GEORGIA INSTITUTE OF TECHNOLOGY
SCHOOL of ELECTRICAL and COMPUTER ENGINEERING

ECE 4550 — Control System Design — Fall 2017

Lab #8: Position Control with AC Synchronous Motors

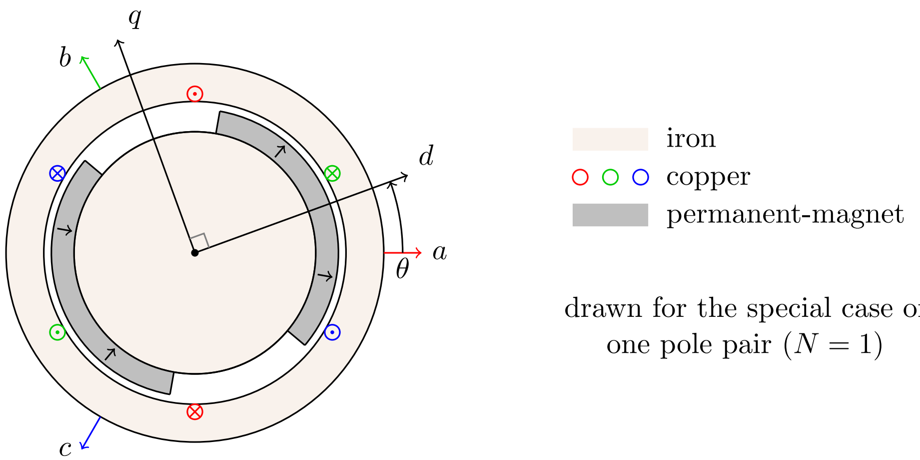
Contents

1 Background Material

1.1 Introductory Comments

In Labs 5 and 6, the concepts of PWM actuation, QEP sensing, and position control were introduced in the context of DC motors. As summarized in the posted notes, there are numerous practical motivations to favor AC motors over DC motors in applications. However, control systems for DC motors and AC motors are quite distinct from one another on the physical level, due to the fact that DC motors are described by single-input linear dynamic models whereas AC motors are described by multi-input nonlinear dynamic models. The objective of this lab is to illustrate how these physical distinctions may be suppressed by clever modeling abstraction and interfacing structure, thus enabling position control with AC motors (specifically permanent-magnet synchronous motors) to closely resemble position control with DC motors on the higher level.

Considering the cross-section diagram below, measurement of θ enables the controller to locate the applied voltage vector, and hence the resulting current vector, in the dq rotor reference frame, with a favorable orientation. What would be a favorable orientation for the current vector in the dq plane? Any vector can be decomposed into a d-axis component and a q-axis component. Consider the effect of these two components separately. The d-axis component of current will establish magnetic poles on the stator that are perfectly aligned with the rotor magnets, so this component of current does not induce any torque. The q-axis component of current will establish magnetic poles on the stator that are maximally unaligned with the rotor magnets, so this component of current has the strongest possible influence on induced torque. Since the converter impresses voltage rather than current, we will need to exploit the physics-based relationship between voltage and current in our controller design if we intend to make use of this field-orientation control principle.



drawn for the special case of
one pole pair ($N = 1$)

1.2 Plant Model

Appropriate models of the motor and the converter are required for controller design. This lab will exploit a motor model that uses the rotating dq axes as the frame of reference for expressing voltages and currents. In the dq frame of reference, the motor will have DC voltages and currents, rather than AC voltages and currents, when operating at constant speed.

1.2.1 Motor Model in Phase Variables

In posted notes, two equivalent versions of the motor model were developed in detail; the simpler of these two models is reproduced below for convenience. In terms of phase voltages v_a, v_b, v_c and phase currents i_a, i_b, i_c, the system of differential equations describing the motor is

$$\frac{d\theta}{dt} = \omega \tag{1}$$

$$\frac{d\omega}{dt} = \frac{1}{J}\left(-\Lambda_m N \left(i_a \sin(N\theta) + i_b \sin(N\theta - \tfrac{2\pi}{3}) + i_c \sin(N\theta + \tfrac{2\pi}{3})\right) - F\omega\right) \tag{2}$$

$$\frac{di_a}{dt} = \frac{1}{L}\left(v_a - Ri_a + \Lambda_m N\omega \sin(N\theta)\right) \tag{3}$$

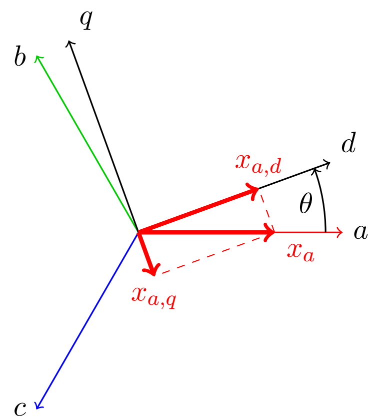
$$\frac{di_b}{dt} = \frac{1}{L}\left(v_b - Ri_b + \Lambda_m N\omega \sin(N\theta - \tfrac{2\pi}{3})\right) \tag{4}$$

$$\frac{di_c}{dt} = \frac{1}{L}\left(v_c - Ri_c + \Lambda_m N\omega \sin(N\theta + \tfrac{2\pi}{3})\right) \tag{5}$$

where N is the number of rotor pole pairs, R is the phase resistance, L is the phase inductance, Λ_m is the permanent-magnet flux, J is the rotational inertia and F is the viscous friction coefficient; the rotor motion is described by angular position θ and angular speed ω. This model captures the internal features of the motor, such as how torque is induced on the rotor and how voltage is induced on the stator; the phase variables are physically-relevant variables, so this model is quite tangible. However, significant simplification arises if we turn our attention to a frame of reference that rotates with the rotor magnets, the dq frame; this is the key idea to reformulating AC behavior into equivalent DC behavior that is easier to analyze.

1.2.2 Motor Model in Rotor-Frame Variables

The phase-variable model (1)–(5) includes a separate state equation for each phase current, driven by the corresponding phase voltage. A much simpler model results by projecting these phase voltages and phase currents onto coordinate axes that rotate with the rotor magnets. The geometric interpretation of such projections is shown in the figure below.



drawn for the special case of one pole pair ($N = 1$)

x_a is phase-a voltage (v_a) or phase-a current (i_a)

Application of trigonometric reasoning to this diagram leads to transformation equations. Choosing the arbitrary scale factor to assure power invariance, the forward change of variables is

$$\begin{bmatrix} v_d \\ v_q \end{bmatrix} = \sqrt{\tfrac{2}{3}} \begin{bmatrix} \cos(N\theta) & \cos(N\theta - \frac{2}{3}\pi) & \cos(N\theta + \frac{2}{3}\pi) \\ -\sin(N\theta) & -\sin(N\theta - \frac{2}{3}\pi) & -\sin(N\theta + \frac{2}{3}\pi) \end{bmatrix} \begin{bmatrix} v_a \\ v_b \\ v_c \end{bmatrix}$$

$$\begin{bmatrix} i_d \\ i_q \end{bmatrix} = \sqrt{\tfrac{2}{3}} \begin{bmatrix} \cos(N\theta) & \cos(N\theta - \frac{2}{3}\pi) & \cos(N\theta + \frac{2}{3}\pi) \\ -\sin(N\theta) & -\sin(N\theta - \frac{2}{3}\pi) & -\sin(N\theta + \frac{2}{3}\pi) \end{bmatrix} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix}$$

and the reverse change of variables is

$$\begin{bmatrix} v_a \\ v_b \\ v_c \end{bmatrix} = \sqrt{\tfrac{2}{3}} \begin{bmatrix} \cos(N\theta) & -\sin(N\theta) \\ \cos(N\theta - \frac{2}{3}\pi) & -\sin(N\theta - \frac{2}{3}\pi) \\ \cos(N\theta + \frac{2}{3}\pi) & -\sin(N\theta + \frac{2}{3}\pi) \end{bmatrix} \begin{bmatrix} v_d \\ v_q \end{bmatrix}$$

$$\begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} = \sqrt{\tfrac{2}{3}} \begin{bmatrix} \cos(N\theta) & -\sin(N\theta) \\ \cos(N\theta - \frac{2}{3}\pi) & -\sin(N\theta - \frac{2}{3}\pi) \\ \cos(N\theta + \frac{2}{3}\pi) & -\sin(N\theta + \frac{2}{3}\pi) \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix}.$$

The result of the change of variables is that the rotor dynamics (1)–(2) gets transformed into

$$\frac{d\theta}{dt} = \omega \tag{6}$$

$$\frac{d\omega}{dt} = \frac{1}{J}\left(\Lambda N i_q - F\omega\right) \tag{7}$$

and the stator dynamics (3)–(5) gets transformed into

$$\frac{di_d}{dt} = \frac{1}{L}\left(v_d - Ri_d + N\omega L i_q\right) \tag{8}$$

$$\frac{di_q}{dt} = \frac{1}{L}\left(v_q - Ri_q - N\omega L i_d - N\omega\Lambda\right) \tag{9}$$

where

$$\Lambda = \sqrt{\tfrac{3}{2}}\,\Lambda_m.$$

The simplifications apparent in the dq model may be summarized as follows; the induced torque in (6)–(7) no longer depends on θ, and the induced voltages in (8)–(9) no longer depend on θ.

1.2.3 Reduced-Order Model for Controller Design

Recall that the DC motor position controller of Lab 6 relied on neglecting the small electrical time constant; by setting that time constant equal to zero in the DC motor model, we obtained and utilized an expression for the steady-state current that flows in response to an applied voltage. The same idea is applicable to AC motor control in this lab. Evaluating (8)–(9) at steady-state so as to neglect the small time constant in the voltage-to-current response, we obtain

$$\begin{bmatrix} v_d \\ v_q \end{bmatrix} \approx \begin{bmatrix} R & -N\omega L \\ N\omega L & R \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} + \begin{bmatrix} 0 \\ N\omega\Lambda \end{bmatrix}$$

from which it follows that

$$\begin{bmatrix} i_d \\ i_q \end{bmatrix} \approx \frac{1}{R^2 + (N\omega L)^2} \begin{bmatrix} R & N\omega L \\ -N\omega L & R \end{bmatrix} \begin{bmatrix} v_d \\ v_q - N\omega\Lambda \end{bmatrix}.$$

Note how the steady-state values of both i_d and i_q are influenced by both v_d and v_q in a coupled way; as a consequence, the steady-state value of magnetic torque is influenced by both v_d and v_q in a coupled way as well, even though magnetic torque depends only on i_q:

$$T_e = \Lambda N i_q \approx \frac{\Lambda N}{R^2 + (N\omega L)^2}\left(Rv_q - N\omega L v_d - RN\omega\Lambda\right).$$

These expressions for current and torque are nonlinear with respect to ω, so if they are used without further modification then the design model would be nonlinear. Since the controller design methods taught in this course rely on a linear design model, an additional linearizing approximation is required. In particular, we will approximate the above expressions for current and torque by assuming that the resistive impedance dominates the reactive impedance, $R \gg N\omega L$, such that

$$\begin{bmatrix} i_d \\ i_q \end{bmatrix} \approx \frac{1}{R} \begin{bmatrix} v_d \\ v_q - \Lambda N\omega \end{bmatrix}, \quad T_e \approx \frac{\Lambda N}{R}\left(v_q - \Lambda N\omega\right). \tag{10}$$

This approximation is linearizing in the sense that current and torque are now expressed as linear functions of v_d, v_q and ω. This linearizing approximation is certainly valid in a low speed range; although not particularly accurate at higher speeds, we will show by simulation and experiment that this linearizing approximation still leads to a simple and satisfactory control system design.

According to (10), i_d may be manipulated by v_d whereas i_q and T_e may be manipulated by v_q; therefore, we will control the motor with voltage vectors satisfying

$$\begin{bmatrix} v_d \\ v_q \end{bmatrix} \equiv \begin{bmatrix} 0 \\ u \end{bmatrix} \tag{11}$$

such that the corresponding current vectors will satisfy

$$\begin{bmatrix} i_d \\ i_q \end{bmatrix} \approx \frac{1}{R} \begin{bmatrix} 0 \\ u - \Lambda N \omega \end{bmatrix} \tag{12}$$

where u denotes the torque-producing component of voltage; this design results in $i_d \approx 0$, which maximizes the efficiency of torque production since the losses due to current flow are $P_{\text{loss}} = Ri_d^2 + Ri_q^2$. Substitution of (12) into (6)–(7) results in

$$\frac{d\theta}{dt} = \omega \tag{13}$$

$$\frac{d\omega}{dt} = -\alpha\omega + \beta u \tag{14}$$

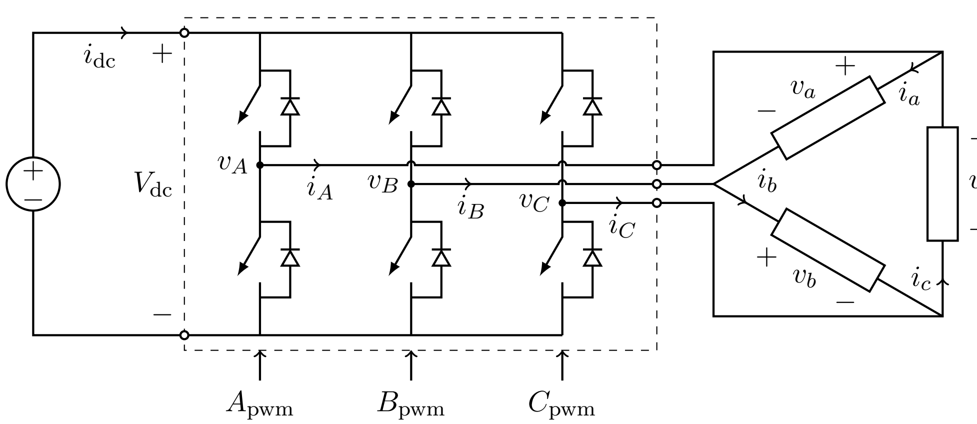
where

$$\alpha = \frac{K^2 + FR}{JR}, \quad \beta = \frac{K}{JR}, \quad K = \Lambda N.$$

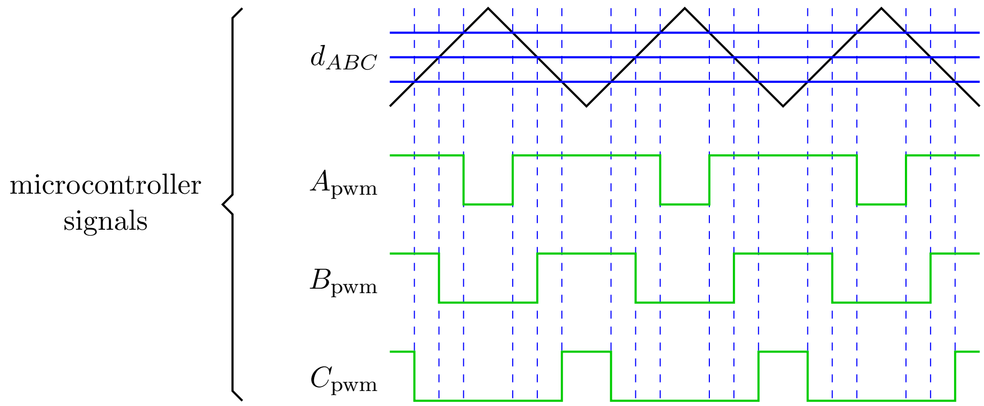
As a consequence of using the dq reference frame, the AC motor design model and the DC motor design model have identical structures.

1.2.4 Power Converter Model

Now that we have an appropriate state-space model of the motor and load, it is time to consider the switched-mode power converter circuit used to provide excitation to the motor. The circuit, which is often referred to as a three-phase inverter to emphasize that it converts DC voltage into AC voltage, has much in common with the full-bridge circuit used to control DC motors. As shown in the schematic diagram below, there are three separate half-bridge legs each consisting of a high-side switch and a low-side switch. By operating the two switches in a given leg in complementary fashion using PWM, it is possible to synthesize any average voltage between V_{dc} and zero at the terminal located between the two switches of that leg. The three converter terminals, labeled A, B and C below, would be connected to a corresponding set of three terminals on the Y or Δ connected motor. Since our motor is Δ connected, only that case has been shown.



To impose a particular set of voltages v_A, v_B, v_C at the motor terminals, referenced to the negative supply terminal, all three converter legs will be operated independently, as depicted below; this is unlike the DC motor case, in which two converter legs operate in coordinated fashion.



Comparison of the independently-assigned duty cycles (three blue lines) with a single triangular reference waveform (black) leads to corresponding command signals (three green waveforms) fed to the power converter. Each switching cycle occurs over the relatively short time interval of length T_{pwm}, i.e. one period of the triangular reference waveform appearing in the PWM diagram. The kth control cycle occurs on the time interval $kT \leq t < (k+1)T$ of length T. During this one control cycle, the average values of the converter leg voltages are determined by the corresponding instantaneous voltage waveforms as dictated by the duty cycles $d_A[k]$, $d_B[k]$ and $d_C[k]$. Assuming that T/T_{pwm} is an integer, the voltage values obtained by averaging over one control cycle will be identical to those obtained by averaging over one switching cycle within that control cycle:

$$\bar{v}_A[k] = \frac{1}{T}\int_{kT}^{kT+T} v_A(t)\,dt = d_A[k]V_{\text{dc}} \tag{15}$$

$$\bar{v}_B[k] = \frac{1}{T}\int_{kT}^{kT+T} v_B(t)\,dt = d_B[k]V_{\text{dc}} \tag{16}$$

$$\bar{v}_C[k] = \frac{1}{T}\int_{kT}^{kT+T} v_C(t)\,dt = d_C[k]V_{\text{dc}}. \tag{17}$$

These converter leg voltages determine the motor phase voltages according to

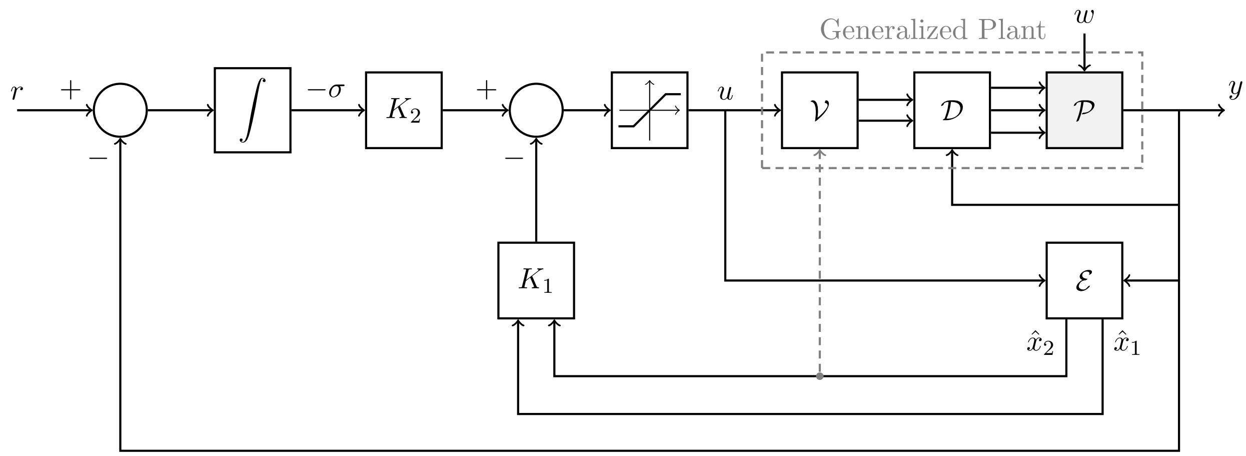
$$\begin{aligned} v_a &= v_A - v_B \\ v_b &= v_B - v_C \\ v_c &= v_C - v_A. \end{aligned}$$

The resulting motor phase currents determine the converter leg currents according to

$$\begin{aligned} i_A &= i_a - i_c \\ i_B &= i_b - i_a \\ i_C &= i_c - i_b. \end{aligned}$$

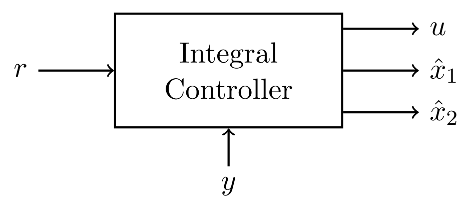
1.3 Design of the Position Control System

The structure of the overall system is displayed in the block diagram shown below. The shaded block labeled $\mathcal{P}$ represents the physical plant which consists of an actuator (the DC-AC converter), a motion system (the AC motor and load) and a sensor (the optical encoder); all unshaded blocks represent embedded code. The embedded code is organized into three parts, described below.



1.3.1 Integral Controller

In the block diagram of the overall system, everything outside the generalized plant constitutes the integral controller. The integral controller is used to provide both internal stability and the command following function; its structure matches that used for the DC motor position control system of Lab 6 (the structure of rotor dynamics is consistent for all types of motors). The controller receives two inputs—the position sensor measurement and the position reference command, which is generated on-chip in this lab—and it uses estimates of position and speed, along with the integrated position error, to generate a voltage command that will ultimately induce the desired torque:



According to (13)–(14), the design model of the AC motor plant may be taken to be

$$\begin{bmatrix}\dot{x}_1(t)\\ \dot{x}_2(t)\end{bmatrix} = \begin{bmatrix}0 & 1\\ 0 & -\alpha\end{bmatrix}\begin{bmatrix}x_1(t)\\ x_2(t)\end{bmatrix} + \begin{bmatrix}0\\ \beta\end{bmatrix}(u(t) - w(t))$$

$$y(t) = \begin{bmatrix}1 & 0\end{bmatrix}\begin{bmatrix}x_1(t)\\ x_2(t)\end{bmatrix}$$

where state variables x_1 and x_2 denote position and speed, input variables u and w denote effective control voltage and effective disturbance voltage, and output variable y denotes the position sensor measurement; load torque, including cogging torque, is treated as disturbance voltage. The full-order estimator for this design model has the form

$$\begin{bmatrix}\dot{\hat{x}}_1(t)\\ \dot{\hat{x}}_2(t)\end{bmatrix} = \begin{bmatrix}0 & 1\\ 0 & -\alpha\end{bmatrix}\begin{bmatrix}\hat{x}_1(t)\\ \hat{x}_2(t)\end{bmatrix} + \begin{bmatrix}0\\ \beta\end{bmatrix}u(t) - \begin{bmatrix}L_1\\ L_2\end{bmatrix}\left(\begin{bmatrix}1 & 0\end{bmatrix}\begin{bmatrix}\hat{x}_1(t)\\ \hat{x}_2(t)\end{bmatrix} - y(t)\right)$$

where L_1 and L_2 are design parameters. The regulator feedback loop has the form

$$u(t) = -K_{11}\hat{x}_1(t) - K_{12}\hat{x}_2(t) - K_2\sigma(t)$$

where K_{11}, K_{12} and K_2 are design parameters. The controller state equations reduce to

$$\begin{aligned}
\dot{\hat{x}}_1(t) &= \hat{x}_2(t) - L_1\left(\hat{x}_1(t) - y(t)\right) \\
\dot{\hat{x}}_2(t) &= -\alpha\hat{x}_2(t) + \beta u(t) - L_2\left(\hat{x}_1(t) - y(t)\right) \\
\dot{\sigma}(t) &= y(t) - r(t).
\end{aligned}$$

After discretization with one-cycle delay, the controller output is initially computed as

$$u^*[k] = -K_{11}\hat{x}_1[k-1] - K_{12}\hat{x}_2[k-1] - K_2\sigma[k-1], \tag{18}$$

then saturated if necessary according to

$$u[k] = \begin{cases} +V_{\max} & , \text{ if } u^*[k] > +V_{\max} \\ -V_{\max} & , \text{ if } u^*[k] < -V_{\max} \\ u^*[k] & , \text{ otherwise} \end{cases} \tag{19}$$

and the discrete-time controller state variables are updated according to

$$\hat{x}_1[k+1] = \hat{x}_1[k] + T\hat{x}_2[k] - TL_1\left(\hat{x}_1[k] - y[k]\right) \tag{20}$$

$$\hat{x}_2[k+1] = \hat{x}_2[k] - T\alpha\hat{x}_2[k] + T\beta u[k] - TL_2\left(\hat{x}_1[k] - y[k]\right) \tag{21}$$

$$\sigma[k+1] = \sigma[k] + T\left(y[k] - r[k]\right). \tag{22}$$

We will place estimator eigenvalues at $s = -\lambda_e$ by using the estimator gains

$$L_1 = 2\lambda_e - \alpha, \quad L_2 = \lambda_e^2 - 2\alpha\lambda_e + \alpha^2, \tag{23}$$

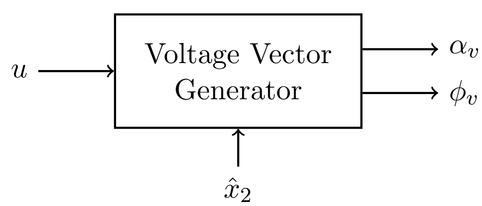
and we will place regulator eigenvalues at $s = -\lambda_r$ by using the regulator gains

$$K_{11} = \frac{1}{\beta}3\lambda_r^2, \quad K_{12} = \frac{1}{\beta}\left(3\lambda_r - \alpha\right), \quad K_2 = \frac{1}{\beta}\lambda_r^3. \tag{24}$$

These gain formulas match those of Lab 6, since the design model formulas on which they are based match those of Lab 6. Thus far, DC motor control and AC motor control are similar, which is a consequence of the fact that an electromagnetic torque accelerates a rotor inertia in both cases.

1.3.2 Voltage Vector Generator

In the block diagram of the overall system, the block labeled $\mathcal{V}$ represents the voltage vector generator. This module receives one input, the desired component of voltage along the rotor-frame quadrature axis—and optionally uses the estimated rotor speed to determine the desired component of voltage along the rotor-frame direct axis, according to a flux weakening[1] algorithm—to ultimately generate the polar coordinate representation of the desired voltage vector:



[1] Flux weakening (not used in this lab) is a strategy for injecting negative d-axis electromagnet flux to counteract the permanent-magnet flux, which is itself aligned with the d-axis. The motivation for weakening the effective d-axis flux in this way is to permit operation at high speeds that would otherwise not be reachable due to voltage limits.

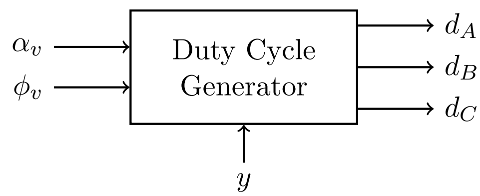
For operation that does not include flux weakening, the cartesian coordinate voltage vector is specified in (11), and the corresponding polar coordinates are computed according to

$$\alpha_v = \begin{cases} |u| & , \text{ if } |u| \leq V_{\max} \\ V_{\max} & , \text{ otherwise} \end{cases} \quad , \quad \phi_v = \begin{cases} +\frac{\pi}{2} & , \text{ if } u \geq 0 \\ -\frac{\pi}{2} & , \text{ otherwise} \end{cases} \quad . \tag{25}$$

The two coordinate systems are related by $(v_d, v_q) = (\alpha_v \cos\phi_v, \alpha_v \sin\phi_v)$.

1.3.3 Duty Cycle Generator

In the block diagram of the overall system, the block labeled $\mathcal{D}$ represents the duty cycle generator. This module receives two inputs—the polar coordinates of the desired rotor-frame voltage vector—and uses the measured rotor position to determine the three-phase duty cycles that will be required to establish the desired rotor-frame voltage vector:



The desired voltage vector has been computed in rotor-frame coordinates, but it must be implemented using the stator windings in stator-frame coordinates, so a position-dependent transformation is necessary. Using the required transformation, it follows that the desired voltage vector will be imposed if the converter leg voltages are assigned according to

$$v_A = \tfrac{1}{2}V_{\text{dc}} + \sqrt{\tfrac{2}{9}}\,\alpha_v \cos\left(Ny + \phi_v - \tfrac{\pi}{6}\right) \tag{26}$$

$$v_B = \tfrac{1}{2}V_{\text{dc}} + \sqrt{\tfrac{2}{9}}\,\alpha_v \cos\left(Ny + \phi_v - \tfrac{\pi}{6} - \tfrac{2\pi}{3}\right) \tag{27}$$

$$v_C = \tfrac{1}{2}V_{\text{dc}} + \sqrt{\tfrac{2}{9}}\,\alpha_v \cos\left(Ny + \phi_v - \tfrac{\pi}{6} + \tfrac{2\pi}{3}\right). \tag{28}$$

To impose these leg voltages in an average sense, we compute duty cycles according to

$$d_A = \frac{v_A}{V_{\text{dc}}}, \quad d_B = \frac{v_B}{V_{\text{dc}}}, \quad d_C = \frac{v_C}{V_{\text{dc}}} \tag{29}$$

at which point all controller computations for the kth sampling interval are complete.

1.3.4 Position Measurement Calibration Procedure

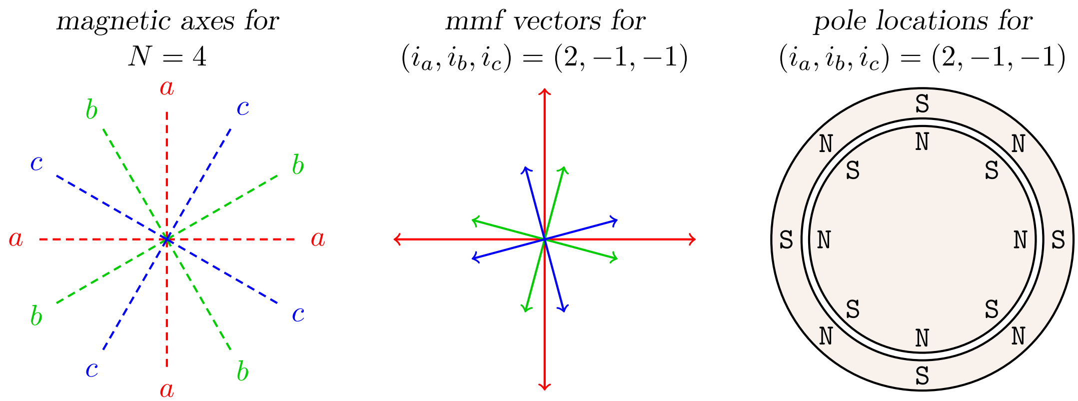
A start-up procedure is necessary to ensure that the encoder measures rotor position as shown in the motor cross-section diagram (our definition of rotor position θ is visually shown on page 2). The idea is to rely on the fact that the rotor will tend to settle into a specific fixed position for a given constant excitation. For example, consider the following constant choices for the applied voltages v_A, v_B and v_C and the resulting voltages v_a, v_b and v_c:

$$\begin{bmatrix} v_A \\ v_B \\ v_C \end{bmatrix} = \begin{bmatrix} 13.2 \\ 10.8 \\ 12.0 \end{bmatrix} \text{ V} \quad \Rightarrow \quad \begin{bmatrix} v_a \\ v_b \\ v_c \end{bmatrix} = \begin{bmatrix} 2.4 \\ -1.2 \\ -1.2 \end{bmatrix} \text{ V}.$$

Since this lab's motor has $R = 1.2\ \Omega$, it follows that the resulting currents will be

$$\begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} = \begin{bmatrix} 2 \\ -1 \\ -1 \end{bmatrix} \text{ A} \quad \Rightarrow \quad \begin{bmatrix} i_A \\ i_B \\ i_C \end{bmatrix} = \begin{bmatrix} 3 \\ -3 \\ 0 \end{bmatrix} \text{ A.}$$

Since this lab's motor has $N = 4$ pole pairs, the result of these constant phase currents on equilibrium rotor position is visualized as shown below ($\theta \to 0$ to align rotor `N` with stator `S`).



This excitation will drive the rotor to the magnetic alignment we have modeled as corresponding to $\theta = 0$ rad. Therefore, if there is not too much load torque and we wait until the rotor settles into place, then the measured position may be safely assigned the value of zero. We do not care which of the four possible alignments per revolution we end up in; all that matters is that the rotor has the correct orientation with respect to phase a, and this will be achieved. After the rotor has settled, the `SWI` field of the `QEPCTL` register is used to reset the measured position to zero.

One simple way to implement the start-up procedure is to place the required code prior to the `while(1)` loop; with this placement, it is essential to allow the start-up procedure to complete (conservatively, 1 s for rotor position settling) prior to starting the timer, since otherwise the controller will begin acting before the correct position sensor orientation has been established. Alternatively, one could enter the `while(1)` loop with the timer running but with a state machine configured to first complete the start-up task and then proceed to the control task.

1.3.5 Over-Current and Over-Temperature Protection

It is critical to understand what happens in a motor drive system if the microcontroller fails to maintain appropriate PWM operation of the converter at all times. Under normal operation, the signals from microcontroller output pins to motor driver chip input pins maintain high-frequency PWM switching at all times, such that the average voltages applied to the motor terminals result in acceptable currents flowing through the power stage of the converter and through the motor windings. If the signals from the microcontroller to the motor driver chip stop performing their intended PWM switching function, either because of faulty code or because of inappropriate user actions during a debug session, then destructive over-current conditions could occur.

For our DC motor, the worst-case situation would be when one converter leg command stays high while the other converter leg command stays low. Since the motor resistance is 7.6 Ω, the largest fault current that could possibly flow would be 3.2 A (24 V across 7.6 Ω), and this level of current would flow only if the rotor is mechanically locked in place. If the rotor is free to rotate, then the sustained fault current would be much smaller, since the rotor would accelerate to some steady-state speed at which the induced voltage would counter the voltage applied by the converter. Since the DRV8312 motor driver chip is specified to handle up to 3.5 A continuous current, we did not concern ourselves with the issue of over-current protection for our DC motor.

For the AC motor used in this lab, the worst-case situation would be when one converter leg command stays high while the other two converter leg commands stay low. Since the phase resistance is 1.2 Ω, the largest fault current would be 40 A (sum of two currents, each being 24 V across 1.2 Ω). Unlike a DC motor, an AC motor would not spin during the fault condition, so there would be no induced voltage to counter the voltage applied by the converter. Since the DRV8312 motor driver chip is specified to handle only up to 3.5 A continuous current, lack of over-current protection for our AC motor can easily lead to catastrophic damage to the motor driver chip.

The DRV8312 motor driver chip provides two over-current protection schemes, selected according to the logic level assigned to pin `M1`. On the Motor Control Motherboard, header pins permit `M1` to be set to `0` or `1` by a jumper (pins `M2` and `M3` are hard-wired to `0` and `1` respectively):

$$\texttt{M1} = \begin{cases} \texttt{0} & \text{, cycle-by-cycle current limiting mode} \\ \texttt{1} & \text{, over-current latching shutdown mode} \end{cases}$$

For our purposes, the over-current latching shutdown mode is preferred as the safer option. Therefore, always take the following precautions to 1) implement the desired form of over-current protection and 2) avoid conditions that will cause excessive currents to flow:

1. Place a jumper between pins 1 and 2 on the header labeled `M1`; note that pin 1 is labeled `H`. With this setting, the DRV8312 will operate in over-current latching shutdown mode.

2. Never use `Pause` or set breakpoints in a debug session when the motor is connected and the converter is enabled, since this would cause the PWM pins to stop switching and thereby potentially create a fault condition; instead, use `Reset CPU` followed by `Restart`.

Excessively large short-term current flows are not the only mechanism that can damage the motor driver chip. Another failure mechanism is an over-temperature condition, which could result from long-term use of moderate current levels in an environment with high ambient temperature.

Over-current protection requires current monitoring; over-temperature protection requires temperature monitoring. The DRV8312 motor driver chip has these (and other) protection schemes built in, as described in §8 of Motor Driver Chip. Two pins on the DRV8312 motor driver chip are active-low output pins that provide the two signals named `FAULT` and `OTW`; these two signals drive two LED indicators on the Motor Control Motherboard, and the microcontroller can read these two signals at GPIO pins if desired. If the DRV8312 motor driver chip experiences a condition that leads to an over-current shutdown or an over-temperature shutdown, a fault will be reported by a `0` on `FAULT`. If the DRV8312 motor driver chip reaches a temperature (125°C) that approaches the critical shutdown temperature (150°C), a warning will be reported by a `0` on `OTW`. If the DRV8312 motor driver chip is experiencing normal operation, both of these output pins will report a `1` and the two indicator LEDs will be turned off.

1.4 Simulation of the Position Control System

The methodical approach to control system design begins with physics-based plant modeling and consideration of sensing and actuating possibilities. Once a control strategy has been selected, the design equations are developed and a simulation study is conducted. The simulation study should precede hardware/software prototyping, so that issues may be resolved before wasting time, effort and money. The need for a simulation study is particularly relevant for this lab, as we have constructed the design model of the plant dynamics with an assumption that $R \gg N\omega L$; this assumption was necessary to obtain a linear design model for linear controller design, and it helped to exhibit some similarity between DC motor control and AC motor control, but—since this approximation is not very accurate at higher speeds—it will be essential to investigate the stability and performance of the resulting control system by simulation before attempting any implementation. Motivated by this philosophy, we begin this lab with a simulation study.

The simulation code posted on tsquare represents a virtual implementation of the position control method defined by the state-space integral controller (18)–(22), feedback gain assignment (23)–(24), voltage vector generator (25) and duty cycle generator (26)–(29), applied to the physical system defined by the phase-variable dynamic model (1)–(5). The simulation code includes monitoring of the measurable converter currents i_{ABC} and the computable dq motor currents i_{dq} to help understand the consequences of the assumptions mentioned above. We already know that converter currents may be computed from abc motor currents i_{abc} according to

$$\begin{bmatrix} i_A \\ i_B \\ i_C \end{bmatrix} = \begin{bmatrix} 1 & 0 & -1 \\ -1 & 1 & 0 \\ 0 & -1 & 1 \end{bmatrix} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix},$$

but the coefficient matrix of this relation is singular and thus cannot be inverted. By incorporating the additional constraint $i_a + i_b + i_c = 0$, it is possible to establish that abc motor currents i_{abc} may be computed from converter currents according to

$$\begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} = \frac{1}{3} \begin{bmatrix} 1 & -1 & 0 \\ 0 & 1 & -1 \\ -1 & 0 & 1 \end{bmatrix} \begin{bmatrix} i_A \\ i_B \\ i_C \end{bmatrix}.$$

This inverse relation and the rotor-position-dependent dq-transformation matrix have been used in the simulation code to obtain values of i_{dq} from measurements of i_{ABC}.

Zoom in on v_{ABC} and i_{ABC} to see the three-phase AC waveforms in detail, and note how the dq-frame variables exhibit DC waveforms. The simulation code does not include integrator anti-windup logic or reference command shaping, but these features may be easily added if desired; the simulation model of the plant does not include the effect of magnetic cogging torque, but this disturbance effect could be added if desired. The parameters used in the simulation code are listed in the tables below; these same parameter values are appropriate for the experimental implementation. Compared to the DC motor labs, *the timer ISR frequency has been increased by a factor of five and the position encoder resolution has been increased by a factor of four*.

Motor Parameters			
Parameter	Value	Units	Description
N	4	–	Pole Pairs
L	1.8×10^{-3}	H	Inductance
R	1.2	Ω	Resistance
J	4.8×10^{-6}	kg m^2	Inertia
F	5×10^{-5}	Nm/(rad/s)	Friction
Λ	11×10^{-3}	Wb	PM Flux

Controller-Related Parameters			
Parameter	Value	Units	Description
V_{dc}	24	V	Supply Voltage
$I_{\text{leg,max}}$	5	A	Max Converter Leg Current
$V_{\max}$	$1.061\,V_{\text{dc}}$	V	Max (dq) Motor Voltage Vector Magnitude
$I_{\max}$	$0.707\,I_{\text{leg,max}}$	A	Max (dq) Motor Current Vector Magnitude
T	2×10^{-4}	s	Controller Period
λ_r	125	rad/s	Regulator Eigenvalue
Q_s	$2\pi/4000$	rad	Sensor Resolution
Q_a	$24/1500$	V	Actuator Resolution

2 Lab Assignment

2.1 Pre-Lab Preparation

Each individual student must work through the pre-lab activity and prepare a pre-lab deliverable to be submitted *by the beginning of the lab session*. The pre-lab deliverable consists of a brief typed statement, no longer than one page, in response to the following pre-lab activity specification:

1. Read through this entire document, and describe the overall purpose of this week's project.

2. Why is a start-up procedure required for encoder-based AC motor position control (Lab 8) but not for encoder-based DC motor position control (Lab 6)?

3. Simulate the control system using the supplied code, in order to provide informed responses to the following questions; use words to convey your observations (don't submit plots).

   (a) Why has a start-up procedure not been used in the supplied code?

   (b) How long does it take to reach steady state, and how does λ_r influence this?

   (c) What are the steady-state voltages at the converter legs?

   (d) What are the peak voltages at the converter legs?

   (e) What are the steady-state power and current delivered by the source?

   (f) What are the peak power and current delivered by the source?

   (g) What is the steady-state magnitude of the integral controller's output signal u?

   (h) What is the peak magnitude of the integral controller's output signal u?

   (i) What is the steady-state magnitude of the d-axis current component i_d?

   (j) What is the peak magnitude of the d-axis current component i_d?

   (k) What is the largest swing in r that will not trip the over-current protection?

Please note that it is not essential to write application code prior to the lab session; the point of the pre-lab preparation is for you to arrive at the lab session with a clear understanding of how the plant should respond to the controller we will implement during the lab session.

2.2 Specification of the Assigned Tasks

2.2.1 Three-Phase Voltage Generation

Develop code that implements three-phase voltage generation as defined by (15)–(17), updating desired average voltage in a timer ISR with period of 200 μs. Initialize the PWM module such that all three PWM pins are commanding 50% duty cycles prior to activating the power stage.

1. The motor must be disconnected from the converter while verifying proper voltage generation. Connect three separate oscilloscope probes to the three separate converter output nodes in order to measure instantaneous converter output voltages v_A, v_B and v_C.

2. Command the following average leg voltages (one set at a time), to verify proper operation:

$$\begin{bmatrix} v_A \\ v_B \\ v_C \end{bmatrix} = \begin{bmatrix} 16 \\ 8 \\ 12 \end{bmatrix}, \begin{bmatrix} 12 \\ 16 \\ 8 \end{bmatrix}, \begin{bmatrix} 8 \\ 12 \\ 16 \end{bmatrix} \text{V}.$$

Instructor Verification (separate page)

2.2.2 Position Control with Step Reference Command

Develop code that implements the position control method defined by the state-space integral controller (18)–(22) with feedback gain assignment (23)–(24), the voltage vector generator (25) and the duty cycle generator (26)–(29). Initialize the PWM module such that all three PWM pins are commanding 50% duty cycles prior to activating the power stage. Use the start-up procedure described in §1.3.4, the protection measures described in §1.3.5, and the parameter values listed in §1.4. The reference command r should alternate abruptly between 0 rad and 2π rad with each position maintained for 0.5 seconds. Verify proper operation by plotting the measured output y along with the commanded input u in Matlab (and compare with simulation).

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GEORGIA INSTITUTE OF TECHNOLOGY
SCHOOL of ELECTRICAL and COMPUTER ENGINEERING

ECE 4550 — Control System Design — Fall 2017

Lab #8: Position Control with AC Synchronous Motors

Instructor Verification Page

Lab Section	Begin Date	End Date
L01, L02	November 7	November 14
L03, L04	November 9	November 16

To be eligible for full credit, do the following:

1. Submissions required by each student (one per student)
   (a) Upload your pre-lab deliverable to tsquare before lab session begins on begin date.
   (b) Upload your `main.c` file for §2.2.2 to tsquare before lab session ends on end date.
2. Submissions required by each group (one per group)
   (a) Submit a hard-copy of this verification page before lab session ends on end date.
   (b) Attach to this page the hard-copy plot requested in §2.2.2.

Name #1: ________________________________

Name #2: ________________________________

Checkpoint: Verify completion of the task assigned in §2.2.1.

Verified: ______________________ **Date/Time:** ______________________

Checkpoint: Verify completion of the task assigned in §2.2.2.

Verified: ______________________ **Date/Time:** ______________________