Operational Transconductance Amplifiers in PWM ICs: Grounded vs. Negative Feedback

AND90312/D

Introduction:

Operational transconductance amplifiers (OTA) are often used in PWM ICs instead of voltage error amplifiers (EA) for the voltage feedback loop. OTAs are simpler to implement, typically offer lower power consumption, and require less die area compared to voltage EAs. It has been shown [Ref. 1, 4] that the compensation network can also be connected as a negative feedback connection to the output voltage divider. This article will review Type−II feedback expressions for both grounded and negative feedback implementations to describe benefits/trade−offs.

Background

Grounded compensation OTA networks are well documented [Ref. 2]. ICs requiring external compensation have an internal protection ESD structure and parasitic capacitance (R_{esd},C_0) located between the package pinout and the internal feedback OTA control signal (Figure 1) which are not normally mentioned in datasheets. Detailed grounded compensation OTA expressions describing the feedback control signal are presented in Ref. 3.

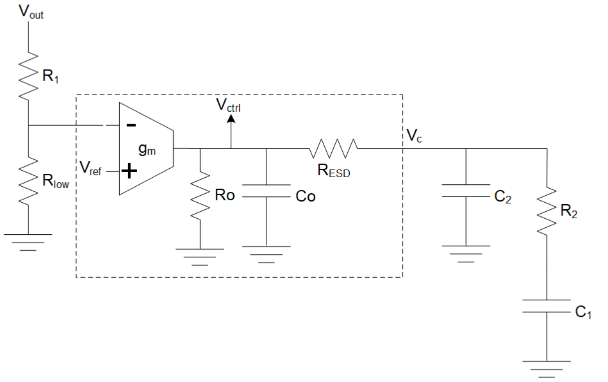

Figure 1. Type−II OTA with Grounded Compensation Network

Output parasitic capacitance C_0 introduces a high frequency pole. R_{esd} introduces an upper mid−band frequency zero. For low R_2 compensation values, ($R_2 <$ ~$5 \cdot R_{esd}$), use of R_2 to set the mid−band gain is limited and feedback loop measurements (OTA response, control−to−output response) begin introducing measurement errors as signals V_{ctrl} (the control loop signal) and V_c (the external IC measurement point) begin to diverge. Open−loop response in closed−loop form remains unchanged.

The ESD resistance introduces a limitation on compensation values and/or performance tradeoffs in applications such as automotive battery connected boost and SEPIC topologies.

When a voltage error amplifier feedback network is connected using a negative feedback configuration, the positive terminal is connected to a voltage reference, rendering the negative terminal as a virtual ground in the gain small−signal transfer function.

When an OTA compensation is configured as negative feedback (Figure 2), the negative terminal virtual ground is lost. Voltage feedback divider resistor R_{low} is present in the error amplifier AC gain transfer function and attenuates mid−band gain. Larger R_2 values are used when targeting similar AC transfer function responses as their grounded feedback counterparts. This overcomes design limitations from the Figure 1 grounded feedback configuration.

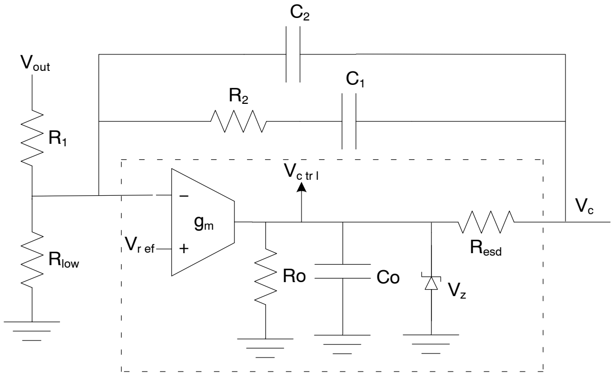

Figure 2. Type−II OTA with Negative Feedback Configuration

Type−II Expressions for a Ground Referenced Compensation Network

Type−II ground referenced compensation network OTA expressions shown in Figure 1 were developed in Ref. 3. The OTA gain may be expressed as a third−order polynomial Equation 1.

$$G_{gnd}(s) = -G_0 \frac{1 + a_{1\_gnd}s + a_{2\_gnd}s^2}{1 + b_{1\_gnd}s + b_{2\_gnd}s^2 + b_{3\_gnd}s^3} \qquad \text{(eq. 1)}$$

Coefficients may be expressed as

$$\begin{aligned} a_{1\_gnd} &= \left(R_2 + R_{esd}\right)C_1 + R_{esd}C_2 \qquad \text{(eq. 2)} \\ &\approx \left(R_2 + R_{esd}\right)C_1 \end{aligned}$$

$$a_{2\_gnd} = R_2R_{esd}C_1C_2 \qquad \text{(eq. 3)}$$

$$\begin{aligned} b_{1\_gnd} &= R_0C_0 + \left(R_0 + R_2 + R_{esd}\right)C_1 + \left(R_0 + R_{esd}\right)C_2 \qquad \text{(eq. 4)} \\ &\approx \left(R_0 + R_2 + R_{esd}\right)C_1 + \left(R_0 + R_{esd}\right)C_2 \end{aligned}$$

$$\begin{aligned} b_{2\_gnd} &= R_0\left(R_2 + R_{esd}\right)C_0C_1 + R_0R_{esd}C_0C_2 + R_2\left(R_0 + R_{esd}\right)C_1C_2 \qquad \text{(eq. 5)} \\ &\approx R_0\left(R_2 + R_{esd}\right)C_0C_1 + R_2\left(R_0 + R_{esd}\right)C_1C_2 \end{aligned}$$

$$b_{3\_gnd} = R_0R_{esd}R_2C_0C_1C_2 \qquad \text{(eq. 6)}$$

For well separated poles and zeros, the Equation 1 quality factor Q is << 1 and may be simplified as cascaded 1st−order terms as in Equation 7.

$$G_{gnd}(s) \approx -G_0 \frac{\left(1 + \frac{s}{\omega_{z1\_gnd}}\right)\left(1 + \frac{s}{\omega_{z2\_gnd}}\right)}{\left(1 + \frac{s}{\omega_{p1\_gnd}}\right)\left(1 + \frac{s}{\omega_{p2\_gnd}}\right)\left(1 + \frac{s}{\omega_{p3\_gnd}}\right)} \qquad \text{(eq. 7)}$$

where,

$$G_0 = \frac{R_{low}}{R_{low} + R_1} g_m R_0 \qquad \text{(eq. 8)}$$

$$\omega_{z1\_gnd} = \frac{1}{a_{1\_gnd}} = \frac{1}{\left(R_2 + R_{esd}\right)C_1} \qquad \text{(eq. 9)}$$

$$\omega_{z2\_gnd} = \frac{a_{1\_gnd}}{a_{2\_gnd}} = \frac{1}{\left(R_2 \| R_{esd}\right)C_2} \qquad \text{(eq. 10)}$$

$$\omega_{p1\_gnd} = \frac{1}{b_{1\_gnd}} = \frac{1}{\left(R_0 + R_2 + R_{esd}\right)C_1 + \left(R_0 + R_{esd}\right)C_2} \qquad \text{(eq. 11)}$$

$$\omega_{p2\_gnd} = \frac{b_{1\_gnd}}{b_{2\_gnd}} = \frac{\left(R_0 + R_2 + R_{esd}\right)C_1 + \left(R_0 + R_{esd}\right)C_2}{R_0\left(R_2 + R_{esd}\right)C_0C_1 + R_2\left(R_0 + R_{esd}\right)C_1C_2} \qquad \text{(eq. 12)}$$

$$\begin{aligned} \omega_{p3\_gnd} &= \frac{b_{2\_gnd}}{b_{3\_gnd}} = \frac{R_0\left(R_2 + R_{esd}\right)C_0C_1 + R_2\left(R_0 + R_{esd}\right)C_1C_2}{R_0R_{esd}R_2C_0C_1C_2} \qquad \text{(eq. 13)} \\ &\approx \frac{R_2\left(R_0 + R_{esd}\right)}{R_0R_{esd}R_2C_0} \end{aligned}$$

To determine the compensation value expressions, we must first express the magnitude of $G_{gnd}(s)$ at a selected crossover frequency f_c.

$$G_{fc\_gnd} \approx -G_0 \frac{\sqrt{1+\left(\frac{f_{z1}}{f_c}\right)^2}}{\sqrt{1+\left(\frac{f_c}{f_{p1}}\right)^2}} \qquad \text{(eq. 14)}$$

$$= -\frac{R_{low}}{R_{low}+R_1} g_m R_0 \frac{\omega_{p1\_gnd}}{\omega_{z1\_gnd}}$$

From Equations 8, 9, and 10 we now determine the compensation network component values.

$$R_2 = \frac{R_{low}+R_1}{R_{low}} \frac{\left|G_{fc_{gnd}}\right|}{g_m} - R_{esd} \qquad \text{(eq. 15)}$$

$$C_1 = \frac{1}{\omega_{z1_{gnd}}\left(R_2 + R_{esd}\right)} \qquad \text{(eq. 16)}$$

$$C_2 = \frac{1}{\omega_{z2_{gnd}}\left(R_2 \| R_{esd}\right)} \qquad \text{(eq. 17)}$$

To validate the similarity of the exact and simplified expressions, desired pole/zero values are provided in Table 1. Component parametric values are from **onsemi** PWM IC NCV8871 [Ref. 5]. Table 2 lists the selected compensation component values. The resulting OTA transfer function is plotted in Figure 3.

Table 1. DESIRED POLE AND ZEROES

g_m	0.00107 Ω^{-1}	NCV8871
R_0	3 MΩ	NCV8871
C_0	10 pF	NCV8871
R_{esd}	542 Ω	NCV8871
f_{z1}	350 Hz	
f_{p2}	37kHz	

Table 2. COMPONENT VALUES FOR GROUNDED COMPENSATION NETWORK

	Calculated	Selected
R_1		53.6 kΩ
R_{low}		2.61 kΩ
C_1	46.98 nF	47 nF
R_2	8.958 kΩ	9.09 kΩ
C_2	470.2 pF	470 pF
f_{z2}	650 kHz	N/A
f_{p3}	19 MHz	N/A

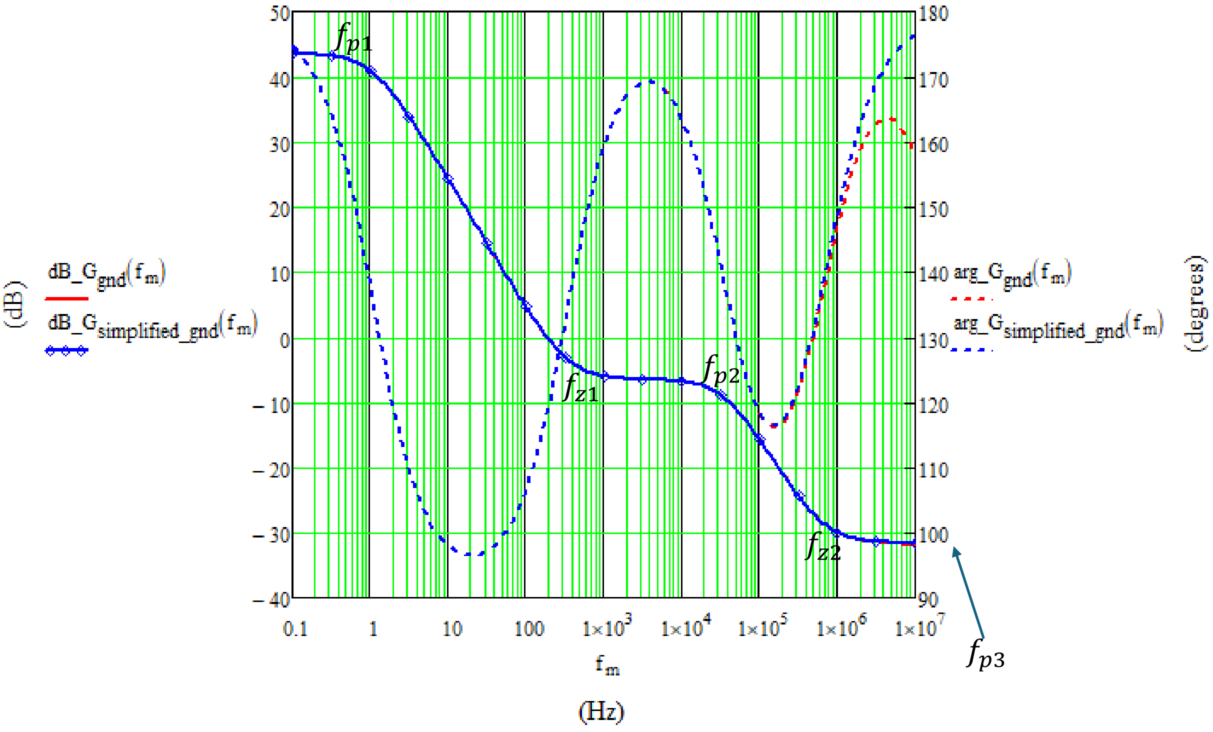

Figure 3. Ground Referenced OTA Transfer Function

The simplified transfer function matches the exact expression up to ~1 Mhz, well beyond the maximum switching frequency of the NCV8871 family (340 kHz). The expressions would remain accurate for a similar NCV898031 2 MHz device [Ref. 6].

Type−II Expressions for a Negative Feedback Compensation Network

OTAs with Type−II negative feedback compensation network do not have a virtual ground because of the g_m voltage-dependent current source. The Figure 2 transfer function may be expressed as a third−order polynomial transfer function in Equation 18.

$$G_{neg}(s) = -G_0 \frac{\left(1 + \frac{s}{\omega_{z1\_neg}}\right)\left(1 - \frac{s}{\omega_{z2\_neg}}\right)}{1 + b_{1\_neg}s + b_{2\_neg}s^2 + b_{3\_neg}s^3} \qquad \text{(eq. 18)}$$

The multiplier preceding s/ω_{z2_neg} is negative because the term is a right−half−plane zero.

$$\omega_{z1\_neg} = \frac{1}{\left(R_2 + R_{esd} - \frac{1}{g_m}\right)C_1} \qquad \text{(eq. 19)}$$

$$\omega_{z2\_neg} = \frac{1}{\left(R_{esd} - \frac{1}{g_m}\right)C_2} \qquad \text{(eq. 20)}$$

$$\begin{aligned} b_{1_{neg}} &= \left(R_0 + R_2 + R_{esd} + (1 + g_mR_0)\left(R_1 \parallel R_{low}\right)\right)C_1 + \left(R_0 + R_{esd} + (1 + g_mR_0)\left(R_1 \parallel R_{low}\right)\right)C_2 + R_0C_0 \\ &\approx \left(R_0 + \left(1 + g_m\left(R_1 \parallel R_{low}\right)\right)\right)\left(C_1 + C_2\right) \end{aligned} \qquad \text{(eq. 21)}$$

$$R_A = R_0 + R_2 + R_{esd} + (1 + g_mR_0)\left(R_1 \parallel R_{low}\right) \qquad \text{(eq. 22)}$$

$$R_B = R_0 + R_{esd} + (1 + g_mR_0)\left(R_1 \parallel R_{low}\right) \qquad \text{(eq. 23)}$$

$$R_C = \frac{R_2 + R_{esd} + \left(R_1 \parallel R_{low}\right)}{1 + g_m\left(R_1 \parallel R_{low}\right)} \qquad \text{(eq. 24)}$$

$$R_D = \frac{R_{esd} + \left(R_1 \parallel R_{low}\right)}{1 + g_m\left(R_1 \parallel R_{low}\right)} \qquad \text{(eq. 25)}$$

$$\begin{aligned} b_{2\_neg} &= R_AC_1\left(\left(R_B \parallel R_2\right)C_2 + \left(R_C \parallel R_0\right)C_0\right) + R_BC_2\left(R_D \parallel R_0\right)C_0 \\ &\approx \left(R_0 + \left(1 + g_m\left(R_1 \parallel R_{low}\right)\right)\right)R_2C_1C_2 \end{aligned} \qquad \text{(eq. 26)}$$

$$\begin{aligned} b_{3\_neg} &= R_A\left(R_B \parallel R_2\right)\left(R_D \parallel R_0\right)C_0C_1C_2 \\ &\approx R_0\left\{1 + g_m\left(R_1 \parallel R_{low}\right)\right)R_2\left(\frac{R_{esd} + \left(R_1 \parallel R_{low}\right)}{1 + g_m\left(R_1 \parallel R_{low}\right)}\right)C_0C_1C_2\right. \end{aligned} \qquad \text{(eq. 27)}$$

For well separated poles, $G_{neg}(s)$ may be simplified.

$$G_{neg}(s) \approx -G_0 \frac{\left(1 + \frac{s}{\omega_{z1\_neg}}\right)\left(1 - \frac{s}{\omega_{z2\_neg}}\right)}{\left(1 + \frac{s}{\omega_{p1\_neg}}\right)\left(1 + \frac{s}{\omega_{p2\_neg}}\right)\left(1 + \frac{s}{\omega_{p3\_neg}}\right)} \qquad \text{(eq. 28)}$$

where,

$$\omega_{p1\_neg} = \frac{1}{b_{1\_neg}} = \frac{1}{\left(R_0 + \left(1 + g_m\left(R_1 \parallel R_{low}\right)\right)\right)\left(C_1 + C_2\right)} \qquad \text{(eq. 29)}$$

$$\omega_{p2\_neg} = \frac{b_{1\_neg}}{b_{2\_neg}} = \frac{1}{R_2\left(C_1 \parallel C_2\right)} \qquad \text{(eq. 30)}$$

$$\omega_{p3\_neg} = \frac{b_{2\_neg}}{b_{3\_neg}} = \frac{1 + g_m\left(R_1 \parallel R_{low}\right)}{\left(R_{esd} + \left(R_1 \parallel R_{low}\right)\right)C_0} \qquad \text{(eq. 31)}$$

To establish compensation component expressions, we must first express the magnitude of $G_{neg}(s)$ at the same crossover frequency f_c as that from the grounded compensation network on page 2.

$$G_{fc\_neg} = G_{fc\_gnd} \qquad \text{(eq. 32)}$$

From Equations 29, 30 and 31,

$$R_2 = R_0\left(1 + g_m\left(R_1 \parallel R_{low}\right)\right)\frac{\omega_{p1\_neg}}{\omega_{z1\_neg}} - R_{esd} + \frac{1}{g_m} \qquad \text{(eq. 33)}$$

$$C_1 = \frac{1}{\omega_{p1\_neg}R_0\left(1 + g_m\left(R_1 \parallel R_{low}\right)\right)} \qquad \text{(eq. 34)}$$

$$C_2 = \frac{1}{R_2\omega_{p1\_neg}} \qquad \text{(eq. 35)}$$

To verify the similarity between the exact and simplified expressions, we use the same pole/zero values from that of Table 1. The calculated transfer function compensation values (C1, R2, C2) are summarized in Table 3 and plotted in Figure 4.

Table 3. COMPONENT VALUES FOR GROUNDED COMPENSATION NETWORK

	Calculated	Selected
R_1		53.6 kΩ
R_{low}		2.61 kΩ
C_1	12.2 nF	12 nF
R_2	37.8 kΩ	38.3 kΩ
C_2	114 pF	120 pF
f_{z2}	3.4 MHz	N/A
f_{p3}	19 MHz	N/A

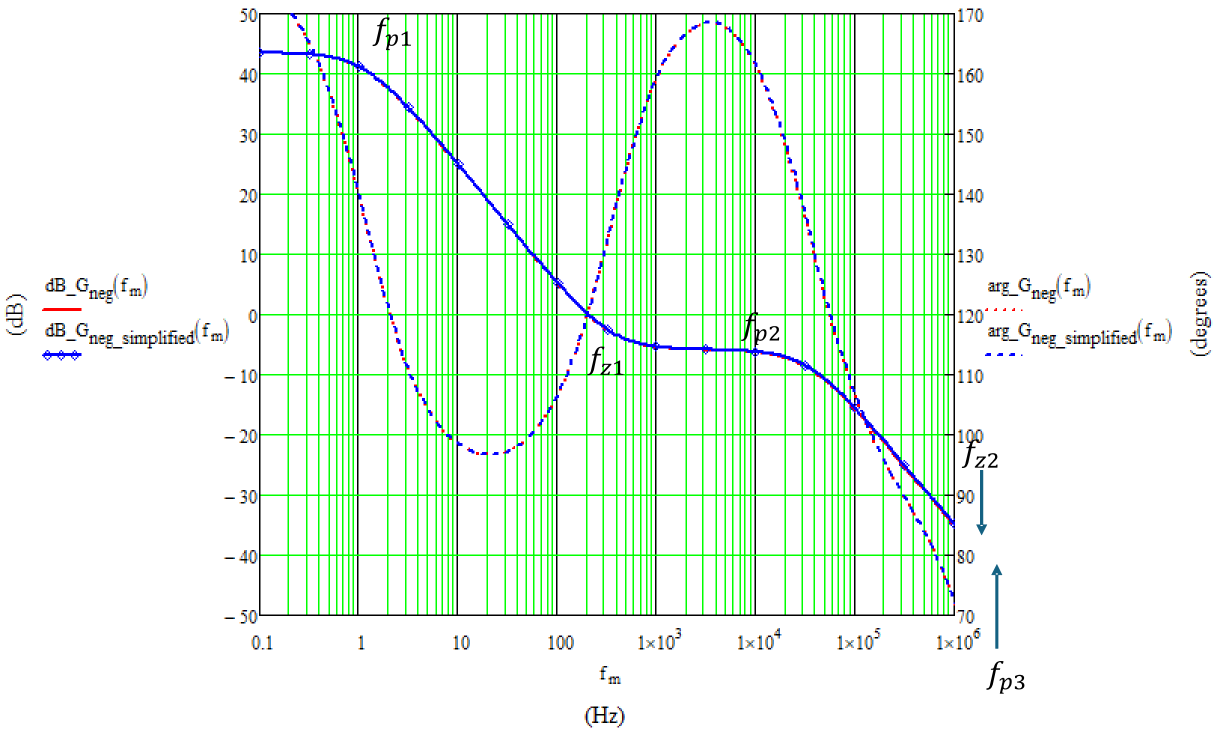

Figure 4. Negative Feedback OTA Transfer Function

Both simplified and exact transfer functions match in a useful range well beyond the maximum switching frequency of 2 MHz controllers.

Comparison of Grounded vs. Negative Feedback OTA Compensation

Figure 5 overlays OTA compensation for grounded and negative feedback compensation networks for examples from prior section. The ESD structure's (R_{esd} and a portion of C_0) influence on the transfer function manifests itself differently at high frequency. Depending on the design compensation requirements, both compensation methods have tradeoffs.

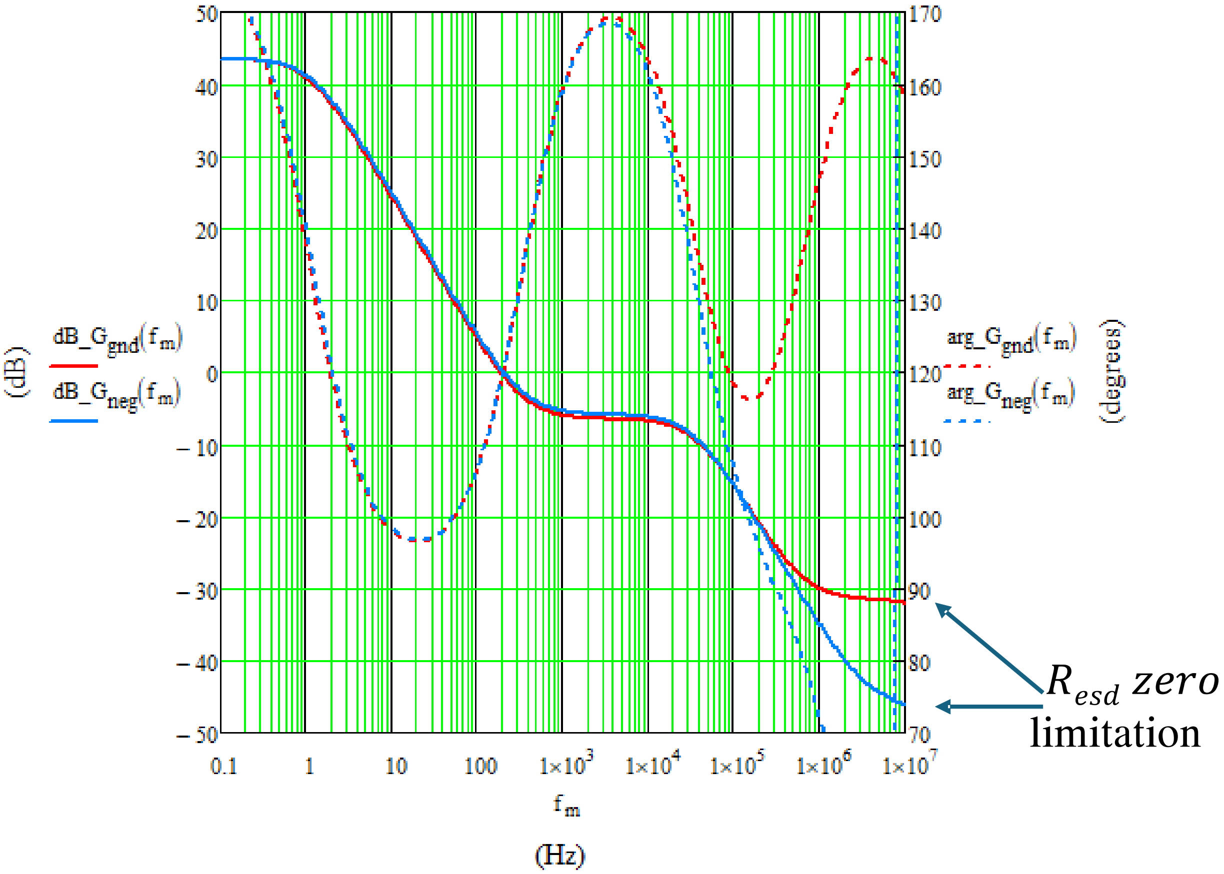

Figure 5. OTA Compensation Comparison: Grounded vs. Negative Feedback

Grounded compensation:

Benefit:

- R_{esd} introduces a high frequency ω_{z2_gnd} zero which provides a high frequency boost in phase. This facilitates higher bandwidth systems having good phase margin.

Disadvantage:

- In designs where the PWM IC's slope compensation is low for the inductor's ripple current, it may be necessary to reduce the OTA mid–band gain significantly with a low value compensation resistor R_2 so as to help provide adequate closed–loop attenuation at half the switching frequency to prevent sub–harmonic oscillation. This can be challenging for designs having high ESR output filter capacitors. R_{esd} limits the high frequency response gain roll–off to a fixed value irrespective of the value of compensation resistor R_2.

Negative feedback compensation:

Benefit:

- Larger R_2 compensation resistor values are used to shape the same mid–band gain frequency response. R_{esd} has significantly less influence on the minimum high–frequency gain rendering the configuration beneficial for difficult designs such as that described in the previous paragraph.

Disadvantage:

- ω_{z2_gnd} introduces a high frequency right–half plane zero making it more challenging to achieve a sufficient phase margin for designs requiring a high bandwidth closed–loop response.

References:

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