Analog and Mixed-Signal Products

Analog Applications Journal

First Quarter, 2003

IMPORTANT NOTICE

Mailing Address:

Texas Instruments
Post Office Box 655303
Dallas, Texas 75265

Contents

To view past issues of the
***Analog Applications Journal*, visit the Web site**
www.ti.com/sc/analogapps

Introduction

Analog Applications Journal is a collection of analog application articles designed to give readers a basic understanding of TI products and to provide simple but practical examples for typical applications. Written not only for design engineers but also for engineering managers, technicians, system designers and marketing and sales personnel, the book emphasizes general application concepts over lengthy mathematical analyses.

These applications are not intended as “how-to” instructions for specific circuits but as examples of how devices could be used to solve specific design requirements. Readers will find tutorial information as well as practical engineering solutions on components from the following categories:

- Power Management
- Amplifiers: Op Amps

Where applicable, readers will also find software routines and program structures. Finally, *Analog Applications Journal* includes helpful hints and rules of thumb to guide readers in preparing for their design.

Load-sharing techniques: Paralleling power modules with overcurrent protection

By Lisa Dinwoodie (Email: lisa_dinwoodie@ti.com)
Power Applications Specialist

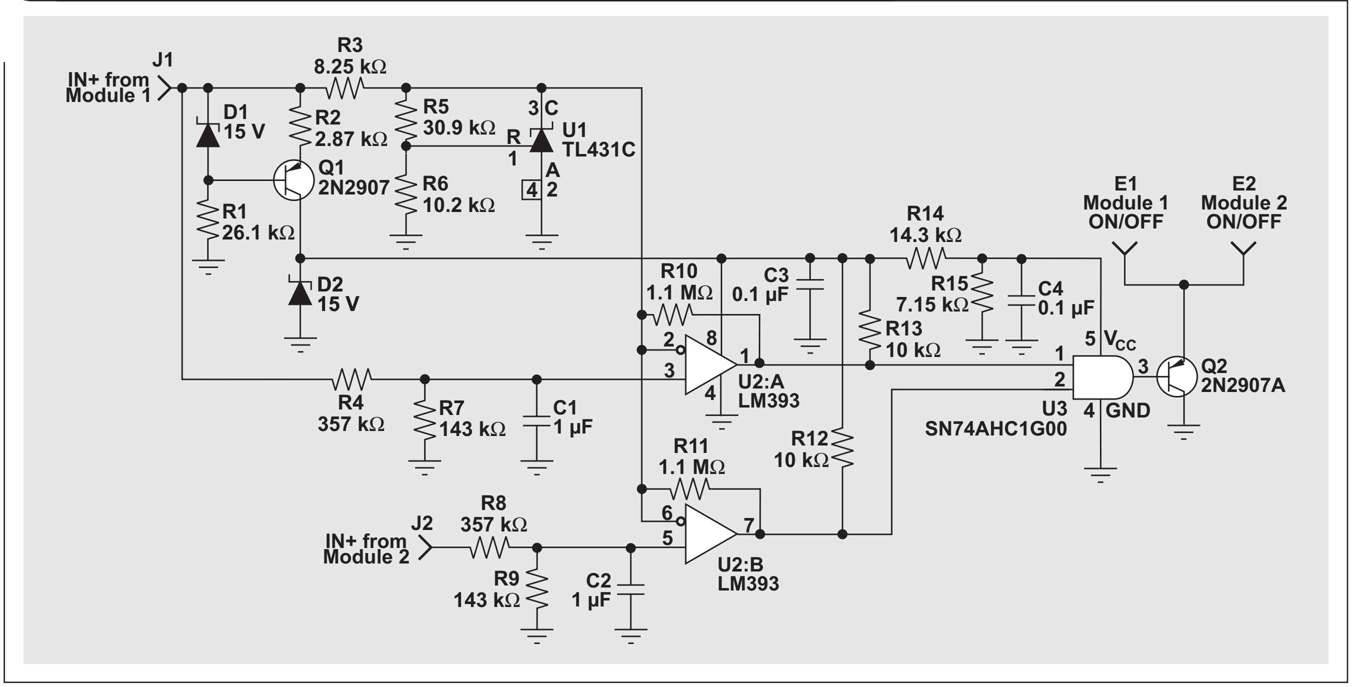
Paralleling low-current, low-voltage power modules for high-current, low-voltage applications has many benefits. Among them are: redundancy for enhanced reliability, hot-swap capability, distributed heat removal, and design flexibility. Paralleling power stages requires load sharing in order to equalize the stresses among the modules. One method of load sharing, based upon the automatic master/slave architecture, is to use a dedicated controller, such as the UCC39002, to provide for equal current distribution of the load current among the parallel-connected power supplies. The power modules must be equipped with true remote-sense capability or an output-adjustment terminal. The output current of each module is measured and compared to a common load-share bus. The positive sense voltage or the voltage of the output voltage adjust pin of each module is adjusted to provide equal current sharing.

Several modules are paralleled so that the entire assembly can support a full load much greater than an individual module would be capable of supplying. Due to manufacturing tolerances and component variations, startup delay times typically vary slightly from module to module. When the modules to be paralleled have an overcurrent protection circuit featuring constant current limit with automatic recovery, starting up fully enabled into the full system load does not pose a problem. Inevitably, one module will have a faster turn-on than the others. The eager module will carry as much of the load as it can, sometimes up to 140% of its individual current capacity, before its output voltage falters. Meanwhile, the next module will come up and contribute to the load. After a brief transition time, all of the modules will be up, the master will be recognized, and accurate load sharing will take place.

When the modules to be paralleled have an overcurrent protection circuit featuring a hiccup mode, starting up fully enabled into full system load, regardless of the load sharing technique used, does pose a problem. The module with the fastest turn-on profile will come up into an overcurrent condition. Immediately, in an act of self-preservation, it will go into hiccup mode, alternately sinking and sourcing current. The next module to come up into the load will also fall into this hiccup mode, sinking current when the other module sources it. Because the load-share circuitry essentially adds a voltage loop to the output of each module, this hiccupping overcurrent protection mode will prevent loop closure. Simultaneously enabling the modules will prevent this hiccup mode from starting, and load sharing can be successfully achieved.

Figure 1 shows a simple comparator circuit that will simultaneously enable two modules and can be expanded to accommodate more if needed. It assumes that the only

Figure 1. Logic circuit to turn on two power modules simultaneously



bias available for the logic gates is the 48-Vdc bus used as the input to the modules themselves. The circuit is designed to short the modules' on/off pins to ground simultaneously when their inputs reach 35 V, assuming that the modules' input range is between 36 V and 75 V and that they have a turn-on threshold of approximately 33 V.

The PNP transistor, Q1, its emitter and base resistors, and the two 15-V Zener diodes provide a 15-V, 5-mA bias to the comparators and the NAND gate from the input line, which could vary from 36 Vdc to 75 Vdc. The TL431 is set up to provide a regulated 10-V reference voltage for the inverted comparator inputs. The non-inverted comparator input signals are derived from resistively dividing the input voltages of the modules. Because the bias and comparator signals are from the same source, capacitors are needed to delay the comparator input signals long enough so that the LM393, U2, is operational and "smart." This circuit is added to the load-share board and successfully turns on the modules simultaneously into a full system load without triggering the overcurrent hiccup mode of the modules.

Related Web sites

analog.ti.com
www.ti.com/sc/device/*partnumber*
Replace *partnumber* with LM393, SN74AHC1G00, TL431 or UCC39002

Using the TPS61042 white-light LED driver as a boost converter

By Jeff Falin (Email: j-falin1@ti.com)

Applications, Portable Power

Introduction

Although designed to be a white-light LED driver, the TPS61042 can be configured as a discontinuous, hysteretically controlled boost converter with a 500-mA peak switch current. For example, Figure 1 shows the TPS61042 configured to provide V_{OUT} = 16.2 V and I_{OUT} = 30 mA, from V_{IN} down to 2.5 V. The LED driver circuitry is either left unconnected (pins 1 and 2) or grounded (pin 7); and pin 5, CTRL, is used as enable.

Operation

As a boost converter, the TPS61042 operates with an input voltage range of 1.8 to 6 V and can generate output voltages of up to 28 V. The device operates in a pulse-frequency modulation (PFM) scheme with constant peak-current control. This control scheme maintains high efficiency over the entire load-current range. With a switching frequency of up to 1 MHz, the device enables the use of very small external components.

The converter monitors the output voltage. When the feedback voltage falls below the reference voltage (typically 0.25 V), the internal switch turns on and the current ramps up. The switch turns off when the inductor current reaches the internally set peak current of 500 mA (typ). Refer to the following paragraph, entitled "Peak-current control," for more information. The second criterion that turns off the switch is the maximum on-time of 6 µs (typ). This limits the maximum on-time of the converter in extreme conditions. As the switch turns off, the external Schottky diode is forward-biased, delivering the current to the output. The switch remains off for a minimum of 400 ns (typ), or until the feedback voltage drops below the reference voltage. Using this peak-current control scheme, the converter operates in discontinuous conduction mode (DCM) where the switching frequency depends on the output current. This results in very high efficiency over the entire load-current range. Inherently stable, this regulation scheme allows a wide selection range for the inductor and output capacitor.

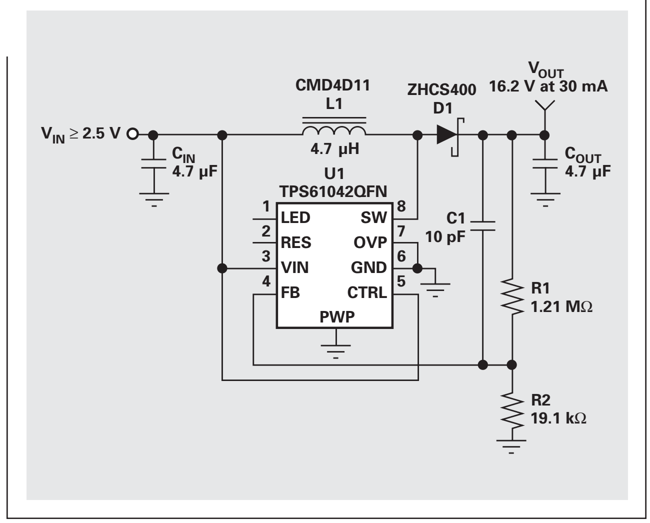
Peak-current control

The internal switch turns on until the inductor current reaches the typical dc current limit (I_{LIM}) of 500 mA. Due to the 100-ns (typ) internal propagation delay, the actual current exceeds the dc current-limit threshold by a small amount. The typical peak-current limit can be calculated by

$$I_P(\text{typ}) = I_{LIM} + \frac{V_{IN}}{L} \times 100\text{ ns} \quad \text{and}$$

$$I_P(\text{typ}) = 500\text{ mA} + \frac{V_{IN}}{L} \times 100\text{ ns}.$$

Figure 1. TPS61042 in a boost configuration



Soft start

All inductive step-up converters exhibit high in-rush current during startup if no special precaution is taken. This can cause voltage drops at the input rail during startup and may result in unwanted or early system shutdown. The TPS61042 limits this in-rush current by increasing the current limit in two steps, starting from $I_{LIM}/4$ for 256 cycles, then up to $I_{LIM}/2$ for the next 256 cycles, and ending with the full current limit.

Inductor selection, maximum load current

Since the PFM peak-current control scheme is inherently stable, the inductor value does not affect regulator stability. The selection of the inductor, together with the nominal load current and the application's input and output voltage, determines the converter's switching frequency. Depending on the application, inductor values between 2.2 to 47 µH are recommended. The maximum inductor value, L_{MAX}, is determined by the maximum on-time of the switch, 6 µs (typ). The peak-current limit must be reached within this 6-µs period for proper operation. L_{MAX} is calculated as

$$L_{MAX} = \frac{V_{IN}(\min) \times 6\,\mu s}{I_P}.$$

DEVICE	CAPACITOR	VOLTAGE RATING (V)	COMPONENT SUPPLIER	COMMENTS
TPS61042	4.7 µF/X5R/0805	6.3	Tayo Yuden JMK212BY475MG	C_{IN}/C_{OUT}
	10 µF/X5R/0805	6.3	Tayo Yuden JMK212BJ106MG	C_{IN}/C_{OUT}
	1.0 µF/X7R/1206	25	Tayo Yuden TMK316BJ105KL	C_{OUT}
	1.0 µF/X5R/1206	35	Tayo Yuden GMK316BJ105KL	C_{OUT}
	4.7 µF/X5R/1210	25	Tayo Yuden TMK325BJ475MG	C_{OUT}

The minimum inductor value, L_{MIN}, is a function of the output voltage, load current, and switching frequency and is calculated as

$$L_{MIN} = \frac{2 \times I_{LOAD} \times [V_{OUT} - V_{IN}(\min) + V_D]}{I_P^2 \times f_{SMAX}},$$

where I_P is the peak current as previously described under "Peak-current control," I_{LOAD} is the maximum load current, V_D is the maximum rectifier diode forward voltage (0.3 V typ), and f_{SMAX} is the maximum switching frequency (1 MHz).

A smaller inductor value gives a higher converter switching frequency but lowers the efficiency.

The best way to calculate the maximum available load current under certain operating conditions is to estimate the expected converter efficiency at the maximum load current. The maximum load current can be estimated by

$$L_{LOAD}(\max) = \eta \frac{V_{IN}(\min) \times I_P}{2 \times V_{OUT}},$$

where η is the expected converter efficiency (typically 85%).

Output capacitor selection

For the best output-voltage filtering, a low-ESR output capacitor is recommended. Ceramic capacitors have a low ESR value; but tantalum capacitors can also be used, depending on the application.

Assuming that the converter does not show double pulses or pulse bursts on the switch node (SW), the output voltage ripple can be calculated as

$$\Delta V_{OUT} = \frac{I_{OUT}}{C_{OUT}} \times \left(\frac{1}{f_S \times I_{OUT}} - \frac{I_P \times L}{V_{OUT} + V_D - V_{IN}} \right) + I_P \times ESR,$$

where I_P is the peak current as previously described under "Peak-current control," L is the selected inductor value, I_{OUT} is the nominal load current, f_S (I_{OUT}) is the switching frequency at the nominal load current as previously calculated, V_D is the rectifier diode forward voltage (0.3 V typ), C_{OUT} is the selected output capacitor, and ESR is the output capacitor ESR value.

Refer to Table 1 for recommended output capacitors.

Input capacitor selection

For good input-voltage filtering, low-ESR ceramic capacitors are recommended. A 4.7-µF ceramic input capacitor is sufficient for most applications. Increasing this value provides better input-voltage filtering. Refer to Table 1 for recommended input capacitors.

Efficiency

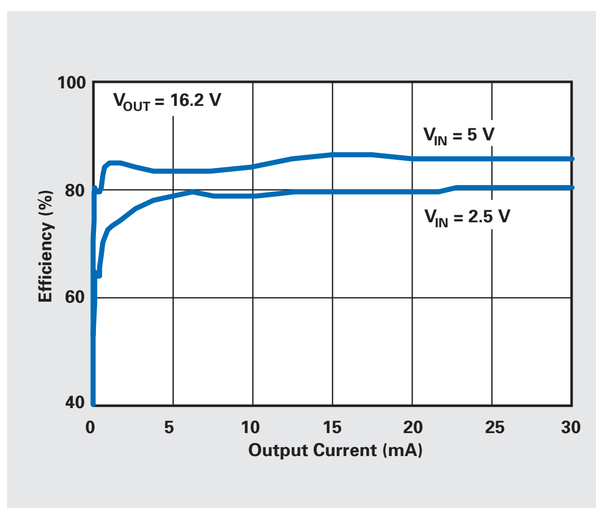
As shown in Figure 2, the TPS61042's efficiency ranges from about 70% to 86% in a boost configuration.

The inductor and diode in Figure 1 were selected to minimize the overall area. A larger inductor and/or diode can improve efficiency.

Related Web sites

analog.ti.com
www.ti.com/sc/device/TPS61042

Figure 2. TPS61042 boost-converter efficiency



RF and IF amplifiers with op amps

By Bruce Carter (Email: r-carter5@ti.com)
Advanced Linear Products, Op Amp Applications

Introduction

Why use op amps for RF design? Traditional RF design techniques using discrete transistors have been practiced successfully for decades. RF designers who are comfortable with things “as is” will scrutinize introduction of a new design technique using op amps. For high-performance RF equipment, however, high-speed op amps have some distinct advantages:

- When discrete transistors are used, the bias and operating points of the transistors interact with the gain and tuning of the stage. With op amps, the bias point is independent of gain and tuning.
- Op amp stages can operate over wide ranges of frequencies because there are no inductors to take into account.
- Transistor parameter drift and beta variation must also be taken into account over the system operating temperature range. When op amps are used, the drift is almost eliminated because gain is determined in stable passive components.

So what type of circuits can realistically be implemented with op amps? The remainder of this article describes wideband RF and narrow-band IF amplifiers.

Choosing the op amp

There is a decision to be made when selecting op amps for RF applications—whether to use voltage- or current-feedback amplifiers. Most analog interface designers are very familiar with voltage-feedback op amps but are needlessly apprehensive about current feedback. RF designers, being relatively new to op amps, probably have no such reservations. Voltage-feedback amplifiers, although a mature technology that is rapidly increasing in speed, still have a significant limitation of gain versus bandwidth imposed by their internal compensation capacitors. This limits their use in RF circuits to high-gain, relatively low-frequency circuits.

Current-feedback amplifiers are often pitched as having no limitation of gain versus bandwidth. Supposedly they are usable to almost their specified –3-dB frequency at just about any gain. This is only partially true. While there is no “–20 per decade” slope to their gain/bandwidth curve, they most definitely are bandwidth-limited at higher gains. Internal parasitics do take a toll on the bandwidth! The stability of current-feedback amplifiers is determined solely by their feedback resistor value. This value, or narrow range of values, is often fairly low—a few hundred ohms at most. This makes the useful gain range fairly low—a voltage gain of 10 or 12 in a single stage.

Wideband RF amplifiers

For wideband RF amplifiers, current-feedback amplifiers are the components of choice. As an example, the THS3202 in Figure 1 was chosen for its wide bandwidth and fast slew rate. The circuit shown was used to produce an amplifier voltage gain of 20 and a stage voltage gain of 10.

Note the simplicity of this circuit compared to traditional RF circuitry. Provide the op amp, the termination and decoupling components, and two resistors—and the circuit is done! The 301-Ω (R_F) and 16.5-Ω (R_G) resistors are all that are required to set the stage gain. This circuit produces the amplitude curve in Figure 2.

Figure 1. Wideband RF amplifier

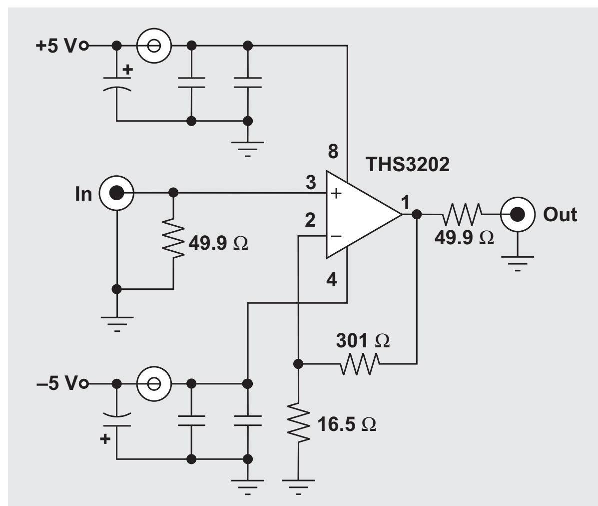


Figure 2. Wideband response for circuit in Figure 1

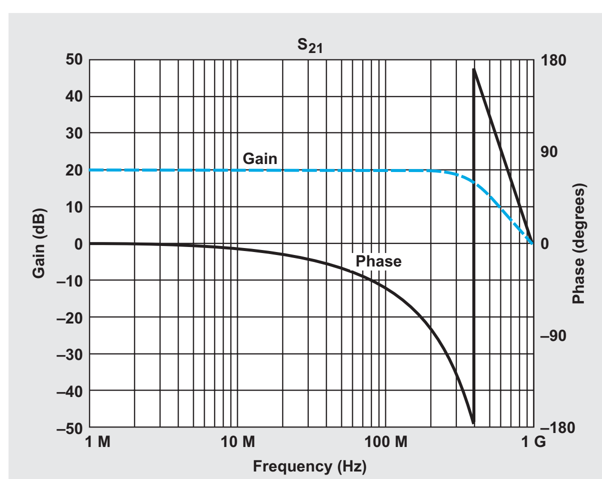
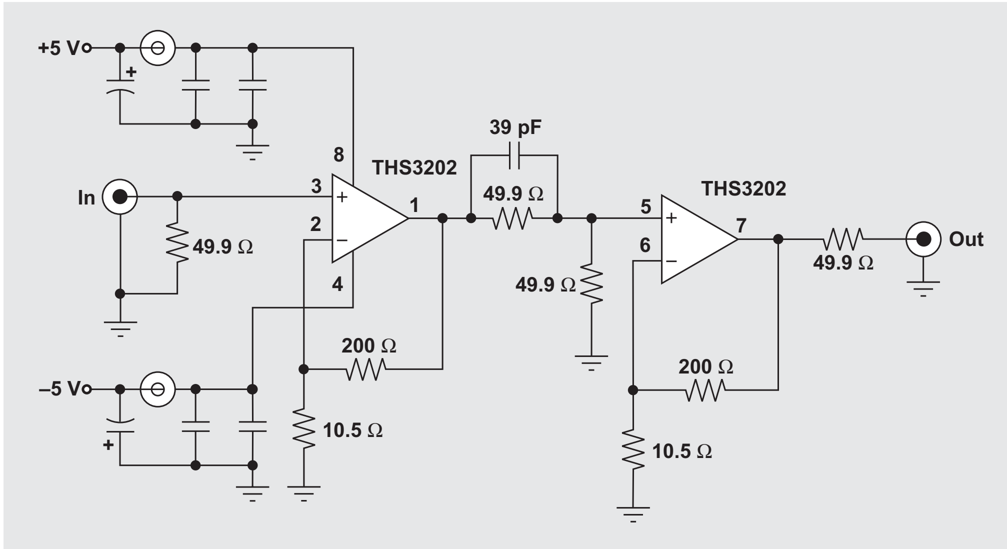


Figure 3. 40-dB wideband RF amplifier



The voltage gain of the op amp stage itself is 20, but this is cut in half by the action of the back termination resistor in combination with the load. The –3-dB point of the RF amplifier is about 390 MHz. If a flat gain over frequency is required, this circuit is only usable to about 200 MHz. Input and output VSWR values are better than 1.01:1 for most of the bandwidth, degrading to only about 1.1:1 near 200 MHz. S_{12} is –75 dB over most of the bandwidth, degrading to only –50 dB near the bandwidth limit.

One might wonder if more gain could be coaxed from the stage by lowering the gain resistor (R_G) even more. The answer is yes, but there is a practical limit. Remember that the feedback resistor (R_F) is the determining factor for current-feedback amplifier stability. Remember also that R_G has to drop proportionally more. One can see that it would not be long until the value of R_G became impractically small. Lab tests were attempted with various values of R_F. The result was that there is no advantage to making it smaller than 200 Ω. Below that, peaking starts to occur regardless of the value of R_G, becoming worse and worse as the resistance is made lower and lower. This is exactly what one would expect, because one thing to avoid in working with current-feedback amplifiers is to make R_F a short.

More gain requires cascading multiple stages of THS3202 op amps. Fortunately for the designer, the THS3202 is a dual device, making a two-stage RF amplifier easy to implement at very little additional cost.

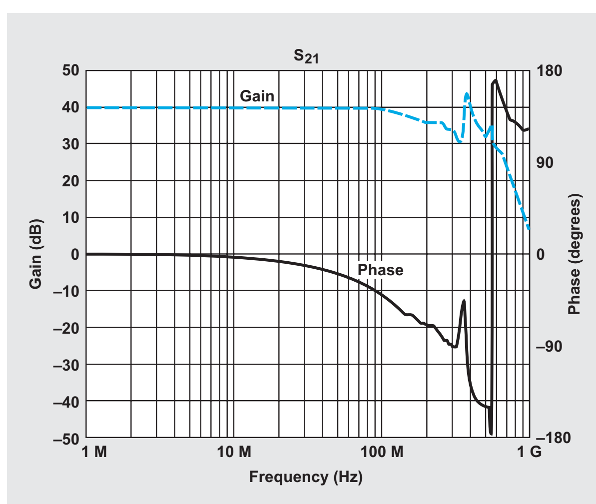
The circuit in Figure 3 shows such a cascaded RF amplifier, used to create a voltage gain of 100 (40 dB).

This circuit requires little explanation. It is obviously composed of two identical gain stages. Isolation is accomplished by using interstage termination resistors. The 39-pF capacitor provides peaking to compensate for some high-frequency roll-off, but better IP3 performance can be achieved by removing it and living with the roll-off. Overall circuit response is shown in Figure 4.

The reader will note that the stage is usable to 100 MHz, but there is significant peaking outside of that range. It is assumed that an amplifier of this high gain will be adequately shielded, and that there will be passive filtering components used to eliminate any problems from out-of-band response. Other S parameters for this circuit are similar to the case of the single op amp.

But what about voltage-feedback amplifiers? Is there any “niche” where they are more suitable than current-feedback amplifiers? There may be one. The stability of a voltage-feedback amplifier is primarily a function of its feedback and gain resistor ratio. Therefore, the designer is not constrained in the selection of a feedback resistor. Selecting a low-value feedback resistor can maximize the speed of the device; indeed, most data sheets recommend low feedback-resistor values. Making the resistor larger, however, does not adversely affect the stability of the

Figure 4. 40-dB RF amplifier response



amplifier. This means that large resistor ratios are possible without making R_G a very low value.

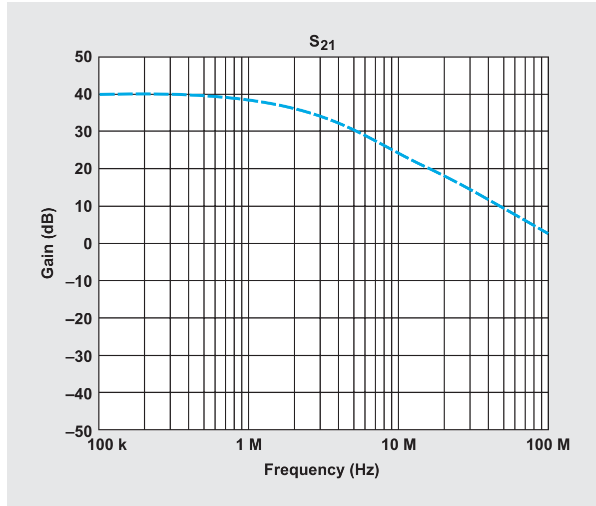
Figure 5 shows the response of a 40-dB RF amplifier constructed from a THS4271. Most notable is the bandwidth limitation; this amplifier is usable to only about 1 MHz. This might be valuable for medium-wave applications, but nothing higher in frequency. Yet the stage was constructed from a single op amp, with an R_F of 10 kΩ and an R_G of 49.9 Ω (that just happens to match the standard termination value). This gives a non-inverting gain of 201, but this gain is divided in two by the back termination resistor and the monitoring instrument. The freedom to use 10 kΩ as R_F allows R_G to be a reasonable value. Thus a single op amp is capable of delivering 40 dB of gain, something a current-feedback amplifier cannot do. As a test, a 455-kHz intermediate-frequency (IF) amplifier was constructed that had 37 dB of gain with a single op amp.

Therefore, there is a range of applications at relatively low frequencies and higher gains that are the exclusive domain of voltage-feedback amplifiers.

Intermediate-frequency (IF) amplifiers

The gain circuit shown in Figure 3 can easily be cascaded with ceramic filters and surface acoustic wave (SAW) filters to form high-performance IF stages. The only design consideration is the insertion loss of the filter, which may not be a constant value from part to part or from batch of parts to batch of parts, etc. If precise gain is needed from the stage, the designer may need to include a trim resistor in one or both stages. This trim adjustment, however, will not affect the tuning of the stage, except for a slight effect on its upper frequency limit.

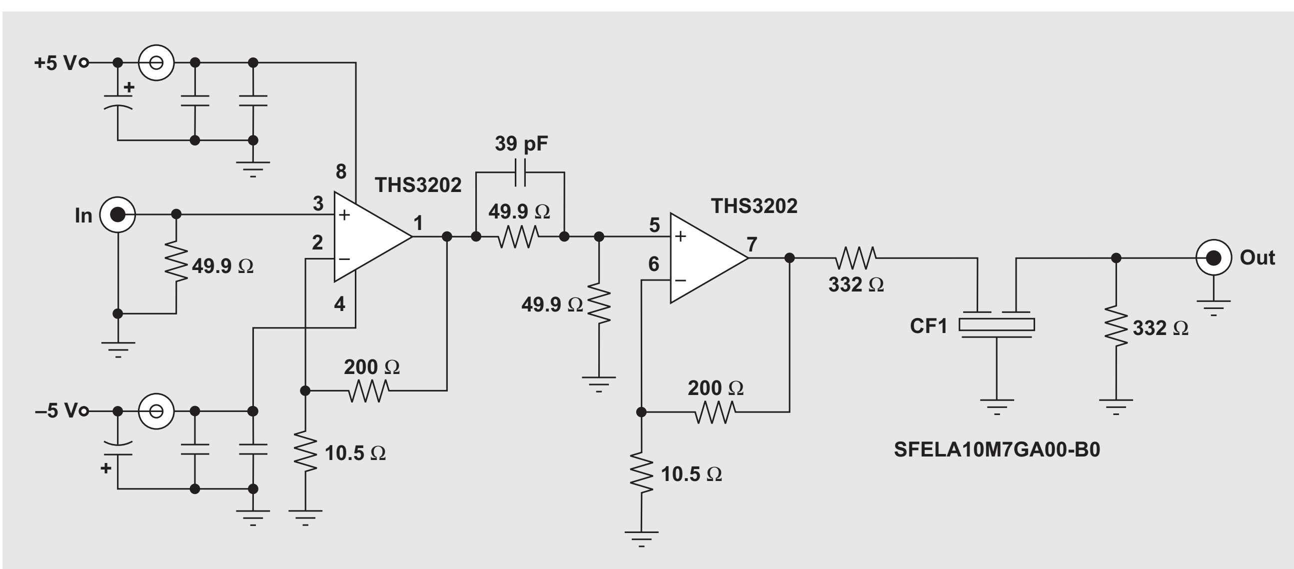
Figure 5. RF amplifier using THS4271



10.7-MHz IF amplifier

There is widespread use of 10.7-MHz IF amplifiers in FM broadcast receivers and cell phone base station receivers (final IF). These products make use of inexpensive ceramic filters, which are available in a variety of bandwidths and insertion losses. The circuit shown in Figure 6 was constructed with the gain stage shown in Figure 3 and a 230-kHz bandwidth filter from Murata, the

Figure 6. 10.7-MHz IF amplifier



SFELA10M7GA00-B0. This filter has a nominal insertion loss of 4 dB, and the circuit response is shown in Figure 7.

The slight nonlinearity in the passband has more to do with the characteristics of the ceramic filter than it does with the op amp. Insertion loss is slightly more than 4 dB. This response has been corrected for the impedances of the circuit, which are different from a nominal load of 50 Ω. The response was actually measured with a 280-Ω resistor in series with the 50-Ω instrument as the load. That made a voltage divider, the effects of which have been compensated for in the figure.

70-MHz IF amplifier

Cellular telephone base stations and satellite communications receivers use 70-MHz IF amplifiers. At these frequencies, SAW filters are used. For our example, an 854660 filter from Sawtek was selected. This unit requires input and output inductors, and operates with standard 50-Ω input and output. Sawtek conveniently provided an EVM board, which greatly simplified prototyping. The circuit shown in Figure 8 produces the 70-MHz response shown in Figure 9.

Figure 7. Response of 10.7-MHz IF amplifier

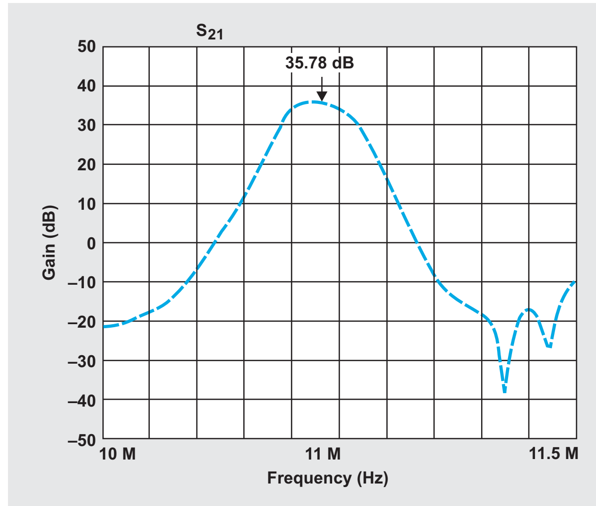
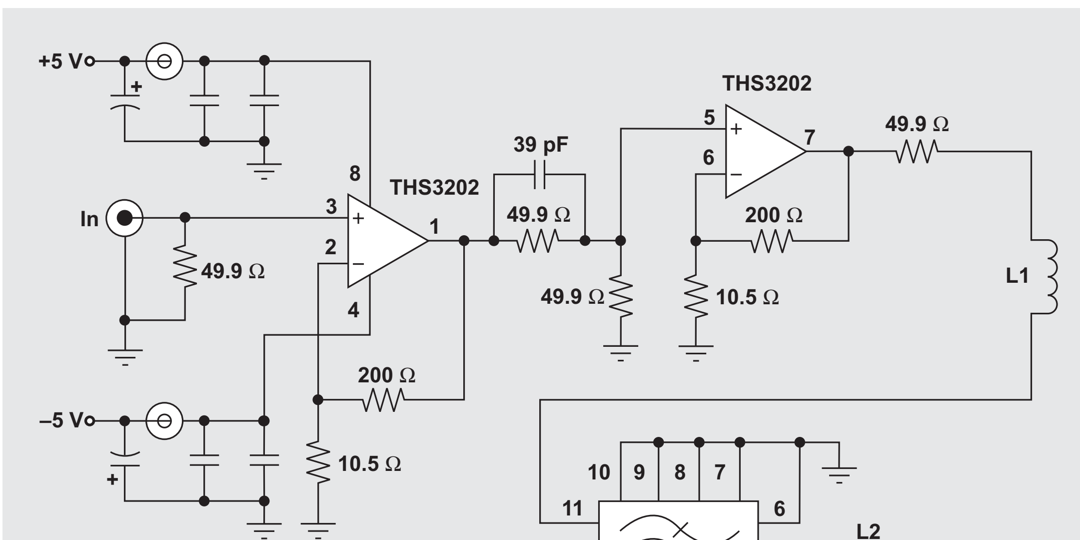


Figure 8. 70-/140-MHz IF amplifier



Device	IF Amplifier 70 MHz	IF Amplifier 140 MHz
L1 (nH)	180	47
L2 (nH)	180	56
Sawtek Filter	854660	854916

The response curve in Figure 9 is almost identical in shape to the curve of the Sawtek filter itself. In other words, the amplifier is providing gain, while not adding undesirable harmonic content. The insertion loss of the SAW filter is about 7 dB.

140-MHz IF amplifier

Cellular telephone base stations also use 140-MHz IF amplifiers. For our example, an 854916 filter from Sawtek was selected along with smaller inductors, as shown in Figure 8. The circuit response is shown in Figure 10.

As was the case for the 70-MHz IF filter, the response curve in Figure 10 is almost identical in shape to the curve of the Sawtek filter itself. Again, the amplifier is providing gain, while not adding undesirable harmonic content. The insertion loss of the SAW filter is only 8 dB maximum; but the gain circuit itself is starting to roll off at this frequency, accounting for the rest of the loss. Careful examination of the passband shows the slight roll-off of the gain stage.

Conclusion

Although inexpensive RF design continues to be the exclusive domain of transistors, there is a class of RF applications where performance—not cost—is the driving factor. These applications stand to benefit tremendously from the excellent RF performance that op amps can provide. By freeing the designer from the interrelated tasks of calculating biasing and gain, decoupling, and peaking, op amps also simplify RF design, even for novices. Troublesome components such as inductors and variable peaking capacitors are eliminated, and circuit gain depends on relatively stable resistors instead of the widely variable parameters of transistors. This makes RF design with op amps repeatable in production and eliminates trimming and aligning test stations when the product is manufactured. Maintenance of the equipment, therefore, is also simplified; no periodic adjustments are needed to maintain top levels of performance.

Related Web sites

analog.ti.com
www.ti.com/sc/device/THS3202
www.ti.com/sc/device/THS4271

Figure 9. Response of 70-MHz amplifier

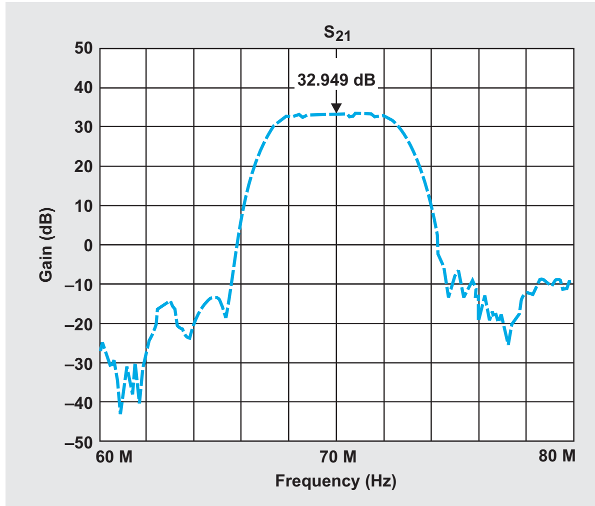
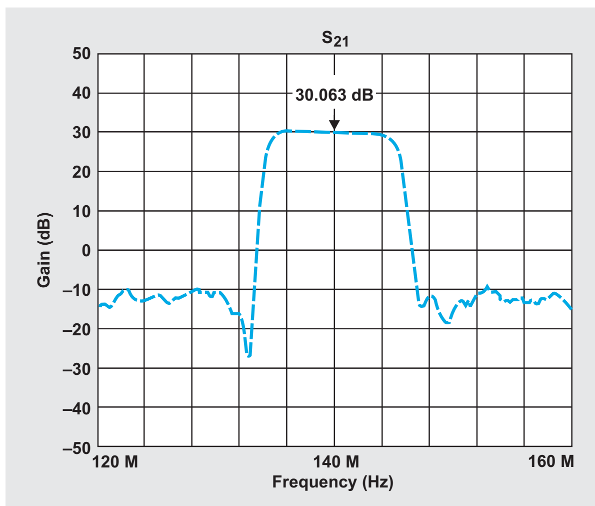


Figure 10. Response of 140-MHz amplifier



Analyzing feedback loops containing secondary amplifiers

By Ron Mancini (Email: rmancini@ti.com)
Staff Scientist, Advanced Analog Products

Introduction

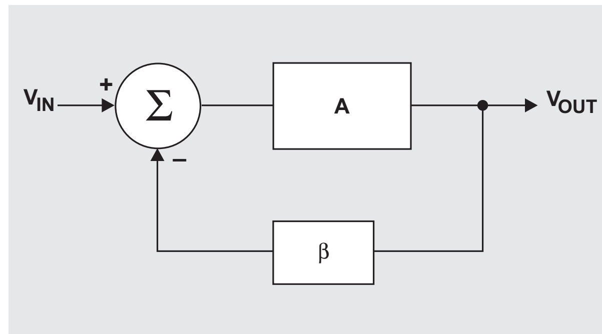
An op amp circuit contains a feedback loop, and sometimes it is advantageous to include a second amplifier within the op amp's feedback loop to modify closed-loop performance. For example, there is a limited selection of precision input/high-drive capability op amps. Because few such op amps exist, many designers connect a power amplifier in a precision op amp's feedback loop to obtain output-drive capability. Current amplifiers, linear-power amplifiers, switching-power amplifiers, summing amplifiers, nonlinear amplifiers, high-voltage amplifiers, and other amplifier variations are included in an op amp's feedback loop to obtain the desired closed-loop performance.

Sometimes the secondary amplifiers don't cause problems; but, more often than not, overshoot, ringing, or oscillation occurs when secondary amplifiers are added. Ringing and overshoot are the first signs of instability (with oscillation being the final sign). This article explains how the secondary op amp causes instability and how that instability can be cured.

Feedback theory

Feedback theory and its application to op amps are discussed in Reference 1. The classic feedback loop is shown in Figure 1.

Figure 1. Classic feedback loop with gain block, A, and feedback block, β

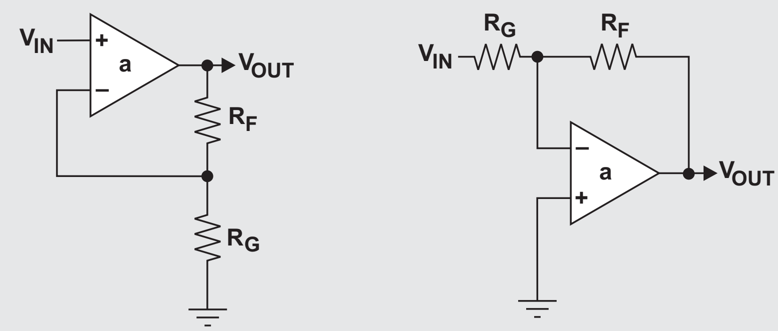


The block labeled "A" is the gain block, the block labeled "β" is the feedback block, and the condition for oscillation is $A\beta = -1 = |1| \angle{-180°}$. To achieve oscillation, the gain magnitude must equal 1 when the loop phase shift equals 0. It seems almost impossible to keep the gain magnitude equal to 1 in an active circuit (especially when the phase angle must be exact), but the op amp output stage introduces nonlinearities that automatically do that.

When an op amp starts to saturate, its output stage becomes nonlinear, thus reducing the overall gain and satisfying the condition for oscillation. This situation enables the oscillation criteria to be written as $A\beta \geq 1 \angle{-180°}$; hence, if the gain magnitude is greater than 1, the circuit oscillates when the amplifier phase shift is –180° because nonlinearities force the gain magnitude to 1. The greater the gain magnitude, the greater the nonlinearity required to achieve oscillation; consequently, the output voltage waveform becomes distorted rather than sinusoidal.

The classic feedback loop and op amp circuits have an inverting amplifier in the feedback loop. These circuits always oscillate at the frequency that yields –180° phase shift (when the gain ≥1) because this is the frequency where the feedback is in phase. Amplifier stability is evaluated at the point where the gain crosses the 0-dB axis (gain = 1). When the gain crosses over the axis, note the frequency and find the phase shift by following that frequency vertically to the phase curve. A measure of stability is phase margin, ϕ, which is calculated by subtracting the accumulated loop phase shift from 180°.

Figure 2. The two most common op amp configurations



(a) Non-inverting Amplifier **(b) Inverting Amplifier**

Real op amps and data sheets

Figure 2 shows the typical schematic for op amp circuits.

The loop gain, $A\beta$, is given in the following equation, which is identical for inverting, non-inverting, and differential circuits.

$$A\beta = \frac{aR_G}{R_G + R_F}$$

Parameter "a" in Figure 2 is the op amp open-loop gain (shown in Figure 3 for a TLC27Lx op amp). As the maximum loop gain or the vertical gain intercept (800,000) decreases, a circuit becomes more stable because it accumulates less phase shift before the gain curve crosses 0 dB. When the TLC27Lx's open-loop gain equals 1, rising vertically on a constant frequency line shows that the op amp's phase shift is approximately 145°. If $R_G = R_F$, the vertical intercept curve for the gain, a (A_{VD} in Figure 3), drops 6 dB from 800,000 to 400,000; thus the phase shift changes from 145° to 130°. This results in a circuit phase margin of $\phi = 180° – 130° = 50°$.

We should note several things at this point: Increasing phase margin increases stability, inverting and non-inverting op amp circuits have the same loop gain, and increasing the closed-loop gain increases stability. If a second TLC27Lx op amp configured as a non-inverting amplifier with a gain of 2 (see Figure 4) is included in the feedback loop, is the circuit stable?

The answer depends on the phase shift of the secondary op amp. With $R_F = R_G = 10\ k\Omega$, the secondary op amp (TLC27Lx) has a measured phase shift of 90° at f = 73 kHz. Figure 3 shows that the primary op amp has 100° phase shift at 73 kHz with a gain of 15, so the complete circuit with the secondary op amp can easily achieve the criteria for oscillation. Actually, the circuit oscillates at 22.7 kHz; the exact frequency of oscillation is extremely hard to predict because there are two op amps contributing phase shift, and the phase/frequency transfer function is nonlinear.

There are three solutions to this problem. First, the loop gain can be reduced by inserting an attenuator in the feedback loop. The reduced loop gain causes a reduced accumulated phase shift at the 0-dB crossover point, and some value of loop gain can be found that yields a phase

Figure 3. Gain/phase curves for the TLC27Lx

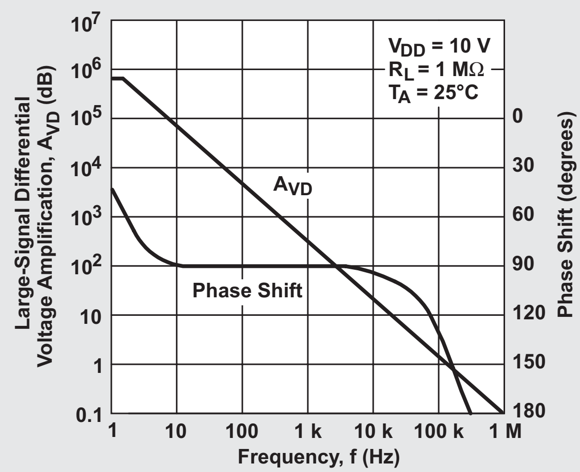
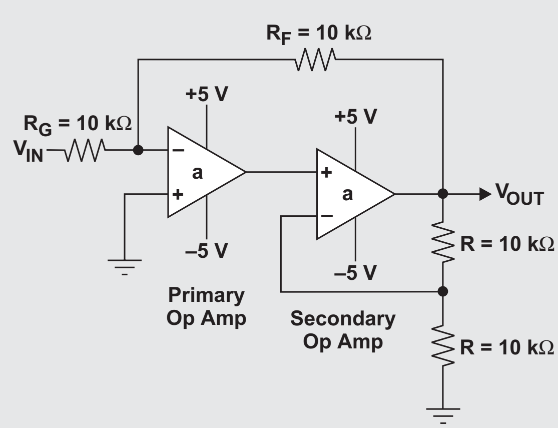


Figure 4. A secondary op amp in the feedback loop can cause oscillations



shift of less than –180°. Reduced loop gain reduces the closed-loop accuracy, so this solution is generally not acceptable. Second, the closed-loop gain can be increased, causing a loop-gain reduction and consequent increase in stability. This solution is preferable if the input signal is small enough to restrict the output voltage swing to the secondary op amp's output voltage range. Third, a secondary op amp can be selected whose phase shift is close to 0 at the 0-dB crossover point.

The TLC27Mx's unity-gain bandwidth or 0-dB crossover frequency (see Figure 5) is 635 kHz compared to the TLC27Lx's unity-gain bandwidth of 110 kHz. The 5.77/1 ratio of unity-gain bandwidths indicates that TLC27Mx's phase-shift contribution should be too small to accumulate –180° phase shift when the TLC27Lx crosses the 0-dB point. The TLC27Mx has 31° phase shift at 266 kHz when it is configured as a non-inverting op amp. Figure 3 shows that the accumulated phase shift is 160° (130° from the primary op amp and 30° from the secondary op amp), so the circuit is stable with a TLC27Mx used as the secondary op amp. Any op amp with a higher unity-gain bandwidth will satisfy the stability criteria, but higher-speed op amps tend to burn more power.

The circuit oscillates at 266 kHz when the TLC27Mx is used as the primary and secondary op amps. The measured non-inverting amplifier phase shift of the TLC27Mx is 90° at a frequency of 500 kHz. Figure 5 shows that the TLC27Mx has accumulated approximately 110° phase shift at 500 kHz with a gain of approximately 5. This circuit meets the criteria for an oscillator, so we find no surprises here. If we select a secondary op amp with a unity-gain bandwidth of approximately five times the TLC27Mx unity-gain bandwidth of 635 kHz, the circuit should become stable. Let's select a TLC07x op amp (see Figure 6a) because it has a unity-gain bandwidth of 10 MHz. The ratio of unity-gain bandwidths is 10/0.635 = 15.7, so the circuit should be very stable; and measurements prove this statement to be accurate.

The measured non-inverting amplifier phase shift of the TLC07x is 15° at a frequency of 442 kHz. The TLC27Mx has a phase shift of 150° at the 0-dB crossover frequency; thus the theory agrees with the measurements because the accumulated phase shift is 165°. The circuit oscillates when the TLC07x is used for the primary and secondary amplifiers. The TLC07x's output-current capability is ±50 mA, while the TLC27Mx's output current is limited to a few milliamps; so adding the TLC07x as the secondary amplifier has increased the loop's ability to drive high-current loads.

Figure 5. Gain/phase curves for the TLC27Mx

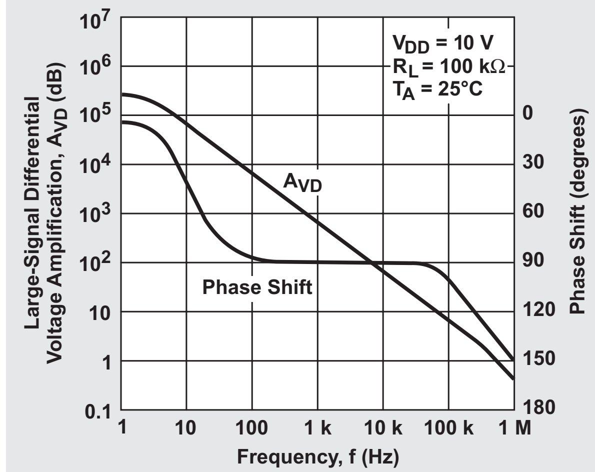
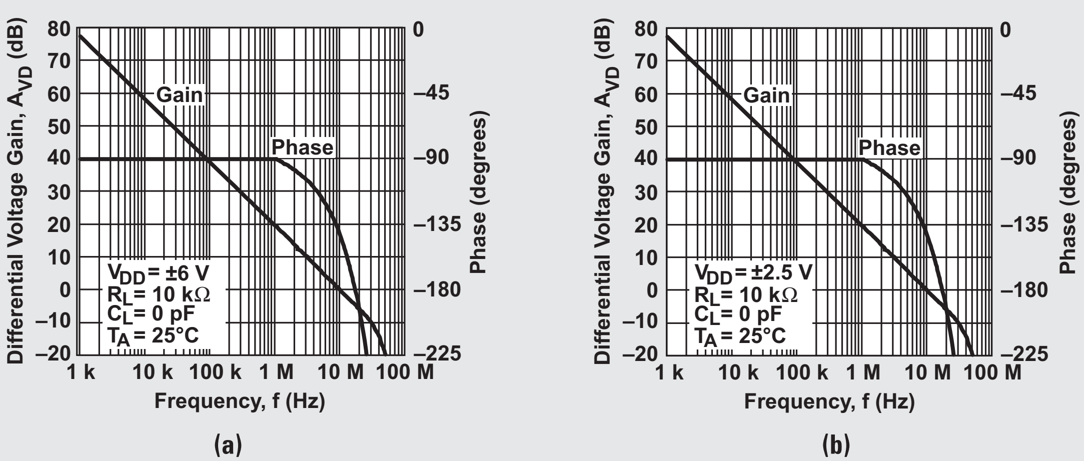


Figure 6. Gain/phase curves for the TLC07x



Conclusion

When two op amps with similar unity-gain bandwidths are included in a feedback loop, oscillations occur. The unity-gain bandwidth of the op amps must be separated by a factor of five (rule-of-thumb) to prevent oscillations. Bandwidth data is not a guaranteed specification, and a phase-versus-frequency calculation is nonlinear; so the rule-of-thumb can't be relied upon for more than a mathematical starting point. All assumptions must be verified in the lab, and significant margins must be left to accommodate parameter changes.

Circuit stability is increased when the closed-loop gain is increased because the loop gain decreases. Circuit stability is decreased when the secondary amplifier gain is increased because the loop gain increases. The primary op amp is usually selected because it has precision or low-noise specifications. The secondary op amp is usually selected because it has excellent voltage/current-drive specifications.

The primary amplifier usually has the higher unity-gain bandwidth, but this is not a rigid criterion. A feedback loop is a linear system; thus the gain blocks can be interchanged within the loop without affecting performance. High-power amplifiers may not be available with high bandwidth, so in that case the primary amplifier is selected as the high-unity-gain bandwidth amplifier to obtain the best speed performance.

Reference

For more information related to this article, you can download an Acrobat Reader file at www-s.ti.com/sc/techlit/*litnumber* and replace *"litnumber"* with the **TI Lit. #** for the materials listed below.

Document Title	**TI Lit. #**
1. Ron Mancini (ed.), *Op Amps for Everyone*, Design Reference	slod006

Related Web sites

analog.ti.com

www.ti.com/sc/device/*partnumber*

Replace *partnumber* with TLC07X, TLC27LX or TLC27MX

Index of Articles

TI Worldwide Technical Support

Internet

TI Semiconductor Product Information Center Home Page

support.ti.com

TI Semiconductor KnowledgeBase Home Page

support.ti.com/sc/knowledgebase

Product Information Centers

Americas

Phone	+1(972) 644-5580
Fax	+1(972) 927-6377
Internet/Email	support.ti.com/sc/pic/americas.htm

Europe, Middle East, and Africa

Phone	
Belgium (English)	+32 (0) 27 45 55 32
Finland (English)	+358 (0) 9 25173948
France	+33 (0) 1 30 70 11 64
Germany	+49 (0) 8161 80 33 11
Israel (English)	1800 949 0107
Italy	800 79 11 37
Netherlands (English)	+31 (0) 546 87 95 45
Spain	+34 902 35 40 28
Sweden (English)	+46 (0) 8587 555 22
United Kingdom	+44 (0) 1604 66 33 99
Fax	+(49) (0) 8161 80 2045
Email	epic@ti.com
Internet	support.ti.com/sc/pic/euro.htm

Japan

Fax	International	+81-3-3344-5317
	Domestic	0120-81-0036
Internet/Email	International	support.ti.com/sc/pic/japan.htm
	Domestic	www.tij.co.jp/pic

Asia

Phone	
International	+886-2-23786800
Domestic	Toll-Free Number
Australia	1-800-999-084
China	108-00-886-0015
Hong Kong	800-96-5941
Indonesia	001-803-8861-1006
Korea	080-551-2804
Malaysia	1-800-80-3973
New Zealand	0800-446-934
Philippines	1-800-765-7404
Singapore	800-886-1028
Taiwan	0800-006800
Thailand	001-800-886-0010
Fax	886-2-2378-6808
Email	tiasia@ti.com
Internet	support.ti.com/sc/pic/asia.htm

A010203

SLYT045