INTRODUCTION

Operational amplifiers (op amps) that drive large capacitive loads tend to have peaking and oscillation problems when they are not properly compensated. Other problems include: reduced bandwidth, lower output slew rate, and higher power consumption.

This application note explains why these problems occur, how to modify the op amp circuit for better performance, and how to quickly compute circuit values.

SIMPLIFIED OP AMP MODEL

In order to understand how capacitive loads affect op amps, we must look at the op amp output impedance and bandwidth. Figure 1 shows a simplified AC model of an op amp configured for a non-inverting gain of $G_N$. The open-loop gain is represented by the dependent source with gain $A_{OL}(s)$, where $s = j\omega = j2\pi f$. The output stage is represented by the resistor $R_O$ (open-loop output resistance).

$f_{2P}$ models the open-loop gain’s reduced phase margin (PM < 90°) at high frequencies due to internal parasitics. Both $f_{2P}$ and the capacitive load ($C_L$) reduce the feedback loop’s phase margin.

The op amp feedback loop ($R_F$ and $R_G$) causes its closed-loop behavior to be different from its open-loop behavior. Gain bandwidth product ($f_{GBP}$) and open-loop output impedance ($R_O$) are modified to give closed-loop bandwidth ($f_{3dBA}$) and output impedance ($Z_{OUT}$). We can analyze the circuit in Figure 1 to give:

$$G_N = 1 + \frac{R_F}{R_G}$$

$$f_{3dBA} = \frac{f_{GBP}}{G_N}$$

$$Z_{OUT} = R_O \left( 1 + \frac{A_{OL}(s)}{G_N} \right) \approx R_O \left( \frac{1 + \frac{s}{\omega_{GBP}}}{\frac{1}{G_N}} \right)$$

Figure 2 shows $Z_{OUT}$’s behavior. At low frequencies, it is constant because the open-loop gain is constant. As the open-loop gain decreases with frequency, $Z_{OUT}$ increases. Past $f_{3dBA}$, the feedback loop has no more effect and $Z_{OUT}$ stays at $R_O$. The peaking at $G_N = +1$ is caused by the reduced phase margin due to $f_{2P}$.

We will include gain bandwidth product ($f_{GBP}$), the open-loop gain’s “second pole” ($f_{2P}$) and the non-inverting gain ($G_N$) in our open-loop gain ($A_{OL}(s)$) model. Low frequency effects are left out for simplicity.

$$A_{OL}(s) = \frac{5 \cdot G_N}{\omega_{GBP} - \frac{s}{\omega_{2P}}}$$

Figure 3 shows a simple AC model that approximates this behavior. The amplifier models the no-load gain and bandwidth, while the inductor and resistor model the output impedance vs. frequency.
The equations for $L_{OUT}$ and $R_{OUT}$ are:

$$L_{OUT} = G_N R_G / (2\pi f_{GBP})$$

$$R_{OUT} \approx R_G \left( 1 - \frac{f_{GBP}}{G_N f_{2P}} \right), \quad \frac{f_{GBP}}{G_N} \leq \frac{f_{2P}}{2}$$

Note that $R_{OUT}$ is larger than $R_G$ in order to include $f_{2P}$'s phase shift effects, especially at low gain ($G_N$).

### Capacitive Load Compensation

Our discussion on compensating capacitive loads is divided into three topics. First, we will show the effect of capacitive loads when there is no compensation. Second, we will show a simple compensation method, and how it improves circuit behavior. Last, we will show how to deal with inverting circuits.

#### Capacitive Load

Figure 4 shows a non-inverting gain circuit with an uncompensated capacitive load. For small capacitive loads and high gains (typically $C_L/G_N < 100$ pF), this circuit works quite well.

We recommend setting $R_F + R_G \gg R_{OUT}$ for better circuit performance. A simplified transfer function is:

$$\frac{V_{OUT}}{V_{IN}} \approx G_N \left( 1 + \frac{s}{\omega_p Q_p} + \frac{s^2}{\omega_p^2} \right)$$

where:

$$\omega_p = 2\pi f_p = 1/\sqrt{L_{OUT} C_L}$$

$$Q_p = R_{OUT} \sqrt{C_L / L_{OUT}}, \quad R_F + R_G \gg R_{OUT}$$

Now that we have estimates of $f_p$ and $Q_p$, we can use the equations in Appendix A to estimate bandwidth ($f_{3dB}$), frequency response peaking ($H_{PK/GN}$) and step response overshoot (%ovrsht). Note that $f_{3dB}$ is not the same as the op amp's no load -3dB bandwidth ($f_{3dBA}$).

### MCP6271 Example (Uncompensated)

The equations above were used to generate the curves in Figure 5 and Figure 6 for Microchip's MCP6271 op amp. The parameters used are (see Appendix B):

- $f_{GBP} = 2.0$ MHz, $f_{2P} = 4.5$ MHz and $R_G = 360$ Ω. As can be seen at $G_N = +1$ V/V and $C_L = 100$ pF, the response peaks enough to be a concern.

#### Figure 5: Estimate of MCP6271's AC response with $G_N = +1$.

#### Figure 6: Estimate of MCP6271's AC response with $G_N = +10$.
The peaking (HPK/GN) should be near 0 dB for the best overall performance. Keeping the peaking below 3 dB usually gives enough design margin for changes in op amp, resistor and capacitor parameters over temperature and process. However, the performance is degraded.

For this example, our formulas give the estimated results shown in Table 1. As C_L increases and gain decreases, there is more peaking.

TABLE 1: RESPONSE ESTIMATES

| Compensated Capacitive Load | The simplest compensation method for capacitive loads is shown in Figure 7. The resistor R_{ISO} is used to load down the LC resonant circuit, which reduces frequency response peaking. As can be seen, R_{ISO} does not change the DC response or power and only costs one additional resistor.

```
TABLE 1: RESPONSE ESTIMATES

<table>
<thead>
<tr>
<th>Circuit</th>
<th>Response</th>
</tr>
</thead>
<tbody>
<tr>
<td>G_N (V/V)</td>
<td>C_L (F)</td>
</tr>
<tr>
<td>----------</td>
<td>---------</td>
</tr>
<tr>
<td>1.0</td>
<td>10p</td>
</tr>
<tr>
<td>100p</td>
<td></td>
</tr>
<tr>
<td>1n</td>
<td>940k</td>
</tr>
<tr>
<td>10.0</td>
<td>100p</td>
</tr>
<tr>
<td>1n</td>
<td>300k</td>
</tr>
<tr>
<td>10n</td>
<td>94k</td>
</tr>
</tbody>
</table>
```

With these equations, we can now find a reasonable R_{ISO} value. When Q_P = 1/\sqrt{2}, the response has the highest possible bandwidth without peaking, and the equations are in their simplest form.

\[
Q_P = 1/\sqrt{2} \approx 0.707
\]

\[
R_{ISO} = 0, \quad C_L \leq L_{OUT}/(2R_{OUT}^2)
\]

\[
R_{ISO} = L_{OUT}/R_{OUT}C_L \sqrt{2R_{OUT}^2C_L - L_{OUT}}, \quad C_L > L_{OUT}/2R_{OUT}^2
\]

MCP6271 Example (Compensated)

These equations were used to compensate the MCP6271 circuits represented in Figure 5 and Figure 6. The same parameters were assumed, with the only change being the addition of R_{ISO}. The improved results are shown in Figure 8 and Figure 9. Table 2 shows much better results than Table 1.

![FIGURE 7: Compensated capacitive load.](image)

**FIGURE 7:** Compensated capacitive load.

The simplified op amp AC model produces the following transfer function, where, again, we require R_F + R_G >> R_{OUT}:

\[
\frac{V_{OUT}}{V_{IN}} = G_N \left(1 + \frac{s}{\omega_P Q_P} + \frac{s^2}{\omega_P^2}\right)
\]

where:

\[
\omega_P = 2\pi f_P = 1/\sqrt{L_{OUT}C_L (1 + R_{ISO}/R_{OUT})}
\]

\[
Q_P = 1/\left(\omega_P \sqrt{L_{OUT}/R_{OUT} + R_{ISO}C_L}\right) \quad R_F + R_G >> R_{OUT}
\]

![FIGURE 8: Estimate of MCP6271's compensated AC response with G = +1.](image)

**FIGURE 8:** Estimate of MCP6271's compensated AC response with G = +1.

![FIGURE 9: Estimate of MCP6271's compensated AC response with G = +10.](image)

**FIGURE 9:** Estimate of MCP6271's compensated AC response with G = +10.
TABLE 2: RESPONSE ESTIMATES.

<table>
<thead>
<tr>
<th>Circuit</th>
<th>Response</th>
</tr>
</thead>
<tbody>
<tr>
<td>G_N (V/V)</td>
<td>C_L (F)</td>
</tr>
<tr>
<td>1.0</td>
<td>10p</td>
</tr>
<tr>
<td>1M</td>
<td>100p</td>
</tr>
<tr>
<td>1n</td>
<td>1</td>
</tr>
<tr>
<td>10.0</td>
<td>100p</td>
</tr>
<tr>
<td>1n</td>
<td>1</td>
</tr>
</tbody>
</table>

Figure 10 shows the R_ISO values for the MCP6271 estimated by the above equations. It is shown versus normalized load capacitance (C_L/G_N) for ease of interpretation. Measured data for one representative part is shown in Figure 11.

The main difference between Figure 10 and Figure 11 is at G_N = +1. The reduced phase margin at low G_N (caused by f_2P) requires additional compensation at low C_L. The simplified equations in this application note give reasonable estimates in this condition, but are not exact.

When large capacitive loads give lower op amp bandwidth than desired, refer to Microchip’s line of Power MOSFET Drivers at www.microchip.com.

Inverting Gains

Inverting gain circuits (see Figure 12) are compensated in the same way as non-inverting gain circuits. Since the inverting input of the op amp is at virtual ground, the load presented to the output by the feedback network is now R_F instead of R_F + R_G. Thus, we need to set R_F >> R_OUT. Use the noise gain:

\[ G_N = 1 + \frac{R_F}{R_G} \]

in the previous equations, even though the inverting gain is -R_E/R_G. For example, an inverting gain of -1 V/V gives G_N = +2. The reasons for this behavior come from op amp feedback theory [1, 3].

FIGURE 12: Compensated inverting gain circuit.

SLEW RATE

In Figure 7 and Figure 12, the op amp will produce an output current (I_OUT) that goes into C_L. This current cannot exceed the op amp’s output short circuit current (I_SC). This current limit causes the output slew rate to be limited (SR_CL). Note that SR_CL is independent of the op amp’s internally-set slew rate (SR). We can derive SR_CL as follows.

\[ \frac{dV_{OUT}(t)}{dt} = \frac{I_{OUT}(t)}{C_L} \]

\[ SR_CL = \max \left( \frac{dV_{OUT}(t)}{dt} \right) = \frac{I_{SC}}{C_L} \]

where:

SR_CL is in units of V/s

Slew Rate and Sine Waves

Sine waves with edge rates faster than SR_CL or SR will cause signal distortion problems. The sine wave

\[ V_{OUT}(t) = V_M \sin(2\pi ft) \]
Slew Rate and Square Waves

Square waves with fast edges can also cause problems with capacitive loads. The maximum edge rate of a square wave with a (10% to 90%) rise time of \( t_r \) and a peak-to-peak voltage of \( V_{PP} \) can be approximated as:

\[
\max \left( \frac{dV_{OUT}(t)}{dt} \right) = \frac{0.8V_{PP}}{t_r}
\]

Thus, we need to keep

\[
0.8V_{PP}/t_r < \min(SR_{CL}, SR)
\]

One solution to this problem is to use square waves with lower edge rates (higher \( t_r \)). Filtering the square waves (lowpass filter bandwidth < 0.35/\( t_r \)) is another approach. Using slower logic gates may be a solution in some cases. It is also possible to add \( R_{ISO} \), as shown in Figure 7 and Figure 12. The maximum current occurs when the ideal output just reaches the new level and \( V_{OUT}(t) \) is still slew rate limited. To keep \( I_{OUT} < I_{SC} \), we need:

\[
R_{ISO} > \frac{V_{PP} - (t_r/0.8)\min(SR_{CL}, SR)}{I_{SC}}
\]

Using \( R_{ISO} \) will both slow the edges down and change the shape of the transitions.

When large capacitive loads cause a lower slew rate than desired, refer to Microchip’s line of Power MOSFET Drivers at www.microchip.com.

Square Wave Example

Let’s use the MCP6271 with \( G = +1 \) V/V and \( C_L = 100 \) nF. In Appendix B, we find \( SR = 0.9 \) V/µs and \( I_{SC} = 25 \) mA. We can then calculate:

\[
SR_{CL} = 0.25V/\mu s
\]

which is significantly slower than SR. With a maximum voltage swing of 5.0V_{PP}, we need an input signal with a rise time > 16 µs.

Filtering the input square wave at the input of the op amp would require a bandwidth less than 22 kHz.

If we use \( R_{ISO} \) to limit the output current, (with a maximum voltage swing of 5.0V_{PP} and an input rise time of 10 µs), we need \( R_{ISO} > 75\Omega \). Setting \( R_{ISO} = 100\Omega \) gives:

\[
Q_P = 0.18
\]

\[
f_{3dB} = 16 \text{ kHz}
\]
Note that if we used the $R_{ISO}$ value for response peaking elimination (24.0Ω), we would achieve a wider small signal bandwidth (92 kHz), but would need to keep $V_{PP} < 3.7V_{PP}$ to avoid output current limiting and reduced rise and fall times.

**POWER DISSIPATION**

It is well known that reactive elements (ideal capacitors and inductors) do not dissipate power. However, an op amp driving a reactive load will dissipate power. This happens because load current in the output stage always flows in a direction that dissipates power. The output transistors rectify the load current.

Figure 7 and Figure 12 show the circuits under discussion. There will be no DC load current because $C_L$ blocks DC. At low frequencies, $I_Q$ (op amp’s quiescent current) and $C_L$ will dominate the output current behavior. At high frequencies, $R_{ISO}$ will dominate.

Given an output voltage of

$$V_{OUT}(t) = V_M \sin(2\pi ft)$$

it can be shown that the average power dissipated by the op amp at low frequencies is:

$$P_{OA} = (V_{DD} - V_{SS})(I_Q + 2V_M f C_L) \frac{I}{2\pi R_{ISO} C_L}$$

The power dissipation increases with frequency because $C_L$ dominates the load.

At high frequencies, the average power dissipated by the op amp becomes constant because $R_{ISO}$ dominates:

$$P_{OA} = (V_{DD} - V_{SS})\left(I_Q + \frac{V_M}{\pi R_{ISO}}\right) - \frac{V_M^2}{R_{ISO}}$$

$$f \approx \frac{I}{2\pi R_{ISO} C_L}$$

In the frequency range where neither $C_L$ or $R_{ISO}$ dominates the load ($f \approx 1/(2\pi R_{ISO} C_L)$), a somewhat conservative estimate of $P_{OA}$ is the minimum value from the two formulas above.

**DESIGN VERIFICATION**

We recommend that you always verify the performance of your circuit design with SPICE simulations and by breadboarding it on the bench. SPICE macro models of Microchip’s op amps are available on the Microchip web site at www.microchip.com for your convenience.

**SUMMARY**

We have seen that op amps that drive large capacitive loads tend to show peaking and oscillation, reduced bandwidth, lower output slew rate, and higher power consumption. Adding one resistor to the circuit can greatly improve the performance. The resulting bandwidth is a little under the no load bandwidth.

Simple formulas were given that allow a circuit designer to quickly evaluate the impact of capacitive loads. The fix is easy to implement and understand.

Designs that need to drive large capacitors at high bandwidth or rise time may benefit from using Microchip’s line of Power MOSFET Drivers.
APPENDIX A: RESPONSE MODEL

In this application note, we have seen transfer functions of the form:

\[ \frac{V_{OUT}}{V_{IN}} \approx K \left(1 + \frac{s}{\omega_p Q_p} + \frac{s^2}{\omega_p^2} \right) \]

This is a 2nd order, low-pass response, which models the op amp circuits in this application note reasonably well. We will show some simple formulas for sine wave and step responses which help evaluate the performance of the circuits in this application note [2,4].

Given \( f_p (\omega_p = 2\pi f_p) \) and \( Q_p \), we can calculate the bandwidth \( f_{3dB} \), peak response frequency \( f_{PK} \) and gain peaking \( H_{PK}/G_N \) for a sine wave as follows:

\[
f_{3dB} = \frac{f_p Q_p}{\sqrt{1 - Q_p^2 + \left(\frac{1 - Q_p^2}{2}\right)^2}}; \quad Q_p < 0.7
\]

\[
f_{3dB} = f_p \left[1 - \frac{1}{2 Q_p^2} + \left(\frac{1 - 1}{2 Q_p^2}\right)^2 + 1\right]; \quad Q_p \geq 0.7
\]

\[
f_{PK} = 0, \quad Q_p \leq 1 / \sqrt{2}
\]

\[
f_{PK} = f_p \left[1 - \frac{1}{2 Q_p^2}\right]; \quad Q_p > 1 / \sqrt{2}
\]

\[
H_{PK} = 1, \quad Q_p \leq 1 / \sqrt{2}
\]

\[
H_{PK} = \frac{Q_p}{\sqrt{4 Q_p^2 - 1}}, \quad Q_p > 1 / \sqrt{2}
\]

The step response overshoot \( \%\text{ovrsht} \) and rise time \( t_r \) are calculated as follows:

\[
\%\text{ovrsht} = 0\%, \quad Q_p \leq 1/2
\]

\[
\%\text{ovrsht} = (100\%)e^{-\pi / \sqrt{4 Q_p^2 - 1}}, \quad Q_p > 1/2
\]

\[
t_r = 0.35 / f_{3dB}
\]

It is relatively simple to extract \( K, f_p \) and \( Q_p \) from frequency response simulations or measurements.

- \( K \) is the gain at low frequencies \( (f << f_{3dB}) \)
- \( f_p \) is the frequency where the phase is -90°
- \( |V_{OUT}/V_{IN}| \) at \( f_p \) is \( KQ_p \) (in units of \( V/V \))

APPENDIX B: MICROCHIP OP AMPS

The performance parameters of some Microchip op amps shown in Table B-1 below are typical and were extracted from the parts’ data sheets. These data sheets contain the officially-supported specifications, and can be found on Microchip’s website at www.microchip.com. This data is current as of August, 2003.

<table>
<thead>
<tr>
<th>Part</th>
<th>( f_{GBP} ) (Hz)</th>
<th>( f_{2P} ) (Hz)</th>
<th>SR (V/( \mu )s)</th>
<th>( R_O ) (( \Omega ))</th>
<th>( I_{SC} ) (mA)</th>
</tr>
</thead>
<tbody>
<tr>
<td>MCP6041</td>
<td>14k</td>
<td>45k</td>
<td>0.0030</td>
<td>37k</td>
<td>21</td>
</tr>
<tr>
<td>TC1034</td>
<td>60k (Note 1)</td>
<td>1.1M</td>
<td>0.035</td>
<td>15k</td>
<td>8</td>
</tr>
<tr>
<td>MCP6141</td>
<td>100k</td>
<td>55k</td>
<td>0.024</td>
<td>28k</td>
<td>21</td>
</tr>
<tr>
<td>MCP606</td>
<td>155k</td>
<td>620k</td>
<td>0.080</td>
<td>4.2k</td>
<td>17</td>
</tr>
<tr>
<td>MCP616</td>
<td>190k</td>
<td>1.1M</td>
<td>0.080</td>
<td>5.0k</td>
<td>17</td>
</tr>
<tr>
<td>MCP6001</td>
<td>1.0M</td>
<td>45M</td>
<td>0.6</td>
<td>780</td>
<td>23</td>
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<td>MCP6271</td>
<td>2.0M</td>
<td>4.5M</td>
<td>0.9</td>
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<td>25</td>
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<tr>
<td>MCP6021</td>
<td>2.8M</td>
<td>20M</td>
<td>2.3</td>
<td>680</td>
<td>20</td>
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<tr>
<td>MCP6281</td>
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<tr>
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<td>124M</td>
<td>7.0</td>
<td>110</td>
<td>30</td>
</tr>
</tbody>
</table>

**Note 1:** These parameters also apply to the TC1026, TC1029, TC1030 and TC1035.

2: \( f_{2P} \) can be estimated from the Open-Loop Gain plots in the data sheet. Estimate the frequency \( (f_{-135}) \) where the Open-Loop Phase is -135° (i.e., the phase margin is 45°). Adjust for the typical capacitive load used in the measurements \( (C_{Ltyp}) \):

\[ \phi_{CLtyp} = \tan(2\pi f_{-135} R_O C_{Ltyp}) \]

\[ f_{2P} \approx f_{-135}/\tan(45° - \phi_{CLtyp}) \]

\[ f_{2P} \approx 12 f_{-135}, \quad \phi_{CLtyp} > 40° \]
APPENDIX C: REFERENCES


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