

## **SECTION 3**

# **SWITCHING REGULATORS**

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### **INTRODUCTION**

Virtually all of today's electronic systems require some form of power conversion. The trend toward lower power, portable equipment has driven the technology and the requirement for converting power efficiently. Switchmode power converters, often referred to simply as "switchers", offer a versatile way of achieving this goal. Modern IC switching regulators are small, flexible, and allow either step-up (boost) or step-down (buck) operation.

When switcher functions are integrated and include a switch which is part of the basic power converter topology, these ICs are called "switching regulators". When no switches are included in the IC, but the signal for driving an external switch is provided, it is called a "switching regulator controller". Sometimes - usually for higher power levels - the control is not entirely integrated, but other functions to enhance the flexibility of the IC are included instead. In this case the device might be called a "controller" of sorts - perhaps a "feedback controller" if it just generates the feedback signal to the switch modulator. It is important to know what you are getting in your controller, and to know if your switching regulator is really a regulator or is it just the controller function.

Also, like switchmode power conversion, linear power conversion and charge pump technology offer both regulators and controllers. So within the field of power conversion, the terms "regulator" and "controller" can have wide meaning.

The most basic switcher topologies require only one transistor which is essentially used as a switch, one diode, one inductor, a capacitor across the output, and for practical but not fundamental reasons, another one across the input. A practical converter, however, requires several additional elements, such as a voltage reference, error amplifier, comparator, oscillator, and switch driver, and may also include optional features like current limiting and shutdown capability. Depending on the power level, modern IC switching regulators may integrate the entire converter except for the main magnetic element(s) (usually a single inductor) and the input/output capacitors. Often, a diode, the one which is an essential element of basic switcher topologies, cannot be integrated either. In any case, the complete power conversion for a switcher cannot be as integrated as a linear regulator, for example. The requirement of a magnetic element means that system designers are not inclined to think of switching regulators as simply "drop in" solutions. This presents the challenge to switching regulator manufacturers to provide careful design guidelines, commonly-used application circuits, and plenty of design assistance and product support. As the power levels increase, ICs tend to grow in complexity because it becomes more critical to optimize the control flexibility and precision. Also, since the switches begin to dominate the size of the die, it becomes more cost effective to remove them and integrate only the controller.

## **SWITCHING REGULATORS**

The primary limitations of switching regulators as compared to linear regulators are their output noise, EMI/RFI emissions, and the proper selection of external support components. Although switching regulators do not necessarily require transformers, they do use inductors, and magnetic theory is not generally well understood. However, manufacturers of switching regulators generally offer applications support in this area by offering complete data sheets with recommended parts lists for the external inductor as well as capacitors and switching elements.

One unique advantage of switching regulators lies in their ability to convert a given supply voltage with a known voltage range to virtually any given desired output voltage, with no “first order” limitations on efficiency. This is true regardless of whether the output voltage is higher or lower than the input voltage - the same or the opposite polarity. Consider the basic components of a switcher, as stated above. The inductor and capacitor are, ideally, reactive elements which dissipate no power. The transistor is effectively, ideally, a switch in that it is either “on”, thus having no voltage dropped across it while current flows through it, or “off”, thus having no current flowing through it while there is voltage across it. Since either voltage or current are always zero, the power dissipation is zero, thus, ideally, the switch dissipates no power. Finally, there is the diode, which has a finite voltage drop while current flows through it, and thus dissipates *some* power. But even that can be substituted with a synchronized switch, called a “synchronous rectifier”, so that it ideally dissipates no power either.

Switchers also offer the advantage that, since they inherently require a magnetic element, it is often a simple matter to “tap” an extra winding onto that element and, often with just a diode and capacitor, generate a reasonably well regulated additional output. If more outputs are needed, more such taps can be used. Since the tap winding requires no electrical connection, it can be isolated from other circuitry, or made to “float” atop other voltages.

Of course, nothing is ideal, and everything has a price. Inductors have resistance, and their magnetic cores are not ideal either, so they dissipate power. Capacitors have resistance, and as current flows in and out of them, they dissipate power, too. Transistors, bipolar or field-effect, are not ideal switches, and have a voltage drop when they are turned on, plus they cannot be switched instantly, and thus dissipate power while they are turning on or off.

As we shall soon see, switchers create ripple currents in their input and output capacitors. Those ripple currents create voltage ripple and noise on the converter’s input and output due to the resistance, inductance, and finite capacitance of the capacitors used. That is the *conducted* part of the noise. Then there are often ringing voltages in the converter, parasitic inductances in components and PCB traces, and an inductor which creates a magnetic field which it cannot perfectly contain within its core - all contributors to *radiated* noise. Noise is an inherent by-product of a switcher and must be controlled by proper component selection, PCB layout, and, if that is not sufficient, additional input or output filtering or shielding.

## **INTEGRATED CIRCUIT SWITCHING REGULATORS**

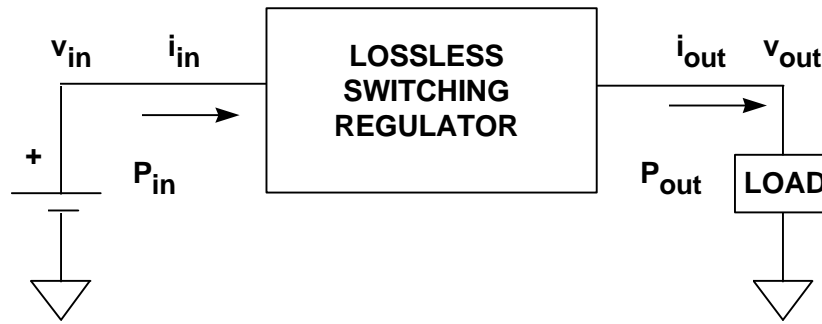
- **Advantages:**
  - ◆ **High Efficiency**
  - ◆ **Small**
  - ◆ **Flexible - Step-Up (Boost), Step-Down (Buck), etc.**
  
- **Disadvantages**
  - ◆ **Noisy (EMI, RFI, Peak-to-Peak Ripple)**
  - ◆ **Require External Components (L's, C's)**
  - ◆ **Designs Can Be Tricky**
  - ◆ **Higher Total Cost Than Linear Regulators**
  
- **"Regulators" vs. "Controllers"**

**Figure 3.1**

Though switchers can be designed to accommodate a range of input/output conditions, it is generally more costly in non-isolated systems to accommodate a requirement for both voltage step-up and step-down. So generally it is preferable to limit the input/output ranges such that one or the other case can exist, but not both, and then a simpler converter design can be chosen.

The concerns of minimizing power dissipation and noise as well as the design complexity and power converter versatility set forth the limitations and challenges for designing switchers, whether with regulators or controllers.

The ideal switching regulator shown in Figure 3.2 performs a voltage conversion and input/output energy transfer without loss of power by the use of purely reactive components. Although an actual switching regulator does have internal losses, efficiencies can be quite high, generally greater than 80 to 90%. Conservation of energy applies, so the input power equals the output power. This says that in step-down (buck) designs, the input current is lower than the output current. On the other hand, in step-up (boost) designs, the input current is greater than the output current. Input currents can therefore be quite high in boost applications, and this should be kept in mind, especially when generating high output voltages from batteries.

**THE IDEAL SWITCHING REGULATOR**

- $P_{in} = P_{out}$
- Efficiency =  $P_{out} / P_{in} = 100\%$
- $v_{in} \cdot i_{in} = v_{out} \cdot i_{out}$
- $\frac{v_{out}}{v_{in}} = \frac{i_{in}}{i_{out}}$
- Energy Must be Conserved!

Figure 3.2

Design engineers unfamiliar with IC switching regulators are sometimes confused by what exactly these devices can do for them. Figure 3.3 summarizes what to expect from a typical IC switching regulator. It should be emphasized that these are typical specifications, and can vary widely, but serve to illustrate some general characteristics.

Input voltages may range from 0.8 to beyond 30V, depending on the breakdown voltage of the IC process. Most regulators are available in several output voltage options, 12V, 5V, 3.3V, and 3V are the most common, and some regulators allow the output voltage to be set using external resistors. Output current varies widely, but regulators with internal switches have inherent current handling limitations that controllers (with external switches) do not. Output line and load regulation is typically about 50mV. The output ripple voltage is highly dependent upon the external output capacitor, but with care, can be limited to between 20mV and 100mV peak-to-peak. This ripple is at the switching frequency, which can range from 20kHz to 1MHz. There are also high frequency components in the output current of a switching regulator, but these can be minimized with proper external filtering, layout, and grounding. Efficiency can also vary widely, with up to 95% sometimes being achievable.

## WHAT TO EXPECT FROM A SWITCHING REGULATOR IC

- **Input Voltage Range: 0.8V to 30V**
- **Output Voltage:**
  - ◆ “Standard”: 12V, 5V, 3.3V, 3V
  - ◆ “Specialized”: VID Programmable for Microprocessors
  - ◆ (Some are Adjustable)
- **Output Current**
  - ◆ Up to 1.5A, Using Internal Switches of a Regulator
  - ◆ No Inherent Limitations Using External Switches with a Controller
- **Output Line / Load Regulation: 50mV, typical**
- **Output Voltage Ripple (peak-peak) :**  
20mV - 100mV @ Switching Frequency
- **Switching Frequency: 20kHz - 1MHz**
- **Efficiency: Up to 95%**

Figure 3.3

## POPULAR APPLICATIONS OF SWITCHING REGULATORS

For equipment which is powered by an AC source, the conversion from AC to DC is generally accomplished with a switcher, except for low-power applications where size and efficiency concerns are outweighed by cost. Then the power conversion may be done with just an AC transformer, some diodes, a capacitor, and a linear regulator. The size issue quickly brings switchers back into the picture as the preferable conversion method as power levels rise up to 10 watts and beyond. Off-line power conversion is heavily dominated by switchers in most modern electronic equipment.

Many modern high-power off-line power supply systems use the distributed approach by employing a switcher to generate an intermediate DC voltage which is then distributed to any number of DC/DC converters which can be located near to their respective loads (see Figure 3.4). Although there is the obvious redundancy of converting the power twice, distribution offers some advantages. Since such systems require isolation from the line voltage, only the first converter requires the isolation; all cascaded converters need not be isolated, or at least not to the degree of isolation that the first converter requires. The intermediate DC voltage is usually regulated to less than 60 volts in order to minimize the isolation requirement for the cascaded converters. Its regulation is not critical since it is not a direct output. Since it is typically higher than any of the switching regulator output voltages, the distribution current is substantially less than the sum of the output currents, thereby reducing  $I^2R$  losses in the system power distribution wiring. This also allows the use of a smaller energy storage capacitor on the intermediate DC supply output. (Recall that the energy stored in a capacitor is  $\frac{1}{2}CV^2$ ).

## SWITCHING REGULATORS

Power management can be realized by selectively turning on or off the individual DC/DC converters as needed.

### POWER DISTRIBUTION USING LINEAR AND SWITCHING REGULATORS

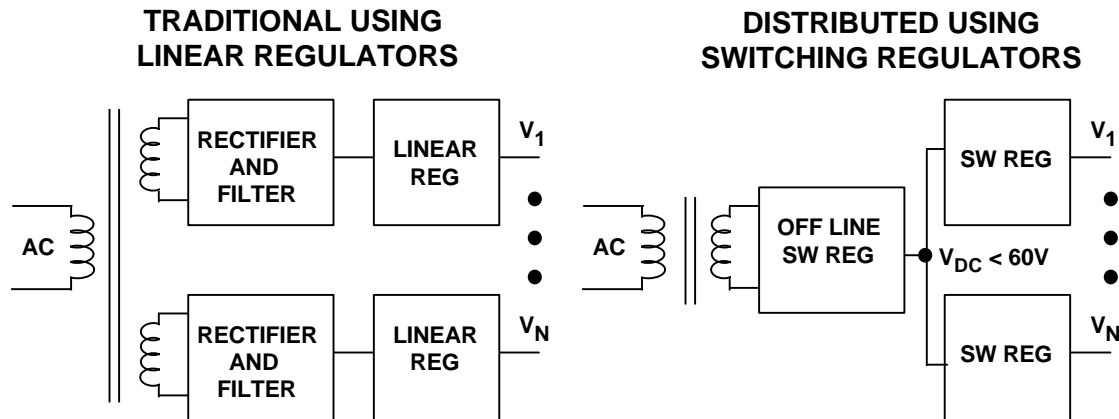


Figure 3.4

### ADVANTAGES OF DISTRIBUTED POWER SYSTEMS USING SWITCHING REGULATORS

- Higher Efficiency with Switching Regulators than Linear Regulators
- Use of High Intermediate DC Voltage Minimizes Power Loss due to Wiring Resistance
- Flexible (Multiple Output Voltages Easily Obtained)
- AC Power Transformer Design Easier (Only One Winding Required, Regulation Not Critical)
- Selective Shutdown Techniques Can Be Used for Higher Efficiency
- Eliminates Safety Isolation Requirements for DC/DC Converters

Figure 3.5

Batteries are the primary power source in much of today's consumer and communications equipment. Such systems may require one or several voltages, and they may be less or greater than the battery voltage. Since a battery is a self-contained power source, power converters seldom require isolation. Often, then, the basic switcher topologies are used, and a wide variety of switching regulators are

available to fill many of the applications. Maximum power levels for these regulators typically can range up from as low as tens of milliwatts to several watts.

Efficiency is often of great importance, as it is a factor in determining battery life which, in turn, affects practicality and cost of ownership. Often of even greater importance, though often confused with efficiency, is quiescent power dissipation when operating at a small fraction of the maximum rated load (e.g., standby mode). For electronic equipment which must remain under power in order to retain data storage or minimal monitoring functions, but is otherwise shut down most of the time, the quiescent dissipation is the largest determinant of battery life. Although efficiency may indicate power consumption for a specific light load condition, it is not the most useful way to address the concern. For example, if there is no load on the converter output, the efficiency will be zero no matter how optimal the converter, and one could not distinguish a well power-managed converter from a poorly managed one by such a specification.

The concern of managing power effectively from no load to full load has driven much of the technology which has been and still is emerging from today's switching regulators and controllers. Effective power management, as well as reliable power conversion, is often a substantial factor of quality or noteworthy distinction in a wide variety of equipment. The limitations and cost of batteries are such that consumers place a value on not having to replace them more often than necessary, and that is certainly a goal for effective power conversion solutions.

### TYPICAL APPLICATION OF A BOOST REGULATOR IN BATTERY OPERATED EQUIPMENT

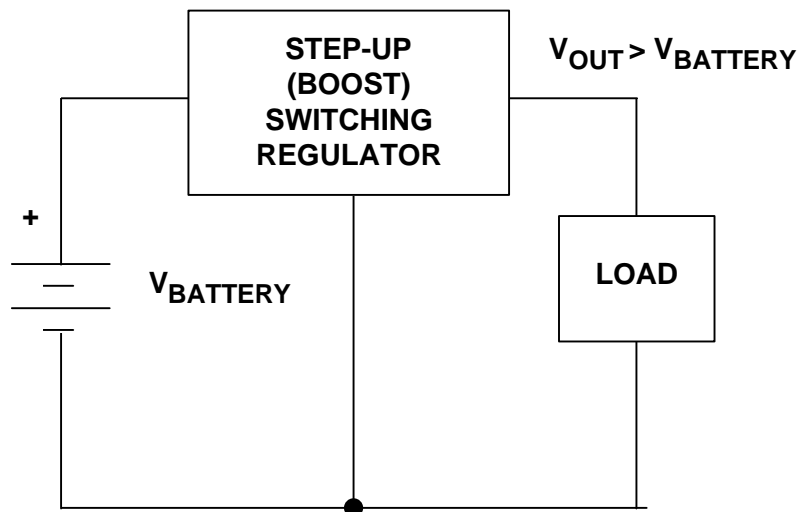
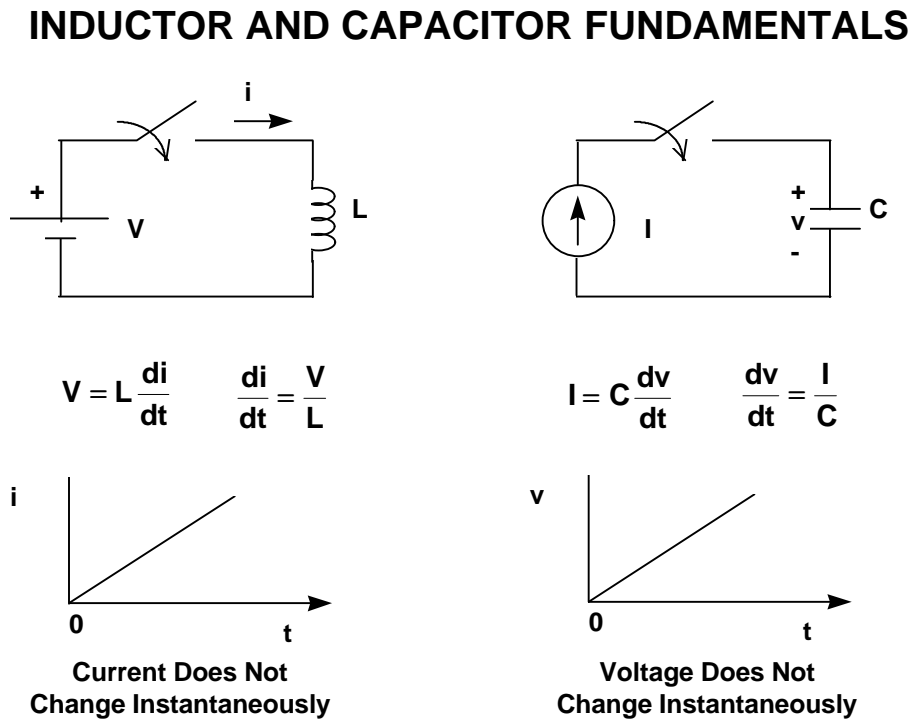


Figure 3.6

## INDUCTOR AND CAPACITOR FUNDAMENTALS

In order to understand switching regulators, the fundamental energy storage capabilities of inductors and capacitors must be fully understood. When a voltage is applied to an ideal inductor (see Figure 3.7), the current builds up linearly over time at a rate equal to  $V/L$ , where  $V$  is the applied voltage, and  $L$  is the value of the inductance. This energy is stored in the inductor's magnetic field, and if the switch is opened, the magnetic field collapses, and the inductor voltage goes to a large instantaneous value until the field has fully collapsed.



**Figure 3.7**

When a current is applied to an ideal capacitor, the capacitor is gradually charged, and the voltage builds up linearly over time at a rate equal to  $I/C$ , where  $I$  is the applied current, and  $C$  is the value of the capacitance. Note that the voltage across an ideal capacitor cannot change instantaneously.

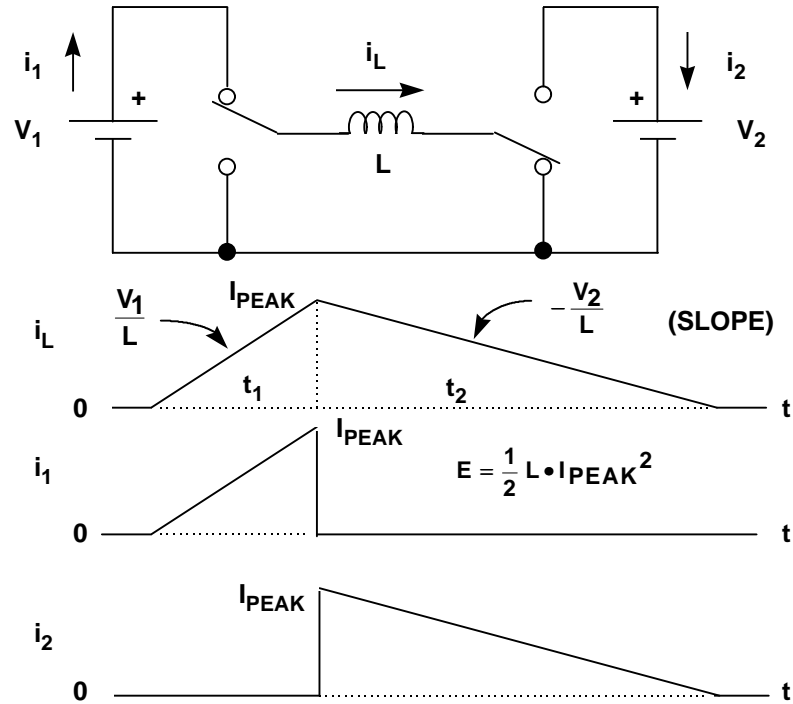
Of course, there is no such thing as an ideal inductor or capacitor. Real inductors have stray winding capacitance, series resistance, and can saturate for large currents. Real capacitors have series resistance and inductance and may break down under large voltages. Nevertheless, the fundamentals of the ideal inductor and capacitor are critical in understanding the operation of switching regulators.

An inductor can be used to transfer energy between two voltage sources as shown in Figure 3.8. While energy transfer could occur between two voltage sources with a resistor connected between them, the energy transfer would be inefficient due to the power loss in the resistor, and the energy could only be transferred from the higher to the lower value source. In contrast, an inductor ideally returns all the energy that



is stored in it, and with the use of properly configured switches, the energy can flow from any one source to another, regardless of their respective values and polarities.

**ENERGY TRANSFER USING AN INDUCTOR**



**Figure 3.8**

When the switches are initially placed in the position shown, the voltage  $V_1$  is applied to the inductor, and the inductor current builds up at a rate equal to  $V_1/L$ . The peak value of the inductor current at the end of the interval  $t_1$  is

$$I_{PEAK} = \frac{V_1}{L} \cdot t_1.$$

The average power transferred to the inductor during the interval  $t_1$  is

$$P_{AVG} = \frac{1}{2} I_{PEAK} \cdot V_1.$$

The energy transferred during the interval  $t_1$  is

$$E = P_{AVG} \cdot t_1 = \frac{1}{2} I_{PEAK} \cdot V_1 \cdot t_1.$$

Solving the first equation for  $t_1$  and substituting into the last equation yields

$$E = \frac{1}{2} L \cdot I_{PEAK}^2.$$

## SWITCHING REGULATORS

When the switch positions are reversed, the inductor current continues to flow into the load voltage  $V_2$ , and the inductor current decreases at a rate  $-V_2/L$ . At the end of the interval  $t_2$ , the inductor current has decreased to zero, and the energy has been transferred into the load. The figure shows the current waveforms for the inductor, the input current  $i_1$ , and the output current  $i_2$ . The ideal inductor dissipates no power, so there is no power loss in this transfer, assuming ideal circuit elements. This fundamental method of energy transfer forms the basis for all switching regulators.

### IDEAL STEP-DOWN (BUCK) CONVERTER

The fundamental circuit for an ideal step-down (buck) converter is shown in Figure 3.9. The actual integrated circuit switching regulator contains the switch control circuit and may or may not include the switch (depending upon the output current requirement). The inductor, diode, and load bypass capacitor are external.

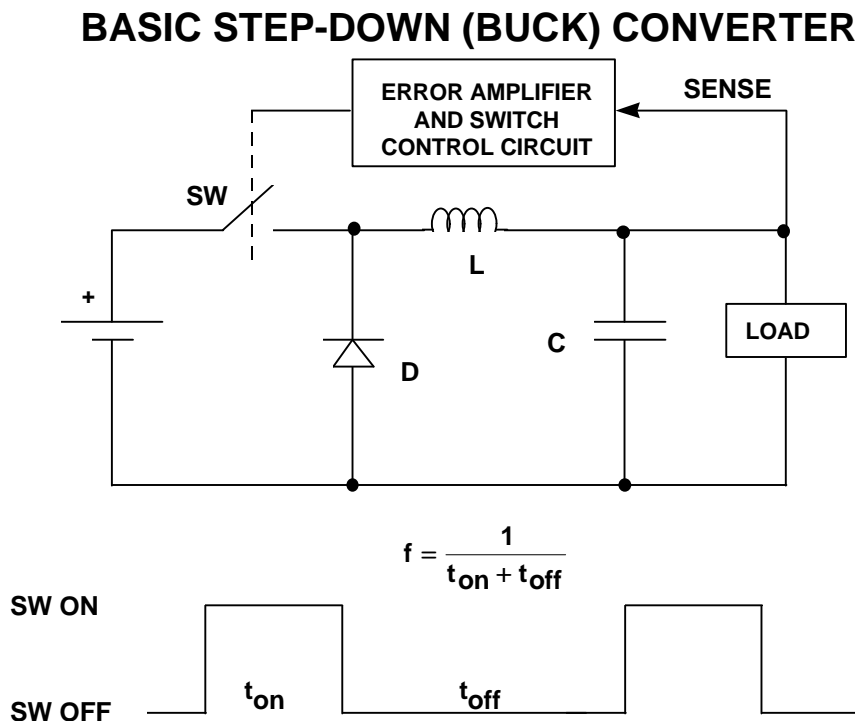


Figure 3.9

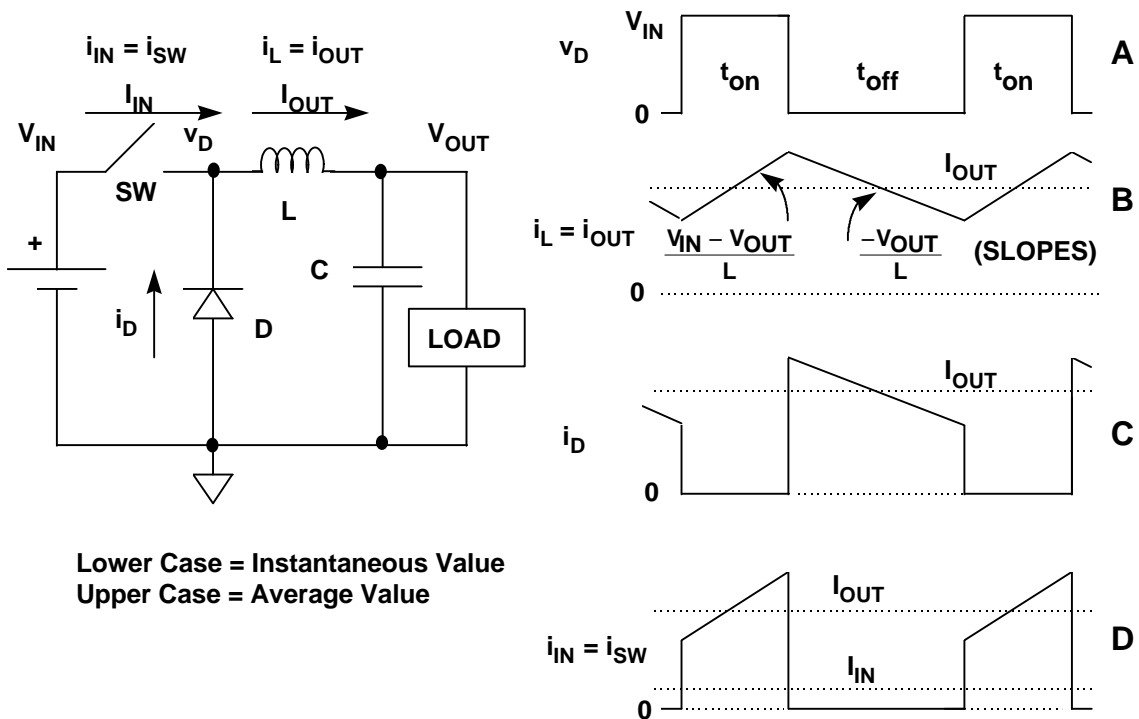
The output voltage is sensed and then regulated by the switch control circuit. There are several methods for controlling the switch, but for now assume that the switch is controlled by a pulse width modulator (PWM) operating at a fixed frequency,  $f$ .

The actual waveforms associated with the buck converter are shown in Figure 3.10. When the switch is on, the voltage  $V_{\text{IN}} - V_{\text{OUT}}$  appears across the inductor, and the inductor current increases with a slope equal to  $(V_{\text{IN}} - V_{\text{OUT}})/L$  (see Figure 3.10B). When the switch turns off, current continues to flow through the inductor and into the load (remember that the current cannot change instantaneously in an inductor), with the ideal diode providing the return current path. The voltage across the inductor is now  $V_{\text{OUT}}$ , but the polarity has reversed. Therefore, the inductor

current decreases with a slope equal to  $-V_{OUT}/L$ . Note that the inductor current is equal to the output current in a buck converter.

The diode and switch currents are shown in Figures 3.10C and 3.10D, respectively, and the inductor current is the sum of these waveforms. Also note by inspection that the instantaneous input current equals the switch current. Note, however, that the average input current is less than the average output current. In a practical regulator, both the switch and the diode have voltage drops across them during their conduction which creates internal power dissipation and a loss of efficiency, but these voltages will be neglected for now. It is also assumed that the output capacitor,  $C$ , is large enough so that the output voltage does not change significantly during the switch on or off times.

**BASIC STEP-DOWN (BUCK) CONVERTER WAVEFORMS**



**Figure 3.10**

There are several important things to note about these waveforms. The most important is that ideal components have been assumed, i.e., the input voltage source has zero impedance, the switch has zero on-resistance and zero turn-on and turn-off times. It is also assumed that the inductor does not saturate and that the diode is ideal with no forward drop.

Also note that the output current is continuous, while the input current is pulsating. Obviously, this has implications regarding input and output filtering. If one is concerned about the voltage ripple created on the power source which supplies a buck converter, the input filter capacitor (not shown) is generally more critical than the output capacitor with respect to ESR/ESL.

## SWITCHING REGULATORS

If a steady-state condition exists (see Figure 3.11), the basic relationship between the input and output voltage may be derived by inspecting the inductor current waveform and writing:

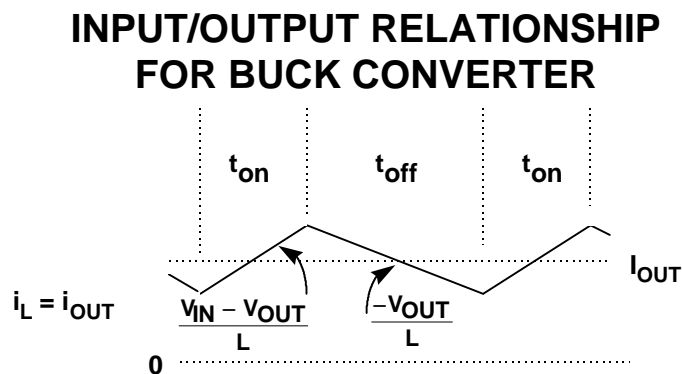
$$\frac{V_{IN} - V_{OUT}}{L} \cdot t_{on} = \frac{V_{OUT}}{L} \cdot t_{off}.$$

Solving for  $V_{OUT}$ :

$$V_{OUT} = V_{IN} \cdot \frac{t_{on}}{t_{on} + t_{off}} = V_{IN} \cdot D,$$

where  $D$  is the switch *duty ratio* (more commonly called *duty cycle*), defined as the ratio of the switch on-time ( $t_{on}$ ) to the total switch cycle time ( $t_{on} + t_{off}$ ).

This is the classic equation relating input and output voltage in a buck converter which is operating with *continuous* inductor current, defined by the fact that the inductor current never goes to zero.



■ Write by Inspection from Inductor/Output Current Waveforms:

■ 
$$\frac{V_{IN} - V_{OUT}}{L} \cdot t_{on} = \frac{V_{OUT}}{L} \cdot t_{off}$$

■ Rearrange and Solve for  $V_{OUT}$ :

■ 
$$V_{OUT} = V_{IN} \cdot \frac{t_{on}}{t_{on} + t_{off}} = V_{IN} \cdot D$$

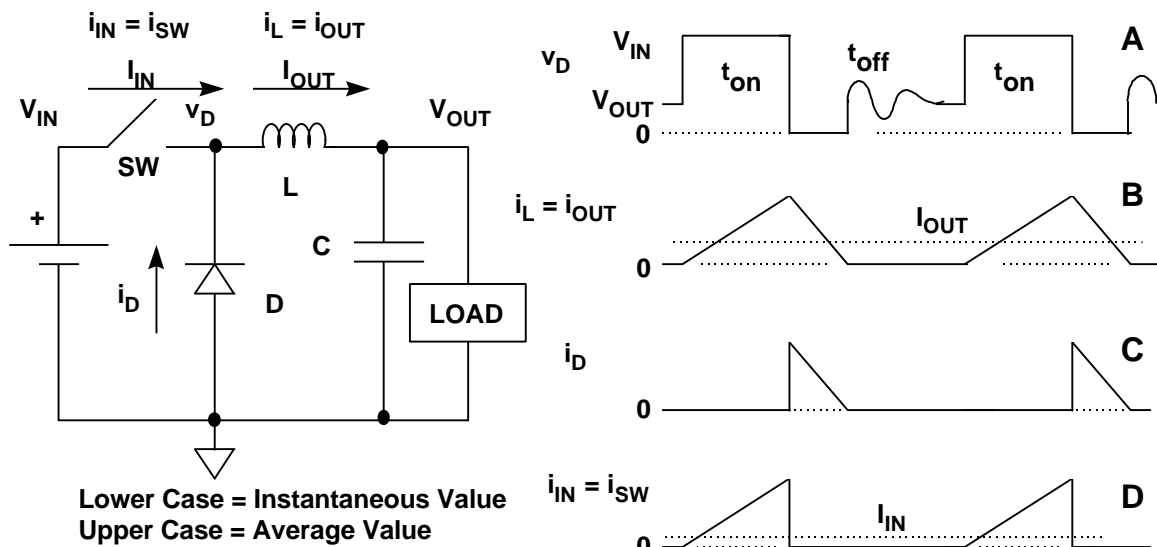
Figure 3.11

Notice that this relationship is independent of the inductor value  $L$  as well as the switching frequency  $1/(t_{on} + t_{off})$  and the load current. Decreasing the inductor value, however, will result in a larger peak-to-peak output ripple current, while increasing the value results in smaller ripple. There are many other tradeoffs involved in selecting the inductor, and these will be discussed in a later section.

In this simple model, line and load regulation (of the output voltage) is achieved by varying the duty cycle using a pulse width modulator (PWM) operating at a fixed frequency,  $f$ . The PWM is in turn controlled by an error amplifier - an amplifier which amplifies the "error" between the measured output voltage and a reference voltage. As the input voltage increases, the duty cycle decreases; and as the input voltage decreases, the duty cycle increases. Note that while the average inductor current changes proportionally to the output current, the duty cycle does not change. Only dynamic changes in the duty cycle are required to modulate the inductor current to the desired level; then the duty cycle returns to its steady state value. In a practical converter, the duty cycle might increase slightly with load current to counter the increase in voltage drops in the circuit, but would otherwise follow the ideal model.

This discussion so far has assumed the regulator is in the *continuous-mode* of operation, defined by the fact that the inductor current never goes to zero. If, however, the output load current is decreased, there comes a point where the inductor current will go to zero between cycles, and the inductor current is said to be *discontinuous*. It is necessary to understand this operating mode as well, since many switchers must supply a wide dynamic range of output current, where this phenomenon is unavoidable. Waveforms for discontinuous operation are shown in Figure 3.12.

**BUCK CONVERTER WAVEFORMS  
DISCONTINUOUS MODE**



**Figure 3.12**

Behavior during the switch on-time is identical to that of the continuous mode of operation. However, during the switch off-time, there are two regions of unique behavior. First, the inductor current ramps down at the same rate as it does during continuous mode, but then the inductor current goes to zero. When it reaches zero, the current tries to reverse but cannot find a path through the diode any longer. So the voltage on the input side of the inductor (same as the diode and switch junction)

## SWITCHING REGULATORS

jumps up to  $V_{OUT}$  such that the inductor has no voltage across it, and the current can remain at zero.

Because the impedance at diode node ( $v_D$ ) is high, ringing occurs due to the inductor,  $L$ , resonating with the stray capacitance which is the sum of the diode capacitance,  $C_D$ , and the switch capacitance,  $C_{SW}$ . The oscillation is damped by stray resistances in the circuit, and occurs at a frequency given by

$$f_0 = \frac{1}{2\pi \sqrt{L(C_D + C_{SW})}}.$$

A circuit devoted simply to dampening resonances via power dissipation is called a *snubber*. If the ringing generates EMI/RFI problems, it may be damped with a suitable RC snubber. However, this will cause additional power dissipation and reduced efficiency.

If the load current of a standard buck converter is low enough, the inductor current becomes discontinuous. The current at which this occurs can be calculated by observing the waveform shown in Figure 3.13. This waveform is drawn showing the inductor current going to exactly zero at the end of the switch off-time. Under these conditions, the average output current is

$$I_{OUT} = I_{PEAK}/2.$$

We have already shown that the peak inductor current is

$$I_{PEAK} = \frac{V_{IN} - V_{OUT}}{L} \cdot t_{on}.$$

Thus, discontinuous operation will occur if

$$I_{OUT} < \frac{V_{IN} - V_{OUT}}{2L} \cdot t_{on}.$$

However,  $V_{OUT}$  and  $V_{IN}$  are related by:

$$V_{OUT} = V_{IN} \cdot D = V_{IN} \cdot \frac{t_{on}}{t_{on} + t_{off}}.$$

Solving for  $t_{on}$ :

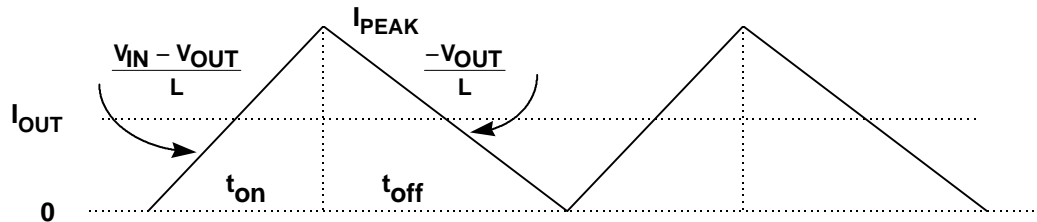
$$t_{on} = \frac{V_{OUT}}{V_{IN}} \cdot (t_{on} + t_{off}) = \frac{V_{OUT}}{V_{IN}} \cdot \frac{1}{f}.$$

Substituting this value for  $t_{on}$  into the previous equation for  $I_{OUT}$ :

$$I_{OUT} < \frac{V_{OUT} \left( 1 - \frac{V_{OUT}}{V_{IN}} \right)}{2Lf} \quad \text{(Criteria for discontinuous operation - buck converter)}$$

**BUCK CONVERTER POINT OF DISCONTINUOUS OPERATION**

**INDUCTOR CURRENT AND OUTPUT CURRENT**



**DISCONTINUOUS MODE IF:**

$$I_{OUT} < \frac{1}{2} I_{PEAK} = \frac{V_{IN} - V_{OUT}}{2L} \cdot t_{on}$$

$$I_{OUT} < \frac{V_{OUT} \left( 1 - \frac{V_{OUT}}{V_{IN}} \right)}{2Lf}, \quad f = \frac{1}{t_{on} + t_{off}}$$

**Figure 3.13**

**IDEAL STEP-UP (BOOST) CONVERTER**

The basic step-up (boost) converter circuit is shown in Figure 3.14. During the switch on-time, the current builds up in the inductor. When the switch is opened, the energy stored in the inductor is transferred to the load through the diode.

The actual waveforms associated with the boost converter are shown in Figure 3.15. When the switch is on, the voltage  $V_{IN}$  appears across the inductor, and the inductor current increases at a rate equal to  $V_{IN}/L$ . When the switch is opened, a voltage equal to  $V_{OUT} - V_{IN}$  appears across the inductor, current is supplied to the load, and the current decays at a rate equal to  $(V_{OUT} - V_{IN})/L$ . The inductor current waveform is shown in Figure 3.15B.

**BASIC STEP-UP (BOOST) CONVERTER**

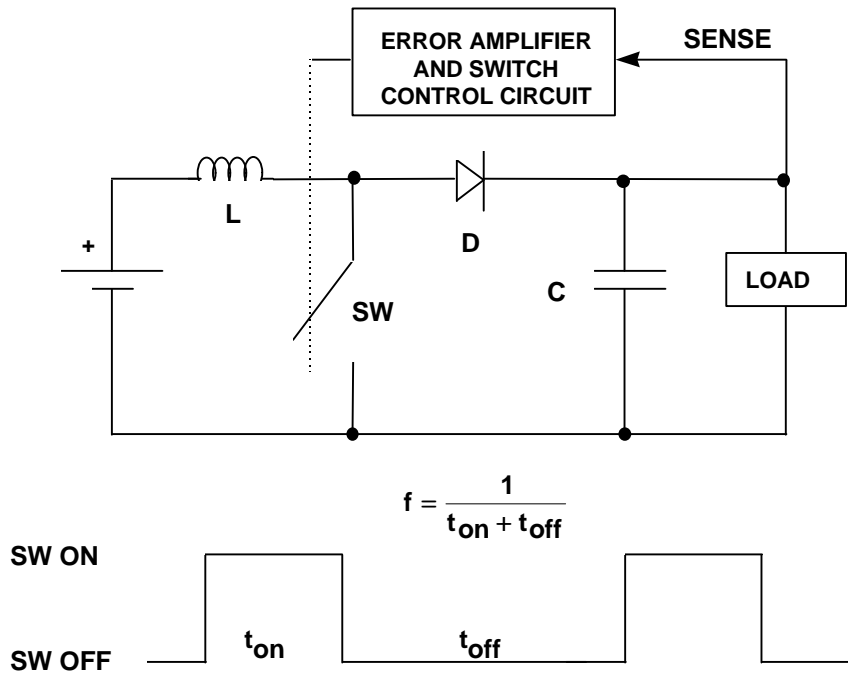


Figure 3.14

**BASIC STEP-UP (BOOST) CONVERTER WAVEFORMS**

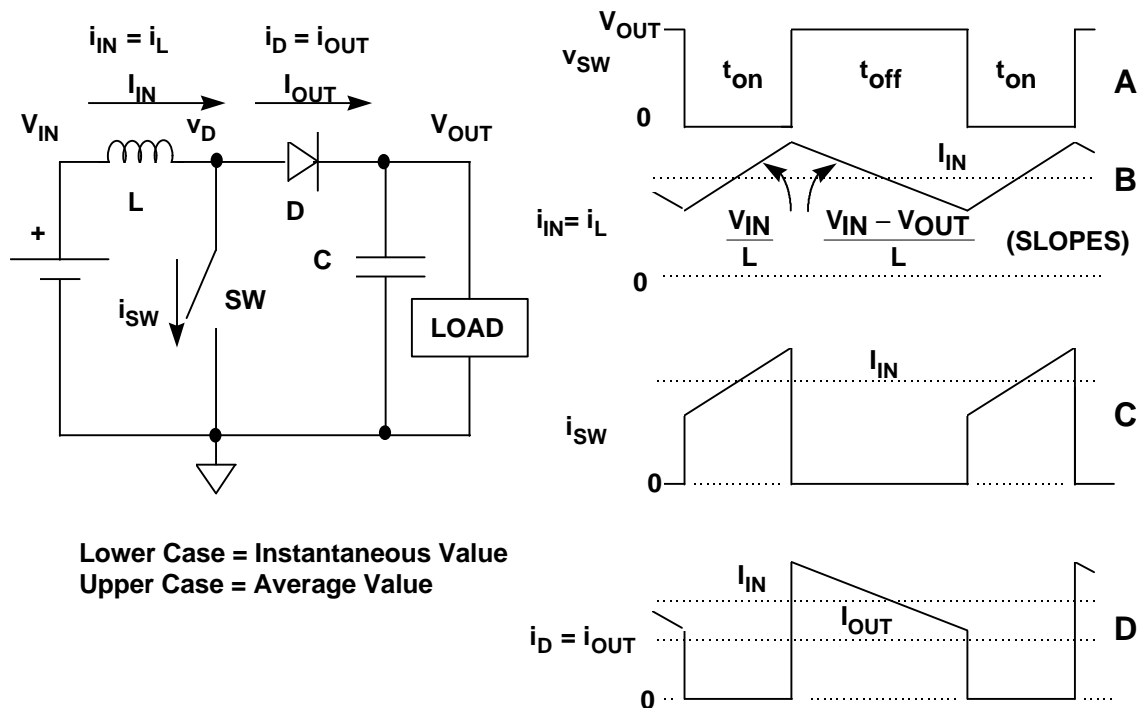


Figure 3.15



Note that in the boost converter, the input current is continuous, while the output current (Figure 3.15D) is pulsating. This implies that filtering the output of a boost converter is more difficult than that of a buck converter. (Refer back to the previous discussion of buck converters). Also note that the input current is the sum of the switch and diode current.

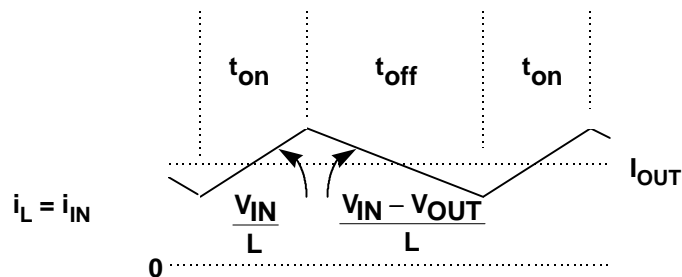
If a steady-state condition exists (see Figure 3.16), the basic relationship between the input and output voltage may be derived by inspecting the inductor current waveform and writing:

$$\frac{V_{IN}}{L} \cdot t_{on} = \frac{V_{OUT} - V_{IN}}{L} \cdot t_{off}$$

Solving for  $V_{OUT}$ :

$$V_{OUT} = V_{IN} \cdot \frac{t_{on} + t_{off}}{t_{off}} = V_{IN} \cdot \frac{1}{1 - D}$$

### INPUT/OUTPUT RELATIONSHIP FOR BOOST CONVERTER



■ Write by Inspection from Inductor/Input Current Waveforms:

■  $\frac{V_{IN}}{L} \cdot t_{on} = \frac{V_{OUT} - V_{IN}}{L} \cdot t_{off}$

■ Rearrange and Solve for  $V_{OUT}$ :

■  $V_{OUT} = V_{IN} \cdot \frac{t_{on} + t_{off}}{t_{off}} = V_{IN} \cdot \frac{1}{1 - D}$

Figure 3.16

## SWITCHING REGULATORS

This discussion so far has assumed the boost converter is in the *continuous-mode* of operation, defined by the fact that the inductor current never goes to zero. If, however, the output load current is decreased, there comes a point where the inductor current will go to zero between cycles, and the inductor current is said to be *discontinuous*. It is necessary to understand this operating mode as well, since many switchers must supply a wide dynamic range of output current, where this phenomenon is unavoidable.

Discontinuous operation for the boost converter is similar to that of the buck converter. Figure 3.17 shows the waveforms. Note that when the inductor current goes to zero, ringing occurs at the switch node at a frequency  $f_0$  given by:

$$f_0 = \frac{1}{2\pi \sqrt{L(C_D + C_{SW})}}$$

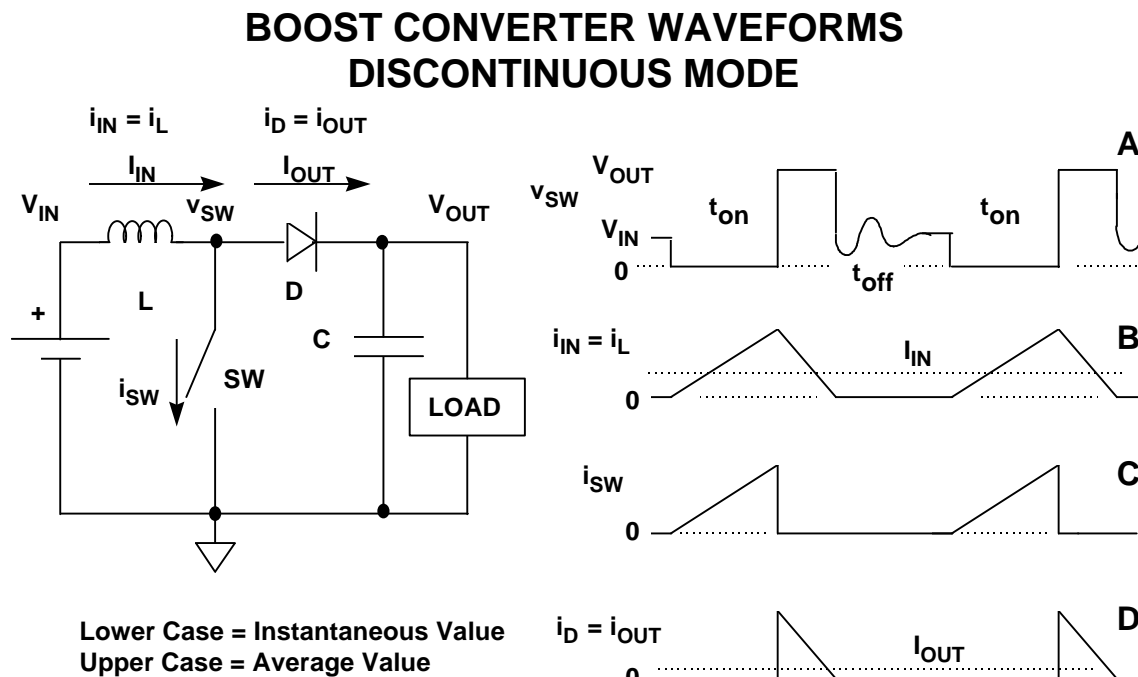
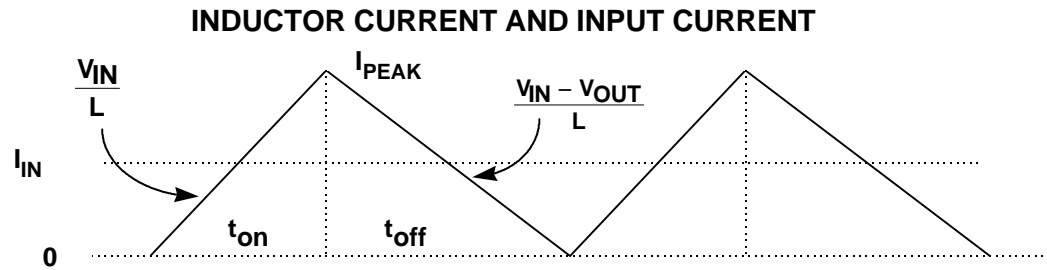


Figure 3.17

The inductor,  $L$ , resonates with the stray switch capacitance and diode capacitance,  $C_{SW} + C_D$  as in the case of the buck converter. The ringing is dampened by circuit resistances, and, if needed, a snubber.

The current at which a boost converter becomes discontinuous can be derived by observing the inductor current (same as input current) waveform of Figure 3.18.

**BOOST CONVERTER POINT OF DISCONTINUOUS OPERATION**



**DISCONTINUOUS MODE IF:**

$$I_{IN} < \frac{1}{2} I_{PEAK} = \frac{V_{OUT} - V_{IN}}{2L} \cdot t_{off}$$

$$I_{OUT} < \frac{V_{IN}^2 (V_{OUT} - V_{IN})}{V_{OUT}^2 \cdot 2Lf}, \quad f = \frac{1}{t_{on} + t_{off}}$$

**Figure 3.18**

The average input current at the point of discontinuous operation is

$$I_{IN} = I_{PEAK}/2.$$

Discontinuous operation will occur if

$$I_{IN} < I_{PEAK}/2.$$

However,

$$I_{IN} = \frac{I_{PEAK}}{2} = \frac{V_{OUT} - V_{IN}}{2L} \cdot t_{off}.$$

Also,

$$V_{IN} \cdot I_{IN} = V_{OUT} \cdot I_{OUT}, \text{ and therefore}$$

$$I_{OUT} = \frac{V_{IN}}{V_{OUT}} \cdot I_{IN} = \frac{V_{IN}}{V_{OUT}} \cdot \frac{(V_{OUT} - V_{IN})}{2L} \cdot t_{off}.$$

However,

$$\frac{V_{OUT}}{V_{IN}} = \frac{1}{1-D} = \frac{1}{1 - \frac{t_{on}}{t_{on} + t_{off}}} = \frac{t_{on} + t_{off}}{t_{off}}.$$

## SWITCHING REGULATORS

Solving for  $t_{\text{off}}$ :

$$t_{\text{off}} = \frac{V_{\text{IN}}}{V_{\text{OUT}}}(t_{\text{on}} + t_{\text{off}}) = \frac{V_{\text{IN}}}{f \cdot V_{\text{OUT}}}.$$

Substituting this value for  $t_{\text{off}}$  into the previous expression for  $I_{\text{OUT}}$ , the criteria for discontinuous operation of a boost converter is established:

$$I_{\text{OUT}} < \frac{V_{\text{IN}}^2 (V_{\text{OUT}} - V_{\text{IN}})}{V_{\text{OUT}}^2 \cdot 2Lf} \quad (\text{Criteria for discontinuous operation - boost converter}).$$

The basic buck and boost converter circuits can work equally well for negative inputs and outputs as shown in Figure 3.19. Note that the only difference is that the polarities of the input voltage and the diode have been reversed. In practice, however, not many IC buck and boost regulators or controllers will work with negative inputs. In some cases, external circuitry can be added in order to handle negative inputs and outputs. Rarely are regulators or controllers designed specifically for negative inputs or outputs. In any case, data sheets for the specific ICs will indicate the degree of flexibility allowed.

### **NEGATIVE IN, NEGATIVE OUT BUCK AND BOOST CONVERTERS**

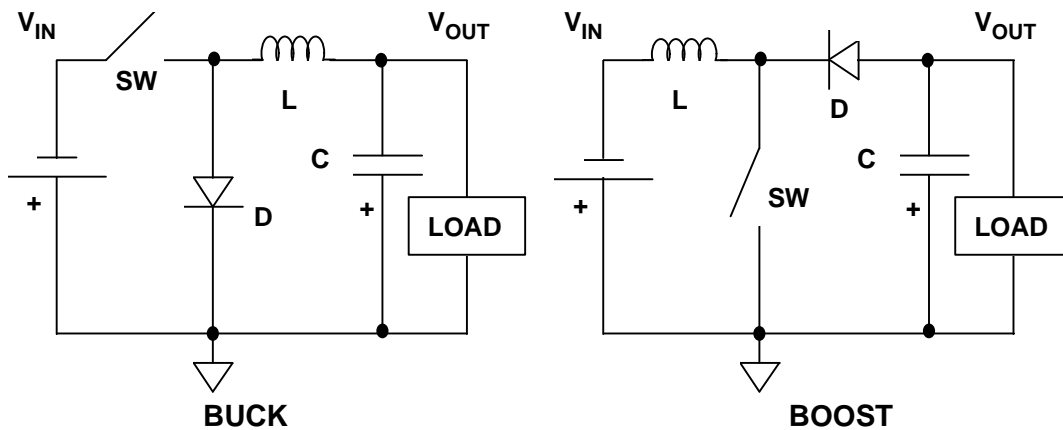


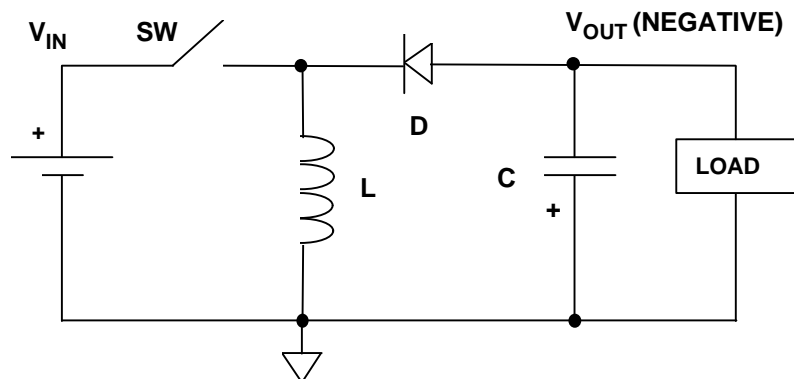
Figure 3.19

## BUCK-BOOST TOPOLOGIES

The simple buck converter can only produce an output voltage which is less than the input voltage, while the simple boost converter can only produce an output voltage greater than the input voltage. There are many applications where more flexibility is required. This is especially true in battery powered applications, where the fully charged battery voltage starts out greater than the desired output (the converter must operate in the buck mode), but as the battery discharges, its voltage becomes less than the desired output (the converter must then operate in the boost mode).

A *buck-boost* converter is capable of producing an output voltage which is either greater than or less than the absolute value of the input voltage. A simple buck-boost converter topology is shown in Figure 3.20. The input voltage is positive, and the output voltage is negative. When the switch is on, the inductor current builds up. When the switch is opened, the inductor supplies current to the load through the diode. Obviously, this circuit can be modified for a negative input and a positive output by reversing the polarity of the diode.

### BUCK-BOOST CONVERTER #1, $+V_{IN}, -V_{OUT}$

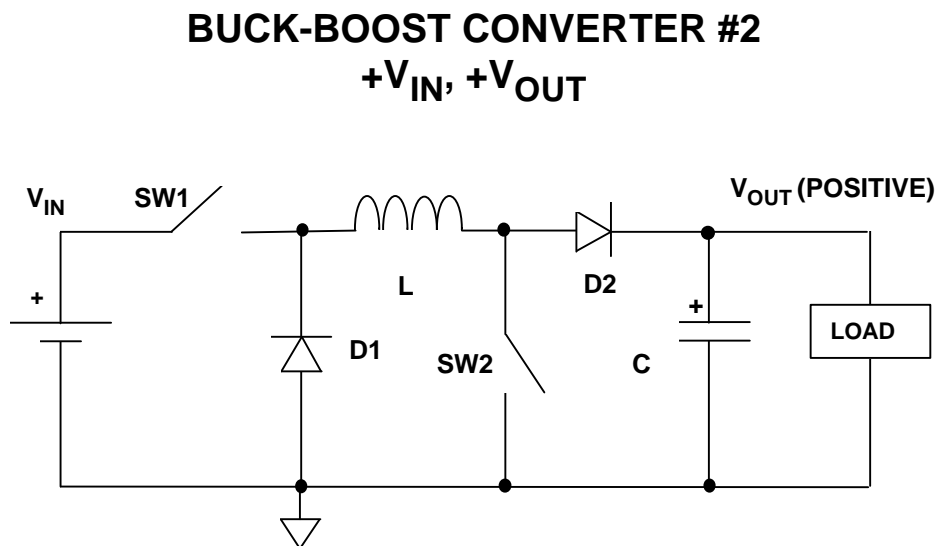


The Absolute Value of the Output Can Be Less Than  
Or Greater Than the Absolute Value of the Input

Figure 3.20

## SWITCHING REGULATORS

A second buck-boost converter topology is shown in Figure 3.21. This circuit allows both the input and output voltage to be positive. When the switches are closed, the inductor current builds up. When the switches open, the inductor current is supplied to the load through the current path provided by D1 and D2. A fundamental disadvantage to this circuit is that it requires two switches and two diodes. As in the previous circuits, the polarities of the diodes may be reversed to handle negative input and output voltages.



**The Absolute Value of the Output Can Be Less Than  
Or Greater Than the Absolute Value of the Input**

**Figure 3.21**

Another way to accomplish the buck-boost function is to cascade two switching regulators; a boost regulator followed by a buck regulator as shown in Figure 3.22. The example shows some practical voltages in a battery-operated system. The input from the four AA cells can range from 6V (charged) to about 3.5V (discharged). The intermediate voltage output of the boost converter is 8V, which is always greater than the input voltage. The buck regulator generates the desired 5V from the 8V intermediate voltage. The total efficiency of the combination is the product of the individual efficiencies of each regulator, and can be greater than 85% with careful design.

An alternate topology is use a buck regulator followed by a boost regulator. This approach, however, has the disadvantage of pulsating currents on both the input and output and a higher current at the intermediate voltage output.

## CASCADED BUCK-BOOST REGULATORS (EXAMPLE VOLTAGES)

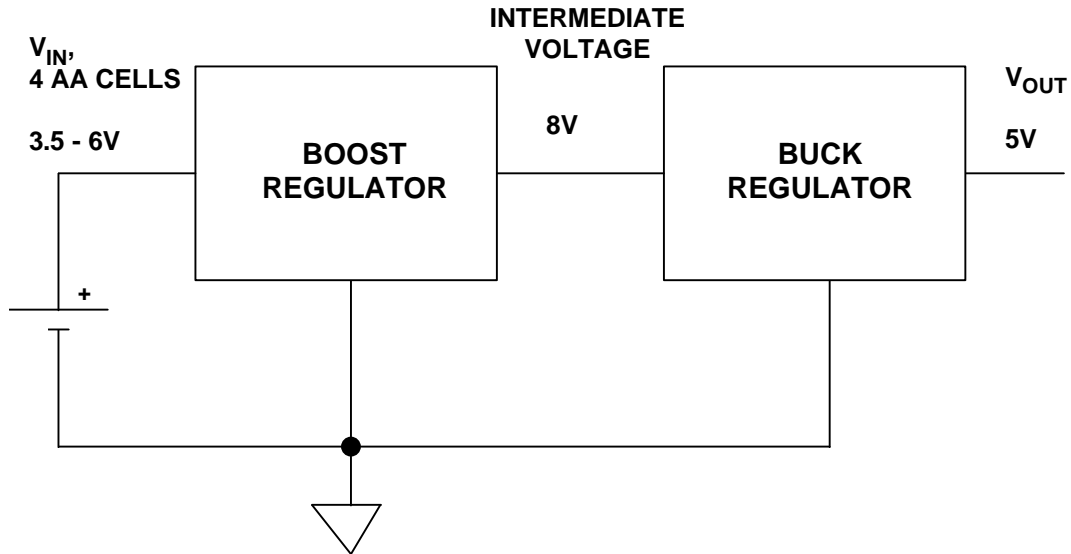


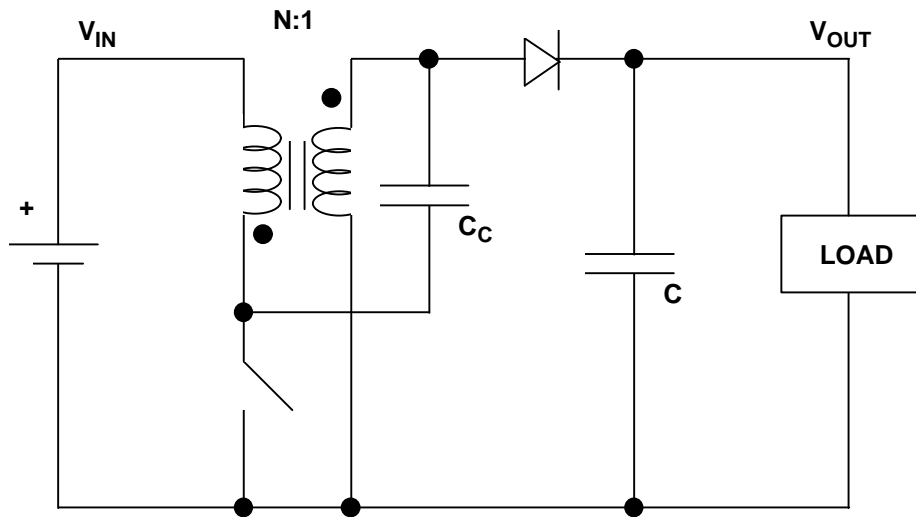
Figure 3.22

## OTHER NON-ISOLATED SWITCHER TOPOLOGIES

The coupled-inductor single-ended primary inductance converter (SEPIC) topology is shown in Figure 3.23. This converter uses a transformer with the addition of capacitor  $C_C$  which couples additional energy to the load. If the turns ratio ( $N =$  the ratio of the number of primary turns to the number of secondary turns) of the transformer in the SEPIC converter is 1:1, the capacitor serves only to recover the energy in the leakage inductance (i.e., that energy which is not perfectly coupled between the windings) and delivering it to the load. In that case, the relationship between input and output voltage is given by

$$V_{OUT} = V_{IN} \cdot \frac{D}{1-D}$$

For non-unity turns ratios the input/output relationship is highly nonlinear due to transfer of energy occurring via both the coupling between the windings and the capacitor  $C_C$ . For that reason, it is not analyzed here.

**SINGLE-ENDED PRIMARY INDUCTANCE CONVERTER (SEPIC)****Figure 3.23**

This converter topology often makes an excellent choice in non-isolated battery-powered systems for providing both the ability to step up or down the voltage, and, unlike the boost converter, the ability to have zero voltage at the output when desired.

The Zeta and Cük converters, not shown, are two examples of non-isolated converters which require capacitors to deliver energy from input to output, i.e., rather than just to store energy or deliver only recovered leakage energy, as the SEPIC can be configured via a 1:1 turns ratio. Because capacitors capable of delivering energy efficiently in such converters tend to be bulky and expensive, these converters are not frequently used.

**ISOLATED SWITCHING REGULATOR TOPOLOGIES**

The switching regulators discussed so far have direct galvanic connections between the input and output. Transformers can be used to supply galvanic isolation as well as allowing the buck-boost function to be easily performed. However, adding a transformer to the circuit creates a more complicated and expensive design as well as increasing the physical size.

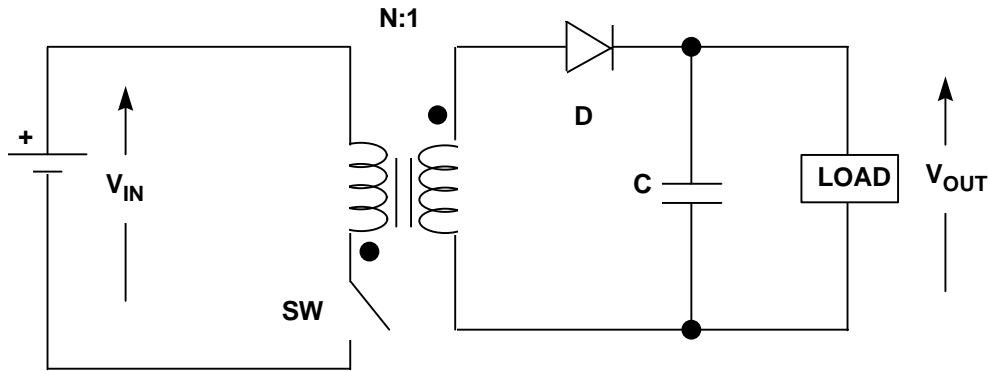
The basic *flyback* buck-boost converter circuit is shown in Figure 3.24. It is derived from the buck-boost converter topology. When the switch is on, the current builds up in the primary of the transformer. When the switch is opened, the current reverts to the secondary winding and flows through the diode and into the load. The relationship between the input and output voltage is determined by the turns ratio,  $N$ , and the duty cycle,  $D$ , per the following equation:

$$V_{OUT} = \frac{V_{IN}}{N} \cdot \frac{D}{1-D}$$



A disadvantage of the flyback converter is the high energy which must be stored in the transformer in the form of DC current in the windings. This requires larger cores than would be necessary with pure AC in the windings.

**ISOLATED TOPOLOGY:  
FLYBACK CONVERTER**



(BUCK-BOOST DERIVED)

$$V_{OUT} = \frac{V_{IN}}{N} \cdot \frac{D}{1-D}$$

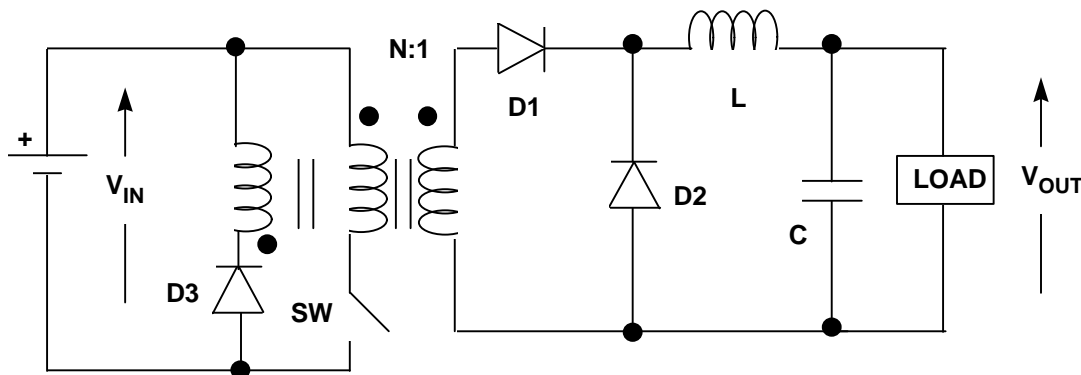
D = Duty Cycle

**Figure 3.24**

The basic *forward* converter topology is shown in Figure 3.25. It is derived from the buck converter. This topology avoids the problem of large stored energy in the transformer core. However, the circuit is more complex and requires an additional magnetic element (a transformer), an inductor, an additional transformer winding, plus three diodes. When the switch is on, current builds up in the primary winding and also in the secondary winding, where it is transferred to the load through diode D1. When the switch is on, the current in the inductor flows out of D1 from the transformer and is reflected back to the primary winding according to the turns ratio. Additionally, the current due to the input voltage applied across the primary inductance, called the *magnetizing current*, flows in the primary winding. When the switch is opened, the current in the inductor continues to flow through the load via the return path provided by diode D2. The load current is no longer reflected into the transformer, but the magnetizing current induced in the primary still requires a return path so that the transformer can be *reset*. Hence the extra *reset* winding and diode are needed.

The relationship between the input and output voltage is given by:

$$V_{OUT} = \frac{V_{IN}}{N} \cdot D.$$

**ISOLATED TOPOLOGY:  
FORWARD CONVERTER****(BUCK DERIVED)**

$$V_{OUT} = \frac{V_{IN}}{N} \cdot D$$

**D = Duty Cycle****Figure 3.25**

There are many other possible isolated switching regulator topologies which use transformers, however, the balance of this section will focus on non-isolated topologies because of their wider application in portable and distributed power systems.

**SWITCH MODULATION TECHNIQUES**

Important keys to understanding switching regulators are the various methods used to control the switch. For simplicity of analysis, the examples previously discussed used a simple fixed-frequency pulse width modulation (PWM) technique. There can be two other standard variations of the PWM technique: variable frequency constant on-time, and variable frequency constant off-time.

In the case of a buck converter, using a variable frequency constant off-time ensures that the peak-to-peak output ripple current (also the inductor current) remains constant as the input voltage varies. This is illustrated in Figure 3.26, where the output current is shown for two conditions of input voltage. Note that as the input voltage increases, the slope during the on-time increases, but the on-time decreases, thereby causing the frequency to increase. Constant off-time control schemes are popular for buck converters where a wide input voltage range must be accommodated. The ADP1147 family implements this switch modulation technique.

## CONTROL OF BUCK CONVERTER USING CONSTANT OFF-TIME, VARIABLE FREQUENCY PWM

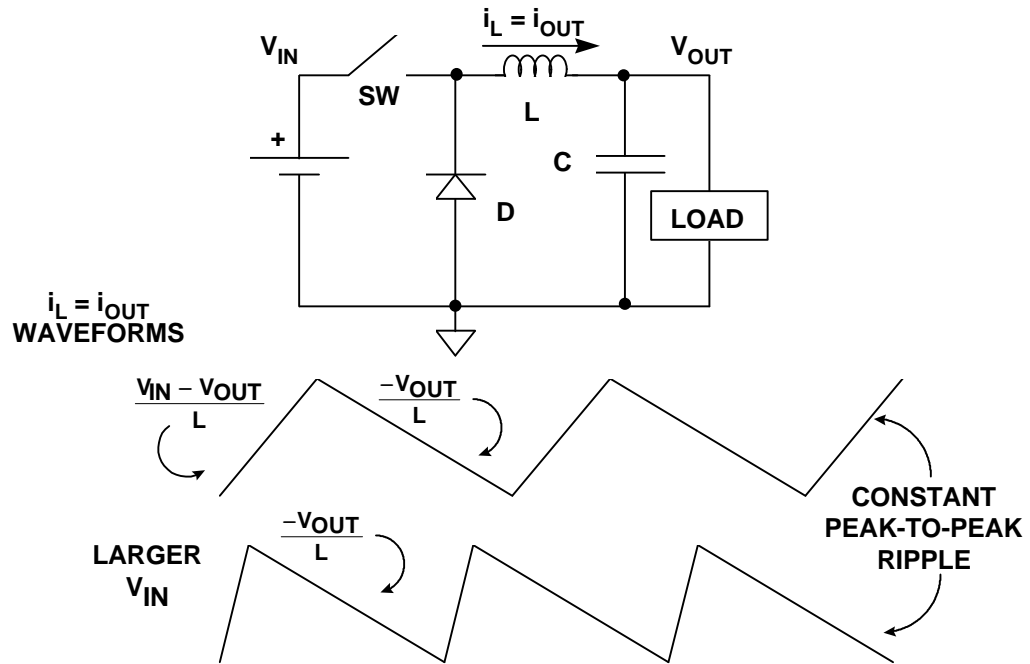


Figure 3.26

In the case of a boost converter, however, neither input ramp slopes nor output ramp slopes are solely a function of the output voltage (see Figure 3.15), so there is no inherent advantage in the variable frequency constant off-time modulation method with respect to maintaining constant output ripple current. Still, that modulation method tends to allow for less ripple current variation than does fixed frequency, so it is often used.

In the case where very low duty cycles are needed, e.g., under short circuit conditions, sometimes the limitation of a minimum achievable duty cycle is encountered. In such cases, in order to maintain a steady-state condition and prevent runaway of the switch current, a pulse skipping function must be implemented. This might take the form of a current monitoring circuit which detects that the switch current is too high to turn the switch on and ramp the current up any higher. So either a fixed frequency cycle is skipped without turning on the switch, or the off-time is extended in some way to delay the turn-on.

The pulse skipping technique for a fixed frequency controller can be applied even to operation at *normal* duty cycles. Such a switch modulation technique is then referred to as *pulse burst modulation* (PBM). At its simplest, this technique simply gates a fixed frequency, fixed duty cycle oscillator to be applied to the switch or not. The duty cycle of the oscillator sets the maximum achievable duty cycle for the converter, and smaller duty cycles are achieved over an average of a multiplicity of pulses by skipping oscillator cycles. This switch modulation method accompanies a simple control method of using a hysteretic comparator to monitor the output voltage versus a reference and decide whether to use the oscillator to turn on the switch for that cycle or not. The hysteresis of the comparator tends to give rise to several cycles of switching followed by several cycles of not switching. Hence, the

## **SWITCHING REGULATORS**

resulting switching signal is characterized by pulses which tend to come in bursts - hence the name for the modulation technique.

There are at least two inherent fundamental drawbacks of the PBM switch modulation technique. First, the constant variation of the duty cycle between zero and maximum produces high ripple currents and accompanying losses. Second, there is an inherent generation of subharmonic frequencies with respect to the oscillator frequency. This means that the noise spectrum is not well controlled, and often audible frequencies can be produced. This is often apparent in higher power converters which use pulse skipping to maintain short-circuit current control. An audible noise can often be heard under such a condition, due to the large magnetics acting like speaker coils. For these reasons, PBM is seldom used at power levels above ~10 Watts. But for its simplicity, it is often preferred below that power level, but above a power level or with a power conversion requirement where charge pumps are not well suited.

## **CONTROL TECHNIQUES**

Though often confused with or used in conjunction with discussing the switch modulation technique, the control technique refers to what parameters of operation are used and how they are used to control the modulation of the switch. The specific way in which the switch is modulated can be thought of separately, and was just presented in the previous section.

In circuits using PBM for switch modulation, the control technique typically used is a voltage-mode hysteretic control. In this implementation the switch is controlled by monitoring the output voltage and modulating the switch such that the output voltage oscillates between two hysteretic limits. The ADP3000 switching regulator is an example of a regulator which combines these modulation and control techniques.

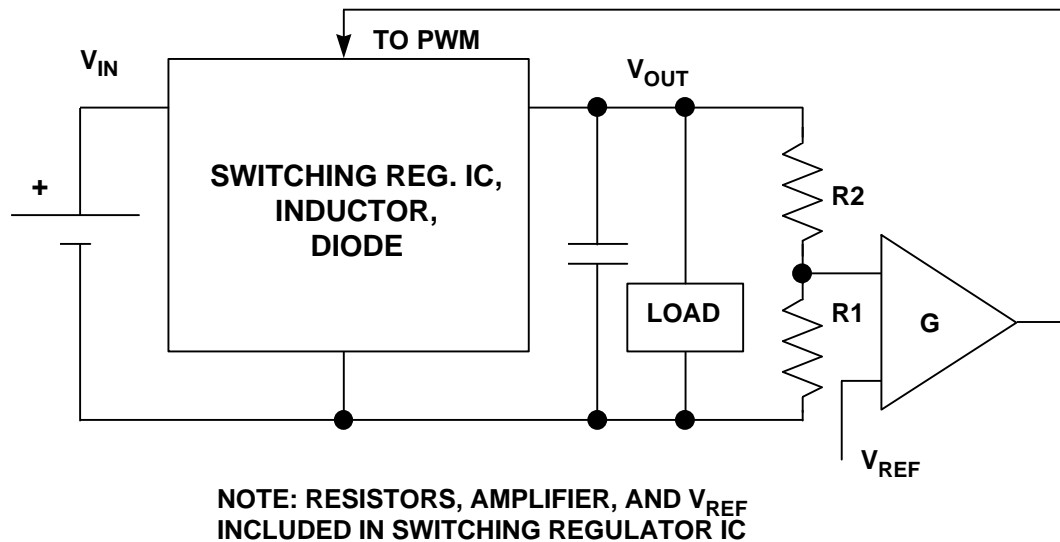
The most basic control technique for use with PWM is *voltage-mode (VM) control* (see Figure 3.27). Here, the output voltage is the only parameter used to determine how the switch will be modulated. An error amplifier (first mentioned in the Buck Converter section) monitors the output voltage, its error is amplified with the required frequency compensation for maintaining stability of the control loop, and the switch is modulated directly in accordance with that amplifier output.

The output voltage is divided down by a ratio-matched resistor divider and drives one input of an amplifier, G. A precision reference voltage ( $V_{REF}$ ) is applied to the other input of the amplifier. The output of the amplifier in turn controls the duty cycle of the PWM. It is important to note that the resistor divider, amplifier, and reference are actually part of the switching regulator IC, but are shown externally in the diagram for clarity. The output voltage is set by the resistor divider ratio and the reference voltage:

$$V_{OUT} = V_{REF} \left( 1 + \frac{R2}{R1} \right).$$

The internal resistor ratios and the reference voltage are set to produce standard output voltage options such as 12V, 5V, 3.3V, or 3V. In some regulators, the resistor divider can be external, allowing the output voltage to be adjusted.

### VOLTAGE FEEDBACK FOR PWM CONTROL



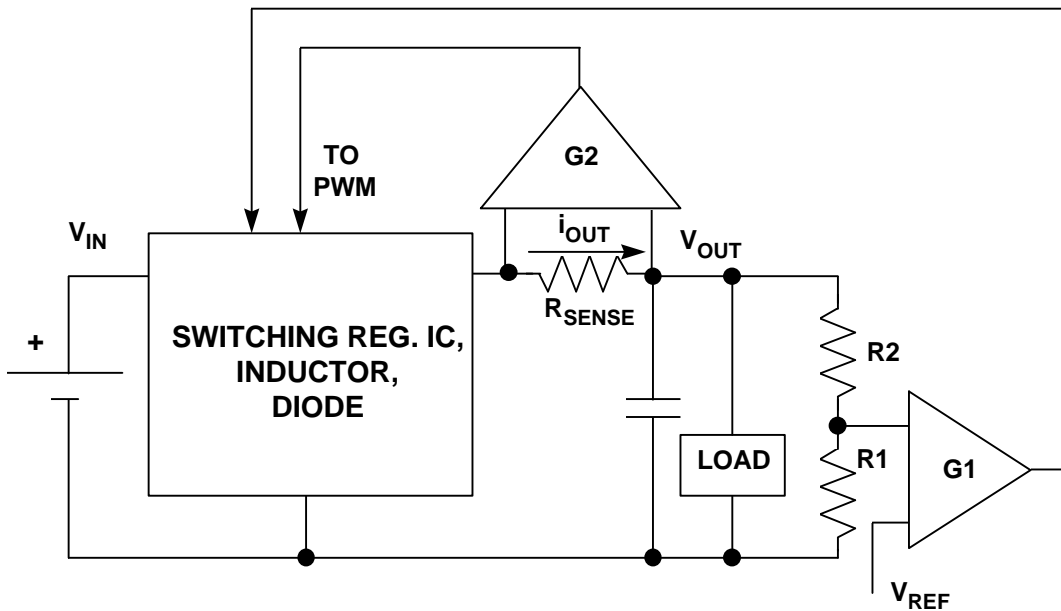
**Figure 3.27**

A simple modification of VM control is voltage *feedforward*. This technique adjusts the duty cycle automatically as the input voltage changes so that the feedback loop does not have to make an adjustment (or as much of an adjustment). Voltage feedforward can even be used in the simple PBM regulators. Feedforward is especially useful in applications where the input voltage can change suddenly or, perhaps due to current limit protection limitations, it is desirable to limit the maximum duty cycle to lower levels when the input voltage is higher.

In switchers, the VM control loop needs to be compensated to provide stability, considering that the voltage being controlled by the modulator is the average voltage produced at the switched node, whereas the actual output voltage is filtered through the switcher's LC filter. The phase shift produced by the filter can make it difficult to produce a control loop with a fast response time.

A popular way to circumvent the problem produced by the LC filter phase shift is to use current-mode (CM) control as shown in Figure 3.28. In current-mode control, it is still desirable, of course, to regulate the output voltage. Thus, an error amplifier (G1) is still required. However, the switch modulation is no longer controlled directly by the error amplifier. Instead, the inductor current is sensed, amplified by G2, and used to modulate the switch in accordance with the command signal from the [output voltage] error amplifier. It should be noted that the divider network,  $V_{REF}$ , G1 and G2 are usually part of the IC switching regulator itself, rather than external as shown in the simplified diagram.

**CURRENT FEEDBACK FOR PWM CONTROL**



**NOTE: RESISTORS, AMPLIFIERS, AND  $V_{REF}$  INCLUDED IN SWITCHING REGULATOR IC**

**Figure 3.28**

The CM control system uses feedback from both the output voltage and output current. Recall that at the beginning of each PWM cycle, the switch turns on, and the inductor current begins to rise. The inductor current develops a voltage across the small sense resistor,  $R_{SENSE}$ , which is amplified by G2 and fed back to the PWM controller to turn off the switch. The output voltage, sensed by amplifier G1 and also fed back to the PWM controller, sets the level at which the peak inductor current will terminate the switch on-time. Since it is inductor current that turns off the switch (and thereby sets the duty cycle) this method is commonly referred to as *current-mode* control, even though there are actually two feedback control loops: the fast responding current loop, and the slower responding output voltage loop. Note that inductor current is being controlled on a pulse-by-pulse basis, which simplifies protection against switch over-current and inductor saturation conditions.

In essence, then, in CM control, rather than controlling the average voltage which is applied to the LC filter as in VM control, the inductor current is controlled directly on a cycle-by-cycle basis. The only phase shift remaining between the inductor current and the output voltage is that produced by the impedance of the output capacitor(s). The correspondingly lower phase shift in the output filter allows the loop response to be faster while still remaining stable. Also, instantaneous changes in input voltage are immediately reflected in the inductor current, which provides excellent line transient response. The obvious disadvantage of CM control is the requirement of sensing current and, if needed, an additional amplifier. With increasingly higher performance requirements in modern electronic equipment, the performance advantage of CM control typically outweighs the cost of

implementation. Also, some sort of current limit protection is often required, whatever the control technique. Thus it tends to be necessary to implement some sort of current sensing even in VM-controlled systems.

Now even though we speak of a CM controller as essentially controlling the inductor current, more often than not the switch current is controlled instead, since it is more easily sensed (especially in a switching regulator) and it is a representation of the inductor current for at least the on-time portion of the switching cycle. Rather than actually controlling the average switch current, which is not the same as the average inductor current anyway, it is often simpler to control the peak current - which is the same for both the switch and the inductor in all the basic topologies. The error between the average inductor current and the peak inductor current produces a non-linearity within the control loop. In most systems, that is not a problem. In other systems, a more precise current control is needed, and in such a case, the inductor current is sensed directly and amplified and frequency-compensated for the best response.

Other control variations are possible, including *valley* rather than peak control, *hysteretic current* control, and even *charge* control - a technique whereby the integral of the inductor current (i.e., charge) is controlled. That eliminates even the phase shift of the output capacitance from the loop, but presents the problem that instantaneous current is not controlled, and therefore short-circuit protection is not inherent in the system. All techniques offer various advantages and disadvantages. Usually the best tradeoff between performance and cost/simplicity is peak-current control - as used by the ADP1147 family. This family also uses the current-sense output to control a *sleep, or power saving mode* of operation to maintain high efficiency for low output currents.

### **GATED OSCILLATOR (PULSE BURST MODULATION) CONTROL EXAMPLE**

All of the PWM techniques discussed thus far require some degree of feedback loop compensation. This can be especially tricky for boost converters, where there is more phase shift between the switch and the output voltage.

As previously mentioned, a technique which requires no feedback compensation uses a fixed frequency gated oscillator as the switch control (see Figure 3.29). This method is often (incorrectly) referred to as the Pulse Frequency Modulation (PFM) mode, but is more correctly called *pulse burst modulation (PBM)* or *gated-oscillator* control.

The output voltage ( $V_{OUT}$ ) is divided by the resistive divider ( $R1$  and  $R2$ ) and compared against a reference voltage,  $V_{REF}$ . The comparator hysteresis is required for stability and also affects the output voltage ripple. When the resistor divider output voltage drops below the comparator threshold ( $V_{REF}$  minus the hysteresis voltage), the comparator starts the gated oscillator. The switcher begins switching again which then causes the output voltage to increase until the comparator threshold is reached ( $V_{REF}$  plus the hysteresis voltage), at which time the oscillator is turned off. When the oscillator is off, quiescent current drops to a very low value

## SWITCHING REGULATORS

(for example,  $95\mu\text{A}$  in the ADP1073) making PBM controllers very suitable for battery-powered applications.

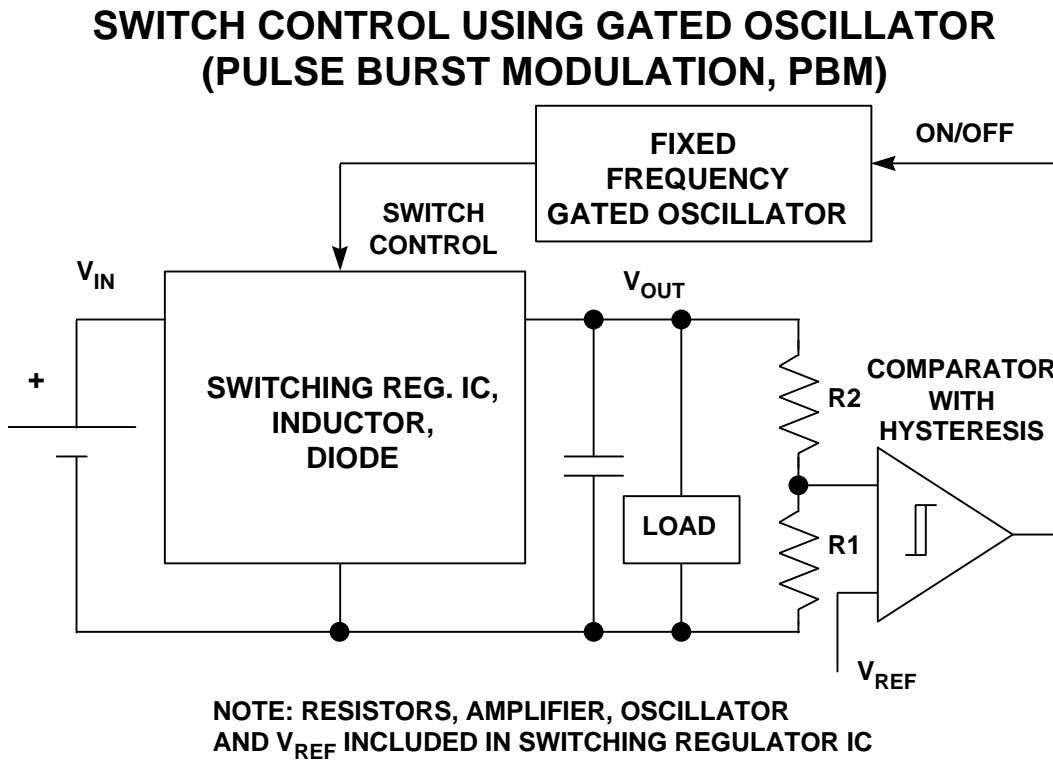


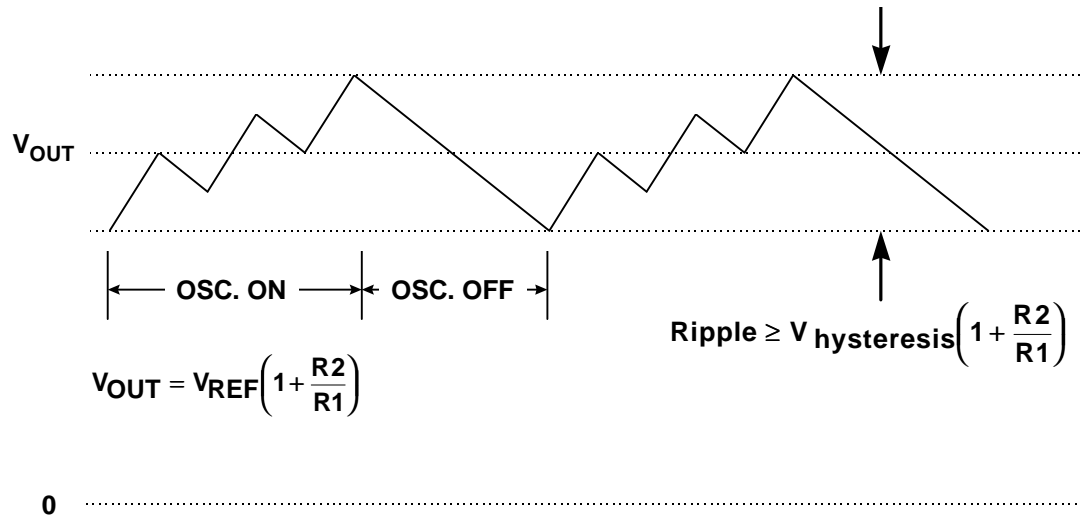
Figure 3.29

A simplified output voltage waveform is shown in Figure 3.30 for a PBM buck converter. Note that the comparator hysteresis voltage multiplied by the reciprocal of the attenuation factor primarily determines the peak-to-peak output voltage ripple (typically between 50 and 100mV). It should be noted that the actual output voltage ripple waveform can look quite different from that shown in Figure 3.30 depending on the design and whether the converter is a buck or boost.

A practical switching regulator IC using the PBM approach is the ADP3000, which has a fixed switching frequency of 400kHz and a fixed duty cycle of 80%. This device is a versatile step-up/step-down converter. It can deliver an output current of 100mA in a 5V to 3V step-down configuration and 180mA in a 2V to 3.3V step-up configuration. Input supply voltage can range between 2V and 12V in the boost mode, and up to 30V in the buck mode. It should be noted that when the oscillator is turned off, the internal switch is opened so that the inductor current does not continue to increase.



**REPRESENTATIVE OUTPUT VOLTAGE WAVEFORM FOR  
GATED OSCILLATOR CONTROLLED (PBM)  
BUCK REGULATORS**



**Figure 3.30**

In the gated-oscillator method, the comparator hysteresis serves to stabilize the feedback loop making the designs relatively simple. The disadvantage, of course, is that the peak-to-peak output voltage ripple can never be less than the comparator hysteresis multiplied by the reciprocal of the attenuation factor:

$$\text{Output Ripple} \geq V_{\text{hysteresis}} \left( \frac{R2}{R1} \right).$$

Because the gated-oscillator (PBM) controlled switching regulator operates with a fixed duty cycle, output regulation is achieved by changing the number of “skipped pulses” as a function of load current and voltage. From this perspective, PBM controlled switchers tend to operate in the “discontinuous” mode under light load conditions. Also, the maximum average duty cycle is limited by the built-in duty cycle of the oscillator. Once the required duty cycle exceeds that limit, no pulse skipping occurs, and the device will lose regulation.

One disadvantage of the PBM switching regulator is that the frequency spectrum of the output ripple is “fuzzy” because of the burst-mode of operation. Frequency components may fall into the audio band, so proper filtering of the output of such a regulator is mandatory.

Selection of the inductor value is also more critical in PBM regulators. Because the regulation is accomplished with a burst of fixed duty cycle pulses (i.e., higher than needed on average) followed by an extended off time, the energy stored in the inductor during the burst of pulses must be sufficient to supply the required energy to the load. If the inductor value is too large, the regulator may never start up, or may have poor transient response and inadequate line and load regulation. On the other hand, if the inductor value is too small, the inductor may saturate during the charging time, or the peak inductor current may exceed the maximum rated switch

## SWITCHING REGULATORS

current. However, devices such as the ADP3000 incorporate on-chip overcurrent protection for the switch. An additional feature allows the maximum peak switch current to be set with an external resistor, thereby preventing inductor saturation. Techniques for selecting the proper inductor value will be discussed in a following section.

## DIODE AND SWITCH CONSIDERATIONS

So far, we have based our discussions around an ideal lossless switching regulator having ideal circuit elements. In practice, the diode, switch, and inductor all dissipate power which leads to less than 100% efficiency.

Figure 3.31 shows typical buck and boost converters, where the switch is part of the IC. The process is bipolar, and this type of transistor is used as the switching element. The ADP3000 and its relatives (ADP1108, ADP1109, ADP1110, ADP1111, ADP1073, ADP1173) use this type of internal switch.

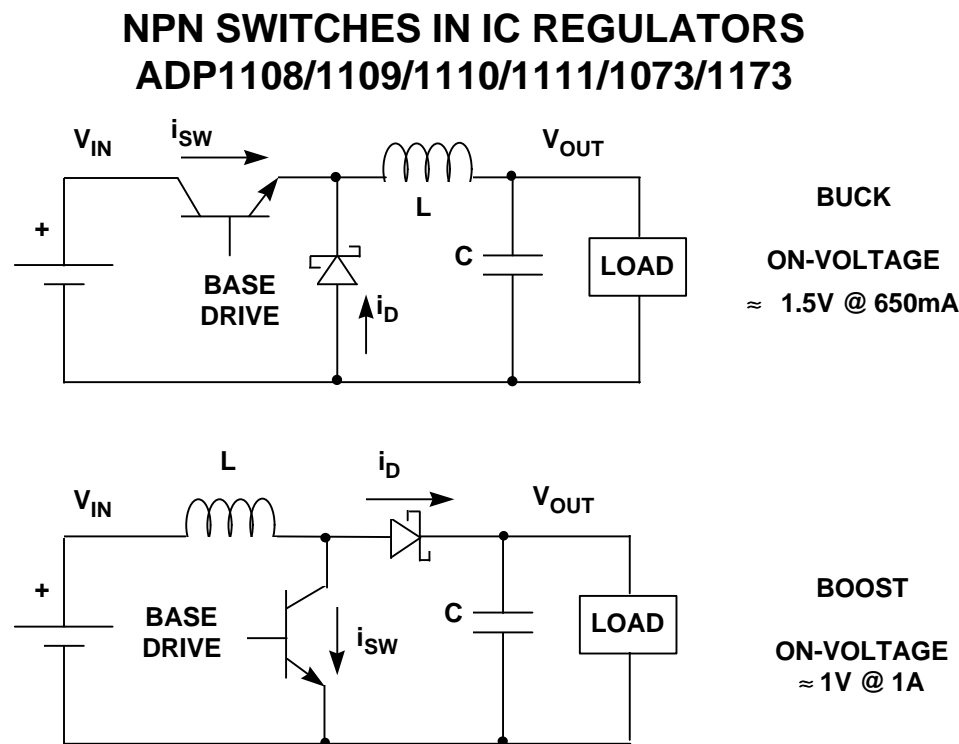


Figure 3.31

The diode is external to the IC and must be chosen carefully. Current flows through the diode during the off-time of the switching cycle. This translates into an average current which causes power dissipation because of the diode forward voltage drop. The power dissipation can be minimized by selecting a Schottky diode with a low forward drop (0.5V), such as the 1N5818-type. It is also important that the diode capacitance and recovery time be low to prevent additional power loss due to charging current, and this is also afforded by the Schottky diode. Power dissipation can be approximated by multiplying the average diode current by the forward voltage drop.

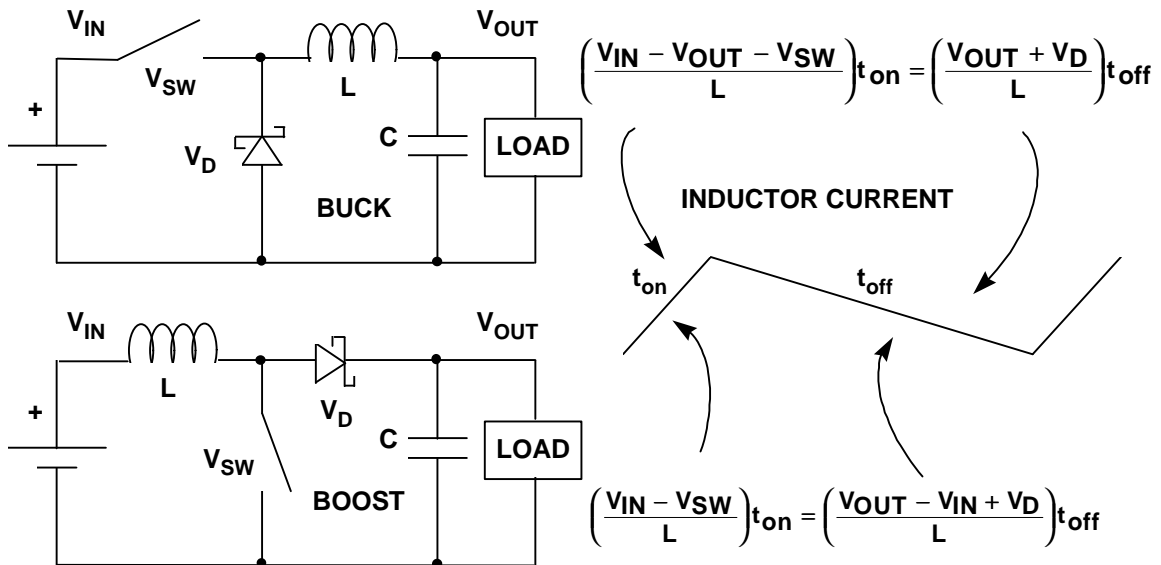
The drop across the NPN switch also contributes to internal power dissipation. The power (neglecting switching losses) is equal to the average switch current multiplied by the collector-emitter on-state voltage. In the case of the ADP3000 series, it is 1.5V at the maximum rated switch current of 650mA (when operating in the buck mode).

In the boost mode, the NPN switch can be driven into saturation, so the on-state voltage is reduced, and thus, so is the power dissipation. Note that in the case of the ADP3000, the saturation voltage is about 1V at the maximum rated switch current of 1A.

In examining the two configurations, it would be logical to use a PNP switching transistor in the buck converter and an NPN transistor in the boost converter in order to minimize switch voltage drop. However, the PNP transistors available on processes which are suitable for IC switching regulators generally have poor performance, so the NPN transistor must be used for both topologies.

In addition to lowering efficiency by their power dissipation, the switching transistors and the diode also affect the relationship between the input and output voltage. The equations previously developed assumed zero switch and diode voltage drops. Rather than re-deriving all the equations to account for these drops, we will examine their effects on the inductor current for a simple buck and boost converter operating in the continuous mode as shown in Figure 3.32.

**EFFECTS OF SWITCH AND DIODE VOLTAGE ON INDUCTOR CURRENT EQUATIONS**



**Figure 3.32**

In the buck converter, the voltage applied to the inductor when the switch is on is equal to  $V_{IN} - V_{OUT} - V_{SW}$ , where  $V_{SW}$  is the approximate average voltage drop across the switch. When the switch is off, the inductor current is discharged into a

## SWITCHING REGULATORS

voltage equal to  $V_{OUT} + V_D$ , where  $V_D$  is the approximate average forward drop across the diode. The basic inductor equation used to derive the relationship between the input and output voltage becomes:

$$\left(\frac{V_{IN} - V_{OUT} - V_{SW}}{L}\right)t_{on} = \left(\frac{V_{OUT} + V_D}{L}\right)t_{off}.$$

In the actual regulator circuit, negative feedback will force the duty cycle to maintain the correct output voltage, but the duty cycle will also be affected by the switch and the diode drops to a lesser degree.

When the switch is on in a boost converter, the voltage applied to the inductor is equal to  $V_{IN} - V_{SW}$ . When the switch is off, the inductor current discharges into a voltage equal to  $V_{OUT} - V_{IN} + V_D$ . The basic inductor current equation becomes:

$$\left(\frac{V_{IN} - V_{SW}}{L}\right)t_{on} = \left(\frac{V_{OUT} - V_{IN} + V_D}{L}\right)t_{off}.$$

From the above equations, the basic relationships between input voltage, output voltage, duty cycle, switch, and diode drops can be derived for the buck and boost converters.

The ADP3000 is a switching regulator that uses the NPN-type switch just discussed. A block diagram is shown in Figure 3.33 and key specifications are given in Figure 3.34.

### **ADP3000 SWITCHING REGULATOR BLOCK DIAGRAM**

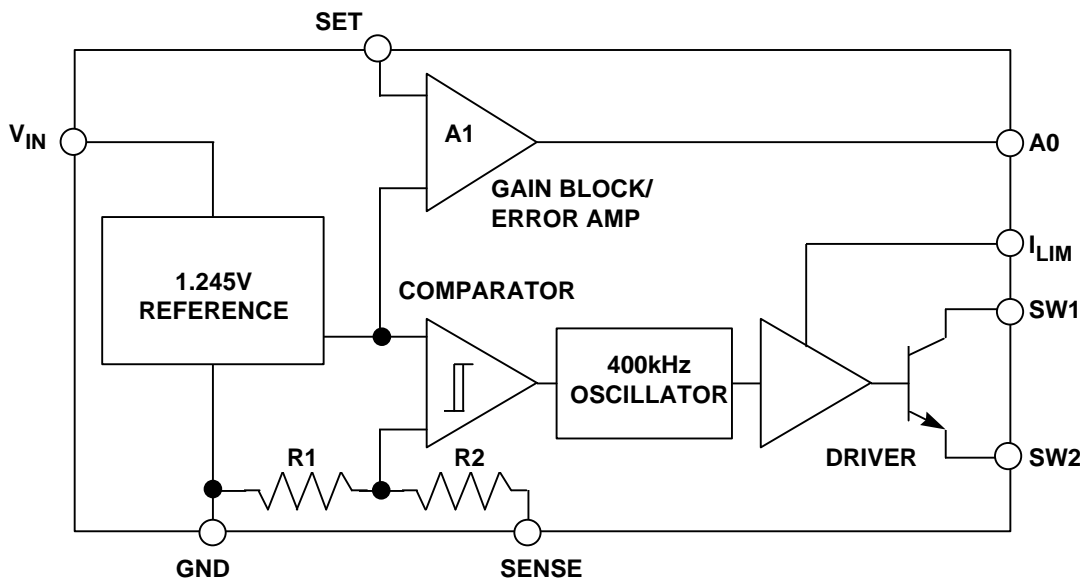


Figure 3.33

## ADP3000 SWITCHING REGULATOR KEY SPECIFICATIONS

- Input Voltages from 2V to 12V (Step-Up),  
2V to 30V (Step-Down)
- Fixed 3.3V, 5V, 12V and Adjustable Output Voltage
- Step-Up or Step-Down Mode
- PBM (Gated Oscillator) Control Simplifies Design
- 50mV Typical Output Ripple Voltage (5V Output)
- 400kHz Switching Frequency Allows Low Value Inductors
- 80% Duty Cycle
- 500 $\mu$ A Quiescent Current
- Output Drive Capability:
  - 100mA @ 3V from 5V Input in Step-Down Mode
  - 180mA @ 3.3V from 2V Input in Step-Up Mode
- 8-Pin DIP or SOIC Package

Figure 3.34

The device uses the gated oscillator, or pulse burst modulation (PBM) feedback control scheme. The internal oscillator operates at a frequency of 400kHz allowing the use of small value inductors and capacitors. The internal resistors, R1 and R2, set the output voltage to 3.3V, 5V, or 12V, depending upon the option selected. A completely adjustable version is also available where the comparator input is brought out directly to the "SENSE" pin, and the user provides the external divider resistors. Total quiescent current is only 500 $\mu$ A. The uncommitted gain block, A1, can be used as a low-battery detector or to reduce output hysteretic ripple limits by adding gain in the feedback loop. A current-limit pin, I<sub>LIM</sub>, allows switch current to be limited with an external resistor. Limiting the switch current on a cycle by cycle basis allows the use of small inductors with low saturation current. It also allows physically small tantalum capacitors with a typical ESR of 0.1 $\Omega$  to achieve an output ripple voltage as low as 40 to 80mV, as well as low input ripple current.

A typical ADP3000 boost application circuit is shown in Figure 3.35. The input voltage can range from +2V to +3.2V. The output is +5V and supplies a load current of 100mA. Typical efficiency for the circuit is 80%. All components are available in surface mount.

The ADP3000 can also be used in the buck configuration as shown in Figure 3.36. The input voltage to the regulator is between 5V and 6V, and the output is 3V at 100mA. Note that in this case, the adjustable version of the ADP3000 is used. The external divider resistors, R1 and R2, are chosen to set the nominal output voltage to 3V. All components are available in surface mount, and the efficiency of the circuit is approximately 75%.

**ADP3000 2V TO 5V BOOST APPLICATION**

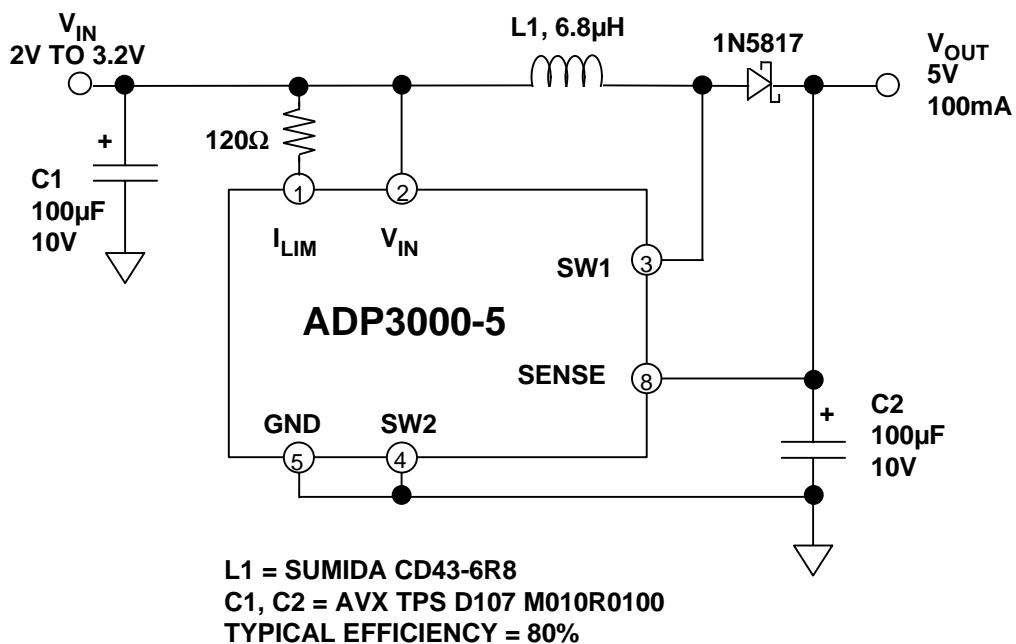


Figure 3.35

**ADP3000 5V TO 3V BUCK APPLICATION**

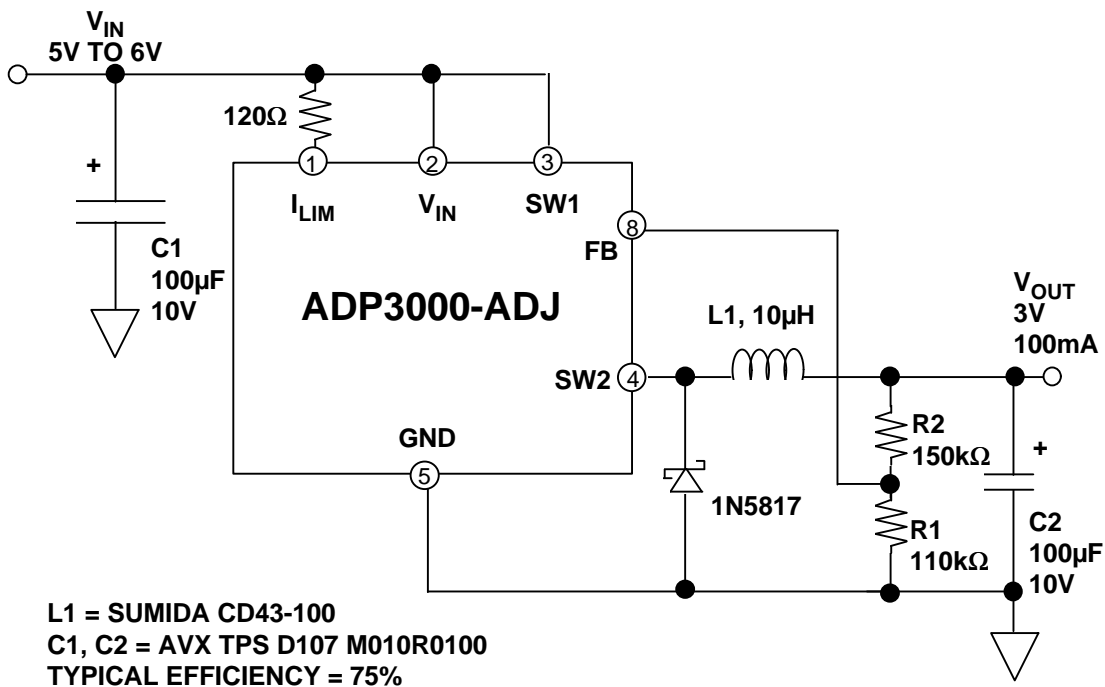
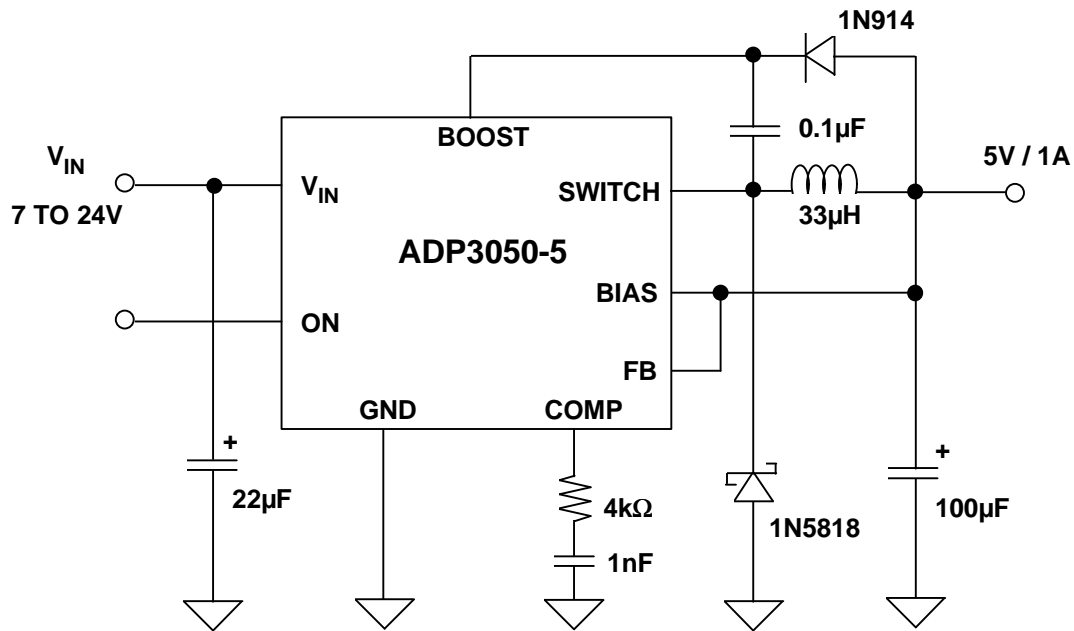


Figure 3.36

The ADP3050 is a 1.5A buck converter with an internal saturable NPN switch. It utilizes PWM current-mode control and operates at a fixed 250kHz switching frequency. An application circuit for the device is shown in Figure 3.37 and key specifications are summarized in Figure 3.38. A special boosted drive stage is used to saturate an NPN power switch, providing a system efficiency higher than conventional bipolar buck switchers. An external diode and capacitor provide the boosted voltage to the drive stage that is higher than the input supply voltage. A shutdown signal places the device in a low power mode, reducing the supply current to under 15 $\mu$ A. The ADP3050 provides excellent line and load regulation, maintaining  $\pm 2.5\%$  output voltage accuracy over an ambient operating range of  $-40^{\circ}\text{C}$  to  $+85^{\circ}\text{C}$ . The ADP3050 package (8-pin SOIC footprint) is thermally enhanced and has a junction-to-ambient thermal resistance of approximately  $90^{\circ}\text{C}/\text{W}$ .

**ADP3050 250kHz, 1.5A BUCK REGULATOR**



**Figure 3.37**

For high current output switchers, external power MOSFETs are often used as switches. The basic buck and boost converter circuits using MOSFETs are shown in Figure 3.39. On-resistances are typically  $0.006\Omega$  -  $0.1\Omega$ , depending upon power and efficiency requirements. The MOSFETs are generally discrete devices and are rarely integrated onto the IC regulator. The regulator generates the appropriate gate drive signal to the MOSFET.

**ADP3050 BUCK REGULATOR KEY SPECIFICATIONS**

- **Input Voltage Range: 3.6V to 24V**
- **3.3V, 5V, and Adjustable Output Versions**
- **0.5Ω Saturating NPN Switch**
- **250kHz Switching Frequency**
- **Current-Mode Control**
- **Cycle-by-Cycle Current Limit**
- **Shutdown Feature Reduces Current to 15μA**
- **8-Pin SOIC Thermally Enhanced Package,**  
 $\theta_{JA} \approx 90^{\circ}\text{C/W}$

**Figure 3.38**

The main selection criteria for the power MOSFET is the peak current rating, threshold voltage, and the on-resistance. The minimum regulator input voltage determines whether a standard threshold or logic-level threshold MOSFET must be used. For input voltages greater than 8V, a standard threshold MOSFET with a threshold voltage of less than 4V can be used. If the input voltage is expected to drop below 8V, a logic-level MOSFET is recommended. In applications involving high current outputs and input voltages less than 8V, it may be necessary to drive the MOSFET gates with circuits which operate on a higher voltage, such as 12V. If this voltage is not available in the system, it can be derived from the input voltage using charge pump techniques (described in a later section) since the current requirements of the drive circuits are typically fairly low.

The  $I^2R$  loss in this type of regulator can be quite low because of the low MOSFET on-resistance, however one source of internal power dissipation which must not be overlooked is the gate charge required to turn the MOSFET on and off. The gate drive signal must overcome the gate capacitance (typically 1000 to 3000pF, and is directly proportional to physical size and the current-handling capability of the MOSFET). This current must be supplied by the input power supply and adds to the overall regulator power dissipation. It can be a significant contributor to efficiency reduction, up to 2 or 3% for output currents of 100 to 200mA.

Note that gate charge loss increases directly with both input voltage and operating frequency. This is the principal reason why highest efficiency circuits which utilize this topology operate at moderate frequencies of 200kHz or less. Furthermore, it argues against using a larger MOSFET than necessary to control on-resistance  $I^2R$  loss at the maximum expected output current.



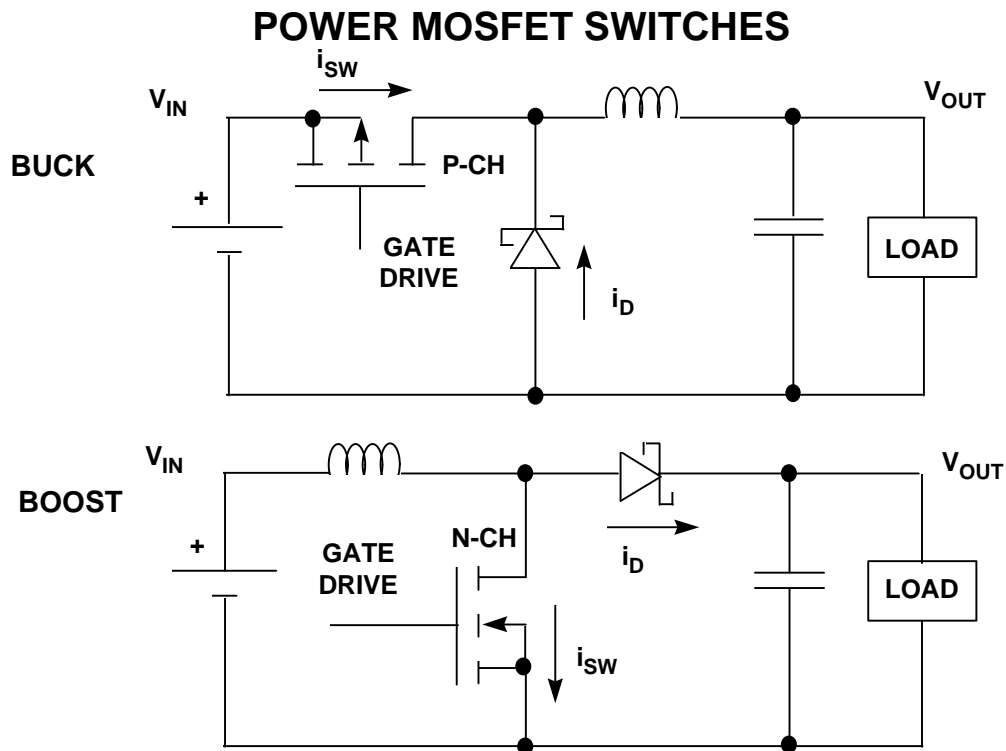


Figure 3.39

Power MOSFET switches allow current levels greater than 1A at high efficiencies (greater than 90%) using ICs such as the ADP1147 buck converter controller. The input voltage for the ADP1147 can range from 3.5V to 14V. Two output voltage versions are available : 3.3V (ADP1147-3.3) and 5V (ADP1147-5). The ADP1147 regulator controller operates in a constant off-time, variable frequency control mode with current-mode control. Operating in the constant off-time mode maintains constant inductor ripple current, thereby easing the output filter design. High efficiency is maintained at low output currents by switching automatically into a power-saving (PBM) mode.

A typical step-down application of the ADP1147 is shown in Figure 3.40. Input voltage in the circuit can range from 5.2V to 14V, and the output is 5V at 2A. The external resistor ( $R_C$ ) and capacitor ( $C_C$ ) connected to the  $I_{TH}$  pin serves to control the frequency response of the voltage feedback loop. The off-time of the regulator is determined by the external  $C_T$  capacitor. Current feedback is obtained from the voltage developed across the external  $R_{SENSE}$  resistor.

Typical efficiency is shown in the composite Figure 3.41, where the contributions of each source of efficiency loss are given (switch  $I^2R$  loss, gate charge loss, quiescent power loss, and Schottky diode loss). The lower curve represents the total efficiency. Note that for output currents between about 100mA and 1A, the power required to drive the MOSFET gate (gate charge) is the largest contributor to efficiency loss. At the higher current levels, the  $I^2R$  loss due to MOSFET on resistance dominates. Key specifications for the ADP1147 are summarized in Figure 3.42.

**HIGH EFFICIENCY STEP DOWN REGULATOR  
USING THE ADP1147 CONTROLLER**

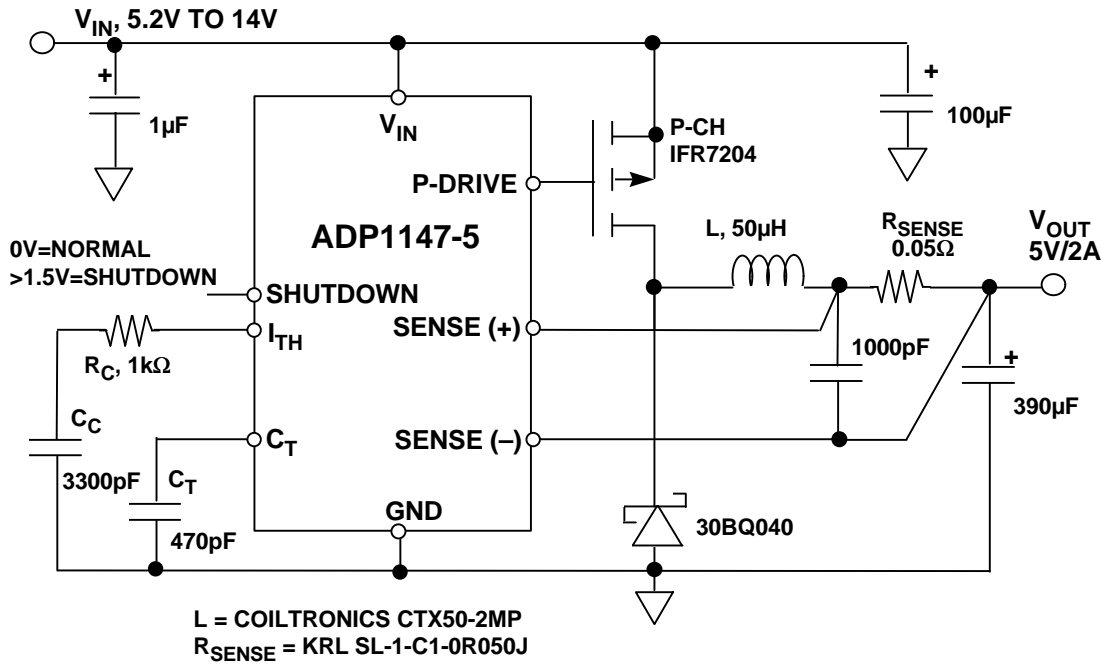


Figure 3.40

**ADP1147 TYPICAL EFFICIENCY LOSSES**

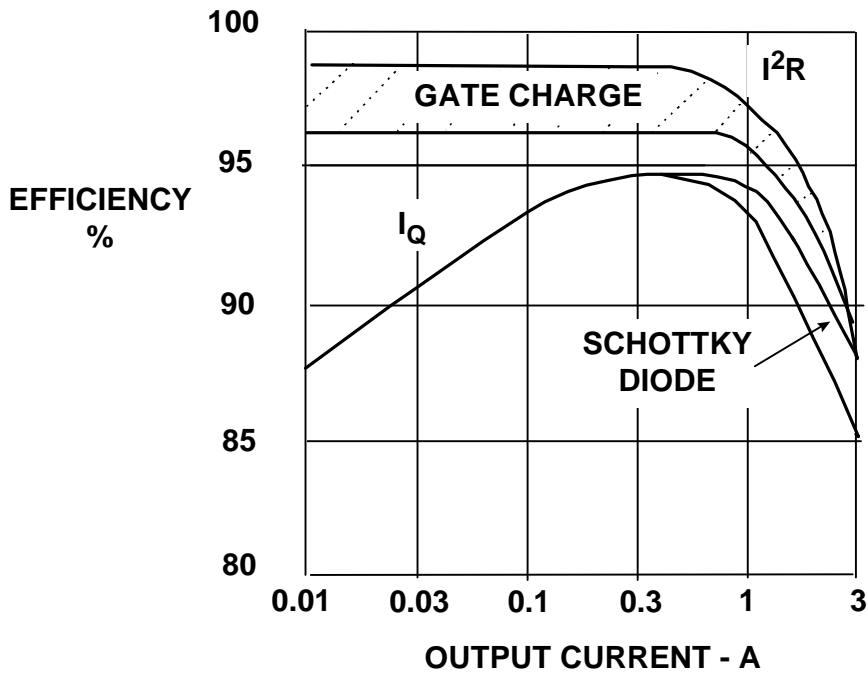


Figure 3.41

## ADP1147 STEP-DOWN REGULATOR CONTROLLER KEY SPECIFICATIONS

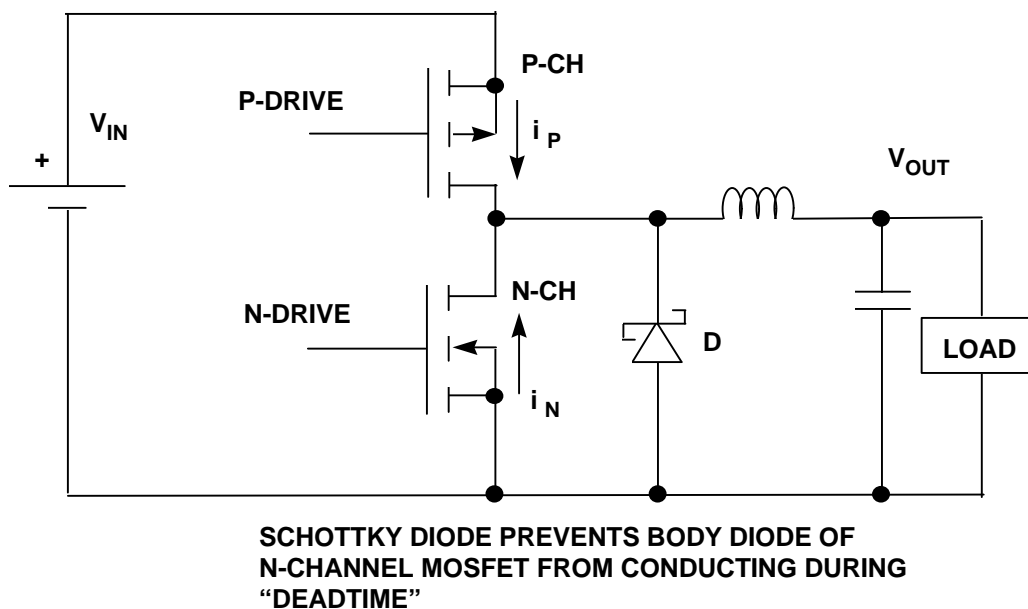
- Input Voltage Range: 3.5V to 14V (16V Max.)
- Output Voltage Options: 3.3V, 5V
- Current-Mode Control Circuit
- Constant Off-Time (5 $\mu$ s), Variable Frequency
- P-Channel MOSFET Gate Drive Output
- Power Saving Mode: 160 $\mu$ A Typical
- Up to 95% Efficiency
- 8-Pin SOIC and DIP Packages

Figure 3.42

In order to achieve even higher efficiency, the Schottky diode can be replaced with an N-channel MOSFET switch as shown in Figure 3.43. This configuration is referred to as a *synchronous rectifier*, or *synchronous switch*, because the switching of the N-channel MOSFET switch must be synchronized to the switching of the P-channel MOSFET switch so that it essentially passes the current in one direction and blocks it in the other direction, just like a rectifier or diode. This terminology does not imply that the switching *frequency* of the regulator is synchronized to an external clock.

The gate drive signals from the controller must be non-overlapping to prevent cross-conduction current spikes in the switches. This means that there is a period of time when both switches are off. The external Schottky diode prevents the body diode of the N-channel MOSFET from conducting during this time. It is not always necessary to add the Schottky diode, but it will increase overall efficiency slightly even if not required.

## BUCK CONVERTER WITH SYNCHRONOUS SWITCH USING P AND N-CANNEL MOSFETS



**Figure 3.43**

The ADP1148 is a high efficiency synchronous step-down switching regulator controller with an input voltage range of 3.5V to 18V. It utilizes a constant off-time, variable frequency current-mode control topology and is available in three versions: the ADP1148-3.3 (3.3V output), the ADP1148-5 (5V output), and the ADP1148 (adjustable output). At low output currents, the device switches into a power-saving mode to maintain high efficiency.

An application circuit for the ADP1148 synchronous step-down regulator controller is shown in Figure 3.44. Operation of the ADP1148 is similar to the ADP1147 with the addition of the drive circuitry for the synchronous N-channel MOSFET. The input voltage can range from 5.2V to 18V, and the output is 5V at 2A. A breakdown of the ADP1148 efficiency losses is shown in Figure 3.45, where the lower curve represents the total efficiency. Key specifications for the device are given in Figure 3.46.

### HIGH EFFICIENCY STEP-DOWN REGULATOR USING THE ADP1148 CONTROLLER

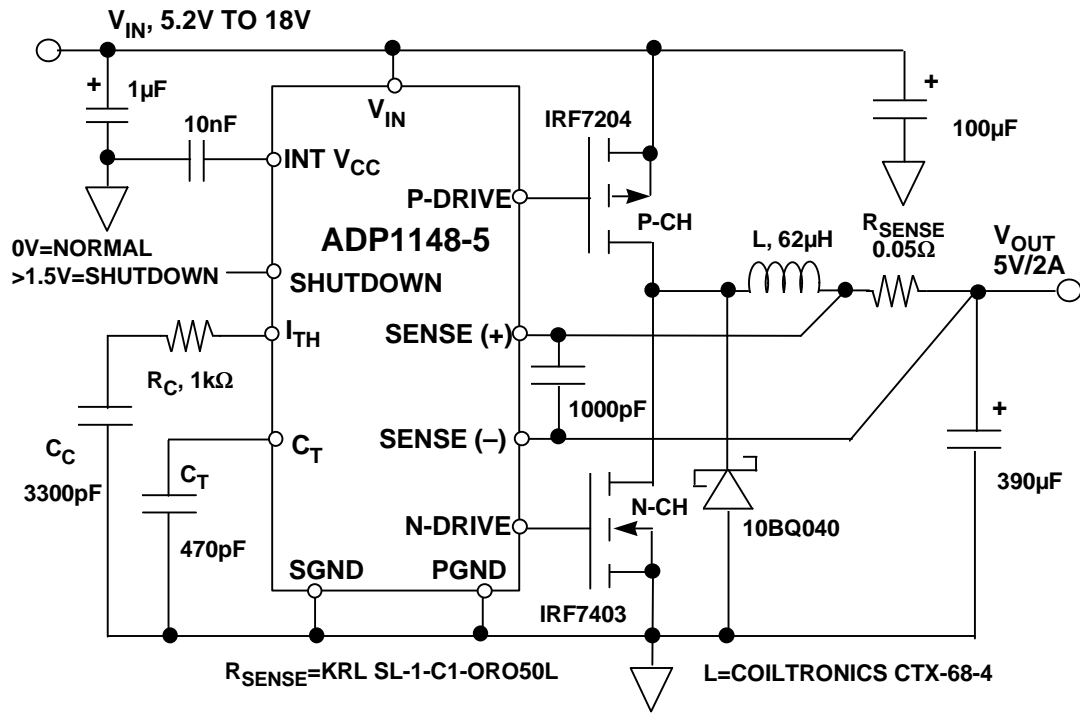


Figure 3.44

### ADP1148 TYPICAL EFFICIENCY LOSSES

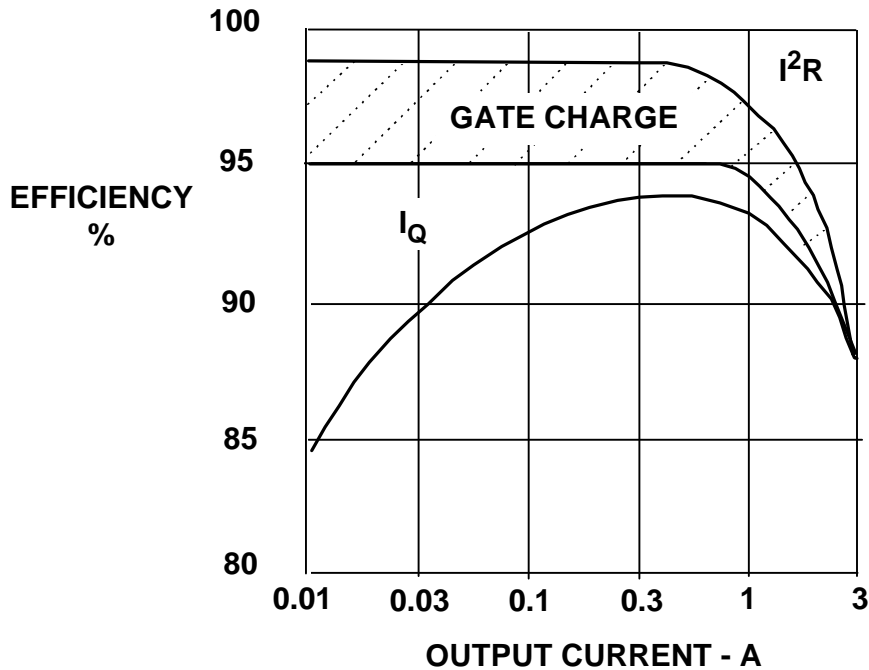


Figure 3.45

## **ADP1148 HIGH EFFICIENCY SYNCHRONOUS SWITCH REGULATOR CONTROLLER KEY SPECIFICATIONS**

- **Input Voltage Range: 3.5V to 18V (20V Max.)**
- **Output Voltage Options: 3.3V, 5V, and Adjustable**
- **Current-Mode Control Circuit**
- **Non-Overlapping P and N-Channel MOSFET  
Gate Drive Outputs**
- **Constant Off-Time (5 $\mu$ s), Variable Frequency**
- **Power Saving Mode: 160 $\mu$ A Typical**
- **Up to 95% Efficiency Possible**
- **14-Pin SOIC and DIP Packages**

**Figure 3.46**

The ADP3153 is a 5-bit programmable synchronous switching regulator controller suitable for the Pentium II processor. An application circuit is shown in Figure 3.47, and key specifications are given in Figure 3.48. The ADP3153 is optimized for applications where 5V is stepped down to a digitally controlled output voltage between 1.8V and 3.5V. Using a 5-bit DAC to read a voltage identification (VID) code directly from the processor, the ADP3153 generates the precise output voltage by using a current mode constant off-time topology to drive two N-channel MOSFETs at a nominal switching frequency of 250kHz. The constant off-time topology maintains constant inductor ripple current, and current mode operation together with an optimal compensation design provide excellent line and load transient response. The current limit level is user programmable with an external current sense resistor.

**ADP3153 POWER SUPPLY CONTROLLER FOR PENTIUM II  
(SIMPLIFIED SCHEMATIC)**

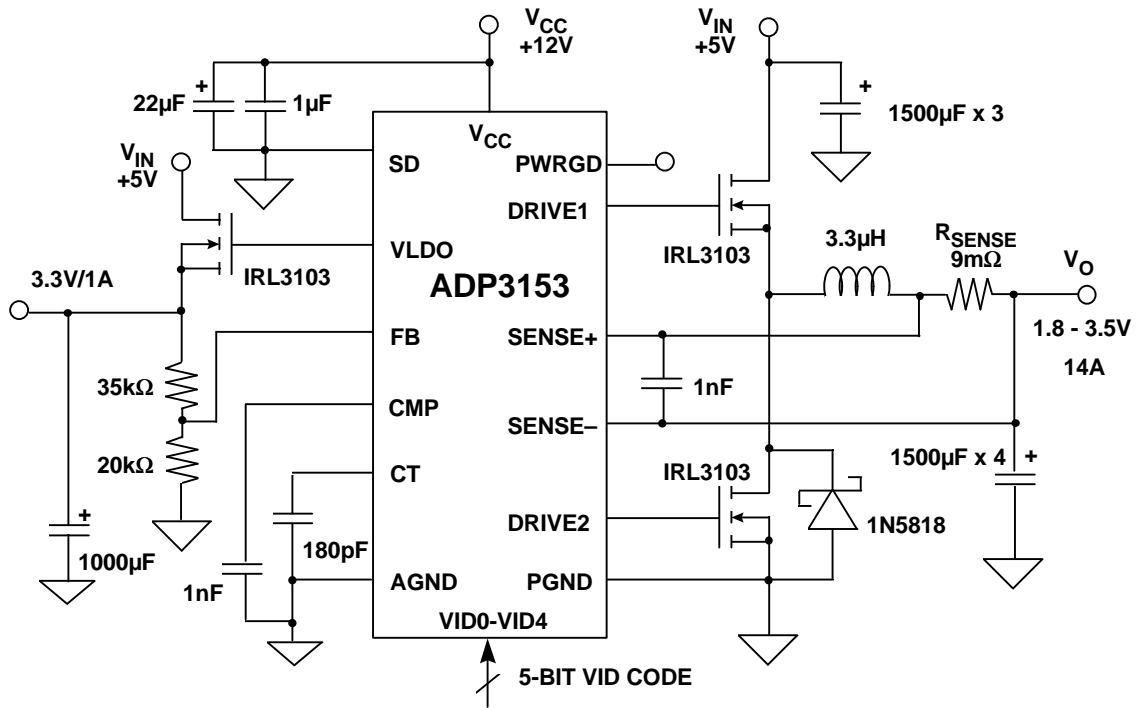


Figure 3.47

**ADP3153 VID-PROGRAMMABLE MICROPROCESSOR  
CONTROLLER KEY SPECIFICATIONS**

- 5-Bit Digitally Programmable 1.8V to 3.5V Output Voltage
- Dual N-Channel Driver Outputs
- Output Accuracy:  $\pm 1\%$  ( $0^{\circ}\text{C}$  to  $+70^{\circ}\text{C}$ )
- Constant Off-Time, Variable Frequency Current-Mode Control
- On-Chip Adjustable Linear Regulator Controller
- 20-Lead TSSOP Package
- Suitable for Pentium II, Pentium Pro, AMD-K6 Processors

Figure 3.48

## INDUCTOR CONSIDERATIONS

The selection of the inductor used in a switching regulator is probably the most difficult part of the design. Fortunately, manufacturers of switching regulators supply a wealth of applications information, and standard off-the-shelf inductors from well-known and reliable manufacturers are quite often recommended on the switching regulator data sheet. However, it is important for the design engineer to understand at least some of the fundamental issues relating to inductors. This discussion, while by no means complete, will give some insight into the relevant magnetics issues.

Selecting the actual value for the inductor in a switching regulator is a function of many parameters. Fortunately, in a given application the exact value is generally not all that critical, and equations supplied on the data sheets allow the designer to calculate a minimum and maximum acceptable value. That's the easy part.

Unfortunately, there is more to a simple inductor than its inductance! Figure 3.49 shows an equivalent circuit of a real inductor and also some of the many considerations that go into the selection process. To further complicate the issue, most of these parameters interact, thereby making the design of an inductor truly more of an art than a science.

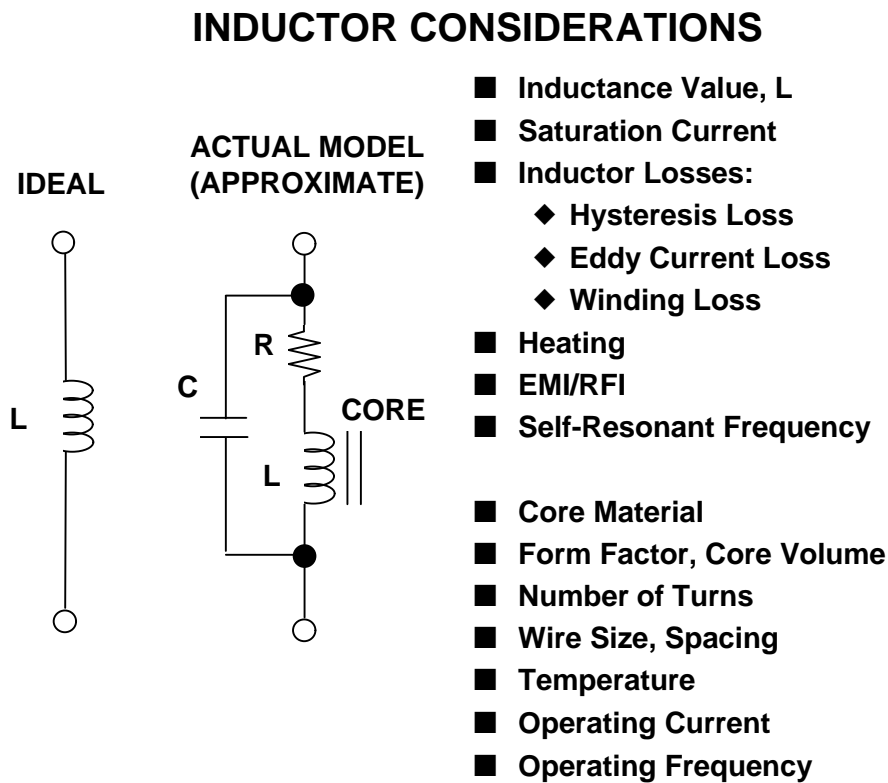


Figure 3.49



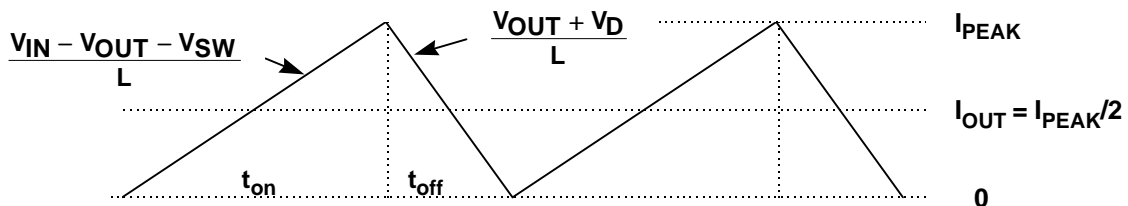
Probably the easiest inductor problem to solve is selecting the proper value. In most switching regulator applications, the exact value is not very critical, so approximations can be used with a high degree of confidence.

The heart of a switching regulator analysis involves a thorough understanding of the inductor current waveform. Figure 3.50 shows an assumed inductor current waveform (which is also the output current) for a buck converter, such as the ADP3000, which uses the gated-oscillator PBM switch modulation technique. Note that this waveform represents a worst case condition from the standpoint of storing energy in the inductor, where the inductor current starts from zero on each cycle. In high output current applications, the inductor current does not return to zero, but ramps up until the output voltage comparator senses that the oscillator should be turned off, at which time the current ramps down until the comparator turns the oscillator on again. This assumption about the worst case waveform is necessary because in a simple PBM regulator, the oscillator duty cycle remains constant regardless of input voltage or output load current. Selecting the inductor value using this assumption will always ensure that there is enough energy stored in the inductor to maintain regulation.

*It should be emphasized that the following inductance calculations for the PBM buck and boost regulators should be used only as a starting point, and larger or smaller values may actually be required depending on the specific regulator and the input/output conditions.*

**CALCULATING L FOR BUCK CONVERTER:  
GATED OSCILLATOR (PBM) TYPE**

**OUTPUT AND INDUCTOR CURRENT:**



$$I_{PEAK} = \left( \frac{V_{IN} - V_{OUT} - V_{SW}}{L} \right) t_{on}$$

$$L = \left( \frac{V_{IN} - V_{OUT} - V_{SW}}{2 I_{OUT}} \right) t_{on}$$

USE MINIMUM  $V_{IN}$

Figure 3.50

## SWITCHING REGULATORS

The peak current is easily calculated from the slope of the positive-going portion of the ramp:

$$I_{PEAK} = \left( \frac{V_{IN} - V_{OUT} - V_{SW}}{L} \right) t_{on}.$$

This equation can then be solved for L:

$$L = \left( \frac{V_{IN} - V_{OUT} - V_{SW}}{I_{PEAK}} \right) t_{on}.$$

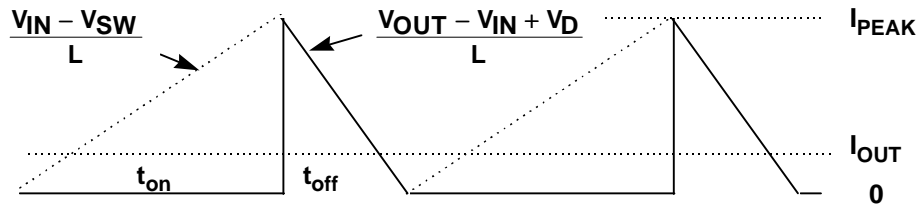
However, the average output current,  $I_{OUT}$  is equal to  $I_{PEAK}/2$ , and therefore  $I_{PEAK}=2I_{OUT}$ . Substituting this value for  $I_{PEAK}$  into the previous equation yields:

$$L = \left( \frac{V_{IN} - V_{OUT} - V_{SW}}{2I_{OUT}} \right) t_{on}. \quad [L \text{ for buck PBM Converter}]$$

The minimum expected value of  $V_{IN}$  should be used in order to minimize the inductor value and maximize its stored energy. If  $V_{IN}$  is expected to vary widely, an external resistor can be added to the ADP3000 to limit peak current and prevent inductor saturation at maximum  $V_{IN}$ .

A similar analysis can be carried out for a boost PBM regulator as shown in Figure 3.51.

### CALCULATING L FOR BOOST CONVERTER: GATED OSCILLATOR (PBM) TYPE OUTPUT CURRENT:



$$DC = \frac{1}{t_{on} + t_{off}}$$

$$I_{OUT} = \frac{I_{PEAK}}{2} (1 - D) \quad I_{PEAK} = \frac{2 I_{OUT}}{1 - D}$$

$$I_{PEAK} = \left( \frac{V_{IN} - V_{SW}}{L} \right) t_{on}$$

$$L = \left( \frac{V_{IN} - V_{SW}}{I_{PEAK}} \right) t_{on} = \left( \frac{V_{IN} - V_{SW}}{2 I_{OUT}} \right) (1 - D) t_{on}$$

USE MINIMUM  $V_{IN}$

Figure 3.51

We make the same assumptions about the inductor current, but note that the output current shown on the diagram is pulsating and not continuous. The output current,  $I_{OUT}$ , can be expressed in terms of the peak current,  $I_{PEAK}$ , and the duty cycle,  $D$ , as:

$$I_{OUT} = \frac{I_{PEAK}}{2}(1 - D).$$

Solving for  $I_{PEAK}$  yields:

$$I_{PEAK} = \frac{2I_{OUT}}{1 - D}.$$

However,  $I_{PEAK}$  can also be expressed in terms of  $V_{IN}$ ,  $V_{SW}$ ,  $L$ , and  $t_{on}$ :

$$I_{PEAK} = \left( \frac{V_{IN} - V_{SW}}{L} \right) t_{on}, \text{ which can be solved for } L:$$

$$L = \left( \frac{V_{IN} - V_{SW}}{I_{PEAK}} \right) t_{on}.$$

Substituting the previous expression for  $I_{PEAK}$  yields:

$$L = \left( \frac{V_{IN} - V_{SW}}{2I_{OUT}} \right) (1 - D)t_{on}. \quad [L \text{ for boost PBM Converter}]$$

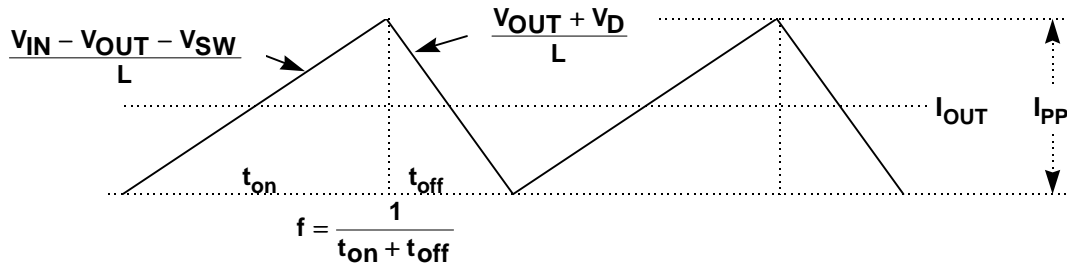
The minimum expected value of  $V_{IN}$  should be used in order to ensure sufficient inductor energy storage under all conditions. If  $V_{IN}$  is expected to vary widely, an external resistor can be added to the ADP3000 to limit peak current and prevent inductor saturation at maximum  $V_{IN}$ .

*The above equations will only yield approximations to the proper inductor value for the PBM-type regulators and should be used only as a starting point. An exact analysis is difficult and highly dependent on the regulator and input/output conditions. However, there is considerable latitude with this type of regulator, and other analyses may yield different results but still fall within the allowable range for proper regulator operation.*

Calculating the proper inductor value for PWM regulators is more straightforward. Figure 3.52 shows the output and inductor current waveform for a buck PWM regulator operating in the continuous mode. It is accepted design practice to design for a peak-to-peak ripple current,  $I_{pp}$ , which is between 10% and 30% of the output current,  $I_{OUT}$ . We will assume that  $I_{pp} = 0.2I_{OUT}$ .

**CALCULATING L FOR BUCK CONVERTER:  
CONSTANT FREQUENCY PWM TYPE**

**OUTPUT AND INDUCTOR CURRENT, CONTINUOUS MODE:**



$$L = \left( \frac{1}{f} \right) \left( \frac{V_{IN} - V_{OUT} - V_{SW}}{V_{IN} - V_{SW} + V_D} \right) \left( \frac{V_{OUT} + V_D}{I_{PP}} \right)$$

**NOMINALLY, MAKE  $I_{PP} = 0.2 I_{OUT}$**

**Figure 3.52**

By inspection, we can write:

$$\left( \frac{V_{IN} - V_{OUT} - V_{SW}}{L} \right) t_{on} = \left( \frac{V_{OUT} + V_D}{L} \right) t_{off}, \text{ or}$$

$$t_{off} = \left( \frac{V_{IN} - V_{OUT} - V_{SW}}{V_{OUT} + V_D} \right) t_{on}.$$

However, the switching frequency,  $f$ , is given by

$$f = \frac{1}{t_{on} + t_{off}}, \text{ or } t_{off} = \frac{1}{f} - t_{on}.$$

Substituting this expression for  $t_{off}$  in the previous equation for  $t_{off}$  and solving for  $t_{on}$  yields:

$$t_{on} = \frac{1}{f} \left( \frac{V_{OUT} + V_D}{V_{IN} - V_{SW} + V_D} \right).$$

However,

$$I_{PP} = \left( \frac{V_{IN} - V_{OUT} - V_{SW}}{L} \right) t_{on}.$$

Combining the last two equations and solving for L yields:

$$L = \left( \frac{1}{f} \right) \left( \frac{V_{IN} - V_{OUT} - V_{SW}}{V_{IN} - V_{SW} + V_D} \right) \left( \frac{V_{OUT} + V_D}{I_{pp}} \right). \quad \text{[L for buck PWM converter, constant frequency]}$$

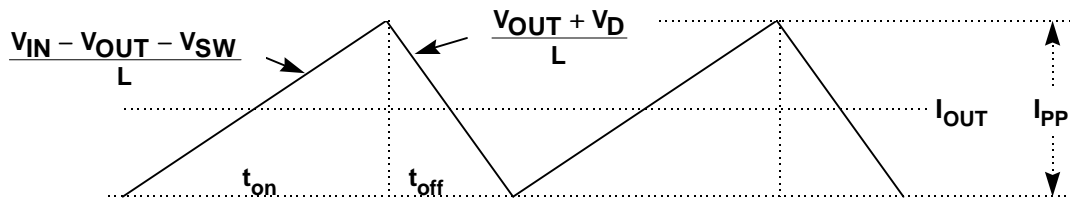
As indicated earlier, choose  $I_{pp}$  to be nominally  $0.2I_{OUT}$  and solve the equation for L. Calculate L for the minimum and maximum expected value of  $V_{IN}$  and choose a value halfway between. System requirements may dictate a larger or smaller value of  $I_{pp}$ , which will inversely affect the inductor value.

A variation of the buck PWM constant frequency regulator is the buck PWM regulator with variable frequency and constant off-time (e.g., ADP1147, ADP1148).

A diagram of the output and inductor current waveform is shown in Figure 3.53 for the continuous mode.

**CALCULATING L FOR BUCK CONVERTER:  
CONSTANT OFF-TIME, VARIABLE FREQUENCY PWM TYPE**

**OUTPUT AND INDUCTOR CURRENT, CONTINUOUS MODE:**



$$\left( \frac{V_{OUT} + V_D}{L} \right) t_{off} = I_{PP}$$

$$L = \left( \frac{V_{OUT} + V_D}{I_{PP}} \right) t_{off}$$

**NOMINALLY, MAKE  $I_{PP} = 0.2 I_{OUT}$**

**Figure 3.53**

The calculations are very straightforward, since the peak-to-peak amplitude of the ripple current is constant:

$$I_{pp} = \left( \frac{V_{OUT} + V_D}{L} \right) t_{off}.$$

## SWITCHING REGULATORS

Solving for L:

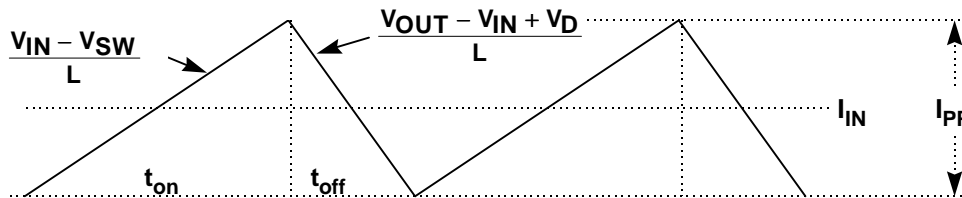
$$L = \left( \frac{V_{OUT} + V_D}{I_{PP}} \right) t_{off} \quad \text{[L for buck PWM constant off-time, variable frequency converter]}$$

Again, choose  $I_{PP} = 0.2I_{OUT}$ , or whatever the system requires.

The final example showing the inductance calculation is for the boost PWM constant frequency regulator. The inductor (and input) current waveform is shown in Figure 3.54.

### **CALCULATING L FOR BOOST CONVERTER: CONSTANT FREQUENCY PWM TYPE**

**INPUT AND INDUCTOR CURRENT, CONTINUOUS MODE:**



$$f = \frac{1}{t_{on} + t_{off}}$$

$$L = \left( \frac{1}{f} \right) \left( \frac{V_{OUT} - V_{IN} + V_D}{V_{OUT} - V_{SW} + V_D} \right) \left( \frac{V_{IN} - V_{SW}}{I_{PP}} \right)$$

$$I_{IN} = \left( \frac{V_{OUT}}{V_{IN}} \right) I_{OUT}$$

**NOMINALLY, MAKE  $I_{PP} = 0.2 I_{IN}$**

**Figure 3.54**

The analysis is similar to that of the constant frequency buck PWM regulator.

By inspection of the inductor current, we can write:

$$\left( \frac{V_{IN} - V_{SW}}{L} \right) t_{on} = \left( \frac{V_{OUT} - V_{IN} + V_D}{L} \right) t_{off}, \text{ or}$$

$$t_{off} = \left( \frac{V_{IN} - V_{SW}}{V_{OUT} - V_{IN} + V_D} \right) t_{on}.$$

However, the switching frequency,  $f$ , is given by

$$f = \frac{1}{t_{\text{on}} + t_{\text{off}}}, \text{ or } t_{\text{off}} = \frac{1}{f} - t_{\text{on}}.$$

Substituting this expression for  $t_{\text{off}}$  in the previous equation for  $t_{\text{off}}$  and solving for  $t_{\text{on}}$  yields:

$$t_{\text{on}} = \frac{1}{f} \left( \frac{V_{\text{OUT}} - V_{\text{IN}} + V_{\text{D}}}{V_{\text{OUT}} - V_{\text{SW}} + V_{\text{D}}} \right).$$

However,

$$I_{\text{pp}} = \left( \frac{V_{\text{IN}} - V_{\text{SW}}}{L} \right) t_{\text{on}}.$$

Combining the last two equations and solving for  $L$  yields:

$$L = \left( \frac{1}{f} \right) \left( \frac{V_{\text{OUT}} - V_{\text{IN}} + V_{\text{D}}}{V_{\text{OUT}} - V_{\text{SW}} + V_{\text{D}}} \right) \left( \frac{V_{\text{IN}} - V_{\text{SW}}}{I_{\text{pp}}} \right). \quad [\text{L for boost PWM, constant frequency converter}]$$

For the boost converter, the inductor (input) current,  $I_{\text{IN}}$ , can be related to the output current,  $I_{\text{OUT}}$ , by:

$$I_{\text{IN}} = \left( \frac{V_{\text{OUT}}}{V_{\text{IN}}} \right) I_{\text{OUT}}.$$

Nominally, make  $I_{\text{pp}} = 0.2I_{\text{IN}}$ .

Note that for the boost PWM, even though the input current is continuous, while the output current pulsates, we still base the inductance calculation on the peak-to-peak inductor ripple current.

As was previously suggested, the actual selection of the inductor value in a switching regulator is probably the easiest part of the design process. Choosing the proper type of inductor is much more complicated as the following discussions will indicate.

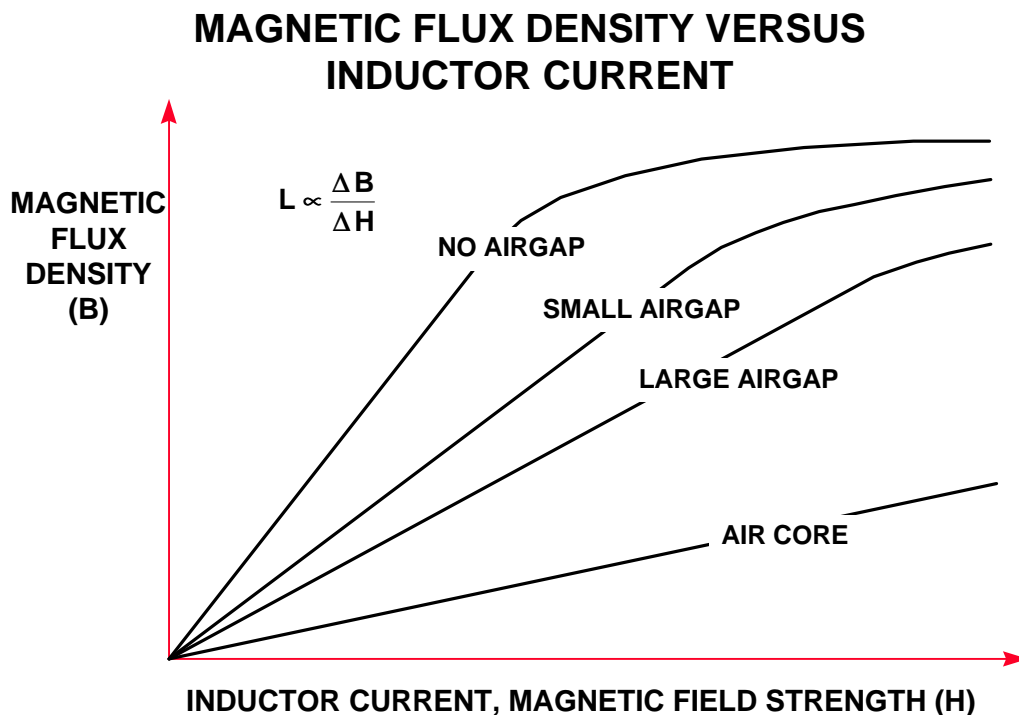
Fundamental magnetic theory says that if a current passes through a wire, a magnetic field will be generated around the wire (right-hand rule). The strength of this field is measured in ampere-turns per meter, or *oersteds* and is proportional to the current flowing in the wire. The magnetic field strength produces a *magnetic flux density* ( $B$ , measured in webers per square meter, or *gauss*).

## SWITCHING REGULATORS

Using a number of turns of wire to form a coil increases the magnetic flux density for a given current. The effective inductance of the coil is proportional to the ratio of the magnetic flux density to the field strength.

This simple air core inductor is not very practical for the values of inductance required in switching regulators because of wiring resistance, interwinding capacitance, sheer physical size, and other factors. Therefore, in order to make a reasonable inductor, the wire is wound around some type of ferromagnetic core having a high *permeability*. Core permeability is often specified as a relative permeability which is basically the increase in inductance which is obtained when the inductor is wound on a core instead of just air. A relative permeability of 1000, for instance, will increase inductance by 1000:1 above that of an equivalent air core.

Figure 3.55 shows magnetic flux density (B) versus inductor current for the air core and also ferromagnetic cores. Note that B is linear with respect to H for the air core inductor, i.e., the inductance remains constant regardless of current.



**Figure 3.55**

The addition of a ferromagnetic core increases the slope of the curve and increases the effective inductance, but at some current level, the inductor core will saturate (i.e., the inductance is drastically reduced). It is obvious that inductor saturation can wreck havoc in a switching regulator, and can even burn out the switch if it is not current-limited.

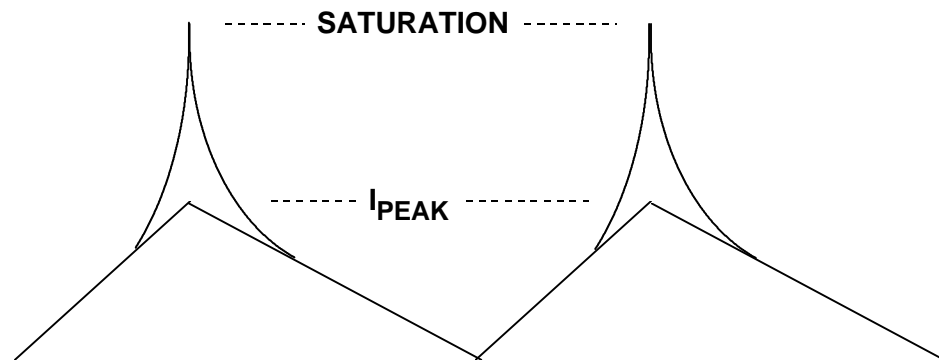
This effect can be reduced somewhat while still maintaining higher inductance than an air core by the addition of an air gap in the ferromagnetic core. The air gap reduces the slope of the curve, but provides a wider linear operating range of inductor current. Air gaps do have their problems, however, and one of them is the tendency of the air-gapped inductor to radiate high frequency energy more than a



non-gapped inductor. Proper design and manufacturing techniques, however, can be used to minimize this EMI problem, so air-gapped cores are popular in many applications.

The effects of inductor core saturation in a switcher can be disastrous to the switching elements as well as lowering efficiency and increasing noise. Figure 3.56 shows a normal inductor current waveform in a switching regulator as well as a superimposed waveform showing the effects of core saturation. Under normal conditions the slope is linear for both the charge and discharge cycle. If saturation occurs, however, the inductor current increases exponentially, corresponding to the drop in effective inductance. It is therefore important in all switching regulator designs to determine the peak inductor current expected under the worst case conditions of input voltage, load current, duty cycle, etc. This worst case peak current must be less than the peak-current rating of the inductor. Notice that when inductor literature does not have a “DC-current” rating, or shows only an “AC amps” rating, such inductors are often prone to saturation.

### EFFECTS OF SATURATION ON INDUCTOR CURRENT



**Figure 3.56**

From a simplified design standpoint, the effects or presence of inductor saturation can best be observed with a scope and a current probe. If a current probe is not available, a less direct but still effective method is to measure the voltage across a small sense resistor in series with the inductor. The resistor value should be 1Ω or less (depending on the inductor current), and the resistor must be sized to dissipate the power. In most cases, a 1Ω, 1W resistor will work for currents up to a few hundred mA, and a 0.1Ω, 10W resistor is good for currents up to 10A.

Another inductor consideration is its loss. Ideally, an inductor should dissipate no power. However, in a practical inductor, power is dissipated in the form of hysteresis loss, eddy-current loss, and winding loss. Figure 3.57 shows a typical B/H curve for an inductor. The enclosed area swept out by the B/H curve during one complete operating cycle is the hysteresis loss exhibited by the core during that cycle. Hysteresis loss is a function of core material, core volume, operating frequency, and the maximum flux density during each cycle. The second major loss within the core is eddy-current loss. This loss is caused by the flow of circulating magnetic currents

## SWITCHING REGULATORS

within the core material caused by rapid transitions in the magnetic flux density. It is also dependent on the core material, core volume, operating frequency, and flux density.

In addition to core loss, there is winding loss, the power dissipated in the DC resistance of the winding. This loss is a function of the wire size, core volume, and the number of turns.

In a switching regulator application, excessive loss will result in a loss of efficiency and high inductor operating temperatures.

### INDUCTOR POWER LOSSES

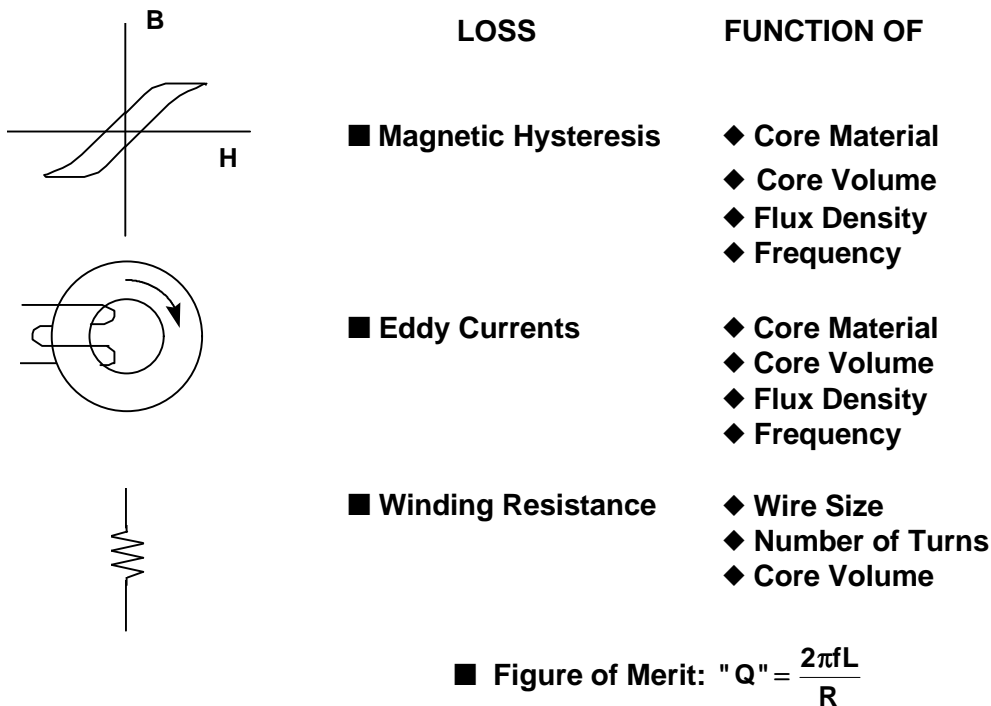


Figure 3.57

Fortunately, inductor manufacturers have simplified the design process by specifying maximum peak current, maximum continuous current, and operating frequency range and temperature for their inductors. If the designer derates the maximum peak and continuous current levels by a factor of 20% or so, the inductor should be satisfactory for the application. If these simple guidelines are observed, then the designer can be reasonable confident that the major sources of efficiency losses will be due to other parts of the regulator, i.e., the switch ( $I^2R$ , gate charge, on-voltage), the diode (on-voltage), and the quiescent power dissipation of the regulator itself.

One method to ensure that the inductor losses do not significantly degrade the regulator performance is to measure the Q of the inductor at the switching frequency. If the Q is greater than about 25, then the losses should be insignificant.

There are many possible choices in inductor core materials: ferrite, molypermalloy (MPP) ferrite, powdered iron, etc. High efficiency converters generally cannot accommodate the core loss found in the low-cost powdered iron cores, forcing the use of more expensive ferrite, molypermalloy (MPP), or “Kool M $\mu$ ”<sup>®</sup> cores.

Ferrite core material saturates “hard”, which causes the inductance to collapse abruptly when the peak current is exceeded. This results in a sharp increase in inductor ripple current.

Molypermalloy from Magnetics, Inc., is a very good, low loss core material for toroids, but is more expensive than ferrite. A reasonable compromise from the same manufacturer is “Kool M $\mu$ ”.

The final consideration is the inductor self-resonant frequency. A practical example would be an inductor of 10 $\mu$ H which has an equivalent distributed capacitance of 5pF. The self-resonant frequency can be calculated as follows:

$$f_{\text{resonance}} = \frac{1}{2\pi} \sqrt{\frac{1}{LC}} = 22\text{MHz.}$$

The switching frequency of the regulator should be at least ten times less than the resonant frequency. In most practical designs with switching frequencies less than 1MHz this will always be the case, but a quick calculation is a good idea.

## **CAPACITOR CONSIDERATIONS**

Capacitors play a critical role in switching regulators by acting as storage elements for the pulsating currents produced by the switching action. Although not shown on the diagrams previously, all switching regulators need capacitors on their inputs as well as their outputs for proper operation. The capacitors must have very low impedance at the switching frequency as well as the high frequencies produced by the pulsating current waveforms.

Recall the input and output current waveforms for the simple buck converter shown in Figure 3.58. Note that the input current to the buck converter is pulsating, while the output is continuous. Obviously, the input capacitor  $C_{IN}$  is critical for proper operation of the regulator. It must maintain the input at a constant voltage during the switching spikes. This says that the impedance of the capacitor must be very low at high frequencies, much above the regulator switching frequency. The load capacitor is also critical in that its impedance will determine the peak-to-peak output voltage ripple, but its impedance at high frequencies is not as critical due to the continuous nature of the output current waveform.

The situation is reversed in the case of the boost converter shown in Figure 3.59. Here the input waveform is continuous, while the output waveform is pulsating. The output capacitor must have good low and high frequency characteristics in order to minimize the output voltage ripple. Boost converters are often followed by a post filter to remove the high frequency switching noise.

**BUCK CONVERTER  
INPUT AND OUTPUT CURRENT WAVEFORMS**

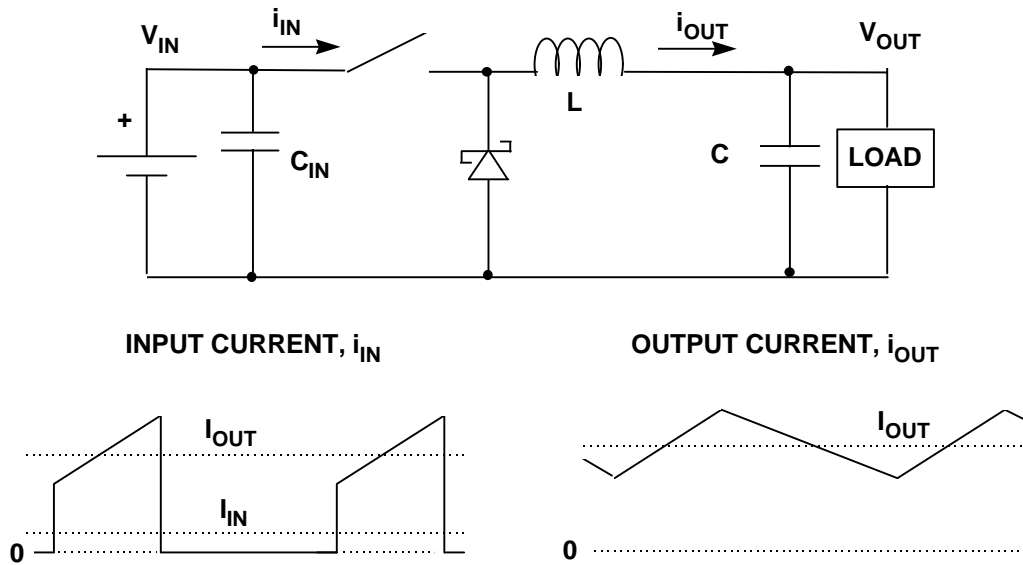


Figure 3.58

**BOOST CONVERTER  
INPUT AND OUTPUT CURRENT WAVEFORMS**

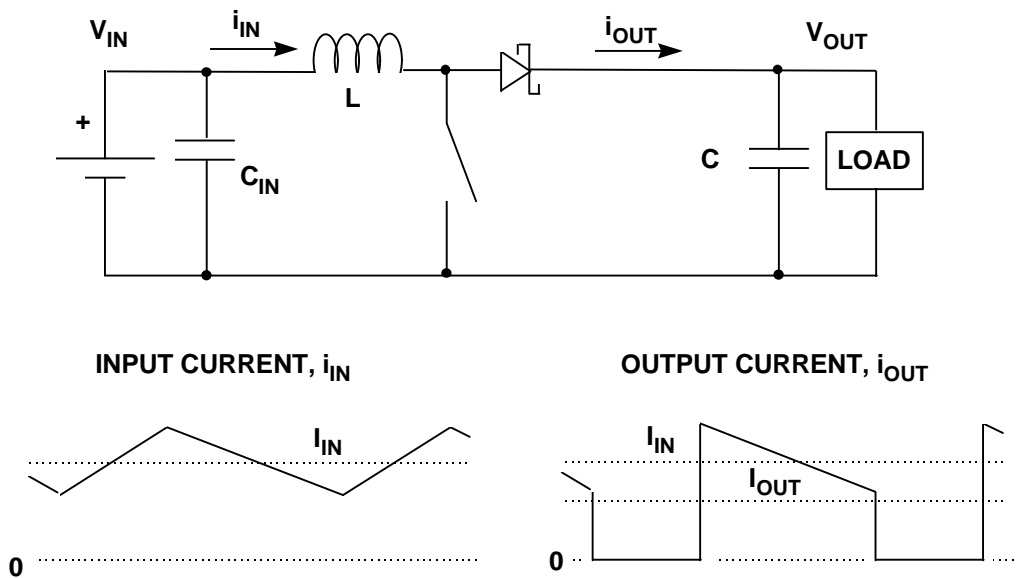
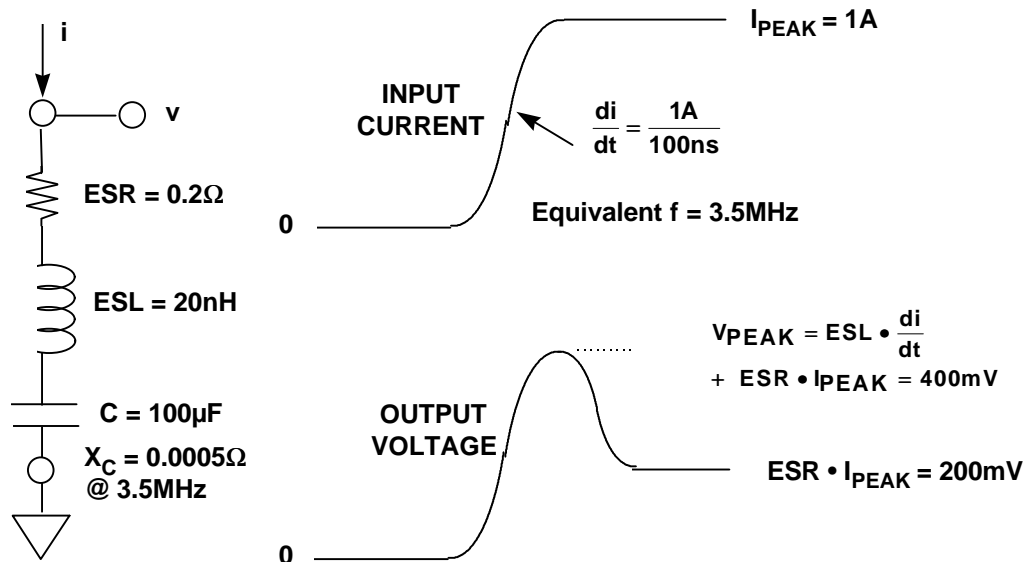


Figure 3.59

Switching regulator capacitors are generally of the electrolytic type because of the relatively large values required. An equivalent circuit for an electrolytic capacitor is shown in Figure 3.60. In addition to the capacitance value itself, the capacitor has some equivalent series resistance (ESR) and equivalent series inductance (ESL). It is useful to make a few assumptions and examine the approximate response of the capacitor to a fast current step input. For the sake of the discussion, assume the input current switches from 0 to 1A in 100ns. Also, assume that the ESR is 0.2Ω and that the ESL is 20nH. ESR and ESL vary widely between manufacturers and are also dependent upon body style (through-hole vs. surface mount), but these values will serve to illustrate the point.

**RESPONSE OF CAPACITOR TO CURRENT STEP**



**Figure 3.60**

Assume that the actual value of the capacitor is large enough so that its reactance is essentially a short circuit with respect to the step function input. For example, 100μF at 3.5MHz (the equivalent frequency of a 100ns risetime pulse) has a reactance of  $1/2\pi fC = 0.0005\Omega$ . In this case, the output voltage ripple is determined exclusively by the ESR and ESL of the capacitor, not the actual capacitor value itself.

These waveforms show the inherent limitations of electrolytic capacitors used to absorb high frequency switching pulses. In a practical system, the high frequency components must be attenuated by low-inductance ceramic capacitors with low ESL or by the addition of an LC filter.

Figure 3.61 shows the impedance versus frequency for a typical 100μF electrolytic capacitor having an ESR of 0.2Ω and an ESL of 20nH. At frequencies below about 10kHz, the capacitor is nearly ideal. Between 10kHz and 1MHz (the range of switching frequencies for most IC switching regulators!) the impedance is limited by the ESR to 0.2Ω. Above about 1MHz the capacitor behaves like an inductor due to the ESL of 20nH. These values, although they may vary somewhat depending upon the actual type of electrolytic capacitor (aluminum general purpose, aluminum

## SWITCHING REGULATORS

switching type, tantalum, or organic semiconductor), are representative and illustrate the importance of understanding the limitations of capacitors in switching regulators.

### TYPICAL ELECTROLYTIC CAPACITOR IMPEDANCE VERSUS FREQUENCY

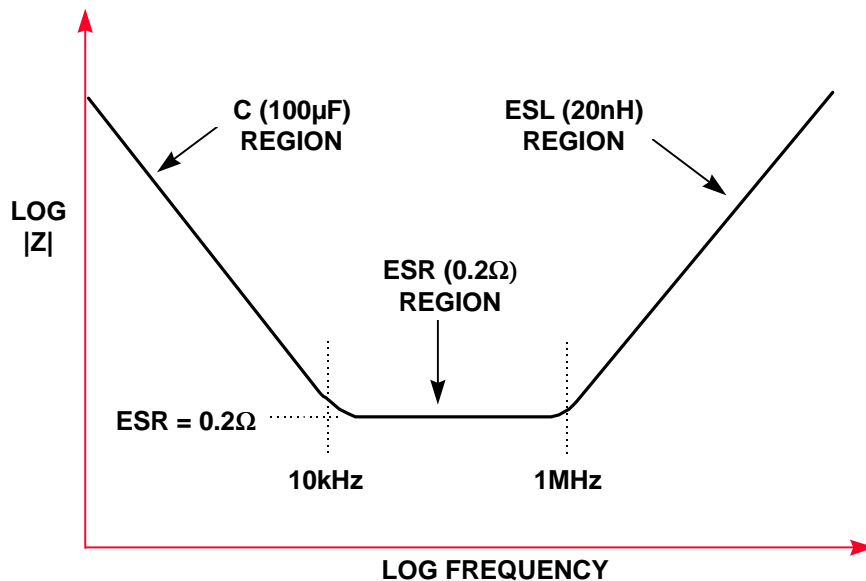


Figure 3.61

From the electrolytic capacitor impedance characteristic, it is clear that the ESR and ESL of the output capacitor will determine the peak-to-peak output voltage ripple caused by the switching regulator output ripple current.

In most electrolytic capacitors, ESR degrades noticeably at low temperature, by as much as a factor of 4-6 times at  $-55^{\circ}\text{C}$  vs. the room temperature value. For circuits where ESR is critical to performance, this can lead to problems. Some specific electrolytic types do address this problem, for example within the HFQ switching types, the  $-10^{\circ}\text{C}$  ESR at 100kHz is no more than  $2\times$  that at room temperature. The OS-CON electrolytics have a ESR vs. temperature characteristic which is relatively flat.

There are generally three classes of capacitors useful in 10kHz-100MHz frequency range, broadly distinguished as the generic dielectric types; *electrolytic*, *film*, and *ceramic*. These can in turn be further sub-divided. A thumbnail sketch of capacitor characteristics is shown in the chart of Figure 3.62.

## CAPACITOR SELECTION

	Aluminum Electrolytic (General Purpose)	Aluminum Electrolytic (Switching Type)	Tantalum Electrolytic	OS-CON Electrolytic	Polyester (Stacked Film)	Ceramic (Multilayer)
Size	100 $\mu$ F	120 $\mu$ F	120 $\mu$ F	100 $\mu$ F	1 $\mu$ F	0.1 $\mu$ F
Rated Voltage	25 V	25 V	20 V	20 V	400 V	50 V
ESR	0.6 $\Omega$ @ 100 kHz	0.18 $\Omega$ @ 100 kHz	0.12 $\Omega$ @ 100 kHz	0.02 $\Omega$ @ 100 kHz	0.11 $\Omega$ @ 1 MHz	0.12 $\Omega$ @ 1 MHz
Operating Frequency (*)	$\cong$ 100 kHz	$\cong$ 500 kHz	$\cong$ 1 MHz	$\cong$ 1 MHz	$\cong$ 10 MHz	$\cong$ 1 GHz

(\*) Upper frequency strongly size and package dependent

Figure 3.62

The *electrolytic* family provides an excellent, cost-effective low-frequency component, because of the wide range of values, a high capacitance-to-volume ratio, and a broad range of working voltages. It includes *general purpose aluminum electrolytic* types, available in working voltages from below 10V up to about 500V, and in size from 1 to several thousand  $\mu$ F (with proportional case sizes). All electrolytic capacitors are polarized, and thus cannot withstand more than a volt or so of reverse bias without damage. They also have relatively high leakage currents (up to tens of  $\mu$ A, and strongly dependent upon design specifics).

A subset of the general electrolytic family includes *tantalum* types, generally limited to voltages of 100V or less, with capacitance of 500 $\mu$ F or less [Reference 8]. In a given size, tantalums exhibit a higher capacitance-to-volume ratio than do general purpose electrolytics, and have both a higher frequency range and lower ESR. They are generally more expensive than standard electrolytics, and must be carefully applied with respect to surge and ripple currents.

A subset of aluminum electrolytic capacitors is the *switching* type, designed for handling high pulse currents at frequencies up to several hundred kHz with low losses [Reference 9]. This capacitor type competes directly with tantalums in high frequency filtering applications, with the advantage of a broader range of values.

A more specialized high performance aluminum electrolytic capacitor type uses an organic semiconductor electrolyte [Reference 10]. The *OS-CON* capacitors feature appreciably lower ESR and higher frequency range than do other electrolytic types, with an additional feature of low low-temperature ESR degradation.

## **SWITCHING REGULATORS**

*Film* capacitors are available in very broad value ranges and different dielectrics, including polyester, polycarbonate, polypropylene, and polystyrene. Because of the low dielectric constant of these films, their volumetric efficiency is quite low, and a 10 $\mu$ F/50V polyester capacitor (for example) is actually a handful. Metalized (as opposed to foil) electrodes does help to reduce size, but even the highest dielectric constant units among film types (polyester, polycarbonate) are still larger than any electrolytic, even using the thinnest films with the lowest voltage ratings (50V). Where film types excel is in their low dielectric losses, a factor which may not necessarily be a practical advantage for filtering switchers. For example, ESR in film capacitors can be as low as 10m $\Omega$  or less, and the behavior of films generally is very high in terms of Q. In fact, this can cause problems of spurious resonance in filters, requiring damping components.

Typically using a wound layer-type construction, film capacitors can be inductive, which can limit their effectiveness for high frequency filtering. Obviously, only non-inductively made film caps are useful for switching regulator filters. One specific style which is non-inductive is the *stacked-film* type, where the capacitor plates are cut as small overlapping linear sheet sections from a much larger wound drum of dielectric/plate material. This technique offers the low inductance attractiveness of a plate sheet style capacitor with conventional leads [see References 9, 10, 11]. Obviously, minimal lead length should be used for best high frequency effectiveness. Very high current polycarbonate film types are also available, specifically designed for switching power supplies, with a variety of low inductance terminations to minimize ESL [Reference 12].

Dependent upon their electrical and physical size, film capacitors can be useful at frequencies to well above 10MHz. At the highest frequencies, only stacked film types should be considered. Some manufacturers are now supplying film types in leadless surface mount packages, which eliminates the lead length inductance.

*Ceramic* is often the capacitor material of choice above a few MHz, due to its compact size, low loss, and availability up to several  $\mu$ F in the high-K dielectric formulations (X7R and Z5U), at voltage ratings up to 200V [see ceramic families of Reference 8]. NP0 (also called COG) types use a lower dielectric constant formulation, and have nominally zero TC, plus a low voltage coefficient (unlike the less stable high-K types). NP0 types are limited to values of 0.1 $\mu$ F or less, with 0.01 $\mu$ F representing a more practical upper limit.

Multilayer ceramic “chip caps” are very popular for bypassing/ filtering at 10MHz or more, simply because their very low inductance design allows near optimum RF bypassing. For smaller values, ceramic chip caps have an operating frequency range to 1GHz. For high frequency applications, a useful selection can be ensured by selecting a value which has a self-resonant frequency *above* the highest frequency of interest.

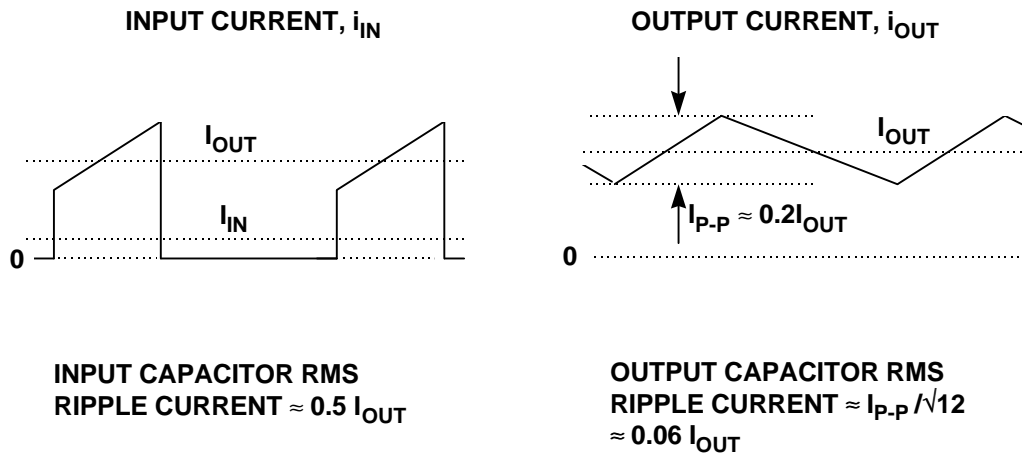
The *ripple-current* rating of electrolytic capacitors must not be ignored in switching regulator applications because, unlike linear regulators, switching regulators subject capacitors to large AC currents. AC currents can cause heating in the dielectric material and change the temperature-dependent characteristics of the capacitor. Also, the capacitor is more likely to fail at the higher temperatures produced by the



ripple current. Fortunately, most manufacturers provide ripple-current ratings, and this problem can be averted if understood.

Calculating the exact ripple current can be tedious, especially with complex switching regulator waveforms. Simple approximations can be made, however, which are sufficiently accurate. Consider first the buck converter input and output currents (refer to Figure 3.63). The rms input capacitor ripple current can be approximated by a square wave having a peak-to-peak amplitude equal to  $I_{OUT}$ . The rms value of this square wave is therefore  $I_{OUT}/2$ . The output capacitor current waveform can be approximated by a sawtooth waveform having a peak-to-peak amplitude of  $0.2I_{OUT}$ . The rms value of this sawtooth is therefore approximately  $0.2I_{OUT}/\sqrt{12}$ , or  $0.06I_{OUT}$ .

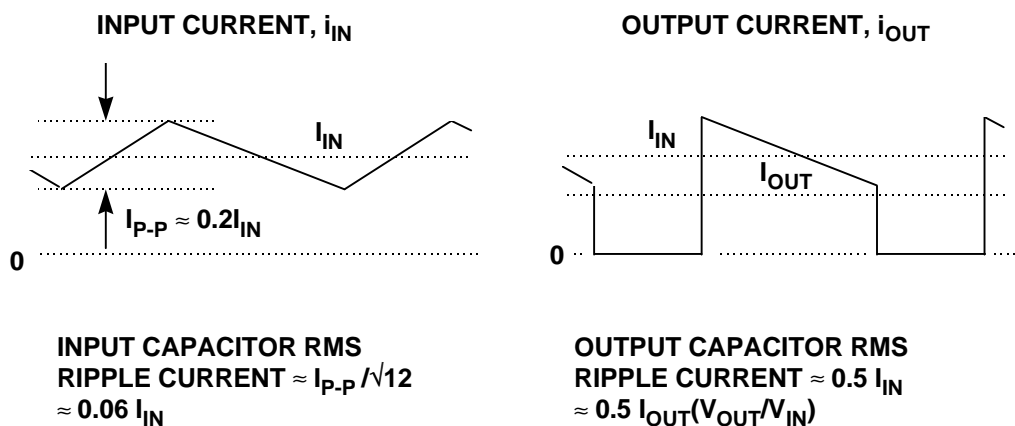
**BUCK CONVERTER INPUT AND OUTPUT CAPACITOR RMS RIPPLE CURRENT APPROXIMATIONS**



**Figure 3.63**

Similarly for a boost converter (see waveforms shown in Figure 3.64), the input capacitor rms ripple current is  $0.06I_{IN}$ , and the rms output current ripple is  $0.5I_{IN}$ . These boost converter expressions can also be expressed in terms of the output current,  $I_{OUT}$ , using the relationship,  $I_{IN} = I_{OUT}(V_{OUT}/V_{IN})$ . In any case, the minimum expected value of input voltage should be used which will result in the largest value of input current.

In practice, a safety factor of 25% should be added to the above approximations for further derating. In practical applications, especially those using surface mount components, it may be impossible to meet the capacitance value, ESR, and ripple current requirement using a single capacitor. Paralleling a number of equal value capacitors is a viable option which will increase the effective capacitance and reduce ESR, ESL. In addition, the ripple current is divided between the individual capacitors.

**BOOST CONVERTER INPUT AND OUTPUT CAPACITOR  
RMS RIPPLE CURRENT APPROXIMATIONS****Figure 3.64**

Several electrolytic capacitor manufacturers offer low ESR surface mount devices including the AVX TPS-series [Reference 14], and the Sprague 595D-series [Reference 15]. Low ESR through-hole electrolytic capacitors are the HFQ-series from Panasonic [Reference 16] and the OS-CON-series from Sanyo [Reference 17].

**SWITCHING REGULATOR OUTPUT FILTERING**

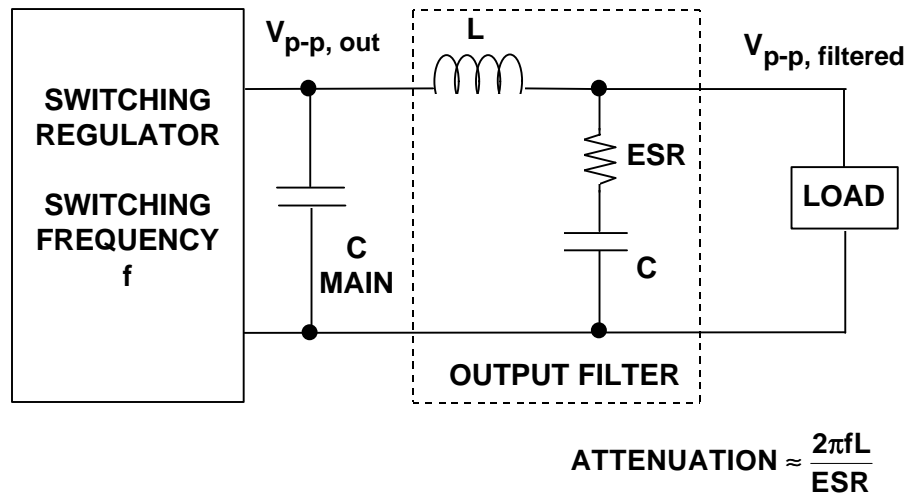
In order to minimize switching regulator output voltage ripple it is often necessary to add additional filtering. In many cases, this is more efficient than simply adding parallel capacitors to the main output capacitor to reduce ESR.

Output ripple current in a boost converter is pulsating, while that of a buck converter is a sawtooth. In any event, the high frequency components in the output ripple current can be removed with a small inductor (2 to 10 $\mu$ H or so followed by a low ESR capacitor). Figure 3.65 shows a simple LC filter on the output of a switching regulator whose switching frequency is  $f$ . Generally the actual value of the filter capacitor is not as important as its ESR when filtering the switching frequency ripple. For instance, the reactance of a 100 $\mu$ F capacitor at 100kHz is approximately 0.016 $\Omega$ , which is much less than available ESRs.

The capacitor ESR and the inductor reactance attenuate the ripple voltage by a factor of approximately  $2\pi fL/ESR$ . The example shown in Figure 3.65 uses a  $10\mu\text{H}$  inductor and a capacitor with an ESR of  $0.2\Omega$ . This combination attenuates the output ripple by a factor of about 32.

The inductor core material is not critical, but it should be rated to handle the load current. Also, its DC resistance should be low enough so that the load current does not cause a significant voltage drop across it.

**SWITCHING REGULATOR OUTPUT FILTERING**



**Example:**       $ESR = 0.2\Omega$  ,  $L = 10\mu\text{H}$  ,  $f = 100\text{kHz}$   
**ATTENUATION = 32**

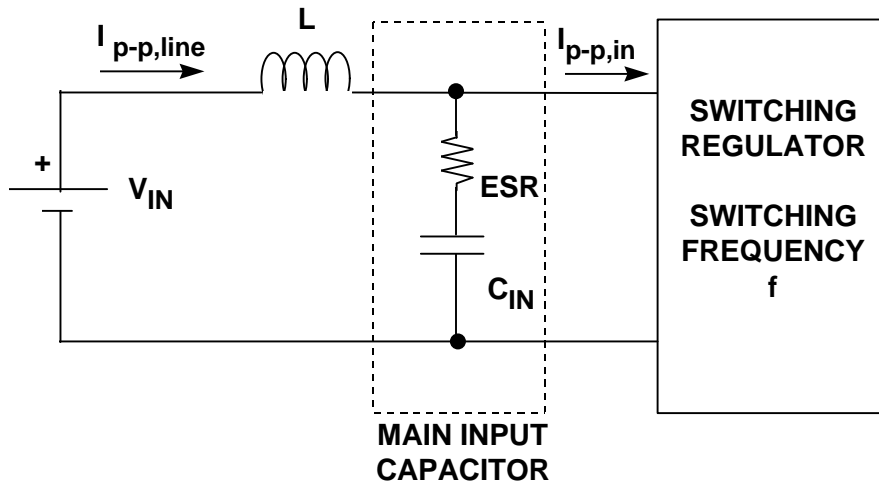
**Figure 3.65**

**SWITCHING REGULATOR INPUT FILTERING**

The input ripple current in a buck converter is pulsating, while that of a boost converter is a sawtooth. Additional filtering may be required to prevent the switching frequency and the other higher frequency components from affecting the main supply ripple current.

This is easily accomplished by the addition of a small inductor in series with the main input capacitor of the regulator as shown in Figure 3.66. The reactance of the inductor at the switching frequency forms a divider with the ESR of the input capacitor. The inductor will block both low and high frequency components from the main input voltage source. The attenuation of the ripple current at the switching frequency,  $f$ , is approximately  $2\pi fL/ESR$ .

**SWITCHING REGULATOR INPUT FILTERING**



$$\text{ATTENUATION} \approx \frac{2\pi fL}{ESR}$$

Figure 3.66

**REFERENCES**

1. Irving M. Gottlieb, **Power Supplies, Switching Regulators, Inverters, and Converters, Second Edition**, McGraw Hill (TAB Books), 1994.
2. Marty Brown, **Practical Switching Power Supply Design**, Academic Press, 1990.
3. Marty Brown, **Power Supply Cookbook**, Butterworth-Heinemann, 1994.
4. John D. Lenk, **Simplified Design of Switching Power Supplies**, Butterworth-Heinemann, 1995.
5. Keith Billings, **Switchmode Power Supply Handbook**, McGraw-Hill, 1989.
6. George Chryssis, **High-Frequency Switching Power Supplies: Theory and Design, Second Edition**, McGraw-Hill, 1989.
7. Abraham I. Pressman, **Switching Power Supply Design**, McGraw-Hill, 1991.
8. Tantalum Electrolytic and Ceramic Families, Kemet Electronics, Box 5828, Greenville, SC, 29606, 803-963-6300.
9. Type HFQ Aluminum Electrolytic Capacitor and Type V Stacked Polyester Film Capacitor, Panasonic, 2 Panasonic Way, Secaucus, NJ, 07094, 201-348-7000.
10. OS-CON Aluminum Electrolytic Capacitor 93/94 Technical Book, Sanyo, 3333 Sanyo Road, Forest City, AK, 72335, 501-633-6634.
11. Ian Clelland, *Metalized Polyester Film Capacitor Fills High Frequency Switcher Needs*, **PCIM**, June, 1992.
12. Type 5MC Metallized Polycarbonate Capacitor, Electronic Concepts, Inc., Box 1278, Eatontown, NJ, 07724, 908-542-7880.
13. Walt Jung, Dick Marsh, *Picking Capacitors, Parts 1 and 2*, **Audio**, February, March, 1980.

## **SWITCHING REGULATORS**

### **Capacitor Manufacturers:**

14. AVX Corporation, 801 17<sup>th</sup> Ave. S., Myrtle Beach, SC 29577, 803-448-9411.
15. Sprague, 70 Pembroke Road, Concord, NH 03301, 603-224-1961.
16. Panasonic, 2 Panasonic Way, Secaucus, NJ 07094, 201-392-7000.
17. Sanyo Corporation, 2001 Sanyo Ave., San Diego, CA 92173, 619-661-6835
18. Kemet Electronics, Box 5828, Greenville, SC 29606, 803-963-6300

### **Inductor Manufacturers:**

19. Coiltronics, 6000 Park of Commerce Blvd., Boca Raton, FL 33487, 407-241-7876.
20. Sumida, 5999 New Wilke Rd. Suite 110, Rolling Meadow, IL. 60008, 847-956-0666.
21. Pulse Engineering, 12220 World Trade Drive, San Diego, CA 92128, 619-674-8100.
22. Gowanda Electronics, 1 Industrial Place, Gowanda, NY 14070, 716-532-2234.
23. Coilcraft, 1102 Silver Lake Rd., Cary, IL 60013, 847-639-2361.
24. Dale Electronics, Inc., E. Highway 50, P.O. Box 180, Yankton, SD 57078, 605-665-9301.
25. Hurricane Electronics Lab, 331 N. 2260 West, P.O. Box 1280, Hurricane, UT 84737, 801-635-2003.

### **Core Manufacturers:**

26. Magnetics, P.O. Box 391, Butler, PA 16003, 412-282-8282.

### **MOSFET Manufacturers:**

27. International Rectifier, 233 Kansas Street, El Segundo, CA 90245, 310-322-3331.
28. Motorola Semiconductor, 3102 North 56<sup>th</sup> Street, MS56-126, Phoenix, AZ 85018, 800-521-6274.
29. Siliconix Inc., 2201 Laurelwood Road, P.O. Box 54951, Santa Clara, CA 95056, 408-988-8000.

**Schottky Diode Manufacturers:**

30. General Instrument, Power Semiconductor Division, 10 Melville Park Road, Melville, NY 11747, 516-847-3000.
31. International Rectifier, 233 Kansas Street, El Segundo, CA 90245, 310-322-3331.
32. Motorola Semiconductor, 3102 North 56<sup>th</sup> Street, MS56-126, Phoenix, AZ 85018, 800-521-6274.