ANALOG DEVICES

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APPLICATION NOTE

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Driving High Power LEDs

by Yuxin Li, Gang Liu, and Alan Li

INTRODUCTION

The ADP3806 is a switching mode power supply (SMPS) controller featuring dual loop constant-voltage and constant-current control; remote and accurate current sensing; and shutdown and programmable, synchronizable switching frequency. For different application requirements, it can be configured in a variety of topologies: buck, boost, buck-boost, SEPIC, and CUK. This application note guides the designs of controller circuits for driving high power LEDs (light emitting diodes) by using the ADP3806 to achieve an efficiency of up to 95%.

Before using this application note for designing the circuits, download the ADP3806 data sheet from the ADI website.

POWERING UP THE ADP3806

The minimum VCC is 6.25 V (undervoltage lockout UVLO voltage) and the maximum VCC should not exceed 23 V, which brings the switch driver voltage BST to 30 V, the junction breakdown voltage. The recommended VCC range is from 6.5 V to 20 V.

To ensure a clean voltage source at the VCC pin, an RC bypass network is used between the input source and the IC. The decoupling capacitor can be in the value range of 0.1 μF to 22 μF. The shutdown control pin $\overline{SD}$ accepts external control logic input. If automatic startup is preferred, a voltage divider from VCC can be used to ensure that the voltage at this pin is below its limit, 10 V, and above its logic high, 2.0 V.

A capacitor connected to the CT pin sets the switching frequency. Recommended switching frequency is from 300 kHz to 750 kHz for the optimized trade-off between the overall system physical size and the efficiency. Higher switching frequency of up to 1 MHz demands more gate-driving power and generates more switching losses, resulting in lower efficiency, while the inductors' inductance can be scaled down to reduce the system's physical size and cost. Another drawback of using a higher switching frequency is that the duty cycle range is reduced so that the output voltage range is narrowed.

REG, REF, and BSTREG pins are output pins of three internal low dropout (LDO) regulators. Decouple them with the recommended capacitors to ensure the stability of the regulators.

The current sensing resistor along with the voltage at the ISET pin determines the output current. The required voltage at the ISET pin should be generated from the 2.5 V precision reference source REF pin. The maximum output current from the REF pin is 500 μA. To generate a higher setting voltage, the voltage divider can run off the REG pin (6 V). Note that the accuracy for REF is 1%, and the accuracy for REG is 3%.

Across CS+ and CS– pins, a filter capacitor of about 220 nF should be placed right next to the pins in the PCB layout to filter out noise.

BUCK CONFIGURATION

To drive LEDs with an output voltage lower than the input power source voltage, a step-down topology Buck can be used.

Figure 1 shows the Buck configuration circuit. This scheme utilizes the synchronous rectification functionality of the controller to provide the highest power conversion efficiency. The controller and the power stage can be powered up by either the same or different power sources. Using the same power source, the input voltage range should be within the range of 6.5 V to 20 V. With different power sources, the power stage input voltage can be as low as 6.5 V, while the maximum should not exceed 20 V.

Using proper MOSFETs, the output current can go up to 4 A, the sense resistor value, and the ISET voltage. The output voltage can go from zero up to the power stage input voltage.

The current sensing resistor, R_{CS}, should be low enough to minimize its power loss but high enough to provide sufficient signal for the control. The voltage across the CS+ and CS– pins should be higher than 50 mV to activate the synchronous rectification function, that is, to switch the lower MOSFET on and off. When (CS+ – CS–) voltage is lower than 50 mV, the low-side MOSFET is not turned on so its body diode conducts, causing extra conduction loss. At the same time, the energy for driving the lower MOSFET is saved. When the output current is lowered far enough, turning off the lower MOSFET may result in higher overall system efficiency.

By using the circuit shown in Figure 1, the output current equals

$$I_{LED} = \frac{V_{ISET}}{25R_{CS}}$$

If the current sensing signal in the design is lower than 50 mV, a bias circuit, shown in Figure 2, is needed to fully activate the synchronous functionality. By using a voltage divider composed of RB1 and RB2, a bias voltage, V_{BIAS}, is generated at the CS+ pin. Thus, the sensing signal needed to activate the synchronous rectification function is V_{BIAS} lower than the 50 mV. If V_{BIAS} equals 50 mV, the synchronous function is effective all the time. With this bias, the output current is

$$I_{LED} = \frac{V_{ISET} - 25V_{BIAS}}{25R_{CS}}$$

The current limit setting accuracy is obviously a little lower with this added offset term. Also, the biasing circuitry should draw minimum current from the VREF pin, and filtering capacitor C4 should now be designed to make a small enough time constant (recommend 0.3/f_{SW}) with RB2 instead of R_{CS}. Thus, C4 is much smaller than 200 nF.

The output capacitor, C_{OUT}, is optional. It smoothes the voltage and current across the LEDs.

The input capacitor, C_{IN}, is needed to absorb the input ripple current. Normally, two 10 μF/25 V ceramic capacitors are enough.

The inductor, L, determines the ripple current. It is recommended that the ripple current be about one-third of the nominal output current to optimize both the system size and the efficiency. The peak inductor current is therefore,

$$I_{LPEAK} = 1.15I_{LEDM}$$

where:

I_{LPEAK} is the peak current of the inductor.

I_{LEDM} is the maximum average current in the LEDs.

The inductor should be chosen based on two criteria. First, the inductor saturation current should be larger than the maximum peak current, I_{LPEAK}, in the inductor. Second, the inductor's maximum dc current rating should be larger than the load dc average current, I_{LED}.

The recommended bootstrap capacitance is 10 nF. A 0.5 A Schottky diode is recommended as the bootstrap diode. This diode must be placed on the PCB as close to the BST and BSTREG pins as possible.

C_{CH}, C_{CL}, and R_C compose the loop compensation network of the closed loop. They need to be designed carefully to ensure system stability and control speed.

The control (duty cycle, d) to inductor current (i_L) transfer function is a constant given by

$$\frac{i_L(S)}{d} = \frac{DV_{IN}}{25R_{CS}f_{SW}L}$$

where:

D is the duty cycle of the steady state.

f_{SW} is the switching frequency of the power stage.

This is valid below half of the switching frequency and with the assumption that the output voltage is constant.

It is obvious that a simple integrator is adequate to compensate the loop with infinite dc gain. To set a bandwidth of $f_C(\omega_C)$, the compensation component C_{CL} needs to be

$$C_{CL} = \frac{g_M DV_{IN}}{25R_{CS}f_{SW}L\omega_C}$$

where:

g_M = 0.002 is the transconductance of the current loop error amplifier.

R_C can be set to zero.

C_{CH} can be set to be open.

LEDs are not critical to the driving transition, so the system loop bandwidth can be low. It is recommended that the bandwidth is designed to be 1/30th of the switching frequency for the buck configuration.

MOSFETs can be chosen based on their dc voltage/current ratings, switching speed (gate charges), and thermal capability.

Note that the BSTREG regulator needs to provide the entire gate-driving energy to drive both the high-side MOSFET and the high-side driver itself. Because the maximum output current, I_{BSTREG}, of the BSTREG regulator is ≤ 3 mA, the gate charge, Q_G, of the high-side MOSFET and the switching frequency should satisfy

$$I_{BSTREG} > Q_G f_{SW}$$

If more gate-driving capability is needed, use an external NPN transistor, QBST, in the emitter follower configuration (see Figure 2) to deliver more current.

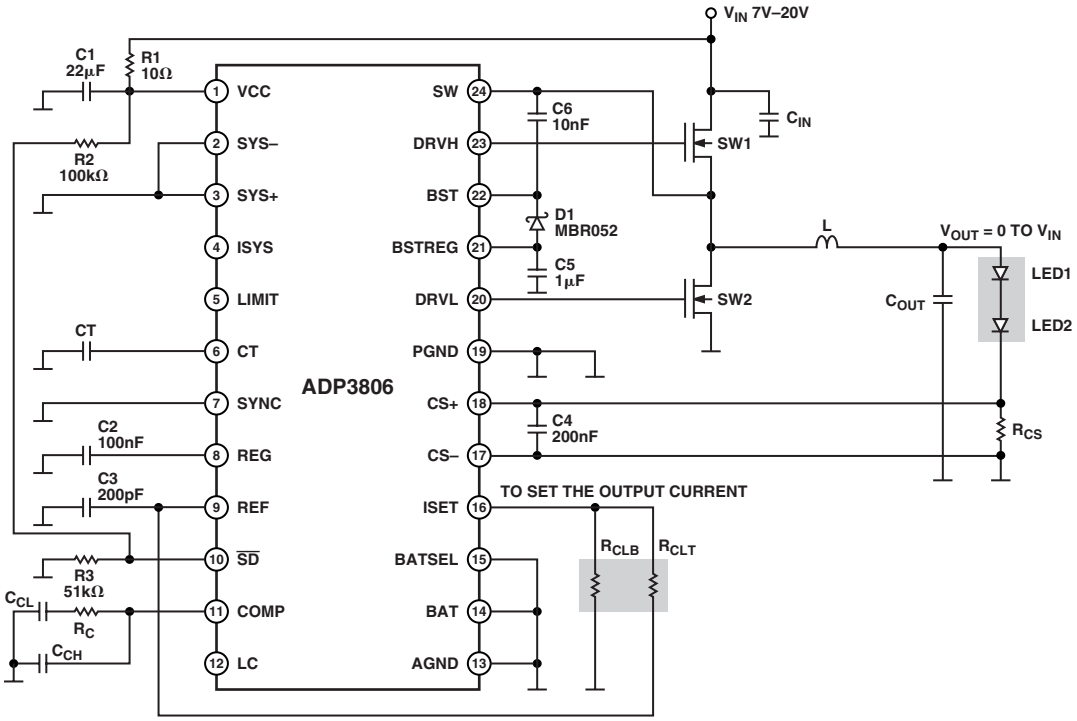

Figure 1. Driving LEDs with the ADP3806 in Buck Configuration

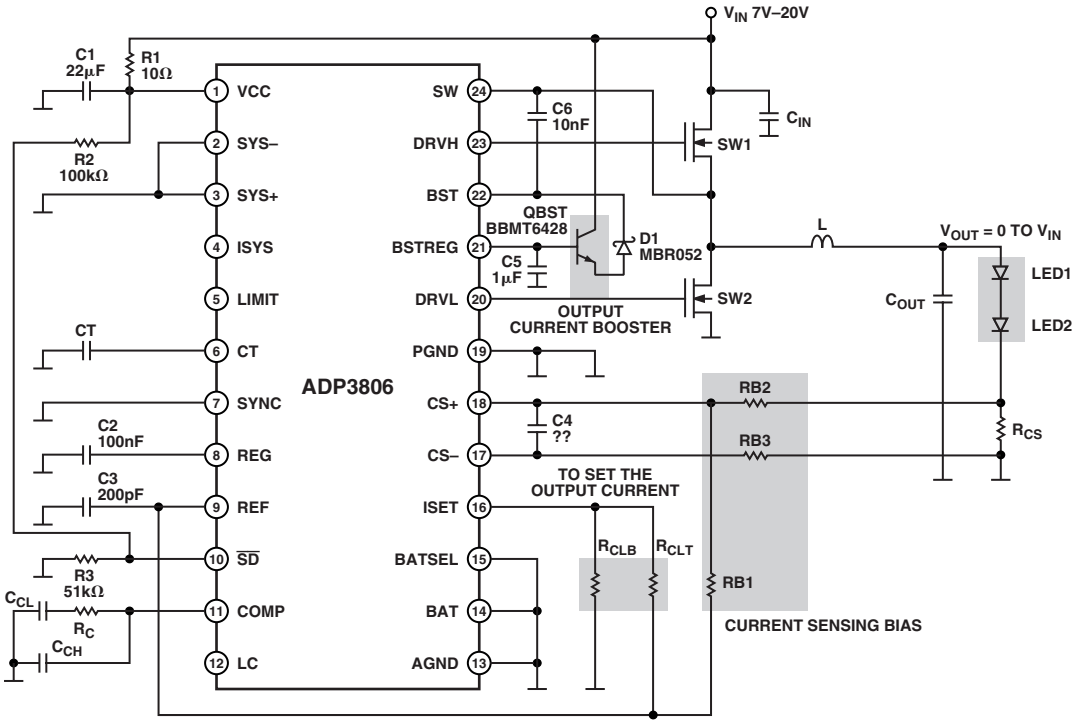

Figure 2. Biasing the Current Sensing Input to Accommodate Low Input Signal

BOOST CONFIGURATION

If the output voltage is higher than the input power source voltage to power a series of LEDs, step-up conversion is used.

Figure 3 shows the boost configuration using the ADP3806 to drive LEDs. This scheme utilizes the main driving output DRVH to control the main switch SW1, with the diode switch SW2 used for rectification.

To avoid overvoltage output, the voltage loop can be connected to set the maximum allowable output voltage, V_{OUTM}. Because the internal reference is 2.5 V, the maximum output voltage is set by

$$V_{OUTM} = 2.5\frac{R_{OVPT} + R_{OVPB}}{R_{OVPB}}$$

Choose appropriate voltage divider resistance to set the overvoltage protection point, V_{OUTM}.

The output capacitor, C_{OUT}, is optional. Without it, the current flowing into the LEDs is discontinuous and fluctuates at a duty of (1 – D), where D is the duty cycle of the PWM control. With the output capacitor, the current in the LEDs can be shaped smoother. Appropriate output capacitance is needed to achieve the required driving current pattern.

The current sensing resistor, R_{CS}, needs to be large enough to generate sufficient signal for the controller. The maximum differential input voltage between CS+ and CS– is 160 mV.

C_{CH}, C_{CL}, and R_C need to be designed carefully to ensure system stability and control speed. This design is different from that of the buck configuration.

With this configuration, the output is always connected to the input power supply via SW2 and the inductor, even when the controller is in shutdown mode. This may pose some issues for specific applications.

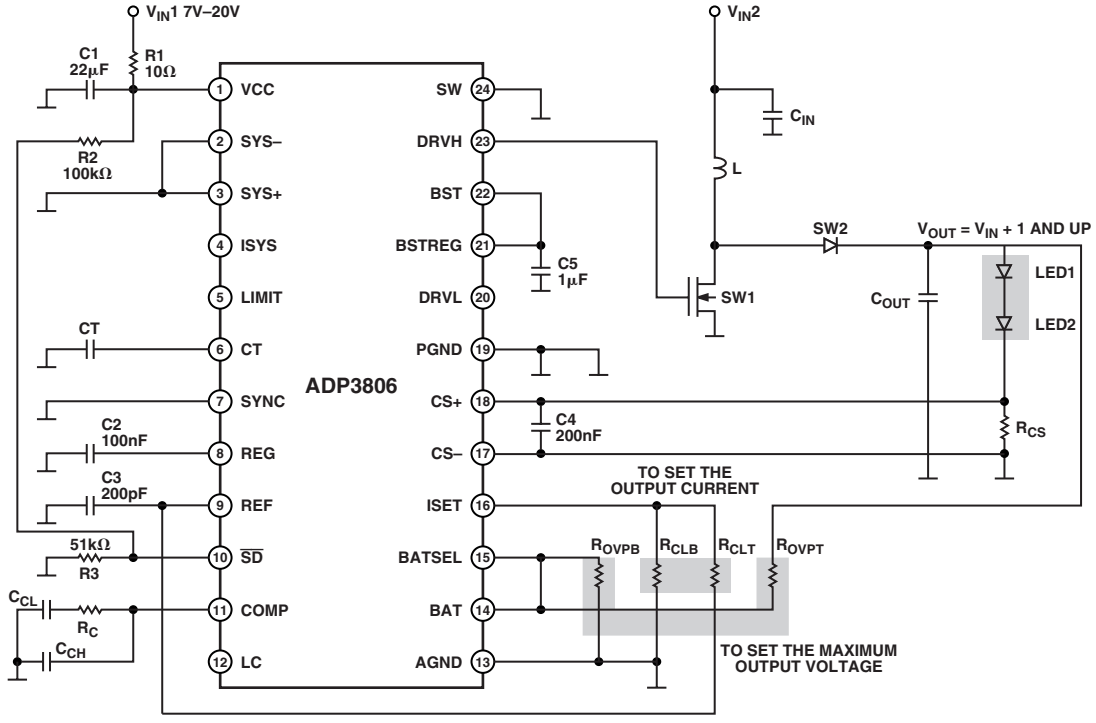

Figure 3. Driving LEDs with the ADP3806 in Boost Configuration

SEPIC CONFIGURATION

The SEPIC configuration provides an output that can be either higher or lower than the input voltage at a cost of an additional inductor and capacitor. A SEPIC configuration using the ADP3806 to drive LEDs is shown in Figure 4.

The output stage is very similar to the boost configuration, except that the rectification switch, SW2, conducts when the main switch, SW1, is on.

The output capacitor, C_{OUT}, is optional.

C_{CH}, C_{CL}, and R_C need to be designed carefully to ensure system stability and control speed. This design is different from that of the buck configuration.

Unlike the boost configuration, the output is isolated to the input source when the controller is in shutdown mode.

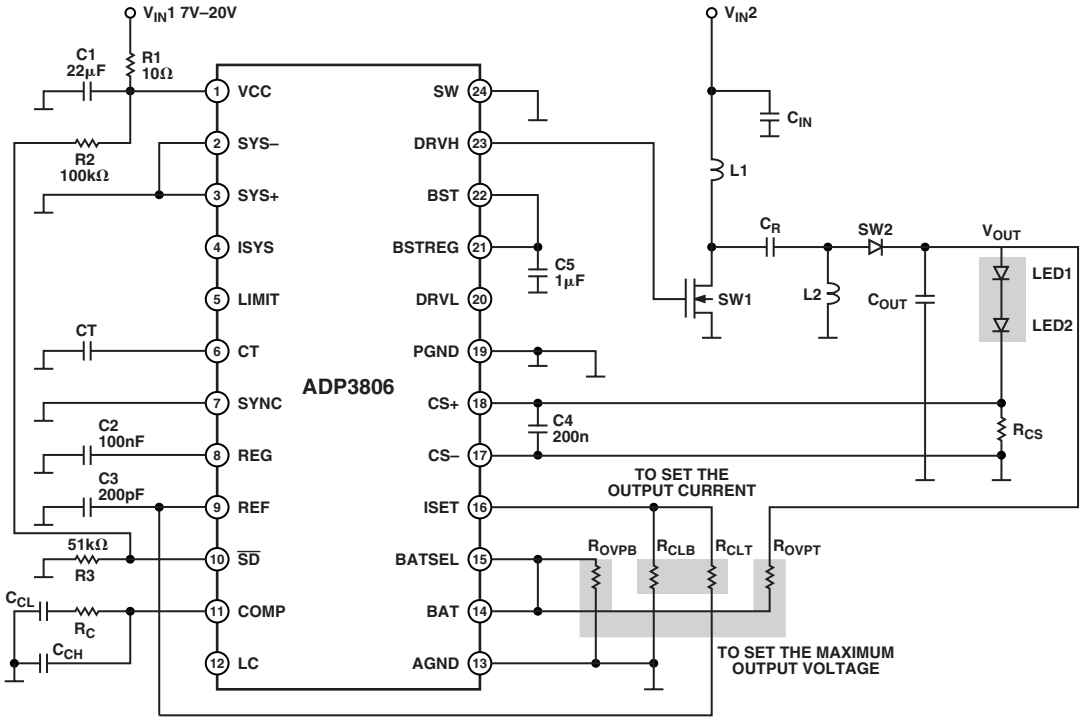

Figure 4. Driving LEDs with the ADP3806 in the SEPIC Configuration

CUK CONFIGURATION

The CUK configuration provides an inverted output that can be either higher or lower in magnitude than the input voltage. Figure 5 shows the CUK configuration using the ADP3806 to drive LEDs.

Similar to the buck configuration, the output current is continuous, with fewer pulsating components compared to both the boost and the SEPIC configurations.

Similarly, the output capacitor, C_{OUT}, is optional.

C_{CH}, C_{CL}, and R_C need to be designed carefully to ensure system stability and control speed. This design is different from that of the buck configuration.

The output is isolated to the input source when the controller is in shutdown mode.

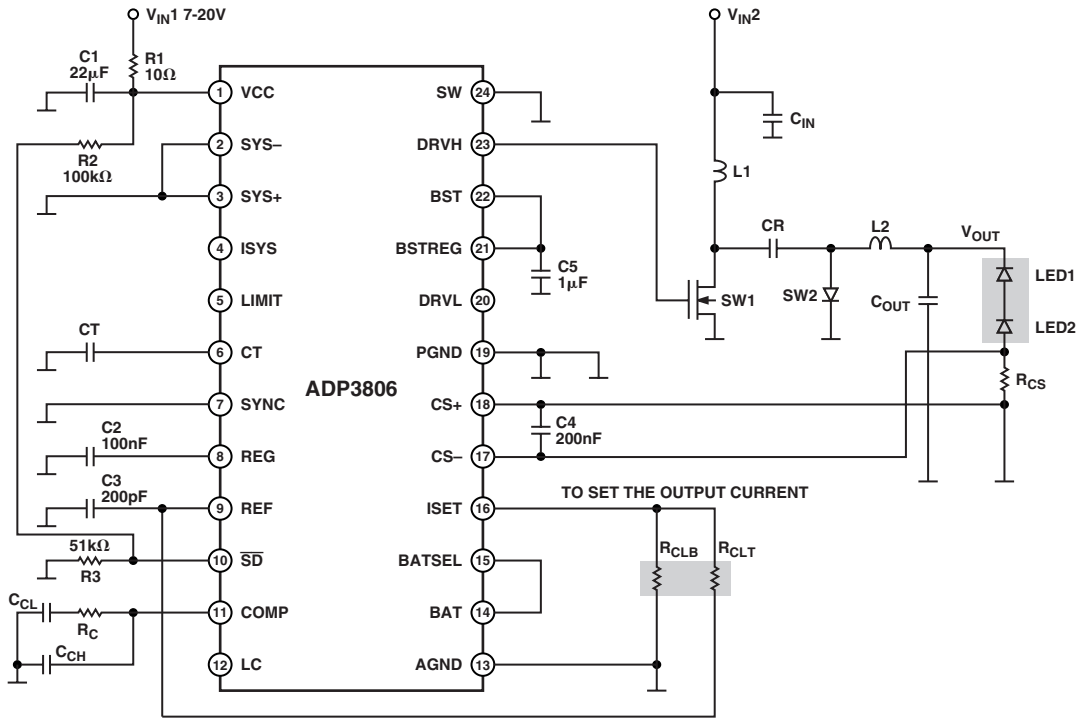

Figure 5. Driving LEDs with the ADP3806 in the CUK Configuration

BRIGHTNESS CONTROL

Brightness control can be implemented easily by controlling the voltage at Pin ISET. Referring to the data sheet, the output current, I_{OUT}, is set by V_{ISET} (voltage at the ISET pin) by

$$I_{OUT} = \frac{V_{ISET}}{25R_{CS}}$$

There are two basic ways to adjust the brightness control. The first is to control the output current by changing V_{ISET} at the ISET pin. This can be done simply by using a potentiometer running off the 6 V V_{REG} rail or 2.5 V V_{REF} reference.

To minimize the board space, layout inflexibility, and reliability concern, users may consider using a digital potentiometer, AD5228, rather than its mechanical counterpart for the brightness control. AD5228 is a low cost, 32-step, manual control digital potentiometer that features a built-in debouncer and zero-scale/midscale selectable preset. As shown in Figure 6, an AD5228 with two external push buttons adds in the circuit flexibility with brightness control.

This control is simple but may be inefficient because the efficiency of LEDs tends to be higher at their full current rating and lower at reduced current. To overcome this problem, a PWM brightness control scheme can be used.

By periodically shorting the ISET pin down to AGND, V_{ISET} shows a pattern of PWM, as does the output current, I_{OUT}. With an open-drain MOSFET connected to the ISET pin, the brightness of the output can be controlled by a digital signal, which can be generated by a microcontroller.

It takes time for the controller to react to the control signal, so the frequency of this brightness control PWM signal should be much lower than the bandwidth of the control loop, which is much slower than the switching frequency. The brightness PWM signal can be in the range of 50 Hz to 500 Hz.

Figure 7 shows a typical high brightness LED's efficiency curve versus its driving current. I_{OPT} is the optimal driving current point with the highest light emitting efficiency. The efficiency optimized brightness control approach is to drive the LEDs with a PWM current pattern. The current changes from zero to I_{OPT} at a duty D_{BRI}. This leads to the maximum light-emitting output at the continuous I_{OPT} driving. When more light output is needed, raise the driving current.

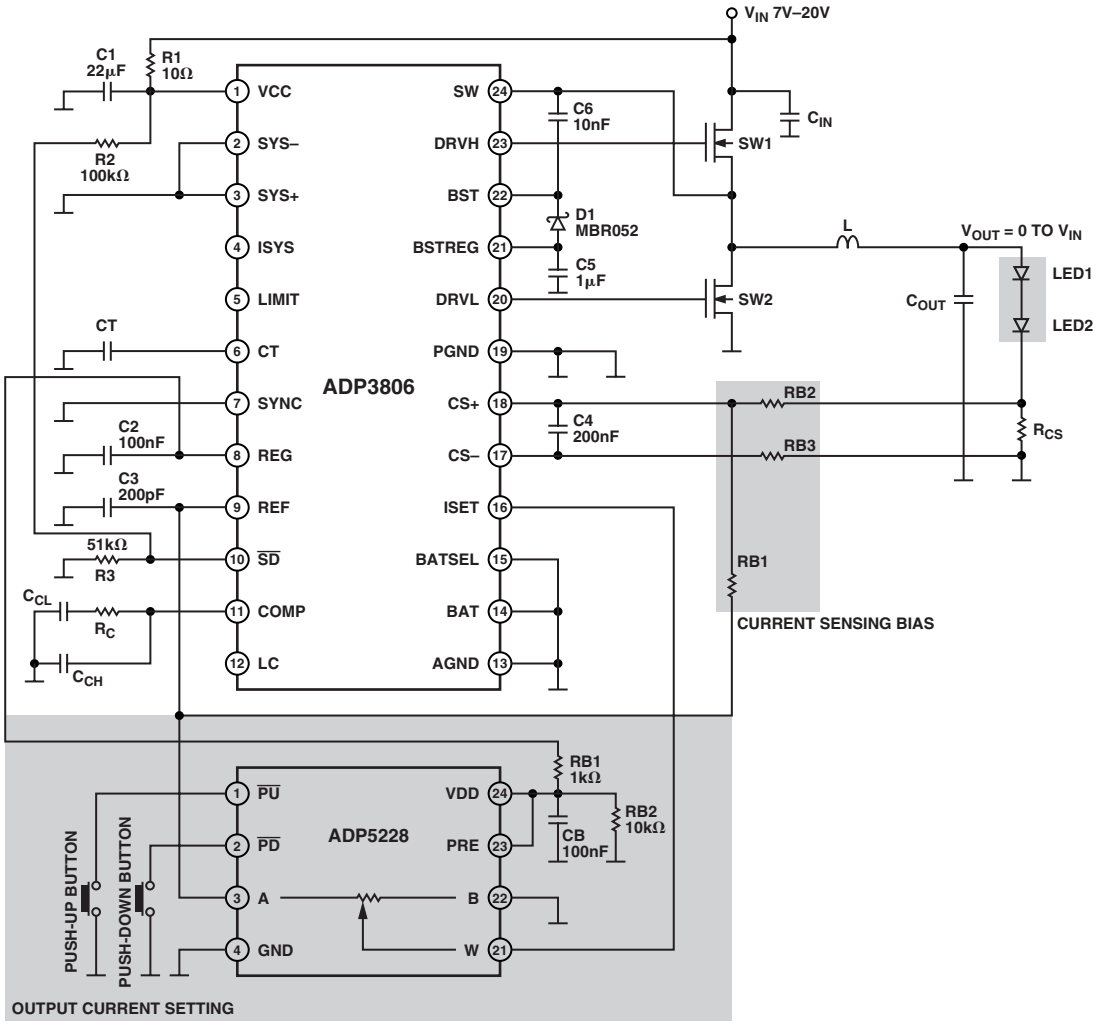

Figure 6. Brightness Control Using AD5228 Manual Control Digital Potentiometer

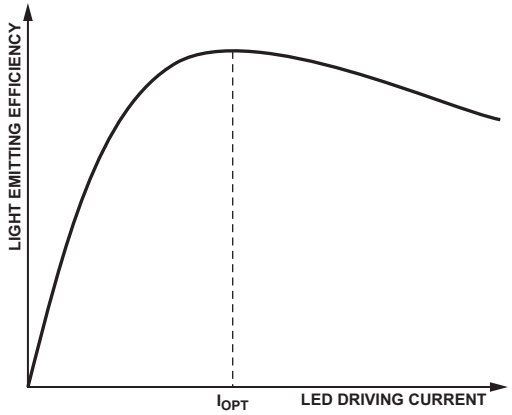

Figure 7. Typical High Brightness LED's Efficiency

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