ABOUT COMMUTATION AND CURRENT CONTROL METHODS FOR BRUSHLESS MOTORS

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ABSTRACT: *For brushless motor drives, current control is often used to improve performance and reliability. This tutorial paper analyzes and compares various different schemes of electronic commutation and current control. Starting from a simple 6-step drive without current control, discussion will include bus current control, sinusoidal commutation, phase current control, and synchronous regulator. Relative advantages of various commutation techniques and current control schemes with respect to dynamic performance and steady-state torque output are discussed. Practically used current sensors, topologies, circuit design and digital control implementation are also presented.*

I. Introduction

Due to the increasing demand for compact and reliable motors and the evolution of low-cost power semiconductor switches and permanent magnet (PM) materials, brushless motors became popular in every application from home appliance to aerospace industry. Unlike brushed DC motors, every brushless motor requires a "drive" to supply commutated current to the motor windings synchronized to the rotor position. In other words, some kind of feedback position sensors are necessary to commutate brushless motors. Some drives are just commutating while others may include voltage control with or without current-loop. Fig. 1.1 shows a block diagram of a typical brushless motor drive system.

In this article, various different schemes of electronic commutation and current control will be discussed. Starting from a simple 6-step drive without current control, discussion will include bus current control, sinusoidal commutation, phase current control, and synchronous regulator. Relative advantages of various commutation techniques and current control schemes with respect to dynamic performance and steady-state torque output are discussed.

<u>Unipolar vs. bipolar drive</u>

Instead of popular 6-switch (bipolar drive) configuration, only 3 switches (unipolar drive) can be used to drive 3 phase brushless motors. Since direction of winding current cannot be reversed, windings are not efficiently utilized in this configuration. Use of unipolar drives are restricted in low-power, low voltage motors due to lower drive costs and lower voltage drop across semiconductor switches. The remaining discussion will be focused on bipolar drives.

<u>Constant or variable bus voltage operation.</u>

Similar to brushed DC motors, brushless DC motors with electronic commutation can operate from a DC voltage source and speed can be adjusted by varying supply voltages.

Variable voltage DC source may be obtained from rectification of a variable voltage transformer output or from the thyrister bridge as shown in Fig. 1.2. In this configuration, DC bus voltage is a function of the firing angle of the thyrister bridge. Instead of varying supply voltage, variable speed operation can also be realized from a constant bus voltage.

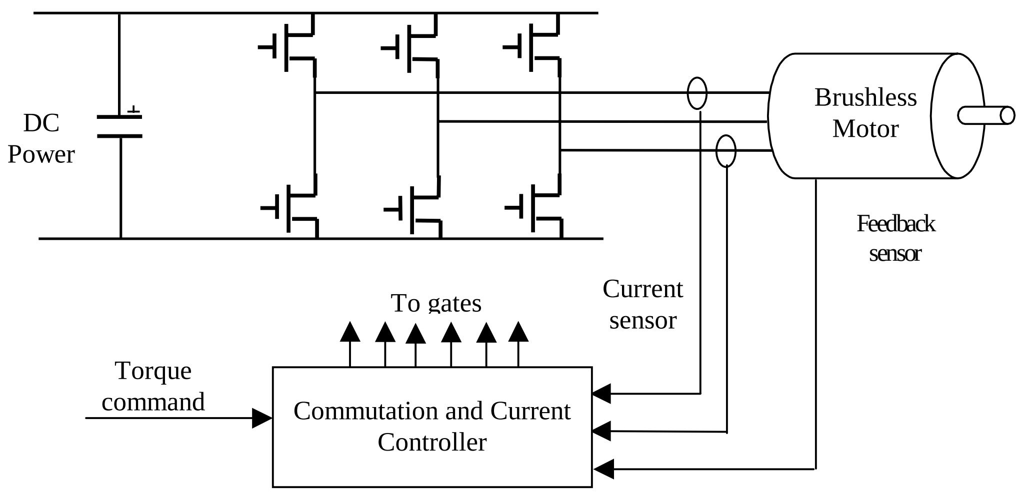


Fig. 1.1. Block Diagram of a Brushless Motor Drive System

In this case, power semiconductor switches of Fig. 1.1 are used not only to commutate but also to control the motor terminal (drive output) voltage via the PWM (pulse width modulation) technique. The technique generates a fixed frequency (usually 2 kHz - 30 kHz) voltage pulse whose on-time duration is controlled. Since the brushless motor is highly inductive, the motor current produced from this switched voltage would be almost identical to that from the fixed voltage whose magnitude is the average of the switched voltage waveform. Although PWM control is now very popular in drives, variable bus voltage control is still used in some applications where dynamic performance is not important.

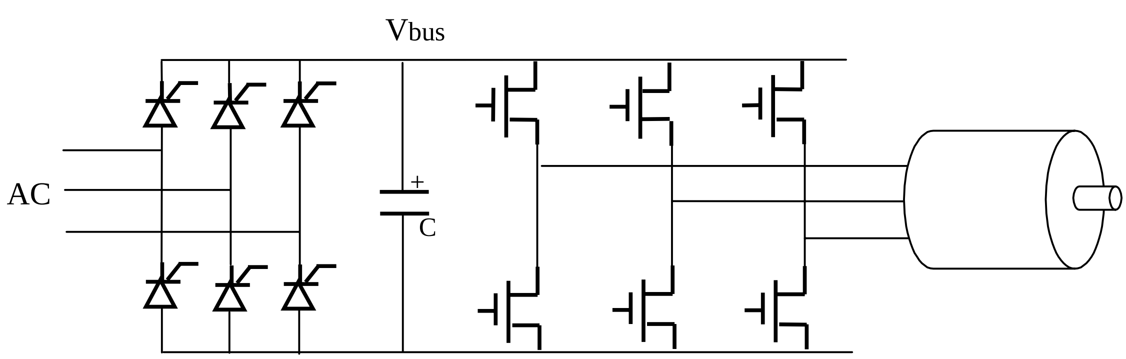


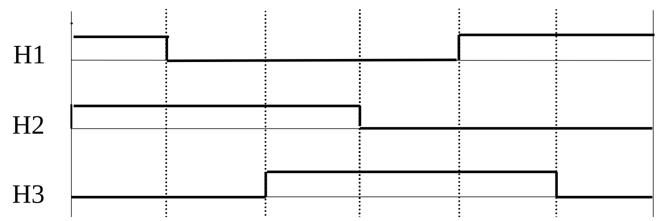
Fig. 1.2 Variable Bus Voltage Operation using Thyristers

Open-loop vs. closed-loop drive.

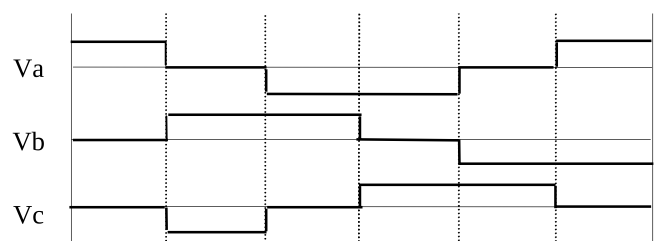
Since motor speed is roughly proportional to the terminal voltage of the motor, variable speed operation is possible by changing the terminal voltage via controlling duty cycle of the PWM. In order to control speed accurately, a closed-loop control may be used where commanded speed is compared with the measured actual speed and the motor is driven by the speed error signal. Without closed-loop control (open-loop drive), motor speed may vary depending on the load, supply voltage, etc.

II. 6-Step Commutation and Current Control Methods

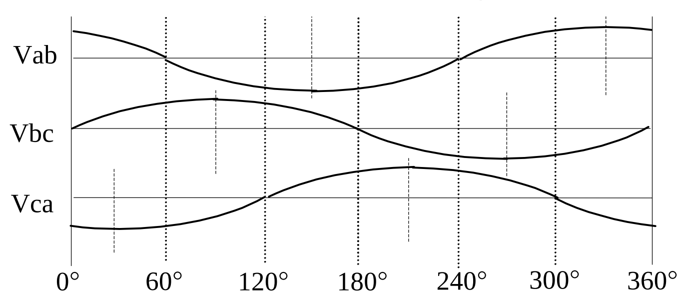
One of the simplest methods of commutating 3 phase brushless motor, commonly known as the "6-step drive" will be discussed in this section. In this method, each phase voltage is energized for 120 deg. (electrical) interval according to its rotor electrical angle as shown in Fig 2.1(b). This may be realized by the switch configuration of Fig. 2.2. Each phase voltage is positive (negative) when the top switch is on (off) and the bottom switch is off (on). No voltage is injected when both switches are off, in which case the actual terminal voltage is governed by the back emf voltage of the motor. In other words, each phase voltage at a time takes one of three states - positive, negative, or float. At every sector, only one phase is energized as positive and one of the other phases is energized as negative in order to maintain current path. In order to commutate properly, the controller needs to know the sector (60 degree interval) position of the shaft angle.



(a) Hall sensor outputs



(b) 6-step phase voltages



(c) Back emf voltage waveform

Fig. 2.1. 120° Conduction Commutation

Three Hall sensor outputs, as shown in Fig. 2.1(a), are often used to detect the shaft position. One advantage of this relation is that during the Hall sensor alignment procedure, we can align the Hall sensor board so that each Hall output is in phase with the corresponding back emf (line-to-line) waveform when the rotor is rotated by an external means.

When the output voltage should also be controlled in addition to commutation, pwm (pulse width modulation) control may be applied to both sides or lower side switches only. When pwm is applied only to the low side switches (Q2, Q4 and Q6), a short-circuited current path is established through one of the free-wheeling diodes of upper switches during pwm-off time. For instance, assume that switches Q1 and Q4 are turned on for commutation and the current is established through A and B winding as shown by the arrows in Fig. 2.2. Since Q4 is doing pwm, current path is still established through {Q1-A winding-C winding-D3} path while Q4 is briefly off. Idealized current waveform at this time is depicted in Fig. 2.3(a). Since this configuration cannot control the current during fast torque reversal, it is only used when the directions of torque and speed are always identical (1-quadrant or 2-quadrant operation).

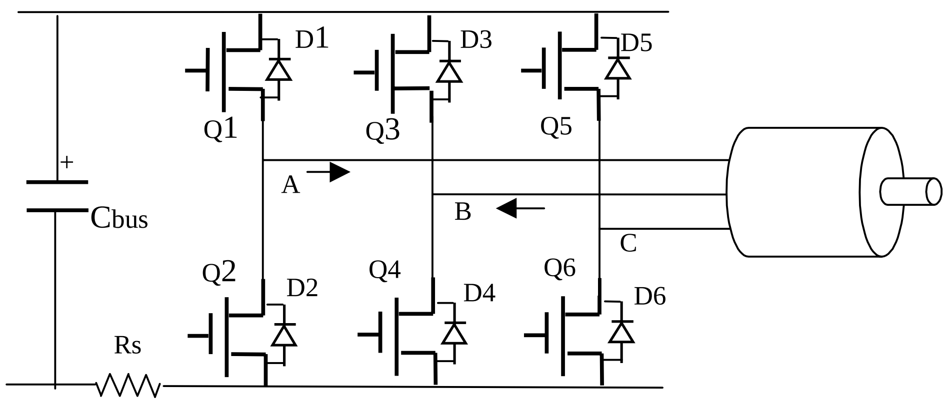


Fig. 2.2. Switch Configuration for a Burshless Motor

When the pwm is applied on both upper and lower side switches, current decay during pwm off times are quicker. In the previous example, since both Q1 and Q4 are doing pwm, current path is though {D2-A winding-C winding-D3} path while Q1 and Q4 are briefly off. This condition is like a reverse bus voltage is applied to the winding and the current decays at a fast rate. A full 4-quadrant operation is possible at this configuration where the directions of torque and speed are arbitrary. This configuration is preferred when fast speed reversal is required as in servo application. Current waveform in this 4-quandrant pwm is shown in Fig. 2.3(b). Note that there are additional penalties in applying the 4-quandrant switching, in addition to higher pwm switching loss. First, the fast decaying current must flow through the bus capacitance, one should select the capacitor with higher capacitance and higher ripple current rating. Second, during fast torque reversal condition, there might be a chance of shoot-through - a short-circuited condition when both upper and lower switches are turned on during switching transition. Therefore, turn-on dead-time of several microseconds have to be applied for safe operation. In 2-quadrant switching, insertion of dead-time is normally not necessary.

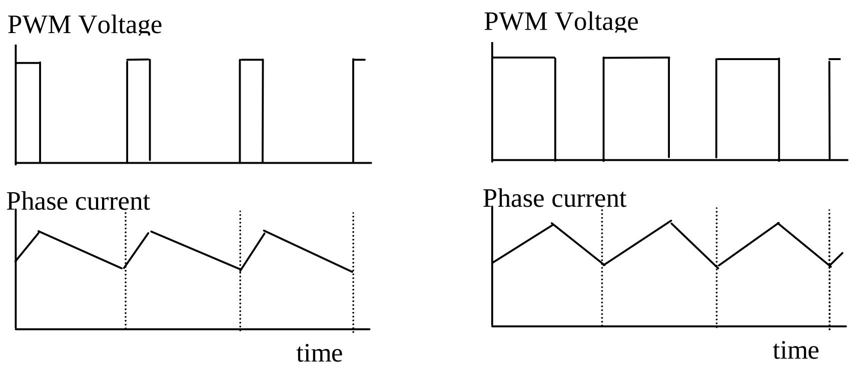
When the brushless motor is operated from a 6-step voltage source drive, permanent magnet DC motor equations of Eq. 2.1-3 are applicable.

$$V = R\,Ia + L\,dIa/dt + Ke\,\omega \qquad (2.1)$$

$$T = Kt\,Ia \qquad (2.2)$$

$$T - Tload = J\,d\omega/dt + B\,\omega \qquad (2.3)$$

In the above equation, applied voltage V produces armature current Ia, resulting in torque T. The produced torque, then, starts to accelerate the motor to a speed ω where the above three equations are satisfied in steady-state ($dIa/dt = 0$ and $d\omega/dt = 0$). Constants R (armature resistance), L (armature inductance), Kt (torque constant), and Ke (voltage constant) are determined by motor design while J (rotor and load inertia) and B (viscous friction) are system constants.

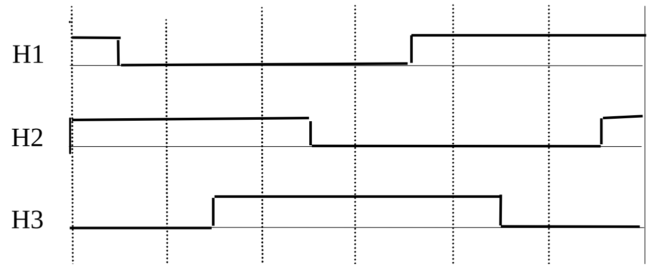


(a) 2-Quad PWM (b) 4-Quad PWM

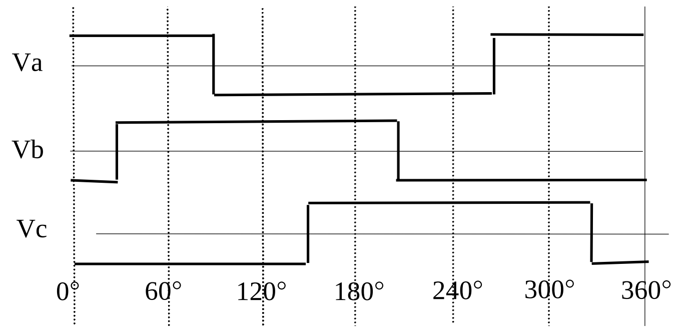
Fig. 2.3. 2-Quad vs 4-Quad PWM

When the motor is controlled by a voltage source, the armature current is solely determined by Eq. 2.1 and over-current situation may occur. In order to protect motors and drives from a destructive over-current, it is common to limit the maximum allowable current. In addition, when we wish to regulate torque (current) or shorten dynamic response time, motor current may be regulated. In either case, one needs to measure the current. Although measurement of all 3 phase currents are ideal, it is, in many applications, too costly to be justified. One of the simplest method of measuring current is inserting a low value resistor (Rs) in the DC link path as shown in Fig. 2.2. This method is very economical up to about 20A and does not require isolation in most case. When operated by the 2-quadrant switching, a positive current is detected during pwm on period and can be used to limit the current by turning pwm off as soon as current limit is detected. For 4-quadrant switching, a positive current is detected during pwm on period, while negative of one of the phase current is detected during pwm off period. Therefore, an absolute value circuit is normally necessary to limit or regulate the current in this case. An alternative method is to capture the current only during pwm on time on digital drives with A/D converters.

In general, 6-step drive produces high torque ripple, especially during transition of commutation, and the overall system efficiency is poor. In addition, audible noise might be an important concern for high power motors. Nevertheless, they are very popular in small size applications where high dynamic performance or accurate speed regulation is not required.



(a) Hall sensor outputs



(b) phase voltages

Fig. 3.1. Square-wave Mode

III. Sinusoidal Commutation and Current Control

Unlike the 6-step mode operation where only one upper and one low switches are turned on at a time, the square-wave mode operation allows more than 2 switches are on at a time as shown in Fig. 3.1. In fact, each phase voltage is either positive or negative, without allowing floating condition except during a brief moment during switching to eliminate shoot-through condition. Note that Hall sensor alignment in this case differs from the 6-step mode by 30 degrees. Although this square-wave mode operation is not popular in practice, it is the basis of the sinusoidal and the space vector pwm [1] operation. Sinusoidal voltage waveform can be generated if the square-wave commutation is pulse width modulated by the sinusoid of the same phase and frequency. Sinusoidal drives produce smooth torque (small torque ripple assuming that the back emf is close to sinusoidal) but requires higher resolution feedback sensors such as resolvers or incremental encoders.

When the phase voltage is controlled to regulate speed or torque, torque response is delayed by the electrical time constant (Te = L/R) of the motor. To shorten the torque response time, one may utilize closed-loop current control to expedite the response time. Practically, it is similar to controlling current directly, resulting in fast torque response. For high dynamic performance servo systems, current-loops are important because the performance of the outer velocity-loop is dependent on how well the current (or torque) is regulated. As a rule of thumb, current-loop bandwidth must at least be 6 - 8 times higher than required velocity-loop bandwidth. Other advantages of current regulation is to reduce the sensitivity of system non-linearities (such as inverter dead-time effect, friction) and disturbances (such back EMF). When bus current measurement is used, current magnitude can be controlled effectively. Unfortunately, this scheme is not very effective in controlling the phase lag of the current. In general, drives with bus current control have low torque at high speeds due to phase lag, compared to phase current controlled drives. Brushless motor drives have another requirement calling for high bandwidth. Unlike DC motors where current command is a constant during steady-state operation and commutation is done mechanically at the motor, commutation is executed electronically at the drive. Traditionally, the commutation function generates balanced 3 phase sinusoidal or trapezoidal current commands with angles synchronized to the rotor position. With this implementation, analog current regulator is implemented for each phase. The same commutation and current control can be realized by a microprocessor-based digital control, although different current regulation schemes such as synchronous regulator are becoming popular. For drives with phase current regulator, phase current command waveforms on the motors are sinusoidal (or with harmonics for some motors) with the commutation frequency proportional to the speed of the motor. For high speed variable motors with a large number of pole-pairs, maximum commutation frequency (f_c) may be several hundred Hz or even higher. Since current regulators on AC motors should produce negligible magnitude and phase errors at all operating frequencies in order to produce desired torque, current regulator bandwidth must be high. In general, it is desirable to have the current loop bandwidth (f_{bw}) at least 4 - 5 times the maximum electrical frequency and phase lag should be compensated by means of "phase advancing technique" [2] at the current command stage. Torque reduction at high speed operation is due to the fact that high back EMF acts as a disturbance and deteriorates the current-loop performance. Therefore, when designing current regulator of an AC drive, desired bandwidth of the current-loop should satisfy requirements not only from velocity control objective but also from the torque specification at high speed. Often, current-loop bandwidth required to satisfy the high speed torque is higher than that to meet velocity-loop performance objective.

Instead of conventional phase current regulator, the "synchronous regulator" [3] based on rotating reference frame may be employed. In this case, measured 3 phase currents are first transformed to d-axis (rotor magnetic axis) and q-axis (axis in quadrature to d-axis) current components based on the rotor angle and then current-loops are closed in order to produce desired d- and q-axis voltage commands. These voltage commands may be inverse transformed back to phase voltage commands for conventional PWM circuits or Space Vector PWM [1]. Block diagram of a synchronous regulator is introduced in Fig. 4.2. In synchronous regulators both d- and q-axis current commands are constant values in steady-state without sinusoidal modulation, much like a separately excited DC motor. When synchronous regulators are used instead of phase current regulators, the bandwidth requirement for commutation vanishes if the current compensators incorporate integral control due to its inherent zero steady-state error tracking capability to step inputs. Nevertheless, a reasonably high bandwidth, with a fast sampling rate for digital systems, is still required for smooth current regulation. Note that the concept of synchronous regulator assumes sinusoidal steady-state current and is limited to motors driven by sinusoidal drives.

A dynamic model of a PM synchronous motor in synchronous frame can be expressed as,

$$d(Iq)/dt = (1/Lq)[\ -Rs\ Iq - \omega\ Ld\ Id - \omega\ \lambda m + Vq\ ] \qquad (3.1)$$

$$d(Id)/dt = (1/Ld)[\ -Rs\ Id + \omega\ Lq\ Iq + Vd\ ] \qquad (3.2)$$

where,

Id (Iq): d- (q-) axis current components.

Vd (Vq): d- (q-) axis voltage components.

Ld (Lq): d- (q-) axis inductance.

Rs: stator resistance

ω: Commutation frequency in rad/sec.

λm : flux linkage due to PM.

On the above cross-coupled multi-variable system model, Vq and Vd are inputs and Id and Iq are outputs while the dominant system time constants are Ld/Rs and Lq/Rs for d-axis and q-axis, respectively. With the synchronous regulator, the induced EMF (back EMF) term of ωλm on q-axis is a slowly varying disturbance proportional to the speed and can simply be compensated by injecting an offset voltage. With back EMF compensation, the magnitude of current error can be kept at a small value and actual current tracks commanded current faster. Eqs.3.1-3.2 also indicates that two current-loop are coupled together. For better dynamic performance, cross-coupling terms (ωLdId and ωLqIq) may also be compensated as in Fig. 4.2. When traditional phase current regulators are used, generated back EMF of AC motors when the motor is rotating acts as a sinusoidal voltage disturbance and is difficult to compensate unless an accurate dynamic model of the motor is incorporated. A similar statement can be applicable to the de-coupling compensation. Whether back EMF compensation is employed or not, back EMF effectively reduces available DC bus voltage, resulting in lower effective gain and a low dynamic current regulator performance at high speed. When a motor is operated with flux-weakening, current regulation must be performed well under low available bus voltage. In general, drives for field-weakening operation should be designed for higher current-loop bandwidth. With practical motor drives using IGBT switches and PWM frequencies less than 25 kHz, 1 - 3 kHz bandwidth may be achieved with careful design efforts.

Consider the dynamic model of the plant for which a current regulators will be designed. For both phase current regulators and synchronous regulators, the current control circuit is composed of a compensator, a PWM amplifier connected with a motor, and a current measurement circuit. Block diagram of a current-regulator with a PI compensator is shown in Fig. 4.3. The time delay block is applicable to a digital current control system where data conversion time and computation time delays are not negligible. When a continuous model is used, the time delay should include a sample and hold delay of one half of the sampling period. An open-loop transfer function from input command $I_c(s)$ to the measured current $I_{fb}(s)$ can be modeled as a second order plus time delay (T_d) system as

$$\frac{I_{fb}(s)}{I_c(s)} = (K/L)\,\frac{1}{(s + \omega_e)(s/\omega_f + 1)}\,\exp(-\,sT_d) \qquad (3.3)$$

where open-loop gain K is proportional to the available bus voltage and L is the motor inductance. For synchronous regulators, L is either Lq or Ld, while for phase current control the inductance may swing between maximum (Lq) and minimum (Ld) value as rotor turns. A slow pole represented by ω_e is the inverse of the electrical time constant L/R as mentioned in Eqs. 3.1-2. The crossover frequency of a low-pass filter denoted by ω_f is usually on the current measurement circuit and/or anti-aliasing filter. When multiple high frequency filters are used, they may be modeled a single low-pass filter with an equivalent cross-over frequency [4].

The above model may be simplified to a first order plus time delay model by converting filter frequency ω_f into equivalent time lag and add it to the time delay. As mentioned before, the purpose of current regulator is to "emulate" a current source amplifier out of a voltage amplifier by increasing the bandwidth of the current-loop to a very high frequency, such as a few kHz.

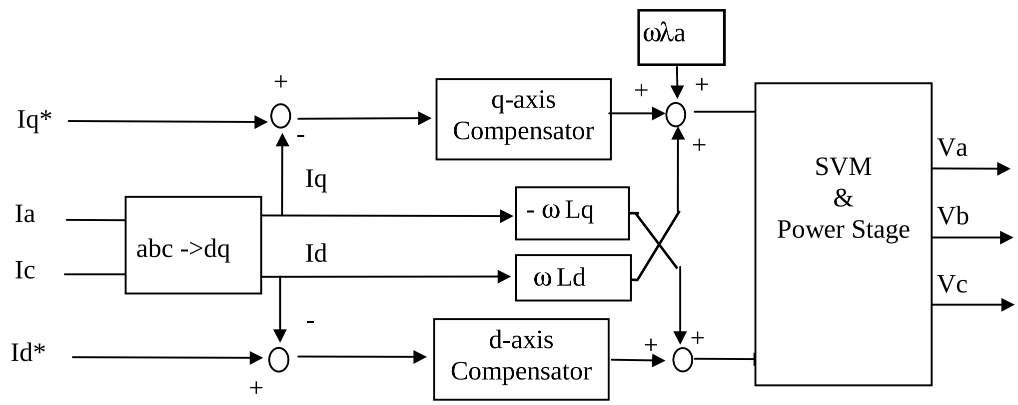


Fig. 4.2 Synchronous Regulator with Decoupling Control

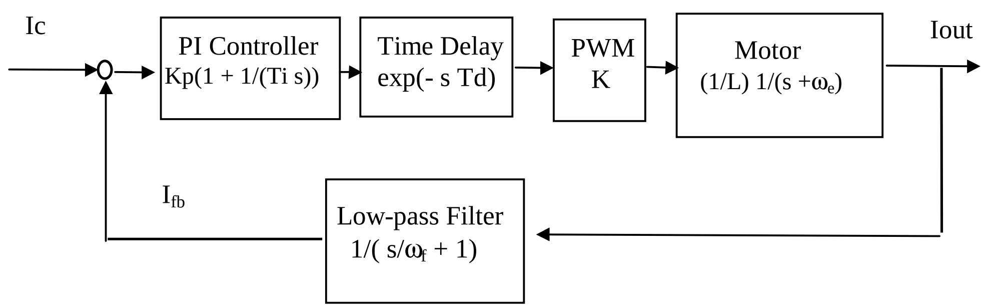


Fig. 4.3 Block Diagram of a Current-loop

There are two major factors limiting high closed-loop bandwidth on a current-loop. One is multiple pulsing phenomenon on analog current control and the other is current-loop resonance. Multiple pulsing occurs when the closed-loop system command response is faster than the slew rate of the triangle waveform in analog PWM method. In this case, the actual PWM rate would be higher than the designed PWM frequency, and could possibly cause overheating of the motor and drive and may result in unexpected drive failure. For a digital current regulator multiple pulsing does not occur due to the fact that current is sampled only once per cycle. It is not too difficult to verify that multiple pulsing frequency (f_p) is about 60% of the PWM frequency when there is no appreciable filtering (cut off frequency lower than PWM frequency) present in the circuit.

Most practical closed-loop systems tend to resonate when controller gain is increased. In current-loops, resonance occurs when unity gain occurs at a particular frequency whose phase delay is -180°. When analog current-loop is designed with minimal phase lag in the current measurement circuit, resonance frequency can be designed even higher than the multiple pulsing frequency. On the other hand, digital current control adds signal processing delay (data conversion delay and computational delay) and inherent sample and hold delay into the current loop, forcing the resonance frequency (f_r) considerably lower than that of the analog PWM circuit. Often, sampling frequency is selected to twice the PWM frequency (double sampling) in order to achieve high bandwidth. The achievable bandwidth with small overshoot is typically less than 1/2 of f_r, as indicted by a closed-loop PID tuning method known as Ziegler-Nicholson tuning rule [5].

Most practical analog and digital current regulators use the popular PI controller due to its simplicity and fast calculation time. Higher order control algorithms such as pole-zero placement control based on state-space concept has been reported [6,7]. Complex control algorithms may be used with digital control but a compromise should be made between the amount of phase lead achieved and additional computation time required to process. When signal processing time is considerably shorter than sampling period, a higher order compensator may improve dynamics. Note that in digital designs, higher sampling frequencies could sometimes be much more effective in obtaining higher bandwidth than using a higher order compensator. The PI controller has major limitations in digital current control. One is an inherent controller zero that causes overshoot. This can be easily corrected if a PDFF [8] controller is used, which is a first order pole-assignment structure with a feed-forward term. It is also called as a PID controller with set point weighing [9]. Next, the PI controller lacks phase leading capability that gives additional damping in the response. Since digital current regulator has significant time delay, a derivative term can improve system response. High order compensators may be used either to increased bandwidth or to decrease sensitivity to parameter variations.

IV. Concluding Remarks

For brushless motor drives, current control is often used to improve dynamic and steady-state performance as well as reliability. This tutorial paper analyzes and compares various different schemes of electronic commutation and associated current control methods. Although a simple 6-step 2-quadrant drive without current control may be sufficient for some application, varying degrees of sophistication on commutation and current control are required to achieve performance requirement governed by application needs. Several basic principles regarding commutation and current control methods are summarized as follows.

(1) Sinusoidal control offers smoother torque, quieter operation and higher efficiency compared to 6-step commutation. It requires higher resolution feedback device such as resolver or incremental encoders.

(2) Bus current control method does not compensate for the inherent phase lag of the line currents, while phase current regulator with sinusoidal commutation reduces the phase lag, resulting in higher torque at high speeds.

(3) Further compensation of phase lag may either be achieved by phase advance technique, or by use of synchronous regulator.

(4) Required current-loop bandwidth of the servo system comes from two performance objectives. First, it should be at least 6-8 times higher than desired velocity loop bandwidth. Second, it should be at least 5 times the maximum excitation frequency in order to amplify current wave-form without too much attenuation.

REFERENCES

[1] J. Holtz, "Pulse-Width Modulation for Electronic Power Conversion," Proceedings of IEEE, Vol.82, No.8, pp. 1194-1214, Aug. 1994.

[2] J. S. Whited, U.S. Patent 4,447,771, "Control Systems for Synchronous Brushless Motors," May 4, 1984.

[3] T. M. Rowan and R. J. Kerkman, "A New Synchronous Current Regulator and Analysis of Current-Regulated PWM Inverters," IEEE Trans. IAS, Vol.IA-22, No. 4, pp. 678-690, Jul/Aug 1986.

[4] F. Froehr and F. Orttenburger, Introduction to Electronic Control Engineering, Siemens Aktiengesell-shaft and Hyden & Son Ltd., 1982.

[5] G. F. Franklin, Digital Control of Dynamic Systems, 2nd Ed., Addison-Wesley, 1990.

[6] F. L. Lewis, Applied Optimal Control & Estimation, Prentice-Hall, 1992.

[7] D. Jouve, J. P. Rognon and D. Roye, "Effective Current and Speed Controllers for Permanent Magnet Machines: A Survey," IEEE CH2853-0/90/0000-0384, 1990

[8] D. Y. Ohm, "Analysis of PID and PDF Compensators for Motion Control Systems," IEEE IAS Annual Meeting, pp. 1923-1929, Denver, Oct.2-7, 1994.

[9] K. Astrom and T. Hagglund, PID Controllers: Theory, Design and Tuning, 2nd Ed., ISA, 1994.

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