

RELATING CONVERTER TRANSIENT RESPONSE
CHARACTERISTICS TO FEEDBACK CONTROL
LOOP DESIGN

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*Originally published Powercon11, April 9-13, 1984***ABSTRACT**

Confusion often surrounds the fundamental relationships of time domain and frequency domain as they pertain to the design of the feedback control portions of power supplies. The purpose of this paper is to demonstrate the relationships between loop gain crossover frequency, phase margin, and transient response so that the design engineer can achieve the best possible dynamic performance from a given power supply.

1. INTRODUCTION

The common practice of using a step load combined with trial and tweek techniques to stabilize switching power supplies has serious drawbacks as compared with using measured transfer functions (Bode plots) to allow proper optimization of the feedback loop. The design engineer needs to be certain that his or her designs will be stable under all conditions of line, load, and environmental conditions. Using transient analysis (Figure 1) the engineer sees only the end result, not the interaction of the various pieces of the circuit under test.

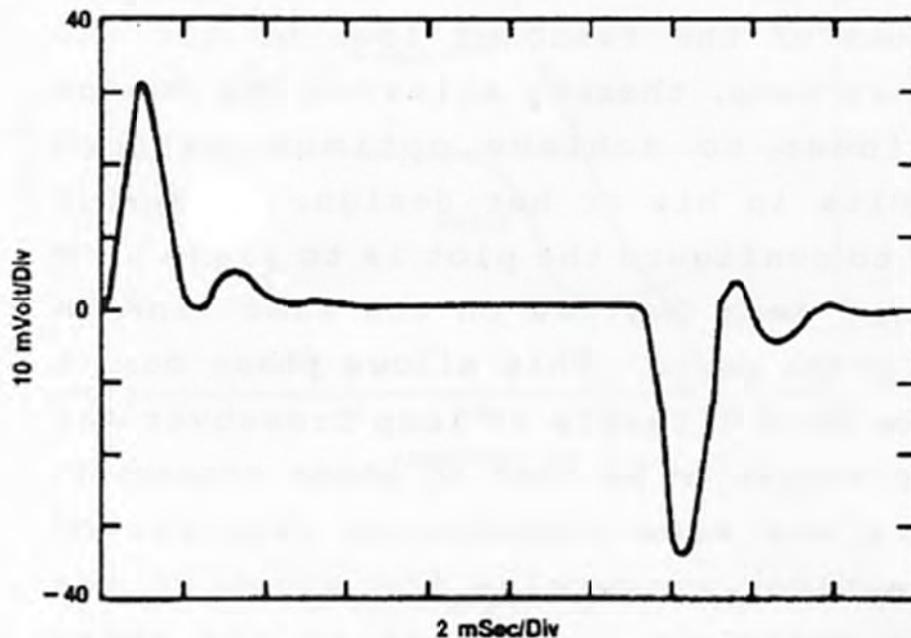


FIGURE 1. TYPICAL TRANSIENT RESPONSE

The Bode plot, on the other hand, gives information on this interaction by giving the relationship of gain and phase shift with frequency of either individual pieces of the loop or the overall loop. By properly optimizing the feedback loop of the power supply, the designer can cut costs in manufacturing and reduce the number of field failures. The fact that his or her power supplies simply perform better than those of the competition comes as a side benefit. In this paper we will explore the relationships between the Bode plot (frequency domain response) and the transient analysis (time domain response) in a clear concise way. But first, let's start by reviewing the basics.

2. WHAT IS A BODE PLOT

A Bode plot is a powerful tool for the design engineer. It consists of a plot of the transfer function (output divided by input) of a circuit, and is usually expressed as log gain (in dB), and linear phase shift (in degrees), on the Y axis versus log frequency on the X axis. An example is shown in Figure 2.

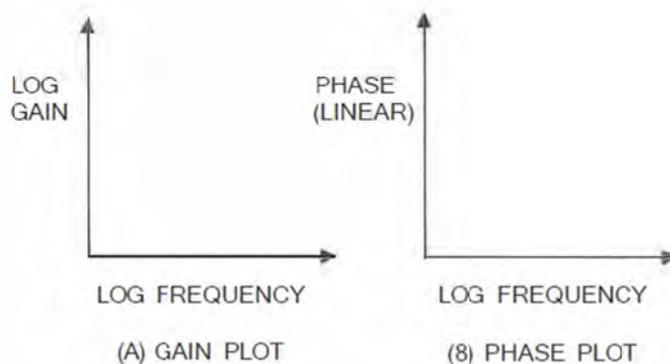


FIGURE 2. GAIN AND PHASE BODE PLOTS

Bode plots can be created either for pieces of the feedback loop or for the entire loop, thereby allowing the design engineer to achieve optimum desired results in his or her design. A useful way to configure the plot is to place zero dB and zero degrees on the same line on the graph paper. This allows phase margin to be read directly at loop crossover and gain margin to be read at phase crossover. There are some interesting features of Bode plots concerning the slope of the gain curve as it relates to the phase shift. If the gain is flat with frequency, it is said to have a 0 slope. If the gain falls at 20 dB per decade (6 dB per octave) it is said to have a -1 slope. Gain falling at 40 dB per decade (12 dB per octave) has a -2 slope and so on. There is a definite phase shift associated with each slope. A 0 slope typically has 0 phase shift. A -1 slope has 90 degrees more phase shift. A -2 slope has 180 degrees more phase shift than a 0 slope. Each minus slope change adds an additional 90 degrees of phase shift. Each plus slope change removes 90 degrees of phase shift. The actual amount of phase shift at any particular frequency is dependent on how near the corner frequency it is, but away from the

corner the phase will settle in on the right shift for its gain slope. An example of an exception to this rule is an inverting amplifier which has 180 degrees from the inversion and -90 degrees from its gain slope for a total of -270 degrees of phase shift. Consistency in format also helps. All of our Bode plots have a factor of ten in gain (20 dB) be the same dimension as a factor of ten in frequency (one decade), so that a 1 slope really is a 45 degree angle on the plot.

2.1 MODULATOR TRANSFER FUNCTION

At Venable Industries, we define the modulator as the power processing section of the circuit. The input of the modulator typically is a point where a DC voltage creates a certain amount of duty ratio or output. For power supplies, this typically is the compensation pin on the PWM chip, or the input to a comparator. The output of the modulator is the output of the supply or a voltage proportional to the variable being controlled. In the modulator exist the various parasitic components and other assorted demons that try to hide until your supply is in production. A typical Bode plot of a modulator is shown in Figure 3. A demonstration power supply was designed and built especially for this paper. We used an SG1525A as a controller and an external op-amp for the error amplifier. The schematic for this power supply is given in Figure 4.

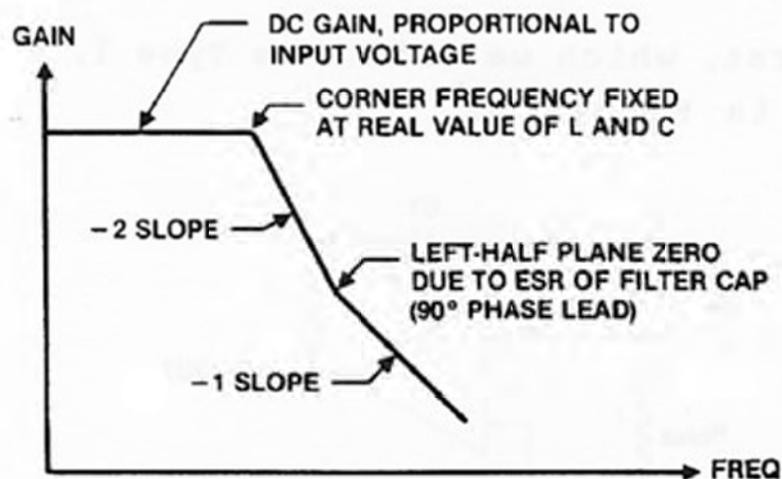
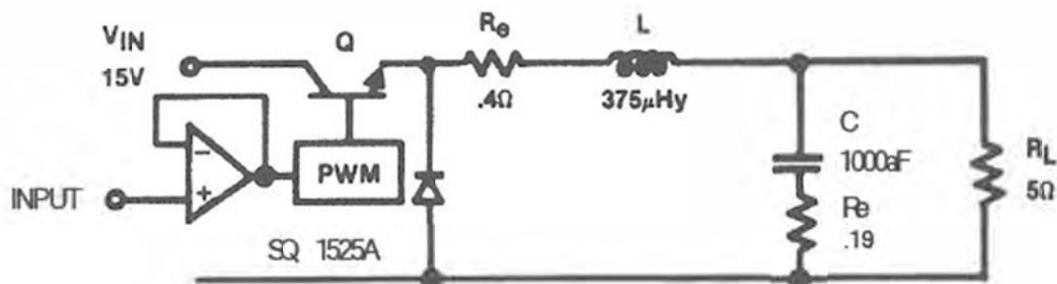


FIGURE 3. TYPICAL BODE PLOT OF A BUCK REGULATOR



$$\text{DC GAIN} = 6 = 15.5\text{dB}$$

$$\text{DOUBLE POLE AT } L/C : f = 260 \text{ Hz}$$

$$\text{ESR ZERO AT } R_{\theta}C : f = 838 \text{ Hz}$$

FIGURE 4. BUCK REGULATOR SCHEMATIC

The reason for using an external op-amp in this power supply is to allow the design engineer the most flexibility and repeatability in the design. The internal op-amp in this IC, and most other similar PWM chips on the market, is not really an op-amp at all but a transconductance amplifier. These amplifiers have a high output impedance and poorly specified internal components that the design engineer must rely upon to stabilize his or her design. These internal components can vary several hundred percent from lot to lot and vendor to vendor. By using the internal amplifier as a non-inverting buffer and using a real op-amp as the error amplifier, the design engineer has total flexibility of design and can choose whatever tolerance of components required by the design criterion. Many times we have found engineers spending thousands of engineering dollars trying to save the cost of a thirty cent IC. Now let's get back to the SG1525A. The voltage delta at the control input to change from zero duty ratio to full duty ratio is 2.5 volts. At zero duty ratio the output voltage is zero, and at full duty ratio the output voltage is V_{in} . The input voltage in this particular case is 15 vdc. Since the gain of the modulator is V_{out}/V_{in} the gain is $15/2.5$, or 6. A gain of 6 is 15.56 dB. This is the gain at DC and frequencies below the L-C corner. As you can see in Figure 5, when we pass the L-C corner frequency (260 Hz in this case) the gain begins to fall off.

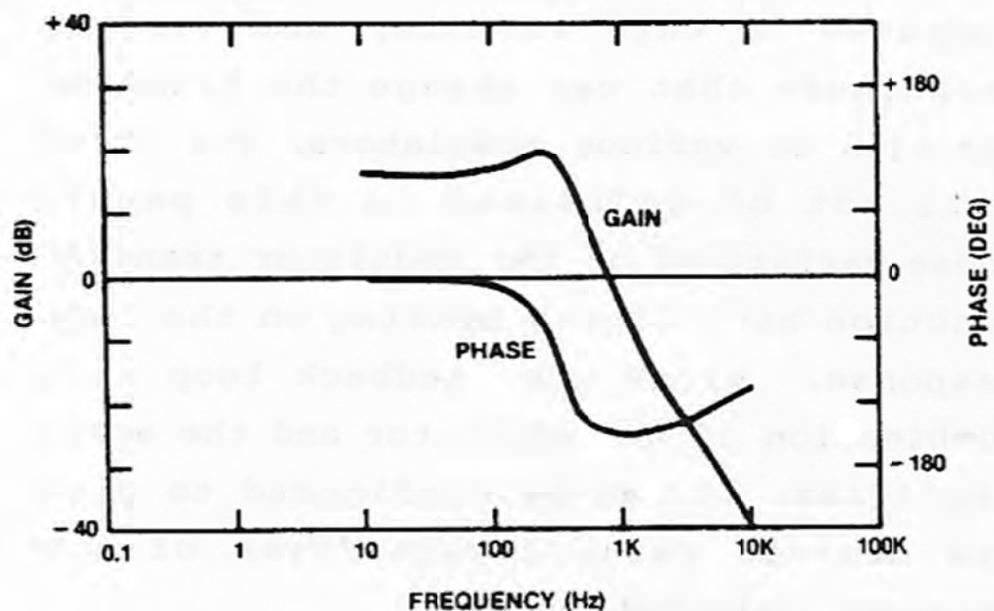


FIGURE 5. BODE PLOT OF BUCK REGULATOR

Since this is a second order filter, the gain falls off at a -2 slope or 40 dB/decade. The phase is also shifting at the same time. It will try to get to -180 degrees, but in this case, it cannot, due to the equivalent series resistance (ESR) of the output filter capacitor. The ESR becomes significant at the frequency where it equals the reactance of the capacitor. Above this frequency the filter becomes an L-R filter instead of an L-C filter. The gain then falls at a -1 slope or 20 dB per decade. The phase shift above this frequency will now start shifting upward toward -90 degrees. If the corner frequency of the ESR and the capacitor is below the L-C corner frequency, then the gain will fall at a 1 slope from the L-R corner frequency and the maximum phase shift will be -90 degrees. This is significant because we have encountered several supplies that have this phenomenon and they can become unstable when the end user connects them to a well filtered bus that has a lower ESR than the output filter capacitor. The additional phase shift due to this will reduce the phase margin, sometimes to the point where the supply oscillates. There are various other topologies that have slight variations in their transfer functions compared to this example, and various techniques that can change the transfer function of various modulators, but these will not be addressed in this paper. These variations of the modulator transfer function have little bearing on the loop response. Since the feedback loop is a combination of the modulator and the error amplifier, it can be configured to give the desired results regardless of the topology selected.

2.2 AMPLIFIER TRANSFER FUNCTION

The error amplifier is the part of your circuit that you create to make the loop do what you want it to do instead of what it wants to do. In normal practice the way to stabilize the feedback loop is simply to create an amplifier with the right amount of gain and phase shift at the chosen loop crossover frequency. There are three basic types of error amplifiers that are useful in the compensation of switch mode power supplies. The first, which we define as Type 1, is shown in Figure 6.

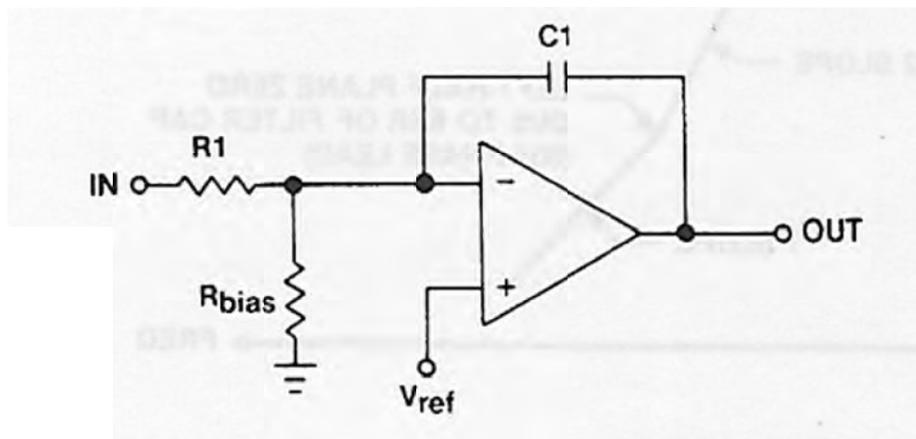


FIGURE 6. TYPE 1 AMPLIFIER SCHEMATIC

Since it is an inverting amplifier, it has 180 degrees of phase shift at DC, and due to the feedback capacitor, it has a pole at the origin which causes an additional 90 degrees of phase lag. This adds to a total of 270 degrees of phase lag. A pole is a point on the s-plane where the denominator of the transfer function goes to zero. From a practical standpoint this means that the slope of the Bode plot breaks downward as shown in Figure 7.

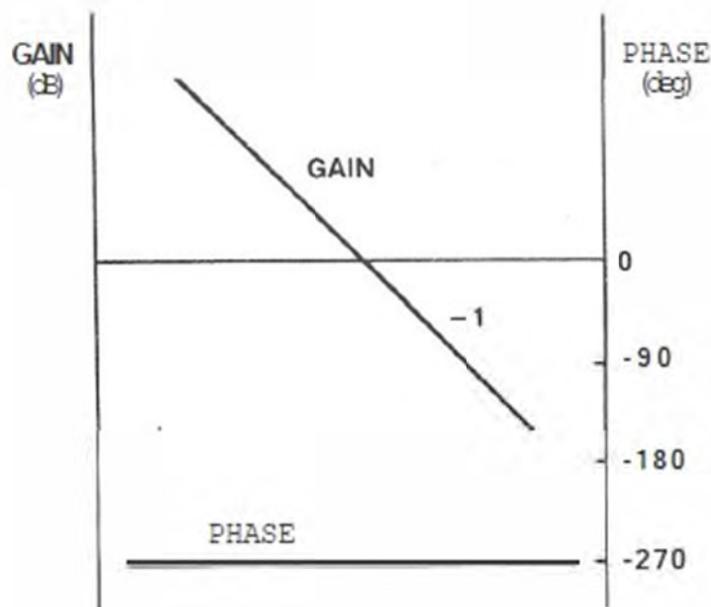


FIGURE 7. TYPE 1 AMPLIFIER TRANSFER FUNCTION

The gain falls at a -1 slope essentially forever, and crosses unity gain at the frequency where $R1$ and $C1$ have the same impedance. The phase is at -270 degrees at frequencies above DC.

A second type of amplifier, which we define as Type 2, also has a pole at the origin, but the resistor R2 in series with C1 causes a zero at the frequency where the impedance of R2 and C1 are equal. See Figure 8.

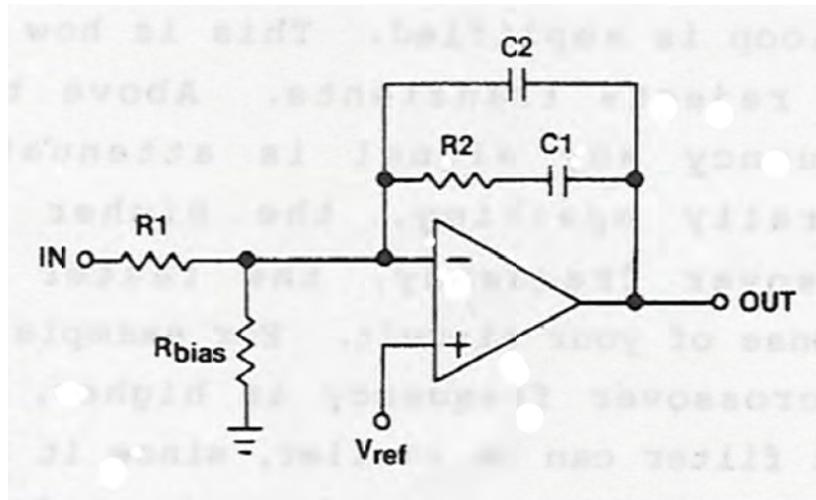


FIGURE 8. TYPE 2 AMPLIFIER SCHEMATIC

A zero is a point on the s-plane where the numerator of the transfer function goes to zero. This causes the Bode gain plot to break upward as shown in Figure 9.

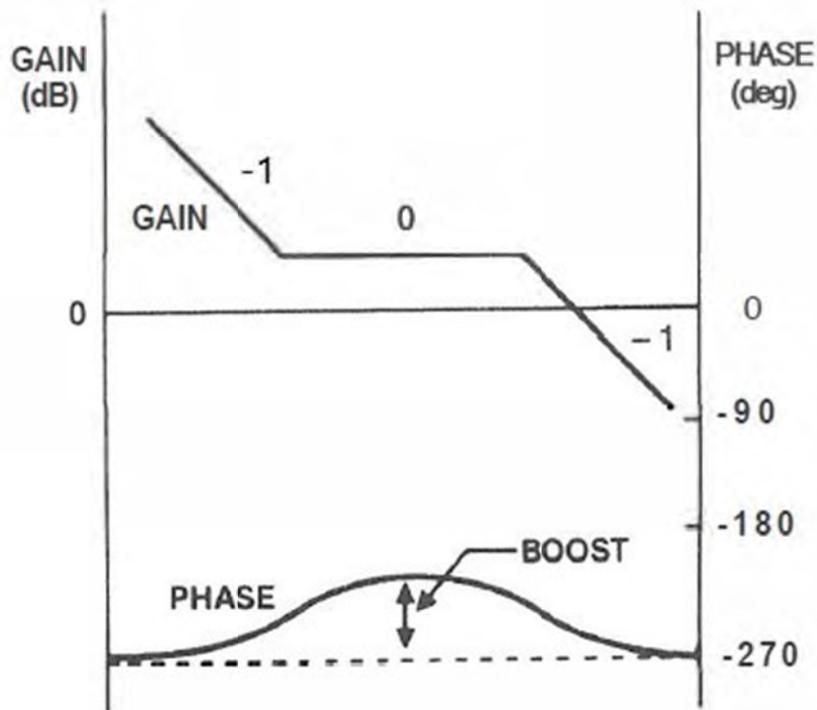


FIGURE 9. TYPE 2 AMPLIFIER TRANSFER FUNCTION

Since the slope below the frequency of the zero was a -1, the slope above the zero is 0. At a higher frequency, the impedance of C2 falls below R2 and another pole is introduced. The slope above this frequency is now -1 again. As we discussed earlier, there is a certain amount of phase shift associated with every slope. When the slope of the amplifier gain changes from -1 (180 degrees from the inversion and 90 degrees from the -1 slope for a total of 270 degrees of phase lag) to a 0 (180 degrees of phase lag due to the inversion only) it is said to have phase boost. The maximum amount of phase boost possible from this amplifier is 90 degrees. It is not practical to get all 90 degrees of boost from this amplifier since the zero and pole would have to be infinitely far apart. It is very important to understand at this point that there is no such thing as phase lead. The phase never really leads at all, instead it just lags less.

The third type of amplifier, which we define as Type 3, starts with a pole at the origin just like the previous two configurations. Instead of the single zero in the Type 2 amplifier, it has two zeros. See Figure 10.

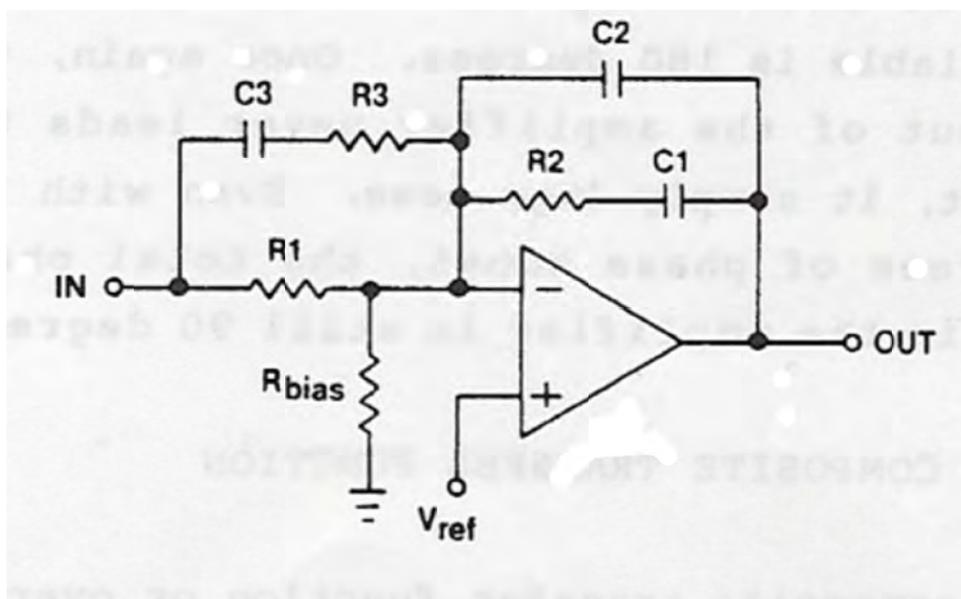


FIGURE 10. TYPE 3 AMPLIFIER SCHEMATIC

The second zero is caused by the impedance of C3 becoming less than R1. A second pole occurs when the impedance of C3 becomes less than impedance of R3. The most effective design calls for the two zeros to be coincident and the two poles to be coincident. A detailed analysis of this was presented at Powercon 10 in the now famous K-Factor paper. Just as in the Type 2 amplifier, the change of slope causes a change in the phase shift. See Figure 11.

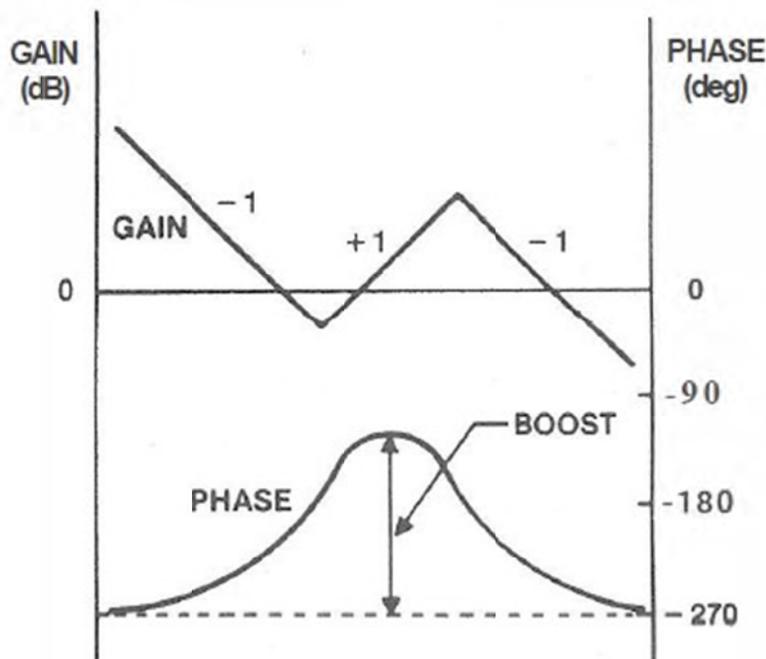


FIGURE 11. TYPE 3 AMPLIFIER TRANSFER FUNCTION

Now that there are two zeros, the slope changes from a -1 to a +1. The two poles then cause the slope to change back to -1. Since each slope change equates to a 90 degree shift in phase, the total boost available is 180 degrees. Once again, the output of the amplifier never leads the input, it simply lags less. Even with 180 degrees of phase boost, the total phase lag in the amplifier is still 90 degrees.

2.3 COMPOSITE TRANSFER FUNCTION

The composite transfer function or overall loop Bode plot can be simply stated as the combination of the modulator and amplifier together. The gains just multiply and the phase shifts add. As was stated earlier, we are expressing gain in dB. Because of this, we can simply add the two gain curves together and add the two phase curves together and we have the composite.

3. GENERAL FEEDBACK LOOP TERM DEFINITIONS

3.1 CROSSOVER FREQUENCY

CROSSOVER FREQUENCY is the frequency at which the loop gain is unity. This is the frequency where phase margin is measured. Below this frequency, any error signal in the loop is amplified. This is how the loop rejects transients. Above this frequency any signal is attenuated. Generally speaking, the higher the crossover frequency, the faster the response of your circuit. For example, if the crossover frequency is higher, the input filter can be smaller, since it does not need to be effective at as low a frequency as would be required with a slower loop.

3.2 PHASE MARGIN

PHASE MARGIN is the difference between the actual amount of loop phase shift and 360 degrees when the loop gain is unity. At DC it is obvious that any control system with negative feedback will be inherently stable. At higher frequencies, where reactive components and time delays cause phase shifts, potentials for instability exist. Due to the inversion caused by negative feedback, we start out with 180 degrees of phase shift. If we add an additional 180 degrees of phase shift from other sources such as the L-C filter, compensation components used in the error amplifier, time delays, etc., a signal progressing around the loop will be shifted a total of 360 degrees and come back in phase **with** the original signal. If this happens when the net gain around the loop is unity, we have an oscillator. The relationships are shown in Figure 12.

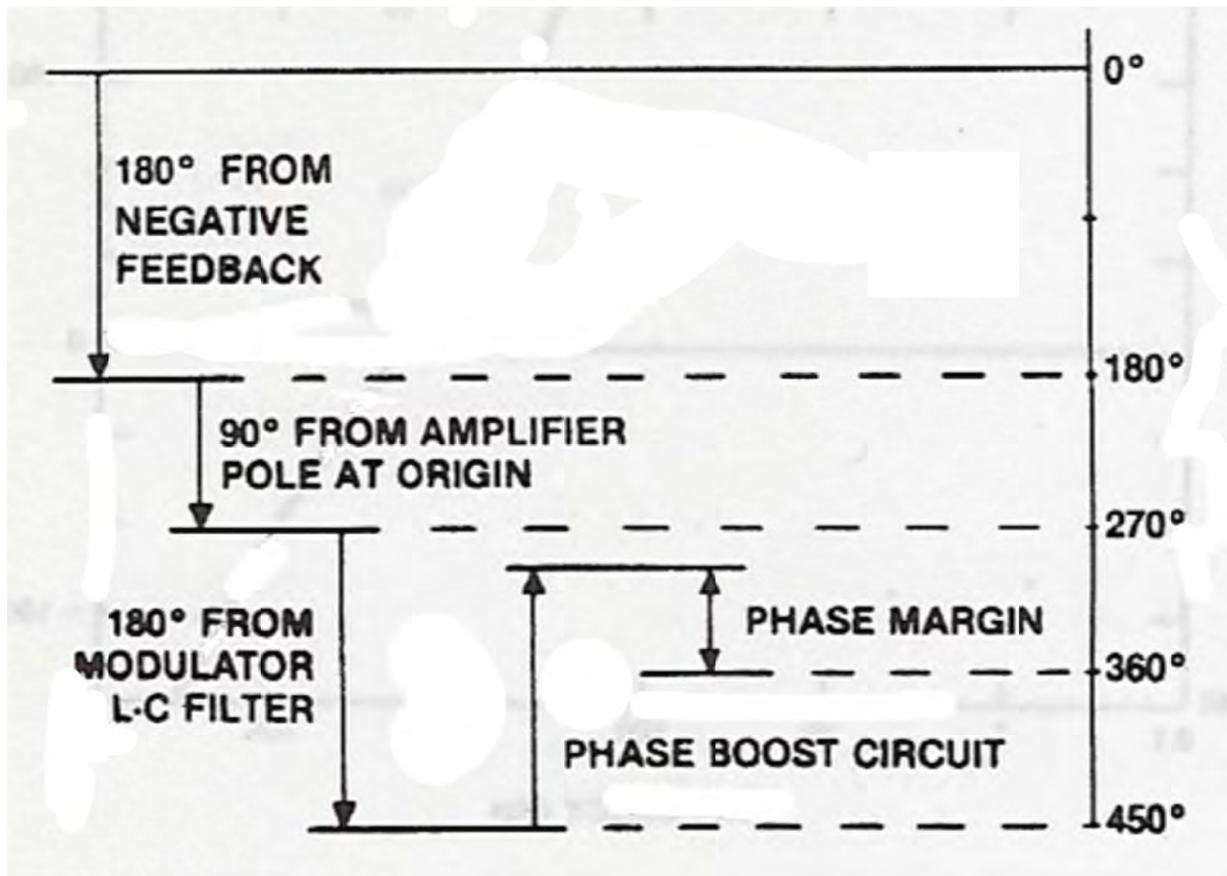


FIGURE 12. LOOP PHASE SHIFT

3.3 GAIN MARGIN

GAIN MARGIN is the amount that the gain has fallen below unity when the phase has shifted 360 degrees. Both gain margin and phase margin can be seen in Figure 13.

4. TEST SETUP AND PROCEDURES

The testing documented in the following section of this paper was performed at the test lab facility of Venable Industries, Inc. The equipment used in the testing was as follows:

- ACDC EL300 Electronic Load Test Instrument
- ACDC EL301 Electronic Load Control Module H-P 1740A Oscilloscope
- H-P 3561A Dynamic Signal Analyzer
- Lambda LP-522-FM Variable DC Power Supply
- Venable Industries Model 250 Frequency Response Analysis System

The testing was done as follows. The Lambda power supply was used to provide the +15 VDC to operate the unit under test (UUT). The ACDC load was used with its controller to perform step load tests when the step load tests were performed at line frequency. When transient testing at other frequencies was desired, we used the Model 250 system to inject a transient into the loop. Real time display of the transient response as well as the verification of the small signal nature of the testing were done on the oscilloscope. The graphic hard copy display of the transient response was generated by the dynamic signal analyzer. These could have been done by the oscilloscope and a camera, but we felt that the graphics plotter output would reproduce in print better. We used the graphics plotter of the Model 250 system to make plots with the signal analyzer. The Bode plots were created by the Venable Industries Model 250 frequency response analysis system. We also used the Model 250 for the design of the error amplifiers, using the built-in **circuit** analysis software, and the necessary computer modeling to verify that the circuit worked as designed.

5. THE RELATIONSHIP OF CROSSOVER FREQUENCY TO TRANSIENT RESPONSE

Figure 13 is an overall loop Bode plot of a buck converter. The crossover frequency of this loop is 2.5 kHz and the phase margin is 60 degrees. Figure 14 is the transient response for the same power supply. Figure 15 is an overall loop Bode plot of the same power supply with different compensation components in the error amplifier to yield 2.5 kHz and 30 degrees of phase margin. Figure 16 is the transient response of this power supply. By comparing Figures 14 and 16 we can see what difference this 30 degree change in phase margin made. Figure 16, which had a lower amount of phase margin, responded to the transient more quickly than did Figure 14. The only difference between Figure 14 and Figure 16 is the feedback loop compensation which changed the phase margin. It would appear the way to speed up a loop would be to simply remove as much

phase margin as possible, but without sufficient phase margin the loop will oscillate at the loop crossover frequency. The loop crossover frequency of these two examples is approximately 10 times the L-C corner frequency. Because the loop has gain at the L-C corner frequency, the loop will reject any oscillations caused by the L-C output filter resonating as a tank circuit.

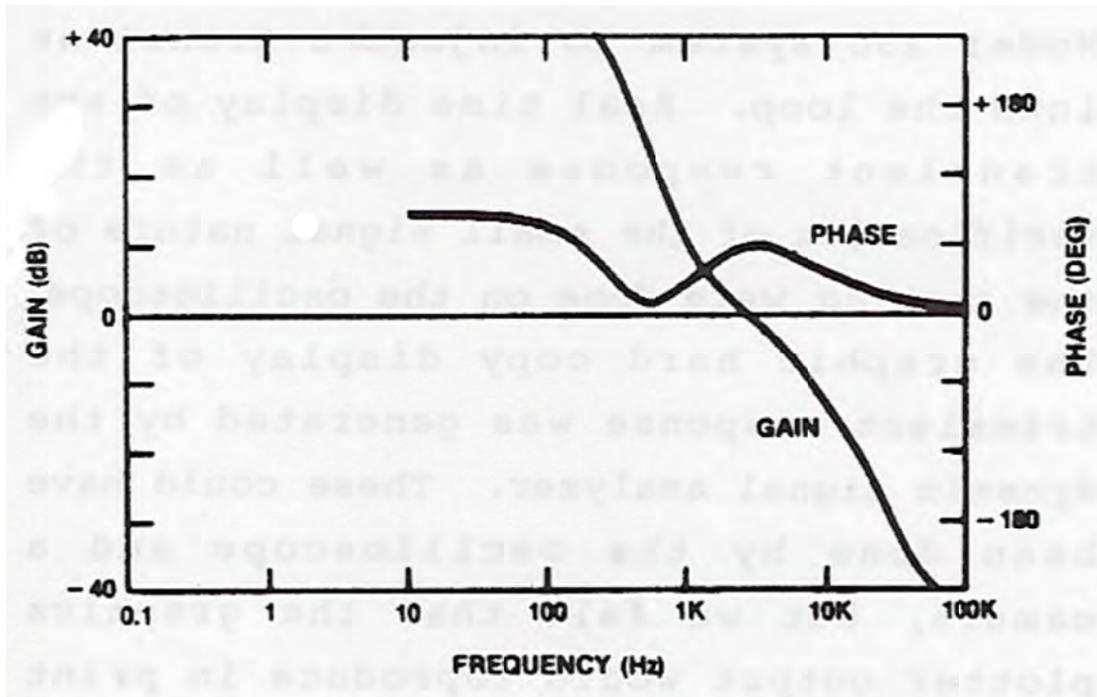


FIGURE 13. OVERALL LOOP BODE PLOT
2.5 KHZ, 15 DEGREES COMPENSATION

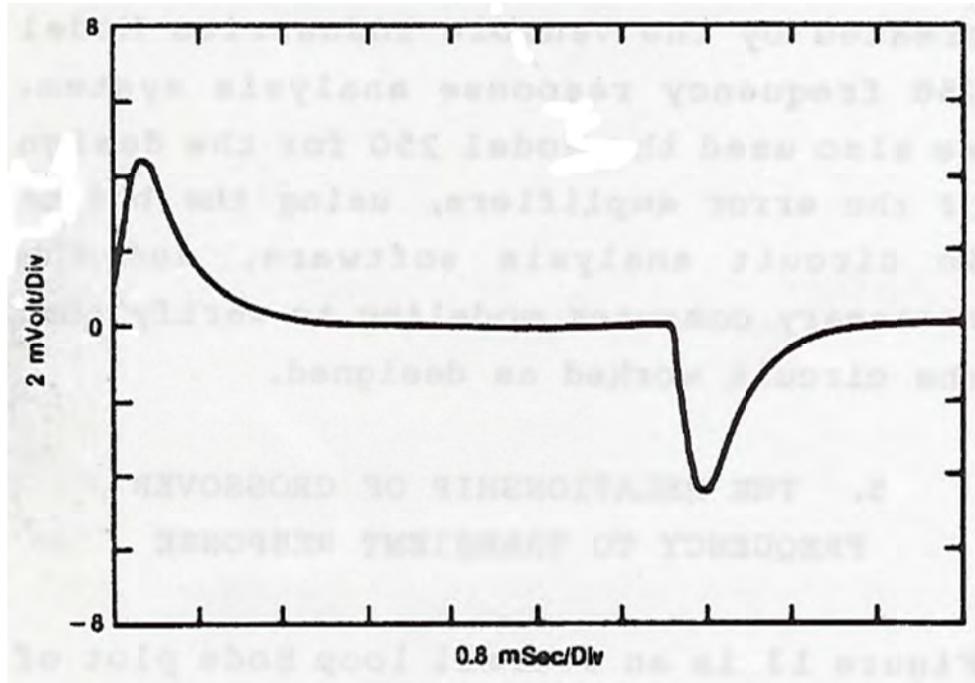


FIGURE 14. TRANSIENT RESPONSE
2.5K KHZ, 60 DEGREE COMPENSATION

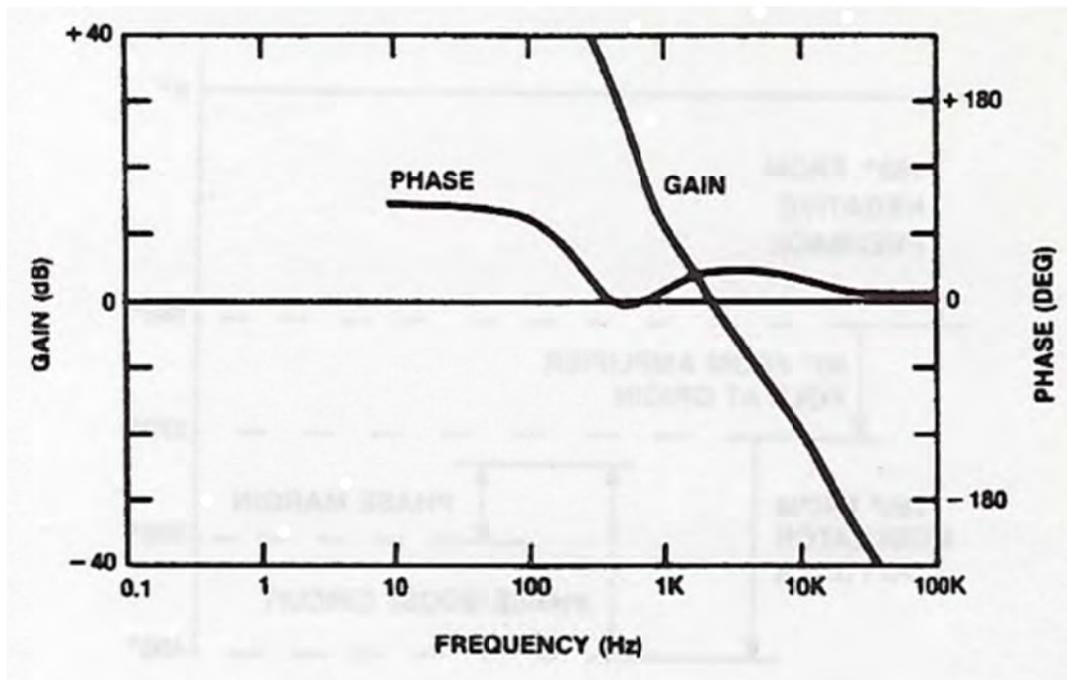


FIGURE 15. OVERALL LOOP BODE PLOT
2.5 KHZ, 30 DEGREES COMPENSATION

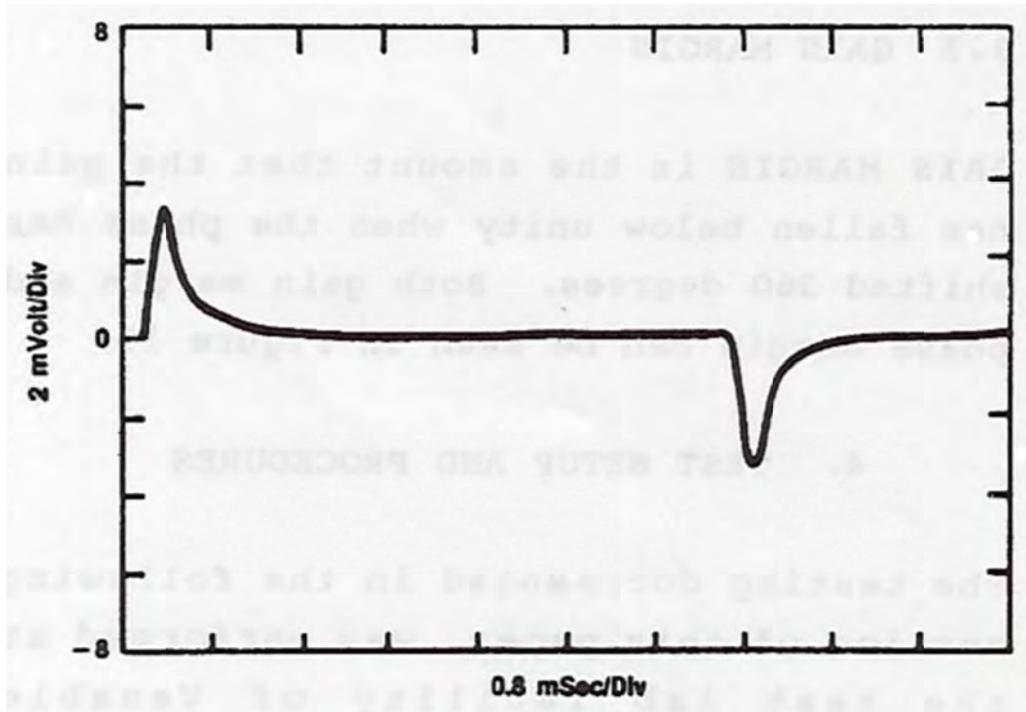


FIGURE 16. TRANSIENT RESPONSE
2.5 KHZ, 30 DEGREES COMPENSATION

Both of these examples are stable systems. Both possess phase margin and gain margin. Figure 15 illustrates one of the more fascinating aspects of feedback control loops, that of conditional stability. Notice that the phase shift is 360 degrees at two different times during the plot. At frequencies in between these two points the phase shift exceeds 360 degrees. In order for the feedback control system to be unstable, the gain must be unity and the phase must be 360 degrees at the same frequency. Conditional stability is defined as the feedback control loop having more than one point where the phase shift is 360 degrees. In unconditionally stable systems the gain can decrease with no effect on the stability. Increasing the gain will cause the phase margin to reach zero and cause instability. A conditionally stable system can become unstable if the gain either increases or decreases. This is the reason to avoid allowing your feedback loops to be conditionally stable. In unusual operating modes the gain may be decreased for a number of reasons. One of the most common reasons is an amplifier, or other stage, of the power supply becoming saturated. If the decrease in gain is sufficient to remove the phase margin, the circuit will oscillate. The circuit may seem to perform well under all other conditions, but a potential disaster awaits the end user. Conditionally stable power supplies are far more common than would normally be expected. This, and reliance on the ESR zero of the output filter capacitor for stability are among the most common causes of power supplies oscillations in the field.

Figure 17 is the overall loop Bode plot of the same power supply with a crossover frequency of 250 Hz and 60 degrees of phase margin. The L-C corner frequency is still 260 Hz. Figure 18 is the transient response for this particular circuit.

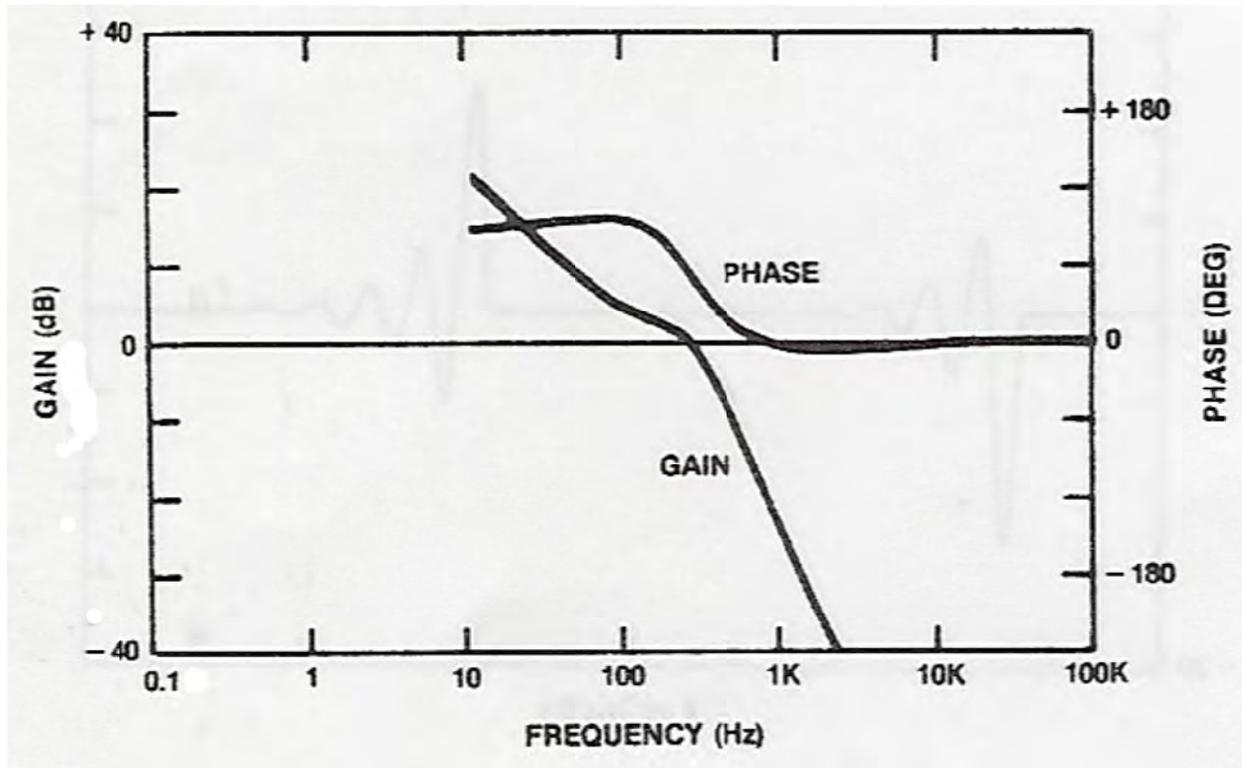


FIGURE 17. OVERALL LOOP BODE PLOT
250 HZ, 60 DEGREES COMPENSATION

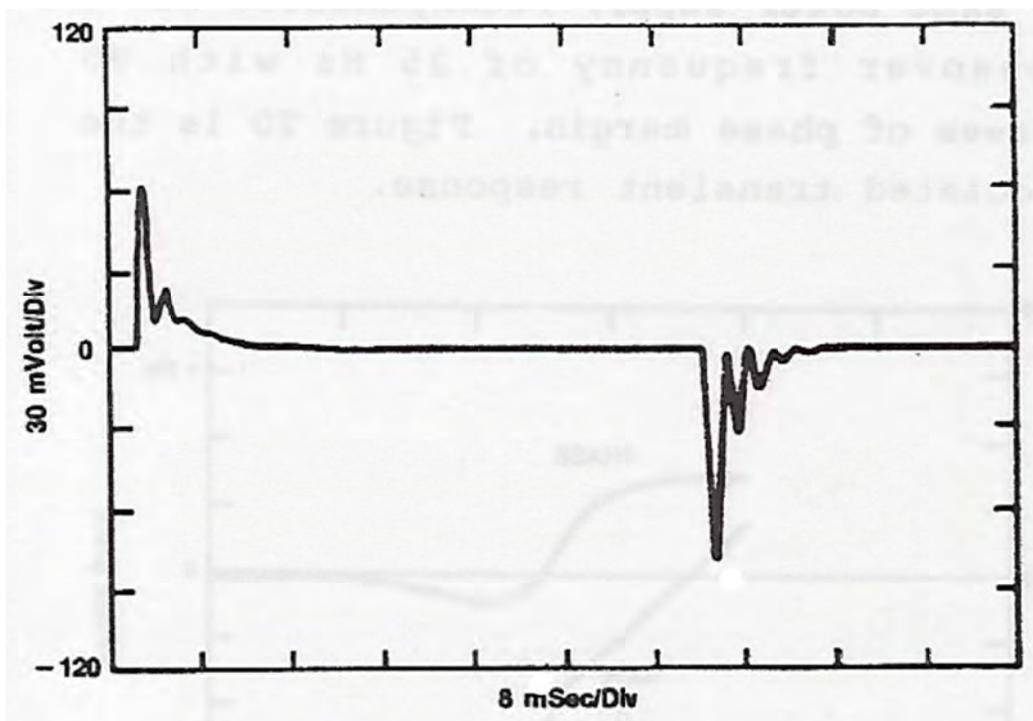


FIGURE 18. TRANSIENT RESPONSE
250 HZ, 60 DEGREES COMPENSATION

The oscillations that occur during the response to the transient are not due to low phase margin. As you can see in Figure 17, the phase margin is 60 degrees which is sufficient to prevent oscillation and overshoot. Referring back to Figure 18 we can see that the oscillations are superimposed on the normal response curve. These oscillations are caused by the resonance of the output filter inductor and capacitor. If we eyeball the curve and average out the oscillations, we will see that the shape of this plot looks very much like Figure 14. These oscillations gave rise to the myth that the power supply will oscillate at the L-C corner frequency instead of the loop crossover frequency. This myth abounds in the industry and is caused by all the supplies that have loop crossover frequencies below the resonance of the output filter. One of the interesting things that came about as a result of this piece of misinformation is the belief that feedback controls lie in some middle ground between black magic and science. Once properly understood, feedback controls can be easy and enjoyable.

Figure 19 is an overall loop Bode plot of the same power supply recompensated for a crossover frequency of 25 Hz with 90 degrees of phase margin. Figure 20 is the associated transient response.

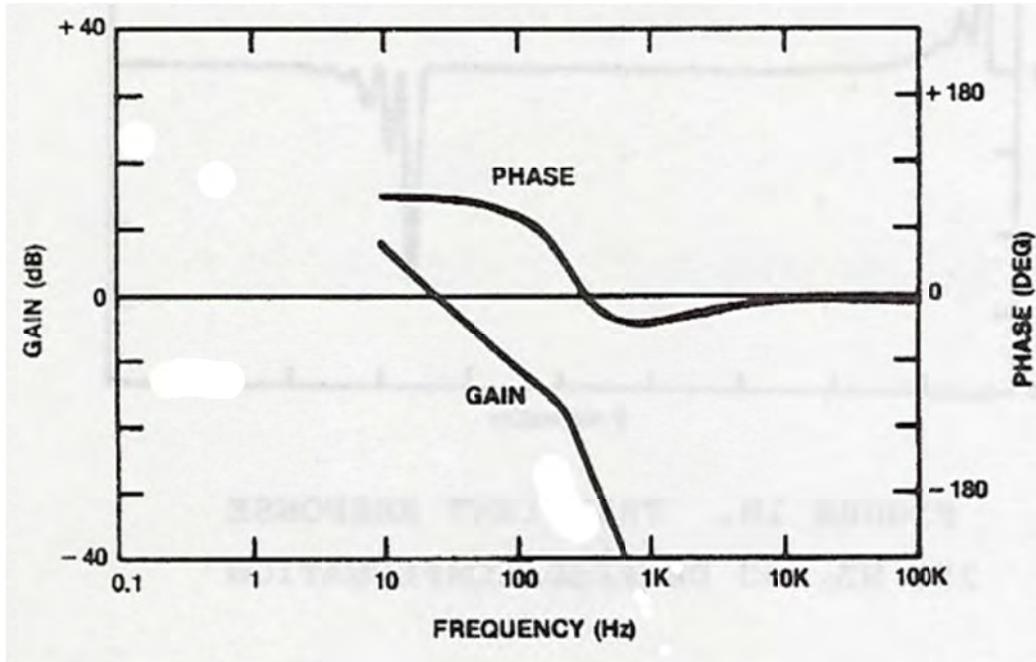


FIGURE 19. OVERALL LOOP BODE PLOT
25 HZ, 90 DEGREES COMPENSATION

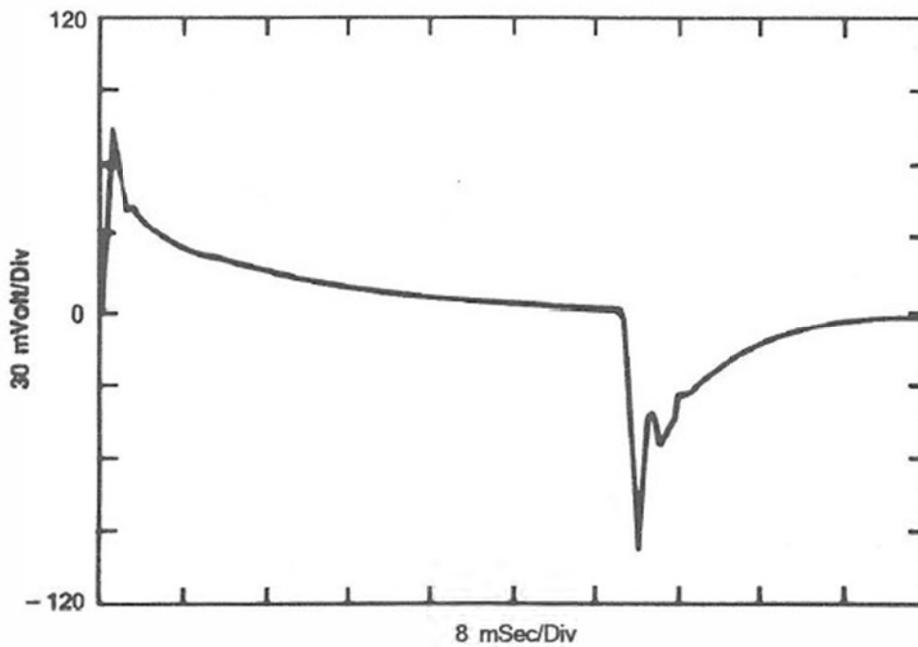


FIGURE 20. TRANSIENT RESPONSE
25 HZ, 90 DEGREES COMPENSATION

The reason it responds more slowly to the transient than the previous ones is the lower crossover frequency. The reason that the crossover frequency affects the speed at which the loop can respond to the transient, is because it affects the amount of low frequency gain. The higher the gain at low frequency, the more quickly the loop can respond to a transient. Although the loop is responding to the transient with no overshoot, we can still see the oscillation of the output filter. If we have insufficient phase margin in the feedback loop we will have ringing at the crossover frequency which is independent of the L-C corner frequency.

Figure 21 is an overall loop Bode plot of this same power supply crossing over at 2.5 kHz with 15 degrees of phase margin. Figure 22 is the transient response with this compensation.

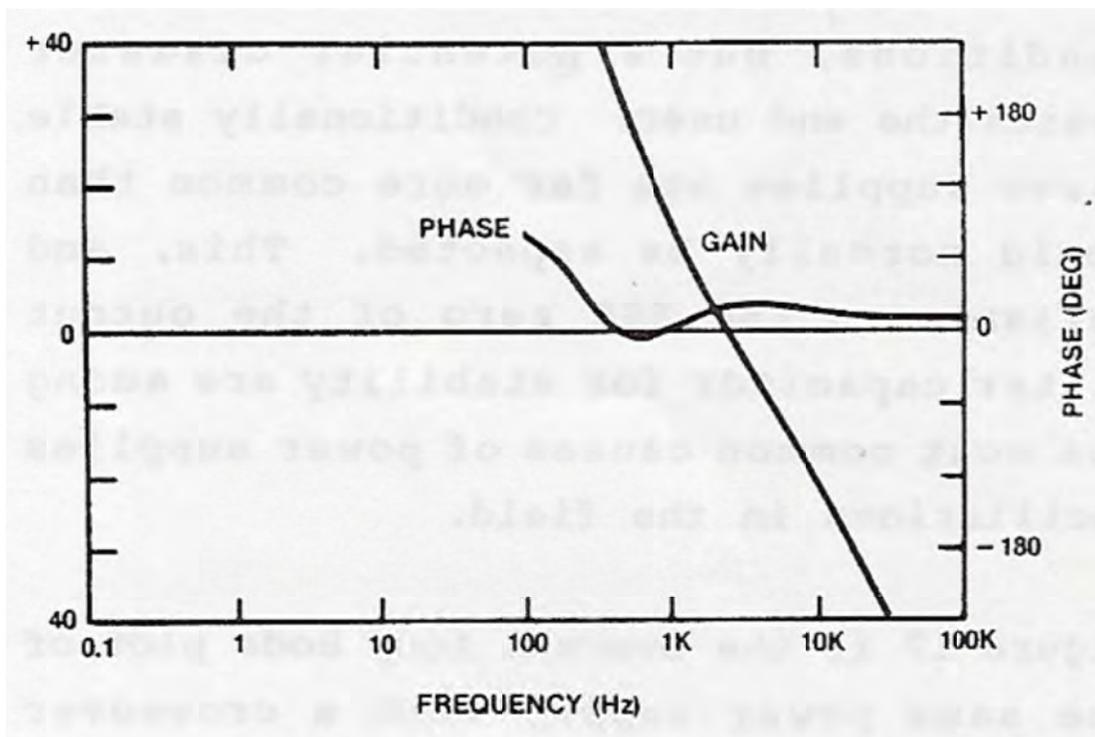


FIGURE 21. OVERALL LOOP BODE PLOT
2.5 KHZ, 15 DEGREES COMPENSATION

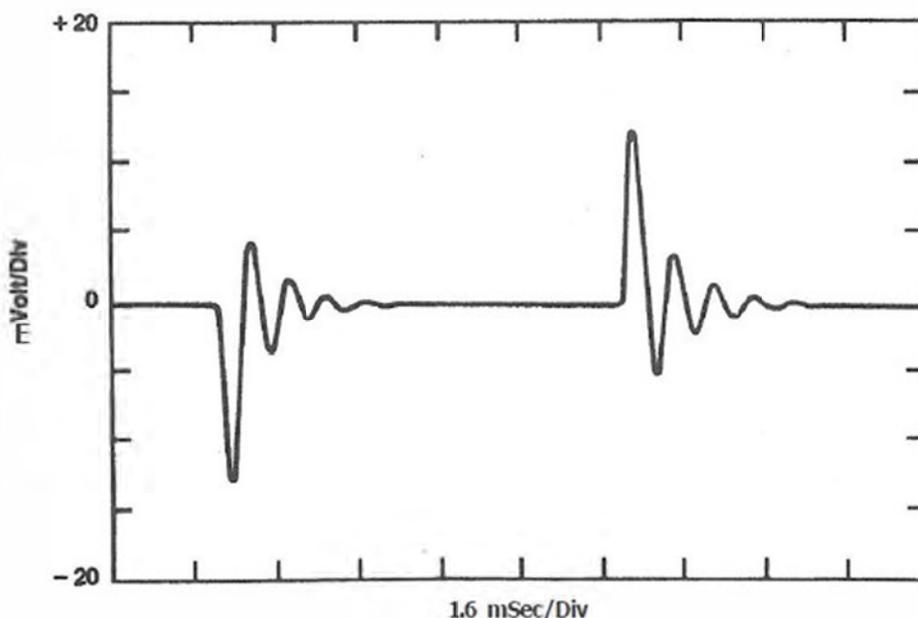


FIGURE 22. TRANSIENT RESPONSE
2.5 KHZ 15, 15 DEGREES COMPENSATION

When hit with the transient, the output rings at the crossover frequency, eventually damping out and recovering.

These examples demonstrate several important facts about feedback loops in power supplies.

5.1 IMPORTANT POINT #1

The higher the crossover frequency, the faster the loop response. This, like most truisms of the world, has to be qualified. Sometimes when pushing the crossover frequency higher and maintaining constant phase margin, we may have less low frequency gain and therefore a slower transient response. It is not obvious at first, but can be explained simply. If we move the crossover frequency higher, through a range in which the phase shift in the modulator is increasing, we will need more phase boost from the error amplifier. The way we get this boost is by increasing the separation between the zeros and poles in the compensation network. As we do this, the actual gain from the amplifier is lower at low frequency and higher gain at high frequency than it would have been with a slightly lower crossover frequency. The composite Bode plot will therefore have lower gain at low frequency and higher gain at high frequency. Both of these situations are generally bad. The depressed gain at low frequency will slow down the loop's ability to respond to a transient. The increased high frequency gain causes more high frequency noise to be present in the feedback loop. If we increase the crossover frequency in a range where the modulator phase shift is not increasing rapidly, we may well gain in the ability to respond to a transient. This was

documented in the Powercon 10 paper on the K-factor which was referenced earlier. Using the equations presented in that paper, the design engineer can determine what truly is the optimum loop crossover frequency for a particular circuit.

5.2 IMPORTANT POINT #2

For a given crossover frequency, the higher the phase margin, the slower the transient response. You can see from Figures 14 and 16 that while nothing but the phase margin changed, the loop responds more quickly with 30 degrees than 60 degrees. This is caused by the decreased low frequency gain that resulted from the increase in phase boost in the amplifier that gave 60 degrees of phase margin. Of course there are practical limits of how little phase margin you can tolerate in your designs. As a design engineer you need to be certain that your circuit will be stable under all conditions of line, load, and temperature. To meet this requirement your circuit will need to have a worst case phase margin of at least 30 degrees.

5.3 IMPORTANT POINT #3

When the loop crossover frequency is at or below the L-C corner frequency, the feedback loop is not capable of rejecting the oscillations caused by the resonance of the output filter inductor and capacitor. These oscillations will appear during the response to a transient, but can be differentiated from loop oscillations by observing the output voltage. If it is oscillating on either side of the nominal output, as in Figure 22, it is a loop oscillation. If it is oscillating during the transient response, as in Figure 20, these are oscillations exclusive of the feedback loop.

6. HOW TO OPTIMIZE TRANSIENT RESPONSE

Optimizing the transient response of the power supply can be done in two ways. The first way that we will discuss is the preferred way. As was stated in section 5.1, we can speed up the response by pushing the loop crossover frequency as high as possible with reasonable phase margin. The low cost of increasing the loop bandwidth is very reasonable on a cost per supply basis. The only cost is the feedback components in the error amplifier. You will probably use the same number of these even if you have a slow loop. Many times the overall parts cost of the supply will go down because of the reduced cost associated with smaller (and possibly fewer) output filter capacitors and smaller input filters. The reason for smaller output filter capacitors is that if the loop reacts fast enough, the capacitor will not have to supply all the power to the load for as long a period of time. The input filter may get smaller because it only needs to be effective above the loop gain crossover. Either the input filter or the loop has to attenuate at any given frequency. The amount of overlap between the two should be minimal. The higher the loop bandwidth, the higher the minimum frequency of the input filter will need to be. The one significant cost associated with this solution to increased transient response is in instrumentation. Computer modeling does not usually work well at high frequencies. Quite often the parasitics are left out of the model. Parasitic series resistance

and inductance are very difficult to predict. Stray capacitance, irregularities due to packaging, and time delays all start to become problems as we get higher in frequency. The only way to be certain that your circuit will be optimum and have sufficient margins for worst case operation is by measurement of the feedback loop characteristics. Time delays, which show up as linear increases in phase shift with frequency, are only a problem if you do not know that they exist. Parasitics can be dealt with in the error amplifier easily if their effect on the gain and phase shift of the modulator is understood. The instrumentation cost is a one-time expenditure that is repaid quickly in reduced engineering time and reduced manufacturing costs.

The second way to improve the transient response is to use oversized output filter capacitors and damp any high-Q L-C filters. The costs here are mostly associated with manufacturing. The increased cost of large capacitors used in the output filter and in damping networks can be tremendous in large production runs when compared with the components that can be utilized with faster loops. The reason for the larger output filter capacitors is to try to maintain the output voltage during the time the loop takes to respond. The damping of the L-C filters is to prevent ringing at resonance at frequencies higher than the loop gain crossover.

7. WHAT ABOUT Q'?

The transfer function of the modulator used in this paper has a Q of 1. If it had a higher value of Q, we would have seen more peaking at the L-C corner frequency, a more abrupt shift in the phase, and more ringing at the L-C corner frequency when the loop gain was insufficient to reject it. The Q of the input filter can also have some exciting effects on the modulator transfer function. As the Q increases, the output impedance at resonance also increases. It is easy to imagine the difficulty of the converter trying to pull power through the input filter if the output impedance is high. Since this will look like a reduction in input line to the converter at the input filter's resonance, the gain of the modulator will have a dip in it. This may cause the power supply to go unstable by causing loop gain to go to unity and the phase shift to 360 degrees. If this dip is deep enough, the power supply will not be able to reject noise or transients at this frequency.

8. HOW TO FIX THE DAMP THING

Damping L-C filters is really easy, even if you leave your calculator at home. All you need is Reactance-Frequency graph paper. Just remember that you are trying to limit the amount of output impedance that the L-C circuit can have at resonance. The easiest way to do this is to put a resistor in parallel with the capacitor. This limits the impedance to the maximum of the resistor's value. There is only one problem with this method. The resistor will get very hot. Since this heating of the resistor is a safety hazard and waste of electricity, you will probably want a blocking capacitor in series with the damping resistor. This is the classic Type 6 damping method as explained in depth in Keith Phelps' and Will Tate's award winning paper at Powercon 6.

We will start our search for dampness by first selecting the damping resistor. The entire process for damping the L-C filter is contained in Figure 23.

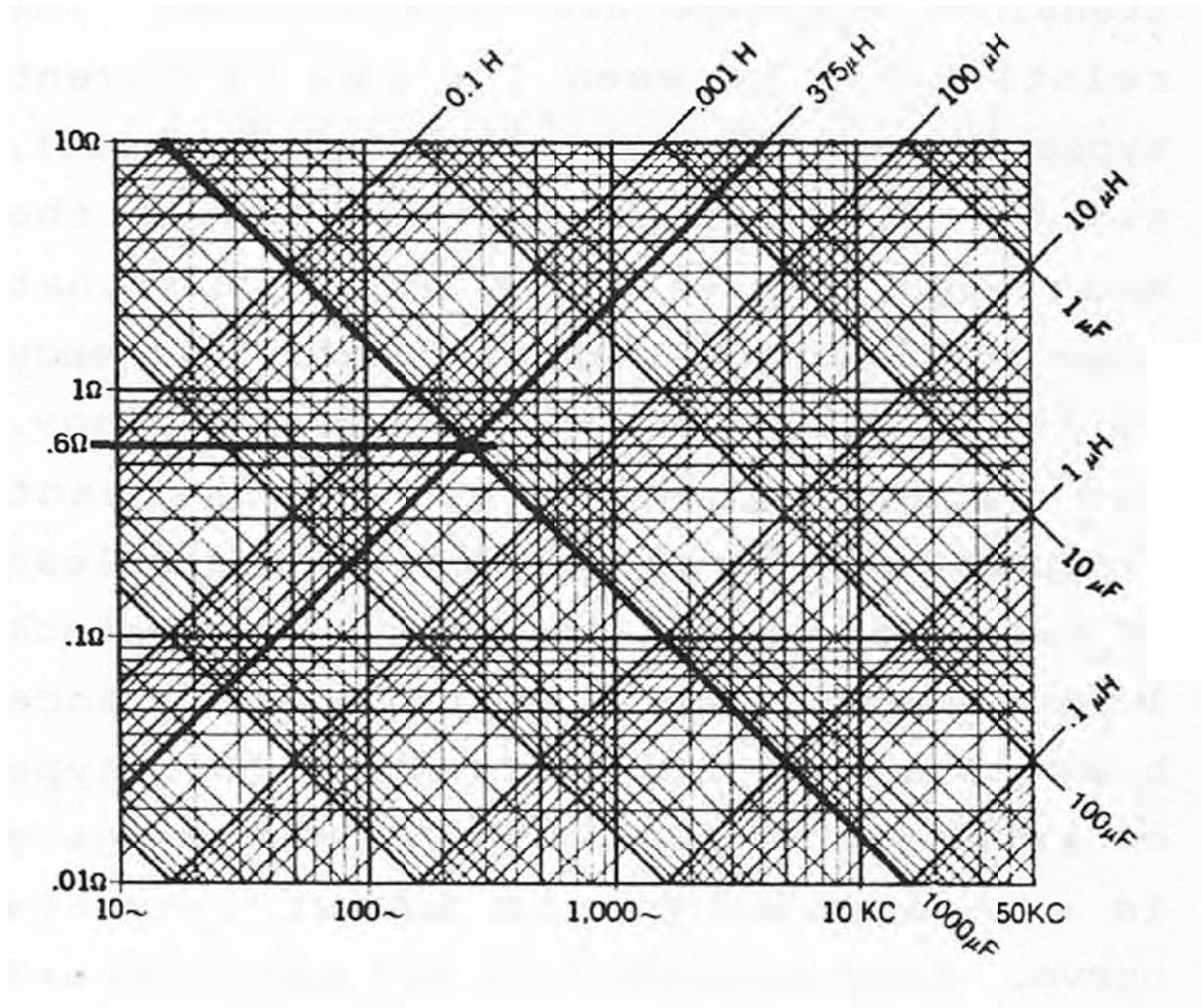


FIGURE 23. USING REACTANCE-FREQUENCY PAPER TO CALCULATE DAMPING RESISTOR VALUE

First, find the constant inductance line on the reactance paper representing the inductor of the filter to be damped (for this example, 375 microhenrys). Highlight this line. Now find the -constant capacitance line of the output filter capacitor and highlight this line (1000 microfarads). The intersection of these two lines determines the impedance at which they will be at resonance (0.6 ohms). For a Q of 1, choose a resistor that is the same impedance as the L and C at their resonant frequency. To choose the blocking capacitor size, a good rule of thumb is to make the blocking capacitor at least four times the capacitance of the filter capacitor. As you can see, we could have chosen any amount of impedance at which to limit the filter at resonance. This means that we can control Q for the input and output filters, all without the aid of a calculator.

8. SUMMARY AND CONCLUSIONS

The composite transfer function and the transient response are interrelated. The relationship between the two different types of test data was explored in detail, and the findings presented. One of the most important of these is the fact that when the feedback loop crossover frequency is lower than the L-C corner frequency, the output can ring at the resonant frequency of the output filter, regardless of the phase margin of the actual feedback loop. It is easy to tell the difference between a loop oscillation and this type of resonant ringing. The resonant ringing is superimposed on the normal response curve. Loop oscillations are superimposed on the normal output voltage. The way to correct both of these phenomena is the same. Cross the loop over at a higher frequency (well above the L-C corner) with sufficient phase margin. If, for some reason, you cannot achieve a high enough crossover frequency, damping the L-C output filter will prevent the output from oscillating independent of the feedback loop. The design trade-off here is the speed of the response and the cost of a very large capacitor (four times the size of the output filter capacitor). The best way to design the feedback portions of any power supply is to measure the transfer function of the modulator at all combinations of line, load, and temperature. Then synthesize an error amplifier with the correct amount of gain and phase boost at the chosen crossover frequency. Use the K-Factor equations to determine if this is optimum for your particular circuit. Close the loop and test the power supply by measuring the overall loop transfer function and performing transient testing to verify that the design meets the specifications.

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