AN2653
Application note

Operational amplifier stability compensation methods for capacitive loading applied to TS507

Introduction

Who has never experienced oscillations issues when using an operational amplifier? Op-amps are often used in a simple voltage follower configuration. However, this is not the best configuration in terms of capacitive loading and potential risk of oscillations.

Capacitive loads have a big impact on the stability of operational amplifier-based applications. Several compensation methods exist to stabilize a standard op-amp. This application note describes the most common ones, which can be used in most cases.

The general theory of each compensation method is explained, and based on this, specific data is provided for the TS507. The TS507 is a high precision rail-to-rail amplifier, with very low input offset voltage, and a 1.9 MHz gain bandwidth product, which is available in SOT23-5 and SO-8 packages.

This document simplifies the task of designing an application that includes the TS507. It spares you the time-consuming effort of trying numerous combinations on bench, and it is also much more accurate than using Spice models which are not designed to study system stability, even though they can give a general trend.

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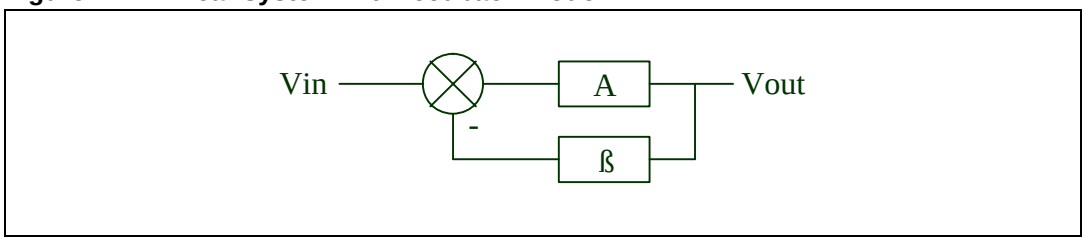
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1 Stability basics

1.1 Introduction

Consider a linear system modeled as shown in *Figure 1*.

Figure 1. Linear system with feedback model



The model in *Figure 1* gives the following equation:

$$V_{out} = \frac{A}{1 + A\beta} \cdot V_{in}$$

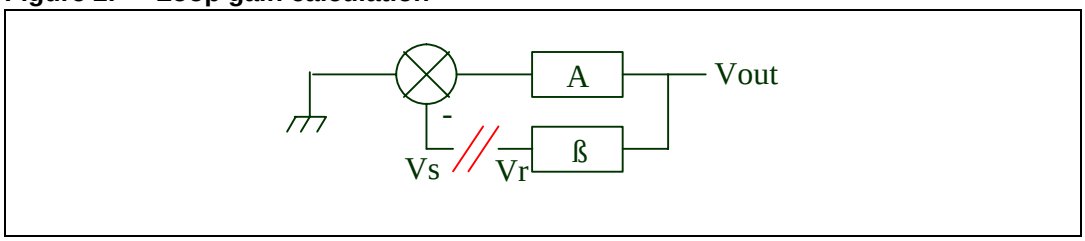
$\frac{A}{1 + A\beta}$ is named **closed loop gain**.

From this equation, it is evident that for $A\beta = -1$, the circuit is unstable (V_{out} is independent of V_{in}).

$A\beta$ is the **loop gain**.

To evaluate it, the loop is opened and $-V_r/V_s$ is calculated as shown in *Figure 2*.

Figure 2. Loop gain calculation



Opening the loop leads to the following equation:

$$-\frac{V_r}{V_s} = A\beta$$

If a small signal V_s is sourced into the system, and if V_r comes back in phase with it with an amplitude above that of V_s (which means that $A\beta$ is a real number greater than or equal to 1) then the system oscillates and is unstable.

This leads to the definition of the **gain margin**, which is the opposite of the loop gain (in dB) at the frequency for which its phase equals -180°. The bigger the gain margin, the more stable the system. In addition, the **phase margin** is defined as the phase of the loop gain plus 180° at the frequency for which its gain equals 0 dB. Therefore, from the value of $A\beta$ it is possible to determine the stability of the system.

1.2 Operational amplifier modeling for stability study

Figure 3 illustrates the definition of phase and gain margins in a gain configuration.

Figure 3. Illustration of phase and gain margins

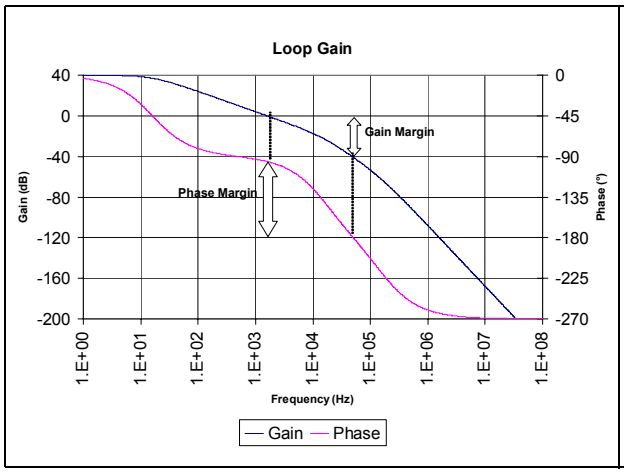
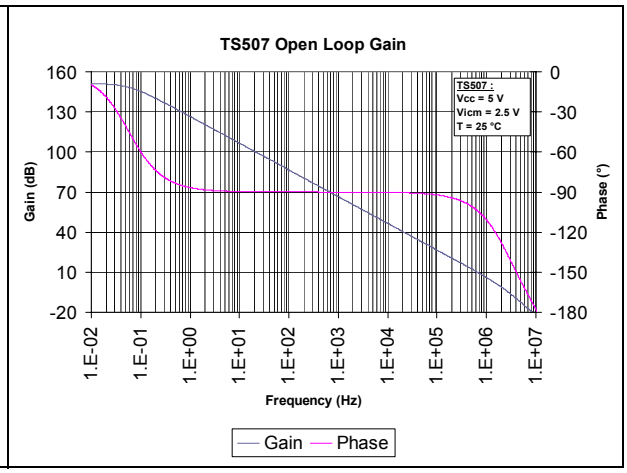
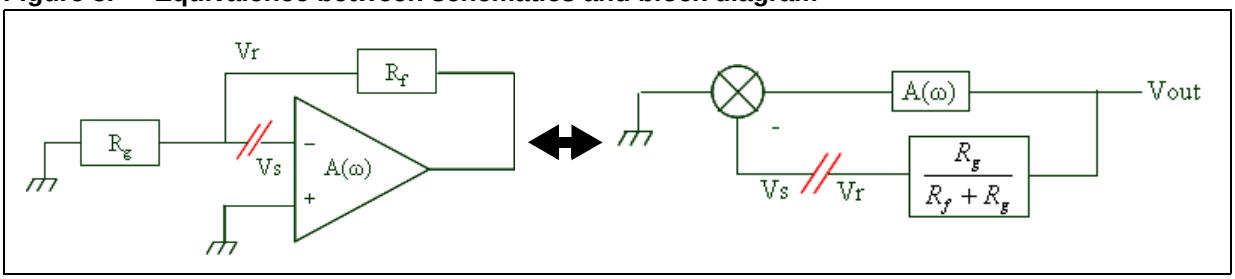


Figure 4. TS507 open loop gain



To apply this stability approach to operational amplifier based applications, it is necessary to know the gain of the operational amplifier when no feedback and no loads are used. It is the **open loop gain** (A(ω)) of the amplifier (shown in *Figure 4* for the TS507). From this parameter, it is possible to model the amplifier and to study the stability of any gain configuration.

Figure 5. Equivalence between schematics and block diagram



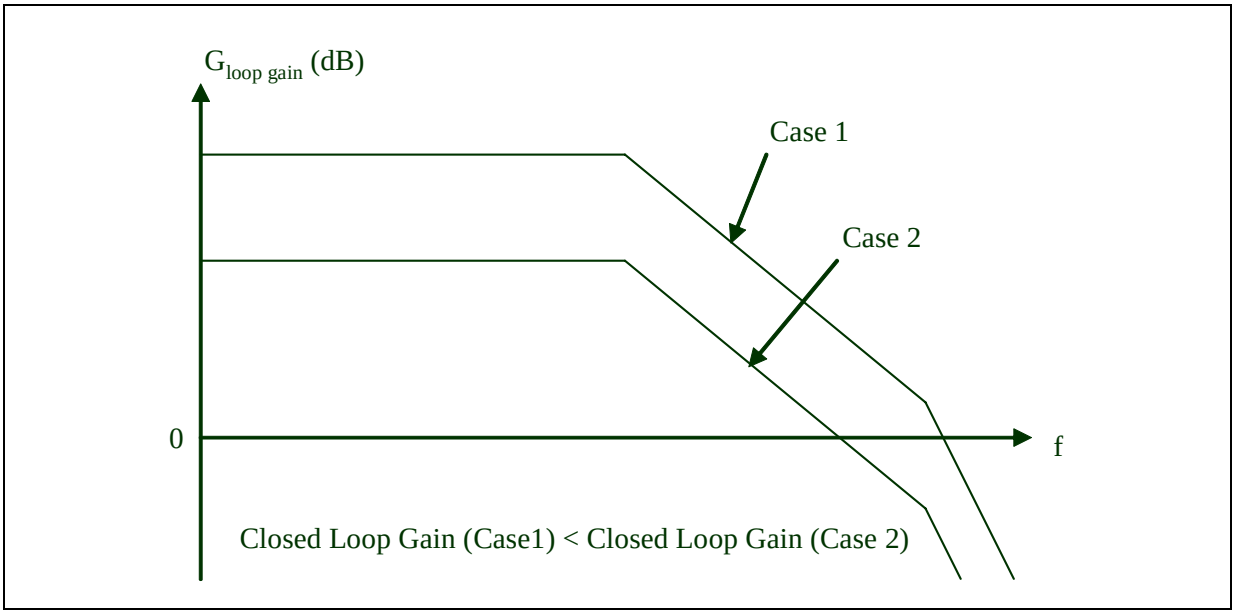
The loop gain is:

$$-\frac{V_r}{V_s} = A(\omega) \cdot \frac{R_g}{R_f + R_g}$$

This equation shows the impact of the gain on the stability: if R_f/R_g increases, the closed loop gain of the system increases and the loop gain decreases. Because the phase remains the same, the gain margin increases and stability is improved.

In addition, if you consider the case of a second order system such as the one shown in *Figure 6*, a decrease of the loop gain allows to pass the 0 dB axis before the second pole occurs. It minimizes the effect of the phase drop due to this pole, and as a result, the phase margin is higher. Therefore, a voltage follower configuration is the worst case for stability.

Figure 6. Impact of closed loop gain on stability



Another parameter that impacts stability is the amplifier **output impedance** Z_o. Including this parameter in the model of the amplifier leads to the model shown in *Figure 7*.

Figure 7. Follower configuration model with capacitive load for loop gain calculation

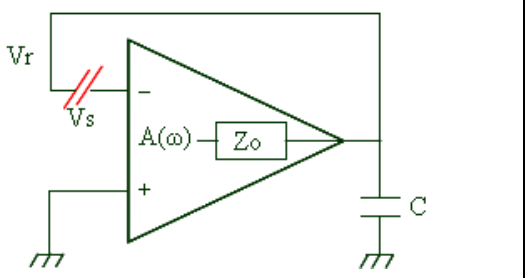
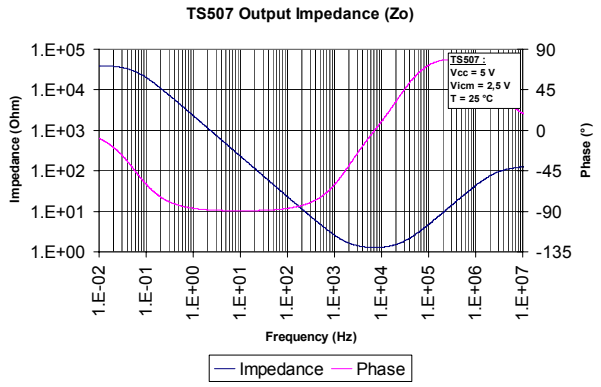


Figure 8. TS507 output impedance Z_o



Z_o is neither constant over frequency nor purely resistive. *Figure 8* shows how the output impedance varies with the frequency in the case of the TS507. These variations complicate the stability study.

Finally, to study the stability of an op-amp based system, two parameters need to be taken into account in order to better fit reality: the amplifier open-loop gain and the amplifier output impedance. Then, a calculation of the loop gain indicates how stable the system is.

2 Stability in voltage follower configuration

This section examines a voltage follower configuration because it is the worst case scenario for stability (compared with a gain configuration).

Figure 9. Voltage follower configuration

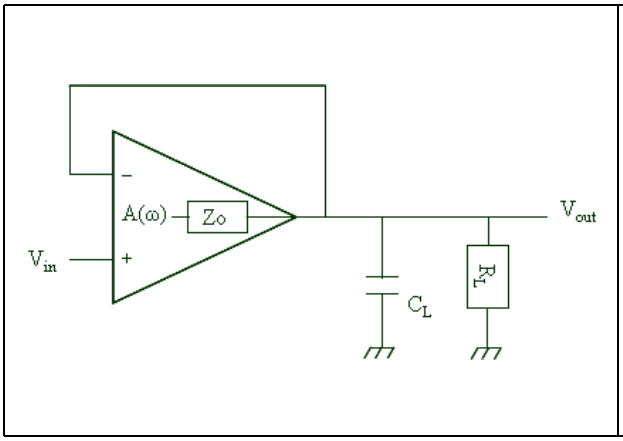
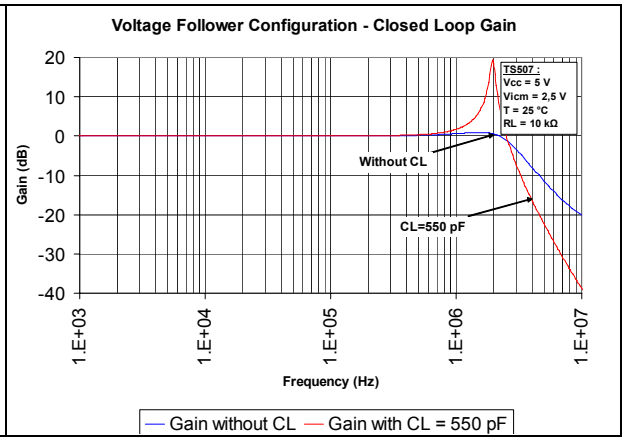


Figure 10. Closed loop gain measured for a voltage follower configuration



In voltage follower configuration, the loop gain is:

$$-\frac{V_r}{V_s} = \frac{A(\omega)}{1 + \frac{Z_0}{R_L} + jZ_0C_L\omega}$$

The capacitive load adds a pole to the loop gain that impacts the stability of the system. The higher the frequency of this pole, the greater the stability. In fact, if the pole frequency is lower than or close to the unity gain frequency, the pole can have a significant negative impact on phase and gain margins. It means that the stability decreases when the capacitive load increases.

Without C_L, the system is stable. However, *Figure 11* and *Figure 12* show, for the TS507, the oscillations due to instability with and without an AC input signal for a capacitive load of 550 pF. The oscillation frequency is in line with the peaking frequency observed in a closed loop gain configuration (approximately 1.9 MHz according to *Figure 10*).

Figure 11. Input and output signals measured with grounded input

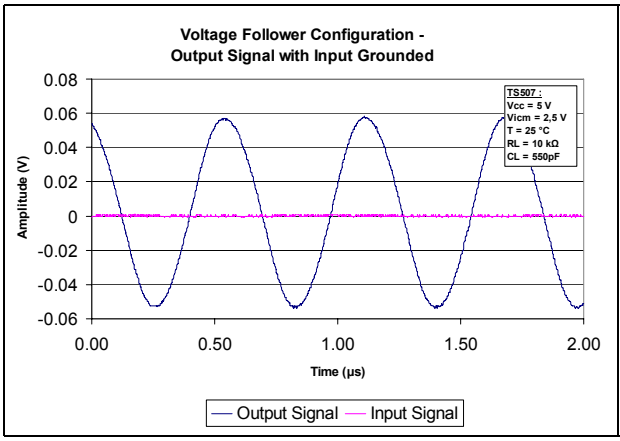
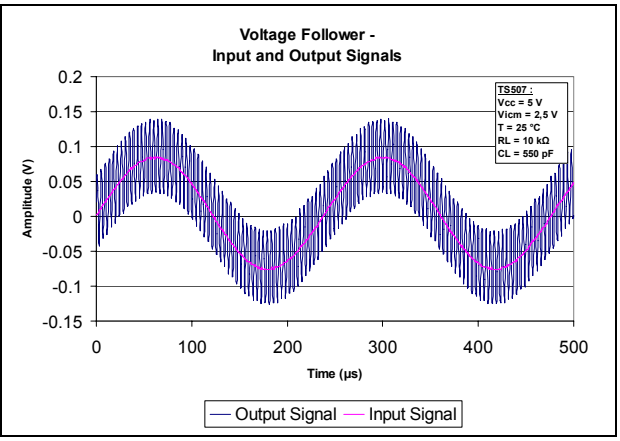


Figure 12. Input and output signals measured for an AC input signal



To remove this instability and work with higher capacitive loads, many compensation methods exist, and this application note examines some of them. By adding zeroes and poles to the loop gain, stability can be improved.

However, compensation components have to be chosen carefully. A compensation scheme can indeed improve stability, but can also lead the system to instability, depending on the choice of component values. Similarly, a compensation configuration can work for a specific load, but modifying this load can affect stability.

3 Out-of-the-loop compensation method

3.1 Theoretical overview

A simple compensation method, using only one extra component, consists in adding a resistor in series between the output of the amplifier and its load (see *Figure 13*). It is often referred to as the **out-of-the-loop** compensation method because the additional component (R_{OL}) is added outside of the feedback loop. The resistor isolates the op-amp feedback network from the capacitive load.

Figure 13. Out-of-the-loop compensation schematics

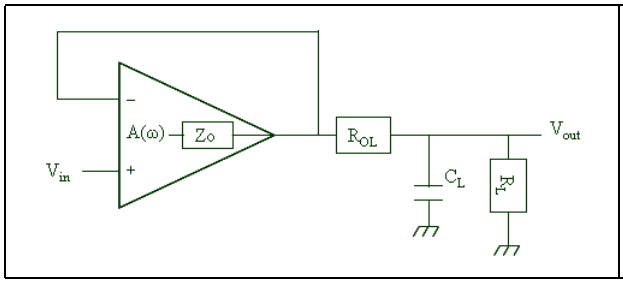
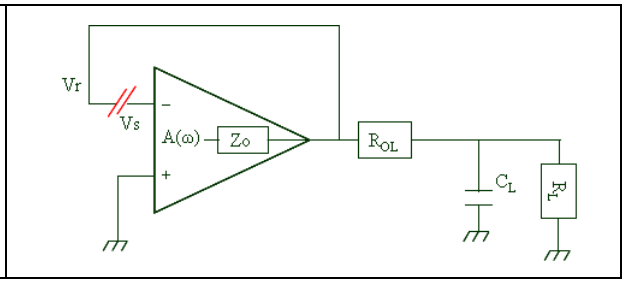


Figure 14. Out-of-the-loop equivalent schematics for loop gain calculation



From *Figure 14*, the loop gain with this compensation method is:

$$-\frac{V_r}{V_s} = \frac{A(\omega)\cdot\left(1+\frac{R_{OL}}{R_L}+jR_{OL}C_L\omega\right)}{1+\frac{R_{OL}+Z_o}{R_L}+j(Z_o+R_{OL})C_L\omega}$$

This compensation introduces a zero in the loop gain, just after the pole caused by the capacitive load, at:

$$\frac{1}{2\pi\cdot R_{OL}\,||\,R_L C_L}$$

This pole is also unfortunately shifted to lower frequencies at:

$$\frac{1}{2\pi\cdot (Z_o+R_{OL})\,||\,R_L C_L}$$

However, due to the zero, the effect of the pole is minimized and the stability is improved. To obtain a good level of stability, R_{OL} must be chosen such that the frequency of the zero occurs at least one decade before unity-gain frequency. It then allows a significant shift of the phase and therefore increase phase and gain margins.

The previous equation shows that if $R_{OL} >> Z_o$, then $-V_r/V_s = A(\omega)$, and the circuit is stable. In that case, pole and zero occur at the same frequency. However, the value of R_{OL} is limited by the load impedance, R_{OL} and R_L acting as a a divider bridge from the operational amplifier output. Therefore, in order to minimize the error on V_{out}, R_{OL} must be very small compared to R_L (for example, a maximum of 1%, but this criterion depends on the required accuracy).

Finally, this compensation method is effective, but the drawback is a limitation on the accuracy of V_{out} depending on the resistive load value.

3.2 Application on the TS507

This compensation method brings very good results in terms of stability, improving strongly the phase and gain margins. *Table 1* and *Table 2* show the results obtained for different load conditions, in the case of voltage follower and gain configurations. Note that R_{OL} is limited to 1% of R_L even though better results can be obtained with higher values of R_{OL}.

Table 1. Results of out-of-the-loop compensation for different load conditions in the case of a voltage follower configuration for TS507

C_L	R_L = 1 kΩ				R_L = 10 kΩ				R_L = 100 kΩ			
	R_{OL} (Ω)	f_u/f_z (1)	Mg (2) (dB)	Mφ (2) (degree)	R_{OL} (Ω)	f_u/f_z (1)	Mg (2) (dB)	Mφ (2) (degree)	R_{OL} (kΩ)	f_u/f_z	Mg (2) (dB)	Mφ (2) (degree)
1 nF			-4.1	-28.5			-5	-34.1			-5.1	-34.4
	10	0.11	-2.5	-16.8	100	1.13	16	26.9	1	11.3	22.4	52.1
10 nF			-22.2	-78.4			-22.9	-79.5			-23	-79.6
	10	1.13	-14	-32.4	100	11.3	23	37	1	112.3	22.6	52.3
100 nF			-34.1	-84.4			-34.4	-84.6			-34.5	-84.6
	10	11.3	17.1	6.8	100	113.3	23.4	39.4	1	1126	22.6	52.3

1. f_u/f_z cells are shaded when the value is lower than 10, which is not the best case due to R_{OL} limitation.
2. Negative values indicate instability.

Table 2. Results of out-of-the-loop compensation for different load conditions in the case of a gain configuration of either -10 or +11 (R_g = 100 Ω and R_f = 1 kΩ) for TS507

C_L	R_L = 1 kΩ				R_L = 10 kΩ				R_L = 100 kΩ			
	R_{OL} (Ω)	f_u/f_z (1)	Mg (2) (dB)	Mφ (2) (degree)	R_{OL} (Ω)	f_u/f_z (1)	Mg (2) (dB)	Mφ (2) (degree)	R_{OL} (kΩ)	f_u/f_z	Mg (2) (dB)	Mφ (2) (degree)
1 nF			17.6	84.7			16.8	85.1			16.7	85.2
	10	0.11	19	84.7	100	1.13	36.9	85.1	1	11.3	43.4	85
10 nF			-0.6	-16.1			-1.3	-25.7			-1.4	-25.9
	10	1.13	7.2	81.2	100	11.3	43.9	81.4	1	112.6	43.4	84.8
100 nF			-13	-69.2			-13.3	-69.8			-13.3	-69.9
	10	11.3	38	41	100	113.3	44.3	80.6	1	1126	43.4	84.8

1. f_u/f_z cells are shaded when the value is lower than 10, which is not the best case due to R_{OL} limitation.
2. Negative values indicate instability.

As expected, *Table 1* and *Table 2* show that the higher the value of R_{OL}, the better the compensation (because the best R_{OL} is always its maximum value R_L/100).

These results also show that, for a voltage follower configuration, this compensation method does not work with low R_L (and low C_L), because the zero frequency cannot be one decade before the unity-gain frequency of the open loop gain. In the case of the TS507, it works well only if the $R_{OL}.C_L$ product is above 10^{-6}.

Figure 15 and *Figure 16* show the loop gain and closed loop gain respectively. These curves are plotted for R_L = 10 kΩ and C_L = 1 nF.

Figure 15. Loop gain

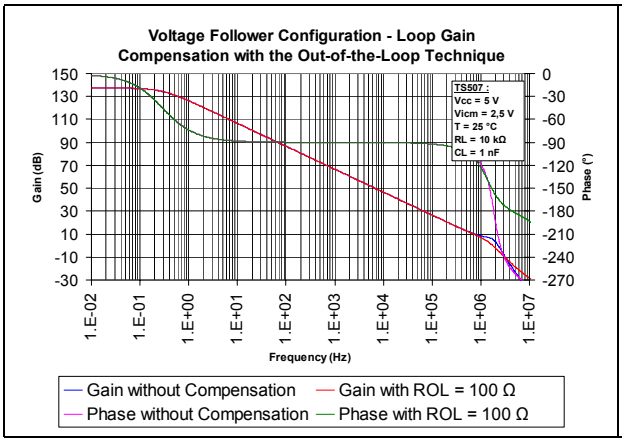
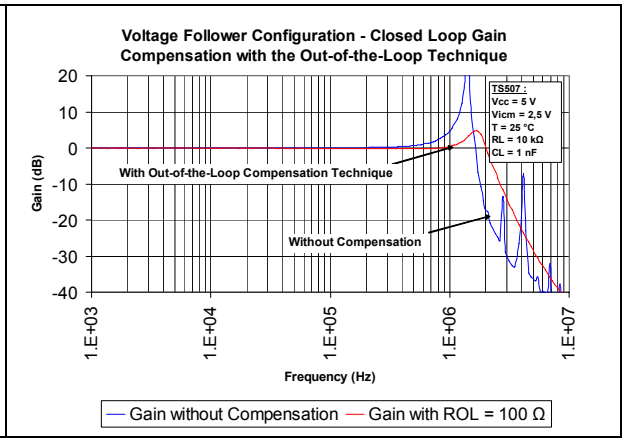


Figure 16. Measured closed loop gain



Both figures further demonstrates the stability improvement.

Note that the fact that Z_0 is almost a self at high frequencies (for the TS507) explains the presence of peaking in the loop gain curve, depending on the load capacitor. This is because the denominator is equal to

$$1 + \frac{R_{OL}}{R_L} + j\left(\frac{L(\omega)}{R_L} + R_{OL}C_L\omega\right) - L(\omega)C_L\omega^2$$

with $Z_0 = jL(\omega)\omega$

It leads to a resonance frequency of approximately

$$\frac{1}{2\pi \cdot \sqrt{L(\omega)C_L}}$$

For the peaking frequency

$$f = \frac{1}{2\pi \cdot \sqrt{L(\omega)C_L}}$$

the damping is given by the term:

$$\frac{L(\omega)}{R_L} + R_{OL}C_L\omega$$

When there is no compensation, it is only:

$$\frac{L(\omega)}{R_L}$$

With the compensation, at the resonance frequency,

$$\frac{L(\omega)}{R_L} \ll R_{OL}C_L\omega$$

therefore the peaking is attenuated.

To help implement the compensation, the abacus given in *Figure 17* to *Figure 20* provide the R_{OL} value to choose for a given C_L and phase/gain margins. These abacus are plotted in the case of a voltage follower configuration and a gain configuration of -10 or +11, with a load resistor of 10 kΩ.

Figure 17. Gain margin abacus in the case of a voltage follower configuration

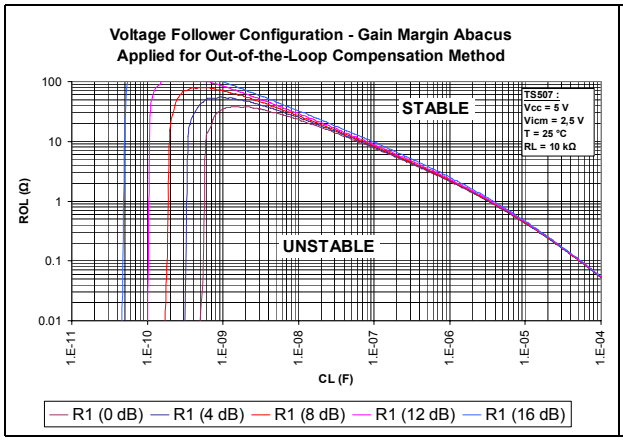


Figure 18. Phase margin abacus in the case of a voltage follower configuration

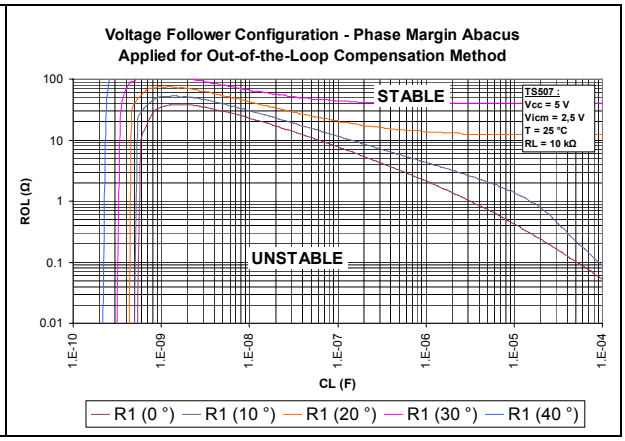


Figure 19. Gain margin abacus in the case of a gain configuration of -10 or +11

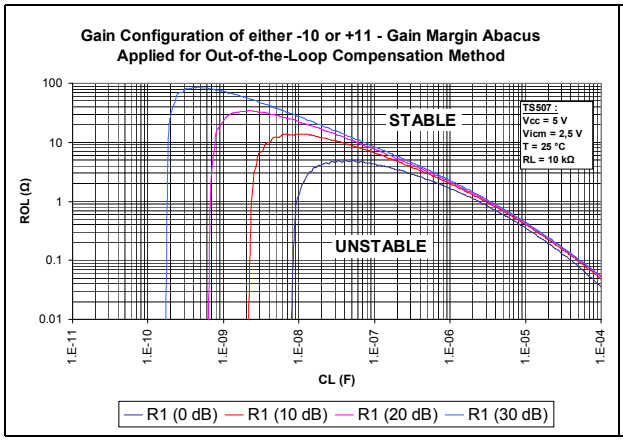
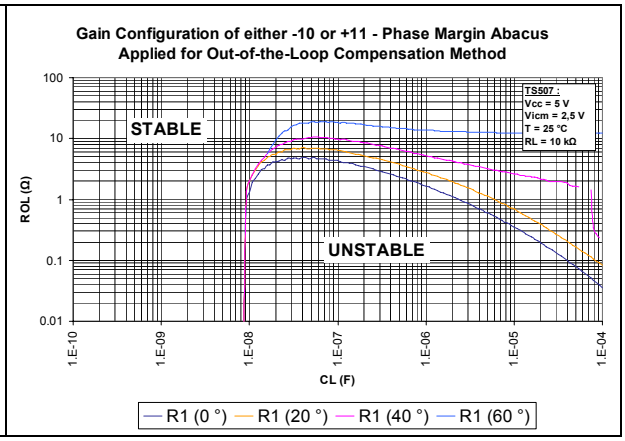


Figure 20. Phase margin abacus in the case of a gain configuration of -10 or +11



4 In-the-loop compensation method

4.1 Theoretical overview

Figure 21 shows a commonly used compensation method, often called *in-the-loop*, because the additional components (a resistor and a capacitor) used to improve the stability are inserted in the feedback loop.

Figure 21. In-the-loop compensation schematics

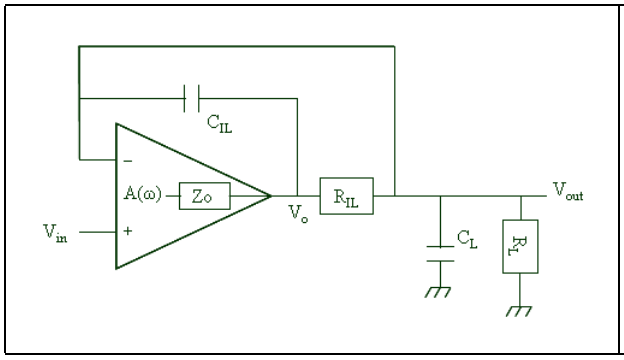
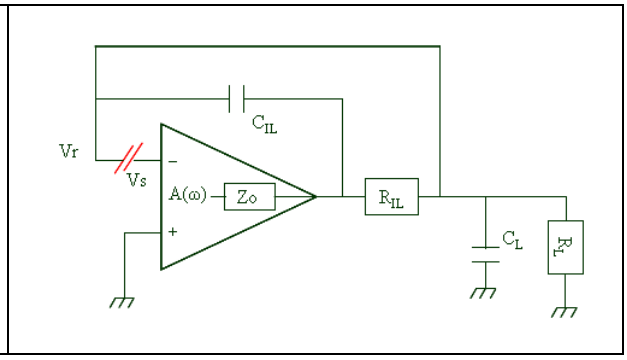


Figure 22. In-the-loop equivalent schematics for loop gain calculation



The loop gain in this configuration, corresponding to *Figure 22*, is the following:

$$-\frac{V_r}{V_s} = \frac{A(\omega)\cdot\ [1 + jR_{IL}C_{IL}\omega]}{1 + \frac{Z_o + R_{IL}}{R_L} + j(Z_o + R_{IL})C_L\omega + jR_{IL}\left(1 + \frac{Z_o}{R_L}\right)C_{IL}\omega - Z_oR_{IL}C_{IL}C_L\omega^2}$$

It adds a zero and splits the pole caused by the capacitive load into two poles in the loop gain. This compensation method allows, by a good choice of compensation components, to compensate the original pole (caused by the capacitive load), and then to improve stability.

The main drawback of this circuit is the reduction of the output swing, because the isolation resistor is in the signal path.

Note that, for the following cases, R_{IL} is limited to 10% of R_L (or R_f // R_L in the case of a gain configuration) even if better results can be obtained with higher R_{IL} values.

But because the feedback loop is taken directly on V_{out}, the R_{IL} / R_L divider bridge does not create inaccuracy on V_{out} as it does with the out-of-the-loop method.

4.2 Application on the TS507

In the case of the TS507, the first pole of the loop gain caused by the feedback occurs around:

$$\frac{1 + \frac{Z_o + R_{IL}}{R_L}}{2\pi\cdot\left[(Z_o + R_{IL})C_L + R_{IL}\left(1 + \frac{Z_o}{R_L}\right)C_{IL}\right]} \approx \frac{1}{2\pi\cdot\ (R_{IL}\,||\,R_L)(C_{IL} + C_L)}$$

The second one occurs at higher frequencies where its impact on stability is limited. The goal of the first pole is to decrease the loop gain to get closer to 0 dB, just before the zero, occurring at

$$\frac{1}{2\pi \cdot R_{IL}C_{IL}}$$

whose goal is to minimize the phase shift caused by the pole. The stability is increased as the loop gain crosses the 0 dB axis with a limited phase shift. It minimizes the effect of the second pole caused by the feedback, which is also pushed toward higher frequencies.

Although this compensation method may seem difficult to set up, it brings very good results, as shown in *Table 3* and *Table 4*, for the TS507 operational amplifier.

Table 3. Results of in-the-loop compensation for different load conditions in the case of a voltage follower configuration for TS507

C_L	R_L = 1 kΩ				R_L = 10 kΩ				R_L = 100 kΩ			
	R_{IL}(1) (Ω)	C_{IL} (nF)	Mg(2) (dB)	Mφ(2) (degree)	R_{IL}(1) (kΩ)	C_{IL} (nF)	Mg(2) (dB)	Mφ(2) (degree)	R_{IL}(1) (kΩ)	C_{IL} (nF)	Mg(2) (dB)	Mφ(2) (degree)
1 nF			-4.1	-28.5 °			-5	-34.1 °			-5.1	-34.4 °
	100	1	4.7	24.5 °	1	0.4	15.2	53.9 °	10	0.2	24.3	71.9 °
10 nF			-22.2	-78.4 °			-22.9	-79.5 °			-23	-79.6 °
	100	2	6	21.9 °	1	1.26	13.6	61.2 °	5	1.26	13.3	79.2 °
100 nF			-34.1	-84.4 °			-34.4	-84.6 °			-34.5	-84.6 °
	79.4	7.9	6.5	34.3 °	0.5	6.3	6.5	66.9 °	0.63	6.3	6.2	70.6 °

1. R_{IL} cells are shaded when its value is clamped to R_L/10.
2. Negative values indicate instability.

Table 4. Results of in-of-the-loop compensation for different load conditions in the case of a gain configuration of either -10 or +11 (R_g = 100 Ω and R_f = 1 kΩ) for TS507

C_L	R_L = 1 kΩ				R_L = 10 kΩ				R_L = 100 kΩ			
	R_{IL}(1) (Ω)	C_{IL} (pF)	Mg(2) (dB)	Mφ(2) (degree)	R_{IL}(1) (Ω)	C_{IL} (pF)	Mg(2) (dB)	Mφ(2) (degree)	R_{IL}(1) (Ω)	C_{IL} (pF)	Mg(2) (dB)	Mφ(2) (degree)
1 nF			17.6	84.7 °			16.8	85.1 °			16.7	85.2 °
	100	126	39	89 °	100	126	39	88.1 °	100	126	39	88 °
10 nF			-0.6	-16.1 °			-1.3	-25.7 °			-1.4	-25.9 °
	39.8	251	40.2	78.6 °	31.6	316	40.3	84.9 °	31.6	316	40.3	84.8 °
100 nF			-13	-69.2 °			-13.3	-69.8 °			-13.3	-69.9 °
	10	631	44.5	66.2 °	10	631	44.5	65.6 °	10	631	44.5	65.6 °

1. R_{IL} cells are shaded when its value is clamped to (R_L// R_f)/10.
2. Negative values indicate instability.

In a gain configuration, when considering the loop gain, the output is loaded by a resistive load of R_L // (R_f + R_g), where R_f and R_g are the resistors used for the gain. If R_f + R_g << R_L, the loop gain and therefore the stability parameters are the same whatever the value of R_L. This is visible in *Table 4* where R_f + R_g = 1.1 kΩ with R_L = 10 kΩ and R_L = 100 kΩ.

Table 5 and *Table 6* help you to choose the best compensation components for different ranges of load capacitors (and with R_L = 10 kΩ) in voltage follower configuration and in a gain configuration of either -10 or +11.

However, each case of load can be improved by choosing specific components (see*Table 3* and *Table 4*).

Table 5. Best compensation components for different load capacitor ranges in voltage follower configuration for TS507 (with R_L = 10 kΩ)

Load capacitor range	R_{IL} (kΩ)	C_{IL} (pF)	Minimum gain margin (dB)	Minimum phase margin (degree)
10 pF to 100 pF	1	251	16.8	54.9 °
100 pF to 1 nF	1	251	15.8	42.1 °
1 nF to 10 nF	1	631	10.9	27 °
10 nF to 100 nF	1	2500	3.8	18.4 °

Table 6. Best compensation components for different load capacitor ranges in a gain configuration of either -10 or +11 (R_g = 100 Ω and R_f = 1 kΩ) for TS507 (with R_L = 10 kΩ)

Load capacitor range	R_{IL} (Ω)	C_{IL} (pF)	Minimum gain margin (dB)	Minimum phase margin (degree)
10 pF to 100 pF	1000	40	39.2	88.8 °
100 pF to 1 nF	39.8	63	37.2	86.8 °
1 nF to 10 nF	63	251	36.5	70.7 °
10 nF to 100 nF	15.8	631	39.1	63.1 °

These tables are very valuable because almost all the follower and gain configuration applications requiring compensation have capacitive loads in the range of 100 pF to 1 nF. Thus, a simple combination of (R_{IL}, C_{IL}), depending on R_L can cover all these cases with a very good stability.

The loop gain shown in *Figure 23*, plotted for a voltage follower configuration with C_L = 1 nF and R_L = 10 kΩ, shows the instability without compensation. This can also be observed in *Figure 24* with the peaking present on closed loop. Both figures show the benefits of compensation.

Figure 23. Loop gain

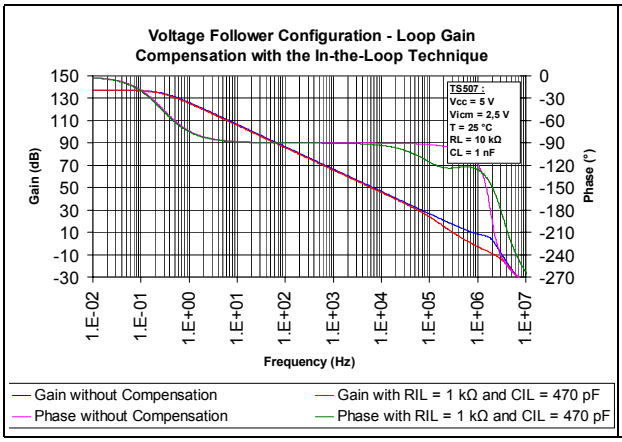
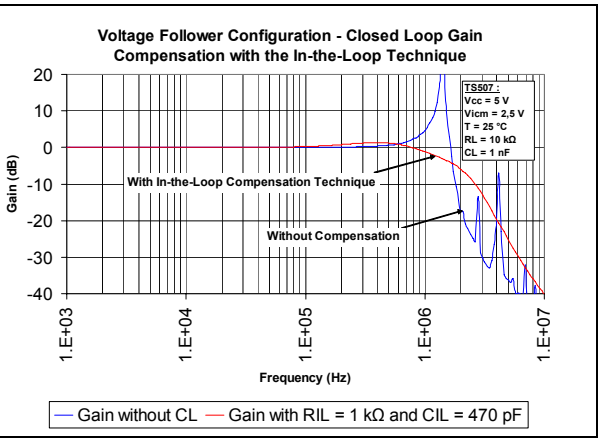


Figure 24. Measured closed loop gain



The zero and the pole introduced by the compensation are visible on the loop gain. Introducing first the pole at

$$\frac{1}{2\pi \cdot (R_{IL} \,||\, R_L)(C_{IL} + C_L)} = 125\text{kHz}$$

leads to a gain fall (with a slope of -40 dB/decade with compensation) which allows to come closer to unity gain. The zero, occurring at

$$\frac{1}{2\pi \cdot R_{IL} C_{IL}} = 400\text{kHz}$$

leads to an upturn of the phase so that when unity gain is reached, the effect of the first poles are limited in terms of phase shifting. Thus, the circuit is stable with a good phase margin. Furthermore, it leads to an excellent gain margin, because the gain keeps falling whereas the phase increases due to the zero, before finally decreasing to reach the -180° point.

5 Snubber network compensation method

5.1 Theoretical overview

Figure 25 shows another way to stabilize an operational amplifier driving a capacitive load. The snubber network compensation method consists in adding an RC series circuit connected between the output and the ground. It is particularly recommended for lower voltage applications, where the full output swing is needed.

Figure 25. Snubber network schematics

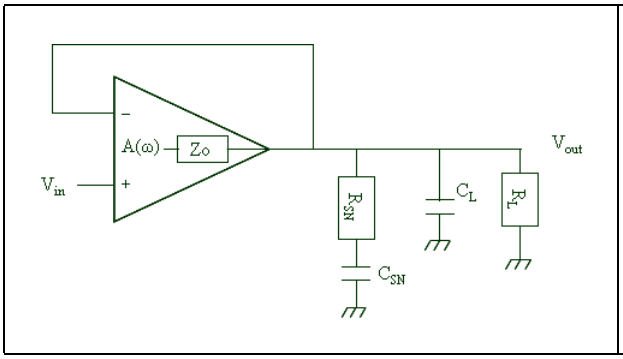
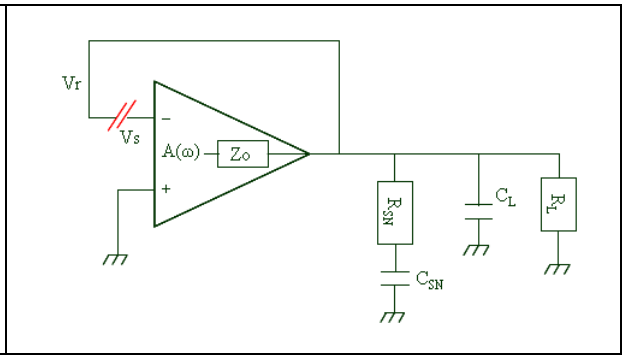


Figure 26. Snubber network equivalent schematics for loop gain calculation



Introducing a second load resistor R_{SN} in the circuit decreases the resistive load, and as a result, pushes the pole caused by the capacitive load to higher frequencies, from

$$\frac{1}{2\pi \cdot (Z_o \,||\, R_L) C_L}$$

to

$$\frac{1}{2\pi \cdot (Z_o \,||\, R_L \,||\, R_{SN}) C_L}$$

Therefore, stability is increased.

Furthermore, adding a serial capacitor C_{SN} with R_{SN} removes the impact of R_{SN} in DC.

On one hand, C_{SN} must be big enough to consider that its impedance is small compared to R_{SN} at the frequencies where R_{SN} plays its role of stabilizer.

On the other hand, R_{SN} being very small, C_{SN} must be small enough because when the frequency increases, the system becomes limited by the current flowing through R_{SN} and C_{SN} (depending on the output voltage swing).

$$I = \frac{V}{R_{SN} - \frac{j}{C_{SN}\omega}}$$

Therefore, in the following examples, C_{SN} is limited to 100 nF in order to not limit the frequency range.

In fact, this compensation introduces a zero and an additional pole into the loop gain:

$$-\frac{V_r}{V_s} = \frac{A(\omega)\cdot\ (1 + jR_{SN}C_{SN}\omega)}{1 + \frac{Z_o}{R_L} + jZ_o(C_L + C_{SN})\omega + j\left(1 + \frac{Z_o}{R_L}\right)R_{SN}C_{SN}\omega - Z_oR_{SN}C_LC_{SN}\omega^2}$$

Because Z_o is not a pure resistance over frequency, choosing the minimum R_{SN} is not always the best case.

5.2 Application on the TS507

For the TS507, according to the abacus in *Figure 27* and *Figure 28*, this compensation method in the case of a voltage follower configuration, works only if the capacitive load is less than 1 nF, in order to obtain (at least) a phase margin of 20°.

Figure 27. Gain margin abacus in a voltage follower configuration

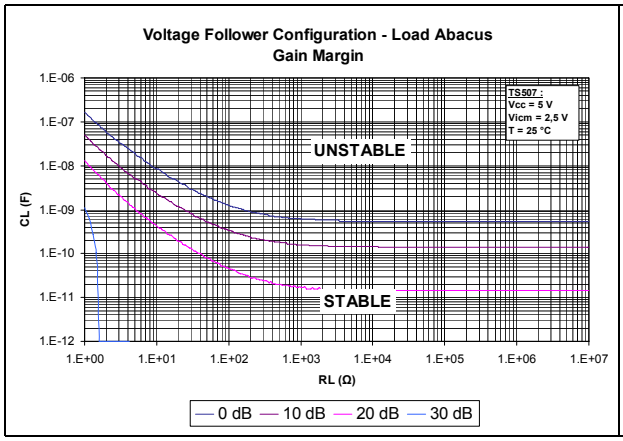


Figure 28. Phase margin abacus in a voltage follower configuration

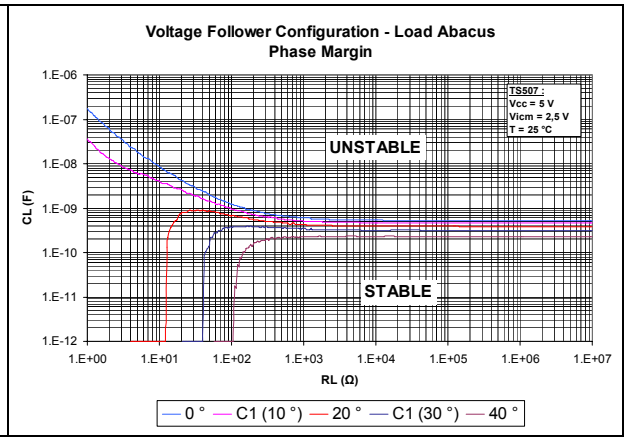


Figure 29 and *Figure 30* are the same abacus in case of a gain of either -10 or +11.

Figure 29. Gain margin abacus in a gain configuration of either -10 or +11

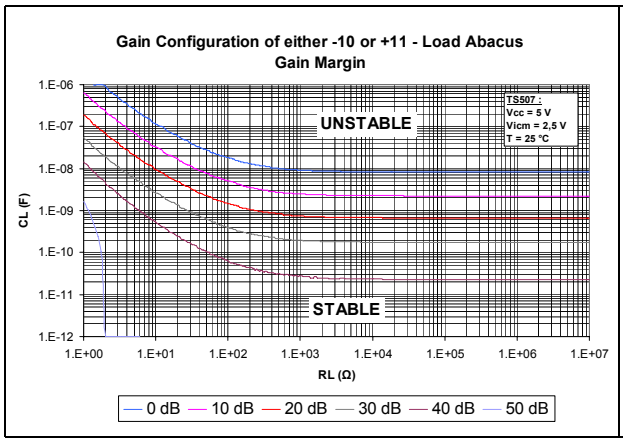
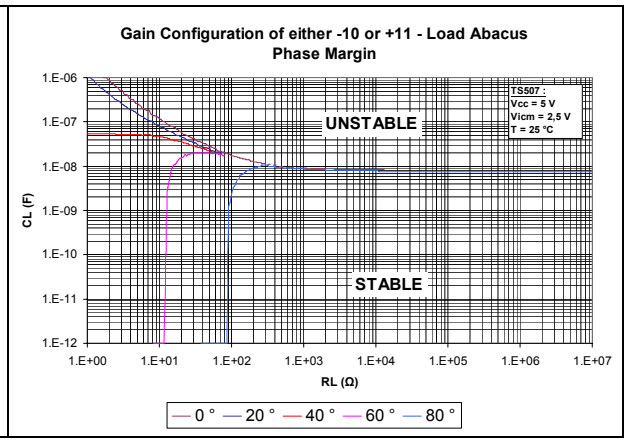


Figure 30. Phase margin abacus in a gain configuration of either -10 or +11



You can use the abacus in *Figure 27* to *Figure 30* to determine the best R_{SN} value.

Table 7 to *Table 10* give the results of compensation in a follower configuration and several gain configurations for different load conditions.

Table 7. Results of snubber network compensation for different load conditions in the case of a voltage follower configuration for TS507

C_L	R_L = 1 kΩ				R_L = 10 kΩ				R_L = 100 kΩ			
	R_{SN} (Ω)	C_{SN} (nF)	$Mg^{(1)}$ (dB)	$M\varphi^{(1)}$ (degree)	R_{SN} (Ω)	C_{SN} (nF)	$Mg^{(1)}$ (dB)	$M\varphi^{(1)}$ (degree)	R_{SN} (Ω)	C_{SN} (nF)	$Mg^{(1)}$ (dB)	$M\varphi^{(1)}$ (degree)
1 nF			-4.1	-28.5 °			-5	-34.1 °			-5.1	-34.4 °
	34.4	100	7.3	16.9 °	31.6	100	7.7	16.7 °	31.6	100	7.7	16.8 °
10 nF			-22.2	-78.4 °			-22.9	-79.5 °			-23	-79.6 °
	13.5	100	-7.5	-18.3 °	13.5	100	-7.6	-18.5 °	13.5	100	-7.7	-18.5 °

1. Negative values indicate instability.

Table 8. Results of snubber network compensation for different load conditions in the case of a gain configuration of either -1 or +2 (R_f = R_g = 1 kΩ) for TS507

C_L	R_L = 1 kΩ				R_L = 10 kΩ				R_L = 100 kΩ			
	R_{SN} (Ω)	C_{SN} (nF)	$Mg^{(1)}$ (dB)	$M\varphi^{(1)}$ (degree)	R_{SN} (Ω)	C_{SN} (nF)	$Mg^{(1)}$ (dB)	$M\varphi^{(1)}$ (degree)	R_{SN} (Ω)	C_{SN} (nF)	$Mg^{(1)}$ (dB)	$M\varphi^{(1)}$ (degree)
1 nF			2.4	28.7 °			1.5	19.4 °			1.4	19.6 °
	58.9	100	11	37.6 °	56.2	100	10.9	37.6 °	58.9	100	10.6	37.6 °
10 nF			-15.7	-70.9 °			-16.5	-72.4 °			-16.6	-72.6 °
	13.5	100	-1.5	-4 °	13.5	100	-1.6	-4.1 °	13.5	100	-1.6	-4.1 °

1. Negative values indicate instability.

Table 9. Results of snubber network compensation for different load conditions in the case of a gain configuration of either -10 or +11 (R_g = 100 Ω and R_f = 1 kΩ) for TS507

C_L	R_L = 1 kΩ				R_L = 10 kΩ				R_L = 100 kΩ			
	R_{SN} (Ω)	C_{SN} (nF)	$Mg^{(1)}$ (dB)	$M\varphi^{(1)}$ (degree)	R_{SN} (Ω)	C_{SN} (nF)	$Mg^{(1)}$ (dB)	$M\varphi^{(1)}$ (degree)	R_{SN} (Ω)	C_{SN} (nF)	$Mg^{(1)}$ (dB)	$M\varphi^{(1)}$ (degree)
1 nF			17.6	84.7 °			16.8	85.1 °			16.7	85.2 °
	171.1	100	21.6	81.8 °	149.1	100	21.6	81.8 °	146.8	100	21.6	81.8 °
10 nF			-0.6	-16.1 °			-1.3	-25.7 °			-1.4	-25.9 °
	26.1	100	11.3	60.6 °	25.7	100	11.3	60.7 °	25.5	100	11.3	60.6 °
100 nF			-13	-69.2 °			-13.3	-69.8 °			-13.3	-69.9 °
	17.1	100	-5.2	-33.3 °	15.3	100	-5.4	-31.7 °	15.3	100	-5.4	-31.7 °

1. Negative values indicate instability.

Table 10. Results of snubber network compensation for different load conditions in the case of a gain configuration of either -30 or +31 (R_g = 100 Ω and R_f = 3 kΩ) for TS507

C_L	R_L = 1 kΩ				R_L = 10 kΩ				R_L = 100 kΩ			
	R_{SN} (Ω)	C_{SN} (nF)	$Mg^{(1)}$ (dB)	$M\varphi^{(1)}$ (degree)	R_{SN} (Ω)	C_{SN} (nF)	$Mg^{(1)}$ (dB)	$M\varphi^{(1)}$ (degree)	R_{SN} (Ω)	C_{SN} (nF)	$Mg^{(1)}$ (dB)	$M\varphi^{(1)}$ (degree)
1 nF			26	88.2 °			25.2	88.4 °			25.1	88.4 °
	198	100	29.8	87.3 °	168.5	100	29.8	87.3 °	166	100	29.8	87.3 °
10 nF			7.9	87.9 °			7.2	88 °			7.1	88.1 °
	30.7	100	19.6	82.7 °	30	100	19.2	82.7 °	30	100	19.1	82.8 °
100 nF			-4.2	-44.7 °			-4.5	-45.8 °			-4.5	-45.9 °
	16.6	100	3.4	63.8 °	16.5	100	3.3	64 °	16.5	100	3.3	64.1 °

1. Negative values indicate instability.

The results are almost the same whether R_L = 1 kΩ, 10 kΩ or 100 kΩ because in all cases, at the frequency range that has a significant impact on stability, the resistive load on the amplifier is R_{SN} // $R_L \cong R_{SN}$.

From *Table 7*, you can see that, in the case of a voltage follower configuration, this compensation method doesn't work for capacitive loads higher than 1 nF because C_{SN} is limited to 100 nF. *Figure 27* and *Figure 28* show that lower R_{SN} values would give better results. However, in this case, these abacus are not valid because the C_{SN} impedance is not negligible compared with R_{SN} at the frequency range where R_{SN} plays its role of stabilizer.

For R_L = 10 kΩ and C_L = 1 nF, the snubber network compensation method gives the loop gain and closed loop gain shown in *Figure 31* and *Figure 32* respectively.

Figure 31. Loop gain

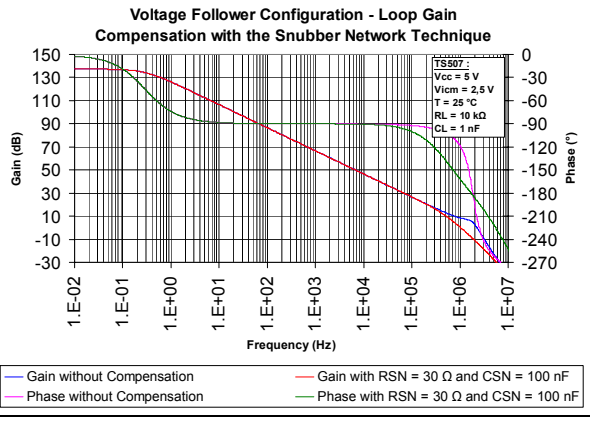
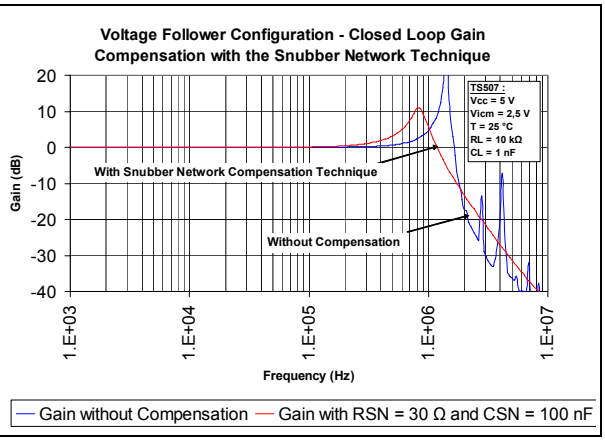


Figure 32. Measured closed loop gain



This compensation has the main advantage of neither reducing the output swing nor the gain accuracy, unlike the two other compensation methods. Nevertheless, for the TS507, this method is limited to capacitive loads lower than 1 nF in a voltage follower configuration, and the stability improvement provided by the compensation is not as good as with the two other methods.

6 Conclusion

Based on the results of the three compensation methods described in this application note, it can be stated that the best compensation solution for the TS507 is the in-the-loop method. This is the most complex solution to understand, but a pick-up table is provided in order to choose the most appropriate components for your application. The main drawback of this compensation method is a limited output swing.

The out-of-the-loop compensation method is easy to implement because it requires only one extra component. The way it works is also easy to understand. However, its main limitation is an inaccuracy on the output voltage because the load is part of a divider bridge. An abacus is provided to choose the component you need for your application.

Finally, the snubber method is easy to understand and use. It does not have the drawbacks of the first two solutions. But because the loop gain cannot be considered as a pure third-order system (because of the open loop gain of the amplifier and the variations of the output impedance which is not purely resistive and constant over frequency), it does not lead to great improvements, and it is limited to load capacitors lower than 1 nF in the case of the TS507. An abacus is also provided for this compensation method. Another drawback of this solution is that it potentially limits the frequency range of the application for large output signals due to a strong current flowing through the compensation elements. This compensation method is not really useful for the TS507.

In conclusion, for the TS507, whenever possible, we recommend to use the in-the-loop compensation method.

Note, that this document provides typical values for the TS507 at ambient temperature. Therefore, your chosen solution must in any case be checked on bench.

7 Revision history

Table 11. Document revision history

Date	Revision	Changes
7-Nov-2007	1	Initial release.

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