

Feedback Loop Design of an LLC Resonant Power Converter

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ABSTRACT

This application note describes an approach to design feedback loop compensation for an LLC resonant half-bridge power converter. The approach described here is based on the measured Bode plots of the modulator generated by a network analyzer. As we know, as long as the modulator Bode plots are obtained, the frequency domain poles and zeros in the feedback loop compensation can be analytically determined, then fine-tuned with a bench test. This measurement is necessary as part of feedback loop design because a practical small-signal model is not available for LLC resonant converters. This document uses the Texas Instruments' [UCC25600](#) as the frequency controller. A detailed description of the UCC25600 and its associated example design fixture, the [UCC25600EVM-341](#), can be found in the product data sheet and the evaluation module user's guide, respectively; both additional documents are available for download at www.ti.com.

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1 Control Loop Description of an LLC Converter

NOTE: For complete details on the design of an LLC resonant power converter, the recommended reference is SEM1900 Topic 3, *Designing an LLC Resonant Half-Bridge Power Converter*. This application note will not repeat the discussion of these details.

Figure 1 shows a typical block diagram of an LLC resonant converter.

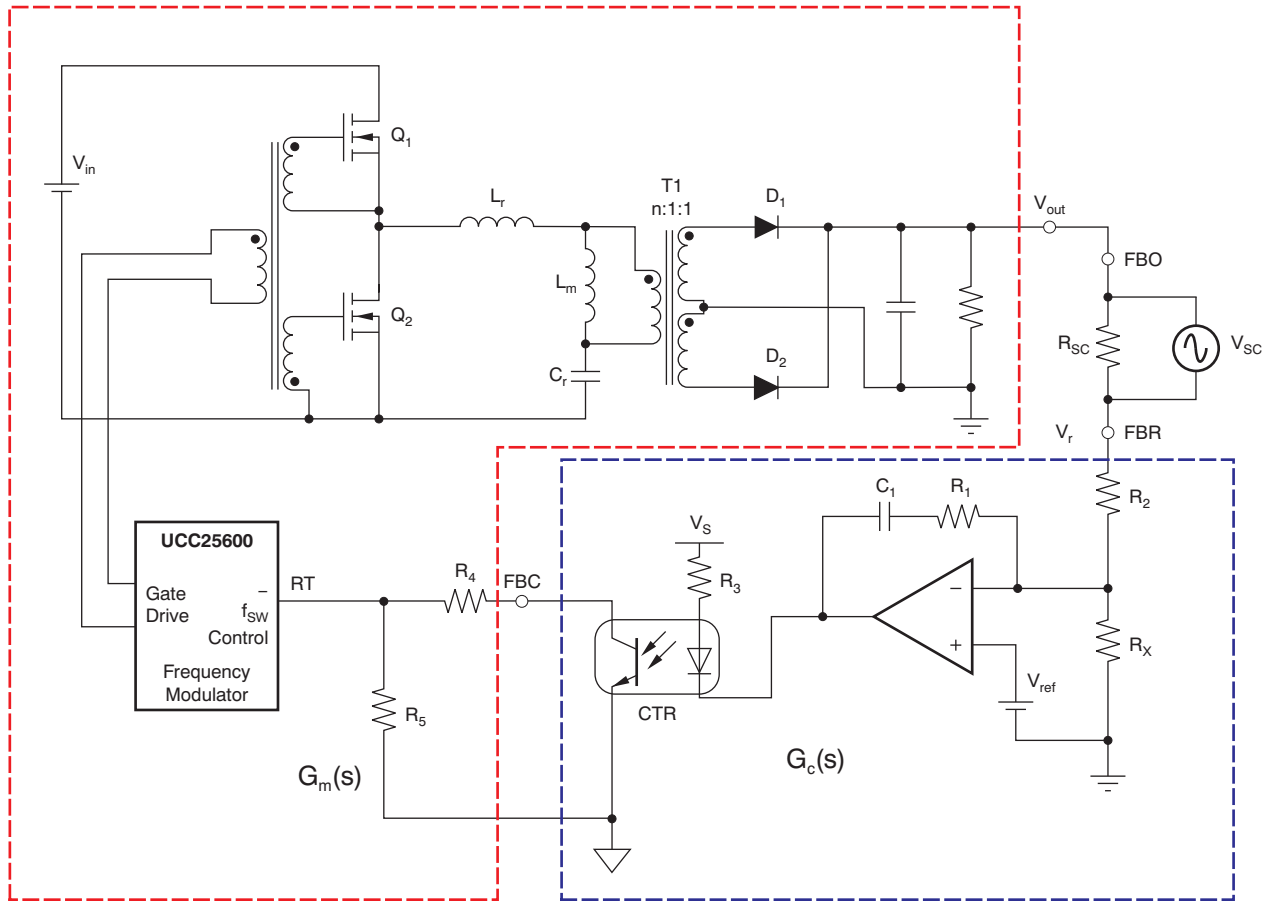


Figure 1. Block Diagram of a Typical LLC Resonant Half-Bridge Converter

This diagram consists of two blocks: the modulator and the compensator, expressed in the respective transfer functions of $G_m(s)$ and $G_c(s)$, or $G_m(j\omega)$ and $G_c(j\omega)$. In Figure 1, the red outline defines the $G_m(s)$ transfer function, and the blue outline defines the $G_c(s)$ transfer function.

$G_m(\omega)$ and $G_c(\omega)$ can be used to express $G_m(j\omega)$ and $G_c(j\omega)$ respectively in shorthand, where $j = \sqrt{-1}$ can be omitted without confusion. Then the loop gain transfer function $G_{lp}(s)$, or $G_{lp}(\omega)$, is expressed as Equation 1.

$$G_{lp}(s) = \frac{V_{out}(s)}{V_r(s)} = G_c(s) \cdot G_m(s) \tag{1}$$

or as Equation 2:

$$G_{lp}(\omega) = \frac{V_{out}(\omega)}{V_r(\omega)} = G_c(\omega) \cdot G_m(\omega) \tag{2}$$

Both $G_m(\omega)$ and $G_c(\omega)$ can be measured with a frequency sweeping signal applied on resistor R_{SC} , as these equations show:

$$G_m(\omega) = \frac{FBO}{FBC} \tag{3}$$

$$G_c(\omega) = \frac{FBC}{FBR} \tag{4}$$

The loop gain transfer function $G_{lp}(\omega)$ is measurable and obtained as Equation 5.

$$G_{lp}(\omega) = \frac{FBO}{FBR} \tag{5}$$

It is obvious that FBC can be assigned at different locations in the converter circuit. For example, it can be moved to the optocoupler input as Figure 2 illustrates. Here, the red area and the blue area represent the $G_m(\omega)$ and $G_c(\omega)$ functions, respectively, as they do in Figure 1.

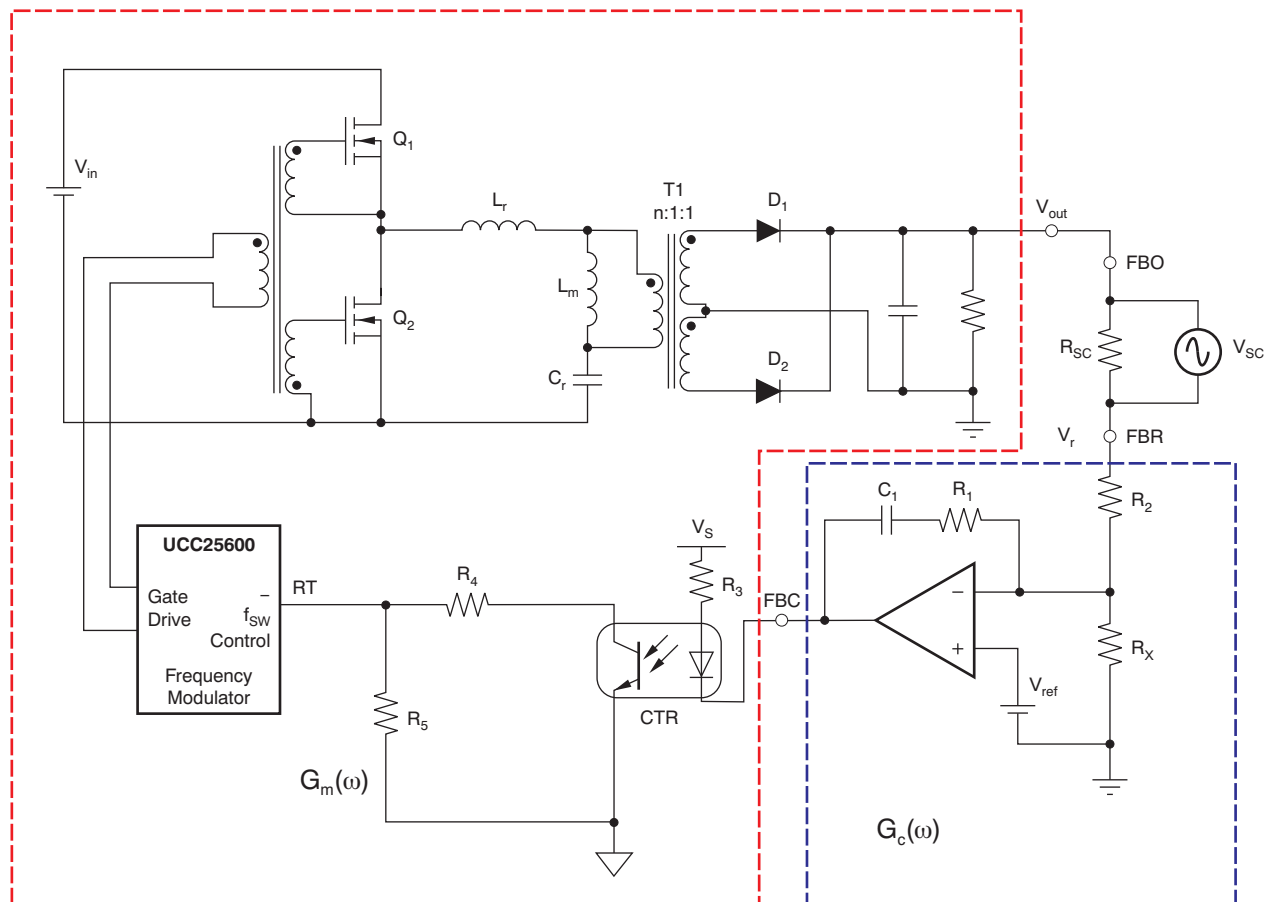


Figure 2. Another Way to Define $G_m(\omega)$ and $G_c(\omega)$

In this position, then, the measured $G_m(\omega)$ and $G_c(\omega)$ are different from those measured in Figure 1.

Note that it is possible to define $G_m(\omega)$ and $G_c(\omega)$ in several different ways. But $G_{lp}(\omega)$ will be the same, regardless of how $G_m(\omega)$ and $G_c(\omega)$ are defined. In this document, we use the definition as shown in Figure 1 to avoid confusion.

2 Loop Compensation Design Approach

Unlike $G_m(\omega)$, $G_c(\omega)$ can be expressed analytically. As shown in [Figure 1](#), if we use Type I compensation, then $G_c(\omega)$ is expressed as [Equation 6](#):

$$G_c(s) = \frac{\frac{s}{1} + 1}{R1 \cdot C1} = \frac{s}{\omega_1} \cdot \left(\frac{s}{\omega_{p_opto}} + 1 \right) \quad (6)$$

Where ω_{p_opto} is the angular frequency of the optocoupler frequency-domain pole. This value can be typically considered to be approximately $\omega_{p_opto} = 2\pi \times 10$ kHz, although the value varies with the particular optocoupler in the circuit as well as with the bias point. CTR is the optocoupler current transfer ratio; the angular frequency ω_1 is the gain of the frequency domain pole at origin; and therefore, we arrive at:

$$\omega_1 = \frac{R_4 \cdot CTR}{R_3 \cdot R_2 \cdot C_1} \quad (7)$$

The angular frequency ω_1 is the 0-dB crossover of $G_c(\omega)$ when we set $R_1 = 0$ and ignore ω_{p_opto} at a low-frequency range. In fact:

$$G_c(s) = \frac{\frac{j\omega}{1} + 1}{R1 \cdot C1} \quad \text{with: } R1 = 0 \quad \omega \ll \omega_{p_opto} \quad \rightarrow \quad G_c(\omega) = \frac{1}{\frac{j\omega}{\omega_1}} = \frac{\omega_1}{j\omega} \quad (8)$$

If we let $|G_c(\omega)| = 1$, then:

$$|G_c(\omega)| = \left| \frac{\omega_1}{j\omega} \right| = 1 \rightarrow \omega = \omega_1 \quad (9)$$

To compensate the LLC converter feedback loop, $G_m(\omega)$ must first be obtained by measurement. To make $G_m(\omega)$ measurable, the feedback loop must be stable, which is our design goal yet to be achieved. So we are now in a circle that we must break in order to achieve our design goal. To break this circle, one common technique is to use a large-value capacitor for C_1 in [Figure 1](#) to push the loop bandwidth (that is, the gain crossover frequency) low enough in the expectation that the loop can be *stable* enough for initial measurement. This idea is workable and allows us to obtain initial $G_m(\omega)$ value. Sometimes, though, we may be confused about how large a capacitor is adequate for a specific application in order to make a stable $G_m(\omega)$ measurement. If an estimated C_1 value is not sufficient, we have a potential risk that the resulting loop may not be stable to measure $G_m(\omega)$, and potential circuit damage can occur.

To avoid these issues, we propose a more reliable approach that combines with the practice of using an initial low bandwidth measurement from a large value C_1 . As we know, to maintain output voltage regulation in an LLC resonant converter, the input voltage must be great enough to achieve the desired output voltage. If the input voltage is not sufficient, the output voltage regulation cannot be achieved because the maximum gain has already reached its limit. However, even if output regulation is not achieved, the converter is stable, and this stability is not related to the feedback loop parameters. During such an operation, bench tests show a $G_m(\omega)$ can continue to be measured in the usual way. Although the measured $G_m(\omega)$ is not exactly the same as that obtained from an operation when the output comes into regulation, it is sufficient to give us an idea of how large the value of C_1 must be and how ω_1 should be designed. Then, we will be able to complete the remaining design and avoid the risks noted earlier.

3 Methodology

To implement this approach, we first must make sure that all other segments of the converter work properly, especially the associated power stage. This operation is usually called an *open-loop test*, which must be passed to assure full functionality. Then we can close the feedback loop and measure $G_c(\omega)$ in this way:

- We first increase the input voltage slowly from zero and observe the output voltage. Because the input voltage is low, the controller in the feedback loop will generate the maximum gain to raise the output voltage as much to the point of regulation as it can.
- We then continue to increase the input voltage until the output voltage is close to the regulation (within 10%). At that point, we stop increasing the input voltage. The converter should now be stable and quite independent of the feedback loop compensation parameters.
- Then, we apply a frequency sweeping signal to R_{SC} and measure the Bode plots between FBO and FBC. This measurement gives us the initial $G_m(\omega)$ value.

We can make several more measurements by further increasing the input voltage in 1% increments if $G_m(\omega)$ does not show good measurement results. However, do **not** risk increasing the voltage to bring the output voltage into regulation yet.

Here, we offer an example to show how to make this type of initial $G_m(\omega)$ measurement, then how to design the subsequent feedback loop once the measurement is complete. We use the [UCC25600EVM-341](#) as our example with the corresponding circuit diagram shown in [Figure 1](#). Complete schematics for this device are found in the [UCC25600EVM-341 User Guide](#).

This converter has these electrical specifications:

- Input voltage: 375 V_{DC} to 405 V_{DC}
- Output power (rated): 300 W
- Output voltage: 12 V_{DC}
- Output current (rated): 25 A
- Output voltage line regulation (with $I_O = 1.0$ A): $\leq 1\%$
- Output voltage load regulation (at $V_{IN} = 390$ V): $\leq 1\%$
- Output voltage peak-to-peak ripple (at $V_{IN} = 390$ V and $I_O = 25$ A): ≤ 120 mV
- Efficiency (at $V_{IN} = 390$ V and $I_O = 25$ A): $\geq 90\%$
- Switching frequency: 70 kHz to 150 kHz in normal operation
- Converter topology: LLC resonant half-bridge converter

3.1 Procedure and Results

Step 1. **$G_m(\omega)$ measurement; C_1 initial value.**

In the expression:

$$\omega_1 = \frac{R_4 \cdot \text{CTR}}{R_3 \cdot R_2 \cdot C_1}$$

(10)

C_1 is only the parameter we must determine initially for a stable $G_m(\omega)$ measurement. The remaining parameters are determined by other design factors and generally have a limited adjustable range. For example, CTR is fixed after an optocoupler is selected. Additionally, R_3 and R_4 are determined by the requirements of the frequency range and the optocoupler configuration.

The value of C_1 is not critical from our proposed approach. For example, an initial C_1 value can be between 0.1 μF and 1 μF ; this range is given only to show some commonly-used values. If C_1 is outside these values, the process should work. In this example, we use $C_1 = 1.0$ μF (note that it does not make much difference to use $C_1 = 0.1$ μF).

Step 2. **$G_m(\omega)$ measurement with output voltage not in regulation.**

As shown in Step 1, the output voltage regulation is 12 V with minimum input voltage 375 V. If we use a voltage much lower than 375 V, the output voltage is not able to enter regulation. A good rule of thumb is to use an input voltage low enough to keep the output voltage out of regulation, but close to

the regulation voltage; say, within 10%. This technique generally should give an acceptable measurement. If it does not, a value closer to regulation may be tested, using a 1% increment each time until a good measurement is obtained. In this example, when we slowly increase V_{IN} to 285 V with $I_O = 1$ A, the output voltage showed to be approximately 11 V, which is 8.3% below 12 V. Then we measured $G_m(\omega)$ with an acceptable result. A 25-mV frequency sweeping signal was used during this test. The $G_m(\omega)$ measurement is shown in [Figure 3](#).

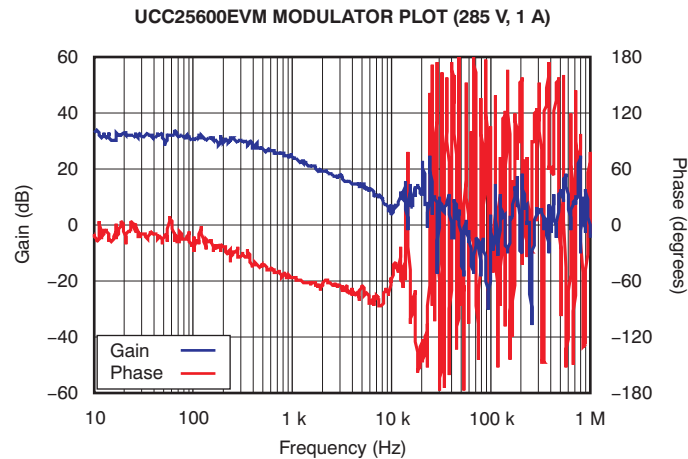


Figure 3. Initial $G_m(\omega)$ Measurement

Step 3. Determine initial $G_c(\omega)$ based on $G_m(\omega)$ measurement.

Now we need to design a proper $G_c(\omega)$ to allow the converter output to enter into regulation. The minimum switching frequency of the converter is specified at 70 kHz. As a rule of thumb, the gain crossover frequency should be below one-fifth of its minimum switching frequency. If we use one-tenth of its minimum switching frequency, that allows approximately 7 kHz to cover the entire operating range. Keep in mind, though, that we are only at the beginning stages and our primary purpose is to establish an initial operating point. The $G_m(\omega)$ obtained so far is also a very preliminary value. As such, we may want to make the target crossover frequency very conservative. Based on the measurement just obtained, it appears acceptable to set the crossover frequency somewhere around 100 Hz. One advantage to selecting 100 Hz is that it has a flat phase angle close to 0°; this response helps to achieve the desired stability as a 90° phase margin can be expected at 100 Hz. Notice, again, this target is only an initial design; the final desired value can be adjusted further.

At 100 Hz, $|G_m(\omega)| = 28$ dB. Therefore, we would need to design $|G_c(\omega)|$ to have:

$$20\log(|G_c(\omega)|)_{\omega = 2\pi \times 100\text{Hz}} = -28 \text{ dB}$$

Because the crossover frequency of 100 Hz is a rough number to be used initially, we can simply do a quick design with only ω_1 while leaving out the zero of R_1-C_1 by making $R_1 = 0$, such that the zero of R_1-C_1 is pushed beyond its effect to the 100-Hz crossover.

$$\omega_1 = 10^{\frac{-28}{20}} \cdot 2\pi \cdot 100(\text{Hz}) = 25.0 \text{ rad/s} \tag{11}$$

$$\omega_1 = \frac{R_4 \cdot \text{CTR}}{R_3 \cdot R_2 \cdot C_1} \tag{12}$$

If $C_1 = 0.22 \mu\text{F}$, we can obtain a $\omega_1 = 25.0$ rad/s, or $f_1 = 3.98$ Hz, with a given $\text{CTR} = 120\%$, $R_4 = 510 \Omega$, $R_2 = 110 \text{ k}\Omega$, and $R_3 = 1.00 \text{ k}\Omega$.

Step 4. Re-measurement of $G_m(\omega)$.

Now we are confident enough to increase the input voltage so the output is in regulation, and then make a stable measurement of $G_m(\omega)$. The measured $G_m(\omega)$ is shown in [Figure 4](#), at 390-V input.

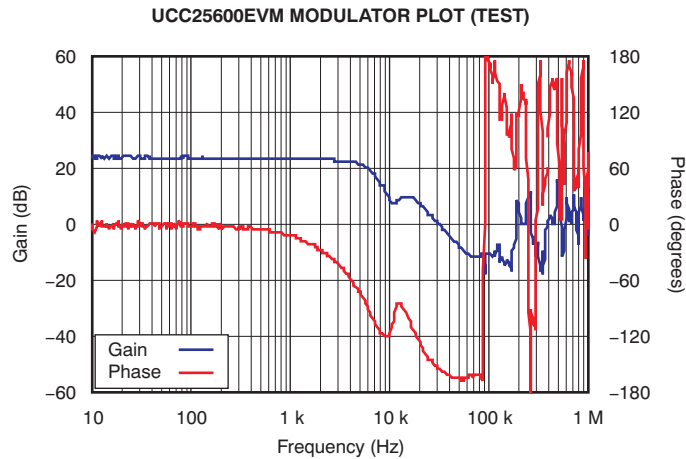


Figure 4. Re-Measurement of $G_m(\omega)$

Certainly, one can make additional measurements between 280 V and 390 V, say in 20-V increments, to gain confidence in what has already been obtained. An important note for test manipulation here is that the input voltage increase should progress slowly until the feedback loop compensation is complete. A slow increase of the input voltage can allow time for the designer to decide to stop if there is any sign that the feedback loop may become unstable.

Step 5. Design $G_c(\omega)$ based on measured $G_m(\omega)$.

To achieve the crossover frequency of 7 kHz with a minimum 45° phase margin, $G_c(\omega)$ would need to be designed to achieve the Bode plot responses shown in [Figure 5](#). This result can be easily accomplished by arranging the pole and the zero shown in [Equation 6](#). Below is one possible set of parameters to achieve the illustrated Bode plots:

- $R_1 = 5.1 \text{ k}\Omega$
- $R_2 = 19.7 \text{ k}\Omega$
- $R_3 = 1.0 \text{ k}\Omega$
- $R_4 = 510 \text{ }\Omega$
- $C_1 = 0.22 \text{ }\mu\text{F}$
- $\text{CTR} = 120\%$

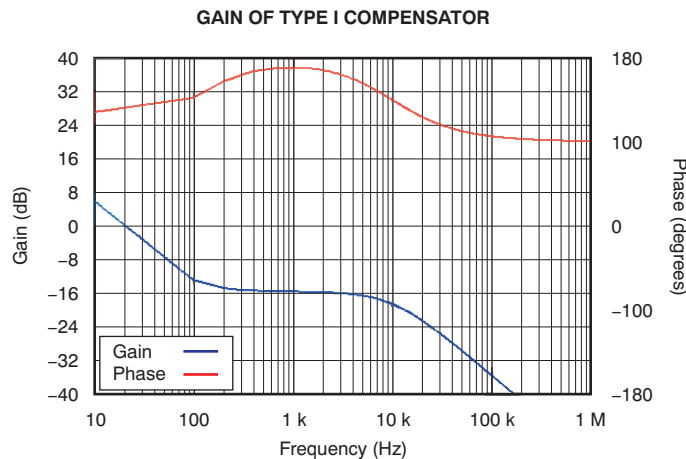


Figure 5. Design of $G_c(\omega)$

Step 6. Bench testing and fine tuning.

The last step is to plug in the obtained parameter values to perform a bench test. Usually, fine tuning is necessary to cover all operating conditions and to adapt to the parameters not modeled in the $G_c(\omega)$ for example parasitic variables.

The compensation values are finalized as $R_1=17.8\text{ k}\Omega$, $R_2 = 19.7\text{ k}\Omega$, $R_3 = 1.0\text{ k}\Omega$, $R_4 = 0.51\text{ k}\Omega$, and $C_1 = 47\text{ nF}$. Figure 6 shows the control loop Bode plots at $V_{IN} = 390\text{ V}$, $V_o = 12\text{ V}$, and a 25-A load.

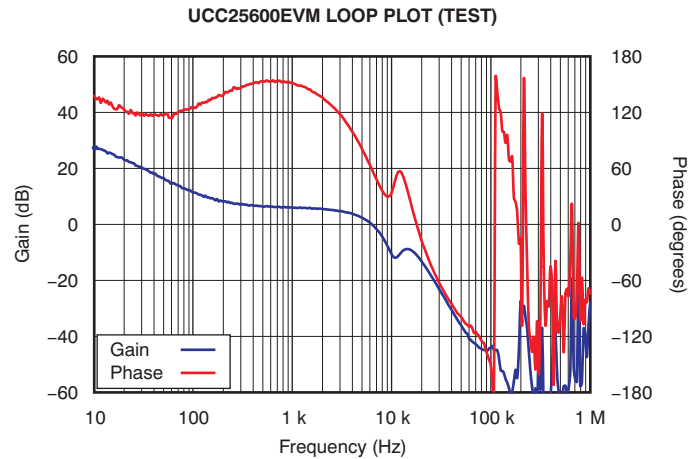


Figure 6. Feedback Loop Bode Plots After Final Values Set

4 References

Unless otherwise noted, these documents are available for download from the TI website (www.ti.com).

1. UCC25600 Product data sheet. Texas Instruments literature number [SLUS846](#).
2. UCC25600EVM User guide. Texas Instruments literature number [SLUU361](#).
3. TI Power Supply Design Seminar SEM1900 Topic 3, *Designing an LLC Resonant Half-Bridge Power Converter*. (2010-11). Texas Instruments literature number SLUP252.

Revision History

Changes from Original (October, 2010) to A Revision	Page
• Corrected Figure 1	2
• Updated Figure 2	3
• Replaced Figure 6	8

NOTE: Page numbers for previous revisions may differ from page numbers in the current version.

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