GEORGIA INSTITUTE OF TECHNOLOGY

SCHOOL of ELECTRICAL and COMPUTER ENGINEERING

ECE 4550 — Control System Design — Fall 2017

Lab #9: Position Control with AC Induction Motors

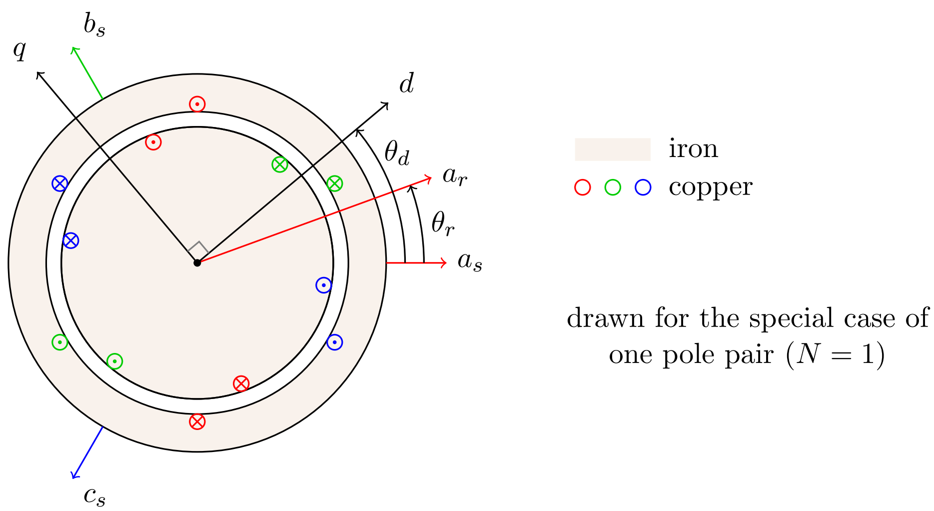
Contents

1 Background Material

1.1 Introductory Comments

Position control systems are found in many applications, and such systems often rely on a converter-fed electric motor to supply the required driving and braking torques. We have gained experience with position control technology using DC motors and AC synchronous motors. In this lab, we conclude our study of position control technology using AC induction motors. AC motors are generally preferred over DC motors, due to their brushless construction. There are just two fundamental types of AC motors; AC synchronous motors have permanent magnets on their rotors whereas AC induction motors have short-circuited coils on their rotors. Our progression through the various types of motors has gone from lower complexity to higher complexity; a DC motor has just one control variable (scalar voltage), an AC synchronous motor has two control variables (vector voltage), and an AC induction motor has three control variables (vector voltage and excitation frequency). AC synchronous motors may have higher power conversion efficiency than AC induction motors, but they are typically more expensive due to their use of high-energy permanent-magnet materials, so a design trade off exists. In the all-electric vehicle space, the Chevy Bolt uses AC synchronous motors whereas the Tesla Model 3 uses AC induction motors.

A cross-section diagram of an AC induction motor is shown below. Its stator is essentially identical to the stator of an AC synchronous motor, but its rotor is completely different, consisting of short-circuited coils instead of permanent magnets. For modeling purposes the rotor of an AC induction motor is assumed to incorporate a three-phase winding, but in reality its conductors are often formed by a casting process that results in a so-called "squirrel cage" assembly. As shown below, the stator and rotor each have three magnetic axes associated with their sets of three phases; the stator magnetic axes are stationary, whereas the rotor magnetic axes rotate. The rotor position θ_r will be measured from the stator phase-a axis to the rotor phase-a axis; however, since the rotor is magnetically symmetric any other convention would work just as well. The modeling equations simplify significantly by transforming stator and rotor variables to a dq reference frame; the reference frame position θ_d will be measured from the stator phase-a axis to the d-axis. Both AC synchronous motors and AC induction motors are modeled using dq variables, but in the former case the reference frame should rotate with the rotor while in the latter case the reference frame should rotate with the applied stator voltages; only then will constant-speed constant-torque operation correspond to constant voltages, constant currents and constant fluxes.



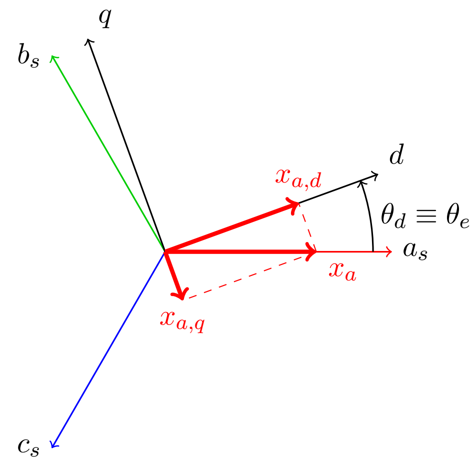
1.2 Plant Model

Appropriate models of the motor and the converter are required for controller design. The development of motor models begins with physical laws applied to the phase variables, but the initial steps are not shown here for sake of brevity. This lab will exploit a motor model expressed in terms of the so-called "synchronous" dq reference frame that rotates with the applied stator voltages. In this frame of reference, constant-speed operation would be achieved by applying DC voltages rather than AC voltages. Rotor speed differs from excitation speed due to the phenomenon known as slip; the rotor coils must experience a time-varying magnetic field in order to induce rotor flux, so slip is essential for this type of motor (torque production requires non-zero slip).

1.2.1 Motor Model in Synchronous-Frame Variables

The process of deriving the synchronous-frame motor model requires consideration of two coordinate transformations; one to project stator variables onto the dq-frame, and one to project rotor variables onto the dq-frame. Since the controller will interface only with the stator, just one of these two coordinate transformations is displayed in the figure below. Note that in this geometric visualization and in the corresponding transformation equations, the reference frame position θ_d

has been assigned the value of θ_e, the stator voltage excitation angle which will appear explicitly in the controller implementation equations that follow.



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

x_a is phase-a stator voltage (v_a) or phase-a stator 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

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

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

and the reverse change of variables is

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

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

The converter-sourced stator magnetic field could be described in terms of a stator current vector or a stator flux vector; similarly, the magnetically-induced rotor magnetic field could be described in terms of a rotor current vector or a rotor flux vector. In the model provided below, the choice has been made to use the stator current and rotor flux vectors for characterizing the state of the electromagnetic parts of the plant. The dq frame rotates at speed ω_e defined by

$$\omega_e = \frac{d\theta_e}{dt}. \tag{1}$$

After transformation to the dq frame, the mechanical subsystem of the plant is modeled by

$$\frac{d\theta_r}{dt} = \omega_r \tag{2}$$

$$J\frac{d\omega_r}{dt} = \gamma N \left(\psi_d i_q - \psi_q i_d\right) - F\omega_r, \tag{3}$$

the rotor flux subsystem of the plant is modeled by

$$\tau_r \frac{d\psi_d}{dt} = -\psi_d + N(\omega_e - \omega_r)\tau_r \psi_q + L_m i_d \tag{4}$$

$$\tau_r \frac{d\psi_q}{dt} = -\psi_q - N(\omega_e - \omega_r)\tau_r \psi_d + L_m i_q, \tag{5}$$

and the stator current subsystem of the plant is modeled by

$$\tau_s \frac{di_d}{dt} = -i_d + N\omega_e \tau_s i_q + \frac{1}{R}\left(\frac{\gamma}{\tau_r}\psi_d + \gamma N \omega_r \psi_q + v_d\right) \tag{6}$$

$$\tau_s \frac{di_q}{dt} = -i_q - N\omega_e \tau_s i_d + \frac{1}{R}\left(\frac{\gamma}{\tau_r}\psi_q - \gamma N \omega_r \psi_d + v_q\right), \tag{7}$$

where θ_r and ω_r denote rotor position and rotor speed, ψ_d and ψ_q denote rotor flux vector coordinates, i_d and i_q denote stator current vector coordinates, and v_d and v_q denote stator voltage vector coordinates. The constant parameters appearing in these state equations are inertia J, friction coefficient F, pole pairs N, mutual inductance L_m, dimensionless factor γ, rotor flux time constant τ_r, stator current time constant τ_s and effective resistance R. These last four parameters are related to more fundamental parameters as follows:

$$\gamma = \frac{L_m}{L_r}, \quad \tau_r = \frac{L_r}{R_r}, \quad \tau_s = \frac{L_s - \gamma^2 L_r}{R_s + \gamma^2 R_r}, \quad R = R_s + \gamma^2 R_r.$$

1.2.2 Reduced-Order Model for Controller Design

In each of our previous position control labs, we simplified the plant model by assuming quasi-steady-state behavior of the electromagnetic portions of the plant, before designing a controller. After simplification, the dynamic order of the plant model matched that of a purely mechanical system, thereby reducing the complexity of controller design. To exploit the same design concept in this lab, we must first examine which variables determine the electromagnetic torque. As indicated above, the interaction of the stator and rotor magnetic fields results in an electromagnetic torque proportional to $\psi_d i_q - \psi_q i_d$. This dependency on controller-influenced variables is far more complex than we have previously seen; for both the DC motor and the AC synchronous motor (expressed in the appropriate dq reference frame), electromagnetic torque depended only on a single scalar current variable. We prefer that same simplicity, so we will impose a field-orientation principle; the desired rotor flux will correspond to a positive constant value of ψ_d and a zero value of ψ_q, so that electromagnetic torque will be proportional to the scalar current component i_q.

For design purposes, we will make the following substitutions in (4)–(7): we will set $\psi_d = \Psi_d$ and $\psi_q = 0$ to impose field orientation; we will set $\tau_s = 0$ to neglect a small time constant; and we will set time derivatives to zero to neglect electromagnetic dynamics. After these substitutions, it follows from (4) and (6) that v_d will be constant and equal to

$$v_d = \left(\frac{R_s}{L_m}\right)\Psi_d, \tag{8}$$

it follows from (7) that v_q will influence torque according to

$$i_q = \frac{v_q - K\omega_r}{R}, \quad K = \gamma N \Psi_d, \tag{9}$$

and it follows from (5) and (7) that ω_e will be determined by

$$\omega_e = \left(\frac{R_s}{R}\right)\omega_r + \left(\frac{\gamma^2 R_r}{R}\right)\frac{v_q}{K}. \tag{10}$$

Substitution of (9) into (2)–(3) along with

$$v_q = u \tag{11}$$

results in

$$\frac{d\theta_r}{dt} = \omega_r \tag{12}$$

$$\frac{d\omega_r}{dt} = -\alpha\omega_r + \beta u \tag{13}$$

where

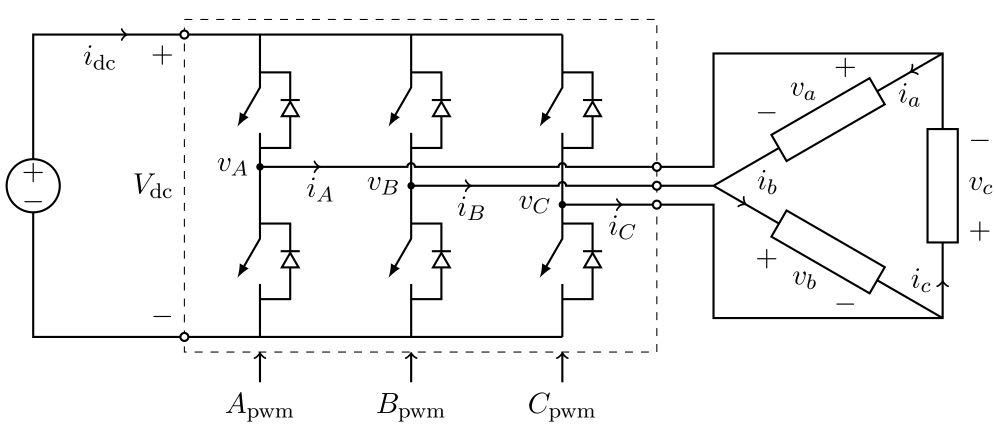
$$\alpha = \frac{K^2 + FR}{JR}, \quad \beta = \frac{K}{JR}.$$

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

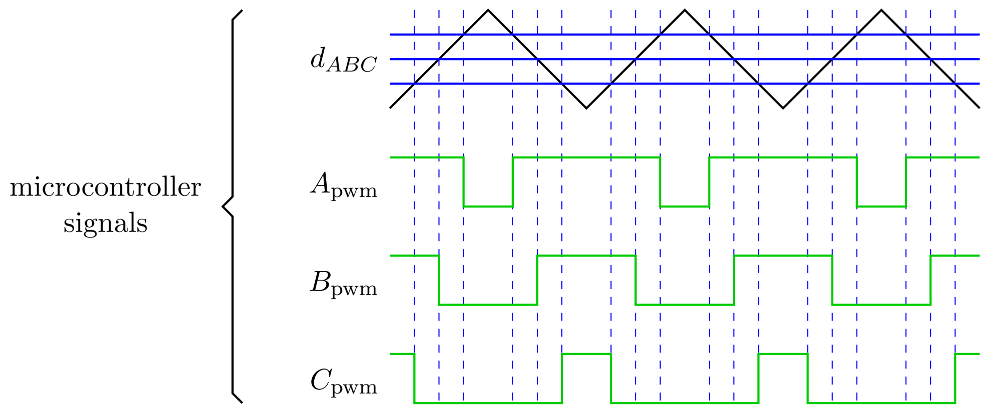
To summarize our design thus far, recall that our control variables are v_d, v_q and θ_e. We have determined a suitable constant value for v_d, the integral controller output will be assigned to v_q, and we will integrate the constraint on ω_e to determine θ_e.

1.2.3 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:

$$\bar{v}_A[k] = \frac{1}{T}\int_{kT}^{kT+T} v_A(t)\,dt = d_A[k]V_{\text{dc}} \tag{14}$$

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

$$\bar{v}_C[k] = \frac{1}{T}\int_{kT}^{kT+T} v_C(t)\,dt = d_C[k]V_{\text{dc}}. \tag{16}$$

These converter leg voltages determine the motor phase voltages according to

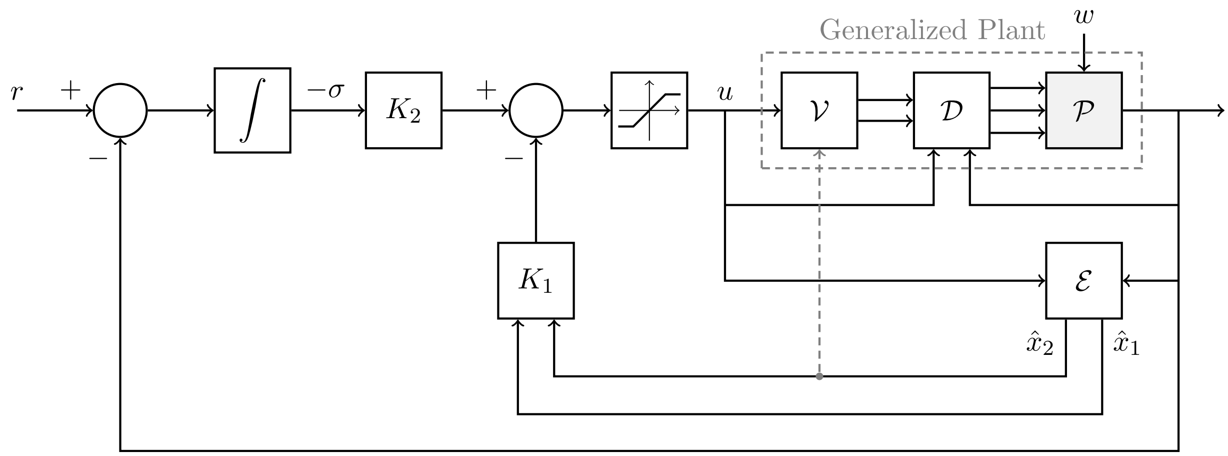
$$v_a = v_A - v_B$$
$$v_b = v_B - v_C$$
$$v_c = v_C - v_A.$$

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

$$i_A = i_a - i_c$$
$$i_B = i_b - i_a$$
$$i_C = i_c - i_b.$$

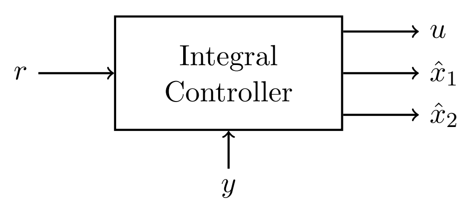
1.3 Design of the Position Control System

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 internal stability and the command following function; its structure matches that used for the DC motor position control system of Labs 6–7 and the AC synchronous motor position control system of Lab 8 (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 (12)–(13), the design model of the AC induction 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 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

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

On the microcontroller, the controller output is initially computed as

$$u^* = -K_{11}\hat{x}_1 - K_{12}\hat{x}_2 - K_2\sigma, \tag{17}$$

then saturated if necessary according to[1]

$$u = \begin{cases} +U_{\max} & \text{, if } u^* > +U_{\max} \\ -U_{\max} & \text{, if } u^* < -U_{\max} \\ u^* & \text{, otherwise} \end{cases} \tag{18}$$

and the controller state variables are updated using forward Euler integration:

$$\hat{x}_1 \leftarrow \hat{x}_1 + T\left(\hat{x}_2 - L_1\left(\hat{x}_1 - y\right)\right) \tag{19}$$

$$\hat{x}_2 \leftarrow \hat{x}_2 + T\left(-\alpha\hat{x}_2 + \beta u - L_2\left(\hat{x}_1 - y\right)\right) \tag{20}$$

$$\sigma \leftarrow \sigma + T\left(y - r\right). \tag{21}$$

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

$$L_1 = 2\lambda_e - \alpha, \quad L_2 = \lambda_e^2 - 2\alpha\lambda_e + \alpha^2, \tag{22}$$

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

$$K_{11} = \frac{1}{\beta}3\lambda_r^2, \quad K_{12} = \frac{1}{\beta}\left(3\lambda_r - \alpha\right), \quad K_2 = \frac{1}{\beta}\lambda_r^3. \tag{23}$$

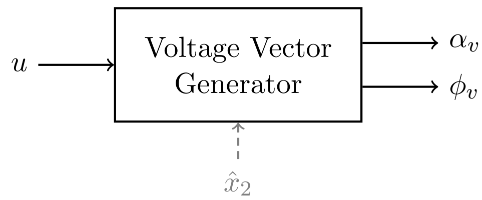
These gain formulas match those of Labs 6–8, since the design model formulas on which they are based match those of Labs 6–8. Thus far, DC motor control, AC synchronous motor control and AC induction motor control are similar, which is a consequence of the fact that an electromagnetic torque accelerates a rotor and load inertia in all 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 synchronous-frame q-axis—and optionally uses the estimated rotor speed to determine the desired component of voltage along the synchronous-frame d-axis, according to a flux weakening[2] algorithm—to ultimately generate the polar coordinate representation of the desired voltage vector:

[1]Since $v_d = V_d$ and $v_q = u$, an upper bound on the magnitude of u would be $U_{\max} = \sqrt{V_{\max}^2 - V_d^2}$.

[2]Flux weakening (not used in this lab) is a strategy for reducing the level of d-axis rotor flux at speeds beyond base speed, thereby increasing the range of speeds available in the presence of voltage limits.



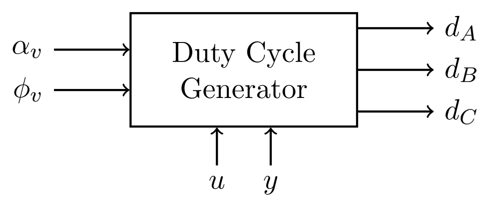
For operation that does not include flux weakening, the cartesian coordinate voltage vector is specified in (8) and (11), and the corresponding polar coordinates are computed according to

$$\alpha_v = \sqrt{v_d^2 + v_q^2}, \quad \phi_v = \operatorname{atan2}(v_q, v_d). \tag{24}$$

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 synchronous-frame voltage vector—and uses the measured rotor position along with the commanded torque-producing voltage to determine the three-phase duty cycles that will be required to establish the desired synchronous-frame voltage vector:



The desired voltage vector has been computed in synchronous-frame coordinates, but it must be implemented using the stator windings in stator-frame coordinates, so an excitation-angle-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

$$v_A = \tfrac{1}{2}V_{\mathrm{dc}} + \sqrt{\tfrac{2}{9}}\,\alpha_v \cos\left(N\theta_e + \phi_v - \tfrac{\pi}{6}\right) \tag{25}$$

$$v_B = \tfrac{1}{2}V_{\mathrm{dc}} + \sqrt{\tfrac{2}{9}}\,\alpha_v \cos\left(N\theta_e + \phi_v - \tfrac{\pi}{6} - \tfrac{2\pi}{3}\right) \tag{26}$$

$$v_C = \tfrac{1}{2}V_{\mathrm{dc}} + \sqrt{\tfrac{2}{9}}\,\alpha_v \cos\left(N\theta_e + \phi_v - \tfrac{\pi}{6} + \tfrac{2\pi}{3}\right) \tag{27}$$

where, by integrating (10), we obtain[3]

$$\theta_e = \left(\frac{R_s}{R}\right) y + \rho, \quad \dot{\rho} = \left(\frac{\gamma^2 R_r}{R}\right)\frac{u}{K}. \tag{28}$$

Using forward Euler integration, the update for ρ would be

$$\rho \leftarrow \rho + T\left(\frac{\gamma^2 R_r}{R}\right)\frac{u}{K}. \tag{29}$$

[3] If y and ρ are expected to reach large values, then modulus operations should be introduced so that the computed value of θ_e remains small; otherwise, the cosine function may not evaluate correctly on a microcontroller.

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

$$d_A = \frac{v_A}{V_{\mathrm{dc}}}, \quad d_B = \frac{v_B}{V_{\mathrm{dc}}}, \quad d_C = \frac{v_C}{V_{\mathrm{dc}}} \tag{30}$$

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

1.3.4 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.

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 simulation code posted on tsquare represents a virtual implementation of the position control method defined by the state-space integral controller (17)–(21), feedback gain assignment (22)–(23), voltage vector generator (24) and duty cycle generator (25)–(30), applied to the physical system expressed in the stator reference frame.

Zoom in on v_{ABC} and i_{ABC} to see the three-phase AC waveforms in detail, and note how u exhibits a DC waveform. The simulation code does not include integrator anti-windup logic or reference command shaping, but these features may be easily added if desired.

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. 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) How long does it take to reach steady state, and how does λ_r influence this?
   (b) What are the steady-state voltages at the converter legs?
   (c) What are the peak voltages at the converter legs?
   (d) What are the steady-state currents at the converter legs?
   (e) What are the peak currents at the converter legs?
   (f) What is the steady-state magnitude of the integral controller's output signal u?
   (g) What is the peak magnitude of the integral controller's output signal u?

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

The parameter values needed to implement the controller on the experimental system may be found in the simulation code. The attached encoder provides 4000 counts per revolution. The encoder cable yellow wire should be connected to the J4 pin identified by a dot.

2.2.1 Three-Phase Voltage Generation

Develop code that implements three-phase voltage generation as defined by (14)–(16), 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 (17)–(21) with feedback gain assignment (22)–(23), the voltage vector generator (24) and the duty cycle generator (25)–(30). Initialize the PWM module such that all three PWM pins are commanding 50% duty cycles prior to activating the power stage. Use the protection measures described in §1.3.4 and the parameter values listed in the simulation code. The reference command r should alternate abruptly between 0 rad and 2π rad with each position maintained for 1 second. 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 #9: Position Control with AC Induction Motors

Instructor Verification Page

Lab Section	Begin Date	End Date
L01, L02	November 14	November 28
L03, L04	November 16	November 30

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:** ______________________