LINFINITY *Application Note*

AN-10

Design Procedure for Microprocessor Buck Regulators

LINFINITY
MICROELECTRONICS

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INTRODUCTION

This document is intended to help designers of computer motherboards and other circuits using Linfinity's LX166x family of buck regulator controllers. Linfinity's family includes adjustable and programmable controllers, some with additional linear regulator functions.

BUCK REGULATOR

A buck (step-down) switching regulator is commonly used in applications such as powering microprocessors. They are ideal for converting a 5V-system voltage to the 2V or so, at up to 20A, that processors require.

The main advantages of a buck regulator are high efficiency; relatively simple design; no transformer; low switch stress and small output filter. The main disadvantage is possible over voltage if the main switch shorts.

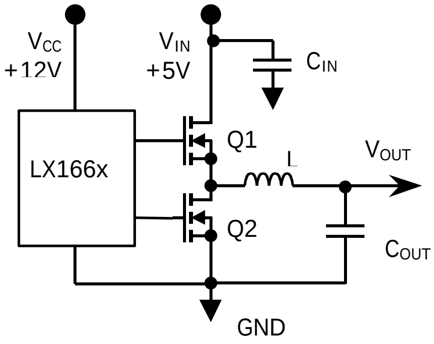

Figure 1: Synchronous Buck Regulator

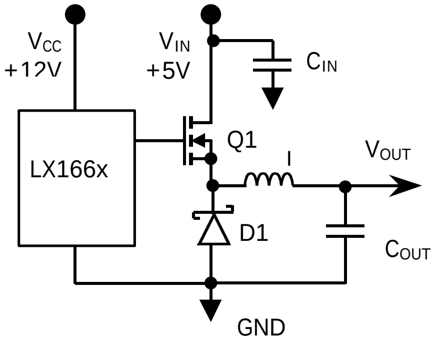

Figure 2: Non-Synchronous Buck Regulator

Figure 1 shows a typical synchronous buck regulator. The current through the inductor is always continuous. When the upper MOSFET, Q1, is on, power is passed to the output through the inductor, L. After Q1 is switched off, there is some deadtime before Q2 turns on. Since the inductor current is continuous, current flows through the body diode of Q2 during the deadtime. Typical waveforms are shown in Figure 14 on page 9.

An alternate configuration is the non-synchronous buck regulator, shown in Figure 2, where the lower MOSFET is replaced by a free-wheeling diode. This usually results in lower efficiency, particularly at low output voltages.

The LX166x family devices are designed to operate in either synchronous or non-synchronous modes. See datasheets for details.

MODULATED CONSTANT OFF-TIME ARCHITECTURE

Switching Frequency

The Modulated Constant Off-Time Architecture is described fully in AN-9. The off-time is kept constant under normal conditions, but is modulated as a function of the output and input voltage in order to keep the operating frequency constant. The architecture is shown in Figure 3.

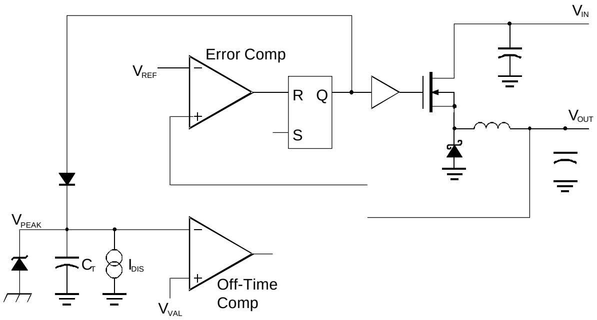

Figure 3: Modulated Constant Off-Time Architecture

Duty Cycle

As with any buck regulator, the inductor volt-second balance conditions must hold.

$$(V_{IN} - V_{OUT}) \times T_{ON} = (V_{OUT} \times T_{OFF})$$

Where V_{IN} is the input voltage (to upper MOSFET drain), V_{OUT} is the output voltage, T_{ON} is the on-time (of the upper MOSFET) and T_{OFF} is the off-time of the upper MOSFET.

The duty cycle, D, equals the ratio of output to input voltage.

$$D = V_{OUT} / V_{IN}$$

The off-time for a particular switching frequency, f_{SW}, is

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$$T_{OFF} = (1 - D)/f_{SW}$$

The switching frequency of the LX1668/1669 is designed to be fixed at 200kHz in steady state conditions.

$$T_{OFF} = 5\mu s \times (1 - V_{OUT}/V_{CC5})$$

Where V_{CC5} is the 5V supply voltage to the IC

Switching Frequency Calculation (LX1660 – 65)
The switching frequency of the LX1660 – 65 devices is controlled differently, and can be calculated using the following equation.

$$f_{SW} = \frac{(1 - V_{OUT}/V_{IN}) \times I_{DIS}}{C_T \times (1.52 - 0.29 \times V_{OUT})} \quad \ldots\ldots(1)$$

Where I_{DIS} is fixed at 200µA and C_T is the timing capacitor. For a 5V input, this can be simplified to:

$$f_{SW} = \frac{0.621 \times I_{DIS}}{C_T} \quad \ldots\ldots(2)$$

INDUCTOR SELECTION

A microprocessor such as the Pentium® II processor requires the power supply to be able to supply a very rapid step demand in current (20 – 30A/µs) as the processor comes out of a stop grant or other low activity state. The inductor value is the main factor in determining how fast the current will increase.

A lower value inductor will increase the ripple current and so require lower ESR (Equivalent Series Resistance) capacitors on the output, but will allow a much faster current change. Selection of the inductor value is a compromise between reducing ripple current, I_{RIPPLE} and improving response time.

An inductor cannot change current instantly. The voltage across an inductor is the product of the inductance and rate of change of current:

$$V = L \times dI/dt \quad \ldots\ldots(3)$$

The inductance required to get a specific response time, T_R, to a step load current change of ΔI can be approximated by:

$$L = (V_{IN} - V_{OUT}) \times T_R / \Delta I \quad \ldots\ldots(4)$$

The peak-to-peak output ripple current is given by the following formula:

$$I_{RIPPLE} = \frac{(V_{IN} - V_{OUT})}{f_{SW} \times L} \times \frac{V_{OUT}}{V_{IN}} \quad \ldots\ldots(5)$$

The input current is a square wave, whose rms value is:

$$I_{INPUT} = I_{OUT} \sqrt{D(1 - D)} \quad \ldots\ldots(6)$$

Where I_{OUT} is the output current.

The ripple current through the input capacitors will be a function of the impedance of the capacitors (ESR) and the impedance of an input inductor. Usually the capacitors will be much lower impedance, so the capacitors will have to handle a large ripple current. This can result in heating of the capacitor and can be a reliability concern if the ripple current rating of the capacitor is exceeded.

The effect of different inductors is shown in Figure 4 – a lower inductance results in faster transient response but greater ripple.

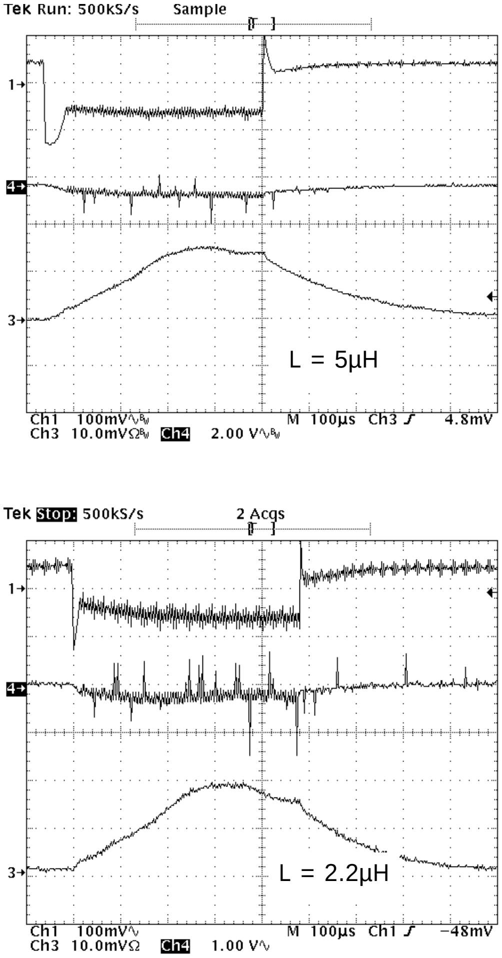


Figure 4: Effect of Different Inductor Values.
Trace 1 = Output voltage; Trace 3 = Input Current;
Trace 4 = Input Ripple Voltage.

A small input inductor (~1μH) can be used to reduce ripple and noise which might affect other 5V blocks.

FILTER CAPACITORS

The output capacitors serve to filter the output voltage. Although a certain amount of bulk capacitance is required, the primary parameter of concern when selecting capacitors is ESR. A model of a capacitor is shown in Figure 5.

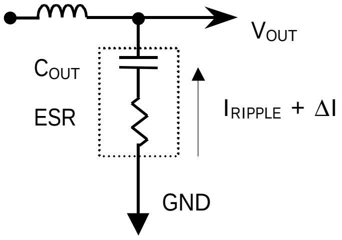


Figure 5: Equivalent Series Resistance

Equation (3) shows that during a heavy load transient, an inductor cannot respond instantaneously. The transient current step, ΔI, has to be provided by the capacitor. The current flowing through the ESR of the capacitor causes a voltage droop whose worst case magnitude is (I_{RIPPLE} + ΔI) × ESR. It is important that the voltage droop does not exceed the dynamic voltage specification, V_{DYN-}, of the processor manufacturer. The requirement for ESR is as follows:

$$V_{DYN-} \geq \left(I_{RIPPLE} + \Delta I\right) \times ESR \quad \text{..........................} \quad (7)$$

Where V_{DYN-} is the lower limit of the dynamic voltage tolerance (usually 100mV for under 2μs).

ADAPTIVE VOLTAGE POSITIONING

The LX166x family (except LX1660) incorporates a 40mV offset into the regulation feedback loop in order to help transient performance. This is shown in Figure 7. The controller regulates point 'A' – at no load, the output will have a **peak output** 40mV above the set point voltage. As current increases, the output voltage will fall, due to the voltage drop in the sense resistor. The benefit of adaptive voltage positioning is increased margin to deal with transient voltage undershoots as shown in Figure 6. Note that the LX166x series uses peak voltage detection, so the dc voltage offset, V_{OFFSET} (as measured with a digital volt meter) will be approximately 25mV instead of 40mV (see Figure 6).

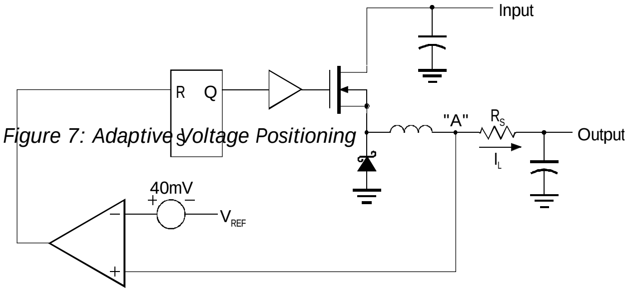
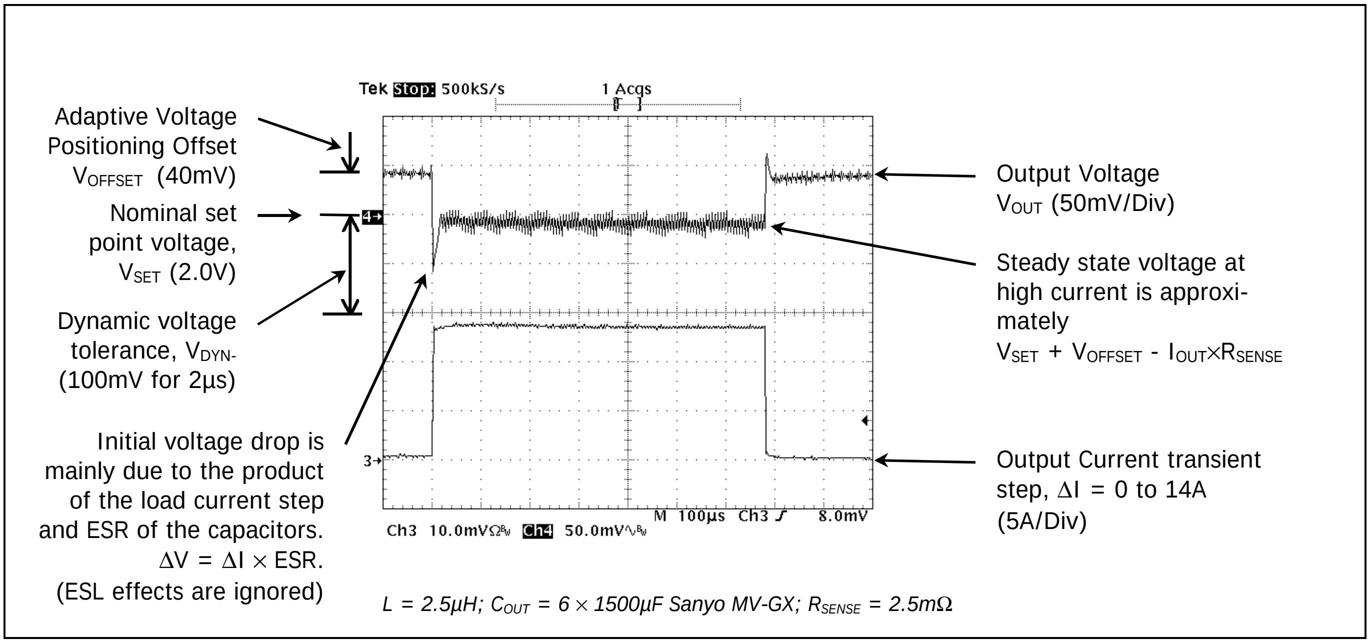


Figure 7: Adaptive Voltage Positioning



Adaptive Voltage Positioning Offset V_{OFFSET} (40mV)

Nominal set point voltage, V_{SET} (2.0V)

Dynamic voltage tolerance, V_{DYN-} (100mV for 2μs)

Initial voltage drop is mainly due to the product of the load current step and ESR of the capacitors. ΔV = ΔI × ESR. (ESL effects are ignored)

Output Voltage V_{OUT} (50mV/Div)

Steady state voltage at high current is approximately $V_{SET} + V_{OFFSET} - I_{OUT} \times R_{SENSE}$

Output Current transient step, ΔI = 0 to 14A (5A/Div)

L = 2.5μH; C_{OUT} = 6 × 1500μF Sanyo MV-GX; R_{SENSE} = 2.5mΩ

Figure 6: Transient Response with Adaptive Voltage Positioning

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The dc offset voltage, V_{OFFSET}, modifies equation (7) as shown below:

$$V_{DYN-} + V_{OFFSET} \geq \left(I_{RIPPLE} + \Delta I\right) \times ESR \quad ....(8)$$

Controllers without adaptive voltage positioning (V_{OFFSET} = 0) will require a lower ESR (i.e. extra capacitors). See design calculation 7 on page 11.

CURRENT LIMIT

The LX1662 – 65 have a current limit function which will hold the current to a maximum limit when the current sense comparator detects an over-current situation. The LX1668/69 have the additional protection of hiccup-mode current limit, whereby the controller goes into low duty-cycle operation in an over-current situation and can reset itself as soon as the short-circuit is removed. Please see application note AN-8 for further details.

Sensing the voltage drop caused by the output current through a resistive element performs current sensing – this can be a sense resistor or the parasitic resistance of the inductor.

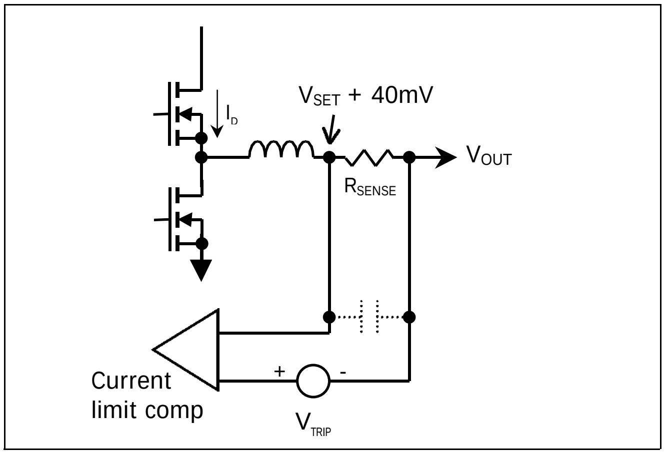

Figure 8: Current Limit Circuit

The sense resistor should be chosen so that the current limit level is not excessively high, nor so low that it interferes with normal operation. The sense resistance should also be as small as possible to reduce voltage droop across it (so that the output voltage does not fall below static voltage limits and also to help reduce power losses).

Current limiting will occur when the output current exceeds the current limit level, I_{CL}:

$$I_{OUT} \geq I_{CL} = V_{TRIP} / R_{SENSE} \quad .... (9)$$

Where V_{TRIP} is the comparator trip voltage.

At higher current levels, such as those demanded by the Pentium II processor, it is desirable to have a lower R_{SENSE} to minimize voltage droop. However, a lower sense resistor will result in a higher over-current trip point, unless the comparator trip voltage is also lowered. The LX1662A – 65A and the LX1668/69 have a 60mV comparator voltage, whereas the LX1660 – 65 have a 100mV voltage. This is shown in Table 1.

Table 1 : Current Sense Comparator Trip Voltages

Device	V_{TRIP}	Application
LX1660/61	100mV	Pentium, PowerPC
LX1662/63/64/65	100mV	Pentium
LX1662A/63A/64A/65A	60mV	Pentium II
LX1668/69	60mV	Pentium II

Sense Resistor

The sense resistor method can use a surface mount power sense resistor; a PCB trace resistance or the parasitic resistance of the inductor (for details of this method, please see application note AN-7). The three alternatives are contrasted in Table 2.

Table 2: Current Sense Resistor Elements

Method	Advantages	Disadvantages
Surface mount power resistor	■ Highly accurate ■ Exposure to air to dissipate heat	■ Expensive ■ Heat dissipation ■ Few values <5mΩ
PCB trace resistor	■ Low cost ■ Flexible resistance value	■ Heat dissipation ■ Accuracy ~20%
Parasitic inductance of the inductor	■ Low cost ■ Low heat dissipation ■ Small space	■ Accuracy ~20% ■ Dependent on type of inductor

A PCB trace resistor can be constructed, as shown in Figure 9. By attaching directly to the relatively large pads for the capacitor and inductor, heat is dissipated more effectively by the larger copper masses. Connect the sense lines as Kelvin connections, to avoid any errors in measurement. Recommended PCB trace dimensions are given in Table 3.

An alternate method for current sensing uses the $R_{DS(ON)}$ of the upper MOSFET. This is much less accurate since $R_{DS(ON)}$ can vary 50 - 100% with temperature. This method also relies on peak current sensing, and so is inflexible for different output voltages.

Table 3 PCB Resistor Dimensions (for 30°C rise at 20A)

Copper Weight	Copper Thickness	Resistor Value	Dimensions (w x l)	
			mm	inches

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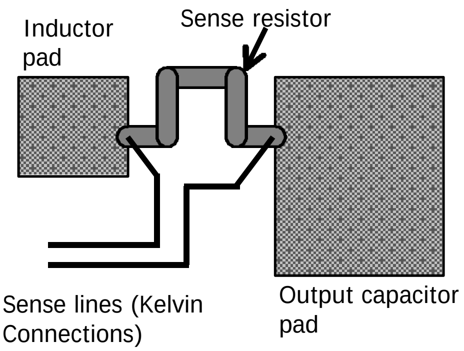

Figure 9 : PCB Trace Sense Resistor Construction

2oz/ft²	68µm	2.5mΩ	2.5 x 22	0.1 x 0.85
		5mΩ	2.5 x 43	0.1 x 1.70

THERMAL ISSUES

Management of heat becomes increasingly important at higher power levels. The following are important factors in managing heat:

1. Airflow – the thermal performance of heatsinks, and even PCB copper is affected greatly by airflow (or lack of).
2. Board layout - neighboring components (such as the processor cartridge) can prevent or reduce significantly the air flow around the voltage regulator. They can also contribute to heat generation – the processor is a significant heat source.
3. Copper and PCB material – outside and thicker copper layers will be better able to dissipate heat.
4. Component selection – a synchronous buck regulator is normally more efficient than a non-synchronous regulator and so creates less heat. Using MOSFET's with lower $R_{DS(ON)}$ will lower heat generation.
5. Ambient temperature – a higher ambient temperature will be reflected in higher silicon junction temperatures (potentially lower reliability).

Thermal Model of a Semiconductor

A power semiconductor device, such as a MOSFET, can be modeled as shown in Figure 10. The top block is the semiconductor junction – the source of heat in the device. Heat generated in the device is dissipated to ambient through a series of thermal resistances and blocks as shown. Its connection to the case can be represented by a resistor, R1. The unit of thermal resistance, Rθ, is °C/W, meaning that for every watt dissipated, there will be a temperature rise of Rθ °C. Values of thermal resistance are specified in datasheets for semiconductors and heatsinks.

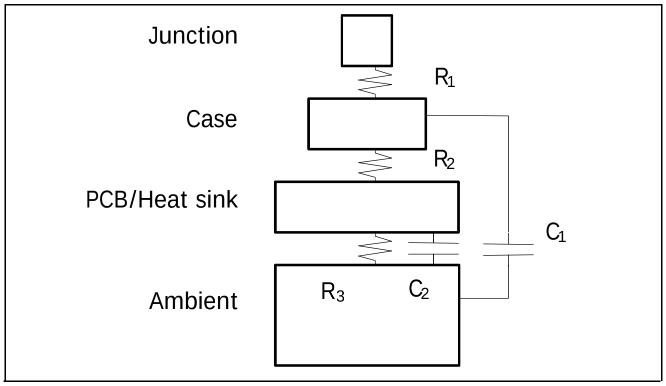

Figure 10: Thermal Model of a Power Semiconductor

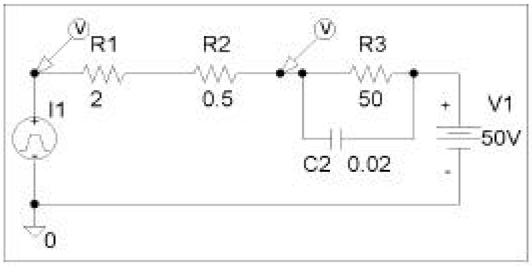

Figure 11: Electrical Equivalent of the Thermal Model

The temperature can be calculated using the electrical equivalent model shown in Figure 11. The counterpart of the temperature in electrical model is the voltage and the heat power in the thermal model is equivalent to a current source in the electrical model.

Figure 11 also gives the typical values for the thermal resistors, where R3 = 50Ω (the unit of thermal resistance is °C/W) is the thermal resistance of the PCB with one square inch of copper. C_1 is neglected and C2 is selected to be 0.02 because the thermal time constant of the heat sink is in the order of one second. The ambient temperature is usually constant therefore is represented by a voltage source V1. When a steady heat is generated in the junction, the junction temperature can be found as,

$$T_j = P_h \cdot \sum_{i=1}^{3} R\theta_i + T_A \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \tag{10}$$

where T_j is the junction temperature, P_h is the heat power, and T_A is the ambient temperature. For example, if 1W heat is generated in the junction, the ambient temperature is 50°C, then the temperature at the junction is T_j = 1W×(2+0.5+50)°C/W +50°C = 102.5°C.

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Selecting a Heatsink

From equation (10), the required heat sink to ambient thermal resistance, $R\theta_{SA}$ can be calculated as follows:

$$R\theta_{SA} \le \frac{(T_J - T_A)}{P_D} - (R\theta_{JC} - R\theta_{CS}) \ldots\ldots\ldots\ldots\ldots\ldots (11)$$

$R\theta_{JC}$ (junction-to-case) can be found from the FET or LDO manufacturer's datasheet. $R\theta_{CS}$, the case-to-heatsink thermal resistance, depends on the mounting method to attach the IC to the heatsink but is usually 0.5 – 1°C/W. T_A is normally assumed to be around 55°C for most computer applications.

Having calculated the required $R\theta_{SA}$, and knowing the airflow and power dissipation, a suitable heatsink can be found in the catalog of a heatsink vendor, such as Aavid, Wakefield or Thermalloy. See the design calculations later in this application note for an example.

Power Dissipated in Upper MOSFET

Power lost in the MOSFET is a combination of the resistive loss due to the $R_{DS(ON)}$ of the FET and switching losses.

$$P_D = (I^2 \cdot R_{DS(ON)} \cdot D) + (0.5 \cdot I \cdot V_{IN} \cdot T_{SW} \cdot f_{SW}) \; (12)$$

Where P_D is power dissipated, D is the duty cycle and T_{SW} is the switching time (~100ns).

Power Dissipated in Lower MOSFET

Power dissipated is due to current flowing through the $R_{DS(ON)}$ of the lower MOSFET.

$$P_D = (I^2 \times R_{DS(ON)}) \times (1 - D) \ldots\ldots\ldots\ldots\ldots\ldots\ldots (13)$$

Care should be taken in the selection of MOSFETs – buying a lower cost FET can result in much higher heat dissipation. It may become necessary to use a heatsink with the FET, so increasing total costs. The lower $R_{DS(ON)}$ FET could be a surface mount device, dissipating heat to the PCB copper.

External Linear Regulator (LX1664/65/68)

The heat generated in the MOSFET used as the regulator's pass element is as follows:

$$P_D = I_{LIN} \times (V_{IN} - V_{OUT}) \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (14)$$

Where I_{LIN} is the current and V_{IN} and V_{OUT} are the drain and source voltages of the linear transistor respectively (i.e. the input and output of the linear regulator).

Internal Linear Regulator (LX1668 only)

The LX1668 has an internal 2.5V fixed LDO. The LDO can have a 3.3V or a 5V power supply, but heat generation is an important consideration – the SOIC or TSSOP package has a high thermal resistance, and cannot dissipate much heat. The heat dissipation is shown in calculation 10 on page 12.

Power Dissipated in Controller IC

Excluding the internal LDO, the power dissipation will be approximately:

$$P_D = I_{OPERATING} \times V_{CC} \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (15)$$

The LX1660 – 65 devices operate from a 12V V_{CC}, whereas the LX1668/69 operates from 5V.

PROGRAMMING THE OUTPUT VOLTAGE

The LX1662 – LX1669 devices have 5 VID inputs which read the 5-bit voltage identification (VID) code to program the output voltage. The VID pins on the LX1662/63/64 and 1665 are not digital compatible – pull up resistors should not be used on the VID bus, or errors may occur.

LOW DROPOUT REGULATORS

The LX1664 and LX1665 have an external linear regulator driver; the LX1668 incorporates a similar linear regulator driver and also has an internal low-power low dropout regulator.

External Linear Regulator Driver

Connecting an external MOSFET to the controller IC can make an adjustable low dropout linear regulator (LDO). The adjust (L_{FB}) pin has a feedback voltage of 1.5V, meaning that no resistors are required for setting the output to 1.5V for GTL+ Bus termination.

The dropout voltage is determined by current and the $R_{DS(ON)}$ of the transistor – for most applications, it is not important to have low $R_{DS(ON)}$.

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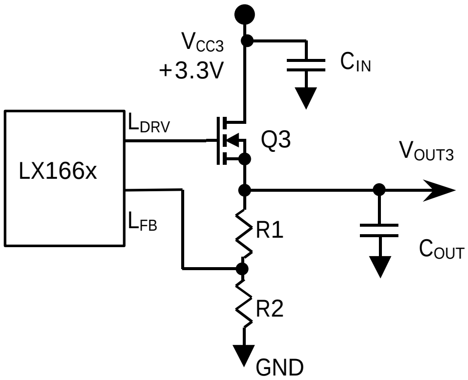

Figure 12: Linear Regulator Driver

The output voltage, V_{OUT3}, is calculated as follows:

$$V_{OUT3} = V_{FB} \times (1 + R1/R2) \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (16)$$

Where V_{FB} is the feedback voltage (1.5V). For V_{OUT3} = 1.5V, R2 = ∞ and R1 can be shorted.

The linear regulator can be disabled by pulling the feedback pin, L_{FB}, up to 3.3V or 5V. See datasheet for details.

Internal Fixed LDO (LX1668 only)

The internal LDO has a fixed output voltage of 2.5V. Although the LDO can handle transient currents as high as 400mA during start-up, normal operation should be limited to 200mA or below.

The LDO can use a 5V or 3.3V input – beware of excessive heat dissipation if using 5V input. See equation (14) on page 7.

LAYOUT CONSIDERATIONS

As with any power device, careful layout is essential. Linfinity's devices are tolerant of noise, but basic precautions should be taken.

Power Traces and Ground Planes

Ensure that power traces on the PCB are as wide as possible, to minimize resistive voltage drops at high currents.

Connect ground points together on a separate plane, as shown in Figure 13.

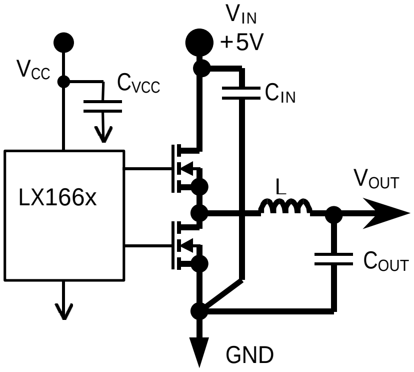

Figure 13: Power Traces

Ensure that the decoupling capacitor, C_{VCC}, in Figure 13 is placed as close to the IC as possible, to isolate the controller from any noise on the V_{CC} rail. Note that in the LX1662 – 65, an under-voltage lockout function can "shut down" the IC during momentary undervoltage situations when the capacitor is too small or too far from the device. Use at least 1µF.

FURTHER INFORMATION

Please see Linfinity's web site at http://www.linfinity.com for the latest datasheets and application notes.

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Figure 14: Typical Buck Regulator Waveforms

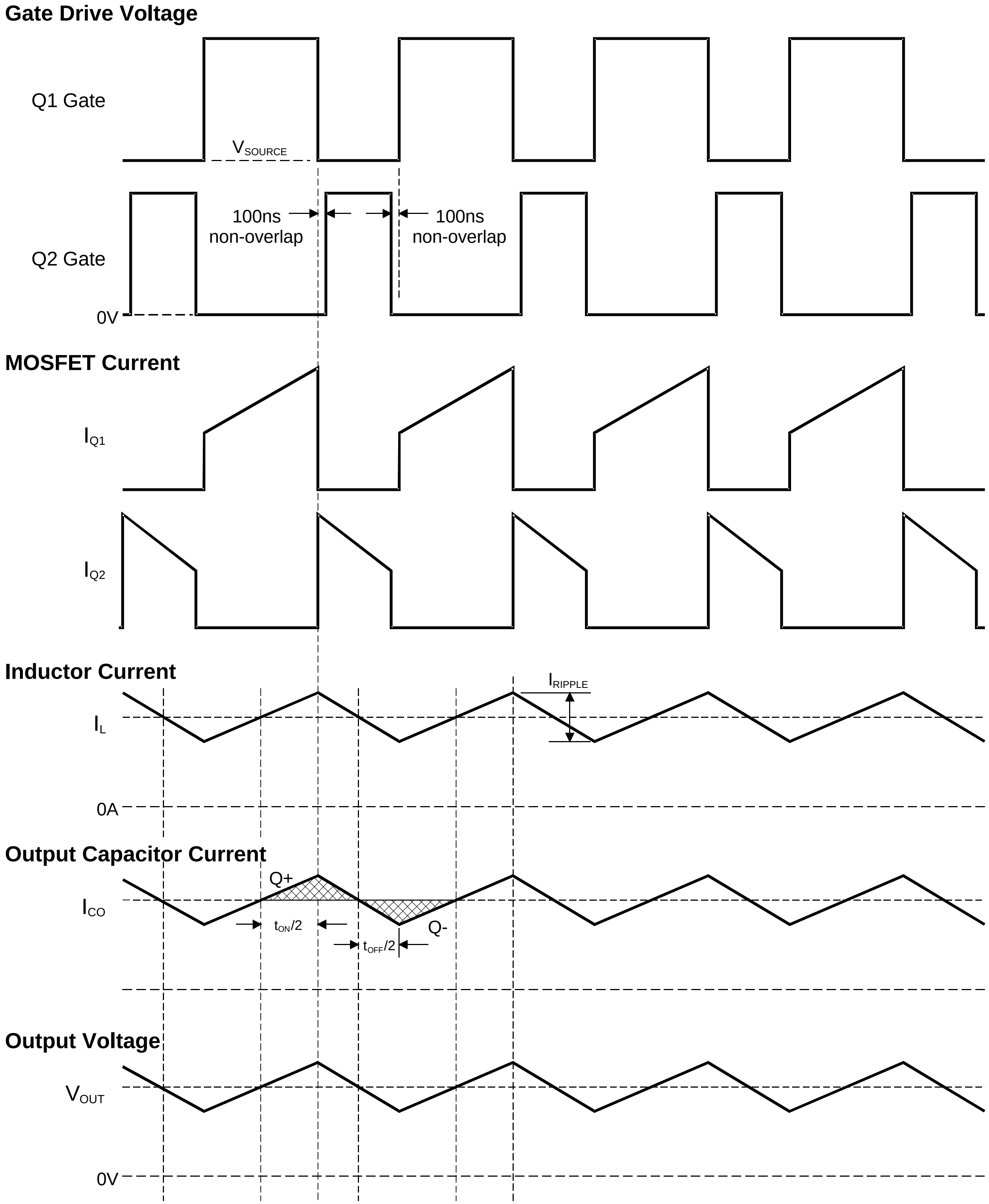

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Table 4 Switching Regulator Selection Guide

Device	Package	5-bit VID	Power Good	OVP Driver	Adaptive Voltage Positioning	Hiccup Mode Current Limit	Synchronous Rectification	External LDO	Internal LDO	Current Sense Threshold (mV)	Production
LX1660	SO-16					•	•			100	Now
LX1661	SO-16				•	•	•			100	Now
LX1662	SO-14	•			•		•			100	Now
LX1662A	SO-14	•			•		•			60	Now
LX1663	SO-16	•	•	•	•		•			100	Now
LX1663A	SO-16	•	•	•	•		•			60	Now
LX1664	SO-16	•			•		•	•		100	Now
LX1664A	SO-16	•			•		•	•		60	Now
LX1665	SO-18	•	•	•	•		•	•		100	Now
LX1665A	SO-18	•	•	•	•		•	•		60	Now
LX1668	SO-20 TSSOP	•(TTL)	•	•	•	•	•	•	•	60	8/98
LX1669	SO-16	•(TTL)	•	•	•	•	•			60	8/98

Design Calculations

This design example is used to calculate the components required for a Pentium® II processor power supply with the following characteristics:

V_{IN}=5V; V_{OUT}=2.0V (programmable, but assumed to be 2.0V for worst-case analysis); $I_{MAX=}$15A maximum steady state load current; $\Delta I_{MAX=}$14A worst case transient load step; $I_{CL=}$20A current limit activation level; $T_{A=}$55°C ambient temperature (with 100 linear ft/min air flow)

1. **Select Controller IC**

Select controller IC from Table 4 on page 10. See also Linfinity Application Note AN-6 *"Power Solutions for Flexible Motherboards"* for reference designs. For a Pentium II supply, select a controller with a 60mV current sense comparator (i.e. LX1662A – 65A or LX1668/69). If GTL+ Bus and clock circuit loads are far from the processor, it may be better to use a single output controller (LX1662/63 or LX1669) with low dropout regulators to power V_{CLOCK} and V_{TT}.

2. **Select Timing Capacitor**

(Not applicable to LX1668/69).

Choose a switching frequency of 200kHz by selecting the appropriate timing capacitor value. From equation (2):

$$C_T = \frac{0.621 \times 200 \cdot 10^{-6}}{200 \cdot 10^{3}} = 621pF$$

3. **Select Inductor**

For most applications, a 12µs inductor response time will give adequate performance. From equation (4):

$$L = \frac{(5-2) \times 12 \cdot 10^{-6}}{14} = 2.57\mu H$$

An inductor in the range of 2.5 to 3µH will give a sufficiently fast transient response. Suitable inductors include the surface mount HM00-97713 and the through-hole version HM00-98637 from BI Technologies.

4. **Input Current**

From equation (6), the input current will be:

$$I_{INPUT} = 15\sqrt{0.4 \times 0.6} = 7.4\text{ A}$$

5. **Output Ripple Current**

From equation (5), the input current will be:

$$I_{RIPPLE} = \frac{(5-2)}{200 \cdot 10^{3} \times 2.5 \cdot 10^{-6}} \times \frac{2}{5} = 2.4\text{ A}$$

6. **Select Input Capacitors**

Select input capacitors based on the input current calculated in the previous step (this will be the absolute worst case capacitor current- see page 3). From the capacitor datasheet, the maximum rating for the Sanyo MV-GX 1500µF capacitor is 2A. Therefore, the design requires 7.4/2, i.e. 3 – 4 capacitors for greatest reliability.

7. **Select Output Filter Capacitors**

The output capacitors have to be selected to meet the ESR specification – see equation (8).

$$ESR \le \frac{0.1 + 0.025}{2.4 + 14} \le 0.0076\Omega$$

Sanyo MV-GX capacitors have a maximum ESR of 44mΩ (at 20°C and 100kHz).

Number of capacitors ≥ 44/7.6 ≥ 6 capacitors.

If a controller without adaptive voltage positioning were used,

$$\text{Number of caps} \ge \frac{0.044}{(0.1+0)/(2.4+14)} \ge 7$$

Adaptive voltage positioning results in the elimination of one capacitor!

8. **Select Current Sense Resistor**

To ensure current limiting does not happen until the current exceeds 20A, R_{SENSE} is selected according to equation (9).

$$R_{SENSE} \le 0.06/20 \le 0.003\Omega$$

We can use a 2.5mΩ resistance.

9. **Construct Resistor**

If using a surface mount sense resistor, the lowest commonly available value is 5mΩ, so use two in parallel.

The lowest cost solution is to construct a sense resistor using a PCB trace. From Table 3, using 2oz/ft²

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copper, suitable dimensions are 2.5mm wide by 22mm long. See Figure 9.

10. Thermal Analysis

Upper MOSFET

Using an IRL3102S ($R_{DS(ON)}$ = 13mΩ), from equation (12), the heat dissipated is:

$$P_D = \left(15^2 \times 0.013 \times 0.4\right) + \left(0.5 \times 15 \times 5 \times 100 \cdot 10^{-9} \times 200 \cdot 10^3\right)$$

$$P_D = 1.17 + 0.75 = 1.92W$$

This can be dissipated using the TO-263 surface mount package soldered to a copper pad for heatsinking.

Lower MOSFET

Using IRL3102S, equation (13) gives the heat as:

$$P_D = \left(15^2 \times 0.013 \times 0.6\right) = 1.755W$$

Again, a surface mount package can be used for this transistor.

If an IRL3303 (26mΩ $R_{DS(ON)}$) is used,

$$P_D = \left(15^2 \times 0.026 \times 0.6\right) = 3.51W$$

This will require the use of a heatsink.

External Linear Regulator

Assuming V_{IN} = 3.3V; V_{OUT} = 1.5V and 3A steady state current, equation (14) gives the heat:

$$P_D = 3 \times (3.3 - 1.5) = 5.4W$$

Internal Linear Regulator

Assuming 200mA steady state current, equation (14) gives the heat:

For 3.3V input:

$$P_D = 0.2 \times (3.3 - 2.5) = 0.16W$$

For 5V input:

$$P_D = 0.2 \times (5 - 2.5) = 0.50W$$

Power Dissipated in Controller IC

LX1668/69 (V_{CC} = 5V), equation (15) gives the heat:

$$P_D = 0.024 \times 5 = 0.12W \text{ (excluding LDO)}$$

LX1660-65 (V_{CC} = 12V), equation (15) gives the heat:

$$P_D = 0.027 \times 12 = 0.324W$$

11. Heat Sink Requirements

Maximum junction temperatures are 150°C for Linfinity LX166x and 175°C for International Rectifier IRL series MOSFET's (see datasheets). Since MTBF decreases with increasing temperature, calculations should ideally use a lower maximum junction temperature such as 125°C. Assume $R\theta_{CS}$ is 0.5°/W. Remember that heatsink performance improves with additional air flow.

Upper MOSFET

The IRL3102S ($R\theta_{JC}$ = 1.4°C/W) could be used surface-mounted with the copper PCB pad as heatsink. If a heatsink is desired, it is specified by equation (11):

$$R\theta_{SA} \le \frac{125 - 55}{1.92} - \left(1.4 + 0.5\right) \le 34.6\ ^\circ\mathrm{C/W}$$

A suitable heatsink would be the Aavid 577002 which has a thermal resistance of 32°C/W.

Lower MOSFET

From equation (11), using IRL3303 ($R\theta_{JC}$ = 2.7°C/W):

$$R\theta_{SA} \le \frac{125 - 55}{3.51} - \left(2.7 + 0.5\right) \le 16.7\ ^\circ\mathrm{C/W}$$

A suitable heatsink would be the Aavid 530613, which has a thermal resistance of 16.7°C/W.

Linear Regulator MOSFET

From equation (11), using IRLZ44N ($R\theta_{JC}$ = 1.4°C/W):

$$R\theta_{SA} \le \frac{125 - 55}{5.4} - \left(1.4 + 0.5\right) \le 11.1\ ^\circ\mathrm{C/W}$$

The 563202 (11.0°C/W will be sufficient to dissipate the heat.

12. Temperature Rise in Controller IC

The LX1668 has two package options, SO-20 and TSSOP-20, with different thermal resistances. The IC will have a different junction temperature rise, depending on the package and whether the internal LDO uses 5V or 3.3V input.

Table 5 shows the temperature rise in the IC for different LDO input voltages and package options. This assumes 200mA steady state current from the LDO, as well as operating current for the IC.

Table 5: Temperature Rise in LX1668

LDO Input Voltage	Total Power Dissipated	Package

		SO-20 85°C/W	TSSOP-20 110°C/W
3.3V	0.28W	23.8	30.8
5.0V	0.62W	52.7	68.2

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