AN1427

High-Efficiency Solutions for Portable LED Lighting

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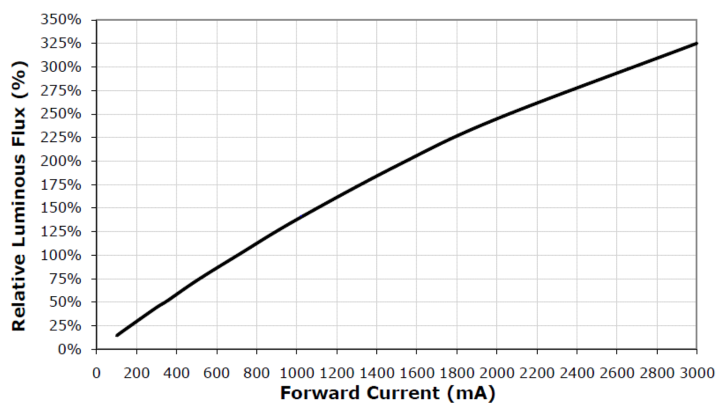
INTRODUCTION

As the LED manufacturing technology advances, parts with higher luminous flux and higher lumen per watt characteristics appear on the market. Applications include: street and roadway lighting, outdoor lighting, indoor commercial and industrial lighting, portable lighting. The drivers used in these kinds of applications need to be cost-effective, but efficiency is also critical. Depending on the type of LED (single die or multiple die), either high voltage or high current is needed. For example, a single die 10W LED will require 3A at 3.3V, and a 40W multi-die LED will require 1A at 40V.

The purpose of this application note is to demonstrate a high-efficiency design for a 10W security type flashlight. The power source is very important, since it dictates the power converter topology. Four AA type batteries can provide loaded voltages ranging from 6.4V (Ni-Zn), 6.0V (Ultimate Lithium and Alkalines) down to 4.8V (Ni-MH). A multi-die LED will require higher voltage and a boost topology, which will usually require a separate controller chip, while a single die LED requires higher current, but a buck converter topology can be easily adapted for this low voltage application.

LED luminous flux is always characterized using forward current, and varies in a linear fashion with current (Figure 1).

FIGURE 1: LED LUMINOUS FLUX VERSUS CURRENT FOR A CREE XM-L LED

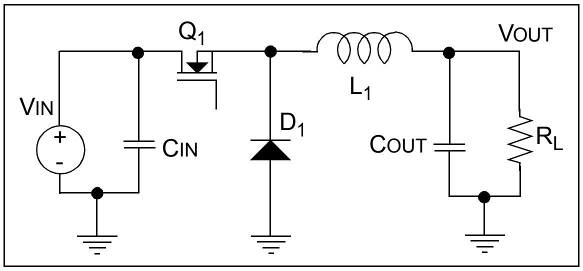


This becomes very important when trying to achieve a small flux variation on the proposed line of products (flashlights). Using Voltage mode and current limiting resistors is not recommended for several reasons. First of all, LED forward voltages may vary significantly, so the current will vary significantly. Also, using resistors at the required forward current will severly compromise efficiency (100 mOhms dissipates 0.9W at 3A). Current mode converters will achieve the highest efficiency when driving LEDs, and will keep luminous flux constant.

BUCK CONVERTER IMPLEMENTATION

The non-synchronous buck converter is very simple and can be driven with most of the PIC® microcontrollers, but there are downsides at high output currents and low output voltages. This is exactly the case with single die LEDs.

FIGURE 2: BASIC BUCK CONVERTER

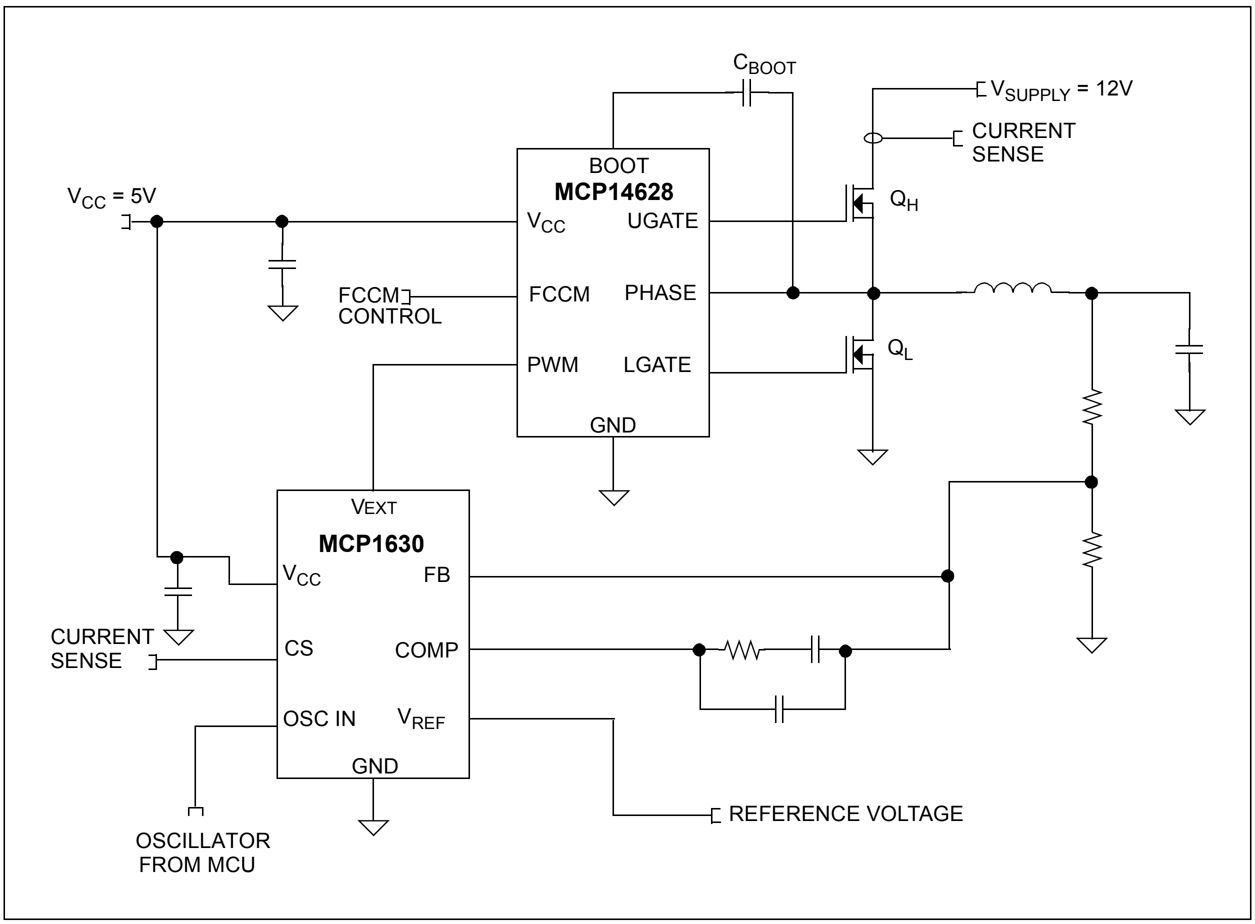


Diode D1 is a Schottky type and has a typical forward current of 0.7V. At a 3A current, power dissipation is around 2W while conducting, which is unacceptable for a battery-powered device, not to mention the thermal management. Higher cost diodes may have a lower forward voltage (down to 0.35V), but even 1W of dissipation is a serious problem.

Another issue is driving the MOSFET correctly to minimize losses. Portable applications usually have space constraints, and so require high operating frequencies to keep the magnetic components small. The higher the switching frequency is, the higher the importance of the MOSFET switching time for overall efficiency. High current pulses from an external driver chip are needed to drive the MOSFET gate properly.

A synchronous rectification switching power supply also needs adequate dead time between the transistor gate signals to avoid current shoot-through, and not to allow the body diode to start conducting. Microchip's MCP14628 has an adaptive dead-time generator, bootstrapped floating high side driver, and a 2A current driving capability, which makes it perfect for this design.

FIGURE 3: MCP14628 VOLTAGE MODE SYNCHRONOUS BUCK CONVERTER

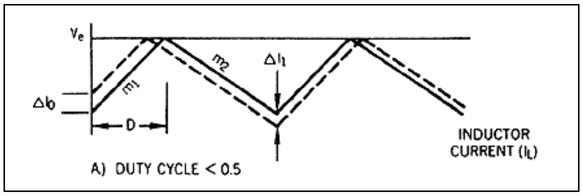


Using an additional PWM controller makes things straightforward enough, but raises cost. A LED driver is also a Current mode controller and this complicates things further. A slope compensation circuit on the current sensing (CS) pin is needed to stabilize the PWM duty cycle.

CURRENT MODE CONTROL

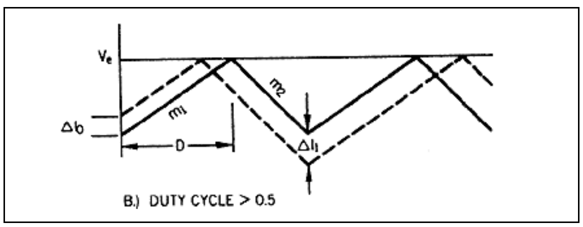
An early discovery in the development of Current mode control was that the current feedback loop became open loop unstable when the duty cycle was increased beyond 50%. This phenomenon has been thoroughly studied and analyzed. Disturbances in the operating point gradually die out when the duty cycle is below 50% (see Figure 4).

FIGURE 4: FOR DUTY RATIO LESS THAN 0.5, DISTURBANCES DIE OUT



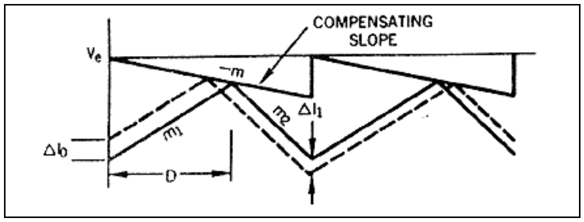
For duty cycles greater than 50% however, a disturbance from the nominal operating point grows larger with each cycle. This leads to large deviations from the nominal operating point and to a phenomenon known as "sub-cycle oscillation." Figure 5 shows the beginning of this process.

FIGURE 5: FOR DUTY RATIO GREATER THAN 0.5, DISTURBANCES GROW



By adding "slope compensation" to the trip level (or to the sensed current signal), the duty cycle at which a disturbance begins to grow can be increased. Figure 6 shows the effects of slope compensation.

FIGURE 6: SLOPE COMPENSATION CAN CAUSE DISTURBANCES TO DIE OUT FOR ANY DUTY RATIO



If the slope of the falling current in the energy storage inductor is called m_2, then a negative slope equal to half the slope of m_2 will, in theory, cause a disturbance to die out for any duty cycle up to 100%. Two other advantages that occur with this particular amount of slope compensation are that the average current is no longer a function of duty cycle, and that line voltage changes are rejected without requiring action by the voltage loop. Figure 6 shows the slope compensation, where $-m = m_2/2$, and the effect on average current. Although the advantages of having $-m = m_2/2$ are significant, they are difficult to achieve. In practice, it is better to have more compensation than $-m = m_2/2$ to assure no oscillation occurs at high duty cycles.

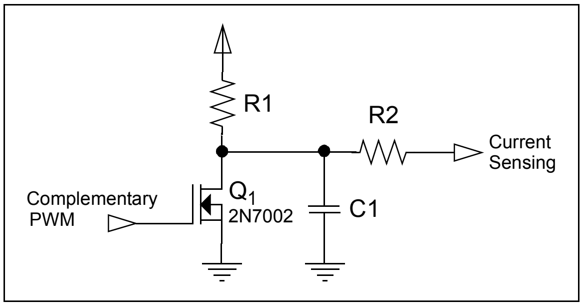
For a buck converter, the m_2 current slope is equal to V_{OUT}/L. Since V_{OUT} is the LED forward voltage, this is easy enough to calculate. Inductor values may have a 20% tolerance, so one should use the highest LED forward voltage and the lowest possible inductor value to calculate the maximum current slope.

CURRENT MODE CONTROL USING SLOPE COMPENSATION

A practical circuit for generating slope compensation uses a complementary PWM output and a transistor that discharges the capacitor from an RC (R1/C1 in Figure 7) element. When the transistor is open, the capacitor charges from V_{CC} through the resistor, and a slope is added to the current sensing input. Keep in mind that only the first portion of the waveform is of interest, so pick the RC constant accordingly.

To obtain a perfect ramp use a current source instead of resistor R1 to charge the capacitor.

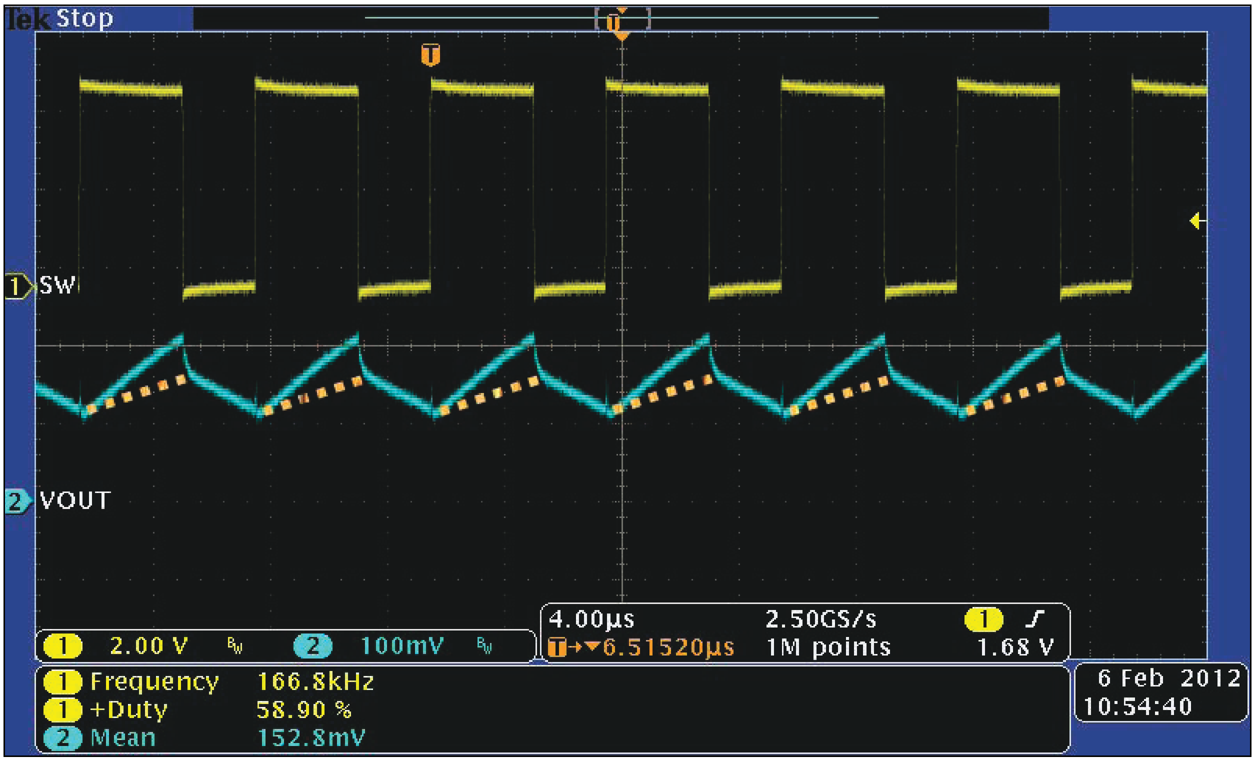
FIGURE 7: PRACTICAL SLOPE COMPENSATION CIRCUIT



Note that this is a fixed frequency implementation and can be used together with any PIC microcontroller that has an Enhanced Capture Compare PWM (ECCP) module and a fast comparator.

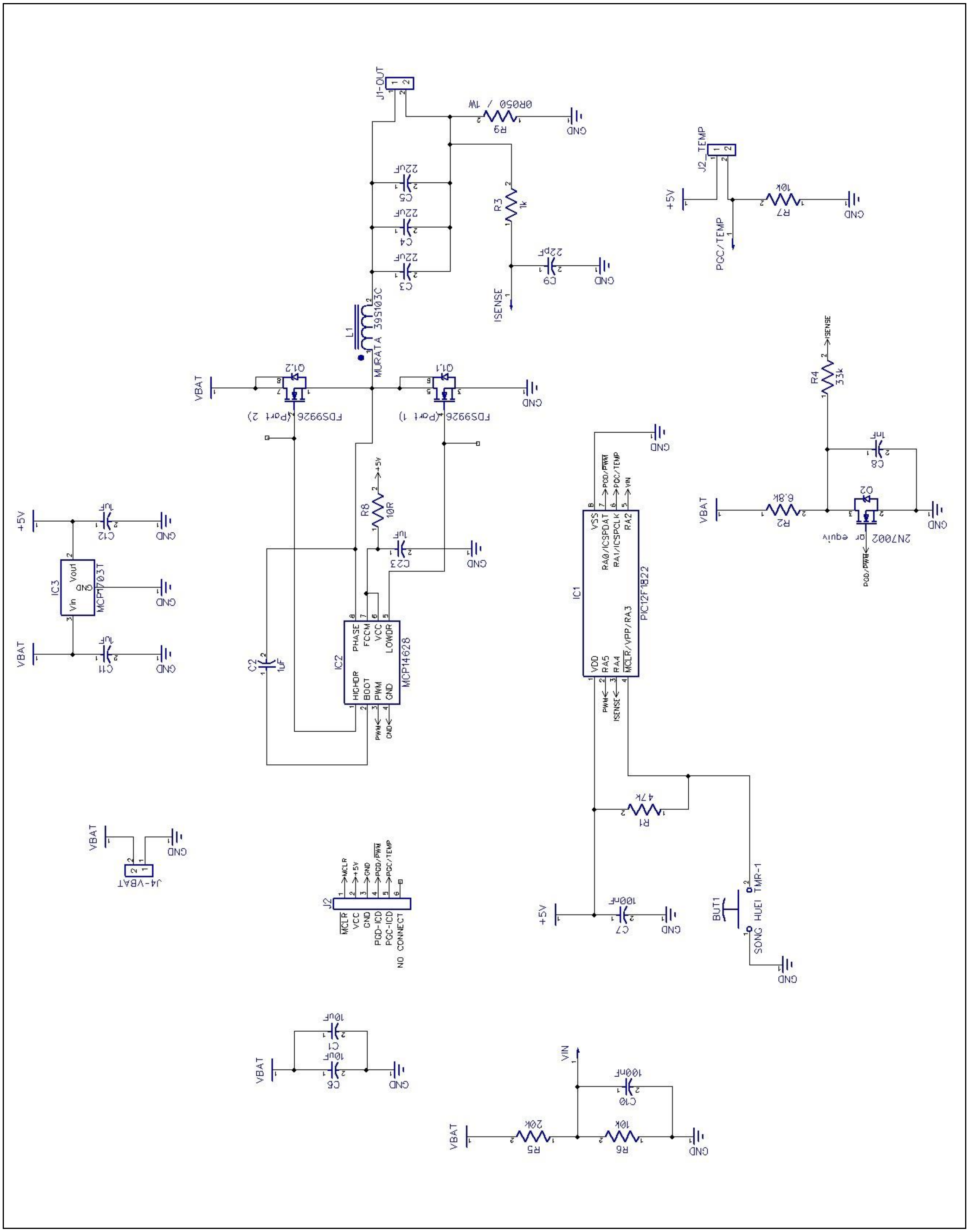
The output current waveform is shown in Figure 8. The dotted lines show how the waveform would look without slope compensation. Note that when the slope compensation transistor is closed (high side transistor open), a sudden drop occurs in the current waveform caused by the removal of the slope. The method is similar to the one shown in Figure 6, but the circuit used for the implementation adds slope compensation to the sensed signal, not to the trip level.

FIGURE 8: OUTPUT CURRENT WAVEFORM WITH SLOPE COMPENSATION



HARDWARE CIRCUIT DESCRIPTION

FIGURE 9: PIC12F1822 CURRENT MODE SYNCHRONOUS BUCK CONVERTER WITH SLOPE COMPENSATION



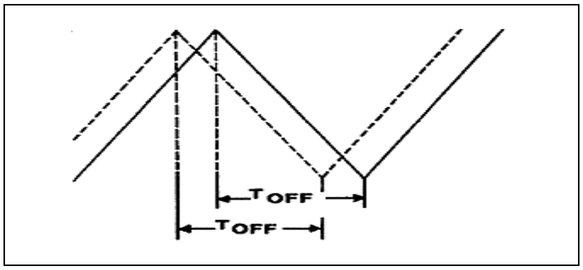
Current limiting is done in a pulse-by-pulse manner using the PIC MCU comparator. The reference is set by the internal DAC and the comparator is fed to the PWM auto-shutdown input. Note that the output capacitors are connected to ground through the shunt resistor. The inductor and output capacitors form an LC filter, so the waveform normally present on a low side shunt resistor (R9 in Figure 9) would only be appropriate for average current control. A high side shunt would also complicate the schematic needlessly. By connecting the capacitors to ground through the shunt, the equivalent series resistance (ESR) will be slightly higher (also slightly increasing power loss), but the current waveform can be successfully used for pulse-by-pulse control schemes.

CURRENT MODE CONTROL USING FIXED OFF TIME

Fixed frequency is by far the most popular mode of operation for PWM converters, because it allows them to be synchronized and simplifies the design of the magnetics. If the application permits a mode other than fixed frequency however, fixed off time presents significant advantages over any other mode of operation.

With fixed off time, disturbances in the nominal operating point die out in one cycle, as they would normally with a slope compensation of $-m = m_2$. Figure 10 shows this effect.

FIGURE 10: DISTURBANCE DIES OUT IN ONE CYCLE FOR FIXED OFF TIME



Changing line voltage causes the current to reach the trip level at a different time than otherwise, but once the transistor turns off, the rate of fall of the inductor current is constant and determined only by the output voltage. With fixed off time, the change in current is constant, therefore the average current does not change. Line ripple is totally rejected, as it would be with normal Current mode and a slope compensation of $-m = m_2/2$. For higher line voltages, the inductor current rise rate increases, so it takes less time to reach the set level. For lower line voltages, the inductor current rise rate decreases, so we can see that switching frequency increases with line voltage.

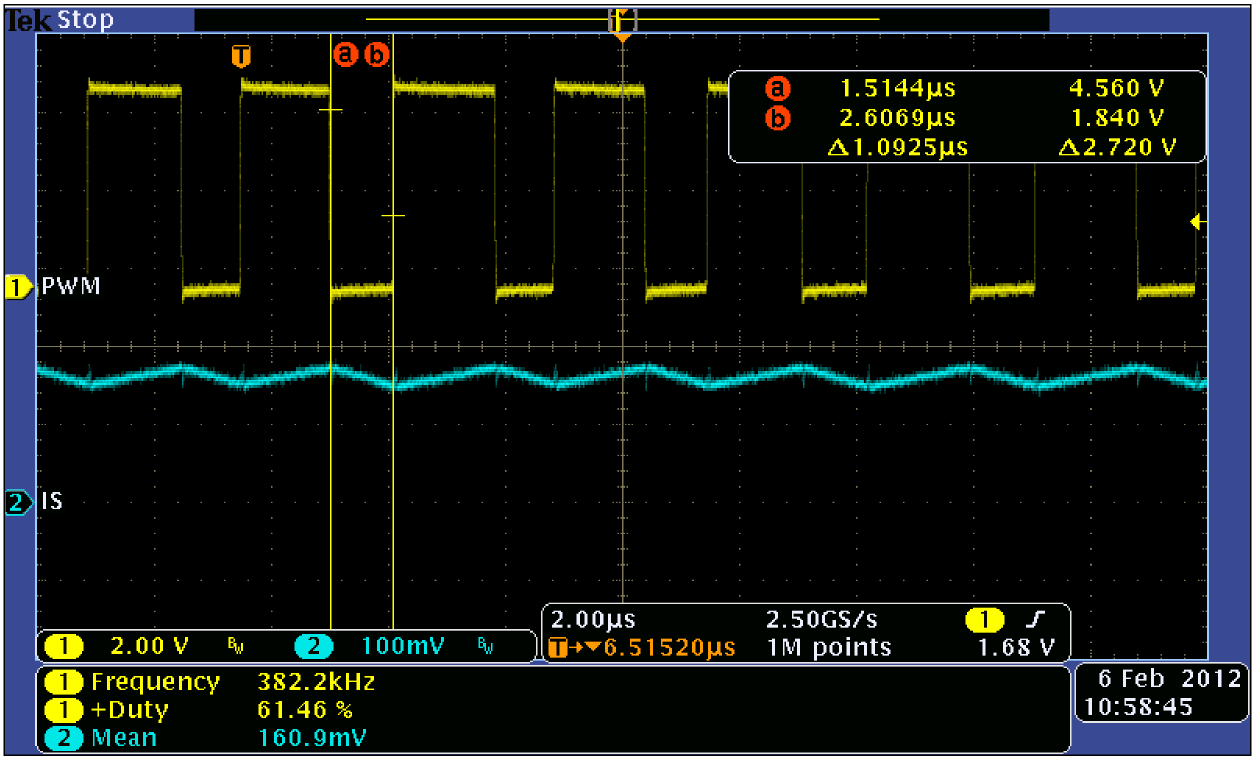
The PIC12F752 complementary output generator (COG) module can be used to implement this control scheme. COG rising and falling inputs are set to a comparator output edge. The shunt voltage is fed to the comparator inverting input and the current limit is set by the DAC, internally connected to the non-inverting input. The schematic is identical to the PIC12F1822 implementation, with the exception of the slope compensation circuit, which is removed, as it is not needed anymore.

A comparator high output shows that the inductor current is below the set point and a low output shows that the current is above the set point.

In this mode of operation, the comparator would turn off the COG output asynchronously when the current set point is reached, and very quickly turn on again (depending on the response time) when the current drops. The trick is to use the rising event blanking register to delay COG output for a set time (basically controlling the off time). Since the blanking register masks the event for a number of clock periods, there will be no edge to turn on the output (the comparator output positive edge has already occurred), and a level type of event needs to be configured for the rising event. Because the level type events are synchronized with the COG clock, this will introduce a timing uncertainty, which is always less than one oscillator period, but it will not cause any problems, because the uncertainty period is at least one order of magnitude smaller than the switching period. Some jitter can be observed on the off-time period, but stability is preserved.

Figure 11 shows converter operation using a fixed off-time control scheme. The PIC12F752 oscillator runs at 8 MHz, and the blanking period is set to eight clocks, resulting in a 1us fixed off time. Depending on the input voltage range, the switching frequency may go very high increasing switching losses, or very low increasing output ripple, so it's very important to choose the blanking period well.

FIGURE 11: OUTPUT CURRENT WAVEFORM WITH FIXED OFF-TIME CONTROL



The off time is fixed and is user-configurable, and the on time depends on the input voltage. Since the inductor current drop caused by the off time is always the same (LED voltage constant), we have:

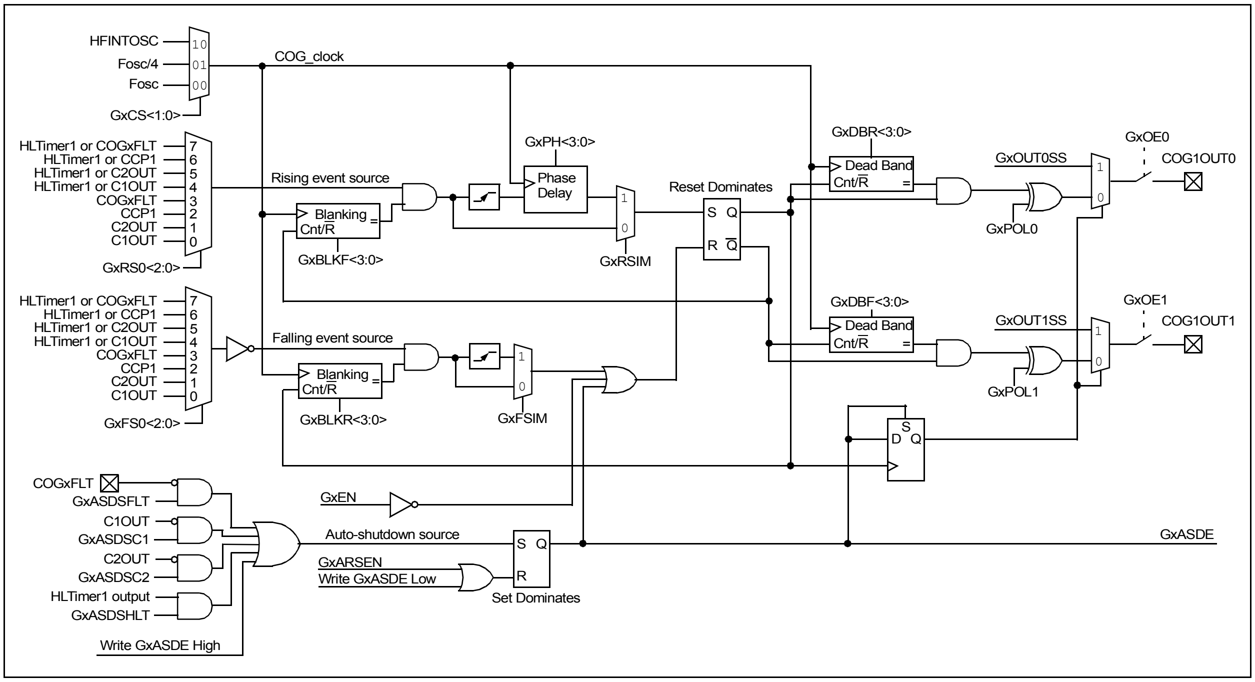
EQUATION 1:

$$\frac{V_{IN} - V_{OUT}}{L} T_{ON} = \frac{V_{OUT}}{L} T_{OFF}$$

The inductor value, LED forward voltage and “off time” are known values, so the “on time” can be calculated using the input voltage value. The inductor current increase during the “on time” is equal to the current decrease during the “off time”.

One very interesting observation is that the fixed “off time” essentially sets the inductor ripple current.

FIGURE 12: PIC12F752 COG MODULE



Fixed frequency schemes have a very simple protection method for a broken feedback loop. The maximum duty cycle is set to certain value and, if the comparator stops functioning for some reason, it will never short the input to the output (100% duty cycle). Unfortunately, this is very likely to happen with the fixed off-time control scheme if only the comparator output is used to control duty cycle.

One of the advantages of the COG module is that certain inputs can be ORed together with a hardware safety timer, HLTimer. Figure 12 shows the COG module structure. The comparator, Fault signal and CCP outputs can be combined with the safety timer. The HLTimer will change the COG output when its period expires and resets itself. So, there are two scenarios:

1. If the comparator output gets stuck output high, then the safety timer will turn the COG output off after the timer expires. Blanking should be always shorter than the timer period. After the blanking period expires, the COG output turns on again, resulting in a period equal to the safety timer period plus the blanking period. The duty cycle is:

EQUATION 2:

$$D = \frac{THLTimer}{THLTimer + TBlanking}$$

2. If the comparator output gets stuck low, then the safety timer will turn on the COG output after the safety timer expires. Since there is no edge to trigger the transition, the COG output will stay high until the safety timer expires again, resulting in a period equal to two safety timer periods and a 50% duty cycle.

The safety timer is also very important for start-up conditions based on edges. For example, when the circuit is powered on, COG output is low, comparator output is high (because the inductor current is lower than the set point), but there was no edge to trigger the transition. The safety timer triggers it after it expires, and the circuit starts functioning normally.

OUTPUT EFFICIENCY

Having synchronous rectification can push efficiency over 90%. The fixed off-time buck implementation has a conversion efficiency of 92% in either of the 10W or 2W modes.

Please note that the lack of I/O ports has made it necessary to hardwire the FCCM pin of the MCP14628 to VDD, essentially forcing the driver to always work in continuous Conduction mode. This is no problem for the 10W mode, where the current is very high and continuous, but it will severly compromise efficiency in the low-power modes, if the inductor current becomes discontinuous.

For this reason, a low-power mode of 2W has been selected, and the 1W and 0.6W modes are obtained by PWM/PFM dimming the 2W setting. Using a lower current setting would risk going into Discontinuous mode depending on the set “off time”.

LAYOUT TIPS

To design a low noise, high-efficiency converter, it is necessary to observe the current paths. Everything should be placed as tightly as possible and connected by generous copper areas. Only use traces instead of copper pours where the current ripple is small (e.g., after the output capacitors). Even in that case, traces should be thick enough to accommodate the passing current.

Because current limiting is done on the current peak in a pulse-by-pulse fashion, it is needed to have a proper inductor current waveform. The inductor and the output capacitors filter the output current nicely and make the waveform unsuitable for the PIC MCU comparator. Normally, a high side current sensor would do the trick, but that would complicate the layout and add a lot of extra components. A quick solution is to connect the output capacitors to the ground through the current shunt. There is a very small loss of efficiency because of the higher ESR, but the current waveform has the correct shape. Even with this trick measured, efficiency goes above 90% at a 3A output current.

Closing ground loops properly means the layout needs to have a short (and clear of obstruction) path between the ground connections of important components: input capacitors, low-side switch and output capacitors (current shunt in this case, because of the trick mentioned earlier). It is ideal to use solid ground connections for these components but, if it really makes it impossible to solder the components by hand, use a thermal relief with very thick spokes (40 mils or more).

Use ground planes on both sides, if possible, and a generous number of vias to decrease the impedance. One side should have unobstructed, short paths between component ground connections. If it is not possible to use ground planes on both sides, use copper pours (on the side you cannot use a plane on), and pack it with vias to minimize the impedance to the other side.

Other things that need to be laid out carefully:

- Connections from the MCP14628 driver to the MOSFET gates should be thick and short, because peak current is 2A. The measured positive and negative edges on the transistor gates are < 10 ns, resulting in a signal bandwidth close to 1 GHz.
- MCP14628 should be decoupled properly with a capacitor placed very close to the power pin. When driving the transistor gates, high current pulses are needed.
- MCP14628 BOOT and PHASE pins should have short and thick connections.
- Capacitor C9 (See Figure 9) from the current sense RC filter should be placed close to the microcontroller comparator pin to properly attenuate high-frequency switching noise.

CONCLUSIONS

The new generation of microcontroller peripherals like the COG allows easy implementation of cost-effective, high-performance switching converters, while allowing a lot of flexibility. Current mode converters have many important applications, especially in LED lighting. Bringing together the flexibility of a PIC microcontroller and a synchronous topology will result in a high-efficiency power supply design (>90%) with a lot of intelligence and room for customization.

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- "*PIC12F752/HV752 Data Sheet*" (DS41576), (http://ww1.microchip.com/downloads/en/DeviceDoc/41576B.pdf)

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ISBN: 9781620761601

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