Closing the Loop with a Popular Shunt Regulator

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Using a popular adjustable shunt regulator to drive an optocoupler, this method optimizes the use of an optocoupler to couple an analog error signal across the boundary.

Most offline and telecom power supplies need a method to isolate the relatively high input voltage from lower voltage outputs to provide safety, isolate from lighting-induced voltage surges, and mitigate ground loop issues. Various methods are used to transfer a signal across this isolation boundary, with the most common using pulse transformers or optocouplers. These devices provide a means to couple a secondary side error signal, timing signal, or gate drive signal across this boundary back to the primary side.

Fig.1 shows a typical flyback power supply using a power transformer and optocoupler to provide high-voltage isolation. A UCC3809 controller provides current mode con-

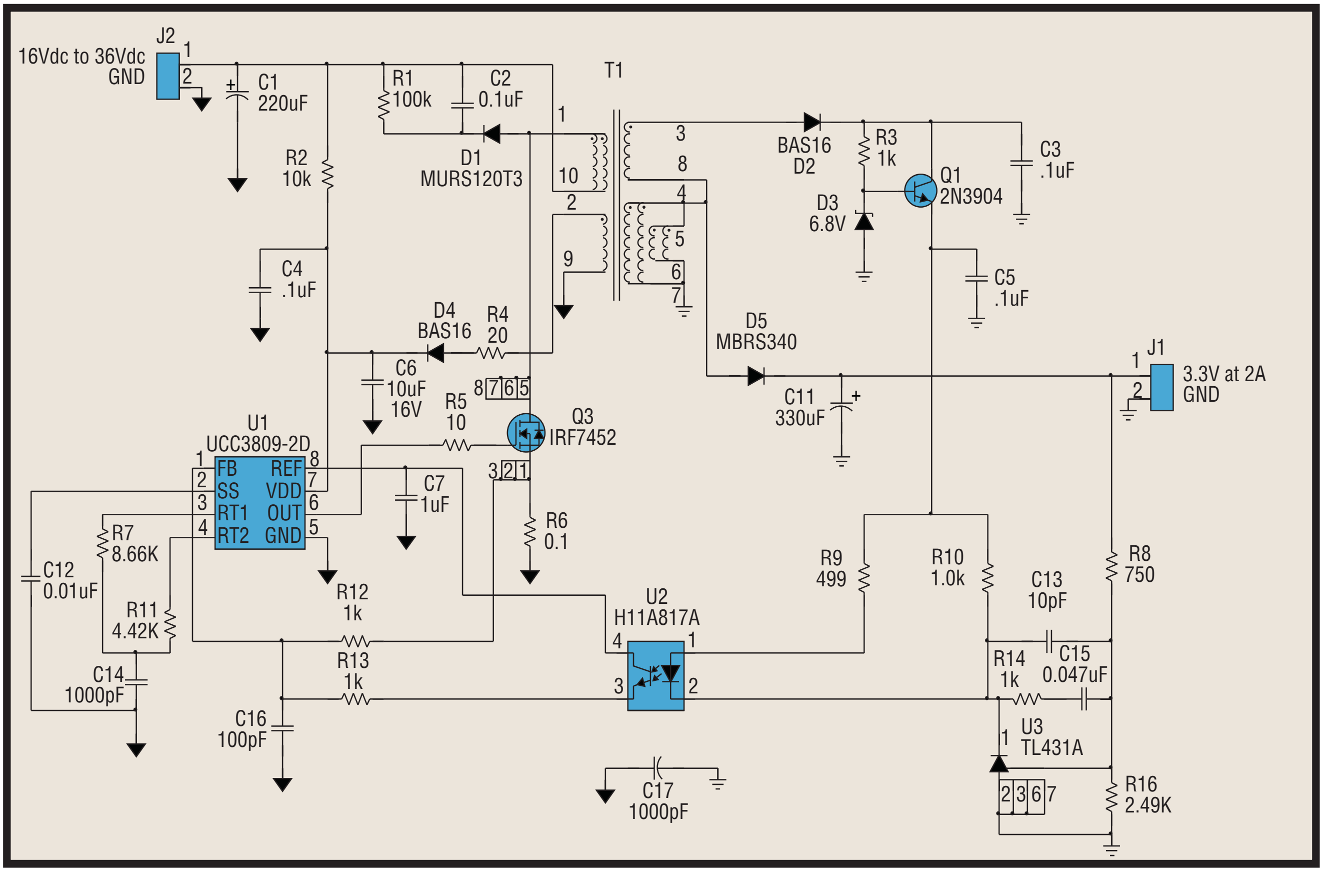

Fig 1. *The TL431 provides a feedback error signal and optocoupler drive.*

trol to the power stage primary. This is accomplished by combining the voltage generated by the primary current sense resistor and the optocoupler output transistor. The combined signal contains current and output voltage error information, and serves to set the power switch primary current and duty cycle. A higher current in the optocoupler output transistor results in an increase in the controller's FB pin voltage, thereby reducing the peak primary current and effectively lowering the output voltage. Consequently, less current in the optocoupler's output transistor, such as that seen at power up, will result in an increased current and output voltage.

To maintain output voltage regulation, the output voltage is compared against the internal voltage reference of the TL431. Differences between the voltage reference and the divided down output voltage are gained up by the internal op-amp, which converts this error voltage into a proportional error current. This current is then passed through the optocoupler diode, transferring the error signal to the primary controller. By creating the error signal on the secondary side rather than trying to transfer a signal to the primary that is proportional to the output voltage, the effects of the

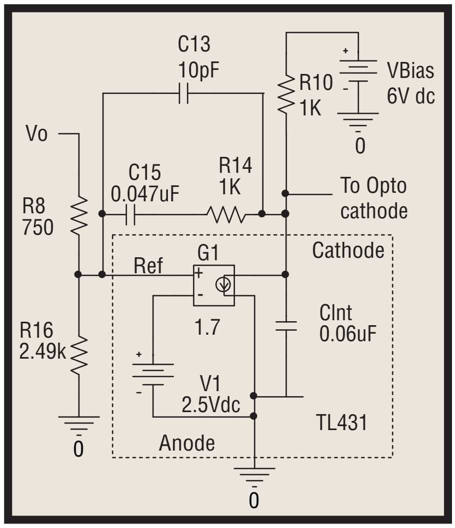

Fig 2. *The TL431 can be modeled as voltage controlled, current source.*

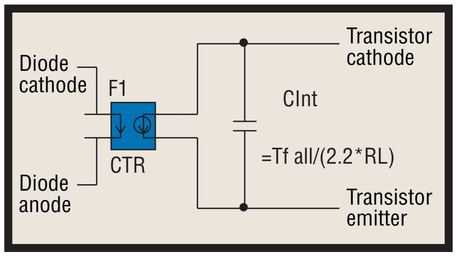

Fig 3. *Optocoupler modeled as current controlled, current source.*

optocoupler's nonlinearity and high gain variation can be minimized. The optocoupler's output current is related to its current transfer ratio and the current generated by the TL431.

The TL431's internal functional blocks consist of a 2.5V reference connected to an op-amp's negative input, with the op-amp output used to drive

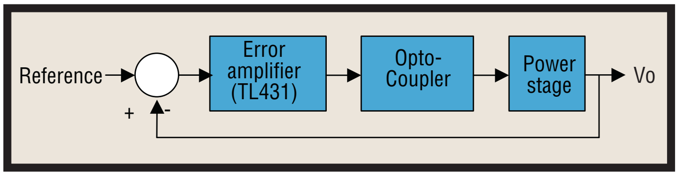

Fig 4. *Simplified control loop block shows high gain in forward path.*

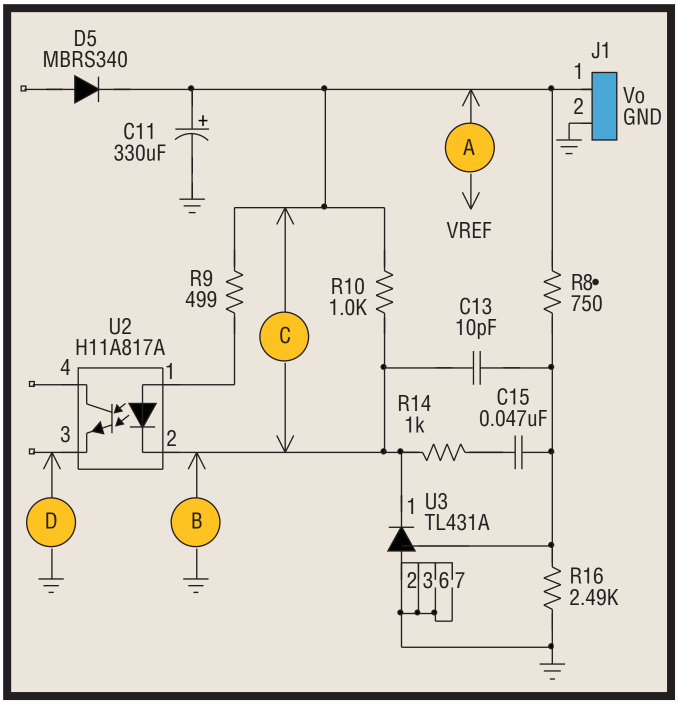

Fig 5. *Alternate optocoupler connection helps stabilize gain variation.*

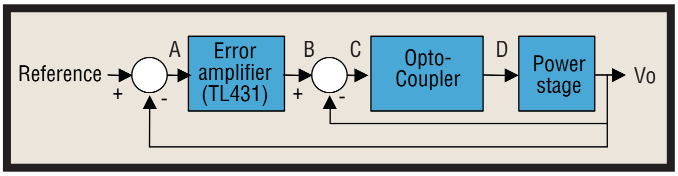

Fig 6. *Alternate optocoupler connection provides second loop.*

an open collector transistor's base. **Fig. 2** shows a simplified SPICE model equivalent of the TL431. In **Fig. 2**, G1 represents a voltage-controlled current-source, the functional equivalent of the TL431's op-amp and output transistor. The voltage-to-current gain is large at low frequencies but decreases at higher frequencies due to the TL431's limited bandwidth. In **Fig. 2's** simplified model, the pole is simulated by the capacitor, CInt. Because the TL431's output can only sink current, a pull-up resistor on the cathode pin guarantees a voltage greater than the 2.5V minimum required for operation, and also provides the minimum required sink current of 1mA. The R8/R16 resistor ratio and the TL431's 2.5V reference set the converter's dc output voltage. While R16 is used to set the converter's dc output voltage, it essentially has no effect on the ac error amplifier gain, which is defined from the TL431 cathode pin to Vout. This is because the REF input of the TL431 (pin 8) is a virtual ground and is held constant over frequency, allowing no ac current flow into R16. For this reason, if a dc output voltage change is desired, the designer should vary R16 rather than R8. R16 can therefore be ignored in the ac model, with only R8 and feedback values R14, C15, and C13 determining the error amp's frequency response.

Fig. 3 shows a simple SPICE model of the optocoupler. The optocoupler can be modeled as a current-controlled current-source, with the CTR defining the gain. Most optocouplers used in power supplies don't have very high bandwidths. To model this, capacitor CInt has been added to **Fig. 3**. It places a mid-frequency pole in the optocoupler's response. Its estimated value can be calculated from **Fig. 3** by finding the optocoupler's fall-time spec and the fall-time test load resistor. The optocoupler voltage gain is set by its CTR, and the equivalent internal capacitance and external components connected to it. For example, in **Fig. 1**, with a 1V swing on the resistor driving the optocoupler anode voltage (holding the TL431 cathode output constant and an optocoupler CTR=1), 2V is seen at U1- pin 1. The ratio of R12 to R9 and the device CTR sets the optocoupler gain, which in this case is approximately 6 dB. Resistor R13 doesn't factor into the gain because the optocoupler is acting as a current source, and only the voltage seen by the controller with respect to ground has an effect on the gain.

As shown in **Fig.4's** block diagram of the gain stages comprising the control loop for the circuit in **Fig. 1**, the error amplifier (TL431) block is in the forward path rather than in the feedback path. In addition, note that there's only a single feedback path in this circuit. The optocoupler itself has no dependency on the changes in the output voltage. The bias circuit, consisting of R3, D3, and Q1, fix a dc voltage in the optocoupler's anode path. This removes any optocoupler dependency on the output voltage, load, and input line to changes in the ac gain characteristics. This independence on the output voltage can simplify the chore of stabilizing the control loop, but with the added expense of the bias circuit components.

Fig. 5 shows an alternate method for connecting the optocoupler. The optocoupler's current is now determined by the difference between the output voltage and the error voltage.

In making this connection, a second inner loop is introduced in the block diagram in **Fig. 6**. The inner loop serves to stabilize the optocoupler's gain variations and speeds up the control loop response by providing an alternate control path so the error signal doesn't have to propagate through the error amplifier.

The simulated ac gain of the inner loop's various blocks is revealed in **Fig. 7**. The power stage transfer function is Vo/D, based on the model described by Vorperian (see references), and shows a low-frequency break established by the output capacitors and load resistance. It also has two zeros, one established by the ESR of the output capacitor and a right half plane zero established by the continuous

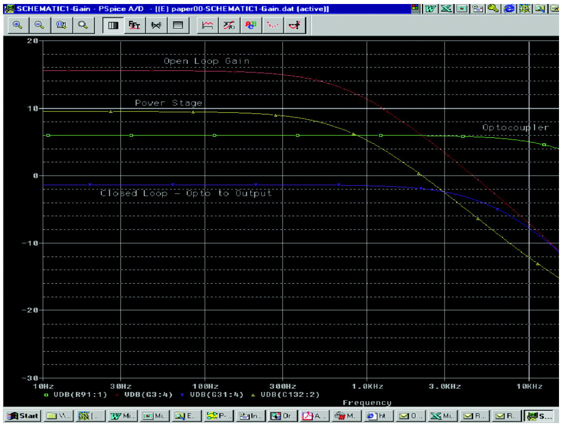

Fig 7. *Compensation of the inner loop (TL431 out of circuit).*

Fig 8. *Compensation of the outer loop (TL431 in circuit).*

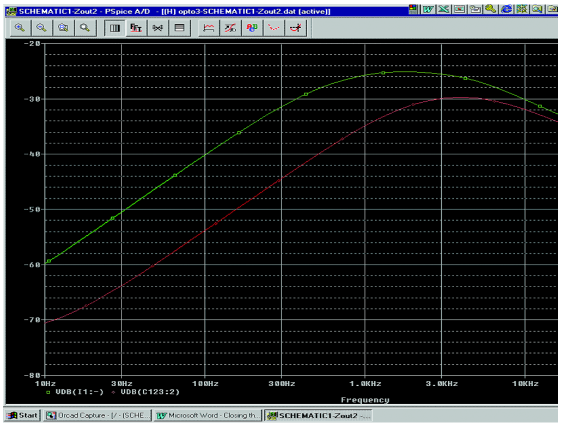

Fig 9. *The two-loop control method lowers the output impedance by 12 dB.*

mode of operation. The optocoupler gain block, defined as D/C, is based on the simplified model of **Fig. 3**. It shows a pole at approximately 20 kHz and is characteristic of this particular optocoupler. In this example, no further compensation of the loop was provided, and these two blocks were just summed together, with Vo/C defining the open loop power/optocoupler stage gain. The effect of closing the inner loop can be seen in the response defined by Vo/B. Since the feedback is unity, the inner loop gain is reduced to near 0 dB at lower frequencies, and at higher frequencies it tracks the open loop gain. Interestingly, closing the loop moves the low frequency pole from 300 Hz to above 3 kHz.

Fig.8 illustrates the responses for closing the outer loop. The inner-loop closed-loop response is repeated, and the error amplifier's planned response on the error amplifier is presented. A type-two amplifier is configured with a low-frequency integrator zero placed at the same frequency as the inner loop pole. In addition, a pole is placed at a frequency much higher than the planned crossover frequency of 5 kHz. This high frequency pole is added to the TL431 compensation to prevent switching noise from being amplified by the TL431 and perturbing the control characteristics. The zero location in the TL431's compensation can be seen in this figure at approximately 3.5 kHz, and was selected to align with the pole of the closed loop power/optocoupler stage gain. When the outer loop is closed around the TL431, the total closed loop gain is the sum of the power/optocoupler closed loop gain (Vo/B) and the error amplifier gain (B/A). The outer closed loop gain is defined as Vo/A and has a bandwidth of approximately 4 kHz.

Fig. 9 presents a key reason to add the complexity of two loops into a TL431 design. It presents the output impedance of two different designs having the same crossover frequency through the optocoupler loop. The output impedance was determined by adding a 1A current source in parallel with the load resistor and measuring the output voltage variation vs. frequency. The green curve represents the output impedance of a power supply that doesn't have the inner loop. The red curve represents the output impedance of a two loop supply. With two loops, there's a fourfold reduction in the output impedance of the power supply. This means that for a given load transient, power supply output variation can be reduced four times without having to push the overall loop bandwidth to higher frequencies. Other advantages to this approach include stabilizing the optocoupler's gain variation by using the first loop and reducing parts count—resulting in improved reliability, smaller size, and reduced cost.

References

1. *Dinwoodie, Lisa, "Design Review: Isolated 50W Flyback Converter Using the UCC3809 Primary Side Controller, Unitrode Application Note U-165," Texas Instruments Power Supply Control Products Data Book.*

2. *Zhang, Jovanovic, Lee; "Design Considerations and Performance Evaluations of Synchronous Rectification in Flyback Converters," IEEE Transactions on Power Electronics, Vol. 13, No. 3, May 1998.*

3. *Ridley, Ray, "Designers Series, Part V, Current-Mode Control Modeling," Switching Power Magazine, 2001.*

4. *Vorperian, Vatche, "Simplified Analysis of PWM Converters Using Model of PWM Switch Part 1: Continuous Conduction Mode," IEEE Transactions on Aerospace and Electronic Systems, Vol. 26, No. 3, May 1990.*

5. *Ridley, R., "A New Small-Signal Model For Current Mode Control," Ph.D. Dissertation, Virginia Polytechnic Institute and State University, Blacksburg, Va., November 1990.*

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