

# **Fully Differential Amplifiers Applications: Line Termination, Driving High-Speed ADCs, and Differential Transmission Lines**

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## ABSTRACT

In high-speed systems, proper line termination requires considering the termination resistors and adjusting the gain-setting resistors to maintain symmetrical feedback. Integrated, fully differential amplifiers are well suited for driving differential ADC inputs. They provide an easy means for anti-alias filtering and for setting the common-mode voltage.

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## 1 Introduction

The August 2000 issue of *Analog Applications Journal* introduced the fully differential amplifiers from Texas Instruments and illustrated their basic operation. The November 2000 issue delved into the topic more deeply, by analyzing gain and noise. This issue investigates some typical applications, such as transmission lines and driving ADC inputs.

To simplify calculations and formulas, assume that the amplifier is used at frequencies where the open-loop gain is very large ( $A_F \gg 1$ ) and will not include its effects in the analysis.

The circuit analysis assumes that symmetrical feedback is being used ( $\beta_1 = \beta_2$ ). Before going into the application circuits, this document will detour briefly into how termination affects the feedback factor, and how to account for it.

## 2 Terminating the Input Source

Double termination is typically used in high-speed systems to reduce transmission-line reflections. With double termination, the transmission line is terminated with the same impedance as the source. Common values are 50  $\Omega$ , 75  $\Omega$ , 100  $\Omega$ , and 600  $\Omega$ . When the source is differential, the termination is placed across the line. When the source is single-ended, the termination is placed from the line to ground.

Figure 1 shows an example of terminating a differential signal source. The situation depicted is balanced so that half of  $V_S$  and half of  $R_S$  is attributed to each input, with  $V_{IC}$  being the center point.  $R_S$  is the source impedance and  $R_t$  is the termination resistor. The circuit is balanced, but there are two issues to resolve: (1) proper termination and (2) gain setting.

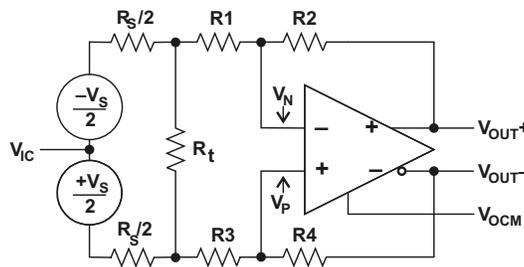


Figure 1. Terminating a Differential Input Signal

As long as  $A_F \gg 1$  and the amplifier is in linear operation, the action of the amplifier keeps  $V_N \approx V_P$ . Thus, to first-order approximation, a virtual short is seen between the two nodes as shown in Figure 2. The termination impedance is the parallel combination:  $R_t \parallel (R_1 + R_3)$ . The value of  $R_t$  for proper termination is calculated as shown in Figure 2.

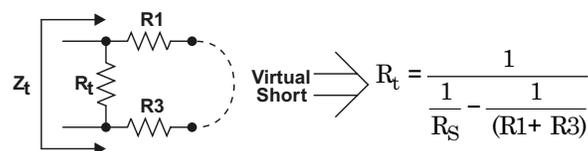


Figure 2. Differential Termination Impedance

Once  $R_t$  is found, the required gain is found by generating a Thevenin equivalent circuit. The circuit is broken between  $R_t$  and the amplifier input resistors  $R_1$  and  $R_3$ .  $V_{IC}$  is not a concern at this point, so leave it out. Then:

$$V_{TH} = V_S \times R_t / (R_t + R_S) \tag{1}$$

and  $R_{TH} = R_S \parallel R_t$  (half is attributed to each side). The resulting Thevenin equivalent is shown in Figure 3.

The proper gain is calculated by:

$$\frac{V_{OUT}}{V_{TH}} = \frac{R_F}{R_G + \frac{R_S || R_t}{2}}$$

where

- where  $V_{OUT} = (V_{OUT+}) - (V_{OUT-})$ . (2)

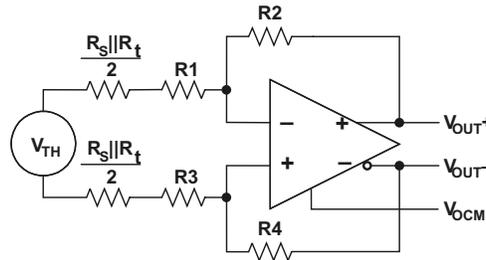


Figure 3. Differential Thevenin Equivalent

Substituting for  $V_{TH}$ , this becomes:

$$\frac{V_{OUT}}{V_S} = \frac{R_F}{R_G + \frac{R_S || R_t}{2}} \times \frac{R_t}{R_S + R_t}$$

where

- $R_F$  is the feedback resistor ( $R_2$  or  $R_4$ )
  - $R_G$  is the input resistor ( $R_1$  or  $R_3$ )
- (3)

Remember: for symmetry, keep the gain equal on the two sides with  $R_2 = R_4$  and  $R_1 = R_3$ .

As an example, suppose you are terminating a 50-Ω differential source that is balanced, and you want an overall gain of 1 from the source to the differential output of the amplifier. Start the design by first choosing the values for  $R_1$  and  $R_3$ , then calculate  $R_t$  and the feedback resistors.

With the voltage divider formed by the termination, assume that a gain of about 2 will be required in the amplifier. Also, feedback resistor values of approximately 500 Ω are reasonable for a high-speed amplifier. Using these starting assumptions, choose  $R_1$  and  $R_3$  equal to 249 Ω. Next calculate  $R_t$  from the formula:

$$R_t = \frac{1}{\frac{1}{R_S} - \frac{1}{(R_1 + R_3)}} = \frac{1}{\frac{1}{50} - \frac{1}{(249 + 249)}} = 55.6 \Omega$$

(4)

(the closest standard 1% value is 56.2 Ω). Then set the gain by calculating the value of the feedback resistors:

$$R_F = \frac{V_{OUT}}{V_S} \times \left( R_G + \frac{R_S || R_t}{2} \right) \times \frac{R_S + R_t}{R_t}$$

$$= 1 \times \left( 249 + \frac{50 || 56.2}{2} \right) \times \frac{50 + 56.2}{56.2} = 495.5 \Omega$$

(5)

(the closest standard 1% value is 499 Ω). The solution is shown in Figure 4 with standard 1% resistor values.

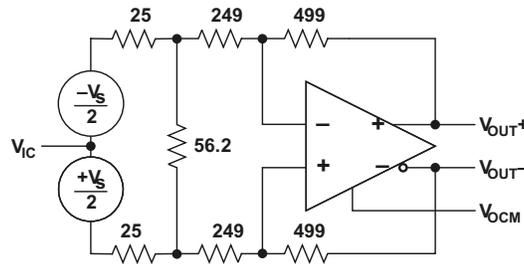


Figure 4. Differential Solution for Gain = 1

Figure 5 shows an example of terminating a single-ended signal source.  $R_S$  is the source impedance, and  $R_t$  is the termination resistor. The circuit is not balanced, so there are three issues to resolve:

- Proper termination
- Gain setting
- Balance

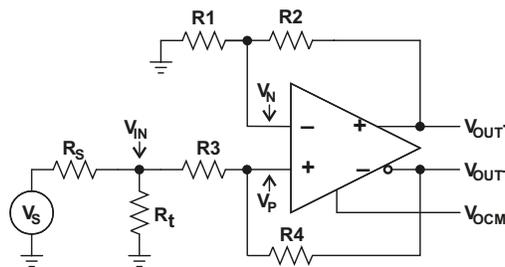


Figure 5. Terminating a Single-Ended Input Signal

To determine the termination impedance seen from the line looking into the amplifier's input at  $V_{IN}$ , remove  $V_S$  and  $R_S$  and short all other sources. As long as  $A_F \gg 1$  and the amplifier is in linear operation, the action of the amplifier keeps  $V_N \approx V_P$ .  $V_N$  will see the voltage at  $V_{OUT+}$  divided by the resistor ratio  $R1 / (R1 + R2)$ . Assuming that the amplifier is balanced,

$$V_{OUT+} = K \times V_{IN} / 2$$

where

- $K$  is the closed-loop gain of the amplifier ( $V_{OCM} = 0$ )

The termination impedance is the parallel combination:  $R_t$  in parallel with:

$$\frac{V_{IN}}{I_{R3}} = \frac{R3}{1 - \frac{K}{2 \times (1+K)}} \tag{7}$$

The value of  $R_t$  for proper termination is then calculated as shown in Figure 6.

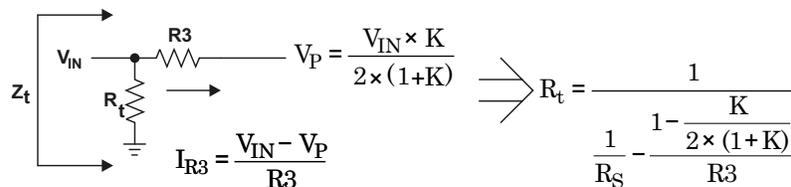


Figure 6. Single-Ended Termination AC Impedance

Once  $R_t$  is found, the required gain is found by generating a Thevenin equivalent circuit. The circuit is broken between  $R_t$  and the amplifier's input resistor  $R_3$ .

$$V_{TH} = V_S \times \frac{R_t}{R_t + R_S} \tag{8}$$

and  $R_{TH} = R_S \parallel R_t$ .

The resulting Thevenin equivalent is shown in [Figure 7](#). The gain is set by:

$$V_{OUT}/V_{TH} = R_F/R_G$$

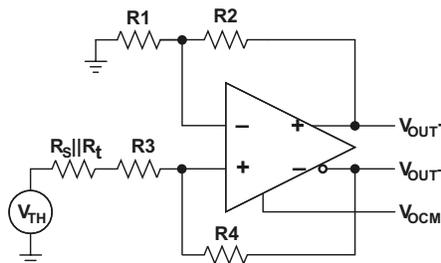
where

- $R_F = R_2 = R_4$
  - $R_G = R_1 = R_3 + R_S \parallel R_t$
  - $V_{OUT} = (V_{OUT+}) - (V_{OUT-})$
- (9)

Substituting for  $V_{TH}$ , this becomes:

$$\frac{V_{OUT}}{V_S} = \frac{R_F}{R_G} \times \frac{R_t}{R_S + R_t} \tag{10}$$

Remember, for symmetry:  $R_2 = R_4$  and  $R_1 = R_3 + (R_S \parallel R_t)$ .



**Figure 7. Single-Ended Thevenin Equivalent**

As an example, suppose you are terminating a 50-Ω single-ended source and want an overall gain of 1 from the source to the differential output of the amplifier. Start the design by first choosing the value for  $R_3$ , then calculate  $R_t$  and the feedback resistors. This will be an iterative process, starting with some initial assumptions that are then refined.

Start with the assumptions that  $R_t = 50 \Omega$  and that a gain of 2 will be required in the amplifier. Also, feedback resistor values of approximately 500 Ω are reasonable for a high-speed amplifier. Using these starting assumptions, choose  $R_1 = 249 \Omega$  and  $R_3 = R_1 - R_S \parallel R_t = 249 \Omega - 25 \Omega = 224 \Omega$ . Next calculate  $R_t$  from the formula:

$$R_t = \frac{1}{\frac{1}{R_S} - \frac{1 - \frac{K}{2(1+K)}}{R_3}} = \frac{1}{\frac{1}{50} - \frac{1 - \frac{2}{2(1+2)}}{224}} = 58.7 \tag{11}$$

Then calculate the value of the feedback resistors:

$$R_2 = \frac{V_{OUT}}{V_S} \times R_1 \times \frac{R_S + R_t}{R_t} = 1 \times 249 \times \frac{50 + 58.7}{58.7} = 460.9 \Omega,$$

$$\begin{aligned} \text{and } R_4 &= \frac{V_{OUT}}{V_S} \times (R_3 + R_S \parallel R_t) \times \frac{R_S + R_t}{R_t} \\ &= 1 \times (224 + 50 \parallel 58.7) \times \frac{50 + 58.7}{58.7} = 464.7 \Omega \end{aligned} \tag{12}$$

The process is iterative because the gain is not 2, but rather  $460.9/249 = 1.85$ ; and  $R_t$  is calculated to be  $58.7 \Omega$ , not  $50 \Omega$ . Iterating through the calculations two more times results in:  $R_3 = 221.9 \Omega$  (the closest standard 1% value is  $221 \Omega$ ),  $R_t = 59.0$  (which is a standard 1% value), and  $R_2 = R_4 = 460.9$  (the closest standard 1% value is  $464 \Omega$ ). The solution is shown in Figure 8 with standard 1% resistor values.

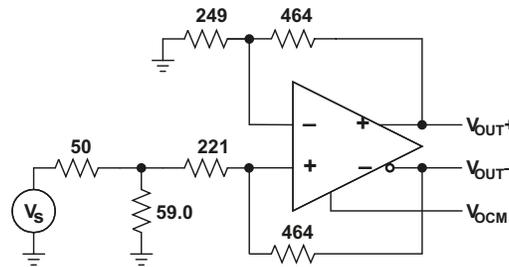


Figure 8. Single-Ended Solution for Gain = 1

Using a spreadsheet makes the iterative process very simple. Also, component values can be easily adjusted to find a better fit to the standard available values.

### 3 Active Anti-Alias Filtering

A major application for fully differential amplifiers is lowpass anti-alias filters for ADCs with differential inputs.

Creating an active first-order low-pass filter is easily accomplished by adding capacitors in the feedback, as shown in Figure 9. With balanced feedback, the transfer function is:

$$\frac{V_{OUT}}{V_{IN}} = \frac{R_F}{R_G} \times \frac{1}{1 + j2\pi f (R_F C_F)}$$

where

- $V_{OUT} = (V_{OUT+}) - (V_{OUT-})$
- $V_{IN} = (V_{IN+}) - (V_{IN-})$

(13)

The pole created in the transfer function is a real pole on the negative real axis in the s-plane.

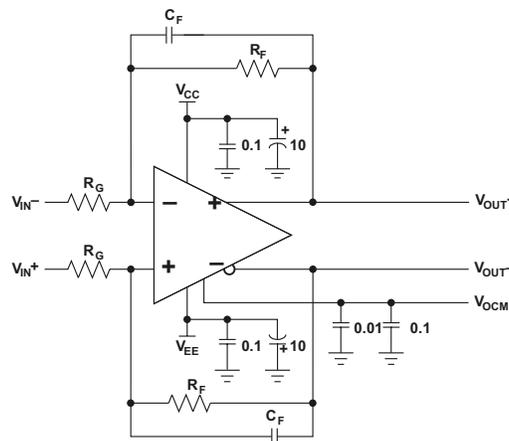


Figure 9. First-Order Active Low-Pass Filter

To create a two-pole low-pass filter, a passive real pole can be created by placing  $R_O$  and  $C_O$  in the output, as shown in Figure 10. With balanced feedback, the transfer function is:

$$\frac{V_{OUT}}{V_{IN}} = \frac{R_F}{R_G} \times \frac{1}{1 + j2\pi f (R_F C_F)} \times \frac{1}{1 + j2\pi f \frac{1}{2} R_O C_O}$$

where

- $V_{OUT} = (V_{OUT+}) - (V_{OUT-})$
- $V_{IN} = (V_{IN+}) - (V_{IN-})$

(14)

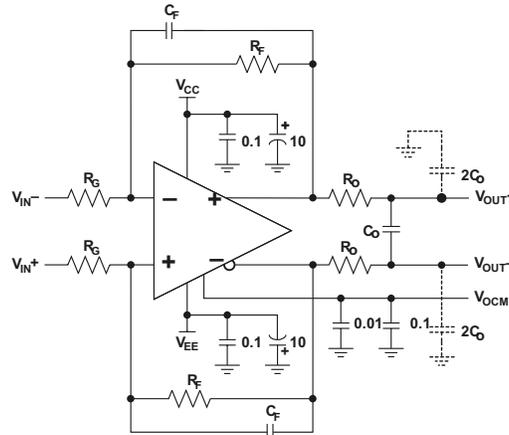


Figure 10. First-Order Active Low-Pass Filter With Passive Second Pole

The second pole created in the transfer function is also a real pole on the negative real axis in the s-plane. The capacitor  $C_O$  can be placed differentially across the outputs as shown in solid lines; or two capacitors (of twice the value) can be placed between each output and ground as shown in dashed lines. Typically,  $R_O$  will be a low value; and, at frequencies above the pole frequency, the series combination with  $C_O$  will load the amplifier. The extra loading will cause extra distortion in the amplifier's output. To avoid this, you might stagger the poles so that the  $R_O C_O$  pole is placed at a higher frequency than the  $R_F C_F$  pole.

The classic filter types like Butterworth, Bessel, Chebyshev, and so forth (second-order and greater), cannot be realized by real poles—they require complex poles. The multiple feedback (MFB) topology creates a complex pole pair, and is easily adapted to fully differential amplifiers, as shown in Figure 11. A third-order filter is formed by adding  $R_4$ s and  $C_3$  at the output.

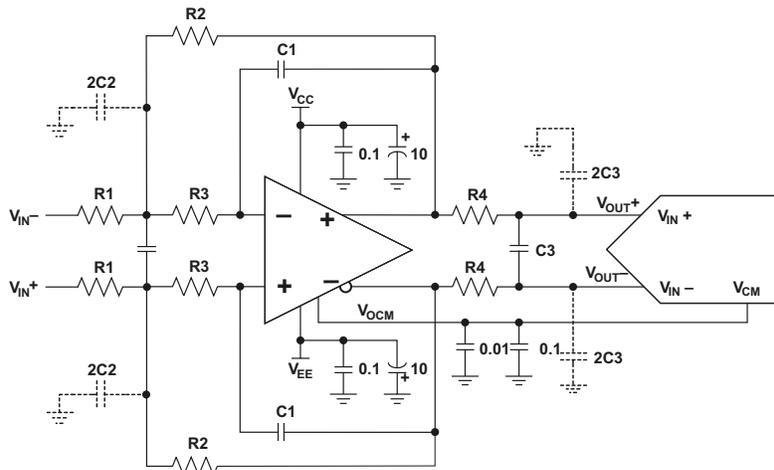


Figure 11. Third-Order Low-Pass Filter Driving an ADC

Capacitors C2 and C3 can be placed differentially across the inputs and outputs, as shown in solid lines. Alternatively, for better common-mode noise rejection, two capacitors of twice the value can be placed between each input or output and ground, as shown in dashed lines.

The transfer function for this filter circuit is:

$$\frac{V_{OUT}}{V_{IN}} = \left[ \frac{K}{-\left(\frac{f}{FSF \times f_C}\right)^2 + \frac{1}{Q} \frac{jf}{FSF \times f_C} + 1} \right] \times \frac{1}{+ j2\pi f \times 2 \times R4C3}$$

where

- $V_{OUT} = (V_{OUT+}) - (V_{OUT-})$
  - $V_{IN} = (V_{IN+}) - (V_{IN-})$
- (15)

$$K = \frac{R2}{R1}, \quad FSF \times f_C = \frac{1}{2\pi\sqrt{2 \times R2R3C1C2}}$$

$$\text{and } Q = \frac{\sqrt{2 \times R2R3C1C2}}{R3C1 + R2C1 + KR3C1} \quad (16)$$

K sets the pass-band gain,  $f_C$  is the cut-off frequency of the filter, FSF is a frequency scaling factor, and Q is the quality factor.

$$FSF = \sqrt{Re^2 + |Im|^2} \quad \text{and } Q = \frac{\sqrt{Re^2 + |Im|^2}}{2Re}$$

where

- Re is the real part of the complex pole pair
  - Im is the imaginary part
- (17)

Setting  $R2 = R$ ,  $R3 = mR$ ,  $C1 = C$ , and  $C2 = nC$  results in:

$$FSF \times f_C = \frac{1}{2\pi RC \sqrt{2 \times mn}} \quad \text{and}$$

$$Q = \frac{\sqrt{2 \times mn}}{1 + m(1 - K)} \quad (18)$$

Start by determining the ratios, m and n, required for the gain and Q of the filter type being designed, then select C and calculate R for the desired  $f_C$ .

R4 and C3 are chosen to set the real pole in a third-order filter. Exercise care when setting this pole. Typically, R4 will be a low value; and, at frequencies above the pole frequency, the series combination with C3 loads the amplifier. The extra loading causes extra distortion in the amplifier output. To avoid this, place the real pole at a higher frequency than the cut-off frequency of the complex pole pair.

**Figure 12** shows the gain and phase response of a second-order Butterworth low-pass filter, with corner frequency set at 1 MHz and the real pole set by R4 and C3 at 15.9 MHz. The components used are:  $R1 = 787 \, \Omega$ ,  $R2 = 787 \, \Omega$ ,  $R3 = 732 \, \Omega$ ,  $R4 = 50 \, \Omega$ ,  $C1 = 100 \, \text{pF}$ ,  $C2 = 220 \, \text{pF}$ ,  $C3 = 100 \, \text{pF}$ , and the THS4141 fully differential amplifier. At higher frequencies, parasitic elements allow the signal to feed through.

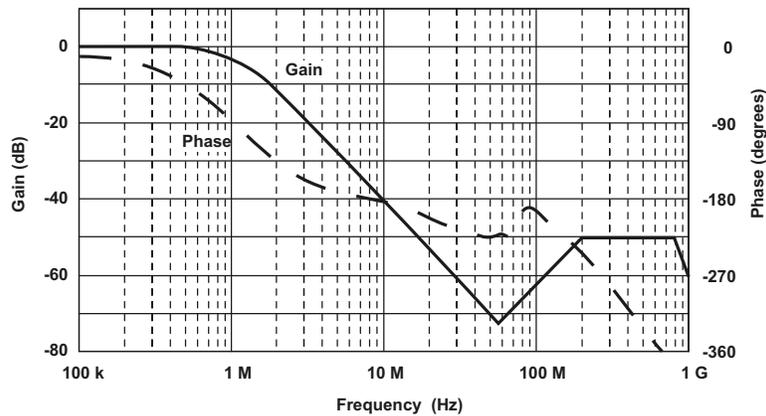


Figure 12. 1-MHz, Second-Order Butterworth Low-Pass Filter With Real Pole at 15.9 MHz

#### 4 $V_{OCM}$

The proper  $V_{OCM}$  is provided as an output by many ADCs with differential inputs. Typically, all that needs to be done is to provide bypass capacitors; 0.1  $\mu\text{F}$  and 0.01  $\mu\text{F}$  are useful choices. If  $V_{OCM}$  is not provided, it can be created by forming a summing node with the plus and minus reference voltages of the ADC to drive  $V_{OCM}$ , as shown in Figure 13. The voltage at the summing node is the midpoint value between  $+V_{REF}$  and  $-V_{REF}$ . Depending on the loading of the  $V_{OCM}$  input, the summing node voltage may need to be buffered.

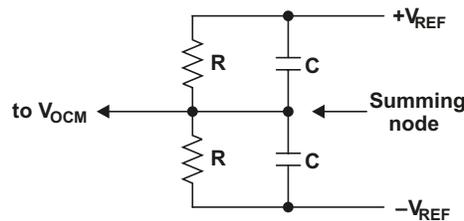


Figure 13. Driving  $V_{OCM}$  from ADC Reference Voltage

#### 5 Power-Supply Bypass

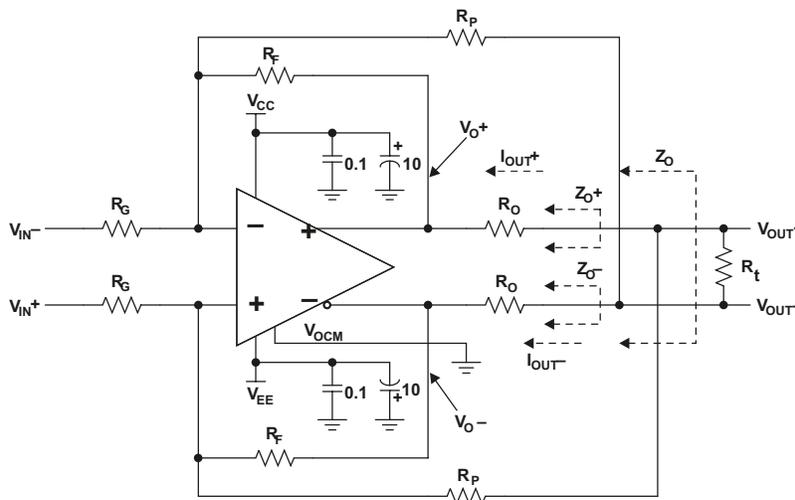
Each power rail should have 6.8- $\mu\text{F}$  to 10- $\mu\text{F}$  tantalum capacitors located within a few inches of the amplifier, to provide low-frequency power-supply bypassing. A 0.01- $\mu\text{F}$  to 0.1- $\mu\text{F}$  ceramic capacitor should be placed within 0.1 inch of each power pin on the amplifier to provide high-frequency power-supply bypassing.

#### 6 Layout Considerations

As with all high-speed amplifiers, minimize parasitic capacitance at the input of the amplifier by removing the ground plane near the pins and near any circuit traces. Also, make trace routing as direct as possible, and use surface-mount components.

#### 7 Using Positive Feedback to Provide Active Termination

Driving transmission lines differentially is a typical use for fully differential amplifiers. Using positive feedback with amplifiers can provide active termination, as shown in Figure 14. Because of the positive feedback, the output line impedance appears larger than the value of output resistor  $R_O$ . The voltage dropped across the resistor depends on its actual value, resulting in increased efficiency.


**Figure 14. Using Positive Feedback to Provide Active Termination**

Use symmetrical feedback with this application.

With double termination, the output impedance of the amplifier,  $Z_O$ , equals the characteristic impedance of the transmission line; and the far end of the line will be terminated with the same value resistor, that is,  $R_t = Z_O$ . For proper balance, half of  $Z_O$  is placed in each half of the differential output, so that  $Z_O = 2 \times Z_{O\pm}$ .

To calculate the output impedance, ground the inputs and insert either a voltage or current source between  $V_{OUT+}$  and  $V_{OUT-}$ .

Due to symmetry,  $Z_{O+} = Z_{O-}$ ,  $V_{OUT+} = -(V_{OUT-})$ , and  $V_{O+} = -(V_{O-})$ . Calculating the impedance of one side provides the solution.

$$Z_{O+} = \left( \frac{V_{OUT+}}{I_{OUT+}} \right) \parallel R_P \parallel I_{OUT+} = \frac{(V_{OUT+}) - (V_{O+})}{R_O},$$

$$\text{and } V_{O+} = (V_{OUT-}) \times \left( \frac{-R_F}{R_P} \right) \quad (19)$$

The output impedance of the amplifier on each side of the line is  $R_O$  divided by 1 minus the gain from the opposite line:

$$Z_{O\pm} = \left( \frac{R_O}{1 - \frac{R_F}{R_P}} \right) \parallel R_P \quad (20)$$

The positive feedback also affects the forward gain. Accounting for this effect and the voltage divider between  $R_O$  and  $R_t \parallel 2R_P$ , the gain from  $V_{IN} = (V_{IN+}) - (V_{IN-})$  to  $V_{OUT} = (V_{OUT+}) - (V_{OUT-})$  is:

$$A = \frac{V_{OUT}}{V_{IN}} = \frac{R_F}{R_G} \times \frac{1}{\frac{2R_O + R_t \parallel 2R_P - R_F}{R_t \parallel 2R_P} - \frac{R_F}{R_P}} \quad (21)$$

Design is easily accomplished if by first choosing the value of  $R_F$  and  $R_O$ . Then, calculate the required value of  $R_P$  to give the desired  $Z_O$ . Then calculate  $R_G$  for the required gain.

For example: if you want a gain of 1, and to terminate a 100-Ω line properly with  $R_F = 1 \text{ k}\Omega$  and  $R_O = 10 \text{ }\Omega$ , the proper value for  $Z_O$  and  $R_t$  is 100 Ω ( $Z_{O\pm} = 50 \text{ }\Omega$ ). Rearranging [Equation 20](#) yields:

$$R_P = \frac{R_F - R_O}{1 - \frac{R_O}{Z_{O\pm}}} = \frac{990 \text{ }\Omega}{1 - \frac{10 \text{ }\Omega}{50 \text{ }\Omega}} = 1.24 \text{ k}\Omega \quad (22)$$

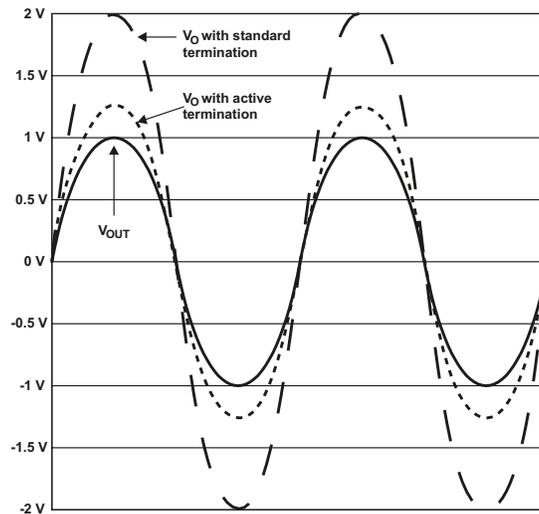
Then, rearranging Equation 21 gives:

$$R_G = \frac{R_F}{A} \times \frac{1}{\frac{2R_O + R_t}{R_t} \left| \frac{2R_P - R_F}{R_P} \right.}$$

$$= \frac{1 \text{ k}\Omega}{\frac{20\Omega + 100\Omega}{100\Omega} \left| \frac{2.48 \text{ k}\Omega}{1.25 \text{ k}\Omega} \right.} = 2.49 \text{ k}\Omega \quad (23)$$

The circuit is built and tested with the nearest standard values to those previously computed:

$R_F = 1 \text{ k}\Omega$ ,  $R_P = 1.24 \text{ k}\Omega$ ,  $R_G = 2.49 \text{ k}\Omega$ ,  $R_t = 100 \Omega$ , and  $R_O = 10 \Omega$ . Compare the output voltage waveforms ( $V_{OUT} = 2V_{PP}$ ) with active termination and standard termination shown in Figure 15: [ $V_O = (V_{O+}) - (V_{O-})$ , and  $V_{OUT} = (V_{OUT+}) - (V_{OUT-})$ ].



**Figure 15. Output Waveforms With Active and Standard Termination**

For standard termination,  $R_F = 1 \text{ k}\Omega$ ,  $R_P = \text{open}$ ,  $R_G = 499 \Omega$ ,  $R_t = 100 \Omega$ , and  $R_O = 50 \Omega$ .

With standard termination, 20 mW of power is dissipated in the output resistors, as opposed to 6.25 mW with active termination, which wastes 69% less power.

Another attractive feature about active termination, especially in low-voltage applications, is the effective increase in output-voltage swing for a given supply voltage.

## 8 Conclusion

In high-speed systems, proper line termination requires considering the termination resistors and adjusting the gain-setting resistors to maintain symmetrical feedback.

Integrated, fully differential amplifiers are well suited for driving differential ADC inputs. They provide an easy means for anti-alias filtering and for setting the common-mode voltage.

Integrated, fully differential amplifiers are also well suited for driving differential transmission lines, and active termination provides for increased efficiency.

## 9 Related Web Sites

[www-s.ti.com/sc/techlit/slyt018](http://www-s.ti.com/sc/techlit/slyt018)

[amplifier.ti.com](http://amplifier.ti.com)

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