



VENABLE TECHNICAL PAPER # 17

Testing and Stabilizing Feedback Loops in Today's Power Supplies

Abstract:

Feedback loops aren't what they used to be. Here are some practical tips for dealing with the problems presented by modern topologies. Gone are the days of a single feedback path, an external or discreet error amplifier, and a direct connection from output to input. Gone also are the days of voltage mode control, even though it is a better choice in some applications. The majority of power supplies manufactured today throughout the world use a similar topology. Multiple outputs with coupled inductors each feed the common summing node of a TL431 "programmable zener" error amplifier. The TL431 drives the diode portion of an opto-coupler, with the transistor portion on the primary (line) side of the isolation barrier controlling the operating point of a current-mode control chip. The internal error amplifier on the current-mode control chip may or may not be used as a buffer and/or gain stage. The questions raised by this topology are "Where is the loop?" and "Where do I connect compensation components if I want to change the performance of the loop?" All too often, the answer to both questions is "I don't know." Step load testing and trial-and-error loop compensation substitute for Bode plots and rational selection of components. The problem with step load testing is that it does not prove that a system is unconditionally stable, and may not prove the system is stable at all if the rise time of the step load is large compared to the feedback loop bandwidth. The problem with trial-and-error compensation, especially if the designer is not sure of the loop path and the components that affect it, is that the final design may not even be close to optimum and many unnecessary or ineffective components may be used which increases the cost of production. Figure 1 shows a simplified version of this type of power supply. The simplification is that only one output is shown instead of multiple outputs with coupled inductors. We will deal with the complications of multiple outputs later. For now, let's go over the operation of this circuit and the signal flow paths.

How it Works

The power supply shown in Figure 1 is a two-transistor forward converter. Transistors Q1 and Q2 turn on and off at the same time. When they are on, energy is delivered through CR3 to inductor L1. When they sue off, leakage inductance energy in transformer T1 is returned to the source through diodes CR1 and CR2, and L1 delivers some of its stored energy to the load by drawing

Inductor L1 and capacitor C1 form a low-pass filter, which removes most of the AC components and passes the DC value of the waveform. The triangular ripple current through L1 passes through the equivalent series resistance (ESR) of C1 and generally creates more ripple voltage than allowed by the specification. A second filter consisting of L2 and C2 reduces the ripple voltage to an acceptable level, but creates an additional phase lag that cannot be compensated for by the feedback loop if the loop were closed at the output of the supply. This problem is sometimes avoided by connecting the optocoupler diode through R5 to the junction between the two filters, before the additional phase lag is incurred. There are two signal flow paths at this point. One is a high-frequency path through R5, and the other a low-frequency path through R7. The high-frequency path is created because C3 and R6 act to make the cathode of programmable zener U2 a virtual AC ground. High-frequency AC voltage on the top of R5 pushes current through R5 and the optocoupler diode without any signal passing through the gate of U2 from R7. This path through R5 is actually more important to the stability of the feedback loop than the low-frequency path through R7.

An internal feedback loop has been created which controls the output current, making it proportional to V_c . There has been a great deal of controversy about the significance of this loop. Our position at Venable Industries is that it is a minor loop; no more significant than the feedback around an op-amp, and the principal reason for examining it is that it can be unstable if improperly compensated. It is a sampled-data feedback loop because the only important instant of time is the moment when the voltage across R_1 is equal to V_c . At all other times, there is no

significance to the relationship between these two voltages. For this reason conventional analog signal measurement techniques, which assume the output is continuously proportional to the input, do not work. The way around this particular limitation is to measure the closed-loop transfer function and to use this data to calculate the open-loop transfer function from the closed-loop data. The open-loop transfer function is the gain around the loop. The closed-loop transfer function is the gain from input to output. In this case, the closed-loop transfer function is a transconductance, the ratio of output current (the current in L1) to control voltage (V_c) as a function of frequency, and can easily be measured using conventional analog techniques.

Measuring Voltage Loop Gain

Neither the low-frequency path through R7 nor the high-frequency path through R5 represents the true feedback loop. In more complex power supplies with additional feedback to the summing node (gate of U2) from other output voltages, this is even more the case. Looking at Figure 1, the only point in the loop where all the paths combine into one is the connection between the op-amp and the comparator. Although this point is available as the compensation pin of U1, it is not possible to break the loop at this point since the connection is made inside an integrated circuit.

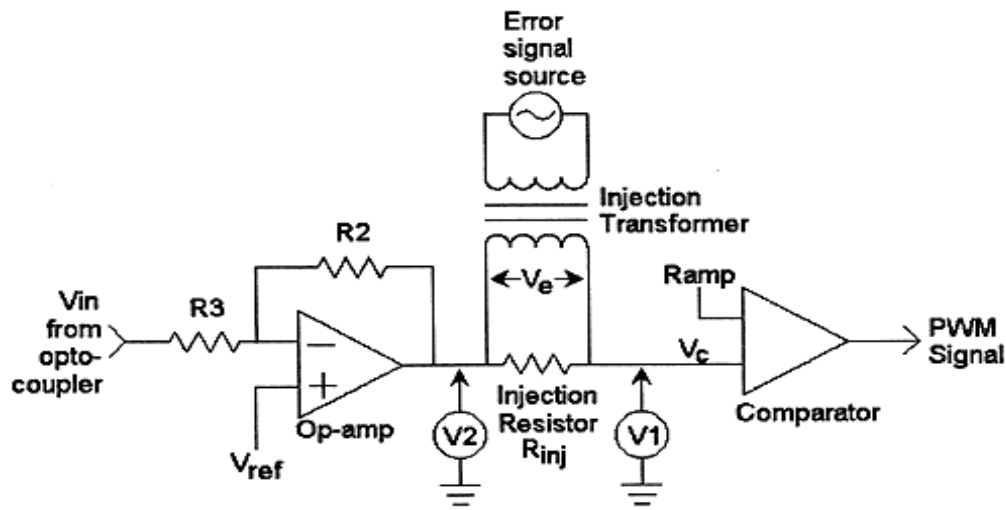


Figure 2. Ideal (desired) injection method

Figure 2 shows the desired method of injecting an error signal in series with the loop so the open loop gain can be measured. The output of the op-amp is the input times the ratio of the feedback resistor to the input resistor, i.e.

$$V2 = V_{in} \times (R2 / R3)$$

An oscillator is connected through a transformer across a small (typically 100-ohm) resistor in series with the feedback loop at a point where the signal is confined to a single path and where the source is low impedance (the output of the op-amp in this case) and the load is a high impedance (the input to a comparator in this case). The voltage across the injection resistor, V_e , is added to the output voltage of the op-amp to form the input voltage to the comparator.

$$V1 = V2 + V_e$$

$V1$ and $V2$ are measured with respect to signal ground. They are vectors (since they rotate, they technically are phasors), with magnitudes and phase angles. Together with the error voltage, V_e , these three voltages form a vector triangle. The ratio of $V2 /$

V1 is the open loop gain. At crossover, the point where the loop gain is one (0 dB), V1 and V2 are equal. They form an isosceles triangle with Ve as the base and V1 and V2 as the sides of the triangle. The phase margin, ϕ_m , is the angle at the apex of the triangle. Figure 3 shows this relationship graphically. As you can see from Figure 3, if the loop has a very small phase margin it is possible to generate very large voltages (V1 and V2) from a small error voltage.

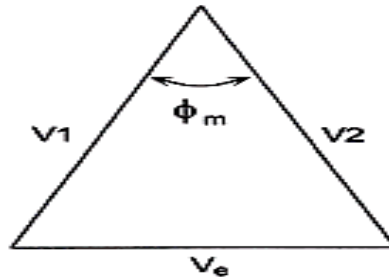


Figure 3. Vector diagram of signal amplitudes at unity gain frequency

The only problem with Figure 2 is that it cannot be implemented because the connection that needs to be broken is a trace inside an integrated circuit. Figure 4 shows an alternate injection method that is mathematically equivalent and is achievable.

Since the currents into the summing node of the op-amp (inverting input) still have to add to zero, the voltage at the point labeled V2 has to be equal to $V_{in} \times (R2 / R3)$. If the injection resistor is connected in series with R2 then the error voltage Ve is connected in series with V2 and V1 is equal to the sum of the two voltages as before. The same conditions apply in Figure 3.

$$V2 = V_{in} \times (R2 / R3) \quad \text{and} \quad V1 = V2 + V_e$$

Injecting in series with the feedback around the internal error amplifier in the current mode control chip is therefore mathematically identical to injecting between the error amplifier and the comparator. This is fortunate because this point is always available. Both the output of the op-amp (compensation pin) and the inverting input are available, even on the most simplified chips. One caveat is that this technique depends on the op-amp having gain. Data is less and less valid as the bandwidth of the op-amp is approached.

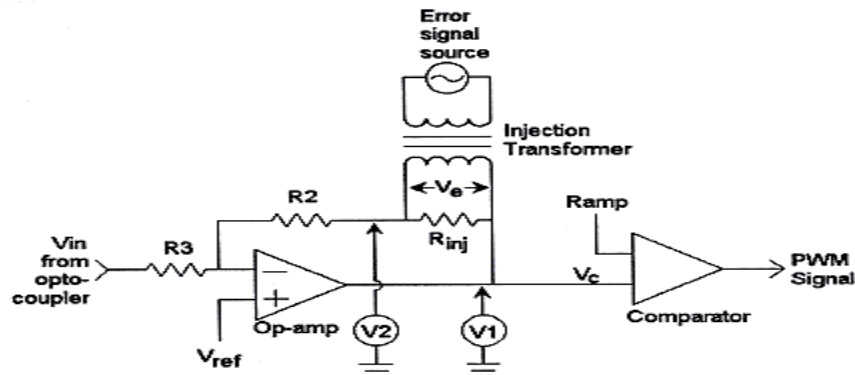


Figure 4. Achievable method of signal injection to measure loop transfer function

This technique doesn't work when the error amplifier is not used at all. Sometimes a manufacturer will couple directly from the opto-coupler into the compensation pin of the control chip, bypassing the error amplifier completely. The advantage is saving one or two resistors, cutting the manufacturing cost by almost a penny. The disadvantages are that design flexibility is reduced and the design is difficult to test. There is some advantage to having gain in the amplifier of the control chip so that the entire loop gain does not have to come from the TL431 acting like an error amplifier. Also, the dynamic range of the opto-coupler is reduced so that it does not have to vary over such a wide range of operating points, especially the low current operating point when the TL431 is near cut-off.

There is a way to test a circuit like the one in Figure 1 even if the error amplifier in the control chip is not used. You can test the fast loop and slow loop independently, by injecting a signal at the top of R5 and R7 respectively. If you call the fast loop gain GAIN1 and the slow loop gain GAIN2, these two transfer functions can be combined using the formula:

$$\text{True loop gain} = \frac{\text{GAIN1} + \text{GAIN2} - (2 \times \text{GAIN1} \times \text{GAIN2})}{1 - (\text{GAIN1} \times \text{GAIN2})}$$

This formula gives the same result as injecting in series with R2 or injecting between the op-amp and the comparator if you could do that. This formula is a built-in part of the Venable Model 350 Frequency Response Analysis System, called Double-to-Single Loop Conversion. It is possible to extend this technique to three loops, but we have not done that yet. Loops that have a third signal path, such as would be the case if a signal from another output voltage were fed into the TL431 gate, can still not be measured if the error amplifier in the control chip is bypassed. If it is not, then R2 can be used as an injection point no matter how many outputs are connected to the TL431 summing node.

To demonstrate the various loops, we modeled the circuit of Figure 1 and made measurements of the model injecting at various places. We didn't put in any rolloff of gain in any of the op-amps or the opto-coupler. We modeled the current mode portion as simply a voltage-controlled current source, again with no rolloff at high frequency. For these reasons, the loop phase shift does not fall off at high frequency as would be the case with hardware, but the model is accurate at low frequency and well past loop gain crossover frequency. Figure 5 is the low frequency loop through R7. It has a bandwidth of about 100 Hz with plenty of phase margin. Figure 6 is the high frequency loop. It has low gain at low frequency. Below 100 Hz, the low frequency loop dominates. Above 100 Hz, the high frequency loop dominates. The true loop is obtained by combining these two loops in accordance with the above formula or by measuring in series with R2. It is shown in Figure 7. The true loop bandwidth is about 4 kHz, which means the power supply will respond to a transient line or load disturbance in about 1/4 of a millisecond. Faster response requires a higher loop bandwidth.

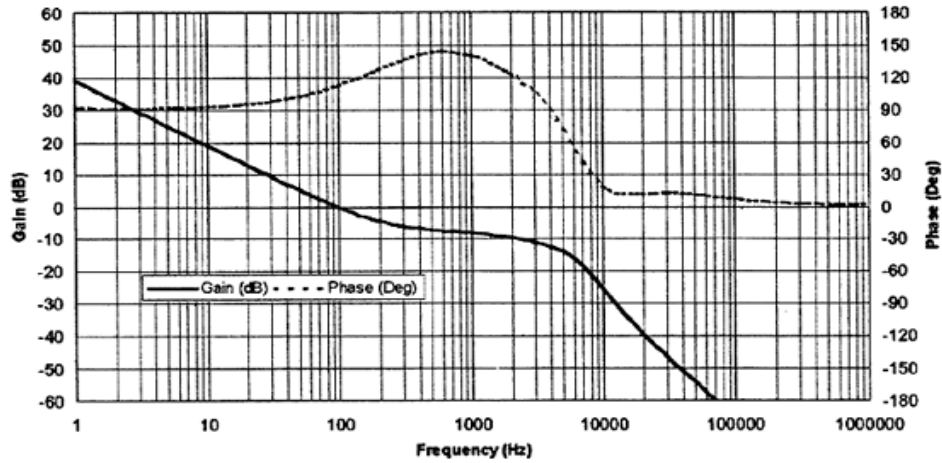


Figure 5. Transfer function of "slow loop" measured in series with R7

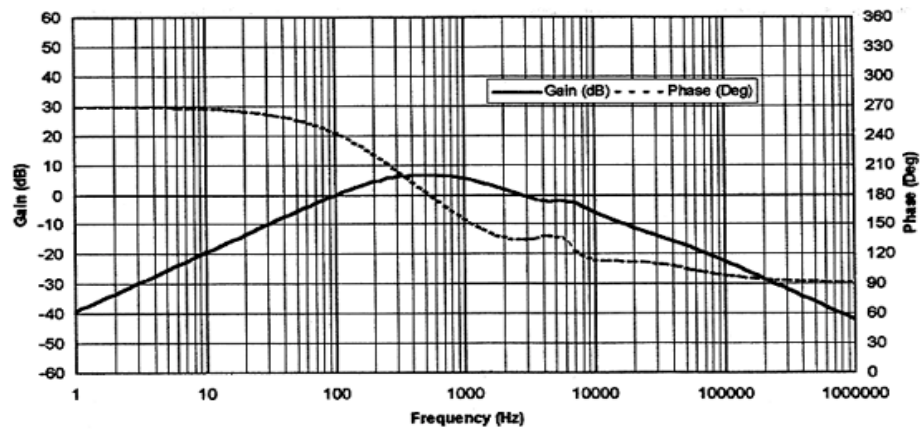


Figure 6. Transfer function of "fast loop" measured in series with R5

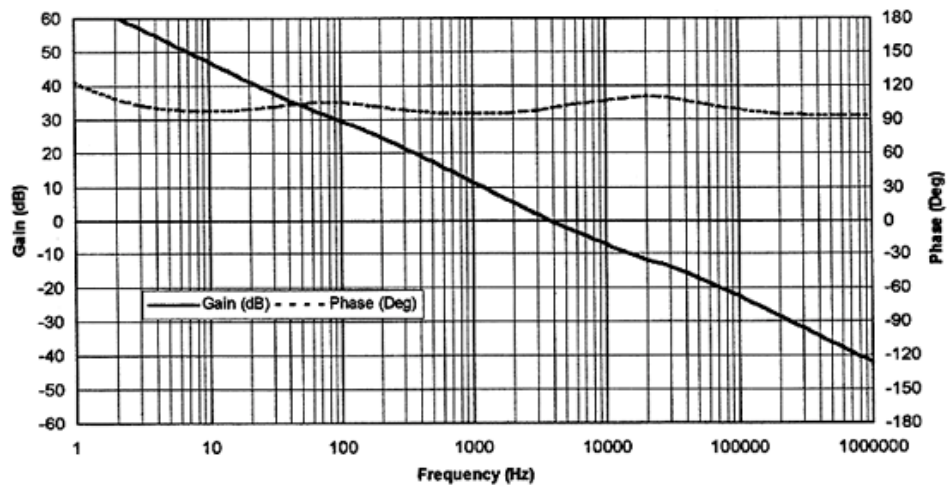


Figure 7. Transfer function of open loop gain of entire loop measured in series with R2

Measuring Current Loop Gain

The discussion so far has primarily concerned measurement of the voltage loop gain, since this is the loop that controls external performance such as transient response. The current loop is an internal loop that makes the output current track the control voltage. In current mode control, the output voltage is regulated by sensing whether the output voltage is high or low, and then decreasing or increasing the output current as required to maintain the output voltage at the desired level. As pioneers in the field of current mode power supply design quickly discovered, the current loop is inherently unstable for duty ratios greater than 0.5. Duty ratio is the ratio of on time to cycle time for the power switching transistors. In forward converters such as the example shown in Figure 1, the duty ratio cannot exceed 0.5 since at least that much time is needed to reset the flux in the core of power transformer T1. At low line voltage the duty ratio does approach 0.5 and the current loop can be on the verge of oscillation at that operating point.

In addition to the question of stability of the current loop, it is also instructive to know the bandwidth of this loop. It is common knowledge that the control-to-output transfer function of current mode control falls at a -1 slope (-20 dB per decade). The reason for this -1 slope may not be so commonly known. It is because the current loop creates a transconductance block that takes a control voltage and puts out a current, and the ratio of current to voltage is constant with frequency (up to the bandwidth of the current loop). This transconductance block drives an impedance consisting of the power supply output filter capacitor in parallel with the load. A current driving an impedance creates a voltage, so the control voltage to output voltage transfer function is the product of the transconductance of the current loop and the impedance of the output filter and load. At frequencies around loop crossover, this impedance is primarily the filter capacitor. Since the transconductance is constant with frequency and the impedance of a capacitor falls at a -1 slope, the product of the two falls at a -1 slope also. This is why the control voltage to output voltage transfer function falls at a -1 slope. Above the bandwidth of the current loop, the transfer function of the voltage to current converter is no longer flat with frequency. Above this bandwidth, the control voltage to output voltage falls at a steeper slope than -1 . The power supply designer must make sure the voltage loop crossover frequency stays well below the bandwidth of the current loop. This is easier to do when the power supply designer knows the bandwidth of the current loop, but this is not usually the case. The current loop is measured by connecting one channel of a frequency response analyzer to the compensation pin of the current mode control IC (V_c) and another channel of the analyzer to the output of a current probe. The current probe is clipped around one lead of the energy storage inductor L1. An appropriate scale factor must be applied if the current probe gain is anything other than one volt per amp. The loop is then disturbed by injecting an error voltage in series with the loop exactly as is the case when measuring the voltage loop. The frequency response analyzer measures the control voltage and output current only at the frequency of signal injection, rejecting all other frequencies and noise. The ratio of current to voltage is then plotted as a function of frequency. This is the plot of transconductance versus frequency, and is the closed-loop gain of the current loop. It is possible to calculate the open-loop gain of the current loop from the closed-loop gain, but in most cases this is not necessary. The two pieces of data of interest are the bandwidth and the phase margin of the current loop. You can tell the bandwidth from the frequency at which the open-loop gain starts to fall. The phase margin can be accurately estimated from the amount of peaking of the closed-loop gain just before the gain starts to fall. Table 1 shows the relationship between peaking in the closed-loop transfer function and phase margin of the open-loop transfer function. The current loop phase margin can be adjusted by varying the amount of slope compensation used in the current mode control circuit.

Table 1. Relationship between open-loop phase margin and closed-loop gain peaking.

ϕ_m (degrees)	Peaking (dB)
20	9.24
30	5.90
40	3.71
50	2.23
60	1.27

Improving the Voltage Loop

If you want to increase the loop crossover frequency as shown in Figure 7, you have to deal with the high frequency path, not the low frequency path. There isn't much you can do with R7, C3, or R6. The cathode of U2 is already a virtual AC ground. The two ways to increase the gain are to decrease the value of R5 or to increase the ratio of R2 / R3. You probably chose R5 to get a particular current through the opto-coupler; so varying that component may not be a good choice. The value of R2 is a good choice, provided that you actually used the error amplifier inside the control chip. If your loop phase shift isn't as benign as the one shown in Figure 7, if increasing the gain would push gain crossover near phase crossover, then you need to add a zero-pole pair to boost the phase at the same time you boost the gain. The best way to do this, but one that isn't thought of very often, is to put a series R-C combination in parallel with R5. This won't alter the DC operating point of the opto-coupler, but will increase the gain and reduce the phase lag over a small range of frequencies. There will be a zero approximately at the frequency where the new capacitor is equal in impedance to R5, and a pole at approximately the frequency where the impedance of the new capacitor is equal to the value of the new resistor in series with it. The design goal is to place the zero and pole symmetrically around the desired loop gain crossover frequency.

Dealing with Multiple Outputs and Coupled Inductors

This one is tough and we don't really have a good answer. The problem is that the coupled inductor is really a transformer also, and although an inductor won't allow its total ampere-turns to be changed instantly, it doesn't care which winding or windings the current is flowing in. If you have 5 volt and 12 volt outputs with coupled inductors, and assuming the output turns ratios are the same as the voltage ratios, then 1.2 amps could be flowing in the 5 volt output, or 0.5 amps in the 12 volt output, or any combination of the two that add up to the same ampere-turns, and the inductor wouldn't know the difference.

The concept of current-mode control is that the control voltage sets an output current, and that current flows through the impedance of an output to produce a voltage, and that voltage is used as a basis of regulation. When there is more than one output, and they are coupled together, there is

no guarantee that the current will flow to the output you are trying to regulate, especially at high frequency. The current will split according to the ratio of the leakage inductances and also the impedance of the individual filters. This is especially a problem when the main load is on an output other than the one you are regulating.

The best solution is to try to make the filter impedances at least comparable. Capacitance reflects back to the primary proportional to the square of the turns ratio, so the 12-volt output should have $(5/12)^2$ or 17.4% of the capacitance of the 5-volt output. Even then, it is unlikely that the leakage impedance ratios will be balanced in the same ratio, so be prepared for some "fiddling" with the loop if you are not loading the same output you are regulating.

Summary

Today's power supplies have multiple feedback loop paths. In the secondary there is frequently a "slow loop" and a "fast loop" created by the topology of the circuit. Neither by itself is the true loop, but the true loop can often be measured on the primary side by injecting in series with the feedback around the operational amplifier of the current mode control chip. If that measurement option isn't available, another possibility is to measure the multiple loops separately and then combine those measurements to find the transfer function of the true loop. This is easy if there are only two loops and you have access to a Venable Frequency Response Analyzer.

With a frequency response analyzer and current probe, you can plot the ratio of output current to control voltage of the internal current loop. The plot of this transconductance versus frequency will tell you immediately what the bandwidth and phase margin of the current loop is. The bandwidth is the frequency where the transconductance starts to roll off as the frequency increases. From the amount of peaking in transconductance just before rolloff and Table 1, you can accurately estimate the phase margin of the current loop.

Knowing the path of the feedback loop at various frequencies, you have a good handle on what to change to improve the loop. Noting that the loop gain is a strong function of the value of resistor R5, you discovered that this is a good place to put compensation components.

Finally, we pointed out the difficulties presented by having multiple outputs with coupled inductors and suggested some solutions to minimize the problem.