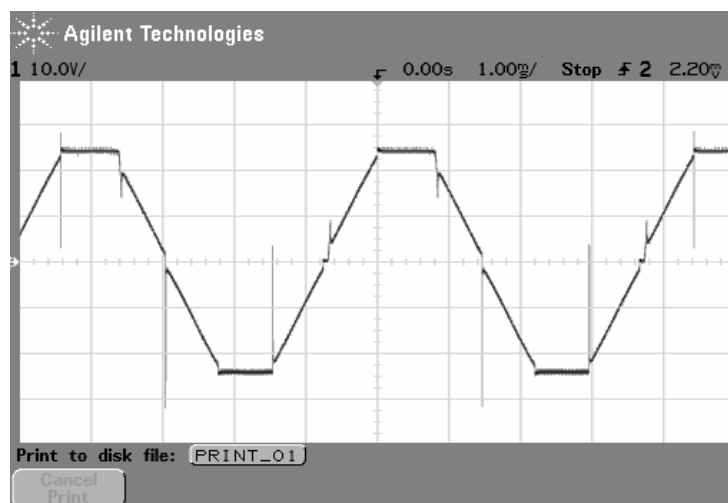


Figure 6.2: Voltage Waveform in 100% PWM

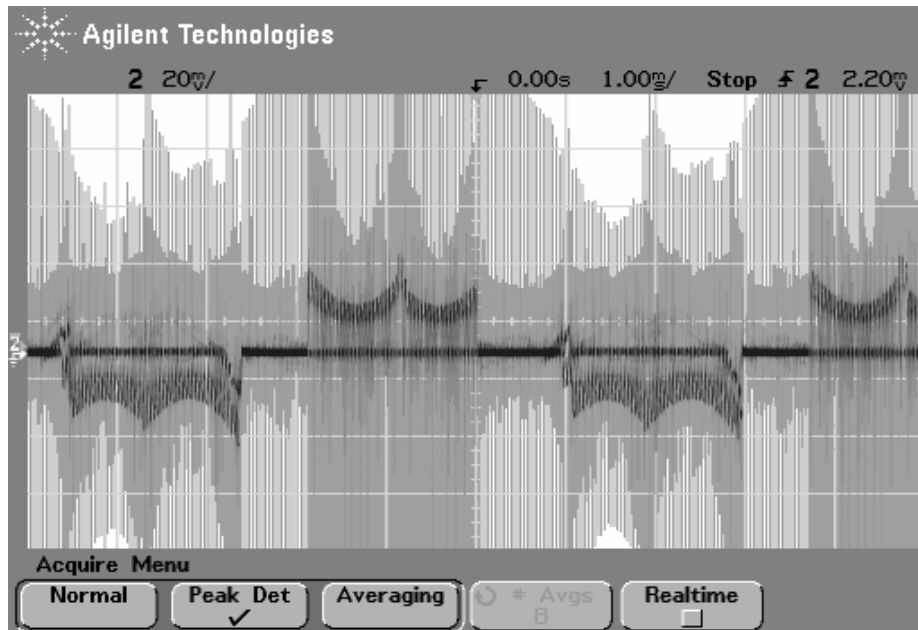


Figure 6.3: Current Waveform

6.1 The Speed Control of the Motor

To perform the speed control to the motor under loaded condition, a system is improved that a generator is coupled to shaft of the brushless motor as is seen in Fig.6.4.

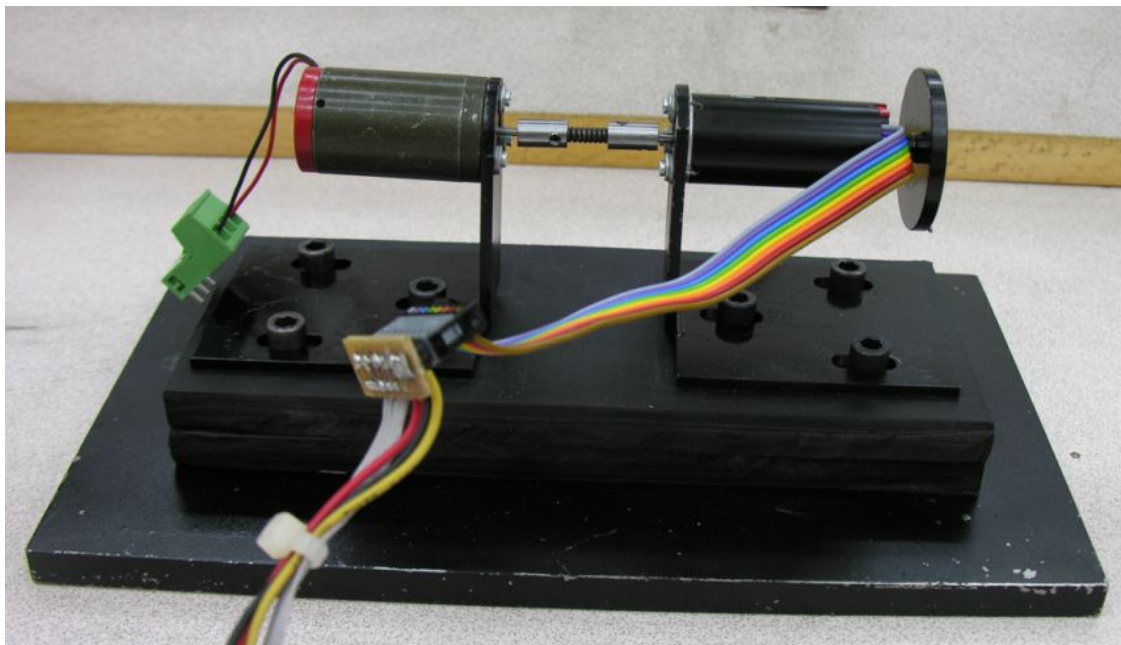


Figure 6.4: The Brushless Motor Loaded with a DC Generator

The motor loading system shown in fig. 6.4 has a problem. It is difficult to provide a proper coupling between motor shaft and generator shaft. without a precise adjustment, the generator loads the motor excessively. To overcome this problem, an improved test setup is built to load the motor properly. In this setup, an aliminium disc is coupled to the motor shaft and a horizontal adjustable magnet is placed in front of the motor. Thus a simple eddy current braking system is builtd. By adjusting the airgap between the magnet and the aliminium disc, the motor could be loaded as required. The improved test setup is shown in Figure 6.5.

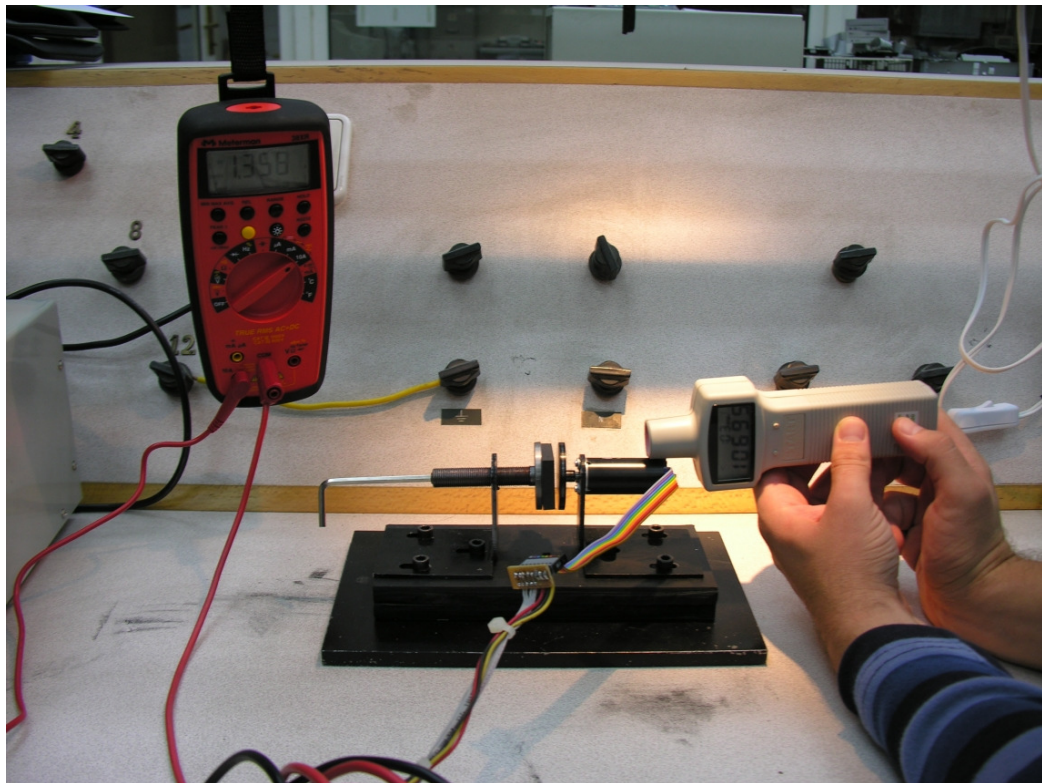


Figure 6.5: The Improved Test Setup (With Eddy Current Braking System)

After the test setup improvement, various exprmental work hae been done by using designed controller. A close view of the line voltage of motor windings at low PWM is shown in Fig. 6.6.

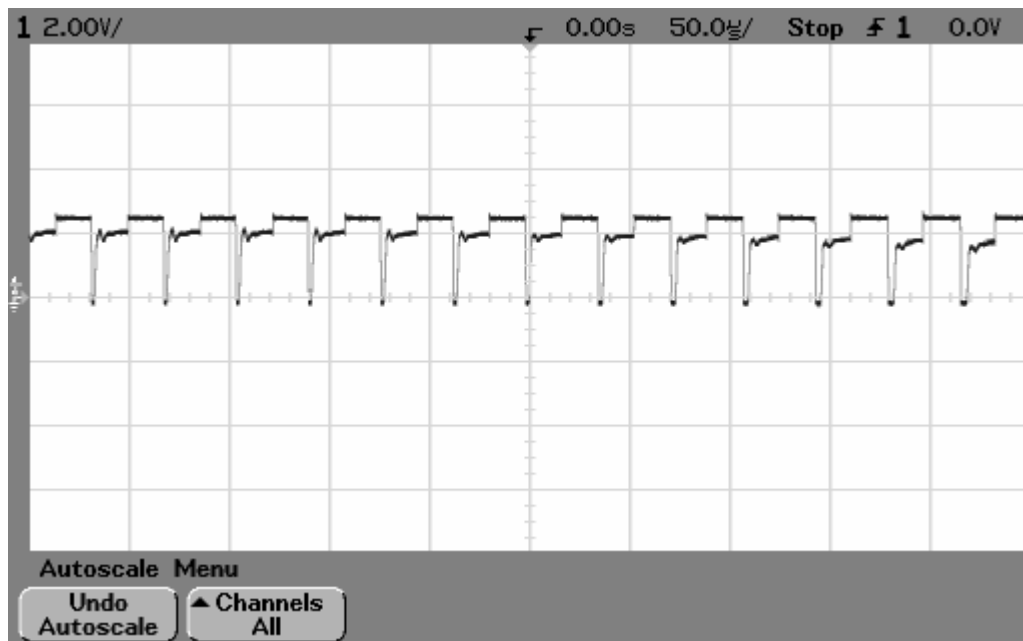


Figure 6.6: The Close View of the Line Voltage of the Motor Windings at Low PWM (Discontinuous Terminal Current Mode)

After loading the motor a closeview of high PWM is shown in the following figure:

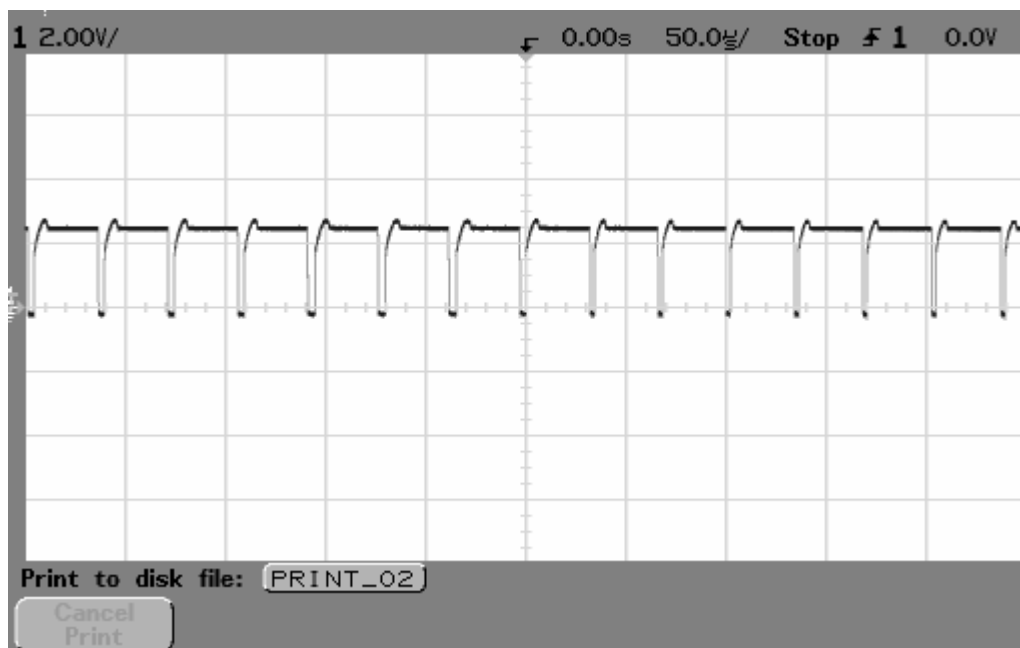


Figure 6.7: The Close View of the Line Voltage of Motor Windings at High PWM. (Continuous Terminal Current Mode)

6.2. Speed Feedback

The speed feedback of the control circuit is provided by MC33039 tachometer IC. The signal outputs from Hall sensors are entered to the MC33039. And this IC produces the speed value as a speed-dependent pulse train. A resistor and a capacitor are used to determine the pulse frequency according to the motor speed. Calculating of R_1 and C_1 for MC33039 pulse frequency:

$$n_{\max} = 14000 \text{ min}^{-1} \quad (\text{no load speed})$$

$$n_{\max} = 234 \text{ s}^{-1}$$

recommended number of pulses during a period = $234 \times 12 = 2808$ per second

$$\frac{1}{2808} = 0.36 \text{ ms}$$

$$C = 22 \text{ nF}$$

$$f_{\text{out}} = \frac{1}{0.36} \text{ ms}$$

Time interval of MC33039 output is determined by the following parameters:

$$C_1 = 22 \text{ nF}$$

$$R_1 = 24 \text{ k}\Omega$$

$$T_{\text{out}} = 0.5 \text{ ms}$$

The frequency output of the motor speed can be seen in Figure 6.

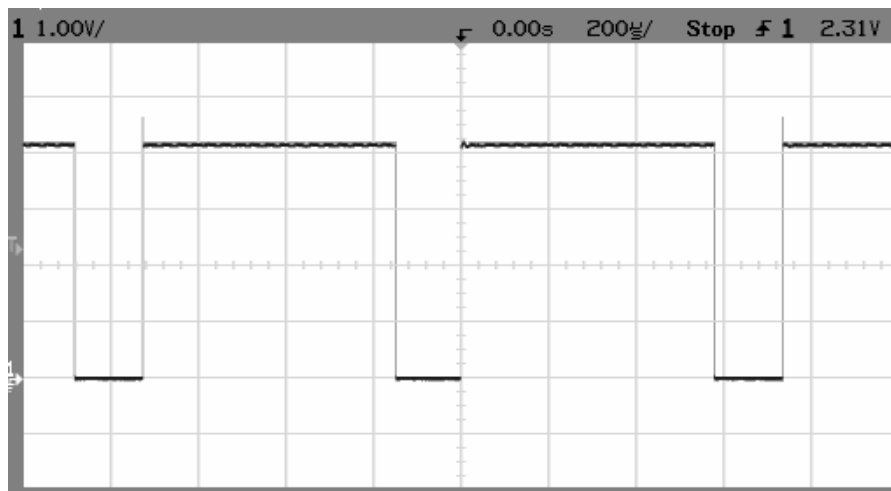


Figure 6.8: The Frequency Output of the Motor Speed.

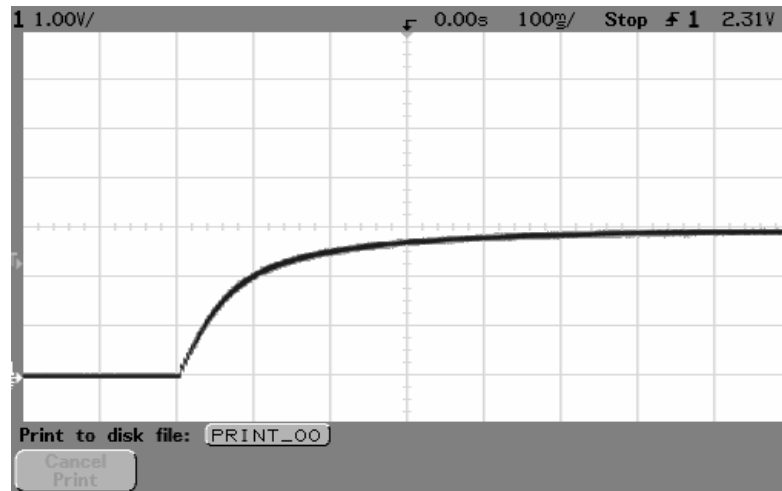


Figure 6.9: Accelerating of motor from standstill to maximum speed.

(Scaling Factor = $4407 \text{ min}^{-1}/\text{V}$)

The output of MC33039 is obtained from the Hall effect sensors by using certain procedures. In the Figure 7.1, the frequency output of the motor speed is shown. The frequency of pulses is increased by accelerating the motor. A low pass filter is used for proper analog speed signal. In Fig.7.2, the speed waveform of accelerating the motor is shown as a voltage waveform.

The speed response of the controller for slightly loaded motor ($I_{bus} = 0.13\text{A}$) is shown in the following figure:

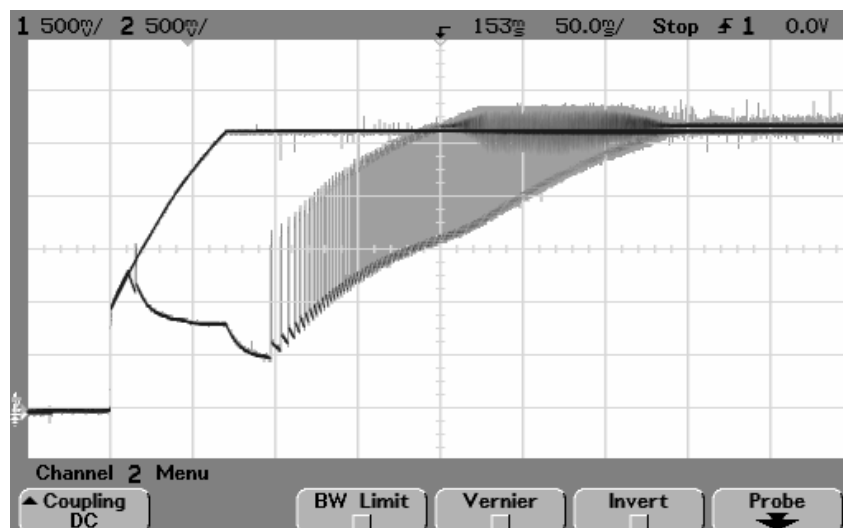


Figure 6.10: The actual motor speed and the command speed as voltage values. The noisy waveform is of MC33039 filtered output (actual motor speed) for slightly loaded motor

The speed response of the controller for highly loaded motor ($I_{bus} = 1.02A$) is shown in the figure.

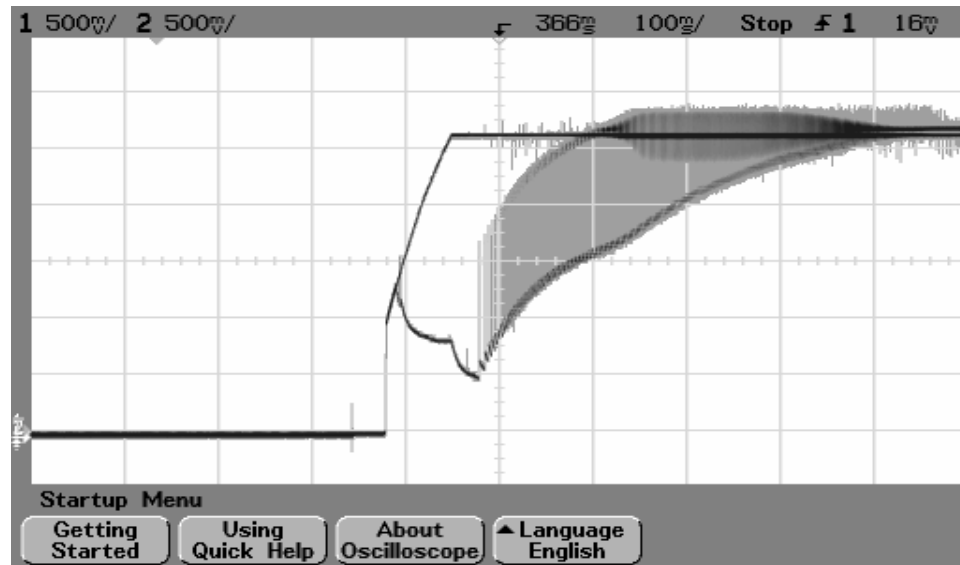


Figure 6.11: The actual motor speed and the command speed as voltage values. The noisy waveform is of MC33039 filtered output (actual motor speed) for highly loaded motor.

The speed control experiment results for a given reference speed are given in Table 6.1. It can be easily seen that the the speed error has a difference range of $\Delta n = 190 \text{ min}^{-1}$ which corresponds the relative speed regulation of 1.8%. This shows that an effective speed control can be achieved by using this circuit.

Table 6.1. Speed Control Experiment Results.

Actual Speed (min^{-1})	Terminal Current (A)	Calculated Electromagnetic Torque (Nm)
10570	0.12	0.02088
10478	0.18	0.03132
10403	0.26	0.04524
10302	0.38	0.06612
10320	0.52	0.09048
10298	0.72	0.12528
10286	0.85	0.14790
10304	1.02	0.17748
10284	1.20	0.20880

7. CONCLUSION

Conventional dc motors are highly efficient and their characteristics make them suitable for use as servomotors. However, their only drawback is that they need a commutator and brushes which are subject to wear and require maintenance. When the functions of commutator and brushes were implemented by solid-state switches, maintenance-free motors were realised. These motors are now known as brushless dc motors.

There are two types for controlling brushless dc motors; sensor and sensorless control. Sensored control type is required less complicated control circuit than sensorless control type. Sensored control is preferred in applications that motor start and stop periodically.

This study is focused to sensored control of a brushless dc motor. A BLDCM speed control driver is designed, manufactured, analyzed and tested for a brushless motor having the parameters of 24 V, 22W, 11.000 rpm (nominal speed). Control circuit has three major parts: MC33035, MC33039 and MSK3003. MC33035 is the brain of the circuit that receive the signals from hall effect sensors for detecting the rotor position and electronically commutate the motor by means of MSK3003 power module pack. MC33039 is a kind of electronic tachometer. MC33035 takes the speed information of the motor from MC33039 and compares them with desired speed data. Thus a closed loop speed control is implemented.

In the application part of the study, brushless dc motor is firstly loaded by a dc generator but this way has a problem that very difficult to provide a proper coupling between the motor shaft and generator shaft, thus a precise adjustment to load the motor can not be achieved. Therefore the brushless motor is loaded by eddy current braking system that is an aluminium disc is coupled to the motor shaft and a horizontal adjustable magnet is placed in front of the motor. By adjusting the airgap between the magnet and the aluminium disc, the motor could be loaded as required properly.

After the test setup is built, several experimental works are performed. In Figure 6.5 the motor is run at different load conditions and the speed of the motor is measured for every load condition thus the following results are obtained:

The motor rpm is 10570 at 0,09A load current

The motor rpm is 10478 at 0,27A load current

The motor rpm is 10403 at 0,859A load current

The motor rpm is 10302 at 1,219A load current

The motor rpm is 10298 at 0.72load current

After observing the above experiment results, it is clearly seen that the motor speed change very slightly at different load conditions and this result proves the system implement PWM operation very well.

In Fig.6.6 and 6.7, line voltage of the motor winding graphs are obtained at high and low PWM for the motor is loaded and unloaded respectively.

In Fig.7.1 the motor speed frequency output is obtained and observed that frequency of pulses is increased by accelerating the motor. In the figure 7.2, a voltage waveform is obtained while the motor is accelerating.

Fig 7.3 and 7.4 are noisy waveforms of MC33039 for slightly and highly loaded motor conditions. These graphs show the speed response of the controller for two different conditions.

In conclusion, an overviewing of experimental works results that the manufactured control circuit drive the motor at nearly constant speed in spite of different load conditions and the speed response of the system at different load conditions is very good. These observations show that the control circuit perform closed loop speed control applications as properly.

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RESUMEE

Alper Kurdoglu has been borned in Istanbul, 1980. He has received his bachelor of Electrical Engineering degree from Istanbul Technical University, Turkey in 2003. From 2003 to 2006, he worked as Avionic Engineer at MNG Technig Aircraft Maintenance Company. He currently works at General Directorate of States Hydrolic Works in Istanbul since 2006, December.

APPENDIX

ISO-9001 CERTIFIED BY DSCC

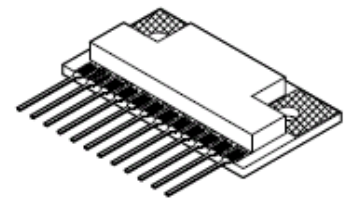


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FEATURES:

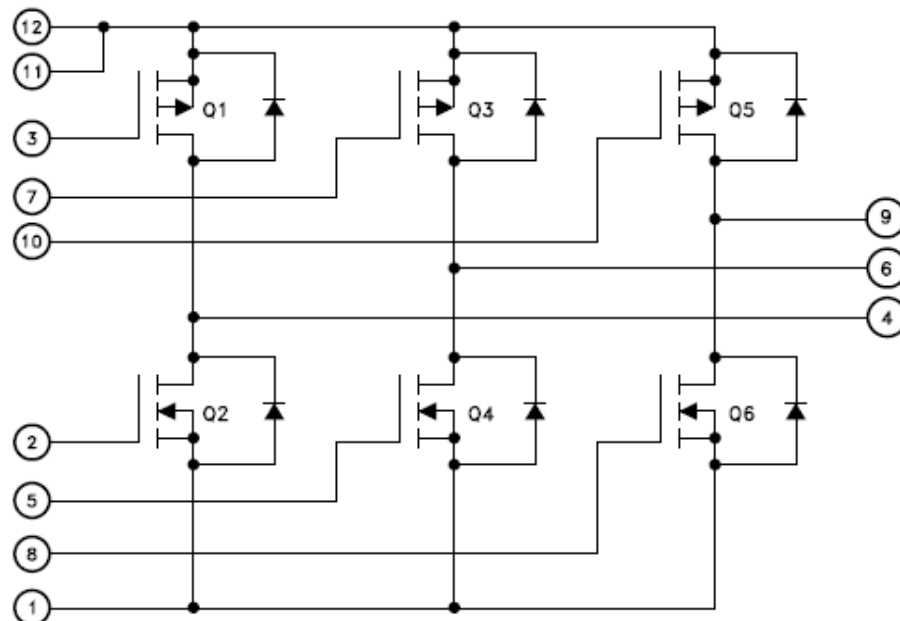
- Pin Compatible with MPM3003
- P and N Channel MOSFETs for Ease of Drive
- Isolated Package for Direct Heat Sinking, Excellent Thermal Conductivity
- Avalanche Rated Devices
- Interfaces Directly with Most Brushless Motor Drive IC's
- 55 Volt, 10 Amp Full Three Phase Bridge



DESCRIPTION:

The MSK 3003 is a three phase bridge power circuit packaged in a space efficient isolated ceramic tab power SIP package. Consisting of P-Channel MOSFETs for the top transistors and N-Channel MOSFETs for the bottom transistors, the MSK 3003 will interface directly with most brushless motor drive IC's without special gate driving requirements. The MSK 3003 uses M.S.Kennedy's proven power hybrid technology to bring a cost effective high performance circuit for use in today's sophisticated servo motor and disk drive systems. The MSK 3003 is a replacement for the MPM3003 with only minor differences in mechanical specifications.

EQUIVALENT SCHEMATIC



TYPICAL APPLICATIONS

- Three Phase Brushless DC Motor Servo Control
- Disk Drive Spindle Control
- Fin Actuator Control
- Az-El Antenna Control

PIN-OUT INFORMATION

1 Source 2,4,6	12 Source 1,3,5
2 Gate 2	11 Source 1,3,5
3 Gate 1	10 Gate 5
4 Drain 1,2	9 Drain 5,6
5 Gate 4	8 Gate 6
6 Drain 3,4	7 Gate 3

ABSOLUTE MAXIMUM RATINGS

VDSS	Drain to Source Voltage	55V MAX	Single Pulse Avalanche Energy	
VDGDR	Drain to Gate Voltage		(Q2,Q4,Q6)	71mJ
	(RGS = 1M Ω)	55V MAX	(Q1,Q3,Q5)	96mJ
VGS	Gate to Source Voltage		TJ	Junction Temperature
	(Continuous)	± 20 V MAX	TST	Storage Temperature
ID	Continuous Current	10A MAX	TC	Case Operating Temperature Range
IDM	Pulsed Current	25A MAX	TL	Lead Temperature Range
RTH-JC	Thermal Resistance			(10 Seconds)
	(Junction to Case)	7.9°C/W		300°C MAX

ELECTRICAL SPECIFICATIONS

Parameter	Test Conditions ④	MSK3003			Units
		Min.	Typ.	Max.	
Drain-Source Breakdown Voltage	VGS = 0 ID = 0.25mA (All Transistors)	55	-	-	V
Drain-Source Leakage Current	VDS = 55V VGS = 0V (Q2,Q4,Q6)	-	-	25	μ A
	VDS = -55V VGS = 0V (Q1,Q3,Q5)	-	-	-25	μ A
Gate-Source Leakage Current	VGS = ± 20 V VDS = 0 (All Transistors)	-	-	± 100	nA
Gate-Source Threshold Voltage	VDS = VGS ID = 250 μ A (Q2,Q4,Q6)	2.0	-	4.5	V
	VDS = VGS ID = 250 μ A (Q1,Q3,Q5)	-2.0	-	-4.5	V
Drain-Source On Resistance ②	VGS = 10V ID = 10A (Q2,Q4,Q6)	-	-	0.15	Ω
	VGS = -10V ID = -7.2A (Q1,Q3,Q5)	-	-	0.28	Ω
Drain-Source On Resistance ③	VGS = 10V ID = 10A (Q2,Q4,Q6)	-	-	0.07	Ω
	VGS = 10V ID = -7.2A (Q1,Q3,Q5)	-	-	0.175	Ω
Forward Transconductance ①	VDS = 25V ID = 10A (Q2,Q4,Q6)	4.5	-	-	S
	VDS = -25V ID = -7.2A (Q1,Q3,Q5)	2.5	-	-	S
N-Channel (Q2,Q4,Q6)					
Total Gate Charge ①	ID = 10A	-	-	20	nC
Gate-Source Charge ①	VDS = 44V	-	-	5.3	nC
Gate-Drain Charge ①	VDS = 10V	-	-	7.6	nC
Turn-On Delay Time ①	VDS = 28V	-	4.9	-	nS
Rise Time ①	ID = 10A	-	34	-	nS
Turn-Off Delay Time ①	RG = 24 Ω	-	19	-	nS
Fall Time ①	RO = 2.6 Ω	-	27	-	nS
Input Capacitance ①	VGS = 0V	-	370	-	pF
Output Capacitance ①	VDS = 25V	-	140	-	pF
Reverse Transfer Capacitance ①	f = 1MHz	-	65	-	pF
P-Channel (Q1,Q3,Q5)					
Total Gate Charge ①	ID = -7.2A	-	-	19	nC
Gate-Source Charge ①	VDS = -44V	-	-	5.1	nC
Gate-Drain Charge ①	VDS = -10V	-	-	10	nC
Turn-On Delay Time ①	VDS = -28V	-	13	-	nS
Rise Time ①	ID = -7.2A	-	55	-	nS
Turn-Off Delay Time ①	RG = 24 Ω	-	23	-	nS
Fall Time ①	RO = 3.7 Ω	-	37	-	nS
Input Capacitance ①	VGS = 0V	-	350	-	pF
Output Capacitance ①	VDS = -25V	-	170	-	pF
Reverse Transfer Capacitance ①	f = 1MHz	-	92	-	pF
BODY DIODE					
Forward On Voltage ①	IS = 10A VGS = 0V (Q2,Q4,Q6)	-	1.3	-	V
	IS = -7.2A VGS = 0V (Q1,Q3,Q5)	-	-1.6	-	V
Reverse Recovery Time ①	IS = 10A di/dt = 100A/ μ S (Q2,Q4,Q6)	-	56	83	nS
	IS = -7.2A di/dt = 100A/ μ S (Q1,Q3,Q5)	-	47	71	nS
Reverse Recovery Charge ①	IS = 10A di/dt = 100A/ μ S (Q2,Q4,Q6)	-	0.12	0.18	μ C
	IS = -7.2A di/dt = 100A/ μ S (Q1,Q3,Q5)	-	0.084	0.13	μ C

NOTES:

- ① This parameter is guaranteed by design but need not be tested. Typical parameters are representative of actual device performance but are for reference only.
- ② Resistance as seen at package pins.
- ③ Resistance for die only; use for thermal calculations.
- ④ TA = 25°C unless otherwise specified.

APPLICATION NOTES

N-CHANNEL GATES (Q2, Q4, Q6)

For driving the N-Channel gates, it is important to keep in mind that it is essentially like driving a capacitance to a sufficient voltage to get the channel fully on. Driving the gates to +15 volts with respect to their sources assures that the transistors are on. This will keep the dissipation down to a minimum level [$R_{DS(ON)}$ specified in the data sheet]. How quickly the gate gets turned ON and OFF will determine the dissipation of the transistor while it is transitioning from OFF to ON, and vice-versa. Turning the gate ON and OFF too slow will cause excessive dissipation, while turning it ON and OFF too fast will cause excessive switching noise in the system. It is important to have as low a driving impedance as practical for the size of the transistor. Many motor drive IC's have sufficient gate drive capability for the MSK 3003. If not, paralleled CMOS standard gates will usually be sufficient. A series resistor in the gate circuit slows it down, but also suppresses any ringing caused by stray inductances in the MOSFET circuit. The selection of the resistor is determined by how fast the MOSFET wants to be switched. See Figure 1 for circuit details.

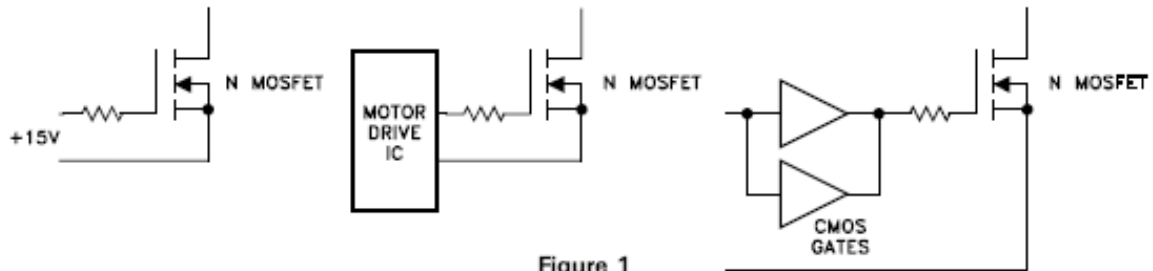


Figure 1

P-CHANNEL GATES (Q1, Q3, Q5)

Most everything applies to driving the P-Channel gates as the N-Channel gates. The only difference is that the P-Channel gate to source voltage needs to be negative. Most motor drive IC's are set up with an open collector or drain output for directly interfacing with the P-channel gates. If not, an external common emitter switching transistor configuration (see Figure 2) will turn the P-Channel MOSFET on. All the other rules of MOSFET gate drive apply here. For high supply voltages, additional circuitry must be used to protect the P-Channel gate from excessive voltages.

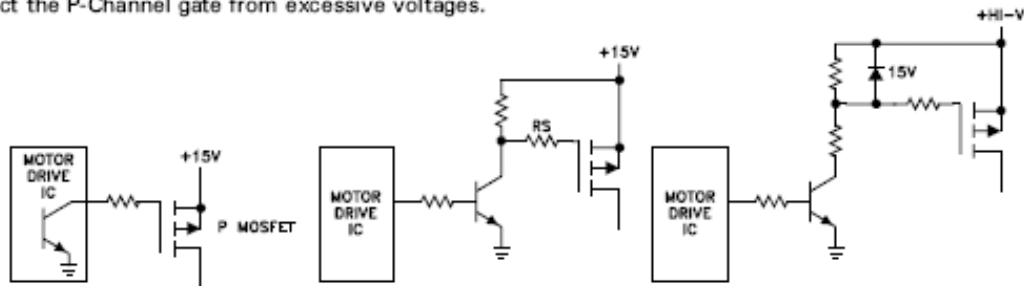


Figure 2

BRIDGE DRIVE CONSIDERATIONS

It is important that the logic used to turn ON and OFF the various transistors allow sufficient "dead time" between a high side transistor and its low side transistor to make sure that at no time are they both ON. When they are, this is called "shoot-through", and it places a momentary short across the power supply. This over stresses the transistors and causes excessive noise as well. See Figure 3.

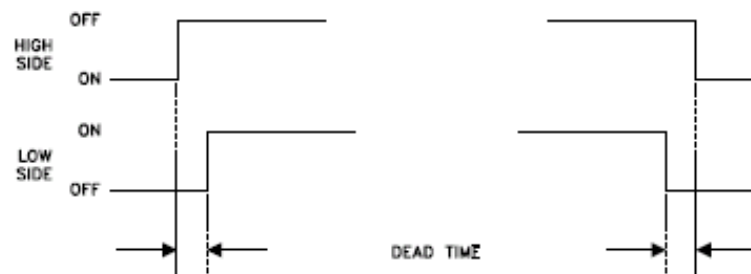
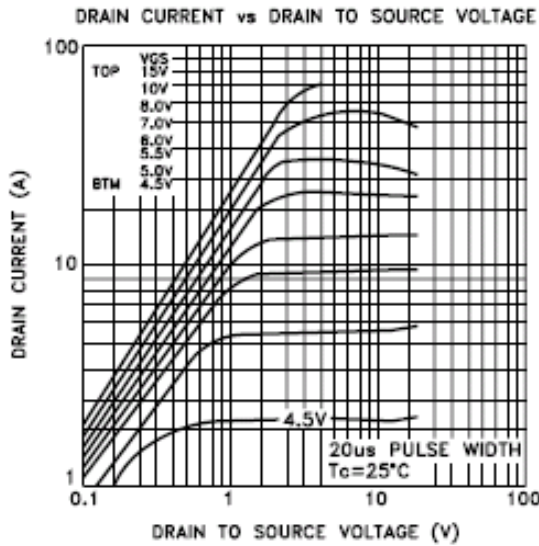


Figure 3

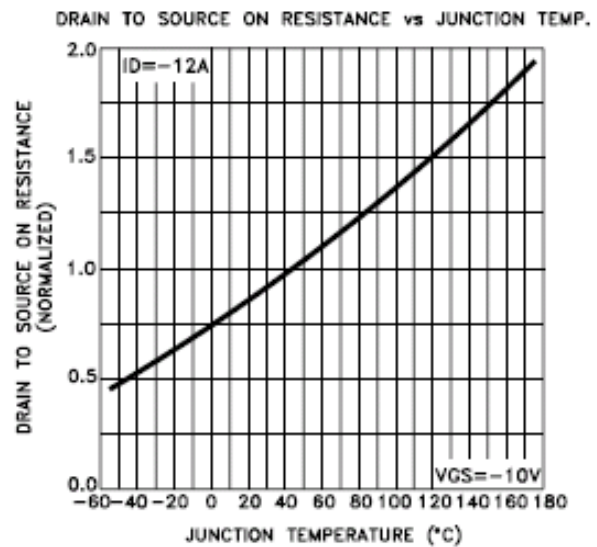
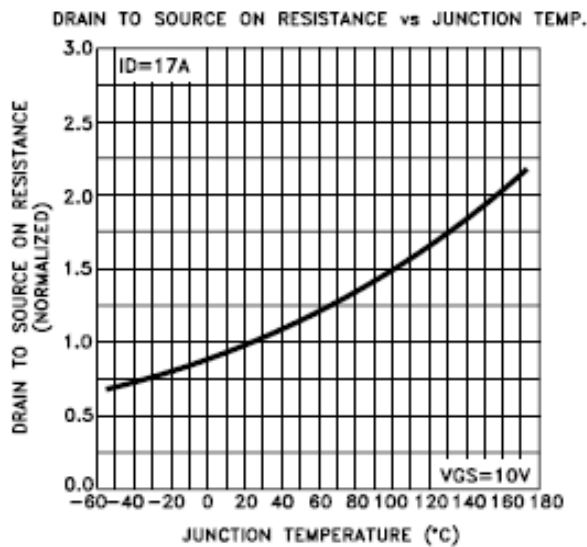
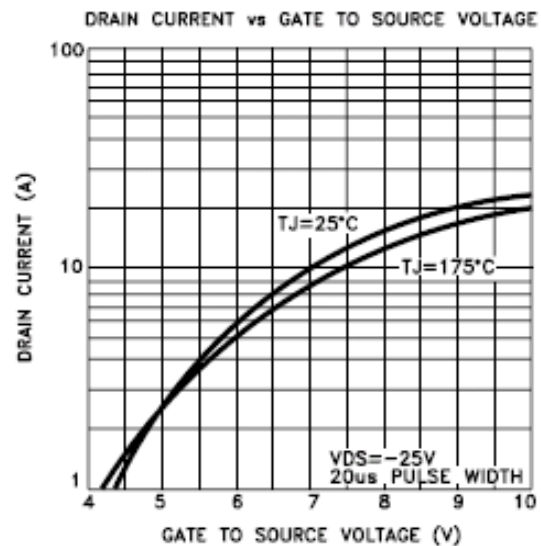
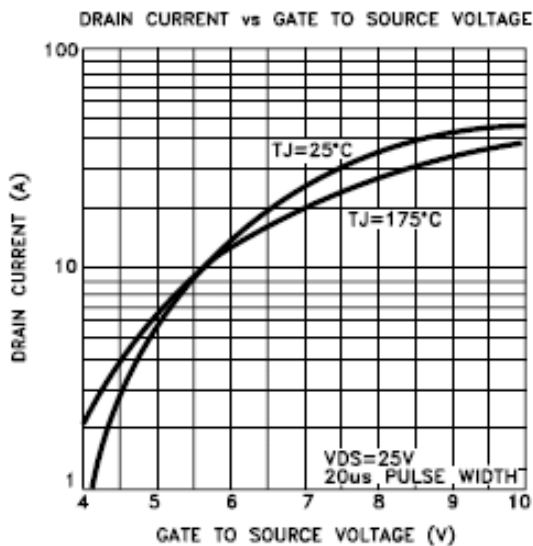
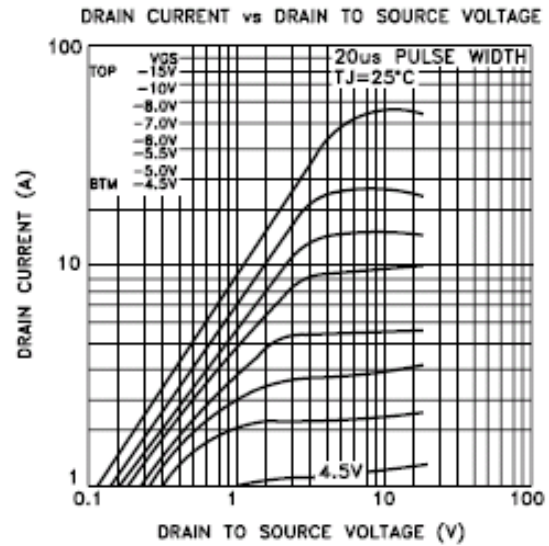
This deadtime should allow for the turn on and turn off time of the transistors, especially when slowing them down with gate resistors. This situation will be present when switching motor direction, or when sophisticated timing schemes are used for servo systems such as locked antiphase PWM'ing for high bandwidth operation.

TYPICAL PERFORMANCE CURVES

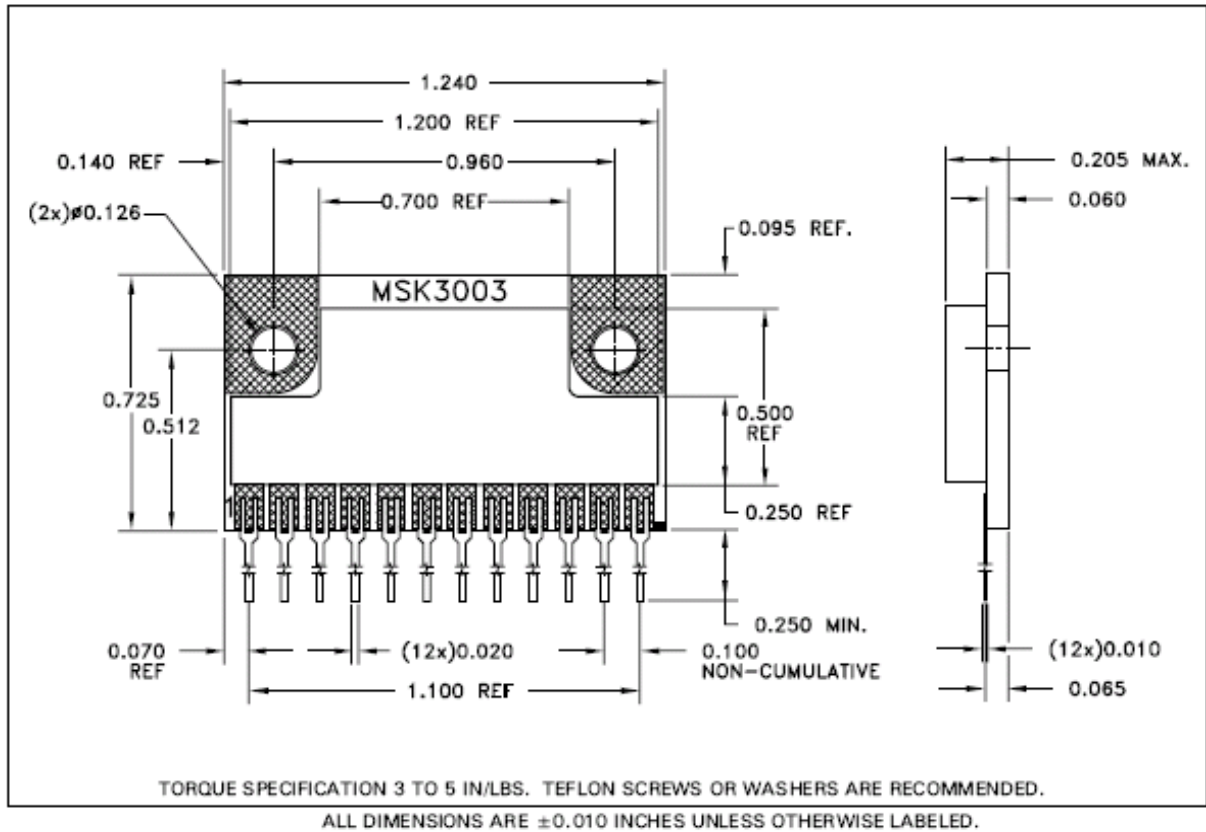
N-CHANNEL DEVICES (Q2,Q4,Q6)



P-CHANNEL DEVICES (Q1,Q3,Q5)



MECHANICAL SPECIFICATIONS



ORDERING INFORMATION

PART NUMBER	SCREENING LEVEL
MSK 3003	Industrial

M.S. Kennedy Corp.
4707 Dey Road, Liverpool, New York 13088
Phone (315) 701-6751
FAX (315) 701-6752
www.mskennedy.com

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