Chapter 5
Speed and Torque Control

Introduction

This chapter describes in detail the speed and torque control structures of the motor controller.

As described in Chapter 1, Overview, the traditional cascade system of inner torque control loop and outer speed control loop is not used. Instead, parallel speed and torque control structures are implemented with the controller switching from the speed to the torque control structure when the current limit is reached.

In this chapter, the simpler torque control structure is first described in detail, then the speed control structure and the mechanism for switching between the two structures are described.

Tuning Parameter Scaling

The tuning parameters are the settings used to set the gains of the various proportional and integral components of the feedback loops of the speed and torque control structures. Their optimum values depend on the motor and load parameters, in particular the motor inductance L, the rotor flux linkage λ_r and the total inertia J.

To make tuning of the feedback loops as easy as possible, the tuning parameters are made dimensionless by scaling them by the right combinations of L, λ_r and J. In this way, the tuning parameters are automatically adjusted when the motor parameters are correctly set.

A particularly important combination of L, λ_r and J is the natural resonance of the motor with added load inertia. When the motor is driven open-loop and unloaded to high speed, with the voltage waveform matching the back emf, the motor speed will tend to oscillate with very little damping. During this oscillation, as the rotor phase angle moves away from the phase angle of the voltage waveform, a restoring torque current is generated with an amplitude depending on L and λ_r. This drives the rotor phase back to equilibrium, but the rotor and load inertia results in overshoot of the rotor phase error causing the oscillation. The frequency of this oscillation is the natural resonance ω_n of the motor and inertia, expressed in radians per second. The value of ω_n is easily calculated by the formula:

$$\omega_n = \frac{\lambda_r}{\sqrt{LJ}} \tag{5.1}$$

Related to the natural resonance is the natural impedance R_n expressed in Ohms, which is a measure of the ratio of the peak phase error to the peak restoring torque during the natural oscillation. Its formula is:

$$R_n = \lambda_r \sqrt{\frac{L}{J}} \tag{5.2}$$

Integral gains in the feedback loops of the controller are set to a constant times ω_n. Damping gains are set to a constant times R_n. This results in the same value of the constants for best dynamic performance independent of the motor and load, provided the correct estimates of L, λ_r, and J are used. This removes the need to manually adjust the drive tuning parameters by trial and error methods.

Note that ω_n and R_n depend on the square roots of inductance L and inertia J. As a result, the drive stability is generally quite insensitive to errors in the estimates of these values. For example, speed control will usually be stable for inertia errors of up to 10:1.

Torque Control Structure

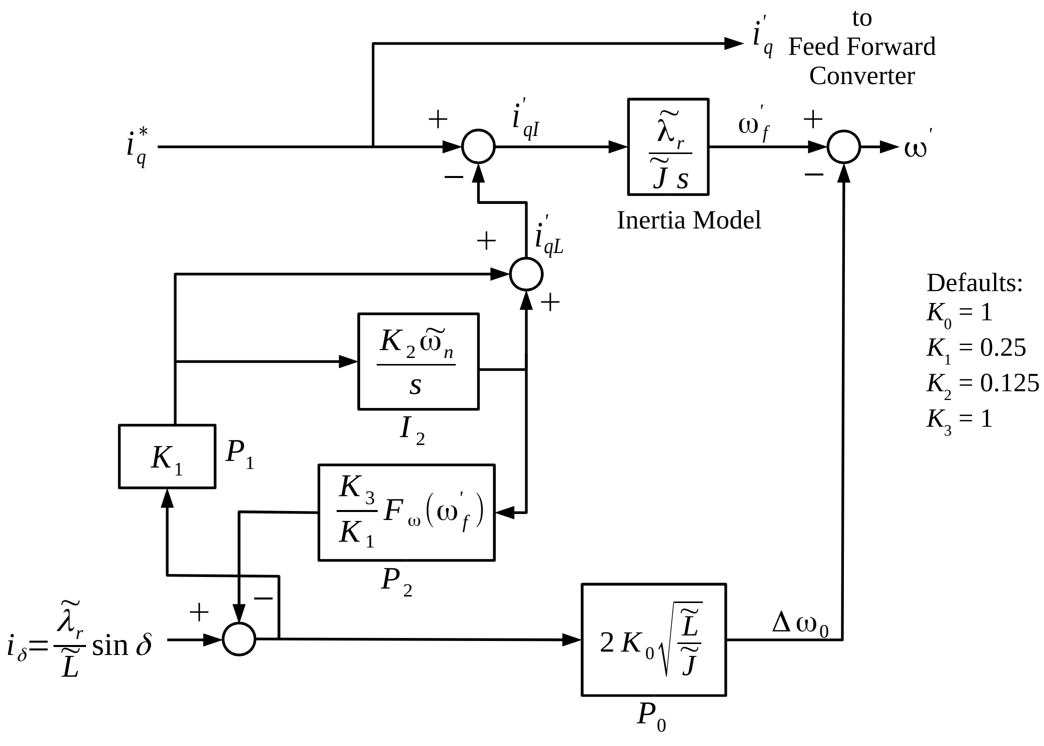

Figure 5.1. Torque control structure with default tuning parameters.

A block diagram of the torque control structure is shown in Figure 5.1. In this structure, the Inertia Model has a transfer function $\tilde{\lambda}_r / \tilde{J}\,s$ where $\tilde{J}$ is an estimate of the 2-pole equivalent inertia (actual inertia divided by the square of the number of pole pairs). The output of the Inertia Model is the speed ω'_f, which can be considered as the filtered version of the applied output speed ω'. The difference is the addition of a stability correction term $\Delta\omega_0$ from block P_0. The controller requires a proportion of the torque current error Δi_q, which is a component of the angle error input i_δ (in terms of equivalent current error), to be subtracted from the speed to stabilise the drive to prevent hunting. The required correction speed $\Delta\omega$ is given by:

$$\Delta\omega = 2\zeta\sqrt{\frac{L}{J}}\, i_\delta \tag{5.3}$$

where ζ is the damping factor, typically set to 1 for critical damping. The required damping is provided by gain block P_0 with gain set by Equation 5.3 and with constant K_0 set to the damping factor, and using estimates of L and J. The output speed correction is $\Delta\omega_0$.

Before being applied to the load model, the command torque current i^*_q is corrected by subtracting the applied load torque current i'_{qL}. Ignoring for now block P_2, which takes effect at low speed, this load torque current is found from a proportional-integral controller, blocks P_1 and I_2 in Figure 5.1, with the angle error i_δ as its input. Any change in actual load torque will cause an angle error to occur, which will then be corrected by the PI controller. In this way, the applied load torque current tracks the real load torque current.

To select the controller gains, a dynamic response analysis is required, which is undertaken in detail in a later chapter of this document. Briefly, it is necessary to recognise that this torque PI controller acts with the stabilising branch P_0 to form a second-order P+I+I^2 controller. The corresponding normalised gain constants are K_0, K_1 and K_2 as shown in Figure 5.1. Constant K_0, which as previously shown, is the damping factor, is typically set to 1. Constants K_1 and K_2 can be set to 0.5 and 0.25 for fastest response characteristics, but are more often set to 0.25 and 0.125 for lower noise sensitivity.

To prevent the output of the integrator I_2 drifting at zero speed when i_δ does not respond to changes in the load torque, the integrator is turned into a low pass filter by adding a feedback path around it, shown as block P_2. This path is only needed at or near zero speed, so its gain is reduced as speed is increased by including the function $F_\omega(\omega'_f)$. This is a simple ramp function, normally staying at 1 up to $\omega'_f = 1.0\,\tilde{\omega}_n$ then ramping down to zero at $\omega'_f = 2.0\,\tilde{\omega}_n$. For maximum flexibility, parameters are included in the software to allow the two inflexion points in this function to be modified.

The output of P_2 is subtracted from the angle error i_δ rather than the input of the integrator I_2 in order to keep the average value of $\Delta\omega_0$ zero, necessary to ensure there is no offset in the filtered speed ω'_f when $F_\omega(\omega'_f)$ is not zero. This is especially important when operating at low and zero speed in the speed control mode.

The DC gain of the integrator at zero speed is the inverse of the gain constant K_3, shown in Figure 5.1. The choice for K_3 depends on the application. When the motor is brought to zero speed, the value stored on the output of the integrator I_2 is the stored load torque from when the motor speed was high enough for the load torque to be measurable, which is when the motor back EMF is high enough to drive a large change in Δi_q when the motor flux phase angle differs from the applied phase θ'. When zero speed is reached, this value slowly returns to zero at a rate depending on the value of K_3. At the same time, the rotor angle settles to an offset value from the applied angle, depending on the standstill load torque and the level of d-axis current i_d that has been applied to hold the rotor in position. For a fixed load torque where the speed is changed slowly through zero, a low value for K_3 is preferred, perhaps 0.1 or less. For a position controller where settling time at zero speed should be as fast as possible, a high value for K_3 is called for, perhaps as high as 1.0. Another consideration is that a low value of K_3 makes a controller driving a 3-phase motor sensitive to PWM dead-band errors, although not for a controller driving a 2-phase motor. K_3 is typically set to 1.0 for a 3-phase motor and 0.25 for a 2-phase motor.

Speed Control Structure with Change-Over

The speed control structure of Figure 1.1 laid over the torque control structure of Figure 5.1 is shown in Figure 5.2.

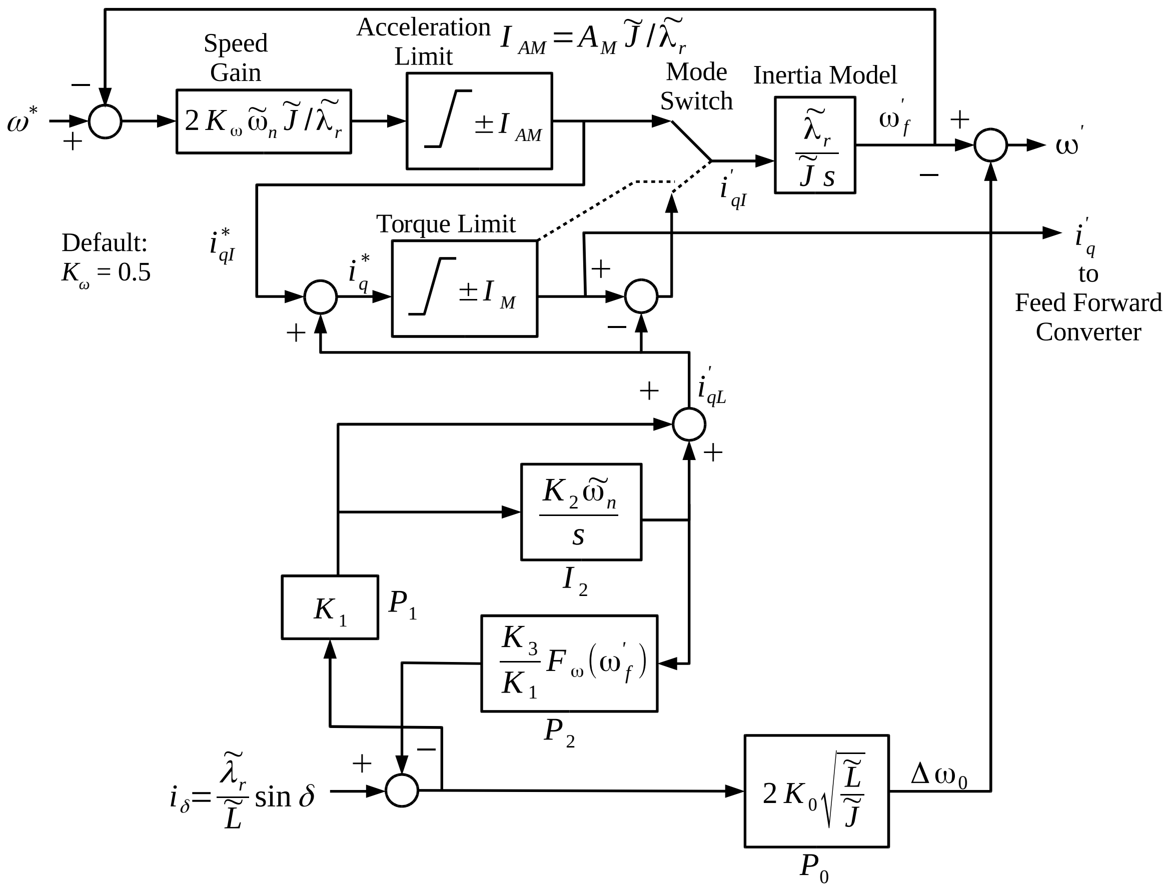

Figure 5.2. Speed control structure added to torque control structure.

As described in Chapter 1, speed is controlled directly, rather than by using a speed feedback loop. The applied speed ω' is integrated directly to the applied rotor angle θ', which rotates the output voltage vector.

The path of the speed signal is shown in Figure 5.2. which shows the mode switch set to speed control mode. The speed command signal ω^* first passes through a local observer feedback structure consisting of the Speed Gain, the Acceleration Limit and the Inertia Model blocks. This structure serves multiple purposes.

First, it acts as a single-pole filter with an output filtered speed signal ω_f'. The roll-off frequency of this filter is set by the speed gain parameter K_ω. The scaling in the Speed Gain block is chosen so that K_ω is the filter roll-off frequency as a proportion of the motor natural frequency ω_n. The default value of $K_\omega = 0.5$ gives an extremely fast response time without over-shoot, with the resulting step speed response time limited mainly by the acceleration limit rather than the speed filter roll-off frequency. As in for the torque control mode, a stability correction signal $\Delta\omega_0$ is added to ω_f' to get the final speed signal ω'.

Second, an inertial model of the motor and load is placed just before the output speed signal ω_f' with the feedback loop forcing the applied q-axis current i_{qI}' on the input to this model to match the inertial component of the motor torque current.

Between the Speed Gain block and the Inertia Model block is the Acceleration Limit block, which limits the value of i_{qI}' to $\pm A_M \widetilde{J}/\widetilde{\lambda}_r$ limiting the rate of change of ω_f' and thus the motor acceleration rate to A_M in radians per second.

As shown in Figure 5.2, in speed control mode, the applied q axis current i_q' is the sum of the estimate of the inertial torque current i_{qI}' picked of from the speed observer loop and the load torque current i_{qL}' obtained as per in the torque control structure shown in Figure 5.1 from the PI control loop that keeps the rotor angle error zero.

As shown in Figure 5.2, the mode control switch is controlled by the Torque Limit block. This block limits the value of i_q' and thus the motor torque current range to a set value $\pm I_M$. Whenever this torque limit is applied, the mode switch is also changed to torque control mode. When the torque current drops back below this limit, the mode switch changes back to speed control mode.

Note that to obtain high speed resolution, the intermediate values in the observer loop must be kept at high resolution. In the prototype software, speed values use 15 bit resolution and i_q' uses 28 bit resolution to obtain a 1 in 30,000 speed resolution and a wide acceleration limit range. To obtain the high resolution, especially at zero speed (where creeping would be normally unacceptable), the integrator I_2 must also be implemented with high resolution to ensure there is no zero error offset in the average value of $\Delta\omega_0$.

Torque Limit Adjustment for Flux Error

The Torque Limit of Fig. 5.2 limits the applied q-axis current i_q' to the maximum value I_M. Normally this is close in value to the q-axis current i_q, so this works to limit i_q to I_M. But if the estimated rotor flux differs from the actual rotor flux, the feedback adjustments will change i_q' away from i_q to correct the q-axis output voltage error caused by the flux error. This results in the Torque Limit limiting i_q to the wrong value. To correct this, the Torque Limit block is modified to that shown in Figure 5.3.

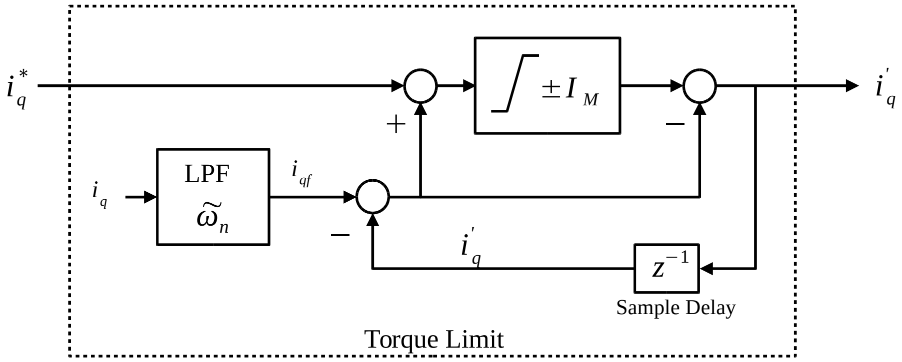

Figure 5.3. Modification of the Torque Limit block for accurate torque current limiting.

In this modified Torque Limit block, the measured q-axis current i_q is first filtered with a low-pass filter, with a roll-off frequency at the natural frequency ω_n, to remove noise and ripple. The filtered output, i_{qf}, is also used for PWM dead-time correction. The command input i_q^*, which is equal to i_q' before limiting, is converted to i_{qf} by adding $i_{qf} - i_q^*$ before the current limiting block then converted back to i_q' by subtracting $i_{qf} - i_q^*$ afterwards.

Step Over-Torque Current Limiting

The over-torque current limiting feature described here is fully developed for stepper motors but is still experimental for high-speed 3-phase motors. It may be turned off when driving these types of motors.

When a step torque exceeding the maximum torque setting is applied, usually due to a mechanical fault, the detected load torque current signal i_{qL}' will rise rapidly until the torque limit setting is reached at which point the controller will switch to torque limit mode. If the step torque is very large, i_{qL}' may overshoot to a level where the motor drive's hard current limit is reached and shut down the drive. To prevent this, the torque control structure of Figure 5.1 is modified to that of Figure 5.4 which incorporates a method of limiting the over-shoot.

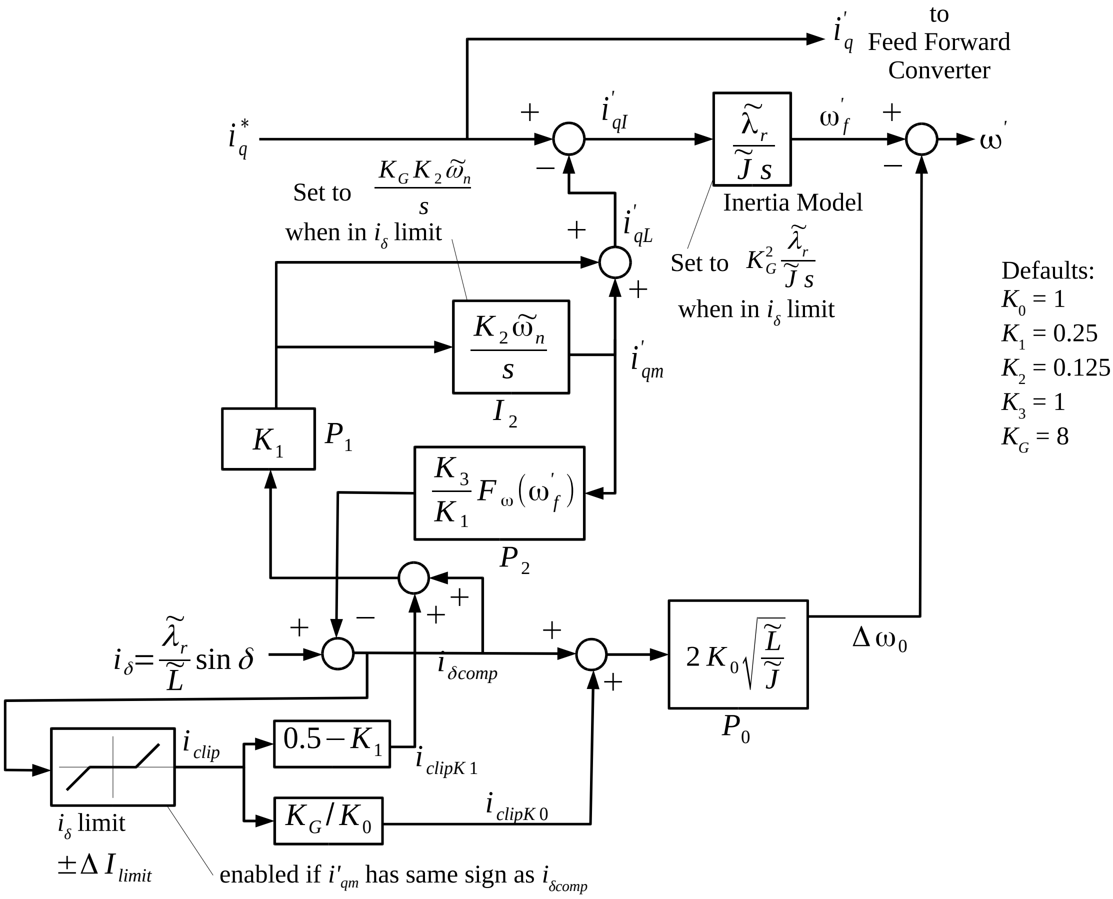

Figure 5.4. Torque control structure with overshoot clipping.

In this method, over-shoot is limited by considerably increasing the feedback gains when the angle error current i_δ exceeds a current limit ΔI_{limit}. The excess current i_{clip} is formed from $i_{\delta comp}$ using the following relation:

$$\begin{aligned} &i_{clip} = i_{\delta comp} && -\Delta I_{limit} \le i_{\delta comp} \le \Delta I_{limit} \\ &i_{clip} = i_{\delta comp} - \Delta I_{limit} && i_{clip} > \Delta I_{limit} \\ &i_{clip} = i_{\delta comp} + \Delta I_{limit} && i_{clip} < -\Delta I_{limit} \end{aligned} \qquad (5.4)$$

For the excess current, the feedback gains are greatly increased to force the rotational speed ω' to change rapidly to quickly cut the output torque and thus the amplitude of the current over-shoot. This is achieved by greatly reducing the value of the estimated inertia used for the gain constants by an amount K_G^2. The default value of K_G is 8, giving a reduction is the estimated inertia used of 1/64. Also, the proportional constant K_1 is increased to its maximum value of 0.5.