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Spec Title: SWITCHING REGULATORS: COMPONENT DESIGN GUIDE - AN50099

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CYPRESS PERFORM

AN50099

Switching Regulators: Component Design Guide

Author: Srinivas N V N S

Associated Project: No

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Abstract

Switching regulators provide a method to convert unfiltered, unregulated DC voltages into filtered, regulated DC outputs. This application note explains how to design components based on the specifications of a switching regulator.

Introduction

The raw power available in automotive batteries, industrial power supplies, and distributed supplies are sources of wide range, unregulated voltages. Switching regulators can easily convert these supplies into voltages required by applications. Better efficiency, simple design, reduced size, and lower costs make switching regulators a viable option for DC-DC converters.

A switching regulator typically consists of a power switch, a freewheeling diode, and an LC filter. The design of these components depends mostly on the input and output specifications, and also on converter specifications as a whole. The converter specifications include operating efficiencies, cost, and form factor. This application note provides the following:

- Introduction to basic switching regulator topologies
- Component design equations for basic topologies
- Design example

Switching Regulator Topologies

Switching regulators are used in almost every piece of electronic equipment; for example, computer and computer peripherals, scientific instruments, televisions, and hand help devices. Most of these devices require DC voltages that are filtered well and regulated. A power supply that converts available raw DC power to usable DC power must perform the following functions at low cost and high efficiency:

- **Voltage Transformation:** Supply the required DC voltage.
- **Filtering:** Smooth the ripple of converted voltage.
- **Regulation:** Control the output voltage level to a constant value, irrespective of line, load, and ambient changes.
- **Protection:** Prevent damaging voltage spikes on the source from reaching the output.

DC-DC converters are widely used to transform and distribute DC power in systems and instruments. DC power is usually available in the form of system power supply or battery, at 3.3 V, 5 V, 9 V, 12 V, 24 V, 36 V, or 48 V. Because these voltages are low, electrical isolation of the output from the input is not usually required until the application specifications require it.

There are several switching regulator designs. These designs are based on available input voltages and the required output voltages and currents. The three basic regulators are:

- **Buck Converter:** A step down regulator with an output voltage that is less than the input voltage. A typical application is reducing the standard automotive bus voltage at 12 V to 5 V to power TTL logic.
- **Boost Converter:** A step up regulator with an output voltage that is more than the input voltage. A typical application is increasing battery voltage from 3.6 V in a hand held device to 5 V to power a wireless transmitter.
- **Buck-Boost Converter:** A regulator where the absolute magnitude of the output voltage is higher or lower than the input voltage. The output voltage has the opposite polarity of the input voltage.

  The main application of this converter is in circuits, where the available input voltage is actually a range of voltages and an output voltage that falls in the same range is required.

For the converter to meet the required voltage and current specifications, you must carefully select the components, such as semiconductor switches, diodes, inductors, and capacitors. The values of the inductance and capacitance can be calculated with the following factors:

- Maximum and minimum input voltage
- Required output voltage
- Maximum allowable ripple in output voltage and output current
- Maximum and minimum load currents
- Desired switching frequency

Switching Regulator: Advantages

Switching regulators offer more advantages when compared to the traditional linear mode regulators. Switching regulators offer higher efficiencies because they dissipate less heat. This entails the fact the component size is smaller and less thermal management is required. The energy stored in the inductor can be used to step down the voltage, step up the voltage, and invert the voltage polarity. However, a disadvantage is that switching regulators may be noisy and require energy control for proper operation of the circuit.

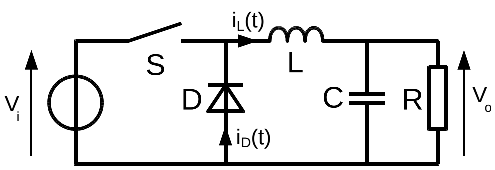
Passive Component Design

It is impossible to perfect low-pass filter at the output end of a DC-DC converter, which allows the DC component to pass through, but removes the components at the switching frequency and its harmonics. As a result, the low-pass LC filter must allow a few high frequency components generated by the switch to reach the output. So in practice, the output voltage and the inductor current consists of a DC component and a small, undesired AC component that arises out of incomplete attenuation of the switching harmonics.

The design of the inductance and capacitance values considers the required attenuation of switching harmonics in terms of ripple percentages. These values depend on the circuit configurations employed and are discussed in the following sections of this application note.

Buck Converter

Figure 1. Buck Converter



The buck converter, shown in Figure 1, is the simplest of all converters. A PWM signal of duty-cycle 'D' is used to drive the semiconductor switch 'S'. When the switch is on, the input voltage V_i is directly applied to one terminal of the inductor L. Assuming that the circuit has reached a steady state, the output voltage is at V_o. As a result, the voltage across the inductor is $(V_i - V_o)$. The inductor current is calculated using the following definition:

$$L\frac{di_L(t)}{dt} = v_L(t) \qquad \text{Equation 1}$$

Figure 2. Buck Converter Operation: Second Sub Interval

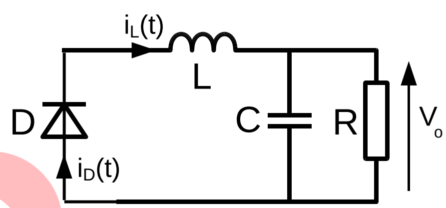
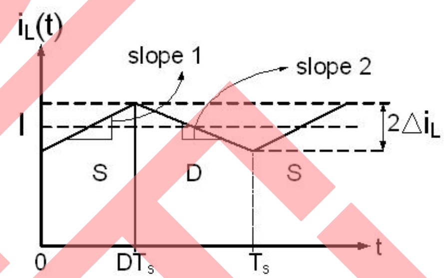


Figure 3. Inductor Current Waveform



Therefore, during the first interval, when $v_L(t)$ is approximately $(V_i - V_o)$, the slope of the inductor current waveform, slope 1, is:

$$\frac{di_L(t)}{dt} = \frac{v_L(t)}{L} = \frac{(V_i - V_o)}{L} \qquad \text{Equation 2}$$

Similar arguments apply during the second sub interval. Figure 2 shows the equivalent circuit when the switch is off. Maintaining the polarity of the inductor voltage, the inductor voltage is calculated using the following equation:

$$v_L(t) = -V_o$$

The slope of the inductor current, slope 2 in Figure 3, is calculated using the following equation:

$$\frac{di_L(t)}{dt} = \frac{v_L(t)}{L} = \frac{-V_o}{L} \qquad \text{Equation 3}$$

The inductor current wave form can now be sketched. During the first sub interval, when the switch is on, the inductor current increases with the slope given in Equation 2. At time $t = DT_S$, the switch closes, and the current decreases with the constant slope given in Equation 2.

The inductor current ripple is now equal to the DC current component I, plus the peak to average ripple, Δi_L.

Because the slope of the inductor currents during both the switching intervals is now known, the magnitude of the inductor current ripple can be calculated. The waveform of the inductor current ripple is symmetrical about DC current component I. During the first sub interval, the inductor current increases by $2\Delta i_L$. So the change in current, $2\Delta i_L$, is equal to slope 1 multiplied by the length of the first sub interval.

$$2\Delta i_L = \left(\frac{V_i - V_o}{L}\right) DT_s \quad \text{Equation 4}$$

The change in current, $2\Delta i_L$, is also equal to slope 2 multiplied by the length of the second sub interval.

$$2\Delta i_L = \frac{-V_o}{L}(1-D)T_s \quad \text{Equation 5}$$

Because the inductor current must be the same at the end of each switching cycle in steady state operation, the net change in current given by the sum of Equation 4 and Equation 5 must be zero.

$$\left(\frac{V_i - V_o}{L}\right) DT_s - \frac{V_o}{L}(1-D)T_s = 0$$

From the previous equation, the input voltage and the output voltage are related by the duty cycle D. This is calculated as:

$$V_o = DV_i \quad \text{Equation 6}$$

From Equation 2 on page 2, the inductance needed for the required ripple current is calculated as:

$$L = \left(\frac{V_i - V_o}{2\Delta i_L}\right) DT_s \quad \text{Equation 7}$$

The value of the inductance directly correlates to the duty cycle 'D' and the time period T_s at a constant ripple current value. As a result, increasing the operating duty cycle and/or decreasing the switching frequency increases the required inductance value.

In practical applications where duty cycle 'D' is adjusted to regulate the output voltage, you must take the maximum possible duty cycle when the inductance value is calculated.

Note that the inductance value depends on the difference between the input and output voltages, ($V_i - V_o$). For a constant required output voltage, the maximum value of the input voltage, V_{imax}, decides the inductance value. Now, you can rewrite Equation 2 as:

$$L = \left(\frac{V_{i\max} - V_o}{2\Delta i_L}\right) D_{\max} T_s \quad \text{Equation 8}$$

Typically, the ripple in the inductor current is restricted to approximately 15 percent – 20 percent of the nominal DC load current. Lighting applications may allow ripple up to 30% of the nominal DC load current.

Output capacitance is required to minimize the voltage overshoot and the ripple present at the output of the converter. Voltage overshoot appears when a full load is suddenly removed from the output. The output capacitor must be large enough to prevent the stored energy in the inductance to raise the capacitor's voltage above the specified maximum output voltage. The energy stored in the inductor at the end of the first sub interval is:

$$E_L = \frac{1}{2} L \left(I + \frac{\Delta i_L}{2} \right)^2$$

If Δv is the allowed overshoot in the capacitor:

$$\frac{1}{2} L \left(I + \frac{\Delta i_L}{2} \right)^2 = \frac{1}{2} C (V_o + \Delta v)^2 - \frac{1}{2} C (V_o)^2$$

This equation means that the energy stored in the inductor is dumped into the capacitor, which increases the stored energy in the capacitor. This in turn increases the terminal voltage.

The value of the capacitor used to prevent the voltage overshoot is:

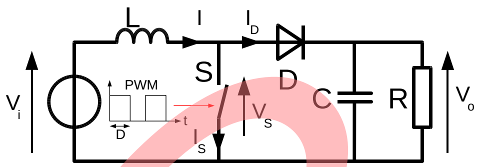
$$C = \frac{\frac{1}{2} L \left(I + \frac{\Delta i_L}{2} \right)^2}{(V_o + \Delta v)^2 - (V_o)^2} \quad \text{Equation 9}$$

The output capacitor ESR dominates the output voltage ripple. The amount of ripple in output voltage is:

$$V_{o_{ripple}} = I_{ripple} \times ESR_C = \Delta i_L \times ESR_C$$

Boost Converter

Figure 4. Boost Converter



The boost converter shown in Figure 4 is a switching converter that gives an output voltage greater or equal to the input voltage. When the switch is on, input voltage V_i is applied to the inductor L. Assuming that the circuit has reached a steady state, the output voltage is at V_o. So the voltage across the inductor is V_i. Using the inductor voltage waveform, the inductor current is calculated by the following definition:

$$L\frac{di_L(t)}{dt} = v_L(t)$$

As a result, during the first interval, when $v_L(t)$ is approximately V_i, the slope of the inductor current waveform is:

$$\frac{di_L(t)}{dt} = \frac{v_L(t)}{L} = \frac{V_i}{L}$$ Equation 10

Similar arguments apply during the second sub interval. Maintaining the polarity of the inductor voltage, the inductor voltage is calculated using the following equation:

$$v_L(t) = V_i - V_o$$

The slope of inductor current is:

$$\frac{di_L(t)}{dt} = \frac{v_L(t)}{L} = \frac{V_i - V_o}{L}$$ Equation 11

During the first sub interval, when the switch is on, the inductor current increases with the slope given in Equation 10 . At time $t = DT_s$, the switch closes, and the current decreases with the constant slope given in Equation 11.

The inductor current ripple is now equal to the DC current component I, plus the peak to average ripple, Δi_L. The DC component I is different from the load current, and in this case it represents the input current of the converter.

The slope of the inductor current during both the switching intervals is already calculated in Equation 10 and Equation 11, so you can now find the magnitude of the inductor current ripple. The waveform of the inductor current ripple is symmetrical about DC current component I. During the first sub interval, the inductor current increases by $2\Delta i_L$. As a result, the change in current, $2\Delta i_L$, is equal to the slope multiplied by the length of the first sub interval:

$$2\Delta i_L = \left(\frac{V_i}{L}\right)DT_s$$ Equation 12

From Equation 10, the inductance required for the ripple current is calculated using:

$$L = \left(\frac{V_i}{2\Delta i_L}\right)DT_s$$ Equation 13

The value of the inductance directly correlates to the duty cycle ‘D’, and the time period Ts at a constant ripple current value. So increasing the operating duty cycle and/or decreasing the switching frequency increases the inductance value required. Similarly, at a constant duty cycle and a constant switching frequency, increasing the current ripple decreases the value of inductance required.

Note that the inductance value depends on the input voltage V_i. For a constant output voltage, the maximum value of the input voltage V_{imax} decides the inductance value. Now Equation 13 is rewritten as:

$$L = \left(\frac{V_{i\max}}{2\Delta i_L}\right)D_{\max}T_s$$ Equation 14

Figure 5. Output Capacitor Current Waveform

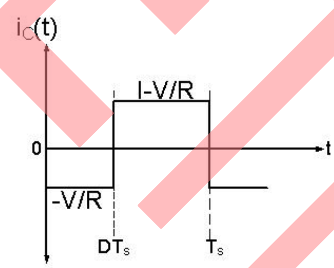
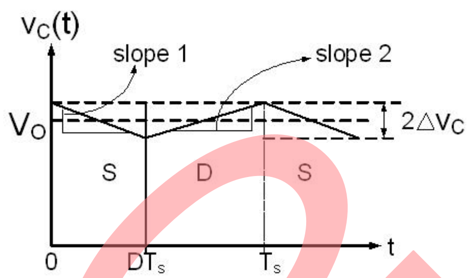


Figure 6. Output Capacitor Voltage Waveform



Similarly, the capacitor voltage $v(t)$ can be sketched and an expression derived for the output voltage ripple peak magnitude Δv. The capacitor current waveform $i_C(t)$ is shown in Figure 5.

During the first subinterval, the slope of the capacitor voltage, slope 1, is:

$$\frac{dv(t)}{dt} = \frac{i_C(t)}{C} = -\frac{V_o}{RC}$$

During the second subinterval, slope 2 is:

$$\frac{dv(t)}{dt} = \frac{i_C(t)}{C} = \frac{I}{C} - \frac{V_o}{RC}$$

he capacitor voltage waveform is shown in Figure 6 on page 4. During the first subinterval, the change in capacitor voltage is – $2\Delta v$, which is equal to the slope multiplied by the length of the subinterval:

$$-2\Delta v = -\frac{V_o}{RC} DT_s$$

Therefore:

$$\Delta v = \frac{V_o}{2RC} DT_s$$

Now, for a given target ripple Δv, the capacitance required is:

$$C = \frac{V_o}{2R\Delta v} DT_s \qquad \text{Equation 15}$$

Choosing an Inductor

Choosing a proper switching frequency is very important when selecting an inductor. The key advantage in choosing a higher frequency operation is smaller component sizes, which results in a smaller inductor. Inductor behavior is governed by the following equations:

$$L\frac{di_L(t)}{dt} = v_L(t)$$

$$E_L = \frac{1}{2} L i_L^2$$

$$L \propto N^2$$

DC-DC converters transfer energy at a controlled rate from an input source to an output load. As the switching frequency increases, the time available for this energy transfer decreases. For example, consider a buck converter operating at 1 MHz with a 100 µH inductor. For most DC-DC converters, changing the frequency to 2 MHz allows use of exactly one half the inductance, that is 50 µH.

The current requirement remains the same, although the inductor size is reduced. Energy storage is reduced to half because the inductor value is also reduced to half. Further number of turns reduces to 70.7 percent of the original value when the inductor value is reduced by half. This means that the copper wire length is reduced to the same extent and so DCR is also reduced to the same extent.

Importance must be given to DCR and ESR at the frequency of operation of the converter. The ESR or the AC resistance contributes to the significant power loss at low operating currents with considerable ripple. ESR typically increases with respect to operating frequency. It is usually a tradeoff between the frequency of operation and the expected efficiency of the converter at low operating currents.

Deciding the value of the inductance leads to the task of selecting the inductor. Instead of selecting off the shelf inductors, a little knowledge of Magnetics can help you choose the optimal inductor for the system. For high frequency operations, inductors with ferrite core are usually preferred.

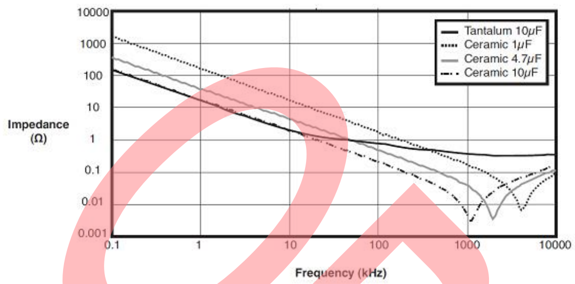
Inductors can be a source of radiated magnetic fields and must be chosen to minimize this effect. The key to reducing inductor emissions is to use high-µ material to keep the field in the core and out of the surrounding space.

The saturation current limit of the inductor, I_{sat}, is defined as the DC current at which the inductance drops 10 percent -30 percent from its value without current, depending on the manufacturer. You must ensure that the inductor current rating I_{sat}, is rated at least to the typical current rating of the switch. The I_{sat} value at 85°C must be considered before deciding the suitability of the inductor. In addition, RMS ratings must be considered to determine the worst case component temperature.

Shielded inductors have more contained fields when compared to unshielded inductors. Unshielded inductors have a wide field surrounding the part that could affect other components situated near it.

Choosing a Capacitor

Figure 7. Impedance vs Frequency Curve



Selecting the right operating frequency also affects the value of the capacitance. Increase in operating frequency reduces the value of capacitance by the same factor. The ripple current requirement remains the same for the converter specifications. It is important that the capacitor impedance is lower at the ripple or the switching frequency.

Figure 7 shows the impedance versus frequency curve of typical capacitors from a data sheet of a capacitor supplier. The curve shows that the impedance hits a minimum at a certain frequency. At this point, the capacitive impedance and the inductive impedance of the capacitor cancel and the capacitor behaves as a resistance. This is usually the ESR of the capacitor. The ESR depends a lot on the construction of the capacitor. Typically, the ESR is lower for the higher value capacitor in the same case size. The ESR is also lower for a same valued capacitor in a higher case size. ESL or the equivalent series inductance of a capacitor is also lower for a capacitor with a smaller case size.

Ceramic capacitors usually have a low ESR when compared to tantalum or electrolytic capacitors. As a result, a ceramic capacitor of low value, for example, 0.1 uF is usually placed across the output bulk capacitors to lower the ESR of the capacitance bank.

Take the closest standard value for the capacitor with the lowest ESR. It is better to deviate from the calculated capacitance value when there is another capacitor value with lower ESR.

Ceramic chip capacitors offer better reliability, performance, no polarity, and allow design engineers to downsize designs with various package options.

Capacitors must be properly rated in terms of temperature. There is an EIA three character code that indicates the temperature coefficient of a capacitor. Manufacturers typically provide these details in the data sheet of the capacitor. For a non-temperature-compensating capacitor, the code consists of three letters. The first character is a letter that gives the low end operating temperature. The second is a digit that gives the high end operating temperature. The third and final letter gives the capacitance change over that temperature range.

Table 1. Capacitor Temperature Rating Naming Convention

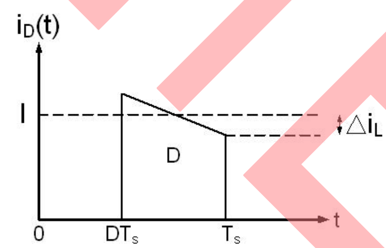
Letter (Low Temp)	Digit (High Temp)	Letter (Change)
X= -55 °C (-67 °F)	2= +45 °C (+113 °F)	D= ± 3.3 %
Y= -30 °C (-22 °F)	4= +65 °C (+149 °F)	E= ± 4.7 %
Z= +10 °C (+50 °F)	5= +85 °C (+185 °F)	F= ± 7.5 %
	6=+105 °C (+221 °F)	P= ± 10 %
	7=+125 °C (+257 °F)	R= ± 15 %
		S= ± 22 %
		T= +22 to -33 %
		U= +22 to -56 %
		V= +22 to -82 %

For applications that require higher operating temperatures and less deviation of capacitance over temperature, the X7R type of capacitors are usually adopted. An X7R capacitor operates from -55 °C to +125 °C with a capacitance change of ±15 %.

Choosing a Freewheeling Diode

The diode's primary function is to provide a freewheeling path for the load current when the switch is off. The current freewheels because the inductor does not allow a sudden change in current, and so must be provided a path to decay. To this effect, the diode must quickly turn on and pick up the load current when the switch turns off. This fast action of the diode protects the switch from experiencing high voltage overshoots and clamps the drain source of the switch to the supply voltage in a buck converter. Figure 8 shows the waveform of the current through the freewheeling diode of a buck converter.

Figure 8. Freewheeling Diode Current Waveform in Buck Converter



A freewheeling diode must be rated to the rms current it conducts. The rms value of the current waveform of the buck converter in Figure 8 is:

$$I_{Drms} = I_{\max}(1 - D_{\min}) \qquad \text{Equation 16}$$

In this equation, I_{max}, is the maximum load current at a minimum duty cycle D_{min}.

At higher currents, designers must also consider the power dissipation the diode can handle.

In average thermal power calculations, the instantaneous voltage drop across the diode at the maximum converter output current must be considered. This must be compared against the value in the manufacturer's data sheet.

Diodes must also turn off appropriately when current transfers to the switch. The reverse recovery time of the diode represents the time it takes for the diode to recover its ability to block reverse voltage. During this time, the diode can conduct current in the reverse direction. If the switch is turned on during this time, substantial reverse currents flow causing power losses. When the stored charge is depleted across the junction, the voltage drop across the diode snaps in the direction of the reverse bias, and after a brief overshoot, settles at the steady state reverse biased value.

To counter the problem of power loss in the diodes, Schottky diodes are a typical choice for switched power supplies. They have a low forward drop voltage and are majority carrier devices, so they do not suffer from the minority carrier storage problems that slow down many other diodes. Schottky diodes also have faster "reverse recovery" than p-n junction diodes. They also tend to have much lower junction capacitance than p-n diodes, which provides for high switching speeds, and can be used in high frequency switching circuits.

Choosing a Sense Resistor

Current mode of control is typically employed in converters. To achieve this, the input current or the output current must be sensed according to the control regime. Sense resistors are used in low cost systems. The value of the sense resistor is typically governed by two factors:

- Minimum and maximum sense voltages the sense amplifier can handle
- Power dissipation in the sense resistance

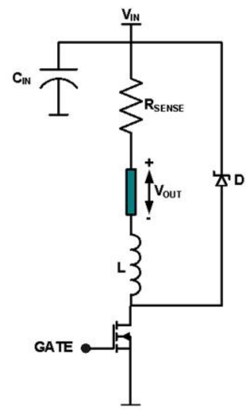
The reason behind choosing a sense resistor is to minimize power dissipation, and also satisfy the sense voltage requirements at the input of the sense amplifier.

This is summarized in the following equation:

$$R_{SENSE} = \frac{V_{SENSE\min}}{I_{SENSE\max}}$$

Design Example

Figure 9. Floating Load Buck Converter



A floating load buck converter is a variant of buck topology, as shown in Figure 9. In such type of converters, capacitors across the output are usually not required, unless the current ripple is a contributing factor to EMI. When compared to a traditional buck topology, the switch is moved to the low side, the load is floating, and the diode is placed between the switch and the high side supply. The converter design equations are the same as that of a normal buck converter.

In case of an LED driver application, the load is typically a string of LEDs. V_{OUT} is then equal to the forward voltage of the LED string at the rated current I_{LED}.

Consider a typical scenario of driving six 350 mA power LEDs with a forward voltage drop of 3 V each. Let the input voltage be in the range of 22 V to 26 V. The current ripple is restricted to 20 percent of the nominal rated current and the switching frequency does not exceed 2 MHz.

In this scenario, n = 6, V_F = 3V, I_{OUT} = 0.35A, ΔI_{OUT} = 0.070 A, V_{INmin} = 22 V, V_{INmax} = 26 V.

Therefore, V_{OUT} = n x V_F = 18 V.

The maximum duty cycle of operation $D_{MAX} = V_{OUT} / V_{INmin}$ = 18/22 = 0.82.

The inductance value required for this operation is calculated from:

$$L = \left(\frac{V_{IN\max} - V_{OUT}}{2\Delta i_L}\right) D_{MAX} T_s$$

In this equation, T_s is the switching frequency time period and is equal to 500 nanoseconds. This is the case when all non idealities are neglected and there is no sensing element in the circuit. When the sensing element and series resistance drops in the circuit are also considered, the amount of sense resistor drop and switch drop must be deducted from the term (V_{INmax} - V_{OUT}).

Inductance value is calculated as L = 46.75 µH. A shielded inductor with ferrite core of value 47 µH can be selected. LPS6235-473MLC from Coilcraft is a good example.

A sense resistor of value 0.1 ohms can be chosen for this application. The net power loss through the sense resistor would be 17.5 mW.

Summary

This application note enables you to quickly figure out the values of the inductor, capacitors, diode, and sense resistor for use in switching regulators. It also helps you to choose the types of components from a good design perspective.

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*B	3934500	MKKU	03/15/2013	Obsolete spec.

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