LX1741 / LX1742 BOOST CONVERTER DESIGN HINT

AN-22

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Microsemi
INTEGRATED PRODUCTS

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INTRODUCTION

The LX1741 and LX1742 Boost Controllers offer high efficiency performance and provide power management circuit designers with the ability to approach a broad range of design applications with a flexible and easy-to-implement solution. This application hint provides an overview of these two products and describes their overall functions in regard to practical design applications. Please refer to the LX1741 and LX1742 data sheets for a complete discussion regarding electrical performance and packaging information. Table 1 provides a comparison of the LX1741 and LX1742 features.

	Internal FET	External FET	Max V_{IN}	I_{SRC} (MAX)	V_{OUT} (MAX)	Package Type	Operating Temp.
LX1741	No	Yes	6.0V	800mA (rms)	Application Dependent	MSOP or MLP	0 ~ 70°C
LX1742	Yes	No	6.0V	500mA (rms)	25V	MSOP	0 ~ 70°C

TABLE 1 – FEATURES

LX1741 AND LX1742

The LX1741 and LX1742 are very similar devices. Both of these controllers implement a Pulse Frequency Modulation-type (PFM) topology and provide designers with a cost effective SMPS controller solution for a variety of real-world battery (e.g., Lithium-Ion) driven applications (e.g., pagers, wireless phones, personal digital assistants, etc…).

The LX1741 or LX1742 support a broad range of output voltages and can source over 100mA of output current depending upon input voltage. One particular low current application that benefits from the selection of either the LX1741 or LX1742 includes Liquid Crystal Display (LCD) biasing.

Each device has 8 functional pins that are designated as IN (voltage input), OUT (voltage output), CS (current sense – used to set the peak inductor current limit), SHDN (active-low shutdown – disables the controller and reduces supply current to < 1mA), GND (circuit ground), FB (feedback – a resistor divider network is connected between this pin and ground to establish V_{OUT}), ADJ (adjust – provides for external control of the output voltage by up to ±15%), and NDRV (n-channel mosfet driver output – LX1741 only) or, SW (switch – LX1742 only: inductor output & diode (anode) input connection – this pin is high impedance in shutdown mode).

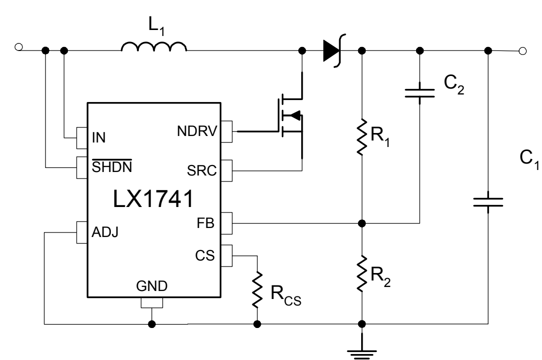

FIGURE 1 – TYPICAL LX1741 APPLICATION CIRCUIT

The LX1741 requires an external N-channel MOSFET to complete the DC-DC converter circuit. This feature provides the designer with maximum flexibility regarding the selection of a device that minimizes switching losses for a particular application. The external MOSFET is driven from the NDRV pin and Figure 1 shows a typical LX1741 application circuit.

The LX1742 simplifies a portion of the design effort by incorporating the N-channel MOSFET device. Applications that have relaxed efficiency and/or increased reliability requirements benefit from this added feature. Figure 2 illustrates a typical LX1742 application. The LX1742 replaces the NDRV pin (found in the LX1741) with the SW (i.e., switch) pin and requires the inductor's output and the diode's anode to be connected to this pin.

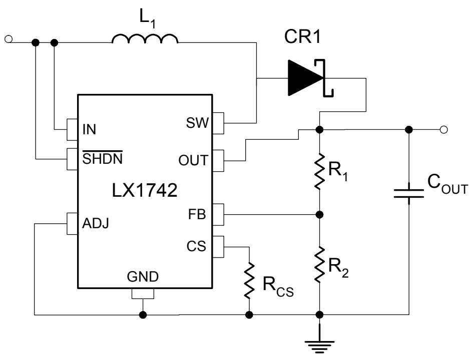

FIGURE 2 – TYPICAL LX1742 APPLICATION CIRCUIT

Other critical design considerations apply to the selection of the inductor, capacitors, diode and transistor. The designer can minimize inductor size, input ripple current, and output ripple voltage by setting the peak inductor current level to 1.5X the expected maximum DC input current. Low ESR capacitors are recommended because they reduce output voltage ripple induced by the inductor's switching current. Multi-layer ceramic capacitors with X5R or X7R dielectric make a superior choice because they feature small size, very low ESR, and a temperature stable dielectric. Low ESR electrolytic capacitors such as solid tantalum or OS-CON types are also acceptable (note: a brief review of capacitor types is provided in the Appendix). When choosing the diode, the designer should consider the device's average and peak current ratings with respect to the application's output and peak inductor current requirements.

Moreover, the diode's reverse breakdown voltage characteristic must be capable of withstanding a negative voltage transition that is greater than VOUT. A properly sized Schottky diode will typically meet these requirements for a broad range of applications. Finally, overall circuit efficiency is further enhanced by selecting a MOSFET device that exhibits a low $R_{DS(ON)}$ and gate charge characteristic. Typical Efficiency versus Output Current Curves for the LX1741 and LX1742 are shown in Figure 3.0 and Figure 4.0 respectively.

LX1741 DESIGN EXAMPLE

Let's work through a typical application using the LX1741. We'll assume that an input voltage supply of 3.6V is available. We need to drive an application that requires an output of 12.0V and 40.0mA (i.e., 480mW). The first step in selecting the desired output voltage is to determine the value of R1 and R2. We'll start with an R2 value less than 100KΩ to minimize V_{FB} offset error (e.g., I_{FB}*R2 = 200nA*50KΩ = 10mV). We will determine R1 using formula 1.0 where V_{REF} = 1.29V or 1.20V for the LX1741 and LX1742 respectively:

Eq 1

$$R_1 = R_2\left(\frac{[V_{OUT} - V_{REF}]}{V_{REF}}\right) =$$

$$49.9K\Omega\left(\frac{[12.0V - 1.29V]}{1.29V}\right) = 414.3K\Omega$$

We'll select an R1 value using a 1% resistor that is the closest to 414.3KΩ (i.e., R1 equals 412KΩ). Now, we need to determine the Peak Inductor Current (I_{PEAK}). Therefore, we will start by determining I_{IN} using the efficiency equation where Output Power (P_{OUT}) is equal to the Efficiency (η) multiplied by the Input Power (P_{IN}):

Eq. 2

$$P_{OUT} = \eta(P_{IN})$$

Recall that power is equal to voltage multiplied by current (i.e., P = V*I). Therefore, we can rewrite the efficiency equation and solve for IIN where, V_{IN} equals 3.6V, V_{OUT} equals 12.0V, I_{OUT} equals 40mA (maximum) and η is estimated from the device's efficiency versus output current curve.

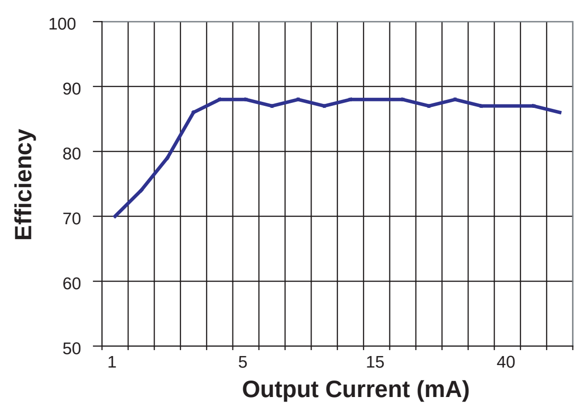

FIGURE 3 – LX1741 EFFICIENCY VS. OUTPUT CURRENT
(V_{IN} = 3.6V, V_{OUT} = 12V, L = 47µH, R_{CS} = 4KΩ)

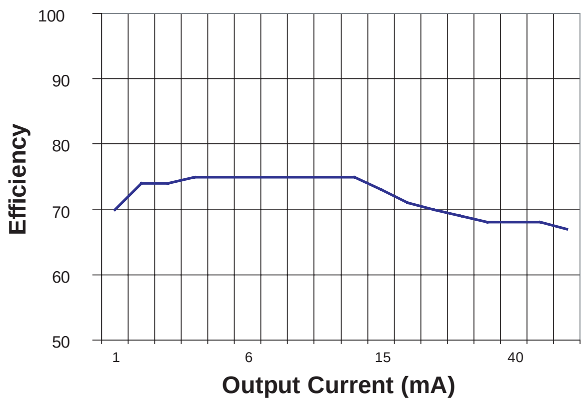

FIGURE 4 – LX1742 EFFICIENCY VS. OUTPUT CURRENT
(V_{IN} = 3.6V, V_{OUT} = 12V, L= 47µH, R_{CS} = 4KΩ)

An efficiency value of approximately 0.85 is found from the curve in Figure 3. Now, I_{IN} may be estimated as follows:

Eq 3

$$I_{IN} = \left(\frac{[I_{OUT} \times V_{OUT}]}{\eta \times V_{IN}}\right) = \left(\frac{[40mA \times 12V]}{0.85 \times 3.6V}\right) = 156mA$$

Now that the input current has been determined, we are ready to calculate the peak inductor current. I_{PEAK} is a function of several parameters, specifically: I_{MIN}, V_{IN}, L, t_D, I_{SCALE}, and R_{CS}. V_{IN} is already defined herein as 3.6V. The ELECTRICAL CHARACTERISTICS section of the LX1741 data sheet provides the values parameters I_{MIN} and t_D. For this example, we will use the LX1741's nominal I_{MIN} and I_{SCALE} value of 145mA and 31mA/kΩ respectively (note: these values change to 104mA and 22mA/kΩ respectively for the LX1742). The parameter t_D is a switching delay related to the operation of the feedback comparator circuit (see Block diagram in data sheet). A typical value for t_D, at 25°C, is 620ns. *Microsemi* recommends using an inductor (L) value of 47µH (i.e., this value works for a broad power conversion range).

A higher inductance value may improve efficiency at the expense of degrading the overall output voltage ripple performance. Inserting a smaller inductance value will degrade efficiency. Moreover, the designer is encouraged to consider I_{PEAK} variation over the input voltage range as a smaller inductance increases I_{PEAK} variation versus a larger inductance. Figure 5 illustrates the R_{CS} versus input voltage relationship as inductance values increase from 27µH to 94µH and the I_{PEAK} value remains fixed at 350mA. Figure 6 shows the I_{PEAK} value versus input voltage relationship as inductance values increase from 27µH to 94µH and the R_{CS} value remains fixed at 4KΩ.

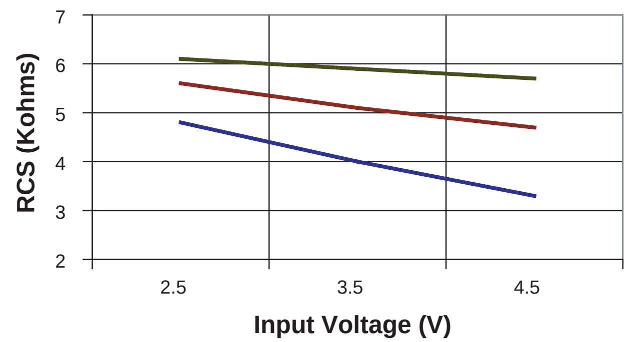

FIGURE 5 – R_{CS} (KΩ) VS. INPUT VOLTAGE
(Note: I_{PEAK} = 350mA, L = 27µH (bottom), 47µH (middle), 94µH (top), t_D = 618ns, I_{MIN} = 145mA)

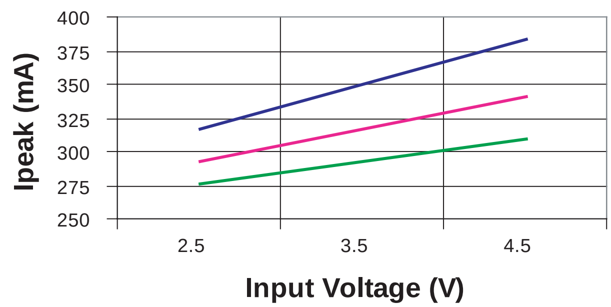

FIGURE 6 – I_{PEAK} (mA) VS. INPUT VOLTAGE
(Note: R_{CS} = 4.02KΩ, I_{SCALE} = 31mA/KΩ, L = 27µH (top), 47µH (middle), 94µH (bottom), t_D = 618ns, I_{MIN} = 145mA)

Using this information, we are ready to determine the R_{CS} value required to set the I_{PEAK} value for our application. From our previous calculation, IIN was determined to be 245mA. We will multiple this number by a factor of 1.5 to ensure that we have sufficient margin (over temperature and device-to-device variability) and reduce the risk of hitting current-limit (continuous-mode operation). Therefore, I_{PEAK} = 1.5(I_{IN}) = 1.5(156)mA = 235mA < 800mA_{RMS} (I_{SRC}). Note: The maximum I_{PEAK} value is limited by the I_{SRC} value (max. = 0.8A_{RMS} : LX1741). Now we can solve for R_{CS} by using formula 5 where:

Eq 4

$$I_{PEAK} = I_{MIN} + \left(\frac{V_{IN}}{L}\right)t_D + \left(I_{SCALE} \times R_{CS}\right) \text{ hence:}$$

Eq 5

$$R_{CS} = \left(\frac{1}{I_{SCALE}}\right)\left(I_{PEAK} - I_{MIN} - \left(\frac{V_{IN}}{L}\right)t_D\right) \text{ or,}$$

$$R_{CS} = \left(\frac{1}{32\text{mA/K}\Omega}\right)\left(235\text{mA} - 145\text{mA}\left(\frac{3.6\text{V}}{47\mu\text{H}}\right)620\text{ns}\right) = 1317\Omega$$

We select an R_CS value using 1% resistors that is the closest to the calculated R_{CS} value; hence, R_{CS} equals 1.37KΩ. Connecting this resistor (R_{CS}) between the CS pin and ground sets the I_{PEAK} value in circuit. Now that we have determined values for R1, I_{PEAK}, and R_{CS}, what about calculating the output ripple voltage? The total output ripple voltage is determined from formula 6:

Eq. 6

$$V_{RIPPLE} = \Delta V_{DROOP} + \Delta V_{OVERSHOOT} + 10\text{mV}$$

Figure 7A illustrates the (ideal) switching waveform relationships with respect to the Droop and Overshoot voltage. The overshoot voltage occurs when the inductor charging cycle ends and the inductor current is released to the load. The overshoot voltage is a function of the inductor, output capacitor, Peak Inductor Current, output current, input voltage, output voltage, and an estimate of the voltage drop across the diode (e.g., 0.5v). The overshoot voltage is improved by increasing the output capacitance but degrades slightly with increasing input voltage. The droop voltage occurs when the output voltage begins to decrease below the feedback threshold. The droop voltage is a function of the inductor, output capacitor, Peak Inductor Current, output current, input voltage, and an estimate of the voltage drop across the inductor and the FET's R_{DS_ON} (e.g., 0.5v). Droop voltage improves by increasing either output voltage or output capacitance (or, both). Finally, there is a 10mV transition error voltage associated with the feedback switching circuit that adds to the total output ripple voltage. Figure 7B shows actual waveforms.

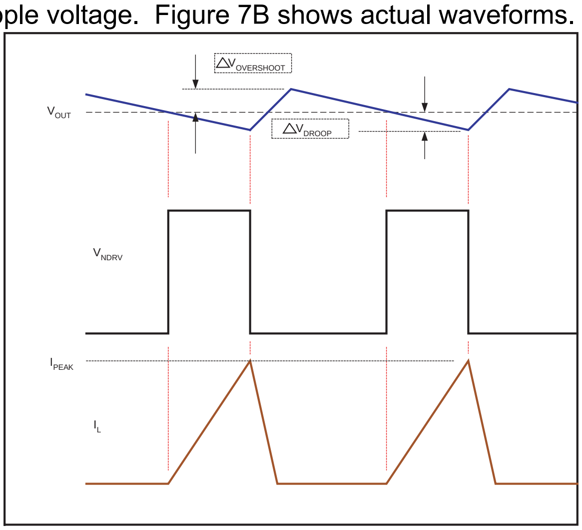

FIGURE 7A – IDEAL SWITCHING WAVEFORMS

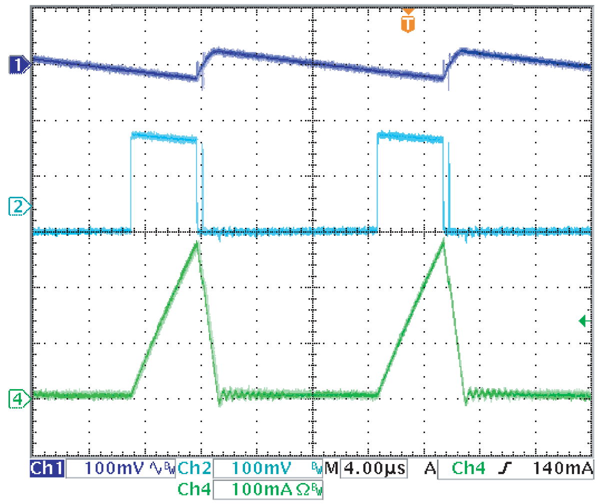

FIGURE 7B – ACTUAL SWITCHING WAVEFORMS
(V_{IN} = 3V, V_{OUT} = 12V, I_{OUT} = 11mA)

The delta (Δ) V_{DROOP} and $V_{OVERSHOOT}$ are determined from equations 7 and 8 respectively:

Eq 7

$$\Delta V_{DROOP} = \frac{\left(L/C_{OUT}\right)\left(I_{PK} \times I_{OUT}\right)}{V_{IN} - 0.5}$$

Eq 8

$$\Delta V_{OVERSHOOT} = \frac{1/2 \times \left(\frac{L}{C_{OUT}}\right)\left(I_{PK} \times I_{OUT}\right)^2}{V_{OUT} + 0.5 - V_{IN}}$$

Let's determine the ΔV_{DROOP} for our application example. Recall that in the previous example, I_{PEAK} was determined to be 368mA. We'll start by using the inductor value of 47mH and choose C_{OUT} equal to 4.7mF. Inserting these values into equation 7 and solving for the ΔV_{DROOP} yields the following result:

$$\Delta V_{DROOP} = \frac{\left(47\mu\text{H}/4.7\mu\text{F}\right)\left(235\text{mA} \times 50\text{mA}\right)}{3.6\text{v} - 0.5\text{v}} = 37.9\text{mV}$$

Now let us determine the $\Delta V_{OVERSHOOT}$ for our application example:

$$\Delta V_{OVERSHOOT} = \frac{1/2 \times \left(\frac{47\mu\text{H}}{4.7\mu\text{F}}\right)\left(235mA \times 50mA\right)^2}{12\text{v} + 0.5\text{v} - 3.6v} = 10.6mV$$

Combining the results of these calculations with formula 6 provides an estimate of 58mV for the output voltage ripple. What if we increased V_{IN} or C_{OUT} while holding the other parameters in formula 7 and 8 constant? Figure 8 and Figure 9 illustrate

Droop and Overshoot Voltage variation versus input voltage for two values of output capacitance. The Droop Voltage curve shows a voltage reduction as V_{IN} increases. The bottom curve (C_{OUT} = 47µF) in Figure 8 demonstrates a significant reduction in Droop voltage versus the top curve (C_{OUT} = 4.7µF). The Overshoot Voltage curves in Figure 9 demonstrate a slight overshoot voltage increase as V_{IN} increases. However, note that the bottom curve (C_{OUT} = 47µF) demonstrates a significant reduction in overshoot voltage value versus the top curve (C_{OUT} = 4.7µF). This exercise provides insight into the criticality of selecting the size of the output capacitor for a particular application. The LX1742 demonstrates similar performance characteristic.

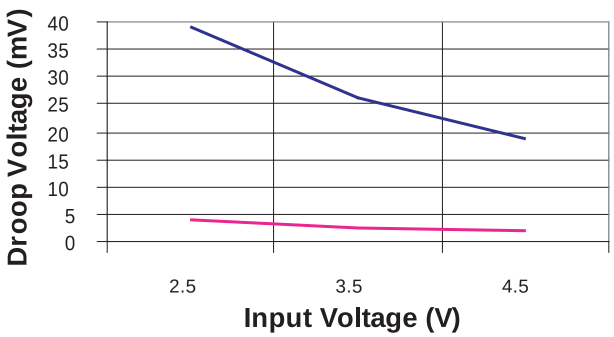

FIGURE 8 – ΔV_{DROOP} VS. INPUT VOLTAGE

(Note: I_{PEAK} = 310mA, I_{OUT} = 25mA, C_{OUT} =4.7µF (top), and 47µF(bottom))

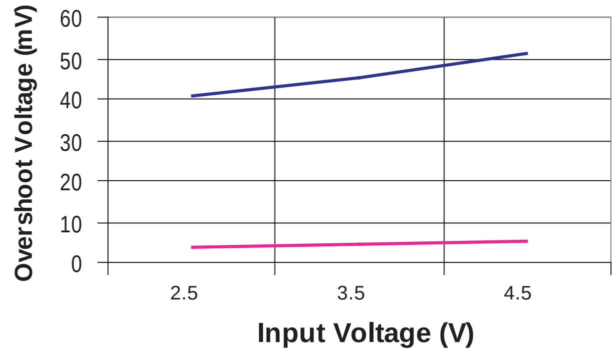

FIGURE 9 – $\Delta V_{OVERSHOOT}$ VS. INPUT VOLTAGE

(Note: I_{PEAK} = 310mA, I_{OUT} = 25mA, C_{OUT} = 4.7µF (top), and 47µF (bottom)).

Design Parameters (I_{OUT} = 40mA)	**Measured Parameters (I_{OUT} = 5mA)**	**Measured Parameters (I_{OUT} = 40mA)**
• V_{IN} = 3.6V • V_{OUT} = 12.0V • I_{PEAK} = 235mA • I_{OUT} = 40mA @ $\eta \geq$ 85% • I_{IN} = 156mA (est.) • Output Ripple < 60mV	• V_{IN} = 3.60V • V_{OUT} = 11.83V • I_{PEAK} = 238mA • I_{OUT} = 5mA @ η = 86.2% • I_{IN} = 19mA • Output Ripple ~ 45mV	• V_{IN} = 3.58V • V_{OUT} = 11.55V • I_{PEAK} = 238mA • I_{OUT} = 40mA @ η = 89.6% • I_{IN} = 144mA • Output Ripple ~ 65mV

TABLE 2 – COMPARISON OF DESIGN VS. ACTUAL PERFORMANCE

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DESIGN SUMMARY

Figure 10 and Figure 11 show the actual switching waveforms for the circuit based upon this design exercise. Channel 2 shows the output ripple voltage, channel 3 shows the NDRV output (pin 8: LX1741), and channel 4 shows the inductor current. Figure 10 shows light load (i.e., I_{OUT} = 5mA) waveforms and Figure 11 shows heavy load waveforms (i.e., I_{OUT} = 40mA). Table 2 highlights the variance between the design and measured performance of the circuit.

These results show that we have achieved our design requirements (at T_A = 25°C). This circuit maintained regulation up to I_{OUT} = 56mA (V_{OUT} = 11.4V) at room temperature. However, some performance variance over the entire operating temperature range is to be expected and should be thoroughly explored by the designer. Finally, the load regulation error - in this example - is approximately 1%.(Note: all scope photos shown in this document were taken using a Tektronix TDS3034B; a Tektronix TCP202 current probe was used for measuring inductor current).

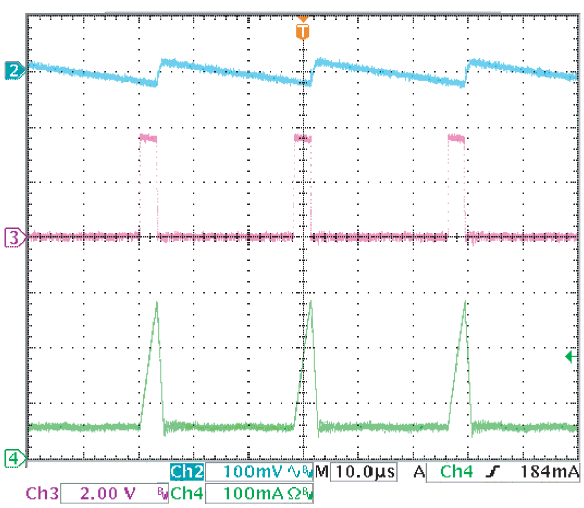

FIGURE 10 – LX1741 WAVE FORMS I_{LOAD} = 5mA
V_{IN} = 3.6V, V_{OUT} = 11.8V, I_{PEAK} = 238mA, η = 86%, V_{RIPPLE} < 50mV (C_{OUT} = 4.7µF)

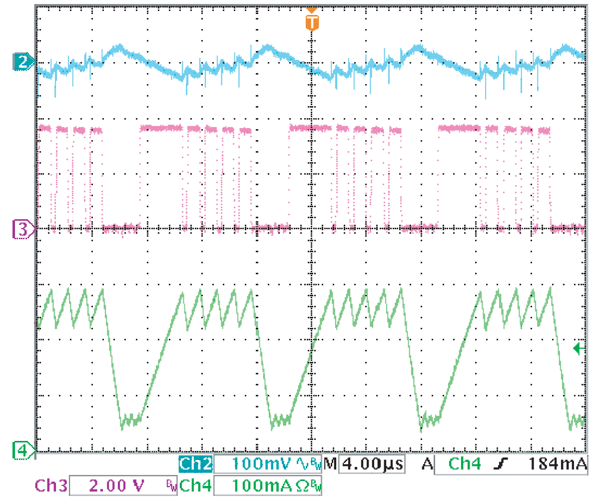

FIGURE 11 - LX1741 WAVE FORMS I_{LOAD} = 40mA
V_{IN} = 3.6V, V_{OUT} = 11.6V, I_{PEAK} = 238mA, η = 90%, V_{RIPPLE} < 100mV (C_{OUT} = 4.7µF)

POWER AND THERMAL CONSIDERATIONS

Designers often examine the maximum output power capability of a DC-DC controller IC. Both LX1741 and LX1742 are available in the 8-pin MSOP package and this package's thermal resistance (Θ_{JA}) is 206°C/W. The device datasheets show a maximum (ambient) junction temperature of (70°C) 150°C. However, at 150°C, degradation to the internal voltage reference bandgap circuit will preclude maintaining optimum output voltage regulation. Moreover, product life-time is reduced when operating at such a high junction temperature. Hence, for practical design considerations, we'll estimate maximum power dissipation using a junction temperature value of 75°C and ambient operating condition of 30°C. Equation 9 describes total power dissipation as a function of maximum junction temperature, ambient temperature, and thermal resistance.

Eq 9

$$P_D = \frac{\left(T_{J(max)} - T_A\right)}{R_{\Theta JA}}$$

Therefore, with respect to these operating conditions, the maximum power dissipation for the LX1741 or LX1742 is calculated to be:

$$P_D = \frac{\left(75°C - 30°C\right)}{206°C/W} = 0.22W$$

Reducing the ambient operating temperature will allow the controller to dissipate more power according to the relationship specified in equation 9.

Reducing the junction temperature improves overall reliability and product life. Equation 10 provides an application specific estimate of the controller's power dissipation where I_Q is the device quiescent current, I_{SRC} is the current through the external FET's source, R_{SRC} is the internal current sense resistor, D is the duty cycle, f_{SW} is the maximum switching frequency and, Qg is the gate charge of the external FET:

Eq 10

$$P_{D(IC)} = \left(V_{IN} \times I_{Q(MAX)} + I_{PEAK}{}^2 \times R_{SRC} \times D + f_{SW} \times V_{IN} \times Q_g\right)$$

Eq 10.1

$$f_{SW} = \left(t_{OFF}\left(\frac{V_{OUT}}{V_{IN}}\right)\right)^{-1}$$

In this design example, V_{IN} = 3.6V and the maximum quiescent current (I_Q) from the LX1741 datasheet is 100mA. The internal current sense resistor's value is 200mΩ (typical). Duty cycle is an estimate and should be maintained to within 85% under full load. Switching frequency is estimated using equation 10.1 where the converter's off-time is typically 300ns. Here, the gate charge is associated with the external MOSFET (e.g., the FDV303N lists a maximum gate charge of 2.3nC). Hence, the LX1741 maximum power dissipation for this example is estimated as:

$$P_{D(IC)} = \begin{pmatrix} 3.6V \times 100\mu A + (235mA)^2 \times 200m\Omega \times \\ 0.85 + 300kHz \times 3.6V \times 2nC \end{pmatrix} = 11.8mW$$

The estimated power dissipation of the IC controller in this application is less than the power value calculated using equation 9 at T_A equals 30°C; therefore, it is safe to proceed with the design.

Equation 11 provides an estimate of the LX1742's controller's power dissipation. Here, the designer must consider the $R_{DS(ON)}$ and gate charge of the internal MOSFET device.

Eq 11

$$P_{D(IC)} = \begin{pmatrix} V_{IN} \times I_{Q(MAX)} + I_{PEAK}{}^2 \times (R_{DS(ON)} + R_{SRC}) \\ \times D + f_{SW} \times V_{IN} \times Q_g \end{pmatrix}$$

DESIGN TOOLS

After reading all this a designer might think, "so many formulas, so little time?" Fortunately, help is on the way. A simple Excel™ spreadsheet is available at our website to help you quickly assess the impact of varying the design parameters for a particular application (i.e., refer to Article 1310 under Application Notes: LX1741 / 1742 Formula Calculator for AN22). The spreadsheet contains two sheets titled 1741 and 1742. The calculator allows the designer to input values for output voltage, output current, input voltage, output capacitance, inductance, output voltage selection resistor R2, and the current limit resistor R_{CS}. The calculator returns values for estimating output voltage ripple (as a function of droop and overshoot), the output voltage selection resistor R1, the peak current, and the output power. The value of peak current is also calculated at R_{CS} = 0Ω for reference. Now some words of advice: the validity of calculator's output is dependent upon the validity of the input data. Therefore, here are some guidelines for selecting input values.

1. The value of R2 should be set so as to minimize error at the VFB input due to offset currents. A value range between 45KΩ and 90KΩ will suffice for most applications.

2. The Inductor (L) value of 47µH is presented as a starting point for most LX1741 and LX1742 application circuits. Remember that selecting the inductor value requires making trade-offs. For example, the inductance value should be sufficient to ensure proper energy storage under worst-case input voltage and on/off-time conditions. Further, the inductor core must not go into saturation. Second, the designer should minimize the device's DC resistance to reduce power loss (thus improving overall efficiency). System-level EMI, cost, and mechanical size are other factors that influence inductor selection criteria. For LX741 and LX1742 designs, inductance values from 20µH to 100µH will support a broad range of applications. Note that small inductor values tend to increase peak current variance due to deviations in the mean value of the comparator delay (t_D).

3. The value of the current limit resistor (R_{CS}) directly affects the value of peak current. The LX1741 and LX1742 have an absolute maximum switch current rating of 800mA_{RMS} and 500mA_{RMS} respectively. Do not exceed these values. Always use the smallest current limit resistor value that your design can tolerate. Setting the peak current excessively high burns away power and reduces overall efficiency (and battery life!). The LX1741 and LX1742 are designed to support applications that have an output requirement of less than 1.5W. Use this as your guideline.

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4. Input voltage is simple. Do not exceed 6.0V. Start-up is guaranteed at 1.6V for very light loads.
5. Cost and the output voltage ripple essentially define the output capacitor type and value. These two constraints will set the calculator's limits (see Appendix).

CONCLUSION

The LX1741 and LX1742 PFM boost-mode controller ICs offer designers a broad range of application solutions. This application note provided a step-by-step design approach for determining critical circuit values such as R1, R2, R_{CS}, L, I_{PEAK}, and C_{OUT}. Moreover, relationship curves for R_{CS}, I_{PEAK}, ΔV_{DROOP}, and $\Delta V_{OVERSHOOT}$ versus Input voltage were provided to aid in the overall understanding of controller performance. Additional device and application information is available from the LX1741 an LX1742 device datasheets available at www.microsemi.com.

Parameter	Units	Typical	Notes
I_{MIN}	mA	**104.0**	from datasheet
I_{SCALE}	A/kΩ	**2.23E-02**	from datasheet
I_{OUT}	mA	**10.0**	Application Parameter
t_D	ns	**773.0**	from datasheet
V_{IN}	V	**3.60**	Application Parameter
V_{OUT}	V	**15.00**	Application Parameter
R_2	Ω	***49900***	Application Parameter
R_{CS}	Ω	**4020**	Application Parameter
L	µH	**47.0**	Application Parameter
C_{OUT}	µF	**20.0**	Application Parameter
V_{REF}	V	**1.200**	from datasheet
P_{OUT}	=	150	mW
R_1	=	573.9	KΩ
I_{PEAK}	=	252.9	mA
$I_{PEAK\ (Rcs=0)}$	=	*163.2*	*mA*
ΔV_{DROOP}	=	2	mV
$\Delta V_{OVERSHOOT}$	=	6	mV
ΔV_{RIPPLE}	=	18	mV

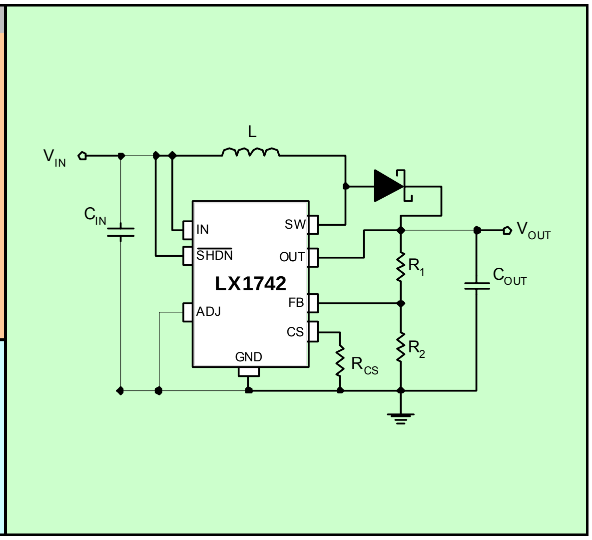

FIGURE 12 – LX1742 DESIGN CALCULATOR SPREADSHEET

LX1741 SEPIC

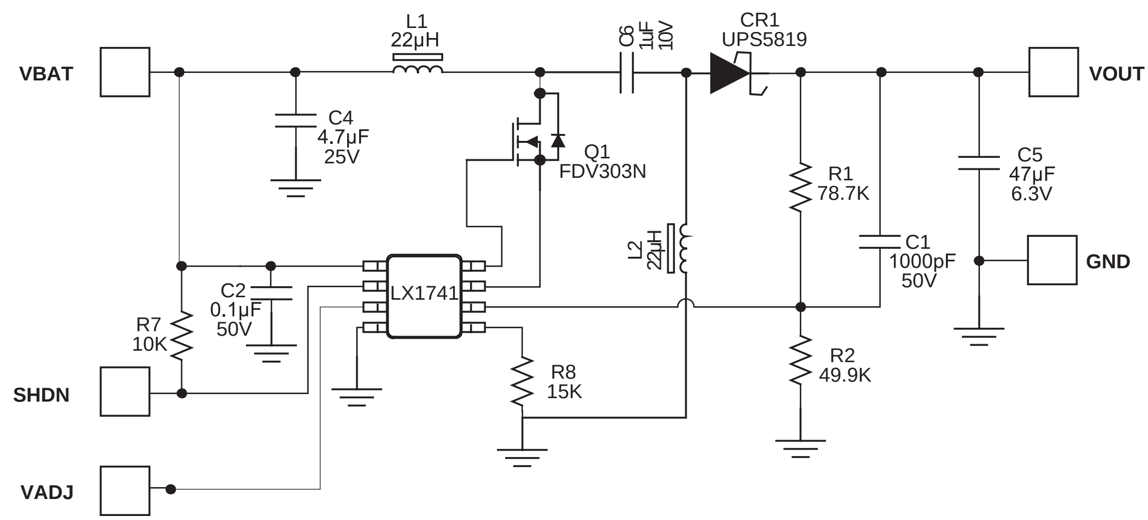

FIGURE 13 – LX1741 SEPIC: TWO-CELL TO 3.3V
(see DN-099A for more information)

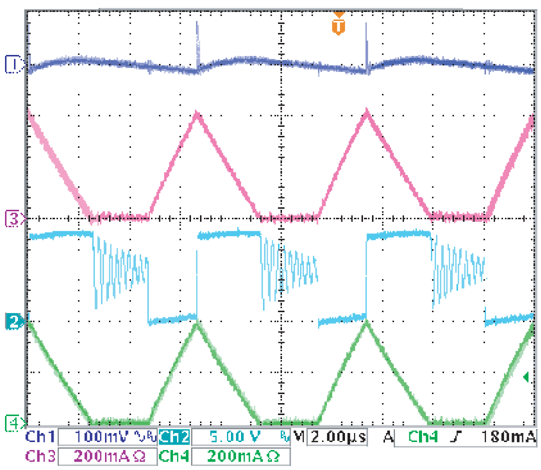

FIGURE 14 – SWITCHING WAVEFORMS
(Configuration: V_{IN} = 5V, V_{OUT} = 3.3V, I_{OUT} = 150mA)Channel 1: V_{OUT} (AC coupled; 100mV/div)Channel 2: Switch voltage (DC coupled; 5V/div)Channel 3: L1 Inductor Current (200mA/div.)Channel 4: L2 Inductor Current (200mA/div.)

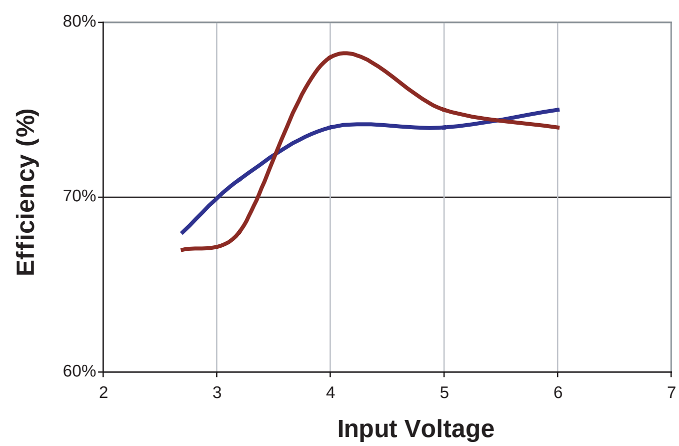

FIGURE 15 – EFFICIENCY VS. INPUT VOLTAGE
(I_{OUT} = 100mA & 200mA)

LX1742

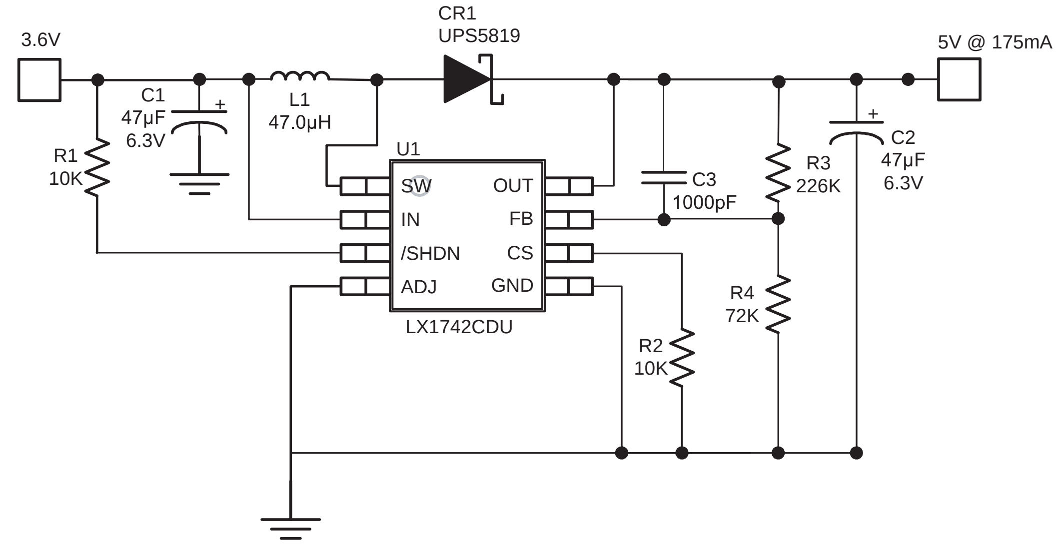

FIGURE 16 – LX1742 3.6V TO 5V BOOST (175mA)
(see DN-099B for more information)

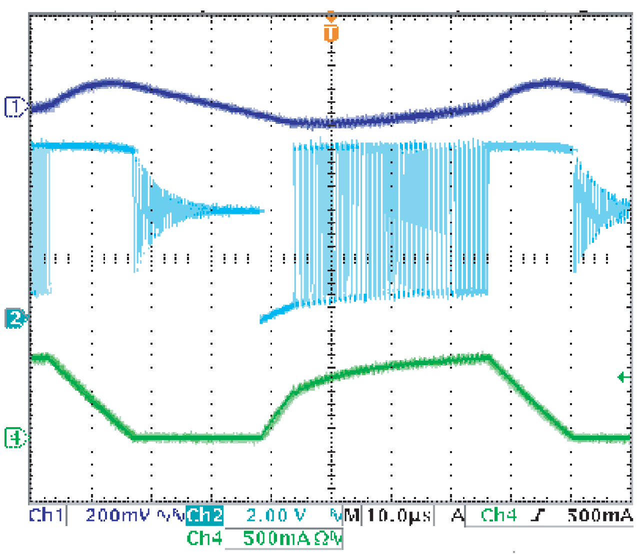

FIGURE 17 – SWITCHING WAVEFORMS

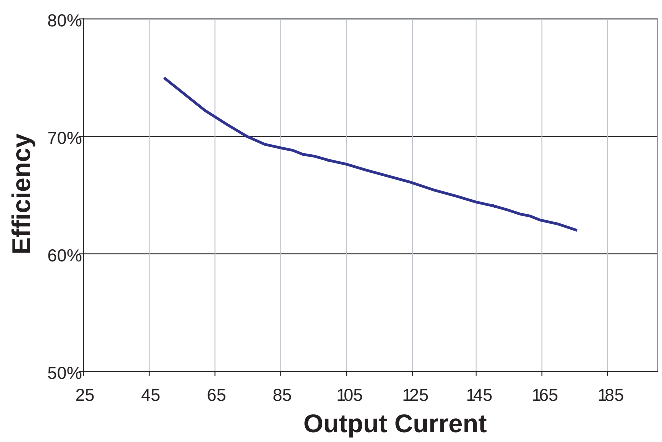

FIGURE 18 – EFFICIENCY VS. OUTPUT CURRENT (mA)

LX1742 OUTPUT DISCONNECT

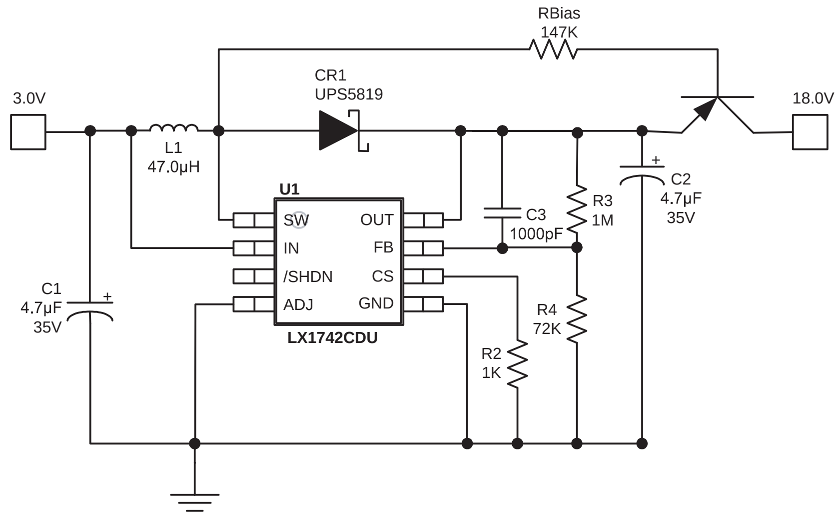

FIGURE 19 – LX7142 WITH OUTPUT DISCONNECT (2N3906)

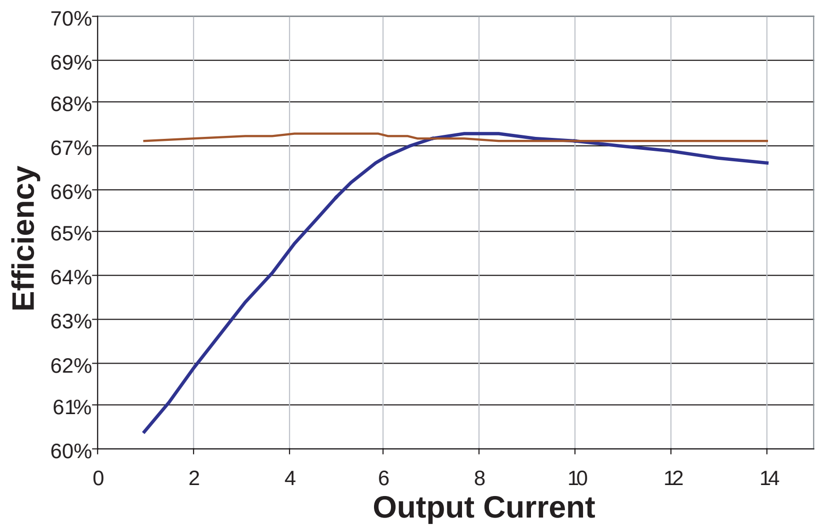

FIGURE 20 – EFFICIENCY VS. OUTPUT CURRENT (mA)

References

Microsemi (2000). LX1741: High efficiency, high voltage boost controller. [Data Sheet], Garden Grove, CA. Author
Microsemi (2000). LX1742: High efficiency, high voltage boost controller. [Data Sheet], Garden Grove, CA. Author

Appendix

A brief summary of the various capacitor types is provided below for the novice designer:

CERAMIC: Multi-layer ceramic capacitors are intended for applications that require a device with a small physical size yet comparatively large electrical capacitance and high insulation resistance. The general-purpose ceramic capacitors, (while not intended for precision applications) are suitable for use as bypass and filtering applications in high frequency circuits where significant changes in capacitance, induced by temperature variation, can be tolerated.

TANTALUM: There are three fundamental types of tantalum capacitors (tantalum foil, wet sintered anode, and solid electrolyte). The designer usually selects tantalum foil capacitors when high voltage components are required or when a substantial reverse voltage is applied to the capacitor (e.g., as in switched mode DC-DC conversion circuits). Wet sintered anode capacitors are often used when low DC leakage is required. Finally, solid electrolyte tantalum capacitors are preferred for their small size versus a given unit of capacitance.

ALUMINUM ELECTROLYTIC: Aluminum electrolytic capacitors are typically preferred for signal filtering and bypass applications when large capacitance values are required and limited board space is available.

FILM: Film capacitors are separated into either film/foil capacitor and metal-film capacitor categories. The film/foil capacitor is characterized by having a high insulation resistance and both excellent current carrying and pulse handling capability. Moreover, these device types are known to provide superior capacitance stability. The metal-film capacitor features high volume efficiency and self-healing properties.

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