

## *Power Supplies for LED Driving*

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30 Corporate Drive, Suite 400, Burlington, MA 01803, USA  
Linares House, Jordan Hill, Oxford OX2 8DP, UK

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**Library of Congress Cataloging-in-Publication Data**  
Application submitted

**British Library Cataloguing-in-Publication Data**  
A catalogue record for this book is available from the British Library.

ISBN: 978-0-7506-3112-2

For information on all Newnes publications  
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Printed in the United States of America  
08 09 10 11 12 13 10 9 8 7 6 5 4 3 2 1

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In most chapters, I will include a section called ‘Common Errors’. This section will highlight errors that engineers have made, and how these can be avoided, with the hope that readers will not make the same mistakes. It is said that people learn from their mistakes, but it is also true that we can learn from the mistakes of others. Our own mistakes are more memorable, but also more costly!

Usually the first problem for a designer is to choose between different topologies. When is a buck preferred to a buck-boost or a boost? Why is a Cuk boost-buck better than a fly-back type? This book will cover these topics at the beginning of the switching supplies section.

Power supply design equations will be given and example designs of practical supplies will be worked through. With switching power supplies, equations are needed to make the correct component choice; a wrong component can make a poor power supply and require a lot of corrective action. Power LEDs generate a lot of heat in a small area, which makes thermal management difficult; an adjacent power supply should be efficient and not add too much heating effect.

The implications of changing the calculated component values into standard values, which is more practical, will be discussed. In many cases, customers want to use standard off-the-shelf parts, because of ease of purchase and cost. Calculations rarely produce a standard value, so a compromise has to be made. In some cases the difference is negligible. In others it may be better to choose a higher (or lower) value. All component value changes will introduce some ‘error’ in the final result.

Having proven worked examples in the book will help the reader to understand the design process: the order in which the design progresses. It will also show how the calculated component value compares with the actual value used, and will include a description of why the choice was made.

## 1.2 Description of Contents

In Chapter 2, the description of some LED applications will show the breadth of the LED driving subject and how LEDs’ physical characteristics can be used to an advantage. It is also important to understand the characteristics of LEDs in order to understand how to drive them properly. One of the characteristics is colour; an LED emits a very narrow band of wavelengths so the colour is fairly pure. The LED color determines the different voltage drop across the LED while it is conducting, and I will

and the opto-coupler's internal LED, the photo-transistor is conducting. If the string goes open-circuit, there is no current through the opto-coupler's LED and the photo-transistor does not conduct.

### 3.3 Testing LED Drivers

Although testing an LED driver with the actual LED load is necessary, it is wise to use a dummy load first. There are two main reasons for this: (1) cost of an LED, especially high power devices, can be greater than the driver circuit; and (2) operating high brightness LEDs for a long time under test conditions can cause eye strain and temporary sight impairment (if LEDs viewed at close range). A further reason is that some dummy loads can be set to limit the current and so enable fault-finding to be made easier.

#### 3.3.1 Zener Diodes as a Dummy Load

Figure 3.9 shows how Zener diodes can be used as a dummy load. This is the simplest and cheapest load. The 1N5334B is a 3.6 V, 5 W Zener diode (3.6 V typical at 350 mA). This is not the perfect dummy load. This reverse voltage is slightly higher than the typical forward voltage of 3.42 V of a Lumileds 'Luxeon Star' 1 W LED. The 1N5334B has a dynamic impedance of 2.5 ohms, which is higher than the Luxeon Star's 1 ohm impedance. The impedance will have an effect on some switching LED drivers that have a feedback loop. For simple buck circuits, the impedance only has a small effect.

$4 \times 1\text{N}5334\text{B} = 4 \times 1\text{ W LED}$

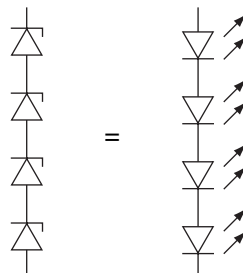


Figure 3.9: Zener Diode Dummy Load.

An active load is more precise. A constant voltage load will have (in theory at least) zero impedance, so simply adding a small value series resistor will give the correct

### 3.5 Conclusions

A voltage regulated LED driver is preferred when there are a number of LED modules that can be connected in parallel. Each module will have its own linear current regulator. An example would be channel lighting, as used in shop name boards.

A current regulated LED driver is preferred when it is desirable to have a number of LEDs connected in series. A series connection ensures that all the LEDs have the same current flowing through them and the light output will be approximately equal.

A switching driver with constant current output is the favored option when driving high power LEDs, for reasons of efficiency. An efficiency of 75–90% can be achieved. If a constant voltage source were used, the LEDs would also need a high current linear regulator in series, which is very inefficient and would increase heat dissipation problems.

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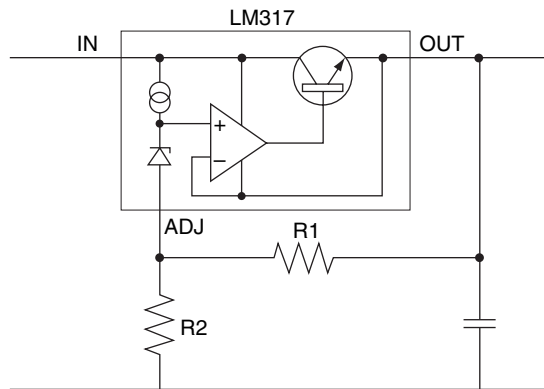


Figure 4.1: LM317 Regulator

A capacitor on the output terminal also helps with stability. The output voltage is given by the equation:

$$V_{OUT} = 1.25 * \frac{1 + R2}{R1} + I_{ADJ} * R2$$

Note,  $I_{ADJ} = 100 \mu\text{A}$ , worst case.

Variations of the LM317 regulator include fixed positive voltage versions (LM78xx) and negative voltage versions (LM79xx), where 'xx' indicates the voltage; i.e. LM7805 is a +5 V 1 A regulator.

The LM317 and its variants need a minimum input to output voltage difference to operate correctly. This is typically in the range 1 V to 3 V, depending on the current through the regulator (higher current requires a higher voltage differential). This input to output voltage difference is equal to the voltage across the internal constant current generator, since the OUT pin is at the same potential as the voltage reference.

Low dropout voltage regulators use a PNP transistor as the power switch, with the emitter connected to the IN terminal and the collector connected to the OUT terminal, see Figure 4.2. They also have a ground pin that enables an internal reference voltage to be generated independent of the input to output voltage differential. A dropout voltage of less than 1 V is possible.

The HV9910B has two current sense threshold voltages – an internally set 250 mV and an external voltage at the LD pin. The actual threshold voltage used during switching will be the lower of the two. The low value of sense voltage allows the use of low resistor values for the current sense, which means high efficiency.

The HV9910B IC operates down to 8 V input, which is required for some automobile applications, and can accept a maximum of 450 V input, which makes it ideal for offline applications. The IC has an internal regulator that supplies 7.5 V to power to the IC's internal circuits from the input voltage, eliminating the need for an external low voltage power supply. The IC is capable of driving the external MOSFET directly, without the need for additional driver circuitry.

## 5.2 Buck Circuits for DC Applications

For DC applications, the schematic shown in Figure 5.3 can be used.

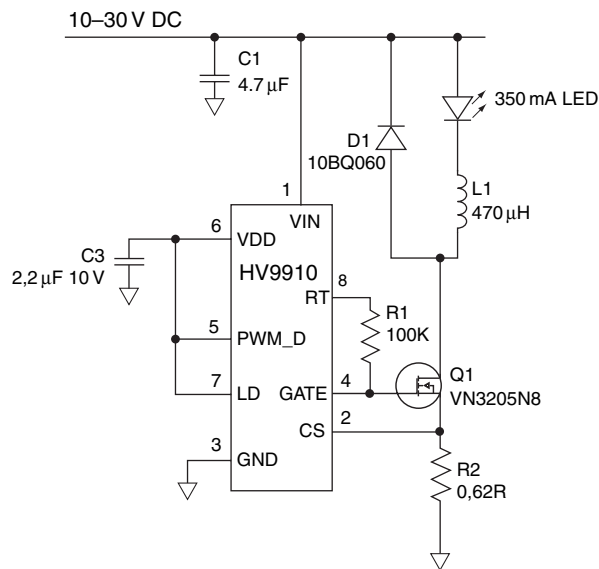


Figure 5.3: Buck Converter for DC Applications.

### 5.2.1 Target Specification

Input voltage = 10 V–30 V

LED string voltage = 4–8 V

LED current = 350 mA

Expected efficiency = 90%

### 5.2.2 Choosing the Switching Frequency and Resistor ( $R_1$ )

The switching frequency determines the size of the inductor  $L_1$ . A larger switching frequency will result in a smaller inductor, but will increase the switching losses in the circuit. A typical switching frequency for low input voltage applications is  $f_s = 150$  kHz, which is a good compromise. From the HV9910B data sheet, the timing resistor between the RT pin and ground that is needed to achieve this frequency is 150 k $\Omega$ .

However, in this case the minimum input voltage is only 80% of the maximum output voltage. In a buck converter, the duty cycle of the MOSFET switch (proportion of the time that the switch is turned on), is given by  $D = \frac{V_{OUT}}{V_{IN}}$  and will also be 80%. However, in continuous conduction mode, instability will result when the duty cycle goes over 50%. To prevent instability, it is necessary to operate in constant off-time mode. This is achieved with the HV9910B circuit by connecting the timing resistor between the RT pin and the gate pin. The timing circuit only charges an internal capacitor when the timing resistor is connected to 0 V; the gate pin is at 0 V when the MOSFET is turned off. Thus the off-time is constant, so the switching frequency varies as the load voltage changes.

If we choose a timing resistor that gives a constant off-time of say 5  $\mu$ s, with an 80% duty cycle the on-time will be 20  $\mu$ s. The switching frequency will be 40 kHz. At the other extreme, with a 30 V supply and a 4 V load, the duty cycle will be just 13.33%, so the on-time will be 767 ns. Now the switching frequency is 173.4 kHz. The average switching frequency will be about 100 kHz, so we can base the selection of other components on this. The timing resistor to give 5  $\mu$ s off-time will be 100 k $\Omega$ .

### 5.2.3 Choosing the Input Capacitor ( $C_1$ )

An electrolytic capacitor is good to hold the voltage, but the large ESR of these capacitors makes it unsuitable to absorb the high frequency ripple current generated by the buck converter. Thus, metallized polypropylene capacitors or ceramic

capacitors in parallel are needed to absorb the high frequency ripple current. The required high frequency capacitance can be computed as

$$C1 = \frac{I_o \times T_{\text{OFF}}}{(0.05 \times V_{\text{min}})}$$

In this design example, the high frequency capacitance required is about 4.7  $\mu\text{F}$  50 V. This capacitor should be located close to the inductor  $L1$  and MOSFET switch  $Q1$ , to keep the high frequency loop current within a small area on the PCB. In practice, two such capacitors with a small inductor between them (to make a Pi filter) are needed to limit EMI emissions.

#### 5.2.4 Choosing the Inductor ( $L1$ )

The inductor value we use depends on the allowed level of ripple current in the LEDs. Assume that 15% ripple (a total of 30%) is acceptable in the LED current.

The familiar equation for an inductor is  $E = L \times \frac{di}{dt}$ . Considering the time when the MOSFET switch is off, so that the inductor is supplying energy to the LEDs,  $E = V_{\text{LED}} = V_{o,\text{max}} = L \times \frac{di}{dt}$ . Another way of writing this is  $L = V_{o,\text{max}} \times \frac{dt}{di}$ . Here,  $di$  is the ripple current  $= 0.3 \times I_{o,\text{max}}$  and  $dt$  is the off-time.

Then, the inductor  $L1$  can be computed at the rectified value of the nominal input voltage as

$$L1 = \frac{V_{o,\text{max}} \times T_{\text{OFF}}}{0.3 \times I_{o,\text{max}}}$$

In this example,  $L1 = 380 \mu\text{H}$  and the nearest standard value is 470  $\mu\text{H}$ . Since this value is a little higher than the calculated value, the ripple current will be less than 30%.

The peak current rating of the inductor will be 350 mA plus 15% ripple:

$$i_p = 0.35 \times 1.15 = 0.4 \text{ A.}$$

The RMS current through the inductor will be the same as the average current (i.e. 350 mA).

through the MOSFET, but the current is limited by the MOSFET current rating, so a lower current rating can reduce the switching losses.

The peak voltage rating of the diode is the same as the MOSFET. Hence,

$$V_{\text{diode}} = V_{\text{FET}} = 562 \text{ V}$$

The average current through the diode is

$$I_{\text{diode}} = 0.5 \times I_{\text{o,max}} = 0.175 \text{ A}$$

Choose a 600 V, 1 A ultra-fast diode. The UF4005 is a low cost ultra-fast type, but for greatest efficiency a faster diode like STTH1R06 should be used. If we assume a forward voltage drop of 1 V at 350 mA, the conduction loss will be less than 350 mW at low duty cycles. The switching loss could be higher than this value, but is less of a problem in faster diodes because the reverse conduction is for a shorter time period.

### 5.3.7 Choosing the Sense Resistor ( $R_2$ )

The sense resistor value is given by

$$R_2 = \frac{0.25}{1.15 \times I_{\text{o,max}}}$$

This is true if the internal voltage threshold of 0.25 V is being used. Otherwise, substitute the voltage at the LD pin instead of the 0.25 V into the equation. A lower voltage could be applied to the LD pin to enable a convenient value of  $R_2$  to be used, as described earlier.

For this design,  $R_2 = 0.625 \Omega$ . The nearest standard value is  $R_2 = 0.62 \Omega$ .

Note that capacitor  $C_3$  is a bypass capacitor for holding up the HV9910B internal supply  $V_{\text{DD}}$  during MOSFET switching, when high frequency current pulses are required for charging the gate. A typical value for  $C_3$  of 2.2  $\mu\text{F}$ , 16 V is recommended, although for AC applications smaller capacitors as low as 0.1  $\mu\text{F}$  have been used successfully. The switching frequency tends to be lower and so the MOSFET gate current requirements are low. Also with a higher voltage on the input supply pin, the voltage drop across the internal regulator during MOSFET switching is unlikely to cause under-voltage drop out.

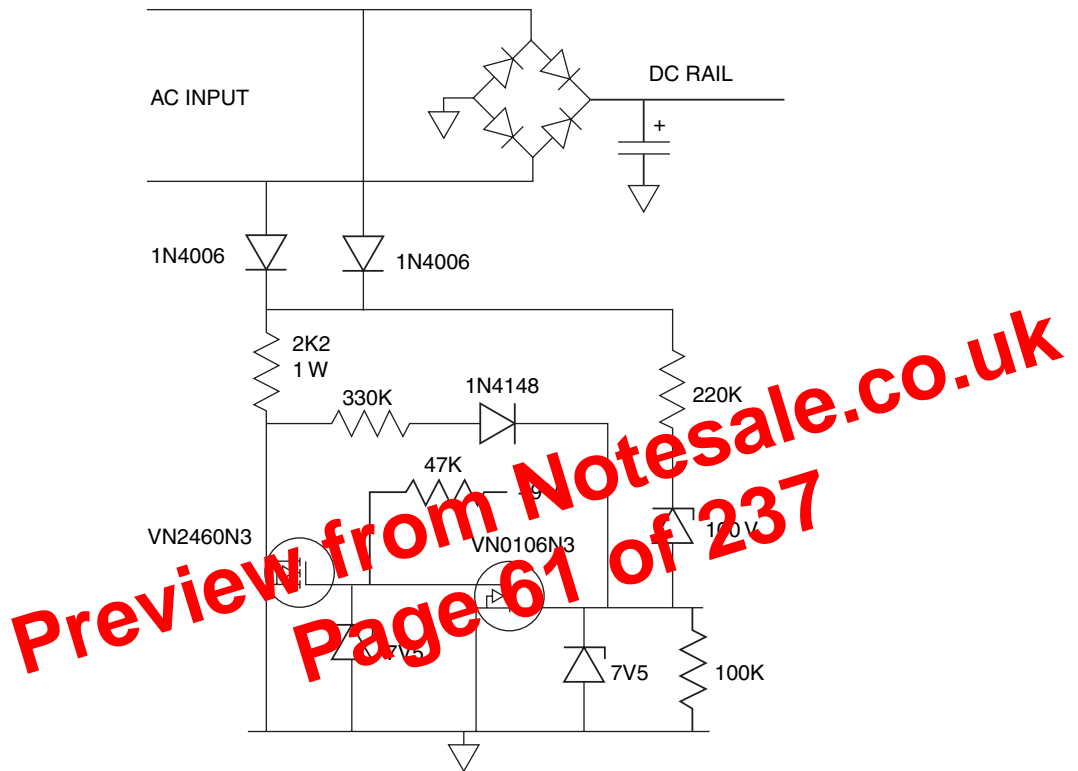


Figure 5.7: Additional Load Switch.

(PWM input). Thus when the triac is off, the LED driver is also off. The combined circuit is shown in Figure 5.8.

## 5.5 Common Errors in AC Input Buck Circuits

The most common error is trying to drive a single LED from the AC mains supply. The duty cycle is  $V_{\text{out}}/V_{\text{in}}$ , so for universal AC input 90 V to 265 V AC, the rectified voltage is about 100 V to 375 V. The worst case is the higher voltage; consider driving a white LED with 3.5 V forward voltage. The duty cycle will be  $3.5/375 = 0.9333\%$  duty cycle. If the switching frequency is 50 kHz, with 0.02 ms second period, the MOSFET on-time will be just 186 ns. This time is too short for the current sense

## 5.7 Hysteretic Buck

As an alternative to the peak current control buck, hysteretic control can be used. This uses a fast comparator to drive the MOSFET switch. The input to the comparator is a high side current sense circuit, where the voltage across a resistor in the positive power feed to the LED load is monitored. This is shown in Figure 5.10.

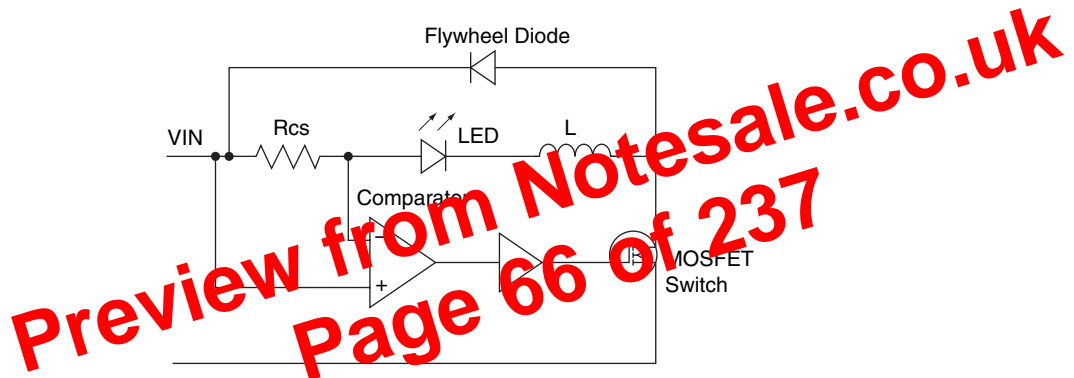


Figure 5.10: Hysteretic Current Control Circuit.

The MOSFET is turned on when the current level is at or below a minimum reference voltage. The MOSFET is turned off when the current is at or above a maximum reference voltage. This is shown in Figure 5.11. By this method, the average LED current remains constant, regardless of changes in the supply voltage or LED forward voltage.

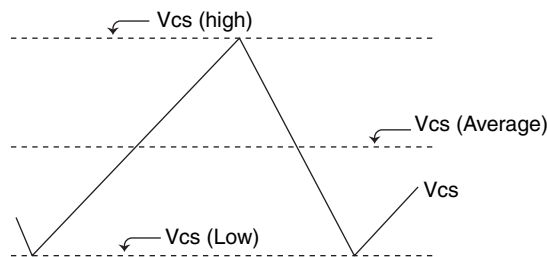


Figure 5.11: Current Sense Voltage (Current in LED Load).

Boost converters have some disadvantages, especially when used as LED drivers, due to the low dynamic impedance of the LED string.

- The output current of the boost converter is a pulsed waveform. Thus, a large output capacitor is required to reduce the ripple in the LED current.
- The large output capacitor makes PWM dimming more challenging. Turning the boost converter on and off to achieve PWM dimming means the capacitor will have to be charged and discharged every PWM dimming cycle. This increases the rise and fall times of the LED current.
- Open loop control of the boost converter to control the LED current (as in the case of an HV9910-based buck control) is not possible. Closed loop is required to stabilize the converter. This also complicates PWM dimming, since the controller will have to have a large bandwidth to achieve the required response times.
- There is no control over the output current during the output short circuit conditions. The current path from the input to the output via a diode and inductor, by turning off the switching MOSFET will have no effect on the short circuit current.
- There will be a surge of current into the LEDs if an input voltage transient raises the input voltage above the LED string voltage. If the surge current is high enough, the LEDs will be damaged.

## 6.1 Boost Converter Operating Modes

A boost converter can be operated in two modes – either continuous conduction mode (CCM) or discontinuous conduction mode (DCM). The mode of operation of the boost converter is determined by the waveform of the inductor current. Figure 6.2(a) is the inductor current waveform for a CCM boost converter whereas Figure 6.2(b) is the inductor current waveform for a DCM boost converter.

The CCM boost converter is used when the maximum step-up ratio (ratio of output voltage to input voltage) is less than or equal to six. If larger boost ratios are required, the DCM boost converter is used. However, in discontinuous conduction mode, the inductor current has large peak values, which increases the core losses in the inductor. Thus DCM boost converters are typically less efficient than CCM boost converters, can create more EMI problems and are usually limited to lower power levels.

### 6.3.1 Design Specification

- Input voltage range = 22–26 V
- LED string voltage range = 40–70 V
- LED current = 350 mA
- LED current ripple = 10% (35 mA)
- LED string dynamic impedance = 18 ohms
- Desired efficiency >90%

### 6.3.2 Typical Circuit

A typical boost converter circuit is shown in Figure 6.5.

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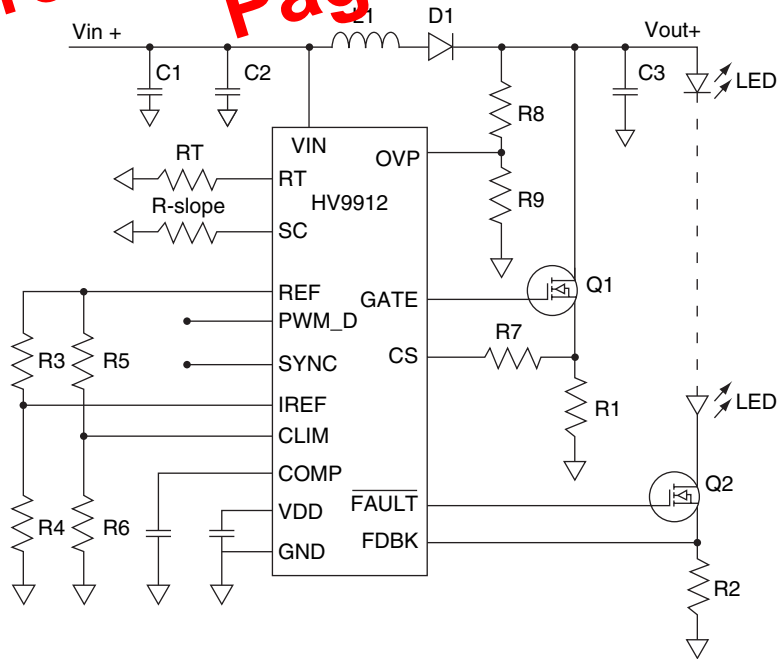


Figure 6.5: Continuous Mode Boost Converter.

The programming resistors can then be calculated as

$$R_{\text{slope}} = \frac{10 \cdot R7 \cdot f_s}{DS(A/\mu S) \cdot 10^6 \cdot R1}$$

Assuming  $R7 = 1 \text{ k}\Omega$ ,

$$\begin{aligned} R_{\text{slope}} &= \frac{10 \cdot 1\text{k} \cdot 200\text{k}}{0.2682 \cdot 10^6 \cdot 0.15} \\ &= 76.62 \text{ k}\Omega \end{aligned}$$

**Note:** The maximum current that can be sourced out of the GC pin is limited to  $100 \mu\text{A}$ . This limits the minimum value of the  $R_{\text{slope}}$  resistor to  $25 \text{ k}\Omega$ . If the equation for slope compensation produces a value for  $R_{\text{slope}}$  less than this value, then  $R7$  would have to be increased accordingly. It is recommended that  $R_{\text{slope}}$  be chosen in the range  $25 \text{ k}\Omega$ – $50 \text{ k}\Omega$ .

Based on this recommendation, the calculated values can be scaled by 0.51. The selected resistor values are

$$\begin{aligned} R7 &= 510, 1/8 \text{ W}, 1\% \\ R_{\text{slope}} &= 39\text{k}, 1/8 \text{ W}, 1\% \end{aligned}$$

### 6.3.16 Setting the Inductor Current Limit ( $R5$ and $R6$ )

The inductor current limit value depends on two factors – the maximum inductor current and the slope compensation signal added to the sensed current. Another resistor divider, connected to the REF pin, sets this current limit. The voltage at the CLIM pin can be computed as

$$V_{\text{CLIM}} \geq 1.35 \cdot I_{\text{in,max}} \cdot R1 + \frac{4.5 \cdot R7}{R_{\text{slope}}}$$

This equation assumes that the current limit level is set at about 120% of the maximum inductor current  $I_{\text{in,max}}$  and that the operating duty cycle is at 90% (maximum for the HV9912).

$$\begin{aligned}
 a &= K_2 \\
 b &= \frac{V_{\text{in}}}{V_o} \cdot K_1 + K_3 \\
 c &= T_{\text{off}}
 \end{aligned}$$

$$\Delta = 12 \cdot \sqrt{3} \cdot \sqrt{\frac{4 \cdot b^3 + 27 \cdot a \cdot c^2}{a}}$$

$$L_2 = \left\{ \frac{1}{6 \cdot a} \left[ (108 \cdot c + \Delta) \cdot a^2 \right]^{1/3} - \frac{2 \cdot b}{\left[ (108 \cdot c + \Delta) \cdot a^2 \right]^{1/3}} \right\}^3$$

The actual off-time  $T_{\text{off,ac}}$  can be computed by substituting the chosen inductor value back into the equations for  $T_r$ ,  $T_f$  and  $T_2$ , to get  $T_{r,ac}$ ,  $T_{f,ac}$  and  $T_{f2,ac}$ .

$$\begin{aligned}
 T_{\text{off,ac}} &= T_{f1,ac} + T_{f2,ac} + T_{f,ac} \\
 &= \frac{V_{\text{in}}}{V_o} \cdot T_{r,ac} + T_{f2,ac} + T_{f,ac}
 \end{aligned}$$

The actual ripple in the inductor current  $\Delta i_{o,ac}$  is

$$\Delta i_{o,ac} = \frac{V_o \cdot T_{\text{off,ac}}}{L_2}$$

#### 7.1.4 Stability of the Boost-Buck Converter

The single-switch boost-buck converter can be considered as separate boost and buck converters (in that order), which are cascaded, and both switches being driven with the same signal (see Figure 7.8).

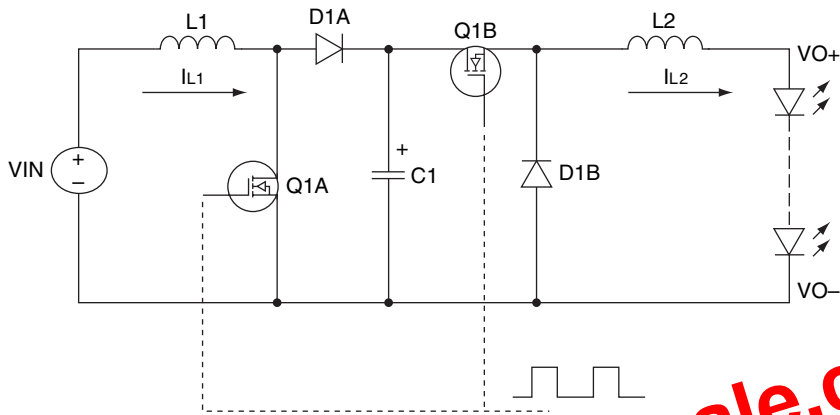


Figure 7.8: Boost-Buck Converter

The relationships between the voltages in the system are

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$$\frac{V_c}{V_{in}} = \frac{1}{1 - D} \quad (\text{boost converter})$$

$$\frac{V_o}{V_c} = D \quad (\text{buck converter})$$

The capacitor voltage  $V_c$  and the input/output relationship can both be derived using the above equations

$$\frac{V_o}{V_{in}} = \frac{V_o}{V_c} \cdot \frac{V_c}{V_{in}} = \frac{D}{1 - D}$$

$$V_c = \frac{V_{in}}{1 - D} = \frac{V_{in}}{1 - V_o/V_c}$$

$$\Rightarrow V_c = V_o + V_{in}$$

For the purposes of designing the damping network, it is easier to visualize the converter in its two-switch format of Figure 7.7 rather than as the single-switch Cuk converter. Hence, for the remainder of this section, the cascaded converter will be used to derive the equations.

cost or availability, a plastic film capacitor such as PET can be used instead, although these are considerably bulkier.

The maximum steady state voltage across the capacitor is 44 V ( $= 28 \text{ V} + 16 \text{ V}$ ), and the maximum transient voltage across the capacitor  $V_{c,\max}$  is 70 V ( $= 28 \text{ V} + 42 \text{ V}$ ). Ceramic capacitors can easily withstand up to 2.5 times their voltage rating for the duration of the load dump voltage. Also, the actual capacitance value of these capacitors reduces based on the bias voltage applied. Ceramic capacitor types X7R and X5R are more stable and the capacitance drop is not more than 20% at full rated voltage.

Thus, a 0.22  $\mu\text{F}$ , 50 V X7R ceramic chip capacitor can be selected.

### Choosing the Switching Transistor (Q1)

The peak voltage across the MOSFET  $Q_1$  is 70 V. Assuming a 30% overhead on the voltage rating to account for leakage inductance spikes, the MOSFET voltage needs to be at least

$$\begin{aligned} V_{\text{FET}} &= 1.3 \cdot V_{c,\max} \\ &= 91\text{V} \end{aligned}$$

The RMS current through the MOSFET will be at maximum level at low input voltage (higher current levels and maximum duty cycle). The maximum RMS current through the MOSFET is

$$\begin{aligned} I_{\text{FET,max}} &= (I_{\text{in,max}} + I_o) \cdot \sqrt{D_{\max}} \\ &= 1.77\text{A} \end{aligned}$$

A typical choice for the MOSFET is to pick one whose current rating is about three times the maximum RMS current. Choose FDS3692 from Fairchild Semiconductors (100 V, 4.5 A, 50 m $\Omega$  N-channel MOSFET). Note that the current rating is normally quoted at 25°C; the current rating reduces as the temperature rises.

The total gate charge  $Q_g$  of the chosen MOSFET is a maximum of 15 nC. It is recommended that the MOSFET total gate charge should not exceed 20 nC, as the large switching times will cause increased switching losses. A higher gate charge would be allowable if the switching frequency can be reduced appropriately.

This power dissipation is a maximum value, which occurs only at minimum input voltage. At a nominal input voltage of 13.5 V, we can compute the input current using the nominal values for the efficiency and the input voltage.

$$I_{in,nom} = \frac{28 \cdot 0.35}{0.8 \cdot (13.5 - 0.5)}$$

$$= 0.942 \text{ A}$$

$$P_{R_{CS1}} = 0.942^2 \cdot 0.228 = 0.2 \text{ W}$$

Thus, at nominal input voltage, the power dissipation reduces by about five times to a reasonable 0.2 W.

Choose the following values for the resistor:

$R_{CS1}$  = parallel combination of three 3.3 kΩ, 1/2 W, 5% resistors

$R_{REF}$  = 10 kΩ, 1/8 W, 1%

$R_{S1}$  = 4.42 kΩ, 1/8 W, 5%

### Input Inductor Current Rating

The maximum current through the input inductor is  $I_{lim,max} = 1.15 \cdot I_{in,lim} = 2.4 \text{ A}$ . Thus, the saturation current rating of the inductor has to be at least 2.5 A. If the converter is going to be in input current limit for extended periods of time, the RMS current rating needs to be 2 A, else a 1.5 A RMS current rating will be adequate.

### Improving Efficiency

The input current sense resistor can be reduced in value, which gives reduced power dissipation (loss). To allow this, it is necessary to add an extra resistor ( $R_A$ ) between the anode of the flywheel diode and the current sense input of the HV9930 (AT9933). This resistor allows a reduction in the hysteresis required by the input comparator. The additional resistor is shown in Figure 7.12.

In Figure 7.12,  $R_{S1} = R_4$ ,  $R_{REF1} = R_7$ , and  $R_{CS1}$  = the parallel combination  $R1//R3$ .

$L1$  is chosen so its total current ripple ( $\Delta I_{L1}$ ) is a fraction ( $\beta = 20\%$  to  $50\%$ ) of  $I_{L1}$ . The worst case for  $\beta$  occurs when  $V_{IN}$  is at maximum, because  $\Delta I_{L1}$  is at maximum when  $I_{L1}$  is at minimum. Assuming  $\beta = 0.5$ :

$$L1_{\min} = \frac{2 \cdot T \cdot (1 - D_{\max}) \cdot V_{IN\_MAX}}{I_{OUT}}$$

Choose a standard value nearest to that calculated for  $L1$ , and make sure its saturation current meets the following condition:

$$I_{L1\_SAT} \gg I_{L1} + 0.5 \cdot \Delta I_{L1} = \frac{Aa_{\min} \cdot I_{OUT} + 0.5 \cdot T \cdot D_{\min} \cdot V_{IN\_MAX}}{L1}$$

The calculation for  $L2$  is similar to that for  $L1$ .

$$L2_{\min} = \frac{2 \cdot T \cdot D_{\min} \cdot V_{IN\_MAX}}{I_{OUT}}$$

$$I_{L2\_SAT} \gg I_{L2} + 0.5 \cdot \Delta I2 = \frac{I_{OUT} + 0.5 \cdot T \cdot D_{\max} \cdot V_{IN\_MAX}}{L2}$$

If  $L1$  and  $L2$  are wound on the same core, you must choose the larger of the two calculated inductor values. Using a single core, the two windings should be bifilar (twisted around each other before being wound on the core) and thus will have the same number of turns and the same inductance values. Otherwise, voltages across the two windings will differ and  $Cp$  will act as a short circuit to the difference. If the winding voltages are identical, they generate equal and additive current gradients. In other words, there will be mutual inductance of equal value in both windings. Thus, the inductance measured across each isolated winding (when there is nothing connected to the other winding) should equal only half of the value calculated for  $L1$  and  $L2$ .

Because no great potential difference exists between the two windings, you can save costs by winding them together in the same operation. If the windings' cross-sections are equivalent, the resistive losses will differ because their currents ( $I_{L1}$  and  $I_{L2}$ ) differ. Total loss, however, is lowest when losses are distributed equally between the two windings, so it is useful to set each winding's cross-section according to the current it carries. This is particularly easy to do when the windings consist of insulated strands

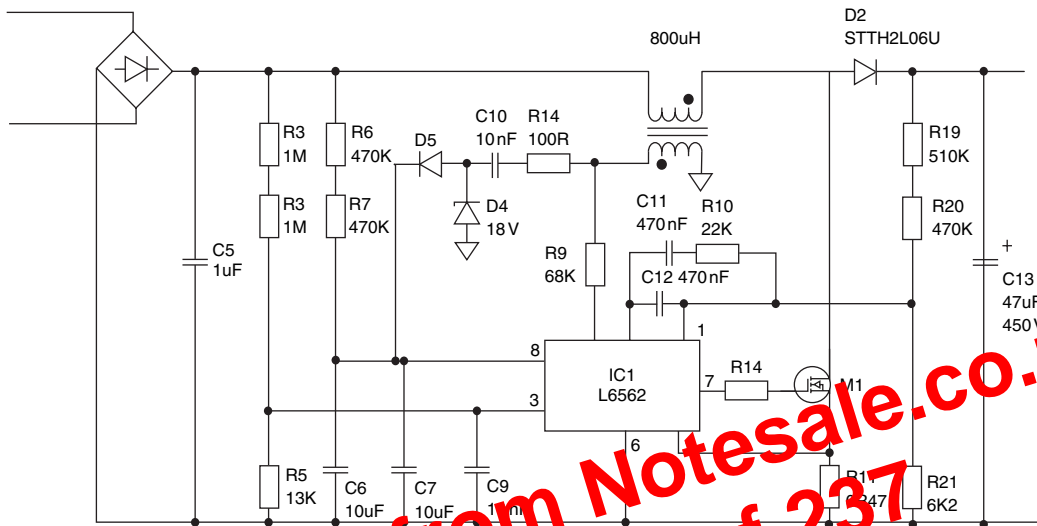


Figure 8.2 PFC Circuit.

## 8.2 Bi-Bred

The Bi-Bred is very similar to the Cuk boost-buck that we described in the previous chapter, see Figure 8.2.

The main difference between the Cuk and the Bi-Bred is that, in a Bi-Bred, the input inductor is in discontinuous conduction mode (DCM) and operation of the output stage is in continuous conduction mode (CCM). The energy stored in each inductor is proportional to the inductance value. This means that in the design, the input inductor  $L1$  must have a small enough energy stored to ensure that conduction stops before the end of each cycle. This means that the input inductor value must be relatively small. The output inductor  $L2$  must have large enough energy stored (large inductance value) so that the current only falls to about 85% of its nominal value at the end of each switching cycle.

When power is first applied, MOSFET  $M1$  is off and waiting for the first clock signal to trigger the gate drive pulse. At this time the storage capacitor  $C3$  immediately begins to charge from the supply voltage through  $D1$  and  $L1$ , although the voltage will not rise very high because, when the MOSFET  $M1$  switches on, the charging current is redirected to the 0 V rail. With  $M1$  conducting, current continues to rise in

## *Fly-Back Converters*

A traditional fly-back converter uses an inductor with at least two windings (really, this is a transformer). Consider two windings; one is the primary, which is connected to the input power supply and a switch to ground; the other is the secondary, which is connected to the load. The circuit is arranged so that magnetic energy is stored in the inductor during the time that the switch is on, when current increases in the primary winding. When the switch is off, the magnetic energy is released by current flowing out of the secondary winding. This is shown in Figure 9.1.

The energy release is the ‘fly-back’, so called because in early television sets with a cathode ray tube, a transformer winding was used to deflect the electron beam back to the starting point on the screen. The electron beam had to ‘fly back’ quickly after completing a scan across the screen, to avoid missing the next line of data to be displayed.

Fly-back power supplies are relatively easy to design, but are more suited to constant voltage outputs. This is because the energy is stored in bursts, in a large reservoir capacitor, and controlling the average voltage across the capacitor can be achieved with simple feedback.

Driving an isolated LED load is then possible if the secondary winding is isolated from the primary winding. Some general-purpose applications can use simple current limit techniques in the primary winding to control the output current from the secondary winding. An opto-coupler will be required, to maintain isolation between primary and secondary, if accurate control of the output current is required.

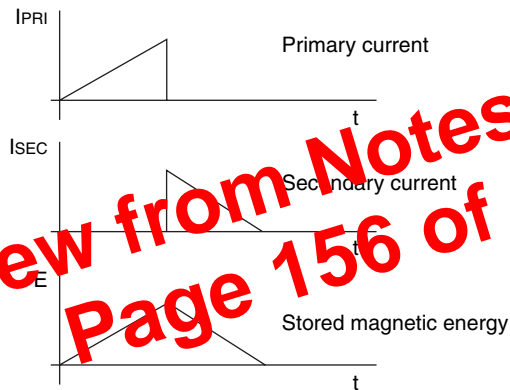
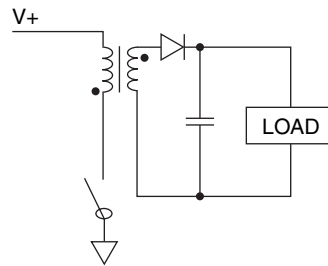


Figure 9.1: Fly-Back Principle.

Some fly-back converters use an inductor with a single winding. These are buck-boost controllers and are an alternative to the boost-buck converters like Cuk and SEPIC types that were discussed in Chapter 7. Clearly, isolation is not possible with this type of converter.

## 9.1 Two Winding Fly-Back

A schematic of a typical fly-back circuit for driving LEDs is shown in Figure 9.2. The dot alongside the transformer winding indicates the start of the winding. In this case the start is connected to the MOSFET drain, which alternates between a ground connection and open circuit. The voltage at the drain, and hence the winding start point, varies considerably during switching. Conversely, the outer layer (end of the winding) is at a fixed high voltage and tends to screen the inner layers, which reduces radiated EMI.

The output power is  $10\text{ V} \times 0.35\text{ A} = 3.5\text{ W}$  and the efficiency can be guessed at as being 85%. The input power is then 4.12 W. The input current at minimum input voltage is then:

$$I_{AV} = \frac{P_{IN}}{V_{IN}} = \frac{4.12}{46} = 0.09\text{ A}$$

$$I_{PK} = \frac{2 \cdot I_{AV}}{D_{MAX}}$$

At 46  $V_{in}$  and 41% duty cycle:

$$I_{PK} = \frac{2 \cdot 0.09}{0.41} = 0.439\text{ A}$$

With 60 kHz switching, the period will be 16.667  $\mu\text{s}$ . With a 41% duty cycle, the switch on-time will be 6.835  $\mu\text{s}$ . So we need the primary current to rise by 439 mA in 6.835  $\mu\text{s}$ .

$$L_{PRI} = \frac{V_{IN} \cdot dt}{I} = \frac{46 \cdot 6.835 \cdot 10^{-6}}{0.439} = 716\text{ } \mu\text{H}$$

The secondary has 1/3 the number of turns compared to the primary, so the inductance of the secondary will be 1/9, or 79.55  $\mu\text{H}$ .

The other design parameter for the transformer is the size and  $AL$  factor of the ferrite core. In a fly-back transformer an air gap between the two halves of the ferrite core are necessary to prevent magnetic saturation, as the air gap increases, the  $AL$  factor reduces. The flux density ( $B$ ) will depend on the cross-sectional area of the core ( $A_C$ ), given in square meters. Suppose in this case we have some E20 cores available from Ferroxcube. For E20/10/6 cores, the core cross-sectional area is 32  $\text{mm}^2$ . So  $A_C = 32 \cdot 10^{-6}\text{ m}^2$ . The number of turns can be calculated, based on the design parameters above and using  $B = 200\text{ mT}$  as the maximum flux density:

$$N = \frac{L_{PRI} \cdot I_{PK}}{A_C \cdot B_{MAX}} \text{ (turns)}$$

$$N1 = \frac{716 \cdot 10^{-6} \cdot 0.439}{32 \cdot 10^{-6} \cdot 0.2} = 49$$

$$A_L = \frac{L_{PRI}}{N1^2} = \frac{716 \cdot 10^{-6}}{2401} = 298\text{ nH}$$

Of course the efficiency may suffer, and so a heatsink will be needed, but this is sometimes very much better than trying to make a switching circuit in terms of cost and physical size.

### 10.2.1 Buck Regulator Considerations

In Chapter 5 we first looked at the simplest switching regulator, the buck converter. In a buck circuit the load voltage must be less than 85% of the supply voltage, otherwise the output becomes difficult to control. Buck circuits are used for mains powered LED drivers, when driving a long string of LEDs. Buck circuits are also used where the input supply voltage is relatively low, say in a 12V DC automotive application, but where just one LED is being driven.

Buck regulators can be very efficient, in fact 90–95%, especially if the load is a long string of LEDs with a moderately high forward voltage (i.e. high duty cycle). This is because the power dissipation in the flywheel diode is a smaller proportion of the total power because the flywheel diode only conducts during the MOSFET off-time, which is a smaller proportion of the total switching cycle. The MOSFET dissipates power during the on-time, when it is conducting, but the voltage drop across the MOSFET switch is usually much lower than the forward drop of a fast rectifier.

In order to operate correctly there must be some ripple in the output current. The output current needs to reduce enough to allow the current sense comparators to be reset. The output ripple current  $\Delta I_O$  is normally designed to be 20–30% of  $I_O$ ; the output current falls far enough in each cycle so that noise in the current sense comparator has little effect. If the ripple current is below 10% of  $I_O$ , the switching of the MOSFET can be erratic. The output current in the LED string ( $I_O$ ) is given by the equation:

$$I_O = \frac{V_{TH}}{R_{SENSE}} - \frac{1}{2} \cdot \Delta I_O$$

Here  $V_{TH}$  is the current sense comparator threshold, and  $R_{SENSE}$  is the current sense resistor. The ripple current can introduce a peak-to-average error in the output current setting that needs to be accounted for. When the constant off-time control technique is used, the ripple current is nearly independent of the input supply voltage variation. Therefore, the output current will remain unaffected by the varying input voltage.

Parasitic elements could potentially cause false triggering of the LED driver IC's current sense comparator, especially if an RC filter is not fitted between the MOSFET source and the current sense (CS) pin. Minimizing parasitic elements is essential for efficient and reliable operation of the buck converter.

Coil capacitance of inductors is typically provided in the manufacturer's data books either directly or in terms of the self-resonant frequency (*SRF*).

$$SRF = 1/(2\pi\sqrt{L \cdot C_L}),$$

Here  $L$  is the inductance value, and  $C_L$  is the coil capacitance. Charging and discharging this capacitance every switching cycle causes high current spikes in the LED string. Therefore, connecting a small capacitor (10 nF) across the LED string is recommended to bypass these spikes, as mentioned earlier.

Using an ultra-fast rectifier flywheel diode is recommended to achieve high efficiency and reduce the risk of false triggering of the current sense comparator. When the MOSFET turns on the diode changes from forward conduction to off (reverse bias), but this cannot happen immediately because charges have to move inside the semiconductor material, which takes time. There is always a reverse recovery current flowing in the opposite direction for a short period,  $T_{RR}$ . Using diodes with shorter reverse recovery time,  $T_{RR}$ , and lower junction capacitance  $C_j$  improves performance. The reverse voltage rating  $V_R$  of the diode must be greater than the maximum input voltage of the LED lamp. The forward voltage-drop of diodes with very fast recovery times is sometimes relatively high and can lead to high conduction losses, so also consider this when making a diode selection.

The total parasitic capacitance present at the DRAIN output of the MOSFET can be calculated as:

$$C_P = C_{DRAIN} + C_{PCB} + C_L + C_j$$

When the switch turns on, the total parasitic capacitance  $C_P$  is discharged into the DRAIN output of the MOSFET. The discharge current is limited to the MOSFET saturation current, so MOSFETs with a high on-resistance and a lower saturation current can sometimes produce lower overall losses. This is especially true if the duty cycle is small, because the switch is conducting for a small proportion of the time and

A fly-back, by definition, is a discontinuous conduction mode converter; energy is taken from the power supply in the first step and then transferred to the output in the second step, as shown in Figure 10.5. This means that EMI must be carefully filtered at both the input and output. The output requires a large storage capacitor to maintain current flow in the LEDs when the converter is on the first step. Dimming the LED light by pulse width modulation (PWM) of the current is very difficult because the stored energy in the capacitor tries to maintain current flow; thus only a modest dimming range is possible.

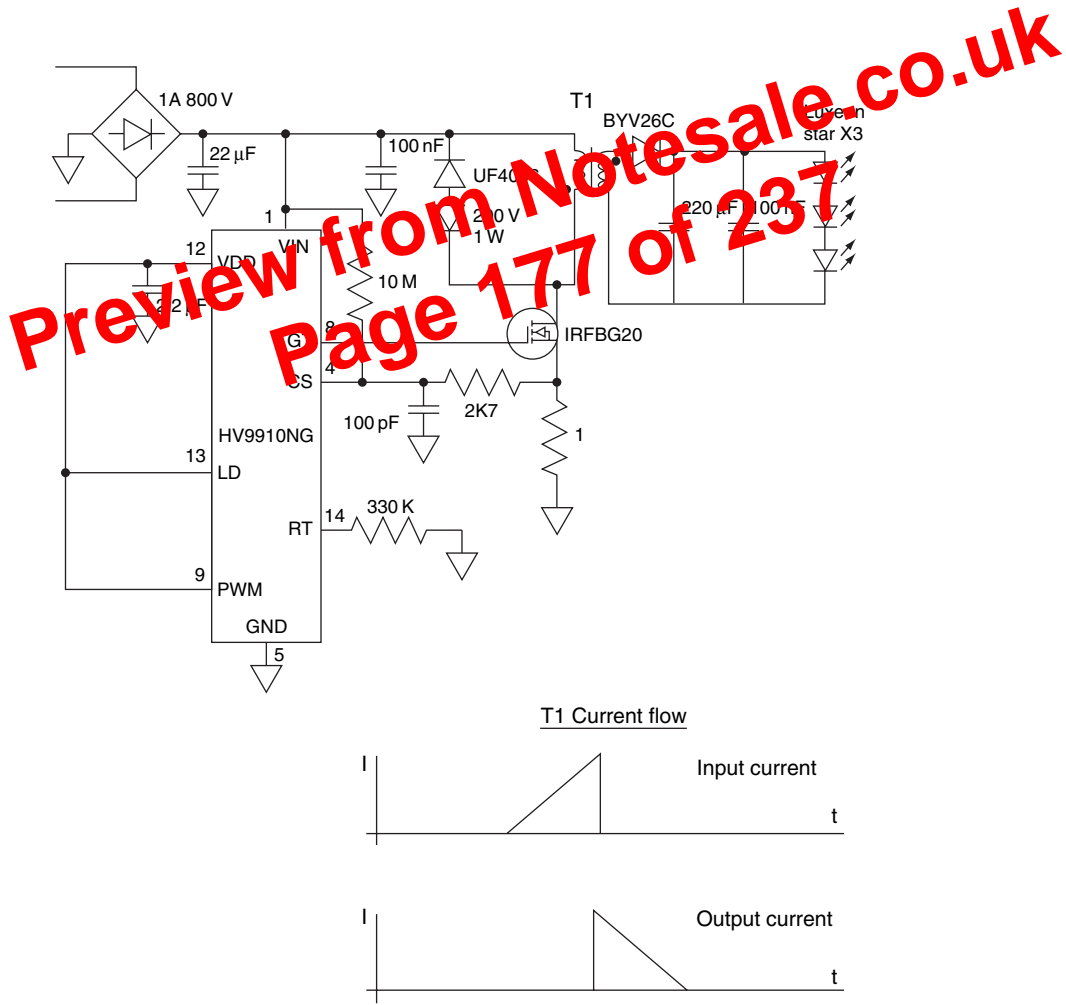


Figure 10.5: Discontinuous Fly-Back Current.

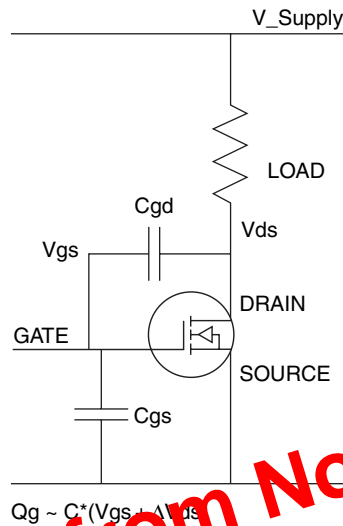


Figure 11.1.1: MOSFET Circuit with Parasitic Capacitance.

The average current into the LED driver IC will be a small quiescent current plus the product of gate charge and switching frequency.

$$I = I_Q + Q_G * F_{SW}$$

This is important when calculating the power dissipation in a MOSFET driver circuit. The power dissipation will be  $V_{Supply} * I$ , where  $I$  is the current calculated using the gate charge.

### 11.1.2 Bipolar Transistors

Bipolar transistors are used in switching and linear LED driver circuits. They operate by a current magnification effect; the collector-emitter current is a multiple of the base-emitter current. The base-emitter voltage is about 0.7 V, being the voltage drop of a forward biased P-N junction. There is some base-emitter resistance, so the forward voltage drop will increase slightly with base current.

Matched transistors can be very useful, particularly in current mirror circuits. A current mirror is one where two or more branches carry identical currents; the current in one branch depends on the current in another, hence the ‘mirror’. Transistors do not have to

An inductor's behaviour is to oppose any change in the current flowing through it. This is because the energy stored in an inductor is given by  $E = \frac{1}{2}LI^2$ . To change the current instantly through an inductor would take infinite power. If we ignore physical imperfections due to the construction of an inductor, when a voltage is applied across it the current will increase linearly. If a load is then applied across the inductor, the current falls linearly. If we alternately switch the voltage source and the load across the inductor, the current will rise and fall, but remain fairly constant.

Inductors can be used to filter the power supply lines in switching LED drivers. Because of their energy storage characteristics, they tend to oppose any change in current, so they present high impedance to unwanted interference. Combined with capacitors that are low impedance to unwanted interference, the resulting 'L' or 'PI' filter considerably reduces the amplitude of high frequency signals.

Inductors can be a source of many problems. High value inductors are bulky. This is because they are usually made up from tens or hundreds of turns of enameled copper wire that is wound on a ferrite core. The winding capacitively couple to each other, which effectively introduces a parallel capacitor across the coil. This capacitance causes switching losses in power supplies, or poor filtering in supply input filters. Above the self-resonant frequency, the impedance of the inductor falls due to the capacitive reactance dominating.

Inductors also possess some series resistance due to the intrinsic resistance of the copper wire used. This resistance will cause losses in the power supply and thus limit the efficiency. Heating effects due to this resistance can cause problems. Choosing an inductor for the correct inductance value, without considering the ESR and self-resonant frequency will give poor results.

Magnetizing (core) losses are also present and are due to the energy required to make the magnetic fields in the core to align with each other. In a switching circuit these losses are continuous and can cause core heating. These losses increase rapidly if the magnetisation is forced to operate outside its linear region. The presence of an air gap in inductor and transformer cores makes them suitable for high magnetic saturation levels. Transformer cores that have no air gap and are prone to saturate easily.

The saturation current ( $I_{\text{sat}}$ ) quoted by manufacturers is usually at the point where the inductance drops by 10%. If the current drops to or near zero each switching cycle, the peak current should be kept well below the saturation level (I suggest  $I_{\text{max}} = 0.5 * I_{\text{sat}}$ , but preferably  $0.25 * I_{\text{sat}}$ ). Take care, because sometimes the

All conductors have some series inductance, simply due to having a certain length. This is typically 6 nH per centimeter. In fact some high frequency circuits just use a thin wire bond to form an inductor. Resistors are conductors and therefore have inductance too. Some types have more inductance than others. Even a thick-film surface mount resistor has inductance, although of considerably lower value than other types. Wire wound resistors have a significant inductance because of their construction, when a wire is wound into a coil its inductance increases in proportion to the number of turns squared. Carbon or metal film resistors that have had a spiral gap cut through their surface will have more inductance than a carbon composition type. All through-hole components also have some inductance due to the wire leads at either end.

Resistors also have capacitance. The two ends have a certain cross-sectional area and are spaced a certain distance apart, separated by a certain dielectric. This capacitance is small, typically 0.2 pF, so has little effect in an LED driver circuit operating up to 1 MHz. At high radio frequencies and at a high impedance circuit node this capacitance can be significant.

### 11.3 The Printed Circuit Board (PCB)

Regulations on the use of tin-lead solder have come into force for most applications, for human health reasons. The notable exceptions are military and (ironically) medical applications, although these will be forced to change due to the lack of RoHS lead-free components. Heavy metals and carcinogenic materials will not be allowed in electronic products, including IC packages. This means that soldering profiles have to change – higher temperatures are needed for lead-free solder.

The circuit board on which the components are connected is important at high frequencies and for surface mount circuits. At high frequencies, for example, capacitance between tracks can cause a lower resonance frequency in a tuned circuit. Surface mount circuits can have reliability problems due to thermal expansion of the circuit board; components firmly attached to the tracks with solder can be stressed if they do not have the same thermal expansion. There are several types of board, with FR4 (fibreglass insulator) being the most common.

Through-hole construction is becoming less common, due to the reduced availability of through-hole components. For slow speed circuit prototypes they are ideal for fault finding and fast construction. For high speed circuits and production, surface mount construction gives better performance and lower cost.

**Table 12.1: Wire Resistance Versus Temperature.**

Temperature	Multiplying factor
20°C	1
40°C	1.079
60°C	1.157
80°C	1.236
100°C	1.314

Unfortunately, the resistance of wire also increases as the frequency of signals passing through it increases. The phenomenon of the ‘skin effect’ is when the magnetic field caused by the current flow tends to force the electrons to flow down the outside of the wire. An alternating magnetic field produced by the current in the wire induces an electric field, strongest at the center of the wire, which repels the electrons and forces them to the outside circumference of the wire. Thus, an increase in current produces a force that opposes this change, which is inductance on a small scale.

The skin depth is given in Table 12.2.

**Table 12.2: Skin Depth Versus Frequency.**

Frequency	Skin depth
50 Hz	9.36 mm
1 kHz	2.09 mm
100 kHz	0.209 mm
1 MHz	0.0662 mm
10 MHz	0.0209 mm

Fortunately, Terman has created a formula for a wire gauge (in millimeters) where the skin effect increases resistance by 10%, which is a nominal limit that allows reasonable losses:

$$D = \frac{200}{\sqrt{F}} \text{ millimeters}$$

## *EMI and EMC Issues*

The first two questions regarding EMI and EMC are: what's the difference between EMI and EMC? And which standards apply? Subsequent questions relate to how equipment can be made to meet the standards. Of course, meeting the standards often costs money (filter components, shielding and suppressors) so the aim is to just meet the standards with a small safety margin.

EMI is electro-magnetic interference. This is the amount of radiation emitted by some equipment when it is operating. EMI is caused by emissions in the radio spectrum, which not only interfere with radio systems but also can cause other equipment to malfunction. One example is interference from portable radio transmitters like CB radios and cell phones; when used near a gasoline station, the pump can indicate the wrong amount being delivered. An often seen warning notice at a gasoline station says 'using a radio transmitter can cause a fire', but in reality the most likely effect is to cause an error in the fuel measurement. I did hear a story that CB radio users could reset the fuel counters by an appropriately timed transmission, but maybe this was just wishful thinking!

So what is EMC? This is electro-magnetic compatibility, and is a measure of how good a system is at rejecting interference from others. Medical systems have a high immunity requirement, because the consequences of a failure are death or injury. Any system connected to the AC mains power line must be immune to transient surges; the degree of immunity depends on the application. Power meters connected to lines where they enter a building are subject to the highest potential surges, so they have very high immunity requirements. Internal lighting and domestic appliances have very much lower immunity requirements.

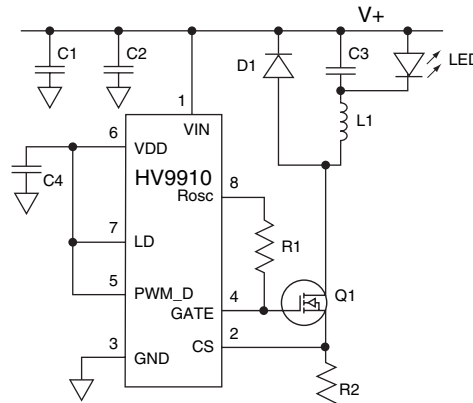


Figure 13.2: Buck Circuit.

latch is reset and the gate drive output is disabled. The MOSFET  $Q1$  turns off but current continues to flow in the LED and inductor due to the flywheel diode  $D1$ . When used in a buck circuit, this IC maintains an almost constant current in the LED.

When the gate pin of the HV9910B outputs a voltage of 7.5 V, the MOSFET  $Q1$  turns on causing current to flow through the inductor  $L1$  and the LED. The drain voltage is very low, just a small voltage due to the current flowing in the drain-source channel of  $Q1$  and in the current sense resistor  $R2$ . When  $Q1$  turns off, current through the inductor cannot stop so it flows through the flywheel diode  $D1$ . When  $D1$  conducts, the drain of  $Q1$  is clamped to the positive supply rail. So the voltage waveform on the MOSFET drain is a rectangular wave. The fast rising and falling edges create a broad spectrum of harmonics.

Current flows are shown in Figure 13.2. Analysis shows that the gate current flows from ground, through  $V_{dd}$  supply capacitor  $C4$ , through the IC and out of the gate drive pin, through the gate and current sense resistor and back to ground. Analysis of the LED current gives a path from ground, through the decoupling capacitors  $C1$  and  $C2$ , through the LED and inductor, through  $Q1$  and the current sense resistor  $R2$ , and back to ground. Both currents have fast rising and falling edges.

Not shown is the current that flows through the flywheel diode  $D1$ . There is a forward current when  $Q1$  is off, due the energy stored in the inductor, which keeps the LED current flowing. There is also a momentary reverse current that flows when  $Q1$  first

current also returns from the cathode of  $D1$ . Thus the high frequency flywheel current loop is from  $Q1$  drain, through diode  $D1$  and back to ground via  $C2$ ; this is in a small area to keep the impedance low and EMI radiation to a minimum. As with the ground connection, the high frequency current path is kept away from the low frequency current flowing into capacitors  $C5$  and  $C1$ .

Figure 13.6 shows both sides of the circuit board overlaid. Notice that the earth plane is below the drain area of  $Q1$  and inductor  $L1$ . Both  $Q1$  and  $L1$  have high frequency, high voltage switching, and a ground plane below helps to reduce the radiation from this area by screening underneath and making the node low impedance. Of course the capacitive coupling adds to the switching losses, but this cannot be avoided.

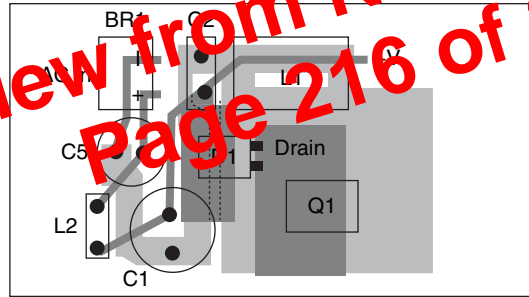


Figure 13.6: PCB Top and Bottom Layers.

### 13.2.2 Cuk Circuit Example

A Cuk circuit is a boost-buck converter that performs well in a DC input application. An example of a Cuk circuit is given in Figure 13.7.

As with the buck circuit already described, and any other switching circuit, the aim in printed circuit board design is to keep the switching currents flowing in as small a loop as possible. An earth plane under the main switching elements will also help reduce radiation.

Radiation couples easily into free space when the impedance of the signal source is similar (the impedance of free space is 377 ohms). Dipole antennas radiate and receive signals easily because their metallic elements are resonant at the transmit frequency and

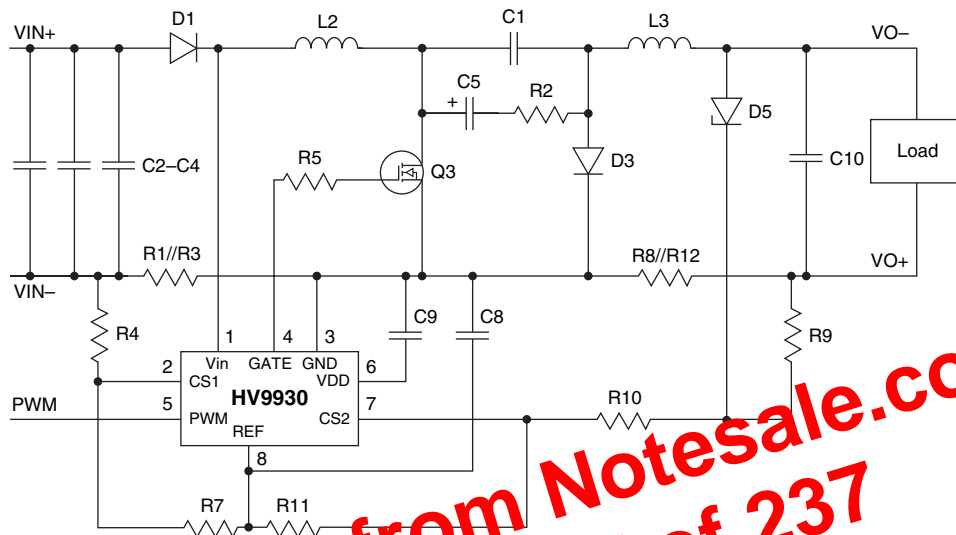


Figure 13.7: Basic Buck Circuit.

thus high impedance at the ends of the elements. Similarly, if the circuit area containing the high voltage switching signals is high impedance it will radiate interference. An earth plane under the circuit lowers the impedance and reduces radiation. The PCB designer should take this into account when designing the circuit board.

High frequency emissions are caused by the fast rising and falling edges of the MOSFET drain voltage. These can be reduced in amplitude by slowing down the switching of the MOSFET. Not only does this reduce high frequency emissions, it also reduces high frequency ringing that is caused by the drain-gate capacitance resonating with stray circuit inductance. A resistor ( $R5$ ) has been connected in series with the gate; this slows the rise-time of the gate drive signal as the MOSFET gate capacitance charges (the resistor and gate form a low-pass RC filter). Slowing the MOSFET switching speed reduces the efficiency of the LED driver circuit, but saves the cost of additional filters.

Filtering the input power connections is likely to be required. Figure 13.8 gives an example of a filter needed to meet demanding automotive specifications.

I will now describe the input filter, starting at the input of the switching circuit and working outwards towards the power source.

Finally, a small value ceramic capacitor  $C11$  is connected differentially across the power supply input to provide a low impedance path at the higher frequencies. I will now discuss the output filter, shown in Figure 13.9.

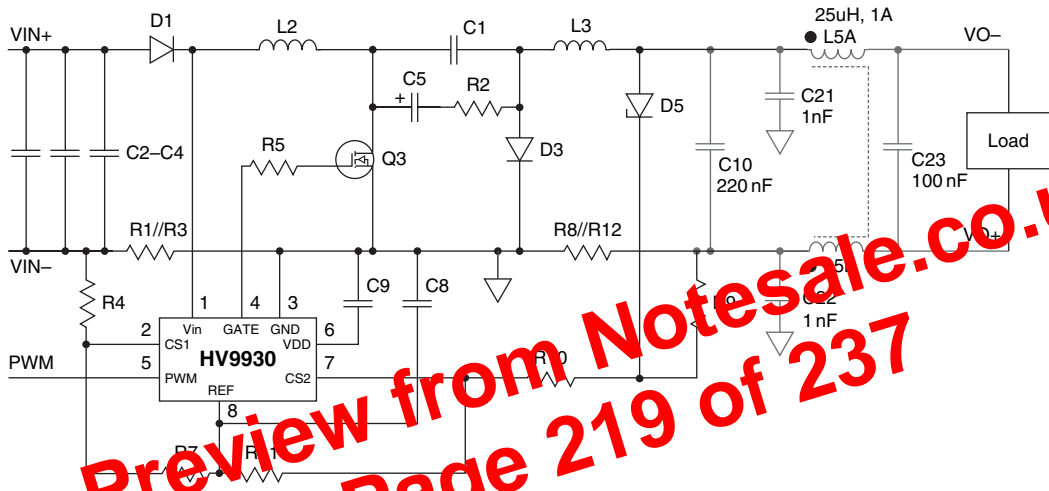


Figure 13.9: Output Filter.

The output may need a filter, particularly when there is a considerable wire length between the driver and the LED load. If the distance is very short, the only filter needed is a differential capacitor ( $C10$ ) across the load. Distances greater than 10 cm (4 inches) can cause common mode signals to be created due to stray coupling between the LED and ground. Thus we may require a common mode inductor,  $L5$ , and a second differential capacitor  $C23$ . Small value ceramic capacitors  $C21$  and  $C21$  provide a shunt path to circuit ground for high frequency signals developed across  $L3$  and the parallel current sense resistors  $R8$  and  $R12$ .

As well as a ground plane on the circuit board to reduce the impedance of the switching circuits at high frequency, a screen over the components may be needed. The position of such a screen is shown in Figure 13.10.

The screen over the switching area, and an earth plane underneath, provides a metal enclosure that stops EMI radiation. However, there will always be some leakage due to signals being carried outside the enclosure by connections to the remainder of the circuit. Even the PWM control wire will radiate unless a simple RC filter is added to it.

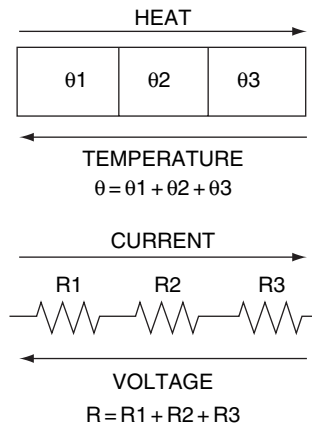


Figure 14.3: Thermal Resistances in Series.

resistance of the package to heat sink interface and the thermal resistance of the heatsink to air interface can all be added to find the total thermal resistance from the silicon junction to air.

Thermal resistance is given as degrees kelvin per watt of heat flow and has symbol  $\theta$ . (Note, a 1 degree kelvin temperature rise = 1 degree Celsius temperature rise.) The end points of the resistance are given as subscripts; for example, from junction to case, the thermal resistance is labelled as  $\theta_{JC}$ . For example, let  $\theta_{JC} = 1.2 \text{ K/W}$ ,  $\theta_{CH} = 0.1 \text{ K/W}$ , and  $\theta_{HA} = 2.4 \text{ K/W}$  (I have given my own notation  $H = \text{heatsink}$ ), so the case to heatsink resistance is  $0.1 \text{ K/W}$ . When the device is dissipating  $10 \text{ W}$ , with a total thermal resistance of  $3.7 \text{ K/W}$ , the silicon junction temperature will be 37 degree hotter than the ambient temperature. If the ambient temperature is  $25^\circ\text{C}$ , the silicon junction will be raised to  $62^\circ\text{C}$ .

Like electrical resistance, having parallel thermal resistance paths reduces the overall resistance (see Figure 14.4). Two paths, each of  $2 \text{ K/W}$ , will create an effect single path of  $1 \text{ K/W}$ . This makes calculation of the exact temperature more difficult, since thermal paths are not as obvious as electrical paths. However, for a first approximation calculating the temperature drop along obvious thermal paths will give a sufficiently accurate result. Less obvious thermal paths usually have much higher thermal resistance and have little effect on the temperature.

Because parallel paths reduce the thermal resistance, in general a large surface area can dissipate heat much better than a small area. Conversely, a small surface area cannot

## *Author Biography*

Steve Winder is now a European Field Applications Engineer for Supertex Inc. Steve works alongside design engineers throughout Europe to design circuits using components made by Supertex, a US-based manufacturer of high voltage MOSFETs and CMOS ICs. A large part of his time is spent helping customers with LED driving applications.

Prior to joining Supertex in 2002, Steve was for many years a team leader at British Telecom research laboratories. Here he designed analogue circuits for wideband transmission systems, mostly high frequency, and designed many active and passive filters.

Steve holds bachelor's and master's degrees. He is a Chartered Electrical Engineer and a member of the Institute of Engineering and Technology (IET), based in London, England.