

Loop Gain Crossover Frequency

by Dr. Ray Ridley

Power supplies are one of the last holdouts of true analog feedback in electronics. For various reasons, including cost, noise, protection, and speed, they have remained this way in the face of the digital control revolution. And it is likely they will stay this way for decades to come. It is still the easiest and most effective control method for PWM converters.

In power supplies, we still measure and characterize the control system in traditional ways—injecting a swept sinusoidal test signal into the components of the system, and measuring the analog response. Unfortunately, most engineers receive little training or guidance in optimizing their systems.

Suppose you have a choice of two power supplies for your system—power supplies A and B. Power supply A has a crossover frequency of its control loop of 1 kHz. Power supply B has a crossover frequency of 20 kHz. Which is a better design?

The answer is that there is not enough information from which to choose. If power supply A and power supply B meet all other criteria, such as cost, size, efficiency, regulation and dynamic loading, the crossover frequency of the control loop is irrelevant. Loop gain crossover has no intrinsic value. It is only useful as a tool to design a power supply that is stable and provides the needed voltage source at its output.

If all other criteria were equal, power supply A with the lower loop crossover is the better choice. A more conservative loop crossover frequency is less prone to noise, component variations, and power supply operating point.

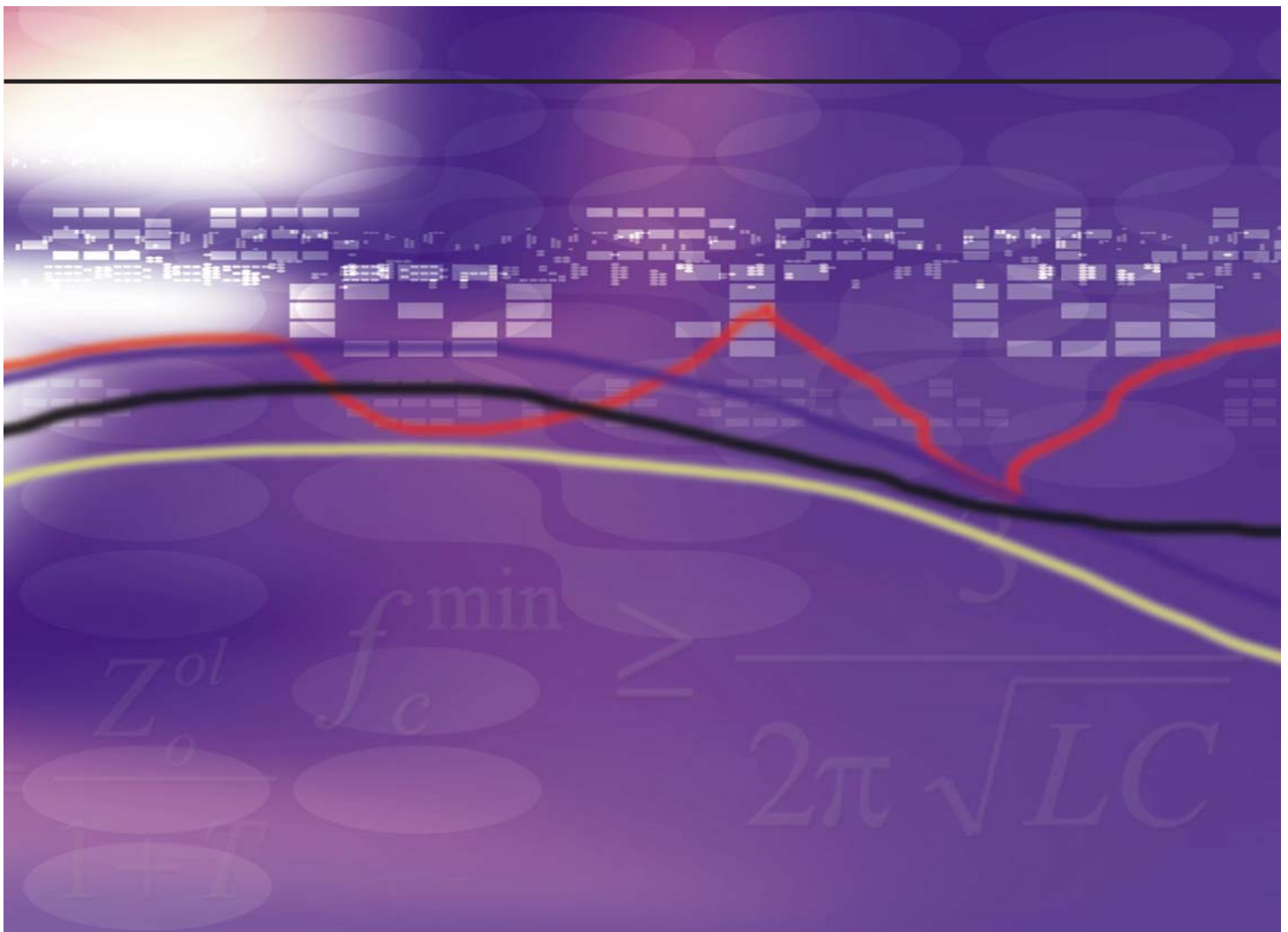
So how should loop gain crossover frequency be selected? The answer is complex, depending on several factors:

- Power supply topology
- Power supply switching frequency
- Control method – voltage-mode or current-mode control
- Output capacitor type and characteristics
- Power stage components

We will lay down the basic boundaries for crossover selection in this article. In the next issue of *Switching Power Magazine*, we'll look at other system issues such as reactive loading, to see how this affects the design of the loop. This is a very important topic with the proliferation of DC-DC converter modules, many of which expect the user to add substantial output capacitance, but do not give access to the control loop.

The Purpose of Loop Gain

Loop gain is not an end in itself—it is a means to achieve other critical performance measures of a power supply. For most power supplies, the objective is to provide a voltage



source to the end user. An ideal voltage source is a fixed voltage regardless of any changing input to the power supply or load.

In reality, a switching power supply is less than ideal. The fixed DC voltage is determined by the quality of reference used, and the dynamic impedance associated with the output will affect load regulation. There is also a dynamic transfer function which transfers noise from the power input through to the output. Loop gain is a tool we use to minimize the real objectives—lowest possible output impedance, and lowest possible transmission of noise to the output.

The output impedance of the closed loop converter is given by the open-loop output impedance, divided by the system characteristic equation (1 plus the loop gain.)

$$Z_o^{cl} = \frac{Z_o^{ol}}{1 + T}$$

The closed loop input to output noise on a converter is given by the open loop noise divided by the characteristic equation:

$$K_a^{cl} = \frac{K_a^{ol}}{1 + T}$$

The closed loop characteristics should be minimized for a good voltage source. That means making the loop gain, T, as large as possible. We cannot arbitrarily increase the loop

gain, however, due to the complex frequency characteristics of typical power converters. In the rest of this article, the limits of crossover frequency for practical converters will be defined.

Topology Equivalence

The equations in this article are defined for the classic buck, boost, and flyback converters. They are also applicable with the other converters in the same family. The buck family includes (but is not limited to) the forward, two-switch forward, half-bridge, full-bridge and push-pull converters. The boost converter is not usually isolated, but if it is, the equations still apply as long as the inductor value is reflected to the secondary side of the isolation transformer. The flyback and buck-boost converters also have the same characteristics. With the flyback, the inductor value is the magnetizing inductance of the transformer reflected to the secondary side.

Resonant Frequency Lower Crossover Limit

If you are using voltage-mode control and operating in continuous conduction mode (CCM), an LC resonant filter is being driven by the controlled switching action. That filter will naturally ring, and the control must eliminate this. For voltage-mode control, there is only one loop. This loop must have significant gain at the power filter resonant frequency. So, the lower bound must be, for the different converters:

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$$f_c^{\min} \geq \frac{3}{2\pi\sqrt{LC}} \quad \text{Buck Converter}$$

$$f_c^{\min} \geq \frac{3}{2\pi\sqrt{L_e C}} \quad \text{Boost, Flyback Converters}$$

$$\text{where } L_e \leq \frac{L}{D^2}$$

For the boost and flyback converters, the resonant frequency moves with input line. The highest value of the frequency occurs at high line operation.

When working with current-mode control, there is no lower bound constraint from the filter resonant frequency. The current feedback loop eliminates the LC filter dynamics. In many cases, a converter with low RHP zero (see RHP article) can only be controlled effectively with current-mode control.

RHP Zero Crossover Limit

For converters with RHP zeros, the crossover is constrained by the power stage dynamics. (See RHP article). It is recommended that the crossover frequency should not exceed 1/3 the RHP zero frequency.

$$f_c^{\max} \leq \frac{R_L}{6\pi D L_e} \quad \text{Flyback Converter}$$

For this converter example, with a design for reasonable ripple as shown in *Figure 1a*, the crossover frequency is restricted by the RHP zero at 20 kHz to 6.8 kHz. Make sure you calculate the allowable crossover frequency at lowest input line (i.e. maximum duty cycle) and maximum load. This is where the RHP zero has its lowest value.

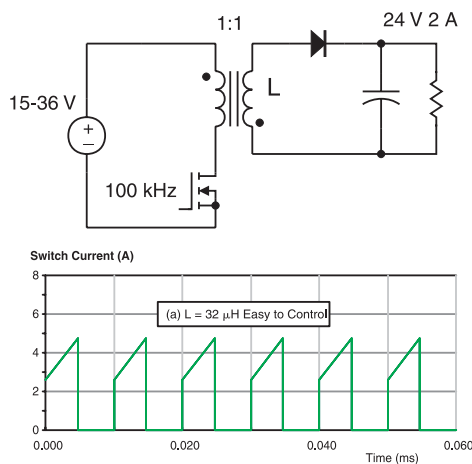


Figure 1a. Inductor current of flyback converter 100 kHz, $L=32 \mu\text{H}$

The controllability of a flyback converter is determined by the amount of ripple current allowed. The larger the inductor, the smaller the ripple, and the harder it is to control. For this converter example, with a design for low ripple as shown in *Figure 1b*, the crossover frequency is constrained to 1 kHz. It is not necessary to take the inductor all the way down to a value which allows only DCM operation, shown in *Figure*

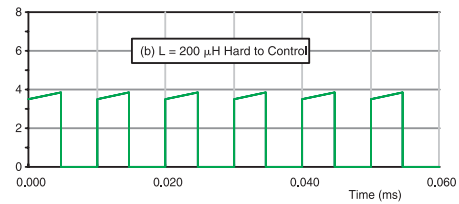


Figure 1b. Inductor current of flyback converter 100 kHz, low ripple $L=200 \mu\text{H}$

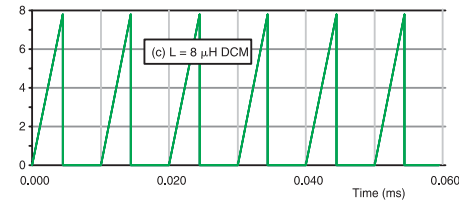


Figure 1c. Inductor current of flyback converter 100 kHz, DCM, $L=8 \mu\text{H}$

1c. This is overdesign, and a penalty will be paid in terms of peak switch current.

The RHP zero also constrains crossover for the boost converter.

$$f_c^{\max} \leq \frac{R_L}{6\pi L_e} \quad \text{Boost Converter}$$

The same rules of design apply to the boost converter as for the flyback converter—allow a reasonable amount of ripple current in the inductor to produce a controllable converter. You need to carefully trade off the control loop crossover frequency which can be achieved versus the input filter requirements for filtering the ripple current.

Switching Frequency Crossover Limit

Switching power supplies are sampled-data systems. The Nyquist frequency (which is half the switching frequency) is the absolute limit of information transfer, but you cannot cross over a loop as high as this. A reasonable limit on the crossover frequency is:

$$f_c^{\max} \leq \frac{f_s}{5}$$

As you push the crossover frequency higher, more issues with noise will arise. As converter switching frequencies have risen in past years from 50 kHz to above 500 kHz, very few power supply designers ever push this upper limit due to the noise complications. Recent developments in fast-response supplies for VRMs (voltage regulator modules for CPU supplies) are starting to push the crossover frequencies closer to this limit.

Capacitor ESR Crossover Limit

In many power supplies, the output capacitor determines how effective raising the crossover frequency will be. If your main performance objective is driven by step load requirements, there is no benefit in raising the crossover frequency above the output capacitor ESR frequency.

$$f_c^{\max} \leq \frac{1}{2\pi C R_c}$$

Right Half Plane (RHP) Zeros

RHP zeros are a property of a special class of active circuits which have a tendency to respond initially in the wrong direction when given a changed input. Eventually, the output will move in the direction commanded by the input. How fast it starts moving in the right direction is determined by the frequency location of the RHP zero. Boost and flyback converters are very popular circuits in use in the industry today, and both have RHP zero characteristics when operating in continuous conduction mode.

The cause of the RHP zero for these converters is the same and intuitive. The output capacitor of the converter is charged by the inductor current, but **ONLY** when the power switch is turned off. An initial increase in duty cycle of the power switch increases the current in the inductor within one cycle, but the net charge to the output capacitor (product of the inductor current and off-time) is initially less. Hence, an increase in command to the system results in a temporary droop in the output voltage.

This can be confusing for a controller that is monitoring the output voltage to make control decisions. There is no alternative but to wait and see where the long-term trend is before adjusting the duty cycle.

There is no other solution to RHP zeros. Constant on-time, or constant off-time controllers have no effect on their location, and do not improve the controllability of these converters. Other converters can fare even worse results— the Sepic and Cuk converters have a pair of complex RHP zeros which can be extremely difficult to compensate. While these converters may sometimes have their niche applications, be prepared for a longer development time and potential increased cost to solve these issues.

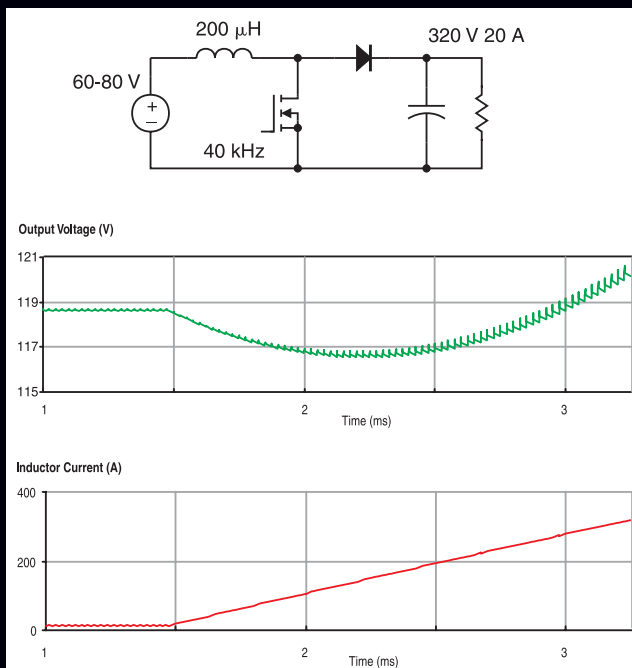


Figure 2. Inductor current and output voltage of boost converter with step duty cycle. Duty cycle step from 0.5 to 0.95

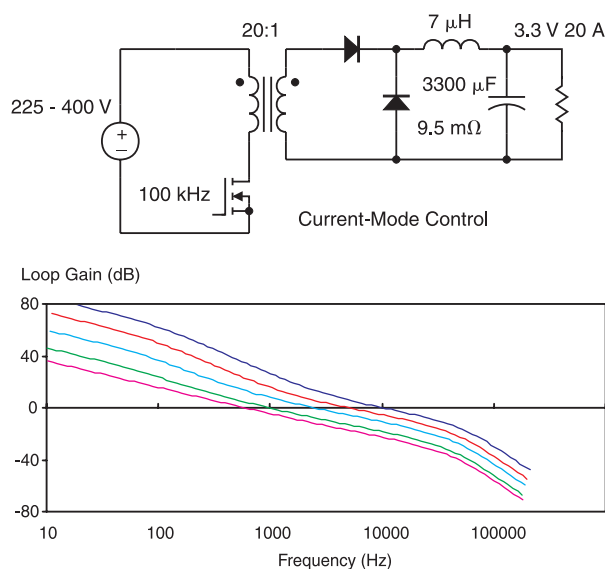


Figure 3a. Loop gains with crossovers from 0.5 to 10 kHz

Figure 3a shows the loop gains for a converter, with crossovers ranging from 500 Hz to 10 kHz. As the crossover frequency is raised from 500 Hz to 5 kHz, the corresponding output overshoot drops significantly from 0.93 V to 0.17 V as shown in Figure 3b. Increasing the loop crossover beyond 5 kHz has no effect on the peak overshoot. At this point, the overshoot is simply determined by the output step current value multiplied by the capacitor ESR value.

Overshoot vs. Crossover Frequency

Crossover Frequency (KHz)	Peak Voltage Transient (V)	Voltage Overshoot (V)
0.5	0.93	4.25
1	0.55	3.87
2	0.30	3.62
3	0.21	3.53
4	0.18	3.50
5	0.17	3.49
7	0.17	3.49
10	0.17	3.49

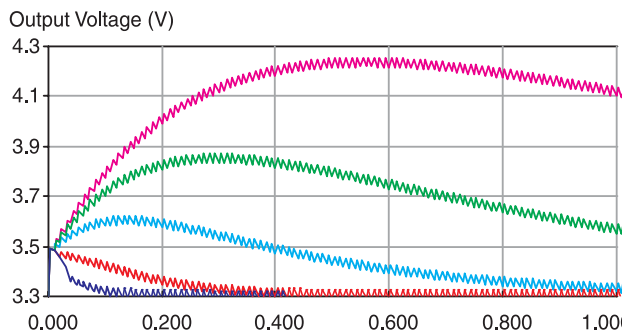


Figure 3b. Transient responses for loop gains above

Error Amplifier Limit

As the loop gain is raised for a given power stage, the error amplifier must provide more gain. Eventually, the limit of the open-loop gain of the amplifier can be reached where no more gain is available. At this point, the feedback transfer function will roll off with one or more poles, depending on the compensation characteristics. The result will be that the desired loop crossover is not achieved, and the phase margin will be less than anticipated.

Figure 4 shows an example converter where this occurs. The desired compensation is shown in Figure 5, together with the open-loop gain bandwidth of the error amplifier, which has a unity gain crossover at 1 MHz, typical for operational amplifiers used in these applications.

The resulting anticipated and actual loop gains and phase are shown in Figure 6.

There are several possible remedies for this. The obvious one of using a higher bandwidth operational amplifier is not always the easiest solution. The increased bandwidth to 10 MHz, for example, can increase the chance of RF oscillations in the circuit. Another solution is to cascade amplifiers, using

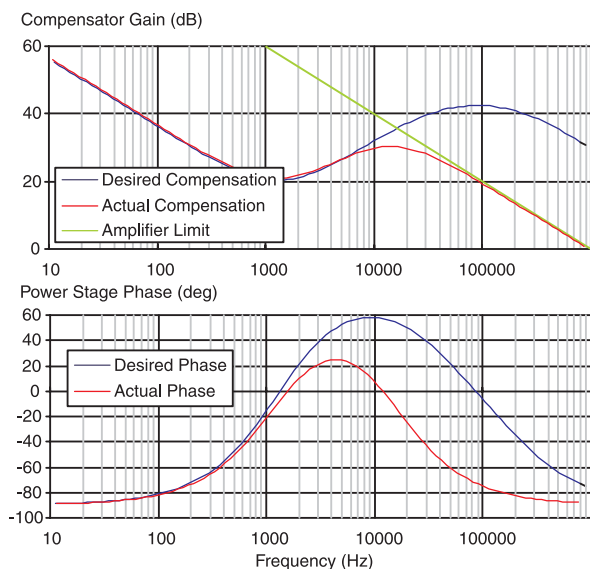


Figure 5. Desired compensation and error amplifier limit

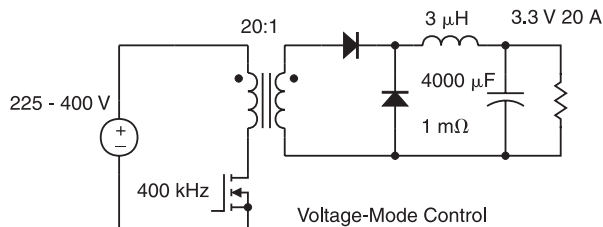


Figure 4. Example converter with desired crossover frequency of 50 kHz.

one for fixed gain (which may be a very wide bandwidth, 20 dB amplifier fixed gain video amplifier, for example) and the other for compensation.

In isolated converters, there is often a spare amplifier available on the primary side of the converter which can be used for this purpose. The secondary side amplifier with the reference, is used for compensation, and the primary side amplifier for increasing the gain. Isolated feedback, however, comes with its own set of problems—the isolation device, be it an optocoupler or transformer isolation of some kind, suffers from limited bandwidth and significant phase delay at high crossover frequencies.

Very High Frequency Converters

Some converters require high packaging density, and the switching frequency is often raised to 1 MHz or higher to achieve this. Very few of these converters take full advantage of the potential control loop opportunities that are available. In fact, many of them deliberately compromise the loop design to allow for variable loads that the user may impose on them.

However, it is possible to aggressively push the crossover of these types of converters to get the full benefit of the increased switching frequency. In the mid-1980s, extensive research was done at the Virginia Power Electronics Center

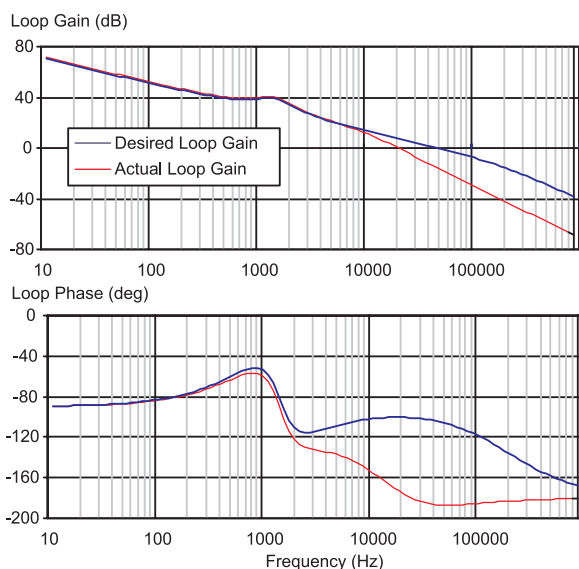


Figure 6. Desired and actual loop gains with 1 MHz amplifier

into converters operating at high switching frequencies. The highest loop gain crossover achieved was at 180 kHz, for a 7 MHz zero-voltage-switched quasi-resonant converter. Noise issues, and error amplifier limits prevented higher crossover frequencies.

The converter power stage schematic is shown in *Figure 7*. Operation of the power stage is given in [1]. The control loop used a modified current-mode control scheme to avoid some of the noise problems of current mode control described in the last issue of *Switching Power Magazine*, and further details of the control circuit are given in [2]. Even using high performance parts, the theoretical phase margin was not achieved with this converter when trying to cross over the loop at 180 kHz. The phase curves should have remained constant for both loop gains, but limits in the feedback components caused additional phase rolloff as shown.

In Summary

Loop gain design is still of great importance to switching power supply designers. Much can be done with the control loop to optimize performance, and reduce costs in the power stage components. Looking at the loop early in the design and checking the potential performance of your power stage with some of the guidelines in this article can help avoid subsequent costly redesign.

There are other issues impacting loop design which are not covered in this paper. The most significant of these are input and output circuitry attached to the converter, usually in the form of input filters and active or reactive loads. These can substantially change the stability of a converter design, and future articles will address these topics.

References

- [1] "Zero-Voltage Switched Buck and Flyback Converters – Experimental Results at 10 MHz", W.A. Tabisz, P. Gradzki and F.C. Lee, *IEEE Transactions on Power Electronics*, Vol. 4, No. 2, April 1989.
- [2] "Multi-Loop Control for Quasi-Resonant Converters", R.B. Ridley, W.A. Tabisz, F.C. Lee and V. Vorperian, *IEEE Transactions on Power Electronics*, Vol. 6 No. 1, January 1991.

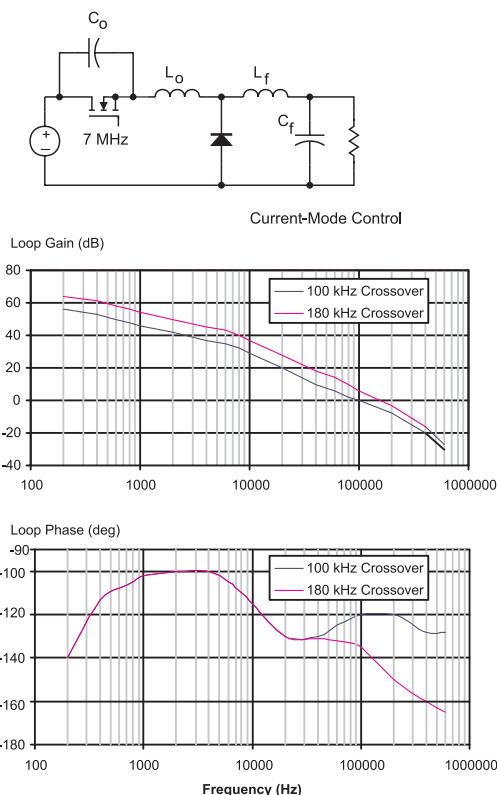


Figure 7: 7 MHz resonant converter loop gains

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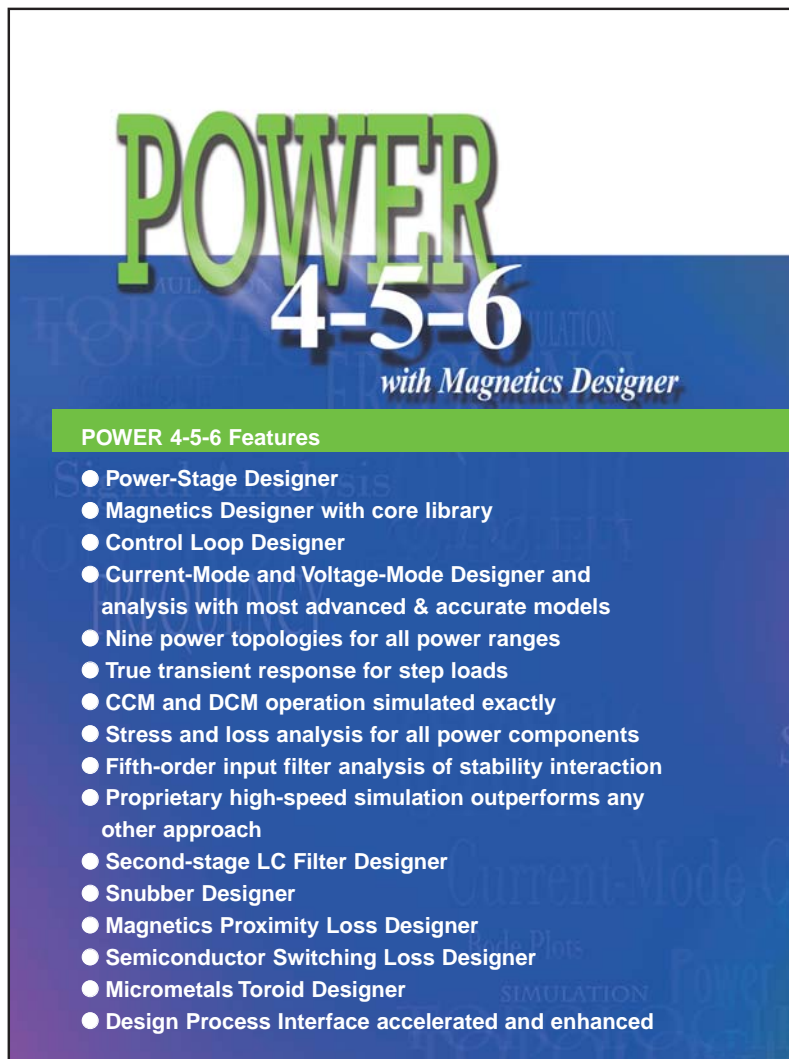
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