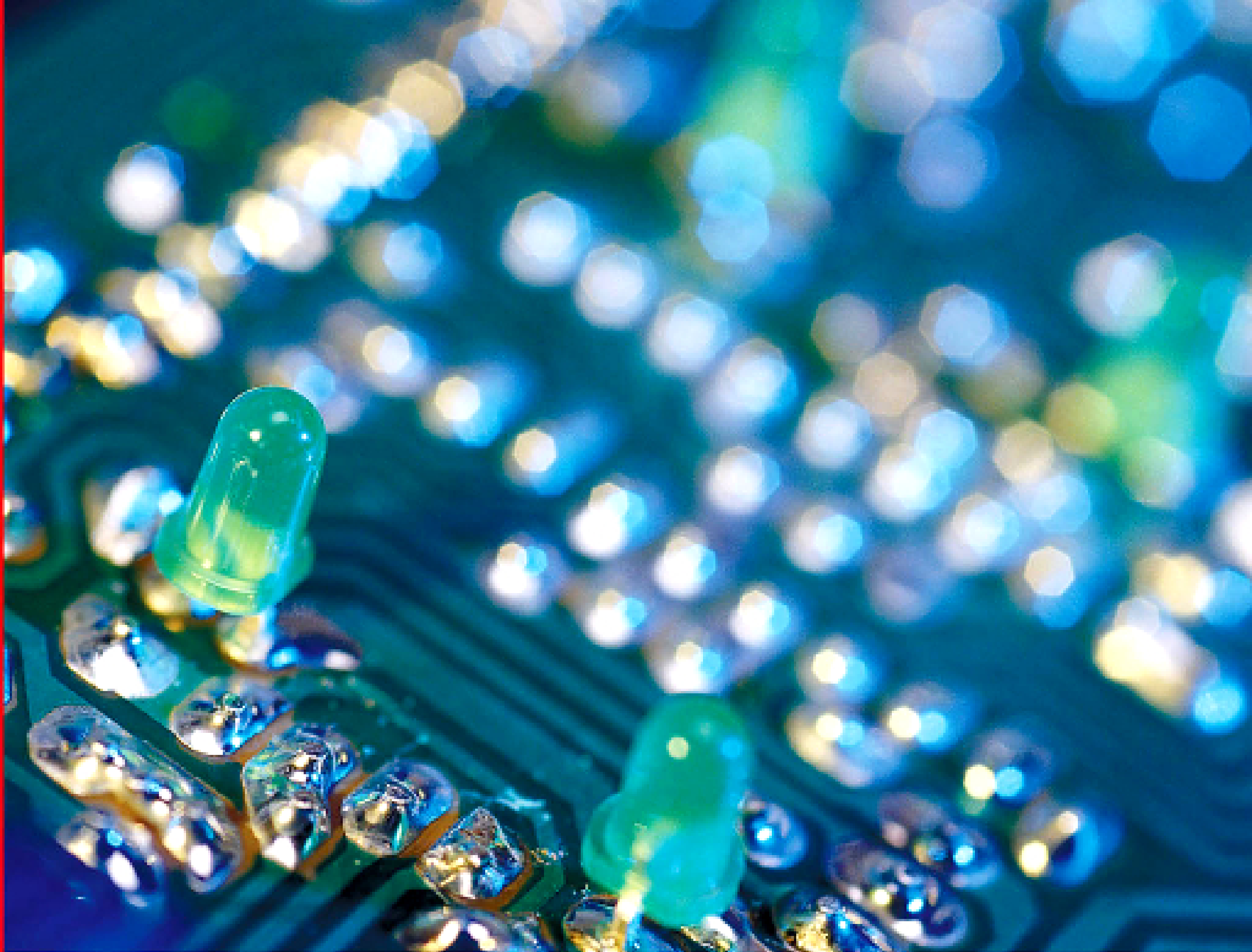

Power Supplies for LED Drivers

Steve Winder

Newnes

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Preface

The LED has been available for many years now, initially as a red colored indicator. Later, yellow/amber, green and finally blue colored LEDs became available, which triggered an explosion in applications. Applications included traffic lights, vehicle lights and wall-washes (mood lighting). Recently blue colored LEDs have been combined with yellow phosphor to create white light. The amount of light available from LEDs has also increased steadily, and now power levels of 1 W, 3 W and 5 W are fairly common.

Driving a single LED, or a long string of LEDs connected in series, has relatively few problems when the current is low (may be 20 mA). High current LEDs are tougher to drive, requiring 350 mA, 700 mA, 1 A or higher. Of course, a simple linear regulator could be used if power dissipation was not an issue, or a simple resistor if current regulation is not critical. However, in most applications, an efficient switching regulator is used. A switching regulator is essential if the LED string voltage is higher than the supply voltage, or if the supply voltage has wide variation. But switching means that electro-magnetic interference (EMI) has to be considered too.

This book describes a number of LED driving methods. The main aims of this book are: (1) to show suitable types of LED driver topologies for given applications; (2) to work through some examples; and (3) to avoid some of the mistakes that some engineers make when creating their own designs. However, the content is not exhaustive and further reading in some peripheral topics will be necessary.

Significant data to create this book have been drawn from the datasheets, application notes, training material and discussions provided by my colleagues in Supertex, particularly Rohit Tirumala and Alex Mednik.

Steve Winder, 2007.

CHAPTER 1

Introduction

As a field applications engineer for one of the pioneering developers of integrated circuits for driving power LEDs, I meet many potential customers who have little or no idea of how to drive an LED properly. The older type of LED requiring a 20 mA supply can be abused to some extent. However, power requirements have been increasing; current ratings of 30 mA, 50 mA, 100 mA, 350 mA and higher are becoming common. There are several manufacturers that produce power levels up to 20 W, and more; these higher powers use LED chip arrays. If a power LED is abused, it tends to die very quickly.

Power LEDs are being used in increasing numbers; in channel lighting (signage), traffic lights, street lights, automotive, mood lighting (colour changing 'wall wash'), theatre lighting for steps and emergency exits. Names such as HB-LEDs (high bright) and UB-LEDs (ultra-bright) are becoming meaningless as the power levels continue to rise. This book will cover all types of LED drivers, from low power to UB-LEDs and beyond.

Is power LED driving simple? No, not usually. In a few cases a linear regulator can be used, which is simple, but most cases require a switching power supply with a constant current output. Linear driving is inefficient and generates far too much heat. With a switching supply, the main issues are EMI and efficiency, and of course cost. The problem is to produce a design that meets legal requirements and is efficient, with minimal cost.

1.1 Objectives and General Approach

The approach of this book will be very practical, although some theory is introduced when necessary for understanding of later chapters. It is important to understand the characteristics of components before they can be used effectively.

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, so 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 show how that varies with the current level. But the current level determines the light output level: higher currents give higher luminosity from a given LED. The light output has the characteristic of intensity and the amount of beam spreading.

Chapter 3 will show that there are several ways to drive LEDs. Because most electronic circuits have traditionally been driven by a voltage source, it is natural for designers to continue this custom when driving an LED. The trouble is that this is not a good match for the LED power requirement. A constant current load needs a constant voltage source, but a constant voltage load (which is what an LED is) needs a constant current supply.

So, if we have a constant voltage supply, we need to have some form of current control in series with the LED. With a series resistor or active regulator circuit we are trying to create a constant current supply. In fact, a short circuit in any part of the circuit could lead to a catastrophic failure so we may have to provide some protection. Detecting an LED failure is possible using a current monitoring circuit. This could also be used to detect an open circuit. Instead of having a constant voltage supply, followed by a current limiter, it seems sensible to just use a constant current supply! There are some merits of using both constant voltage supply and a current regulator, which will be described in Chapter 4.

Chapter 3 continues describing features of constant current circuit. If we have a constant current source, we may have to provide some voltage limiting arrangement, just in case the load is disconnected. Open circuit protection can take many forms. A failure (short) would make no difference to the current level, so voltage monitoring would be a preferred failure detection mechanism. If the circuit failed open the voltage would rise up to the level of the open circuit protection limit, which could also be detected.

Chapter 4 describes linear power supplies, which can be as simple as a voltage regulator configured for constant current. Advantages include no EMI generation, so no filtering is required. The main disadvantage is heat dissipation and the limitation of having to ensure that the load voltage is lower than the supply voltage; this leads to a further disadvantage of only allowing a limited supply voltage range.

Chapter 5 describes the most basic of switching LED drivers: the buck converter. The buck converter drives an output that has a lower voltage than the input; it is a step-down topology. The reader will be taken through the design process, followed by an example design.

Chapter 6 describes boost converters. These are used in many applications including LCD backlights for television, and computer and satellite navigation display screens.

The boost converter drives an output that has a higher voltage than the input; it is a step-up topology. The reader will be taken through the design process, followed by an example design, for both continuous mode and discontinuous mode drivers.

Chapter 7 describes boost-buck converters. These have the ability to drive a load that is either higher or lower voltage compared to the input. However, this type of converter is less efficient than a simple buck or boost converter.

Chapter 8 describes specialist converters: buck-boost and buck (BBB), and Bi-Bred. These converters are intended for AC input applications, such as traffic lights, street lights and general lighting. They combine power factor correction with constant current output, but in many cases can be designed without electrolytic capacitors and so are useful for high reliability applications. This extra functionality does come at a cost – the efficiency is much lower than a standard off-line buck converter.

Chapter 9 describes fly-back converters. This chapter describes simple switching circuits that can be used for constant voltage or constant current output. The use of two windings or more in an inductor permits isolation of the output. A single winding inductor is a non-isolated buck-boost circuit that is sometimes used for driving LEDs, although the Cuk and SEPIC generally produce less EMI (at the cost of an additional inductor).

Chapter 10 covers topics that are essential when considering a switch mode power supply. The most suitable topology for an application will be discussed. The advantages, disadvantages and limitations of each type will be analyzed in terms of supply voltage range and the ability to perform PFC (power factor correction). Discussion will include snubber techniques for reducing EMI and improving efficiency, limiting switch-on surges using either in-rush current limiters or soft-start techniques.

Chapter 11 describes electronic components for power supplies. The best component is not always an obvious choice. There are so many different types of switching elements: MOSFETs, power bipolar transistors and diodes, each with characteristics that affect overall power supply performance. Current sensing can be achieved using resistors or transformers, but the type of resistor or transformer is important; similarly with the choice of capacitors and filter components.

Magnetic components are often a mystery for many electronic engineers, and these will be briefly described in Chapter 12. First, there are different materials: ferrite cores, iron dust cores and special material cores. Then there are different core shapes and sizes. One of the most important physical characteristics from a power supply design point of view is magnetisation and avoiding magnetic saturation.

EMI and EMC issues are the subjects of Chapter 13. It is a legally binding requirement that equipment should meet EMI standards. Good EMI design techniques can reduce the need for filtering and shielding, so it makes sense to carefully consider this in order to reduce the cost and size of the power supply. Meeting EMC standards is also a legal requirement in many cases. It is no use having an otherwise excellent circuit that is destroyed by externally produced interference. In many areas, EMC practices are compatible with EMI practices.

Chapter 14 discusses thermal issues for both the LEDs and the LED driver. The LED driver has issues of efficiency and power loss. The LED itself dissipates most of the energy it receives (volts times amps) as heat: very little energy is radiated as light, although manufacturers are improving products all the time. Handling the heat by using cooling techniques is a largely mechanical process, using a metal heatsink and sometimes airflow to remove the heat energy. Calculating the temperature is important because there are operating temperature limits for all semiconductors.

Another legal requirement is safety, which is covered in Chapter 15. The product must not injure people when it is operating. This is related to the operating voltage and some designers try to keep below SELV (safety extra low voltage) limits for this reason. When the equipment is powered from the AC mains supply, the issues of isolation, circuit breakers and creepage distances must be considered.

CHAPTER 2

Characteristics of LEDs

Most semiconductors are made by doping silicon with a material that creates free negative charge (N-type), or free positive charge (P-type). The fixed atoms have positive and negative charge, respectively. At the junction of these two materials, the free charges combine and this creates a narrow region devoid of free charge. This 'intrinsic region' now has the positive and negative charge of the fixed atoms, which opposes any further free charge combination. In effect, there is an energy barrier created; we have a diode junction.

In order for a P-N junction to conduct, we must make the P-type material more positive than the N-type. This forces more positive charge into the P-type material and more negative charge into the N-type material. Conduction takes place when (in silicon) there is about 0.7 V potential difference across the P-N junction. This potential difference gives electrons enough energy to conduct.

An LED is also made from a P-N junction, but silicon is unsuitable because the energy barrier is too low. The first LEDs were made from gallium arsenide (GaAs) and produced infrared light at about 905 nm. The reason for producing this color is the energy difference between the conduction band and the lowest energy level (valence band) in GaAs. When a voltage is applied across the LED, electrons are given enough energy to jump into the conduction band and current flows. When an electron loses energy and falls back into the low energy state (the valence band), a photon (light) is often emitted. See Figure 2.1.

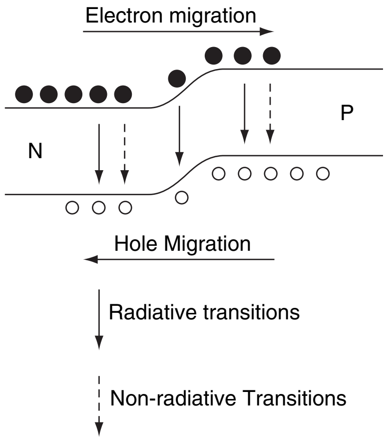


Figure 2.1: Band Diagram of P-N Junction Semiconductors.

2.1 Applications for LEDs

Soon new semiconductor materials were developed and gallium arsenide phosphide (GaAsP) was used to make LEDs. The energy gap in GaAsP material is higher than GaAs, so the light wavelength is shorter. These LEDs produced a red color light and were first just used as indicators. The most typical application was to show that equipment was powered, or that some feature such as 'stereo' was active in a radio. In fact it was mainly consumer products like radios, tape recorders and music systems that used red LEDs in large numbers.

When yellow and green LEDs became available, the number of applications increased. Now the color could change, to give additional information, or could indicate more urgent alarms. For example, green = OK, yellow = requires attention, red = faulty. Most important was the ability to have LED lamps in traffic lights.

One characteristic of the light from an LED is that it occupies a narrow spectrum about 20 nm wide; the color is fairly pure. By contrast, a semiconductor laser used for telecommunications occupies a spectrum about 2 nm wide. The very narrow spectrum of a laser is important because, when used with optical fiber systems, the narrow spectral width allows a wide system bandwidth. In general-purpose LED applications, the spectral width has very little effect.

Another important characteristic of LED light is that current is converted into light (photons). This means that doubling the current doubles the light amplitude. So dimming lights by lowering the current is possible. It should be noted that the specified wavelength emitted by an LED is at a certain current; the wavelength will change a little if the current is higher or lower than the specified current. Dimming by pulse width modulation (PWM) is a viable alternative used by many people. PWM dimming uses a signal, typical frequency 100 Hz–1000 Hz, to turn the LED on and off. The pulse width is reduced to dim the light, or increased to brighten the light.

The ‘holy grail’ was blue LEDs, which are made from indium gallium nitride (InGaN). When adding colored light, red, green and blue make light that appears white to the human eye. The reason for only ‘appearing’ white is that the eye has receptors (cones) that detect red, green and blue. There are big gaps in the color spectrum, but the eye does not notice. White LEDs are sometimes made using blue LEDs with a yellow phosphor dot over the emitting surface. The yellow phosphor creates a wide spectrum and, when combined with the blue, appears white.

An interesting application for blue LEDs is in dentistry. Illuminating modern resins used in tooth filling materials with blue light will harden the resin. The 465 nm wavelength has been found to be close to optimum for this application, although the intensity of the light must be high enough to penetrate through the resin.

Some interesting applications rely on the purity of the LED color. The illumination of fresh food is better with LEDs, because they emit no ultraviolet light. Photographic dark rooms can use colors where film is insensitive – traditionally dark rooms have been illuminated by red colored incandescent lamps. Even traffic lights must emit a limited range of colors, which are specified in national standards.

It should be noted that the color of an LED would change as the LED’s temperature changes. The temperature can change due to ambient conditions, such as being housed adjacent to hot machinery, or due to internal heating of the LED due to the amount of current flowing through it. The only way to control ambient temperature is to add a cooling fan, or by placing the LED away from the source of heat. Mounting the LED on a good heatsink can control internal heating.

The early LEDs were all rated at 20 mA and the forward voltage drop was about 2 V for red, higher for other colors; later low current LEDs were created that operated from a 2 mA source. Over time the current rating of LEDs has increased, so that 30 mA, 50 mA and even 100 mA are fairly common. Lumileds was created by HP and Philips in 1999 and produced the first 350 mA LED. Now there are a number of power LED manufacturers, rated at 350 mA, 700 mA, 1 A and higher. Power LEDs are being used in increasing numbers; in channel lighting (signage), traffic lights, street lights, automotive, mood lighting (color changing 'wall wash'), and also in theaters for lighting steps and emergency exits.

Channel lighting is so called because the LEDs are mounted in a channel; see Figure 2.2. Typically this channel is used to form letters, for illuminated company name signs. In the past, channel lighting used cold-cathode or fluorescent tubes, but these had reliability problems. Health and safety legislation, like the RoHS Directive, banned some materials like mercury that is used in the construction of cold-cathode tubes. So, to cope with the shapes and environmental conditions, the most viable technology is LED lighting.

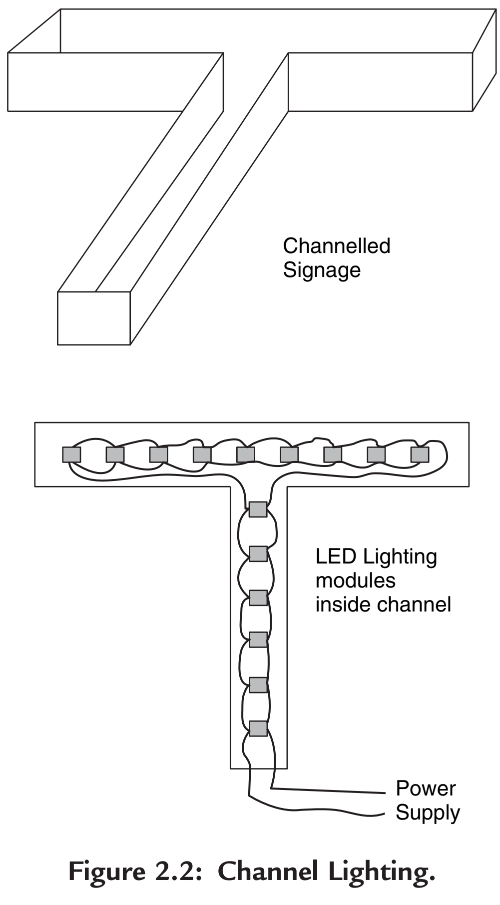


Figure 2.2: Channel Lighting.

Traffic lights have used low power LEDs for some years, but now some manufacturers are using a few high power LEDs instead. One problem with traffic lights is controlling the wavelength of the yellow (amber) light. Yellow LEDs suffer from a greater wavelength shift than other colors, and this can cause them to operate outside their permitted spectral range. Another problem is making them fail-safe – authorities permit some degree of failure, but if more than 20% of the LEDs fail, the entire lamp must be shut down and a fault reported to maintenance teams.

High ambient temperatures inside the lamp housing can lead to LED driver failures. This is particularly true if the LED driver circuit contains electrolytic capacitors, which vent when hot and eventually lose their capacitance. Some novel LED drivers have been developed that do not need electrolytic capacitors and can operate for several years at high temperatures. Failing LED drivers can give LED lights a bad name – why have LEDs that can work for over 100,000 hours if the LED driver fails after 10,000 hours' operation?

Street lights have been built using medium and high power LEDs. Although this would seem to be a simple application, high ambient temperatures and relatively high power LEDs can give rise to driver problems. In some cases, white and yellow LEDs are used together to create a 'warm-white' light. The problem with white LEDs, made using a blue LED and a yellow phosphor, is that the high blue content produces a 'cold-white' light.

Automotive lighting has many applications; internal lights, headlights, stoplights, daylight running lights (DRL), rear fog lights, reversing lights, etc. The greatest problem with automotive applications is that the EMI specifications demand extremely low levels of emissions, which are difficult to meet with a switching circuit. Linear drivers are sometimes used if the efficiency is not a critical requirement. Connecting a linear driver to the metal body of the vehicle can be used to dissipate the heat generated.

Automotive stoplights using LEDs have a significant safety advantage over those using filament lamps. The time from current flow to light output in an LED is measured in nanoseconds. In a filament lamp the response time is about 300 ms. At 60 mph (100 km/h), a vehicle travels 1 mile (1.6 km) per minute, or 88 feet per second. In 300 ms, a car will travel over 26 feet (8 meters). Stopping 300 ms sooner, having seen the previous car's brake lights earlier, could avoid death or injury. Also, LED brake lights are less likely to fail.

Mood lighting is an effect caused by changing the color of a surface and uses human psychology to control people's feelings. It is used in medical facilities to calm patients, and on aircraft to relax (or wake up!) passengers. Generally mood lighting systems use red, green and blue (RGB) LEDs in a 'wall wash' projector to create any color in the spectrum. Other applications for these RGB systems include disco lights!

Backlighting displays, such as flat screen televisions, also use RGB LED arrays to create a 'white' light. In this case the color changes little – ideally not at all. However, a control system is required to carefully control the amount of red, green and blue, to create the exact mix for accurate television reproduction. Cold cathode tubes are sometimes used to backlight computer screens, but here the exact color is not important.

2.2 Light Measure

The total light flux is measured in units of lumens. The lumen is the photometric equivalent of 1 watt, weighted to match the normal human eye response. At 555 nm, in the green-yellow part of the spectrum where the eye is most responsive, 1 W = 683 lumen.

The term candela is also used. This is the light produced by a lamp, radiating in all directions equally, to produce 1 lumen per steradian. As an equation, 1 cd = 1 lm/sr.

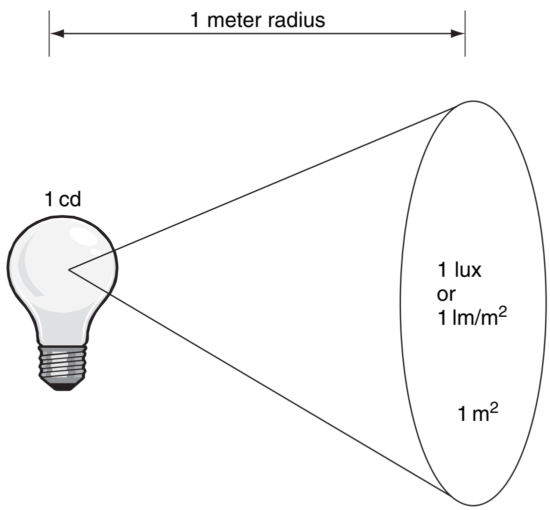


Figure 2.3: Light Measurement.

A steradian has a projected area of 1 square meter, at a distance of 1 meter from the light source. The light from a 1 cd source, at meter distance, is 1 lux, or 1 lm/m^2, see Figure 2.3.

Light emission efficiency (luminous efficacy) from LEDs is described in terms of lumens per watt. There is some competition between LED manufacturers to get the highest luminous efficacy, but when comparing results it is important to make a note of the electrical power levels used. It is easier to make an efficient 20 mA LED, than an efficient 700 mA LED.

2.3 Equivalent Circuit to an LED

An LED can be described as a constant voltage load. The voltage drop depends on the internal energy barrier required for the photons of light to be emitted, as described earlier. This energy barrier depends on the color; thus the voltage drop depends on the color. Will every red LED have the same voltage drop? No, because production variations will mean that the wavelength (color) will not be the same, and thus the voltage drop will have differences. The peak wavelength has typically a ±10% variation.

If there are temperature differences between two LEDs, this will give a color change and hence differences in voltage drop. As the temperature rises, it is easier for electrons to cross the energy barrier. Thus the voltage drop reduces by approximately 2 mV per degree as the temperature rises.

Since the semiconductor material is not a perfect conductor, some resistance is in series with this constant voltage load, see Figure 2.4. This means that the voltage drop will increase with current. The ESR (equivalent series resistance) of a low power 20 mA LED is about 20 ohms, but a 1 W 350 mA LED has an ESR of about 1–2 ohm (depending on the semiconductor material used). The ESR is roughly inversely proportional to the current rating of the LED. The ESR will have production variations too.

The ESR can be calculated by measuring the increase in forward voltage drop divided by the increase in current. For example, if the forward voltage drop increases by from 3.5 V to 3.55 V (a 50 mV increase) when the forward current goes from 10 mA to 20 mA (a 10 mA increase), the ESR will be 50 mV/10 mA = 5 ohms.

In Figure 2.4, the Zener diode is shown as a perfect device. In reality, Zener diodes also have ESR, which can be higher than the ESR of an LED. For initial testing of an LED driver, a 5 W, 3.9 V Zener diode can be used to replace the (white) LED. If the

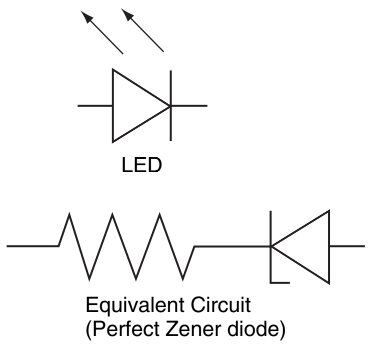


Figure 2.4: Equivalent Circuit for an LED.

driver is not working as planned the Zener diode may be destroyed, but this is far less costly than destroying a power LED. Since the Zener diode does not emit light, the test engineer will not be dazzled.

2.4 Voltage Drop Versus Color and Current

The graph in Figure 2.5 shows how the forward voltage drop depends on the light color and on the LED current. At the point where conduction begins, the forward

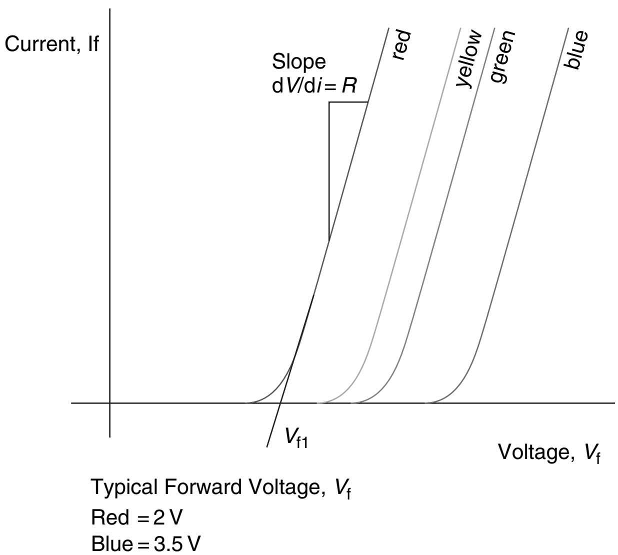


Figure 2.5: Forward Voltage Drop Versus Color and Current.

voltage drop, V_f, is about 2 V for a red LED and about 3.5 V for a blue LED. The exact voltage drop depends on the manufacturer, because of different dopant materials and wavelengths. The voltage drop at a particular current will also depend on initial V_f, but also on the ESR.

2.5 Common Mistakes

The most common mistake is to base a design on the typical forward voltage drop of the LED, $V_{f\,typ}$. This includes connecting strings of LEDs in parallel, with the assumption that the forward voltage drops are equal and the current will share equally between the two or more strings. In fact, the tolerance on the forward voltage drop is very high. For example, a 1 W white Luxeon Star has a typical $V_f = 3.42\,V$, but the minimum voltage is 2.79 V and the maximum is 3.99 V. This is over $\pm15\%$ tolerance on the forward voltage drop!

CHAPTER 3

Driving LEDs

3.1 Voltage Source

We have seen in Chapter 2 that an LED behaves like a constant voltage load with low equivalent series resistance (ESR). This behavior is like a Zener diode – in fact Zener diodes make a good test load, rather than using expensive high power LEDs!

Driving a constant voltage load from a constant voltage supply is very difficult, because it is only the difference between the supply voltage and the load voltage that is dropped across the ESR. But the ESR is very low value, so the voltage drop will also be low. A slight variation in the supply voltage, or the load voltage, will cause a very large change in current; see curve A in Figure 3.1.

If the variation in supply voltage and forward knee voltage (V_f) is known, the variation is current can be calculated. Remember that there are variations in LED

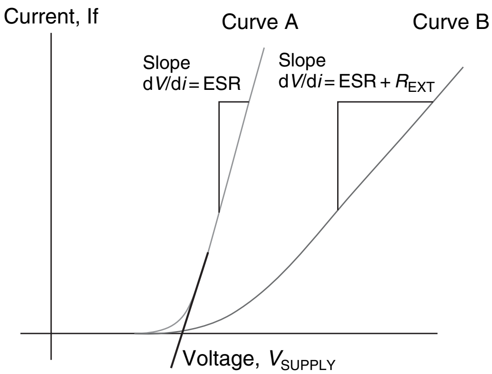


Figure 3.1: LED Current Versus Supply Voltage.

voltage drop due to manufacturing tolerances and operating temperature. Most supply voltages from a regulated supply have a 5% tolerance, but from unregulated supplies like automotive power, the tolerance is far greater.

$$I_{\text{MIN}} = \frac{V_{\text{SOURCE\_MIN}} - V_{\text{F\_MAX}}}{\text{ESR}}$$
$$I_{\text{MAX}} = \frac{V_{\text{SOURCE\_MAX}} - V_{\text{F\_MIN}}}{\text{ESR}}$$

These equations assume that ESR is constant. In practice, the V_f and voltage drop across ESR are combined, since manufacturers quote the voltage drop at a certain forward current. The actual V_f can be determined from graphs, or measured.

If there is a large difference between the source and load voltage, and a high ESR, there is very little difference between the maximum and minimum LED current. This may be perfectly adequate for low current LEDs, up to 50 mA. However, in high power LED circuits, a large voltage drop across a series resistor will be inefficient and may cause heat dissipation problems. Also, the ESR of LEDs is lower as the power rating increases. A standard 20 mA LED may have an ESR of 20 ohms, but a 350 mA LED will have an ESR of 1–2 ohms typically. Thus a 1 V difference in supply voltage could increase the LED current by 1 A in a power LED. Even in low current LEDs, the proportional change in current can be high.

3.1.1 Passive Current Control

Although the LED voltage drop shifts the curve of the graph to the right, the slope of the graph is just due to the ESR. Low current loads can have a relatively high value resistance added in series, in order to reduce the slope of the current versus voltage graph; see curve B in Figure 3.1.

With a series resistor added we are able to calculate the variation in current, provided that the variation in supply voltage and load voltage is known. In the equations below, the load voltage includes the voltage drop across ESR, at the rated current, so only the external resistor value is needed.

$$I_{\text{MIN}} = \frac{V_{\text{SOURCE\_MIN}} - V_{\text{LOAD\_MAX}}}{R_{\text{EXT}}}$$
$$I_{\text{MAX}} = \frac{V_{\text{SOURCE\_MAX}} - V_{\text{LOAD\_MIN}}}{R_{\text{EXT}}}$$

As an example, let us drive from an automotive supply; this is a nominal 13.5 V, but for this exercise we can set the limits at 12 V to 16 V. Let us select a red LED for tail-lights (Lumileds Superflux HPWA-DDOO), with a forward voltage drop of 2.19 V to 3.03 V at 70 mA forward current. Choosing to connect two LEDs in series, with a series resistor, we have a typical voltage drop of 5 V. So the typical voltage drop at 70 mA needs to be 8.5 V; this means that the series resistor should be 121.43 ohms. The nearest standard value resistor is 120 ohms, rated at 1 W since we will have a typical power dissipation of 588 mW.

$$I_{\mathrm{MIN}} = \frac{V_{\mathrm{SOURCE\_MIN}} - V_{\mathrm{LOAD\_MAX}}}{R_{\mathrm{EXT}}} = \frac{12 - 6.06}{120} = 49.5\,\mathrm{mA}$$

$$I_{\mathrm{MAX}} = \frac{V_{\mathrm{SOURCE\_MAX}} - V_{\mathrm{LOAD\_MIN}}}{R_{\mathrm{EXT}}} = \frac{16 - 4.38}{120} = 96.83\,\mathrm{mA}$$

At the high limit of source voltage, the LED is overdriven by 38%. But there is almost a 2:1 ratio between I_{MAX} and I_{MIN}, so if we increase R by 38% the worst case current levels are 70 mA maximum, but only 35.78 mA minimum.

In the previous calculations, the voltage drop across ESR (0.672 V) was included in the minimum and maximum load voltage values, so we ignored ESR. From the manufacturer's data sheet of the Lumileds HPWA-DDOO LED, graphs show that the ESR is about 9.6 ohms. Suppose we now want to operate at a lower current. Using the same example, but operating with a typical LED current of 50 mA, we must modify the results. Now the voltage at the current knee is $V_{\mathrm{f}} = 1.518$ V to 2.358 V. With a typical 13.5 V supply and 50 mA, the value for V_{f} is 1.828 V. The total resistance needed is 196.88 ohms, but we already have 9.6 ohms ESR. An external resistor value of 180 ohms is the nearest preferred value for a current of 50 mA.

$$I_{\mathrm{MIN}} = \frac{V_{\mathrm{SOURCE\_MIN}} - V_{\mathrm{LOAD\_MAX}}}{\mathrm{ESR} + R_{\mathrm{EXT}}} = \frac{12 - 4.716}{189.6} = 38.42\,\mathrm{mA}$$

$$I_{\mathrm{MAX}} = \frac{V_{\mathrm{SOURCE\_MAX}} - V_{\mathrm{LOAD\_MIN}}}{\mathrm{ESR} + R_{\mathrm{EXT}}} = \frac{16 - 3.036}{29.6} = 61.85\,\mathrm{mA}$$

The series resistor has a higher value, so the variation in current is reduced to 1.6:1 ratio. The maximum current is now below the LED current rating of 70 mA.

Unless the LEDs are matched (or 'binned') to ensure the same forward voltage drop, the current through one string could be considerably different from the current through another.

When multiple LEDs are used to provide lighting for an application, they are frequently connected in an array, consisting of parallel strings of series connected LEDs. Since the LED strings are in parallel, the voltage source for all strings is the same. However, due to variations in forward voltage for each LED, the total voltage drop of each string differs from the other strings in the array. The forward voltage also depends on the ambient temperature. To ensure uniform light output for all LEDs, equal current should be designed to flow through each string of LEDs.

The traditional way is to connect a current limiting resistor in series with each string and power all the strings using a single voltage source. A substantial voltage needs to be dropped across the resistor to ensure that the current will stay in the desired range in the presence of temperature and device-to-device voltage variations. This method is inexpensive, but suffers from power inefficiency and heat dissipation. It also requires a stable voltage source.

A better way of powering the LED array is to regulate the total current through all the strings and devise a means to divide that total current equally among the LED strings. This is active current control and is the subject of the next subsection.

3.1.2 Active Current Control

Since a series resistor is not a good current control method, especially when the supply voltage has a wide tolerance, we will now look at active current control. Active current control uses transistors and feedback to regulate the current. Here we will only consider limiting LED current when the energy is supplied from a voltage source; driving LEDs using energy from current sources will be discussed in Section 3.2.

A current limiter has certain functional elements: a regulating device such as a MOSFET or bipolar transistor; a current sensor such as a low value resistor; and some feedback (with or without gain) from the current sensor to the regulating device. Figure 3.2 shows these functions.

The simplest current limiter is a depletion mode MOSFET; it has three terminals called gate, drain and source. Conduction of the drain-source channel is controlled from the gate-source voltage, like any other MOSFET. However, unlike an enhancement MOSFET, a depletion mode MOSFET is 'normally-on' so current flows when the gate-source voltage is zero. As the gate voltage becomes negative with respect to the source, the device turns off, see Figure 3.3. A typical pinch-off voltage is –2.5 V.

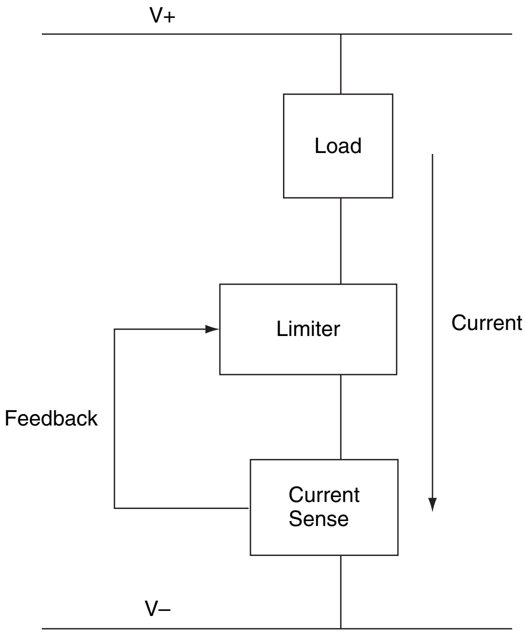


Figure 3.2: Current Limiter Functions.

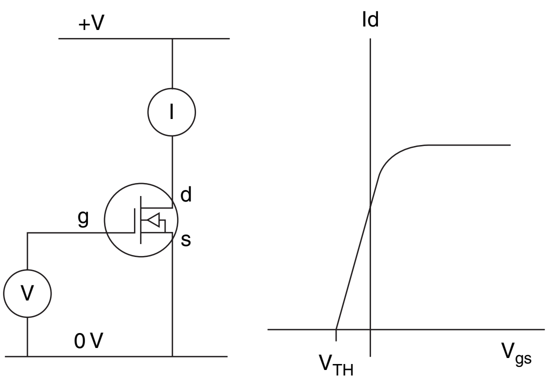


Figure 3.3: Depletion MOSFET Characteristics.

A current limiting circuit with a depletion mode MOSFET uses a resistor in series with the source to sense the current (see Figure 3.4). The gate is connected to the negative supply (0 V). As current flows through the resistor, the voltage drop across it

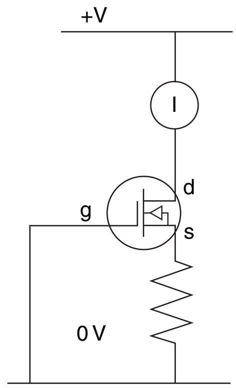


Figure 3.4: Depletion MOSFET Current Limiter.

increases. The voltage at the MOSFET source increases in potential compared to the 0 V rail and the MOSFET gate. In other words, compared to the MOSFET source, the gate becomes more negative. At a certain point, when the voltage drop approaches the MOSFET pinch-off voltage, the MOSFET will tend to turn off and thus regulate the current.

The main drawback of using depletion-mode MOSFETs is that the gate threshold voltage (V_{th}) has a wide tolerance. A device with a typical V_{th} of –2.5 V will have threshold range of –1.5 V to –3.5 V. However, the advantage is that high drain-source breakdown voltages are possible and so a limiter designed using a depletion-mode MOSFET can protect against short transients (longer periods of high voltage would tend to overheat the MOSFET).

A simple integrated current limiter is a voltage regulator in the place of the depletion-mode MOSFET, as shown in Figure 3.5. This uses an internal voltage reference and so tends to be quite accurate. The disadvantage is that there is a minimum dropout voltage of about 3 V. This circuit can be used for current sink or current source regulation, depending on whether the load is connected to the positive or negative supply rail.

The LM317 has a feedback pin called 'REF', and this controls the regulation of the current. When the voltage drop across the resistor tries to exceed 1.25 V, the current through the LM317 is reduced until the output terminal (OUT) is reduced below 1.25 V.

If accurate current limiters are used, parallel strings of LEDs can be connected to the same voltage source and then each string will have approximately the same current.

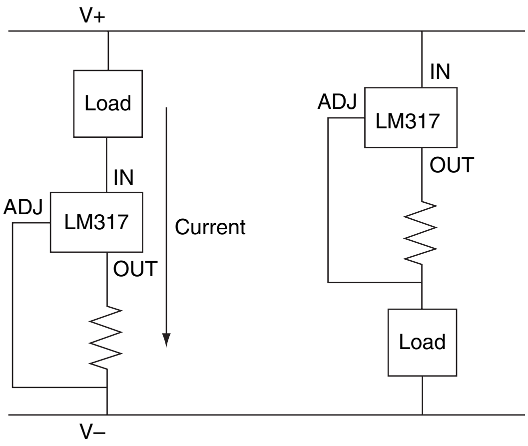


Figure 3.5: Linear Regulator as Current Limiter.

With the same current flowing through each LED, the light produced will be almost the same for each LED and thus no 'bright spots' will be seen in the LED array.

The current limiters described here are purely to show how LEDs can be driven from a constant voltage supply. Further linear regulators are described in Chapter 4. Switching regulators are described in Chapters 5–10.

3.1.3 Short Circuit Protection

The current limiting circuits described in the previous subsection will provide automatic short circuit protection. If the LED goes short circuit, a higher voltage will be placed across the current limiter. Power dissipation is the main issue that needs to be addressed.

If the power dissipation cannot be tolerated when the load goes short circuit, a voltage monitoring circuit will be needed. When a higher than expected voltage is placed across the current limiter, the current must be reduced to protect the circuit. In the LM317 circuit previously described, the regulator itself has thermal shutdown.

3.1.4 Detecting Failures

If we have a short circuit condition in the LEDs, the voltage across the current limiter will increase. We can use this change to detect a failure. In the circuit shown in

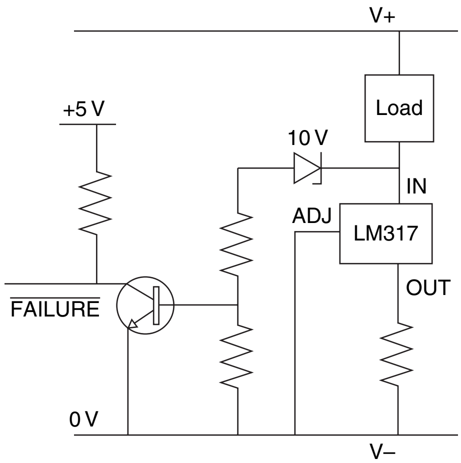


Figure 3.6: Shorted Load Indication.

Figure 3.6, a 10 V Zener diode is used in series with the base of an NPN transistor. When the voltage at the 'IN' terminal of the LM317 reaches about 11 V, the Zener diode conducts and turns on the transistor. This pulls the 'FAILURE' line to 0 V and indicates a short circuit across the LEDs.

3.2 Current Source

Since an LED behaves like a constant voltage load, it can be directly connected to a current source. The voltage across the LED, or string of LEDs, will be determined by the characteristics of the LEDs used. A pure current source will not limit the voltage, so care must be taken to provide some limit; this will be covered in more detail in the next subsection.

If the current source produces much more current than the LED requires, current-sharing circuits will be required. The simplest of these is a current mirror, which shares the current equally between strings based on the current flowing through the primary string.

Figure 3.7 shows a simple current mirror. The basic principle relies on the fact that matched transistors will have the same collector current if their base-emitter junctions have the same voltage across them. By connecting all the bases and all the emitters

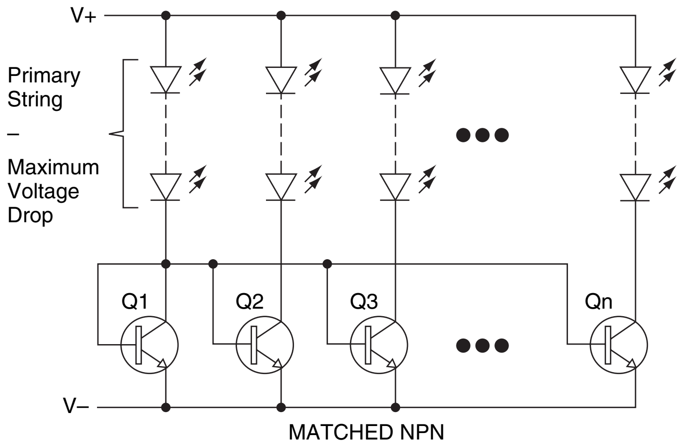


Figure 3.7: Current Mirror.

together, every base-emitter junction voltage must be equal and therefore every collector current must be equal.

The primary LED string is the one that controls the current through the other strings. Since the collector and base of transistor Q_1 are connected, the transistor will be fully conducting until the collector voltage falls low enough for the base-emitter current to limit. Other transistors (Q_2 to Q_n) have their base connections joined to Q_1, and will conduct exactly the same collector current as Q_1 since the transistors are matched. The total current through Q_1 to Q_n will equal the current source limit.

The voltage drop across the LEDs in the primary string must be higher than any other string in order for the current mirror to work correctly. In the slave strings, some voltage will be dropped across the collector-emitter junction of the transistors $Q_2 - Q_n$. The slave circuits adjust the current by raising or lowering this surplus voltage drop across the transistor.

3.2.1 Self-Adjusting Current Sharing Circuit

As an alternative, the current sharing circuit shown in Figure 3.8 automatically adjusts for string voltage.

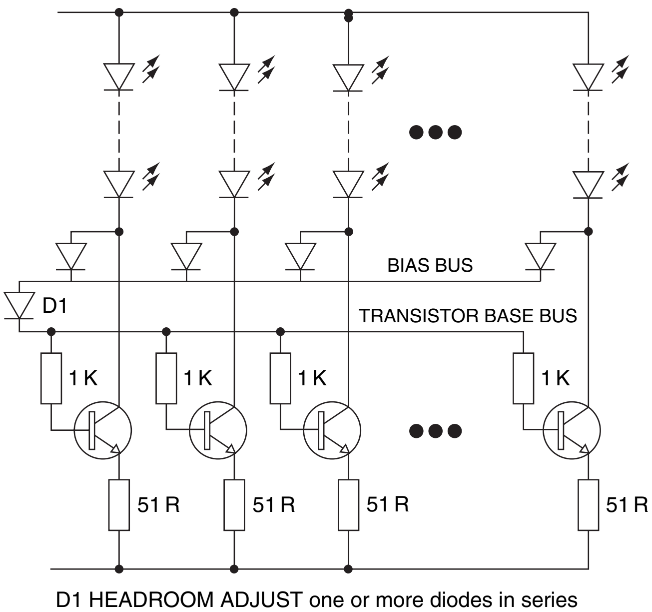


Figure 3.8: Self-Adjusting Current Sharing Circuit.

Assuming that the LED array is driven from a current source, there will be equal current division among all connected branches. If any branch is open due to either a failure or no connection by design, the total current will divide evenly among the connected branches. Unlike the simple current mirror, this one automatically adjusts for the maximum expected voltage difference between strings of LEDs, which is a function of the number of LEDs in the string and the type of LED used. The components must be able to dissipate the heat generated by the sum of each string current and the headroom voltage drop across the regulator for that string.

In high reliability applications, the failure of a single LED should not materially affect the total light output. The use of the current divider will help the situation. When an LED fails short, the voltage of the string containing the shorted LED will have less voltage. The current divider will accommodate the change in voltage and still distribute the current equally. When the failed LED string opens, the current divider will automatically redistribute the total current among the remaining strings, thus maintaining the light output. In this application, an extra diode string can be added for redundancy, so that any single failure will not cause the remaining LEDs to operate in an over-current condition.

Equality of current division among the branches is dependent on the close matching of the transistors, which are in close vicinity (ideally a single package with several matched transistors). When any of the transistors saturate due to large variation of the string voltages, equal current division will be lost.

Diodes connected to each collector detect the voltage of each branch. The highest branch voltage (corresponding to the LED string with the lowest forward voltage) is used to bias the transistors in the linear operating region. The cathode of each diode is connected to a common 'bias bus'.

To accommodate variations in string voltages and keep the current divider transistors from saturation, diodes are connected between the 'bias bus' and the 'transistor base bus'. More than one external diode can be used to accommodate large voltage variations. If the string voltage variation is less than one diode drop, the two buses can be joined.

When a branch is not connected, there will be higher base current flowing in the associated regulating transistor. This could interfere with the current division in the connected branches, so a resistor (about 1 kohm) is connected from the 'transistor base bus' to each transistor base to ensure correct operation of the overall circuit.

3.2.2 Voltage Limiting

In theory, the output voltage of a constant current driver is not limited. The voltage will be the product of the current and load resistance in the case of a linear load. In the case of an LED load, the voltage limit will depend on the number of LEDs in a string.

In practice, there will be a maximum output voltage, because components in the current source will break down eventually. Limiting the LED string voltage is necessary to prevent circuit damage and the voltage level will depend on the particular circuit.

Safety regulations will be covered in Chapter 10, but Underwriters Laboratories (UL) Class 2 and Safety Electrical Low Voltage (SELV) requirements limit any potential to 60 V DC, or 42.4 V AC, so equipment designed to meet these requirements should consider both mains supply isolation (if applicable) and output voltage limiting. The number of LEDs in a string will be restricted in this case, so that the total string voltage remains below 60 V.

3.2.3 Open Circuit Protection

Some constant current drivers, especially switching boost converters, will produce a sufficiently high voltage to destroy the driver circuit. For these types of driver a shutdown mechanism is required. Using a Zener diode to give feedback when the output voltage exceeds a certain limit is the standard method. Some over-voltage detectors within integrated circuits (ICs) have a latched output, requiring the power supply to be turned off and then on again before LED driver functions are enabled. Other circuits will auto-restart when the open circuit condition is removed (i.e. when the LEDs are reconnected).

Some ICs have an over-voltage detector (internal comparator) that disables the LED driver circuit when the voltage at the input exceeds the reference voltage. A potential divider comprising two resistors is usually used to scale down the output voltage to the reference voltage level.

3.2.4 Detecting LED Failures

In a constant current circuit, a failure of an LED can mean that either a whole string is off (open circuit LED) or a single LED is off (short circuit LED).

In the case of an open circuit LED, the load is removed and so the output voltage from the current source rises. This rise in voltage can be detected and used to signal a failure. In circuits where over-voltage protection is fitted, this can be used to indicate a failure.

If a current mirror is used to drive an array of LEDs with a number of strings, the result of an open circuit LED will depend on which string the LED is located. In a basic current mirror, as shown in Figure 3.7, a failure in the primary string will cause all the LEDs to have no current flow and not be lit. Detection of the rise in output voltage would be a solution. However, if the failure were in a secondary string, there would be higher current flowing in the other strings and the output voltage would not rise very much (only due to the extra current flowing through the ESR). The voltage at the transistor collector of the broken string would fall to zero since there is no connection to the positive supply, and this could be detected.

Another technique, for low current LEDs, is to connect the LED of an opto-coupler in series with the LED string. A basic opto-coupler has an LED and a photo-transistor in the same package. Current through the opto-coupler LED causes the photo-transistor to conduct. Thus when current is flowing through the LED string 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.

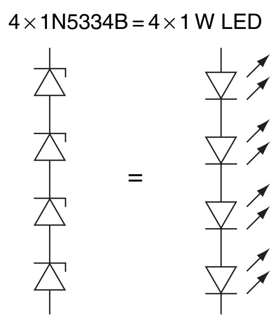


Figure 3.9: Zener Diode Dummy Load.

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 impedance. Commercial active loads can be set to have constant current or constant voltage – a constant voltage setting is required to simulate an LED load.

A constant voltage load built using a low cost discrete solution is shown in Figure 3.10. This is a self-powered load and so can be isolated from ground. The Zener diode can be selected to give the desired voltage (add 0.7 V for the emitter-base junction of the transistor). The transistor should be a power device, mounted on a heatsink.

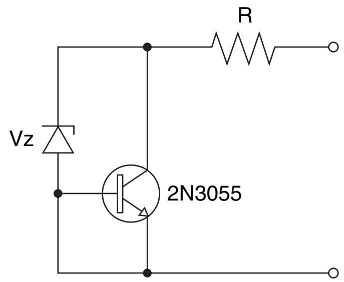


Figure 3.10: Active Dummy Load.

The circuit is Figure 3.10 has low impedance. Although the Zener diode does have a few ohms impedance, the current through it is very small and the effect of the transistor is to reduce the impedance by a factor equal to the gain HFE. Suppose the transistor HFE = 50 at 1 A and the Zener diode impedance $Z_d = 3$ ohms. Changing the collector current from 500 mA to 1 A will cause the base current to rise from 10 mA to 20 mA. A 10 mA change in current through the Zener diode will cause 30 mV voltage rise. This change at the transistor collector is equivalent to an impedance of 30 mV/0.5 A = 0.06 ohm. In other words, the circuit impedance is equal to the Zener diode impedance divided by the transistor gain.

An impedance of 0.06 ohm is unrealistically low, but a power resistor can be added in series to give the desired load impedance. Because of the potentially high load current, both the transistor and series resistor should be rated for high power. The transistor should be mounted on a large heatsink.

3.4 Common Mistakes

The most common mistake is to use expensive high power LEDs when testing a prototype circuit. Instead, 3.6 V, 5 W Zener diodes should be used in place of each LED. Only once the circuit has been tested under all conditions should LEDs be used.

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.

CHAPTER 4

Linear Power Supplies

4.1 Introduction

Linear power supplies for driving LEDs are preferred for a number of reasons. The complete absence of any EMI radiation is one important technical reason. Lowest cost is an important commercial reason. However, they also have disadvantages: in some applications they have low efficiency and hence the introduction of thermal problems; in other applications, such as when powered from the AC mains supply, they have the disadvantage of large size.

4.1.1 Voltage Regulators

Many voltage regulators are based on the LM317 originally from National Semiconductor, but which is now made by a number of manufacturers. Inside the LM317 are: (1) a power switch, which is an NPN transistor; (2) a voltage reference set to produce 1.25 V and (3) an operational amplifier (op-amp) to control the power switch, as shown in Figure 4.1. The op-amp tries to keep the voltage at the output equal to the voltage at the adjust (ADJ) pin minus the reference voltage.

To produce a certain output voltage, a feedback resistor is connected from the output (OUT) to the ADJ pin and a sink resistor is connected from the ADJ pin to ground, thus creating a potential divider. Usually the feedback resistor is set to 240 ohms, in order to draw a minimum of 5 mA from the regulator and help to maintain stability.

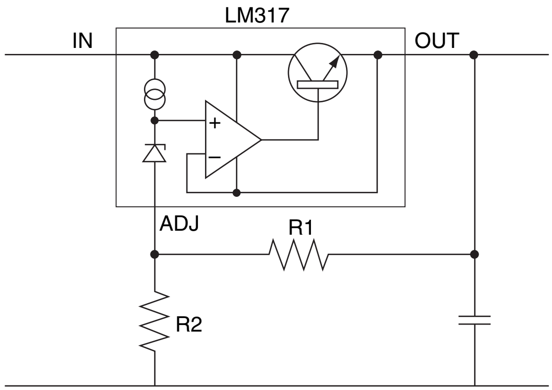


Figure 4.1: LM317 Regulator.

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

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

Note, $I_{\mathrm{ADJ}} = 100\,\mu\mathrm{A}$, 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.

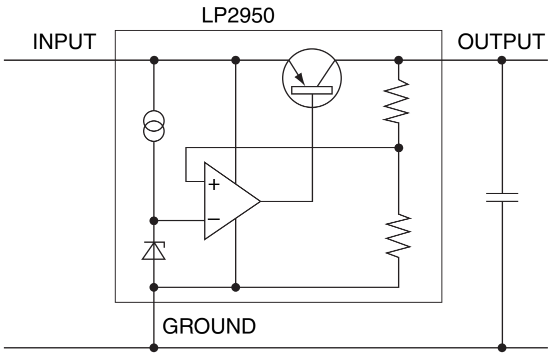


Figure 4.2: Low Dropout Voltage Regulator.

4.1.2 Voltage Regulators as Current Source or Sink

In Figure 4.3 are shown two circuits using a voltage regulator as a current limiter, one is configured as a current source and the other as a current sink.

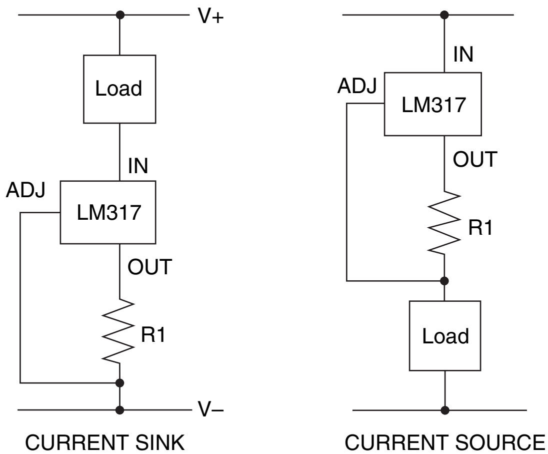


Figure 4.3: Constant Current Circuits Using the LM317.

As previously described, the LM317 regulates when there is +1.25 V between the OUT and ADJ pins. In Figure 4.3, a current sense resistor ($R1$) is connected between the OUT and ADJ pins. Current flowing through $R1$ will produce a voltage drop, with the

OUT pin becoming more positive than the ADJ pin. When the voltage drop across $R1$ reaches 1.25 V, the LM317 will regulate the current. Thus the current limit is

$$I = \frac{1.25}{R1}$$

.

4.1.3 Constant Current Circuits

There are many constant current circuits; some using integrated circuits, some using discrete components, and others using a combination of both ICs and discrete devices. In this subsection, we will examine some examples of each type.

A simple constant current sink uses an op-amp with an input voltage range that extends to the negative rail, as shown in Figure 4.4. In order to set the current level, a voltage reference is required. The voltage drop across a current sensing resistor is compared to the reference voltage and the op-amp output voltage rises or falls to control the current. The voltage reference can be a temperature compensated precision reference, or a Zener diode. A Zener diode generally has a smallest temperature coefficient and lowest dynamic impedance at a breakdown voltage of 6.2 V.

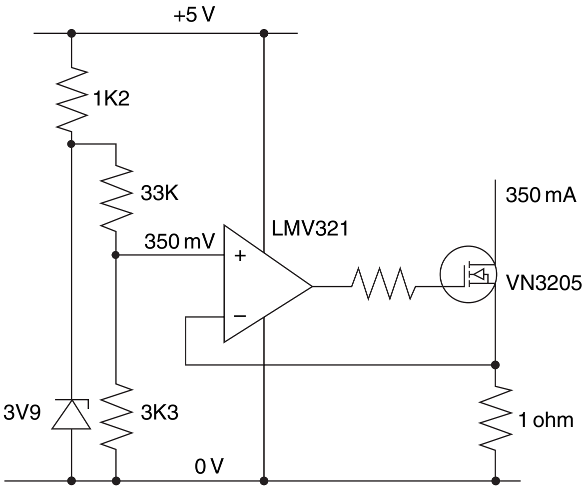


Figure 4.4: Constant Current Sink Using Op-amp.

4.2 Advantages and Disadvantages

The advantage of linear power supplies is that they produce no EMI radiation. This advantage cannot be overstated.

A switching power supply may appear to have few components, but this does not take into account the EMI filtering and screening. These additional circuits can double the overall cost of the LED driver. If the LEDs are distributed, such as in channel lighting where there is no opportunity to shield any EMI, both common mode and differential filtering are required. And common mode chokes are expensive!

One disadvantage of a linear LED driver can be low efficiency, which is the ratio of the LED voltage to the supply voltage. The efficiency is low only if the supply voltage is somewhat higher than the LED voltage. In these cases, poor inefficiency causes the introduction of thermal problems. A heatsink may be required, which is bulky and moderately expensive. It should be noted that where the supply voltage is only a little higher than the LED voltage, the efficiency of a circuit using linear regulator could be higher than one using a switching regulator.

Linear mains powered LED drivers have the disadvantage of large size, because a step-down transformer is almost always required (unless the LED string voltage is very near to the peak AC supply voltage). A 50 Hz or 60 Hz mains transformer is bulky and heavy. Smoothing capacitors after the bridge rectifier are also very bulky. The efficiency will vary as the AC supply voltage rises and falls over a long period, because the difference between the rectified voltage and the LED string voltage will change.

4.3 Limitations

The main limitation of a linear supply is that the LED voltage will always be lower than the supply voltage. Linear voltage and current sources cannot boost the output voltage so that the output is higher than the input. Where the output voltage could be higher than the input voltage a switching regulator is necessary. These will be discussed in the next few chapters.

4.4 Common Errors in Designing Linear LED Drivers

The most common error is to ignore the power dissipation. Power dissipation is simply the voltage drop across the regulator multiplied by the current through it.

If the voltage drop is high, the current must be limited to stay within the device package power dissipation limits. A surface mount D-PAK package may be limited to about 1 W, even when there is some copper area soldered to the tab terminal. Heatsinks are now available for surface mount packages, which eases the problem.

Another error is to ignore the start-up conditions. The voltage rating of the regulator must be high enough to allow for the output being connected to 0 V (ground). This is because at start-up, the output capacitor will be uncharged and thus at 0 V. Only after operating for a short period does the output capacitor charge, which reduces the voltage drop across the regulator. The voltage rating of the regulator should always be greater than the maximum input voltage expected.

CHAPTER 5

Buck-Based LED Drivers

The first switching LED driver that we will study is the buck converter. The buck converter is the simplest of the switching drivers, and is a step-down converter for applications where the load voltage is never more than about 85% of the supply voltage. The limit of about 85% is due to switching delays in the control system. In a buck converter circuit, a power MOSFET is usually used to switch the supply voltage across an inductor and LED load connected in series. The inductor is used to store energy when the MOSFET is turned on; this energy is then used to provide current for the LED when the MOSFET is turned off. A diode across the LED and inductor circuit provides a return path for the current during the MOSFET off time. A simple schematic is shown in Figure 5.1.

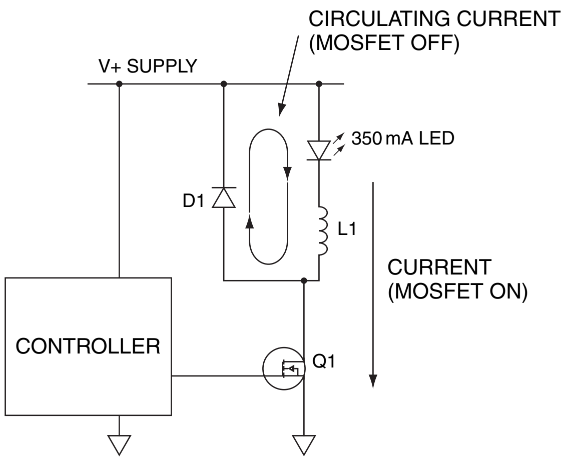


Figure 5.1: Buck LED Driver.

Buck converters are an attractive choice for LED drivers in offline and in low voltage applications as they can produce a constant LED current at very high efficiencies. A peak-current-controlled buck converter can give reasonable LED current variation over a wide range of input and LED voltages and needs no design effort in feedback control design. Coupled with the fact that these converters can be designed to operate at above 90% efficiencies, the buck-based driver becomes an attractive solution to drive high brightness LEDs.

5.1 An Example Buck Converter Control IC

The Supertex HV9910B integrated circuit was designed especially for LED driving. It is a good example of a low cost, low component count solution to implement the continuous mode buck converter (the IC itself needs just three additional components to operate). Linear or PWM dimming can also be easily implemented using the IC. A diagram of the HV9910B is shown in Figure 5.2.

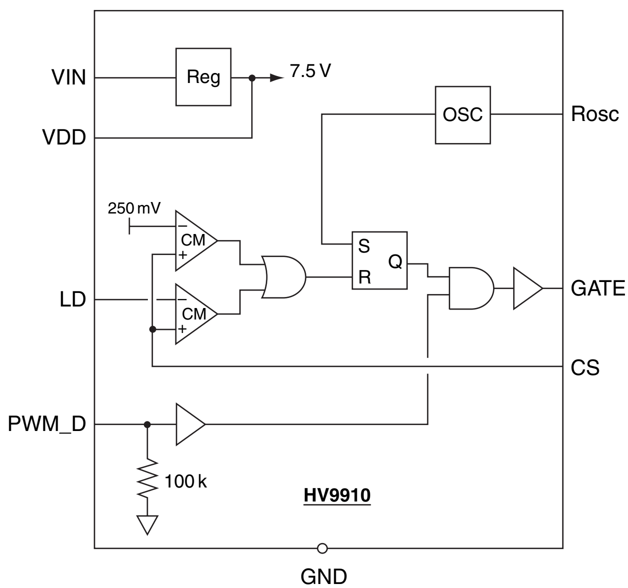


Figure 5.2: Supertex HV9910B.

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.

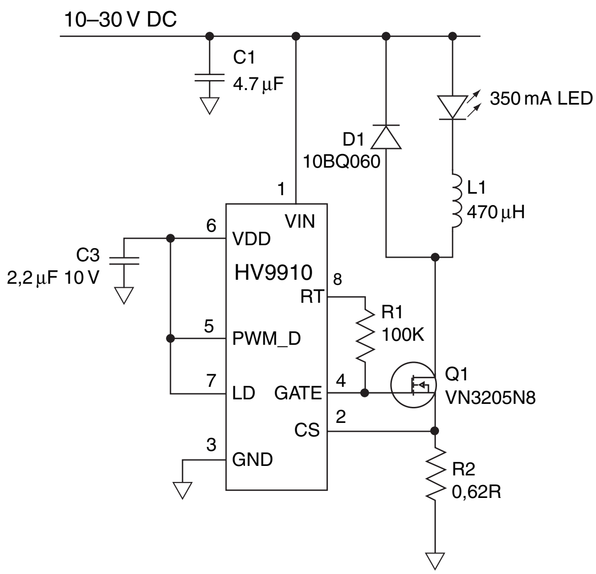


Figure 5.3: Buck Converter for DC Applications.

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 ($R1$)

The switching frequency determines the size of the inductor $L1$. 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 datasheet, the timing resistor between the RT pin and ground that is needed to achieve this frequency is 150 kΩ.

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 μs, with an 80% duty cycle the on-time will be 20 μs. 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 μs off-time will be 100 kΩ.

5.2.3 Choosing the Input Capacitor (C1)

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

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

In this design example, the high frequency capacitance required is about 4.7 µF 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, $E = V_{LED} = V_{o,max} = L \times \frac{di}{dt}$. Another way of writing this is $L = V_{o,max} \times \frac{dt}{di}$. Here, di is the ripple current $= 0.3 \times I_{o,max}$ and dt is the off-time.

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

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

In this example, $L1 = 380$ µH and the nearest standard value is 470 µH. 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).

5.2.5 Choosing the MOSFET ($Q1$) and Diode ($D2$)

The peak voltage seen by the MOSFET is equal to the maximum input voltage. Using a 50% safety rating,

$$V_{FET} = 1.5 \times 30\,V = 45\,V$$

The maximum RMS current through the MOSFET depends on the maximum duty cycle, which is 80% in our example. Hence, the current rating of the MOSFET is $I_{FET} \approx I_{o,max} \times 0.8 = 0.28\,A$.

Typically a MOSFET with about three times the current is chosen to minimize the resistive losses in the switch. For this application, choose a 50 V, >1 A MOSFET; a suitable device is a Supertex part, VN3205N8, rated at 50 V 1.5 A.

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

$$V_{diode} = V_{FET} = 45\,V$$

The average current through the diode under worst case conditions (minimum duty cycle) is

$$I_{diode} = 0.87 \times I_{o,max} = 0.305\,A$$

Choose a 60 V, 1 A Schottky diode. The International Rectifier 10BQ060 is a suitable type.

5.2.6 Choosing the Sense Resistor ($R2$)

The sense resistor value is given by

$$R2 = \frac{0.25}{1.15 \times I_{o,max}}$$

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. Note that the current limit is set to 15% above the maximum required current, due to the total 30% ripple specified.

For this design, $R2 = 0.625\,\Omega$. The nearest standard value is $R2 = 0.62\ \Omega$.

If a standard value is not close to the value calculated, or if a lower power dissipation in the sense resistor is required (perhaps to increase efficiency), a potential divider can be connected to the LD pin to set it at a lower voltage. Say we want to use a $0.47\,\Omega$ resistor; then we would scale the 0.25 V at the LD pin by $0.47/0.625 = 0.752$, so that it becomes 188 mV.

Note that capacitor $C3$ is a bypass capacitor for holding up the HV9910B internal supply V_{DD} during MOSFET switching, when high frequency current pulses are required for charging the gate. A typical value for $C3$ of 2.2 µF, 16 V is recommended, although in this design the MOSFET gate charge is very low, so a 1 µF, 16 V can be used instead.

5.2.7 Common Errors in Low Voltage Buck Design

1. Using an inductor that has too high inductance.

   Although increasing the inductor value may seem to be the answer to reduce current ripple, it actually causes problems because the current does not fall enough between switching cycles for proper control by the controller IC. The voltage seen across the current sense resistor at switch-on will be almost at the current sense comparator reference voltage. At switch-on there will be a current surge, caused by the flywheel diode reverse current and the current through the inductor's parasitic capacitance. The smallest current surge will create a voltage spike across the current sense resistor and hence the current sense comparator will trip. This means that the MOSFET will switch off almost immediately after switch-on.

   A typical switching pattern is one proper switching cycle, where energy is stored in the inductor, followed by one short switching pulse. This switching pulse provides very little energy to the inductor, but generates high switching losses. The result is a less efficient circuit that could suffer from overheating and EMI problems.

2. Using the wrong type of flywheel diode.

   A Schottky diode has a low forward voltage drop, which will give low power dissipation. However, in low duty cycle applications the LED current is flowing in the flywheel diode most of the time. A forward voltage of say 0.45 V at 350 mA results in 157.5 mW conduction losses, so an SMA size package works well, but for higher current applications a large SMB or SMC package should be considered. Note that the forward voltage drop of Schottky diodes increases with their current rating, so a 30 V Schottky has much lower V_f than a 100 V Schottky.

5.3 Buck Circuits for AC Input

I will now discuss the design of a buck-based LED driver using the HV9910B with the help of an AC mains input application example. The same procedure can be used to design LED drivers with other input voltage ranges. The schematic is shown in Figure 5.4.

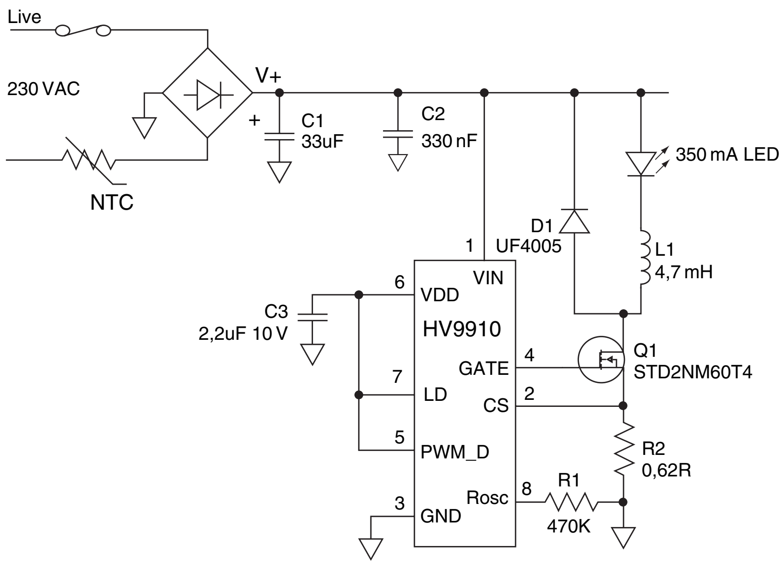


Figure 5.4: Universal Mains Input Buck Circuit.

Designs for an AC input have two problem areas to address. In addition to considering the LED driving aspects, we must also consider the low frequency and, usually, high voltage supply. Because we are applying a low frequency sinusoidal high voltage supply, high value input capacitors are needed to hold up the supply voltage during the cusps between each half-cycle of the input. Applying high voltage across high value capacitors creates a large inrush current that can cause damage, so an inrush limiter (negative temperature coefficient thermistor) is required.

5.3.1 Target Specification

Input voltage = 90 V to 265 V AC (nominal 230 V AC)

LED string voltage = 20–40 V

LED current = 350 mA

Expected efficiency = 90%

5.3.2 Choosing the Switching Frequency and Resistor (*R*1)

The switching frequency determines the size of the inductor $L1$. A larger switching frequency will result in a smaller inductor, but will increase the switching losses in the circuit. A typical switching frequency for high input voltage applications is $f_s = 80\,\text{kHz}$, which is a good compromise. From the HV9910B datasheet, the timing resistor needed to achieve this is 470 kΩ.

5.3.3 Choosing the Input Diode Bridge (*D*1) and the Thermistor (NTC)

The voltage rating of the diode bridge will depend on the maximum value of the input voltage. A 1.5 multiplication factor gives a 50% safety margin.

$$V_{\text{bridge}} = 1.5 \times \left(\sqrt{2} \times V_{\text{max,ac}}\right) = 562\,\text{V}$$

The current rating will depend on the highest average current drawn by the converter, which is at minimum input voltage (DC level, allowing for a 'droop' across the input capacitor between the AC line voltage peaks) and at maximum output power. The minimum input voltage must be more than half the maximum LED string voltage, to make sure that the duty cycle stays below 50% and thus remains stable. For this example, the minimum rectified voltage should be

$$V_{\text{min,dc}} = 2 \times \text{V}_{\text{o,max}} = 80\,\text{V}.$$

$$I_{\text{bridge}} = \frac{V_{\text{o,max}} \times I_{\text{o,max}}}{V_{\text{min,dc}} \times \eta} = \frac{14}{72} = 0.194\,\text{A}$$

For this design, using a 230 V AC supply, choose a 600 V 1 A diode bridge.

The thermistor should limit the inrush current to not more than five times the steady state current, assuming maximum voltage is applied. The required cold resistance is:

$$R_{\text{cold}} = \frac{\sqrt{2} \times V_{\text{max,ac}}}{5 \times I_{\text{bridge}}}$$

This gives us a 380 Ω resistance at 25°C. The calculations suggest that we choose a thermistor whose resistance is around 380 Ω and RMS current greater than 0.2 A, but in practice a 120 Ω thermistor rated at 1 A would suffice.

5.3.4 Choosing the Input Capacitors (C1 and C2)

The first design criterion to meet is that the maximum LED string voltage must be less than half the minimum input voltage. This is to satisfy the stability requirements when operating at a constant switching frequency. As we have already seen, the minimum rectified voltage should be

$$V_{\text{min,dc}} = 2 \times V_{\text{o,max}} = 80\,\text{V}$$

The hold-up capacitor required at the output of the diode bridge will have to be calculated at the minimum AC input voltage. The capacitor can be calculated as

$$C1 \geq \frac{V_{\text{o,max}} \times I_{\text{o,max}}}{\left(2 \times V^2_{\text{min,ac}} - V^2_{\text{min,dc}}\right) \times \eta \times freq}$$

In this example,

$$C1 \geq 26.45\,\mu\text{F}$$

The voltage rating of the capacitor should be more than the peak input voltage.

$$V_{\text{max,cap}} \geq \sqrt{2} \times V_{\text{max,ac}}$$
$$\Rightarrow V_{\text{max,cap}} \geq 375\,\text{V}$$

Choose a 450 V, 33 µF electrolytic capacitor.

The 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, a metallized polypropylene capacitor is needed in parallel with the electrolytic capacitor to absorb the high frequency ripple current. The required high frequency capacitance can be computed as

$$C2 = \frac{I_{\text{o,max}} \times 0.25}{f_{\text{s}} \times (0.05 \times V_{\text{min,dc}})}$$

In this design example, the high frequency capacitance required is about 0.33 µF, 400 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.

5.3.5 Choosing the Inductor (*L*1)

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 $E = L \times \frac{\mathrm{d}i}{\mathrm{d}t}$. Considering the time when the MOSFET switch is off, so that the inductor is supplying energy to the LEDs, $E = V_{\text{LED}} = V_{\text{o,max}} = L \times \frac{\mathrm{d}i}{\mathrm{d}t}$. Another way of writing this is $L = V_{\text{o,max}} \times \frac{\mathrm{d}t}{\mathrm{d}i}$.

Here, $\mathrm{d}i$ is the ripple current $= 0.3 \times I_{\text{o,max}}$ and $\mathrm{d}t$ is the off-time $\mathrm{d}t = \frac{\left(1 - \frac{V_{\text{o,max}}}{\sqrt{2} \times V_{\text{ac,nom}}}\right)}{f_{\text{s}}}$. Note, a buck circuit duty cycle is given by $D = \frac{V_{\text{out}}}{V_{\text{in}}}$, so the off-time is $\mathrm{d}t = \frac{(1-D)}{f_{\text{s}}}$.

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

$$L1 = \frac{V_{\text{o,max}} \times \left(1 - \frac{V_{\text{o,max}}}{\sqrt{2} \times V_{\text{ac,nom}}}\right)}{0.3 \times I_{\text{o,max}} \times f_{\text{s}}}$$

In this example, $L1 = 4.2\,\text{mH}$. The nearest standard value is 4.7 mH. 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).

Note that with a large inductance value, the parasitic capacitance across the coil could be significant and will affect switching losses.

5.3.6 Choosing the MOSFET (*Q*1) and Diode (*D*2)

The peak voltage seen by the MOSFET is equal to the maximum input voltage. Using a 50% safety rating,

$$V_{\text{FET}} = 1.5 \times \left(\sqrt{2} \times 265\right) = 562\,\text{V}$$

The maximum RMS current through the MOSFET depends on the maximum duty cycle, which is 50% by design. Hence, the current rating of the MOSFET is

$$I_{\text{FET}} \approx I_{\text{o,max}} \times \sqrt{0.5} = 0.247\,\text{A}$$

Typically a MOSFET with about three times the current is chosen to minimize the resistive losses in the switch. For this application, choose a 600 V, >1 A MOSFET; a suitable device is an ST part, STD2NM60, rated at 600 V 2 A. This MOSFET has 2.8 Ω on-resistance. With 350 mA being passed up to 50% of the time, the conduction losses will be 171 mW.

Although a MOSFET with lower on-resistance could be used to reduce the conduction losses, the switching losses, which are caused by parasitic capacitance and diode reverse recovery current, will then be higher. The diode D2 passes current in the reverse direction for a short period: imagine a mechanical value that is passing a fluid – when the pressure reverses it takes a short time for the valve to close and shut off the reverse flow. The analogy can be applied to diodes, because they have free electrons in their conduction band that have to be swept out by the reverse potential before current flow stops. Each time the MOSFET turns on, a current spike passes 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 that 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

$$R2 = \frac{0.25}{1.15 \times I_{\text{o,max}}}$$

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, $R2 = 0.625\,\Omega$. The nearest standard value is $R2 = 0.62\,\Omega$.

Note that capacitor *C*3 is a bypass capacitor for holding up the HV9910B internal supply V_{DD} during MOSFET switching, when high frequency current pulses are required for charging the gate. A typical value for *C*3 of 2.2 µF, 16 V is recommended, although for AC applications smaller capacitors as low as 0.1 µF 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.

5.4 Buck Circuits Powered by an AC Phase Dimmer

An LED driver powered by an AC phase dimmer needs special additional circuits. These additional circuits are required because of the phase dimmer circuit. Phase dimmers usually use a triac activated by a passive phase shift circuit. Because of switching transients, which would otherwise cause serious EMI problems, the triac is bypassed by a capacitor (typically 10 nF) and has an inductor in series with its output. The phase dimmer circuit is shown in Figure 5.5.

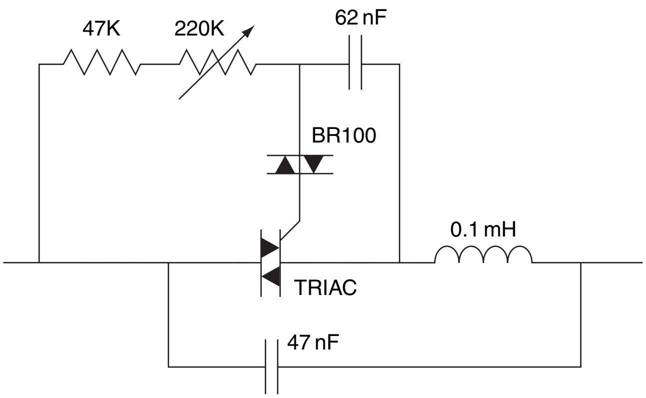


Figure 5.5: Phase Dimmer Circuit.

The input of an inactive LED driver is high impedance, with a large capacitor on the DC side of the bridge rectifier. The capacitor across the triac allows a small current to flow through the bridge rectifier and the smoothing capacitor starts to charge. When the voltage builds up, the LED driver will try to operate. The result is an occasional flicker of the LED.

What is required is a discharge circuit, to keep the smoothing capacitor voltage below that required to start the LED driver. A 390 Ω resistor was found to keep the smoothing capacitor voltage below 5 V. To prevent high power loss when the circuit is active, a simple voltage detector can be used to disconnect the 390 Ω resistor when a voltage above about 8 V is detected. This circuit is shown in Figure 5.6.

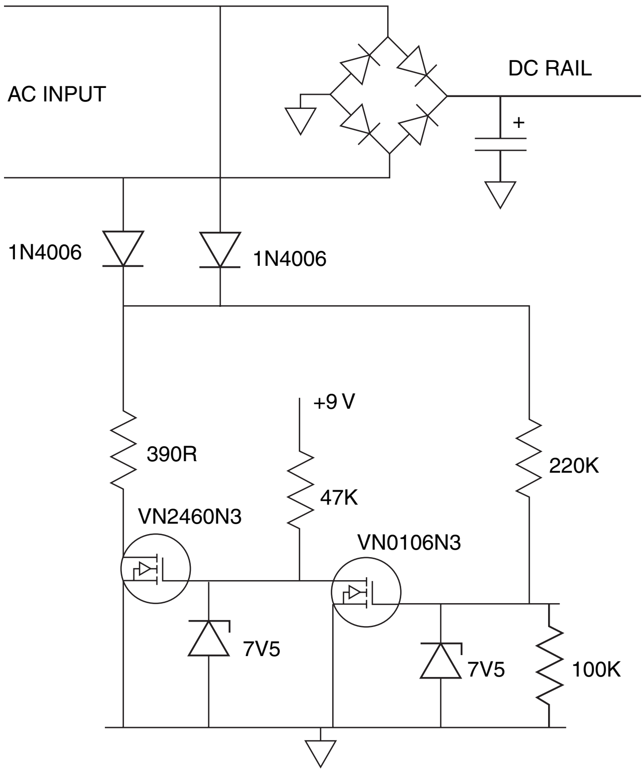


Figure 5.6: Smoothing Capacitor Discharge Circuit.

The triac needs to see a load. Once a triac is triggered, it is the load current that keeps it switched on; the triac is a self-sustaining switch. However, an LED driver provides no load until the input voltage has risen above the LED voltage, and it takes a little time for this current to be stable at a sufficiently high level to keep the triac turned on. For this reason, an additional load must be switched across the LED driver input at low voltages.

Tests have shown that a 2K2 Ω resistor works as a triac load and that it should remain in circuit until the supply voltage has risen to about 100 V, but should then be switched off until the rising edge of the next half-wave. A latching circuit to provide this function is shown in Figure 5.7.

These circuits can be combined. The voltage detector for the smoothing capacitor discharge circuit can also be used to provide an enable signal for the LED driver

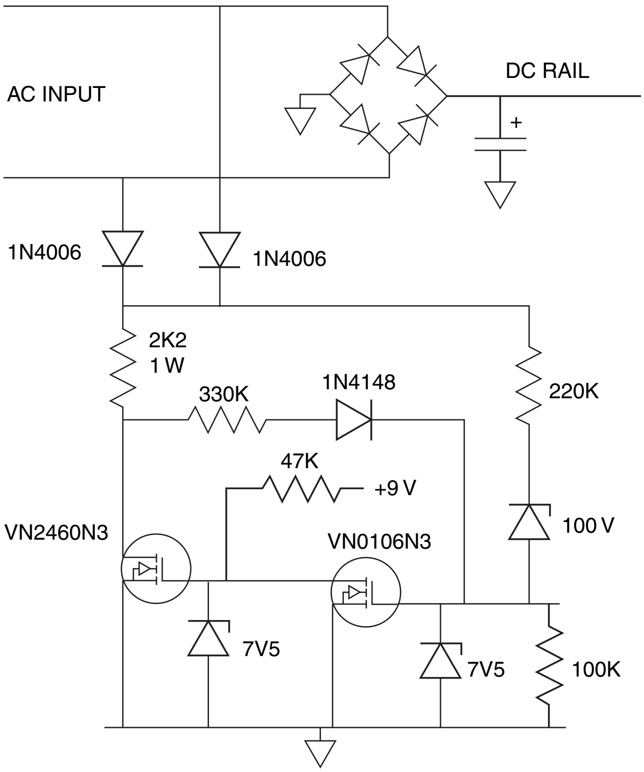


Figure 5.7: Additional Load Switch.

(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_{out}/V_{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

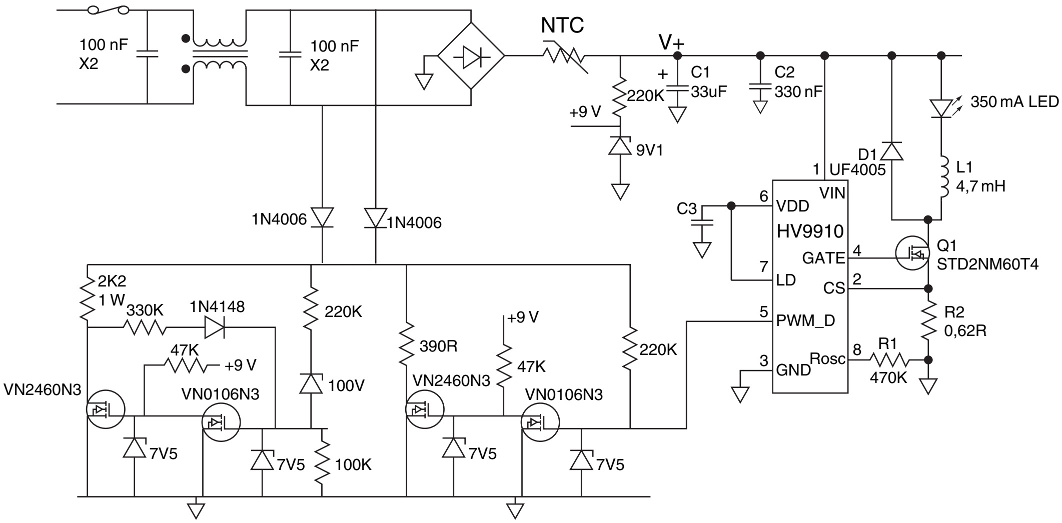


Figure 5.8: Complete Phase Dimmable LED Driver.

circuit to react; it needs to be at least 300 ns. Operating at 20 kHz will give an on-time of 466 ns, which is close to the limit for accurate control. A double buck may be needed (see next section).

Another error is not taking into account the parasitic capacitance of the inductor windings and the reverse current in the flywheel diode. These factors can be ignored in low voltage DC applications, but not in AC applications where the rectified supply is high voltage. The current peak through the MOSFET can be high enough to trip the current sense circuit, resulting in erratic switching. An RC filter between the current sense resistor and the current sense input of the integrated circuit may be necessary. A 2.2 kΩ series resistor followed by a 100 pF shunt capacitor to ground should be sufficient.

5.6 Double Buck

The double buck is an unusual design, as shown in Figure 5.9. It uses one MOSFET switch, but two inductors ($L2$ and $L3$) in series. Diodes steer the current in $L2$, which must operate in discontinuous conduction mode (DCM) for correct operation.

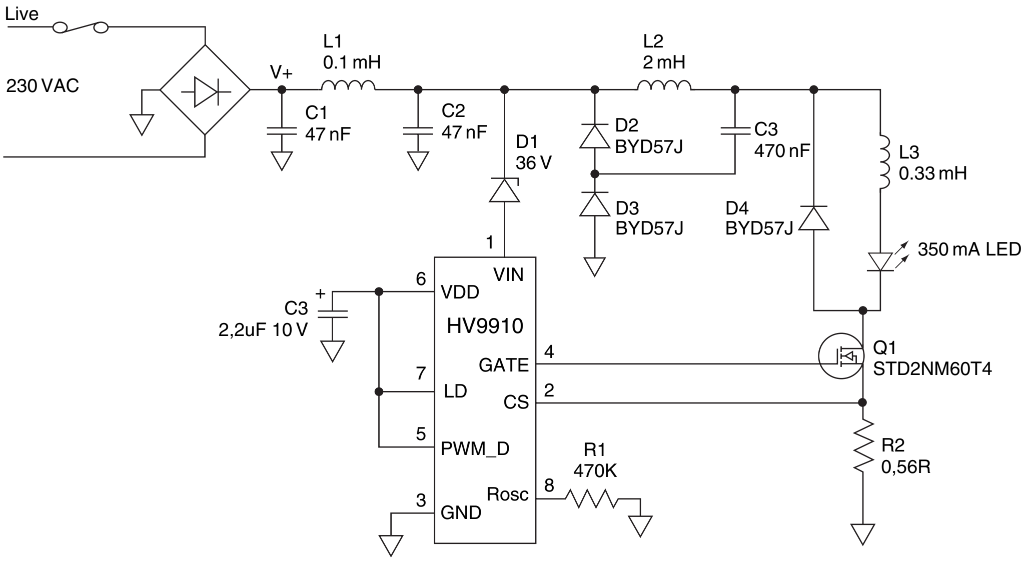


Figure 5.9: Double Buck.

The double buck is used when the output voltage is very low and the input voltage is high. An example is driving a single power LED from an AC supply line. A single buck stage cannot work easily because the on-time of the buck converter is too small, unless a very low switching frequency is used.

Assume the maximum duty cycle, D_{max}, is less than 0.5; also assume that the first stage ($L2$) is in boundary conduction mode (BCM) at D_{max}. Boundary conduction mode means that the current through the inductor only just falls to zero and the next switching cycle begins.

$$V_{\text{in min}} = \frac{Vo}{D_{\max}{}^2}$$

Or transposed, this becomes:

$$D_{\max} = \sqrt{\frac{Vo}{V_{\text{in min}}}}$$

This assumes that $L2$ is in BCM and $L3$ is in continuous conduction mode (CCM); at the minimum operating input voltage ($V_{\text{in min}}$).

The storage capacitor voltage at $V_{\text{in min}}$ and D_{max} is given by the equation:

$$V_{\text{c min}} = V_{\text{in min}} * \mathrm{D}_{\text{min}}$$

The peak current through the input stage inductor, at $V_{\text{in min}}$ equals:

$$\begin{aligned} I_{L2\_\text{pk}} &= 2 * I_{L2\_\text{avg}} \\ &= 2 * \frac{Vo * Io}{V_{\text{c min}}} \end{aligned}$$

Thus the primary stage inductor $L2$ has a value given by:

$$L2 = \frac{(V_{\text{in min}} - V_{\text{c min}}) * D_{\text{max}} * T_{\text{s}}}{I_{L2\_\text{pk}}}$$

The transfer ratio for a DCM buck converter (where R is load resistor seen by the converter) is given by:

$$\frac{V_{\text{c}}}{V_{\text{in}}} = \frac{2}{1 + \sqrt{1 + \frac{8 \times L2}{R \times T_{\text{s}} \times D^2}}}$$

The resistor R seen by the first stage (and assuming second stage is in CCM) is given by:

$$\begin{aligned} R &= \frac{{V_{\text{c}}}^2}{P_{\text{o}}} \\ \Rightarrow R \times D^2 &= \frac{(V_{\text{c}} \times D)^2}{P_{\text{o}}} = \frac{{V_{\text{o}}}^2}{P_{\text{o}}} \end{aligned}$$

Combining the previous two equations (which turn out to be a constant):

$$\frac{V_{\text{c}}}{V_{\text{in}}} = K = \frac{2}{1 + \sqrt{1 + \frac{8 \times L2 \times P_{\text{o}}}{T_{\text{s}} \times {V_{\text{o}}}^2}}}$$

We find that D is inversely proportional to V_{in}:

$$D = \frac{V_{\text{o}}}{V_{\text{c}}} = \frac{V_{\text{o}}}{K \times V_{\text{in}}}$$

And we can now show that the peak inductor current through $L2$ is a constant over the operating input voltage:

Setting $D = K'/V_{in}$, $K' = V_o/K$

K' is a constant, since V_o is constant.

$$\begin{aligned} i_{L2,pk} &= \frac{(V_{in} - V_c) \times D \times T_s}{L2} \\ &= \frac{V_{in} \times (1 - K) \times \frac{K'}{V_{in}} \times T_s}{L1} \\ &= \frac{(1 - K) \times K' \times T_s}{L2} \end{aligned}$$

We can now define the average input voltage as the maximum input voltage ($\sqrt{2}\mathrm{V}_{ac\,max}$) and the minimum operating input voltage:

$$V_{in\,avg} = \frac{(V_{in\,max} + V_{in\,min})}{2}$$

The storage capacitor value is computed based on 10% voltage ripple on the capacitor at $V_{in\ min}$ and D_{max}:

$$C = \frac{0.5 * I_{L2\_pk} * (1 - D_{max}) * T_s}{0.1 * V_{c\,min}}$$

The voltage across the storage capacitor, with average voltage input, is given by:

$$C_{c\,avg} = K * V_{in\,avg}$$

We can now compute the average duty cycle (at average input voltage):

$$\mathrm{D_{avg}} = \frac{\mathrm{Vo}}{\mathrm{V_{c\,avg}}}$$

Computing the value of $L3$:

$$L3 = \frac{(V_{c\,avg} - V_o) * D_{avg} * T_s}{\Delta \mathrm{I_{L3}}}$$

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.

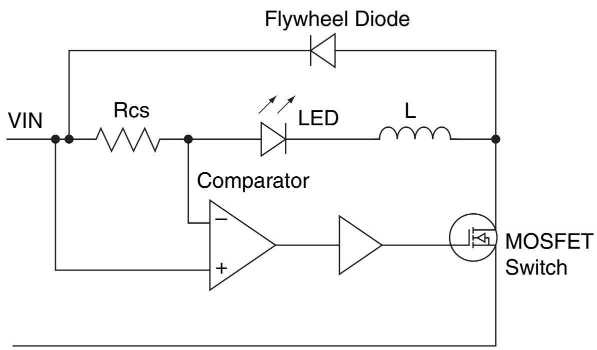


Figure 5.10: Hysteretic Current Control Circuit.

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.

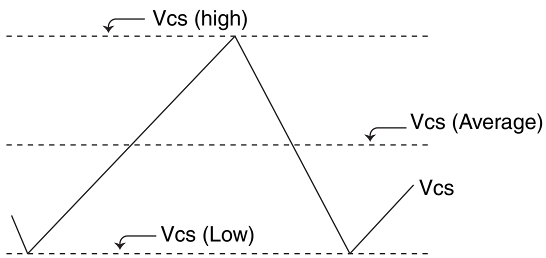


Figure 5.11: Current Sense Voltage (Current in LED Load).

The current level is set by a suitable resistor value, given by:

$$R_{\text{SENSE}} = \frac{1}{2} \cdot \frac{\left(V_{\text{CS(high)}} + V_{\text{CS(low)}}\right)}{I_{\text{LED}}}$$

In words, the average current sense voltage (midway between the high and low levels) divided by the average LED current required. The datasheet of the hysteretic controller being used will give the upper and lower current sense voltage levels that the comparator uses.

CHAPTER 6

Boost Converters

Boost converters (see Figure 6.1) are ideal for LED driver applications where the LED string voltage is greater than the input voltage. Normally, a boost converter would only be used when the output voltage minimum is about 1.5 times the input voltage.

- The converter can easily be designed to operate at efficiencies greater than 90%.
- Both the MOSFET and LED string are connected to a common ground. This simplifies sensing of the LED current, unlike the buck converter where we have to choose either a high side MOSFET driver or a high side current sensor.
- The input current can be continuous, which makes it easy to filter the input ripple current and thus easier to meet any required conducted EMI standards.

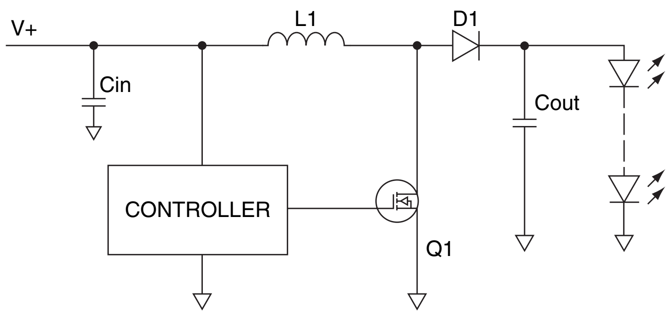


Figure 6.1: Simplified Boost Converter Circuit.

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. There is a path from the input to the output via a diode and inductor, so 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.

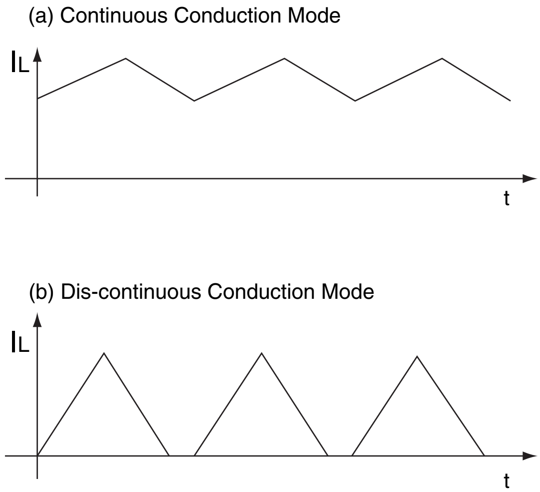


Figure 6.2: Inductor Current CCM and DCM.

6.2 HV9912 Boost Controller

Supertex's HV9912 integrated circuit is a closed-loop, peak current controlled, switch-mode converter LED driver. The HV9912 has built-in features to overcome the disadvantages of the boost converter. In particular, it features a disconnect MOSFET driver output. The external MOSFET driven from this output can be used to disconnect the LED strings during short circuit, or input over-voltage, conditions. This disconnect MOSFET is also used by the HV9912 to dramatically improve the PWM dimming response of the converter (see PWM Dimming section). The Linear Technology LTC3783 has similar functionality, although this part operates from a low voltage supply (6–16 V input).

The most significant functions within the HV9912 are shown in Figure 6.3.

The internal high voltage regulator in the HV9912 provides a regulated 7.75 V VDD from a 9–90 V input, which is used to power the IC. This voltage range is good for most boost applications, but the IC can also be used in buck and SEPIC circuits when accurate current control is required. In a high voltage buck application, a Zener diode could be added in series with the input to allow an even higher operating voltage, or to reduce the power dissipated by the IC.

The VDD pin of the IC can be overdriven (if necessary) with an external voltage source fed through a low voltage (>10 V), low current diode. The diode will help to

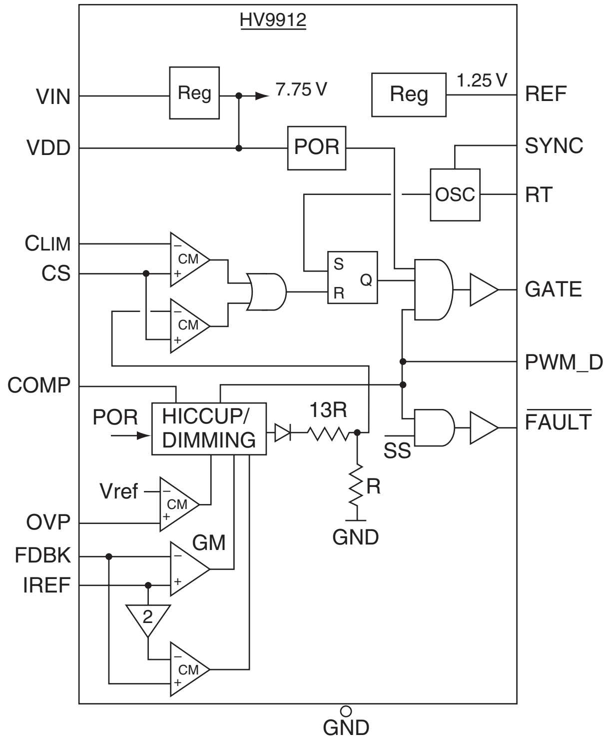


Figure 6.3: HV9912 Internal Structure (Simplified).

prevent damage to the HV9912 if the external voltage becomes less than the internally regulated voltage. The maximum steady state voltage that can be applied to the HV9912 VDD pin is 12 V (with a transient voltage rating of 13.5 V). Allowing for the diode forward voltage drop a 12 V ± 5% power supply would be ideal.

The HV9912 includes a buffered 1.25 V, 2% accurate reference voltage. This reference voltage can be used to set the current reference level as well as the input current limit level, by connecting potential divider networks between the REF pin and the IREF and CLIM pins. This reference is also used internally to set the over-voltage set point.

Using an external resistor, we can set the oscillator timing of the HV9912. If the resistor is connected between the RT and GND pins, the converter operates in a constant frequency mode, whereas if it is connected between the RT and GATE pins, the converter operates in a constant off-time mode (slope compensation is not necessary to stabilize the converter operating in a constant off-time). In both cases, the clock period or off-time can be set to any value between 2.8 µs to 40 µs using the equation given in Section 6.3.12.

Multiple HV9912 ICs can be synchronized to a single switching frequency by connecting the SYNC pins of all the ICs together. This is sometimes necessary in RGB lighting systems, or when EMI filters are designed to remove a certain frequency.

Closed loop control is achieved by connecting the output current sense signal to the FDBK pin and the current reference signal to the IREF pin. The HV9912 tries to keep the feedback signal equal to the voltage on the IREF pin. If the feedback is too high, indicating that the current is above the required level, the MOSFET switching is stopped. When the feedback falls below the voltage at the IREF pin, switching is started again.

The compensation network is connected to the COMP pin (output of the transconductance op-amp). What is not shown in Figure 6.3 is that the output of the amplifier has a switch controlled by the PWM dimming signal. When the PWM dimming signal is low, this switch disconnects the output of the amplifier. Thus, the capacitor(s) in the compensation network hold the voltage while the PWM signal is low. When the PWM dimming signal goes high again, the compensation network is reconnected to the amplifier. This ensures that the converter starts at the correct operating point and a very good PWM dimming response is obtained without having to design a fast controller.

The $\overline{\text{FAULT}}$ pin is used to drive an external disconnect MOSFET (see Figure 6.4). During the start-up of the HV9912, the $\overline{\text{FAULT}}$ pin is held low and once the IC starts up the pin is pulled high. This connects the LEDs in the circuit and the boost

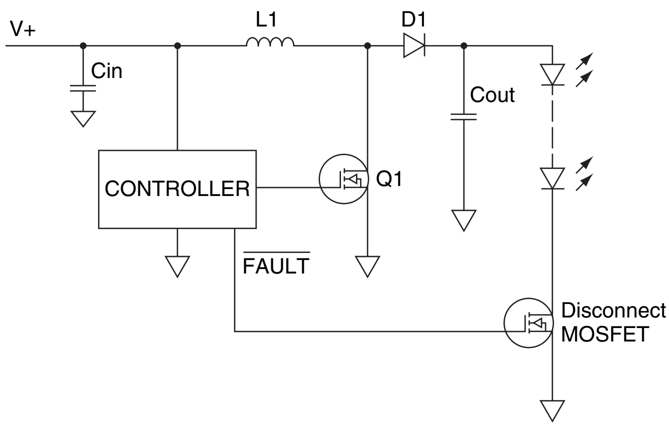


Figure 6.4: Disconnect MOSFET.

converter powers up the LEDs. In case of an output over-voltage condition or an output short circuit condition, the $\overline{\text{FAULT}}$ pin is pulled low and an external MOSFET switched off to disconnect the LEDs.

The $\overline{\text{FAULT}}$ pin is also controlled by the PWM dimming signal, so that the pin is high when the PWM dimming signal is high and vice versa. This disconnects the LEDs and makes sure that the output capacitor does not have to be charged/discharged every PWM dimming cycle. The PWM dimming input to the $\overline{\text{FAULT}}$ pin and the output of the protection circuitry are logically AND'ed to make sure that the protection circuit overrides the PWM input to the $\overline{\text{FAULT}}$ pin.

Output short circuit protection is provided by a comparator that triggers when the output current sense voltage (at the FDBK pin) is twice that of the reference voltage (at the IREF pin). The output over voltage protection is activated when the voltage at the OVP pin exceeds 5V. Both these fault signals are fed into the hiccup control. The output of this hiccup control turns off both the GATE pin and the $\overline{\text{FAULT}}$ pin when a fault condition occurs. Once the IC goes into the fault mode, either by an output over-voltage condition or a short circuit, the hiccup control is activated. The hiccup control turns off the gate drives to both MOSFETs. At the same time, a timer is started to keep the output turned off for a short period (determined by the capacitance on the COMP pin). Once this time period elapses, the HV9912 attempts to restart. If the fault condition persists, the output is turned off again and the timer is reset. This repeats until the fault condition has been removed and the HV99112 returns to normal operation.

Linear dimming is achieved by varying the voltage level at the IREF pin. This can be done either with a potentiometer from the REF pin or from an external voltage source and a resistor divider. This allows the current to be linearly dimmed. However, a minimum output voltage limit has been deliberately added to the output of the GM amplifier, to prevent false triggering of the fault condition that could otherwise place if very low voltages are applied to the IREF pin. This output voltage limit restricts the linear dimming range to about 10:1.

The features included in the HV9912 help achieve a very fast PWM dimming response in spite of the shortcomings of the boost converter. The PWM dimming signal controls three nodes in the IC.

- Gate signal to the switching MOSFET
- Gate signal to the disconnect MOSFET
- Output connection of the transconductance op-amp

When PWMD is high, the gates of both the switching MOSFET and the disconnect MOSFET are enabled. At the same time, the output of the transconductance op-amp is connected to the compensation network. This allows the boost converter to operate normally.

When PWMD goes low, the GATE of the switching MOSFET is disabled to stop energy transfer from the input to the output. However, this does not prevent the output capacitor from discharging into the LEDs causing a large decay time for the LED current. This discharge of the capacitor also means that when the circuit restarts, the output capacitor has to charge again, causing an increase in the rise time of the LED current. This problem becomes more prominent with larger output capacitors. Thus, it is important to prevent the discharge of the output capacitor. This is done by turning off the disconnect MOSFET. This causes the LED current to fall to zero almost instantaneously. Since the output capacitor does not discharge, there is no necessity to charge the capacitor when PWMD goes high. This enables a very fast rise time as well.

So what happens if our controller does not have a switch on the output of the feedback amplifier? When PWMD goes low, the output current goes to zero. This means that the feedback amplifier sees a very large error signal across its input terminals, which would cause the voltage across the compensation capacitor to increase to the positive rail. Thus, when the PWMD signal goes high again, the large voltage across the compensation network, which dictates the peak inductor current value, will cause a large spike in the LED current. The current will come back into regulation depending on the speed of the controller.

The HV9912 disconnects the output of the amplifier from the compensation network when PWMD goes low, which helps to keep the voltage at the compensation unchanged. Thus, when PWMD goes high again, the circuit will already be at the steady state condition, eliminating the large turn-on spike in the LED current.

6.3 Design of a Continuous Conduction Mode Boost LED Driver

As a reminder, continuous conduction mode is valid when the output voltage is between 1.5 and 6 times the input voltage.

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.

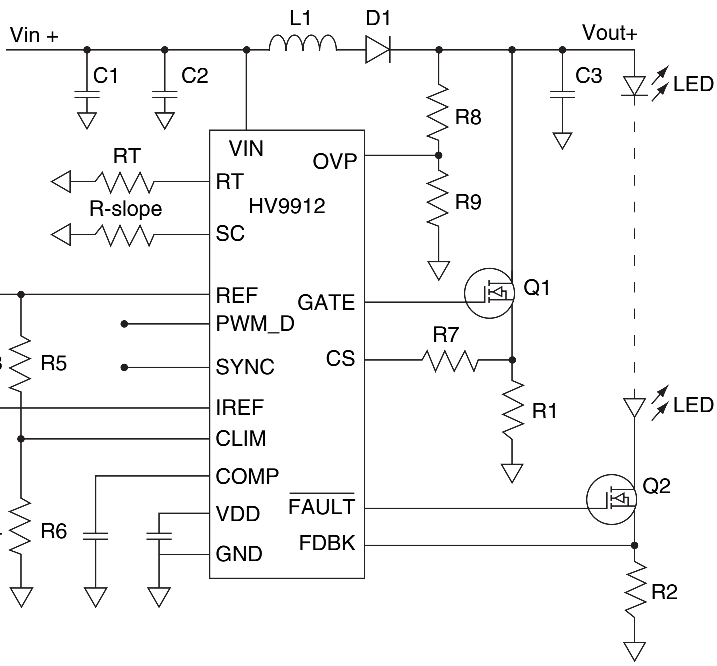


Figure 6.5: Continuous Mode Boost Converter.

6.3.3 Selecting the Switching Frequency (f_s)

For low voltage applications (output voltage < 100 V), and moderate power levels (<30W), a switching frequency of $f_s = 200$ kHz is a good compromise between switching power loss and size of the components. At higher voltage or power levels, the switching frequency might have to be reduced to lower the switching losses in the external MOSFET.

6.3.4 Computing the Maximum Duty Cycle (D_{max})

The maximum duty cycle of operation can be computed as

$$\begin{aligned} D_{max} &= 1 - \frac{\eta_{min} \cdot V_{in\,min}}{V_{o\,max}} \\ &= 0.717 \end{aligned}$$

Note: If $D_{max} = 0.85$, the step-up ratio is too large. The converter cannot operate in continuous conduction mode and has to be operated in discontinuous conduction mode to achieve the required step-up ratio.

6.3.5 Computing the Maximum Inductor Current ($I_{in\,max}$)

The maximum input current is

$$\begin{aligned} I_{in\,max} &= \frac{V_{o\,max} \cdot I_{o\,max}}{\eta_{min} \cdot V_{in\,min}} \\ &= 1.24\text{A} \end{aligned}$$

6.3.6 Computing the Input Inductor Value ($L1$)

The input inductor can be computed by assuming a 25% peak-to-peak ripple in the inductor current at minimum input voltage.

$$\begin{aligned} L1 &= \frac{V_{in\,min} \cdot D_{max}}{0.25 \cdot I_{in\,max} \cdot f_s} \\ &= 254\,\mu\text{H} \end{aligned}$$

Choose a standard 330 μH inductor. To achieve 90% efficiency at the minimum input voltage, the power loss in the inductor has to be limited to around 2–3% of the total output power. Using a 3% loss in the inductor

$$\begin{aligned} P_{\text{ind}} &= 0.03 \cdot V_{\text{o max}} \cdot I_{\text{o max}} \\ &= 0.735\text{W} \end{aligned}$$

Assuming an 80–20% split in the inductor losses between resistive and core losses, the DC resistance of the chosen inductor has to be less than

$$\begin{aligned} \text{DCR} &< \frac{0.8 \cdot P_{\text{ind}}}{I_{\text{in max}}{}^2} \\ \Rightarrow \text{DCR} &< 0.38\,\Omega \end{aligned}$$

The saturation current of the inductor has to be at least 20% higher than its peak current; otherwise the core losses will be too great.

$$\begin{aligned} I_{\text{sat}} &= 1.2 \cdot I_{\text{in max}} \cdot \left(1 + \frac{0.25}{2}\right) \\ &= 1.7\text{A} \end{aligned}$$

Thus $L1$ is a 330 μH inductor with a DC resistance about 0.38 Ω and a saturation current greater than 1.7 A.

> **Note:** Choosing an inductor with an RMS current rating equal to $I_{\text{in max}}$ would also yield acceptable results, although meeting the minimum efficiency requirement might not be possible.

6.3.7 Choosing the Switching MOSFET ($Q1$)

The maximum voltage across the MOSFET in a boost converter is equal to the output voltage. Using a 20% overhead to account to switching spikes, the minimum voltage rating of the MOSFET has to be

$$\begin{aligned} V_{\text{FET}} &= 1.2 \cdot V_{\text{o max}} \\ &= 84\text{V} \end{aligned}$$

The RMS current through the MOSFET is

$$\begin{aligned} I_{FET} &\approx I_{in\ max} \cdot \sqrt{D_{max}} \\ &= 1.05\text{A} \end{aligned}$$

To get the best performance from the converter, the MOSFET chosen has to have a current rating about three times the MOSFET RMS current with minimum gate charge Q_g. The higher current rating gives low conduction losses, even at high silicon junction temperatures (resistance increases with temperature). It is recommended that for designs with the HV9912, the gate charge of the chosen MOSFET be less than 25 nC.

The switching device chosen for this application is a 100 V, 4.5 A MOSFET with a Q_g of 11 nC.

6.3.8 Choosing the Switching Diode ($D1$)

The voltage rating of the diode is the same as the voltage rating of the MOSFET (100 V). The average current through the diode is equal to the maximum output current (350 mA). Although the average current through the diode is only 350 mA, the diode carries the full input current $I_{in\,max}$ for short durations of time. Thus, it is a better design approach to choose the current rating of the diode somewhere in between the maximum input current and the average output current (preferably closer to the maximum input current). Thus, for this design, the diode chosen is a 100 V, 1 A Schottky diode.

6.3.9 Choosing the Output Capacitor (C_o)

The value of the output capacitor C_o (labeled $C3$ in Figure 5.8) depends on the dynamic resistance of the LED, the ripple current desired in the LED string and the LED current. In designs using the HV9912, a larger output capacitor (lower output current ripple) will yield better PWM dimming results. The capacitor required to filter the current appropriately will be designed by considering the fundamental component of the diode current only.

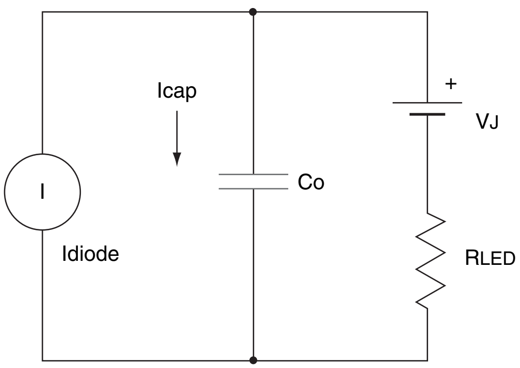


Figure 6.6: Model of Boost Converter Output.

The output stage of the boost converter is modeled in Figure 6.6, where the LEDs are modeled as a constant voltage load with series dynamic impedance.

The output impedance (parallel combination of R_{LED} and C_{o}) is driven by the diode current. The waveform of the capacitor current in steady state is shown in Figure 6.7; the capacitor is charged during the off-time, as the energy stored in the inductor is transferred to the capacitor. While the MOSFET is turned on and energy is being stored in the inductor, the capacitor is discharged by the load.

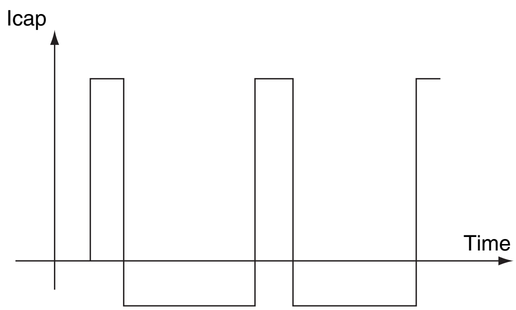


Figure 6.7: Charge and Discharge Cycle of Output Capacitor.

Using the 10% peak-to-peak current ripple given in the design parameters table, the maximum voltage ripple across the LED string has to be

$$\begin{aligned}\Delta v_{\mathrm{p-p}} &= \Delta I_{\mathrm{o}} \cdot R_{\mathrm{LED}} \\ &= 0.63\mathrm{V}\end{aligned}$$

Assuming a constant discharging current of 350 mA when the switch is ON, the equation for the voltage across the capacitor can be written as

$$I_{\mathrm{o\,max}} = C_{\mathrm{o}} \cdot \frac{\Delta v_{\mathrm{p-p}}}{D_{\max} \cdot T_{\mathrm{s}}}$$

Substituting values into the above equation, we can calculate the value for C_{o}.

$$\begin{aligned} C_{\mathrm{o}} &= \frac{I_{\mathrm{o\,max}} \cdot D_{\max}}{\Delta v_{\mathrm{p-p}} \cdot f_{\mathrm{s}}} \\ &= 1.99\,\mu\mathrm{F} \end{aligned}$$

The RMS current through the capacitor can be given by

$$\begin{aligned} I_{\mathrm{rms}} &= \sqrt{D_{\max} \cdot {I_{\mathrm{o\,max}}}^2 + (1 - D_{\max}) \cdot (I_{\mathrm{in\,max}} - I_{\mathrm{o\,max}})^2} \\ &= 0.56\mathrm{A} \end{aligned}$$

In this case, a parallel combination of two 1 μF, 100 V metal polypropylene capacitors is chosen.

> **Note:** The proper types of capacitors to use are either metal film capacitors or ceramic capacitors, since they are capable of carrying this high ripple current. Although ceramic capacitors are smaller in size and capable of carrying the ripple current, they cause a lot of audible noise during PWM dimming since they have a piezo-electric effect. Also, high value ceramic capacitors are normally only rated up to 50 V. Thus metal polypropylene (or any other metal film) capacitors are the ideal choice for LED drivers if PWM dimming is required.

6.3.10 Choosing the Disconnect MOSFET ($Q2$)

The disconnect MOSFET should have the same voltage rating as the switching MOSFET $Q1$. The on-state resistance of the MOSFET at room temperature ($R_{\mathrm{on,25C}}$) has to be chosen based on a 1% power loss in $Q2$ at full load current. Thus,

$$\begin{aligned} R_{\mathrm{on,25C}} &= \frac{0.01 \cdot V_{\mathrm{o\,max}}}{I_{\mathrm{o\,max}} \cdot 1.4} \\ &= 1.43\,\Omega \end{aligned}$$

The 1.4 multiplication factor is included to account for the increase in the on-resistance due to rise in junction temperature. In this case, a MOSFET with high gate charge, Q_g, can be chosen if desired (as it is not switching regularly). A high Q_g MOSFET will slow down the turn-on and turn-off times. In this case, the MOSFET chosen is a 100 V, 0.7 Ω, SOT-89 MOSFET with a Q_g of 5 nC.

6.3.11 Choosing the Input Capacitors (C1 and C2)

The values of input capacitors $C1$ and $C2$ have to be calculated to meet closed loop stability requirements. The connection from the power source to the boost converter circuit will have some resistance, R_{source}, and some inductance, L_{source}. These feed across the input capacitors ($C1$ and $C2$) and so form an LC resonant circuit. To prevent interference with the control loop, the resonant frequency should be arranged to be 40% or less of the switching frequency.

How do we determine the inductance L_{source}? A pair of 22AWG connecting wires 1 foot (30 cm) long will have an inductance of about 1 µH. This is a good starting point. If necessary, the wires can be twisted together to reduce the inductance.

With a 200 kHz switching frequency, the resonant frequency should be less than 80 kHz.

$$C_{IN} \geq \frac{1}{(2 \cdot \pi \cdot f_{LC})^2 \cdot L_{SOURCE}} = 3.95\,\mu F$$

$C1 = C2 = 2.2\,\mu F$, 50 V ceramic.

The magnitude of the reflected converter impedance at the LC resonant frequency is given by:

$$R_{EQ} = (1 - D_{MAX})^2 \cdot R_{LED}$$
$$R_{EQ} = (1 - 0.717)^2 \cdot 18$$
$$R_{EQ} = 1.4416\,\Omega$$
$$R_{SOURCE,MAX} = 1.44\,\Omega$$

6.3.12 Choosing the Timing Resistor (R_T)

The HV9912 oscillator has an 18 pF capacitor charged by a current mirror circuit. An external timing resistor R_T provides a reference current for the current mirror. When R_T is connected to 0 V, current flows and the timing process begins. When charged to a certain voltage, the RS flip-flop is set, the capacitor is discharged, and the timing process starts again. The timing resistor value can be calculated by using the equation:

$$\frac{1}{f_s} \approx R_T \cdot 18\,\text{pF}$$

In this case, for a constant 200 kHz switching frequency, the timing resistor value works out to 274 kΩ. This resistor needs to be connected between the R_T pin and GND as shown in the typical circuit.

6.3.13 Choosing the Two Current Sense Resistors (*R*1 and *R*2)

The value of output current sense resistor *R*2 is calculated to limit its power dissipation to about 0.15 W, so that a 1/4 W resistor can be used. Using this criterion,

$$\begin{aligned} R2 &= \frac{0.15\text{W}}{{I_{\text{o max}}}^2} \\ &= 1.22\,\Omega \end{aligned}$$

In this case, the resistor chosen is a 1.24 Ω, 1/4 W, 1% resistor.

The MOSFET current sense resistor *R*1 is calculated by limiting the voltage across the resistor to about 250 mV at maximum input current.

$$\begin{aligned} R1 &= \frac{0.25}{1.125 \cdot I_{\text{in max}}} \\ &= 0.18\,\Omega \end{aligned}$$

The power dissipated in this resistor is

$$\begin{aligned} P_{R1} &= {I_{\text{FET}}}^2 \cdot R1 \\ &= 0.2\text{W} \end{aligned}$$

Thus, the chosen current sense resistor is a 0.18 Ω, 1/2 W, 1% resistor.

6.3.14 Selecting the Current Reference Resistors (*R*3 and *R*4)

The voltage at the current reference pin IREF can be set either by using the reference voltage provided at the REF pin (through a voltage divider) or with an external voltage source. In the present design, it is assumed that the voltage at the IREF pin is set using a voltage divider from the REF pin. The current reference resistors *R*3 and *R*4 can be computed using the following two equations:

$$R3 + R4 = \frac{1.25\text{V}}{50\,\mu\text{A}} = 25\,\text{k}\Omega$$

$$\frac{1.25\text{V}}{R3 + R4} \cdot R4 = I_{\text{o max}} \cdot R2$$

For this design, the values of the two resistors can be computed to be

$$R_{r2} = 8.68\,\text{k}\Omega,\ 1/8\,\text{W},\ 1\%$$

$$R_{r1} = 16.32\,\text{k}\Omega,\ 1/8\text{W},\ 1\%$$

6.3.15 Programming the Slope Compensation (R_{slope} and *R*7)

Since the boost inductor being designed is operating at constant frequency, slope compensation is required to ensure the stability of the converter. The slope added to the current sense signal has to be one-half the maximum down slope of the inductor current to ensure stability of the peak current mode control scheme for all operating conditions. This can easily be achieved by the proper selection of the two slope compensation resistors R_{slope} and *R*7.

For the present design, the down slope of the inductor current is

$$\begin{aligned} DS &= \frac{V_{\text{o max}} - V_{\text{in min}}}{L} \\ &= 0.145\,\text{A}/\mu\text{s} \end{aligned}$$

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 SC pin is limited to $100\,\mu\text{A}$. This limits the minimum value of the R_{slope} resistor to $25\,\text{k}\Omega$. If the equation for slope compensation produces a value R_{slope} less than this value, then $R7$ would have to be increased accordingly. It is recommended that R_{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).

For this design,

$$\begin{aligned} V_{\mathrm{CLIM}} &= 1.35 \cdot 1.24 \cdot 0.18 + \frac{4.5 \cdot 510}{39\mathrm{k}} \\ &= 0.36\,\mathrm{V} \end{aligned}$$

We need a potential divider to give 0.36 V from a 1.25 V reference. Using a maximum current sourced out of the REF pin of 50 μA the two resistors in series should be >25 kΩ, and can be calculated as:

$$R5 = 20\mathrm{k}, 1/8\,\mathrm{W}, 1\%$$

$$R6 = 8.06\mathrm{k}, 1/8\,\mathrm{W}, 1\%$$

Note: It is recommended that no capacitor be connected at the CLIM pin.

6.3.17 Capacitors at VDD and REF Pins

It is recommended that bypass capacitors be connected to both VDD and REF pins. For the VDD pin, the capacitor used is a 1 μF ceramic chip capacitor. If the design uses switching MOSFETs that have a high gate charge ($Q_g > 15\,\mathrm{nC}$), the capacitor at the VDD pin should be increased to 2.2 μF.

For the REF pin, the capacitor used is a 0.1 μF ceramic chip capacitor.

6.3.18 Setting the Over-Voltage Trip Point (*R*8 and *R*9)

The over-voltage trip point can be set at a voltage 15% higher than the maximum steady state voltage. Using a 20% margin, the maximum output voltage during open LED condition will be

$$\begin{aligned} V_{\mathrm{open}} &= 1.2 \cdot V_{\mathrm{o\,max}} \\ &= 84\mathrm{V} \end{aligned}$$

Then, the resistors that set the over-voltage set point can be computed as

$$R8 = \frac{(V_{\text{open}} - 5)^2}{0.1}$$
$$= 64\,\text{k}\Omega$$

The above equation will allow us to select a 1/8 W resistor by limiting the power dissipation in the resistor.

$$R9 = \frac{R8}{(V_{\text{open}} - 5)} \cdot 5V$$
$$= 3.95\,\text{k}\Omega$$

The closest 1% resistor values are

$$R8 = 68\text{k},\ 1/8\,\text{W},\ 1\%$$
$$R9 = 3.9\text{k},\ 1/8\,\text{W},\ 1\%$$

Note: The actual over-voltage point will vary from the desired point by $\pm 5\%$ due to the variation in the reference (see datasheet). For this design, it varies from 80 V to 88.2 V.

6.3.19 Designing the Compensation Network

The compensation needed to stabilize the converter could be either a Type-I circuit (a simple integrator) or a Type-II circuit (an integrator with an additional pole-zero pair). The type of the compensation circuit required will be dependent on the phase of the power stage at the crossover frequency.

The loop gain of the closed loop system is given by

$$\text{Loop Gain} = R_s \cdot G_m \cdot Z_c(s) \cdot \frac{1}{15} \cdot \frac{1}{R_{cs}} \cdot G_{ps}(s)$$

Where G_m is the transconductance of the op-amp (435 μA/V), $Z_c(s)$ is the impedance of the compensation network, and $G_{ps}(s)$ is the transfer function of the power stage. Please note that although the resistors give a 1:14 ratio, the overall effect when including the diode drop is effectively 1:15.

For the continuous conduction mode boost converter in peak current control mode and for frequencies less than one tenth of the switching frequency, the power stage transfer function is given by

$$G_{ps}(s) = \frac{(1-D_{max})}{2} \cdot \frac{1 - s \cdot \dfrac{L1}{(1-D_{max})^2 \cdot R_{LED}}}{1 + s \cdot \dfrac{R_{LED} \cdot C_o}{2}}$$

For the present design, choose a crossover frequency 0.01^*f_s, $f_c = 2\,\text{kHz}$. The low crossover frequency will result in large values for C_c and C_z, which will indirectly provide a soft-start for the circuit. Since the HV9912 does not depend on the speed of the controller circuit for the PWM dimming response, the low crossover frequency will not have an adverse effect on the PWM dimming rise and fall times.

$$G_{ps}(s) = \frac{0.283}{2} \cdot \frac{1 - s \cdot \dfrac{330 \cdot 10^{-6}}{(0.283)^2 \cdot 18}}{1 + s \cdot \dfrac{18 \cdot 2 \cdot 10^{-6}}{2}}$$

$$G_{ps}(s) = 0.1415 \cdot \frac{1 - s \cdot 2.28912 \cdot 10^{-4}}{1 + s \cdot 1.8 \cdot 10^{-5}}$$

Substituting $s = i \cdot (2\pi \cdot f_c)$, where $f_c = 2$ kHz, $s = i \cdot 12566$.

$$G_{ps}(s) = 0.1415 \cdot \frac{1 - i \cdot 2.8766}{1 + i \cdot 0.226188}$$

At this frequency, the magnitude and frequency of the power stage transfer function (obtained by substituting $s = i \cdot (2\pi \cdot f_c)$ in the previous equation) are

$$\left|G_{ps}(s)\right|\Big|_{fc=2\text{kHz}} = A_{ps} = 0.40996$$
$$\angle G_{ps}(s)\Big|_{fc=2\text{kHz}} = \phi_{ps} = -83.57^\circ$$

To get a phase margin of about $\phi = 45°$ (the recommended phase margin range is 45°–60°), the phase boost required will be

$$\begin{aligned}\phi_{\text{boost}} &= \phi_{\text{m}} - \phi_{\text{ps}} - 90° \\ &= 38.57°\end{aligned}$$

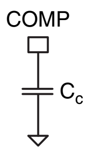
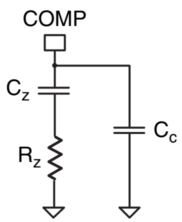
Based on the value of the phase boost required, the type of compensation can be determined.

$$\begin{aligned}\phi_{\text{boost}} \leq 0° &\Rightarrow \text{ Type} - \text{I controller} \\ 0° \leq \phi_{\text{boost}} \leq 90° &\Rightarrow \text{ Type} - \text{II controller} \\ 90° \leq \phi_{\text{boost}} \leq 180° &\Rightarrow \text{Type} - \text{III controller}\end{aligned}$$

Type-III controllers are usually not required to compensate an HV9912-based boost LED driver and thus will not be discussed here.

The implementations for the Type-I and Type-II systems for use with the HV9912 are given in Table 6.1.

Table 6.1: Compensation Networks.

<table>
<tr><th>Type</th><th>Circuit diagram</th><th>Transfer function</th></tr>
<tr><td>I</td><td>
</td><td>$Z_c(s) = \frac{1}{sC_c}$</td></tr>
<tr><td>II</td><td>
</td><td>$Z_c(s) = \frac{1}{s(C_c + C_z)} \cdot \frac{1 + s \cdot R_z \cdot C_z}{1 + s \cdot \frac{C_z \cdot C_c}{C_z + C_c} \cdot R_z}$</td></tr>
</table>

Designing with Type-I controllers is simple – adjust C_c so that the magnitude of the loop gain equals 1 at the crossover frequency. For the present design, however, we need to use a Type-II controller. The equations needed to design the Type-II controller are given below:

$$K = \tan\left(45^\circ + \frac{\phi_{\text{boost}}}{2}\right)$$
$$= 2.077$$

$$\omega_z = \frac{1}{R_z \cdot C_z} = \frac{2 \cdot \pi \cdot f_c}{K}$$
$$= 6050 \text{ rad/sec}$$

$$\omega_p = \frac{C_z + C_p}{C_z \cdot C_p \cdot R_z} = (2 \cdot \pi \cdot f_c) \cdot K$$
$$= 26100 \text{ rad/sec}$$

One more equation can be obtained by equating the magnitude of the loop gain to 1 at the crossover frequency.

$$R_s \cdot G_m \cdot \left(\frac{1}{2 \cdot \pi \cdot f_c \cdot (C_z + C_c)} \cdot \frac{\sqrt{1 + K^2}}{\sqrt{1 + (1/K)^2}}\right) \cdot \frac{1}{15} \cdot \frac{1}{R_{cs}} \cdot A_{ps} = 1$$

$$C_z + C_c = 10\,\text{nF}$$
$$C_c = (C_z + C_c) \cdot \frac{\omega_z}{\omega_p}$$
$$= 2.32\,\text{nF}$$

$$C_z = 7.68\,\text{nF}$$

$$R_z = \frac{1}{\omega_z \cdot C_z}$$
$$= 21.522\,\text{k}\Omega$$

Choose

$C_c = 2.2\,\text{nF}$, 50 V, C0G capacitor

$C_z = 6.8\,\text{nF}$, 50 V, C0G capacitor

$R_z = 22.0\text{k}$, 1/8 V, 1% resistor

6.3.20 Output Clamping Circuit

One problem encountered with a continuous mode boost converter, when operating with $V_{out} < 2 \times V_{in}$, is L-C resonance between the inductor and C_{out}. Clamping the output to the input by a diode from V_{in} to V_{out} can prevent this resonance. This diode is shown as $D2$ in Figure 6.8 Diode $D2$ can be a standard recovery time diode like 1N4002; this type of diode is better at handling surge currents that could be present at switch-on.

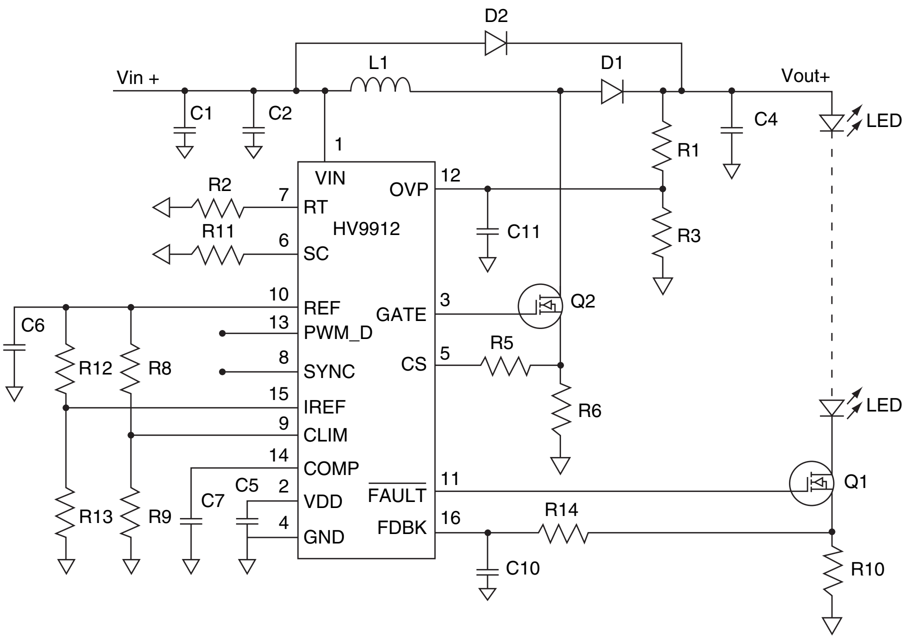


Figure 6.8: Boost Converter with Clamping Diode.

This completes the design of the HV9912-based boost converter operating in continuous conduction mode.

6.4 Design of a Discontinuous Conduction Mode Boost LED Driver

As a reminder, discontinuous mode is used when the output voltage is more than six times the input voltage.

6.4.1 Design Specification

Input voltage range = 9–16 V

LED string voltage range = 30–70 V

(Note, with 9 V input and 70 V output, the V_o/V_{in} *ratio is approximately 7.8)*

LED current = 100 mA

LED current ripple = 10% (10 mA)

LED dynamic impedance = 55 ohms

Efficiency > 85%

6.4.2 Typical Circuit

A typical circuit for a discontinuous mode boost converter, using the HV9912 IC identical to the continuous mode circuit shown in Figure 6.5, but repeated here for convenience, in Figure 6.9.

6.4.3 Selecting the Switching Frequency (f_s)

For low voltage applications (output voltage < 100 V), and moderate power levels (<30 W), a switching frequency of $f_s = 200$ kHz is a good compromise between switching power loss and size of the components. At higher voltage or power levels, the switching frequency might have to be reduced to lower the switching losses in the external MOSFET.

6.4.4 Computing the Maximum Inductor Current ($I_{in\,max}$)

The maximum input current is

$$\begin{aligned} I_{in\,max} &= \frac{V_{o\,max} \cdot I_{o\,max}}{\eta_{min} \cdot V_{in\,min}} \\ &= 0.915\text{A} \end{aligned}$$

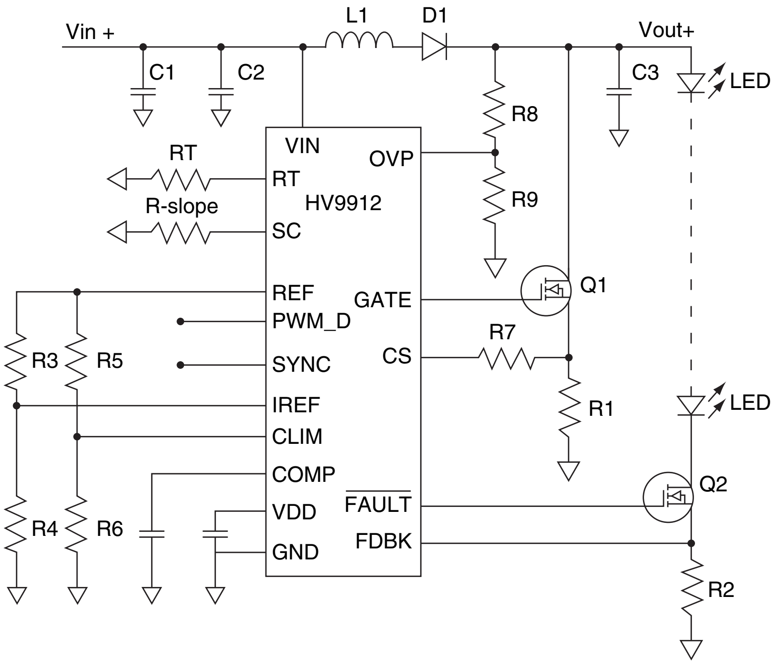


Figure 6.9: Discontinuous Mode Boost Converter.

6.4.5 Computing the Input Inductor Value (*L*1)

Assuming that the sum of the on-time of the switch and the on-time of the diode is 95% of the total switching time period at $V_{\text{in min}}$,

$$L1 \cdot i_{\text{Lpk}} \cdot \left(\frac{1}{V_{\text{in min}}} + \frac{1}{V_{\text{o max}} - V_{\text{in min}}} \right) = \frac{0.95}{f_s}$$
$$= 4.75\,\mu s$$

where i_{LPk} is the peak input current (see Figure 6.10).

$V_{\text{in}}/L1$ controls the rate at which current increases and the rising period is determined by the on-time of the MOSFET, which is the duty cycle multiplied by the switching period. The rate of fall is controlled by $(V_o - V_{\text{in}})/L1$ and the falling period is the time that the diode is conducting.

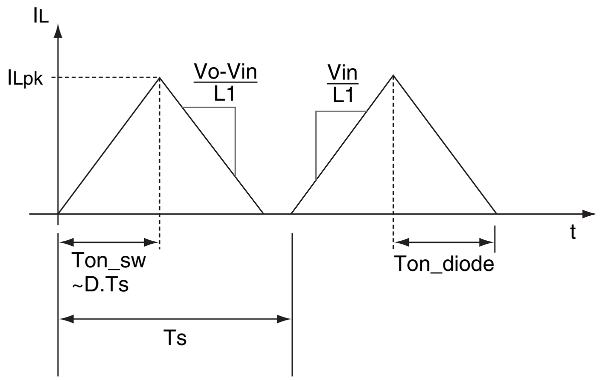


Figure 6.10: Inductor Current Waveform in DCM.

The average input current at the minimum input voltage is equal to the average inductor current and can be computed from

$$\begin{aligned} I_{\text{in max}} &= \frac{1}{2} \cdot i_{\text{Lpk}} \cdot \frac{4.75\,\mu s}{5\,\mu s} \\ &= 0.475 \cdot i_{\text{Lpk}} \end{aligned}$$

Transposing the equation, the peak input current is

$$\begin{aligned} i_{\text{Lpk}} &= \frac{I_{\text{in max}}}{0.475} \\ &\approx 1.93\text{A} \end{aligned}$$

Substituting for i_{Lpk} in the equation for $L1$

$$\begin{aligned} L1 &= \frac{0.95}{200\text{k}} \cdot \frac{9\text{V} \cdot (70\text{V} - 9\text{V})}{70\text{V} \cdot 1.93\text{A}} \\ &= 19.3\,\mu\text{H} \end{aligned}$$

Note that the value of L1 computed is the absolute maximum value for the inductor. Assuming a $\pm 20\%$ variation in the inductance, the nominal inductor value has to be

$$\begin{aligned} L1_{\text{nom}} &= \frac{L1}{1.2} \\ &= 16.08\,\mu\text{H} \end{aligned}$$

The closest standard value is a $15\,\mu$H inductor.

The RMS current through the inductor is

$$I_{\text{Lrms}} = i_{\text{Lpk}} \cdot \sqrt{\frac{0.9}{3}}$$
$$= 1.057\text{A}$$

Choose a 15 µH inductor (±20% tolerance). A custom inductor would work best for this application given the large swings in the inductor flux. However, if a standard value inductor is preferred, the saturation current rating of the inductor should be at least 1.5 times the peak current computed, to keep the core losses to an acceptable value.

The inductor chosen in this case is a 15 µH inductor with an RMS current rating of 1.4 A and a saturation current rating of 3 A.

6.4.6 Computing the On and Off Times of the Converter

The on-time of the switch can be computed as

$$t_{\text{on\_sw}} = \frac{L1_{\text{nom}} \cdot i_{\text{Lpk}}}{V_{\text{in min}}}$$
$$= 3.22\,\mu\text{s}$$

The on-time of the diode is

$$t_{\text{on\_diode}} = \frac{L1_{\text{nom}} \cdot i_{\text{Lpk}}}{V_{\text{o max}} - V_{\text{in min}}}$$
$$= 467\,\text{ns}$$

The maximum duty cycle can then be computed as

$$D_{\max} = t_{\text{on\_sw}} \cdot f_{\text{s}}$$
$$= 0.644$$

The diode conduction time ratio can be expressed as

$$D1 = t_{\text{on\_diode}} \cdot f_{\text{s}}$$
$$= 0.0934$$

6.4.7 Choosing the Switching MOSFET ($Q1$)

The maximum voltage across the MOSFET in a boost converter is equal to the output voltage. Using a 20% overhead to account for switching spikes, the minimum voltage rating of the MOSFET has to be

$$\begin{aligned} V_{FET} &= 1.2 \cdot V_{o\,max} \\ &= 84\,V \end{aligned}$$

The RMS current through the MOSFET is

$$\begin{aligned} I_{FET} &\approx i_{Lpk} \cdot \sqrt{\frac{D_{max}}{3}} \\ &= 0.895\,A \end{aligned}$$

To get the best performance from the converter, the MOSFET chosen has to have a current rating about three times the MOSFET RMS current with minimum gate charge Q_g. It is recommended that for designs with the HV9912, the gate charge of the chosen MOSFET be less than 25 nC.

The MOSFET chosen for this application is a 100 V, 4.5 A MOSFET with a Q_g of 11 nC.

6.4.8 Choosing the Switching Diode ($D1$)

The voltage rating of the diode is the same as the voltage rating of the MOSFET (100 V). The average current through the diode is equal to the maximum output current (350 mA). Although the average current through the diode is only 350 mA, the peak current through the diode is equal to i_{Lpk}. Thus, it is a better design approach to choose the current rating of the diode somewhere in between the peak input current and the average output current (preferably closer to the peak input current). Thus, for this design, the diode chosen is a 100 V, 2 A Schottky diode.

6.4.9 Choosing the Output Capacitor (C_o)

The value of the output capacitor depends on the dynamic resistance of the LED string as well as the ripple current desired in the LED string. In designs using the HV9912, a larger output capacitor (lower output current ripple) will yield better PWM dimming results. The capacitor required to filter the current appropriately will be designed by considering the fundamental component of the diode current only.

The output stage of the boost converter is modeled in Figure 6.11, where the LEDs are modeled as a constant voltage load with series dynamic impedance.

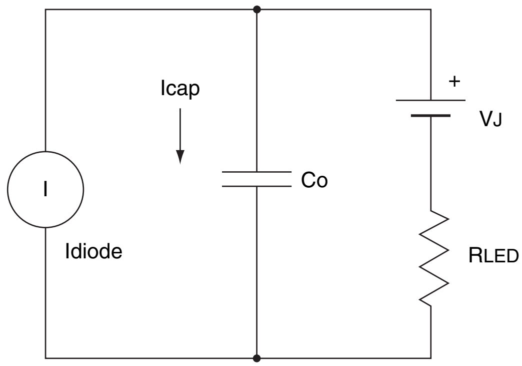


Figure 6.11: Model of Boost Converter Output.

The waveform of the capacitor current in steady state is shown in Figure 6.12.

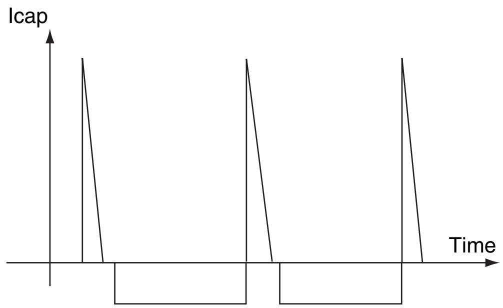


Figure 6.12: Output Capacitor Current.

Using the 10% peak-to-peak current ripple given in the design parameters table, the maximum voltage ripple across the LED string has to be

$$\begin{aligned}\Delta v_{\mathrm{p-p}} &= \Delta I_{\mathrm{o}} \cdot R_{\mathrm{LED}} \\ &= 0.55\,\mathrm{V}\end{aligned}$$

Assuming a constant discharging current of 350 mA when the diode current is zero, the equation for the voltage across the capacitor can be written as

$$I_{\mathrm{o\,max}} = C_{\mathrm{o}} \cdot \frac{\Delta v_{\mathrm{p-p}}}{D_{\max} \cdot T_s}$$

Substituting values into the above equation,

$$\begin{aligned} C_{\mathrm{o}} &= \frac{I_{\mathrm{o\,max}} \cdot D_{\max}}{\Delta v_{\mathrm{p-p}} \cdot f_{\mathrm{s}}} \\ &= 0.585\,\mu\mathrm{F} \end{aligned}$$

The RMS current through the capacitor can be given by

$$\begin{aligned} I_{\mathrm{rms}} &= \sqrt{(1 - D1) \cdot I_{\mathrm{o\,max}}{}^2 + \frac{D1}{3} \cdot (i_{\mathrm{Lpk}} - I_{\mathrm{o\,max}})^2} \\ &= 0.34\,\mathrm{A} \end{aligned}$$

In this case, a parallel combination of two 1 μF, 100 V metal polypropylene capacitors is chosen.

> **Note:** The proper type of capacitor to use is either metal film capacitors or ceramic capacitors, since they are capable of carrying this high ripple current. Although ceramic capacitors are smaller in size and capable of carrying the ripple current, they cause a lot of audible noise during PWM dimming. High value ceramic capacitors are usually limited to 50 V rating. Thus metal polypropylene (or any other metal film) capacitors are the ideal choice for LED drivers if PWM dimming is required.

6.4.10 Choose the Disconnect MOSFET ($Q2$)

The disconnect MOSFET should have the same voltage rating as the switching MOSFET $Q1$. The on-state resistance of the MOSFET at room temperature ($R_{\mathrm{on,25C}}$) has to be calculated based on a 1% power loss in $Q2$ at full load current. Thus,

$$\begin{aligned} R_{\mathrm{on,25C}} &= \frac{0.01 \cdot V_{\mathrm{o\,max}}}{I_{\mathrm{o\,max}} \cdot 1.4} \\ &= 5\,\Omega \end{aligned}$$

The 1.4 multiplication factor is included to account for the increase in the on-resistance due to rise in junction temperature. In this case, a high Q_{g} MOSFET can be chosen if desired (as it is not switching regularly), but a high Q_g MOSFET will slow down the turn-on and turn-off times (which might be allowable based on PWM dimming frequency). In this case, the MOSFET chosen is a 100 V, 0.7 Ω, SOT-23 MOSFET with a Q_g of 2.9 nC.

6.4.11 Choosing the Input Capacitors (*C*1 and *C*2)

The values of input capacitors *C*1 and *C*2 have to be calculated to meet closed loop stability requirements. The connection from the power source to the boost converter circuit will have some resistance, R_{source}, and some inductance, L_{source}. These feed across the input capacitors (*C*1 and *C*2) and so form an LC resonant circuit. To prevent interference with the control loop, the resonant frequency should be arranged to be 40% or less of the switching frequency.

A pair of 22AWG connecting wires 1 foot (30 cm) long will have an inductance of about 1 μH. This is a good starting point. If necessary, the wires can be twisted together to reduce the inductance.

With a 200 kHz switching frequency, the resonant frequency should be less than 80 kHz.

$$C_{IN} \geq \frac{1}{(2 \cdot \pi \cdot f_{LC})^2 \cdot L_{SOURCE}} = 3.95\,\mu F$$

$C1 = C2 = 2.2\,\mu F$, 50 V ceramic.

The maximum source impedance is found using:

$$M = \frac{V_{O,MAX}}{V_{IN,MIN}} = \frac{70}{9} = 7.778$$

$$R_{SOURCE,MAX} = \frac{M - 1}{M^2 \cdot (M - 2)} \cdot R_{LED} = 1.404\,\Omega$$

6.4.12 Choosing the Timing Resistor (R_T)

The HV9912 oscillator has an 18 pF capacitor charged by a current mirror circuit. An external timing resistor R_T provides a reference current for the current mirror.

When R_T is connected to 0 V, current flows and the timing process begins. When charged to a certain voltage, the RS flip-flop is set, the capacitor is discharged, and the timing process starts again. The timing resistor can be calculated by using the following equation:

$$\frac{1}{f_s} \approx R_T \cdot 18\text{pF}$$

In this case, for a constant 200 kHz switching frequency, the timing resistor value works out to 274 kΩ. This resistor needs to be connected between the R_T pin and GND as shown in the typical circuit.

6.4.13 Choosing the Two Current Sense Resistors (*R*1 and *R*2)

The value of the output current sense resistor $R2$ can be calculated by limiting its voltage drop to below 0.4 V. Using this criterion,

$$\begin{aligned} R2 &= \frac{0.4V}{I_{o\,max}} \\ &= 4\,\Omega \end{aligned}$$

The power dissipation will be 0.4 V*$I_{o\ max}$ = 0.04 W. In this case, the resistor chosen is a 3.9 Ω, 1/8 W, 1% resistor.

The MOSFET current sense resistor $R1$ is calculated by limiting the voltage across the resistor to about 250 mV at maximum input current.

$$\begin{aligned} R1 &= \frac{0.25}{i_{Lpk}} \\ &= 0.12\,\Omega \end{aligned}$$

The power dissipated in this resistor is

$$\begin{aligned} P_{R1} &= I_{FET}{}^2 \cdot R1 \\ &= 0.096\,\text{W} \end{aligned}$$

Thus, the chosen current sense resistor is a 0.12 Ω, 1/4 W, 1% resistor.

6.4.14 Selecting the Current Reference Resistors ($R3$ and $R4$)

The voltage at the current reference pin IREF can be set either by using the reference voltage provided at the REF pin (through a voltage divider) or with an external voltage source. In the present design, it is assumed that the voltage at the IREF pin is set using a voltage divider from the REF pin. The current reference resistors $R3$ and $R4$ can be computed using the following two equations:

$$R3 + R4 = \frac{1.25\text{V}}{50\,\mu\text{A}} \leq 25\,\text{k}\Omega$$

$$\frac{1.25\text{V}}{R3 + R4} \cdot R4 = I_{\text{o max}} \cdot R2 = 0.1 \cdot 3.9 = 0.39\text{V}$$

For this design, the values of the two resistors can be computed to be

$$R3 = 19.1\,\text{k}\Omega,\ 1/8\text{W},\ 1\%$$
$$R4 = 8.66\,\text{k}\Omega,\ 1/8\text{W},\ 1\%$$

6.4.15 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 from the REF pin ($R5$ and $R6$) is connected to the CLIM pin and sets the maximum inductor current. The voltage at the CLIM pin can be computed as

$$V_{\text{CLIM}} \geq 1.2 \cdot i_{\text{Ipk}} \cdot R1$$

This equation assumes that the current limit level is set at about 120% of the maximum inductor current $I_{\text{in max}}$.

For this design,

$$\begin{aligned} V_{\text{CLIM}} &= 1.2 \cdot 1.93 \cdot 0.12 \\ &\geq 0.278\,\text{V} \end{aligned}$$

Using a maximum current sourced out of REF pin of 50 µA, the two resistors can be calculated as

$$R5 = 20\text{k},\ 1/8\text{W},\ 1\%$$
$$\text{R6} = 6.04\text{k},\ 1/8\text{W},\ 1\%$$

No capacitor should be connected at the CLIM pin, because this will affect the circuit at start-up.

6.4.16 Capacitors at VDD and REF Pins

It is recommended that bypass capacitors be connected to both VDD and REF pins. For the VDD pin, the capacitor used should be a 10 V ceramic chip capacitor. For low power designs, a 1 µF is adequate. If the design uses high gate charge switching MOSFETs ($Q_g > 15$ nC), the capacitor at the VDD pin should be increased to 2.2 µF.

For the REF pin, the capacitor used is a 0.1 µF ceramic chip capacitor.

6.4.17 Setting the Over-Voltage Trip Point (*R*8 and *R*9)

The over-voltage trip point can be set at a voltage 15% higher than the maximum steady state voltage. Using a 15% margin, the maximum output voltage during open LED condition will be

$$\begin{aligned} V_{\text{open}} &= 1.15 \cdot V_{\text{o max}} \\ &= 80.5\text{V} \end{aligned}$$

Then, the resistors that set the over-voltage set point can be computed as

$$\begin{aligned} R8 &= \frac{\left(V_{\text{open}} - 5\right)^2}{0.1} \\ &= 57\,\text{k}\Omega \end{aligned}$$

The above equation will allow us to select a 1/8 W resistor by limiting the power dissipation in the resistor.

$$\begin{aligned} R9 &= \frac{R8}{\left(V_{\text{open}} - 5\right)} \cdot 1.25\text{V} \\ &= 3.77\,\text{k}\Omega \end{aligned}$$

The closest 1% resistor values are

$$R8 = 56.2\text{k},\ 1/8\text{W},\ 1\%$$
$$R9 = 3.74\text{k},\ 1/8\text{W},\ 1\%$$

Note: The actual over-voltage point will vary from the desired point by $\pm 5\%$ due to the variation in the reference (see datasheet). For this design, it varies from 76.67 V to 84.52 V.

6.4.18 Designing the Compensation Network

The compensation needed to stabilize the converter could be either a Type-I circuit (a simple integrator) or a Type-II circuit (an integrator with an additional pole-zero pair). The type of the compensation circuit required will be dependent on the phase of the power stage at the crossover frequency.

The loop gain of the closed loop system is given by

$$\text{Loop Gain} = R_s \cdot G_m \cdot Z_c(s) \cdot \frac{1}{15} \cdot \frac{1}{R_{cs}} \cdot G_{ps}(s)$$

Where G_m is the transconductance of the op-amp (435 µA/V), $Z_c(s)$ is the impedance of the compensation network, and $G_{ps}(s)$ is the transfer function of the power stage. Please note that although the resistors give a 1:14 ratio, the overall effect when including the diode drop is effectively 1:15.

To compute the transfer function for the discontinuous conduction mode boost converter in peak current control mode, we need to define a couple of factors.

$$M = \frac{V_{o\max} \cdot I_{o\max}}{V_{o\max} \cdot I_{o\max} - 0.5 \cdot L1_{nom} \cdot i_{Lpk^2} \cdot f_s}$$

$$M = \frac{70 \cdot 0.1}{70 \cdot 0.1 - 0.5 \cdot 15 \cdot 10^{-6} \cdot 1.93^2 \cdot 200 \cdot 10^3}$$

$$M = \frac{7}{1.41265} = 4.9552$$

$$G_R = \frac{M-1}{2 \cdot M - 1} = \frac{3.95522}{8.9104} = 0.4439$$

For frequencies less than one tenth of the switching frequency, the power stage transfer function is given by

$$G_{ps}(s) = 2 \cdot \frac{I_{o\,max}}{i_{Lpk}} \cdot \frac{G_R}{1 + s \cdot R_{LED} \cdot C_o \cdot G_R}$$

$$G_{ps}(s) = 2 \cdot \frac{0.1}{1.93} \cdot \frac{0.4439}{1 + s \cdot 55 \cdot 2 \cdot 10^{-6} \cdot 0.4439} = \frac{0.4439}{1 + s \cdot 48.829 \cdot 10^{-6}}$$

For the present design, choose a crossover frequency $\sim 0.01\, f_s$, or $f_c = 2$ kHz. The low crossover frequency will result in large values for C_C and C_Z, which will indirectly provide a soft start for the circuit. Since the HV9912 does not depend on the speed of the controller circuit for the PWM dimming response, the low crossover frequency will not have an adverse effect on the PWM dimming rise and fall times. By substituting $s = i \cdot (2\pi \cdot f_c) = i \cdot 12566$ into the transfer function, we get:

$$G_{ps}(s) = \frac{0.046}{1 + s \cdot 0.6136}$$

The magnitude and frequency of the power stage transfer function are:

$$\left|G_{ps}(s)\right|\Big|_{fc=2\mathrm{kHz}} = A_{ps} = 0.039$$
$$\angle G_{ps}(s)\Big|_{fc=2\mathrm{kHz}} = \phi_{ps} = -31.5^\circ$$

To get a phase margin of about $\phi_m = 45^\circ$ (the recommended phase margin range is 45°–60°), the phase boost required will be

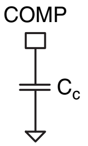
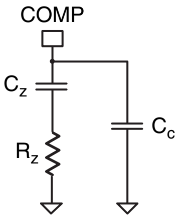
$$\begin{aligned}\phi_{boost} &= \phi_m - \phi_{ps} - 90^\circ \\ &= 45^\circ + 31.5^\circ - 90^\circ \\ &= -13.5^\circ\end{aligned}$$

Based on the value of the phase boost required, the type of compensation can be determined.

$$\phi_{boost} \le 0^\circ \Rightarrow \text{Type} - \text{I controller}$$
$$0^\circ \le \phi_{boost} \le 90^\circ \Rightarrow \text{Type} - \text{II controller}$$
$$90^\circ \le \phi_{boost} \le 180^\circ \Rightarrow \text{Type} - \text{III controller}$$

Type-III controllers are usually not required to compensate a HV9912-based boost LED driver and thus will not be discussed here. The implementations for the Type-I and Type-II systems for use with the HV9912 are given in Table 6.2.

Table 6.2: Compensation Networks.

Type	Circuit diagram	Transfer function
I		$Z_c(s) = \frac{1}{sC_c}$
II		$Z_c(s) = \frac{1}{s(C_c + C_z)} \cdot \frac{1 + s \cdot R_z \cdot C_z}{1 + s \cdot \frac{C_z \cdot C_c}{C_z + C_c} \cdot R_z}$

For the present design, a simple Type-I controller will suffice. All that is needed is to adjust the gain of the loop gain to be 1 at the crossover frequency.

One more equation can be obtained by equating the magnitude of the loop gain to 1 at the crossover frequency.

$$R2 \cdot G_m \cdot \left(\frac{1}{2 \cdot \pi \cdot f_c \cdot C_c}\right) \cdot \frac{1}{15} \cdot \frac{1}{R1} \cdot A_{ps} = 1$$

Transposing, we get:

$$C_C = R2 \cdot G_m \cdot \left(\frac{1}{2 \cdot \pi \cdot f_c}\right) \cdot \frac{1}{15} \cdot \frac{1}{R1} \cdot A_{ps}$$

$$C_C = 3.9 \cdot 435 \cdot 10^{-6} \cdot \left(\frac{1}{12566}\right) \cdot \frac{1}{15} \cdot \frac{1}{0.12} \cdot 0.039 = 2.92\,\text{nF}$$

Choose a $C_C = 3.3\,\text{nF}$, 50 V, C0G capacitor

This completes our DCM boost converter design.

6.5 Common Mistakes

1. The most common mistake is not having adequate over-voltage protection at the output. If the LEDs are disconnected while the circuit is operating, the output voltage will rise until components start to break down.
   The over-voltage limit set at the output of the boost converter should be lower than the breakdown voltage of any component connected across it.

2. Testing the circuit with a short string of LEDs. The forward voltage drop may be lower than the supply voltage, and in this case there is little to prevent the LEDs being destroyed by the high current that will flow.

6.6 Conclusions

Boost converters are used when the minimum output voltage is at least 1.5 times the input voltage. Continuous conduction mode should be used when the output voltage is a maximum of six times the input voltage. Discontinuous conduction mode is necessary if the output voltage is more than six times the input voltage. The EMI produced by a discontinuous mode boost converter is higher than for a continuous conduction mode boost converter of similar power output.

CHAPTER 7

Boost-Buck Converter

A boost-buck converter is a single-switch converter, which consists of a cascade of a boost converter followed by a buck converter. The power train of typical boost-buck circuit topology (used as an LED driver) is shown in Figure 7.1.

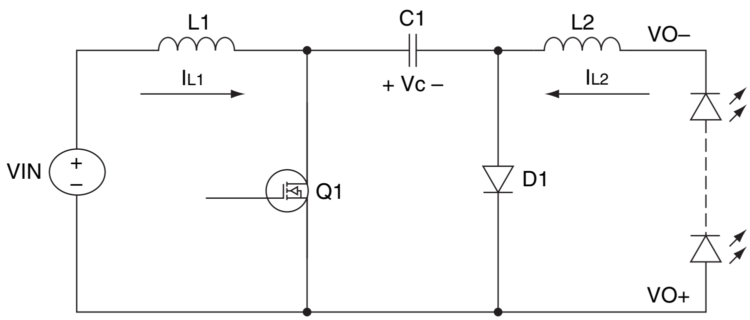


Figure 7.1: Boost-Buck (Cuk) Power Train.

The converter has many advantages:

- The converter can both boost and buck the input voltage. Thus, it is ideal for cases where the output LED string voltage can be either above or below the input voltage during operation. This condition is most common in automotive applications, or when a customer wants a single driver design to cover a wide range of voltage supply and load conditions.

- The converter has inductors on both the input and output sides. Operating both stages in continuous conduction mode (CCM) will enable continuous currents in both inductors with low current ripple, which would greatly reduce the filter capacitor requirements at both input and output. Continuous input current would also help greatly in meeting conducted EMI standards at the input.
- All the switching nodes in the circuit are isolated between the two inductors. The input and output nodes are relatively quiet. This will minimize the radiated EMI from the converter. With proper layout and design, the converter can easily meet radiated EMI standards.
- One of the advantages of the boost-buck converter is the capacitive isolation. The failure of the switching transistor will short the input and not affect the output. Thus, the LEDs are protected from failure of the MOSFET.
- The two inductors $L1$ and $L2$ can be coupled together on one core. When coupled on a single core, the ripple in the inductor current from one side can be transferred completely to another side (ripple cancellation technique). This would allow, for example, the input ripple to be transferred completely to the output side making it very easy for the converter to meet conducted EMI standards.

7.1 The Cuk Converter

In spite of the many advantages of the Cuk converter, a couple of **significant disadvantages** exist which prevent its widespread use.

- The converter is difficult to stabilize. Complex compensation circuitry is often needed to make the converter operate properly. This compensation also tends to slow down the response of the converter, which inhibits the PWM dimming capability of the converter (essential for LEDs).
- An output current controlled boost-buck converter tends to have an uncontrolled and undamped resonance due to an L-C pair ($L1$ and $C1$). The resonance of $L1$ and $C1$ leads to excessive voltages across the capacitor, which can damage the circuit.

The damping of $L1$ and $C1$ can easily be achieved by adding a damping R-C circuit across $C1$. However, the problem of compensating the circuit so that it is stable is more complex.

The Supertex HV9930 solves the problem of compensation and achieving a fast PWM dimming response by using hysteretic current mode control. This uses fast comparators to control a MOSFET gate by setting upper and lower limits, which ensures fast response and accurate current levels. However, a simple hysteretic current mode control would not work, as the converter would not be able to start up. To overcome this problem, the HV9930 has **two** hysteretic current mode controllers – one for the input current and another for the output current.

During start-up, the input hysteretic controller dominates and the converter is in input current limit mode. The MOSFET turns on and the input current rises until the input current limit is reached, it then turns off so that the input current drops until a lower current limit is reached. This cycle continues until the output current has built up to the required value and the output hysteretic controller can take over. The output current is then maintained between the set upper and lower current limits. Unlike peak current mode controller, hysteretic control ensures that the average output current remains constant under a wide range of input and output voltage conditions.

The hysteretic approach will also help in limiting the input current during start-up (thus providing soft-start); also current is limited in the case of an output overload or input under-voltage condition. Three resistors (for each of the two hysteretic controllers) are required to set both the current ripple and the average current, which enables a simple controller design. Thus six resistors determine the input and output performance.

This section will detail the operation of the boost-buck converter and the design of an HV9930-based converter. The design example is specifically designed for automotive applications, but it can also be applied for any DC/DC applications. At the time of writing, there is only one other device with the same functionality as the HV9930, which is the AT9933. The AT9933 has an automotive temperature specification (up to 125°C operation), whereas the HV9930 has an industrial temperature range.

7.1.1 Operation of a Cuk Boost-Buck Converter

The diagram of the power train for a Cuk boost-buck converter was shown previously, in Figure 7.1.

In steady state, the average voltages across both $L1$ and $L2$ are zero. Thus, the voltage, V_c, across the middle capacitor $C1$ is equal to the sum of the input and output voltages.

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

When switch $Q1$ is turned on, the currents in both inductors start ramping up (see Figure 7.2).

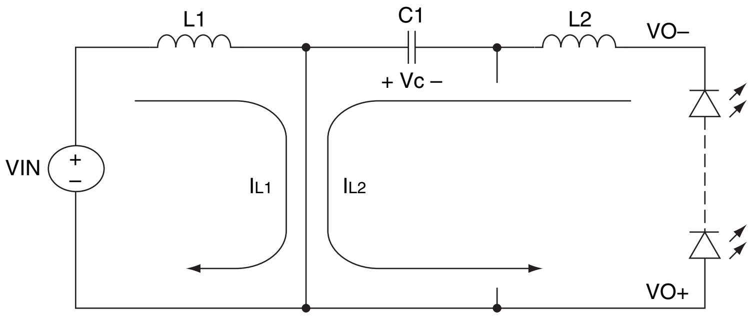


Figure 7.2: Cuk Circuit, MOSFET On.

$$L_1 \frac{di_{L1}}{dt} = V_{in}$$

$$L_2 \frac{di_{L2}}{dt} = V_c - V_o = V_{in}$$

When switch $Q1$ is turned off, the currents in both inductors start ramping down (see Figure 7.3).

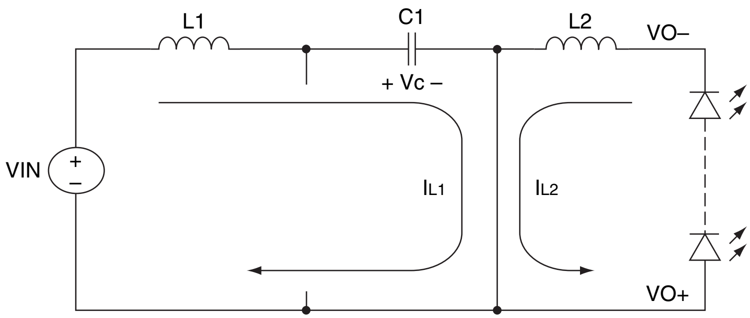


Figure 7.3: Cuk Circuit, MOSFET Off.

$$L_1 \frac{\mathrm{d}i_{L1}}{\mathrm{d}t} = V_{\mathrm{in}} - V_{\mathrm{c}} = -V_{\mathrm{o}}$$

$$L_2 \frac{\mathrm{d}i_{L2}}{\mathrm{d}t} = -V_{\mathrm{o}}$$

Assuming that the switch is ON for a duty cycle D and using the fact that, in steady state, the total volt-seconds applied across any inductor is zero, we get

$$V_{\mathrm{in}} \cdot (D) = V_{\mathrm{o}} \cdot (1 - D)$$
$$\Rightarrow \frac{V_{\mathrm{o}}}{V_{\mathrm{in}}} = \frac{D}{1 - D}$$

Thus, the voltage transfer function obtained for the boost-buck converter will give buck operation for $D < 0.5$ and boost operation for $D > 0.5$. The steady state waveforms for the converter are shown in Figure 7.4.

The maximum voltage seen by $Q1$ and $D1$ is equal to the voltage across the capacitor $C1$.

$$V_{Q1} = V_{D1} = V_{\mathrm{c}}$$

The standard boost-buck converter is modified, by adding three additional components, for proper operation of the HV9930 (see Figure 7.5).

A damping circuit R_{d}-C_{d} has been added to damp the $L1$-$C1$ pair. These additional components stabilize the circuit.

An input diode ($D2$) has been added. This diode is necessary for PWM dimming operation (in case of automobile applications, this could be the reverse polarity protection diode). This diode helps to prevent capacitors $C1$ and C_{d} from discharging when the gate signals for $Q1$ are turned off. Thus, when the HV9930 is enabled, the steady state output current level will be reached quickly.

7.1.2 Hysteretic Control of the Boost-Buck Converter

Hysteretic control refers to the control scheme where the controlled variable (in this case, the inductor current i_{L2}) is maintained between pre-set upper and lower boundaries. As previously shown in Figure 7.4, the inductor current ramps up at a

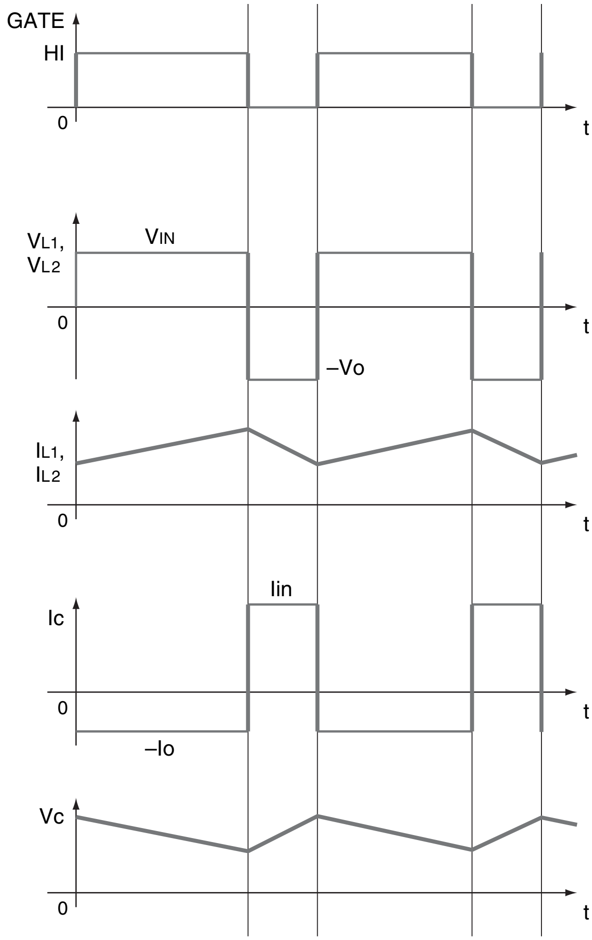


Figure 7.4: Cuk Converter Steady State Waveforms.

rate of $V_{\mathrm{in}}/L2$ when the switch is ON and ramps down at a rate of $-V_o/L2$ when the switch is OFF. Thus, the hysteretic control scheme turns the switch OFF when the inductor current reaches the upper limit and turns the switch ON when it reaches the lower limit.

The average current in inductor $L2$ is then set at the average of the upper and lower thresholds. The ON and OFF times (and thus the switching frequency) vary as the input and output voltages change to maintain the inductor current levels. However,

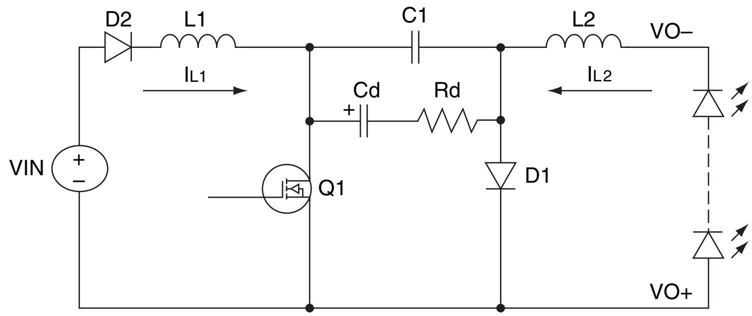


Figure 7.5: Modified Boost-Buck Circuit.

in any practical implementation of hysteretic control, there will be comparator delays involved. The switch will not turn ON and OFF at the instant the inductor current hits the limits, but after a small delay time, as illustrated in Figure 7.6.

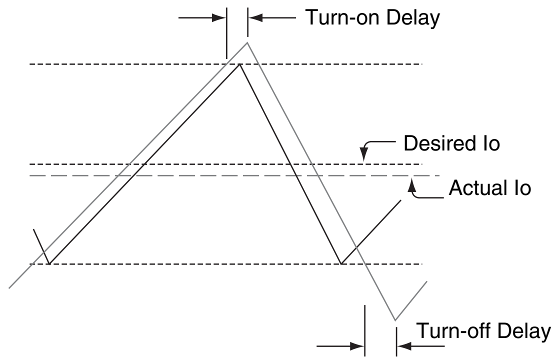


Figure 7.6: Current in the Output Inductor *L*2.

7.1.3 The Effects of Delay in Hysteretic Control

This delay time introduces two unwanted effects:

- It alters the average output current value. For example, if the delay on the down slope of the inductor current is more than the delay on the up slope, then the average current value decreases.
- It decreases the switching frequency, which may make it more difficult for the circuit to meet EMI regulations.

These effects will have to be taken into consideration when choosing the output inductor value and the setting the current limits.

Assume a peak-to-peak current ripple setting of Δi_o (using the programming resistors) and a desired average current l_o. A hysteretic current controlled boost-buck converter acts as a constant-off-time converter as long as the output voltage is fixed, and the off-time is theoretically independent of the input voltage. Thus, the converter is designed assuming a constant off-time T_{off} (the method to determine the off-time will be discussed later).

For the HV9930, as long as the switching frequencies are less than 150 kHz, these delay times have a negligible effect and can be ignored. In these cases, the output inductor can be determined by

$$L_2 = \frac{V_o \cdot T_{off}}{\Delta i_o}$$

If the inductor chosen is significantly different from the computed value, the actual off-time $T_{off,ac}$ can be recomputed using the same equation.

However, in automotive applications, it is advantageous to set the switching frequency of the converter below 150 kHz or in the range between 300 kHz and 530 kHz. This will place the fundamental frequency of the conducted and radiated EMI outside of the restricted bands making it easier for the converter to pass automotive EMI regulations. In cases where the switching frequency is more than 300 kHz, the delay times cannot be neglected and have to be accounted for in the calculations. Figure 7.7 illustrates the output inductor current waveform and the various rise and fall times.

From this figure,

$$\begin{aligned} T_{off} &= T_{f1} + T_{f2} + T_f \\ &= \frac{V_{in}}{V_o} \cdot T_r + T_{f2} + T_f \end{aligned}$$

The desired output current ripple Δi_o and the down-slope of the inductor current m_2 determine T_{f2}. The delay times of the HV9930 determine T_r and T_f. For the HV9930, the delay time of the comparators is related to the overdrive voltage

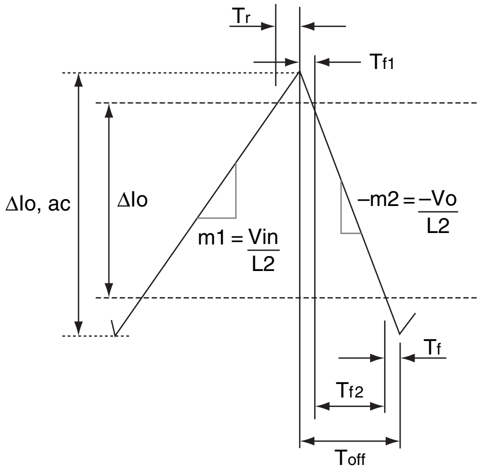


Figure 7.7: Hysteretic Control with Comparator Delays.

(voltage difference between the two input terminals of the current sense comparator) applied as

$$T_{\text{delay}} \approx \frac{K}{\sqrt[3]{m \cdot 0.1/\Delta i_o}}$$

Where '*m*' is the rising or falling slope of the inductor current.

$$T_r = \frac{6\,\mu}{\sqrt[3]{\dfrac{V_{\text{in}} \cdot 0.1}{\Delta i_o}}} \cdot \sqrt[3]{L_2} = K_1 \cdot \sqrt[3]{L_2}$$

$$T_{f2} = \frac{\Delta i_o \cdot L_2}{V_o} = K_2 \cdot L_2$$

$$T_f = \frac{6\,\mu}{\sqrt[3]{\dfrac{V_o \cdot 0.1}{\Delta i_o}}} \cdot \sqrt[3]{L_2} = K_3 \cdot \sqrt[3]{L_2}$$

To find the value of $L2$ using the time delay equations above results in a cubic equation. This cubic has one real root and two complex roots. The inductor value is the real root of the cubic raised to the third power.

$$a = K_2$$
$$b = \frac{V_{\text{in}}}{V_{\text{o}}} \cdot K_1 + K_3$$
$$c = T_{\text{off}}$$

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

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

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 $T_{\text{f}2}$, to get $T_{\text{r,ac}}$, $T_{\text{f,ac}}$ and $T_{\text{f}2\text{,ac}}$.

$$\begin{aligned} T_{\text{off,ac}} &= T_{\text{f}1\text{,ac}} + T_{\text{f}2\text{,ac}} + T_{\text{f,ac}} \\ &= \frac{V_{\text{in}}}{V_{\text{o}}} \cdot T_{\text{r,ac}} + T_{\text{f}2\text{,ac}} + T_{\text{f,ac}} \end{aligned}$$

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

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

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).

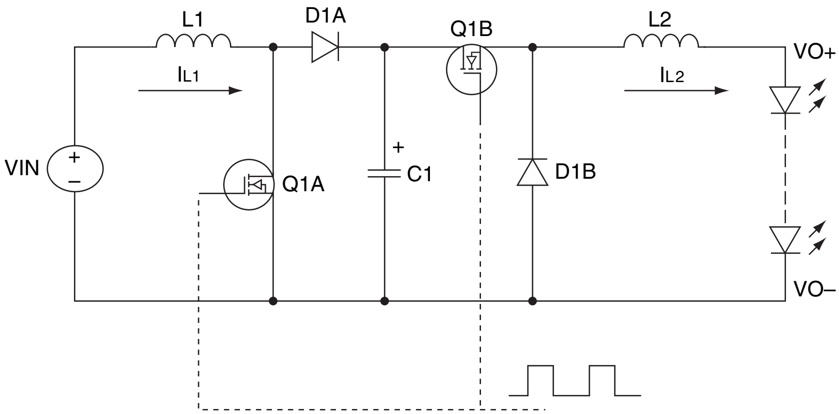


Figure 7.8: Boost-Buck Converter.

The relationships between the voltages in the system are

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

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.

In hysteretic control of the boost-buck converter using the HV9930, the output buck stage is controlled and the input boost stage is uncontrolled. An equivalent schematic of the HV9930 controlled boost-buck converter is shown in Figure 7.9.

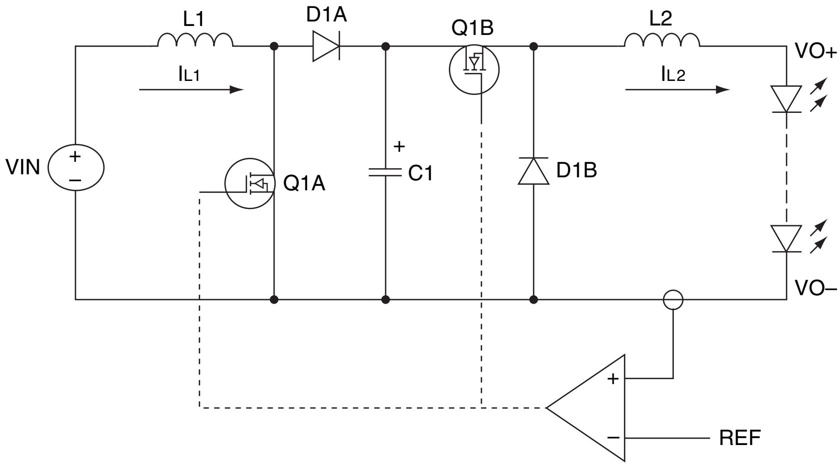


Figure 7.9: Boost-Buck Controller.

The hysteretic control of the buck stage ensures that the output current i_{L2} is constant under all input transient conditions. So, for the purposes of average modeling, the load seen by the capacitor $C1$ can be modeled as a current source equal to $d \cdot l_o$, where d is the instantaneous duty cycle and l_o is the constant output current. The continuous conduction mode buck stage also imposes one more constraint:

$$V_o = d \cdot v_c$$

where d and v_c are the time dependent duty cycle and capacitor voltage and V_o is the constant output voltage. For the system to be stable, it is necessary that the control system will act to reduce any disturbance in capacitor voltage.

The loop gain of the system for a boost-buck converter without damping has a negative phase margin (i.e. the phase is less than $-180°$ when the magnitude crosses 0 dB). This is due to the undamped LC pole-pair and causes the system to be unstable. Thus, any disturbance to the capacitor voltage will get amplified and keep increasing till the components breakdown. When testing the circuit, if it is close to becoming unstable, the switching frequency rises and falls with a low frequency beat and a low frequency ripple in the average output current can be seen.

The addition of *R*-*C* damping of this undamped pole pair can stabilize the system and make sure that the disturbance input is properly damped. Also, the presence of *Cd* ensures that *Rd* will not see the DC component of the voltage V_c across it, reducing the power dissipated in the damping resistor (*Cd* blocks the DC component of the voltage).

Assuming $Cd \gg C1$, the loop gain transfer function of the *R*-*C* damped boost-buck converter can be derived as

$$G(s)H(s) = \frac{D}{1-D} \cdot \frac{(1 + s \cdot Rd \cdot Cd) \cdot \left(1 - s \cdot \dfrac{D}{(1-D)^2} \cdot \dfrac{L1 \cdot I_o}{V_o}\right)}{(1 + s \cdot Rd \cdot C1) \cdot \left(1 + s \cdot Rd \cdot Cd + s^2 \cdot \dfrac{L1 \cdot Cd}{(1-D)^2}\right)}$$

Thus, the loop has a DC gain of $D/(1-D)$ and includes:

1. Damping (and ESR) zero at $\omega_Z = \frac{1}{Rd \cdot Cd}$.
2. RHP zero at $\omega_{RHP} = \frac{(1-D)^2}{D} \cdot \frac{Vo}{L1 \cdot Io}$.
3. Complex double pole with natural resonant frequency $\omega_o = \frac{1-D}{\sqrt{L1 \cdot Cd}}$ and damping factor $\delta = (1-D) \cdot Rd \cdot \sqrt{\frac{Cd}{L1}}$.
4. High frequency pole at $\omega_P = \frac{1}{Rd \cdot C1}$.

In order to achieve stable loop, the 0 dB crossing (ω_c) must be placed such that $\omega_c \ll \omega_{RHP}$ and $\omega_c \ll \omega_p$. The latter condition is easily met by selecting $Cd \gg C1$.

We can easily obtain approximate values of *Cd* and *Rd* for the case of $\omega_c \gg \omega_o$. This condition is usually met for the worst-case calculations at minimum input voltage, since the DC gain is the highest at this condition. Set $\omega_c = \omega_{RHP}/N$, where $N \gg 1$. Then ω_o can be approximately calculated from

$$\omega_o = \omega_c \cdot \sqrt{\frac{1-D}{D}} = \frac{\omega_{RHP}}{N} \cdot \sqrt{\frac{1-D}{D}}$$

Substituting for ω_o and ω_{RHP} in (21) gives the equation for computing Cd:

$$Cd = \frac{N^2 \cdot D^3}{(1-D)^3} \frac{L1 \cdot {I_o}^2}{{V_o}^2}$$

Selecting Rd such that $\omega_z = \omega_c$ results in a good phase margin with minimum power dissipation. Then, using equations for ω_z and ω_{RHP} gives a solution for Rd.

$$Rd = \frac{N \cdot D}{(1-D)^2} \frac{L1 \cdot I_o}{Cd \cdot V_o}$$

Using the equations above, the approximate values for the damping network can be computed using the following equations:

$$Cd = 9 \cdot \left(\frac{D}{1-D}\right)^3 \cdot L1 \cdot \left(\frac{I_o}{V_o}\right)^2$$

$$Rd = \frac{3 \cdot D}{(1-D)^2} \cdot \frac{L1 \cdot I_o}{Cd \cdot V_o}$$

Note that the damping resistor value includes the ESR of the damping capacitor. In many cases, the damping capacitor is chosen to be an electrolytic capacitor, which will have a significant ESR (sometimes a few ohms). In such cases, the damping resistor can be reduced accordingly.

7.1.5 Dimming Ratio Using PWM Dimming

The linearity in the dimming ratio achievable with the boost-buck depends on both the switching frequency and the PWM dimming frequency.

For a converter designed to operate at a minimum switching frequency of 300 kHz, one switching time period equals 3.33 µs. This is the minimum on-time of the PWM dimming cycle. At a PWM dimming frequency of 200 Hz (5 ms period), 3.33 µs equals a minimum duty cycle of 0.067%. This corresponds to a 1:1500 dimming range. However, the same converter being PWM dimmed at 1 kHz (1 ms time period) will have a minimum duty ratio of 0.33% or a PWM dimming range of 1:300.

If the minimum on-time of the PWM dimming cycle is less than the switching time period, the LED current will not reach its final value. Hence the average current will be less. Thus, the LEDs will dim, but there will be a loss of linearity between the average LED current and the duty cycle of the PWM input.

7.1.6 Design of the Boost-Buck Converter with HV9930

Specification

Input voltage: 9–16 V (13.5 V typical)

Transient voltage: 42 V (clamped load dump rating)

Reverse polarity protection: 14 V

Output voltage: 28 V maximum

Output current: 350 mA

LED resistance: 5.6 ohms

Estimated efficiencies: 72% minimum, 82% maximum (80% typical)

These efficiency values do not take into account the power loss in the reverse blocking diode. A Schottky diode will drop about $V_d = 0.5\,\mathrm{V}$ across it and thus will dissipate power in the range 0.4–0.6 W. This diode voltage drop will be taken into account while designing the converter.

The efficiency values used in this design are typical values for the given input voltages and output power level. Higher efficiencies can be obtained at lower input current levels (i.e. higher input voltages): the efficiency drop at lower input voltages is due to conduction losses caused by the correspondingly larger input currents. The efficiency values will depend on the operating conditions and, except in very high power designs, these values can be used as a good approximation.

Efficiencies higher than 85% can easily be achieved with the HV9930 controlled Cuk converter if the operating frequency is kept below 150 kHz. However, because of automotive EMI requirements, the higher efficiencies are traded off for higher switching frequencies (which increase switching losses in the system).

Consider a boost-buck converter circuit as shown in Figure 7.10.

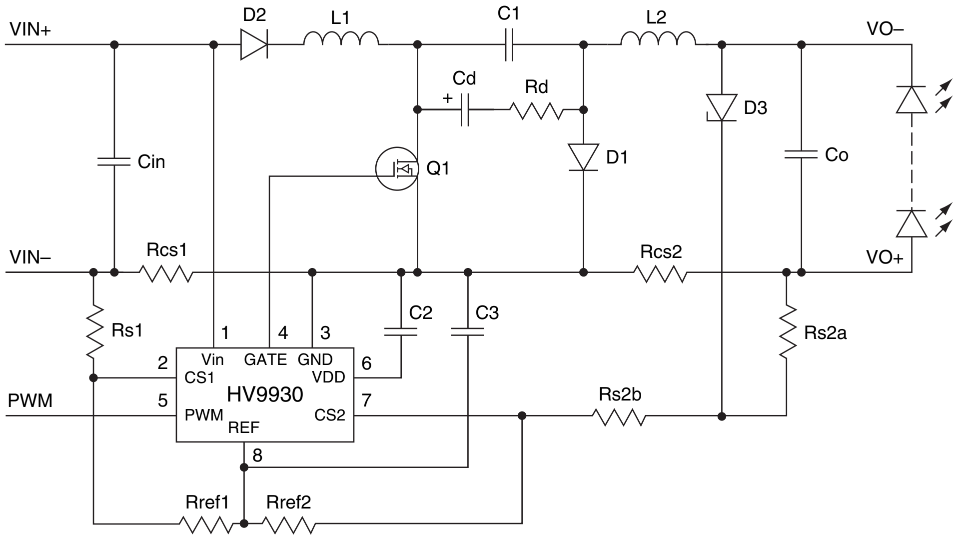


Figure 7.10: Boost-Buck Converter Using HV9930.

Switching Frequency at Minimum Input Voltage

Although the HV9930 is a variable frequency IC, the selection of the minimum switching frequency is important. In the case of automotive converters, designing with a switching frequency in the range between 300 kHz and 530 kHz would avoid the restricted radio broadcast bands and make it easier to meet the conducted and radiated EMI specifications. So, for this application we choose a minimum switching frequency of 300 kHz (which occurs at minimum input voltage).

Calculating the Duty Cycle

The switch duty cycle will have to be computed at the minimum input voltage.

$$D_{\max} = \frac{1}{1 + \frac{\eta_{\min} \cdot (V_{\text{in,min}} - V_{\text{d}})}{V_{\text{O}}}}$$
$$= 0.821$$

Calculating the Input Current

The input current level at the minimum input voltage should be calculated first, because this gives the highest current level. The value obtained will be used to work out the current ratings of the various components.

$$\begin{aligned} I_{\text{in,max}} &= \frac{V_{\text{o}} \cdot I_{\text{o}}}{\eta_{\text{min}} \cdot (V_{\text{in,min}} - V_{\text{d}})} \\ &= 1.601\text{A} \end{aligned}$$

Calculating the Output Inductor

The first step is to compute the off-time. The off-time of the converter can be calculated as

$$\begin{aligned} T_{\text{off}} &= \frac{1 - D_{\text{max}}}{f_{\text{s,min}}} \\ &= 598\,\text{ns} \end{aligned}$$

Assuming a 25% peak-to-peak ripple in the output current ($\Delta i_{\text{o}} = 87.5\,\text{mA}$), and accounting for the diode drop in the input voltage by substituting $V_{\text{in,min}} - V_{\text{d}}$ in place of V_{in}, yields

$$598\,\text{ns} = 0.887\,\mu \cdot \sqrt[3]{L_2} + 3.125\,m \cdot L_2 + 1.89\,\mu \cdot \sqrt[3]{L_2}$$

Solving for L_2 gives

$$L_2 = (0.052)^3 = 145\,\mu\text{H}$$

The closest standard value is a 150 µH, 0.35 A RMS, and 0.4 A saturation inductor.

Since the inductance value is different from the computed value, the actual off-time will also change as

$$\begin{aligned} T_{\text{off,ac}} &= 2.777\mu \cdot \sqrt[3]{L_{2,\text{ac}}} + 3.125\,m \cdot L_{2,\text{ac}} \\ &= 616\,\text{ns} \end{aligned}$$

The actual ripple in the output current is given by

$$\Delta i_{\mathrm{o,ac}} = \frac{V_{\mathrm{o}} \cdot T_{\mathrm{off,ac}}}{L_{2,\mathrm{ac}}}$$
$$= 0.115A$$

Note that although the ripple in the output current was assumed to be about 25% (or 87.5 mA), the actual ripple is almost double that value. This increase in the ripple is due to the delays of the comparators. A capacitor will be required at the output of the converter (across the LEDs) to reduce the ripple to the desired level. This capacitor will be very small, as the switching frequencies are large, but the capacitor will also help to reduce output EMI. Large output capacitors are to be avoided in applications that use PWM dimming, because the stored charge will reduce the dimming ratio that can be obtained.

It is also useful to calculate the ripple overshoot and undershoot beyond the programmed limits. This will help determine how the average current changes due to the delays.

$$\Delta i_{\mathrm{over}} = \frac{V_{\mathrm{o}}}{L_{2,\mathrm{ac}}} \cdot \left(\frac{V_{\mathrm{in,min}} - V_{\mathrm{d}}}{V_{\mathrm{o}}} \cdot K_1 \right) \cdot \sqrt[3]{L_{2,\mathrm{ac}}}$$
$$= 8.3\,\mathrm{mA}$$

$$\Delta i_{\mathrm{under}} = \frac{V_{\mathrm{o}}}{L_{2,\mathrm{ac}}} \cdot K_3 \cdot \sqrt[3]{L_{2,\mathrm{ac}}}$$
$$= 19\,\mathrm{mA}$$

Thus, the average output current will be reduced from the set value by about 10.7 mA.

In most cases, due to the inductor values available, the actual off-time will differ from the computed value significantly. Thus, it is better to use the actual value of the off-time calculated in order to work out the rest of the component values.

If the switching frequency is less than 150 kHz, the equation $L_2 = \frac{V_{\mathrm{o}} \cdot T_{\mathrm{off}}}{\Delta i_{\mathrm{o}}}$ can be used to calculate the output inductance ($L2$) value, simplifying the procedure greatly.

Calculating the Input Inductor

We can assume a 15% peak-to-peak ripple in the input current at minimum input voltage (this low input ripple will minimize the input filtering capacitance needed). The off-time previously calculated can be used to find the value of the input inductor.

$$L_1 = \frac{V_o \cdot T_{off,ac}}{0.15 \cdot I_{in,max}}$$
$$= 72\,\mu H$$

The closest standard value inductor is an 82 µH inductor. The current rating of this inductor will be decided in the final stages after the input current limit has been set.

The peak-to-peak ripple in the input current is

$$\Delta I_{in} = \frac{V_o \cdot T_{off,ac}}{L_{1,ac}}$$
$$= 0.21\text{A}$$

Calculating the Value of the Middle Capacitor (C1)

Assuming a 10% ripple across the capacitor at minimum input voltage ($\Delta v_c = 0.1 \cdot (V_{in,min} - V_d + V_o) = 3.65$ V), capacitor $C1$ can be calculated as

$$C_1 = \frac{I_{in,max} \cdot T_{off,ac}}{\Delta v_c}$$
$$= 0.257\,\mu F$$

$$I_{rms,C1} = \sqrt{I^2_{in,max} \cdot (1 - D_{max}) + I^2_o \cdot D_{max}}$$
$$= 0.72 \text{ A}$$

The voltage rating and type of this capacitor have to be chosen carefully. This capacitor carries both the input current and the output current. Thus, to prevent excessive losses and overheating of the capacitor, it must have a very low ESR. Ceramic capacitors are an ideal choice for this application due to their low ESR and high transient voltage limit. If a ceramic capacitor cannot be used for reasons of 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 V + 16 V), and the maximum transient voltage across the capacitor $V_{c,max}$ is 70 V (= 28 V + 42 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 μF, 50 V X7R ceramic chip capacitor can be selected.

Choosing the Switching Transistor (Q1)

The peak voltage across the MOSFET $Q1$ 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

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

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

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

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 mΩ 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.

A resistor in series with the MOSFET gate reduces EMI by slowing down the turn-on time. Current transients are limited when the MOSFET turns on slowly, but this reduces efficiency. A PNP transistor to discharge the MOSFET gate helps to minimize the reduction in efficiency without significantly increasing EMI.

Choosing the Switching Diode

The maximum voltage rating of the diode $D2$ is the same as the MOSFET voltage rating. The average current through the diode is equal to the output current.

$$I_{\text{diode}} = I_{\text{o}} = 350\,\text{mA}$$

Although the average current of the diode is only 350 mA, the actual switching current through the diode goes as high as 1.95 A ($I_{\text{in,max}} + I_{\text{o}}$). (Note: the calculations were for 360 mA, to allow for 10 mA drop because of delays, but the actual average current is 350 mA.) A 500 mA diode will be able to carry the 1.79 A current safely, but the voltage drop at such high current levels would be extremely large, increasing the power dissipation. Thus, we need to choose a diode whose current rating is at least 1 A. A 100 V, 2 A Schottky diode would be a good choice. Choosing a voltage rating significantly higher than required is not a good idea, since generally the forward voltage drop increases as the reverse voltage rating increases and this causes higher conduction losses.

Choosing the Input Diode

The input diode serves two purposes:

1. It protects the circuit from a reverse polarity connection at the input.
2. It helps in PWM dimming of the circuit by preventing $C1$ from discharging when the HV9930 is turned off.

The current rating of the device should be at least equal to $I_{\text{in,max}}$. The voltage rating of the device should be more than the reverse input voltage rating. A higher current rating often gives a lower forward voltage drop. In this case, a 30BQ015 (15 V, 3 A Schottky diode) would be a good choice.

If neither reverse protection or PWM dimming is required, removing the input diode from the LED driver circuit will increase the input supply voltage at the converter, which will slightly increase the efficiency and slightly reduce the maximum input current.

Calculating the Input Capacitance

Some capacitance is required on the input side to filter the input current. This capacitance is mainly responsible for reducing the 2nd harmonic of the input current ripple (which in this case falls in the AM radio band). According to the SAE J1113 specifications, the peak limit for narrowband emissions in this range is 50 dBµV to meet Class 3 at an input voltage of $13 \pm 0.5\,\mathrm{V}$. Assuming a saw tooth waveform for the input current as a conservative approximation, the RMS value of the 2nd harmonic component of the input current ($I_{\mathrm{in,2}}$) can be computed as

$$I_{\mathrm{in,2}} = \frac{\Delta I_{\mathrm{in}}}{2 \cdot \sqrt{2} \cdot \pi} = 0.024\ \mathrm{A}$$

The switching frequency of the converter at 13 V input can be computed as

$$\begin{aligned} D_{\mathrm{nom}} &= \frac{1}{1 + \dfrac{\eta_{\mathrm{nom}} \cdot (V_{\mathrm{in,nom}} - V_{\mathrm{d}})}{V_{\mathrm{o}}}} \\ &= \frac{1}{1 + \dfrac{0.8 \cdot (13.5 - 0.5)}{28}} \\ &= 0.73 \\ f_{\mathrm{s,nom}} &= \frac{1 - D_{\mathrm{nom}}}{T_{\mathrm{off,ac}}} \\ &= 414\ \mathrm{kHz} \end{aligned}$$

$$\begin{aligned} C_{\mathrm{in}} &= \frac{I_{\mathrm{in,2}}}{4 \cdot \pi \cdot f_{\mathrm{s,nom}} \cdot 10^{-6} \cdot 10^{50/20}} \\ &= 14.6\,\mu\mathrm{F} \end{aligned}$$

Choose a parallel combination of three 4.7 µF, 25 V, X7R ceramic capacitors.

Calculating the Output Capacitance

The output capacitance is required to reduce the LED current ripple from 115 mA to $\Delta I_{LED} = 70$ mA (20% peak to peak ripple) can be approximately calculated by using only the first harmonic in the inductor current. A 70 mA peak-to-peak ripple in the LED results in a 392 mV ($\Delta v_o = \Delta I_{LED} \cdot R_{LED}$) peak-to-peak ripple voltage. Then

$$\frac{\Delta v_o}{2} = \frac{8}{\pi^2} \cdot \left(\frac{\Delta i_{L2}}{2}\right) \cdot \frac{R_{LED}}{\sqrt{1 + (2 \cdot \pi \cdot f_{s,min} \cdot R_{LED} \cdot C_o)^2}}$$

The output capacitance required can then be calculated from this

$$C_o = \frac{\sqrt{\left(\frac{16 \cdot R_{LED}}{\pi^2} \cdot \frac{\Delta i_{L2}}{\Delta v_o}\right)^2 - 1}}{2 \cdot \pi \cdot f_{s,min} \cdot R_{LED}} = 0.178\,\mu\text{F}$$

Use a 0.22 μF, 35 V ceramic capacitor.

Calculating the Theoretical Switching Frequency Variation

The maximum and minimum frequencies (using steady state voltage conditions) can be now be worked out:

$$f_{s,min} = \frac{1 - \dfrac{1}{1 + \eta_{min} \cdot (V_{in,min} - V_d)/V_o}}{T_{off,ac}} = 291\,\text{kHz}$$

$$f_{s,max} = \frac{1 - \dfrac{1}{1 + \eta_{max} \cdot (V_{in,max} - V_d)/V_o}}{T_{off,ac}} = 506\,\text{kHz}$$

The theoretical frequency variation for this design is 398 kHz $\pm$ 27%.

Design of the Damping Circuit

The values for the damping network can be calculated as follows

$$\begin{aligned} Cd &= 9 \cdot \left(\frac{D_{\max}}{1 - D_{\max}}\right)^3 \cdot L_{1,\mathrm{ac}} \cdot \left(\frac{I_\mathrm{o}}{V_\mathrm{o}}\right)^2 \\ &= 11\,\mu\mathrm{F} \\ Rd &= \frac{3 \cdot D_{\max}}{(1 - D_{\max})^2} \cdot \frac{L_{1,\mathrm{ac}} \cdot I_\mathrm{o}}{Cd \cdot V_\mathrm{o}} \\ &= 7.16\,\Omega \end{aligned}$$

The power dissipated in Rd can be computed as

$$\begin{aligned} P_{Rd} &= \frac{\Delta v_\mathrm{c}{}^2}{12 \cdot R_\mathrm{d}} \\ &= \frac{3.65^2}{12 \cdot 7.16} = 0.155\,\mathrm{W} \end{aligned}$$

The RMS current through the damping capacitor will be

$$i_{Cd} = \frac{\Delta v_\mathrm{c}}{2 \cdot \sqrt{3} \cdot R_\mathrm{d}} = 0.147\,\mathrm{A}$$

Choose a 10 μF, 50 V electrolytic capacitor that can allow at least 150 mA RMS current. An example would be EEVFK1H100P from Panasonic (10 μF, 50 V, Size D). This capacitor has about a 1 Ω ESR, so Rd can be reduced to about 6.2 Ω.

Internal Voltage Regulator of the HV9930

The HV9930 includes a built-in 8–200 V linear regulator. This regulator supplies the power to the IC. This regulator can be connected at either one of two nodes on the circuit as shown in Figure 7.11.

In the normal case, when the input voltage is always greater than 8 V, the VIN pin of the IC can be connected to the cathode of the input protection diode (as shown in

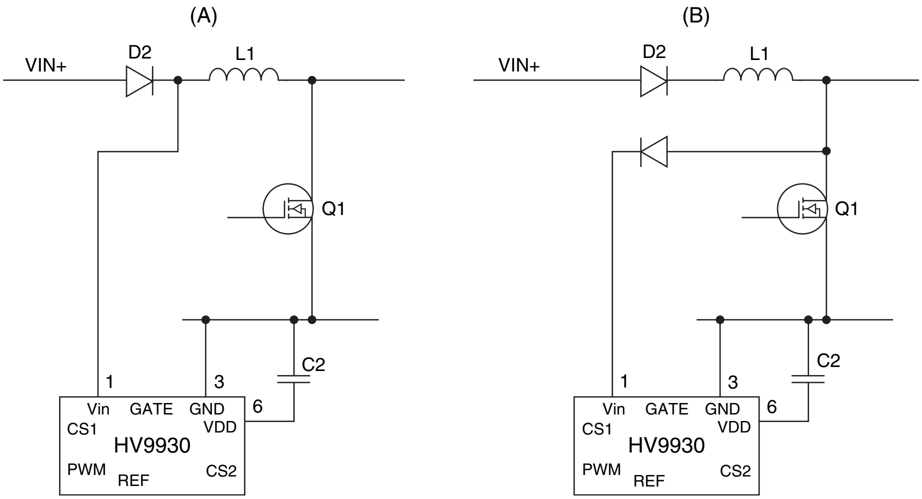


Figure 7.11: Connection Points for VIN.

Figure 7.11A). If reverse protection is not provided, the VIN pin can be connected directly to the positive supply.

In conditions where the converter needs to operate at voltages lower than 8 V, once the converter is running (as in the case of cold-crank operation), the VIN pin of the HV9930 can be connected as shown in Figure 7.11B. In this case, the drain of the MOSFET is at $V_{in} + V_o$, and hence even if the input voltage drops below 8 V, the IC will still be functioning. However, in this case, more hold-up capacitance will be required at the VDD pin to supply the power to the IC when the MOSFET is ON.

In both cases, a 2.2 μF or greater value ceramic capacitor is recommended at the VDD pin.

Internal Voltage Reference

The HV9930 includes an internal 1.25 V ($\pm$3%) reference. This reference can be used to set the current thresholds for the input and output hysteretic comparators. It is recommended that this pin be bypassed with at least a 0.1 μF ceramic capacitor.

Programming the Hysteretic Controllers and Over Voltage Protection

The input and output current levels for the hysteretic controllers are set by means of three resistors for each current – one current sense resistor and two divider resistors. The equations governing the resistors are the same for both the input and output sides and are given as

$$\frac{R_s}{R_{ref}} = \frac{0.05 \cdot \frac{\Delta i}{I} + 0.1}{1.2 \cdot \frac{\Delta i}{I} - 0.1}$$

$$R_{cs} = \frac{1.2 \cdot \frac{R_s}{R_{ref}} - 0.05}{I}$$

These equations assume that the 1.25 V reference provided by the HV9930 is used to set the current. In cases where linear dimming of the LEDs is required, it is recommended that the input current thresholds be based on the 1.25 V reference and the output current thresholds are modified using the variable input voltage available. In such a case, assuming the maximum external voltage V_{LD} as the reference, the above two equations can be modified as

$$\frac{R_s}{R_{ref}} = \frac{0.05 \cdot \frac{\Delta i}{I} + 0.1}{(V_{LD} - 0.05) \cdot \frac{\Delta i}{I} - 0.1}$$

$$R_{cs} = \frac{(V_{LD} - 0.05) \cdot \frac{R_s}{R_{ref}} - 0.05}{I}$$

In this design example, it is assumed that linear dimming is not required and the 1.25 V reference is used for both the input and output programming.

Note: The HV9930 cannot operate the boost-buck converter in the discontinuous conduction mode. The minimum external voltage is given by

$$V_{LD} = 0.1 \cdot \frac{R_{ref\,2} + R_{s2}}{R_{s2}}.$$

The programming of the output side is also linked to the over-voltage protection. The boost-buck converter is not inherently programmed against open LED conditions, so external protection is required. This is achieved by adding Zener diode $D3$, and by splitting the resistor $Rs2$ into two parts – $Rs2a$ and $Rs2b$. In normal operation, the inductor current will flow only through $Rcs2$ and the voltage drop across $Rcs2$ is sensed through $Rs2a$ and $Rs2b$ in series.

When there is an open LED condition, the inductor current will flow through diode $D3$. This will then clamp the output to the Zener breakdown voltage. However, since the diode cannot take the full design current, the current level has to be reduced to more manageable levels. During open LED conditions, the current will flow though both $Rcs2$ and $Rs2a$. Thus, the effective current sense resistor seen by the IC is $Rcs2 + Rs2a$ and the voltage drop across both of these will be sensed through $Rs2b$. This, in effect, will reduce the programmed current level and thus prevent the high LED currents from flowing into the Zener diode.

Choosing the Output Side Resistors

For the output current, $I_o = 0.36$ A (to compensate for the 10 mA drop due to the delay times) and $\Delta I_o = 87.5$ mA. Note that we are using the values assumed and not the actual values computed for the ripple current. Using these values in the above equations,

$$\frac{R_{s2a} + R_{s2b}}{R_{ref\,2}} = 0.534$$

$$R_{cs2} = 1.64\,\Omega$$

$$P_{Rcs2} = 0.35^2 \cdot 1.64 = 0.2\,\text{W}$$

Before we complete the design of the output side, we also have to design the over-voltage protection. For this application, choose a 33 V Zener diode. This is the voltage at which the output will clamp in case of an open LED condition. For a 350 mW diode, the maximum current rating at 33 V works out to about 10 mA. Using a 5 mA current level during open LED conditions, and assuming the same R_s/R_{ref} ratio,

$$R_{s2a} + R_{cs2} = 120\,\Omega$$

Choose the following values for the resistors:

$$R_{cs2} = 1.65\,\Omega,\ 1/4\ W,\ 1\%$$

$$R_{ref\,2} = 10\,k\,\Omega,\ 1/8\ W,\ 1\%$$

$$R_{s2a} = 100\,\Omega,\ 1/8\ W,\ 1\%$$

$$R_{s2b} = 5.23\,k\,\Omega,\ 1/8\ W,\ 1\%$$

Design of the Input Side Resistors

For the input side, we first have to determine the input current level for limiting. This current level is dictated by the fact the input comparator must not interfere with the operation of the circuit, even at minimum input voltage.

The peak of the input current at minimum input voltage will be

$$\begin{aligned} I_{in,pk} &= I_{in,max} + \frac{\Delta I_{in}}{2} \\ &= 1.706\,\mathrm{A} \end{aligned}$$

Assuming a 30% peak-to-peak ripple when the converter is in input current limit mode, the minimum value of the input current will be

$$I_{lim,min} = 0.85 \cdot I_{in,lim}$$

We need to ensure that $I_{lim,min} > I_{in,pk}$ for proper operation of the circuit. Assuming a 5% safely factor, i.e.,

$$I_{lim,min} = 1.05 \cdot I_{in,pk}$$

We can compute the input current limit to be $I_{in,lim} = 2.1$ A. Allowing for a 30% peak-to-peak ripple, we can calculate

$$\frac{R_{s1}}{R_{ref1}} = 0.442$$

$$\begin{aligned} R_{cs1} &= 0.228\,\Omega \\ P_{Rcs1} &= I_{in,lim}^2 \cdot R_{cs1=1\,\mathrm{W}} \end{aligned}$$

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.

$$I_{\text{in,nom}} = \frac{28 \cdot 0.35}{0.8 \cdot (13.5 - 0.5)}$$
$$= 0.942\,\text{A}$$

$$P_{\text{Rcs1}} = 0.942^2 \cdot 0.228 = 0.2\,\text{W}$$

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 resistors:

$$R_{\text{cs1}} = \textit{parallel combination of three } 0.68\,\Omega,\ 1/2\,\text{W},\ 5\%\ \textit{resistors}$$
$$R_{\text{ref 1}} = 10\,\text{k}\,\Omega,\ 1/8\,\text{W},\ 1\%$$
$$R_{\text{s1}} = 4.42\,\text{k}\,\Omega,\ 1/8\,\text{W},\ 1\%$$

Input Inductor Current Rating

The maximum current through the input inductor is $I_{\text{lim,max}} = 1.15 \cdot I_{\text{in,lim}} = 2.4\,\text{A}$. 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 (RA) 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, $R_{\text{S1}} = R4$, $R_{\text{REF1}} = R7$, and $R_{\text{CS1}} =$ the parallel combination $R1//R3$.

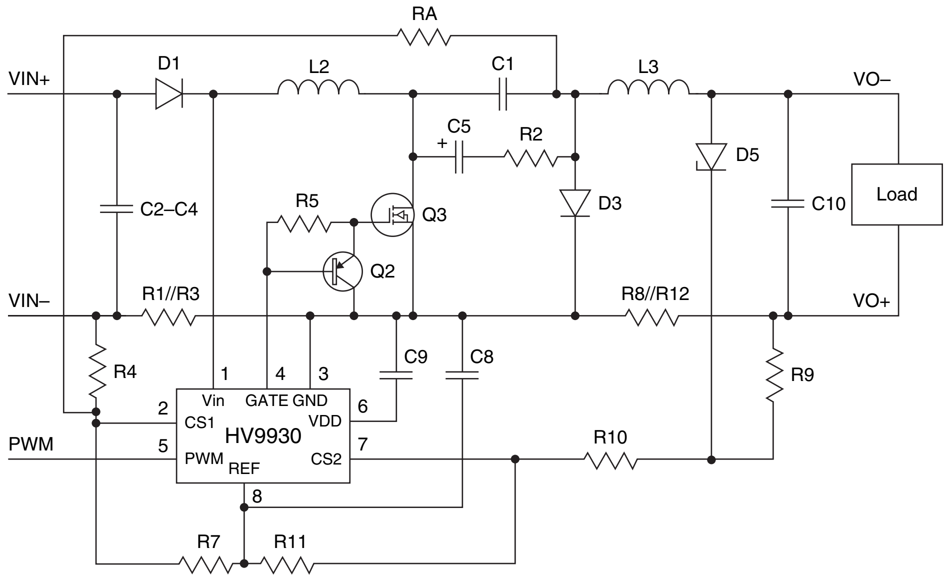


Figure 7.12: Modification of the Cuk Circuit.

Consider the circuit during the period when the MOSFET is ON, so that the input current through $L1$ is increasing by $\Delta I_{IN}/2$, until it reaches $I_{IN,LIM} + \Delta I_{IN}/2$. With the MOSFET turned ON, the positive side of the capacitor $C1$ is grounded and the other side of $C1$, which is connected to resistor RA, is at potential $-V_{C1}$. Note that the potential $-V_{C1,NOM} = V_{IN,NOM} + V_O$. The voltage reference for the comparator input at $CS1$ is 0 V. Now consider the node at $CS1$ in terms of current flow; $CS1$ is high impedance input, so the sum of currents equal zero:

$$\frac{V_{REF}}{R7} = \frac{V_{C1,NOM}}{RA} + \frac{\left(I_{IN,LIM} + \frac{\Delta I_{IN}}{2}\right) \cdot R1//R3}{R4}$$

Now consider the circuit when the MOSFET is OFF. Now the flywheel diode $D3$ is conducting, so the negative side of capacitor $C1$ is grounded (the small forward voltage of the diode can be ignored). With the MOSFET turned OFF, the voltage reference for the comparator input at $CS1$ is 100 mV.

$$\frac{V_{\text{REF}} - 0.1\text{V}}{R7} = \frac{0.1\text{V}}{RA} + \frac{0.1\text{V} + \left(I_{\text{IN,LIM}} - \frac{\Delta I_{\text{IN}}}{2}\right) \cdot R1//R3}{R4}$$

Since $R4$ is a very large value and the voltage across it is small, we can ignore its effect to simplify the calculations:

$$\frac{V_{\text{REF}} - 0.1\text{V}}{R7} = \frac{0.1\text{V} + \left(I_{\text{IN,LIM}} - \frac{\Delta I_{\text{IN}}}{2}\right) \cdot R1//R3}{R4}$$

We can thus ignore the addition of RA during the period that the MOSFET is turned OFF. Clearly, the value of $R1//R3$ can be reduced if the current $\left(I_{\text{IN,LIM}} - \frac{\Delta I_{\text{IN}}}{2}\right)$ can be increased, or if $R4$ can be decreased, or both.

The maximum current sense voltage occurs when the MOSFET is first turned ON.

$$V_{\text{SENSE,MAX}} = \left(I_{\text{IN,LIM}} + \frac{\Delta I_{\text{IN}}}{2}\right) \cdot R1//R3$$

This is a function of the voltage across the capacitor $C1$. If we take another look at the equation for current flow when the MOSFET is turned ON:

$$\frac{V_{\text{REF}}}{R7} = \frac{V_{\text{C1,NOM}}}{RA} + \frac{\left(I_{\text{IN,LIM}} + \frac{\Delta I_{\text{IN}}}{2}\right) \cdot R1//R3}{R4}$$

In a Cuk topology, $V_{\text{C1}} = V_{\text{IN}} + V_{\text{OUT}}$. At start-up, $V_{\text{OUT}} = 0\,\text{V}$, so $V_{\text{C1,MIN}} = V_{\text{IN,MIN}}$. The highest input current occurs at $V_{\text{IN,MIN}}$.

$$\frac{V_{\text{REF}}}{R7} = \frac{V_{\text{IN,MIN}}}{RA} + \frac{\left(I_{\text{IN,LIM}} + \frac{\Delta I_{\text{IN}}}{2}\right) \cdot R1//R3}{R4}$$

If we set the maximum current $\left(I_{\text{IN,LIM}} + \frac{\Delta I_{\text{IN}}}{2}\right)$ in the modified circuit to be equal to the inductor $L1$ saturation current, I_{SAT}, we get

$$\frac{V_{\text{REF}}}{R7} = \frac{V_{\text{IN,MIN}}}{RA} + \frac{I_{\text{SAT}} \cdot R1//R3}{R4}$$

In practice we start with the design of an unmodified circuit, so $\left(I_{IN,LIM} + \frac{\Delta I_{IN}}{2}\right)$ are the values calculated before the addition of RA is considered. In the modified circuit, I_{SAT} (of $L1$) must be much higher than these values in order to gain the loss reduction benefit, which gives a higher input ripple at start-up.

$$RA = \frac{\left(V_{IN,NOM} + V_{OUT}\right) - \dfrac{V_{IN,MIN} \cdot (I_{IN,LIM} + \Delta I_{IN})}{I_{SAT}}}{\dfrac{V_{REF1}}{R7} \cdot \left(1 - \dfrac{(I_{IN,LIM} + \Delta I_{IN})}{I_{SAT}}\right)}$$

$$R4(\text{mod}) = \frac{0.1\text{V}}{\dfrac{V_{REF1} - 0.1\text{V}}{R7} - \dfrac{(I_{IN,LIM} - \Delta I_{IN})}{I_{SAT}} \cdot \left(\dfrac{V_{REF1}}{R7} - \dfrac{V_{IN,MIN}}{RA}\right)}$$

$$R1//R3(\text{mod}) = \frac{R4(\text{mod})}{I_{SAT}} \cdot \left(\frac{V_{REF1}}{R7} - \frac{V_{IN,MIN}}{RA}\right)$$

$V_{REF1} = 1.25\,\text{V}$ in the standard configuration.

Meeting Conducted and Radiated EMI

Due to the nature of the boost-buck converter, it is easy to meet conducted and radiated EMI specifications. A few precautions need to be taken during design and PCB layout to be able to meet the EMI standards.

1. In some cases, when the input current ripple is too large or the switching frequency of the converter is above 150 kHz, it might not be possible to meet the conducted EMI standards using only capacitors at the input. In such cases, an input PI filter might be required to filter the low frequency harmonics.

2. Shielded inductors or toroidal inductors should always be preferred over unshielded inductors. These inductors will minimize radiated magnetic fields.

3. During layout, the IC and MOSFET ground connection should be connected to a copper plane on one of the PCB layers with the copper plane extending under the inductors.

4. The loop consisting of $Q1$, $C1$ and $D1$ should be as small as possible. This would help greatly in the meeting the high frequency EMI specifications.

5. The length of the trace from GATE output of the HV9930 to the GATE of the MOSFET should be as small as possible, with the source of the MOSFET and the GND of the HV9930 being connected to the GND plane. A low value resistor (10–47 ohms) in series with GATE connection will slow down the switching edges and greatly reduce EMI, although this will cause efficiency to decrease slightly. A PNP transistor to discharge the gate quickly helps to limit the decrease in efficiency, without adding any significant EMI.
6. An *R-C* damping network might be necessary across diode *D*1 to reduce ringing due to the undamped junction capacitance of the diode.

This concludes the Cuk converter design. We can now consider a closely related circuit; the SEPIC.

7.2 SEPIC Buck-Boost Converters

The abbreviation SEPIC comes from the description Single Ended Primary Inductance Converter. A SEPIC is a boost-buck converter, like a Cuk, so its input voltage range can overlap the output voltage. SEPIC circuits can be designed for constant voltage or constant current output.

The SEPIC topology has been known for some time, but only recently has there been a revival in its application because: (a) it needs low ESR capacitors and these are now widely available and (b) it can be used to create AC input power supplies with power-factor correction that are used to meet worldwide EMI standards.

In automotive and portable applications, batteries are used as a power source for DC/DC converters. A 12 V supply used in automotive applications can have a wide range of terminal voltage, typically 9 V to 16 V during normal operation using a lead-acid battery, but can go as low as 6.5 V during cold-crank and as high as 90 V during load-dump (when the battery is disconnected). The peak voltage is usually clamped to about 40 V, using a voltage dependent resistor to absorb the energy.

Lithium batteries have been very successful in portable applications, thanks mostly to their impressive energy density. A single lithium cell provides an open voltage of 4.2 V when fully charged, and replaces up to three of the alternative NiCd or NiMH cells. During discharge the cell still retains some energy down to 2.7 V. This input voltage range can be both above and below the output of many DC/DC converters and so discounts the possibility of using boost or buck converters.

International standards for power supplies rated above 75 W require power-factor correction (PFC). Having a good power factor means that the current waveform from the AC line is sinusoidal and in phase with the voltage. Most PFC circuits use a simple step-up converter as the input stage, implying that the input stage output must exceed the peak value of the input waveform. In Europe AC inputs of 190–265 V RMS are found, which impose an output of at least 375 V, forcing the following converters to work with elevated input voltages. Typically a PFC input stage has a 400 V output.

By using a SEPIC topology, which has a boost-buck topology, the boost section provides PFC and the buck section produces a lower output voltage. This provides a compact and efficient design. It provides the required output level even if the peak input voltage is higher.

7.2.1 Basic SEPIC Equations

The boost or step-up topology, as shown in Figure 7.13, is the basis for the SEPIC converter. The boost-converter principle is well understood: first, switch $Q1$ conducts during the on-period, TON, which increases the current in $L1$ and thus increases the magnetic energy stored there. Second, the switch stops conducting during the off-period, TOFF, but the current through $L1$ cannot change abruptly – it continues to flow, but now through diode $D1$ and into C_{out}. The current through $L1$ decreases slowly as the stored magnetic energy decreases. Capacitor C_{out} filters the current pulse that was generated by $L1$ when $Q1$ turned off.

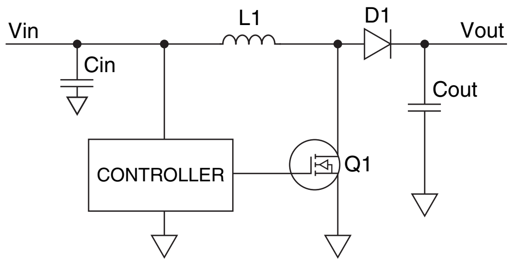


Figure 7.13: This Boost-Converter Topology is the Basis for SEPIC Power-Supply Circuits.

The diode $D1$ has to switch very quickly, so a diode with a short reverse recovery time (Trr less than 75 ns) is needed. In cases where V_{out} is relatively low, the efficiency can be improved by using a Schottky diode with low forward voltage (about 400 mV) for *D*1.

Note that a boost converter has one major limitation: V_{out} must always be higher than V_{in}. If V_{in} is ever allowed to become greater than V_{out}, *D*1 will be forward biased and nothing can prevent current flow from V_{in} to V_{out}.

The SEPIC scheme of Figure 7.14 removes this limitation by inserting a capacitor (*Cp*) between *L*1 and *D*1. This capacitor blocks any DC component between the input and output. The anode of *D*1, however, must connect to a known potential. This is accomplished by connecting *D*1 to ground through a second inductor (*L*2). *L*2 can be separate from *L*1 or wound on the same core, depending on the needs of the application.

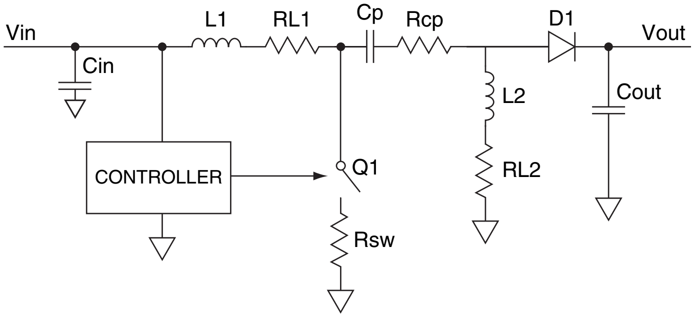


Figure 7.14: SEPIC Topology.

If *L*1 and *L*2 are wound on the same core, which is simply a transformer, one might argue that a classical fly-back topology is more appropriate. However, the transformer leakage inductance, which is no problem in SEPIC schemes, often requires a snubber network in fly-back schemes. Snubber networks are described later in this chapter; put simply they require additional components that must be carefully selected to minimize losses.

Parasitic resistances that cause most of the conduction losses in a SEPIC are *RL*1, *RL*2, *RSW* and *RCP*, and are associated with *L*1, *L*2, *SW*, and *CP* respectively. These parasitic components are also shown in Figure 5.27.

An advantage of the SEPIC circuit, besides buck and boost capability, is a capacitor (*Cp*) that prevents unwanted current flow from V_{IN} to V_{OUT}. Thus the limitation of the simple boost converter, that V_{IN} had to always be less than V_{OUT}, has been overcome.

Though it has very few elements, the operation of a SEPIC converter is not so simple to describe by equations; some assumptions have to be made. First, assume that the values of current and voltage ripple are small with respect to the DC components. Second, assume that at equilibrium there is no DC voltage across the two inductances $L1$ and $L2$ (neglecting the voltage drop across their parasitic resistances). By using these assumptions, Cp sees a DC potential of V_{in} at one side (through $L1$) and ground on the other side (through $L2$). The DC voltage across Cp is:

$$V_{CP}(\text{mean}) = V_{IN}$$

The period of one switching cycle is $T = 1/\text{frequency}$. The portion of T for which switch $Q1$ is closed is the duty cycle, D, and the remaining part of the period is thus $1 - D$. Because the mean voltage across $L1$ equals zero during steady state conditions, the voltage seen by $L1$ during $D * T$ (i.e. the MOSFET 'ON' period) is exactly compensated by the voltage seen by $L1$ during $(1 - D) * T$ (i.e. the MOSFET 'OFF' period):

$$D \cdot T \cdot V_{IN} = (1 - D) \cdot T \cdot (V_{OUT} + V_D + V_{CP} - V_{IN})$$

Where V_D is the forward voltage drop of $D1$ for a direct current of $(IL1 + IL2)$, and V_{CP} is equal to V_{IN}. Simplifying this we get:

$$D \cdot T \cdot V_{IN} = (1 - D) \cdot T \cdot (V_{OUT} + V_D)$$

Transposing this, we get:

$$\frac{(V_{OUT} + V_D)}{V_{IN}} = \frac{D}{1 - D} = Ai$$

Ai is called the amplification factor, where 'i' represents the ideal case for which parasitic resistances are null. Neglecting V_D with respect to V_{OUT} (as a first approximation), we see that the ratio of V_{OUT} to V_{IN} can be greater than or less than 1, depending on the value of D (with equality obtained for $D = 0.5$).

The more accurate expression Aa (amplification, actual) accounts for parasitic resistances in the circuit:

$$Aa = \frac{V_{\mathrm{OUT}} + V_{\mathrm{D}} + I_{\mathrm{OUT}} \cdot (Ai \cdot Rcp + RL2)}{V_{\mathrm{IN}} - Ai \cdot I_{\mathrm{OUT}} \cdot (RL1 + Rsw) - Rsw \cdot I_{\mathrm{OUT}}}$$

This formula allows computation of the minimum, typical and maximum amplification factors for V_{in} (Aa_{min}, Aa_{typ}, and Aa_{max}). The formula is recursive ('Aa_{xxx}' appears in both the result and the expression), but a few iterative calculations lead to the solution. The expression neglects switching losses due to the switch $Q1$ and reverse recovery current in $D1$. Those losses are usually negligible, especially if $Q1$ is a fast MOSFET and its drain-voltage swing ($V_{\mathrm{in}} + V_{\mathrm{out}} + V_{\mathrm{d}}$) remains under 30 V.

In some cases, you should also account for losses due to the reverse recovery current of D1, and for core losses due to high-level swings in stored magnetic energy. You can extrapolate the corresponding values of D:

$$D = Aa/(1 + Aa)$$

Or more generally:

$$D_{\mathrm{xxx}} = Aa_{\mathrm{xxx}}/(1 + Aa_{\mathrm{xxx}}), \text{ where xxx is min, typ or max.}$$

The DC current through Cp is zero, so the mean output current can only be supplied by $L2$:

$$I_{\mathrm{OUT}} = IL2$$

The power-dissipation requirement for $L2$ is eased, because the mean current into $L2$ always equals I_{OUT} and does not depend on variations of V_{IN}.

To calculate the current into $L1$ (I_{L1}), we can use the fact that no DC current can flow through Cp. Thus, the coulomb charge flowing during $D * T$ is perfectly balanced by an opposite coulomb charge during $(1 - D) * T$. When the switch is closed (for an interval $D \cdot T$) the potential at the switch node is fixed at 0 V. Since the capacitor Cp was previously charged to voltage V_{in}, the anode of $D1$ will now have a potential of $-V_{\mathrm{IN}}$, which reverse-biases $D1$. Current through Cp is then $IL2$. When the switch is open during $(1 - D) * T$, current I_{L2} flows through $D1$ while I_{L1} flows through Cp:

$$D \cdot T \cdot I_{L2} = (1 - D) \cdot T \cdot I_{L1}$$

Knowing that $I_{L2} = I_{\mathrm{OUT}}$,

$$I_{L1} = Aa\_\mathrm{xxx} \cdot I_{\mathrm{OUT}}$$

Input power equals output power divided by efficiency, so I_{L1} depends strongly on V_{IN}. For a given output power, I_{L1} increases if V_{IN} decreases. Knowing that I_{L2} (and hence I_{OUT}) flows into Cp during $D{*}T$, we choose Cp so that its ripple delta Vcp is a very small fraction of Vcp (gamma = 1% to 5%). The worst case occurs when V_{in} is minimal.

$$Cp \rangle \frac{I_{\mathrm{OUT}} \cdot D_{\mathrm{min}} \cdot T}{gamma \cdot V_{\mathrm{IN\_MIN}}}$$

By using a high switching frequency, small multi-layer ceramic capacitors can be used for Cp. However, ensure that Cp is able to sustain the power dissipation (Pcp) due to its own internal equivalent series resistance (Rcp):

$$Pcp = Aa\_\mathrm{min} * Rcp * I_{\mathrm{OUT}}{}^2$$

The MOSFET switch drain-to-source resistance, in series with a current sense resistor for limiting the maximum current, is given by the term Rsw. This incurs the following loss:

$$Psw = Aa\_\mathrm{min} * (1 + Aa\_\mathrm{min}) * Rsw * I_{\mathrm{OUT}}{}^2$$

Losses P_{RL1} and P_{RL2}, due to the internal resistances of $L1$ and $L2$, are easily calculated:

$$P_{RL1} = Aa\_\mathrm{min}^2 * R_{L1} * I_{\mathrm{OUT}}{}^2$$
$$P_{RL2} = R_{L2} * I_{\mathrm{OUT}}{}^2$$

When calculating the loss due to $D1$, the average power loss is due to the output current and the forward voltage drop of $D1$:

$$P_{D1} = V_{\mathrm{D}} * I_{\mathrm{OUT}}$$

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

$$I_{L1\_SAT} \rangle\rangle \ I_{L1} + 0.5 \cdot \Delta I_{L1} = \frac{Aa\_\min \cdot I_{OUT} + 0.5 \cdot T \cdot D \min \cdot V_{IN\_MIN}}{L1}$$

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

$$L2\_\min = \frac{2 \cdot T \cdot D_{\max} \cdot V_{IN\_MAX}}{I_{OUT}}$$

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

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 of wire (Litz) for counteracting the skin effect. Finally, the core size is chosen to accommodate a saturation current much greater than ($I_{L1}+I_{L2}+\Delta I_{L1}$) at the highest core temperature anticipated.

The purpose of the output capacitor (C_{OUT}) is to average the current pulses supplied by $D1$ during T_{OFF}. The capacitor must be able to handle high-level repetitive surge currents with low ESR and low self-inductance. Fortunately, ceramic and plastic film capacitors meet these requirements. The minimum value for C_{OUT} is determined by the amount of ripple (ΔV_{OUT}) that can be tolerated:

$$C_{\mathrm{OUT}} \geq \frac{Aa\_\min \cdot I_{\mathrm{OUT}} \cdot D\min \cdot T}{\Delta V_{\mathrm{OUT}}}$$

The actual value of the output capacitor may need to be much larger than that calculated using the above equation, especially if the load current is composed of high energy pulses. The input capacitor can be very small, thanks to the filtering properties of the SEPIC topology. Usually, C_{IN} can be one tenth the value of C_{OUT}:

$$C_{\mathrm{IN}} = C_{\mathrm{OUT}}/10$$

Overall efficiency η can be predicted from V_{IN} and Aa. The result can be misleading, because it doesn't account for the switch-transition losses or core losses and the real efficiency could be much lower:

$$\eta = V_{\mathrm{OUT}}/AaV_{\mathrm{IN}}$$

Finally, the switch SW and diode $D1$ should be rated for breakdown voltages with a 15% margin:

$$V_{\mathrm{DS}}(\text{switch}) > 1.15(V_{\mathrm{OUT}} + V_{\mathrm{D}} + V_{\mathrm{IN}})$$

$$V_{\mathrm{R}}(\text{diode}) > 1.15(V_{\mathrm{OUT}} + V_{\mathrm{IN}})$$

Example

Let $V_{\mathrm{IN}} = 50 - 150\,\mathrm{V}$ and $V_{\mathrm{OUT}} = 15\,\mathrm{V}$ at 1 A maximum. Let us operate at 200 kHz switching frequency, so that $T = 5\,\mu\mathrm{s}$. Now $\frac{V_{\mathrm{OUT}}}{V_{\mathrm{IN}}} = \frac{D}{1-D}$, so $D_{\max} = 0.231$ and $D_{\min} = 0.091$.

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

$$L1_{\min} = 10^{-5} * 0.769 * 150/1 = 1.15\,\text{mH, let } L1 = 1.5\,\text{mH}$$

$$L2_{\min} = 2TD_{\max} V_{\text{IN\_MAX}}/I_{\text{OUT}}$$

$$L2_{\min} = 10^{-5} * 0.231 * 150/1 = 0.347\,\text{mH, let } L2 = 0.47\,\text{mH}$$

$$Cp > I_{\text{OUT}} \cdot D_{\min} T/(\text{gamma} \cdot V_{\text{IN\_MIN}})$$

$$Cp > 1 * 0.091 * 2 * 10^{-5}/(0.05 * 50) = 728\,\text{nF, let } Cp = 1\,\mu\text{F}.$$

Now $D_{xxx} = Aa_{xxx}/(1 + Aa_{xxx})$, where xxx is min, typ or max. So $Aa_{\min}$ occurs at $D_{\min} = 0.091$ and $Aa_{\min} = 0.1$.

$$C_{\text{OUT}} \geq Aa\_\min \cdot I_{\text{OUT}} \cdot D_{\min} \cdot\ T / \Delta V_{\text{OUT}}$$
$$C_{\text{OUT}} > 0.1 * 1 * 0.091 * 2 * 10^{-5}/0.1.$$
$$C_{\text{OUT}} \gg 1.82\,\mu\text{F. Let } C_{\text{OUT}} = 100\,\mu\text{F}.$$
$$C_{\text{IN}} > C_{\text{OUT}}/10.\ \text{Let } C_{\text{IN}} = 10\,\mu\text{F}.$$

So, the fundamental component values have been calculated. Now what remains for the designer is the choice of suitable (and available) parts.

7.3 Buck-Boost Topology

Unlike the boost-buck circuits used by the Cuk and SEPIC topologies, the buck-boost uses a single inductor. It is a fly-back circuit and hence will be covered in Chapter 9.

7.4 Common Mistakes in Boost-Buck Circuits

Boost-buck circuits operate with both inductors in continuous conduction mode. Hence the inductor should be chosen with a value higher than that calculated, to allow for tolerances and for saturation effects (the inductance falls with increasing current). Calculate the value, add 20%, and then pick the next highest standard value.

Current ratings of inductors are given for a certain temperature rise in the core, typically 40°C. So if temperature rise is an issue, pick a component with a higher current rating.

7.5 Conclusions

The boost-buck is an ideal topology where the LED load voltage can be higher or lower than the supply voltage. It should also be used when the supply voltage is no more than 20% difference (worst case) from the LED load voltage. So if the LED voltage (maximum) is 20 V and the supply voltage (minimum) is 23 V, the difference is 3 V, and 3/20 = 0.15 or 15%, so a Cuk or SEPIC should be used. If the supply voltage is more than 20% higher, use a buck topology. If the supply voltage is more than 20% lower, use a boost topology. The boost-buck is less efficient compared to buck or boost topologies.

CHAPTER 8

LED Drivers with Power Factor Correction

8.1 Power Factor Correction

Power factor correction, or PFC, is a term used with AC mains powered circuits. A good power factor is when the AC current is in phase with the AC voltage. A pure resistive load has a power factor of 1, but active loads tend to have power factors closer to 0.5, unless special measures are taken to 'correct' this.

The most common power factor correction circuit is a boost converter. The AC line voltage is boosted to about 400 V and the amplitude of the current pulses into a storage capacitor is arranged to be sinusoidal. This is achieved by switching the current on for short but constant periods: as the supply voltage rises and falls, so does the amplitude of the current. A typical PFC circuit is shown in Figure 8.1.

A simple alternative is to use a fly-back supply. It is common to switch the primary current off when a certain current level is reached, but this leads to constant average current. To give a good power factor, the primary current should be switched with a constant 'on-time', so that the current amplitude rises and falls in phase with the supply voltage. The secondary current will rise and fall at double the AC line frequency and so a large secondary capacitor is required to absorb this ripple, to prevent significant ripple in the output voltage.

Driving an LED from a power factor corrected supply usually requires a simple buck converter, since the voltage source tends to be very high (about 400 V). However, alternative solutions exist; these are the Bi-Bred and the Buck-Boost-Buck (BBB).

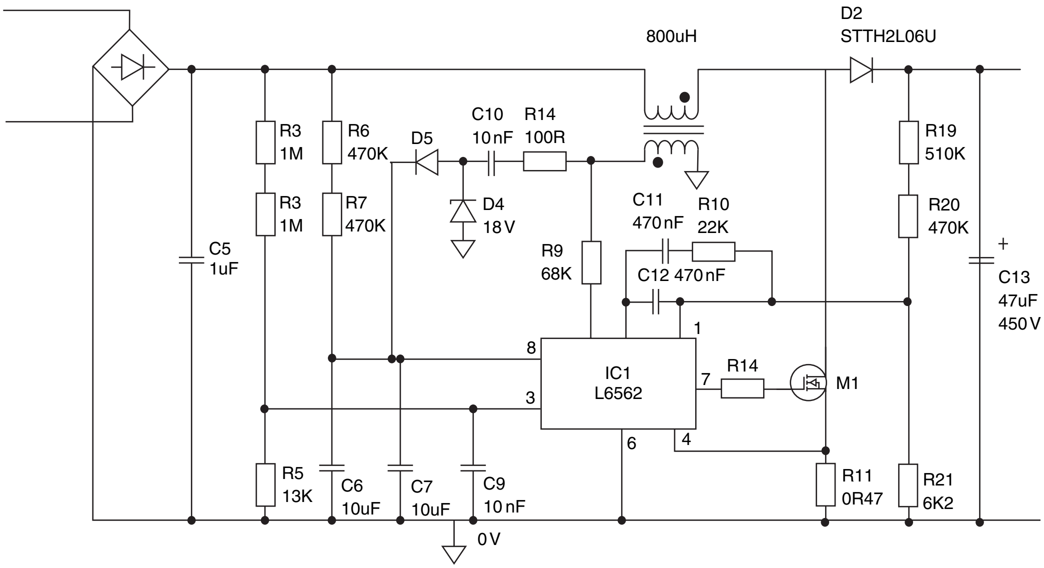


Figure 8.1: PFC Circuit.

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

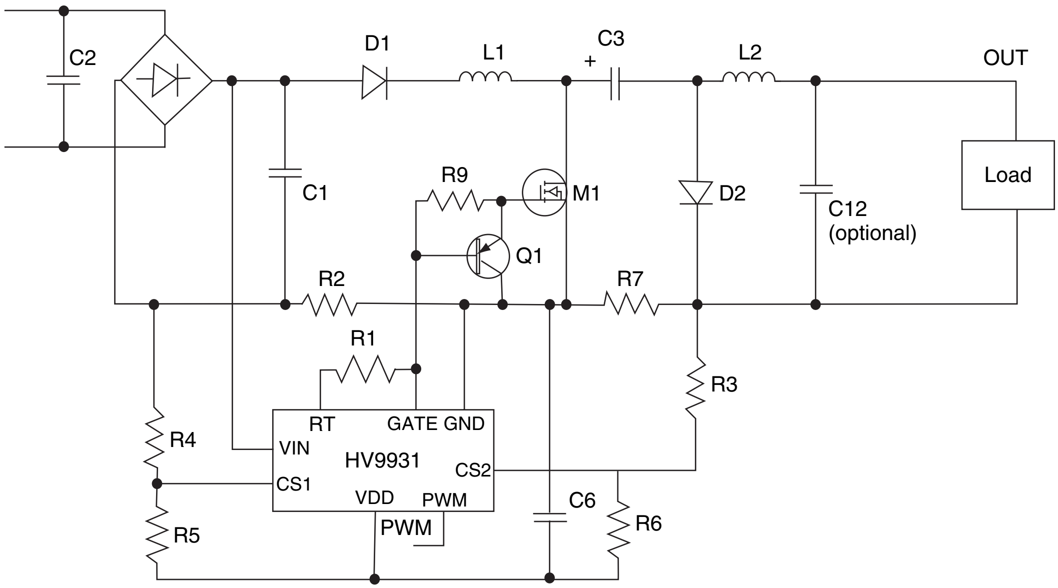


Figure 8.2: Bi-Bred Circuit.

amplitude through the inductor *L*1 until the voltage drop across *R*2 is sufficient for the internal comparator inside the HV9931 to trigger, which turns *M*1 off. Now the input circuit acts like a boost converter because the current through *L*1 cannot change immediately and it charges *C*3 to a high voltage.

The energy in *C*3 is used to drive current through the LED load the next time that *M*1 switches on. The current rises in inductor *L*2 and the load until the voltage drop across resistor *R*7 is sufficient to trip a second internal comparator and turn *M*1 off again. The current flow through *L*2 passes through *D*2 to keep current flowing in the LED load. Notice that the current sense resistor is not in this path, because the current level measurement is not required until the MOSFET turns on again; this minimizes power loss.

The output of the Bi-Bred is configured as a buck stage. Energy is supplied from a bulk storage capacitor, *C*3, with sufficiently large capacitance to provide a more or less constant supply voltage over an AC line cycle period. A constant capacitor voltage supplying power to the buck stage means a constant switch duty cycle when it is driving the LED load. The Bi-Bred draws a more or less sinusoidal AC line input current when driven from a switch operating at constant duty cycle, hence a large capacitance value for *C*3 helps to produce a good power factor.

The duty cycle of the switching is given by $\frac{V_O}{V_I} = \frac{D}{1-D}$.

Or put another way, $D = \frac{V_O}{V_I + V_O}$. So if $V_{in} = 350\,V$ and $V_o = 3.5\,V$ (a typical white LED), $D = \frac{3.5}{350+3.5} = \frac{3.5}{353.5} = 0.99\%$. This is close to the 1% expected duty cycle for a simple buck converter and can be difficult to switch properly. This means that a Bi-Bred is not really suitable for driving short LED strings.

8.3 Buck-Boost-Buck (BBB)

The Buck-Boost-Buck (BBB) is a proprietary circuit, patented by Supertex, and is illustrated in Figure 8.3. It resembles the Bi-Bred in some respects, except for two current steering diodes *D*1 and *D*2.

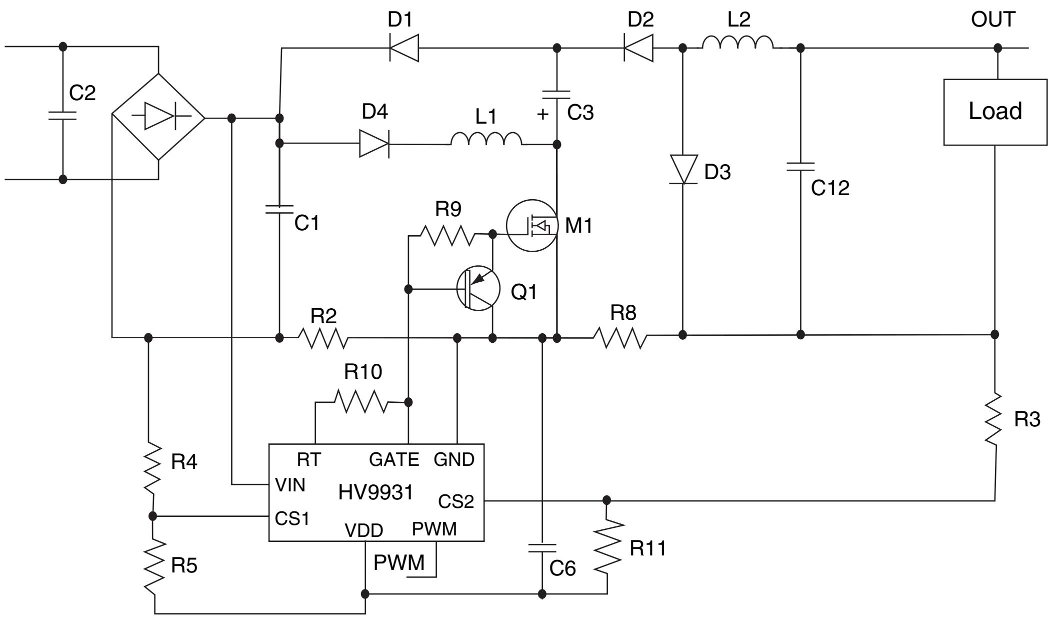


Figure 8.3: Buck-Boost-Buck Circuit.

Like the 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 $C3$ is not charged. With $M1$ conducting, current begins to rise in amplitude through the inductor $L1$ until the voltage drop across $R2$ is sufficient for the internal comparator inside the HV9931 to trigger, which turns $M1$ off. Now the input circuit is in flywheel mode, because the current through $L1$ cannot change immediately and it charges $C3$ to a moderately high voltage. The voltage is typically midway between the input and output voltage levels.

The energy in $C3$ is used to drive current through $D2$, $L2$ and the LED load the next time that $M1$ switches on. The current rises in inductor $L2$ and the load until the voltage drop across resistor $R8$ is sufficient to trip a second internal comparator and turn $M1$ off again. The current flow through $L2$ passes through $D2$ to keep current flowing in the LED load. Notice that, like in the Bi-Bred, the current sense resistor is not in this path, because the current level measurement is not required until the MOSFET turns on again; this minimizes power loss.

The output of the Buck-Boost-Buck (BBB) is the buck stage. Energy is supplied from a bulk storage capacitor, $C3$, with sufficiently large capacitance to provide a more or less constant supply voltage over an AC line cycle period. A constant capacitor voltage supplying power to the buck stage means a constant switch duty cycle when it is driving the LED load. The BBB draws a more or less sinusoidal AC line input current when driven from a switch operating at constant duty cycle, hence a large capacitance value for $C3$ helps to produce a good power factor.

In practice there is a limit to the value of $C3$, particularly if plastic film capacitors are used. This means that there will be some voltage ripple across $C3$, at a frequency double that of the AC line (i.e. 120 Hz when driven from a 60 Hz line). The effect of this ripple voltage is to generate second harmonic signals in the input current, which reduces the power factor. By adding a simple circuit, the second harmonic can be reduced; the MOSFET off-time is modulated by the ripple voltage and this acts like negative feedback to reduce the second harmonic. The additional circuits are given in Figure 8.4.

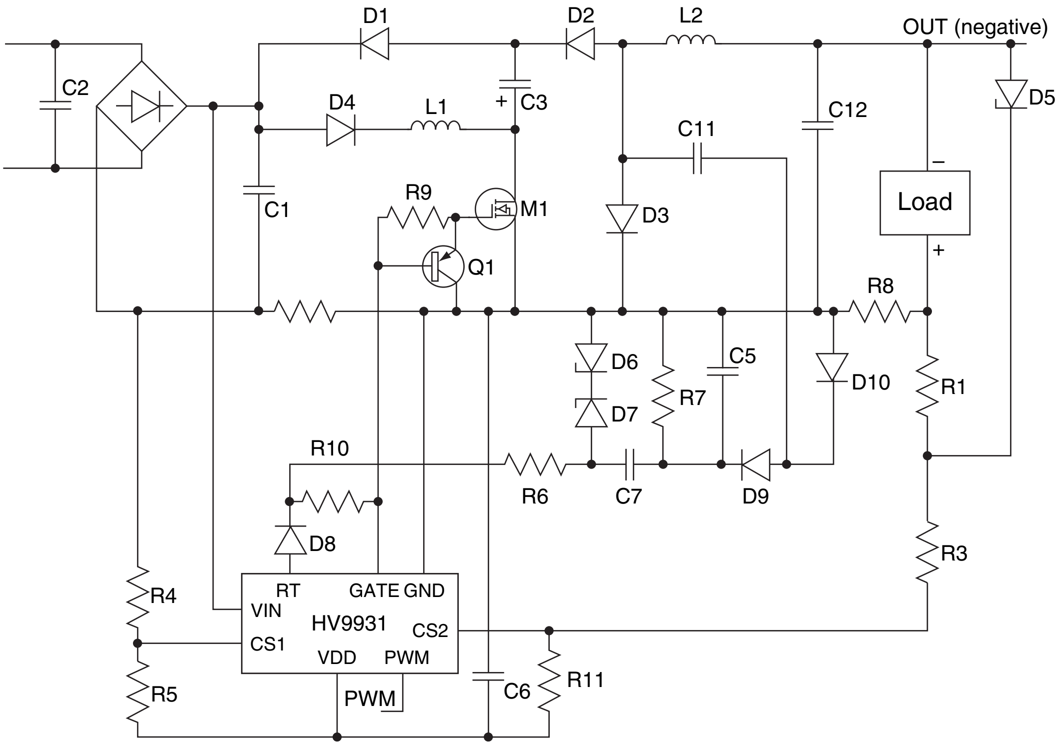


Figure 8.4: Buck-Boost-Buck with Harmonic Reduction.

When MOSFET $M1$ is conducting, the voltage across $C3$ is applied to $C11$, which charges $C5$. Between each switching cycle, resistor $R7$ discharges capacitor $C5$. The ripple voltage across $C3$ will modulate the average voltage across $C5$. Capacitor $C7$ acts as a DC block, to allow just the modulation across $C5$, rather than any DC level, to vary the MOSFET off-time. As the voltage across $C5$ rises and falls, current through $R6$ rises and falls, thus shortening or lengthening the off-time.

The duty cycle of the switching in a BBB converter is given by $\frac{V_O}{V_I} = \frac{D^2}{1-D}$.

Or put another way, $D = \frac{-V_O \pm \sqrt{V_O^2 + 4 \cdot V_I \cdot V_O}}{2 \cdot V_I}$. So if $V_{in} = 350\,\text{V}$ and $V_o = 3.5\,\text{V}$ (a typical white LED), $D = \frac{-3.5 \pm 70}{700} = \frac{66.5}{700} = 9.5\%$. This is a considerably greater duty cycle than the Bi-Bred or the buck converter with a similar low voltage load. This means that the BBB converter is most suitable for driving short LED strings.

8.4 Common Mistakes with PFC Circuits

The most common mistake is to use a standard inductor for $L1$. Inductors are sized for their magnetic saturation level and for resistive heating. Thus an inductor may be specified as $I(\text{av}) = 500\,\text{mA}$, $I(\text{sat}) = 400\,\text{mA}$. This inductor can pass 500 mA with a temperature rise of, say, 40°C. It can pass 400 mA before the inductance is reduced by, say, 10%. If this inductor were used in a PFC stage with a peak current of 400 mA it would overheat. Using an inductor with a much higher saturation current rating will be necessary, to give a reasonably low temperature rise during operation.

Inductor manufacturers do not normally specify magnetizing losses. The magnetic saturation levels are material dependent; the maximum flux density of ferrite is usually 200 mT, other materials can be higher. So when designing a ferrite-based inductor, a manufacturer will make his design based on this level. When considering magnetizing losses, a flux density of about 50 mT would be a better choice for a ferrite-based inductor.

8.5 Conclusions

Detailed design analysis has not been given for the PFC circuits. This chapter has been intended to show readers some options and point out limitations. For example, driving a single LED would require a Buck-Boost-Buck circuit, but longer strings can be driven from a Bi-Bred or a PFC stage followed by a buck converter.

Application notes from ST Microelectronics and Supertex cover the PFC, Bi-Bred and Buck-Boost-Buck circuits in detail. These are proprietary and specialized solutions that are still evolving; interested readers should consult these application notes for the latest designs.

CHAPTER 9

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.

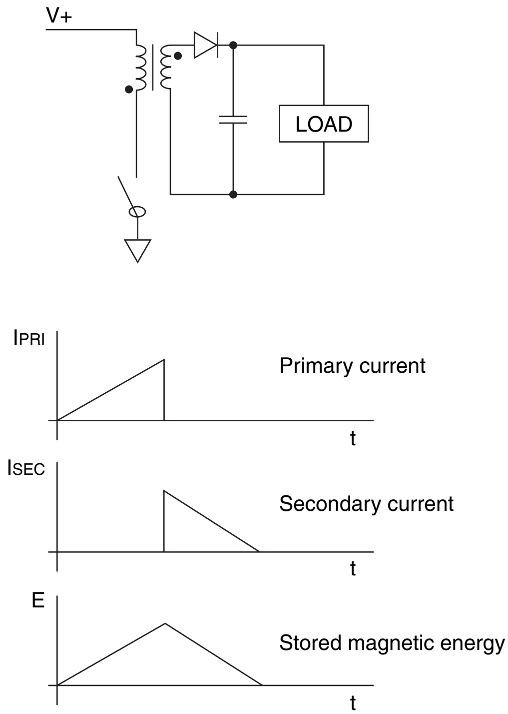


Figure 9.1: Fly-Back Principle.

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.

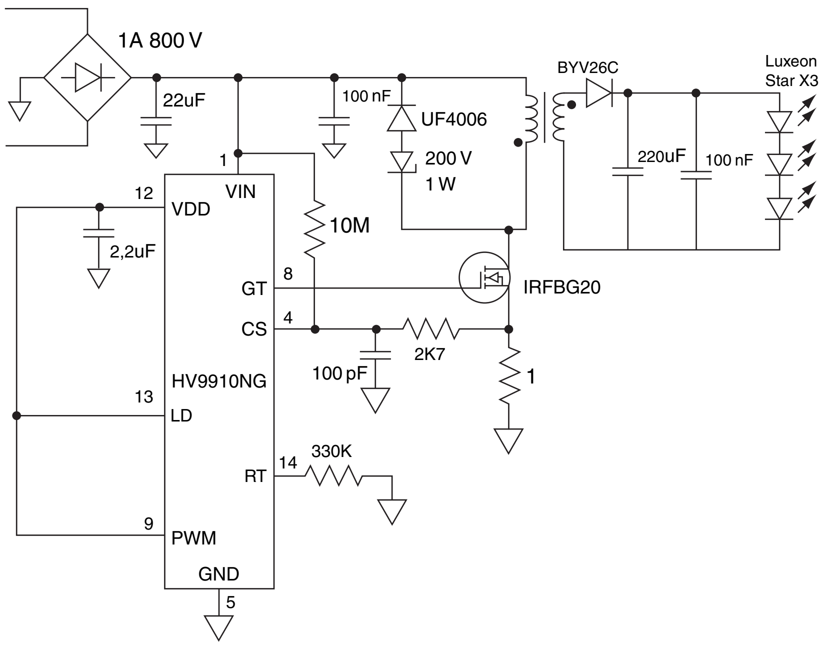


Figure 9.2: Fly-Back Circuit for LEDs.

The secondary winding start point is connected to the output diode, which prevents conduction when the MOSFET is on. The start point of the secondary is connected to the output diode, but the end point is connected to ground and this tends to screen the secondary winding for minimal EMI radiation. Energy that is stored during the MOSFET on-time is released during the off-time, by current flowing through the output diode and into the load.

Calculation of the transformer characteristics, like inductance value and primary to secondary turns ratio, are very important in the design. In order for complete power transfer from the primary to secondary, the volt-seconds must be equal. The equation is:

$$\frac{V_{PRI} \cdot T_{ON}}{N_{PRI}} = \frac{V_{SEC} \cdot T_{OFF}}{N_{SEC}}$$

9.1.1 Fly-Back Example

Let us make an isolated 3 W lamp by connecting three white power LEDs in series.

Suppose we have a primary voltage of 48 V, an on-time of 5 microseconds, and the primary to secondary turns ratio is 1:0.1. If we are driving a 10 V LED load, the off-time will be 240/100 microseconds (2.4 μs). Thus the switching period must be greater than 12.4 μs in order to allow complete removal of the magnetic energy in the transformer core. A switching frequency of below 65 kHz will be satisfactory, say 60 kHz to give some margin.

With 60 kHz switching, the period will be 16.667 μs. If the average output current is 350 mA, the average in 2.4 μs will be 2.43 A. Since this current decays linearly from the transformer winding, the peak secondary current will be double this: 4.86 A. The secondary inductance will be $E = -L \cdot \frac{di}{dt}$.

$$L = E \cdot \frac{dt}{di} = 10 \cdot \frac{2.4 \cdot 10^{-6}}{4.86} = 4.94\,\mu H$$

Since the primary has ten times the turns of the secondary, the primary inductance will be 100 times that of the secondary (the turns ratio, N, is squared). In other words, the primary inductance will be 494 μH.

Most current-mode power supplies control the switching so that the MOSFET turns off when a certain peak current is reached in the primary winding. Since the peak current in the secondary is 4.86 A and the turns ratio is 10:1, we need a peak primary current of 486 mA. [Check: $E = -L \cdot \frac{di}{dt}$, so $E = 494 * 10^{-6} * 0.486/(5 * 10^{-6}) = 48$ V].

The problem with the design that we have is that the LED current will change if the LED voltage changes, because we have based our design on a certain output voltage. Actually this gives a constant power output, assuming a constant voltage input, which is fine for non-critical designs. But what if the input voltage changes?

A higher input voltage will mean that the current limit will be reached in a shorter time. This means that the duty cycle will be reduced and hence the number of volt-seconds on the primary will be unchanged. In practice, inherent delays in the current sense comparator will cause the input current to 'overshoot' the reference level. This overshoot increases with increasing input voltage, this is because the delay is constant but the rate of current rise is increasing with input voltage. Compensation of this overshoot can be achieved by connecting a resistor between the supply voltage rail and the current sense pin. This resistor injects a small DC bias that increases with increasing supply voltage and thus triggers the comparator earlier as the supply voltage rises.

The 1:0.1 turns ratio and 10 V output used in the above example cause a reflected voltage of 100 V in the primary winding when the secondary conduction takes place. This reflected voltage adds to the supply voltage, so a MOSFET with a 200 V or higher voltage rating is required when powering this circuit from a 48 V supply.

The design example does not allow for efficiency. In practice a fly-back converter has about 90% efficiency, so the input current must be increased by about 10% to allow for this.

If we were designing a constant voltage circuit, we would allow the peak primary current to be higher than that given in the example. This margin allows for the input voltage variations. We would then use feedback from the output to control the switching, to reduce the power in the primary, as necessary.

9.2 Three Winding Fly-Back

Some fly-back power supplies use a third winding, called a bootstrap or auxiliary winding, as shown if Figure 9.3. This is used to power the control IC, once the circuit is operating. The bootstrap winding has the same orientation as the secondary winding and the voltage is simply determined by the turns ratio of the bootstrap compared to the secondary. In our example of a 10 V output from the secondary, the bootstrap could have the same number of turns and thus give (approximately) 10 V for the powering the control IC.

At start-up, there is no power available from the bootstrap winding, so a start-up regulator is required. Example start-up regulators are the LR645 and the LR8 from Supertex; these give a low voltage, low current output from an input with a voltage as high as 450 V. Once the bootstrap produces power, the start-up regulator turns off. The HV9120 shown in Figure 9.3 has a start-up regulator built-in.

9.2.1 Design Rules for a Fly-Back Converter

This section gives design rules for a fly-back converter based on either turns ratio selection determined by the maximum duty cycle allowed (case 1), or by the optimum turns ratio based on the maximum working voltage of the MOSFET switch (case 2). In case 1, a design based on the maximum duty cycle (at the lowest input voltage)

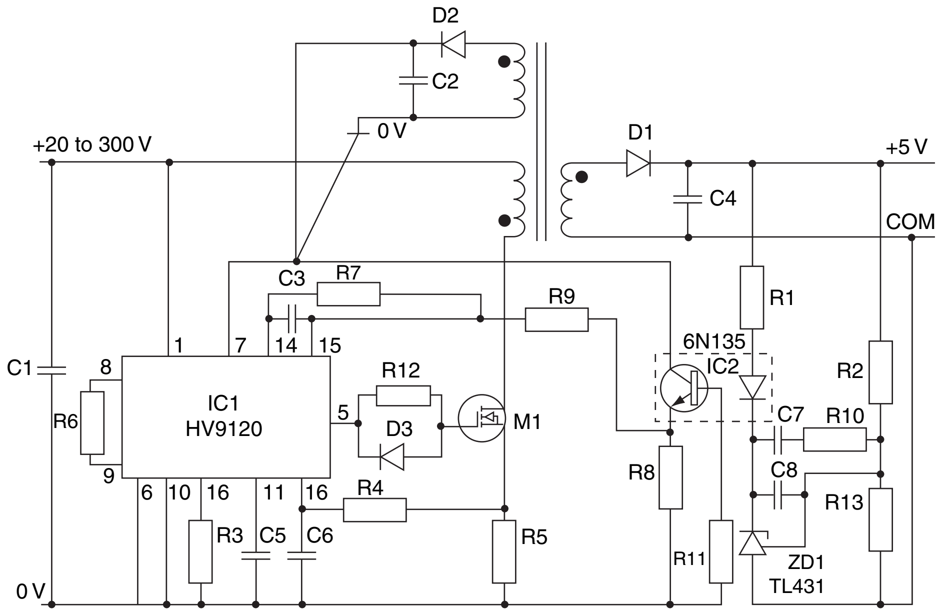


Figure 9.3: Fly-Back Using a Three-Winding Transformer.

allows the widest input voltage range. In case 2, a design based on the maximum voltage across the MOSFET allows a potentially lower cost solution. Alternatively, a fly-back design based on an already available transformer with a known (and fixed) turns ratio may be considered.

The transfer function of a fly-back converter is:

$$\frac{V_O}{V_I} = \frac{D}{(1-D)} \cdot N$$

So the duty cycle can be found by transposing the equation:

$$V_O \cdot (1-D) = V_I \cdot D \cdot N$$

$$V_O = V_I \cdot D \cdot N + V_O \cdot D = D \cdot (V_I \cdot N + V_O)$$

$$D = \frac{V_O}{(V_I \cdot N) + V_O}$$

Case 1: Turns Ratio Based on Maximum Duty Cycle

Given the minimum input voltage V_{I_MIN}, output voltage V_O and maximum duty cycle D_{MAX}, the turns ratio N can be calculated:

$$N = \frac{V_O \cdot (1 - D_{MAX})}{V_{I\_MIN} \cdot D_{MAX}}$$

D_{MAX} is typically chosen as 45% (0.45) for a PWM controller with a maximum 49% duty cycle. With $D_{MAX} < 50\%$, the system is inherently stable and there is no complex compensation required.

If we take the earlier example of 48 V input (say, 46 V minimum), 10 V output (add 0.6 V for the output diode) and allow 45% duty cycle, we get:

$$N = \frac{10.6 \cdot (0.55)}{46 \cdot 0.45} = 0.282$$

This is the minimum value. A transformer with a convenient turns ratio of 1:0.33 (3:1) could be used. The maximum duty cycle would then be:

$$D = \frac{V_O}{(V_I \cdot N) + V_O} \frac{10.6}{(15.33 + 10.6)} = 0.41 \quad (41\%)$$

Case 2: Turns Ratio Based on Maximum Switch Voltage

The output voltage across the secondary winding is induced into the primary and magnified by the turns ratio N. This was illustrated at the beginning of this chapter, when a 10 V output caused 100 V to be induced into the primary winding of a 1:0.1 turns ratio transformer. Considering that the supply voltage was only 48 V, this forced us to use a 200 V MOSFET as the primary switch. The aim here is to minimise the MOSFET switch operating voltage requirement.

Because the voltage reflected into the primary often has some ringing, a snubber circuit is used to limit the voltage across the primary winding. Ringing is due to resonance between the MOSFET drain capacitance, parasitic capacitance in the circuit and parasitic inductance of the transformer primary. Parasitic inductance in the transformer is often referred to as ‘leakage inductance’ because it is the proportion of the primary inductance that is not coupled into the secondary, so the magnetic field ‘leaks out’.

A Zener diode is sometimes used as a snubber. The voltage across the Zener diode will be greater than the voltage induced into the primary from the secondary (output) voltage, otherwise power dissipation and losses will both be very high.

$$V_O = N \cdot (V_{SW} - V_Z - V_{IN\_MAX})$$

In order to find the secondary winding voltage, the forward voltage drop of the output diode, V_F, must be added to the output voltage.

$$N = \frac{V_O + V_F}{(V_{SW} - V_Z - V_{IN\_MAX})}$$

As a safety margin,$(V_{SW} - V_Z - V_{IN\_MAX}) \geq 10\text{V}$.

In the example we used earlier, with 48 V input, we could have used a 100 V switch and a 33 V Zener diode. The output is 10 V, so allowing for V_F this becomes 10.6 V across the secondary winding:

$$N = \frac{10 + 0.6}{(100 - 33 - 48)} = \frac{10.6}{19} = 0.558$$

We could use a transformer with 1:0.5 turns ratio ($N = 0.5$). The primary voltage induced from the secondary winding will be 21.2 V, which is below the Zener diode voltage by 11.8 V, which is a reasonable margin to minimise power dissipation. The peak voltage across the MOSFET drain will be limited to 48 V + 33 V = 81 V.

With a turns ratio of 1:0.5, the maximum duty cycle with a 46 V minimum input voltage will be:

$$D = \frac{V_O}{(V_I \cdot N) + V_O} = \frac{10.6}{23 + 10.6} = 0.315 \quad (31.5\%)$$

Inductance Calculations

Now we have the turns ratio (by either means described above) and the maximum duty cycle, we can now determine the inductance and switching current. Let us use case 1, with 41% as the maximum duty cycle.

$$P_{IN} = \frac{P_{OUT}}{\eta}$$

The output power is 10 V × 0.35 A = 3.5 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 µs. With a 41% duty cycle, the switch on-time will be 6.835 µs. So we need the primary current to rise by 439 mA in 6.835 µs.

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

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 µH.

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 (Ac), 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 mm^2. So $Ac = 32 * 10^{-6}$ m^2. The number of turns can be calculated, based on the design parameters above and using $B = 200$ mT as the maximum flux density:

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

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

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

Refer to core manufacturer's specifications and choose a core with a lower AL value (larger gap) than calculated using the above equation. A suitable core (3C90 material, 160 µm air gap) has an AL value of 250 nH. The number of turns can then be calculated as:

$$N = \sqrt{\frac{L}{A_L}}, \text{ when L is expressed in nH. Thus } 716\ \mu\text{H} = 716{,}000\ \text{nH}.$$

$N_{PRI} = 54$ (rounding up to the next highest value). This quite conveniently gives the secondary turns as $N_{SEC} = 18$, since it is 1/3.

9.3 Single Winding Fly-Back (Buck-Boost)

In the buck-boost converter, a single inductor winding is used for the primary and secondary. This is shown in Figure 9.4.

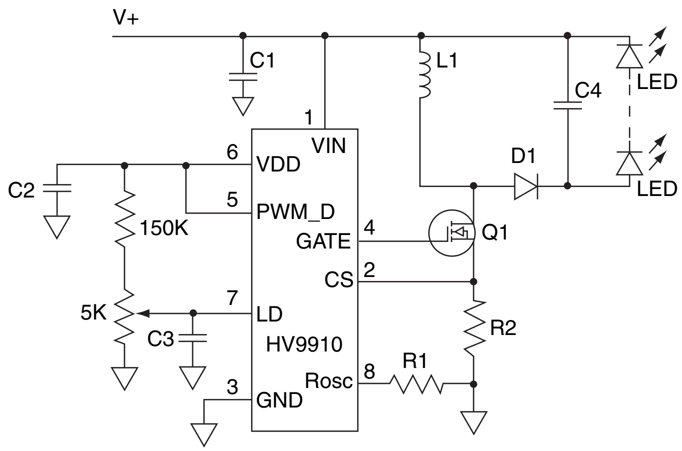


Figure 9.4: Buck-Boost Converter.

Current is forced through the inductor by a MOSFET connecting the inductor across the power supply rail. The current level rises almost linearly with time. At a predetermined current level, the MOSFET is turned off and the current is forced to flow through a diode to charge the output capacitor and drive the load. The current in the inductor falls back to zero and so discharges the energy stored in the magnetic core. Like the two-winding fly-back, the single winding fly-back can be calculated from the number of volt-seconds on the charge cycle equalling the number of volt-seconds on the discharge cycle.

The duty cycle of a buck-boost converter (continuous conduction mode) is given by the equation:

$$\frac{V_O}{V_I} = \frac{D}{1-D}$$

$$V_O \cdot (1-D) = V_I \cdot D$$

$$V_O = V_I \cdot D + V_O \cdot D = D \cdot (V_I + V_O)$$

$$D = \frac{V_O}{V_I + V_O}$$

So, if we have $V_{in} = 24\,\text{V}$ and $V_{out} = 30$ V, $D = 30/54 = 0.555$.

In practice we want discontinuous conduction mode, because continuous conduction mode is difficult to stabilise. This means that the inductor current falls to zero at the end of each cycle. So, assume we want 350 mA output and 100 kHz switching frequency. The period is 10 µs, so the on-time is 5.55 µs and the off-time is 4.45 µs. During the off-time, the current in the inductor falls linearly from a peak level to zero. To average 350 mA output, the average current during the off-time must be $350/0.445\,\text{mA} = 786.5\,\text{mA}$, so the peak current must be double this, or 1.573 A. This means that during the on-time, the current must rise from zero to 1.573 A.

The voltage from the power supply is 24 V, so using the familiar equation:

$$E = -L \cdot \frac{\mathrm{d}i}{\mathrm{d}t}$$

$$L = E \cdot \frac{\mathrm{d}t}{\mathrm{d}i} = 24 \cdot \frac{5.55 \cdot 10^{-6}}{1.573} = 84.67\,\mu\text{H}$$

In practice there should be some dead time allowed, when the inductor carries no current, to ensure discontinuous conduction mode. This dead time is to allow for power supply tolerances, inductor tolerances, etc. Too much dead time means that the peak current is higher and this reduces the efficiency of the power supply.

Suppose we allow 25% tolerance, so that the on-time is 4.44 μs; this will reduce the inductance by 25%.

$$L = E \cdot \frac{dt}{di} = 24 \cdot \frac{4.44 \cdot 10^{-6}}{1.573} = 68\,\mu\text{H}$$

The off-time will be reduced unless the peak current is increased in proportion.

$$E = -L \cdot \frac{di}{dt}$$

$$-30 = 68 \cdot 10^{-6} \cdot \frac{di}{4.45 \cdot 10^{-6}}$$

$$di = \frac{-30 \cdot 4.45 \cdot 10^{-6}}{68 \cdot 10^{-6}} = 1.963\,\text{A}$$

Increasing the peak current by 25% gives the desired result. The peak current is set by the value of current sense resistor between the MOSFET source and ground.

CHAPTER 10

Essentials of Switching Power Supplies

This chapter will examine the advantages and disadvantages of the various driver techniques, which have already been described. The issues of efficiency, EMI, cost and other requirements that are additional to the basic function of the LED driver.

10.1 Linear Regulators

In Chapter 4 we saw how the use of linear regulators caused a heat dissipation problem because of low efficiency. A linear LED driver is generally less efficient than a switching driver. Sometimes a linear driver can be more efficient. For example, if you have a 12 V power source and three LEDs each having a 3.5 V forward drop, by connecting them in series the total drop is 10.5 V. The efficiency of a linear driver, dropping only 1.5 V will be 87.5%. It would be difficult for a switching LED driver to achieve this level of efficiency. And there is no EMI to be filtered.

On the other hand, driving one LED from a 12 V supply would give an efficiency of $3.5/12 = 29\%$ with a linear LED driver. Here a buck switcher would give closer to 90% efficiency. See Figure 10.1. Efficiency is important where heat dissipation must be minimized. Otherwise cost usually takes precedence and the cost of a switching regulator with EMI filters would be somewhat higher.

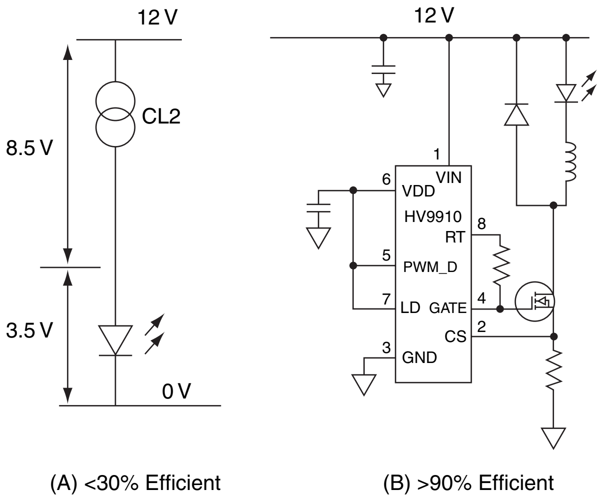


Figure 10.1: Linear vs Switching Solutions.

10.2 Switching Regulators

In Chapters 5 to 9 we looked at switching regulators, which have much higher efficiency, but can generate electro-magnetic interference (EMI) which has to be suppressed by careful circuit board design, screening and filtering. The EMI reducing techniques are described in Chapter 13.

Although Supertex's LED driver integrated circuits are used in examples, similar drivers from other manufacturers can also be used. For example, the Linear Technology LTC3783 has similar functions to the Supertex HV9912. The National Semiconductor LM5020 is a buck controller, like the HV9910B. However, Supertex devices have an internal high voltage regulator, which makes them more versatile.

Switching power supplies have the disadvantage of producing electromagnetic interference (EMI). EMI must be limited, to prevent interference with other systems. This is a legal requirement and product cannot be sold unless the equipment meets the standards laid down in law. Details of EMI techniques are given in Chapter 13.

Conversely, where EMI requirements are very demanding, such as medical and automotive applications, linear LED driver techniques can be used instead.

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 12 V DC automotive application, but where just one LED is being driven.

Buck regulators can be very efficient, maybe 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 ΔI_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.

Adding a filter capacitor across the LED string can reduce the output current ripple, thus allowing a lower inductor value or an apparently more 'constant' current. This capacitor reduces EMI at the output by providing a bypass path for any switching current spikes, which may also improve the LED lifetime. However, keep in mind that the peak-to-average current error is affected by the variation of the MOSFET off-time, T_{OFF}. Therefore, the initial output current accuracy might be sacrificed with large ripple current levels in the inductor.

Another important aspect of designing an LED driver is related to certain parasitic elements of the circuit, including distributed coil capacitance of the inductor C_L, junction capacitance C_J, and reverse recovery of the flywheel diode, capacitance of the printed circuit board traces C_{PCB} and output capacitance C_{DRAIN} of the MOSFET. These parasitic elements affect the efficiency of the switching converter because they cause switching losses. These parasitic elements are shown in Figure 10.2.

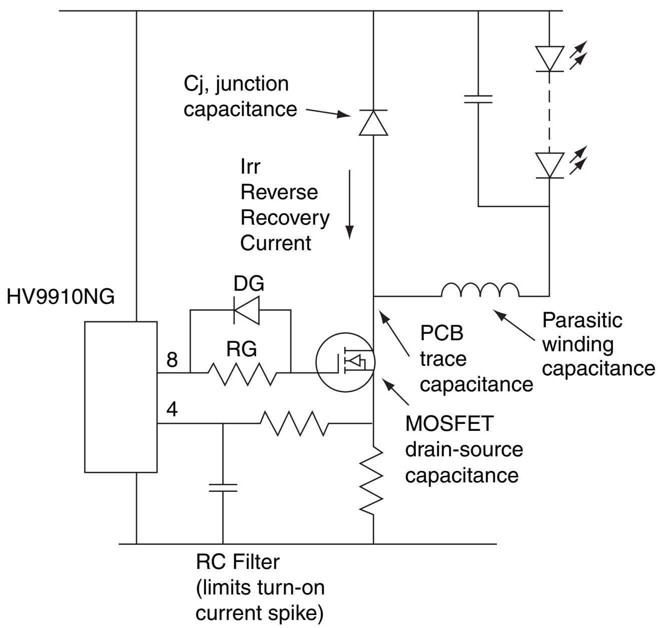


Figure 10.2: Parasitic Elements.

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 C_O ($\sim$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 hence the conduction losses will not be significant. Note that the saturation current in a MOSFET becomes lower at increased junction temperature.

The duration of the leading edge current spike can be estimated as:

$$T_{SPIKE} = \frac{V_{IN} \cdot C_P}{I_{SAT}} + t_{rr}$$

In order to avoid false triggering of the current sense comparator, C_P must be minimized in accordance with the following expression:

$$C_P < \frac{I_{SAT} \cdot \left(T_{BLANK(MIN)} - t_{rr}\right)}{V_{IN(MAX)}}$$

The factor $T_{BLANK(MIN)}$ is the minimum blanking time, which depends on the control IC and is in the order of 300 ns. When the MOSFET gate drive is activated, the control IC disables the current sense input for this time period, to avoid false triggering from the switch-on current surge, previously described. The factor $V_{IN(MAX)}$ is the maximum instantaneous input voltage.

Discharging the parasitic capacitance C_P into the DRAIN output of the MOSFET is responsible for the bulk of the switching power loss. It can be estimated using the following equation:

$$P_{SWITCH} = \left(\frac{C_P V_{IN}^2}{2} + V_{IN} I_{SAT} \cdot t_{rr}\right) \cdot F_S$$

where F_S is the switching frequency, I_{SAT} is the saturated DRAIN current of the MOSFET. The switching loss is the greatest at the maximum input voltage.

The switching frequency of a buck converter having constant off-time operation is given by the following:

$$F_S = \frac{V_{IN} - \eta^{-1} \cdot V_O}{V_{IN} \cdot T_{OFF}}$$

where η is the efficiency of the power converter. This value for F_S based on typical values for V_{IN} and V_O can be used in the previous equation if a value of constant switching frequency is not available.

The switching power loss associated with turn-off transitions of the DRAIN output can be disregarded. Due to the large amount of parasitic capacitance connected to this switching node, the turn-off transition occurs essentially at zero voltage.

Conduction power loss in the MOSFET can be calculated as

$$P_{COND} = D \cdot I_O^{\,2} \cdot R_{ON}$$

where $D = V_O/\eta V_{IN}$ is the duty ratio and R_{ON} is the ON resistance.

Buck Converter AC Input Stage

An off-line LED driver requires a bridge rectifier and input filter; selecting an input filter is critical to obtaining good EMI.

We may use an aluminum electrolytic capacitor after the bridge rectifier, in order to prevent interruptions of the LED current at zero crossings of the input voltage (the cusps in the rectified sine-wave, or haversine, waveform). As a 'rule of thumb', $2 \sim 3\,\mu F$ per each watt of the input power is required. An electrolytic capacitor is often used and has the added ability of being able to absorb voltage surges that may be present on the AC line.

Large values of input capacitor will cause unacceptably high current surges when power is first applied. These current surges can damage the electrolytic capacitor, reducing its life expectancy, and also damage the switch or electrical connectors at the AC line. Inrush current limiters, usually a negative temperature coefficient (NTC) thermistor rated for high current, are often connected in series with the AC line to prevent the current surge.

An inductor in series with the supply rail, after the input capacitor, is needed to present high impedance to switching frequency signals, as shown in Figure 10.3. The current rating of this inductor needs to be higher than the expected current level in normal operation. The value of the inductor depends on the level of signal attenuation required, when combined with the input capacitor shunt impedance, to meet the EMI standards in force.

The impedance of an inductor is given by: $X_L = 2 \cdot \pi \cdot F_S \cdot L$, so if we needed 200 ohms impedance at 100 kHz to give us our desired attenuation, $L = 0.318\,mH$. A $330\,\mu H$ filter inductor could be used.

A capacitor connected between the switching side of the filter inductor and ground, albeit of small value, is necessary in order to ensure low impedance to the high

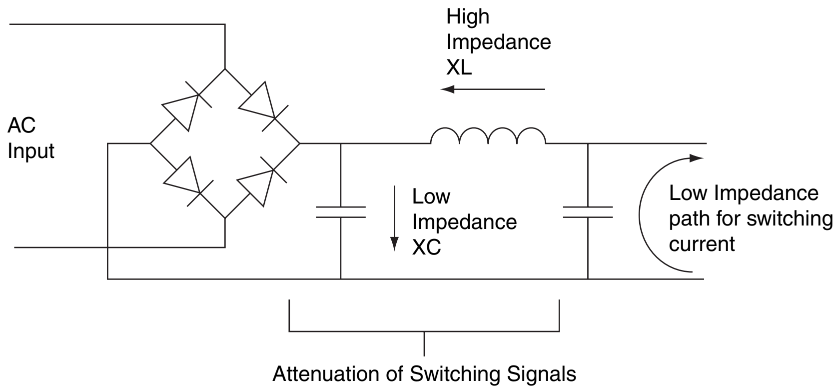


Figure 10.3: Input Filter Functions.

frequency switching currents of the converter. As a rule of thumb, this capacitor should be approximately 0.1–0.2 µF/W of LED output power. A 100 nF capacitor can be used in a circuit that drives a single 1 W LED.

10.2.2 Boost Regulator Considerations

The output voltage in a boost circuit must always be higher than the input voltage by about 20% or more, and this was discussed in Chapter 6. Ignoring PFC applications, a boost converter driving LEDs will always be powered from a low voltage DC supply.

For example, the backlight in a cell phone with a color LCD display usually employs low cost white light LEDs. A boost regulator is used in this application to drive a string of 20 mA LEDs from a 3–4 V battery.

As another example, in flat-screen television backlighting high power red, blue and green (RGB) LEDs are used to create a white light that exactly matches the LCD and produces true colors. In this application a boost converter powered from a 12 V or 24 V DC supply is used to drive many 350 mA LEDs connected in series, with a forward voltage in the range 40–80 V.

Boost regulators should always be provided with over-voltage protection, in case the LED load is disconnected. Otherwise the output voltage will continue to rise and eventually cause component breakdown. In Safety Electrical Low Voltage (SELV) systems, the output voltage would normally be kept below 42 V.

10.2.3 Boost-Buck Regulator Considerations

To operate in an environment where the input voltage could be higher or lower than the output voltage, a buck-boost (or boost-buck) circuit is necessary. Boost-buck circuits were described in Chapter 7. The situation of having a load voltage range that overlaps the supply voltage range is commonly found in automotive applications. The battery voltage rises and falls with a large variation, as the engine speed and battery conditions change.

The two types of converters often found in boost-buck applications are known as SEPIC and Cuk. These converters are similar, but the Cuk converter has an inverted output, which means that the LED anode is connected to the ground rail. Like boost converters, over-voltage protection should be provided to prevent excessively high voltage in case of an open-load condition.

Because there are inductors in series with the input and the output, and both operate in continuous conduction mode (CCM), high frequency signals at the central node where the switching takes place are automatically filtered. Shunt capacitors across the input and output strengthen this filtering, and provide a low impedance path for the circulating currents. Consequently, Cuk and SEPIC circuits require minimal external filtering. Sometimes common mode chokes are added at the input side, to reduce the radiated signals from the whole circuit. Common mode chokes are only required on the output side if the length of wire to the LED load is more than about 0.5 m long.

10.2.4 Circuits with Power Factor Correction

Power factor is an indication of the relative phase of the power line voltage and the power line current. A power factor of 1 indicates that the voltage and current are in-phase and have low harmonic content. A power factor of 0 indicates that the voltage and current are 90 degrees out-of-phase.

In semiconductor circuits powered from the AC mains, a bridge rectifier converts the AC power into DC. The current through the bridge rectifier tends to occur close to the peak voltage, as shown in Figure 10.4, because charging of a large smoothing capacitor takes place each half cycle. These short charging current pulses at the crest of each input cycle cause the power factor to be typically in the 0.3–0.6 range. Power factor correction is an active or passive circuit designed to correct phase errors and

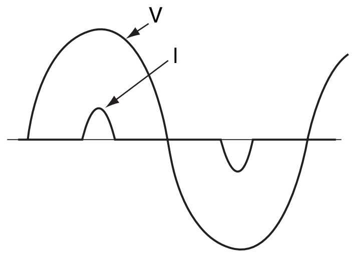


Figure 10.4: Active Circuit AC Input Current.

reduce harmonics, and make the power factor closer to 1. Power factor correction (PFC) is required in higher power LED drivers.

A circuit having a good power factor, approaching 1, has an input current that has low harmonic content with a wave shape that closely follows the sinusoidal input voltage. Circuits that provide a good power factor were described in Chapter 8.

10.2.5 Fly-Back Converter Considerations

Transformer coupled switching regulators can be designed for a very wide range of supply and output voltages. The most common is a fly-back converter, although forward converters are also popular in higher power applications. Fly-back converters were described in Chapter 9.

Fly-back converters allow an isolated LED driver design with about 90% efficiency, but have added cost and complexity. If a wide tolerance can be accepted for the current regulation, a simpler and cheaper circuit can be built. High accuracy requires isolated feedback, usually via an opto-coupler and employing an adjustable shunt regulator such as a TL431 or similar, along with a few passive components.

Fly-back converters have the advantage of stepping up or down the output voltage compared with the supply (buck-boost). This also applies to the single winding inductor version, although since the same winding is used for the primary and secondary side, the turns ratio is 1:1 and the design specification are more restricted than for a two-winding inductor. A single winding inductor is usually much lower cost.

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.

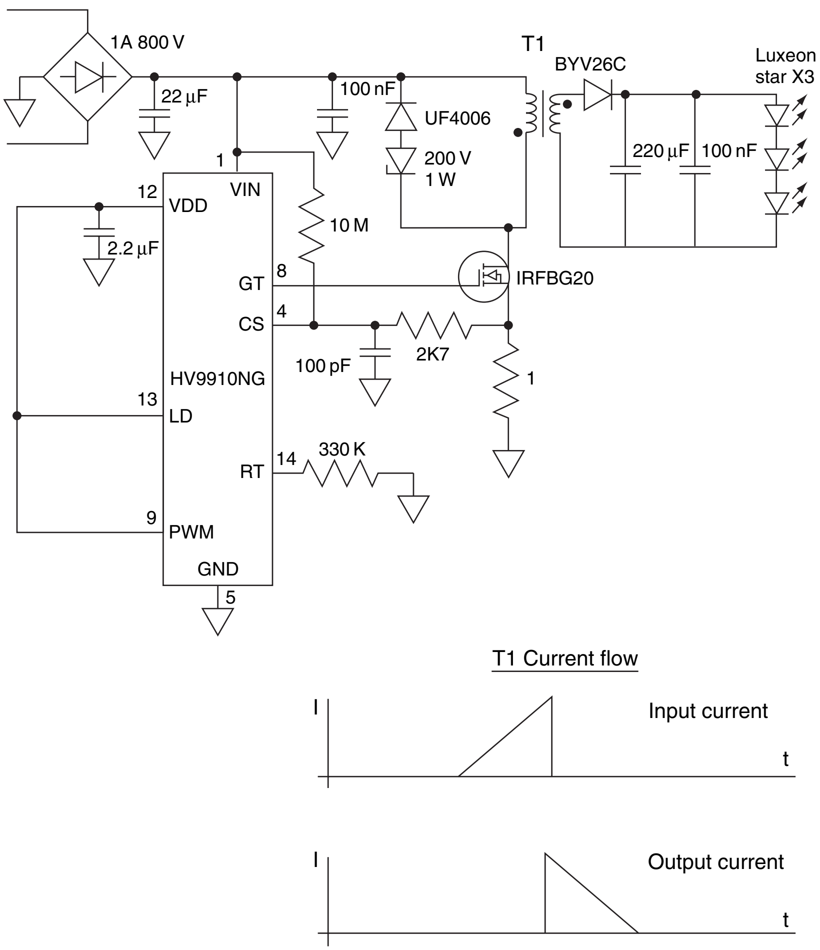


Figure 10.5: Discontinuous Fly-Back Current.

10.2.6 Inrush Limiters

Because almost all circuits have decoupling capacitors, when a power source is connected there will be an inrush current. This current can be very high, causing temporary heating in the capacitor and possible damage to switch contacts or components connected in series. Inrush current limiting using passive or active components can be provided to reduce this risk.

For AC mains applications, an NTC thermistor designed to carry high current is often used. In the active state, the flowing current warms the thermistor and hence the resistance falls to a low level to reduce losses. See Figure 10.6.

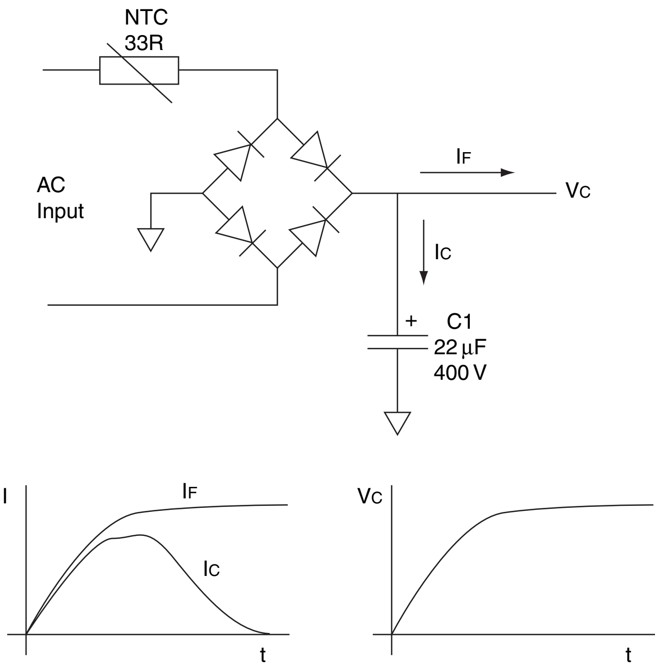


Figure 10.6: NTC Inrush Circuit.

For DC applications, an active inrush limiter is more common because the losses can be minimized during normal operation, when inrush limiting is not needed. This is shown in Figure 10.7.

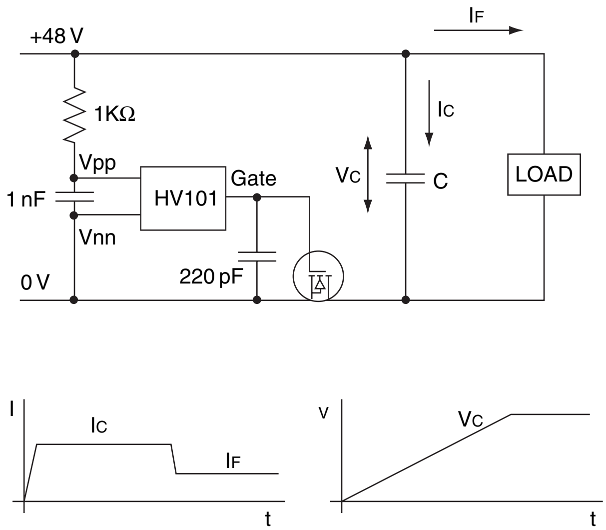


Figure 10.7: Active Inrush Circuit.

10.2.7 Soft-Start Techniques

Some applications need the input current to be controlled, to prevent high current spikes when power is first applied. This could be to reduce damage to switch contacts by the risk of sparking. Clearly the inrush techniques just described could be used, but sometimes it is necessary to control the output power instead.

For example, a circuit for driving one or two power LEDs from the AC mains could use a double-buck topology. But typical applications for this circuit are inside lamp housings, where an electrolytic capacitor cannot be used because of short lifetime or physical size. But using a polyester film capacitor means that the voltage dips between switching cycles; since the output power is normally constant, this means that the input current will peak as the input voltage dips. The peaks in input current give rise to considerable EMI and mean that the power factor is very poor. If the output current was controlled, i.e. reduced as the supply voltage dipped, the input current would remain constant when switching. The addition of a Zener diode in series with the supply to the controller IC would further improve the power factor.

Soft-start can also be implemented by connecting an RC filter to the analogue dimming input (e.g. linear dimming pin of HV9910B). The current level starts low

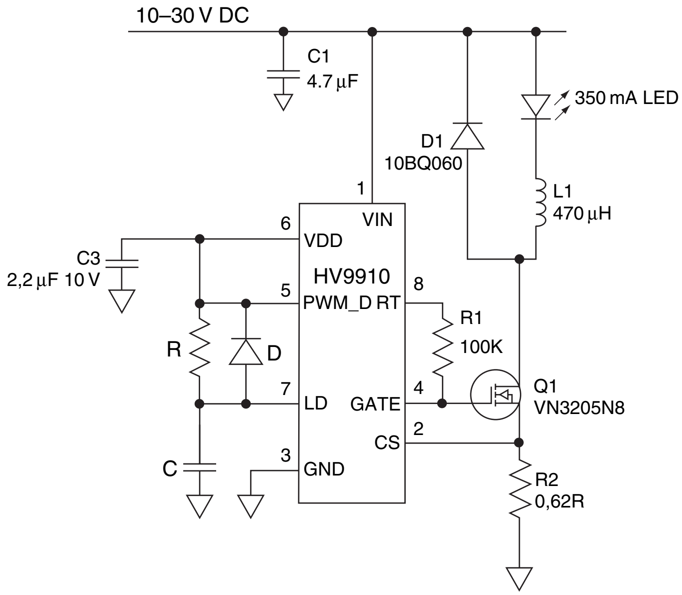


Figure 10.8: Soft-Start with HV9910B.

and grows as the capacitor charges. Clearly a method of discharging this capacitor is needed when the power to the IC is disconnected – a diode to the V_{dd} could be used to reduce the discharge time, see Figure 10.8.

CHAPTER 11

Selecting Components for LED Drivers

This chapter will be very practical in orientation. It will describe how different materials and component types can affect the performance of LED drivers. This will be detailed, showing how the physical construction of components could have an effect.

11.1 Discrete Semiconductors

Atoms of materials have a core (nucleus) of positively charged proton and uncharged neutrons. They have negatively charged electrons orbiting around this nucleus, like planets around the Sun. When atoms combine, they share electrons in their outer orbit (the valence band). Lighter atoms, like silicon, are most stable when there are eight electrons in their outer orbit. Semiconductors are (usually) made from silicon, which has four electrons in its outer orbit.

The addition of a small amount of material (dopant) with either three or five electrons in their atom's outer orbit can create an imbalance because, when combined with the four electrons of silicon, there are either seven or nine electrons in the outer orbit. When doped with material having three electrons in the valence band (boron (B), aluminium (Al), gallium (Ga) or indium (In)), the resultant outer orbit has seven electrons and a 'hole' where an electron is missing. This hole appears as a free positive charge and is called P-type semiconductor. This is shown in Figure 11.1, diagram A.

When doped with material having five electrons in the valence band (phosphorous (P), arsenic (As) or antimony (Sb)), the resultant outer orbit has nine electrons which means that there is a 'free' negatively charged electron and the material is called N-type semiconductor. This is shown in Figure 11.1, diagram B.

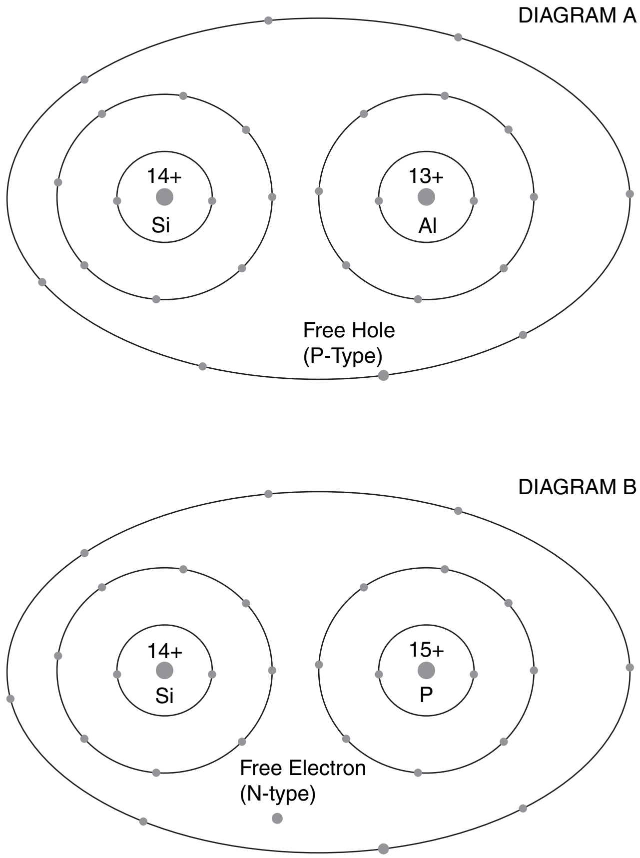


Figure 11.1: P-Type and N-Type Semiconductors.

When P-type and N-type semiconductor form a junction, the free electrons and holes combine and are destroyed. The fixed nuclei have a net negative and positive charge, respectively, and thus repel the combination of further free electrons and holes. Thus there is an energy barrier created; we have a diode junction. This is shown in Figure 11.2.

In order for a P-N junction to conduct, we must make the P-type material more positive than the N-type. This forces more positive charge into the P-type material and more negative charge into the N-type material. Conduction takes place when (in silicon) there is about 0.7 V potential difference across the P-N junction. This potential difference gives electrons enough energy to conduct.

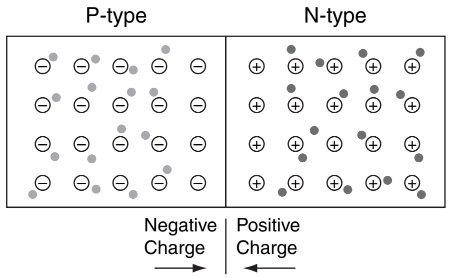


Figure 11.2: P-N Junction Diode.

11.1.1 MOSFETs

Metal oxide silicon field effect transistors (MOSFETs) are used as electronic switches in switching and linear LED driver circuits. They operate by using the 'field effect' in semiconductors; where an electric field attracts or repels free electrons in doped silicon. A MOSFET has three terminals – gate, drain and source; a fourth 'body' terminal is internally connected to the source. A diagram showing the physical construction of the MOSFET is shown in Figure 11.3.

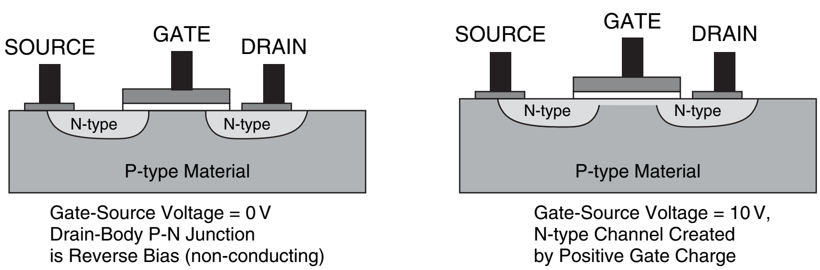


Figure 11.3: N-Channel MOSFET Construction.

Notice that the source and body are connected together by the metallized contact at the source. Also notice that there is a parasitic diode due to the P-type material of the body and the N-type material of the drain. This parasitic diode is reverse biased normally, because the drain is more positive than the body (and source), so does not need to be considered in all applications.

To create a conducting channel in the body of the MOSFET requires a certain amount of gate potential. MOSFETs are specified with a certain gate threshold voltage, usually at the point where the drain current reaches 1 mA, but this varies between manufacturers. Because the gate-body isolation is a dielectric, gate-source and gate-drain capacitance values are usually found in the datasheet.

Typical gate thresholds are in the range 4 V to 7 V; however, a number of 'logic-level' devices are now available. A 'logic-level' device is defined as one that switches fully on at *Vgs* equal to 5 V; this means that the gate threshold is typically about 2 V. So-called 'standard devices' are defined as being fully switched on at *Vgs* equal to 10 V. A logic-level device can also be operated with *Vgs* equal to 10 V or higher, in which case the on-resistance is lower. Many logic level devices have a higher gate capacitance compared to standard devices, for a comparable saturation current rating.

MOSFETs have two current ratings – peak current and continuous current. Continuous current ratings depend on the on-resistance of the MOSFET and are based purely on thermal considerations. Peak current ratings are the maximum current that is able to flow. When designing a switching LED driver circuit, the circuit current is pulsed and so the peak current rating is important. However, note that this current is normally quoted at 25°C; at 100°C the peak current is about half this value. As a rule of thumb, always use a MOSFET that has a peak current rating that is three times the value needed in the application.

When the MOSFET is connected to a load, but turned off, the drain is at high voltage. When the gate voltage rises, the MOSFET turns on and the drain voltage falls close to the ground (0 V) potential. The gate-drain capacitance thus sees a large voltage fall on the drain side and a slight rise on the gate side. At the gate pin, the gate-drain capacitance appears to be much larger than it really is; this is known as the Miller effect, named after the engineer who discovered this phenomenen. Figure 11.4 shows the parasitic capacitance in a simple MOSFET circuit.

Instead of considering the gate-drain capacitance and the gate-source capacitance, we can consider the gate charge. This is the total charge needed to turn the MOSFET on. In switching circuits, the gate charge is most significant and is usually quoted in nano-coulombs (nC). The average gate current is given by the equation:

$$I_{\mathrm{G}} = Q_{\mathrm{G}} * F_{\mathrm{SW}}$$

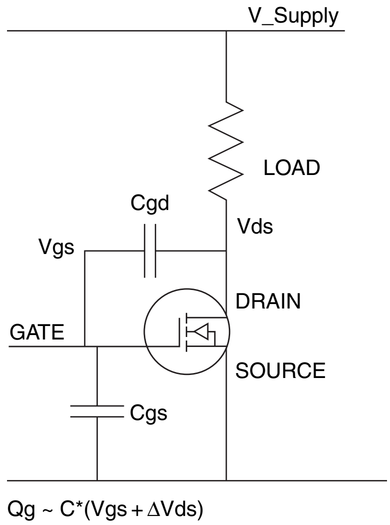


Figure 11.4: MOSFET Circuit with Parasitic Capacitance.

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 be matched to make a current mirror. Transistors of the same type have very similar characteristics, so by adding a low value resistor between the emitter and ground any variation in the base-emitter voltage (*Vbe*) is negligible; see Figure 11.5.

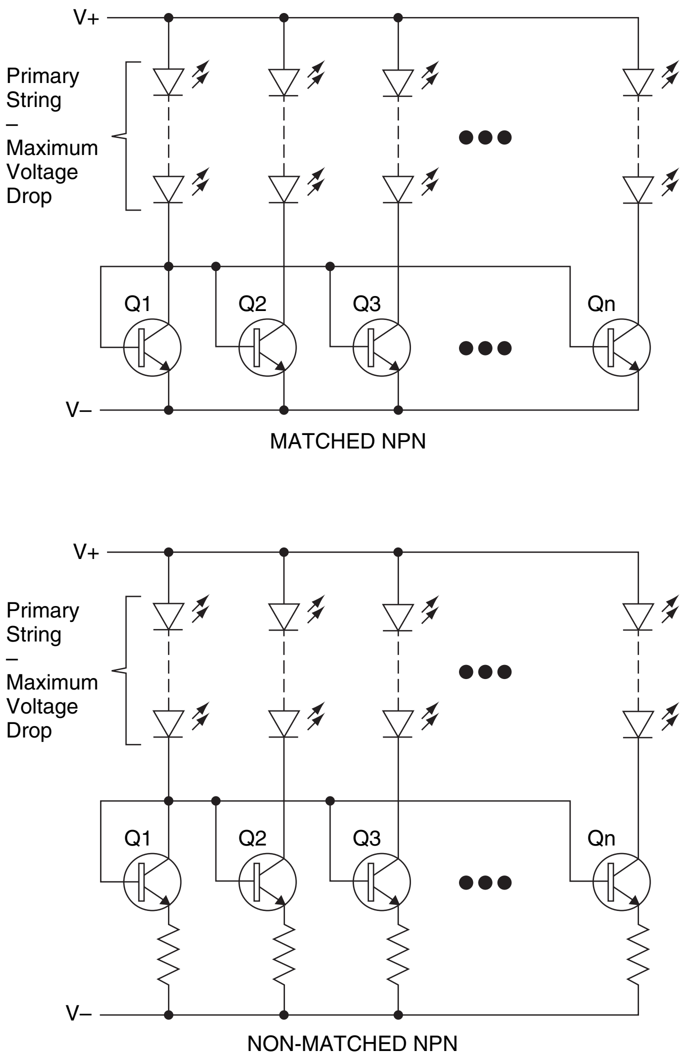


Figure 11.5: Current Mirror Circuits.

11.1.3 Diodes

There are many different diodes (rectifiers). Important parameters include: reverse breakdown voltage, forward current rating (average and peak), forward voltage drop, reverse recovery time and reverse leakage current.

Schottky diodes have the lowest forward voltage drop and the shortest reverse recovery time, but they are more expensive than standard diodes and generally have a limited reverse breakdown voltage range, although the company Cree has recently introduced high voltage Schottky diodes. Instead of a P-type and N-type semiconductor junction, the Schottky diode has an N-type semiconductor and metal junction. Reverse leakage is higher than in most P-N junction diodes. They are used for many applications, including reverse polarity protection and as flywheel diodes in low voltage switching circuits. Note that the forward voltage drop across a Schottky junction tends to increase with diode voltage rating, so use the lowest voltage rating suitable to keep the conduction losses to a minimum.

Diodes are sometimes labeled by their reverse recovery time. When the voltage across a diode is suddenly reversed, an initial current flow will occur in the reverse direction. Reverse recovery time (Trr) is the time taken to stop conducting when the diode is reverse biased. The labels fast, ultra-fast and hyper-fast are sometimes given. A standard rectifier diode like 1N4007 has a typical reverse recovery time of 30 microseconds, but an ultra-fast version UF4007 has $Trr = 75\,\text{ns}$, which is about 500 times faster. More recent devices are much faster, for example the STTH1R06 600V 1A rectifier with $Trr \sim 30\,\text{ns}$.

Shorter reverse recovery times cause lower switching losses. This is because the reverse current often flows through the MOSFET switch when the voltage across the MOSFET is high, so the less time when this happens gives lower losses. However, a 'snappy' diode can sometimes generate radio interference (EMI). In some applications a 'soft-recovery' diode should be used, where the turn-off speed in the reverse biased condition is fast but at a controlled rate of change.

In flyback power supplies, an RC snubber circuit is placed across the primary winding to prevent very high voltages when the MOSFET switch turns off. This snubber often has a medium speed diode in series so that the diode is still conducting for a period and allows any ringing current to flow through the RC network and thus decay quickly.

11.1.4 Voltage Clamping Devices

Voltage clamping devices are used to limit the voltage across a circuit, as part of a voltage regulator or a transient suppressor. These devices are typically semiconductors: Zener diodes, Transorb suppressors or voltage dependent resistors (VDRs).

Zener diodes behave like regular diodes in the forward conducting direction, but break down and conduct at a defined voltage in the reverse direction. Low voltage Zener diodes rated below 6 V have a soft knee in their current versus voltage graph; the conduction increases gradually. High voltage Zener diodes (avalanche diodes), rated above about 6 V, have a sharp knee and conduction increases very rapidly. Zener diodes can exhibit some noise when breaking down and are often used with a small capacitor in parallel to reduce this effect.

Transorb suppressors are like Zener diodes but are designed to handle high current peaks. Transorbs can be uni-directional or bi-directional and rated from low voltage ~5 V up to several hundred volts. A Transorb designed for 275 V AC operation will limit the peak surge voltage to below 600 V, even at high transient current levels.

A voltage dependent resistor (VDR) has high resistance at low voltage and low resistance at high voltage. Thus conduction increases gradually as the voltage across it increases. A VDR can absorb high surge energy; the devices are often rated in joules rather than watts, because the surge energy is short lived. A VDR rated at 275 V AC will break down and limit the voltage to about 710 V at high transient current levels.

11.2 Passive Components

11.2.1 Capacitors

In an LED driver, the key function of a capacitor (symbol *C*) is energy storage. There are two types of storage, slow storage and fast storage.

Slow storage is required across the DC terminals of a bridge rectifier, when the LED driver is powered from a low frequency AC supply. The purpose of this storage is to supply energy to the LED driver between the peaks of the AC voltage, which is twice every cycle. The AC frequency is typically 50–60 Hz, although 400 Hz is used in some aircraft, so the capacitor must supply energy and hold up the supply for as long as 10 ms.

For slow storage, an aluminium electrolytic capacitor is often used because it has a high energy storage density (they take up less space for an equivalent amount of storage, compared to other dielectric types). These capacitors are made using aluminium foil with a wet dielectric material. Because of this construction, they cannot be used in a high temperature environment for long periods; the dielectric dries out and the capacitor eventually fails.

Fast storage is required in switching driver circuits, where the switching frequency is often in the range 50–500 kHz. The energy only has to be stored for a short time, as short as a few microseconds, so the main characteristic of the capacitor for this function is to have the ability to store and discharge energy quickly. This means low self-inductance (high self-resonant frequency). Surface mount components generally have lower self-inductance because they have no added lead inductance. Typically, the capacitors used for fast storage are ceramic or plastic film types.

Capacitors are constructed from two conducting surfaces (known as plates) separated by an insulator (known as a dielectric). The metal plates are made from a thin metal film that has been deposited onto the insulation material. The dielectric can be a number of materials including ceramic, mica and plastic film. The capacitor type is usually known by the dielectric, thus there are '(aluminium) electrolytic' capacitors, 'ceramic' capacitors, 'polyester' capacitors, etc.

Ceramic and mica capacitors are made using flat dielectric sheets; the simplest construction uses just one insulating layer with a conducting plate on either side. Mica capacitors are very rarely used, but ceramic are fairly common. Higher valued devices use several insulating layers with interleaving layers of metal film. The metal film layers are bonded alternatively to side A, side B, side A, side B, etc.

Plastic film capacitors, such as polyester, polypropylene, polycarbonate, etc., use two layers of metallized plastic film. One form of construction is identical to that of ceramic capacitors, where flat sheets of metallized film are used. This type of construction is often found in surface mount polyester capacitors.

Another form of construction for plastic film capacitors uses rolled films. Two metallized layers are placed one above the other and then rolled, so that the two conductors spiral around each other with insulating layers in between. The films are laterally offset from one another so that the conductor of 'side A' protrudes from one side, and the conductor of 'side B' protrudes from the other side (this technique is sometimes known as extended foil). It is then relatively easy to bond lead wires to the ends of the resulting cylindrical body. The rolled form of construction provides a metal film around the body of the capacitor; this can be connected to earth or the 'earthy' side of a circuit to reduce external electric field pickup. The outer foil connection is marked on the case of some film capacitors.

A capacitor's behaviour is not ideal. Capacitors are formed from two conducting layers separated by an insulator. Every capacitor will have some series inductance; this is due to the plate conductors and the lead wires attached to them. This self-inductance can be a problem at frequencies close to the self-resonant frequency, and above. Each capacitor will also have series resistance due to both the conductors and the dielectric of the insulator, this is known as equivalent series resistance or ESR. The ESR will create losses. An equivalent circuit for a capacitor is shown in Figure 11.6.

Figure 11.6: Capacitor Equivalent Circuit.

Generally, ESR and self-inductance is more of a problem with aluminium or tantalum electrolytic capacitors. These types of capacitors are normally used to decouple power supplies. Digital circuit designers have long since become accustomed to connecting 10 nF ceramic capacitors across tantalum devices used for power supply decoupling. This is because the higher value tantalum capacitor absorbs low frequency transient currents, while the ceramic absorbs the high frequency transient currents.

Dissipation factor (DF) and loss tangent are terms used to describe the effect of ESR. The value of DF is given by the equation:

$$\text{Loss tangent} = DF = \frac{ESR}{Xc}$$

where Xc is the capacitor's reactance at some specific frequency. This is the tangent of the angle between the reactance vector Xc, and the impedance vector $(Xc + ESR)$, where the ESR vector is at right angles to the reactance vector.

One of the most notable problems with capacitors is self-resonance. Self-resonance occurs due to the device construction: leads are inductors (albeit low value) and wound capacitors can have some inductance because the currents circulate through the capacitor's plates. Consider the self-resonant frequency of capacitors, of various dielectrics, having a lead length of 2.5 mm (or 0.1 inch): a 10 nF disc ceramic has a self-resonance of about 20 MHz; the same value of polyester or polycarbonate capacitor also has a self-resonance of about 20 MHz.

A rough idea of the self-resonant frequency can be found by calculating the inductance of a component lead. For example, a 0.5 mm diameter lead that is 5 mm long (2.5 mm for each end of the component) has an inductance of 2.94 nH in free space. When combined with a 1 nF capacitor, the self-resonant frequency is calculated to be about 93 MHz. Replacing the 1 nF capacitor in previous calculations with a 10 nF capacitor, results in the self-resonant frequency falling to 29 MHz.

Earlier, I wrote that the self-resonant frequency of a 10 nF capacitor with 2.5 mm leads was about 20 MHz, not 29 MHz. The reason for the discrepancy between the calculated frequency and the actual frequency is that inductance in the plates was not taken into account. Adding the inductance of the plates gives a lower self-resonant frequency. As the value of the capacitor increases, the inductance of its plates also increases and so does the discrepancy between the calculated and the actual self-resonant frequency.

For small value capacitors of less than 1 nF the self-resonant frequency can be approximately calculated by the following equations: $f_R = \frac{1}{2\pi\sqrt{LC}}$, where L is the lead inductance. For a wire in free space, $L = 0.0002b\{[\ln\left(\frac{2b}{a}\right)] - 0.75\}$ µH, where a equals the lead radius and b equals the lead length. All dimensions are in millimeters (mm) and the inductance is in µH.

Using the formulae, if $a = 0.25$ mm (0.5 mm diameter) and $b = 5$ mm (2.5 mm each leg), the inductance is 2.94×10^{-3} µH. This is 2.94 nH. When substituted into the frequency equation, with a 1 nF capacitor, the self-resonant frequency is calculated to be 92.8 MHz.

Surface mount capacitors are in common use now because of their small size. In the past they were often used for high frequency circuits because there is no lead inductance to worry about. This reduction in inductance has benefits for switching power supplies too; where fast pulse rise and fall times are needed. The most popular type of surface mount capacitor is the multi-layer ceramic; its conducting plates are planar and interleaved – they have very little inductance. Some conventional leaded ceramic capacitors are surface mount devices with wire leads attached. They are usually dipped in epoxy resin or similar before having their capacitance value and voltage rating marked on the outside.

Ceramic capacitors generally have a temperature coefficient that is zero or negative. The terms NP0 or C0G are used to describe ceramic capacitors with a zero temperature coefficient (NP0 = Negative Positive Zero). Other ceramic dielectrics are described by the temperature coefficient; N750 describes a dielectric that has a negative temperature coefficient of $-750\,\text{ppm}/°\text{C}$. More exotic dielectrics are X7R and Y5U, which have a higher dielectric coefficient and are used to make capacitors with high capacitance values. The X7R and Y5U capacitors have a wide tolerance on the component value.

Apart from NP0/C0G capacitors, ceramic capacitors exhibit a piezo-electric effect. A high voltage AC signal can generate acoustic noise. The acoustic output increases with physical size, so a surface mount 1206 size capacitor will generate more noise than a 0805 capacitor in the same circuit. The piezo-electric effect will also cause the capacitance value to change with applied voltage.

Polystyrene and polypropylene capacitors have a negative temperature coefficient that fortunately closely matches the positive temperature coefficient of a ferrite-cored inductor. They are thus ideal for making LC filters. Unfortunately, with these dielectrics, capacitors tend to be physically large for a given capacitance value.

Polyester and polycarbonate capacitors are very common. Polyester capacitors are the worst in that they have a poor power factor (high ESR) and a poor (and positive) temperature coefficient. Polyester capacitors are popular because they have a high capacitance density (high capacitance value devices are small). Polycarbonate capacitors have a better power factor and a slightly positive temperature coefficient. Another useful feature of polycarbonate capacitors is that they are 'self-healing': in the event of an insulation breakdown due to over-voltage stress, the device will return to its non-conducting state, rather than become short circuit.

Capacitors used across the AC mains supply must be rated X2. For universal AC input, 275 V AC X2 rating is normally used. These capacitors are available in polyester and polypropylene, and 100 nF is a typical value found across the supply connections. This capacitor reduces EMI emissions and absorbs fast transient surges from the mains supply. In a typical application, a voltage dependent resistor (VDR, or varistor) is connected in parallel.

Capacitors from each AC power line to ground (earth) are sometimes used and must be 250 V AC Y2 rated. These capacitors typically have a dielectric of ceramic, polyester or polypropylene. Capacitance values are readily available in the range 1 nF to 47 nF. A value of 2.2 nF is commonly found in power supply designs.

11.2.2 Inductors

This section will describe 'off-the-shelf' inductors and transformers. Details of custom-made components will be covered in Chapter 12.

Inductors (symbol *L*) are used to store energy in switching LED driver circuits. A length of wire creates inductance, but winding insulated wire into a coil can magnify this; the wire is normally soft copper covered with a thin plastic film. The magnetic field produced by a wire then couples to adjacent wires; the inductance is proportional to the number of turns squared.

Although a simple coil creates inductance, if a magnetic material is placed within the coil the inductance increases considerably. The coil can be wound around a short ferrite or iron-dust rod to increase their inductance, but with this type of core a magnetic field will radiate and may cause interference (EMI). The advantage of this type of inductor is that the saturation current level is very high, considering the inductor size. A typical application for this type of inductor is in the power filter at the input of an LED driver. Many low value inductors look like wire ended resistors, with colored bands marking their inductance value.

Alternatively, the coil can be wound around a toroidal (doughnut)-shaped ferrite or iron-dust core. The toroidal shape keeps the magnetic field contained. Some toroidal materials have a distributed air gap, in which case the saturation current is very high. Toroidal inductor winding is not easy, and so these types of inductors can be more expensive than their bobbin wound counterparts.

As surface mount devices, shielded bobbin cores are popular. The coil is wound on a bobbin, within a closed ferrite material. These are low cost and small physical size, with the option of surface mount construction. The central ferrite core inside the coil often has an air gap to increase the saturation current rating, although this reduces the inductance value.

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 'T' 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 windings 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_{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 $I_{max} = 0.5 * I_{sat}$, but preferably $0.25 * I_{sat}$). Take care, because sometimes the current rating given by manufacturers is the DC current that causes a certain amount of heating, due to the winding resistance; the saturation current could be a lower current value. Some manufacturers quote a saturation current at the point where the inductance has fallen to 60% of the zero current value.

Sometimes an inductor data sheet will give a 'Q' value at a certain frequency. This is the voltage or current magnification value in a tuned circuit. It indicates the equivalent series resistance of an inductor, $Q = \frac{\omega L}{R}$, which is more accurate than the DC resistance measurement. This is because of the 'skin effect'.

The 'skin effect' raises the resistance of wire at high frequencies. The effect is due to an inductive force concentrated at the center of the wire, which forces the electrons to travel down the outside surface (hence 'skin' effect). This can be a serious problem for inductors working at a few hundred kHz, and is alleviated by the use of multiple stands of insulated copper wire, twisted together. Originally, the wire strands were covered overall with a cotton braid and called Litz wire. This is the type of wire used to make ferrite rod antennas for radios working in the low and medium frequency range (LF and MF). It comprises several strands of enameled copper wire inside a cotton braid. This wire has a lower skin effect because the current is shared down each of the strands; the surface area of all the strands combined is considerably larger than the equivalent diameter of the solid copper wire.

Off-the-shelf transformers are available with double or multiple windings, with or without an air gap in the magnetic core. Fly-back power supplies, including isolated LED driver circuits, use gapped cores; the air gap allows high magnetic flux density within the core – the energy is stored and then released. A forward converter is a popular power supply topology, which uses ungapped cores because magnetic energy is not stored in the core – it is immediately transferred to the secondary winding. Forward converter power supplies are rarely used in LED driving.

Transformers with multiple windings are used to create a step-up or step-down primary-to-secondary turns ratio. This allows the duty cycle of the switching circuit to set within a certain range. Very small duty cycles less than 5% should be avoided, because of the difficulty in controlling the switching (due to delays in the system). Duty cycles greater than 50% can cause instability unless external compensation circuits are added. In some cases, such as where the input voltage range is very wide, a wide range of duty cycle may be unavoidable.

Another reason for an additional winding is to create a 'bootstrap'. A bootstrap circuit creates a power supply for the switching circuit, typically in the range 8–15 V. The switching circuit will be powered from the main power source initially, but this can be inefficient if the power source is high voltage. Once switching starts, the voltage developed on the bootstrap winding can be used to self-power the switching circuit. Suppose the device needs 2 mA to operate, when powered from a 300 V DC supply it will dissipate 600 mW, but when powered from a bootstrap winding at, say, 10 V it will only dissipate 20 mW.

11.2.3 Resistors

There are several types of resistor. Wire wound devices are rarely used and would not normally be placed in a circuit that carried high switching current because they have a high self-inductance. They are used at the AC power input of some power supplies to provide some impedance for fast transients and surges. Carbon composition resistors tend to be noisy and have a poor temperature coefficient, but are good in switching power circuits because of their low inductance construction. They are constructed using carbon particles set in a clay rod and the resistance depends on the surface area of the touching particles. Carbon film and metal film devices are most common; surface mount film devices are usually thick film construction.

Carbon film resistors are low noise devices with a negative temperature coefficient. Component tolerances of 1% and 5% are standard, although 0.1% are available albeit more expensive. Through-hole resistors are constructed by applying a carbon film onto a ceramic rod, and then cutting a spiral gap in the film to increase the resistance. The spiral conductor is actually a lossy inductor. Surface mount devices have a carbon film applied to one side of a ceramic layer and a laser is used to cut across the film to alter the resistance. The short length of carbon film has very little inductance.

Metal film resistors have a lower noise than carbon film types, and a lower temperature coefficient. Component tolerances of 1% are standard, although precision devices in an E96 range of values with 0.1% tolerance and 15 ppm temperature coefficient are available at a higher cost. These resistors are constructed by applying a number of metal film layers, of different metals, to a ceramic former to achieve the correct resistance and a low temperature coefficient. In through-hole resistors, a spiral gap is sometime cut around the metal film to increase the resistance value and this increases the inductance slightly.

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 ceramic 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.

11.3.1 Through-Hole PCBs

It is usual for an RF or high speed digital circuit to have an earth plane on the printed circuit board (PCB) component side. In many cases, an LED driver can be considered as a high speed digital circuit.

The earth plane serves two purposes; it screens the components from tracks passing underneath, and it provides part of a low loss transmission line. By using FR4 board in a standard thickness of 1.6 mm, 50 ohm transmission lines can be created by making the printed circuit tracks 2.5 mm wide. A transmission line is formed between the earth plane and the track.

The technique of providing an earth plane on high speed PCBs may cause problems when an inductor is placed on the board, because of the capacitive coupling between the ends of the inductor and the earth plane. This capacitance forms a parallel tuned circuit with the inductance and may cause the filter to be detuned. One solution is to remove the earth plane from the area below the inductor. An alternative solution is to mount the inductor on spacers above the board, so reducing the capacitance.

11.3.2 Surface Mount PCBs

Surface mount components are used extensively in LED driver circuits. Ceramic capacitors are common but can be damaged by stress due to circuit board expansion. One method of minimizing this problem is to use physically small devices: devices larger than 1812 (0.18×0.12 inches) should be avoided.

Ceramic capacitors should be protected with a moisture resistant coating. If moisture is absorbed into the ceramic material, the capacitance value will change. Moisture can also be absorbed into plastic packages, so a conformal coating over the whole board is preferred. Some consideration should be given to storage of components; metallized sealed bags should be used, perhaps with desiccant material. This will prevent moisture being trapped into an assembled board and avert the risk of damage during soldering (as the moisture boils off).

Through-hole PCBs have plated through-holes that are 1 mm or larger in diameter. Surface mount boards do not need holes large enough for component leads; hence they tend to be smaller in diameter. Metallized ‘via’ holes 0.3 mm in diameter are common (used to connect two tracks rather than for component leads).

A problem with via holes arises when the board is heated. Glass and epoxy board, e.g. FR4 type, has a high coefficient of expansion at temperatures above 125°C. Above 125°C the board goes through its glass transition temperature and its coefficient of expansion is greater than normal; Z axis expansion increases the thickness of the board, and can cause fractures between the tracks and the via hole pads.

Soldering causes a problem due to the heat applied to the board; in wave soldering the board is heated to about 300°C, which is way above the glass transition temperature. To reduce the problem of 'via-hole' damage, all plated through-holes should have a wall thickness of 35 μm or more. Temperature cycling of completed boards also causes problems.

On the surface, there is a temperature coefficient mismatch between components and the board. Leadless chip carrier (LCC) devices have an expansion coefficient of 6 ppm/°C, but for the board it is 14 ppm/°C (below the glass transition temperature) in the X-Y plane. Above the glass transition temperature the PCB has a coefficient of expansion of 50 ppm/°C. Again, temperature cycling strains the solder joints and can lead to failure. A small gull-wing IC does not have a problem in this respect, because the leads can bend slightly.

Printed circuit boards built on aluminium sheet are often used with power LEDs. They can also be used for higher power LED driver circuits. Traditionally, copper clad invar has been used within some PCBs to restrain expansion and to distribute heat. This should be used with polyamide boards, rather than glass and epoxy types.

Solder resist can be used to restrain solder, but this can create large blobs on the lead or pad area. Surface mount ICs use smaller packages than conventional leaded devices, and thin tracks of solder resist between the pads are not practical.

PCBs that have a fine track pitch sometimes have 0.05 μm gold plating. If the gold is thicker it causes embrittlement. Gold or nickel plating gives a flat surface and makes surface mount component placing easier.

11.4 Operational Amplifiers and Comparators

The operational amplifier has DC characteristics that may change with temperature, but those most affected are the DC offset, bias current, etc. The AC characteristics are less affected by temperature.

The greatest problem is that the op-amp is not ideal. The ideal op-amp has infinite input impedance, zero output impedance and a flat frequency response with linear phase. Most practical op-amps have very high input impedance and this does not cause us many problems. The output impedance is not zero, and can be up to about 100 Ω. This is not often a problem because negative feedback is used to limit the gain of the op-amp, and this also makes the effective output impedance close to zero. There is, however, an assumption that the gain-bandwidth of the op-amp is far higher than that required by the circuit. If the gain-bandwidth product limit is approached, the output impedance rises.

If the op-amp has insufficient gain-bandwidth product, excessive phase shifts occur and the circuit can show peaking in the frequency response. Gains of 20 dB close to the cut-off frequency can occur unless care is taken in the design. A good frequency response can be obtained by utilizing an op-amp that has a gain-bandwidth product many times that of the circuit's bandwidth. A rule-of-thumb value is 10 to 100 times the bandwidth.

Comparators are used in many LED drivers to detect the current level in a sense resistor. Comparators can be described as an op-amp with a digital output; they compare two voltages on their input and set the output high or low depending on whether the non-inverting input is higher or lower than the inverting input. Often a comparator has some inbuilt hysteresis to prevent jitter when the two inputs are at or near the same potential.

One weakness of a comparator is that they invariably have some input offset voltage; this results in an error in the switching and limits the minimum voltage that can be used for a reference. For example, the current sense comparator in the HV9910B LED driver has an offset of about 10 mV, and the maximum threshold for switching is 250 mV, so the threshold range is 10–250 mV, and could potentially give a dimming range of over 20:1. In practice, operation at the low voltage end of the range would be very noisy and a minimum threshold voltage of about 25 mV is recommended.

It is possible to build a comparator by using an op-amp with positive feedback. However, the output stage has been designed as a linear circuit and the slew rate is slower than a comparator's output.

CHAPTER 12

Magnetic Materials for Inductors and Transformers

Standard off-the-shelf transformers and inductors were described in Chapter 11. This chapter will describe magnetic materials and techniques for constructing custom transformers and inductors. The primary design requirement is to minimise losses, but to do this we have to consider copper losses, core losses, magnetic saturation, size and construction. Since this book is about designing LED drivers, only the basics of magnetic materials will be given here. For more detail, the reader should consult specialist books on the subject.

An inductor can be made from a coil of wire, wound on a bobbin, and surrounded by a soft magnetic core material. By soft, the meaning is that magnetisation is easy and demagnetisation occurs when the magnetising force is removed. A hard magnetic core is like a permanent magnet; it has high 'remnance' (magnetic field remaining once the magnetising force has been removed). Most magnetic materials have some remnance and the field strength required to return the magnetic flux to zero, to overcome this remnance, is called the 'coercivity'. On a graph showing magnetic flux versus field strength, the curve follows an italic 'S' shape. But when the magnetic field is reversed, the flux does not follow the same curve; it needs more field strength (more energy) to return to the same point and thus forms a 'fat S' shape. The fatter the S, the higher the magnetising losses.

The core can be rectangular or cylindrical in cross-section with two halves that separate to allow the bobbin to be inserted. When the inductor is assembled, two spring-steel clips (or adhesive) hold the two halves of the core together. This form of inductor is suitable for values of a few micro-henries up to about one henry.

The advantage of making a custom inductor is that they can be made to almost any value. Remembering that inductance is proportional to the number of turns squared, the number of turns required is given by the simple formula: $N = \sqrt{\frac{L(nH)}{AL}}$. Here L is the required inductance in nano-henries and AL is the core's inductance factor (nano-henries per turn). Each core type has an AL value determined by the core manufacturer, which will be given in the manufacturer's datasheet or catalogue. The AL factor is the inductance, in nano-henries, that will be produced for a single turn of wire.

The core's AL value is related to the permeability of the magnetic material used. Different magnetic materials are used, depending on the frequency at which the inductor is operating. If a particular AL value is required, it can be obtained by removing some of the magnetic material from the center of the core, thus creating an air gap. Note: an air gap in the center of the core, rather than in the outer material, reduces the emission of magnetic fields because the outer material behaves like a shield. The air gap has a lower permeability than the ferrite material, so increasing the gap reduces the overall AL value. A typical core gap is 0.1 to 0.5 mm, although it may be larger or smaller depending on the magnetic material permeability and the required AL value. The larger the air gap, the higher the magnetizing force that be achieved without saturating the core.

The presence of an air gap in inductor and transformer cores makes them suitable for high magnetic saturation levels. An example application for this is inductors in power factor correction (PFC) circuits, which have a discontinuous magnetising force. In PFC circuits, the current is switched on and off at high frequency with zero current flow between each pulse. The amplitude of the current pulse is made to rise and fall in proportion to the instantaneous AC voltage, so the average current is sinusoidal. Thus the power factor is close to unity (true sine wave).

Transformer cores that have no air gap are prone to saturate easily; their AL is normally far higher than an inductor core made from a similar magnetic material, but with an air gap. Gapless inductor cores are often used in forward converters, in which the secondary current flows at the same time as the primary current. In a forward converter, there is no stored energy in the transformer.

If the coupling between windings must be very close, bifilar winding is often used. A bifilar winding has two insulated strands of wire twisted together before winding.

Trifilar and higher order windings use multiple strands. However, if high voltage insulation is required between the windings, bifilar techniques cannot be used (unless special winding wire with high voltage insulation is available, e.g. Rubadue wire).

Sometimes multiple winding strands are used to reduce the equivalent series resistance, because at high switching frequencies the skin effect must be considered. Remember that the skin effect forces current to flow through the outside surface of a conductor, so if insulated strands are used the effective surface area is very large. A type of winding wire with multiple twisted strands is called Litz wire; each strand has a thin polymer film surrounding the conductor, for insulation.

12.1 Ferrite Cores

Ferrite cores are available in many shapes and material types. These cores are quite brittle and can break if dropped or struck with a hard object. Ferrite is usually a compound made from magnesium and zinc, or from nickel and zinc. Most ferrites have very poor electrical conductivity, which limits any eddy currents in the core.

Nickel-zinc ferrites are used in inductors intended for EMI filters, because they have high losses at high frequency – the core absorbs most of the energy above 20 MHz, up to about 1 GHz.

Manganese-zinc cores have losses that rise above 10 MHz, but have little effect on signals above 80 MHz. This characteristic makes them almost useless for EMI filtering.

Manufacturers' data should be studied for details of the switching losses and optimum switching frequency. Ferrite is less effective at very low or very high frequencies. Generally, frequencies in the range 10 kHz to 1 MHz are suitable for ferrite cores.

12.2 Iron Dust Cores

Iron dust cores (also called iron powder cores) are sometimes often made toroidal (doughnut) shaped. The iron dust is ferrous oxide and is mixed with clay-like slurry, which sets when baked. The result is ceramic material with soft-magnetic properties and with high magnetic saturation levels.

These cores are good for switching frequencies up to about 400 kHz. From about 10 MHz up to 20 MHz, the core is very lossy. Above 20 MHz the core has little effect and so cannot be used in EMI filtering applications.

12.3 Special Cores

Proprietary compounds are used to make special cores. An example is MPP (molypermalloy powder). This has the ability to operate with high flux density of typically 800 mT, rather than 200 mT of conventional ferrite cores.

Molypermalloy powder (MPP) cores are distributed air gap toroidal cores made from a 79% nickel, 17% iron and 4% molybdenum alloy powder for the lowest core losses of any powder core material.

MPP cores possess many outstanding magnetic characteristics, such as high electrical resistance (thus, low eddy current losses), low hysteresis (magnetizing) losses, excellent inductance stability after high DC magnetization or under high DC bias conditions and minimal inductance shift when subject to flux densities up to 2000 gauss (200 mT) under AC conditions.

12.4 Core Shapes and Sizes

For custom inductors and transformers, E-cores are popular. An E-core has two halves that look like a capital E. The center segment is designed to pass through the middle of a bobbin on which the windings are wound. This center segment can be machined to create an air gap, as shown in Figure 12.1, to allow high magnetic flux without saturation of the core.

Variations on E-cores are EF and EFD cores. The EFD core is shaped so that the center segment is thinner than the main body of the core, so that the bobbin has a rectangular cross-section, rather than square.

Pot-cores have a round body with a central spigot, so that a round bobbin drops inside the cavity. However, the area on the circuit board is essentially square. This means that the ferrite core has less material and does not provide the maximum use of the space. These cores are rarely used except in a tuned filter, when an adjuster is provided in the central spigot.

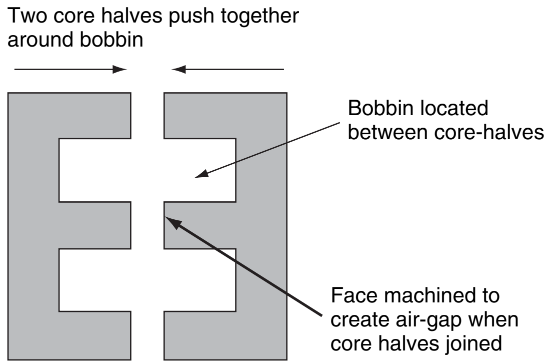


Figure 12.1: E-Core.

Toroidal (doughnut-shaped) cores are good from an EMI point of view, because the magnetic field is fairly well kept in the ferrite core; there are no ‘corners’ in the core where magnetic flux is prone to leak out. However, toroids are difficult to wind, since the wire must loop many times through the central hole. Special coil winders are available for toroidal cores. Magnetic saturation can be a problem, so MPP and iron powder tend to be used because they have the ability to carry a high flux density. A toroidal core is shown in Figure 12.2.

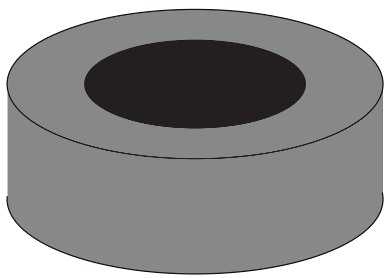


Figure 12.2: Toroidal Core.

12.5 Magnetic Saturation

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 magnetization is forced to operate outside its linear region. Generally, the magnetic flux density should be limited to about 200 mT (200 Weber/m^2).

If an inductor or transformer has a large discontinuous current flow, as in certain fly-back transformers and input inductors, the magnetic flux density may need to be lower than 200 mT. Ferrite core manufacturers recommend that flux variation due to ripple current or discontinuous mode operation should be limited to 50 mT. Inductors requiring the ability to handle high levels of flux variation sometimes use special cores with low losses at high flux density, in which flux levels much greater than 50 mT are used. This allows a much smaller inductor size.

The flux density is given by the equation: $B = \frac{L\,I}{N\,Ae}$. Here, L is the inductance, I is the peak current, N is the number of turns and Ae is the effective core area. The inductance and peak current are calculated in the design of the LED driver circuit. We do not know the core area or the number of turns at this stage, but through iteration we can find something suitable.

The approach for choosing a suitable core is to select a core with a known effective area (Ae value), find the number of turns, and then calculate the maximum AL value that can be used with that size core. The number of turns can be found by transposing the previous equation: $N = \frac{L\,I}{B\,Ae}$. The equation for the maximum AL value is: $AL\,(\mathrm{max}) = \frac{L * 10^9}{N^2}$.

Cores are usually available with standard AL sizes. If a core is available with a slightly lower AL value than the maximum previously calculated, it should be selected and then a new value for the number of turns should be used, $N_1 = \frac{L\,I}{B\,Ae}$. However, if a lower AL value is not available, a larger core size with a higher Ae value should be selected and the above process repeated. A simple spreadsheet can be created to make this process quick and simple.

12.6 Copper Losses

Copper loss is the term used to describe the energy dissipated by resistance in the wire used to wind a coil. In 99.9% of cases this wire will be made of copper, whose resistivity at 20°C is about 1.73×10^{-8} ohm meter. However, coils often have to operate above room temperature and will be heated by the operating losses in any case. The wire resistance at any temperature can be estimated from Table 12.1, developed by Mullard (now Philips).

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 surface of the wire. Thus changes in current produces a force that opposes those changes, 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:

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

For example, suppose we are operating at 100 kHz, then $D = 0.63$ mm. Using a larger diameter wire than this does not give very much benefit, because the current will not be carried in the center of the wire. In fact, in an LED driver (or any PWM power supply) there are harmonics at many times the switching frequency. In the case above, a significant proportion of the signal will have a frequency of 300 kHz.

In some cases, it is necessary to suffer higher copper losses that desirable, in order to have a transformer of a reasonable size. The use of Litz wire may be justified (although it is expensive) if low copper loss is essential at high switching frequency.

CHAPTER 13

EMI and EMC Issues

The first two questions regarding EMI and EMC are: what is 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, screening 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.

Before we look at EMI and EMC standards, and the design techniques used to meet them, it is important that we understand signals. Fourier analysis shows that any signal that is not a pure sine wave can be considered as a fundamental signal plus higher frequency harmonics, which are a multiple of the fundamental frequency. For example, a square wave with a 50/50 duty cycle has a fundamental signal at the switching frequency plus a 3rd harmonic of 1/3 amplitude, plus a 5th harmonic at 1/5 amplitude, plus a 7th harmonic at 1/7 amplitude, etc. If the signal is not 50/50 duty cycle, or if the switching edges have some slope (as all practical signals do), then there will be both odd and even harmonics present and the amplitude of harmonics will be less predictable. Typically, this is like the signal across a MOSFET switch in an LED driver circuit.

13.1 EMI Standards

13.1.1 AC Mains Connected LED Drivers

Any LED driver connected to AC mains supply has to meet the limited specified in harmonic current emissions standard IEC/EN 61000-3-2. Within this standard there are several classes and the one related to lighting is Class C. The harmonic emission limits specified in IEC/EN 61000-3-2, Ed. 2: 2000, up to the 40th harmonic, are listed in Table 13.1.

Table 13.1

Harmonic order 'N'	Maximum current, Class C (percentage of fundamental current)
2	2%
3	(30 × Power factor) %
4–40 (even)	Not specified
5	10%
7	7%
9	5%
11–39 (odd)	3%

Conducted emission limits in the 150 kHz to 30 MHz frequency range are specified in the standard IEC/EN 61000-6-3.

13.1.2 General Requirements for all Equipment

All LED drivers have to meet the radiated emissions standards. The standard is IEC/EN 61000-6-3, which covers the frequency range 30 MHz to 1 GHz. This standard uses limits previously set by CISPR22 in the USA and by the European Norm EN55022. The limits given in CISPR22 and EN55022 standards were intended for computers and communications related equipment, but these have been adopted as generic limits for all electronic products, including lighting.

The emission levels to meet EN55022/CISPR22 Class B are 30 dBμV/m in the frequency range 30 MHz to 200 MHz. From 200 MHz to 1 GHz the emission level increases to 37 dB μV/m. These are the signal levels measured at a range of 10 meters from the equipment under test (EUT). Since the signal power is proportional to $1/R^2$; for example, at 1 meter from the EUT the emission limit will be 20 dB higher (100 times the power), at 50 dB μV/m and 57 dB μV/m, respectively.

13.2 Good EMI Design Techniques

It is important to look at the circuit diagram and determine where the possible sources of EMI are located. This should happen before the printed circuit board (PCB) is designed. The center point for EMI sources must be the MOSFET switch. This turns on very quickly and so has sharp edges with high frequency content. When looking at the circuit schematic, consider the effect of high frequencies (1–200 MHz).

At very high frequencies, an inductor that was thought to block AC signals suddenly behaves like a capacitor that passes AC signals very easily. Similarly, a capacitor thought to have a low impedance characteristic behaves like an inductor at very high frequency; a good example of this is an electrolytic capacitor. So, check the component datasheets and look at the frequency response curves showing impedance versus frequency; see where the resonant frequency is – you will be surprised!

13.2.1 Buck Circuit Example

Let us take a look at a simple buck circuit to see where the EMI can arise. Figure 13.1 shows a typical buck circuit. The integrated circuit is a PWM controller. Internally, a clock signal triggers a latch, causing the gate drive output to be activated. The MOSFET $Q1$ turns on and the current increases at a fairly constant rate, due to the inductance of $L1$. When the CS pin is raised above 250 mV, due to current in $R2$, the internal

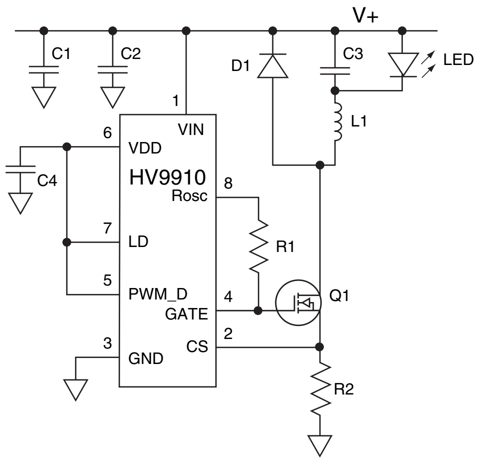


Figure 13.1: Buck Circuit.

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 thorough 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

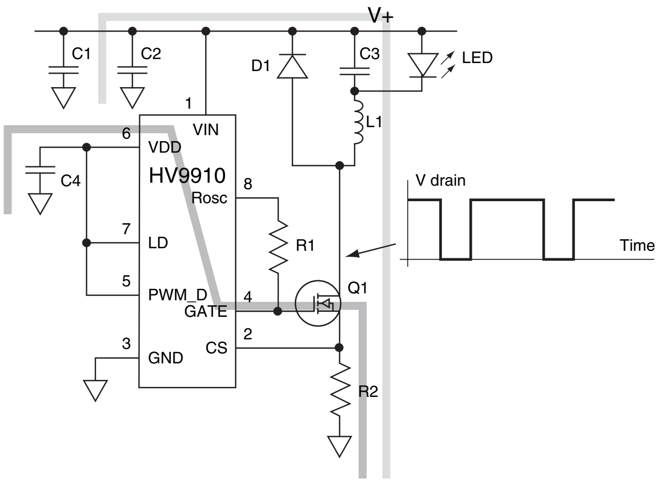


Figure 13.2: Buck Circuit Current Flows.

turns on. This reverse current flows for a short time (typically 75 ns or less) and creates a current spike that can cause false current sense triggering at the integrated circuit. A small part of this current is due to the junction capacitance, but the main part is reverse recovery current.

Reverse recovery current is caused by a diode junction that has a forward current and is then subject to a sudden reverse polarity. The free electrons in the junction take some time to clear and thus create a depletion region inside the silicon. In a buck circuit, the flywheel diode is in forward conduction when *Q*1 first turns on and so has an associated reverse recovery current.

The choice of capacitors in the circuit is important. Capacitor *C*2 must have low impedance at high frequency, for handling the high frequency current to the power switching circuit. The capacitor dielectric could be ceramic for low voltage supplies, or metalized plastic film such as polyester.

The capacitor *C*3 across the LED terminals carries the high frequency signals due to the capacitance of the inductor windings. The inductor winding capacitance is simply due to insulated wires being wound over each other in a coil. Some inductors have more self-capacitance than others, due to differences in construction. The capacitor *C*3 must withstand the voltage across the LED, or the supply voltage if there is a chance that the LED could be disconnected. It must be low impedance and able to carry high frequency signals. A typical value is 100 nF.

The *VDD* capacitor *C*4 should be a ceramic dielectric type, value typically 2.2 µF. This can be a low voltage type; a 16 V rating is commonly used.

We briefly mentioned the inductor *L*1. We discussed the interwinding capacitance, which affects performance by causing current spikes when *Q*1 turns on. But the magnetic field must be considered too; a shielded inductor or a toroidal construction should be used to minimize radiating magnetic fields.

When considering filters, we need to raise the impedance of the current path into the power source by adding an inductor *L*2; this is shown in Figure 13.3. A small capacitor *C*5 on the power source side of *L*2 shunts any small signals that manage to pass through *L*2. Basically adding *L*2 and *C*5 creates a low-pass filter to attenuate (reduce) and high frequency signals from the switching element *Q*1.

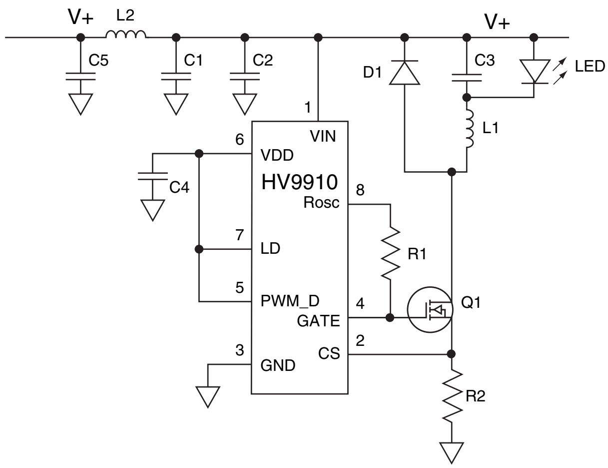


Figure 13.3: Buck Circuit with Filter.

If a filter has been added, but emissions are too high, consider placing a resistor in series with the MOSFET gate. A value in the range 10 ohms to 100 ohms is likely to be sufficient. The resistor slows down the gate charging when switching on and switching off, so the high voltage switching now has sloped edges that has fewer high frequency harmonics.

The paths for the switching currents must be kept short and compact when laying out a printed circuit board (PCB). If components cannot be co-located, so the path length is a little longer than desired, a return path should be placed alongside to ensure that the magnetic field from the current loop is minimized. Using the circuit schematic of Figure 13.3, we can look at the PCB design. In Figure 13.4, the track layout of the bottom layer is shown.

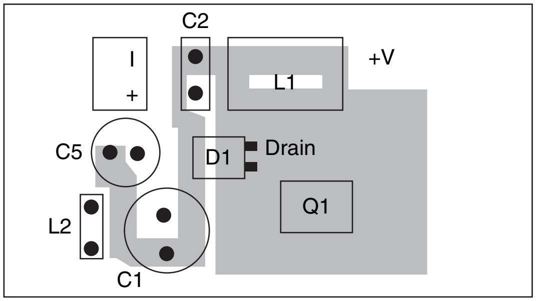


Figure 13.4: PCB Bottom Layer.

Notice how the ground connection goes from *C*5 to *C*1 and then on to *C*2 before reaching the ground plane. By avoiding the direct connection between the ground plane and *C*1, the current flow is steered in the direction we want. High frequency signals are taken from the grounded side of *C*2, which is low impedance at high frequency. The capacitors *C*5 and *C*1 hold up the input voltage during the cusps of the rectified AC input and are not intended to supply the high frequency current pulses needed for the LED load.

Figure 13.5 shows the tracks on the PCB component side. The positive supply from the bridge rectifier *BR*1 flows to *C*5 and onto filter inductor *L*2. From the other side of *L*2, it passes to *C*1 and then *C*2. Notice that the *C*2 connection is a node where

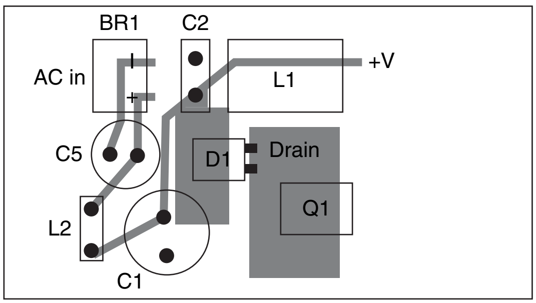


Figure 13.5: PCB Top Layer.

current also returns from the cathode of *D*1. Thus the high frequency flywheel current loop is from *Q*1 drain, through diode *D*1 and back to ground via *C*2; 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 *C*5 and *C*1.

Figure 13.6 shows both sides of the circuit board overlaid. Notice that the earth plane is below the drain area of *Q*1 and inductor *L*1. Both *Q*1 and *L*1 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.

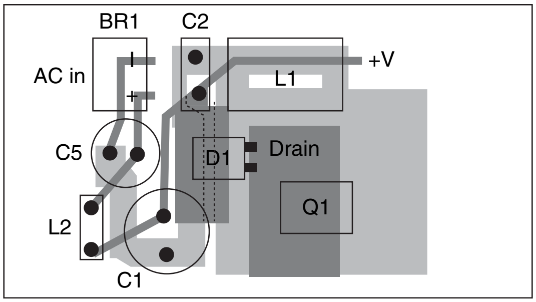


Figure 13.6: PCB Top and Bottom Layers.

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

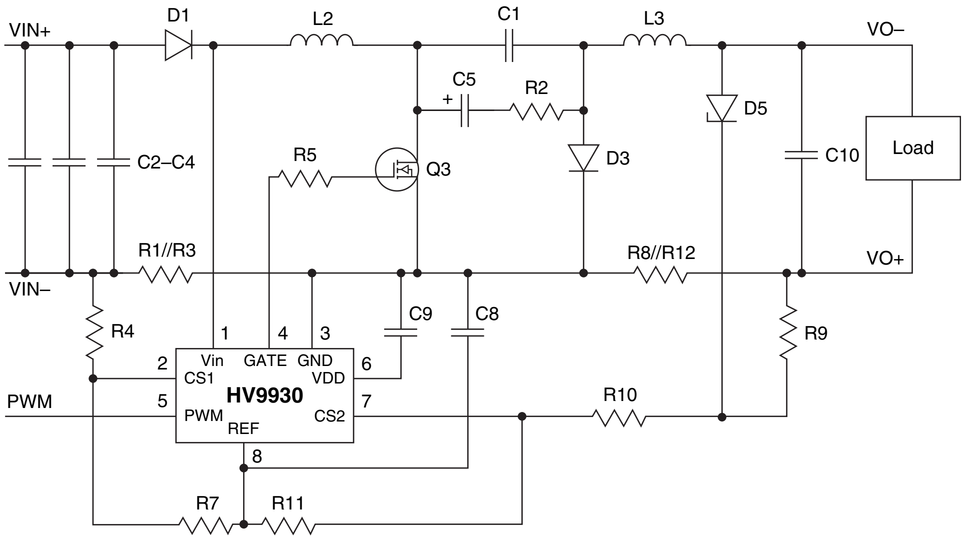


Figure 13.7: Basic Cuk Circuit.

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 (*R*5) 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.

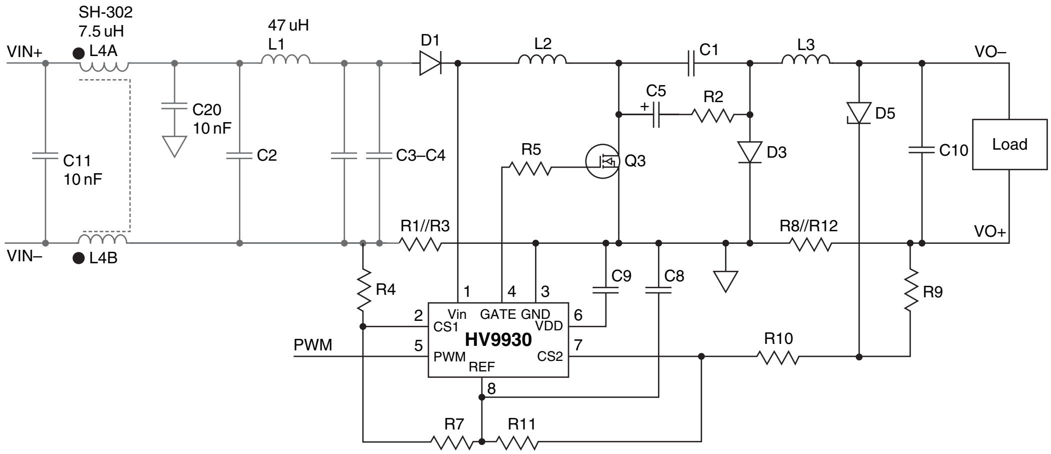


Figure 13.8: Input Filter.

Capacitors $C3$ and $C4$ provide the high frequency current source; these are ceramic capacitors and have very little high frequency ripple across them. Inductor $L1$ and capacitor $C2$ form a low pass filter for any ripple that appears across $C3$ and $C4$.

The circuit ground is not the same as the supply ground, because two parallel resistors $R1$ and $R3$ break the ground connection. This means that a path for any high frequency signals is needed from the positive input to both the circuit ground and the supply ground. The path to circuit ground is provided by $C20$. This small value ceramic capacitor does not affect current sensing at the switching frequency. The path to supply ground is provided by $C2$, which is a high value ceramic capacitor.

Stray coupling from the LED driver circuit to ground can create a common mode signal that is present equally on positive and negative inputs. This means that a differential capacitor, like $C2$, has no effect since the voltage is the same on both sides of the capacitor and no current flows through it. For this type of signal, a common mode inductor (choke) is required.

A common mode inductor $L4$ has two windings on a common magnetic core. Differential currents produce opposing magnetic fields, so the result is no net inductance. Common mode currents produce magnetic fields that add together and thus have a high inductance. A common mode signal will present high impedance to common mode signals and reduce radiation.

Finally, a small value ceramic capacitor *C*11 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.

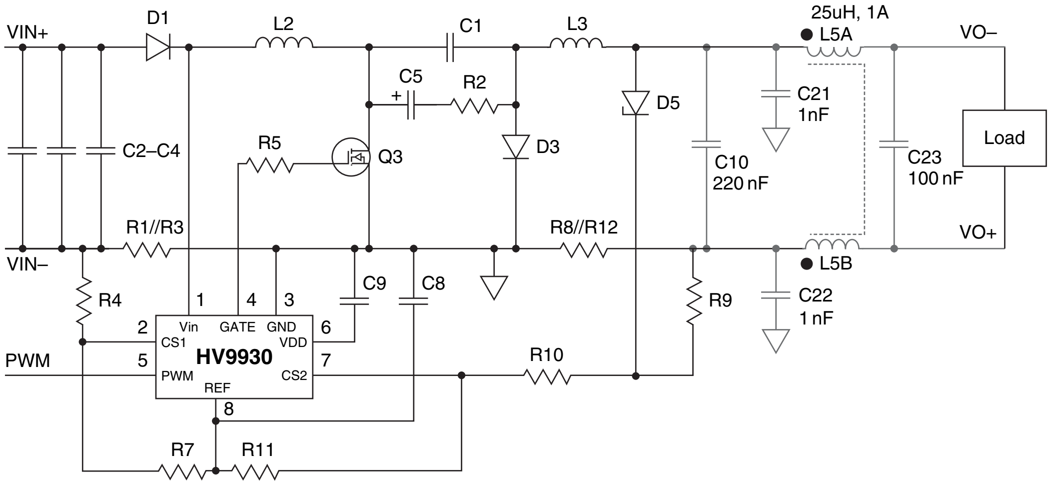


Figure 13.9: Output Filter.

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 (*C*10) 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, *L*5, and a second differential capacitor *C*23. Small value ceramic capacitors *C*21 and *C*21 provide a shunt path to circuit ground for high frequency signals developed across *L*3 and the parallel current sense resistors *R*8 and *R*12.

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.

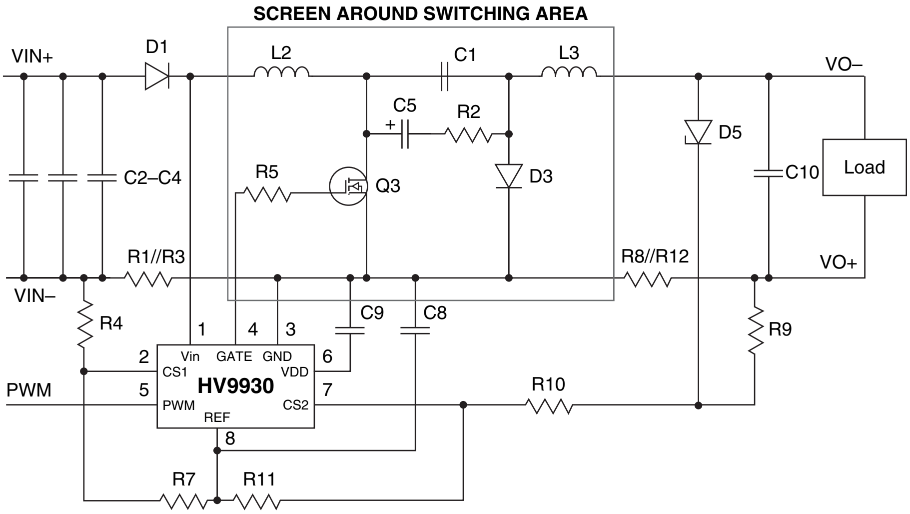


Figure 13.10: Screen.

13.3 EMC Standards

The EMC performance is often automatically assured by the EMI precautions previously described. If radio frequency signals cannot get out of some equipment, they cannot get in either. However, ESD (electro-static discharge) and surge immunity are two areas that are not taken into account in EMI practices.

People generate high electrostatic voltages during normal activities, such as walking across a carpet or opening a plastic envelope. A charged person touching electrical equipment can cause damage or malfunction. Thus equipment must be protected against high voltage discharge. Testing is carried out as specified in IEC/EN 61000-4-2 using an ESD gun. The standard voltage levels are 4 kV for a contact discharge and 8 kV for an air discharge.

Any equipment connected to the AC mains supply must withstand surge pulses, as specified in IEC/EN 61000-4-5. Each surge pulse has an open circuit rise time of 1.2 µs and a fall time of 50 µs. In domestic equipment, the peak surge voltage is 1 kV, which is added to the AC mains supply. In addition, 2 kV surges are applied between the inputs and ground (earth). The test pulses are positive and negative, and are applied at 0, 90, 180 and 270 degree phases of the AC mains voltage.

Another form of surge test is the fast transient burst (FTB), as specified in IEC/EN 61000-4-4. This comprises ±2 kV pulses with a rise time of 5 ns and 50% decay at 50 ns.

These pulses are repeated at a 5 kHz rate (200 microseconds between pulses), for 15 ms. There are 75 pulses in each burst, and the bursts are repeated every 300 ms, for 1 minute. Testing is usually carried out by first applying ±250 V bursts, then ±500 V, then ±1 kV and then finally ±2 kV.

13.4 EMC Practices

Equipment connected to AC mains power lines must be surge tested. The surges are applied, which are added to the normal AC voltage, at times to coincide with different phases of the AC line. The source impedance of the surge test pulse generator is a nominal 50 ohms. The energy in surge pulses can be absorbed or reflected to limit its damaging effects in the equipment under test. Absorbing the energy in surge pulses is the most common method of preventing damage.

A varistor, which is a voltage dependent resistor made from a metal oxide, is commonly used to absorb energy by clamping the voltage. In a varistor rated at 275 V AC, the clamping voltage is typically 710 V, although conduction begins at about 430 V. The amount of energy absorbed in a varistor depends on its physical size. A varistor is usually wire ended and disc shaped; the diameter of the disc is related to the maximum energy (usually given in joules). For example, a 9 mm disc varistor from Epcos that is rated for 275 V AC has a transient energy rating of 21 joules and a peak current rating of 1200 amps.

Another energy absorbing device is a transient voltage suppressor (TVS or TransZorb™). This device is a Zener diode made in silicon and has a stronger clamping action. These are available with either bi-directional or uni-directional breakdown. In AC systems a bi-directional breakdown is required, but in automotive and other DC applications, a uni-directional breakdown is sufficient. TransZorb devices are usually rated in peak power (watts); 600 W and 1500 W devices are commonly available.

The oldest technology, and still sometimes used, is the gas discharge tube (GDT). This has a glass tube filled with inert gas and metal electrodes at either end. When the voltage across the electrodes is high enough, the gas ionizes and conducts to clamp the voltage.

A plastic film capacitor is often connected across the AC line (typically 100 nF, 275 V AC X2 rated). This not only helps to reduce EMI emissions and susceptibility, it also helps to absorb some of the energy in surge pulses. Surge suppressors take some time to respond to impulse voltages; so fast transients can sometimes pass with little loss and can cause damage.

Many systems have a large electrolytic capacitor across the power rails, after a bridge rectifier. This capacitor will absorb surge energy; however, electrolytic capacitor construction results in some inductance that will have high reactance to fast rising surge pulses. A plastic film capacitor connected in parallel with the electrolytic capacitor will help to absorb high frequency energy. A clamping device such as a varistor directly across the AC line is still a good idea because it limits the surge before it reaches the bridge rectifier.

A fuse should be fitted to every piece of equipment powered from the AC mains supply. This provides a means of limiting energy into an anti-surge device, like a varistor. When a high-energy surge causes the varistor to break down, the fuse will blow. Some people fit a high power wire-wound resistor between the AC line and the varistor, to limit the current from surges and prevent burnout of the varistor.

When laying out a PCB, the spacing between tracks should be carefully considered. The breakdown voltage of an air gap is about 1 kV per mm, so at the potentially high voltage input of a power supply, sufficient gap should be allowed. An air gap of 3.2 mm is the minimum to prevent breakdown and a potential fire hazard. On a PCB the gap between conductors is known as the creepage distance. The air gap from a live part of the circuit to any other parts of the enclosure is known as the clearance distance.

Integrated circuits that can be powered directly from the rectified AC supply usually have ‘no connect’ or NC pins adjacent to the high voltage pin. This is designed to give a suitable creepage distance. Where no gap exists, a slot can be cut in the PCB, or the contact pins can be coated with a conformal coating or a resin to increase the insulation.

CHAPTER 14

Thermal Considerations

14.1 Efficiency and Power Loss

People sometime refer to LEDs as being a cold light source. This is true in the sense that an element is not heated to thousands of degrees Celsius in order to produce light. However, LEDs do indeed generate heat and this has been the cause of failure of several designs. As a first approximation, the heat generated is voltage drop multiplied by current flow. A white LED with a 3.5 V drop at 350 mA will produce about 1.225 W of heat. Actually the emission of photons (light) will reduce this power a little, but it is better to design a larger heatsink to be on the safe side.

Power LEDs should always be mounted on a heatsink. For example, a traffic light using six or seven 1 W LEDs could be mounted alongside the driver electronics on a 6 inch diameter circular PCB. A heatsink could be mounted on the backside of the PCB for removing heat from both the LEDs and the driver circuit. Since traffic lights may have to work in high ambient temperatures, a good thermal conductivity is required; electrolytic capacitors in the driver circuit should be avoided in this case, for long-term reliability.

When designing analog or switching power sources, we discuss efficiency. This is the ratio power out/power in, and is usually expressed as a percentage. What designers sometimes overlook is that input power minus output power equals power loss in the LED driver circuit; see Figure 14.1. Loss in the driver must be dissipated as heat. A switching LED driver with 90% efficiency, driving a 10 W load will require an input power of 10 W/0.8 = 11.1 W. This means that 1.1 W is power loss and must be dissipated in the LED driver.

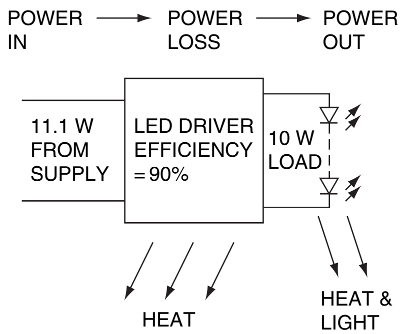


Figure 14.1: Power Loss in Driver Circuit.

14.2 Calculating Temperature

The temperature of a device can be calculated using simple 'Ohm's law' type mathematics. Temperature can be seen as being equivalent to a voltage. Thermal resistance can be equated to electrical resistance. Heat flow (watts) can be regarded as the equivalent to electrical current, see Figure 14.2.

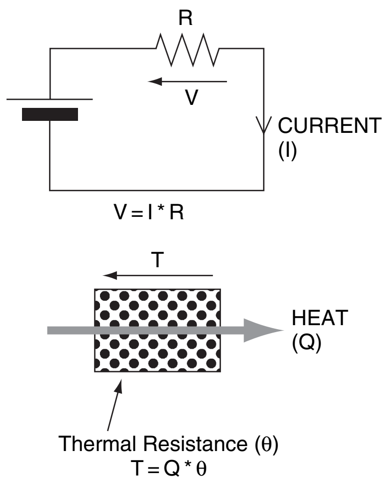


Figure 14.2: Electrical Equivalent Calculations.

Like electrical resistance, thermal resistances can be added when connected in series (see Figure 14.3). Consider a TO-220 package mounted onto an aluminum heatsink. The thermal resistance between the silicon die and the package, added to the thermal

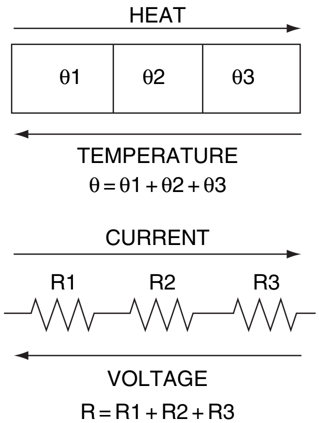


Figure 14.3: Thermal Resistances in Series.

resistance of the package to heatsink 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 θ. (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 θ_{JC}. For example, let $\theta_{JC} = 1.2$ K/W, $\theta_{CH} = 0.1$ K/W, and $\theta_{HA} = 2.4$ K/W (I have given my own notation H = heatsink), so the case to heatsink resistance is 0.1 K/W. When the device is dissipating 10 W, with a total thermal resistance of 3.7 K/W, the silicon junction temperature will be 37 degree hotter than the ambient temperature. If the ambient temperature is 25°C, the silicon junction will be raised to 62°C.

Like electrical resistance, having parallel thermal resistance paths reduces the overall resistance (see Figure 14.4). Two paths, each of 2 K/W, will create an effect single path of 1 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

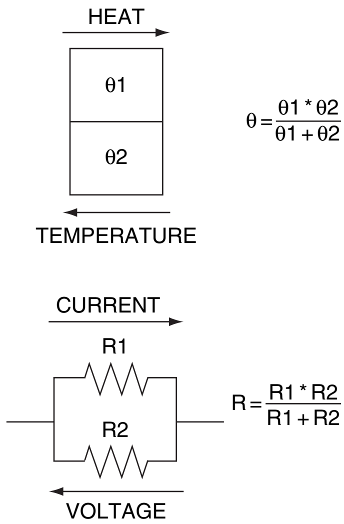


Figure 14.4: Thermal Resistances in Parallel.

dissipate a lot of heat. For this reason, a small driver is rarely able to drive a high power load; remember this when the marketing department asks you to design a smaller driver!

Semiconductor component manufacturers usually specify the minimum and maximum junction operating temperatures for their devices. It is usual for the temperature range to be −40°C to +125°C, but this is not the ambient temperature. Commercial device ambient temperature ratings are 0°C to 70°C, industrial device ratings are −40°C to +85°C. Military and automotive devices are rated for ambient temperatures of −55°C to +125°C, but these require special processing of the silicon material and packaging to achieve a +150°C junction operating temperature and hence are usually more expensive.

Component manufacturers also specify power dissipation (usually based on 25°C ambient temperature). Most manufacturers also provide thermal resistance information in their datasheets and some provide notes on heatsink requirements, which can be very useful to designers.

14.3 Handling Heat – Cooling Techniques

Heat must be dissipated somehow. If there is high thermal resistance from the source, the source temperature will rise until sufficient heat is dissipated (or until components are destroyed). High temperatures will reduce the reliability of components, so temperatures should be reduced somehow. One obvious cooling technique is to reduce the thermal resistance, and thus dissipate heat easier, by using a heatsink. This is fine if the heat is all generated in one place (like in a MOSFET or a voltage regulator).

Surface mounted power MOSFETs are usually in a D-PAK or D2-PAK housing, which have a tab for dissipating heat. However, this means that the tab must be soldered to a copper surface area on the component side of the PCB, or otherwise a surface mount heatsink is required, see Figure 14.5. A surface area of one square inch (25 mm × 25 mm) on a standard FR4 fiberglass board can give a thermal resistance of $\theta_{JA} = 30\,K/A$ with a D-PAK device. Surface mount heatsinks are sometimes made from tinned brass and are soldered to the PCB, either side of the MOSFET body. These reduce the thermal resistance to about $\theta_{JA} = 15\,K/A$.

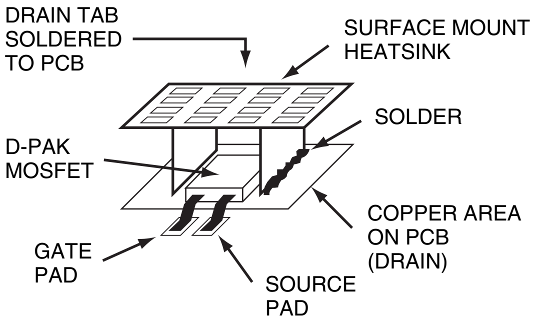


Figure 14.5: Surface Mount Heatsink.

Through-hole MOSFETs in a TO-220 package can be fitted to various heatsinks with a wide range of sizes. A small heatsink can be fitted, supported by the MOSFET pins or bolted onto the PCB through the TO-220 tab. Larger heatsinks could increase parasitic capacitance and cause an increase in switching losses, but this can be prevented if the heatsink is connected to the ground plane. Grounding the heatsink also prevents undue EMI radiation, but the MOSFET should be electrically isolated from the heatsink using a thermal conducting pad (made from a flexible material to give a large surface contact). Switching losses will be due to the capacitance between the MOSFET drain (tab) and the heatsink.

Even where electrical isolation between the MOSFET and heatsink is not required, a thermally conductive pad or paste is a very good idea. This is because the surface of the MOSFET tab and the heatsink surface are not smooth. Without the thermally conductive pad or paste, the actual area where good contact can be made is just a small fraction of the surface available. Micro-cavities between the two surfaces create air pockets that have high thermal resistance, see Figure 14.6. The thermally conductive pad or paste fills these cavities to create a uniform surface with low thermal resistance.

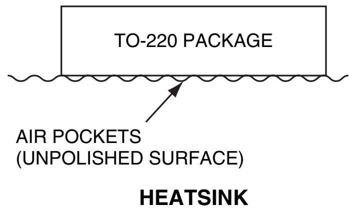


Figure 14.6: Thermal Resistance Created by Air Pockets.

When several devices on a circuit board generate the heat, a solution could be to use a fan to blow air across the circuit board. Cooler air from outside the equipment can be blown over warm components to reduce their temperature. Airflow will reduce the effective thermal resistance of the air interface.

Careful placing of cooling fans can make a big difference to the performance. Large objects like electrolytic capacitors will tend to block the flow and may steer the cooling air away from areas of the PCB. If the air flows in the direction of heatsink fins, it will be more effective. Air flowing across the fins will only cool the front and rear fins, see Figure 14.7.

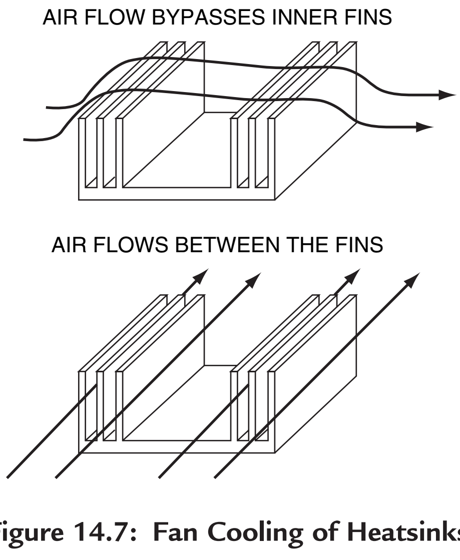


Figure 14.7: Fan Cooling of Heatsinks.

If mounting a fan at the top of equipment make sure that the air flows upwards, with the fan blowing air outwards, so that the fan aids the natural buoyancy of the hot air. Fans mounted in the side of equipment are much more effective if two fans are used, one on either side of the enclosure. In a wide enclosure, both fans could be mounted on the rear panel; one fan should blow in and the other should blow out so that air circulates around the components inside.

Fans do have a reliability issue, so consider adding a fail-safe mechanism in case the fan fails to operate. A fail-safe mechanism should monitor the temperature of sensitive components on the circuit board. Driving the LEDs at a lower power or turning them off when the temperature rises too high may be a solution.

CHAPTER 15

Safety Issues

This chapter discusses electrical safety and readers are advised to obtain the latest requirements from their regulatory body or safety consultant. The information here is to show that many topics must be considered, rather than as a reference for design work. Optical safety is a concern, but it is outside the scope of this book and readers are advised to consult technical data supplied by LED manufacturers.

15.1 AC Mains Isolation

Safety isolation can only be achieved with a transformer. This transformer can be placed on the AC mains supply, or as part of the switching regulator circuit. Transformer isolation on the AC mains supply is bulky because the AC signal is operating at 50 Hz or 60 Hz.

Conversely, a transformer that isolates the *output* of a switching regulator can be very small because it is operating at the switching regulator frequency of typically 50 kHz or more. If accurate current control is needed, additional electronics to control the LED current is needed and some form of isolated feedback is required.

For products connected to AC mains supplies, 1500 V RMS (50 Hz or 60 Hz) isolation is usually required. Products for medical applications usually require a higher isolation voltage; LED lamps are sometimes found in operating theatres and in other medical applications.

15.2 Circuit Breakers

In the event of an over-current, the most common circuit breaker is a fuse. This is basically a piece of wire that is heated by the current flow. The wire eventually melts, thus breaking the circuit. Sometimes two wires are joined by solder, one wire being a weak spring; as the heat softens the solder joint sufficiently the spring wire pulls the joint apart. Fuses tend to be slow and are rated so that a current twice the normal load may be needed to blow the fuse.

Electronic circuit breakers are also available. These usually latch in the off state when a fault is detected, so cycling the power supply off and on again is generally required.

Tyco produce a fuse that becomes high impedance when an over-current is detected, due to the current's heating effect, but then remake the electrical connection once the fuse has cooled down.

15.3 Creepage Distance

In most electrical circuits connected to the AC mains supply, creepage distance is a concern. The concern is two-fold: electrocution or fire – for example, a loose piece of solder could short out a pin carrying high voltage to another low voltage point in the circuit; or moisture and dust could bridge the gap and allow a current to flow. In either example, the current may not be high enough to blow the fuse, but could be lethal to the user through electrocution or toxic smoke inhalation.

The requirements for creepage distance depend upon the application. Some devices have 'no connect' (NC) pins between high and low voltage pins, so that a small piece of solder cannot bridge any two points. I have seen customers cut a slot in their circuit board to allow them to reduce the overall circuit board size. The creepage distance in air is much less than the creepage distance on a PCB. One way around this is to apply conformal coating, usually a silicone-based elastomer or a polyurethane varnish.

15.4 Capacitor Ratings

Capacitors connected across the AC line must be 'X-rated', usually X2. These tend to be more expensive than standard capacitors, because they are rated to withstand voltage surges. The typical DC operating voltage of X2 capacitors is 760 V, whereas the maximum DC level normally expected from a rectified 265 V AC supply is 375 V. Polyester or polypropylene (MKP) is the usual dielectric in X2 capacitors.

Capacitors connected from the AC line to earth must be 'Y-rated', usually Y2. The typical DC operating voltage of Y2 capacitors is 1500 V. These capacitors normally have low capacitance (say 2.2 nF) and are usually made with a ceramic or polypropylene dielectric.

After the bridge rectifier, standard capacitors, rated at 400 V or 450 V are used. Since these are not rated for operation above their nominal working voltage, they are often smaller and lower cost compared to X2 capacitors. For this reason some engineers will place the EMI filter after the bridge rectifier. An EMI filter before the bridge rectifier will tend to prevent voltage surges from reaching more sensitive components and is thus preferred. It will also reduce transients from the bridge rectifier, caused by sudden changes in current flow as capacitors on the DC side of the bridge are charged.

15.5 Low Voltage Operation

The UL1310 Class 2 regulations and the European EN60950 safety standard (also known as IEC 60950) are generally applicable to any electronic circuit. The EN60950 standard was originally intended for information technology equipment (i.e. computers and associated hardware) but, since it is one of the few 'harmonized' standards that have been agreed by all of Europe and many other countries in the world, it has been used as a reference for most safety regulations. If equipment complies with EN60950, it is deemed that due diligence has been performed in legal cases of alleged neglect.

The European Low Voltage Directive (LVD) is a safety regulation in Europe that covers all products operating from voltages of 50–1000 V AC and 75–1500 V DC. There is a further 'catch-all' General Product Safety Directive. These directives require a CE mark to be placed on all goods offered for sale. But to get permission to use the CE mark they must comply with safety standards like EN60950. Note that sub-modules do not require CE marking, but the overall equipment does. Clearly sub-modules ought to be safe to operate and not so high in EMI that the final equipment cannot easily pass testing, otherwise the equipment assembler may decide to look elsewhere for his sub-modules!

To ease the burden in safety testing, many people ensure that their products operate at low voltage. The SELV (safety extra low voltage) requirements are that no touchable conducting parts have a voltage (relative to ground, or across any two points) above 60 V DC, or 42.4 V peak/30 V RMS AC. For example, a DC powered (boost-buck) Cuk converter, with 24 V DC input must not have an output above 36 V. This is because the Cuk produces an inverted output, so the difference between the input and the output is the two voltages added together.

An AC mains powered LED lamp must be isolated to meet these regulations – in addition to the output voltage being limited to 60 V if the electrical connections are ‘touchable’. If the equipment has an isolated cover, this is not enough to ignore the voltage limit since the user could remove the cover. However, the voltage limit can be ignored if the cover has a micro-switch to disable the equipment in the event of the cover being removed. A double-fault (cover broken or removed *and* micro-switch broken or disabled) has to occur before the user can touch a potentially lethal voltage.

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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.