A modern variable-speed motor drive is a system that combines a power-electronic circuit and a motor. The electronics in the drive system control the behavior of the mechanical shaft of the motor. For example, the electronics might be configured to provide the possibility of directly varying the shaft speed or torque based on a low-power, "signal level" command input. The earliest variable-speed drives avoided the need for (then unavailable) highly capable power electronic devices by cascading motors and generators, i.e., an M-G set. In an M-G set, a prime mover, such as a steam turbine or an induction motor running from a fixed-frequency AC source, would be used to turn the shaft of a generator. The generator's terminal voltage could be controlled through a relatively low power field or control winding. The generator's variable terminal voltage could then be used to alter the speed of a motor powered by the generator.

An M-G set involves a significant investment in rotating electric machines. For very high power applications, cascades of electric machines, e.g., a "Ward-Leonard" style system [1], are still occasionally used or maintained. The advent of high performance power-electronic components, however, has generally made it possible to develop electrical terminal drives that can directly control the flow of power through a motor from an electrical source to the machine shaft. Power-electronic circuits permit the construction of compact and highly efficient motor drive systems that are increasingly used in applications from the milliwatt to the megawatt range, including variable fan drives in ventilation systems, pumps and compressors, spindle controls in machine tools, and wheel or shaft drives in electric or hybrid vehicles.

A power-electronic circuit can control either the voltage or the current waveform applied to the terminals of a motor. It can control two basic features of this waveform: the average value or magnitude of the waveform, and the overall frequency content or shape of the waveform. This means that the circuit can drive the machine to control the shaft torque, speed, or position. For AC machines like the induction motor and the permanent-magnet synchronous machine ("brushless" DC motor), the power circuit also provides a specific waveform shape necessary to sustain the conversion of electrical energy into mechanical energy. In the case of the brushless DC machine,
for example, the power-electronic drive effectively serves as an electrical commutator that replaces the action of the mechanical commutator in a conventional DC machine.

The next section explores the distinction between linear amplifiers and high efficiency switching power amplifiers. The following section examines the trade-offs associated with current versus voltage drives in the specific context of the DC motor. Models of DC motor servomechanisms with current and voltage drives will be developed that will expose fundamental design issues associated with all types of DC and AC motor drive systems. The DC motor is used to explore these issues because of the simplicity of the circuitry needed to create a basic DC motor drive. The final section explores useful power-electronic circuits for operating AC machines, either open loop or with active feedback control.

1 Power Converters: Linear versus Switching

The purpose of a power converter is to process or “condition” electrical power, by interfacing a system with a given electrical specification (voltage and/or current) to another system with a different specification. Ideally, this conditioning is accomplished with the use of low-loss components including semiconductor switches, inductors, and capacitors. This distinguishes a switching power supply from a linear regulator: a linear regulator is designed to process power in one direction, from a large reservoir of power to a small consumer, while in the process using a (possibly substantial) fraction of the power drawn from the source to accomplish regulation. The high efficiency of switching regulators (typically over 70 percent, with the state of the art approaching 85 to 95 percent) makes them desirable for powering high density loads with high power requirements and difficult thermal management problems.

The first part of this section examines linear regulators that could, for example, be used to drive a DC motor. The next subsection examines the canonical switching cell DC-DC converter, first presented in [5]. The canonical cell serves as a unifying circuit structure from which the common DC-to-DC high frequency switching converter designs, i.e., the buck, boost, buck-boost, and boost-buck (Cuk) converters, may be derived [6].

1.1 Linear Converters

Figure 1(a) shows an “open-loop,” linear amplifier that uses an NPN bipolar-junction transistor configured as an emitter-follower to provide a controlled voltage across a load. The circuit is called a “linear amplifier” because the transistor is operated in the forward-active region, i.e., the collector current $i_c$ is linearly related to the base current $i_b$ by a constant, $\beta$:

$$i_c = \beta i_b$$

In Fig. 1, the load is a resistor with value $R_{load}$, but this circuit could also be used to drive the terminals of a permanent-magnet DC motor, for example. The voltage $V_{ref}$ is the command voltage that sets the voltage to be applied to the load. Assuming that the transistor is in the forward-active region (not saturated or cut-off), and also making the typical assumption that the forward drop across the base-to-emitter junction diode is about 0.6 volts (optimistically low for a power transistor), Kirchhoff’s voltage law around the loop formed by $V_{ref}$, the base-emitter junction of the transistor, and the load resistor reveals that:

$$V_{load} = V_{ref} - 0.6.$$
Figure 1: Linear Voltage Amplifier.

If the power transistor has a relatively high value of $\beta$ (again, optimistic for a conventional power transistor; a Darlington combination of two transistors or a single Darlington power transistor would probably be essential in practice) then a negligible current in comparison to the load current flows out of the $V_{re}$ source. Hence, a low-power, signal-level voltage could be used to command the voltage across the load. This low power voltage command might come from the digital-to-analog converter of a microcontroller, or from an operational amplifier circuit, for example. As long as we remember to account for the base-emitter voltage drop, the source $V_{re}$ essentially commands the load voltage. The bulk of the power provided to the load comes from the $V_{dc}$ voltage source. The value of this source can fluctuate substantially without significantly affecting $V_{load}$, as long as the transistor stays in the forward-active region.

Assuming that the transistor $\beta$ is large (100 or more), the base current is negligible compared to the load current, and Kirchoff’s current law therefore reveals that

$$I_{load} \approx I_{dc}.$$ 

This approximation can be used to determine the efficiency of the amplifier. Efficiency is commonly defined as the ratio of the output power delivered to the load, to the input power:

$$\eta = \frac{P_{out}}{P_{in}} = \frac{V_{load}I_{load}}{V_{dc}I_{dc}}.$$ 

Recognizing that the load current and the input current are approximately equal,

$$\eta \approx \frac{V_{load}}{V_{dc}} = \frac{V_{out}}{V_{in}}.$$ 

This is a general result for linear power amplifiers: the efficiency is the ratio of the output voltage over the input voltage. If, for example, the circuit is operating from an input source $V_{dc}$ with a nominal voltage of 10 volts, while delivering 5 volts to the load (i.e., $V_{re} \approx 5.6$ volts), the circuit will be 50% efficient. If it delivers 50 watts to the load (10 amps at 5 volts), it must draw about 100 watts from the input source. Where do the other 50 watts go? The power is dissipated as heat in the transistor! Hence, in this example, the case of the transistor would have to be connected to a heat sink adequate to dissipate 50 watts without allowing the temperature of the transistor to rise above its maximum specification.
In situations where the output voltage is fixed and known during the circuit design (i.e., a regulation, not a tracking application), and where the designer can pick the nominal value of an unregulated input voltage that is guaranteed not to fluctuate too much, it is possible to configure the linear amplifier so that its input voltage is only slightly above the output voltage. In this case, relatively high-efficiency operation can be achieved. However, this arrangement is unforgiving if the input voltage fluctuates substantially, or if a wide range of output voltage levels are needed. In general, therefore, most linear amplifiers operate with relatively poor efficiencies in comparison to switching power converters. This disappointing situation is common to all linear amplifiers. For example, a push-pull amplifier constructed from two transistors might be used to provide an AC voltage drive for a load. As long as the transistors in the push-pull stage are operated in the linear, forward-active region, the efficiency of the power stage will still be the ratio of the output and input voltages, and may be distressingly low from the standpoint of thermal management.

Nevertheless, for relatively low power applications, i.e., around 100 watts, the thermal management of a linear amplifier does not pose insurmountable demands, and linear amplifiers are often used in many consumer applications, including stereo amplifiers and small motor drives. Both DC
and AC linear drives are possible. Linear supplies may be especially valuable when electromagnetic interference (EMI) considerations are paramount or are perceived to be important, as in the case of audio amplifiers. They may serve as an effective way to generate a drive waveform with very low total harmonic distortion, for special machines that cannot tolerate high harmonic components in the drive voltage (machine magnetic flux). They are also quick solutions for low power (5 to 10 watts) regulation applications. For example, the popular 7800 series three-terminal voltage regulators essentially consist of the circuit shown in Fig. 1(b). The base of the power transistor is driven by a low power voltage reference, shown in the figure as a stack of diodes. This could also be a zener diode, or another type of precision reference. Something like everything to the left of the dots in Fig. 1(b) is provided in the package of a three-terminal regulator, e.g., the LM7805 five-volt regulator [7]. The user provides the load and the input source $V_{dc}$.

The basic linear amplifier circuit can be modified to provide other improvements or operating modes. For example, a high-gain operational amplifier and a closed-loop feedback arrangement are used in Fig. 2 to ensure that the load voltage precisely follows the command voltage $V_{ref}$. The feedback loop minimizes the effect of the base-emitter voltage drop. This technique is used in many monolithic voltage regulators to ensure good output voltage regulation.

Also, it is possible to reconfigure the circuit as shown in Fig. 3 to provide a current-source amplifier. In this case, the transistor is arranged as a common-emitter amplifier. The command voltage $V_{ref}$ sets the base current through the resistor $R_b$. The transistor draws a collector current through the load resistor that is $\beta$ times the base current. Hence, in this circuit, the command voltage $V_{ref}$ sets the load current, rather than the load voltage, with a gain that is a function of the value of the base resistor, $R_b$, and $\beta$.

### 1.2 DC-DC Switching Converters

A DC-DC switching regulator is a circuit that can provide a controlled or regulated DC output voltage from an unregulated DC input. It serves as an interface between two or more DC systems (hence the name DC-to-DC switching power supply), and can generally be designed to operate with high efficiency in comparison to a linear regulator. In the next section, for example, a “buck”-type switching regulator, or “chopper”, will be used to provide a controllable voltage across the armature terminals of a DC motor given an unregulated DC input voltage source. These converters might also be used to drive the field windings in AC or DC machines.

To get a better understanding of the distinction between different types of converters, consider how two DC systems might be interfaced. Start with a simple two port model. This model is depicted in Fig. 4. The interface (the two port box in Fig. 4) is presumed to contain no significant internal sources or sinks of power.

For now, no particular circuit topology is being derived here. Rather, the goal is to expose common features which relate all of the high-frequency, square-wave switching topologies. In this simple example, ignore any control or implementation issues. Also, assume that the terminal voltages and currents are constant with the polarities shown in Fig. 4. Under these conditions, the steady-state power sinked at one terminal must be sourced at the other terminal and vice versa. Notice that either port may either source or sink power. Applying Kirchhoff’s current and voltage laws (KCL and KVL) reveals that the series switch must have average DC components of 50 volts across it and 15 amps through it. Similarly, the shunt switch must withstand an average of 150 DC volts across it and 5 amps through it.
Ideal switches provide a simple way to create consistent average waveforms in this configuration. They are particularly attractive because, ideally, when “off” they withstand a voltage without passing any current, and when “on” they pass a current with no voltage across the switch terminals. Hence, the product of voltage and current for the ideal switch is zero at all times and no power is dissipated.

To be consistent with KVL and KCL under the terminal conditions specified, the switches are operated to create voltage and current waveforms with appropriate average values. Specifically, during normal operation over a “switch period” of time $T$, the series switch turns on for a period of time $\frac{3T}{4}$. Then, this switch turns off and the shunt switch turns on and remains on for a time $\frac{T}{4}$. At the end of this time, the shunt switch turns off and the cycle repeats itself. In the jargon of the trade, the series switch has a duty cycle of 75 percent and the shunt switch has a duty cycle of 25 percent. At no time are both switches open, as this would result in a cessation of power flow.

When the series switch is on, it carries a current of 20 amps to feed the 150 volt terminals. When the series switch is off, the shunt switch is on and the series switch is connected across the 200 volt input. Hence, the series switch withstands 200 volts. With the timing scheme described in the previous paragraph, we see that the current and voltage waveforms for the series switch are as shown in Figs. 5 and 6 respectively.

The average values of these waveforms are 15 amps and 50 volts respectively, as indicated in the figures (average variables are marked with <> symbols). A similar analysis of the shunt switch yields equally agreeable results. This simple switching scheme is apparently one option for interfacing two DC systems. Unfortunately, while the average values of the waveforms conform to their required values so that Kirchoff's Laws are satisfied (on average), this scheme superimposes a substantial AC component to the terminal waveforms. Even though the waveforms in Figs. 5 and 6 have the necessary average values, their instantaneous values ripple.

Filtering components must be added to smooth the waveforms. To maintain high efficiency, only inductors and capacitors are added. To help stiffen the bus voltage, a capacitor is added in parallel with the left terminals. An inductor is used to isolate the 150 volt terminal pair on the side with the shunt switch. Of course, the inductor will pass a continuous DC current but will tend to block AC current. A revised version of the simple two port interface is shown in Fig. 7.
This minimal topology “interface” is in fact the skeleton of four basic types of high frequency DC-DC switching converters. This simple form, shown in Fig. 7, is so fundamental that it has been called the canonical switching cell [5]. It is often redrawn as shown in Fig. 8 [5], [6].

As long as the switches are ideal (so that currents and voltage may flow or be blocked in any direction), power may flow bidirectionally through the circuit. By varying the interconnection of the designated input and output terminals and the implementation of the ideal switches with real devices (which may or may not limit the flow of current or voltage in certain directions), all four basic high frequency topologies may be derived. When the canonical cell is connected so as to allow power to flow directly from one terminal to another when one of the switches is closed, the buck or boost converters are formed. The type of converter formed depends on the direction of power flow; in the case of the buck converter, or chopper, power flows from a high voltage source to a lower voltage load when the controllable switch is closed. In the case of the boost converter, power flows from a low voltage source to a load operating at a higher voltage when the diode is conducting. The classic buck and boost implementations of the canonical cell are shown in Figs. 9 and 10.

The braces in Figs. 9 and 10 indicate the canonical cell portions of the buck and boost converters.
Figure 7: Simple Topology for a DC-DC converter

Figure 8: The Canonical Switching Cell

Note that actual implementations invariably make modifications to the canonical cell. For example, an extra output capacitor was added to make the buck topology shown in Fig. 9. Also, for simplified discussions, the input capacitor is usually discarded from the buck configuration of the canonical cell, i.e., the voltage source is presumed to be “stiff,” or to exhibit little internal impedance.

The buck-boost and boost-buck converters are implemented by configuring the canonical cell so that power never flows directly from one port terminal to another; energy is accumulated in one of the storage elements during part of the cycle and then is removed from the element to power the load on another part of the cycle. In the buck-boost topology, the most common of the “indirect” power supplies, the inductor is the intermediary energy storage element. Also, the capacitance in the canonical cell is typically split between the input and output terminals. A simple buck-boost converter is shown in Fig. 11. A discussion of the less-common boost-buck converter is deferred to the references [14],[16],[22].

For now, all of the switching power supply circuits have been drawn with a resistive load. Assuming that the switches, inductors, and capacitors in the power supplies are ideal, the resistive load is the only element in the circuits that can dissipate time-average power. This means that, in theory, any power taken from the input source is ultimately delivered to the load. In other words, the circuits are theoretically 100% efficient. Of course, the components would not be ideal in practice, but it is commonplace to achieve efficiencies in excess of 90% in actual implementations.
The output voltage or current provided to the load can be actively controlled to a particular value or reference waveform by appropriately varying the duty cycle of the controllable switch. The next section examines a basic scheme for controlling the output voltage of a buck converter.

1.3 Buck Converter with Voltage Control Loop

To understand how a switching power amplifier might be used in a servomechanical drive, consider the problem of making an output voltage that tracks a command reference using the buck converter shown in Fig. 12. This is identical to the task accomplished by the linear regulator shown in Fig. 2, whose output follows the command $V_{ref}$. Properly controlled, the buck converter should also be able to create an output voltage that follows a command reference, but with a higher efficiency.
over the full tracking range than was achieved with the linear amplifier.

![Buck Converter diagram](image)

**Figure 12: Buck Converter.**

Our goal is to control the output voltage of the converter. Given a particular input voltage source and a fixed switch frequency, our only possibility for controlling the output of the converter is to vary the duty cycle of the controllable switch. First, reconsider the open-loop behavior of the buck converter. Suppose as in the previous section that the controllable switch is operated with a fixed switch period and a duty cycle D that we can pick. This approach is called pulse-width modulation (PWM), and is illustrated in the top plot in Fig. 13(a). The controllable switch is “on” for a time DT each cycle, and “off” for the remainder of the cycle. For now, focus on steady-state operation, and assume that the inductor is large enough that it is in continuous conduction — that is, there is always current in the inductor. When the switch is on, the input voltage will keep the diode reverse biased, i.e., “off.” When the switch turns off, the inductor will force the diode on in order to keep current flowing. Therefore, the instantaneous voltage across the diode, $V_m(t)$, therefore, looks like the bottom trace in Fig. 13(a).

When the load on the switches (e.g., the circuit driven by $V_m(t)$ in Fig. 12) is inherently low-pass in nature — that is, when the natural frequencies of the load are relatively slow compared to the switch frequency — the effect of a PWM drive is conveniently analyzed in terms of average variables. One definition for an average variable value in this context is

$$< w(t) > = \frac{1}{T} \int_{t-T}^{t} w(\tau) d\tau$$

where $< w(t) >$ indicates the average value of a variable $w(t)$ over the period $T$.

The waveform $V_m(t)$ can be thought of as the sum of two waveforms. One is the near-DC or average component, $< V_m(t) > = DV_m$. The remainder is $V_m(t) - < V_m(t) >$, i.e., the “AC” component of the diode voltage waveform. If, for example, $D = 0.5$, then this AC component looks like a zero-centered square wave, with a maximum value of $\frac{V_m}{2}$ and a minimum value of $-\frac{V_m}{2}$.

In a well-designed buck converter, the values of the inductor and capacitor, $L$ and $C$, will have been chosen so that, for a typical range of load values $R$, the transfer characteristic of the LRC low-pass filter formed by the “back-end” of the buck converter will look something like the magnitude Bode plot shown in Fig. 13(b). That is, the filter will pass the low-frequency DC and near-DC components of $V_m(t)$ straight through to the output. The values of $L$ and $C$ will have been selected, however, so that the AC components of $V_m(t)$ (generally at and above the switch frequency, $\frac{1}{T}$) will be severely attenuated by the low-pass filter. This means that $V_{out}(t)$ will look essentially like the near-DC $< V_m(t) >$, with a small amount of AC ripple that makes it through the second-order low pass filter.

Setting a fixed duty cycle for the buck converter in Fig. 12 is therefore somewhat similar to setting $V_{ref}$ in the linear amplifier shown in Fig. 1(a). Of course, the buck converter will generally
(a) Switch State

\[
\begin{array}{c|c|c|c|c|c}
V_m & V_{in} & DT & T & T+DT & 2T \hline
0 & 0 & 0 & 0 & 0 & 0 \\
\end{array}
\]

\[\langle V_m \rangle = \frac{V_{in}D}{t} \]

(b) Graph of \( 20\log\left| \frac{V_{out}}{V_{in}} \right| \) vs. log(frequency)

(approaches) Slope \(-40\text{db/decade}\)

(c) Block diagram of Buck Converter

\[D = K (V_{ref} - V_{out})\]

(d) Block diagram of Buck Converter with transfer function

\[P(s) = \frac{1/LC}{s^2 + s/RC + 1/LC}\]
operate with a higher efficiency than the linear regulator, unless special steps have been taken to limit the difference between the input and output voltage. Notice, contrary to the situation in the linear amplifier, that the open-loop buck converter running with a fixed duty cycle provides absolutely no rejection or attenuation of variations at the output due to changes in the input source \( V_{in} \). With a fixed value of \( D \), the output voltage of the buck converter is essentially \( DV_{in} \). If the input voltage changes, so will the output voltage.

This may not be of great concern if the buck converter is part of a servomechanical control system driving, for example, a DC motor. In this case, the buck converter and motor will be embedded in a closed loop attempting to control speed or some other mechanical quantity. Variations in the input voltage to the buck converter will be handled (by a well-designed control loop) as disturbances that get rejected.

If the buck converter is being used to drive a motor or some other mechanical or thermo-mechanical system such as a heating element, it may not be necessary to include the complete \( LC \) output filter shown in Fig. 12. Two motor-drive examples involving permanent-magnet DC motors are illustrated in Fig. 14. In Fig. 14(a), the low-pass output filter of the buck converter has been eliminated entirely. This circuit is sometimes called a “chopper”. In this case, the DC motor will see an average voltage at its armature terminals equal to \( DV_{in} \). It will also, of course, see a large ripple voltage around this average. However, if the motor is connected to a mechanical subsystem at its shaft that is substantially low-pass in character, e.g., a large inertia (flywheel) with some mechanical friction, the shaft speed will not respond to the rippling armature voltage if the switch frequency is high enough. It will only respond to the average value of the armature voltage set by the duty cycle command.

Taking advantage of a separation in time scales or time constants can be possible and valuable in other systems as well. Consider, for example, driving a resistive thermal heater to warm a large bath. Imagine that this load is suddenly driven with an input power of \( P \) watts by a linear, DC amplifier that provides “flat” waveforms of voltage and current. If the heater is warming a substantial mass, it might be several seconds or more before a measurable change in temperature is observed. Imagine that a switching amplifier is employed that provides \( 2P \) watts for half a switch period, and 0 watts for the other half, cycling periodically and rapidly in comparison to the time scale for response of the thermal mass. In this case, the net effect on the temperature of the mass would be essentially the same as if the amplifier delivered \( P \) watts continuously. The thermal system simply cannot follow the rapid changes in input power. Its “low-pass” character is incapable of following high frequency variation in the input, averaging the effects of high and low input levels over time. Of course, this would not be the case if the switch period were not sufficiently short. A long switch period would provide significant time for the mass to “overheat” during the time when \( 2P \) watts were provided, and would also permit significant cooling of the mass during the fraction of the switch period when 0 watts were delivered. An example of this is switching with an inexpensive bimetallic strip to establish control for electric heating elements in residential kitchen stoves.

Even with a “low-pass” mechanical load, however, the situation in Fig. 14(a) may not be entirely acceptable, because the ripple voltage at the armature electrical terminals may cause a large ripple current to flow. This ripple current may cause large ripples in the shaft torque. If the mechanical system is slow, these ripples may result in only small oscillations of the speed. However, the large peak currents may exceed the switch current ratings in the chopper. They may also be responsible for unacceptable levels of acoustic noise from the motor windings, and could exceed the peak current.
specification of the armature electrical port. Generally, these problems show up most acutely in machines with relatively low armature impedances, e.g., high performance “servodisk” motors. In such cases, one solution is to add sufficient inductance externally to the armature circuit to limit the voltage (current, torque) ripple, as shown in Fig. 14(b). External inductance must be added with care in the case of a DC motor in order to avoid creating unacceptable discharging in the mechanical commutator. Raising the switch frequency, if possible, can reduce the size of the impedance that must be added.

![Diagram of Chopper Circuits](image)

**Figure 14:** Chopper Circuits.

A simple switching circuit that might be employed as a bidirectional chopper is shown in Fig. 15. Notice that both switches can never be on at the same time, or a very low impedance path will form between the positive and negative thirty-volt rails. The diodes provide “free-wheeling” paths for current when the controllable switches are both off. These free-wheeling or “catch” diodes are essential if the load has non-zero inductance.

![Diagram of Bipolar Chopper](image)

**Figure 15:** Bipolar Chopper.

Finally, if for some reason it is essential that the buck converter accurately follow an output voltage command reference and actively reject input voltage disturbances, it is possible to close a feedback loop that varies the duty cycle to achieve a desired output voltage waveform. An actively controlled converter could also serve as a component block in a larger servomechanism. One possibility for controlling the output voltage of the buck converter, shown for the case of a resistive load, is illustrated in Fig. 13(c). The duty cycle is computed by a proportional compensator as the error or difference between a command reference and the actual output voltage, times a gain \( K \). This error computation could be accomplished, for example, by an operational amplifier configured as a subtractor with a gain of \( K \). The duty cycle is compared to a sawtooth waveform by a comparator. The sawtooth has a period of \( T \), a peak voltage of one volt, and a minimum voltage of zero volts. When the sawtooth waveform is below the value of the \( D \) output waveform
computed by the subtractor, the output of the comparator is high. When the sawtooth rises above $D$, the comparator output goes low. Hence, the output of the comparator produces a PWM drive waveform that could be used to directly drive a controllable switch (e.g., a transistor or MOSFET) in the buck converter.

The performance of the actively-controlled buck converter can be analyzed with the feedback diagram shown in Fig. 13(d). This block diagram contains all of the functionality of Fig. 13(c). However, ripple voltage has been presumed to be small and is ignored in Fig. 13(d). That is, the block diagram describes only the relatively slowly varying dynamics (in comparison to the ripple frequency) of the output voltage. The difference between $V_{\text{ref}}$ and $V_{\text{out}}$ is computed by the subtractor, and fed to the proportional compensator with gain $K$ to compute the duty cycle, $D$. The product of the duty cycle and the input voltage $V_{\text{in}}$ produces $< V_{\text{in}}(t) >$, which passes through the transfer function $P(s)$ of the LRC output filter of the buck converter shown in Fig. 12. Using Black’s formula [8], the overall input-output transfer function for the actively-controlled buck converter is

$$C(s) = \frac{V_{\text{out}}}{V_{\text{ref}}}(s) = \frac{K V_{\text{in}}}{s^2 + \frac{s}{RC} + \frac{K V_{\text{in}} + 1}{LC}}$$

The second order denominator of this transfer function has all positive coefficients as long as the circuit is loaded with a finite value of output resistance. Hence, the closed-loop system poles are guaranteed to be in the left half-$s$-plane (Routh criterion) and the system will be stable, in principle. For this proportional compensator, the steady-state error in response to a step change in the command input $V_{\text{ref}}$ decreases as the gain $K$ increases. However, the damping of the system also decreases as $K$ increases. For larger values of $K$, the closed-loop system will therefore overshoot higher and oscillate more times than for lower gains. The proportional compensator is by no means the most sophisticated or flexible choice of series compensation for the buck converter, although it can produce reasonable steady-state performance. It is also possible to close a feedback loop around the buck converter that varies the duty cycle to ensure that a specific value of current is delivered to the load. That is, a current-source output could be made using the buck converter and an active control loop.

The next section will examine the effects of different power amplifier choices (current versus voltage drive) in permanent-magnet DC motor velocity and position servo systems.

2 DC Motor Servomechanisms for Velocity and Position Control

Many electromechanical systems are tasked to provide precise control of a position or a velocity. A system could, for example, be required to regulate a position to a specific value or location, as in the case of a position controller for an antenna. It could also be required to accurately track a time-varying position or velocity reference, e.g., when following an aircraft with a radar dish, or serving as a drive motor in an electric vehicle. Direct-current machines are often used as actuators in such systems. DC machines provide rotary motion and torque, and can also provide linear motion and force through clever mechanical arrangements such as lead-screw mechanisms.

DC machines may be somewhat more expensive than comparably rated machines of other types (induction, stepping, and variable reluctance machines), and may also be more difficult to maintain. Because of limitations on the mechanical commutator, DC machines generally cannot be used at high altitudes or in vacuum. However, from a power electronics standpoint, DC machines are
relatively easy to control compared to many other motors, e.g., induction machines. Well-designed DC machines provide a smooth, nearly continuous motion with little “cogging,” as would be found in stepper motors, for example. For these and other reasons, DC machines have been used in servomechanisms for over a century, and are still popular in many applications.

Because of the relative simplicity of the power electronics needed to drive a DC motor, this section reviews the basic issues of constructing velocity and position servomechanisms using the DC motor as an actuator. The following section will examine the power-electronic requirements for AC machines. However, the basic approach for controlling most AC machines is, to some extent, to first make them “look like” a commutator DC machine, at least in a mathematical sense. The basic approach to designing closed-loop controllers for position and velocity servos, therefore, is generally the same for both DC and also most AC machines (given an appropriate AC power amplifier).

2.1 Circuit Analogue

First, this section will quickly review the basic wound-field DC-machine model developed in more detail in this text. The field winding in a DC machine is reasonably accurately modeled as a resistance in series with an inductance. The armature winding consists of wire coiled around a high-permeability rotor. The armature electrical port, therefore, has some electrical loss that could be modeled as a resistance in series with the back-electromotive force (back-EMF) source. It also has some inductance (typically small, in comparison to the field winding), which could be incorporated as an inductor in series with the resistor and back-EMF source. The mechanical shaft could incorporate elements to account for load torque, viscous damping, and other mechanical effects and components (e.g., a gear box). A reasonable circuit model of a DC machine is shown in Figure 16. The voltage source labeled \( V_{bem f} \) in Fig. 16 is a speed-dependent voltage source that represents the back-EMF of the motor. This “source” is a transducer, therefore, connecting the mechanical subsystem (shaft of the motor) to the electrical subsystem (armature terminals). The torque produced by the shaft is a function of the field current and the armature current:

\[
\tau_m = G I_f I_a
\]  

(2)

The motor constant \( G \) is a function of the machine’s construction (e.g., number of armature turns, rotor radius, etc.). More complicated models accounting for nonlinear electrical effects like saturation and nonlinear mechanical effects like windage (air flow around the rotor) could be developed as needed. The model in Figure 16 is often very satisfactory for control design, however.

![Figure 16: Circuit Model.](image)

Starting with the model summarized in Fig. 16, we can use basic circuit analysis techniques to develop a concise third-order dynamic (differential equation) model of the DC machine. The
field circuit can be described by a first-order, linear, time-invariant differential equation relating the field current \(i_f\) to the field terminal voltage \(v_f\):

\[
\frac{di_f}{dt} = \frac{1}{L_f} v_f - \frac{R_f}{L_f} i_f.
\]

The armature current \(i_a\) is described by a nonlinear differential equation that depends on both field current \(i_f\) and rotor speed \(\omega\):

\[
\frac{di_a}{dt} = \frac{1}{L_a} v_a - \frac{R_a}{L_a} i_a - \frac{G i_f \omega}{L_a}.
\]

Finally, the mechanical subsystem is described by Newton’s second law. The DC machine makes a machine torque \(\tau_m\) described by Eq. 2. The motor shaft has inertia \(J\) and is connected to a load (motor operation) or prime mover (generator operation) that exerts a torque \(\tau_l\) on the shaft. (This torque could be, and often is, a function of speed.) For now, we ignore other possibly complicating details such as the presence of a gear box. Angular acceleration of the machine rotor results from differences between the two torques \(\tau_m\) and \(\tau_l\):

\[
\frac{d\omega}{dt} = \frac{1}{J}(\tau_m - \tau_l) = \frac{1}{J}(G i_f i_a - \tau_l)
\]

One simple model for a common load torque is a friction \(\tau_f\) that is linearly dependent on angular velocity, \(\tau_f = \beta \omega\), where \(\beta\) is a constant.

In the case of a permanent-magnet (PM) DC machine, the field winding is replaced by a permanent magnet. This situation is functionally equivalent to driving the field circuit in Fig. 16 with a constant-current source of value \(I_f\) sufficient to create a comparable air-gap magnetic field to that produced by the magnet. The PM DC machine is reasonably accurately modeled by a second-order model consisting of Eqs. 4 and 5, with \(i_f = I_f\). Block diagrams of a PM DC machine appropriate for control design are shown in Fig. 17. Laplace transforms of Eqs. 4 and 5 have been employed in the block diagrams, and the system is assumed to start from initial rest condition. The constant \(K_m\) equals \(GI_f\), the product of the motor constant and the DC “field current” that represents the effect of the permanent magnet.

The first diagram in Fig. 17(a) shows how an ideal voltage-source drive on the armature would affect the shaft speed and position. A practical implementation of such a power amplifier could be made with a linear amplifier like the emitter-follower or a switching amplifier like the chopper. Notice that, in particular in a high quality machine with low armature resistance and inductance, a fixed armature voltage will approximately set the level of the steady-state back-EMF. Since the back-EMF depends on the shaft speed, setting the armature voltage to a particular value in a PM DC motor approximately sets the shaft speed. The next block diagram in Fig. 17(b) shows how an ideal current-source driving the armature terminals produces torque, speed, and position at the machine’s mechanical shaft. A current-source amplifier, in theory, directly commands armature current and therefore shaft torque. The details of the armature electrical circuit, i.e., armature resistance and inductance and motor back-EMF, are negligible to the extent that the current-source amplifier has sufficient voltage compliance to command current in the armature winding. The block diagram for the ideal current-source drive, therefore, is simpler than for the ideal voltage-source drive in Fig. 17(a).

In practice, it is not possible to drive the machine with an ideal current source. The power amplifier’s ability to command current in a winding will always be limited by its ability to command
(a) Assuming a frictional load torque linearly proportional to velocity

(b) 

(c)
an instantaneous terminal voltage necessary to set the current. This terminal voltage will not be
infinite in magnitude or negligible in rise time, and the current command cannot, therefore, be
arbitrary. This can be seen in any of the amplifiers, linear or switching, that have been examined
in previous sections. Consider, for instance, the common-emitter amplifier shown in Fig. 3. Imagine
that the load resistor is replaced by a winding that consists of some inductance and some resistance.
In response to a step change in the base current, i.e., a sudden change in the command voltage $V_{ref}$,
the best that the transistor can do is saturate, i.e., turn on “fully”. The rise time of the current
will then be governed by the L-R time constant of the load winding.

A more realistic model for a current-source drive, therefore, is shown in Fig. 17(c). Here, the
current amplifier accepts a command labeled $i_{ref}$, which is compared to the actual armature current
$i_a$. A proportional compensator with gain $K_a$ drives an armature voltage $v_a$ based on the error
between the requested and actual current. This arrangement does not directly represent any of the
circuits from previous sections, but it could be created with either the linear or switching amplifiers.
It also does not account for nonlinear effects like saturation in the amplifier. However, the block
diagram in Fig. 17(c) will serve to bring out some of the “real-life” issues that arise with a practical
current-source amplifier.

2.2 Velocity Servo – Current Source Drive

Because it is analytically simplest, let us begin by examining how to make a closed-loop system
to control motor velocity using a current-source drive. A block diagram of a feedback loop for
controlling motor speed with a current-source power amplifier is shown in Fig. 18(a). The gain of
the tachometer is presumed to be one, but any constant gain or transfer function could be added
if appropriate. The block with transfer function $GC(s)$ represents a series compensator chosen to
yield closed-loop stability and performance. This block includes the ideal current-source power
amplifier. This section will consider several possible compensation options. First we consider a
proportional compensator, i.e., $GC(s)$ is a constant gain.

A circuit schematic of a unidirectional demonstration system based on the block diagram in
Fig. 18(a) is shown in Fig. 19. The current-source amplifier is implemented with two NPN bipolar-
junction transistors connected in a Darlington configuration. This arrangement increases the overall
current gain of the amplifier, minimizing the loading of the LM358 operational amplifier driving
the transistors. The Darlington-connected transistors are used in a common-emitter arrangement
with a “power” level (20 Volt) DC voltage source for the motor’s armature circuit. A flyback or
“catch” diode provides a free-wheeling path for the armature current in the event that the control
loop is deactivated suddenly while the machine is running. Operational amplifiers implement
the remaining functions in the feedback loop. An LM358 configured as a non-inverting amplifier
provides a variable, proportional compensation gain. The error signal between the tachometer
voltage $v_{tach}$ and the speed reference $V_{ref}$ is computed by an LM358 arranged as a subtractor. The
voltage reference, which might be made by a potentiometer connected across the power supply rails,
is buffered by an op-amp follower. The tachometer feedback signal $v_{tach}$ might also be buffered by
a follower if necessary (not shown).

The closed-loop transfer function for the velocity servo loop can be found by applying Black’s
formula [8] twice to the inner and outer loops shown in Fig. 18(a). For a proportional compensator,
$GC(s) = K$, a constant that includes the op-amp and transistor amplifier gains in Fig. 19. In this

18
case, the transfer function relating motor speed to commanded speed is

\[
\frac{\omega}{\omega_{ref}}(s) = \frac{KK_o}{\beta s + (1 + KK_o/\beta)}
\]  

(6)

Ideally, for a constant speed reference, this transfer function would approach unity in steady-state (zero frequency). That is, for DC excitations, the output should track the input or reference command perfectly. Unfortunately, for the proportional controller with a finite proportional gain, there is always a steady-state error or difference between the commanded and actual speeds.

In principle, the compensation gain could be increased indefinitely to make this error arbitrarily small. A typical root-locus of the closed-loop system pole locations for increasing positive gain, shown in Fig. 20, indicates that the system should remain stable as the gain is increased. Furthermore, a Bode plot of the magnitude and phase of the closed-loop transfer function, Eq. 6, indicates not only that the low-frequency behavior of the transfer function will approach unity as the compensation gain increases, but also that the closed-loop tracking bandwidth will increase. In practice, however, increasing the compensation gain indefinitely will most probably become seriously detrimental after a certain point. Higher closed-loop bandwidths demand excessive peak power requirements from the power amplifier. Also, as the gain is increased, we become more likely to "find" unmodeled poles. That is, the comforting stability argument made by the root-locus diagram is unlikely to be true when the effects of other, unmodeled system poles (e.g., poles from the tachometer and power amplifier) are included.

We could instead try an integral compensator, \( G_i(s) = \frac{K}{s} \). The integral compensator provides a frequency-dependent gain that is, in principle, infinite at zero frequency. In this case, the closed-loop transfer function is

\[
\frac{\omega}{\omega_{ref}}(s) = \frac{KK_o/\alpha}{s^2 + s^{1+KK_o/\beta\alpha}}
\]  

(7)
where

\[ \alpha = \frac{J}{\beta}. \]

Figure 20: Root Locus: Proportional Compensation.

Notice that the steady-state value of the closed-loop system under integral compensation is unity, i.e., the system tracks DC commands perfectly, regardless of the value of the gain \( K \). The Routh criterion [8] can be used to show that the poles of Eq. 6 are always in the left half-s-plane, i.e., the system is always stable in principle. A root locus of the system poles for positive integral gain \( K \) is shown in Fig. 21. With integral compensation, there is little advantage to excessively high values of the gain \( K \). After a certain point, the closed-loop bandwidth of the system essentially stops increasing, and the transient response of the system becomes more and more oscillatory.

There is a wide range of options for compensating this servo system to achieve desired performance; see [10], [11], and [12] for a more complete discussion of the possibilities.

### 2.3 Velocity Servo – Voltage-Source Drive

Figure 18(b) shows a block diagram of a closed-loop velocity drive that employs a voltage-source amplifier to drive the motor. Because the amplifier is a voltage drive, the block diagram now shows explicit dependence on the armature impedance and the back-EMF. The loop with voltage drive will be shown to have an additional pole in comparison to a similarly compensated loop with current drive. This pole arises from the armature electrical subsystem.

The series compensator \( G_c(s) \) includes the voltage amplifier. For now, assume that the system is compensated with a proportional gain, so that \( G_c(s) = K \), a constant that includes the gain of the voltage power amplifier. Employing Black’s formula repeatedly on the block diagram in Fig. 18(b), the closed-loop transfer function can be shown to be:

\[
\frac{\omega}{\omega_{ref}}(s) = \frac{\frac{K K_m}{R_o \beta}}{(\tau_e s + 1)(\alpha s + 1) + \frac{K^2}{R_o \beta} + \frac{KK_m}{R_o \beta}} \tag{8}
\]

The time constant \( \tau_e = \frac{L_m}{R_a} \) is sometimes called the electrical time constant of the machine. In
relatively small machines, this time constant may be small compared to the mechanical time constant \( \alpha \). If the electrical time constant is negligible (near 0), then the voltage velocity loop with proportional compensation is essentially first order, and has a root locus similar to the one shown in Fig. 20. If the electrical time constant is not ignored, then the root locus will look something like the one shown in Fig. 22.

Whether or not the electrical time constant is ignored, the proportional compensator always leaves some steady-state error in response to a step input. To eliminate this steady-state error, we might again consider using an integral compensator, \( G_c(s) = \frac{K}{s} \), where \( K \) is a gain to be chosen. Ignoring the electrical pole leads to a system that exhibits zero steady-state error in response to a step input, and which has a root locus similar to the one shown in Fig. 21.

If we do not ignore the electrical pole, an integral compensator still leads to a closed-loop system with zero steady-state error in response to a step input. However, this system, with three poles (one from the integral compensator at the origin, one from the mechanical subsystem, and one from the electrical subsystem) will have a root locus like the one shown in Fig. 23. For modest gains, the dominant poles (closest to the origin) of the system shown in Fig. 23 behave much like the dominant poles shown in Fig. 21. For higher gains, disaster ensues! If too high a gain \( K \) is selected, the system in Fig. 23 will be unstable. Even if the electrical pole is ignored during the design process, it or some other unmodeled pole is very likely to be present in the actual system. Great care must always be taken, therefore, when selecting compensator structures and gains. It is perhaps a too common practice to ignore the supposedly fast electrical poles in favor of the mechanical poles when designing a servomechanical drive system. Especially in very large machines, AC or DC, the electrical poles can actually be slower than the mechanical poles. Ignoring these poles in this case will almost certainly lead to an unstable servo system.

2.4 Velocity Servo – Practical Current-Source Drive

Figure 17(c) shows a block diagram of a PM DC machine with a practical current-source drive, i.e., one with some dynamic limitation on the terminal voltage compliance. With some effort, the block diagram in Fig. 17(c) can be manipulated to reveal that the open-loop transfer function relating
Figure 22: Root Locus: Voltage Loop with Significant $\tau_e$.

Figure 23: Root Locus: Voltage Loop with Significant $\tau_e$ and Integral Compensation.
the output speed to the input current command is

\[
\frac{\omega}{i_{ref}}(s) = \frac{K_e K_w}{(\tau_e s + 1 + \frac{K_e}{K_a})(\alpha s + 1 + \frac{K_w}{K_a})}
\]  

(9)

This transfer function is second order. Closing a proportional loop around this system to regulate speed would result in a system with finite steady-state error in response to a step, and with a root locus similar to the one shown in Fig. 21. Notice, however, the effect of the “minor loop” that attempts to force the actual armature current to follow the reference command \(i_{ref}\). The electrical pole begins substantially deeper in the left half-s-plane. That is, as long as it does not saturate, the current minor loop has effectively decreased the electrical time constant, making the system “look more first order”, as we would have expected from the prior analysis with an ideal current-source drive. Also note that an integral compensator wrapped around this system to regulate speed would result in zero steady-state error, and a root locus similar to the one shown in Fig. 23. However, because the minor current loop starts the electrical pole deeper in the left half-s-plane, the system with integral compensation might be able to support a larger range of variation of the integral gain \(K\) before going unstable.

2.5 Position Control Loops

Notice from Fig. 17 that the difference between a DC motor model that describes a velocity output at the mechanical shaft and one that describes position is the absence or presence, respectively, of a final integration block in the diagram (to integrate velocity into position). In a block diagram with a series cascade of blocks, the order of the blocks is irrelevant in determining the through transfer function. Hence, any closed-loop transfer function derived for a velocity servo loop with an integral compensator will be identical to one derived for a position servo loop with a proportional compensator. This means, for example, that the transfer function Eq. 7 describes not only a velocity servo loop with a current-source drive and integral compensator, but also a position servo loop with a current-source drive and proportional compensator. Conclusions drawn in the previous three sections for velocity loops with integral compensators all apply, therefore, to position loops with proportional compensators. The exploration of other compensation possibilities are left to the reader and the references [10], [12].

3 Power Electronics for AC Drive Systems

In a DC machine, the mechanical commutator ensures that current flows in the machine windings in a manner that will produce useful torque, even when the rotor changes position and speed. From a modeling and control standpoint, the presence of the commutator significantly eases the problem of designing appropriate power electronic amplifiers for driving DC machines. Essentially, the problem becomes one of developing circuitry that can create flexible levels of voltage or current, without too much concern for the specific waveshape – either a linear amplifier or a chopper would generally produce adequate results, for example. Only the relatively slowly-varying, average values of the terminal waveforms prove to be of concern in a well-designed system. In an AC machine, there is no mechanical commutator, and the electrical excitation of the stator must be appropriate to ensure sustained torque production. For a controllable motor drive, this generally means that
the drive electronics must be capable of producing AC waveforms with controllable frequency and amplitude.

![Figure 24: Three Phase Rectifier.](image)

Switching power electronic drives for AC machines are often (but not always) constructed as inverters, which operate from a DC input voltage and which produce a controlled AC output voltage waveform or waveforms. A DC bus that serves as the input voltage to the inverter can be created by rectifying a fixed-frequency AC utility service, for example. Figure 24 shows a full-wave rectifier set operating from a three-phase utility connection. A DC output voltage with relatively low ripple is produced across the capacitor. If necessary, the level of the DC output voltage can be controlled by replacing the diodes with controllable devices, such as silicon controlled rectifiers. Controlling the firing angle of these devices permits control of the magnitude of the output voltage. Of course, a DC bus can be created in other ways. For a single-phase utility connection, either a single-phase, full-wave rectifier or a firing angle-controlled rectifier might be used. In an electric vehicle drive system, the DC bus would come from a battery rack, and no rectification would be required.

![Figure 25: Inverter.](image)

An inverter uses the DC bus to create AC waveforms with controllable frequency and amplitude. Figure 25 shows a typical inverter configuration driving a balanced, three-phase, inductive load.
This load could be the wye-connected stator of an induction or PM synchronous motor, for example. There are a variety of schemes for operating the switches Q1 – Q6 to produce desired AC waveforms. Typically, the ultimate goal of an inverter drive for an AC machine is to make the machine appear from an electrical port to be a current-controlled torque source, just like the PM DC machine. Two approaches for operating the switches in a three-phase inverter will be discussed briefly here; others may be found in [13] and especially [14]. The goal of this section is to reveal how useful AC waveforms can be produced by an inverter given a DC bus. The waveform analysis presented here is summarized from the excellent discussion in [13].

In a “six-step”, continuous current inverter, the controllable switches are operated as shown in the top six traces in Fig. 26. One leg or phase of the inverter consists of a top and bottom switch stacked together, e.g., Q1 and Q4. The stator connections \( a, b, \) and \( c \) can each be connected to either the top or the bottom of the DC input voltage by the switches. In any particular leg of the inverter, the two switches are never turned on at the same time in order to avoid shorting the input source. Also, in the continuous current inverter, one of the switches in each leg is always turned on to provide a path for phase current to flow. To emulate the behavior of a balanced, three-phase sinusoidal voltage source, the top or high-side switch in each leg is turned on 120 electrical degrees before the top switch in the next leg, and remains on for half of the electrical cycle.

When the high-side switch in a leg is on, the winding connected to that leg is connected to the top of the DC source. When the high-side switch is off, the low-side switch in the leg is on, and the winding is connected to the bottom of the DC source. The voltage between each stator terminal and point \( g \), therefore, has a waveshape that looks like the switch state for the high-side switch in that leg. For example, the voltage \( V_{ag} \) has a waveshape like the Q1 switch state trace in Fig. 26.

This fact can be used to determine the line-to-line and line-to-neutral voltages seen by the load. For example, the line-to-line voltage \( V_{ab} \) will have a waveshape that looks like the difference of the Q1 and Q2 waveshapes. This line-to-line voltage is plotted in the seventh trace in Fig. 26. To determine the line-to-neutral voltage for phase \( a \), notice that, for stator terminal \( a \), we can use Kirchoff’s voltage law to discover that

\[
V_{ag} = V_{as} + V_{sg},
\]

and so on for the other two phases. Because the inverter and load are balanced and have three phases, we know that

\[
V_{as} + V_{bs} + V_{cs} = 0.
\]

Therefore, the voltage

\[
V_{sg} = \frac{1}{3}(V_{ag} + V_{bg} + V_{cg}).
\]

Substituting Eq. 11 into Eq. 10 reveals that the \( a \)-phase line-to-neutral voltage is

\[
V_{as} = \frac{2}{3}V_{ag} - \frac{1}{3}(V_{bg} - V_{cg}).
\]

The waveshapes of \( V_{ag}, V_{bg}, \) and \( V_{cg} \) are identical to the Q1, Q2, and Q3 wave traces in Fig. 26. Equation 12 and the traces Q1, Q2, and Q3 are used to produce the line-to-neutral voltage waveform \( V_{as} \) for phase \( a \) shown in the last trace in Fig. 26. The six-step inverter produces a line-to-neutral voltage that has a substantial sinusoidal component at the fundamental frequency, with some obvious harmonic distortion present at higher, odd harmonics of the fundamental.
The frequency of the output waveforms can of course be changed by varying the time allotted to complete one electrical cycle. In a PM synchronous machine or “brushless DC motor”, the operation of the switches in the inverter is often “slaved” or synchronized to the rotor position, possibly by Hall-effect switches that sense the location of the rotor. This ensures that the AC waveform produced by the inverter will have a significant constant component when viewed in the rotor frame, as is necessary to sustain torque production. In a two-pole machine, for instance, the inverter would complete one electrical cycle for every revolution of the rotor. In essence, the inverter operates as an electrical commutator. The inverter can also be used to drive an induction machine. This drive could be “open-loop,” i.e., the inverter can provide the induction motor with a fixed-frequency, balanced voltage set. It could also be synchronized to the position of the rotor, as would be essential in the implementation of a field-oriented controller.

In the case of the PM synchronous machine, once the inverter operation is synchronized to the rotor position, the machine essentially behaves like a conventional PM DC machine from the standpoint of the DC input to the inverter. Raising the inverter input voltage will increase the speed of the machine. The current flowing out of the DC input indicates the level of torque produced at the shaft of the machine. It is important, therefore, to be able to control the magnitude of the voltage or current applied to the machine. This can be done in at least two ways. The first approach would be to vary the level of the DC input voltage to the inverter. This might be done either to vary the machine terminal voltage directly, or perhaps to control the current injected into the machine with a minor loop. The second, pulse-width modulation approach uses the inverter switches to chop the voltage applied to the stator. The stator voltages can always be set to zero by turning on all three high-side, or all three low-side, switches in the inverter (but never the high-side and low-side switches at the same time for a DC voltage input). The PWM switch frequency would be set significantly higher than the six-step electrical frequency. Varying the duty cycle will vary the average voltage applied to the stator terminals, again permitting voltage control or current control with a minor control loop.

References


