POWER CONVERTER THEORY

Modern switching AC/DC and DC/DC power converters represent the most recent fulfillment of the need for stable, well-regulated, direct current (dc) power for a variety of electronic instruments, equipment and systems. In the not too distant past, linear power converters were the mainstay of power conversion and voltage regulation. Switching power converters were still an infant technology usually relegated to low-volume, high-cost, commercial or military applications. Over the past 15 years, advances in technology have allowed modular power converter designers to take advantage of circuit configurations, components and materials simply not available prior to that time. These advances have allowed modular power converters to increase their electrical and thermal performance while decreasing their size and cost by utilizing switching power-conversion techniques. Consequently, modern modular power conversion has become a multi-billion dollar industry whose products have attained commodity status.

Linear Voltage Regulators

Linear regulators produce a stable dc output voltage from a less-than-stable dc input voltage. In normal operation, a linear regulator provides a constant output voltage regardless of anticipated variations in input (line) voltage and/or output (load) current. The regulator also attenuates any input voltage ripple variations, not only at their fundamental frequency but also at their first 5-10 harmonics. Lastly, the regulator provides a protection feature for the load in the form of current limiting.

Most linear regulators function as closed-loop servo systems. This operation is easily understood by referring to the typical circuit shown in Figure 1. In this circuit, transistor Q1 is the output-current pass element; A1 is the control-loop error amplifier; V_{REF} is a precision reference voltage source; R_S is the current-limit sense resistor; Q2 is the current-limit control transistor; and R1 and R2 are the loop gain-setting resistors. When the circuit is regulating, error amplifier A1's output responds as necessary to keep the difference between its + and – inputs at zero.

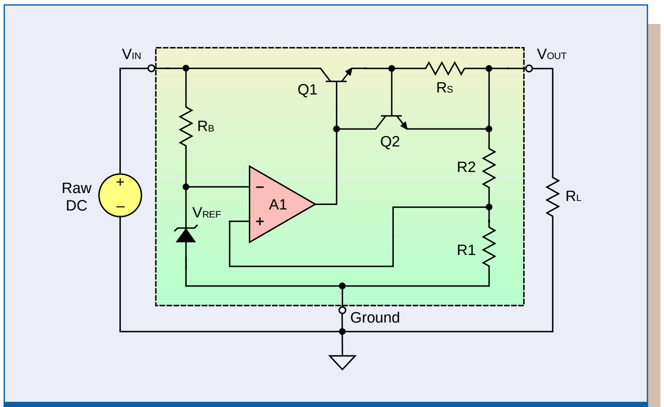


Figure 1. 3-terminal linear regulators produce a stable, regulated, dc output voltage from a higher-potential, perhaps unstable, dc input voltage.

A fraction of the regulator's output voltage (via the R1R2 voltage divider) is constantly compared to the reference (V_{REF}) at the error amplifier's + input. If the regulator's output voltage rises, A1's output voltage decreases. If the regulator's output voltage falls, A1's output voltage increases. It is imperative that the reference have excellent line, load and temperature-response characteristics.

When an excessive current is drawn from the output terminal, sufficient voltage develops across R_S to turn on transistor Q2, which in turn steals base current from Q1 and limits the maximum output current to approximately V_{BE}(for Q2 on)/R_S or 0.6V/R_S.

There are an impressive variety of high-performance linear regulators available today, both in fixed and adjustable-output configurations, as so-called "three-terminal" regulators. Regulators in this class include the LM78XX series positive regulators, the LM79XX series negative regulators, the LM317 positive adjustable regulator and the LM337 negative adjustable regulator.

Tutorial

The advantages of using linear regulators are low parts count, excellent regulation characteristics, low output noise, no generated EMI and excellent transient response. The disadvantages include single outputs, no “step-up” designs, low power density and poor efficiency.

Linear Regulator Efficiency

The efficiency of a linear regulator is defined as the ratio of the output power consumed by the load to the input power delivered to the regulator.

$$Eff. = P_{OUT}/P_{IN}$$

Where $P_{OUT} = V_{OUT} \times I_{OUT}$, $P_{IN} = V_{IN} \times I_{IN}$, $I_{IN} = I_{OUT} + I_Q$. I_Q is the quiescent (no-load) current required by the regulator during normal operation. Substituting values yields:

$$Eff. = (V_{OUT} \times I_{OUT})/(V_{IN}(I_{OUT} + I_Q))$$

If we consider a 5V, fixed-output, three-terminal regulator operating with an input voltage of 10Vdc, an output current of 1Adc and a quiescent current of 0.025Adc, the efficiency is:

$$Eff. = (5V \times 1A)/(10V \times 1.025A) = 0.49 \text{ or } 49\%.$$

If the input voltage is decreased to 7Vdc, the efficiency rises to 70%. It is obvious that the efficiency of this type of linear regulator is not constant over a range of input voltages. This means that adequate heat sinking may need to be provided for the regulator under the worst-case input-voltage and output-current conditions. This is why linear voltage regulators are sometimes referred to as “dissipative” regulators.

Other Linear Regulator Characteristics

Linear regulators not only offer many benefits, but many potential problems as well. Referring back to Figure 1, if the voltage on transistor Q1’s collector (i.e., the input voltage) becomes too low, insufficient drive current will be provided to Q1’s base. The result will be that Q1’s emitter voltage will begin to fall and reach an equilibrium point somewhere below the designated output regulation range. This is obviously not desirable, and the designer must guarantee that the input voltage is never allowed to go below the value of $V_{OUT} + (V_{IN} – V_{OUT})$ minimum, where $(V_{IN} – V_{OUT})$ minimum is defined as the “dropout voltage” or the minimum input-to-output voltage differential that must be maintained for proper operation.

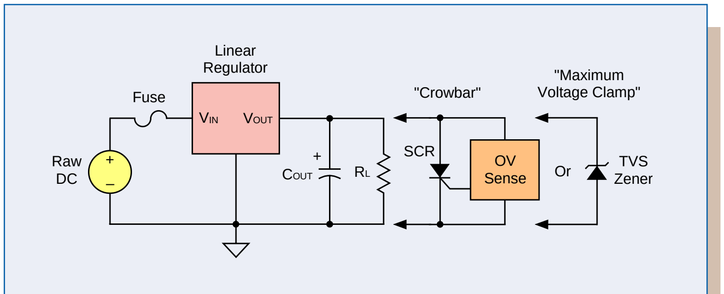


Figure 2. To protect the load against excessive voltage stress upon the failure of Q1, the regulator output voltage can be either shorted to ground or clamped.

For most monolithic voltage regulators, the dropout voltage is usually on the order of 2 Volts. This means that for a 5V regulator, the input voltage must never be allowed to drop below 7V, otherwise specified regulation performance cannot be guaranteed. There is a special class of linear regulators known as low dropout (LDO) devices whose minimum input-to-output differential can be as low as 0.5V. These devices, unfortunately, are less stable than their higher dropout counterparts, and they additionally require special external components and pc-board layouts to be properly implemented.

Tutorial

DATEL, Inc., 11 Cabot Boulevard, Mansfield, MA 02048-1151 (USA)
Tel: (508)339-3000, (800)233-2765 Fax: (508)339-6356 • Email: sales@datel.com • Internet: www.datel.com

Another potential problem with linear regulators is that, upon the failure of pass transistor Q1 (which usually manifests itself as a collector-to-emitter short circuit), the load can become short circuited to the input voltage. In most instances, this can quickly destroy the load or severely impair its operation. It is obviously not acceptable to have such severe consequences caused by the failure of a single element. This situation is remedied with the addition of a maximum-voltage-clamp (or “crowbar”) circuit at the output terminal of the regulator and the addition of a fuse in series with the input terminal, as shown in Figure 2.

To protect the load against excessive voltage stress caused by the failure of Q1, the output voltage can either be clamped to a maximum voltage excursion with a shunt current sink (which will only allow a maximum voltage of V_{OUT}(max.) to be present at the regulator output terminal) or sensed in such a way that if V_{OUT}(max.) occurs, an SCR (or similar latching current sink) is activated, effectively shorting (“crowbarring”) the output of the regulator to ground. In both cases, sufficient current will likely flow through the shorted regulator to open the protection fuse F1 and isolate the circuit from the power source.

Switching Voltage Regulators

Unlike linear dissipative regulators, switching voltage regulators exploit the energy-storage characteristics of passive magnetic and capacitive circuit elements. Unlike the linear regulator, whose pass element continuously transfers energy from the input voltage source to the load, the switching regulator takes discrete packets of energy from the input voltage source. It temporarily stores the energy as a magnetic field in an inductor or as an electric field in a capacitor and then transfers the energy to the load. A block diagram of a simplified switching regulator is shown in Figure 3.

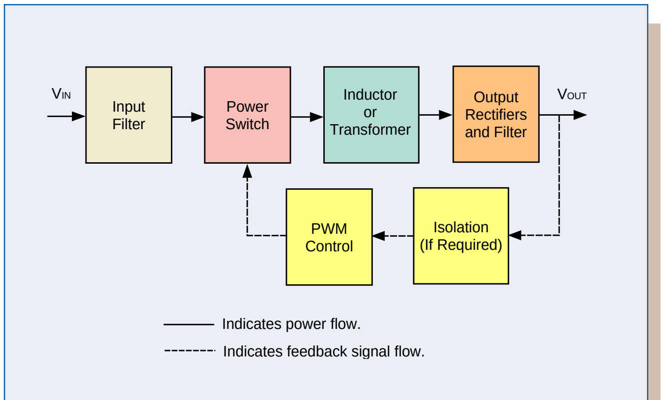


Figure 3. For switching regulators, the regulated output voltage is directly proportional to the PWM duty cycle.

The ability to perform the energy transfer from the source to the load in measured amounts requires a more complex control technique than utilized by the linear regulator. In the most common control technique, known as pulse width modulation (PWM), the packets of energy removed from the input power source are varied in duration, within a fixed operating period, as necessary to maintain an average energy transfer. The duty cycle (δ) of the PWM is the ratio of its “on” time (t_{ON}, the time during which energy is removed from the source) to the total switching period (T).

For switching regulators, the regulated output voltage is directly proportional to the PWM duty cycle, and the control loop utilizes the “large-signal” duty cycle as the controlling signal to the main power switch. This is quite different from the linear regulator which utilizes a “small-signal” dc servo loop to control the current through the main power pass element.

Switching Frequency and Size

The physical size of power-switching and energy-storage elements in switching DC/DC converters is directly affected by the operating frequency. The power coupled by a magnetic element is $P(L) = 1/2(LI^2f)$. As operating frequency rises, the inductance required to maintain a constant power decreases proportionally. Since inductance is related to the area of magnetic material and the number of turns of wire, the physical size of an inductor can be decreased .

Tutorial

The power coupled by a capacitive element is $P(C) = 1/2(CV^2f)$, so a similar size reduction can be realized for the energy-storage capacitors. These size reductions are very important to both the power supply designer and the system designer, as they allow the switching power supply to occupy less volume and printed-circuit-board (pcb) area.

Switching Converter Topologies

Topology refers to the various configurations of power-switching and energy-storage elements that can be used to transfer, control and regulate power (voltage) from an input voltage source. The many different switching-regulator topologies can be grouped into two basic categories: non-isolated, in which the input source and the output load share a common current path during operation, and isolated, in which the energy transfer is achieved by a mutually-coupled magnetic element (a transformer), and the coupling from the source to the load is achieved by means of a magnetic flux rather than a common current. One topology is selected over the other based upon the cost goals, performance objectives and input-line/ output-load characteristics of the system in which it is to operate. Any one topology is not “better” than another. Each has desirable characteristics and shortcomings, and selection is a matter of properly applying the correct power converter to the system requirement.

Non-Isolated Switching Converters

There are four basic, non-isolated, switching-regulator topologies applicable to modular DC/DC converters. They are: the buck, or step-down regulator; the boost, or step-up regulator; the flyback, or buck-boost regulator; and the Cuk. The following analyses assume a closed-loop feedback path is established with a PWM control circuit such that the proper duty cycle is achieved for a desired output voltage (V_{OUT}). Also, ideal power switching elements (power switches, diodes, capacitors and inductors) are assumed in determining the transfer characteristics of each topology.

The Buck Converter

The buck, or step down, switching regulator converts an input voltage to a regulated, lower-valued, output voltage. A simplified circuit diagram and associated operating waveforms are shown in Figure 4. In this circuit, with power switch S1 closed, inductor current (I_{L1}) builds up linearly at a rate approximately equal to $(V_{IN} - V_{OUT})/L1$. The inductor current flows into the load (R_L) and the output capacitor (C1).

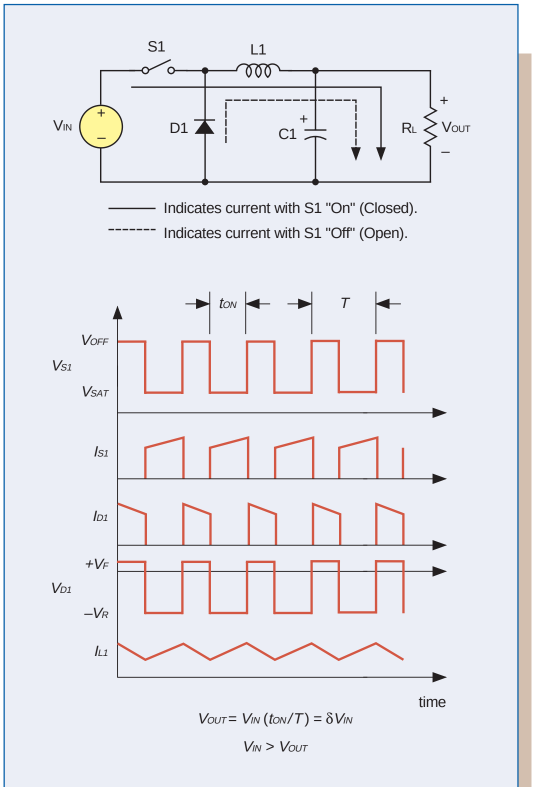


Figure 4. The buck, or step down, switching regulator converts an input voltage to a regulated, lower-valued, output voltage.

Tutorial

As the voltage across C1 rises out of the PWM's upper regulation tolerance window, S1 opens and terminates the current flow from the input voltage source. Even with S1 open, the current still flows in L1 and is "caught" by the rectifier diode D1. The inductor current now decreases linearly at a rate approximately equal to $V_{OUT}/L1$. When the voltage across capacitor C1 falls out of the PWM's lower regulation tolerance window, S1 closes and the cycle repeats.

The input-to-output transfer function can be derived by equating the inductor volt-second product (the energy) during the "on" time to the volt-second product during the "off" time. These two products should be equal in order to satisfy conservation of energy. The on-time energy is given by: $E_{ON} = (V_{IN} - V_{OUT})t_{ON}$, and the off-time energy is given by: $E_{OFF} = (V_{OUT})t_{OFF}$, where $t_{OFF} = T - t_{ON}$. Substituting yields:

$$(V_{IN} - V_{OUT})t_{ON} = V_{OUT}(T - t_{ON})$$

$$(V_{IN})t_{ON} = (V_{OUT})T$$

$$V_{OUT} = V_{IN}(t_{ON}/T),\ t_{ON}/T = \text{duty cycle } (\delta)$$

$$V_{OUT}/V_{IN} = \delta$$

The Boost Converter

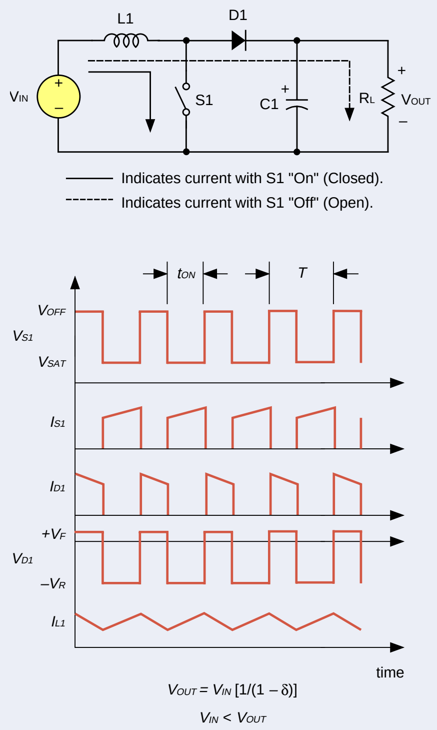


Figure 5. The boost, or step up, switching regulator converts an input voltage into a regulated, greater-valued, output voltage.

The boost, or step up, switching regulator converts an input voltage into a regulated, greater-valued, output voltage. A simplified circuit diagram and associated operating waveforms are shown in Figure 5. In this circuit, with power switch S1 closed, inductor current (I_{L1}) builds up linearly at a rate approximately equal to $V_{IN}/L1$. During this time, load current (through R_L) is drawn from output capacitor C1. As the voltage across C1 drops out of the PWM's lower regulation tolerance window, S1 opens and stops shunting input current. Even with S1 open, the current still flows in L1 and is "caught" by rectifier diode D1. The current is now steered into the load circuit (C1 and R_L). The inductor current now decreases linearly at a rate approximately equal to $(V_{OUT} - V_{IN})/L1$, and when the voltage across capacitor C1 rises above the PWM's upper regulation tolerance window, S1 closes and the cycle repeats.

For the input-to-output transfer function, the on-time energy is given by: $E_{ON} = (V_{IN})t_{ON}$, and the off-time energy is given by: $E_{OFF} = (V_{OUT} - V_{IN})t_{OFF}$, where $t_{OFF} = T - t_{ON}$. Substituting yields:

$$(V_{IN})t_{ON} = (V_{OUT} - V_{IN})(T - t_{ON})$$

$$(V_{IN})T = V_{OUT}(T - t_{ON})$$

$$V_{OUT} = V_{IN}(T/(T - t_{ON}))$$

$$V_{OUT} = V_{IN}(1/(1 - \delta))$$

$$V_{OUT}/V_{IN} = 1/(1 - \delta)$$

The Flyback Converter

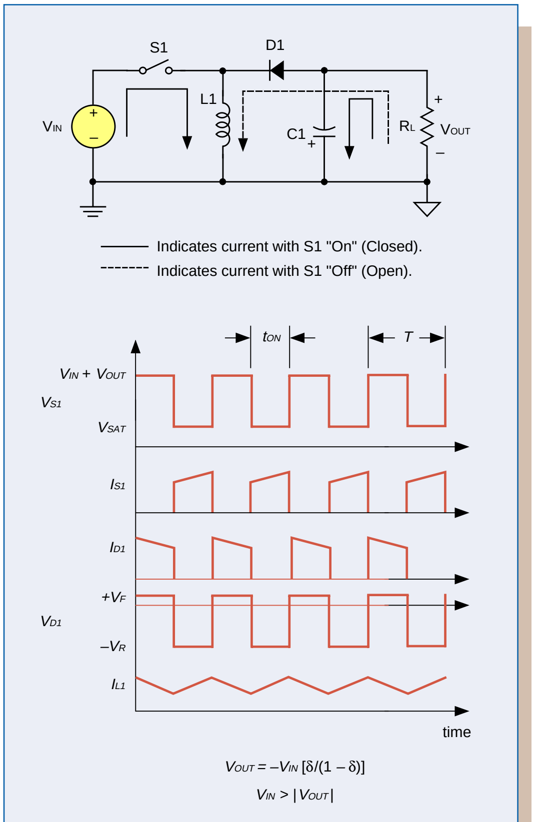


Figure 6. The flyback, or buck-boost, switching regulator converts an input voltage into a regulated, inverted, lower-valued output voltage.

The flyback, or buck-boost, switching regulator converts an input voltage into a regulated, inverted, lower-valued output voltage. A simplified circuit diagram and associated operating waveforms are shown in Figure 6. In this circuit, with power switch S1 closed, inductor current (I_{L1}) builds up linearly at a rate approximately equal to V_{IN}/L1. During this time, load current (through R_L) is drawn from output capacitor C1. As the voltage across C1 drops out of the PWM's lower regulation tolerance window, S1 opens and terminates the current flow from the voltage source. After S1 opens, current still flows in L1 and is "caught" by rectifier diode D1. The current is now steered into the load circuit (C1 and R_L). The inductor current begins to decrease linearly at a rate approximately equal to V_{OUT}/L1. When the voltage across capacitor C1 rises above the PWM's upper regulation tolerance window, S1 closes and the cycle repeats. Please note that because of the direction of current flow in L1, the resulting output voltage will be negative with respect to ground. Thus, a negative output voltage is obtained from a positive input.

Concerning the input-to-output transfer function, the on-time energy is given by: $E_{ON} = (V_{IN})\,t_{ON}$, and the off-time energy is given by: $E_{OFF} = (-V_{OUT})\,t_{OFF}$, where $t_{OFF} = T - t_{ON}$. Substituting yields:

$$(V_{IN})\,t_{ON} = -V_{OUT}\,(T - t_{ON})$$

$$V_{OUT} = (-V_{IN})(t_{ON}/(T - t_{ON}))$$

$$V_{OUT} = -V_{IN}\,(\delta/(1 - \delta))$$

$$V_{OUT}/V_{IN} = -\delta/(1 - \delta)$$

The Cuk Converter

The Cuk (pronounced "Chook") switching regulator converts an input voltage into a regulated, inverted, lower-valued or higher-valued output voltage depending upon the duty cycle. A simplified circuit diagram and associated operating waveforms are shown in Figure 7. In this circuit, with power switch S1 closed, inductor current (I_{L1}) builds up linearly at a rate approximately equal to V_{IN}/L1. Also during this time, inductor current (I_{L2}) flows "through" capacitor C1 and through the load circuit (C2 and R_L).

Note that capacitor C1 provides an important function in this circuit as it is the chief means of energy transfer from the input to the output. The value of C1 is chosen such that the voltage across it in the steady-state is essentially constant.

As the voltage across C2 rises out of the PWM's upper regulation tolerance window, S1 opens and terminates the current flow from the voltage source. When S1 opens, the current still flows in L1 and is "caught" by transfer capacitor C1 and rectifier diode D1. The inductor current now decreases linearly at a rate approximately equal to ($V_{IN} - V_{C1}$)/L1. During this time, the current flowing in L2 is "caught" by rectifier diode D1, and this current decreases linearly at a rate approximately equal to V_{OUT}/L2. When the voltage across capacitor C2 falls below the PWM's lower regulation tolerance window, S1 closes and the cycle repeats. Note that because of the direction of current flow in L1, the resulting output voltage will be

Tutorial

DATEL, Inc., 11 Cabot Boulevard, Mansfield, MA 02048-1151 (USA)
Tel: (508)339-3000, (800)233-2765 Fax: (508)339-6356 • Email: sales@datel.com • Internet: www.datel.com

negative with respect to ground. Thus, a negative output voltage is obtained from a positive input.

Concerning the input-to-output transfer function, the Vt product must be considered for both inductors in the circuit. For L1, the on-time energy is given by: $E_{ON} = (V_{IN})t_{ON}$, and the off-time energy is given by: $E_{OFF} = (V_{C1} - V_{IN})t_{OFF}$, where $t_{OFF} = T - t_{ON}$. For L2, the on-time energy is given by: $E_{ON} = (V_{C1} + V_{OUT})t_{ON}$, and the off-time energy is given by $E_{OFF} = (-V_{OUT})t_{OFF}$, where $t_{OFF} = T - t_{ON}$. Substituting and solving for V_{C1} yields:

$$(V_{IN})t_{ON} = (V_{C1} - V_{IN})(T - t_{ON})$$

$$V_{C1} = (V_{IN})(T/(T - t_{ON}))$$

$$V_{C1} = V_{IN}(1/(1 - \delta)), \text{ and}$$

$$(V_{C1} + V_{OUT})t_{ON} = -V_{OUT}(T - t_{ON})$$

$$V_{C1} = (-V_{OUT}T)/t_{ON}$$

$$V_{C1} = -V_{OUT}/\delta$$

Equating the expressions yields:

$$-V_{OUT}/\delta = V_{IN}(1/(1 - \delta))$$

$$V_{OUT}/V_{IN} = -\delta/(1 - \delta)$$

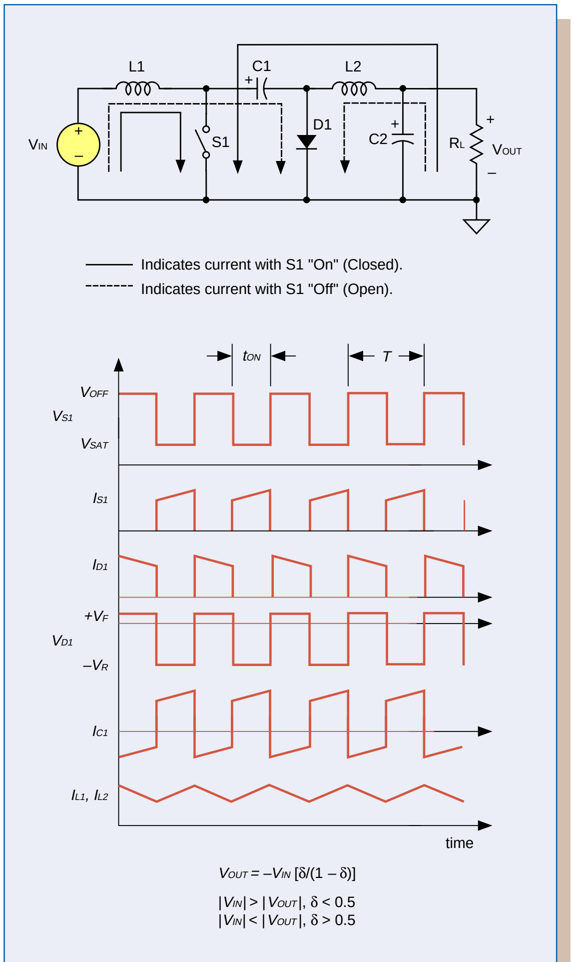


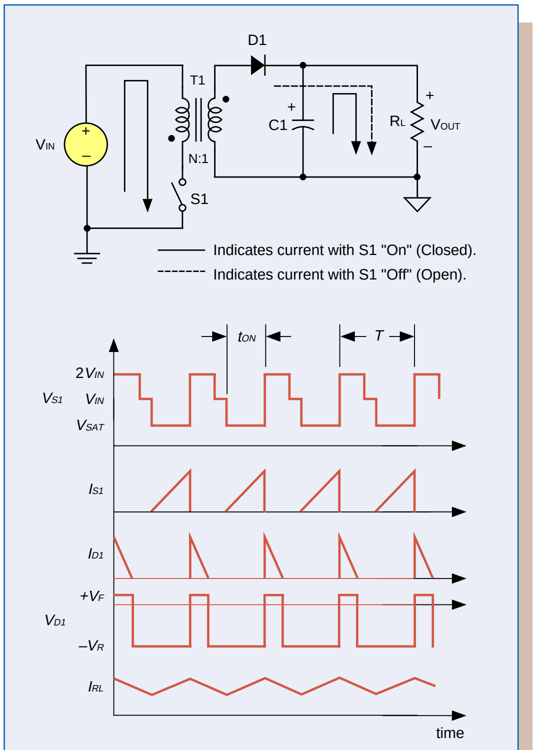
Figure 7. The Cuk switching regulator converts an input voltage into a regulated, inverted, lower-valued or higher-valued output voltage depending on the duty cycle.

Isolated Switching Converters

There are many isolated switching converter topologies, however, there are but three that are applicable to a discussion of modern modular DC/DC converters. They are the flyback, forward and push-pull power converters. For these circuits, all energy transfer from the input power source to the load is achieved via a transformer or other flux-coupled magnetic element. As was the case for the non-isolated converters, the presence of a PWM control circuit to maintain the proper duty cycle and the use of ideal power-switching elements (power switches, diodes, capacitors, inductors, etc.) is assumed.

The Flyback Converter

The flyback switching regulator converts an input voltage into a regulated, lower or higher-valued output voltage depending on its transformer's turns ratio. A simplified circuit diagram and associated operating waveforms are shown in Figure 8. In this circuit, with power switch S1 closed, primary current (I_{S1}) builds up linearly at a rate approximately equal to V_{IN}/ L(primary). During this time, because of the phasing of the transformer (more aptly



$V_{OUT} = V_{IN}(1/N)[\delta/(1 - \delta)]$

$V_{IN} > V_{OUT}$ or $V_{IN} < V_{OUT}$

Figure 8. The flyback switching regulator converts an input voltage into a regulated, lower or higher-valued output voltage.

referred to as a coupled inductor), no energy (current) is supplied to the load by the transformer secondary. Load current, during this interval, is supplied by C1.

As the voltage across C1 falls below the PWM's lower regulation tolerance window, S1 opens and terminates the current flow from the voltage source. When S1 opens, the magnetic field in the transformer collapses, and the voltages at the primary and secondary undergo a polarity reversal. The energy in the primary is now available to the secondary, and secondary current flows and decreases linearly at a rate approximately equal to V_{OUT}/L(secondary). When the voltage across C1 rises above the PWM's upper regulation tolerance window, S1 closes and the cycle repeats.

Concerning the input-to-output transfer function, the on-time energy is given by: $E_{ON} = (V_{IN}/N)\,t_{ON}$, and the off-time energy is given by: $E_{OFF} = (V_{OUT})\,t_{OFF}$, where $t_{OFF} = T - t_{ON}$ and N is the transformer's turns ratio. Substituting yields:

$$(V_{IN}/N)t_{ON} = V_{OUT}(T - t_{ON})$$

$$(V_{IN})\,t_{ON} = NV_{OUT}(T - t_{ON})$$

$$V_{OUT} = (V_{IN})\,t_{ON}/(NT - Nt_{ON}),\ t_{ON}/T = \text{duty cycle } (\delta)$$

$$V_{OUT}/V_{IN} = \delta/(N - N\delta)$$

$$V_{OUT}/V_{IN} = (1/N)(\delta/(1 - \delta))$$

The Forward Converter

The forward switching regulator converts an input voltage into a regulated, lower or higher-valued output voltage depending on its transformer's turns ratio. A simplified circuit diagram and associated operating waveforms are shown in Figure 9. In this circuit, with power switch S1 closed, transformer primary current I_{S1} increases linearly at a rate approximately equal to V_{IN}/L(primary). Also during this time, due to the input-to-output winding phasing of T1, a voltage whose magnitude is V_{IN}/N is present at the transformer's secondary winding. The secondary current (which flows through rectifier diode D1 and output inductor L1) increases linearly at a rate approximately equal to (V_{IN}/N)/L1. This current also flows into the load R_L and the output capacitor C1.

As the voltage across C1 rises out of the PWM's upper regulation tolerance window, S1 opens and terminates the current flow from the voltage source. When S1 opens, the secondary winding undergoes a voltage phase reversal and no longer provides current through D1. However, current still flows in L1 and is "caught" by rectifier diode D2. The inductor current now decreases linearly at a rate approximately equal to V_{OUT}/L1, and when the voltage across capacitor C1 falls out of the PWM's lower regulation tolerance window, S1 closes and the cycle repeats.

Tutorial

DATEL, Inc., 11 Cabot Boulevard, Mansfield, MA 02048-1151 (USA)
Tel: (508)339-3000, (800)233-2765 Fax: (508)339-6356 • Email: sales@datel.com • Internet: www.datel.com

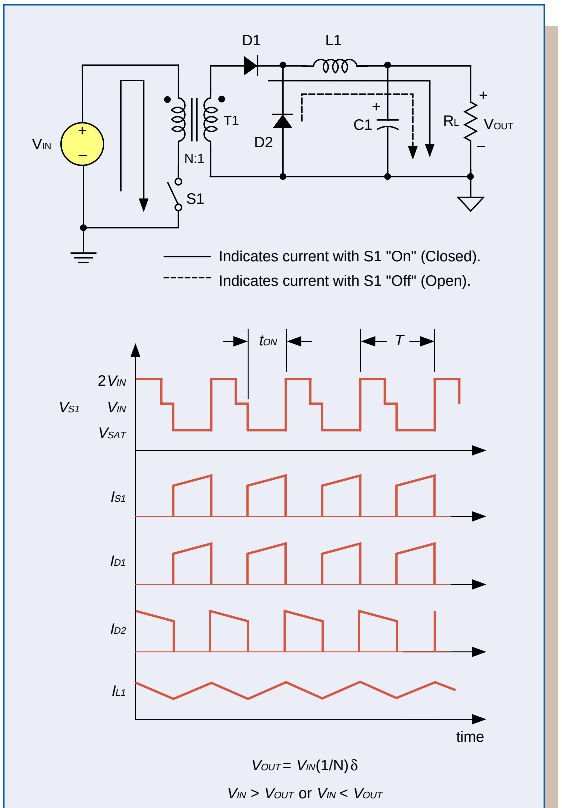


Figure 9. The forward switching regulator converts an input voltage into a lower or higher-valued output depending on the transformer turns ratio.

In determining the input-to-output transfer function, the on-time energy is given by: $E_{ON} = (V_{IN}/N - V_{OUT})\, t_{ON}$, and the off-time energy is given by: $E_{OFF} = (V_{OUT})\, t_{OFF}$, where $t_{OFF} = T - t_{ON}$ and N is the transformer's turns ratio. Substituting yields:

$$(V_{IN}/N - V_{OUT})\, t_{ON} = V_{OUT}(T - t_{ON})$$

$$V_{OUT} = (V_{IN}/N)(t_{ON}/T)$$

$$V_{OUT}/V_{IN} = (1/N)\delta$$

The Push-Pull Converter

The push-pull switching regulator converts an input voltage into a regulated, lower-valued output voltage. A simplified circuit diagram and associated operating waveforms are shown in Figure 10. With power switch S1 closed, transformer primary current I_{S1} increases linearly at a rate approximately equal to V_{IN}/L(primary). Also during this time, due to the input-to-output winding phasing of T1, a voltage whose magnitude is V_{IN}/N is present at the transformer's secondary winding. The secondary current (which flows through rectifier diode D1 and output inductor L1) increases linearly at a rate approximately equal to $(V_{IN}/N - V_{OUT})/L1$. This current also flows into the load R_L and the output capacitor C1.

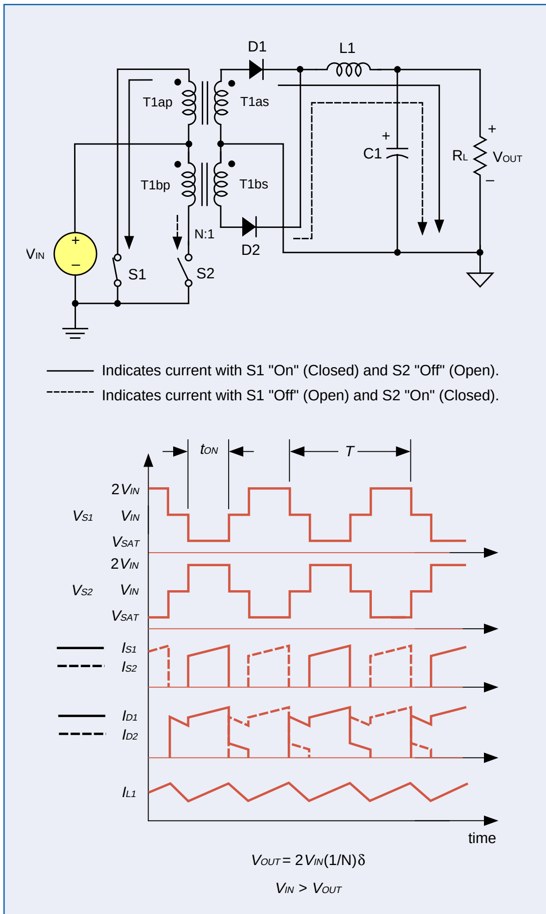


Figure 10. In the push-pull converter, S1 and S2 both have a 50% duty cycle, and switching must be "break-before-make."

Tutorial

As the voltage across C1 rises out of the PWM's upper regulation tolerance window, S1 opens and terminates the current flow from the voltage source. When S1 opens, the secondary winding undergoes a voltage phase reversal and no longer provides current through D1. However, current still flows in L1 and is "caught" by rectifier diode D2. The inductor current now decreases linearly at a rate approximately equal to $V_{OUT}/L1$, and when the voltage across capacitor C1 falls out of the PWM's lower regulation tolerance window, S2 closes and the cycle repeats.

For the input-to-output transfer function, the on-time energy per switch closure is given by: $E_{ON} = (V_{IN}/N - V_{OUT})t_{ON}$, and the off-time energy per switch closure is given by: $E_{OFF} = (V_{OUT})t_{OFF}$, where $t_{OFF} = (T/2 - t_{ON})$ and N is the transformer's turns ratio. The value $T/2$ is used because each transistor turns on once per PWM clock cycle, so each switch closure provides energy for $T/2$ of the total period T. Substituting yields:

$$(V_{IN}/N - V_{OUT})t_{ON} = (V_{OUT})(T/2 - t_{ON})$$

$$V_{OUT} = 2V_{IN}/N(t_{ON}/T)$$

$$V_{OUT}/V_{IN} = (2/N)\delta$$

Please note that the maximum duty cycle for each switch (S1, S2) is 50%, and it is very important to guarantee that both switches cannot turn on simultaneously. This would cause dangerous levels of current to flow into an effectively shorted transformer and cause the destruction of the power switches. Therefore, an adequate "dead time" must be implemented between the opening of the "on" switch and the closure of the "off" switch.

There is another problem that plagues the push-pull architecture, that being "flux walk." Because the push-pull converter utilizes the full bipolar B-H characteristic of the power transformer, any subtle differences in the performance of the two power switches (on-state voltage, switching times, etc.) can result in the application of a dc flux bias (flux asymmetry) to the transformer. With each switching cycle, this flux biasing becomes a cumulative event because the transformer cannot fully reset its flux bias to zero, and the previous dc bias becomes the starting point for the next B-H curve excursion. The transformer eventually saturates, normal energy transfer becomes impaired, and one or both of the power switches will be destroyed by high current levels flowing in the saturated primary since a saturated transformer primary no longer exhibits inductive characteristics.

The Efficiency of Switching Power Converters

Determining the efficiency of a switching power converter is more complicated than for a linear voltage regulator. The linear regulator has a fixed set of dc losses, and most of the power dissipation is in the pass transistor. A switching regulator not only has dc losses but also ac losses in the switching elements and in the energy-storage elements as well.

For example, the total power-switch losses are the sum of the on-state power dissipation, the switching losses during turn-on and turn-off, and the off-state power dissipation. In the case of a transformer, the total power loss is the sum of the ac (core) loss and the dc (Ohmic) loss. The transformer core loss is primarily related to the magnetic flux density and the core material. The dc losses are a result of the I^2R losses in the wires making up the windings. In order to calculate the converter efficiency, the losses for each circuit element are totaled and then averaged over a PWM conversion cycle (T).

Switching power converters exhibit high efficiencies because the power-switching elements are on for only a fraction of the total switching cycle (they have high peak power but low average power) and because the losses in the magnetic and capacitive elements can be controlled and minimized (through careful magnetics design and material selection). DC/DC power converter efficiencies in the range of 75% to 90% are readily achievable with switching designs. Non-isolated topologies tend to be more efficient because they have fewer power handling components and also do not have the losses of the power transformer. Even with their added complexities, isolated converters are still able to achieve efficiencies in the 80% and greater range.

Switching Converter Components

The particular components used within switching power converters greatly affect their performance. The selection of the switching and rectifying elements, the magnetic components and the filter capacitors greatly influences the switching frequency and the efficiency of the converter. Throughout all the previous analyses, the power switches, rectifier diodes, transformers, inductors and capacitors were all considered ideal elements. Real-life elements have parasitic characteristics that the power converter circuit designer must be familiar with and make accommodations for.

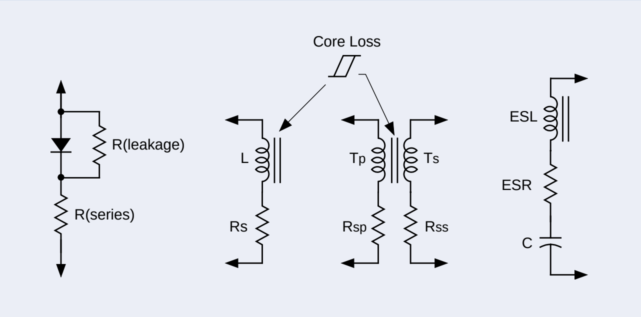


Figure 11. Real-world diodes, inductors, transformers and capacitors all have parasitic elements that degrade efficiency.

Real-world semiconductor power switches all have non-ideal operating characteristics. These include drive requirements, switching speed losses, on-state losses and off-state losses. Similarly, rectifier diodes have on and off-state losses. Magnetic elements have core and conductor losses. Filter capacitors have parasitic effective series inductance (ESL) and effective series resistance (ESR) which contribute to losses (see Figure 11).

PWM Control Techniques

There are two methods for controlling the feedback loop and regulation characteristics of a switching DC/DC power converter: voltage-mode control, in which the duty cycle of the converter is proportional to the error differential between the actual and ideal output voltages; and current-mode control, in which the duty cycle is proportional to the error differential between the nominal output voltage and a function of the converter's controlling current. The controlling current can be either the switch current in a non-isolated topology or the transformer primary current in an isolated topology.

Voltage-mode control responds (and adjusts the duty cycle of the converter) only to changes in output (load) voltage. This means that in order for the converter to respond to changes in load current or input line voltage, it must "wait" for a corresponding change in load voltage (load regulation). This wait/delay affects the regulation characteristics of the converter in that the "wait" is typically one or more switching cycles. Depending upon the load or line perturbation, there will be a corresponding (though not necessarily proportional) output voltage perturbation.

A typical voltage-mode PWM control circuit is shown in Figure 12. In this circuit, A1 is the loop error amplifier, A2 is the PWM comparator, and A3 is an optional output driver (to interface with the power switch). The ramp oscillator provides an output voltage V_{OSC} which linearly ramps in value from 0V to some maximum value (corresponding to the maximum duty cycle) during a converter switching cycle, t_S. The error amplifier A1 compares the difference between a precision, temperature-compensated reference (V_{REF}) and a fraction of the converter output voltage, V_{OUT}(R2/(R1 + R2)).

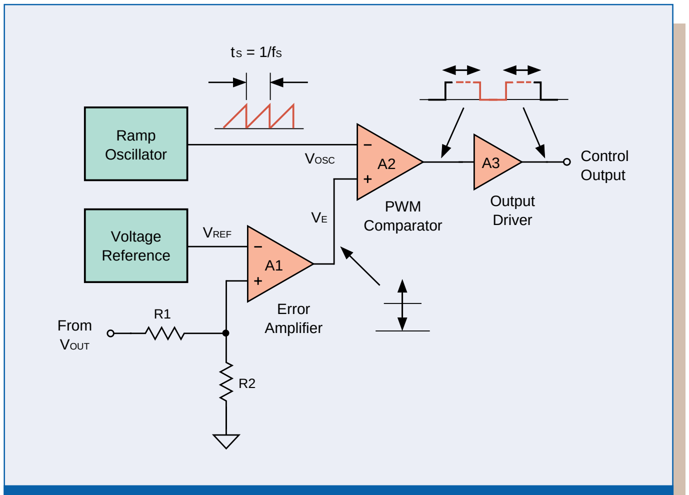


Figure 12. Voltage-mode PWM control can not respond to changes in line voltage or load current until they manifest themselves as a change in output voltage.

The output of A1 (in the form of V_E) assumes a value proportional to the difference between the reference voltage and V_{OUT} . If the output voltage is zero, the output of A1 is at it's maximum value, which is the same value as the oscillator's output ramp maximum value. With this condition at the input to the PWM comparator (A2), the output voltage of A2 will remain at it's maximum value for the entire converter switching cycle period. Therefore, when V_{OUT} is at it's minimum, the duty cycle is at it's maximum.

Now, if the actual output voltage exceeds the regulation range of V_{OUT}, the output of A1 will be at (or near) zero. Under this condition, the output of A2 remains at it's minimum value for the entire switching-cycle period. It is this inverse relationship between the output voltage and the converter duty cycle (i.e. output voltage too low yields maximum duty cycle and output voltage too high yields minimum duty cycle) that provides the stable feedback mechanism for the power converter's control loop.

If there could be a mechanism whereby the PWM control could respond to changes in the load current within a single conversion cycle, the problem of the "wait" and the respective load-regulation penalty associated with voltage-mode control could be eliminated. This is possible with current-mode control.

Current-mode control divides the power converter into two control loops ... current control via an inner control loop and voltage control via an outer control loop. The result is that changes in not only load voltage but also load current can be responded to on a switching pulse-by-pulse basis. A typical current-mode PWM control circuit is shown in Figure 13. In this circuit, A1 is the voltage loop error amplifier, A2 is the PWM comparator, and A3 is an optional output driver (to interface with the power switch). The oscillator provides essentially narrow synchronization pulses at the switching frequency, f_S. It sets the output (Q) of the PWM latch (G1) to logic high and identifies the start of another conversion cycle.

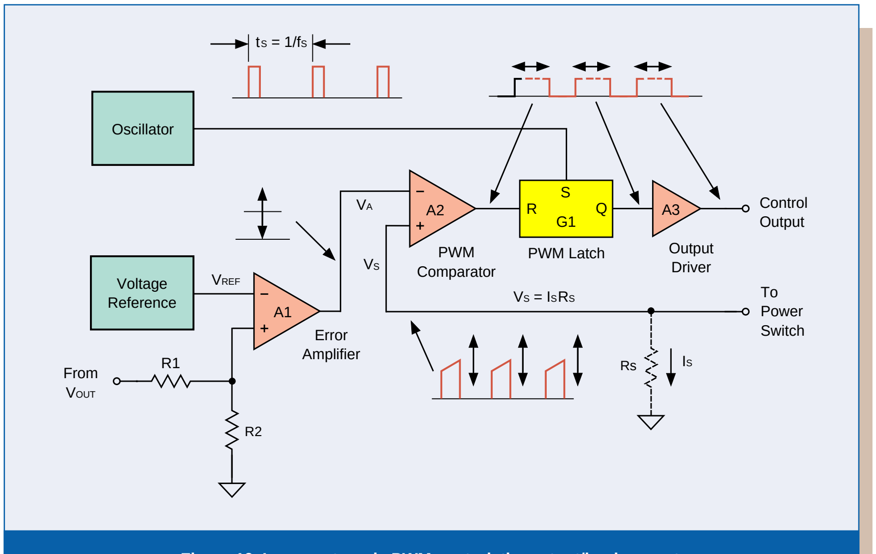


Figure 13. In current-mode PWM control, the output/load current is controlled on a pulse-by-pulse basis.

As was the case in voltage-mode control, the error amplifier A1 compares the difference between a precision, temperature-compensated reference (V_{REF}) and a fraction of the converter output voltage, V_{OUT}(R2/(R1 + R2)). The output of A1 assumes a value proportional to the difference between the reference voltage and V_{OUT}.

DATEL, Inc., 11 Cabot Boulevard, Mansfield, MA 02048-1151 (USA)
Tel: (508)339-3000, (800)233-2765 Fax: (508)339-6356 • Email: sales@datel.com • Internet: www.datel.com

If the output voltage is zero, the output of A1 is at it's maximum value. If the output voltage exceeds the regulation range of V_{OUT}, the output of A1 will be at (or near) zero. Therefore, when the converter output is regulating, the output of A1 assumes some average value (V_A) between the maximum and the minimum values. This value is present at the inverting input to the PWM comparator A2 and essentially becomes the reference for the current feedback signal.

Note that A2's output voltage is at it's minimum value (logic low) if the voltage at its – input is greater than the voltage at the + input. Assuming that the switch or primary current is sensed by a resistor R_S, the voltage present at the + input of A2 will be $I_S R_S$, which is a voltage (V_S) proportional to the switch current. When the value of V_S reaches the value of V_A, the output of A2 will switch to it's maximum value (logic high) and reset the PWM latch G1, causing its output to switch to a logic low. This action determines the length of time, within the total switching period, that the output of G1 is high, and defines the duty cycle of the converter.

Thus, current-mode control shares the same inverse relationship between output voltage and duty cycle as does voltage-mode control, but it also has the advantage that the outer (voltage) control loop sets the threshold at which the inner (current) loop regulates the peak current in the switch or the primary circuit. Since the output current is proportional to the switch or primary current, the output current is controlled on a pulse-by-pulse basis, and it exhibits superior line and load-regulation characteristics over voltage-mode control.

Power Converter Regulation

Although both voltage-mode and current-mode control schemes afford good-to-excellent output voltage regulation, performances can degrade if proper design features are not incorporated into the converter. When a DC/DC power converter, especially a high-output-current model, is applied in an end-use situation, the performance can be degraded by the series resistance present in the output interconnect pins. Additionally, the manner in which the load is connected to the converter can greatly affect the in-application load regulation performance. If each of the package pins has a series resistance R_S, the output voltage at the load will be reduced by an amount $2(I_{OUT})R_S$ Volts as illustrated in Figure 14. This decrease in output voltage depends upon an unspecified parameter ... the resistance of the output pins. If the load is not located physically close to the power converter, there will be

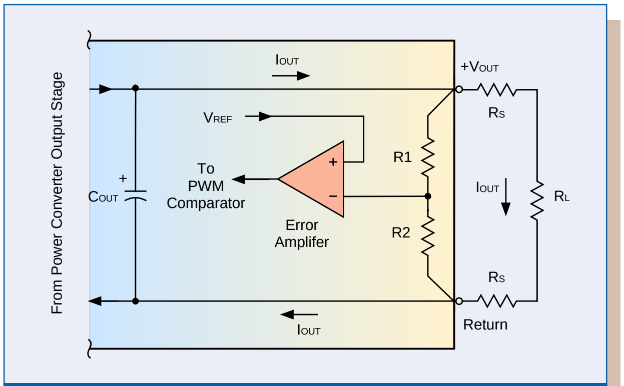


Figure 14. In high-current applications, the series resistance of package pins can adversely affect output voltage regulation.

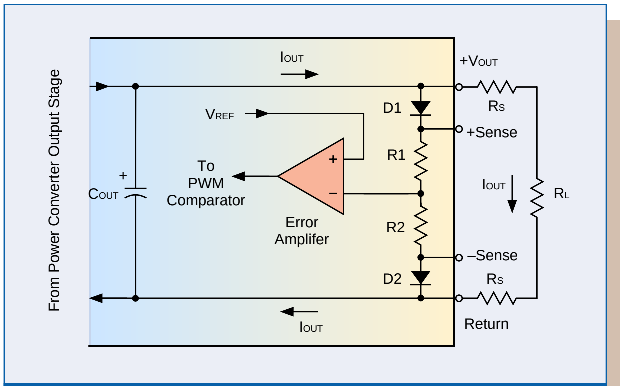


Figure 15. Kelvin or "4-wire" sensing can eliminate the adverse effects of both package-pin and interconnect series resistances.

Tutorial

further unwanted resistance in the form of wiring or pc-etch resistance which will further degrade the load regulation.

This load-regulation problem can be corrected for by the power converter designer if the feedback loop (including ground) is left unconnected within the converter package, and feedback "sense" connections are provided as user-connected input pins, as shown in Figure 15. This scheme, also referred to as "Kelvin" or "4-wire" sensing, provides a valuable feature to the end-user and permits higher-power modular power converters to enjoy excellent load regulation similar to that of lower-power models.

Power Converter Protection

Most power converters require some form of protection, usually to limit the maximum output current or the maximum output power. Otherwise, there is a risk of destroying the converter with a momentary output overload or overstress. The most common form of protection is current limiting, in which the maximum output current deliverable by the converter is electronically limited to a safe value. This value is typically 110% to 150% of the converter's maximum specified/rated load current.

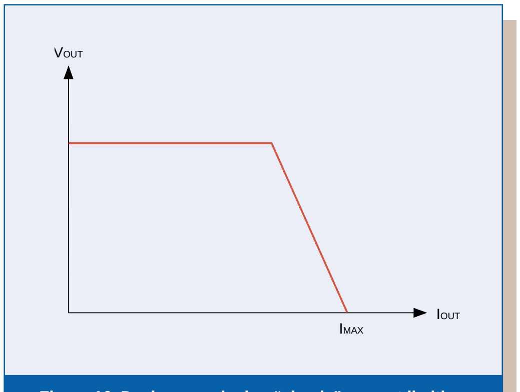


Figure 16. Devices employing "simple" current limiting may only have temporary protection against sustained short circuits.

The output characteristic for simple current limiting is shown in Figure 16. This form of simple current limiting has its limitations. If the power converter's output is subjected to a continuous short circuit to ground, the current-limited value of output current continues to flow. This continuous generation of the short-circuit current causes the converter to dissipate considerably more power than during normal operation. If the converter is to survive a short-circuit event undamaged, it must be properly designed or have adequate heat sinking such that the maximum specified case temperature (or component junction temperatures) is never exceeded.

An alternative to the simple current limiting scheme is foldback current limiting. This technique allows the output current to reach some maximum value, I_{MAX} (typically 110% to 120% of the maximum specified/rated output current). Then, as the output voltage decreases (towards ground), the output current is driven back linearly to some lower short-circuit value, I_{SC}. This lower short-circuit current enables the converter to dissipate a more reasonable power level during the short-circuit condition. The foldback output characteristic is show in Figure 17.

Another protection technique is switched-mode current limiting or "hiccup" control. In this approach, when the maximum current-limit threshold is exceeded, the output current is interrupted for a period of time t then reapplied for a time period $(T - t)$. The output current can be thought of as "hiccuping," as it no longer exhibits a steady-state output characteristic. This action associates a duty cycle $(t/(T - t))$ with the current-limit event, and the duty cycle can be chosen such that the maximum converter power dissipation is not exceeded during the current-limit interval.

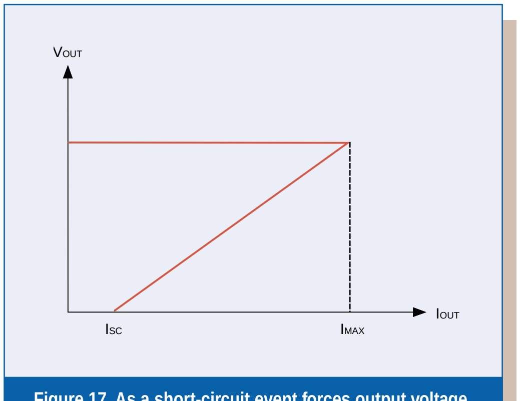


Figure 17. As a short-circuit event forces output voltage to fall, foldback current limiting drives the output short-circuit current down to a sustainable level.

Tutorial

DATEL, Inc., 11 Cabot Boulevard, Mansfield, MA 02048-1151 (USA)
Tel: (508)339-3000, (800)233-2765 Fax: (508)339-6356 • Email: sales@datel.com • Internet: www.datel.com

Two additional protection features often used in conjunction with one another are undervoltage lock-out (UVLO) and converter soft start. UVLO is a fail-safe function which disables the power converter if the input voltage drops below some minimum specified value. A UVLO feature ensures that the maximum duty cycle of the PWM circuit will be limited to a safe value. Oftentimes, there is a hysteresis (several Volts) incorporated with the UVLO turn off and turn on such that threshold noise is eliminated.

The soft-start feature is utilized as a means of controlling the input current and load voltage/current when the converter first turns on (perhaps following a UVLO event). In the absence of soft start, the converter's duty cycle is set to the maximum by the PWM controller, and the output voltage rises at its fastest possible rate. This dV_{OUT}/dt may cause large currents to flow in the converter's output capacitor (and any other distributed load capacitance) which in turn can cause a large "inrush" current into the converter. It can also lead to false triggering of current-limiting protection circuitry. With a soft-start feature, the duty cycle is forced to start at it's minimum value and allowed to linearly ramp-up over time to its nominal operating duty cycle. This action causes the output voltage to linearly rise with time until the nominal output voltage value is achieved.

Converter Input Range Constraints

The working input voltage range of a given DC/DC power converter is constrained by the converter's circuit topology and the components used. The input voltage to the converter and the duty cycle are inversely proportional. An increase in input voltage yields a corresponding decrease in the duty cycle. The minimum duty cycle is determined by the peak current-carrying capability and/or the maximum off-state withstanding voltage of the power switch. When the duty cycle is low/short, the peak-to-average input current is high, and in most topologies, the peak off-state voltage of the power switch is at it's highest value. Theoretically, all switching power converters can operate down to 0% duty cycle. In practice, the minimum duty cycle is limited to 5-10%. This means there is also a corresponding limit to the maximum converter input voltage.

The maximum duty cycle is determined by the maximum current capability of the power switch and/or the saturation characteristics of the transformer or other power magnetic components. When the duty cycle is high/long, the peak-to-average input current is low, and the average input current is at it's maximum value. Also, the power magnetic elements require a finite amount of time to reset the magnetic flux within their cores, otherwise the cores saturate. Therefore, though all converters can theoretically operate with duty cycles up to 100%, in practice the maximum duty cycle is limited to 85-90%. Accordingly, this imposes a lower limit on the converter's input voltage.

The duty cycle limitations usually determine a power converter's maximum input voltage range, and this range is sometimes defined as the ratio of the high to the low value. For example, the isolated forward converter topology can support an input voltage range of 4:1 (e.g. 18-72V around a nominal value of 48V) when careful design techniques are applied.

Filtering and Noise

An important consideration in the design and application of switching power converters is electrical noise in the forms of electro-magnetic and radio-frequency interference (EMI and RFI). All switched-mode power converters generate EMI and RFI noise as a result of their operation. The noise is not necessarily the artifact of their operation at some fixed switching frequency (typically between 50kHz and 1MHz). Rather, it results because switching converters rely on large voltage swings at extremely high rates of change (dV/dt) which inadvertently produce very high-frequency voltage spikes at the leading and trailing edge of the power switching event. Therefore, although the noise frequency spectrum has the switching frequency fundamental as its major contributor, the noise is quite broad band ranging from dc to several hundred MHz.

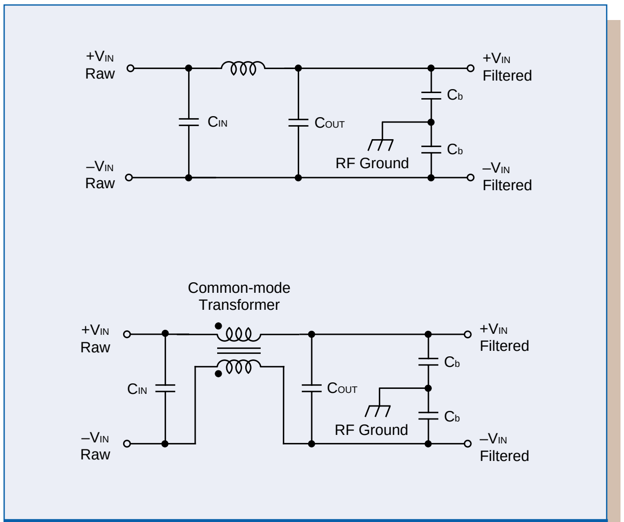


Figure 18. The effects of a power converter's reflected ripple current on the rest of the power system can be greatly reduced by single or double-ended input "pi" filters.

These voltage slew rates and spikes induce large ac displacement currents in both physical and parasitic circuit elements. The displacement currents can cause noise in two forms ... conducted noise (in the connecting busses, wires and ground planes) which is conducted to other circuitry, and radiated noise (capacitively coupled current to the case or heat sink) which is transformed to voltage which is then radiated to other circuitry. These noises can never be fully eliminated, however, they can be greatly attenuated (by proper electrical and packaging design) to levels conforming to accepted regulatory agency limits.

Conducted noise has two major contributors. The first results from reflected ripple current at the input to the converter. The second is due to the output voltage switching noise. The effects of the converter's reflected input ripple current on the rest of the power system current can be greatly reduced by incorporating a single-ended "pi" filter at the input of the converter for differential-mode conducted noise, or a double-ended "pi" filter for common-mode noise reduction. These filter types are demonstrated in Figure 18. Differential-mode noise is that which is present between a signal and ground (e.g. between the input voltage source and the input ground) and common-mode noise is that which exists between the grounds at the various physical locations in the circuit (due to the physical impedance of the conductors and ground planes).

The effects of output switching noise on the load circuitry can be reduced by making the output capacitor as "electrically" large as possible and by selecting a capacitor with a low equivalent series resistance (ESR), so as to reduce the output ripple (see Figure 19) to some acceptable level. In some instances, it may be necessary to synchronize the operation of the load circuitry to the power converter's switching frequency such that the load circuit is momentarily inactive or quiescent during the rising and falling edges of a noisy power switching event.

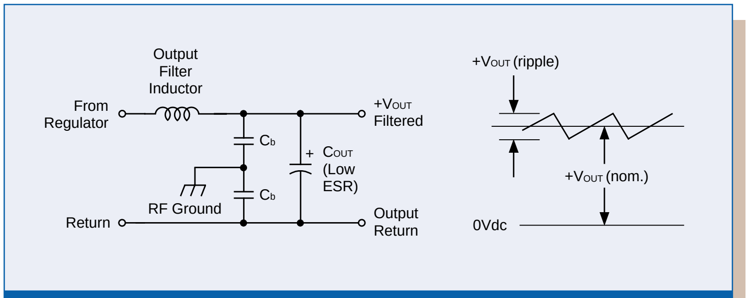


Figure 19. The effects of switching converter ripple and noise on load circuitry can be reduced by making output capacitors as electrically large as possible.

Occasionally, designers will incorporate an rf (radio-frequency) ac ground plane into the switching converter's construction. This allows the converter designer access to an electrical node (which is usually common with the case and the output return) to capacitively shunt noise signal paths to a common ground.

Radiated noise is sometimes controlled by shielding the power converter with a five or six-sided Faraday boundary. This Faraday boundary is usually the case, which can be made of aluminum, steel, copper, zinc or some other electrically conductive material. The case is then connected to the input or output return forcing it to

DATEL, Inc., 11 Cabot Boulevard, Mansfield, MA 02048-1151 (USA)
Tel: (508)339-3000, (800)233-2765 Fax: (508)339-6356 • Email: sales@datel.com • Internet: www.datel.com

"ground" potential so that it and/or the internal circuitry may not radiate. Unfortunately, any shielding strategy must be considered as part of a system-wide noise-reduction strategy, so each application must be dealt with on a case-by-case basis, even though provisions may be incorporated within the switching power converter itself.

Finally, recall that a switching power converter is a high-gain loop servomechanism, and it has a susceptibility to noise from external sources. This noise can degrade or impair the performance of the power converter, and even the incorporation of input filters and shielding is no guarantee that external noise susceptibility will be eliminated. Again, each application must be dealt with on a case-by-case basis.