Design Guide: TIDA-060029

LCR Meter Analog Front-End Reference Design

TEXAS INSTRUMENTS

Description

This reference design, TIDA-060029, demonstrates an analog signal chain solution for LCR Meter applications using an auto balancing impedance measurement method. Auto balancing impedance measurement circuits can often be challenging to stabilize because the circuit stability is dependent on both the value and type of component to be measured. Therefore, it is imperative to have a circuit solution that is inherently stable irrespective of the measured component's type and value. This design presents an analog signal-chain solution that is both inherently stable and accurate to 0.1% for LCR meter applications.

Resources

TIDA-60029	Design Folder
OPA2810	Product Folder
OPA810	Product Folder
BUF634A	Product Folder

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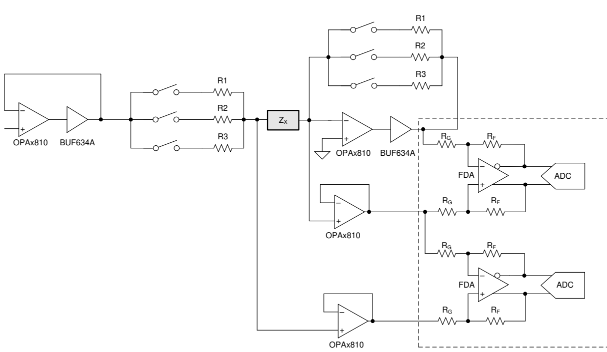
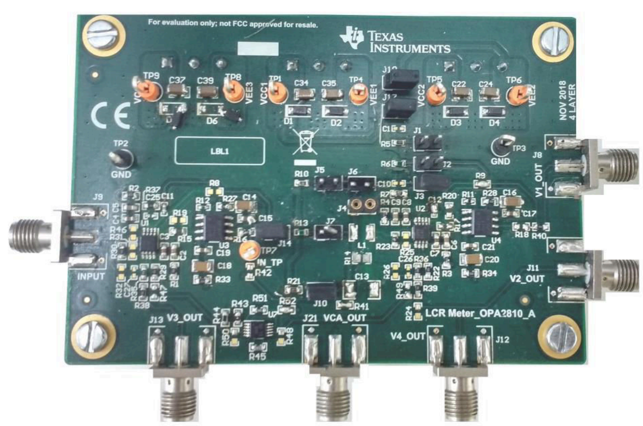
Features

- Measures wide range of components (L, C, R) with impedance values ranging from 1 Ω to 10 MΩ
- Frequency of operation up to 100 kHz
- Tested at 100 Hz, 1 kHz, 10 kHz, 100 kHz
- Impedance accuracy of 0.1 %
- Inherently stable operation of the signal chain

Applications

- Digital Multimeter (DMM)
- Impedance and Vector Network Analyzer
- Semiconductor Manufacturing
- Semiconductor Test





1 System Description

The goal of any test and measurement system is to measure a device under test (DUT) as simply as possible, while only introducing errors significantly smaller than those present in the measured device. For impedance measurements, there are several existing techniques that provide various tradeoffs between measurement accuracy, complexity, and frequency range. For this design, the auto balancing circuit method was chosen because it provides good accuracy over a wide impedance measurement range without any tuning requirements. Table 1-1 lists the advantages and disadvantages of several common impedance measurement techniques along with their frequency ranges and typical applications.

Table 1-1. Impedance Measurement Methods

METHOD	ADVANTAGES	DISADVANTAGES	APPLICABLE FREQUENCY RANGE	COMMON APPLICATION
Bridge method	• High Accuracy • Wide frequency range with different types of bridges	• Manual balancing needed • Narrow frequency coverage with single bridge	DC to 300 MHz	Standard Lab
Resonant method	• Good Q measurement accuracy up to high Q	• Tuning required • Low impedance measurement accuracy	10 kHz to 70 MHz	High Q device measurement
Network analysis method	• Wide frequency coverage • Good accuracy	• Narrow impedance measurement range	5 Hz to above	RF component measurement
Auto balancing method (**Method used in this design**)	• Good accuracy over wide range of impedances • Grounded device measurement	• High frequency ranges are not available	20 Hz to 120 MHz	Generic component measurement

The auto balancing technique is very useful for a wide range of impedance measurements at a frequency range of 20 Hz to 120 MHz. The auto balancing technique uses an op-amp as shown in Figure 1-1.

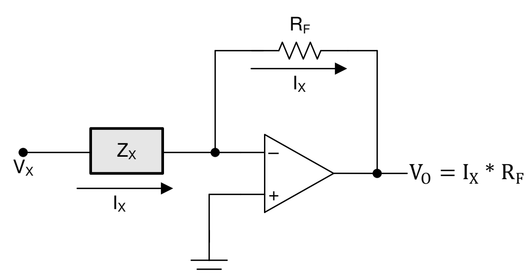


Figure 1-1. Auto Balancing Circuit Amplifier Configuration

The fundamental idea in this technique is to convert the current (I_X) through unknown impedance (Z_X) into voltage (V_O). The unknown impedance value is determined using the value of current flowing through it. The non-ideal properties of the amplifier and circuit play a very crucial role in the design of an LCR meter. For example, the parasitic capacitance at the inverting input of the amplifier will cause instability for a high value of R_F. The circuit's stability is also sensitive to both the type of component and value used for Z_X. The circuit is particularly prone to instability when capacitive impedances are measured. In this design, these stability problems are addressed using a multi-path capacitive compensation technique. This design illustrates the analog signal chain of an LCR meter which is tested up to 100 kHz.

1.1 Key System Specifications

Table 1-2. Key System Specifications

PARAMETER	SPECIFICATIONS
Resistance Range	1 Ω to 10 MΩ
Capacitance Range	1.76 pF to 1.59 mF
Inductance Range	2.59 µH to 1432 H
Frequencies of Operation	100 Hz, 1 kHz, 10 kHz, 100 kHz
R_G – R_F Settings	100Ω, 5 kΩ, 100 kΩ
Best % Accuracy	0.1%
Power Supply	+/- 12 V

2 System Overview

2.1 Block Diagram

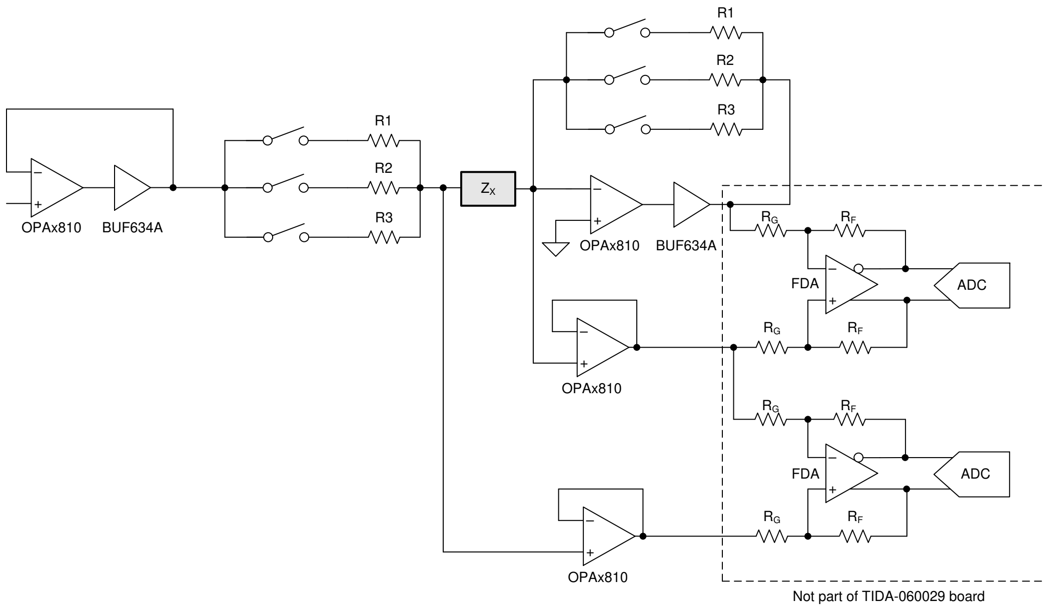


Figure 2-1. TIDA-060029 Block Diagram

2.2 Highlighted Products

2.2.1 OPA2810

The OPA2810 is a dual-channel, FET-input, voltage-feedback operational amplifier with low input bias current of 2 pA. The extremely low input bias current is very useful in this application because this current will flow through unknown impedance which can go high as 10 MΩ. OPA2810 is unity-gain stable with a small-signal unity-gain bandwidth of 105 MHz, and offers excellent DC precision and dynamic AC performance at a low quiescent current. This device has a DC open loop gain equal to 120 dB. With a gain-bandwidth product (GBW) of 70 MHz, the OPA2810 has Aol greater than 60 dB at all the frequencies less 100 kHz. The high Aol of the op amp reduces the error in measurement because as the Aol increases the voltage at the inverting input approaches to zero. Thus, this is very important specification of this device to make it well suited for use in this application. The supply voltage of OPA2810 can go up to +/- 13.5 V. This high voltage operation provides optimal distortion performance in the LCR meter signal chain. The voltage noise of this amplifier is 6 nV/√Hz.

2.2.2 BUF634A

The BUF634A is a high speed wide bandwidth unity gain buffer. It is used in composite loop with the OPA2810 to increase the output current capability from 100mA to 250 mA. The BUF634A has two bandwidth options of 35 MHz and 210 MHz. It is optional to use in this application.

2.3 Design Considerations

2.3.1 Existing architecture

The fundamental concept of this design is the conversion of current through Z_X into a voltage using an amplification factor of R_F. The output of amplifier A2 is given in Equation 1

$$V_O = \left(-\frac{R_F}{Z_X}\right) * V_{IN} \tag{1}$$

If R_F is known then Z_x can be estimated using Equation 1. Figure 2-2 illustrates this architecture.

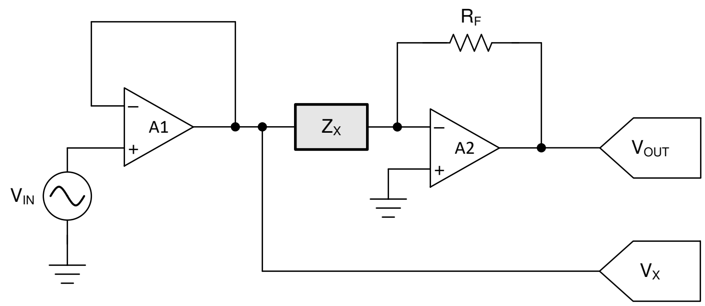


Figure 2-2. Basic Impedance Measurement Circuit

In this method, multiple values of R_F can be used for multiple ranges of impedance, as shown in Figure 2-6. This approach of using multiple R_F values improves the accuracy.

2.3.1.1 Circuit Stability Issue

When the unknown impedance is capacitive as shown in Figure 2-3, the feedback transfer function can be calculated using Equation 2.

$$\frac{1}{\beta} = \frac{V_O}{V_F} = 1 + R_F * C_X * S \tag{2}$$

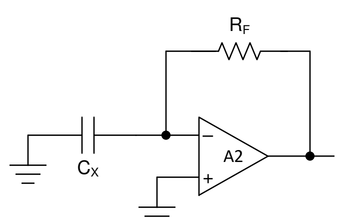


Figure 2-3. Capacitor Measurement Circuit

The transfer function implies that a zero is formed in 1/β. The frequency of this zero can be calculated using Equation 3.

$$\omega_Z = \frac{1}{R_F * C_X} \tag{3}$$

It can be seen that the zero frequency depends on the unknown capacitance, C_X. In Figure 2-4 it can be seen that Aolβ has a rate of closure of 40dB/dec. When the zero frequency is more than a decade below f_{CL}, the phase margin will reduce to zero making the circuit unstable.

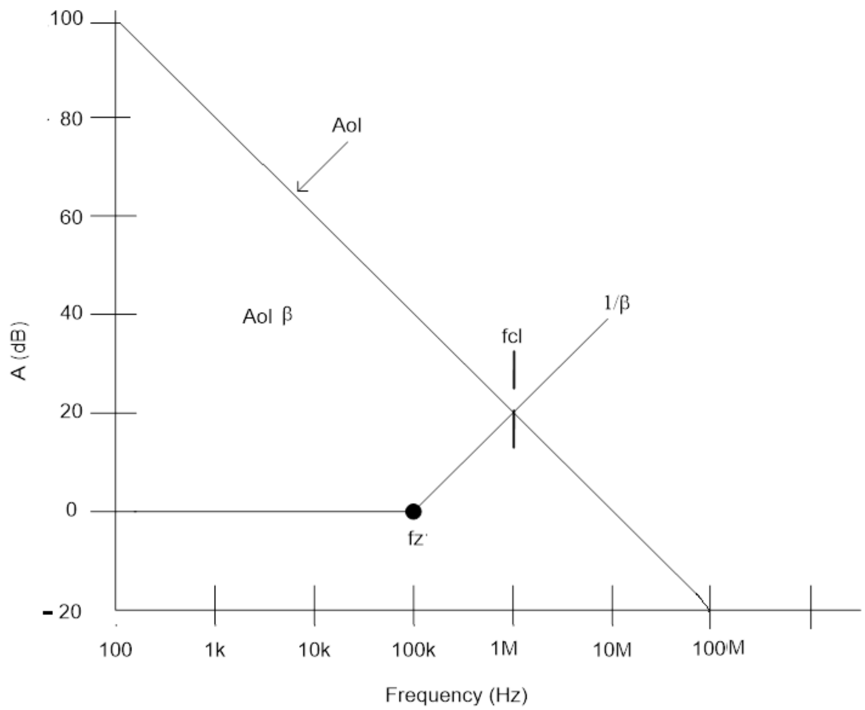


Figure 2-4. Bode Plot of Capacitor Measurement

2.3.1.2 Solution in Existing Architecture (Compensation Cap)

The circuit instability issue is resolved in the existing architecture with the help of compensation capacitor C_F in parallel with R_F as shown in Figure 2-5. The required value of C_F also varies with unknown capacitance C_X. Hence, it becomes impossible to find single value of C_F which can stabilize the circuit for complete range of Z_X.

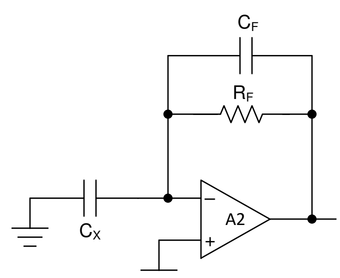


Figure 2-5. Capacitor Measurement Circuit w/ Compensation Capacitor

2.3.2 Proposed Design

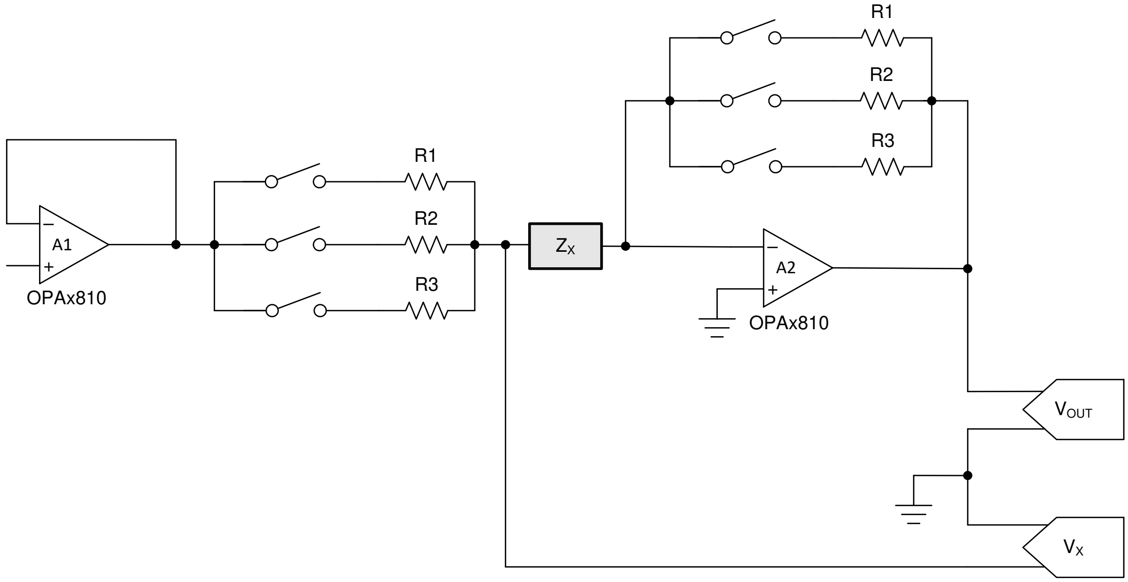


Figure 2-6. Impedance Measurement Design

In this method, three different combinations of R_G – R_F (labeled as R1, R2 & R3 in Figure 2-6) are selected for three ranges of impedance Z_X. The ranges can be seen in Table 2-1. The architecture in this method is very similar to the existing architecture explained in last section. The only difference is that the R_G is added in series with Z_X. Also the value of R_G is equal to R_F. The stability analysis in Section 2.3.2.1 explains the advantage of this kind of setting.

2.3.2.1 Stability Analysis of the Proposed Design

When the unknown impedance to be measured is capacitive i.e. C_X, it forms the circuit shown in Figure 2-3. The transfer function of V_F is given in Equation 4.

$$\frac{V_F}{V_O} = \frac{1}{\beta} = \frac{1 + R_F * C_X * S}{1 + (R_F + R_G) * C_X * S} \tag{4}$$

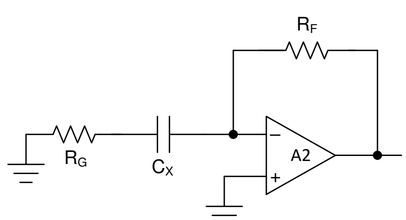


Figure 2-7. Capacitive Measurement with Series Resistance

In comparison with Equation 2, Equation 4 shows that due to presence of R_G, there is a pole-zero combination in the feedback path. The zero and pole frequencies in 1/β are given by,

$$\omega_Z = \frac{1}{(R_F + R_G) * C_X} \tag{5}$$

$$\omega_P = \frac{1}{R_G * C_X} \tag{6}$$

The pole and zero frequencies hold the relation $\omega_P = 2^*\omega_Z$ because R_G is equal to R_F in every R_G - R_F setting.

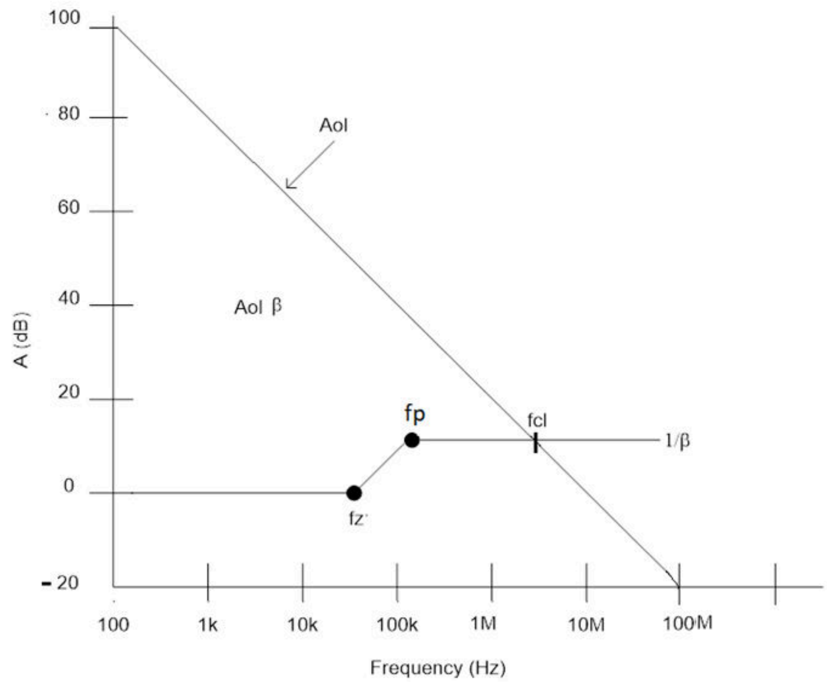


Figure 2-8. Bode Plot of Capacitive Measurement with Series Resistance

This provides the advantage of an inherent pole to cancel the zero. Figure 2-8 shows that the rate of closure of Aolβ is 20 dB/dec for almost all the C_X. The exception for this fact is when f_{CL} lies between f_Z and f_P. The R_G - R_F settings are selected such that this situation is avoided. This allows for a key factor of this design where ω_P = 2*ω_Z is independent of the value of C_X. The measurement can be done in two ways as explained below,

2.3.2.1.1 Without Measurement of Voltage at Inverting Node of A2

In this method of measurement, the inverting node of A2 is not measured. The assumption behind this method is that the inverting node of A2 is 0V since it is equal to the non-inverting node. This case is only possible in the ideal scenario where the Aol of A2 is infinity. But due to the finite open loop gain of op amp there will always be some small voltage at the inverting node of A2. This voltage is inversely proportional to Aol. As practical op amps have gain decay with respect to frequency, Aol will decrease significantly as the frequency increases. It makes this method of measurement erroneous at high frequencies. Hence the Aol of the amplifier plays a very crucial role in this method of measurement and should be as high as possible. Figure 2-9 explains this method of measurement.

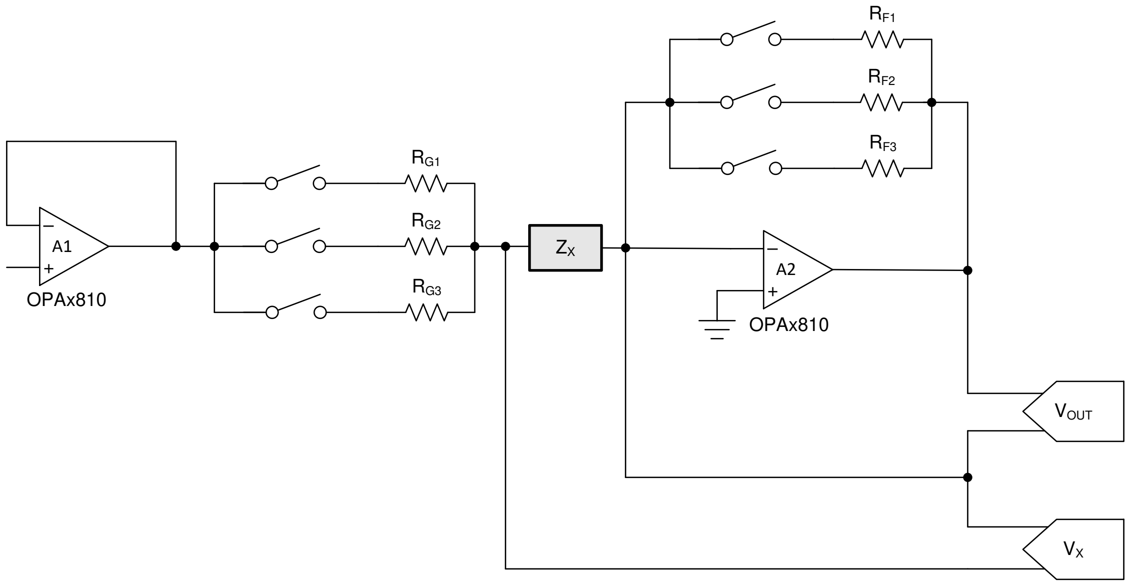


Figure 2-9. Method 1 of Impedance Measurement

As this method measures both voltages with respect to ground, data can be acquired using a single-ended ADC.

2.3.2.1.2 With Measuring Voltage at Inverting Node of A2

Figure 2-6 shows the method to measure the difference voltage across both Z_X and R_F.

This method nullifies the error due to voltage at the inverting node of A_2. Since the voltages across Z_X and R_F are being measured, this method demands a differential ADC be used for data acquisition. In this design, the accuracies of both methods were verified and found to be the same in the proposed frequencies of operation. High value of Aol (more than 60 dB) of OPA2810 is responsible for this performance.

2.3.2.2 R_G = R_F Settings and Respective Impedance Ranges

Table 2-1 explains the ranges for different components with respect to the R_G - R_F and frequency settings.

Table 2-1. Board Setting for Respective Impedance Range

PARAMETERS		**R_G = R_F SETTING**		
Frequency (Hz)	**Component**	**100 Ω**	**5 kΩ**	**100 kΩ**
100	R	1 Ω to 900 Ω	500 Ω to 50 kΩ	10 kΩ to 10 MΩ
	L	1.59 mH to 2.38 H	2.27 H to 79.5 H	72.9 H to 1432 H
	C	1.05 µF to 1.59 mF	31.78 nF to 1.11 µF	1.76 nF to 34.7 nF
1 k	R	1 Ω to 900 Ω	500 Ω to 50 kΩ	10 kΩ to 10 MΩ
	L	159 µH to 238 mH	227 mH to 7.95 H	7.29 H to 143.23 H
	C	106 nF to 159 µF	3.178 nF to 111 nF	176 pF to 3.47 nF
10 k	R	1 Ω to 900 Ω	500 Ω to 50 kΩ	10 kΩ to 10 MΩ
	L	25.9 µH to 23.8 mH	22.6 mH to 795 mH	729 mH to 14.3 H
	C	10.6 nF to 15.9 µF	317.8 pF to 11.1 nF	17.6 pF to 347 pF
100 k	R	1 Ω to 900 Ω	500 Ω to 50 kΩ	10 kΩ to 10 MΩ
	L	2.59 µH to 2.38 mH	2.26 mH to 79.6 mH	72 mH to 1.43 H
	C	1.06 nF to 1.59 µF	31.78 pF to 1.11 nF	1.76 pF to 34.7 pF

In the 100 kΩ setting, the parasitic cap at the inverting pin can make the circuit unstable. To overcome this problem a 5 pF capacitor is added in parallel with R_F = 100 kΩ.

2.3.2.3 Impedance Measurement Procedure

The impedance measurement procedure includes a one-time calibration procedure which consists of four different calibrations measurements named as:

1. Short Cal
2. Impedance Cal
3. 100k Setting Cal
4. Open Cal

It must be noted that one time calibrations are done at all four frequencies of operation. Also all the calibrated values of R_F are phasor quantities and the phase will be used in estimation of the unknown impedance.

2.3.2.3.1 Short Cal

In this calibration, Z_X is shorted and the ratio between V_O and V_{IN} is measured. This measurement is called as G_{CAL},

$$G_{CAL} = \frac{V_O}{V_{IN}} \tag{7}$$

Where V_{IN} is the voltage across R_G + Z_X. In order to measure $V_{IN,}$ R_{41} should be removed and a 0-ohm resistor added to R_{42}. For all other measurements, the default configuration should be used. This calibration is needed only in R_G = R_F = 100k setting, as seen in the next steps.

2.3.2.3.2 Impedance Cal

In this calibration, a known resistance of 500 Ω is used as Z_X. V_O is given by,

$$V_O = \left(-\frac{R_F}{500}\right) * V_{IN} \tag{8}$$

This calibration is used to calculate the exact value of R_F, both in the 100 Ω and 5 kΩ settings. It should be noted that the value of known resistance (R_{CAL}) is selected to be 500 Ω in order to get the best possible calibration accuracies in both 100 Ω and 5 kΩ R_G = R_F settings. User can use other values for R_{CAL}. The accuracy of R_{CAL} will, however, directly affect the calibration accuracy.

2.3.2.3.3 100k Setting Calibration

In this calibration process, the first step is to set R_G = 100 kΩ , R_F = 5 kΩ and to short Z_X. This will give,

$$G_1 = -\frac{R_F}{R_G} \tag{9}$$

With G1 being the measured gain, and R_F being the calibrated value of 5kΩ found in the previous step. Using this, the calibrated value of R_G = 100kΩ can be found. After calibrating for R_G = 100 kΩ, we can then use this to calibrate R_F = 100 kΩ using G_{CAL} from the short calibration step. In this way, we have obtained calibrated values of RF in all the three settings.

2.3.2.3.4 Open Cal

In this calibration, Z_X is kept open. G_O is given by,

$$G_O = -\frac{R_F}{Z_O} \tag{10}$$

Where Z_O is an open circuit impedance. The significance of this calibration is mainly at higher frequencies when the parasitic capacitance in parallel with Z_X is large enough to affect the measurement significantly.

2.3.2.3.5 Calculations

To estimate the value of an unknown impedance, Equation 11 can be used. Z_X is the unknown impedance, V_X is the voltage across Z_X and V_O is the voltage across R_F.

$$\frac{V_O}{V_X} = -\frac{R_F}{Z_X} \tag{11}$$

As the calibrated value of R_F is known for all settings, Z_X can be solved for.

2.3.2.3.6 Correction in Z_X

It should be noted that Z_X is an effective impedance formed by the parallel combination of an actual unknown impedance and Z_O (Open circuit Impedance). Let the actual value of the unknown impedance be Z_X', then

$$Z_X = Z_O \,||\, Z_X' \tag{12}$$

As both Z_X and Z_O are known, Z_X' can be estimated using Equation 13.

$$Z_X{}' = \frac{Z_O - Z_X}{Z_O * Z_X} \tag{13}$$

Note All the impedances are phasor quantities so the subtraction will be phasor subtraction.

2.3.2.3.7 Data Acquisition and Processing

The voltages are acquired using a two channel differential ADC and processed in the following form. The following two steps can be implemented in software to obtain the magnitude and phase of any voltage to be measured in this application:

1. Modulation of the signal by multiplying the signal with a unity magnitude square wave of 0 degree phase and taking the average of the resulting signal
2. Modulation of the signal by multiplying the signal with a unity magnitude square wave of 90 degree phase and taking the average of the resulting signal

2.3.2.3.8 Mathematical Explanation

Let $V=V_O*Sin(\omega_t+\alpha)$ be any signal, if it is multiplied by a unity magnitude square wave with 0° phase then the resultant output is as shown in Figure 2-10.

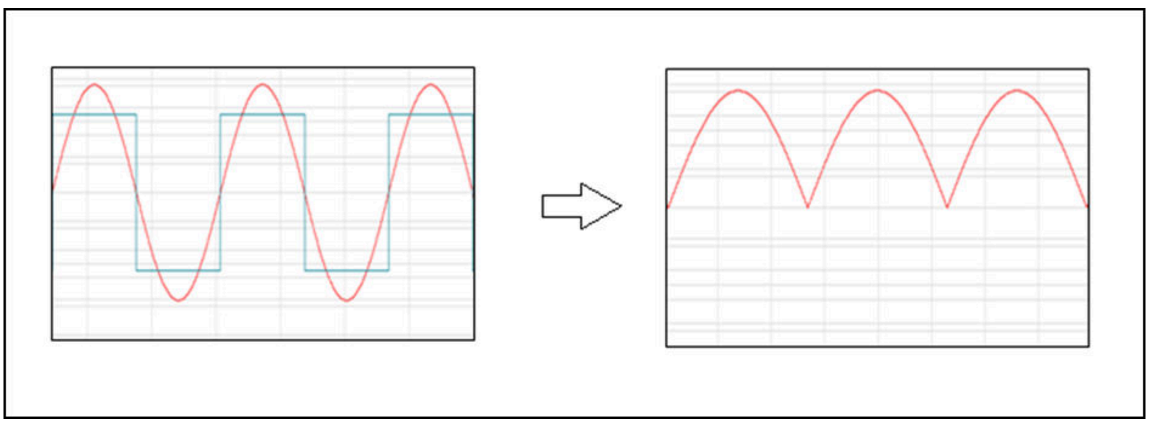


Figure 2-10. Square Wave Modulation

Let the average value of the output be V1,

$$V_1 = k * V_O * Cos(\alpha) \tag{14}$$

Similarly when V is multiplied by square wave with 90° phase, the resultant output is as shown in Figure 2-11.

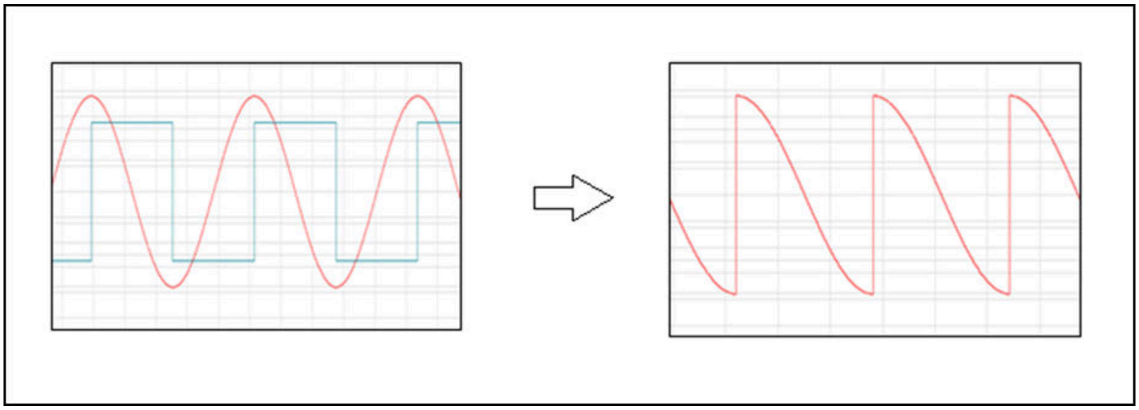


Figure 2-11. Square Wave Modulation with 90° phase

Let the average value of the output be V2,

$$V_2 = k * V_O * Sin(\alpha) \tag{15}$$

Where k is equal to 4/π.

Using Equation 16 and Equation 17, we get

$$|V| = Average\ of\ V = \sqrt[2]{{V_1}^2 + {V_2}^2} \tag{16}$$

$$\alpha = \tan^{-1}\left(\frac{V_2}{V_1}\right) \tag{17}$$

In this way, both the magnitude '|V|' and phase 'α' of any signal V is estimated.

3 Hardware, Software, Testing Requirements, and Test Results

3.1 Required Hardware and Software

3.1.1 Hardware

Figure 3-1 and Figure 3-2 illustrate the schematic and board connections of the TIDA-060029 board.

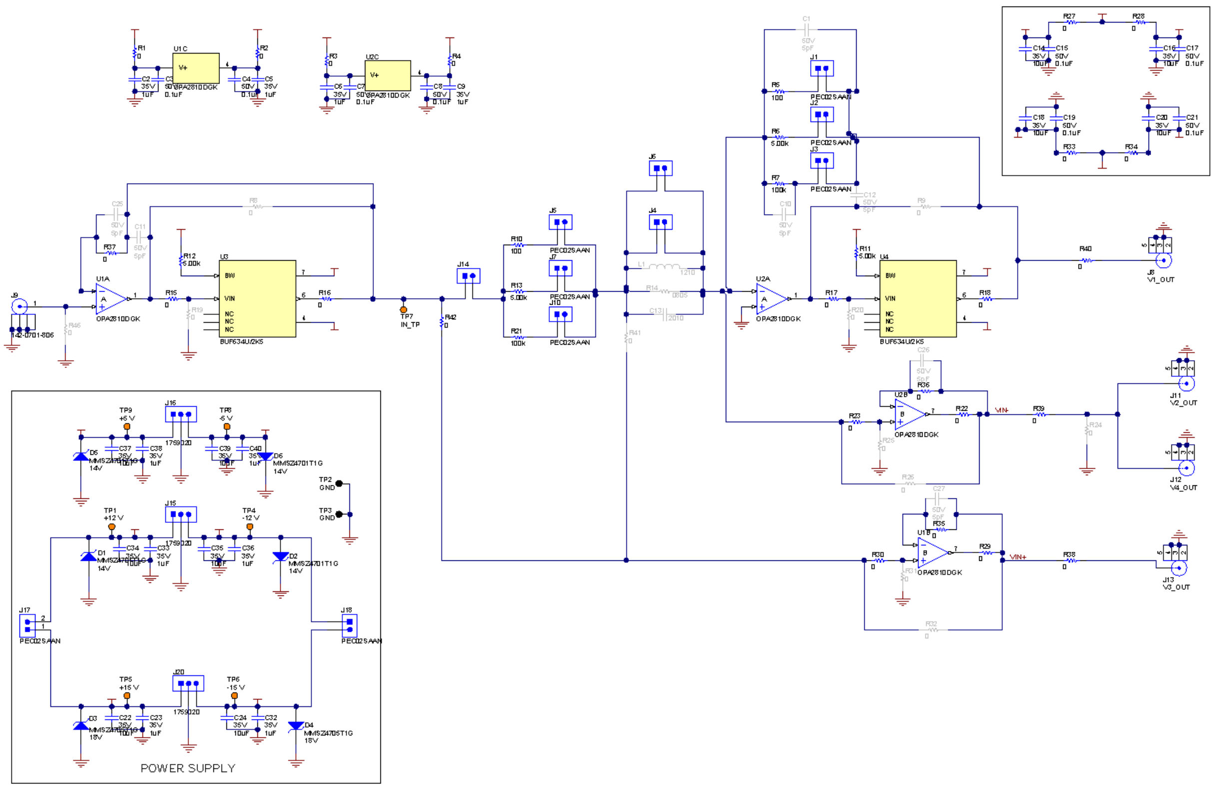


Figure 3-1. Hardware Schematic

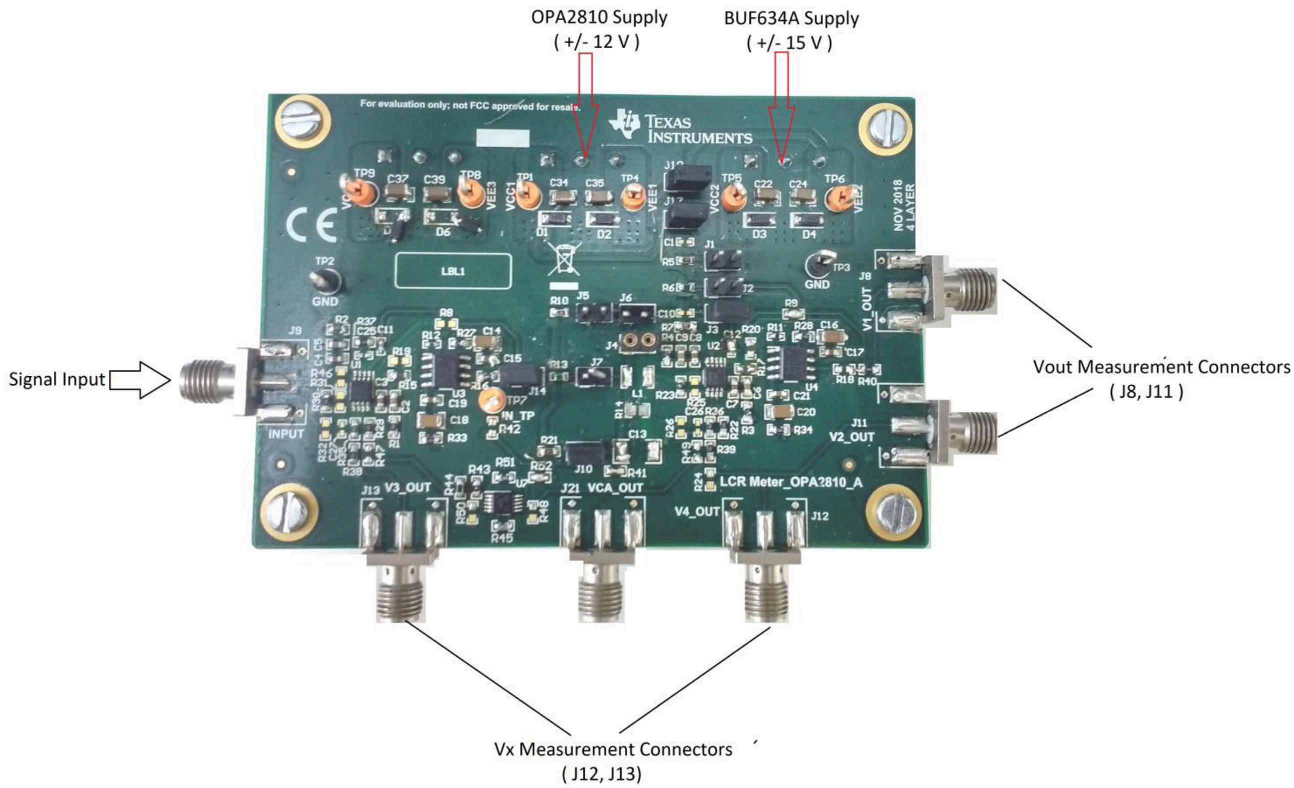


Figure 3-2. Board Connections

Table 3-1. Connector Details

CONNECTOR	DESCRIPTION
J1, J2, J3, J5, J7, J10	R_G = R_F Settings connectors
J6	Offset Calibration connector
J14	Offset Cal Connector
J15	OPA2810 Supply
J20	BUF634 Supply
J16	VCA 821 Supply
J17, J18	Jumpers between J15 and J20
J9	Input Connector
J12, J13	V_X Measurement Connectors
J8, J11	V_{OUT} Measurement Connectors

3.2 Testing and Results

3.2.1 Test Setup

1. Short the jumper J6. It should be noted that jumper J6 is for offset calibration and not used in this design.
2. Always keep one of the jumpers among J1, J2, J3 short according to R_G – R_F setting needed. This prevents the saturation of U2A.
3. Connect both +/- 12 V and +/-15 V power supplies to J15 and J20 respectively.
4. Set the required R_G – R_F setting, refer to Table 2-1 and Table 3-2 to determine and set desired connections.
5. Do calibration of each setting at all 4 frequencies according to the calibration process explained in Section 2.3.2.3. Use Table 3-2 to make connections according to required calibration.
6. Use the calibration results to estimate the unknown impedance as explained in Section 2.3.2.3.

Configurations for different connections on the board are given inTable 3-1

Table 3-2. Connector Configurations

CONDITION	CONNECTOR CONFIGURATION					
Short Calibration	J6 = Short, R41 = Open, R42 = Short					
Open Calibration	J6 = Open					
Impedance Calibration	J6 = Open					
RG = RF =100 Setting	J1	J2	J3	J5	J7	J10
	Short	Open	Open	Short	Open	Open
RG = RF =5 k Setting	J1	J2	J3	J5	J7	J10
	Open	Short	Open	Open	Short	Open
RG = RF =100 k Setting	J1	J2	J3	J5	J7	J10
	Open	Open	Short	Open	Open	Short

Table 3-3 provides recommended operating voltages on connectors.

Table 3-3. Operating Voltages

DESCRIPTION	RECOMMENDED VOLTAGE
OPA2810 Supply (J15)	+/- 12 V
BUF634A Supply (J20)	+/- 15 V
VCA Supply (J16)(NOT USED)	+/- 5 V

3.2.2 Test Results

The following example shows the unknown capacitive impedance measurement in detail.

Component : C = 100 nF

Measured value of the C = 99.472 nF

Frequency of Test = 1 kHz

R_G = R_F Setting = 100 Ω

Calibrated value of R_F = 99.97686

R_F/Z_X = 0.062412398 and α = 90.125° (phase of the ratio)

Thus, Z_X = 1601.875005 and θ_X = 90.125°

$$X_C = Z_X * Sin(\theta_x) \tag{18}$$

$$X_C = 1601.875 * \sin(-90.125) \tag{19}$$

$$X_C = 1601.87 \tag{20}$$

$$C = \frac{1}{2 * \pi * f * X_C} \tag{21}$$

Using Equation 21 we get, C = 99.3556 nF

$$\%\ Error = \frac{(99.472 - 99.356) * 100}{99.472} \tag{22}$$

Thus the % Error = 0.116 %

All other components are measured in the same way. The results are shown in Table 3-4.

It should be noted that the errors are estimated with respect to the value estimated by Keysight Technologies' E4980A precision LCR Meter. For testing, an input of 3.6 V_{pp} was used and results were measured with a seperate board utilizing the THS4551 and ADS9224R.

Table 3-4. Board Measurement Results

Parameters		**RG = RF Setting**					
Frequency (Hz)	**Component**	**100 Ω**	**Error(%)**	**5 kΩ**	**Error(%)**	**100 kΩ**	**Error(%)**
100	R	1 Ω – 900 Ω	0.74	500 Ω – 50 kΩ	0.11	10 kΩ – 10 MΩ	0.3
	L	1.59 mH – 2.38 H	1.18	2.27 H – 79.5 H	-	72.9 H – 1432 H	-
	C	1.05 µF – 1.59 mF	3	31.78 nF - 1.11 µF	0.62	1.76 nF - 34.7 nF	0.36
1k	R	1 Ω – 900 Ω	0.72	500 Ω – 50 kΩ	0.12	10 kΩ – 10 MΩ	0.56
	L	159 µH – 238 mH	0.47	227 mH – 7.95 H	-	7.29 H – 143.23 H	-
	C	106 nF – 159 µF	0.12	3.178 nF – 111 nF	0.39	176 pF – 3.47 nF	0.1
10k	R	1 Ω – 900 Ω	0.71	500 Ω – 50 kΩ	0.12	10 kΩ – 10 MΩ	2.49
	L	25.9 µH – 23.8 mH	0.57	22.6 mH – 795 mH	1.81	729 mH – 14.3 H	-
	C	10.6 nF – 15.9 µF	0.94	317.8 pF – 11.1 nF	0.4	17.6 pF – 347 pF	0.22
100k	R	1 Ω – 900 Ω	0.47	500 Ω – 50 kΩ	0.87	10 kΩ – 10 MΩ	14
	L	2.59 µH – 2.38 mH	0.71	2.26 mH – 79.6 mH	4.8	72 mH – 1.43 H	-
	C	1.06 nF – 1.59 µF	0.17	31.78 pF - 1.11 nF	1.8	1.76 pF- 34.7 pF	5.5

4 Design Files

4.1 Schematics

To download the schematics, see the design files at TIDA-60029.

4.2 Bill of Materials

To download the bill of materials (BOM), see the design files at TIDA-60029.

4.3 PCB Layout Recommendations

This design follows the guidelines found in the Layout section of the OPA2810 data sheet.

4.3.1 Layout Prints

To download the layer plots, see the design files at TIDA-60029.

4.4 Altium Project

To download the Altium Designer® project files, see the design files at TIDA-60029.

4.5 Gerber Files

To download the Gerber files, see the design files at TIDA-60029.

4.6 Assembly Drawings

To download the assembly drawings, see the design files at TIDA-60029.

5 Software Files

To download the software files, see the design files at TIDA-60029.

6 Related Documentation

1. Texas Instruments, OPA2810 Dual-Channel, 27-V, Rail-to-Rail Input/Output FET-Input Operational Amplifier data sheet
2. Texas Instruments, BUF634A 36-V, 210-MHz, 250-mA Output, High-Speed Buffer data sheet

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7 Revision History

NOTE: Page numbers for previous revisions may differ from page numbers in the current version.

Changes from Revision A (20200901) to Revision B (December 2021)	Page
• Updated Xc = 1599.99 to Xc = 1601.87	15

Changes from Revision * (June 2020) to Revision A (September 2020)	Page
• Changed *Capacitive Measurement with Series Resistance* image	7
• Changed *Square Wave Modulation* image	11