Chapter 20: Permanent Magnet Brushed DC Motor Applications

20.1: Introduction

In Chapter 19, our discussion was focused on motor construction and on characterizing motor behavior and performance. These are essential to determining what specifications are required for a motor in a given application, and in selecting the right motor for each occasion, but it doesn’t describe much about how to integrate a motor into a system once it has been selected, nor how to control it once it is integrated. This chapter focuses entirely on the practical issues related to designing with permanent magnet brushed DC motors and successfully applying them. The practical information in the following discussion is just as critical as the more analytical discussion in Chapter 19, and will make the difference between a successful system design and a well-specified but unsuccessful one.

In this chapter, you will learn:

1) Non-steady-state characteristics of motors: how they respond when they are switched on and off
2) What inductive kickback is, and how to manage it
3) How to implement uni-directional and bi-directional control of brushed DC motors
4) What components are useful for designing motor control circuitry, and how to select them
5) How to implement speed control of motors using pulse width modulation (PWM).

20.2: Inductive Kickback

A very simple implementation of a permanent magnet brushed DC motor that can be switched on and off is one in which one lead is connected to a power source and the other lead is connected to a switch to ground, as in Figure 20.1(a). This has the disadvantage of requiring physical interaction with the switch to power the motor on and off. In order to switch the motor electronically, you might be tempted to replace the switch with a transistor (one of many possible approaches), as shown in Figure 20.1(b).
When the switch is closed or the transistor is “ON”, current flows from the power source, through the motor, through the switch or transistor, and finally to ground. When the switch is open or the transistor is “OFF,” no current may flow. Unfortunately, the approach shown in Figure 20.1(b) has a terrible fatal flaw, after a few cycles of switching the motor on and off, the transistor will most likely be destroyed by the effects of inductive kickback – the inevitable effect of trying to abruptly turn off the current flowing in an inductor, such as those made up by the motor’s coil windings.

One of the primary practical issues that must be addressed when designing drive circuitry for permanent magnet brushed DC motors is a phenomenon known as “inductive kickback.” Inductive kickback is a large voltage spike that occurs whenever the current flowing through an inductor is abruptly switched off, and can result in the destruction of drive circuit components if not handled correctly. Fortunately, inductive kickback can be managed effectively with the addition of a few inexpensive components.
Recall that the voltage potential across an inductor is given by:

\[ V = L \frac{\partial I}{\partial t} \quad [\text{Volts}] \quad \text{Eq. 20.1} \]

The current flowing through an inductor cannot be instantaneously changed. Depending on the magnitude of the inductance \( L \), an abrupt increase in the voltage applied across the leads of an inductor causes the current \( I \) to ramp up in an exponential fashion, as shown in the time period between \( 0 \mu s \) and about \( 400 \mu s \) in Figure 20.2. In this portion of the figure, the applied voltage is initially 0V, until time \( t_0 = 0 \mu s \), when the voltage is abruptly switched to 12V. The resulting current doesn’t switch on immediately in response to the change in voltage, rather it rises exponentially to its final value which is determined by Ohm’s law \( (V = I \cdot R_{\text{inductor}}) \). The time constant characterizing the current rise time in an inductor is:

\[ \tau = \frac{L}{R} \quad [\text{seconds}] \quad \text{Eq. 20.2} \]
During the relatively lengthy time (for an electronic timescale, that is) that the current is rising, the inductor is establishing a magnetic field. This accompanying magnetic field is what affords the inductor its characteristic behavior: changes in current are resisted by the magnetic field as it adjusts to the changing current. Energy is traded back and forth between the current flow and the magnetic field while changes occur. If the current flow increases, the energy in the magnetic field must also increase, and energy will be drawn from the current flow to increase the magnetic field energy. If the current flow decreases, energy from the magnetic field must also decrease, and energy will be contributed from the field to the current flow. In this way, *inductors act in such a way that they resist changes in current flow, whether that change is an increase or a decrease.*

Figure 20.2: Current Response in an Inductor for a Step Increase in Applied Voltage at $t = 0 \mu s$ and a Step Decrease in Applied Voltage at $t = 400 \mu s$
In the case where the applied voltage across an inductor is abruptly switched off (which happens every time a motor is turned off) the magnetic field established by an inductor (e.g., the motor’s coil windings) will contribute its stored energy in an effort to force current to continue to flow. The energy stored in the inductor’s magnetic field is given by the expression:

\[ E = \frac{1}{2} I^2 L \]  

Eq. 20.3

For a brief time (a few microseconds, depending on the coil’s inductance and resistance), the current flowing in the inductor will continue, first at a level equal to the current that was flowing immediately prior to the voltage being switched off, and then decaying rapidly as the energy from the magnetic field is exhausted. The rapid decrease in current shown in Figure 20.2 at \( t = 400 \mu s \) illustrates the current decay characteristics that result when the applied voltage is switched off. The time required for the magnetic field to collapse is short relative to the time required to establish the magnetic field (that is, the rise time of the current), but it is significant.

In order to maintain this current flow in the absence of any external applied voltage, the inductor establishes a very large voltage differential across its terminals, using the energy from the magnetic field. For a brief time (equal to the amount of time it takes for the current to cease flowing, typically a few microseconds), the voltage across the inductor will spike to very high levels, often reaching well above 1kV. Figure 20.3 shows the resulting voltage response. This effect is known as “inductive kickback,” “kickback voltage” or “flyback voltage.”

The energy stored in the inductor’s magnetic field must go somewhere, either by being bled off through some leakage mechanism in one of the system components or through arcing. In all cases, the energy
won't go anywhere until there is current flow. As a result, the maximum voltage achieved by the inductive kickback spike is only limited by the first opportunity for some current to flow.

![Graph showing inductive kickback voltage](image)

**Figure 20.3: Inductive Kickback Voltage that Results when the Applied Voltage Across an Inductor is Abruptly Switched Off**

Inductive kickback, which results in voltage spikes that may exceed 1kV, can exceed the specifications of many of the electronic components used to control motors. Even in a very simple configuration, such as the one shown in Figure 20.1(a), the inductive kickback can cause problems, since most manual switches are not rated for more than a few hundred volts across their terminals. Most switches will function well for several cycles, but under these extreme circumstances arcing will occur which will eventually damage the switch. If, however, a typical field-effect transistor (FET) or a bipolar junction transistor (BJT) is used to control a motor, such as the circuit shown in Figure 20.1(b), the transistor will soon be destroyed by
exposure to the inductive kickback, since most silicon-based components aren’t rated to withstand more than 30V-50V across their terminals.

Fortunately, dealing with inductive kickback is straightforward. In general, the strategy is to add components to the control circuit that provide a low-voltage current path, thereby reducing, or “snubbing,” the magnitude of the kickback voltage to safe levels.

The most straightforward means of reducing inductive kickback to safe levels is to place a diode across the terminals of the motor in a reverse-biased orientation. This limits the voltage spike caused by inductive kickback to the forward voltage drop of the diode (about 0.6V) plus the positive voltage supply (+V). A diode used in this configuration is called a “kickback diode,” “flyback diode,” or “diode snubber.” Figure 20.4 shows a kickback diode in place across the terminals of a DC motor.

Figure 20.4: DC Motor Circuit with Kickback Diode
For normal operation of the motor, the diode is reverse-biased and no current will flow through the diode. The diode only conducts when the collector voltage of the transistor exceeds the diode’s forward voltage drop of 0.6V, which occurs only during inductive kickback.

With the kickback diode in place, the voltage spike is effectively limited to the positive voltage supply level plus the forward voltage drop of the diode. Figure 20.5 shows a simulation of the circuit shown in Figure 20.4, with the kickback diode in place. Note that the maximum voltage calculated is approximately 13V. This voltage level is well within the capabilities of all circuit components.

![Collector Voltage with Diode Snubber](image)

Figure 20.5: Kickback Diode Snubbing Inductive Kickback

When selecting a kickback diode, two characteristics should be considered. First, the diode should be able to switch very rapidly from the forward-biased condition, when it is conducting current (as when inductive
kickback is occurring and a large voltage spike is developing across the terminals of the motor), to the case where it is reverse-biased and blocking the flow of conducting (as occurs when the motor is powered up and operating normally). This specification is called the “reverse recovery time,” and is usually given the symbol $t_{rr}$. Reverse recovery time specifications vary widely for different types of diodes. Diodes that are appropriate for this application will often be labeled “fast recovery” or even “ultra-fast recovery.” The faster, the better – diodes with reverse recovery times of a few hundreds nanoseconds or less are good candidates. The second important diode specification to consider is the maximum allowable current flow. The maximum current that the diode will experience under normal conditions is the maximum current through the motor: the stall current. A very conservative choice would be to specify a diode that can tolerate the stall current continuously. In practice, this is unnecessarily conservative: even if the motor is switched on and off very rapidly (which happens any time pulse width modulation, or “PWM,” speed control is used, as discussed below in Section 20.4), the diode will only experience these peak current levels a small fraction of the time. A diode that has a peak intermittent current specification that exceeds a motor’s stall current will usually suffice.

There are several alternative methods of snubbing the inductive kickback. A single kickback diode is the simplest solution, but higher overall system performance can be achieved using slightly more complex arrangements.

One avenue of increasing system performance is to focus on reducing the time required to dissipate the energy stored in the inductor, which is the root cause of the inductive kickback. In the case of the simple kickback diode as illustrated in Figures 20.4 and 20.5, the duration that the kickback diode is forward-biased and conducting is about 200µs. During this time, the energy from the collapsing magnetic field causes current to flow in a loop that includes the motor and the kickback diode. This energy is dissipated both in the motor, since torque is generated by the current flow and $I^2R$ losses occur in the coil windings, and as $V*I$ losses across the diode. This approach is straightforward and gets the job done, but it results in
maximum voltage levels and power dissipation levels that are well below the maximum allowable values for most components. An expression for the dissipation of this energy would be:

$$E_{\text{STORED}} = \int Power(t) \, dt \quad [\text{Joules}] \quad \text{Eq. 20.4}$$

Restating the expression with the assumption that the power dissipation has an average value:

$$E_{\text{STORED}} = Average \, Power \times \Delta T \quad [\text{Joules}] \quad \text{Eq. 20.5}$$

Minimizing $\Delta T$, the time required to dissipate the energy, while ensuring that the kickback voltage never rises above the specified maximum levels for any of the system components, and that the average power dissipation limits of all components is never exceeded, will result in improved system response. By limiting the kickback voltage and power dissipation to the highest possible safe values, the duration required to dissipate the energy will be minimized and this aspect of system performance will be optimal.

One means of implementing this strategy is to add a resistor to the diode in the kickback loop, as shown in Figure 20.6(a). The additional resistance increases the energy dissipation rate both by increasing the resistance $R$ in the $I^2R$ loss term, and by increasing the voltage $V$ in the $V*I$ term. Figure 20.6(b) shows the results of a simulation of this approach, and predicts that the kickback spike will be dissipated in about 100$\mu$s – about a factor of 2 faster than a diode alone.
Since the maximum voltage observed using this approach is higher than that seen with a simple kickback diode, care must be taken when selecting the value of the resistor in the kickback loop so that the maximum voltage specifications of the other components in the system are not exceeded. Also, the power rating of the resistor must be adequate for the task.

Further performance gains can be realized by replacing the resistor illustrated in Figure 20.6 with a zener diode arranged as shown in Figure 20.7(a). Note that it is forward biased with respect to the overall power supply, but that since the standard diode is reverse-biased, no current will flow in the kickback loop under normal operating conditions. However, when the kickback voltage exceeds the zener’s reverse breakdown voltage plus the forward drop of the standard diode, both diodes will begin to conduct. The zener diode will hold the voltage drop across its anode and cathode to the reverse breakdown voltage (12V for the zener diode used in the simulation shown in Figure 20.7(b)), and the standard diode in forward conduction will have a voltage drop of approximately 0.6V. One benefit of this approach is that the diodes allow the designer to very accurately and confidently set the maximum voltage that the kickback surge will attain. Zener diodes are available with a very large variety of reverse-breakdown voltages, ranging from 2V to...
200V, so there are many options when setting the limit for the kickback voltage. Standard and zener diodes are not ohmic devices – that is, they have no “resistance,” only an inherent voltage drop when conducting. Because of this, the energy stored in the inductor’s magnetic field is dissipated as $I^2R$ losses across the motor windings, as $V_{\text{DIODE}}I$ losses across the diode, and as $V_{\text{ZENER}}I$ losses across the zener diode. As shown in Figure 20.7(b), this approach results in the dissipation of the energy in approximately 50µs, or about 4 times faster than the case for the snubber diode alone.

![Figure 20.7(a)](image1)  
Kickback Diode w/ Zener

![Figure 20.7(b)](image2)  
Resulting Inductive Kickback

The selection criteria for the standard diode and the zener diode in this application are essentially the same as the criteria for the configuration with the standard diode alone. For the standard diode in this case, select a fast recovery diode (one with a short reverse recovery time specification) and ensure that it can safely handle the motor stall current. The typical choice is to select a diode with a peak, intermittent forward current specification equal to or greater than the motor’s stall current. For the zener diode, ensure that the power dissipated when the diode is conducting in the reverse-breakdown mode is less than the diode’s rated power dissipation. This process is a little more involved for zener diodes than for standard diodes, as zener diodes that have peak intermittent current ratings in excess of typical motor stall currents are not as easy to
find or as inexpensive. For zener diodes, a more detailed analysis of average power dissipation is required to ensure that its power dissipation specifications are not exceeded.

Recall from Eq. 20.3 that the energy stored by an inductor is $E = \frac{1}{2} I^2 L$. From this, we can develop an expression for the average power dissipation resulting from kickback that takes into account how frequently a motor is switched on and off. In practice, motors may be switched on and off very frequently, as when pulse width modulation (PWM) drive is used. PWM is discussed below in Section 20.4.

$$P = \left( \frac{1}{2} I_{\text{STALL}}^2 L \right) \cdot \text{Frequency} \quad [\text{Watts}] \quad \text{Eq. 20.6}$$

Note that this is the average power that will be dissipated during kickback, and that the dissipation will be distributed among the motor, the standard diode, and the zener diode. A zener diode that is rated for average power dissipation in excess of the levels calculated with Eq. 20.6 in a specific application will be a good choice, with a comfortable margin of safety.

A zener diode can also be used to directly limit the voltage across the transistor. The example circuit used throughout this discussion has used an NPN bipolar junction transistor (BJT) to control a DC motor. There is a limit to the voltage that can be applied across the BJT’s emitter (the terminal connected to ground in Figure 20.8(a)) and collector (the terminal connected to the DC motor in Figure 20.8(a)) without damaging the component. Most NPN BJT’s can tolerate a maximum “collector-emitter voltage” of 30V to 60V. When the motor is on and running steady-state, the collector-emitter voltage will be very small (about 0.8V, depending on the specifications of the BJT). When the motor is off and in steady-state, the collector-emitter voltage will be the power supply voltage, +V. When the motor is switched off and inductive kickback is occurring, much higher collector-emitter voltages will be observed, and the BJT is likely to be
destroyed. Placing a zener diode across the collector and emitter of the transistor in a reverse biased orientation will limit the voltage across the transistor to the zener’s reverse breakdown voltage. In Figure 20.8(a), a zener diode with a reverse breakdown voltage of 34V has been selected. As before, care must be taken when selecting the zener diode to ensure that it is capable of dissipating the heat that will be generated when it is conducting during kickback. To ensure that a particular zener diode is suitable, Eq. 20.6 may be applied to calculate the average power dissipation requirements. Select zener diodes with power rating specifications above the calculated average power dissipation.

A drawback to this approach, despite its simplicity, is that whenever inductive kickback occurs, the power supply ground will be required to absorb a short but significant current spike. Depending on the quality of the power supply and the wiring used to connect the motor and zener, this could cause substantial noise in the power supply’s ground. While it’s not likely to cause a problem for the motor or the transistor, other components connected to the power supply, such as analog circuitry or microcontrollers, may not tolerate the results well. (See Chapter 16: Noise, Grounding & Isolation for a discussion of these issues and remedies.)
20.2.1: Inductive Kickback Summary

Turning off the current flow through a motor as rapidly as possible will result in improved performance, however the complexity and cost of the design must also be considered when deciding how to deal with inductive kickback. Figure 20.9 compares the time required to stop the current flow using each of the techniques discussed above. The goal is to get as close as possible to the case where no kickback snubbing is performed, which is the fastest, without allowing the kickback to destroy any circuit components. Each of the kickback snubbing techniques presented will protect the circuit components, but each technique has a different effect on the turn-off time of the kickback current, and differs in cost and complexity. The diode-only approach has the benefit of simplicity and low cost, but results in the slowest turn-off time and therefore the least optimal motor performance. This approach is perfectly adequate for a wide variety of applications, however improvements are possible. Adding a zener diode to the standard diode across the motor leads, for example, results in the fastest current turn-off time and therefore the most desirable motor performance. The other configurations discussed (the reverse-biased diode plus a resistor in a loop around the motor leads, and the zener diode across the collector and emitter of the switching transistor) result in current turn-off times that lie between these extremes. The best method of snubbing inductive kickback depends on the application, but some method must always be used.
20.3: Bi-Directional Control of Motors

Any of the circuits above that include kickback suppression are good choices for uni-directional control of DC motors, but what about applications that will require a motor to spin in both directions (clockwise – CW, and counter-clockwise – CCW)?

For the circuits shown in Figures 20.4, 20.6, 20.7, and 20.8, current can only flow from the positive terminal of the power source to ground, so the motor can only turn in one direction. Enabling a motor to turn either CW or CCW requires the capability to cause current to flow through the motor in either direction. A different, more complex drive circuit will be required to accomplish this. One possibility is to make use of a “dual” power supply, that is, one with both a positive voltage supply and a negative voltage supply. By switching one or the other on, current can flow through the motor in either direction. Figure 20.10 illustrates this configuration. Note that kickback diode protection must be included for each of the transistors, preventing inductive kickback from exceeding the maximum collector-emitter voltage for each.
For this circuit, when the voltage at the Input node has a value of about +0.6V or higher, the NPN transistor will be “ON” and conducting, and current will flow from the +V terminal through the motor to ground. The PNP transistor will be “OFF” under these conditions.

If the input voltage has a value of -0.6V or lower, the PNP transistor will be “ON” and the NPN transistor will be “OFF.” Under these conditions, current will flow from ground (0V) to the negative supply terminal (-V), which has a voltage potential lower than that of ground.

For this example, current flows from left to right across the motor if the NPN transistor is on, and from right to left across the motor if the PNP transistor is on. This is one way to accomplish bi-directional current flow, but in practice it is not common to have a high current negative voltage supply available. It would be more practical to design a circuit capable of causing current to flow in either direction through a motor using only a single positive power supply. The most common way of accomplishing this is through the use of a network of transistors, arranged in what is known as an “H-bridge” configuration. The shape
of the circuit resembles an “H,” which gives rise to the distinctive name. Figure 20.11 illustrates an H-bridge.

![Diagram of an H-bridge circuit](image)

Figure 20.11: A typical H-bridge circuit, including kickback diodes to protect against inductive kickback.

Recall that the objective of this circuit is to enable current to flow through the motor in either direction (left-to-right or right-to-left) using only a single positive power supply (shown in Figure 20.11 as +V). The diodes that surround the motor are in place to protect the transistors from the inductive kickback, as described above in Section 20.2. Since current may flow in either direction across the motor, the high voltage inductive kickback can occur on either motor terminal, and must be snubbed to prevent the destruction of the transistors. Though any of the snubbing techniques discussed above are applicable with H-bridges, by far the most common practice is the use of standard diodes.
Examining Figure 20.12(a) first, we see that we have established the situation where the left PNP transistor is on, the right PNP transistor is off, the left NPN transistor is off, and the right NPN transistor is on. For these conditions, current can flow from the positive voltage terminal of the power supply, through the left PNP transistor, through the motor from left to right, and finally through the right NPN transistor to ground.

In Figure 20.12(b), the opposite conditions have been established: by reversing the state of each of the transistors, current can now flow from the positive terminal of the power supply, through the PNP transistor on the right side, through the motor (from right to left this time) and finally trough the left NPN transistor to ground.

Other possibilities exist for combinations of transistors states, of course. For instance, if both transistors on either the left side or the right side of the “H” are switched on, current will flow without any resistance, directly from power to ground. In this case, a short circuit is effectively established between power and ground, since there is no load or other component in the path to resist the flow of current. This results in
very high levels of current flow. If it is sustained for any significant length of time, one of the components involved is very likely to be destroyed by the excessive current levels. This unrestrained current is referred to as “shoot-through current.” Shoot-through current can occur under normal circumstances, such as while an H-bridge is in the process of switching from one desirable configuration to another (for instance, when reversing the direction of the current flow through a motor). This may occur if the transistors are not switched simultaneously, or if they don’t have identical turn-on and turn-off times. Under these conditions, brief bursts of shoot-through current may be experienced. Shoot-through current happens, but it is NOT desirable. Care should be taken to ensure that it is minimized if it is unavoidable, and avoided where possible.

![Figure 20.13: Shoot-through current in an H-bridge, where current flows unimpeded from power to ground, usually with disastrous effects](image-url)

Another, more useful, configuration can be used to “dynamically brake,” or resist the rotation of, a motor. When the leads of a spinning motor are shorted together, the current that is generated by the back-EMF (recall that the spinning motor is acting as a generator) causes current to flow in a loop, as shown in Figure 20.14. The direction of current flow during dynamic braking depends on the direction of the motor’s
rotation. Current will flow from one of the motor’s terminals, through one of the kickback diodes, around
to one of the transistors that are in the “on” state, and finally around to the motor’s other terminal. The
internal resistance of the motor coils, the forward voltage drop across the diode, and the saturation voltage
of the transistors serve as the load for this current, and the electrical power being generated by the rotating
motor will be dissipated as heat. In the absence of a load caused by dynamic braking, the only forces
acting to slow the rotation of a motor are the frictional resistance of the motor bearings and other
mechanical systems coupled to the motor. Dynamic braking can be performed with an H-bridge by turning
on both the upper transistors while turning off both the lower transistors, or turning off both the upper
transistors while turning on both the lower transistors. This effectively shorts both leads of the motor
together and results in dynamic braking. This is useful when switching a motor off in circumstances where
it is desirable for the motor to stop rotating more quickly than would otherwise occur if it were simply
allowed to slowly decelerate and come to a full stop.
Figure 20.14: An H-bridge configured to short both leads of motor together, creating dynamic braking. Depending on the direction of rotation of the motor, the current flow will be either clockwise or counterclockwise, as shown.

Bi-directional switching of current through a load, such as a motor, is such a common need in electronic design that there are several integrated circuits on the market that implement H-bridges in a single package, so that designers don’t have to wire one up out of discrete components every time the need arises. H-bridges are available in a wide variety of current drive capabilities, package sizes and features. They all perform the same basic task, however: switching current through a load in two directions. Some integrated solutions you may wish to investigate that are the Texas Instruments SN754410NE, the ST L293, the ST L298, and the National Semiconductor LMD18200, all of which are discussed in detail in Chapter 13: Digital Outputs and Power Drivers.
Example 20.1:

Design drive circuitry based on an L293B (manufactured by ST – SGS Thomson) H-bridge integrated circuit that is capable of controlling a Maxon RE-13 (Part #: 118638) motor using a 12V power supply. The motor has terminal resistance $R = 14.1 \, \Omega$ and coil inductance $0.48 \, mH$. It will be switched on and off at a maximum rate of $1kHz$. To protect the circuit from excessive inductive kickback voltages, use a standard diode in combination with a zener diode. Ensure that the specifications for all components are met by the design.

In order to satisfy the constraints, the L293B must be capable of switching the maximum current possible for this motor powered at 12V. Since the terminal resistance is given, we can solve for the stall current ($\omega = 0$):

$$I_{STALL} = \frac{V}{R} = \frac{12V}{14.1\, \Omega} = 0.851 \, A$$

The specifications for the L293B show that it is capable of handling up to 1A per channel. Since the stall current is the maximum current the system will experience, the L293B will be able to safely handle controlling this motor.
The next issue to address is the inductive kickback. The problem specifies that a design using both a standard diode and a zener diode is required. Beginning with the standard diode, we must identify a candidate that has a suitable reverse recovery time specification and appropriate current capabilities. A diode that is very commonly employed as a kickback diode is the 1N4935, which is often identified as a “fast recovery diode” in the list of features found in its datasheets. Recall that should have a reverse recover time specification of a few hundred nanoseconds or less, and the 1N4935 has \( t_{rr} < 200 \text{ ns} \). The requirement for small reverse recovery time has been met. Recall also that good design practice dictates that the diode’s specification for peak intermittent current should be above the maximum current that the circuit will encounter, which is the stall current calculated above (0.851 A). The peak intermittent current specification for the 1N4935 is 10A, which is more than 10 times the stall current expected for this circuit. The 1N4935 is therefore a good choice.

Next, we need to choose a zener diode. We need to select both an appropriate reverse breakdown voltage (so we can set the maximum voltage that the kickback will be allowed to reach), and ensure that the power dissipation requirements are met.
Unfortunately, the specifications for the L293B do not call out the highest allowable voltage that the outputs can withstand during inductive kickback. This is a very common issue for designers: a specification we wish to design to is not given. In these cases, we need to be able to pick a reasonable limit based on the information available. One particularly helpful specification is the limits of the allowable voltages that may be used to power the chip, 36V. Given that the chip MUST be able to survive having 36V connected across both logic power (VSS), motor power (VS) and ground, then 36V should be a reasonable limit for kickback voltage as well. We will choose to limit the kickback voltage to ~33V, which is a few volts lower than 36V. If we use a zener diode with a 20V reverse breakdown voltage in combination with a standard diode that has a forward voltage drop of 0.6V-1V and the 12V power supply specified in the problem statement, the kickback voltage will be limited to approximately 33V (20V+1V+12V = 33V).

Finally, we must ensure that the zener diode can safely dissipate the power generated during kickback. For this analysis, we can use Eq. 20.6:

\[ P = \left( \frac{1}{2} I_{\text{STALL}}^2 L \right) \cdot \text{Frequency} \]

\[ P = \left( \frac{1}{2} \right) (0.851 A)^2 (0.48 mH)(1 kHz) = 0.174 W \]

This indicates that any zener diode rated above 0.174 W will be sufficient. ½-Watt zener diodes are very common and inexpensive (typically $0.10 - $0.25 each), so we’ll chose one of these. The 1N5280B is a ½-Watt, 20V zener diode, and is a good choice for this application.

At this point, all of the components have been selected, and the schematic may be drawn:
The inputs “Half1” and “Half2” each control one side, or leg, of the H-bridge. Each side of the H-bridge is called a “half bridge.” A logic-level high to an input (such as 1A) connects the corresponding output (1Y) to the motor supply voltage, minus a voltage drop due to losses (called the “saturation voltage”). This drop is $V_{CE\text{satH}}$ (saturation voltage, high) and is specified to have a value of about 1.4V in the L293B datasheet. A logic-level low to an input connects the corresponding output to ground, again with losses due to the saturation voltage across the active switch. This is the low side saturation voltage ($V_{CE\text{satL}}$) and is specified to be about 1.2V in the datasheet. Setting Half1 to a logic-level high and Half2 to a logic level low will drive current through the motor in one direction and cause the motor to turn. Reversing the inputs drives the current in the opposite direction, so the motor also turns the other direction. Setting Half1 and Half2 both to logic-level low at the same time (or both to logic-level high) connects the terminals of the motor together and causes dynamic braking. The input called “Enable” controls current flow through the half-bridges. Setting the Enable pin low inhibits current flow and turns the motor off, allowing it to coast to a stop.

### 20.4: Speed Control with Pulse Width Modulation (PWM)

Another essential aspect of controlling motors is the ability to change the speed of rotation and the amount of torque produced. One simple but usually impractical means of achieving this is to adjust the supply.
voltage up and down as required. However it is more effective and common to use the concept of “pulse width modulation,” or “PWM.” This is a control technique where power to the motor is switched on and off rapidly, at rates high enough that the effects of the switching are negligible. The resulting effective voltage is then the average fraction of the time the power is on. This technique is also used in many other applications, and is discussed elsewhere in the text.

Figure 20.15 illustrates the concept. The drive signal is switched on and off with a given period and is in the “on” state at voltage $V_{ON}$ for a fixed fraction of the period. This “on”-time is referred to as the “duty cycle” and is stated as a percentage, calculated as:

$$Duty\ Cycle\ (%) = \frac{On\ Time}{Period} \times 100$$  \hspace{1cm} \text{Eq. 20.7}

Figure 20.15 shows a duty cycle of 50%, with the resulting average perceived voltage of this PWM signal equal to 50% of the maximum voltage. The frequency of the PWM drive signal is calculated by taking the reciprocal of the period:

$$PWM\ Frequency = \frac{1}{Period}$$  \hspace{1cm} \text{Eq. 20.8}
Changing the duty cycle of a PWM signal changes the average, or perceived, voltage level. For instance, adjusting the duty cycle shown in Figure 20.15 so that it is in the “on” state for 80% of the period, the situation will change to that shown in Figure 20.16.

Figure 20.15: Pulse Width Modulation with 50% Duty Cycle

Figure 20.16: Pulse Width Modulation with 80% Duty Cycle
By increasing the duty cycle to 80%, the perceived voltage increases to 80% of V\text{ON}.

If PWM is implemented at a frequency that is too low, the result will be a jerky, stop-start response. Instead, the desired result is to approximate the “perceived voltage” as closely as possible with a minimum of perceptible ripple. When driving a permanent magnet brushed DC motor with PWM, the smoothing or filtering is very effectively performed by the physical inertia of the mechanical system. When done properly, the switching occurs too rapidly for the mechanical system to follow. For applications other than controlling the speed of a DC motor (for example: approximating an analog output voltage with a microcontroller digital output pin), filtering circuits may be used to perform the required smoothing of the PWM output.

A typical mechanical time constant for a permanent magnet brushed DC motor is a few milliseconds, without a gearhead or other components attached. When selecting the frequency of the PWM drive signal, the designer must take into consideration the physical characteristics of the mechanical drive system, as well as the electrical characteristics of the motor. Usually, a fairly wide range of PWM frequencies will give reasonable results.

At the lower limit of acceptable PWM frequencies, the mechanical time constant of the physical system (that is, the amount of time the system requires to respond to a change in an input) will dominate. In the extreme case where a very long period is chosen – much longer than the mechanical time constant – the motor will very obviously start and stop. Consider the case of PWM period of 2 second (which is a 0.5 Hz PWM frequency): the motor will almost certainly accelerate and decelerate noticeably during each period. This is the undesirable “torque ripple” that will become much less noticeable if higher PWM frequencies are used. Usually, PWM frequencies in the range of 100 Hz to 1,000 Hz will give good results.
Regardless of the PWM frequency, torque ripple (and, by extension, **current ripple**) will occur to some degree. Higher PWM frequencies result in less ripple, but since current flow in the inductive motor coils is never able to follow the crisp edges of the PWM drive signal, ripple is unavoidable. Figure 20.17 illustrates how a steady state, low duty cycle PWM signal switching between a 0V and \( U_{dc} \) with an average value of \( E \), induces a current flow in the motor’s windings that has a minimum value of \( I_m \) and a maximum value of \( I_M \). Shorter periods, \( T \), will reduce the magnitude of the ripple, and hence the difference between \( I_m \) and \( I_M \). Again, the rise and fall times of the current levels are governed by the motor winding’s inductive time constant, \( \tau = L/R \), Eq. 20.2. The amount of ripple current that can be tolerated will differ depending on the requirements of each application. One manufacturer, Maxon Precision Motors, recommends that ripple current be limited to 10% of their motors’ maximum continuous current specifications.¹

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When the PWM duty cycle is in the process of changing and isn’t in steady state, the current flowing in the motor’s coils will rise or fall in response to the changing average perceived voltage. The current can’t immediately reflect instantaneous change in the duty cycle, just as it can’t perfectly follow the crisp edges of the PWM signal as it switches on and off. Figure 20.18 shows how the motor current responds to changes in PWM duty cycle, again dictated by the inductive rise time of the motor’s coil windings.

![Figure 20.18: Motor current levels changing in response to a non-steady state PWM signal.](image)

PWM speed control can be readily combined with position or velocity feedback to create a servomotor controller (also called a servomotor amplifier). We’ll cover this concept in more detail in Chapter 23: Basic Closed-Loop Control.

**Example 20.2:**

A motor will be required to spin continuously clockwise, and must supply 0.25 oz·in. torque at 5000 rpm when powered by 12V. The motor has $K_T = 1.657$ oz·in./A, $K_e = 1.230$ V/krpm and terminal resistance $R = 20.3$ Ω. Given that bi-directional control is not required, use a DS3658 integrated circuit, manufactured by National Semiconductor, and described as a “high-current peripheral driver” in the datasheet, to drive the motor.

1) Design a circuit to drive the motor and show that none of the component specifications will be violated.
2) Second, specify the PWM duty cycle that will be required to obtain the specified torque and speed. Is there any uncertainty about what the resulting speed and torque will be? If so, where is this uncertainty introduced?

To answer the first question, we need to understand the function of the DS3658. It incorporates four independent switches, each of which can connect a load to ground or disconnect it from ground, and each of which is controlled by a corresponding logic-level input. In the case of a motor or similar load, when one terminal is connecting to a power supply and the other terminal is connected to one of the DS3658’s output pins, the motor will be turned on when the switch connects the output to ground and turned off when it is disconnected. One detail to look out for, though: the output pins are not perfect and can’t connect a load to ground without some small losses. These losses manifest themselves as a slight rise in voltage from ground to the level at the output pin when it is on. This voltage rise is called the “saturation voltage,” and is labeled on the data sheet as “Output Low Voltage.”

The circuit we need to design must take into account that we will be using PWM drive to control the speed and torque of the motor (so this will require that we switch the motor on and off rapidly), that we must protect the components from inductive kickback, and that we must ensure that the components are capable of switching the current flowing through the motor. The circuit below will be effective for controlling the switching, but the remaining specifications will need to be checked to ensure that none are being violated by the design:
Examining the data sheet for the DS3658 shows that each of the chip’s four channels can safely switch 600mA. Conveniently, the chip has built-in kickback diodes (CLAMP 1 and CLAMP 2), each capable of handling up to 800mA of peak current. When making use of the incorporated kickback diodes, they should be clamped to supply voltages less than or equal to 35V. The “output low voltage” specification is 0.35V typical, 0.70V maximum, and no minimum is given. Since we can’t know in advance what the actual output low voltage will be for our system, we will assume it is the “typical” value, and will take into consideration the maximum and minimum values if they are going to have an effect. In order to determine whether the design meets the specifications or not, the second part of the problem will need to be solved.

First, it will be useful to understand if the torque and speed requirements can be met without using PWM. This is, effectively, the case where PWM with 100% duty cycle is used. Once we have established that the system requirements call for a duty cycle less than 100%, we know we can create a successful design. The maximum current the system will need to safely manage is the motor’s stall current, which will occur any time the motor is switched on. This will happen very frequently when PWM control is used, so we will need to ensure that the system can do this safely. With a 12V power supply and including the effects of the typical losses across the DS3658, the stall current is easy to determine using Ohm’s Law ($V=I_{STALL}*R$). Here, the voltage across the motor is the power supply voltage (12V) minus the output low voltage (0.35V):
\[ V_{MOTOR} = V_{SUPPLY} - V_{SAT} \]
\[ I_{STALL} = \frac{V_{MOTOR}}{R} = \frac{12V - 0.35V}{20.3\, \Omega} = 0.574\, A \]

This is close to the specified maximum allowable current flow for a DS3658 channel, 0.6 A, but is within the specifications. Also, the clamp diodes can safely handle up to 0.8 A, so this is within their capabilities as well. However, this is not the worst case. The worst case will be when the output low voltage has a minimum value (resulting in the maximum current flow). Since no minimum value is stated in the datasheet, the safest assumption is that it can have a minimum value of 0V. The results of that calculation are that \( I_{STALL,\, MAX} \) is 0.591 A, which is still below the chips maximum current capabilities. The DS3658 can safely control this motor under these conditions.

For 100% duty cycle, the speed and torque produced can be found using Eq. 19.10 and entering the conditions that are given to solve for the rotational speed that will result:

\[ \omega = \frac{V_{MOTOR}}{K_e} - \frac{R}{K_T K_e} T \]
\[ \omega = \frac{(12V - 0.35V)}{1.23\, V/\, krpm} - \frac{(20.3\, \Omega)(0.25\, oz.\cdot\, in.)}{(1.657\, oz.\cdot\, in./\, A)(1.23\, V/\, krpm)} \]
\[ \omega = 6981\, rpm \]
Since this is a higher rotational speed than we require, we know that an average voltage below the 12V supply voltage will result in the desired output. A PWM duty cycle less than 100% will achieve this. In order to solve for the appropriate duty cycle, the required average voltage must be determined. Rearranging Eq. 19.10 to solve for the required voltage gives:

\[
V_{\text{MOYOR}} = K \left( \omega + \frac{RT}{K_T} \right) = K \omega + \frac{RT}{K_T}
\]

\[
V_{\text{MOYOR}} = V_{\text{MOYOR}} = K \omega + \frac{RT}{K_T}
\]

\[
V_{\text{MOYOR}} = (1.230V / \text{krpm})(5 \text{krpm}) + \frac{(20.3 \Omega)(0.25 \text{oz} \cdot \text{in.})}{1.657 \text{oz} \cdot \text{in.} / A}
\]

\[
V_{\text{MOYOR}} = 9.21V
\]

This is the voltage that must be applied across the terminals of the motor, but recall that there is uncertainty about what the output low voltage at the pin of the DS3658 will be. It could have a minimum value of 0V, a maximum value of 0.7V, and a typical value of 0.35V. For the purposes of this example, we will use the typical value, and we’ll refer to the effective, average voltage that we will achieve by applying PWM to the supply voltage as “\( V_{\text{AVE}} \):

\[
V_{\text{MOYOR}} = V_{\text{AVE}} - V_{\text{SAT}}
\]

\[
V_{\text{AVE}} = V_{\text{MOYOR}} + V_{\text{SAT}} = 9.21V + 0.35V = 9.56V
\]

In order to achieve an average voltage of 9.56V, the 12V power source should be switched on with duty cycle:
\[
Duty Cycle = \frac{V_{AVE}}{V_{SUPPLY}} = \frac{9.56V}{12V} \times 100\% = 79.7\% 
\]

To sum up: we’ve ensured that all the specifications are satisfied, and we will use a PWM duty cycle of 79.7% to achieve the desired operating point. There is uncertainty associated with the DS3658’s output low voltage, and this will affect the speed and torque generated. Great care should always be taken to choose the output low voltage specification that results in the most conservative design. Often, this is one of the extremes, the maximum or minimum, and not the typical value.

### 20.5: Summary

In this chapter, several practical issues related to applying permanent magnet brushed DC motors were addressed. Many of the concepts and techniques presented here also apply to applications involving other categories of high-current loads, especially inductive loads.

The first practical issue addressed in this chapter was inductive kickback, which is the result of abruptly switching off the voltage supplied to an inductive load (such as a DC motor). Its characteristics and potentially destructive results were explored. Several methods of addressing inductive kickback were presented, including a diode snubber, a diode in combination with a resistor or a zener diode, and finally a zener diode across a switching transistor. The relative performance of these snubbing methods was compared.

Circuits for controlling current in DC motors were presented. Circuits enabling bi-directional control of motors were introduced for both single and dual power supplies. The most common and flexible of these circuits, the H-bridge, was explored in detail, as were the various ways in which H-bridges can be used. H-
bridges can be used to spin a motor CW or CCW, and to cause dynamic braking by shorting the motor’s terminals together.

Finally, speed control of DC motors via pulse width modulation (PWM) of a fixed-voltage power supply was introduced. Current flow characteristics of DC motors under PWM control were explored, and guidelines for picking PWM frequencies for satisfactory system performance were given. Minimizing perceptible torque ripple was stated as a primary objective.

Once you have mastered the concepts contained in this chapter, you should be able to:

3) Design control circuitry to switch DC motors on and off and control the motor shaft rotation clockwise and counterclockwise.
4) Effectively mitigate inductive kickback with an appropriate snubbing circuit.
5) Identify and apply integrated circuits that incorporate H-bridge drive circuitry.
6) Use pulse width modulation (PWM) to control the speed and torque produced by a DC motor with a fixed voltage supply.
7) Select the appropriate PWM frequency and duty cycle for a given application.

Chapter 20 Problems:

1. A sensor will be mounted on a small platform that will be required to rotate continuously at a slow, steady rate. A motor with the following specifications is proposed for turning the platform: R = 102 Ω, \( \omega_{NL} \) (at 36 V) = 4080 rpm at 36V, \( T_{STALL} \) (at 36V) = 29 mNm. The application calls for the motor to deliver 1 oz.-in. of torque at 300 rpm. If the motor will be powered by 12V and switched by a BJT (bi-polar junction transistor) rated to handle a maximum of 500mA continuously.

   a) Can the BJT safely switch the required current?

   b) Will the design meet the requirements for torque and speed? If not, how can the requirements be met with this motor and the 12V power supply?

   c) What is the current required when running at the design point?
d) What duty cycle is required when running at the design point?

2. A Pittman 14203S002 DC motor is to be used in an application where it is required to produce up to 3.0 oz.-in. The motor has a no-load speed of \( \omega_{NL} = 4230 \) rpm, stall torque \( T_S = 82.8 \) oz.-in., torque constant \( T_K = 7.44 \) oz.-in./A, speed constant \( K_e = 5.5 \) krpm/V, and resistance \( R = 2.79 \) \( \Omega \). It is to be driven by a 6.0V power supply. Design a circuit to drive the motor using an ST L298 H-bridge, and incorporate kickback diode protection. The design should take in logic level signals for enabling the motor (turning it on and off) and changing the direction of the motor’s rotation. Show that all specifications of the motor, the kickback diodes and the L298 are satisfied.

3. A motor with a stall current of 1.85A is driven with PWM frequency of 1.5kHz. The kickback protection for the drive circuit consists of a standard diode (a 1N4935) in combination with a 24V, ½-Watt zener diode (a 1N5252B). Will this circuit operate within the specifications of these diodes?

4. Design a circuit that performs bi-directional control of a DC motor that uses only the following components: a 2-position, on-on type, double-pole, double-through switch; a DC motor and standard diode(s). Show a wiring diagram of your circuit and be sure to include kickback protection with the standard diodes.

5. Sketch the current response vs. time for a motor with \( L = 0.5 \)mH and \( R = 5 \)\( \Omega \) under PWM control for the following scenarios. Include a sketch of the PWM drive signal on a separate axis for reference:

   a) \( \text{PWM frequency} = 25\text{Hz}, \text{duty cycle} = 90\% \)

   b) \( \text{PWM frequency} = 25\text{Hz}, \text{duty cycle} = 10\% \)

   c) The motor is initially off. At time \( t_0 \), a 20kHz drive signal is enabled at 60% duty cycle.

   d) The motor is initially running at 100% duty cycle. At time \( t_1 \), the PWM duty cycle is changed to 25%.

   e) The frequency of the PWM drive signal is increased steadily from 25Hz to 1kHz. The duty cycle is held constant at 20%.
6. A motor with $K_T = 1.657 \text{ oz.-in./A}$, $K_e = 1.230 \text{ V/krpm}$ and terminal resistance $R = 20.3 \Omega$ is driven at 14V. How fast will the motor spin under 0.15 oz.-in. load if it is driven with a 500Hz PWM signal at 25% duty cycle? At 85 duty cycle%.

7. Design kickback protection circuitry that incorporates a standard diode and a resistor for a motor that has stall current of 900mA. Limit the inductive kickback voltage spike to 36V. Ensure that the diode and the resistor can safely manage the current and resulting heat dissipation. The motor is driven by an NPN bi-polar junction transistor (BJT). Draw a schematic diagram of the complete circuit.