

Feedback Circuit Improves Hysteretic Control

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Modifying the conventional hysteretic converter circuit's feedback network results in a predictable switching frequency with less variation relative to external components.

The hysteretic power supply is simpler than the voltage- or current-mode closed-loop-control dc-dc switchers, although its simplicity may be a bit deceiving due to component variations and potential sources of "injected" feedback voltage.

Along with its simplicity, the hysteretic power supply is popular today for its low cost, inherently stable operation (with no need to perform loop analysis), extremely fast load response time that requires no compensation, and suitability for applications requiring low dropout because the main switch is a p-channel power MOSFET that can be driven up to 100% duty cycle.

The hysteretic controller IC is the key element in the conventional hysteretic power supply. And, even better results are possible with a simple modification to the conventional hysteretic controller IC circuit. That modification requires only the addition of a single capacitor and resistor

in the feedback circuit of the controller IC. To describe the modification, we will first look at the performance of the conventional hysteretic controller.

Principles of Operation

As shown in Fig. 1, at the heart of the hysteretic controller IC is a comparator with a small amount of voltage hysteresis (V_{HYS}). When the comparator feedback voltage exceeds the internal reference voltage (V_{REF}) plus its hysteresis, the comparator output turns off the main power switch. Then, when the feedback voltage drops below V_{REF} minus V_{HYS} , the power switch turns on and the cycle repeats itself. Thus, exceeding the hysteresis voltage determines the switching frequency and overall power-supply performance.

For predictable switching frequency operation within a desired range, the IC's comparator needs a reasonably clean and well-controlled triangular ramp voltage superimposed on the dc feedback voltage, which essentially matches the comparator's hysteresis. Unfortunately, this feedback signal can vary significantly due to the input-supply voltage, output capacitance (C_{OUT}), C_{OUT} equivalent series resistance (ESR), inductor value and board layout.

Board layout is critical with any switching supply, and this is perhaps even more important with hysteretic controllers due to high sensitivity to ground and voltage feedback noise that can directly affect switching frequency. Careful adherence to recommended board layout is essential to avoid drastically varying operating frequencies that may result in excessive output-voltage ripple and poor regulation. Additionally, variations in external components, especially C_{OUT} and its ESR, can cause large

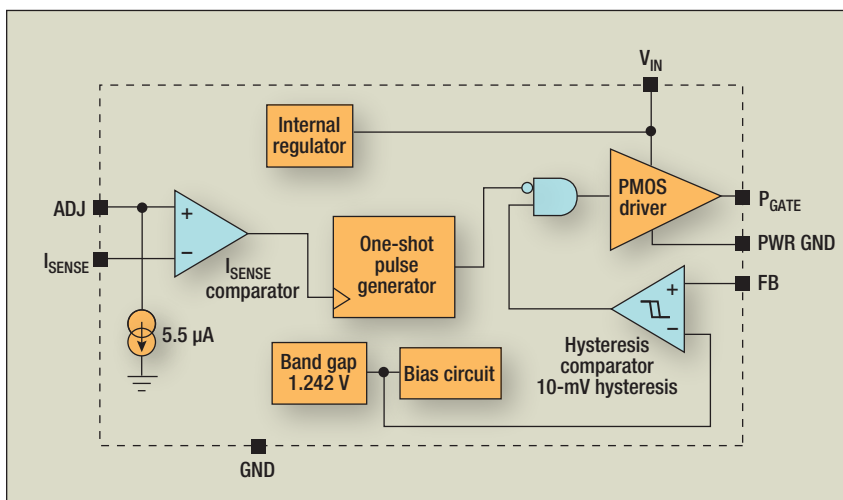


Fig. 1. The LM3485, a high-efficiency PFET switching regulator controller IC, employs a hysteretic control architecture that does not require any control-loop stability concerns with a wide variety of external components.

shifts in operating frequency. Fortunately, there are some techniques that can circumvent these pitfalls.

Referring to **Fig. 1**, a typical hysteretic buck controller IC includes a comparator, PMOS driver, one-shot, short-circuit protection and inrush control, and does not require an oscillator to control the switching frequency. The one-shot limits the duty cycle and external component power dissipation during an overcurrent event. Basic operation requires feedback ripple voltage (**Fig. 2**) to be compared with an internal V_{REF} (1.242 V) to toggle the high-side PMOS driver whenever the comparator threshold exceeds its 10-mV hysteresis. The conventional hysteretic controller generates this ripple voltage by using the inductor ripple current conducting through C_{OUT} and some amount of ESR.

Common methods to create ripple voltage in **Fig. 2** include:

- Rely on ESR inherent in C_{OUT}
- Insert an external ESR in series with C_{OUT} that is sufficiently larger.
- Insert a known series resistor between C_{OUT} and inductor.

Determining Switching Frequency

The drawback to using only the ESR of C_{OUT} is that its value is not well controlled and may vary significantly. Adding a series resistor to C_{OUT} adds to V_{OUT} ripple, and inserting it between C_{OUT} and inductor reduces efficiency. Furthermore, low resistance values (e.g., 100 mΩ) add cost. When using any of the previously mentioned methods, the feed-forward capacitor (C_{FF}) is recommended to reduce output ripple voltage by providing an ac-coupling path from V_{OUT} to the feedback pin. This reduces V_{OUT} ripple required by the controller because feedback resistors R1 and R2 will not divide the output ripple voltage down. Eq. 1 estimates the switching frequency for methods 1, 2 and 3:

$$\text{Frequency} = \frac{V_{OUT}}{V_{IN}} \times \left\{ \frac{(V_{IN} - V_{OUT}) \text{ESR}}{[(V_{HYS} \times \alpha \times L) + (V_{IN} \times T_D \times \text{ESR})]} \right\}, \quad (\text{Eq. 1})$$

where α equals $(R1+R2)/R1$ and equals 1 when using C_{FF} , T_D equals the approximate comparator delay of 110 ns and L equals the inductor inductance (henries).

Several factors determine the switching frequency in Eq. 1. For example, C_{OUT} and its ESR are dominant factors that impact the switching frequency. Changes in C_{OUT} from part to part, C_{OUT} trace inductance or additional capacitance at the point of load alter the effective ESR. Calculations based on Eq. 1 are rough estimations, and design engineers must also empirically determine circuit operation with a known set of components, board layout and operating conditions. The problem is that all variations are not easily accounted for during testing. The design challenge is not just multiple variables, but sensitivity to variable tolerances.

Fortunately, a new design approach can improve switching frequency operation and take much of the guesswork

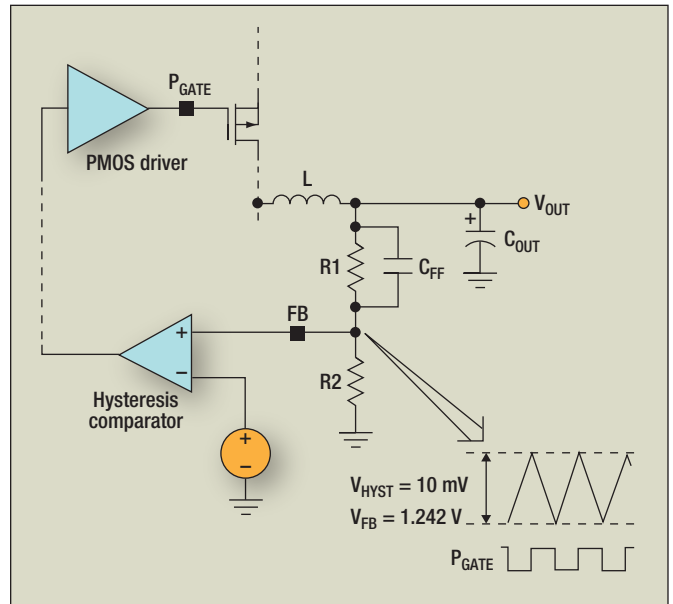


Fig. 2. After inductor ripple current passes through the output capacitor (C_{OUT}), ripple voltage then passes through the feed-forward capacitor (C_{FF}) and is reduced by the R1 and R2 voltage divider.

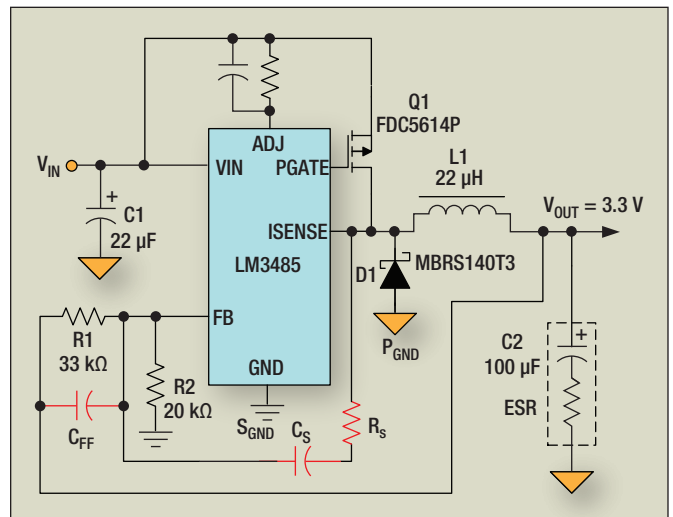


Fig. 3. The combination of R_S and C_{FF} take the voltage swings from the output switch node and produce an emulated ripple voltage that is similar to the triangular ramp required for a conventional hysteretic circuit operation. This improves the hysteretic converter's possible switching frequency variation.

and unpredictability out of designing a hysteretic switching regulator. Instead of using the net output ripple voltage created in the final output capacitor, use the source that produces ripple voltage beforehand. In a sense, we can use an “emulated” output ripple voltage by taking the voltage swings from the switch node that produces the same triangular ramp required for operation as shown in **Fig. 2**. The buck regulator circuit in **Fig. 3** uses the LM3485 hysteretic controller and highlights the components in red that emulate V_{OUT} ripple voltage.

Instead of relying on output ripple voltage at C2 and feeding this back to the comparator, the switch node volt-

V_{IN} (V)	Frequency (kHz)	Frequency deviation (%)
8	134	-16.3
10	149	-6.9
12	156	-2.5
13.7	160	—
16	166	3.8

Table 1. Operating frequency versus the input voltage for the traditional hysteretic converter with a 3.3-V output and a 10- Ω load (expected frequency = 376 kHz).

age creates a current source. During the on time of Q1, a constant current of $(V_{IN} - V_{OUT})/R_S$ charges C_{FF} , and during Q1 off time, C_{FF} ramps down from the current sink of approximately V_{OUT}/R_S to produce the required ramp signal based on $I = Cdv/dt$. Choose capacitor C_{FF} so its impedance is much less than feedback resistor R1 at the desired switching frequency. Capacitor C_S serves as an ac coupling path and its size is approximately 20 times the value of C_{FF} . Note that Eq. 2 no longer includes the output ESR or inductor value, and is derived by analyzing the integration components and comparator at a given duty cycle (D) determined by V_{IN} and V_{OUT} . Rearranging this equation and solving for R_S in Eq. 3 helps to set the desired frequency using R_S as a dependent variable:

$$\text{Frequency} = \left[\frac{D(1-D)(V_{IN} - V_{OUT})}{C_{FF} \times V_{HYS} \times R_S \times (1-D) + T_D(V_{IN} - V_{OUT})} \right] \quad (\text{Eq. 2})$$

$$R_S = \frac{\left\{ \frac{D(1-D)}{\text{Frequency}} - T_D(V_{IN} - V_{OUT}) \right\}}{[C_{FF}(V_{HYS})(1-D)]} \quad (\text{Eq. 3})$$

An example using an LM3485 evaluation board will help solidify the design process and operating performance. Measurements in **Table 1** were done using an unmodified board (**Fig. 3** without C_S or R_S and with $C_{FF} = 100$ pF). The V_{IN} range was chosen to be 8 V to 16 V, V_{OUT} equals 3.3 V and output load equals 10 Ω . The output load has a negli-

V_{IN} (V)	Frequency (kHz)	Frequency deviation (%)
8	312	-8.2
10	335	-1.5
12	342	0.6
13.7	340	—
16	332	-2.3

Table 2. Operating frequency versus the input voltage for a modified hysteretic converter.

gible effect on the switching frequency, provided the circuit maintains the continuous mode. If I_{OUT} is less than one-half the inductor ripple current, the discontinuous mode will cause the frequency to reduce as required.

Another benefit of the LM3485-type controller is that under light loads, very low regulator quiescent current is achievable since the p-channel MOSFET only switches when necessary without the need for a charge pump or boot-strap circuit.

With this design, no additional ESR other than that inherent in C_2 (C_{OUT}) is used to set the switching frequency. Although frequency deviation with respect to V_{IN} is acceptable, it is important to emphasize that very wide variations will occur from board to board mainly because of variations in the C_{OUT} ESR and inductor tolerance. For example, if C_{OUT} is a Sanyo 6TPC100M 100 μ F with a rated ESR of 45 m Ω at 100 kHz, then from Eq. 1 the calculated frequency is 376 kHz, where α equals 1, ESR equals 45 m Ω , V_{HYS} equals 10.5 mV, T_D equals 110 ns, L equals 22 μ H, V_{IN} equals 13.7 V and V_{OUT} equals 3.3 V.

The measured frequency is actually 160 kHz because C_{OUT} ESR is much lower than rated and it varies with temperature, further widening the frequency spread. **Table 1** shows the variation in switching frequency for input voltages from 8 V to 16 V. Here, the frequency deviation percentage is relative to 160 kHz.

Emulated Ripple Voltage Example

Next, we modified the same board to include emulated ripple voltage circuitry with C_{FF} , C_S and R_S as shown in **Fig. 3**. For example, calculated values of C_{FF} , C_S and R_S are for the desired switching frequency of approximately 330 kHz, and measurements were taken to confirm the results.

For the design procedure:

- Set C_{FF} -based feedback resistor R1 (33 k Ω) to be $\ll$ impedance at 330 kHz.
- Select C_{FF} equals 2.2 nF for an impedance of $1/(2\pi F_c) = 219 \Omega$.
- C_S equals approximately 10 to 20 times C_{FF} , so choose 68 nF, which is not critical.
- Duty cycle can be calculated based on a volts-seconds balance across the inductor for on and off cycles (measured at 26% for nominal V_{IN} of 13.7 V and V_{OUT} of 3.3 V).
- Calculate required R_S for desired frequency using V_{HYS}

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equals 10.5 mV and T_D equals 110 ns, from the LM3485 data sheet using Eq. 3.

Note that a T_D of 110 ns is primarily the propagation delay of the LM3485 comparator. For a larger input capacitance of the selected MOSFET, between 10 ns to 20 ns may need to be added:

$$R_s = \frac{\left[\frac{0.26(1-0.26)(13.7-3.3)}{330 \text{ kHz}} - 110 \text{ ns}(13.7-3.3) \right]}{2.2 \text{ nF}(10.5 \text{ mV})(1-0.26)}$$

(Eq. 4)

$$R_s = 287 \text{ k}\Omega$$

As shown in **Table 2**, test results of the same evaluation board, but with modified circuitry, came very close to what was expected, using R_s equals 287 k Ω , C_s equals 68 nF and C_{FF} equals 2.2 nF.

In this design, the C_{OUT} ESR does not appreciably impact the switching frequency. Actually, we don't require this type of capacitor when using the emulated ripple voltage method, and a smaller and less-expensive ceramic capacitor with very low ESR between 10 μ F to 22 μ F is preferable. Note that if an output capacitor with too large an ESR is selected, it will tend to increase the frequency from the summation of emulated and feedback ripple voltage.

Table 2 shows the variation in switching frequency for input voltages from 8 V to 16 V. The frequency deviation percentage is relative to 340 kHz.

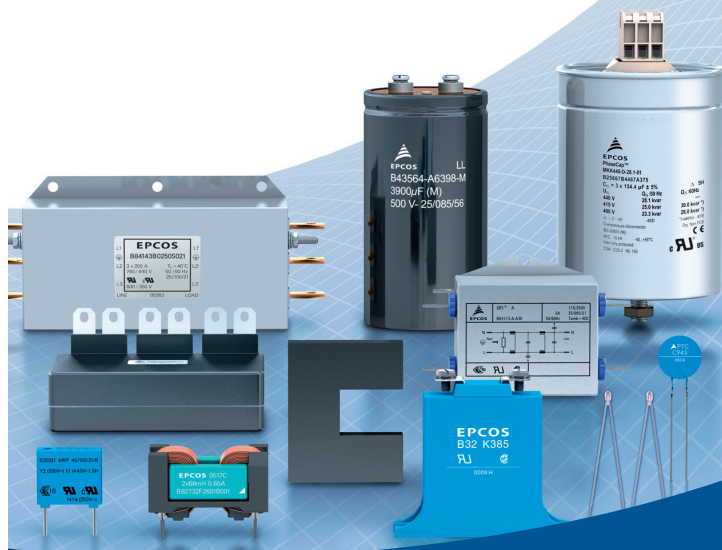
PC-Board Design

Board layout is a separate topic, but it is critical to achieving desirable operation of the hysteretic controller, so a few points should be emphasized:

- Make the feedback trace thin and keep it well away from the inductor and high di/dt traces of the input and switch node.
- Place the feedback resistors very close to the controller.
- Connect the feedback network ground directly to the controller ground pin and run the controller signal ground path completely separate from the power ground return to the input source ground.
- Connect C_1 (C_{IN}) and its ground as close as possible to the anode side of D1 to contain the high di/dt loop. For additional points, the LM3485 evaluation board (AN-1227) serves as an excellent example of proper layout.

The most important objectives are to quantify and narrow the switching frequency variations that can occur due to component characteristics as well as board layout. Compared with previous methods of creating feedback ripple voltage, the emulated approach provides a much-improved method to initially set frequency, and removes the two largest variables (C_{OUT} ESR and inductance) from the equation. Better control of the frequency improves overall system cost and reliability. This occurs because you do not have to size the design for a relatively low frequency; therefore, the inductor does not require as high of a peak current rating, and it allows for the optimization of input and output capacitors.

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