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TI Precision Designs: Reference Design
10 µA-10 mA, Single-Supply, Low-Side, Current Sensing Solution

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INA326	Product Folder

Circuit Description

This single-supply low-side current sensing solution can accurately detect load currents from 10 µA to 10 mA. The linear range of the output is from 100 mV to 4.9 V. A unique yet simple gain switching network was implemented in order to accurately measure the three-decade load current range.

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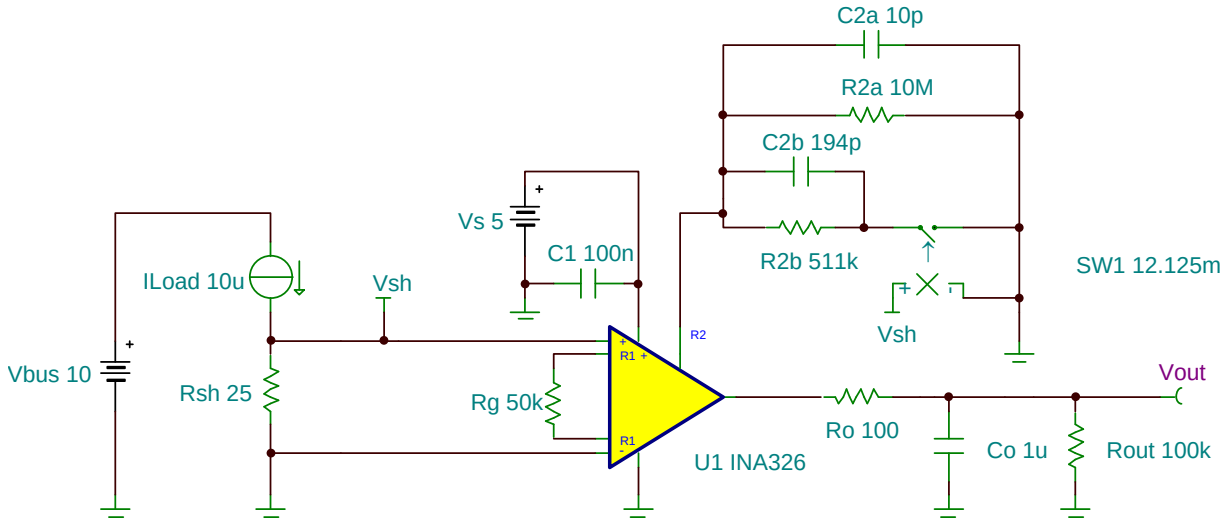

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1 Design Summary

The design requirements are as follows:

- Supply Voltage: 5 V
- Input: 10 µA – 10 mA
- Output: 100 mV – 4.9 V
- Maximum Shunt Voltage: 250 mV

The design goals and performance are summarized in Table 1. Figure 1 depicts the simulated transfer function of the design.

Table 1. Comparison of Design Goals and Simulated Performance

	Goal	Simulated
Error (%FSR_{error})	0.1%	0.088%
Relative Error (I_{load}=10 µA)	40%	29.6%
Relative Error (I_{load}=100 µA)	4%	4.04%
Relative Error (I_{load}=1 mA)	0.4%	0.4%

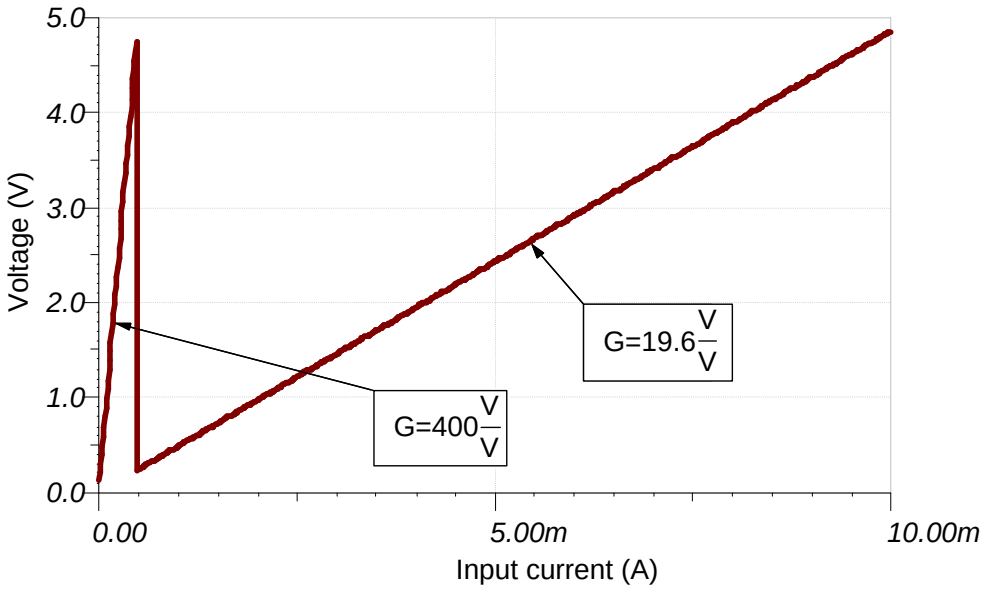

Figure 1. Simulated Transfer Function

2 Theory of Operation

Low-side sensing is desirable because the common-mode voltage is near ground. Therefore the current sensing solution is independent of bus voltage (V_{bus}), which allows for the use of single-supply, rail-to-rail input/output (RRIO) amplifiers. Low-side current sensing places a shunt resistor (R_{sh}) between the system load and ground. The current drawn by the load (I_{load}) generates a voltage across the shunt resistor (V_{sh}) that is proportional to R_{sh}. As shown in Figure 2, the output voltage (V_{out}) of the circuit is equal to the product of V_{sh} and the gain of the amplifier (G).

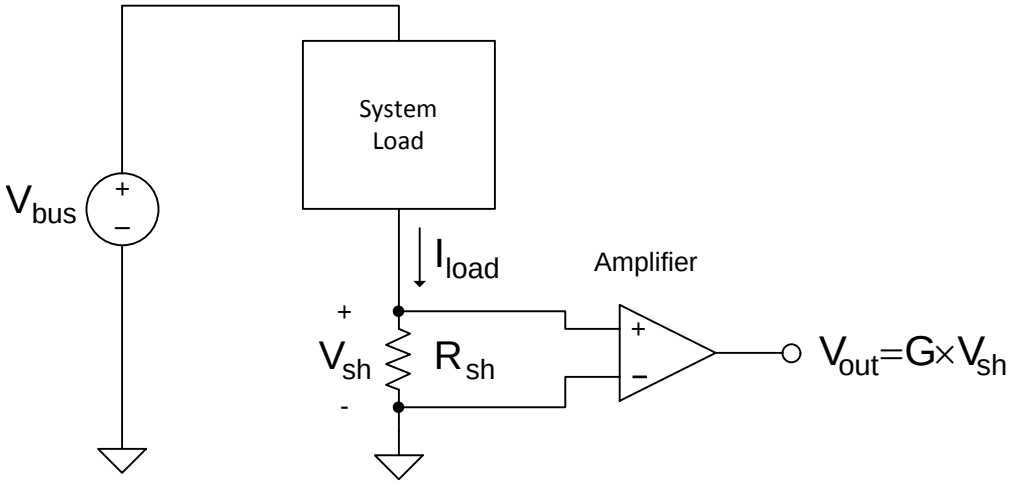

Figure 2. Low-Side Current Sensing Topology

As shown in Figure 2, the value of V_{sh} is the ground potential for the system load. If the value of V_{sh} is too large it may cause issues when interfacing with systems whose ground potential is truly 0 V. Therefore it is important to limit the voltage drop across the shunt resistor. Equation 1 can be used to calculate the maximum value of R_{sh}.

$$R_{sh(max)} = \frac{V_{sh(max)}}{I_{load(max)}} = \frac{250mV}{10mA} = 25\Omega \quad (1)$$

It is recommended to use the maximum shunt resistance to minimize relative error at minimum load current. Relative error is discussed in Section 4.3.

The gain(s) required for this design depend on the maximum output swing of the amplifier, shunt resistor, and the load current range. It is recommended to use the maximum gain to ensure full utilization of the linear operating range of the device. Equation 2 shows how to calculate the maximum gain for the maximum load current.

$$G_{Iload(max)} = \frac{V_{out(max)}}{V_{sh(max)}} = \frac{V_{out(max)}}{R_{sh} \times I_{load(max)}} = \frac{4.9V}{25\Omega \times 10mA} = 19.6\frac{V}{V} \quad (2)$$

To determine if the design requires more than one gain, the minimum load current that can be measured given $G_{Iload(max)}$ and $V_{out(min)}$ must be calculated as shown in Equation 3 and Equation 4.

$$V_{sh(min)} = \frac{V_{out(min)}}{G_{Iload(max)}} \quad (3)$$

$$I_{load(min)} = \frac{V_{sh(min)}}{R_{sh}} = \frac{V_{out(min)}}{G_{Iload(max)} \times R_{sh}} = \frac{100mV}{19.6\frac{V}{V} \times 25\Omega} = 204.08\mu A \quad (4)$$

Since $I_{load(min)}$ does not meet the minimum load current specification of 10 µA, a second gain is required. It is recommended to now calculate the maximum gain for the minimum load current ($G_{Iload(min)}$) as shown in Equation 5.

$$G_{Iload(min)} = \frac{V_{out(min)}}{V_{sh(min)}} = \frac{V_{out(min)}}{R_{sh} \times I_{load(min)}} = \frac{100mV}{25\Omega \times 10\mu A} = 400\frac{V}{V} \quad (5)$$

Equation 6 and Equation 7 show how to calculate the maximum load current that can be measured given $G_{lload(min)}$.

$$V_{sh(max)} = \frac{V_{out(max)}}{G_{lload(min)}} \tag{6}$$

$$I_{load(max)} = \frac{V_{sh(max)}}{R_{sh}} = \frac{V_{out(max)}}{G_{lload(min)} \times R_{sh}} = \frac{4.9V}{400\frac{V}{V} \times 25\Omega} = 490\mu A \tag{7}$$

Since the minimum load current that can be measured when G=19.6 V/V overlaps with the maximum load current for G=400 V/V, only two gains are required to measure the entire load current range.

If more than two gains are required, the design of the switching network may become overly complicated. For such applications it is recommended to investigate the use of alternate topologies and/or devices, such as programmable gain amplifiers (PGAs).

3 Component Selection

3.1 Amplifier

When selecting an amplifier for a low-side current sensing solution it is desirable to have a RRIO device with low offset voltage relative to the minimum value of V_{sh}. In addition, ensure that the input bias current is insignificant with respect to the load current and that the bandwidth of the device is sufficient for the application. While it is common to utilize op amps in low-side current sensing solutions, error can be introduced due to any parasitic impedance between the shunt resistor and ground. Implementing a low-side current sensing solution using an instrumentation amplifier, however, addresses this issue because the measurement is differential. Most instrumentation amplifiers utilize traditional 3-op amp or 2-op amp architectures. Such architectures typically have input common-mode voltage and output swing limitations that are not suited for low-side current sensing, especially when powered with a single-supply. The INA326, however, utilizes a unique current-mirroring topology that allows for true rail-to-rail input voltage swing even with a single supply. The linear input voltage range of the INA326 is from (V-)-20 mV to (V+)+100 mV. The output stage is also considered rail-to-rail for it can swing to within 75 mV of the supplies. The offset voltage is 100 µV (max) and the input bias current is +/-2 nA. If the application is dc, the INA326's 1 kHz of bandwidth is sufficient. Therefore the INA326 is well suited for this application.

3.2 Shunt Resistor

The ideal shunt resistor was calculated in Equation 1 to be 25 Ω. This value is a standard value and is available with a tolerance of 0.1%. As previously mentioned it is recommended to use the maximum shunt resistor to minimize relative error at minimum load current.

3.3 Gain-Setting Resistors

The INA326 utilizes two resistors (R_1 and R_2) to set the gain as shown in Figure 3.

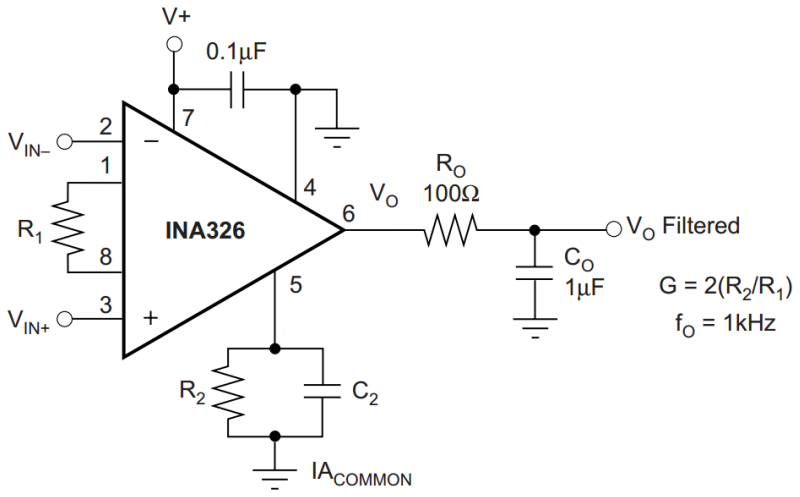

Figure 3. Setting the Gain of the INA326

A switch can be used to change the value of either R_1 or R_2. Since R_1 is connected to pins that are sensitive to parasitic capacitance, this design will change the value of R_2. One approach to changing the value of R_2 is shown in Figure 4.

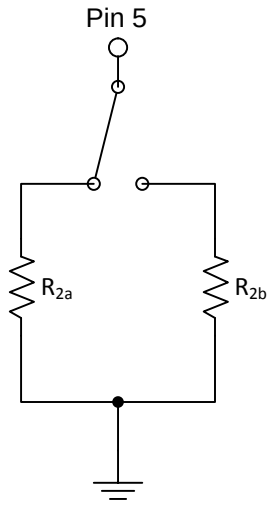

Figure 4. Changing the Gain Using SPDT Switch

Using a single-pole double-throw (SPDT) switch, however, can cause inaccurate results during the switch's transition time. Therefore and alternate approach that utilizes a single-pole single-throw (SPST) switch was implemented as shown in Figure 5.

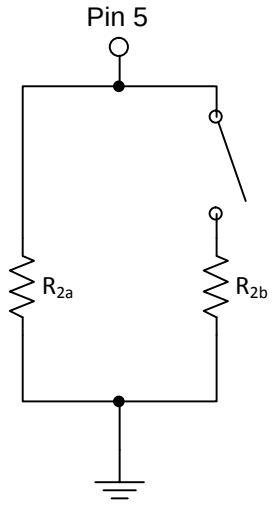

Figure 5. Changing the Gain Using SPST Switch

This approach ensures that pin 5 of the INA326 is always connected to ground through a known impedance. While the gain calculation for R_1 and R_{2a} is straightforward, the second gain calculation will utilize the parallel combination of R_{2a} and R_{2b}.

Setting R_1=50 kΩ we can calculate the value for R_{2a} for G=400 V/V as shown in Equation 8.

$$R_{2a} = \frac{R_1 \times G}{2} = \frac{50k\Omega \times 400\frac{V}{V}}{2} = 10M\Omega \tag{8}$$

To calculate the value of R_{2b} the parallel combination of R_{2a} and R_{2b} must be used in the calculation.

$$R_{2a} \,||\, R_{2b} = \frac{R_1 \times G}{2} = \frac{50k\Omega \times 19.6\frac{V}{V}}{2} = 490k\Omega \tag{9}$$

Where

$$R_{2a} \parallel R_{2b} = \frac{R_{2a} \times R_{2b}}{R_{2a} + R_{2b}} = 490k\Omega \quad (10)$$

Finally, R_{2b} can be calculated as shown in Equation 11.

$$R_{2b} = \frac{490k\Omega \times R_{2a}}{R_{2a} - 490k\Omega} = 515.25k\Omega \quad (11)$$

Since gain and the value of R_2 are directly related, a standard resistor value less than 515.25 kΩ must be selected to ensure the maximum gain of 19.6 V/V is not exceeded. Therefore R_{2b} was selected to be 511 kΩ and the ideal gain now becomes 19.4463 V/V.

3.4 Filtering

The INA326 datasheet values for R_o and C_o were used to create a low-pass filter on the output at a -3 dB point of 1.59 kHz.

$$f_{-3dB} = \frac{1}{2 \times \pi \times R_o \times C_o} = \frac{1}{2 \times \pi \times 100\Omega \times 1\mu F} = 1.59kHz \quad (12)$$

Figure 6 shows the gain and filtering network.

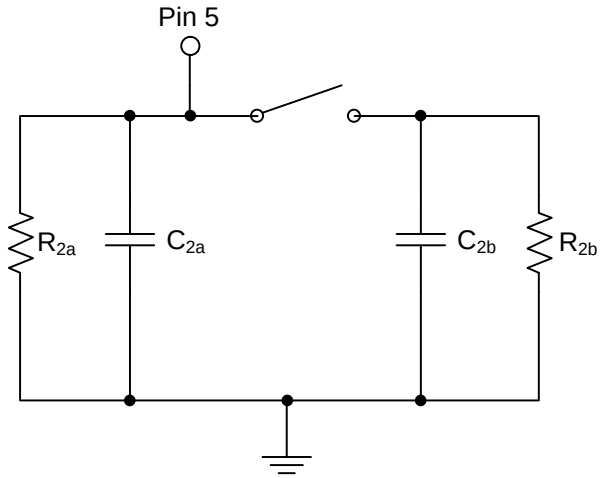

Figure 6. Gain and Filtering Network

When the switch is open, C_{2a} can be calculated using Equation 13.

$$C_{2a} = \frac{1}{2 \times \pi \times R_{2a} \times f_{-3dB}} = \frac{1}{2 \times \pi \times 10M\Omega \times 1.59kHz} = 10pF \quad (13)$$

After the switch closes, R_{2a}, R_{2b}, C_{2a}, and C_{2b} will be in parallel. The effective capacitance of C_{2a} in parallel with C_{2b} is the sum of the two capacitances. Therefore C_{2b} can be calculated as shown in Equation 14.

$$C_{2b} = \frac{1}{2 \times \pi \times (R_{2a} \parallel R_{2b}) \times f_{-3dB}} - C_{2a} = \frac{1}{2 \times \pi \times (10M\Omega \parallel 511k\Omega) \times 1.59kHz} - 10pF = 196pF \rightarrow 194pF \quad (14)$$

The closest standard capacitor value to 196 pF is 194 pF.

3.5 Passive Components

The critical passive components for this design are R_{sh}, R_1, R_{2a}, and R_{2b}. In order to meet the design objective of 0.1% FSR_{error}, 0.1% tolerance was selected for these components. Other passive components in this design may be selected for 1% or greater as they will not directly affect the transfer function or add significant error to this design.

4 Simulation

The TINA-TI™ schematic shown in Figure 7 includes the circuit values obtained in the design process.

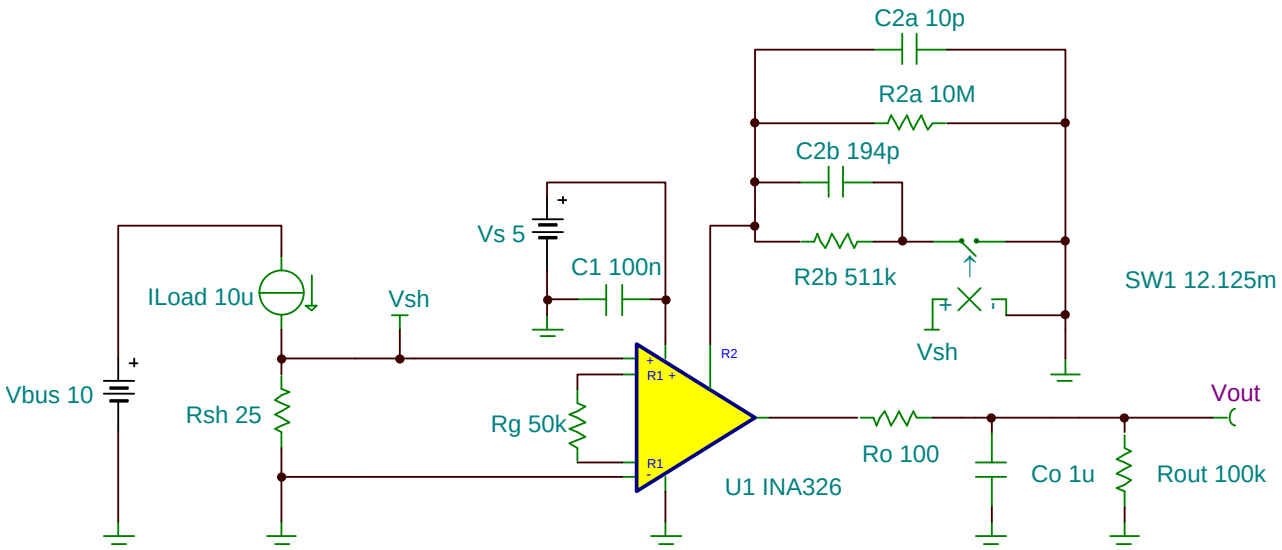

Figure 7. TINA-TI™ Schematic

Notice that a voltage-controlled switch (SW1) was inserted to control the gain of the circuit. By default, the switch is open, which equates to an ideal gain of 400 V/V. The set point of the switch can be calculated as shown in Equation 15. Note that 485 µA was used instead of 490 µA (the maximum load current before the output saturates when G=400 V/V) to allow for design margin and that the set point is referred to the input.

$$V_{SW1} = 485\mu A \times 25\Omega = 12.125mV \quad (15)$$

While an ideal switch was used in this design, it represents a real-world control system. This most likely solution would include measuring V_{out} using an analog-to-digital converter (ADC) of a microcontroller. Once the threshold is reached, the microcontroller would use a general purpose input/output (GPIO) pin to control the switch. It is recommended to use a switch with low on-resistance (R_{on}) so as to preserve accuracy. The TS5A3166 would work well for this application.

4.1 Transfer Function

The result of the dc transfer function simulation is shown in Figure 8.

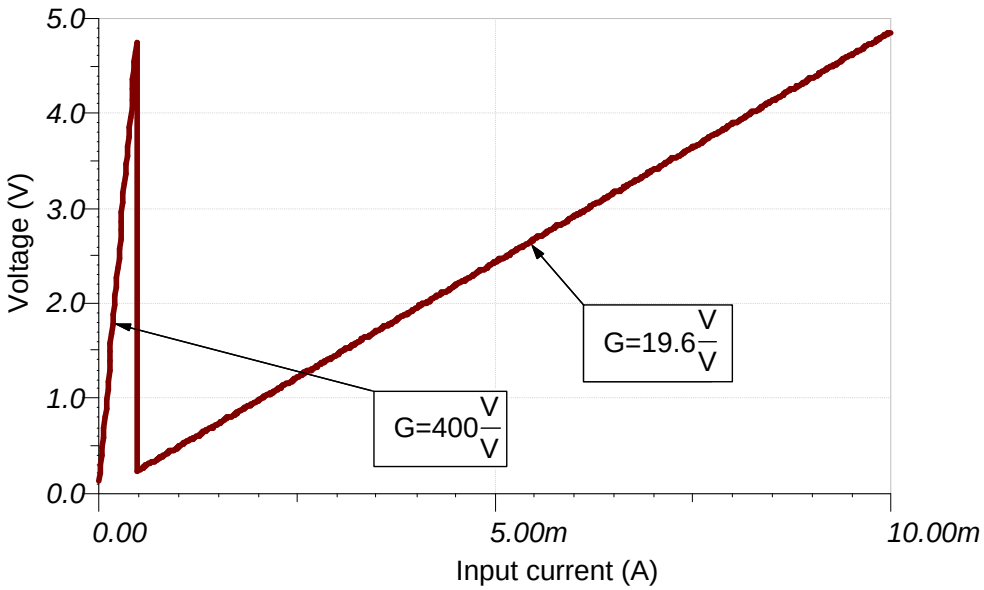

Figure 8. Transfer Function

4.2 Frequency Response

The circuit shown in Figure 9 was utilized to perform an ac analysis for each gain of the circuit depending on whether the switch was open or closed. The simulation results are shown in Figure 10.

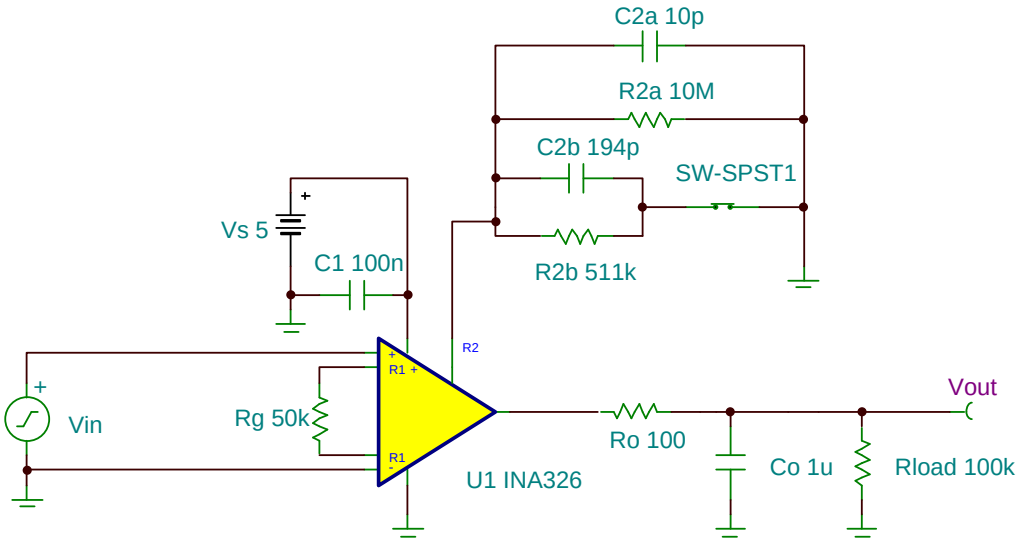

Figure 9. AC Analysis Schematic

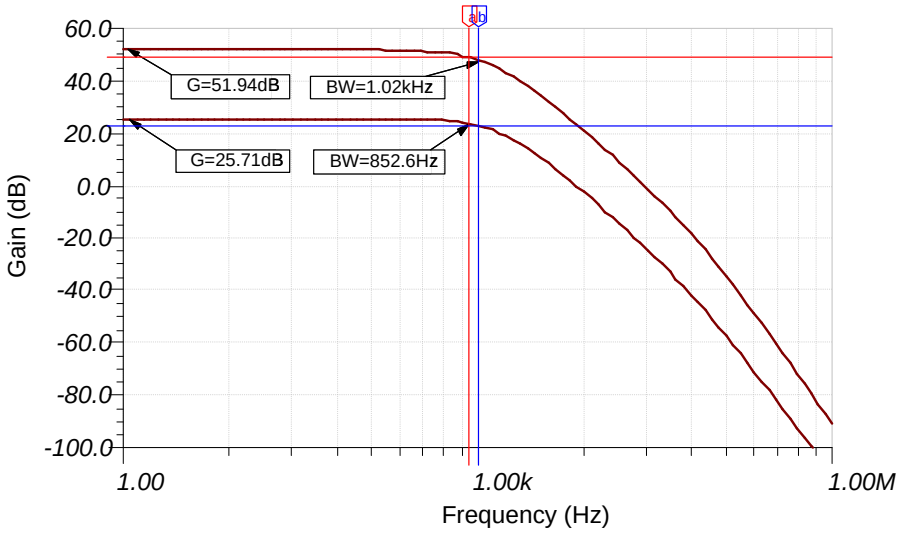

Figure 10. AC Analysis

The dc gains of the simulation were found to be 51.94 dB and 25.71 dB, which equate to 395.37 V/V and 19.3 V/V, respectively. The bandwidth of each gain configuration was 852.6 Hz and 1.02 kHz, respectively.

4.3 Error Analysis

The data from Figure 8 was exported to a spreadsheet in order to calculate the error as a percent of the full-scale range (%FSR_{error}). Equation 16 was used to calculate %FSR_{error}.

$$\%FSR_{error} = 100 \times \frac{I_{load(sim)} - I_{load(ideal)}}{I_{load(max)} - I_{load(min)}} = 100 \times \frac{\left[\frac{V_{out}}{G} \times \frac{1}{25\Omega}\right] - I_{load(ideal)}}{10mA - 10\mu A} = 100 \times \frac{\left[\frac{V_{out}}{G} \times \frac{1}{25\Omega}\right] - I_{load(ideal)}}{9.99mA} \quad (16)$$

In order to accurately calculate the error, the gain was switched from 400 V/V to 19.4463 V/V once the load current reached 485 µA. Figure 11 shows %FSR_{error} as a function of ideal load current.

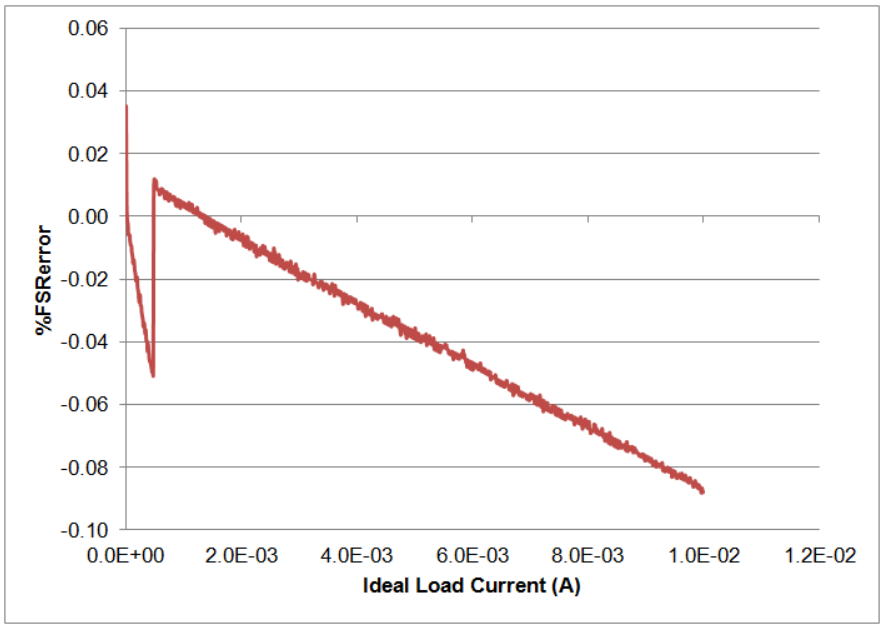

Figure 11. Full Scale Error

The maximum $\%FSR_{error}$ was found to be 0.088%, which meets our design goal of 0.10%.

The error relative to the ideal load current (relative error) was also calculated for each load current decade. Equation 17 shows how to calculate relative error.

$$\%RelError = 100 \times \frac{V_{os(max)}}{V_{sh}} \tag{17}$$

The maximum offset voltage of the INA326 is 100 µV. The maximum relative error for each decade of load current will occur at the minimum load current for each range (10 µA, 100 µA, and 1 mA). For example, the maximum relative error for a load current of 10 µA is calculated in Equation 18.

$$\%RelError_{10\mu A} = 100 \times \frac{V_{os(max)}}{V_{sh}} = 100 \times \frac{100\mu V}{25\Omega \times 10\mu A} = 40\% \tag{18}$$

The maximum relative error for 100 µA and 1 mA can be calculated in a similar manner. Table 3 summarizes the results of the calculations.

Equation 19 calculates the simulated relative error using the data from Figure 8. The output voltage was referred to the input by dividing by the ideal gain (400 V/V or 19.4463 V/V).

$$\%RelError_{sim} = 100 \times \frac{V_{os(max)}}{V_{out}} \times G \tag{19}$$

Table 2 depicts the simulated output voltage at the minimum load current of each range.

Table 2. Simulated Output Voltage for Various Load Currents

Load Current	Simulated Output Voltage
10 µA	135.15 mV
100 µA	990.65 mV
1 mA	486.31 mV

The simulated relative error for a load current of 10 µA can be calculated as shown in Equation 20.

$$\%RelError_{sim-10\mu A} = 100 \times \frac{100\mu V}{135.15mV} \times 400\frac{V}{V} = 29.6\% \quad (20)$$

The simulated relative error for 100 µA and 1 mA can be calculated in a similar manner. Table 3 summarizes the results of the calculations.

4.4 Simulated Results Summary

Table 3 summarizes the simulated performance of the design.

Table 3. Comparison of Design Goals and Simulated Performance

	Goal	Simulated
Error (%FSR_{error})	0.1%	0.088%
Relative Error (I_{load}=10 µA)	40%	29.6%
Relative Error (I_{load}=100 µA)	4%	4.04%
Relative Error (I_{load}=1 mA)	0.4%	0.4%

5 Modifications

The components selected for this design were based on the design goals outlined at the beginning of the design process. An instrumentation amplifier was selected because the input is differential, which reduces errors due to parasitic impedances associated with the shunt resistor's ground connection. The amplifier's input common-mode range must include ground, which limits our choice of instrumentation amplifiers. One alternative to the INA326 is the INA122. The INA122 is a 2-op amp instrumentation amplifier with a common-mode voltage range that includes V-. Table 4 is a brief comparison of the two instrumentation amplifiers.

Table 4. Brief Comparison of Instrumentation Amplifiers

Amplifier	Max Supply Voltage (V)	Max Offset Voltage (µV)	Max Offset Drift (µV/°C)	Quiescent Current (µA)	Bandwidth (MHz, G=5)
INA326	5.5	100	0.4	3400	0.001
INA122	36	500	5	85	0.12

Traditional op amps are also used for low-side current sensing. They are especially useful for higher-frequency measurements because they can have significantly greater bandwidth than the INA326 and INA122. Table 5 is a brief comparison of op amps that may be considered for low-side current sensing.

Table 5. Brief Comparison of Op Amps

Amplifier	Max Supply Voltage (V)	Max Offset Voltage (µV)	Max Offset Drift (µV/°C)	Quiescent Current (µA)	Bandwidth (MHz)
OPA333	5.5	10	0.05	25	0.35
OPA330	5.5	50	0.25	35	0.35
OPA378	5.5	50	0.25	150	0.90
OPA320	5.5	150	5	1750	20
OPA365	5.5	200	1	5000	50
OPA376	5.5	25	1	950	5.5
OPA188	36	25	0.085	510	2

6 About the Author

Pete Semig is an Analog Applications Engineer in the Precision Linear group at Texas Instruments. He supports Texas Instruments' difference amplifiers & instrumentation amplifiers. Prior to joining Texas Instruments in 2007, he earned his B.S.E.E. and M.S.E.E. from Michigan State University in 1998 & 2001, respectively. From 2001-2007 he was a faculty member in Michigan State University's Department of Electrical & Computer Engineering where he taught a variety of courses and laboratories.

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