AN874

Buck Configuration High-Power LED Driver

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INTRODUCTION

The circuit and firmware described in this application note demonstrates a minimal parts count driver/controller for a high-power (1W or greater) LED. The circuit is based on a buck topology switching power supply using the on-chip comparator peripheral within the PIC12F675 PIC® microcontroller. The switching power supply design ensures efficient power transfer between the system battery and the output LED.

The Flash PIC microcontroller is used to implement intensity control, automated intensity compensation for low battery conditions, and the ability to playback pre-programmed Flash sequences. The Flash PIC microcontroller also allows the creation of a custom PC-based graphical user interface, using the PICkit™ 1 Flash Starter Kit for programming the pre-programmed Flash sequences. The combination of the switching power supply design and the microcontroller results in an efficient circuit with advanced automated features while keeping the circuit simple and inexpensive.

High-Power LEDs

The new generation of high-power LED modules represent a significant advancement in LED design. Ranging from 1 watt to 10 watts, current high-power LED modules are capable of delivering between 10 and 50 Lumens/watt of light output. This level of light output is comparable to most incandescent lamps and even halogen bulbs in single-color applications. In addition, high-power LEDs are available in a variety of colors from multiple manufacturers, in discrete or pre-built modules.

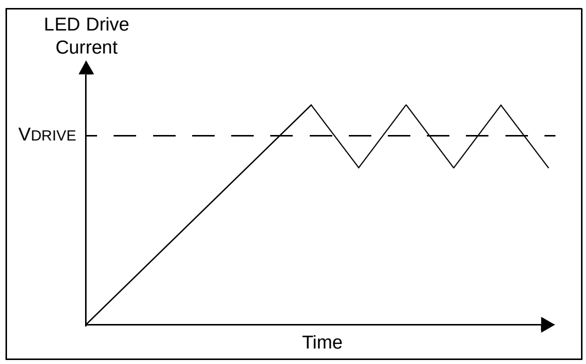
This application note will focus on driving two different high-power LED modules, a 1 watt and a 3 watt.

Voltage and Current

The 1 watt chosen module requires a typical drive of 3.4V at 350 mA to produce its full output brightness. The maximum current drive for the unit is 500 mA. Exceeding the maximum current specification for the module, even by short duration pulses, may result in damage. Therefore, the drive circuitry for the module must deliver a DC or mostly DC current to the LED module to produce a full brightness output. The drive requirements for the 3 watt module are similar, with full brightness output at 700 mA and a maximum current drive of 1A.

The circuit described in this application note delivers a mostly DC drive current with a small triangular ripple waveform superimposed at the drive level (V_{DRIVE}) (see Figure 1). The triangular waveform is the result of the switching nature of the driver, and if kept small in relation to the DC drive current, will not exceed the instantaneous maximum current specification for the LED. In fact, the example designs have been designed with a safety margin to insure that the instantaneous current delivered to the LED is always less than the maximum rating, even when delivering the full-rated current to the device.

FIGURE 1: LED CURRENT



Buck Topology Switching Power Supply

The Buck Topology Switching Power Supply is an efficient voltage regulator that translates a high-source voltage into a lower output voltage. It accomplishes this by rapidly switching the input of an inductor/capacitor (LC) network between source voltage and ground and then back to the source voltage (see Figure 2). While the PWM switch is in position 1, L1 is connected to the source voltage, the power supply is in its charging phase, and an increasing current flow (I_L) passes from the source, through the inductor, to the load and C_{OUT}. While the charging current is flowing through the inductor to the load, part of its energy is stored in the inductor as a magnetic field. When the PWM switch changes to position 2, the power supply enters its discharge phase and the magnetic field around the inductor collapses, continuing the current flow to the load. See the graph of the inductor current in Figure 2. When I_L drops to zero, the PWM switch is switched to position 1 and the charge/discharge cycle starts over. The result of this switching cycle is an inductor current that ramps up and down over the course of a cycle.

The capacitor (C_{OUT}), in the LC network, acts to smooth I_L into a DC current flow to the load. When I_L is greater than the load current (part B of the capacitor current graph in Figure 2), the load current is supplied by I_L and any surplus current (I_C) flows into C_{OUT}, charging the capacitor. When I_L falls below the load current requirement (part A of the capacitor current graph), the current flow to C_{OUT} reverses and I_C supplements I_L to make up the difference between I_L and the required load current.

When describing the operation of the switch control in a switching power supply design, it is typically the charge time and total switching time which are specified. The charge time is specified as a percentage of the total switching time and is referred to as the duty cycle of the system, or D in the design equations.

EQUATION 1:

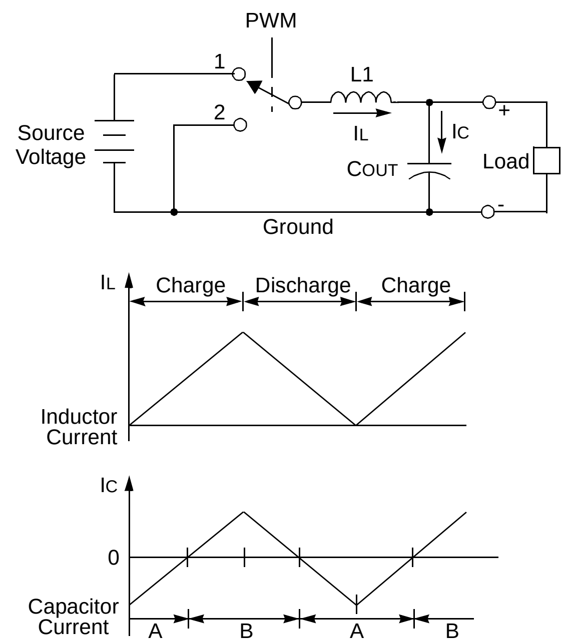
$$D = \frac{\text{Charge Time}}{\text{Charge Time + Discharge Time}}$$

The total switching time (represented by T in the equations) can either be specified directly or by its reciprocal, the switching frequency (F_{SWX}).

EQUATION 2:

$$Fswx = \frac{1}{\text{Charge Time + Discharge Time}} = \frac{1}{T}$$

FIGURE 2: BUCK TOPOLOGY SWITCH CYCLE



A = C_{OUT} supplies current to the load.

B = The inductor supplies all of the load current and any surplus current charges C_{OUT}.

A feedback circuit regulates the switching in the switching power supply. This circuit monitors the load voltage and compares it to a stable reference. Based on the result of the comparison, the circuit adjusts the switching duty cycle to compensate for any discrepancies. The feedback circuit cancels out any errors in the load voltage due to component or timing tolerance and it adjusts the duty cycle to compensate for changes in the load current. The result is a self-regulating step-down voltage regulator that produces a stable load voltage over a variety of load currents.

Note:	A more in-depth explanation of switching power supplies can be found in the reference materials listed at the end of this application note.

Driver Theory of Operation

One switching power supply design concept, not previously mentioned, is the idea of continuous versus discontinuous inductor current. In a discontinuous design, the current flow in the inductor drops to zero at the end of each discharge cycle as previously discussed. However, in a continuous current design, the inductor current does not drop to zero. Instead, the inductor maintains a DC current flow throughout the switch cycle. The resulting inductor current has both AC and DC components to its waveform. The DC component equals the average current flow during the cycle and is determined by the reference voltage VDRIVE. The AC component is a triangular shaped waveform super-imposed on the DC component and is caused by the switching action of the driver. The advantage of continuous current design is that the inductor current flows to the output continuously, which reduces the charge storing requirements on COUT.

The driver presented in this application note is designed to take advantage of the DC component of a continuous inductor current design. See Figure 3 for the block diagram of the driver. Notice that the circuit is very similar to the buck topology switching power supply in Figure 2. The MOSFET (Q1) and Schottky diode (D) form the switch, the inductor is the same as the buck topology switching power supply, and the LED is the load connected to the output. The only differences are the lack of an output capacitor (COUT), and feedback is based on the load current instead of load voltage.

The Buck topology LED driver also has a similar charge/discharge cycle to the buck topology switching power supply. The charge phase of the cycle starts by turning on MOSFET, Q1. This starts an increasing current flow from the battery, through Q1, the inductor, the LED, and R8 (see Figure 3). When the current through R8 generates a voltage greater than the VDRIVE reference voltage, the comparator turns Q1 off, ending the charge phase and starting the discharge phase. During the discharge phase, current flows through the Diode D, the inductor L1, and R8. The inductor current ramps down until the current flowing through R8 generates a voltage less than VDRIVE. When this occurs, the comparator turns Q1 back on and the next charge phase begins. The resulting current flow through the LED/inductor is a DC level with a small triangular waveform, super-imposed on top, that is synchronized to the charge/discharge cycle (see Figure 4).

Because the current through R8 generates a voltage that toggles about the VDRIVE level, controlling VDRIVE controls the current output to the LED, making it the control input to the driver circuit.

FIGURE 3: BUCK CONFIGURATION LED CURRENT CONTROL

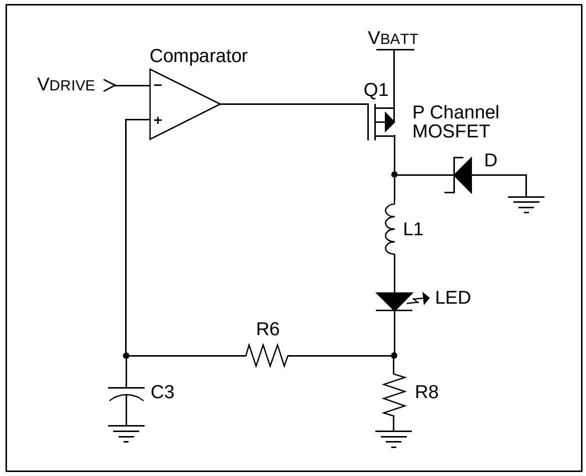
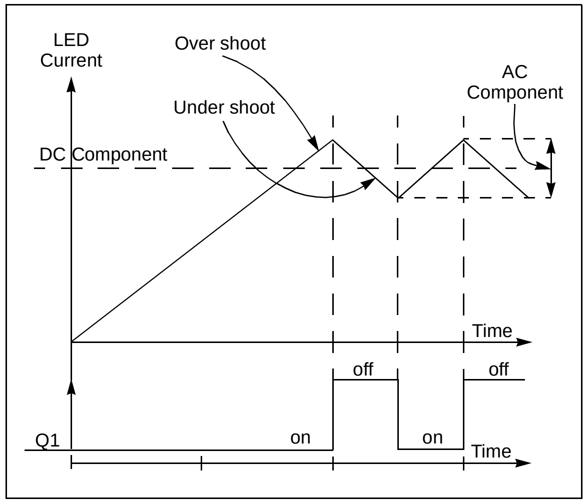


FIGURE 4: INDUCTOR/LED CURRENT FLOW



If the driver is built as described above, it will operate at a frequency determined by the switching speed of the MOSFET and the propagation delay of the comparator. Unfortunately, this will switch the MOSFET so rapidly that the switching time of the MOSFET will be a significant percentage of the switch time, resulting in increased switching losses in the transistor. Therefore, an RC low-pass filter (R6 and C3) is added to the feedback path to introduce a time delay. The resulting low-pass filter does two things for the driver circuit:

1. It reduces the amplitude of any fast voltage transients generated by switching Q1.
2. And, it increases the charge and discharge periods of the switch cycle, such that the inductor/LED current will over and undershoot the VDRIVE level.

Driver Component Selection

The design of the buck topology driver requires the selection of 6 components critical to the operation of the driver. These components are:

- MOSFET Q1
- Schottky Diode D
- Inductor L1
- Current-sensing resistor R8
- Time-delay components R6 and C3

THE MOSFET TRANSISTOR

The selection of the MOSFET transistor is determined by 4 factors:

Current Rating

The transistor must handle the maximum current output of the driver. Therefore, the minimum current rating for the transistor (I_D) is determined by the combined peak value of the DC and AC components of the LED current waveform.

The current drive for maximum normal output of the 1 watt module is 350 mA and the absolute maximum current is 500 mA. For the 5 watt module, the maximum normal drive is 700 mA and the absolute maximum is 1A. Therefore, a MOSFET transistor with an I_D rating of 1A minimum should handle the maximum possible charge-phase current for both modules.

On-resistance R_{DSON}

The on-resistance will determine the transistor's power dissipation during the charge phase of the cycle. Therefore, R_{DSON} should be minimized to limit the power dissipation of the transistor to less than 3% to 5% of the total power for the driver. To drive the 1 watt module, a maximum drive current of 350 mA is required. Therefore, the R_{DSON} of the transistor should not dissipate more than 50 mwatts (1 watt x 5%). This yields an upper limit on R_{DSON} of 408 mΩ (50 mwatt/(350 mA x 350 mA)).

To drive the 3 watt module, the maximum drive current is 700 mA, subsequently the R_{DSON} of the transistor should not dissipate more than 150 mwatts (3 watts x 5%). This yields an upper limit on R_{DSON} of 306 mΩ (150 mwatts/(700 mA x 700 mA)).

Total Gate Charge

The total gate charge (Qg) and the drive capability of the pin driving the gate determine the upper-frequency limit of the MOSFET switching speed. Therefore, the lowest total gate charge possible is preferred. The pin driving the gate of the MOSFET is a standard I/O pin, which can source and sink 20 mA. The transistor switching time should be limited to less than 5% of the total switch period to keep the transistor switching efficiently. Thus, for a switching frequency of 100 kHz, Qg for the MOSFET must be less than 10 nC (20 mA X 5% x 1/100 kHz)). (nano – Culombs = nano-amp-seconds).

Cost

As with any design, the lowest possible cost is preferred and a consideration in the transistor's selection. For this design, an arbitrary cost limit of $0.75 is chosen for a 1000-piece quantity.

A transistor which meets the requirements of the design is the International Rectifier IRF7524D1.

I_D = 1.7A

R_{DSON} = 270 mΩ

Qg = 8.2 nC

Cost < $0.70 (Digi-Key®, 1000-count, cut reel)

THE SWITCHING DIODE D

The selection of the switching diode is determined by 3 factors:

Speed

The diode should switch quickly to prevent high reverse currents, when Q1 turns on and the switching speed of a diode is largely determined by the physical construction of the diode. Therefore, the selection for D should be limited to Schottky or FRED (Fast Recovery Diode) type diodes.

Power Dissipation

Because the diode will have to carry the inductor/LED current during the discharge phase of the switching cycle, it should be able to dissipate the heat generated during conduction. The capability of the diode to dissipate power should be sufficient for a 50% duty-cycle current of 1A without additional heat sinking. Assuming a Schottky diode with a forward voltage of 0.5V, then the diode should be able to dissipate 0.25 watt (0.5V x 50% x 1A).

Cost

As with any design, the lowest possible cost is preferred. Therefore, this factor should be considered in the diode's selection. For this design, an arbitrary cost limit of $0.30 is chosen for a 1000-piece quantity.

The diode chosen is the on-chip Schottky diode in the IRF7524D1 FETKEY transistor. The diode has a forward voltage of 0.39V, for a total power dissipation of 0.195 watt. Together with the power dissipation of the MOSFET results in a total power dissipation for the package of 0.327 watt, which is well below the package rating of 0.800 watt.

Power Dissipation of the diode

0.39V x 1000 mA x 1000 mA x 50% = 0.195 watt

Power Dissipation of the MOSFET

0.27 ohms x 1000 mA x 1000 mA x 50% = 0.135 watt

Because the diode is included in the cost of the MOSFET, the cost is $0.00.

THE INDUCTOR L1

The selection of the inductor is determined by multiple factors.

- Switching frequency (F_{SWX})
- Saturation current (I_{SAT})
- Cost
- System battery voltage (V_{BATT})
- LED forward voltage (V_{FOR})
- Peak-to-peak amplitude of the maximum AC component of the current flow (I_{RP})

The first parameters to determine are:

- Inductance
- Switching frequency
- Peak-to-peak AC component (I_{RP})

The battery voltage for this design is V_{BATT} = 6V.

The 1 watt module has a typical V_{FOR} = 3.4V and the 3 watt module has a V_{FOR} = 3.7V.

Recall that the current drive required for full output brightness from a 1 watt module is 350 mA, with an absolute maximum of 500 mA. The full output brightness current drive for a 3 watt module is 700 mA, with an absolute maximum of 1000 mA. Thus, for 1 watt operation, a current drive with a DC component of 350 mA and an AC ripple current (I_{RP} = 100 mA peak-to-peak) will still limit the maximum instantaneous current to 400 mA (350 mA + (100 mA/2)). This gives a 100 mA safety margin between the full brightness current and the absolute maximum rating for the drive.

For the 3 watt module, a DC component of 700 mA and an I_{RP} = 200 mA peak-to-peak will limit the maximum instantaneous current to 800 mA (700 mA + (200 mA/2)). This gives a 200 mA safety margin between the maximum driver current and the 1000 mA absolute maximum rating for the LED.

The switching frequency (F_{SWX}) is determined by several factors, both subjective and quantitative:

- Size of the inductor
- Weight of the inductor
- Availability from a specific supplier or multiple suppliers
- Limiting ultrasonic sound emissions
- The series resonant frequency (SRF) of the available inductors
- Trade-off of inductance versus size/cost

For this design, a frequency between 50 kHz and 100 kHz was desired. The frequency range is reasonable for two manufacturers' families of small surface-mount inductors and it is above the hearing range of most animals, and it is also well below the SRF rating for both families of inductors.

Given the information above, a range of inductor values (L) can be determined using Equation 3.

EQUATION 3:

$$L = \frac{V_{BATT} - V_{FOR}}{I_{RP} \times F_{SWX} \times 2}$$

	1 watt	3 watt
50 kHz	315 µH	86 µH
100 kHz	158 µH	43 µH

To keep a common inductor selection for 1 watt and 3 watt operation, a value of 220 µH was selected, assuming a switching frequency of 50 kHz. For the 3 watt operation, this will result in a slightly lower I_{RP} value and oscillating frequency.

The two inductors meeting these drive requirements are the DO5022-22P-224 from Coilcraft™ or PM5022-221M-B from J.W. Miller™.

CURRENT SENSING RESISTOR R8

Resistor R8 is used to measure the current flowing in the inductor/LED combination, as a result, it has two conflicting requirements. It should be large enough to generate a reasonable feedback voltage, but it should also be small enough to limit its power dissipation.

For 1 watt operation, a resistance of 1Ω was chosen. At 350 mA, a 1Ω resistor generates a 350 mV signal and only dissipates 123 mwatts (11% of the total power). The 1Ω resistor will also generate a 100 mV peak-to-peak voltage, corresponding to the 100 mA ripple current.

For 3 watt operation, a resistance of 0.5Ω was chosen. At 700 mA, a 0.5Ω resistor generates a 350 mV signal and only dissipates 123 mwatts (4% of the total power). The 0.5Ω resistor will generate a 100 mV peak-to-peak voltage for a 200 mA ripple current.

Both resistors should have 0.25 watt power ratings with carbon composition elements to limit inductive effects.

TIME DELAY COMPONENTS R6 AND C3

Selection of R6 and C3 is now simple, the switching frequency F_{SWX} is known, so a suitable combination of values can be determined from Equation 4.

EQUATION 4:

$$Fswx = \frac{.225}{R6 \times C3}$$

Note:	This equation was derived empirically from the performance of the circuit and assumes a 100 mV peak-to-peak feedback signal.

A value of 1000 pF for C3 and a value of 5.1K for R6 produce a switching frequency of approximately 50 kHz. Because the frequency of the circuit is influenced by a number of factors, including the average current in the inductor, the actual frequency will shift up and down during operation and may be as much as 25% off from the predicted value.

OTHER IMPORTANT COMPONENTS

Resistor R2 (see Figure 8) is important to the proper operation of the driver during power-up. It keeps transistor Q1 off until the microcontroller can configure the on-chip comparator. Without R2, any charge present on the gate of Q1 could turn the transistor on, creating a high current level that could permanently damage both the MOSFET transistor and the LED.

Because the gate of Q1 is capacitive, Resistor R7 is needed as a current limiter to prevent abnormally high transient currents when the transistor is switched on and off. If the switching frequency of the driver is increased, R7 may need to be reduced to allow faster MOSFET switching times.

Thermal Considerations

Due to the amount of heat dissipated by the LED, it is critical that the LED have the proper substrate for cooling the device. While the individual LED emitters are available from most manufacturers, it is recommended that prototyping be conducted with complete LED modules due to the stringent mounting and cooling requirements. The modules are constructed with the LED emitter mounted on a thick aluminum substrate which facilitates heat dissipation. Heat dissipation is furthered by mounting the modules to a heat sink (integrated into the PCB design), using 4-40 screws and a short length of hook-up wire.

LED emitters can be used in the finished design to reduce cost, however, this requires that the designer pay much closer attention to thermal dissipation considerations. Refer to the manufactures thermal dissipation guidelines for both LED modules and LED emitters (for examples, see reference 2 in **Section "References"**).

Firmware

The microcontroller firmware is responsible for several aspects of the system's operation.

1. Generating the V_{DRIVE} reference voltage for the driver circuit (see Figure 2).
2. Monitoring the battery voltage and adjusting the V_{DRIVE} reference voltage to prevent dimming during low battery conditions.
3. Disabling the LED drive when the battery voltage falls below the minimum safe operating voltage for the circuit.
4. Decoding and executing the various system commands input from the push button.
5. Decoding and executing the user selected pre-programmed Flash sequences.
6. Enabling and disabling the driver circuit in response to power on and off commands from the user.

The firmware is designed using a multi-tasking state machine design methodology that allows the system to accomplish the five separate tasks simultaneously.

MULTI-TASKING

Each task is implemented as a separate state machine with a system of variables to provide communications between the tasks. Figure 5 shows the various task state machines and the main communications paths. The state machines are called from a central control loop, which interleaves their execution, allowing each state machine execution time to perform its function.

KEY TASK

This task monitors the push button for the system, debouncing the switch and decoding three different key-press durations. These virtual key presses are the SHORT_KEY, the LONG_KEY and the HOLD_KEY. The SHORT_KEY is entered by a push button press of less than 1.5 seconds. The LONG_KEY is entered by a push button press of less than 3.0 seconds, but longer than 1.5 seconds. The HOLD_KEY is entered by a push button press greater than 3.0 seconds. The HOLD_KEY also automatically repeats every 3 seconds after the initial 3-second hold.

The debounce algorithm uses a counter with hysteretic open and close limits to filter the raw input from the button and remove any mechanical bounce. Once the debounced state of the push button is known, the task then decodes the input into the three virtual keys. Decoding of the push button input is based on the elapsed time that the push button is held. The decoding of the virtual keys is performed when the push button is released and the duration of the button press is determined. The one exception is the HOLD_KEY, which generates a virtual key press when the push button has been held for the first 3 seconds. Every 3 seconds after the first HOLD_KEY is decoded, another HOLD_KEY output is generated until the push button is released.

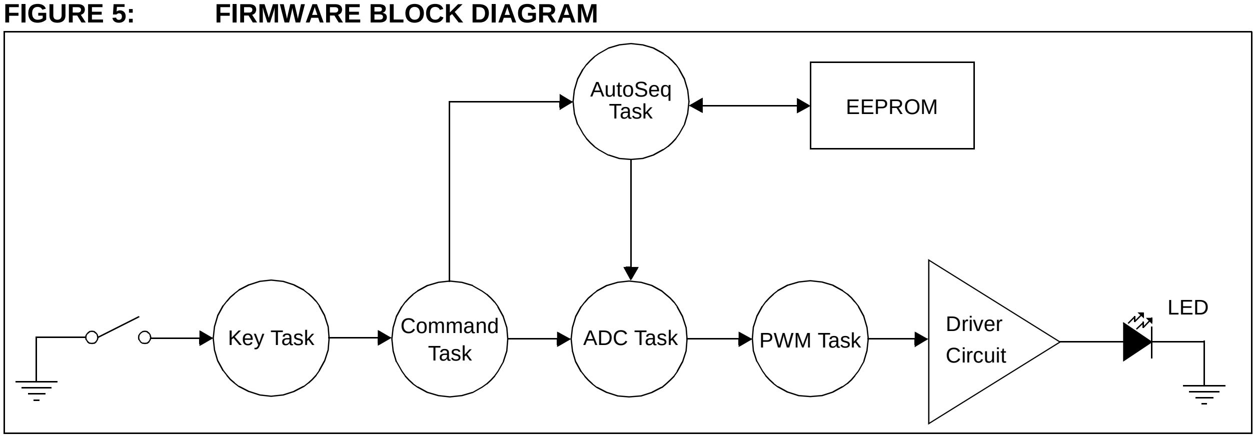
COMMAND TASK

The system has been designed to operate in five different modes. These modes determine the behavior of the system and identify the commands available to the user. Mode 1 is a continuous intensity mode and is analogous to a standard incandescent flashlight. Modes 2 through 5 select and execute one of the four pre-programmed Flash sequences programmed into the EEPROM.

The Command task interprets the virtual key input from the Key task and executes the appropriate system commands based on which mode is currently active. There are five possible commands:

- Increment Intensity Command
- Decrement Intensity Command
- Increment/Decrement Toggle Command
- Power-off Command
- Change Mode Command

FIGURE 5: FIRMWARE BLOCK DIAGRAM



Increment Intensity Command

The Increment Intensity command, only available in Mode 1, is selected by a SHORT_KEY input when the system is in Mode 1 and the increment direction is selected. The SHORT_KEY input will increment the current intensity by one step each key press. The range of intensity settings is 0 to 15, with 0 corresponding to no light output and 15 corresponding to full brightness. Once full brightness has been reached, however, additional Increment Intensity commands will be ignored until the brightness setting is reduced below full brightness.

Decrement Intensity Command

The Decrement Intensity command, only available in Mode 1, is also selected by a SHORT_KEY input when the system is in Mode 1. However, the decrement direction must be selected using the Increment/Decrement Toggle command prior to the Decrement Intensity command. As with the Increment Intensity command, the decrement in intensity will occur with the decoding of each SHORT_KEY input. The range of intensity is the same as the Increment Intensity command, and the Decrement Intensity command cannot decrement the intensity below 0.

Increment/Decrement Toggle Command

The Increment/Decrement Toggle command, only available in Mode 1, is selected by a LONG_KEY input when the system is in Mode 1. The Increment/Decrement Toggle command reverses the direction of any subsequent intensity commands (increment versus decrement). If the current direction for intensity commands is increment, an Increment/Decrement Toggle command will make all subsequent intensity commands decrement. If the current direction is decrement, the Increment/Decrement Toggle command will toggle the direction to increment. As there is no visible indication of the current direction (increment or decrement) the direction is always reset to increment at power-up and any mode changes.

Note:	The current intensity setting is retained through power-down and any mode change. The intensity setting will not affect any pre-programmed Flash sequence, but the intensity will be reset to the last valid setting upon return to Mode 1.

Power-off Command

The Power-off command, available in all modes, is a special condition of the Increment/Decrement Toggle command. Entering two Increment/Decrement Toggle commands in series, without another command in between, will select the Power-off command. Once executed, the driver circuit will be disabled and the microcontroller will be put to Sleep. To cancel a Power-off command, the user must press the system push button.

Note:	Decoding is not performed on the button press used to wake the system. This prevents inadvertent changes to the system intensity or mode on power-up.

Change Mode Command

The Change Mode command, available in all modes, is selected by a HOLD_KEY input. The first Change Mode command will turn off the LED, increment the System mode by 1, and display the new mode. After 3 seconds, the next Change Mode command will increment the System mode again and display the new mode. The Change Mode command will repeat this until the push button is released. If a Change Mode command is executed when the current mode is Mode 5, the system will roll over to Mode 1 and continue incrementing modes until the push button is released.

At each mode change, the system will Flash the LED to indicate the current active mode. The LED will Flash once for Mode 1, twice for Mode 2, and so on. The last mode change prior to the release of the push button is selected and will remain active until it is changed with another Change Mode command. The selected mode is also retained through a Power-off command, and will reset the system to the intensity or pre-programmed Flash sequence active prior to power-down, upon the next power-up.

ADC TASK

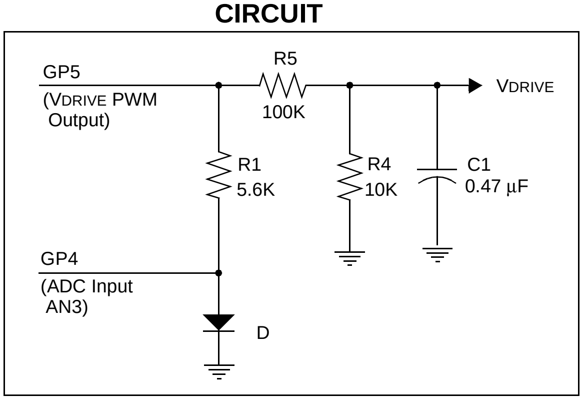
This task performs three functions; it monitors the battery voltage, compensates the VDRIVE PWM output for any reduction in the supply voltage, and shuts down the system when the microcontroller supply voltage drops below 3.5V. The battery voltage is determined by measuring the voltage drop across a silicon diode with the on-chip ADC. The voltage drop across the diode is fixed, however, the reference for the ADC is the supply voltage. As a result, as the supply voltage drops, the value measured for the voltage drop across the diode increases giving an inverse reading of the supply voltage.

The battery monitoring circuit is composed of R1, D, and the appropriate software (see Figure 6). To measure the battery voltage, the ADC task drives the VDRIVE PWM output high, momentarily. While the output is high, diode D is forward biased and the voltage across D is measured by the on-chip ADC. The VDRIVE PWM output is used to bias D so that during power-down, the VDRIVE output can be disabled and the current draw through the diode eliminated from the circuit.

The resulting ADC conversion value is then used to calculate a compensation constant. The compensation constant scales the duty cycle value presented to the PWM task. The result is a VDRIVE PWM output from the PWM task that does not change with power supply voltage shifts. The ADC task performs a conversion on the power supply 10 times a second. Each conversion generates a new compensation constant and the current duty cycle is scaled using the compensation constant. Changes in the intensity, either through a preprogrammed sequence or manual intensity change, can also trigger the ADC task to scale the duty cycle with the compensation constant.

The final function of the ADC task is to determine when the battery voltage falls below the 3.5V minimum operating voltage and shut down the system. To insure that a momentary drop in voltage due to switching in the driver circuit does not shut down the system, a minimum of 32 low battery events must be in sequence to cause the shutdown.

FIGURE 6: BATTERY MONITORING CIRCUIT



PWM TASK

The PWM task performs two functions for the system. It generates the VDRIVE reference voltage for the driver circuit, and it controls the timing of the other tasks in the system. The VDRIVE reference voltage is generated by the combination of software Pulse-width Modulation (PWM) routine and an external RC low-pass filter (see Figure 6). The timing for the system is determined by the roll over of TMR0 and a group of software timers (skip timers) which gate the execution of the other state machines.

The following list is a time-line of the PWM task's execution.

1. The task begins by waiting for the TMR0 interrupt flag T0IF (TMR0 rollover) and clearing it when it is set.
2. When TMR0 rolls over, the software PWM section of the task generates the appropriate width output pulse.
3. The task then decrements the various task skip timers, one for each of the other tasks in the system.
4. If the skip timer for a specific task decrements to zero, the skip timer is then reset to its default value and the associated task state machine enable flag is set.
5. The PWM task returns to the calling loop so other state machines can execute.

The VDRIVE signal is generated by a combination of software and hardware. Following the TMR0 roll over, the software determines whether the required duty cycle is greater or less than 50% and generates the shorter portion of the pulse. The output of the pulse is then passed through an RC filter formed by R4, R5, and C1. The filter smooths the PWM output into a DC voltage and attenuates the voltage to a level appropriate for the driver. Because the start of the pulse is tied to the TMR0 roll over, the period of the pulse is equal to the period of TMR0. In addition, because the software always generates the shorter portion of the pulse, the time remaining for the execution of the other tasks is certain to be at least 50% of the cycle.

The external hardware for the PWM task is composed of R4, R5 and C1, which act to smooth and attenuate the VDRIVE signal. The attenuation of the VDRIVE signal is necessary to reduce the VDD to VSS range output down to the expected feedback voltage from the driver. The supply voltage to the microcontroller, typically 3.5V-5.5V determines the maximum PWM output voltage. The maximum feedback voltage from the driver for the system is 0.35V. Therefore, a variable attenuation of 15.7:1, down to 10:1, is required to reduce the PWM output to the feedback level of the driver.

The resistor ratio of R5 to R4 generates an attenuation of 11:1, this ratio is not quite high enough for the supply of 5.5V and it is too high for a supply voltage of 3.5V. The discrepancies in the attenuation ratios are deliberate. For a supply of 3.5V to 5.5V (V_{BATT} = 4V to 6V), the 11:1 attenuation ratio leaves room for some of the attenuation to be accomplished in software by scaling the PWM value.

The ADC task performs the scaling based on the current battery voltage. By scaling the PWM value up as the battery voltage drops, the design can produce nearly constant output intensity for the life of the battery.

Because the 11:1 ratio is too high for the 3.5V supply, the software compensation of the intensity will reach its limit at a battery voltage of 4.3V, causing the intensity to dim during the last portion of the battery life. The dimming is an indication to the user that the batteries are nearly exhausted.

The other function that the PWM task performs is to regulate the execution rate of the other task state machines. It accomplishes this through a group of software timers called skip timers. The skip timers are decremented on each pass through the PWM task. When a skip timer reaches zero, it is reset to its default value and a flag enabling the associated state machine is set.

Because the skip timers run at fixed intervals, and because the default value for the skip timers are multiples of a common value, it is possible to limit the number of state machines that execute in a given pass by offsetting the initial values at Reset. Starting each timer with a 1-count, 2-count or 3-count difference from each other, ensures that only one of the skip timers will reach zero on a specific pass through the PWM task. This type of priority control is referred to as passive, due to the automatic nature of the offset.

Note: For a passive priority system to operate, the default value for the skip timers must be multiples of a common value. For example, the default values of 25, 55 and 75 are all multiples of the common value 5.

AUTOSEQ TASK

This task is responsible for decoding and executing all user selected pre-programmed Flash sequences (see the next section).

Automated Flash Sequences

The system has the ability to execute up to 4 automated Flash sequences stored in the on-chip EEPROM storage area. The sequences are stored as lists of 8-bit commands in the EEPROM memory and include commands to set intensity, delay fixed time periods, repeat a group of commands, jump to a specific command in the list, and power-down the system. Table 1 lists the values required for configuring an automated sequence in the microcontroller's EEPROM storage.

TABLE 1: EEPROM MEMORY MAP

Address	Value
0x00	Default Intensity
0x01	Current Mode (1 to 5)
0x02	Number of Modes (max = 5) 1 = Flashlight mode 2 to 5 = sequences 1 through 4
0x03	Start address of sequence 1
0x04	Start address of sequence 2
0x05	Start address of sequence 3
0x06	Start address of sequence 4
0x07	First command of sequence 1

The next section discusses each command in greater detail and outlines any limitations on the design of an automated Flash sequence.

AUTOSEQUENCE LANGUAGE

The autosequence language is comprised of 4 basic commands:

1. **Intensity Set**: Sets the LED intensity to 0 of 63 levels, (0 = off and 63 = full brightness).

   The Intensity Set command is a 0x00 plus the desired intensity (0x00-0x3F).
2. **Time Delay**: Delays the execution of the next command for up to 6.3 seconds, in 0.1-second increments.

   The Time Delay command is a 0x40 plus the desired delay in 1/10th's of a second (0x01-0x3F).
3. **Repeat/Return**: The commands create a loop in the sequence that repeats all instructions between the start (Repeat) and the end (Return). The enclosed sequence of commands can be repeated 1 to 63 times.

   The Repeat command is a 0x80 plus the desired number of execution cycles (0x01-0x3F).

   The Return command is a 0x80.
4. **GOTO/Shutdown**: Forces the system to start executing commands at line numbers 1 to 63. Executing a `GOTO 0` command will cause the system to shutdown.

   The `GOTO` command is a 0xC0 plus the line number of the destination command (0x01-0x3F).

   The Shutdown command is a 0xC0.

Any combination of the commands can be used to create an automated sequence (see Tables 1, 2 and 3). However, several system limitations must be observed or the system will execute erratically.

1. No command sequence may be over 63 commands in length (note the Repeat command requires two commands).
2. The total for all 4 sequences must be less than 120 commands because of the on-chip EEPROM limitations.
3. Repeat commands can be nested no more than 4 levels deep. Nesting additional Repeat commands will result in the loss of the highest level command.
4. `GOTO` commands must jump to valid line numbers in their own pre-programmed sequence. `GOTO` commands must not try to jump to another sequence.
5. It is recommended that any unused sequences should contain a Shutdown or `GOTO` command (`GOTO 0`) on the chance that the sequence is executed due to a system error.

TABLE 2: FLASH ON AND OFF AT A RATE OF 1-FLASH/SECOND

Line	Instruction	Data	Opcode
1	INTENSITY	63	0x3F
2	TIME DELAY	0.5	0x45
3	INTENSITY	0	0x00
4	TIME DELAY	0.5	0x45
5	GOTO	1	0xC1

TABLE 3: FLASH 3 TIMES AT 1-FLASH/SECOND, THEN DELAY 2 SECONDS

Line	Instruction	Data	Opcode
1	REPEAT	3	0x83
2	INTENSITY	63	0x3F
3	TIME DELAY	0.5	0x45
4	INTENSITY	0	0x00
5	TIME DELAY	0.5	0x45
6	RETURN	—	0x80
7	TIME DELAY	2.0	0x54
8	GOTO	1	0xC1

TABLE 4: FLASH S-O-S 4 TIMES A MINUTE

Line	Instruction	Data	Opcode
1	REPEAT	3	0x83
2	INTENSITY	63	0x3F
3	TIME DELAY	0.5	0x45
4	INTENSITY	0	0x00
5	TIME DELAY	0.5	0x45
6	RETURN	—	0x80
7	TIME DELAY	0.3	0x43
8	REPEAT	3	0x83
9	INTENSITY	63	0x3F
10	TIME DELAY	1.5	0x4F
11	INTENSITY	0	0x00
12	TIME DELAY	0.5	0x45
13	RETURN	—	0x80
14	TIME DELAY	0.3	0x43
15	REPEAT	3	0x83
16	INTENSITY	63	0x3F
17	TIME DELAY	0.5	0x45
18	INTENSITY	0	0x00
19	TIME DELAY	0.5	0x45
20	RETURN	—	0x80
21	TIME DELAY	2.0	0x54
22	GOTO	1	0xC1

Programming the System

Because of the very limited user interface for the system, a PC-based Graphical User Interface, or GUI, has been developed to aid in the programming of the pre-programmed Flash sequences. The GUI operates with the Microchip PICkit™ 1 Flash Starter Kit to program the EEPROM within the microcontroller in the system.

Note:	Creating a pre-programmed Flash sequence with the GUI is reasonably simple. However, a description of the GUI program's purpose is beyond the scope of this application note. Individuals interested in learning more can download the GUI program and review the Lumileds Driver Web Seminar.
	The Lumileds Driver Web Seminar is archived under the "20 Minute Web Seminars" link located on Microchip's web page.

To program a new Flash sequence:

1. Load the GUI interface program on a PC with Windows® 98SE or later, operating system loaded.
2. Connect the PICkit 1 programmer to the PC using a USB cable.
3. Create the desired Flash sequence using the interface buttons and windows.
4. Remove jumper 1 from the demo board.
5. Insert the board into the 14-pin header on the right side of the PICkit 1 programmer.
6. Program the board using the simboost.asm source file.
7. Remove the board and replace jumper 1.
8. Power-up the system.

Memory Usage

The total amount of memory used by the project's firmware is 800 words (1400 bytes) of program memory and 48 bytes of RAM. The amount of EEPROM required depends of the number and size of pre-programmed Flash sequences included in the firmware, minimum usage is 10 bytes.

Power Usage

In Sleep mode (Flashlight off), the microcontroller current draws less than 5 µA. If 4 alkaline AAA batteries are used, this means the shelf life of the circuit is equal to the shelf life of the batteries. Operating at full intensity, the life of 4 AAA batteries is 6-1/2 hours.

Efficiency

In operation at 0% light output, the circuit draws less than 800 µA. At full intensity, the circuit draws an average current of 190 mA, with a DC LED current of 285 mA and a forward average voltage of 3.4V. This results in a driver efficiency of 85%.

EQUATION 5:

$$Efficiency = \frac{285\ mA \times 3.4V}{190\ mA \times 6V} = 85\%$$

CONCLUSION

The system described in this application note shows a minimal part count control for a high-power LED driver based upon a simple buck regulator topology. Using the on-chip comparator in the microcontroller decreases cost and adds the ability to implement custom features.

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FIGURE 7: BUCK LED DRIVER BOARD LAYOUT

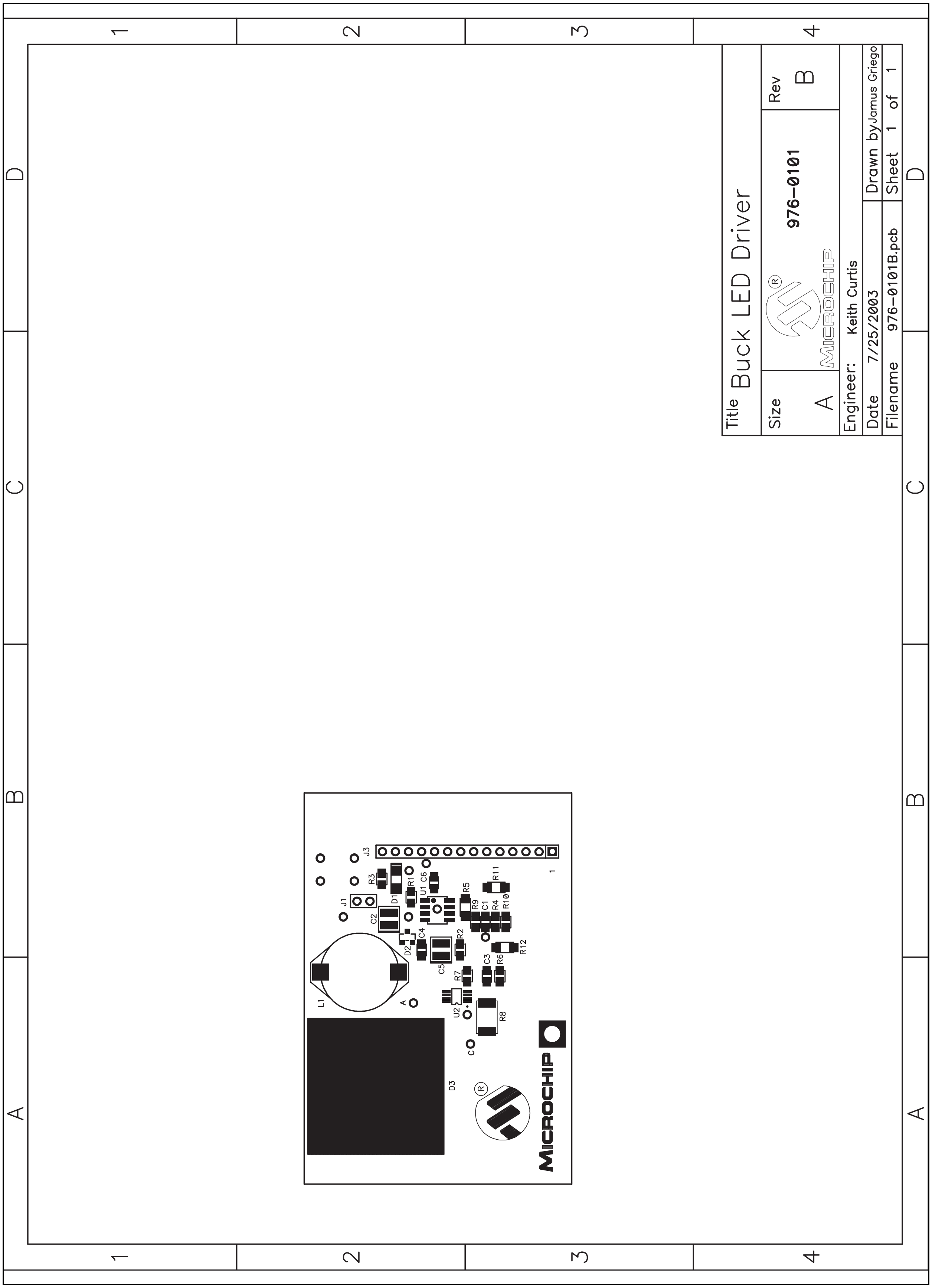
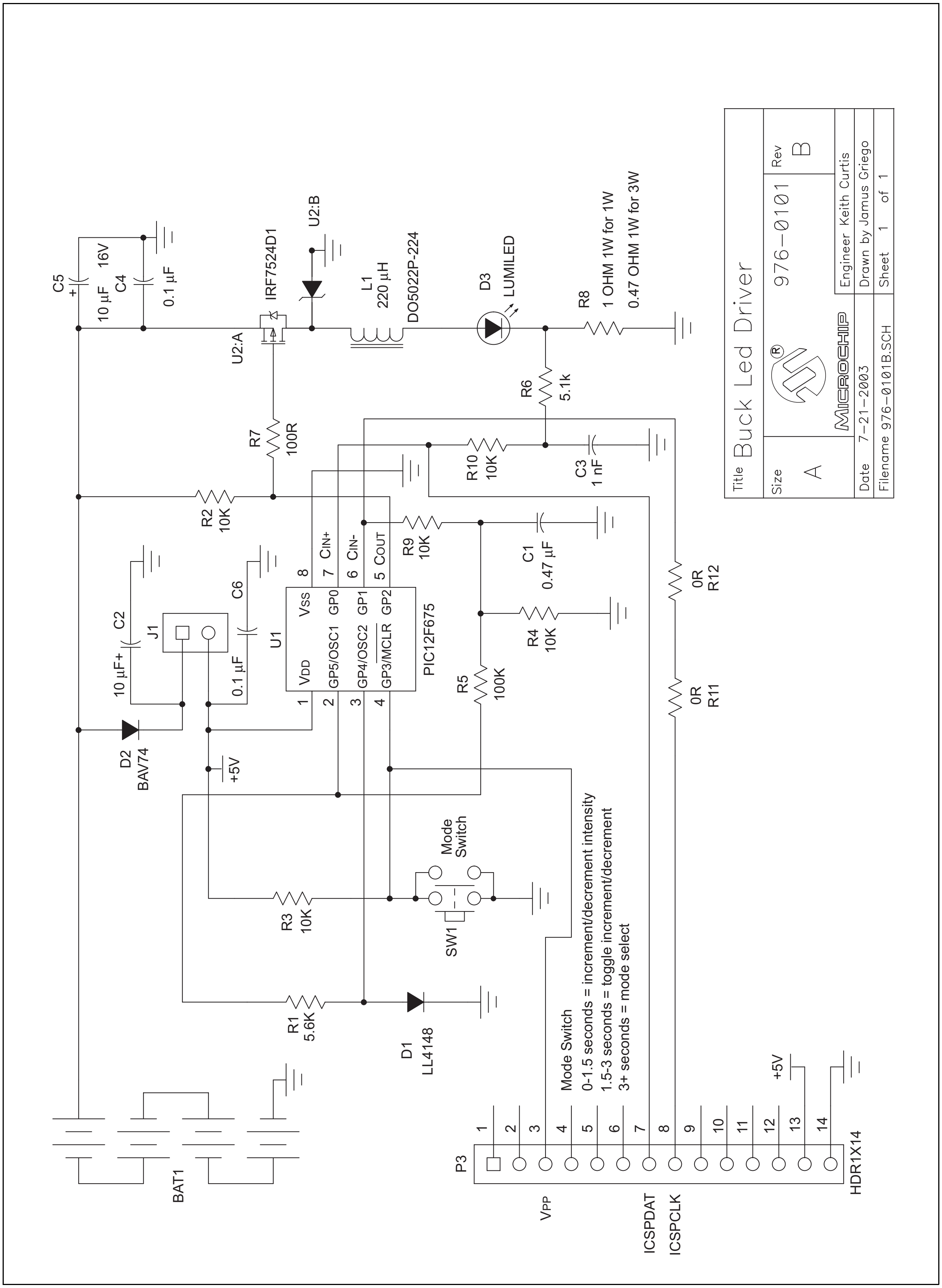


FIGURE 8: BUCK LED DRIVER SCHEMATIC



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