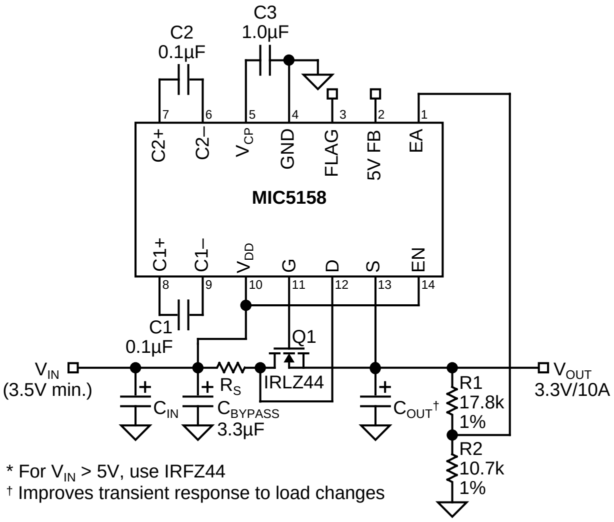
Application Note 25

High-Side Switch Using LDO Regulator Controller

by Jerry Kmetz

General Description

The Micrel MIC5158 Super LDO™ Voltage Regulator Controller can be used as a high-side switch driver that features a relatively accurate current limit. The part is normally intended for driving the MOSFET pass device of a voltage regulator. It also has constant-current output limiting (see Figure 1, current sensing resistor R_S) which becomes active when the drop across R_S is approximately 35mV. The similarity that this circuit shares with high-side MOSFET drivers is that it includes a charge-pump for driving N-channel MOSFETs. An advantage is that its current limit is more accurate than that of most high-side drivers.



* For V_{IN} > 5V, use IRFZ44
† Improves transient response to load changes

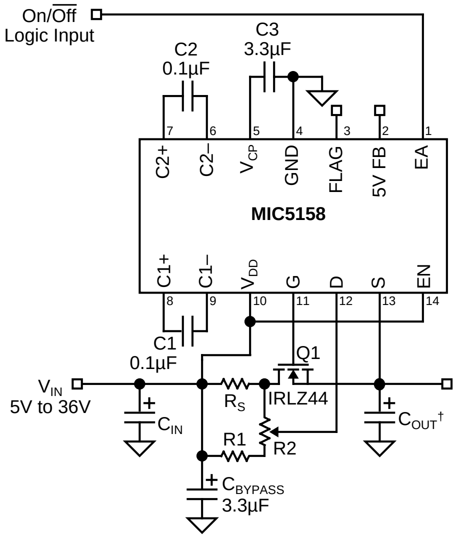
$$R_S = \frac{0.035V}{I_{LIMIT}} \qquad V_{OUT} = 1.235V\left(1+\frac{R1}{R2}\right)$$

Figure 1. Super LDO™ Voltage Regulator

MIC5158-Based High-Side Driver

Figure 2 illustrates a high-side driver that offers adjustable (trimmable) current-limit. Note the similarity to the voltage regulator circuit of Figure 1. This circuit normally operates with a fully enhanced (saturated) MOSFET because, unlike the voltage regulator, there is no feedback voltage divider. The controller then acts like a switch, driving the MOSFET gate fully on or fully off in response to a digital control signal applied to the error amplifier (EA) feedback input pin. A logic "1" input greater than V_{REF} (1.235V) turns the switch off. This provides much faster switching than using the voltage regulator enable (EN) input because the EN input disables charge pump operation, which also requires some time to generate the required gate drive voltage. Otherwise, the EA and EN inputs may be used interchangeably. Faster switching is consistent with reducing power dissipation in the MOSFET (switch element) and the higher switching speeds needed for pulse-width modulation applications. If the on/off switching signal is via the EN pin, the EA pin in Figure 3 will act as a disable if a logic-high level is applied.

The nominal threshold value for V_{LIM} is 35mV, but its range is from 28mV to 42mV (35mV±20%). Resistor R1 and potentiometer R2 are needed only if this relatively wide spread of V_{LIM} does not support the required accuracy for a given current-limit value.



$R_S = \frac{0.042V}{I_{LIMIT}}$ R2 trims I_{LIMIT} $V_{OUT} = V_{IN} - I_{OUT} R_{DS(on)}$

Figure 2. High-Side MOSFET Driver

Resistor Calculations

Given the tolerances involved, and to provide a useful adjustment range for R2, sense resistor R_S should be designed to provide a voltage drop at least as large as the maximum specified value for V_{LIM} (42mV). Because the drain input (D) of the MIC5158 is high impedance, R1 and R2 can be several orders of magnitude larger than R_S, which makes their contribution to the current-sensing resistance negligible.

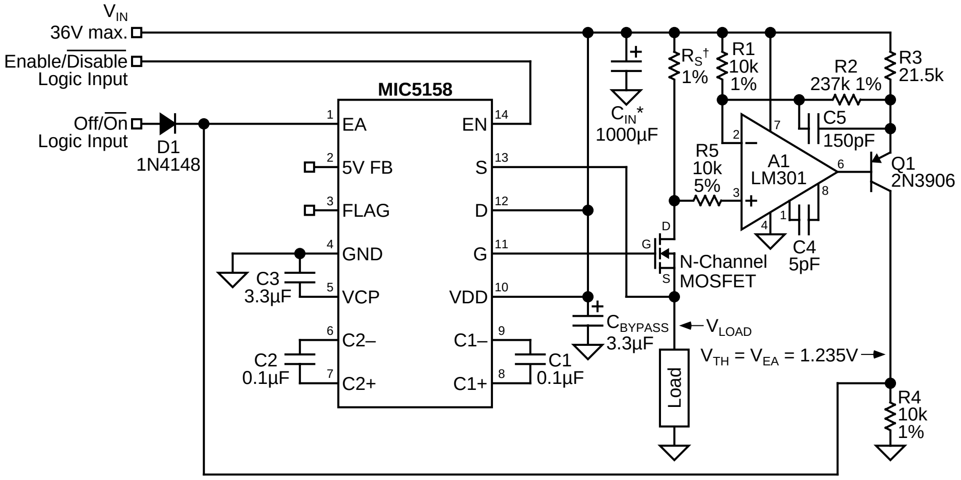
Then:

$$R_S = \frac{V_{LIM(max)}}{I_{LIMIT}}$$

where:

$V_{LIM(max)} = 0.042V$

I_{LIMIT} = desired current-limit value.



* C_{IN} should be large enought to prevent excessive sagging of the input voltage during turn-on transients.
† R_S sets the current-limit value. Limiting occurs when approximately 100mV is dropped across R_S.

Figure 3. Adjustment-Free High-Side MOSFET Driver

Since:

$$V_{R2(drop)} = V_{LIM(max)} - V_{LIM(min)} = 0.042V - 0.028V = 0.014V$$

$$V_{R1(drop)} = V_{LIM(min)} = 0.028V = 2 \cdot V_{R2(drop)}$$

then:

$$R1 = 2 \times R2$$

R_S determines the current-limit value; the adjustment range provided by R1 and R2 accommodates the range of parameter V_{LIM}. Resistor R2 should be adjusted by driving a test load impedance sufficiently low to assure current limiting

Adjustment-Free High-Side Driver

If additional current-limit setting accuracy is required and the calibration adjustment provided by potentiometer R2 in Figure 2 is an unacceptable inconvenience, the circuit of Figure 3 eliminates the need for this adjustment. The internal current-limit circuitry is defeated by connecting the MIC5158 drain input (pin 12) directly to V_{IN} (V_{DD}, pin 10). The current-limit function is now provided by external components: an operational amplifier, some precision resistors, and a bipolar transistor. The selected op amp offers a common-mode range that includes the positive supply, a feature needed by this circuit. The amplifier circuitry provides non-inverting gain for the voltage developed across current-sensing resistor R_S. When the drop across R_S is 100mV (V_{SENSE}) the gain

$$A_V = 1 + \frac{R2}{R1} = 24.7$$

is such that a 2.47V drop is produced across R3. Level translation by transistor Q1 and resistor R4 inverts and references this signal to ground. The base-drive current of Q1 is negligible, so essentially all of the emitter current also flows through R4 and develops the required threshold voltage ($V_{TH} = V_{EA} = 1.235V$) for the current-limit control loop. When the voltage across R4 attempts to exceed V_{TH} the loop will provide reduced gate drive to the MOSFET, increasing its $R_{DS(on)}$, and thereby establishing constant-current limiting. Capacitors C4 and C5 are required for stability.

Upon comparing the values of R3 and R4 one may observe their values are not a simple R-2R relationship (which might have been assumed from the above description of their common current and their 2:1 voltage drops). R3 is larger because it is paralleled by the sum of R1 and R2; the value of this combination is 19.78kΩ (within about 1% of 20kΩ).

Switch Performance

The switching characteristics of the basic circuit of Figure 3 are shown in Figure 4. Note that there is a response-time aspect to the current-limiting behavior. Because of system bandwidth limitations, the output voltage, and load current, overshoot the desired setpoint for approximately 50µs. Since this is a very brief interval, and will represent a very low duty-cycle in many applications, this behavior will be well tolerated by many types of loads. The overshoot does not appear as a very low duty cycle in Figure 4 because it was also desired to illustrate that the MOSFET can be switched at a relatively high rate: on and off times shown here are approximately 500µs.

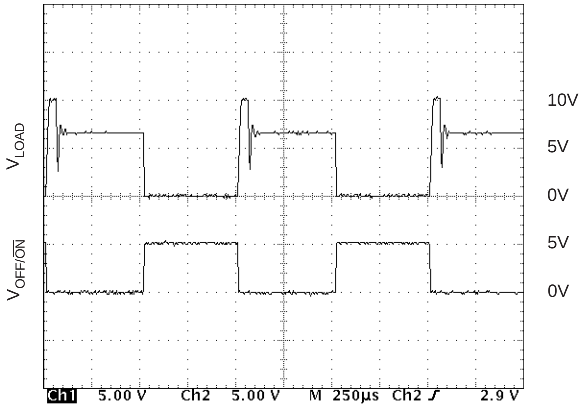
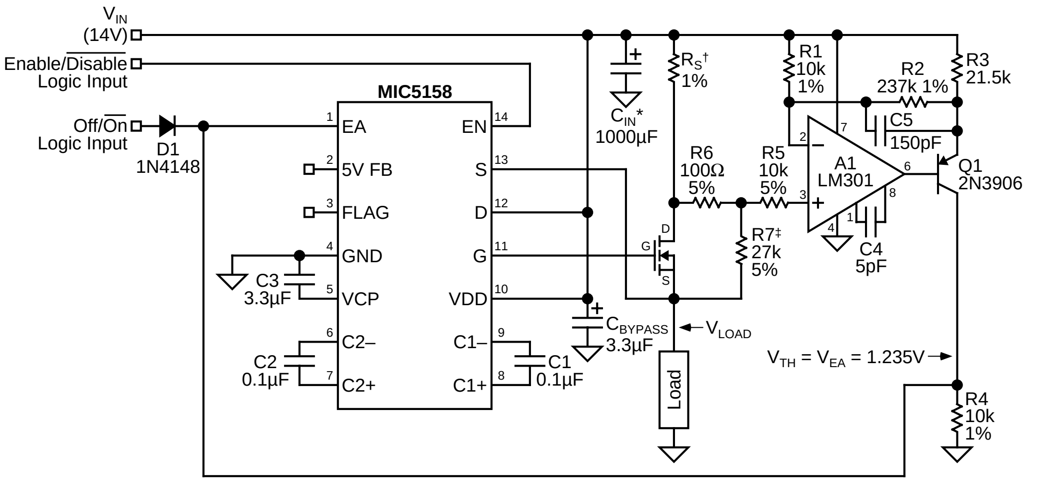


Figure 4. Switch Waveforms



* C_{IN} should be large enought to prevent excessive sagging of the input voltage during turn-on transients.

† R_S sets the current-limit value. Limiting occurs when approximately 100mV is dropped across R_S.

‡ $R7 = \left(\frac{V_{IN}}{V_{R6}}\right)R6 \quad V_{R6} = \frac{0.1V}{\left(\frac{I_{LIMIT}}{I_{SHORT}}\right)}$; see text.

Figure 5. Power Switch Featuring Foldback Current Limiting

Providing Foldback Current Limiting

The addition of resistors R6 and R7 to the circuit, as shown in Figure 5, adds the feature of foldback current limiting. This helps to further protect the MOSFET switch in the event of a shorted output. Foldback current limiting allows for more efficient utilization of the switch or pass element safe operating area (SOA) by reducing short-circuit output current. With a constant-current limiting scheme, such as provided by the circuits of Figures 1 through 3, the MOSFET sustains higher power dissipation in a short-circuited condition than in a foldback-limited scheme.

The circuit of Figure 5 is essentially identical to that of Figure 3, but with two additional resistors (R6 and R7). When the system is not in current limiting, the voltage drop across R_S is less than V_{SENSE} (100mV) and the MOSFET source voltage (output voltage) is nearly as high as the MOSFET drain voltage, less the small amount equal to $I_{OUT} \times R_{DS(on)}$. However, if the voltage drop across the MOSFET channel becomes significant, such as would occur if the output voltage was shorted or if the load impedance was decreased, a fraction of the drain-to-source voltage is added to the voltage dropped across R_S. Voltage divider R6 and R7 attenuates the drain-to-source voltage; the voltage drop across R6 is additive with the voltage drop across R_S. The result is a larger input to the current-limit control amplifier and a correspondingly lower current-limit value.

Foldback Characteristic

Figure 6 illustrates the nature of the foldback current limiting provided by the component values shown in the circuit of Figure 5. Note that the system provides a constant-voltage output until the current-limit threshold is crossed. When in current limiting the output voltage is reduced. The amount of reduction is inversely proportional to load impedance; the greatest reduction (to zero volts) will occur with a shorted output (zero ohms). Unlike constant-current limit schemes, a foldback current-limit circuit includes output voltage as part of the control signal driving the current-control loop. The net result produces the characteristic shown in Figure 6.

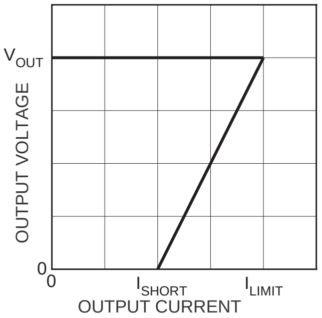


Figure 6. Foldback Current-Limiting Characteristics of Figure 5

As in Figure 3, the initial current-limit (I_{LIMIT}) is established by choosing R_S to generate a V_{SENSE} of 100mV across R_S. Then $R_S = 100mV/I_{LIMIT}$. The design value for I_{SHORT} should be chosen as a function of how much power the MOSFET and its heat sink can *safely* dissipate. In the example of Figure 5, the design goal was

$$I_{SHORT} = \frac{I_{LIMIT}}{2} \; (\textit{when } V_{IN} = 14V).$$

An I_{LIMIT} / I_{SHORT} ratio of 2 is typical; making this ratio too large can cause a start-up (latched-off) problem. It is important to note that this circuit is sensitive to input voltage; if the input voltage surges high enough the foldback current-limit capability could potentially shut down the output for the duration of the surge. Whether or not this is a desirable situation it does reduce power dissipation in the MOSFET.

To implement the desired I_{LIMIT} / I_{SHORT} ratio, attenuator R6 and R7 is designed to generate a specific voltage drop across R6 in a shorted-output condition. In Figure 4 the desired I_{LIMIT} / I_{SHORT} ratio is 2.

Let:

$$R6 = 100\Omega$$

$$V_{SENSE} = 100mV$$

set:

$$V_{R6} = \frac{V_{SENSE}}{\left(\frac{I_{LIMIT}}{I_{SHORT}}\right)} = \frac{100mV}{2} = 50mV$$

and:

$$V_{R7} = \frac{V_{IN}}{V_{R6}} \cdot R6 = \frac{14V}{50mV} \cdot 100\Omega = 280 \cdot 100\Omega = 28k\Omega$$

Since R6 drops 50mV, R_S now has only to drop another 50mV before current-limiting begins. The net result is that output current is limited at $I_{LIMIT} / 2$ if the output is shorted. The circuit of Figure 5 uses 27kΩ (closest standard value) for R7. This impact is slight.

Since:

$$V_{R6} \approx \frac{14V}{27.1k\Omega} \cdot 100\Omega = 52mV$$

and:

$$V_{SENSE} = 100mV - V_{R6} = 100mV - 52mV = 48mV$$

Then:

$$\frac{I_{SHORT}}{I_{LIMIT}} \propto \frac{48mV}{100mV} = 0.48 \qquad \text{(only 4\% error).}$$

Power Switch Featuring Overvoltage Protection

The circuit of Figure 7 combines the *constant-current* limited high-side driver of Figure 3 with the feedback voltage divider of Figure 1 to provide a voltage-clamped output power switch. Such a circuit may be useful in automotive applications, where transient overvoltage conditions exist that can be detrimental to the circuitry being powered.

When the input voltage (V_{IN}) to this circuit is less than a user-specified $V_{OUT(max)}$, feedback voltage divider R6 and R4 contributes less than V_{REF} (1.235V) to the error amplifier (EA) input of the MIC5158. The circuit then operates like a switch (MOSFET saturated) and the output voltage essentially follows the input voltage. When the input voltage equals or exceeds the specified $V_{OUT(max)}$, the circuit then operates like a voltage regulator and its line-regulation capability protects the load circuit from the input overvoltage condition.

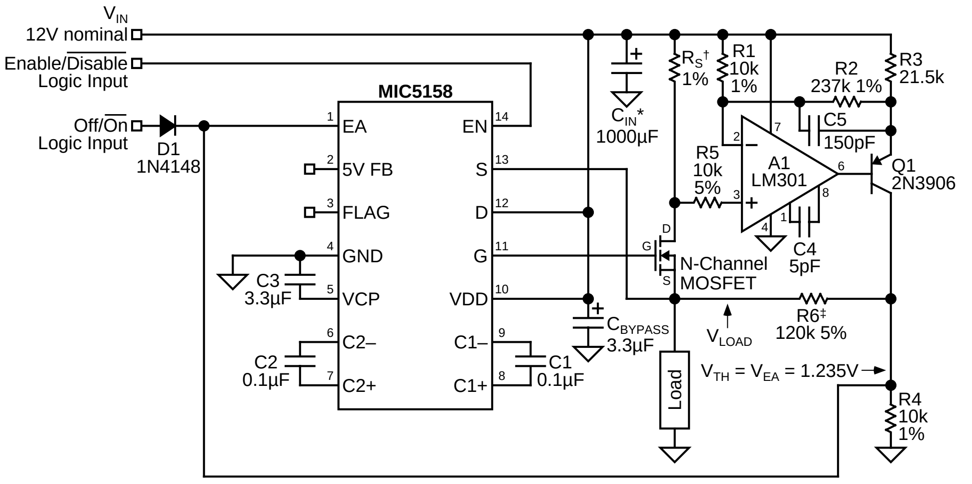
The voltage divider is designed in the familiar manner:

$$V_{OUT(max)} = V_{REF}\left(1 + \frac{R6}{R4}\right)$$

Substituting 10kΩ for R4 and 1.235V for V_{REF} yields:

$$R6 = 1 \times 10^4 = V_{REF}\left(\frac{V_{OUT(max)}}{1.235 - 1}\right)$$

Using 16V for V_{OUT}(max) in the above equation produces 120kΩ for R6. Since R4 is shared with the current-limit circuitry some interaction between the input overvoltage protection and the current limit functions is to be expected. The closer the output voltage is to $V_{OUT(max)}$ the less collector current is required from Q1 to produce $V_{TH} = V_{EA} = 1.235V$ across R4. The net result is a lower current-limit value.



* C_{IN} should be large enought to prevent excessive sagging of the input voltage during turn-on transients.

† R_S sets the current-limit value. Limiting occurs when approximately 100mV is dropped across R_S.

‡ $$R6 = 1 \times 10^4\left(\frac{V_{OUT(max)}}{1.235V - 1}\right)$$

Figure 7. Overvoltage Protected Power Switch

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