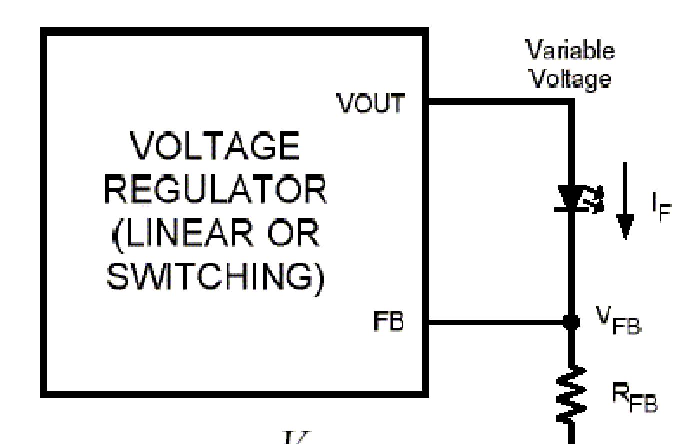
Driving high brightness LEDs with switching regulators

By Chris Richardson,
Applications Engineer,
Power Management Division
National Semiconductor Corporation

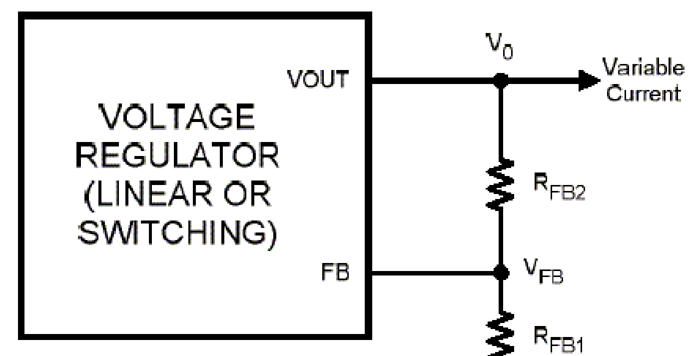
The first LEDs produced in the 1960's were feeble infrared devices that suffered from short lifetimes and poor efficiency. The semiconductor processes used to make LEDs expanded from early silicon carbide (SiC) and gallium arsenic phosphide (GaAsP) devices to gallium phosphide (GaP) and indium gallium phosphide (InGaP) which produced red and later orange, yellow, and greenish-yellow. LEDs at that point were bright enough to be indicators, and even at this early stage they had better luminous efficiency than incandescent light bulbs. Their overall light output was still much too low and their colour range too narrow for other uses, however. The concept of general illumination, the lighting of cars and buildings, was limited both by the lumens output and by the lack of a blue LED to complete the RGB spectrum and make white light. Research into SiC led to blue LEDs with luminous flux that was far too weak to be useful. Still, as the number of processes increased, so did the luminous flux, or brightness, the colour, and the power dissipation. Gallium aluminium arsenide (GaAlAs) and indium aluminium gallium phosphide (InGaAlAs) LEDs evolved into the first "super bright" LEDs. Nichia Corporation in Japan took early but important steps towards the ideal of general illumination with semiconductor lighting in the early 1990's. Research done by Dr. Shuji Nakamura[1] led to the first commercial blue LEDs. Nichia also pioneered efforts to produce a white LED by adding a phosphor to blue LEDs, which emits yellow light. Although the white LEDs cast a cool white, they were quickly adopted in clusters as replacements for low light incandescent applications.

LEDs designed for true illumination emerged in the early 1990's, with devices made by Nichia, Osram Opto-Semiconductor, and Lumileds. These devices were rated in lumens, common to light fixtures, as opposed to candela, which is normally used for single point light sources. The new High Brightness LEDs (HBLEDs) were also packaged like power



$$I_F = \frac{V_{FB}}{R_{FB}}$$

Figure 1b: Constant current regulator



$$V_O = V_{FB}\frac{R_{FB1}+R_{FB2}}{R_{FB1}}$$

Figure 1a: Constant voltage regulator

semiconductors, using surface mount technology and thermal pads. Standard LEDs encapsulated in epoxy suffer from poor junction to ambient thermal resistance and from loss of light due to the gradual yellowing of the material. The new HBLEDs replaced the epoxy with long lasting silicone-based materials.

Single-die, white HBLEDs are available today that deliver 30-40 lumens at a dissipation of 1W. RGB LEDs with three or more HBLED dice placed in a single power package deliver as much as 200 lumens. HBLED designs are used or are under development in automotive, industrial, and commercial lighting, as well as in backlighting for LCD monitors and televisions, as the actual pixels in outdoor/stadium video screens, and in optical communication.

Constant current sources

Regardless of type, colour, size or power, all LEDs work best when driven by a constant current source. Light output, measured in lumens, is proportional to current, and hence LED manufacturers specify the characteristics (such as lumens, beam pattern, colour) of their devices at a specified forward current, IF, not at a specific forward voltage, VF. LEDs are PN junction devices with a steep I-V curve, hence driving an LED with a voltage source can lead to large swings of forward current in response to even the smallest changes in voltage.

Most power supply ICs are designed to provide constant voltage outputs over a range of currents, (Figure 1a) and it is not always straightforward to adapt a voltage regulator to provide constant current. With an array of more than one LED, the main challenge is to match the drive currents through each LED. Placing all the LEDs in a series string is a common way to ensure that exactly the same current flows through each device.

Heat and light

The luminous efficiency of HBLEDs has already eclipsed the incandescent light bulb. A typical 1W white LED delivers an optical efficiency of 30 lm/W, whereas a typical 60W light bulb delivers 15 lm/W. For the current generation of LED products, retrofitting a system designed for standard light bulbs or the ability to make the LED and driver a "drop-in" replacement is critical. Perhaps the biggest technical challenge is to create a complete LED solution (driver, optics, heat sink) that can withstand the high temperatures

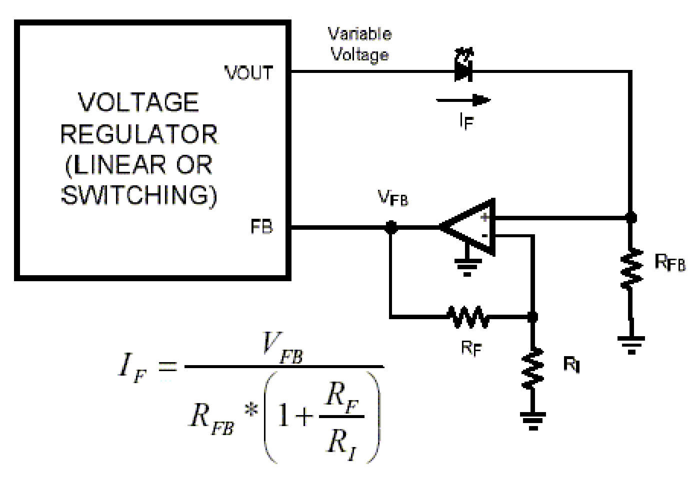


Figure 2a: Low side sense with amplification

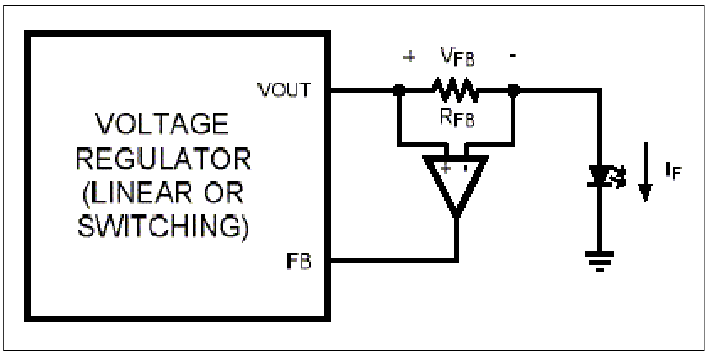


Figure 2b: Generic high side sense

and confined spaces formerly occupied by light bulbs. Where an incandescent bulb can operate at temperatures as high as 200°C, semiconductor LED drivers are characterized to a maximum of 125°C. Ratings of 150°C to 175°C may be available, but come at a premium in price. Dissipating a specific amount of power in a confined space results in a fixed temperature rise, and if the LED and driver must fit into the same space as a light bulb, yet maintain a significantly lower ambient temperature, then the LED and driver must be more efficient.

Why use a switcher?

A switching regulator is the best choice for driving an HBLED when high efficiency and low power dissipation are required in a constant current source. When input voltage is lower than output voltage a switcher must always be used, and when input voltage is much greater than output voltage the advantages of switching regulators over linear regulators becomes apparent.

At 350 mA, a typical forward current for 1W HBLEDs, a constant current source could be built with an IC-based linear regulator or even a discrete linear regulator, however the high forward current coupled with the wide differences between the input voltage and the LED forward voltage place too great a demand for dissipation on the pass device. Consider, for example, a linear regulator in a TO-220 package driving a single 1W HBLED in a passenger car. This application is one of the 'early adopters' of HBLEDs for map lights and general white-light interior illumination. The circuit must operate correctly over a range of 9V to 16V, but must also be able to survive for 2 minutes with an input of 28V in the case that a jump-start is applied from the double battery system common on tow-trucks. Output voltage is approximately 3.5V, hence the maximum steady-state pass-transistor voltage is 11.5V. At a continuous current of 350 mA the pass transistor must dissipate approximately 4W. Assuming a typical junction-to ambient thermal resistance of 53°C/W and multiplying by the power dissipation yields a die temperature rise of over 200°C. Even at an ambient temperature of 25°C the linear regulator would require a large, expensive heat sink. Space is limited in this application, and the circuit must operate at ambient temperatures as high as 85°C. Making matters worse, the LED itself often shares a metal core PCB (MCPCB) with the driving circuitry.

The operating characteristics and especially the lifetime of the LED depend upon keeping the junction temperature below a specified value, typically 90°C to 135°C. A switching regulator offers the advantage of efficiency, reducing dissipation in the semiconductor that drives the LED as well as reducing the heat transferred to the LED itself.

High side and low side current sensing

Sensing relatively high current is not new to the world of switching regulators, but sensing current with the accuracy required in an HBLED drive is not found in products designed for controlled output voltages. In general, to meet the needs of a driver for an HBLED, the current output tolerance must be in the ±5% to ±15% range. This stands in contrast to the tolerance of existing current sense functions like current limit or peak current sensing found in voltage regulators. A series resistor for current sensing becomes mandatory when the tolerance of output current is on par with the tolerance of output voltage.

The challenge of current sensing is to make it accurate and efficient, two goals that are in direct conflict. Referring to Figure 1b, the higher the sense voltage, VFB, the higher the signal-to-noise ratio, but the higher the power lost in RFB. The lower the sense voltage the more power saved, but this comes at the expense of noise sensitivity and wider tolerance in the sense voltage itself. Silicon bandgaps are the logical choice for references in switching constant current sources, but their voltage of approximately 1.2V means that at 350 mA RFB would dissipate 0.42W. Current HBLEDs are driven at 700 mA and 1A, requiring resistors rated to 1W and above. Single-die HBLEDs are projected to run with forwards currents in excess of 2A. High power dissipation combined with high accuracy and low tempco are features that make for expensive resistors. A lower sense voltage must be obtained. Whether the reference voltage is divided down or the current sense signal is amplified, additional error will be introduced, and the signal-to noise ratio reduced.

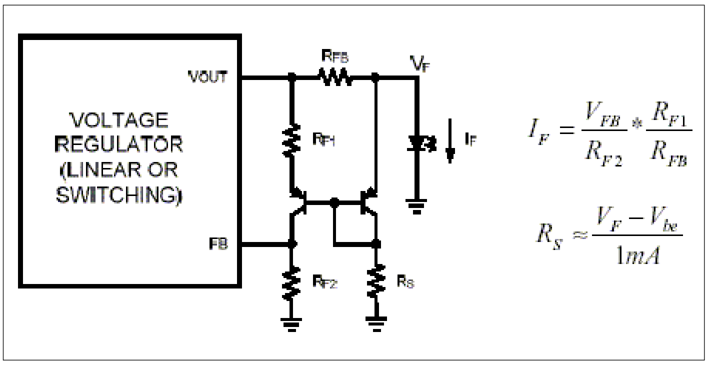


Figure 3: High side sense with PNP current mirror

The circuit of Figure 1b shows the simple method of controlling current with a voltage regulator. This method, low side sensing, requires that the sensing resistor RFB dissipate a significant amount of power and also depends upon the tolerance of RFB to set the accuracy of IF. Power resistors tend to be large and expensive, and as noted in the first section, heat generated in RFB can transfer to the LED itself. Figure 2a shows a simple method for amplifying the cur-

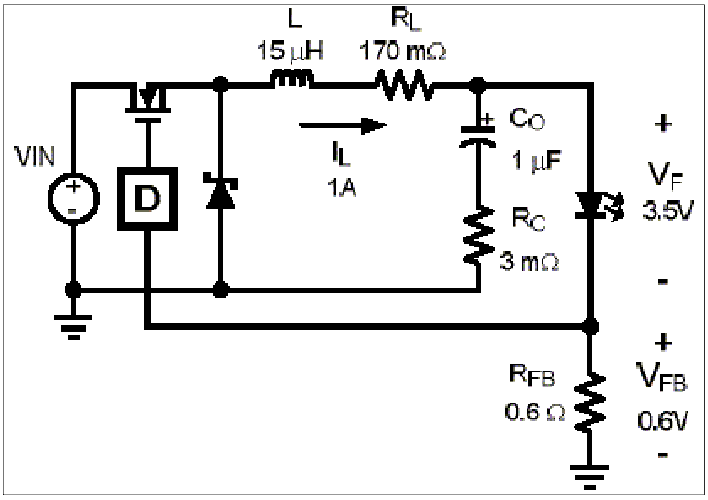


Figure 4a: Buck regulator with output capacitance

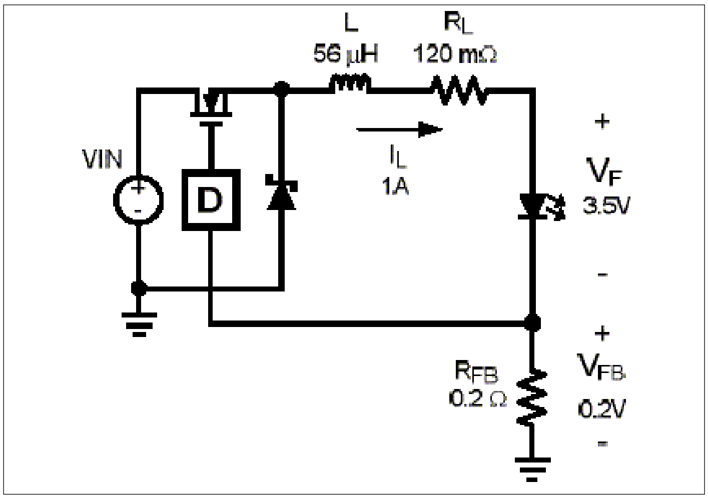


Figure 4b: Buck regulator without output capacitance

rent sense signal using a single-supply op-amp. This method reduces thermal stress on RFB and allows the user to select RFB based on the desired dissipation and current sense SNR while setting the average value of IF with the gain of the op-amp.

The current sensing resistor can also be placed between the converter output and the anode of the LED array, a system analogous to the high side sensing used in many peak current mode buck converters. Shown in Figure 2b, the main advantage of this topology is the ability to connect the cathode of the last LED to ground. For systems that drive an LED array remotely but still distribute system ground, such as an automobile, this saves a connection and makes the system simpler and less expensive. High side sensing requires a differential amplifier of some kind. An IC op-amp can be used, but it must combine high CMRR with low input offset. Low cost differential sensing can be accomplished with discrete transistors as well, and an example using two matched PNP transistors appears in Figure 3. Figure 2a: Low side sense with amplification

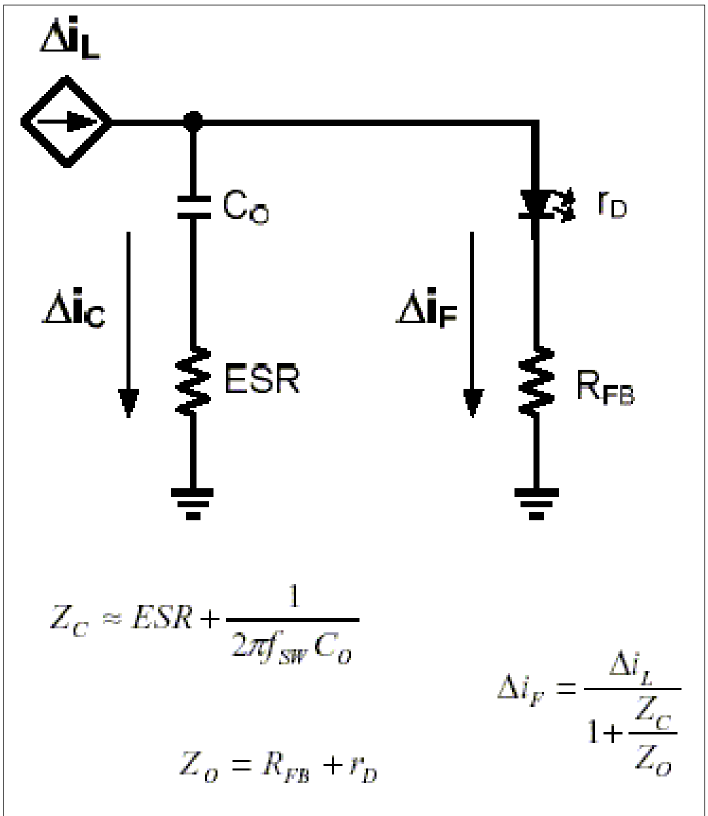


Figure 5: Output Impedances and Ripple Calculations

The buck regulator is best

The standard buck converter, shown in Figure 4a, is the ideal choice for driving a constant current because of the output inductor. Inductor current ripple, ΔiL, is a known, controlled quantity in buck converter designs. Only the buck converter has an average inductor current that is equal to the average load current, or IF for LED drivers. Regardless of the control method, the fact that output current cannot change instantaneously during any part of the switching cycle makes the conversion of a constant-voltage source into a constant current source much easier. With attention to detail, many buck converter-based constant current circuits can run without any output capacitor.

As with voltage regulator circuits, the design of a current source using a switching regulator has two main criteria: the accuracy of the average level of output current, and the ripple, or AC current (ΔiF) that rides along with the DC portion. The tolerance of the average output current depends on the accuracy of the current sensing circuitry and that of the reference. LED ripple current depends on input voltage, output voltage, switching frequency, and the inductance itself.

While it is possible to control output current in a circuit that runs in Discontinuous Conduction Mode (DCM), the buck converter is well characterized and understood when it functions in Continuous Conduction Mode, which maintains a positive current through the inductor as all times. Since the inductor shares the same DC current as the LED array, and keeping current through the LEDs constant is the prime goal, CCM operation is the preferred way to run a buck converter

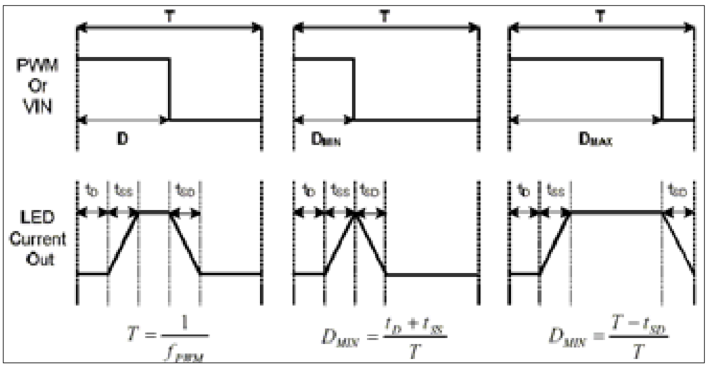


Figure 6: Dimming Response of IC Enable Pin

	With Co	Without Co
ΔiL(%)	40	10
ΔiLED(%)	10	10
L calc (μH)	13.4	53.4
L actual (μH)	15	56
Isat	1.5	1.7
Inductor Dimension (mm)	8.9 x 6.1 x 5.0	12.3 x 12.3 x 6.0

Table 1Ripple and inductance tradeoffs

that drives a constant current. The buck converter operates without a right-half-plane zero in CCM and DCM, giving it another advantage over boost, flyback, SEPIC, or Cuk converters. A buck regulator-based current source in CCM can also function without an output capacitor, an advantage detailed below, in the section Running Without an Output Capacitor.

Running without an output capacitor

Almost any voltage regulator with an adjustable output can be modified to control output current using the method of Figure 1b. The buck regulator is unique because the output capacitor is not needed to provide current to the load during the power switch ON time or OFF time. It filters a portion of the AC current ripple created by the output inductor and the switching action, and provides a well of charge to supply the load during fast transients. Constant current LED drivers are free of load transients by design, and the output capacitor is needed only if the design requires a lower amplitude current ripple through the LED array. Figure 4 shows two versions of an LED drive circuit designed to drive 1A at 3.5V from a 12V input at 500 kHz. Both circuits control Δ iF to 100 mAP-P, or 10% of the average current. Inductor ripple current (ΔiL) in a fixed frequency buck converter can be estimated with two familiar expressions found in buck converter IC datasheets:

$$\Delta i_L = \frac{V_{IN}-V_O}{L * f_{SW}} * D,\quad D = \frac{V_O+V_D}{V_{IN}+V_D},\ V_D$$

Schottky Diode Forward Voltage equation

For constant current circuits like the two shown in Figure 4, output voltage can be estimated as VO = VFB + VF. To calculate the ripple current that flows through the LED and RFB in Figure 4a the impedance of both LED and output capacitor branches are estimated and shown in Figure 5. This circuit can use lower inductance while relying upon the low AC impedance of CO to filter the majority of the ripple current.

Care must be taken in determining LED dynamic resistance, rD, as dividing forward voltage by forward current will give a value that is 5x to 10x the true rD. This parameter is provided in some manufacturer's datasheets, and can also be estimated by taking the inverse slope of forward current vs. forward voltage. Dynamic resistance can also be measured with a network analyzer; however this method is time consuming and requires expensive equipment. The LED in the circuits of Figure 4 is a Lumileds Luxeon III, white, with a typical rD of 0.8Ω Measurement of rD with a network analyzer in the actual test circuit yielded an rD of 0.65., and this value was used for ripple current calculations.

Based on the equations above, to filter 30% of the inductor current out of the LED an output capacitor of 770 nF is required. This calculation assumes that the capacitor used will be ceramic, that the ESR and ESL can be ignored, and that the triangular shape of the ripple current is close enough to a sinusoid to make the reactance calculation of ZC meaningful. A 1 ìF ceramic capacitor with X7R dielectric and 6.3V or 10V rating provides more than enough ripple filtering in a small package at low cost.

The circuit of Figure 4b saves a component and increases the output impedance of the driver, but depends entirely upon inductance to control the ripple through the LED, as ΔiL = ΔiF. Table 1 shows a comparison of ripple and inductance tradeoffs between these two circuit topologies.

Other converters such as the boost, flyback, and SEPIC rely on the output capacitor to maintain the output voltage during a portion of their switching cycle. These regulators cannot be used without at least some capacitance at their outputs. This makes them more difficult to dim via PWM, a topic which is detailed in the following section.

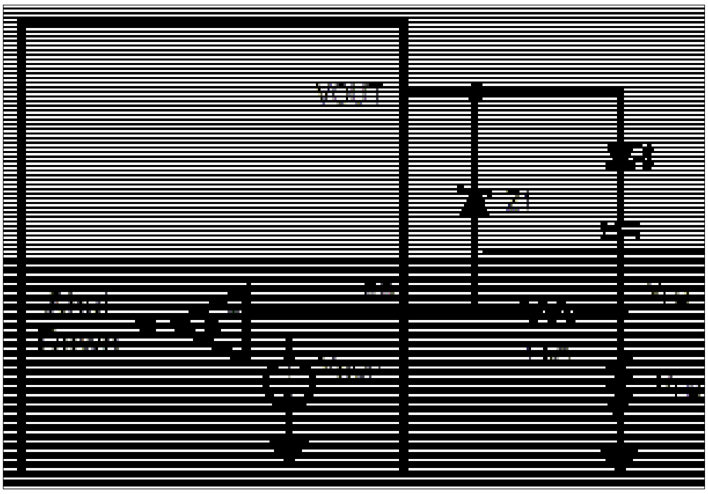


Figure 7a: Internal OVP with Zener Diode

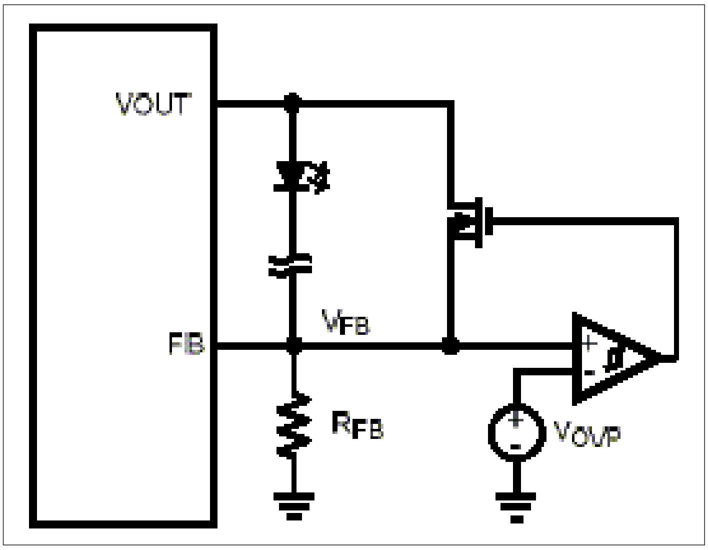


Figure 7b: External OVP with MOSFET

Methods for PWM dimming

Dimming with PWM is the accepted standard for reduction of light output in LED lighting systems. Light output from LEDs changes in a linear fashion as IF is varied, however the dominant frequency of light emitted also shifts. Linear dimming is used in applications where this change in colour is judged to be acceptable. Examples of these applications could be flashlights, reading lights, and systems with few LEDs, where a disparity in colour or brightness from light to light is not noticeable. For applications such as automotive brake lights, LCD backlighting, and direct view RGB, brightness and colour requirements are too strict for this approach. To reduce light output in these applications while maintaining tight control over colour, a known and controlled current

Ambient Temp (°C)	30	35	40	45	50	55	60	65	70	75	80	85	90	95	100
LED Current (mA)	924	924	922	922	920	920	916	872	830	790	744	700	658	618	568

Table 2. Lab-tested method for linear de-rating of LED current

must be run through the LED array string and then chopped with a controlled duty cycle.

Many switching regulators have an enable pin that can accept a PWM dimming signal. Few ICs, however, were designed with constant toggling of this pin in mind. They perform a power-on reset and soft-start when they are enabled, and when disabled the control ICs shut down as many internal circuits as possible to save quiescent current. The delay from soft-start and from the time needed for power-on reset places a limit on both the maximum frequency of PWM dimming and on the range of duty cycle that can be used. Soft-start and soft-shutdown circuits also introduce error into the average value of light output by slowing the slew rate of the LED current, making the pulses less than ideal.

Another method for PWM of the output of a current source is to chop the power input voltage, also know as "theatre dimming" and commonly used in passenger vehicles. Generally this happens at low frequency, in the range of 50 Hz to 1000 Hz. Both this approach and the PWM of a regulator's enable pin must be analyzed carefully in order to prevent interaction between the PWM frequency and switching frequency. PWM of input voltage or an enable pin can cause a switcher to oscillate. In general, making sure that the switching frequency is at least two orders of magnitude above the PWM frequency will eliminate interaction.

Figure 6 shows the exaggerated response of a switching regulator to a PWM signal applied at the enable pin or a PWM of the power input voltage. The quantities tD, tSS, and tSD represent the start-up delay, rise time (soft-start time), and fall time (shutdown time), respectively. For example, a buck regulator with hysteretic control might have tD + tSS = tSD = 40 ìs in order to drive a single white LED from 0 mA to 350 mA and back. Following the calculations in Figure 6, at a dimming frequency fPWM of 100 Hz this regulator could respond to a duty cycle range of 0.4% to 99.6%. If the PWM dimming frequency were increased to 1000 Hz, the range of duty cycle would become 4% to 96%.

HBLEDs have enabled a range of new applications that require high precision dimming or pulsing in a frequency range that can equal or exceed the switching frequency of the driver. Optical communication and infrared scanning for obstacle detection are examples of circuits that must pulse in the 10 to 20 MHz range. Next generation LCD televisions are projected to dim at around 1000 Hz but with a duty cycle resolution of 12 bits, requiring minimum duty cycles of 0.025% and minimum on-times of 25 ìs. For these circuits a power switch in parallel with the LED is needed. This arrangement maintains a constant current flow from the converter, eliminating the need to turn the control loop on and off. A typical N-channel enhancement mode MOSFET capable of carrying 2A continuously can be switched on and off in 10 ns, permitting a wide dimming ratio even at frequencies in the megahertz range. The dimming frequency is limited by the MOSFET switching time and the ability of the control loop to respond to output voltage transients.

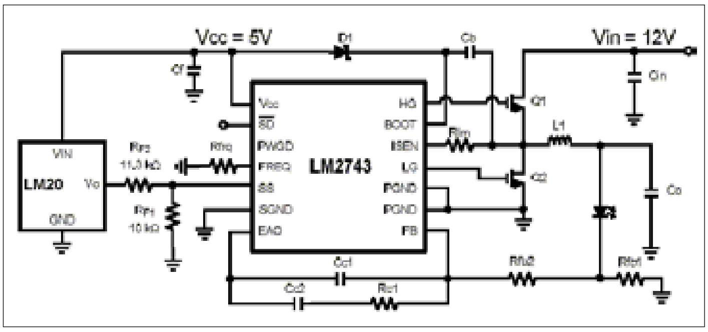


Figure 8: Temperature De-rating of LED Current with LM2743 and LM20

Dealing with output open circuit conditions

Switching regulators configured as constant current sources have no problem dealing with short-circuits across the LED array, but will drive their output voltage to the system limits if the LED fails as an open circuit. Open circuit failure is much more common in HBLEDs and comes from the ten-

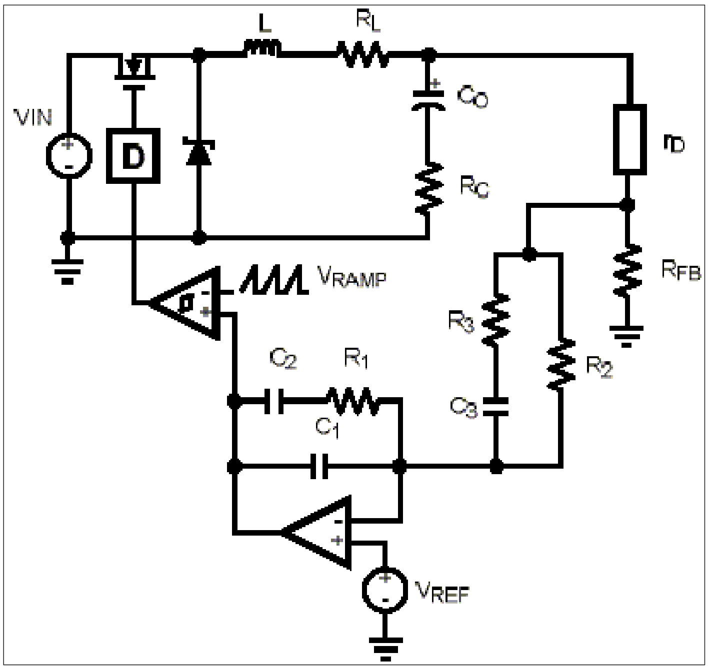


Figure 9a: Voltage Mode, Type III

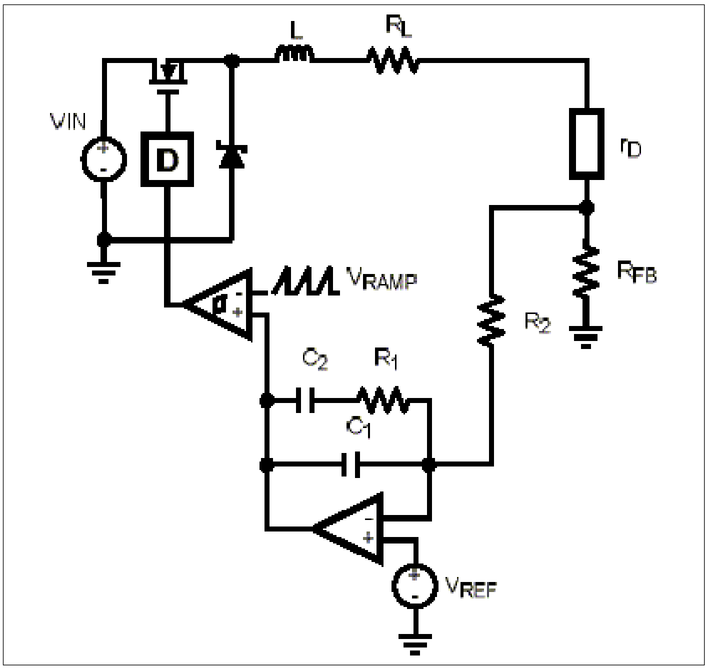


Figure 9b: No Output Capacitor, Type II

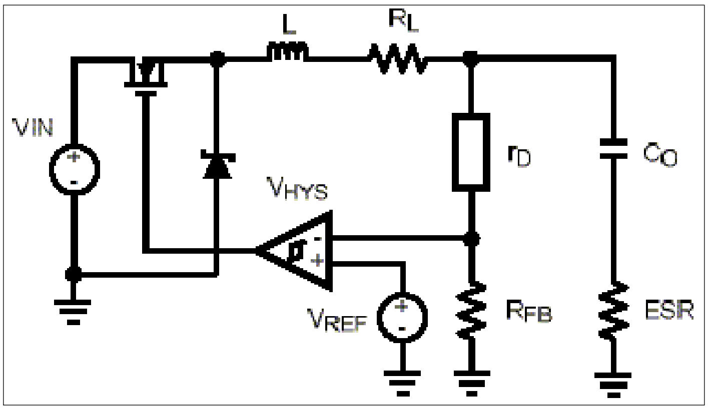


Figure 10a: Additional ESR

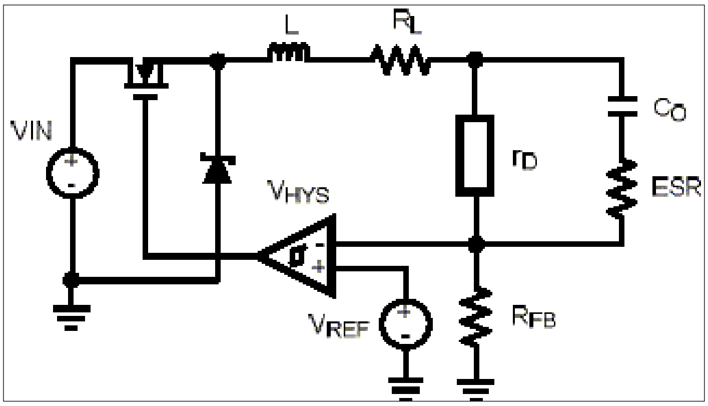


Figure 10b: CO Across LED

dency of the bond wires to break during overcurrent or thermal breakdown. In the case of a buck regulator the output can rise as high as the input voltage multiplied by the maximum duty cycle limit. As many buck regulator based current sources use little or no output capacitance, the user may elect to rate the output circuit to handle the maximum input voltage, and simply allow the output to rise during an open circuit condition. With little to no current flowing, this situation offers little chance of thermal stress. Currently a typical, complete switching regulator circuit costs less than a single HBLED, leading to production of modules which will be thrown away rather than repaired in the case of an LED failure.

In the case of boost converters and buck-boost converters, the output voltage of an open circuit system is undefined and can exceed the voltage ratings of the regulator IC, the power semiconductors, and the output filter. A simple method for preventing output voltage runaway is to place a zener diode from the output to feedback input, taking advantage of the over-voltage comparator (OVP) that is found on many switching regulators. Figure 7a shows this zener diode protection method, and Figure 7b shows an alternative method that takes advantages of the parallel MOSFET used in systems with high frequency PWM dimming. As shown these methods will place the system into a 'hiccup' mode where output voltage rises, falls due to protection circuitry, and then rises again when the protection circuitry shuts off. Users may add circuitry to make the protection feature latch, and many switching regulator ICs have a latch built into their OVP comparators.

Temperature de-rating

In order to achieve the long lifetime projection for LEDs the die temperature must remain below a limit set by the manufacturer. The Luxeon LED from Lumileds, for example, specifies a lumen maintenance of 70% after 50,000 hours provided that the die temperature is controlled to 90°C or less. HBLED product datasheets provide specific curves for the derating of forward current based on ambient temperature rise. Few switcher regulators include built-in de-rating of the output. The silicon processes suited to temperature sensing are rarely the same processes suited to switching regulator control ICs. However, the range of semiconductor temperature sensors is well suited to HBLED drives. The circuit of Figure 8 shows a lab-tested method for linear de-rating of LED current using a power supply tracking input and a low-cost external temperature sensor. The track pin is in effect a second non-inverting input to the error amplifier which supersedes the reference voltage whenever the track pin voltage is less than the reference voltage. The same method could be used with NTC thermistors. Results of the test are shown in Table 2.

Voltage mode buck regulators

A voltage mode buck regulator with output capacitance and a true op-amp for error amplification are shown in Figure 9a. Controlling output current makes one significant change to the compensator design, as rD is several orders of magnitude lower than the top feedback resistor normally associated with output voltage control. Resistor R2 of Figure 9a is used to increase the resistance from the output to the error amplifier's inverting input, keeping the compensation capacitor values in the pF to nF range. The high impedance of R2 also provides a convenient point to inject an AC signal for control loop analysis with a network analyzer. A DC attenuation created by RFB/(RFB + rD) can be used to match simulations to Bode plots measured by injection at the junction of R2 and rD.

High DC gain for output accuracy is needed, but by definition the majority of LED drive applications do not need fast transient response. For these circuits a low bandwidth, 2-pole, 1-zero (Type II) compensation is adequate. Applications that switch one or more LEDs in and out of the loop with parallel MOSFETs require faster transient response and critical damping. A 3-pole, 2-zero system (Type III) is needed for these circuits. Various papers and application notes are available on the subject of compensation component selection, any of which are valid for constant current regulation with the assumption that the input impedance to the error amplifier is equal to R3 (Type II) or R3//(R2 + ZC3).

Voltage-mode buck regulators used without an output capacitor enjoy a single-pole power stage transfer function and can be compensated with Type II integrators. For these systems, the control-to-output transfer function Gvd can be approximated as:

$$G_{vd} = \frac{V_{IN}}{V_{RAMP}} * \frac{1}{s+2\pi f_P}, \quad f_P = \frac{r_D + R_L}{L}$$

Current mode buck regulators

Several models for the Gvd of a current mode buck regulator exist, and although these models vary in their treatment of high frequency effects, they all predict a single, low frequency pole. A current mode buck used as a current source has this pole, but the low value ceramic output capacitors typical of current sources push this pole to much higher frequency, creating a flat gain that often extends into the range of the desired control loop bandwidth. A current mode buck that runs without any output capacitance can exhibit flat gain from DC up to nearly one-half of the switching frequency. From a small-signal perspective, a pure integrator would make an excellent compensator. Large signal concerns frequently make this impractical.

Current mode regulators are compensated almost exclusively with transconductance (gM) amplifiers, which by their definition are current sources. Connecting a compensation capacitor directly from the gM output to ground forms an integrator, but charging and discharging of the compensation capacitor limits the movement of the gM output. A resistor must be placed in series with the compensation capacitor in order to keep the large signal slew rate of the error amplifier from interfering with the small signal response. The response of a gM error amplifier with an R-C branch to ground is a low frequency pole and higher frequency zero, common to most current mode regulators. Various papers and application notes provide detailed approaches for the selection of the compensation resistor and capacitor. Many of these are valid for current mode buck regulators controlling output current as long as the effects of flat gain out to high frequency are taken into account.

Hysteretic and COT buck regulators

Hysteretic and Constant On-Time (COT) converters control their outputs by comparing the feedback signal directly with a reference voltage. The hysteresis of this comparator sets a window around the reference voltage within which the output is maintained. Hysteretic and COT switching regulators are attractive for LED driver control because of their basic stability regardless of the power stage. Unlike PWM converters, hysteretic converters do not have error amplifiers or small signal control loops and are not analyzed with respect to gain and phase. Controlling output current, adding DC gain through an op-amp, and the subtraction of the output capacitor do not change this behaviour. Hysteretic control provides transient response that is faster than any PWM regulator, and this ability gives hysteretic regulators another advantage over voltage mode and current mode when dimming with PWM. Hysteretic buck regulators turn on, reach regulation, and turn off faster, minimizing the delay factors introduced in Figure 6 that limit PWM dimming frequency and resolution.

Pure hysteretic converters use their comparators to turn the main power switch on and off. One challenge in the design of all pure hysteretic switching regulators is that both switching frequency and the inductor current ripple change with input voltage, output voltage, inductance, and the AC voltage ripple (&DeltavFB) at the comparator input. Component selection for pure hysteretic regulators can be done with the addition of the terms ZO and ZC from Figure 5 to the equations found in the voltage source hysteretic product datasheets and app notes:

$$f_{SW} = \frac{V_F}{V_{IN}} * \frac{(V_{IN} - V_F)(Z_S)}{(V_{HYS} * L * \frac{R_{FB} + r_D}{R_{FB}}) + (V_{IN} * t_D * Z_S)},$$

Equation 3

$$\Delta i_L = \frac{V_{IN} - V_F}{L} * \frac{D}{f_{SW}}, \; \Delta v_O = \Delta i_L * Z_S, \; Z_S = \frac{Z_O * Z_C}{Z_O + Z_C}, \; \Delta v_{FB} = \Delta v_O * \frac{R_{FB}}{R_{FB} + r_D}$$

Equation 4

The preceding equations are first-order only, and are suitable as a starting point for design. In practice, hysteretic design requires more bench testing and iteration than fixed frequency converters. If the converter is run without an output capacitor the term ZC can be omitted from the above calculations. COT converters use one-shot timers to fix either the power switch on-time or off-time and use their hysteretic comparators for either the power switch turn-on or turn-off. This added control helps keep switching frequency more constant over changes in input voltage. For either type of converter to operate properly, ΔvFB must have sufficient amplitude and must be in phase with the switching of the power transistor. LED drive circuits that do have output capacitance use ceramic capacitors almost exclusively due to their small size and low cost. The impedance ZC of ceramic capacitors is small, reducing the SNR of ΔvFB. More importantly, ZC is dominated by capacitive reactance, which introduces a phase delay between the power switch waveform and ΔvFB. This can result in sub-harmonic oscillation that causes larger output voltage ripple and output current ripple. To combat this problem additional resistance can be added in series with the output capacitor to increase the real axis component of ZC. (Figure 10a) The circuit shown in Figure 10b uses the output capacitor connected across the LED, reducing the ripple current through the LED while forcing ΔiL to flow through RFB. Switching frequency and the AC voltage and currents for Figure 10b can be approximated with the following:

$$f_{SW} = \frac{V_F}{V_{IN}} * \frac{(V_{IN} - V_F)(R_{FB})}{(V_{HYS} * L * \alpha) + (V_{IN} * t_D * R_{FB})}, \; t_D = \text{comparator delay}$$

Equation 5

$$\alpha = \frac{R_{FB} + \frac{2\pi f_{SW} C_O r_D ESR + r_D}{2\pi f_{SW} C_O (r_D + ESR) + 1}}{R_{FB}}, \; \Delta i_F = \frac{\Delta i_L}{1 + \frac{r_D}{Z_C}}$$

Equation 6

'Floating' buck regulators

The 'floating' buck regulator, also called a negative buck regulator or input referenced buck regulator, is a switching regulator with the same transfer function, inductor volt-second balance, and output capacitor charge-balance equations as a standard buck converter. Shown in Figure 11, this arrangement of the power switches, inductor, and output capacitor uses a main power switch that is ground referenced. It simplifies the power switch driving circuitry and the peak inductor/switch current sensing circuitry while producing an output that is referenced to the input voltage instead of system ground. Floating buck constant current sources can also be used without an output capacitor, and as with the standard buck, they must run in CCM in order to do so.

The primary advantage of floating buck converters with controller ICs and external MOSFETs is the ability to run from input voltages higher than the voltage limits of the control IC. This technique can be done with other low-side FET controllers, and uses a resistor and zener diode to limit the IC input voltage. (Figure 11). Schottky diodes should be used whenever possible for the recirculating diode in a floating buck converter.

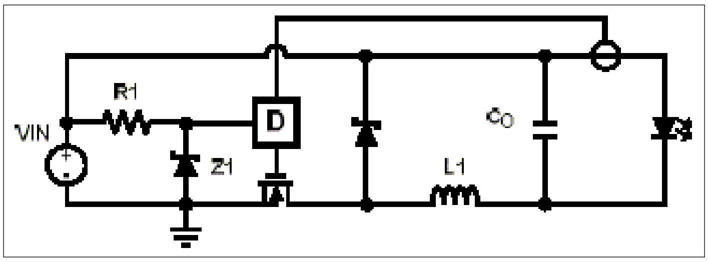


Figure 11 Floating Buck with Zener Protected Input

$$I_{L(peak)} = I_{L(avg)} + 0.5 * \Delta i_L$$

Equation 7

The low forward voltage drop of a Schottky diode improves efficiency in all types of switching converters, but for the floating buck a second efficiency gain comes from the near-zero reverse recovery time of a Schottky. In offline converters and systems with input voltages above 100V, where Schottky diodes are not available, using an output capacitor can reduce reverse recovery losses and operating in DCM, where inductor current drops to zero before the recirculating diode enters reverse bias.

Non-isolated voltage regulators are expected to produce an output that is referenced to ground, however an LED driver has no such requirement. Current control in a floating buck regulator can be done with high side sensing methods like those shown in Figure 2b and Figure 3. If the LED current need not be controlled tightly, and the input voltage tolerance is low, then a low side controller using peak current mode control can be adapted to control average current by working back from the familiar relationship between peak current and average current through the inductor:

Boost and buck-boost regulators

As HBLEDs propagate into more and more general lighting applications the number of lumens required has increased. The luminous efficiency of HBLEDs is increasing, but a single HBLED still cannot provide enough light to replace an automotive headlight or a fluorescent tube. Safety standards for DC voltages prevent many circuits from putting more than ten HBLEDs in a single chain, but applications for true non-isolated boost converters exist. A combination of the tolerance of input voltage, the tolerance of LED forward voltage, and the need for series-parallel arrays create a requirement for buck-boost regulators as well. Several factors make boost and buck-boost regulators more difficult to adapt to output current control. The average inductor current in boost and buck-boost converters is not equal to the average output current, and varies with both input and output voltage. HBLEDs cannot be placed in series with the inductor as with buck regulators. The low-side controllers and regulators used for boost and buck-boost use current mode PWM control with few exceptions. The boost and buck-boost converters have not been analyzed as deeply as buck, and their RHP zeroes force their control loops bandwidth to be low.

Conclusion

The circuits and equations presented in this paper were developed to aid engineers in driving HBLEDs with the switching power supplies available today. General illumination with opto-semiconductors is the goal of HBLED manufacturers, and to meet this goal the constant current power supplies needed to drive them are becoming more powerful and more sophisticated. Manufacturers of power semiconductors and power supply controllers are releasing the first generation of dedicated HBLED drivers, many of which are based on existing voltage regulators with modifications to ease their use as current regulators. Future generations of switch-mode HBLED drivers will require design from the beginning as current sources to meet the increasing requirements of current tolerance, current matching between different arrays, temperature correction, and dimming. By learning from the shortcomings of circuits built with the switching voltage regulators of today, the foundations of switching current regulators of tomorrow are laid.

Appendix

1. S. Nakamura, et al., "InGaN-based multi-quantum-well-structure laser diodes," Japan J. Appl. Phys. 2, 35(1B):L74-1, 1996

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