



Application Note 9019

October, 2001

Motor Drive System Using SPM Inverter

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1. Introduction

The terms “energy-saving” and “quiet-running” are becoming very important in the world of variable speed motor drives. Inverter technology is being accepted by a wide range of users in the design of these products, and their use is increasing.

For low-power motor control, there are increasing demands for compactness, built-in control, and lower overall-cost. An important consideration, in justifying the use of inverters in these applications, is to optimize the total-cost-performance ratio of the overall drive system. In other words, the systems have to be less noisy, more efficient, smaller and lighter, more advanced in function and more accurate in control with a very low cost.

In order to meet these needs, Fairchild has developed a new series of compact, high-functionality, and high efficiency power semiconductor devices called “SPM (Smart Power Module)”. SPM-based inverters are now considered an attractive alternative to conventional discrete-based inverters for low-power motor drives, specifically for appliances such as washing machines, air-conditioners etc.

SPM combines optimized circuit protection and drive matched to the IGBT’s switching characteristics. Highly effective short-circuit current detection/protection is achieved through the use of advanced current sensing IGBT chips that allow continuous monitoring of the IGBT’s current. System reliability is further enhanced by the integrated under-voltage protection function. The high speed built-in HVIC provides an opto-coupler-less IGBT gate driving capability that further reduces the overall size of the inverter system design. Additionally, the incorporated HVIC allows the use of a single-supply drive topology without negative bias. It should be noted that one could build a typical block diagram of an SPM-inverter as shown in Fig. 1.1.

Small power drive systems using 3-phase voltage-fed inverters, which are frequently used in home appliances, are being used in an increasing number of applications. Induction motors and BLDC motors are mostly used in these drive systems. Induction motors are used more often because they have the advantages of price and durability. In home appliances, except where high performance control is a requirement, the constant V/Hz principle that maintains a constant ratio between the stator voltage magnitude and the stator voltage frequency is generally used. This keeps the stator magnetic field magnitude constant. The advantage of this method is a simple control scheme that allows for the design of a low priced system. The BLDCM has high output power per volume due to good heat radiation. It also has the merits of high efficiency, low audible noise, and ease of control. The output torque of a trapezoidal back emf (electromotive force) BLDCM is proportional to the motor input current regardless of the rotor position. Although the output torque has a ripple component, the BLDCM has a merit that the driver can be designed with low cost because the drive circuit is relatively simple. A sinusoidal back emf BLDCM is frequently applied in precision servo control applications because it does not have the output torque ripple. But the drive system can be complicated because the sinusoidal current with respect to the absolute rotor position should be applied to each phase. The simple speed control system of the trapezoidal back emf BLDCM using a hall sensor, which can detect the rotor position, is shown.

PWM signals are pulse trains that have a variable width, constant frequency, and a constant magnitude. A PWM signal has one pulse per one PWM period, and its width varies with respect to the modulating signal. The frequency of a PWM signal should be greater than that of a modulating signal. The most frequently used continuous voltage modulation method is the symmetric PWM, which means that each pulse is symmetric with respect to the PWM period. The offset voltage injection symmetric SVPWM that maximizes the DC link voltage utilization, with a relatively light computation load, is shown.

The objective of this application note is to show the details of SPM power circuit design and its drive applications to SPM users. This note describes some designs that should enable users in deploying the SPM expeditiously and shorten their development time. It will make inverter designers very familiar with the SPM and help them in incorporating it in their designs.

2. SPM-Inverter System Overview

2.1 Introduction

An Induction and brush less DC motor drive system using an SPM (Smart Power Module) has been designed. In this chapter, an overview of the entire system is given, and the more details, including drive performance, are described in another application note.

Fig. 2.1 shows the block diagram of the drive system and Fig. 2.2 is the external view of the system. From Fig. 2.1, it may be seen that the control system consists of three parts. The first part is composed of the line filter, rectifier, and SMPS (Switching Mode Power Supply) for power circuit blocks. The second is the SPM block for three-phase inverter circuit. Finally there are the processor and user interface blocks for the control part. The system is fully assembled on a board and can be connected to the AC power source and motors as shown in Fig. 2.2. It is appropriate for home appliance applications. The system is configured for the operation of induction motors and brush less DC motors without any external circuit. The operation of a brush less DC motor can be achieved even without a mechanical position sensor as required by numerous other low cost applications. The power requirements for a large number of applications, such as industrial tools and home appliances, range between 500W and 1kW. A typical application is the one-horse power or 750W drive the inverter is designed for. TMS320F2406 is used for the control processor, which is one of the C2000 series. It is designed for digital motor control.

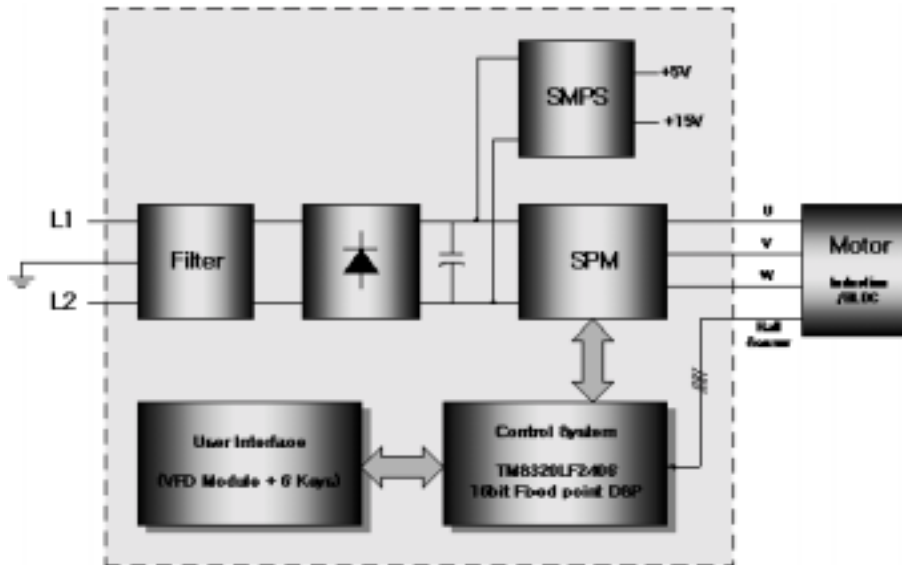


Fig. 2.1 Block diagram of the drive system using an SPM-inverter



(a) Control board



(b) Control board with BLDC motor

Fig. 2.2 External view of the drive board

The present configuration allows the immediate operation of an induction motor with 750W of power on the shaft. An equivalent brush less DC motor requires the proper adjustment of the wire connection located on the motor input part.

The inverter can be operated at 220/110V with 50/60Hz. The rectified voltage from the 220V mains optimally requires a motor with a line-to-line voltage of 220V due to the space-vector based PWM method implemented. The rms current amounts to about 3A for providing an output power of 750W. Similarly, the current at the 110V mains increases to 6A.

The motor control program running on the micro controller is implemented by using a 16 digit * 2 line VFD (Vacuum Fluorescent Display) and 6 keys. The keypad allows users to set up the switching frequency, dead time, acceleration time, deceleration time, base voltage, base frequency, frequency reference, torque boost etc. Three kinds of pattern operation are possible. A manually controlled volume also supports the variable speed operation of the motors.

2.2 Line Filter

Power electronic circuits, by switching large amounts of current at high voltages, can generate electrical signals that affect other electronic systems. These unwanted signals give rise to electromagnetic interference (EMI), also known as radio frequency interference (RFI), since they occur at higher frequencies. These signals can be transmitted by radiation through space or conduction along cable.

Apart from emitting EMI, the control circuit of power systems can also be affected by EMI generated by its own power circuitry, by other circuits or by natural phenomena. When this occurs the system is said to be susceptible to EMI.

Any system, which does not emit EMI above a given level, and is not affected by EMI, is stated to have achieved electromagnetic compatibility (EMC).

There are three elements to any EMC system. The source of the EMI, the media through which it is transmitted, and the receptor, which is any system that suffers adversely due to the received EMI. Therefore electromagnetic compatibility can be achieved by reducing the EMI levels from the source, blocking the propagation path of the EMI signals, or by making the receiver less susceptible to the received EMI signals.

The source of the EMI is primarily any system where the current or voltage changes rapidly (for example, the breaking of current by relay contacts, arcing of motor commutators, high-frequency switching such as the rapid turn-on and turn-off of IGBTs).

EMI can be radiated through space as electromagnetic waves, or it can be conducted as a current along a cable. Conduction can take the form of common-mode or differential-mode currents. In differential-mode, the currents are equal and opposite on the two wires and are caused primarily by other users on the same lines. Common-mode currents are almost equal in amplitude on the two lines, but travel in the same direction. These currents are mainly caused by the coupling of radiated EMI to the power lines and by stray capacitive coupling to the body of the equipment.

Emissions can be classified as broadband and narrowband. In broadband emission, the signal bandwidth is greater than the reference bandwidth, and the pulse-repetition frequency is less than that of the reference bandwidth. The reference bandwidth, for EMC purposes, may be considered to be the equipment being interfered with, or a test receiver. In a narrowband emission, the signal bandwidth is less than the reference bandwidth, and the pulse-repetition frequency is greater than that of the reference bandwidth.

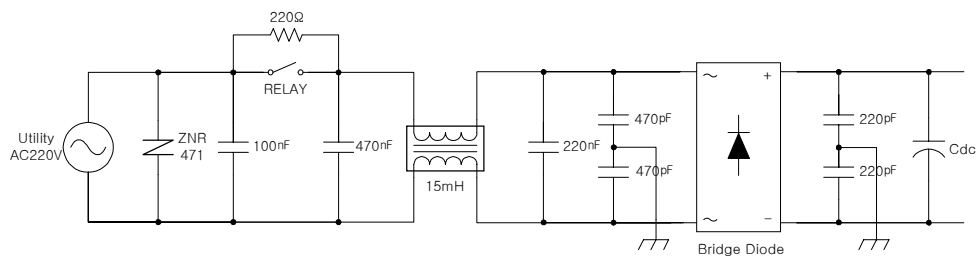


Fig. 2.3 Input power stage with the line-filter

EMI sources can be broadly divided into two categories, natural and man-made. Naturally caused EMI, below 10MHz, is mainly due to atmospheric noise resulting from electrical storms. Above 10MHz they are primarily a result of cosmic noise and solar radiation. Man-made EMI can be intentional or unintentional. It is the variation of the voltage and current which produces EMI, whose magnitude depends on the value of the current, the length of the conductors, the rate of change of the voltage and current, and the physical position of the conductors relative to each other and any earth plains.

Protection against conducted EMI being transmitted along a cable is achieved by means of suppression filters, which consist basically of inductive and capacitive elements. A variety of such filters exist since one type, which suppresses interference completely from one system, may be quite useless in another.

The location of the filter is also important. It should generally be placed directly at the source of interference, and the output and input leads should never be bundled together. The components chosen for the filters should also be carefully designed. The inductors must have low stray capacitors, so they should not be multi-layer, and the capacitors must have low series inductance. The filter should be enclosed in a screened box as much as possible. This is connected to the wall of the shielded enclosure, so that interference signals from the noisy side do not mix with the quiet side of the system.

Fig. 2.3 shows an example of one of the basic filter arrangements, which are used in the designed drive system for suppressing the conducted interference. This includes the rectifying bridge diode, the varistor to suppress an input spike voltage, and the relay circuit for the initial charging of the DC-link capacitor. It may be noted that there are choke coil and Y-connected capacitors against common-mode noise, and AC capacitors against differential-mode noise.

2.3 Initial Charging Circuit

On turning on the system power, there would be high current at the system input line. In order to prevent an over-current damage of the bridge diode and the DC-link capacitor, the initial charging circuit is designed as shown in Fig. 2.4.

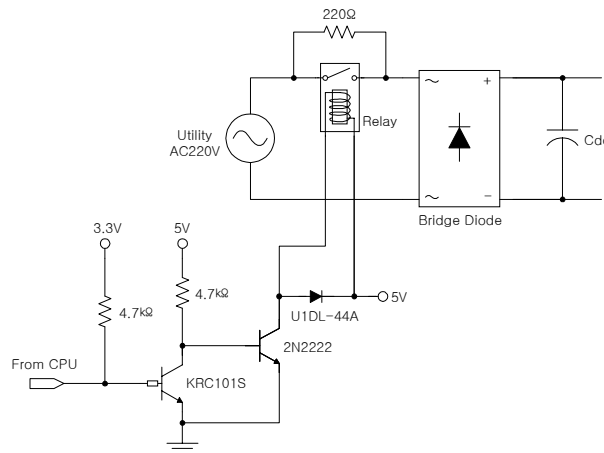


Fig. 2.4 Initial charging circuit

Depending on the bridge diode rating and the electrolytic capacitor maximum ripple current, the relay is driven after 800ms of charging duration from power on. The series resistor connected in parallel with the relay should be able to endure the power loss during the charging period. The simulation waveform is shown below.

Simulation condition

Load : 1Hp(750W)
Input : AC220V, single phase
DC-link capacitors: $470\mu\text{F}/250\text{WV} \times 2\text{EA}$ connected in series
Charging resistor: $220\Omega/3\text{W}$

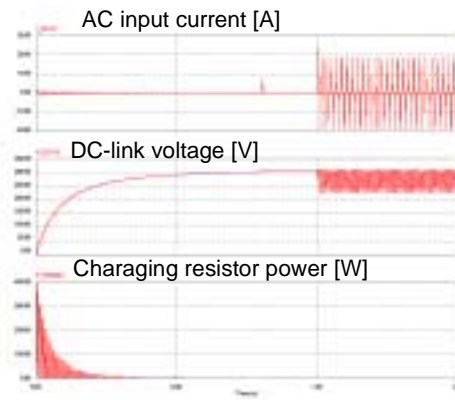


Fig. 2.5 Initial charging simulation

2.4 Dynamic Brake (DB) Circuit

In adjustable-speed motor drives, the machines may be subjected to electrical braking for reduction of speed. In electrical braking, the motor is operated in the generating mode and the kinetic energy stored in the system inertia is converted to electrical energy. In the designed system a resistor is used to dissipate the energy, which is well known as dynamic braking resistor. Dynamic braking power depends on the DC-link voltage and resistor value. The DB resistor used is a $200\Omega/200W$ metal resistor. An IGBT with a rating of $2A/600V$ is used for the DB. Fig. 2.6 shows the detailed DB circuit.

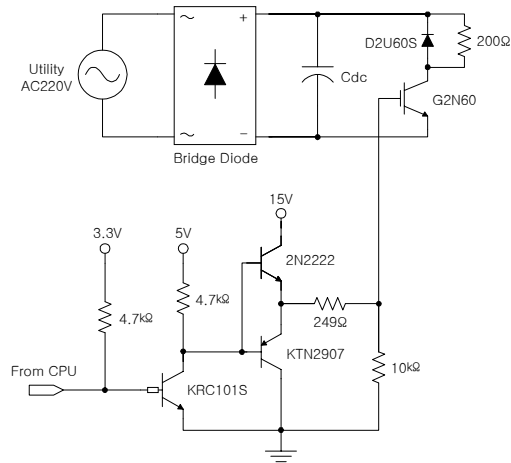


Fig. 2.6 Dynamic braking circuit of the designed system

2.5 SMPS (Switching Mode Power Supply)

The Flyback type of SMPS is designed for the isolated dc-to-dc converter. The detailed schematic is shown in Fig. 2.7. It is designed to generate two values of 15V and 5V for control supply, and their current ratings are 0.3A and 0.7A, respectively.

For the SMPS control, the adopted main power switch is KA5H0280R made by Fairchild Semiconductor, and its major specifications are shown in Table 2.1.

Table 2.1 Specification of KA5H0280R

Vdmax [V]	Ipeak [A]	Pin (max) [W]	Fopr [kHz]	Rds (on) max [Ω]	PKG
800	1.2	25	100/70	7.0(I _d =1.0A)	TO-220F-4L

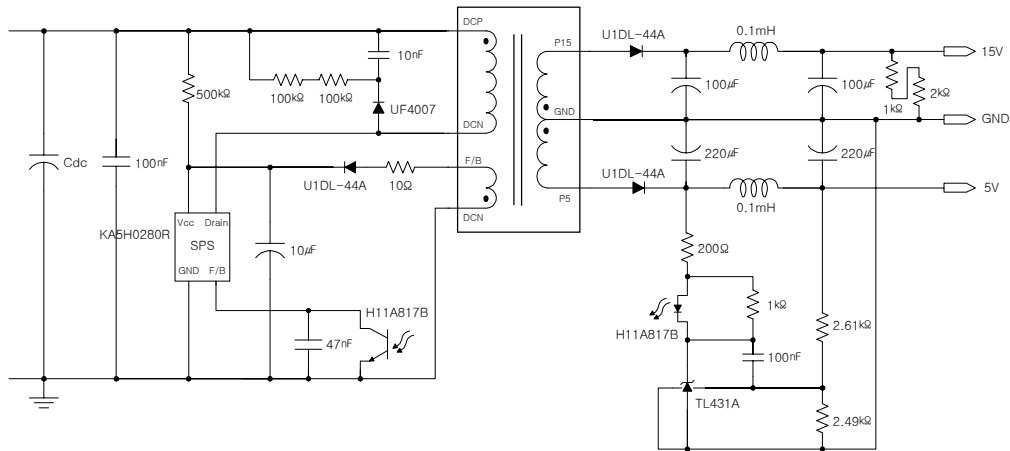


Fig. 2.7 SMPS circuit of the designed system

Table 2.2 shows the winding specification for the SMPS transformer. The turn ratio of the primary winding to the secondary one for 5V supply is 120:7. 0.4 copper wire is used for 5V winding and 0.2 copper wire is used for the others. There are 3 turns of 0.050mm thick polyester taping to ensure that there is insulation between each winding.

Table 2.2 Winding specifications

No.	Pin (Start → Finish)	Wire	Turns	Winding Method
N_{p1}	1 → 2	0.20φ × 1	60	Space Winding
Insulation : Polyester Tape t=0.050mm, 3Layer				
N_{5V}	6 → 7	0.40φ × 1	7	Space Winding
Insulation : Polyester Tape t=0.050mm, 3Layers				
N_{15V}	6 → 8	0.20φ × 1	17	Space Winding
Insulation : Polyester Tape t=0.050mm, 3Layers				
N_{FB}	5 → 4	0.20φ × 1	20	Space Winding
Insulation : Polyester Tape t=0.050mm, 3Layers				
N_{P2}	2 → 3	0.20φ × 1	60	Space Winding
Insulation : Polyester Tape t=0.050mm, 3Layers				

Note) NP: Primary winding NFB: Flyback winding
 N5V: 5V output winding N15V: 15V output winding

Table 2.3 Electrical characteristics of the transformer

	Pin	Spec.	Remarks
Inductance	1 – 3	1.3mH ± 5%	100kHz, 1V
Leakage	1 – 3	120μH (max.)	Secondary Short

The winding order shown in Fig. 2.8 is applied considering the coupling coefficient.

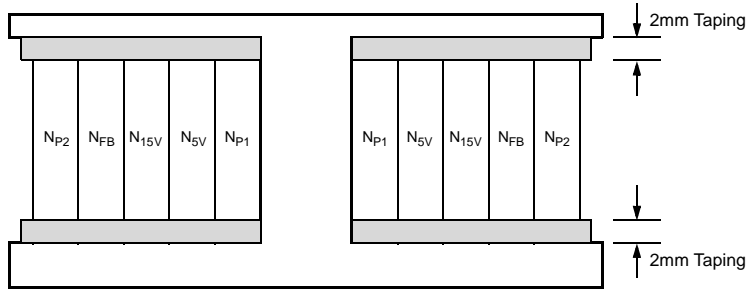


Fig. 2.8 The winding order (EI2218 core and bobbin are used)

3. SPM Product Guide

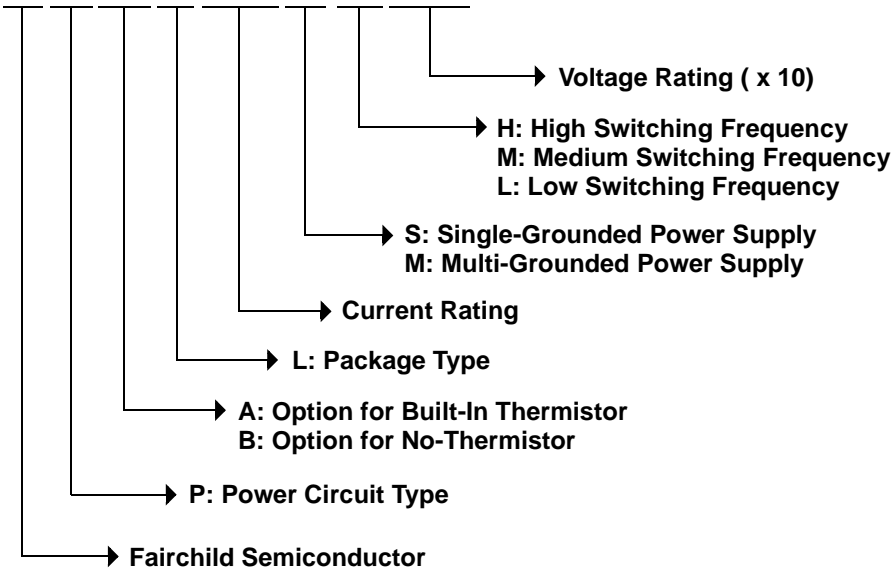
3.1 Product Line-Up

Part Number	Rating	Switching Frequency	Primary Application
FPAL10SH60	10A / 600V	High Speed	Washing Machines ($\geq 8\text{kHz}$)
FPBL10SH60	10A / 600V	High Speed	
FPAL15SH60	15A / 600V	High Speed	
FPBL15SH60	15A / 600V	High Speed	
*FPAL15SM50	15A / 500V	Medium Speed	Air Conditioners (3kHz ~ 9kHz)
*FPBL15SM50	15A / 500V	Medium Speed	
FPAL15SM60	15A / 600V	Medium Speed	
FPBL15SM60	15A / 600V	Medium Speed	
*FPAL20SM50	20A / 500V	Medium Speed	
*FPBL20SM50	20A / 500V	Medium Speed	
FPAL20SM60	20A / 600V	Medium Speed	
FPBL20SM60	20A / 600V	Medium Speed	
FPAL15SL60	15A / 600V	Low Speed	Air Conditioners ($\leq 4\text{kHz}$)
FPBL15SL60	15A / 600V	Low Speed	
FPAL20SL60	20A / 600V	Low Speed	
FPBL20SL60	20A / 600V	Low Speed	
FPAL30SL60	30A / 600V	Low Speed	
FPBL30SL60	30A / 600V	Low Speed	

* under development

3.2 Ordering Information

FPAL15SH60



4. Induction Motor Drive (V/Hz)

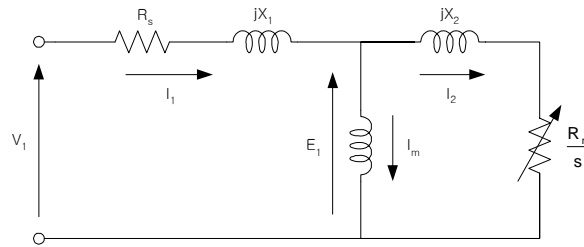


Fig. 4.1 Single-phase equivalent circuit of a polyphase induction motor

The fundamental steady-state equivalent circuit of the induction motor is shown in Fig. 4.1. The rotating airgap flux induces an emf (electromotive force), E_1 , in the stator winding. This emf has the value of applied voltage V_1 minus the voltage drop due to the stator leakage impedance, $I_1(R_S + jX_1)$. If the rotating flux wave has a sinusoidal space distribution, and the flux linking each stator turn has a sinusoidal time variation, the instantaneous flux linking a full-span stator turn is

$$\phi = \Phi_1 \sin \omega_1 t$$

where, ω_1 , is $2\pi f$ the angular frequency of the supply voltage,
 Φ_1 is the fundamental flux per pole

The induced emf per turn is, therefore,

$$e_1 = \frac{d\phi}{dt} = \omega_1 \Phi_1 \cos \omega_1 t$$

and the rms stator emf is given by

$$E_1 = \frac{\omega_1 \Phi_1 k_w N_1}{\sqrt{2}} = 4.44 k_w f_1 n_1 \Phi_1$$

where, N_1 is the number of series turns per phase,
 k_w is the winding factor.

If the winding factor is unity, the usual transformer emf equation is obtained; hence, for a motor or transformer, Φ_1 is proportional to E_1/ω_1 or E_1/f_1 . When the ratio is constant, a constant airgap flux is obtained. If the voltage drop across the stator leakage impedance is small, V_1 and E_1 have almost the same value. Consequently, the airgap flux is nearly constant when the ratio V_1/f_1 has a fixed value. This is the constant terminal V/Hz mode that is commonly used in simple open-loop control systems. The inverter that incorporates the voltage and frequency control scheme provides the linear output voltage-frequency characteristics.

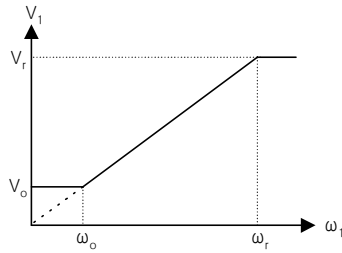


Fig. 4.2 Approximate Voltage/Frequency characteristics

Around the rated frequency it is valid that the stator voltage drop, $I_1(R_S + jX_1)$ is negligible. But the stator voltage drop developed by the rated current stays constant even though the output frequency is reduced. This drop occupies a large portion of the terminal voltage. The airgap emf and flux decreases significantly, causing torque reduction. This problem can be resolved by boosting the voltage above the V/Hz ratio under low frequency. There are two commonly used methods. One is to impose a lower limit at frequency. And the other is to impose a lower limit at voltage command without frequency limit. The V/Hz profile should also be modified in the upper rated frequency area. In this area, the V/Hz ratio-corresponding voltage cannot be imposed in order to avoid insulation breakdown. The stator voltage must be maintained under the rated voltage. The V/Hz profile is shown in Fig. 4.2 under the above two conditions.

5. BLDC Motor Drive

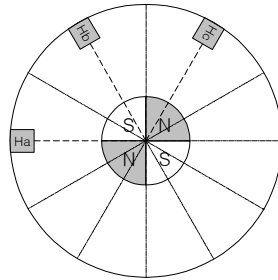


Fig. 5.1 Structure of the 4 pole trapezoidal BLDCM used in this note

A 4 pole magnet trapezoidal BLDCM, that has 60 degree spatially distributed hall sensors, is used. Each hall sensor generates 2 pulses per revolution. The switch pair is determined at each 30 mechanical degree. The trapezoidal BLDC motor generates trapezoidal back emf at each phase depending on the rotor position. The motor torque is in proportion to the product of phase back emf and phase current. Therefore, constant torque control is possible by applying constant current at constant back emf duration. When the current path is changed, the phase current needs a little time to establish itself. This allows the current to not be in a flat shape. The generated output torque contains a ripple component as shown in Fig. 5.3.

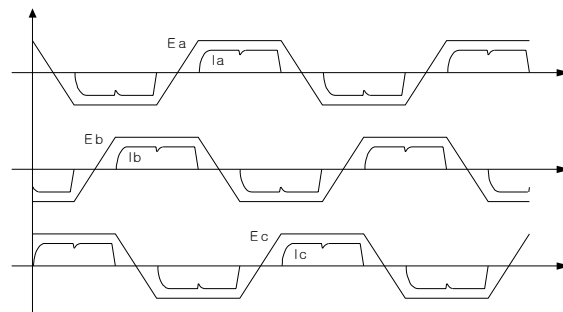


Fig. 5.2 Back emf and current of each phase

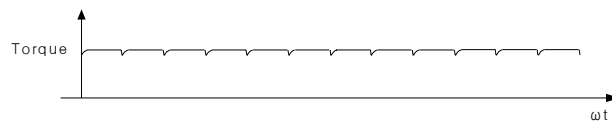


Fig. 5.3 Produced torque

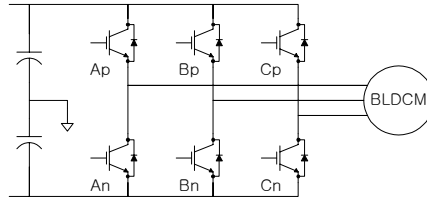
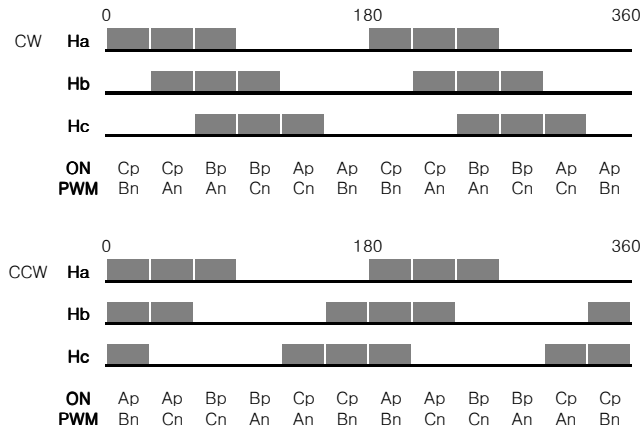


Fig. 5.4 Used voltage-controlled inverter topology

In a BLDCM drive, it is necessary to let each stator current flow in a certain pattern depending on the rotor position. Fig. 5.4 shows the structure of the inverter used in this application. In this application, the mechanical rotor position is detected by three hall sensors. The phase current is changed at each 30 mechanical degree.

The hall sensor signal and the corresponding switching pattern of a 3-phase 2 excitation PWM is shown below. A pair of switches is selected at each 30 mechanical degree duration. In order to control the stator rms voltage, the low side switch of the selected pair performs PWM. In order to avoid the localization of device stress, it is also used to perform PWM with the newly selected switch of the activated switch pair.



CW From Back End

Hall Sensor Output			Switch On	Phase Current		
A	B	C		A	B	C
0	0	1	Ap Cn	+	Off	-
0	0	0	AP Bn	+	-	Off
1	0	0	Cp Bn	Off	-	+
1	1	0	CP An	-	Off	+
1	1	1	Bp An	-	+	Off
0	1	1	Bp Cn	Off	+	-

CCW From Back End

Hall Sensor Output			Switch On	Phase Current		
A	B	C		A	B	C
0	1	1	Cp Bn	Off	-	+
1	1	1	AP Bn	+	-	Off
1	1	0	Ap Cn	+	Off	-
1	0	0	Bp Cn	Off	+	-
0	0	0	Bp An	-	+	Off
0	0	1	Cp An	-	Off	+

Fig. 5.5 Hall sensor signal and corresponding switch pair

6. Space Vector Pulse Width Modulation (SVPWM)

In this note, a symmetric SVPWM using the offset voltage injection method is used. This lowers the software load, but maximizes the utilization of the DC link voltage. Its brief theory follows.

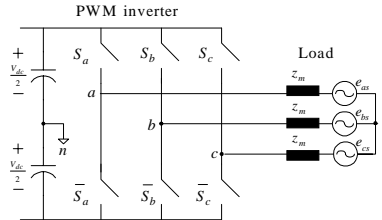


Fig. 6.1 Three-phase voltage source inverter

Fig. 6.1 shows the basic structure of a 3phase Voltage-controlled inverter. Pole voltage, phase voltage and offset voltage relationships are as follows.

$$V_{cn} = V_{cs} + V_{sn} \quad (6.1)$$

$$V_{an} = V_{as} + V_{sn} \quad (6.2)$$

$$V_{bn} = V_{bs} + V_{sn} \quad (6.3)$$

An arbitrary value can be assigned to offset voltage V_{sn} . Various voltage modulation methods can be realized by this value. This means that the DC component injected into the pole voltage does not appear in the output phase voltage because the 3-phase output voltage sums to zero. In other words, the offset voltage is the hidden 'degree of freedom' of 3-phase voltage modulation. All the following equations of the pole voltage effective range should be satisfied.

$$-\frac{V_{dc}}{2} \leq V_{an}^* \leq \frac{V_{dc}}{2} \quad (6.4)$$

$$-\frac{V_{dc}}{2} \leq V_{bn}^* \leq \frac{V_{dc}}{2} \quad (6.5)$$

$$-\frac{V_{dc}}{2} \leq V_{cn}^* \leq \frac{V_{dc}}{2} \quad (6.6)$$

If the maximum, medium and minimum value of the phase voltages reference are defined as follows

$$V_{max}^* = \max(V_{as}^*, V_{bs}^*, V_{cs}^*) \quad (6.7)$$

$$V_{mid}^* = \text{mid}(V_{as}^*, V_{bs}^*, V_{cs}^*) \quad (6.8)$$

$$V_{min}^* = \min(V_{as}^*, V_{bs}^*, V_{cs}^*) \quad (6.9)$$

The range of offset voltage can be obtained as follows.

$$-\frac{V_{dc}}{2} - V_{min}^* \leq V_{sn} \leq \frac{V_{dc}}{2} - V_{max}^* \quad (6.10)$$

Now the pole voltage reference (V_{an}^* , V_{bn}^* , V_{cn}^*) can be determined from the phase voltage reference (V_{as}^* , V_{bs}^* , V_{cs}^*) by selecting the appropriate offset voltage. Let us obtain the maximum modulation index (MI) of this method. As phase voltage reference has periodic characteristics, it is sufficient to consider only one case: voltage reference is in sector 1. In this case,

$$V_{\max}^* = V_{as}^* \quad (6.11)$$

$$V_{\min}^* = V_{cs}^* \quad (6.12)$$

Let us express the phase voltage reference with the voltage modulation index.

$$V_{as}^* = MI \cdot (V_{dc}/2) \cdot \cos(\theta) \quad (6.13)$$

$$V_{cs}^* = MI \cdot (V_{dc}/2) \cdot \cos(\theta + 2\pi/3) \quad (6.14)$$

$$\theta \in (0, \pi/3)$$

From the equation (4.10), the following equation should be satisfied in order to select the effective

$$V_{\max}^* - V_{\min}^* \leq V_{dc} \quad (6.15)$$

$$V_{\max}^* - V_{\min}^* > V_{dc} \quad (6.16)$$

offset voltage. The maximum voltage modulation index obtained from the equation (6.15), (6.13) and (6.14) is $2/\sqrt{3}$, and is equal to that of the existing space vector voltage modulation method. If the phase voltage reference is so large that the following equation is satisfied, then the appropriate offset voltage that matches the equation (6.10) cannot be selected. This occurs in overmodulation mode. This means that the reference voltage vector is outside of the hexagon. In this case, the pole voltage should be reassigned through the proper overmodulation technique.

Now we should determine switching time using pole voltage. Pole voltage is basically determined by the switching state. It is much easier to determine the switching time from pole voltage than from other voltages. Fig. 6.2 shows the relationship between pole voltage and gating time. Gating time is linear with

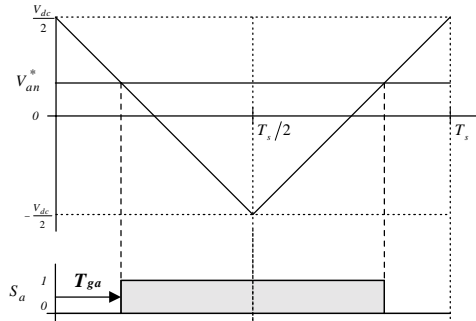


Fig. 6.2 Relationship between pole voltage and gating time

pole voltage, and is shown in the equation (6.17).

$$T_{ga} = \frac{T_s}{4} - \frac{V_{an}^* T_s}{V_{dc} 2} \quad (6.17)$$

After determining the offset voltage, the gating time can be calculated easily by substituting the pole voltage that is obtained by adding the offset voltage to the phase voltage. Various existing voltage modulation methods can be determined easily by using the offset voltage expressed as V_{\max}^* , V_{mid}^* , V_{\min}^* . The principle of symmetric space vector voltage modulation is to locate the effective voltage vector ($V_{\bar{n}}, V_{\bar{n}+1}$) in the center of one modulation cycle. As shown in Fig. 6.2, this occurs when the absolute value of the maximum and minimum pole voltage are the same. Hence, the existing symmetric space vector voltage modulation can be realized by setting the offset voltage as a certain value that makes the maximum and minimum pole voltage have the same value.

$$V_{\max}^* + V_{\text{sn}} = -(V_{\min}^* + V_{\text{sn}}) \quad (6.18)$$

So now, the offset voltage is as follows.

$$V_{\text{sn}} = -\frac{V_{\max}^* + V_{\min}^*}{2} \quad (6.19)$$

Fig. 6.3 shows the pole voltages that are the sum of phase voltage and offset voltage as shown in the equation (6.19).

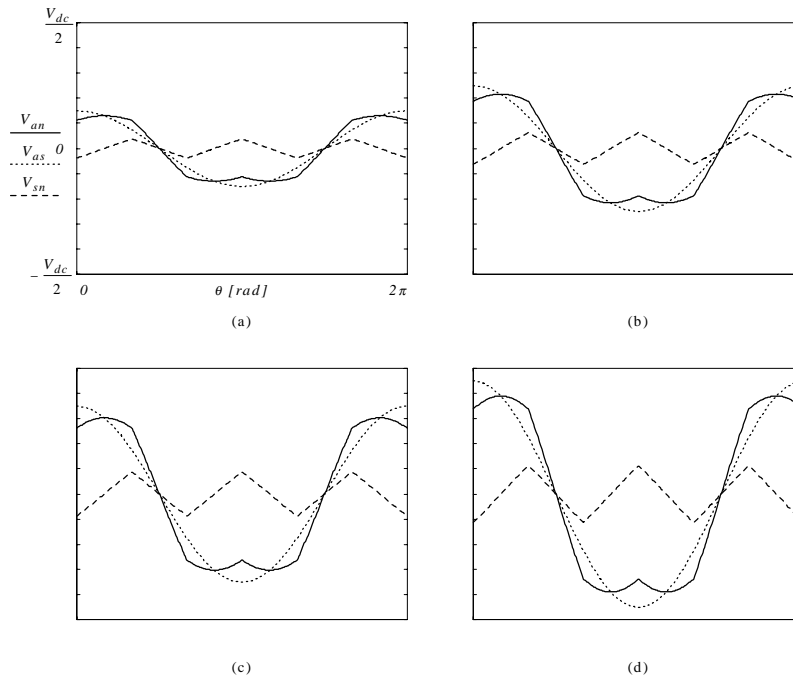


Fig. 6.3 Pole voltage of SVPWM.h
(a) MI = 0.3 (b) MI = 0.5 (c) MI = 0.7 (d) MI = 0.9

7. Real Implementation

An open-loop speed control system for a 3-phase AC induction motor using the V/Hz principle, and a speed controlled drive system for a 4 pole 3-phase BLDC motor is implemented using a 3-phase voltage-controlled inverter. The outline of the control algorithm for each system is described below.

IM

- Obtain target frequency from ADC input.
- Calculate output frequency.
- Obtain output voltage magnitude.
- Obtain output voltage phase, θ .
- Calculate $\sin\theta$, $\cos\theta$.
- Perform 2 phase/3 phase transformation
- Load compare register with phase voltage corresponding timer value.

BLDC

- Obtain speed reference from ADC input.
- Calculate current speed from hall sensor signal.
- Obtain reference voltage from speed error PI.
- Load compare register with phase voltage corresponding timer value.

7.1 Block diagram

IM

Determine the target frequency from volume resistor voltage. Calculate the output frequency after considering the accelerate/decelerate time. Obtain the target output voltage magnitude from the V/Hz profile. Obtain \sin , $\cos\theta$ from the software table 2 phase/3 phase conversion. Load the compare register with the calculated phase voltage-corresponding timer value.

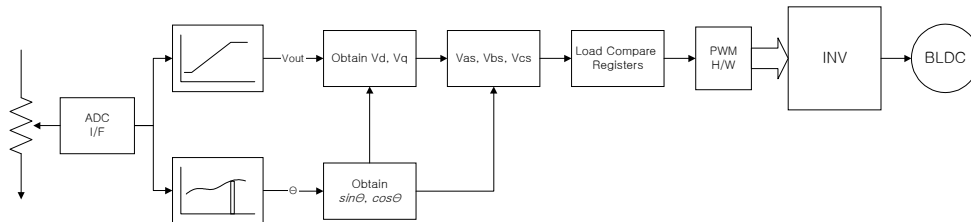


Fig. 7.1 Block diagram of V/Hz control loop for induction motor

BLDC

Determine the speed reference from the volume resistor voltage. Calculate the rotor speed from the hall sensor signal. Obtain the reference voltage by applying PI control to the speed error. Load the compare register with the calculated phase voltage-corresponding timer value.

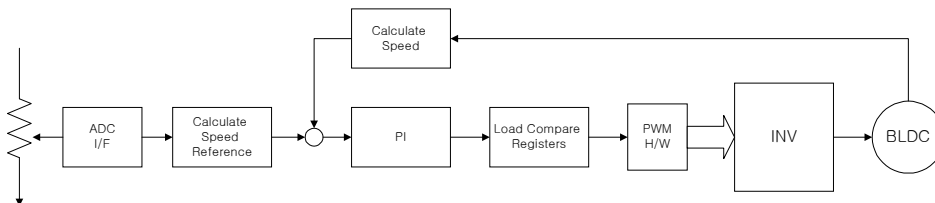


Fig. 7.2 Block diagram of speed control loop for BLDC motor

7.2 Flow Chart

The following figure is the flow chart of the two main interrupt routines - 1ms interrupt routine and the PWM interrupt routine. The software includes routines for both IM and BLDCM.

For IM, obtain the output frequency and the corresponding output voltage magnitude in the 1ms interrupt routine. Calculate the current output voltage phase in the PWM interrupt routine. First, the target frequency is obtained in the 1ms routine. If the current operating mode is manual, then read the variable resistor voltage and calculate the target voltage. In the auto mode, read the preset table. Then determine the output frequency from the current frequency and target frequency considering the acceleration/deceleration time. Calculate the output phase transition amount() from the output frequency and obtain the output voltage magnitude from the V/Hz profile. Finally, drive the DB (Dynamic Brake) IGBT according to the DC-link voltage. In the PWM interrupt routine, obtain of the current output voltage vector from, and calculate \sin , $\cos\theta$. Make the axis conversion, obtain each phase voltage, and then load the compare register with the value calculated from the PWM algorithm.

For the BLDCM, in the 1ms interrupt routine, obtain the output reference voltage by applying PI control to the rotor speed error. In the PWM interrupt routine, determine the switching pattern according to the hall sensor signal. Then calculate the switching time and load the timer register in order to PWM the low side IGBT.

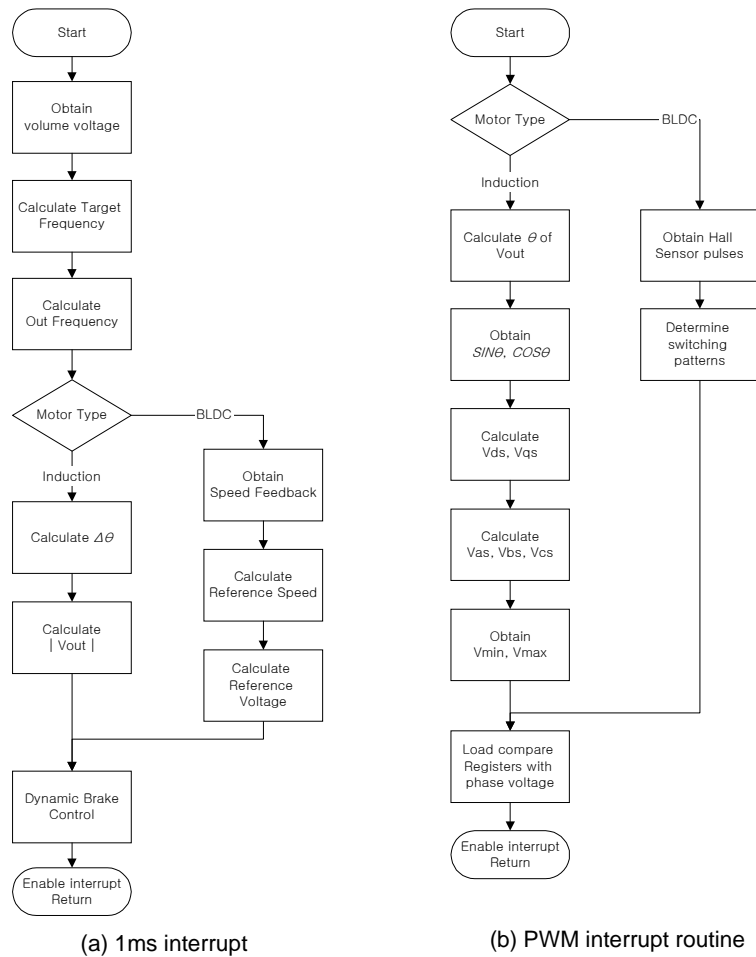


Fig. 7.3 The Flow Chart

7.3 The Extract of the source code

The main structure of the implemented software is shown below.
A DSP of type TMS320F2406 is used.

```
/* =====*
 *           Main Module                               *
 *   Main program initialization and Main loop execution *
 * =====*/

void main()
WORDi, j;
  mainInit();    /* H/W Initialization */
  InitVar();     /* S/W Initialization */
  InitVFD();    /* VFD(Display) Initialization */
  EI;

  while (1)
  {
    Keypad();
    VariUpdate();
  }
}

void mainInit(void)
{
  iolnit();
  Tmr1mInit();
  adc_init();
  da_init();
  SPIInit();
}

void adclnt(void)
{
  ADCCTRL2 |= BIT_14;    /* reset SEQ1 */
  EVAIFRA = BIT_1;      /* eva1int flag reset */
  ADCCTRL2 |= BIT_9;    /* clear INT_FLAG_SEQ1 */
}
```

```

/* =====*
*                               PWM_INT Module                               *
* =====*/

int Fwd[8] = {                  /* BLDC switching pattern */
    4028,                       /* ACTRA = 0000 1111 1011 1100 b */
    3263,                       /* ACTRA = 0000 1100 1011 1111 b */
    4095,                       /* ACTRA = 0000 1111 1111 1111 b */
    3323,                       /* ACTRA = 0000 1100 1111 1011 b */
    3068,                       /* ACTRA = 0000 1011 1111 1100 b */
    4095,                       /* ACTRA = 0000 1111 1111 1111 b */
    3023,                       /* ACTRA = 0000 1011 1100 1111 b */
    4043                        /* ACTRA = 0000 1111 1100 1011 b */
};

int Rev[8] = {
    4043,                       /* ACTRA = 0000 1111 1100 1011 b */
    3023,                       /* ACTRA = 0000 1011 1100 1111 b */
    4095,                       /* ACTRA = 0000 1111 1111 1111 b */
    3068,                       /* ACTRA = 0000 1011 1111 1100 b */
    3323,                       /* ACTRA = 0000 1100 1111 1011 b */
    4095,                       /* ACTRA = 0000 1111 1111 1111 b */
    3263,                       /* ACTRA = 0000 1100 1011 1111 b */
    4028                        /* ACTRA = 0000 1111 1011 1100 b */
};

void pwmInt(void)
{
    if(wMotorType==0){          /* for IM */

        if(Dir)                 /* CCW */
            lAngle += IDTheta;  /* Obtain Angle */
        else                    /* CW */
            lAngle -= IDTheta;

        CALC_SIN_COS();        /* Obtain Sin(theta), Cos(theta) */
        iVqeRef = wOutV;
        iVdeRef = 0;

        DQE_DQS(iVdsRef, iVqsRef, iVdeRef, iVqeRef, iSin, iCos);
        /*=====
                               Sync -> Stat
                               =====
                               ds = cos * de - sin * qe
                               qs = sin * de + cos * qe
                               -----*/

        DQS_ABC(iVasRef, iVbsRef, iVcsRef, iVdsRef, iVqsRef);
        /*=====
                               2 -> 3
                               =====
                               as = ds
                               bs = -1/2 * ds + sqrt(3)/2 * qs
                               cs = -as - bs
                               -----*/
    }
}

```

```

Max_MinABC(Max,Min,iVasRef,iVbsRef,iVcsRef);
Offset = (Max+Min)>>1;

iVasRef -= Offset;
iVbsRef -= Offset;
iVcsRef -= Offset;

wPwmU = (wPwmAmp) + MulDiv((wPwmAmp),iVasRef,wBaseV);
wPwmV = (wPwmAmp) + MulDiv((wPwmAmp),iVbsRef,wBaseV);
wPwmW = (wPwmAmp) + MulDiv((wPwmAmp),iVcsRef,wBaseV);

CMPR1 = wPwmU;          /* Load compare register */
CMPR2 = wPwmV;
CMPR3 = wPwmW;
}

else{                    /* for BLDCM */
Ha = (PADATDIR>>3 & 1);
Hb = (PADATDIR>>4 & 1);
Hc = (PADATDIR>>5 & 1);

Read_Speed(delT,Ha,Hb,Hc); /* for Speed Read */
if(!flag.Run){
BuildTimer = 0;
ACTRA = 0x0fff;          /* All switch : forced high */
}
else{
if(Dir)                  /* ----- */
ACTRA = Rev[((PADATDIR>>3)&7]; /* | IOPA5 | IOPA4 | IOPA3 | */
Else                      /* |-----| */
ACTRA = Fwd[((PADATDIR>>3)&7]; /* | Hc | Hb | Ha | */
/* ----- */

wPwmU = wVOut4BLDC;

CMPR1 = wPwmU;          /* Load compare register */
CMPR2 = wPwmU;
CMPR3 = wPwmU;
}
}
EVAIFRA = BIT_9;      /* T1UFINT Flag Reset */
}

```

```

/* =====*
*                               Tmr1m Module                               *
* =====*/

void Tmr1m(void)
{
    tic++;
    VolIn = MulDiv(VOL_IN,10000,65000);    /* Scale the potentiometer */
    if(flag.CmdRun) {
        wTarFreq = MulDiv(VolIn,wBaseFreq,10000);
        if(DirCmd!=Dir){
            wTarFreq=0;
            if(wMotorType){                /* BLDC */
                if(wSpeedFb<100) Dir = DirCmd;
            }
            else{                            /* IM */
                if(wCurFreq<wStartFreq) Dir = DirCmd;
            }
        }
    }
    else wTarFreq = 0;
    AccDecCalc();                          /* Calculate wOutFreq */
    IDTheta = wOutFreq * IDThetaScale;     /* Obtain delta theta */
    if(wOutFreq>wBaseFreq){
        wOutV = wBaseV;                    /* Limit Max Voltage */
    }
    else{                                    /* Calculate wOutV with respect to V/Hz profile */
        VHz_profile(wOutV,wOutFreq,wBaseV,wBaseFreq,wTrqBoost);
    }
    /* 4 BLDC */
    wSpeedFb = MulDiv(750,wCarFreq,deIT);
    wRefSpeed = MulDiv(wCurFreq,2500,wBaseFreq);
    wRefV4BLDC = pi(wRefSpeed,wSpeedFb,&ISpdPIITerm,Ki,Kp,0,MaxOutV4BLDC,15);
    wVOut4BLDC = MulDiv(wRefV4BLDC,(wPwmAmp<<1),wDCVolt);

    if(wDCVolt>380) DB_ON;
    if(wDCVolt<360) DB_OFF;

    EVBIFRA = BIT_9;                       /* T3UFINT Flag Reset */
}

```

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