

TI Designs: TIDA-060020

HART Enabled PLC Analog Input Module Reference Design



Description

This reference design is a single-channel, 4-mA to 20-mA analog input module that supports the highway addressable remote transducer (HART) communication protocol, enabling bidirectional digital communication between field transmitters and programmable logic controllers (PLCs). The loop current, representing the primary variable, flows through a current-sense resistor producing a proportional voltage that is digitized by a precision delta-sigma analog-to-digital converter (ADC) (ADS1260). A DAC8740H HART® modem is used in this design to implement HART communication across the current loop. The ADS1260 and DAC8740H are interfaced with an external microcontroller that controls the ADS1260 sampling and includes a HART software stack to support the HART protocol. This design also features digital isolation, a wide power-supply range, and a high-precision external voltage reference for the ADC.

Features

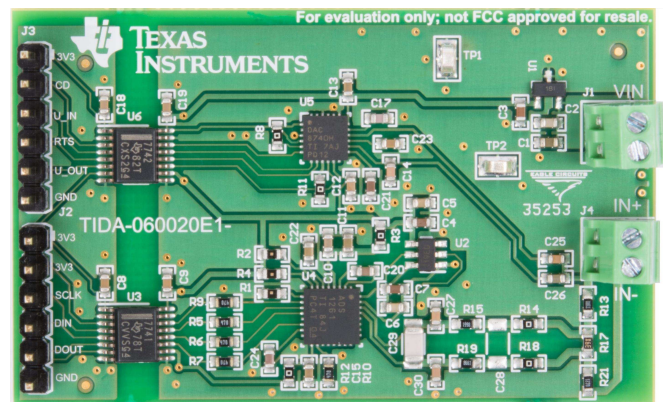
- 4-mA to 20-mA Analog Input With HART Modem
- High Accuracy Loop Current Measurement < 0.05% FSR TUE
- HART Physical Layer Compliance Tested
- 6-V to 28-V Power-Supply Input
- 20 Effective Bits of Resolution for Loop-Current Measurement

Applications

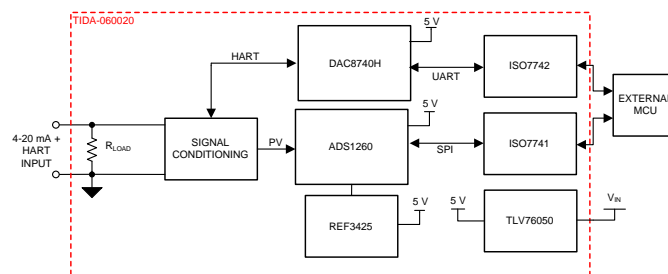
- [Factory Automation and Control](#)
- [Building Automation](#)

Resources

TIDA-060020	Design Folder
DAC8740H	Product Folder
ADS1260	Product Folder
REF3425	Product Folder
TLV760	Product Folder
ISO7741	Product Folder
ISO7742	Product Folder



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1 System Description

This analog input module reference design accepts traditional 4-mA to 20-mA inputs and supports HART enabled transmitters with the HART communication protocol. The key components are the ADS1260 for sensing the loop current and the DAC8740H HART modem, which provides digital frequency shift key (FSK) communication over the current loop. The input module acts as the HART master and requests data from the field transmitter (i.e. the HART slave). The ADC and HART modem are supported by a low-dropout (LDO) power supply, a precision voltage reference, and digital isolators. These components, along with the ADC and HART modem, create a complete system-level design for HART-enabled analog input modules. This design guide covers the key considerations to achieve HART compliance with highly accurate loop-current measurements.

1.1 Key System Specifications

[Table 1](#) lists key specifications for this reference design.

Table 1. Key System Specifications

PARAMETER	SPECIFICATIONS	DETAILS
Supply input voltage	6 V - 28 V	Section 2.3.4
Load resistance	260 Ω	Section 2.2.2
ADC effective resolution	20 bits	Section 3.2.2.9
ADC sampling rate	20 SPS	Section 2.2.3
Total unadjusted error (DC loop current)	<0.05% FSR at 25°C	Section 2.2.2
Current input range	3.5 mA - 23.5 mA	Section 2.2.2
Power consumption	220 mW at 24 V	Section 2.3.4
Board dimensions	58 mm x 36 mm	Section 4.3.1
HART signal amplitude	400 mV _{PP} to 800 mV _{PP}	Section 3.2.2.1
HART FSK frequencies	1200 Hz \pm 1%, 2200 Hz \pm 1%	Section 3.2.2.1
HART PHY compliance	Physical layer tests completed for low impedance device	Section 3.2.1.1

2 System Overview

2.1 Block Diagram

Figure 1 shows a block diagram of this reference design.

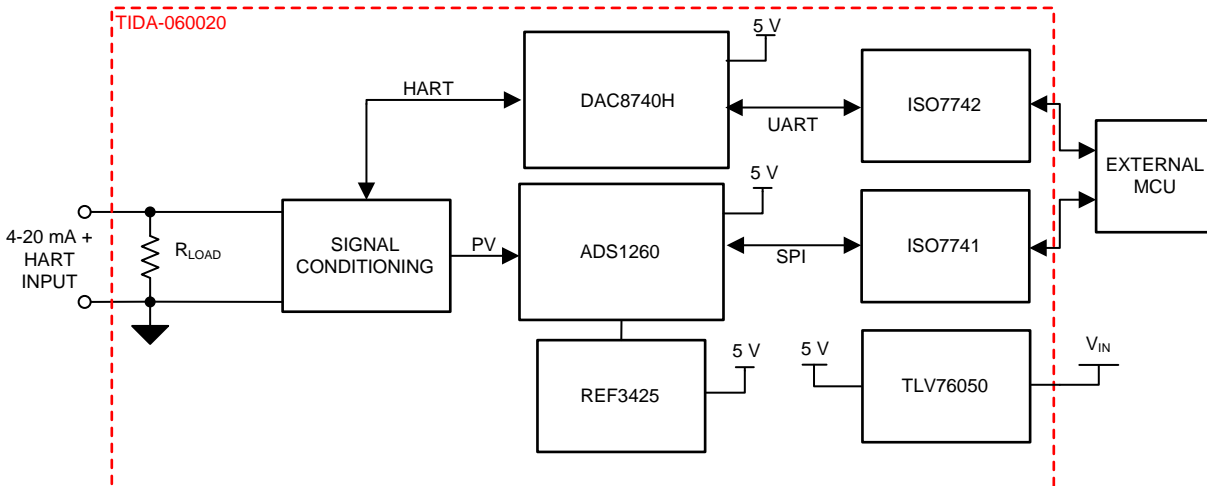


Figure 1. TIDA-060020 Block Diagram

2.2 Design Considerations

2.2.1 Overview of a HART Enabled Analog Input Module

In 4-mA to 20-mA systems, the analog input module is responsible for digitizing the loop current value by sensing the voltage across a known load resistor. The value of the DC loop current represents the primary variable and the transmitter regulates this current based on the measurement of a sensor. The analog input module can be used with both 2-wire and 3-wire loop current transmitters.

Figure 2 shows a simplified circuit diagram of a HART enabled analog input module. When the current flows across the load resistor, R1, a proportional voltage is induced across R1. This voltage is equal to the loop current multiplied by the load resistance. The ADC samples and converts this voltage into a digital value that can be read by a microcontroller. R2 and C1 form a simple low-pass filter to remove out-of-band noise present at the ADC input. C2 and C3 isolate the DC from the modem input and output to the high side of the load resistor.

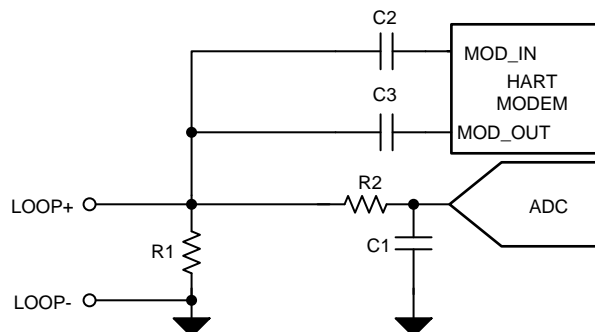


Figure 2. Simplified Analog Input Module

Because the transmitter regulates the loop current, the HART signal is transmitted as a 1-mA_{pp} current waveform that is AC-coupled to the DC loop current value. The input module directly couples the HART voltage signal across the load resistor when the input module is transmitting HART. The HART modem demodulates the voltage signal created across R1 by the 1-mA_{pp} HART current waveform.

Specific filtering is required to separate the analog signal from the HART signal at the analog input module. The analog signal is restricted to very low frequency, imposing a restriction on the transmitter analog rate of change. The HART modem requires a band-pass filter to remove both the analog signal and limit high frequency noise. Figure 3 shows a simplified illustration of the analog and HART signal bands.

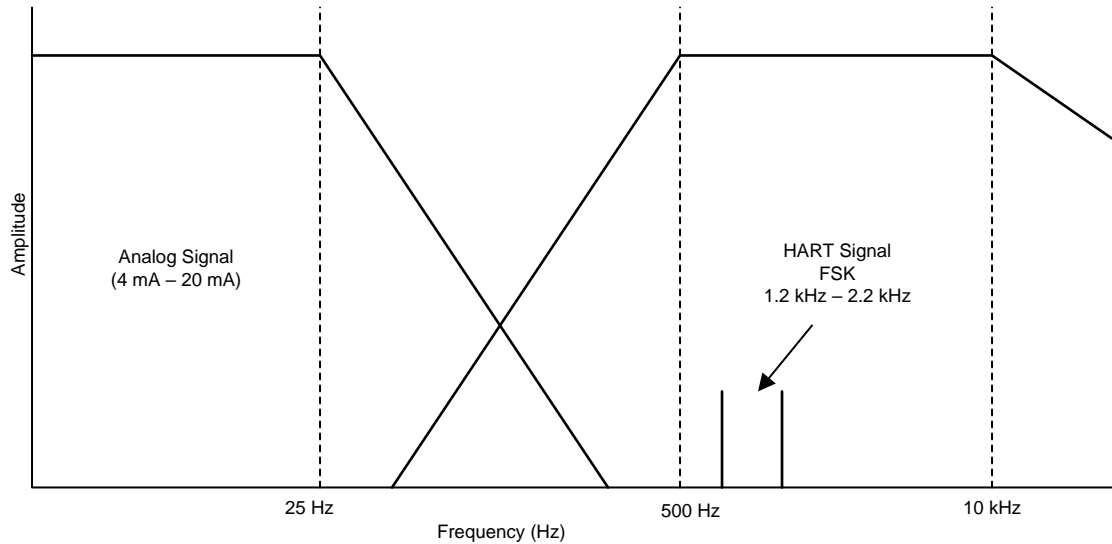


Figure 3. Analog Signal and HART Frequency Bands

2.2.2 Load Resistor Considerations

The load resistor is used both for sensing the loop current and for converting the HART current waveform into a voltage signal. The HART specification requires a load resistance between $230\ \Omega$ and $600\ \Omega$. The simplified configuration illustrated in Figure 2 uses a single load resistor but, at full-scale current, creates a high voltage at the input of the ADC that may exceed the absolute input range. At a 23.5-mA input, the voltage at LOOP+ is $250\ \Omega \times 23.5\ \text{mA} = 5.875\ \text{V}$, which is above the voltage range of many precision ADC inputs. The solution is to split the load resistance across multiple resistors in order to limit the voltage at the ADC inputs, but still maintain the total load resistance required for HART communication.

Figure 4 illustrates the load resistance split across three discrete resistors: R1, R2, and R3. In this configuration, the HART signal and the loop current are both present across the total load resistance ($R1 + R2 + R3$). The ADS1260 only measures the voltage across R2; therefore, R2 is the only resistor that requires high precision. Because of the high input impedance presented by the internal PGA, the high and low sides of R2 can be connected to the ADS1260 through only a discrete low-pass antialiasing filter. The amplifier characteristics of the PGA impose specific voltage range limitations in order to maintain linear operation. The combination of three resistors allows the design to adapt to varying load currents and input voltage requirements.

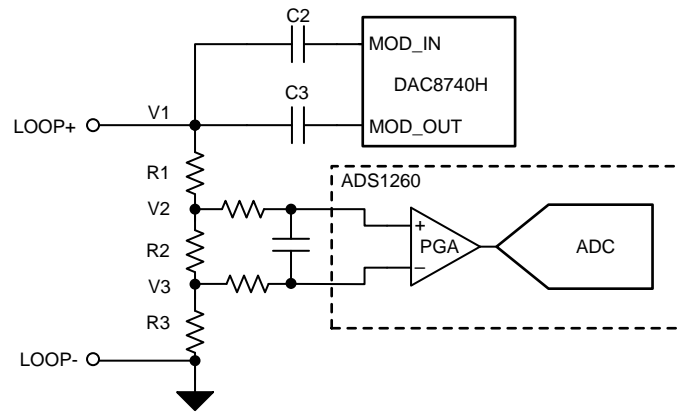


Figure 4. Schematic With Split Load Resistors

In addition to providing a high input impedance, the PGA is used to add gain to the differential input voltage before being converted by the ADC. This process ensures that the design uses most of the ADC full-scale range. Using the PGA also allows for a smaller sense resistor (R2), giving more flexibility when choosing R1 and R3 to meet the device input voltage requirements. Figure 5 shows the relationship between the voltages at the positive and negative PGA inputs (V_{AINP} and V_{AINN}) and their respective outputs (V_{OUTP} and V_{OUTN}). The outputs of the PGA amplifiers are limited to 300 mV away from each supply rail, which translates to a limitation on the absolute input voltage, differential input voltage, and usable PGA gain.

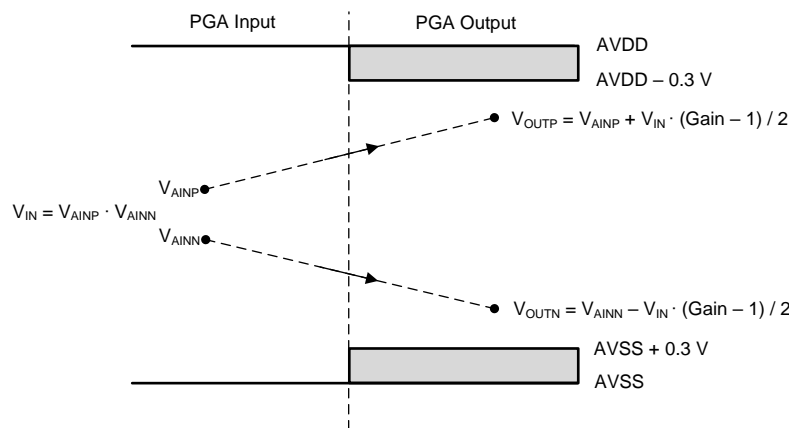


Figure 5. PGA Input and Output Range

The first step in determining R1, R2, and R3 is to determine the full-scale input range of the ADC. For most precision delta-sigma ADCs, the full-scale range depends on the reference voltage, V_{REF} . The ADS1260 can convert differential input voltages between $-V_{REF}$ and $+V_{REF}$. For devices with an internal PGA, the full-scale range must be referred to the input as $-V_{REF} / \text{Gain}$ to $+V_{REF} / \text{Gain}$. Select the PGA gain and R2 resistance value to span most of the full-scale range over the desired input current range. Using a PGA gain greater than 1 V/V not only reduces the required sense resistor value, but also reduces the input-referred noise of the ADC. The main advantage of using the PGA gain is reducing the sense resistance required, thus reducing self-heating and drift that contributes to error. Equation 1 shows the differential input voltage to the ADC after the internal PGA as a function of loop current.

$$V_{ADC} = I_{LOOP} \times R2 \times \text{GAIN} \tag{1}$$

This design uses a mid-supply reference voltage of 2.5 V from a REF3425. In order to support the maximum expected current of 23.5 mA, R2 must be less than $2.5 \text{ V} / 23.5 \text{ mA} = 106.4 \Omega$. Increasing the PGA gain to 4 V/V decreases the minimum R2 value to 26.6Ω , and also reduces the input-referred noise of the ADS1260. R2 is set to 24.9Ω to allow for some headroom, resulting in a span of 348.6 mV to 2.34 V into the ADC.

When the sense resistor is chosen, R3 must be selected based on Figure 5 to satisfy the PGA input requirements. Equation 2 calculates the absolute input voltage requirements for both PGA inputs (V_{AINP} , V_{AINN}). These limits are based on the analog supply rails (AVDD, AVSS), the differential input voltage ($V_{IN} = \text{Loop Current} \times R2$), and the gain of the PGA. The voltage at the negative PGA input (V_{AINN}) is equal to R3 times the loop current. The voltage at the positive PGA input is equal to $V_{AINN} + V_{IN}$. Equation 3 calculates the input limits for the minimum input current of 3.5 mA and Equation 4 calculates the input limits for the maximum input current of 23.5 mA.

$$AVSS + 0.3 \text{ V} + V_{IN} \times \frac{\text{Gain} - 1}{2} < V_{AINP}, V_{AINN} < AVDD - 0.3 \text{ V} - V_{IN} \times \frac{\text{Gain} - 1}{2} \tag{2}$$

$$0.4307 \text{ V} < V_{AINP}, V_{AINN} (3.5 \text{ mA}) < 4.569 \text{ V} \tag{3}$$

$$1.178 \text{ V} < V_{AINP}, V_{AINN} (23.5 \text{ mA}) < 3.804 \text{ V} \tag{4}$$

R3 is selected to ensure the voltage A_{INN} is above the lower limit at the minimum loop current of 3.5 mA. Equation 5 calculates the minimum R3 value as $430.7 \text{ mV} / 3.5 \text{ mA} = 123 \Omega$. A value of 127Ω allows for additional margin in the design. Equation 6 shows the voltage at the minimum loop current for the chosen resistance value of 127Ω . This value satisfies the minimum PGA requirement. R1 is finally selected to ensure the sum of the resistance is greater than 230Ω to meet the HART specification. Equation 7 shows the voltage at V2 based on R2, R3, and the maximum loop current. R1 must be at least 78Ω to account for the remaining required load resistance in order to comply with the HART protocol, 110Ω is therefore selected.

$$R3 \geq \frac{0.4307 \text{ V}}{3.5 \text{ mA}} \tag{5}$$

$$V3 (3.5 \text{ mA}) = 3.5 \text{ mA} \times 127 \Omega = 0.4445 \text{ V} \tag{6}$$

$$1.178 \text{ V} < V_{AINP}, V_{AINN} (23.5 \text{ mA}) < 3.804 \text{ V} \tag{7}$$

The PGA inputs must stay within the limits set by Equation 2 across the entire current input range. Checking the corner cases ensures that the minimum PGA input is not violated at the minimum input current and that the maximum PGA input is not violated at the maximum input current. Figure 6 illustrates A_{INN} , A_{INP} , and the PGA input limitations over the input current range. Figure 6 shows that the inputs are within range during all conditions.

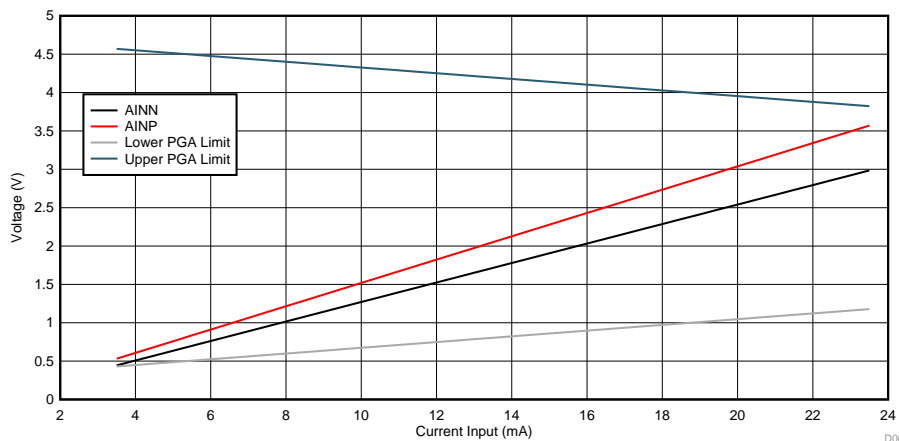


Figure 6. PGA Inputs vs Loop Current

2.2.3 ADC Input Filter and Internal Sinc Filter

The loop-current measurement encompasses two types of signal filtering: a discrete antialiasing filter and a digital decimation filter. The digital filter is responsible for low-pass filtering and decimating the samples from the delta-sigma modulator into a high-resolution result. The ADS1260 offers multiple sinc and finite impulse response (FIR) filter options with a variable frequency response based on the data rate, filter order, and clock frequency. The response of the digital filter returns to unity gain at multiples of the modulator sampling frequency, f_{MOD} .

The filter type and filter order both offer a tradeoff between conversion latency and stop-band attenuation. Lower latency allows for multiple sensors to be multiplexed into a single ADC with less delay, whereas lower stop-band attenuation reduces the level out-of-band signals that alias into the pass band, thus improving the noise performance of the system. In addition to noise, the filter in this design must reject the 1.2-kHz and 2.2-kHz FSK signals imposed onto the current loop by the field transmitter. Increasing the filter order to a second order lowers the stop-band attenuation to at least -80 dB at those frequencies and adds only one additional sample of conversion latency. Finally, the loop current is nearly a DC signal, which requires minimal signal bandwidth. This design uses an output data rate (f_{DR}) of 20 samples-per-second (SPS), yielding a -3 -dB bandwidth of 6.3 Hz. [Figure 7](#) shows the frequency response of the sinc2 filter with a data rate of 20 SPS.

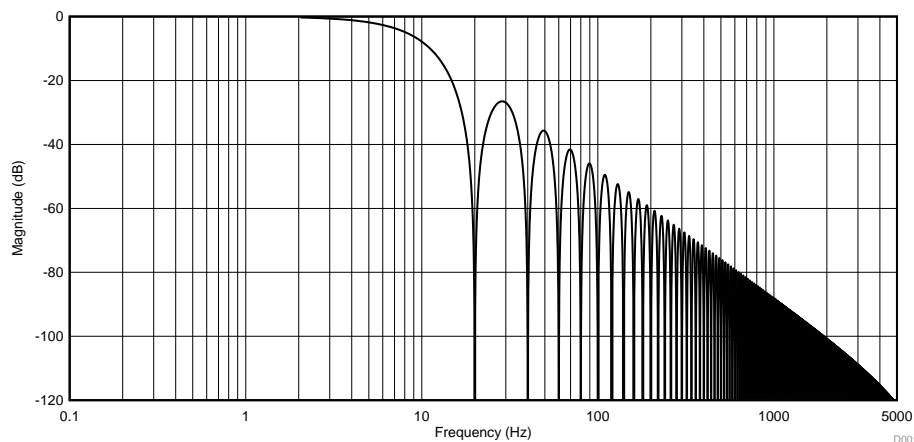


Figure 7. Frequency Response of ADS1260 Sinc2 Filter at 20 SPS

An analog antialiasing filter supplements the attenuation of the digital filter by rejecting both differential and common-mode noise. The second lobe of the sinc2 frequency response reaches approximately -36 dB and continues to roll off until $f_{MOD} / 2$ where the response repeats. At f_{MOD} , the digital filter response returns to unity gain, providing no attenuation. Because the stop-band rejection of the sinc2 filter is high, the discrete antialiasing filter has almost six decades in frequency before it is needed to attenuate noise around f_{MOD} . A single-pole, low-pass filter with a -3 -dB cutoff frequency conservatively placed around 100 Hz yields nearly -80 dB of rejection at a modulator frequency of 921.6 kHz, which is more than sufficient in this design. Common-mode capacitors that are each ten times smaller than the differential capacitor can be added from each channel input to ground to help filter common-mode noise without affecting the system common-mode rejection. [Figure 8](#) illustrates the simulation schematic used to simulate the frequency response of the antialiasing filter at the input of the ADS1260 PGA (labeled V_{IN}) and [Figure 9](#) illustrates the resulting frequency response. The analog antialias filter formed at the PGA output contributes an additional 24-dB rejection at f_{MOD} .

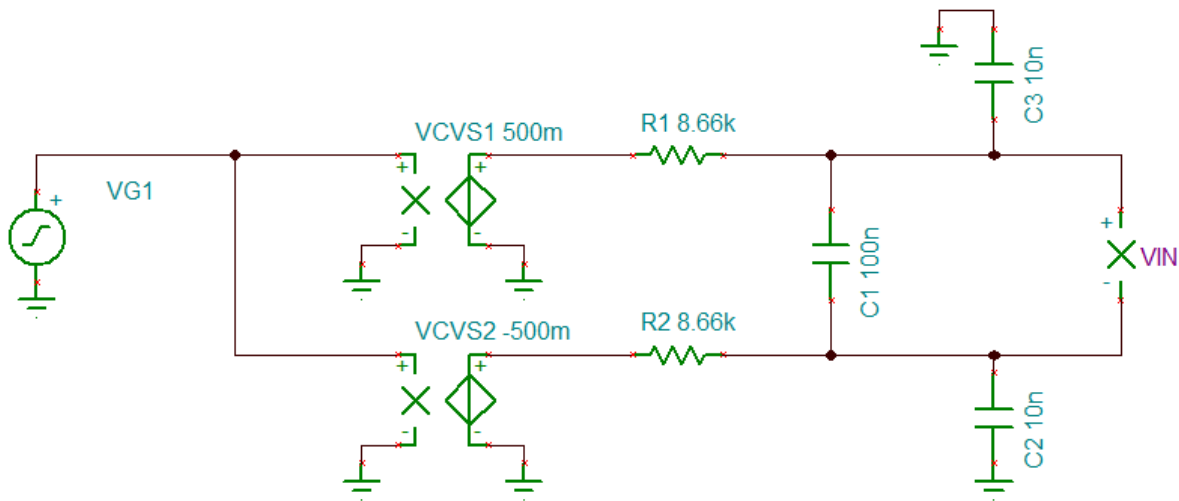


Figure 8. TINA-TI™ Simulation Schematic for Antialiasing Filters

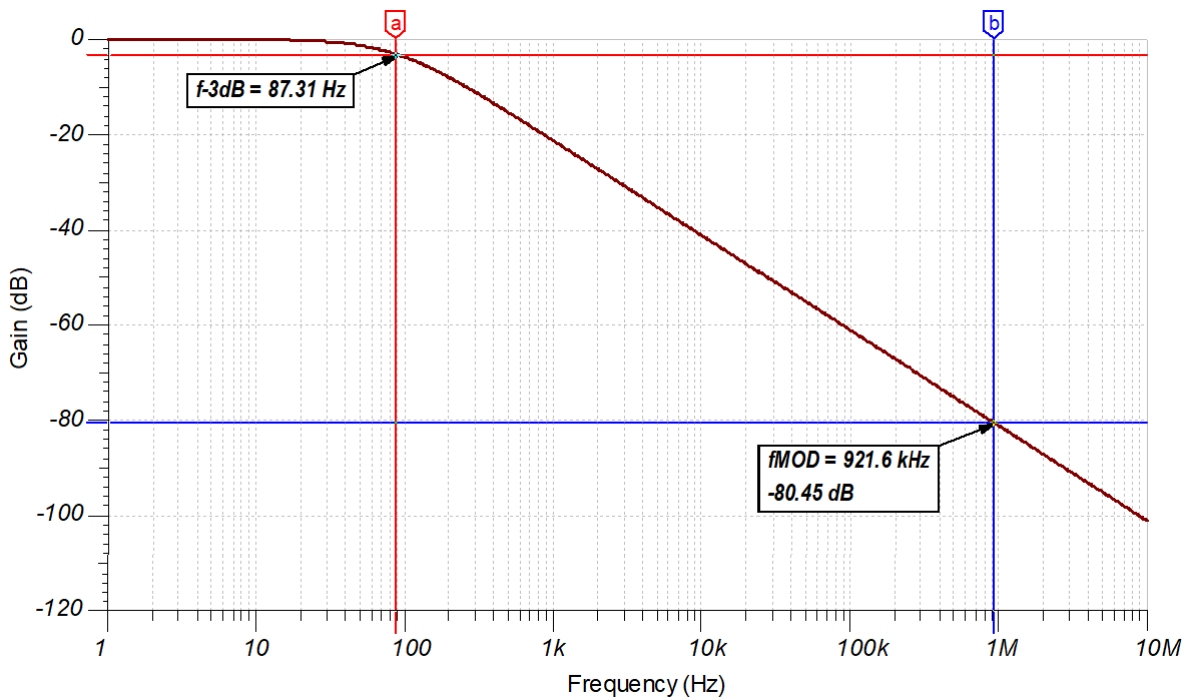


Figure 9. Low-Pass Antialiasing Filter Frequency Response

2.2.4 HART Considerations

The highway addressable remote transducer (HART) send and receive signals are AC-coupled directly to the load resistor. The input module receives the HART signal as a 1-mA_{PP} current waveform that is converted to voltage by the sense resistor and AC-coupled to the HART modem. Alternatively, when the input module transmits HART data a voltage signal is coupled across the sense resistor and the modulation of loop voltage is detected by the transmitter. The HART protocol is a half-duplex communication scheme and the devices have a master and slave relationship. The input module is the HART master and is responsible for requesting HART data from the current transmitter or slave device.

Figure 10 illustrates how the DAC8740H is connected in this input module design. MOD_OUT is coupled to the load resistor with a large capacitance value to avoid attenuating the HART signal. C1 and RL form a high-pass filter that can attenuate the HART signal if the cutoff frequency is too high. The cutoff frequency can be calculated based on the RC time constant formed by RL and C1.

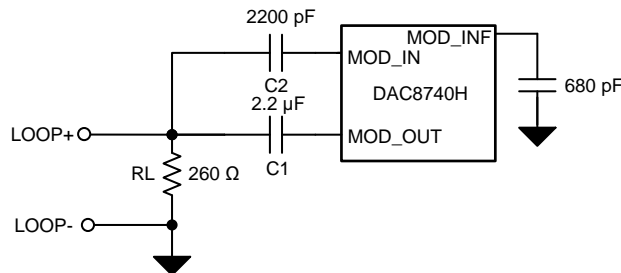


Figure 10. DAC8740H Input Module Coupling

The input HART signal is coupled to the DAC8740H MOD_IN pin by C2. The recommended value for C2 is 2200 pF because this capacitor affects the DAC8740H input filter response. Figure 11 shows the band-pass filter (BPF) formed by the AC-coupling capacitor internal resistors and the 680 pF on the MOD_INF pin. Figure 12 shows the passive filter response. This passive filter is followed by an internal second-order active high-pass filter (HPF) to meet the requirements for filtering the analog signal and high-frequency noise. The resulting filter has a 3rd-order, high-pass response below 500 Hz and a 1st-order, low-pass response above 10 kHz. The DAC8740H minimizes the external filtering components and filter design with this internal band-pass filter and ensures the HART filtering requirements are met.

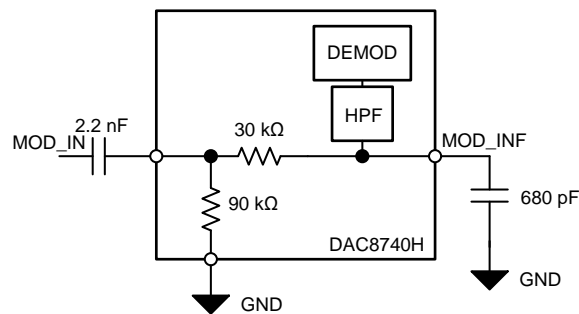


Figure 11. DAC8740H Internal Filter

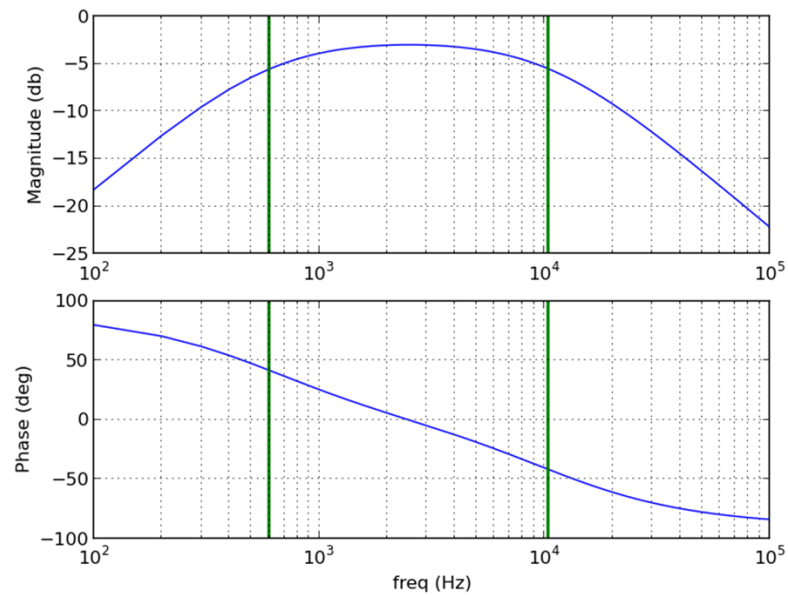


Figure 12. Passive BPF Response

The DAC8740H acts as a UART-to-HART direct feedthrough converter. The device transmits HART when the RTS ("Request-To-Send") pin is held low and interrupts the host processor with the CD pin when HART RX data are detected. In this design, an external TI MSP430FR5969 LaunchPad™ development kit is used to interface with the HART modem over the universal asynchronous receiver/transmitter (UART).

2.3 Highlighted Products

2.3.1 DAC8740H

The DAC8740H is a HART physical layer compliant modem with a UART interface. The DAC8740H integrates all required circuitry to operate as the physical layer HART modem in both slave and master configurations. The device features an internal reference, oscillator, and HART input filter requiring only a few passive components and a host processor for operation. The DAC874xH family of devices also function as FOUNDATION Fieldbus™ and Profibus PA modems.

2.3.2 ADS1260

The ADS1260 is a precision, 40-kSPS, delta-sigma, analog-to-digital converter with an integrated programmable gain amplifier (PGA) and internal fault monitors. The ADCs are comprised of an input signal multiplexer, a low-noise PGA providing a gain from 1 to 128, a 24-bit delta-sigma modulator, and a programmable digital filter. The high-impedance PGA inputs reduce measurement error caused by sensor loading. The ADS1260 supports three differential or five single-ended inputs. The ADS1260B integrates a 9-ppm/°C reference over a temperature range of 0°C to +85°C.

2.3.3 REF3425

The REF3425 device is a low temperature drift (6 ppm/°C), low-power, high-precision CMOS voltage reference, featuring $\pm 0.05\%$ initial accuracy and low operating current with power consumption less than 95 μA . This device also offers very low output noise of 3.8 $\mu\text{V}_{\text{PP}}/\text{V}$, which enables its ability to maintain high signal integrity with high-resolution data converters in noise-critical systems. With a small SOT-23 package, the REF3425 offers enhanced specifications and a pin-to-pin replacement for the MAX607x and ADR34xx devices. This reference was chosen because the REF3425 has better drift performance than the ADC internal reference. The REF3425 provides a balance between cost, accuracy, and drift.

2.3.4 TLV760

The TLV760 is an integrated linear-voltage regulator featuring operation from an input as high as 30 V. The TLV760 has a maximum dropout of 1.2 V at the maximum 100-mA load across operating temperature. Standard packaging for the TLV760 is the 3-pin, SOT-23 package. The TLV760 is available in 3.3-V, 5-V, 12-V, and 15-V options. The SOT-23 packaging of the TLV760 series allows the device to be used in space-constrained applications. The TLV760 is a small size alternative to the LM78Lxx series and similar devices.

2.3.5 ISO774X

The ISO774x devices are high-performance, quad-channel digital isolators with 5000 V_{RMS} (DW package) and 3000 V_{RMS} (DBQ package) isolation ratings per UL 1577. This family of devices has reinforced insulation ratings according to VDE, CSA, TUV, and CQC. The ISO774x devices provide high electromagnetic immunity and low emissions at low power consumption.

3 Hardware, Software, Testing Requirements, and Test Results

3.1 Required Hardware and Software

3.1.1 Hardware

This reference design features multiple headers to externally interface the onboard DAC8740H HART modem and the ADS1260 ADC through digital isolators. The ADS1260EVM is used to interface with the ADC to test the input current measurement aspects of the design. The DAC8740H is interfaced with both the DAC8740HEVM and an MSP430FR5969 LaunchPad™ containing a HART software stack to conduct the physical layer HART testing. The additional external test equipment required includes an oscilloscope, 8.5 digit digital multimeter (DMM), precision current source, HART filter tools, and HART test modem.

3.2 Testing and Results

3.2.1 Test Setup

3.2.1.1 HART Physical Layer Test Setup

Figure 13 shows the test setup for most physical layer HART test cases. This test setup is used to evaluate the reference design. Any waveform measurements or scope captures are taken across the test load. The current source and DC offset are set to replicate 4-mA to 20-mA operating conditions. Section 3.2.2 specifies if another test setup is used.

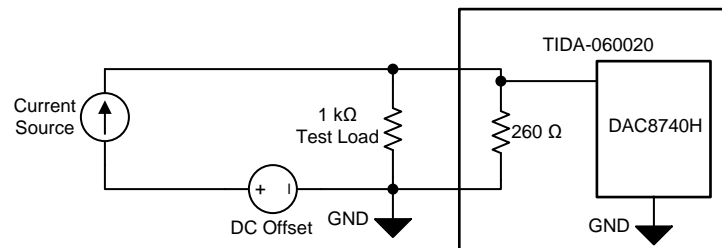


Figure 13. HART Current Input Device Test Setup

3.2.1.2 DC Current Test Setup (ADC)

Figure 14 shows how the ADC is tested. The DAC8760EVM is used as a precision DC current source to produce the loop current through the load resistors. An 8.5 digit DMM is used to measure the voltage across the current-sense resistor and calculate the loop current from the measured sense resistance. The sense resistor was measured using a 4-wire measurement for accuracy. This measured current value is compared to the value sampled by the ADC, producing a total unadjusted error (TUE) curve. Using the DAC8760EVM allows the HART signal to be coupled, thus ensuring the HART waveform does not impact the DC current measurement by the ADC.

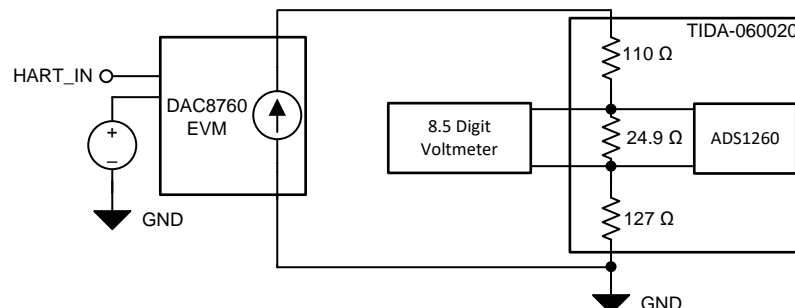


Figure 14. ADC Test Setup

3.2.2 Test Results

This section contains the test results for the HART physical layer tests and the ADC DC current measurements.

3.2.2.1 HART Waveform Test Results

Figure 15 and Figure 16 show the measured waveforms across the test load for both the 1200-Hz and 2200-Hz HART signals, respectively. Table 2 shows the measured amplitude and frequency compared to the pass criteria for a low impedance device.

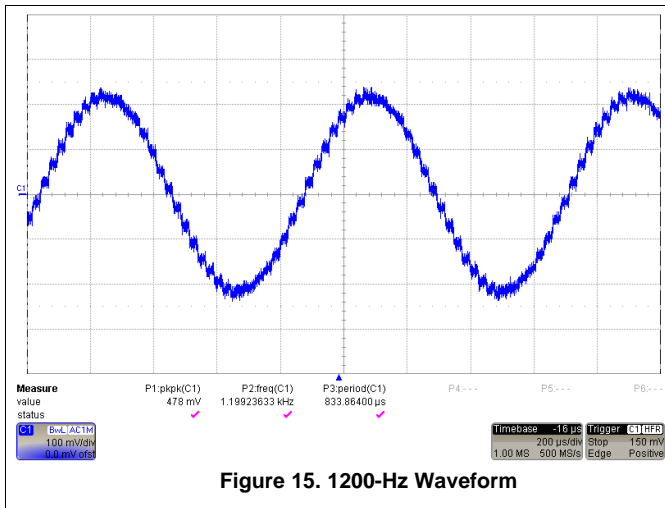


Figure 15. 1200-Hz Waveform

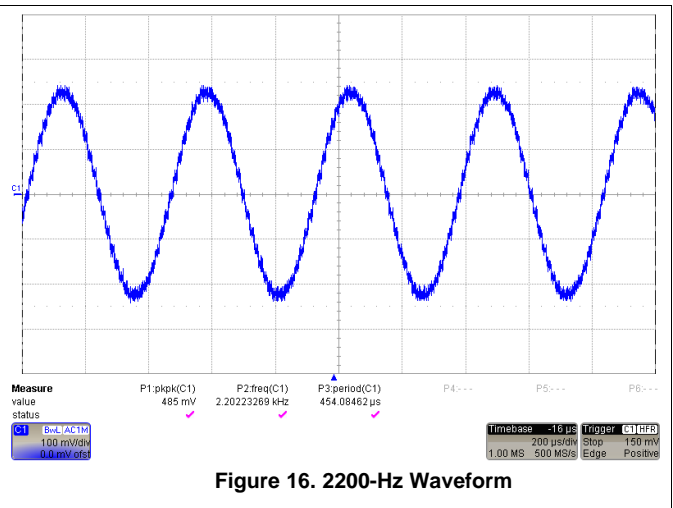


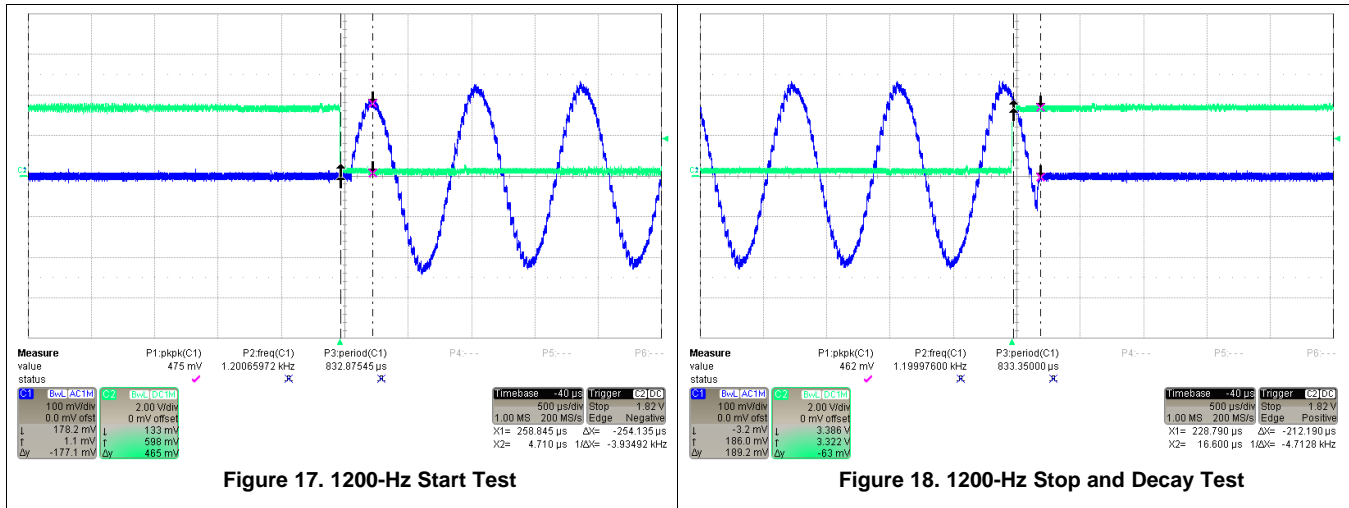
Figure 16. 2200-Hz Waveform

Table 2. HART Waveform Test Results

TEST	MEASURED	PASS CRITERIA
1200-Hz waveform amplitude	478 mV	400 mV - 800 mV; low Z
2200-Hz waveform amplitude	485 mV	400 mV - 800 mV; low Z
1200-Hz waveform frequency	1199.2 Hz	1188 Hz - 1212 Hz
2200-Hz waveform frequency	2202.2 Hz	2179 Hz - 2222 Hz

3.2.2.2 HART Carrier Start, Stop, Decay Test Results

The carrier must start within 5 bit times (1 bit time = 1 period at 1200 Hz), stop within 3 bit times, and decay within 6 bit times. The carrier is tested at both 1200 Hz and 2200 Hz. Figure 17 and Figure 18 show an example of the start and stop results with a 1200-Hz carrier waveform. Channel 1 is measured at the test load and channel 2 is the RTS signal. For both carrier frequencies, the stop, start, and decay tests were less than 1 bit time, satisfying the requirement.



3.2.2.3 Carrier Start and Stop Transient Test Results

The carrier start and stop transient test ensures that a large spike does not occur in the analog signal domain when the HART communication begins and ends. Figure 19 shows the test setup. The analog test filter is connected to the reference design and the output is displayed on the oscilloscope. The output through the filter must not create a spike greater than 100 mV. Figure 20 and Figure 21 illustrate the start and stop transients. There is no noticeable transient at the output of the analog test filter. Channel 1 is measured across the test load, channel 2 is the RTS pin, and channel 3 is the voltage after the analog filter.

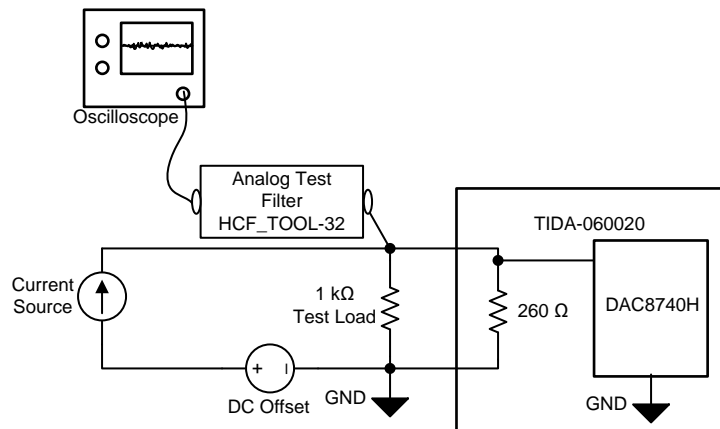


Figure 19. Carrier Start and Stop Transient Test Setup

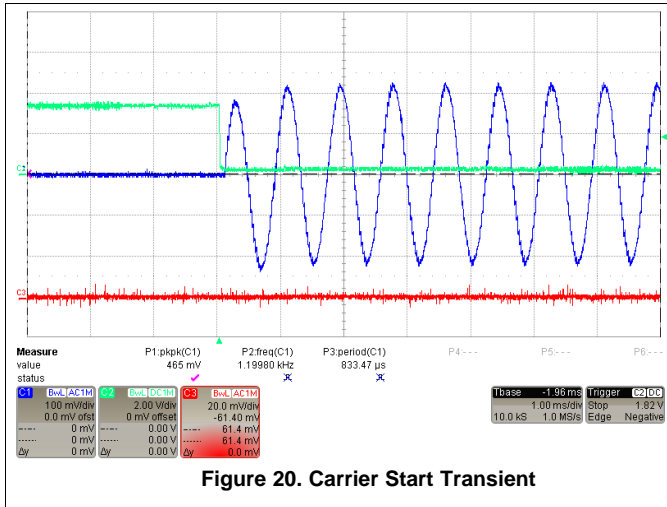


Figure 20. Carrier Start Transient

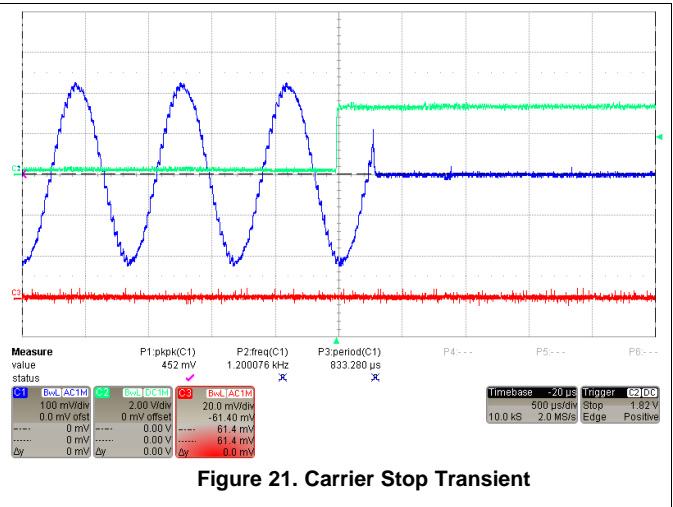


Figure 21. Carrier Stop Transient

3.2.2.4 HART Output Noise During Silence Test Results

The output noise during silence test ensures that the input module does not contribute unacceptable noise levels to the loop when HART is not being transmitted. Figure 22 shows the test setup. The TIDA-060020 is configured in an inactive state. No loop current is flowing to ensure the measured noise is the only contribution from the TIDA-060020 circuitry. First, the oscilloscope is connected to the reference design directly to measure broadband noise levels. Then the oscilloscope is connected through a digital test filter to measure in-band noise levels. Figure 23 illustrates the scope capture of the noise. Channel 1 shows the broadband noise without the digital test filter and channel 2 shows the in-band noise with the digital test filter. If the measured broadband noise without the filter is less than $2.2\text{mV}_{\text{RMS}}$, then measuring the in-band noise through the digital test filter is not required. Table 3 illustrates the measured noise results compared to the pass criteria.

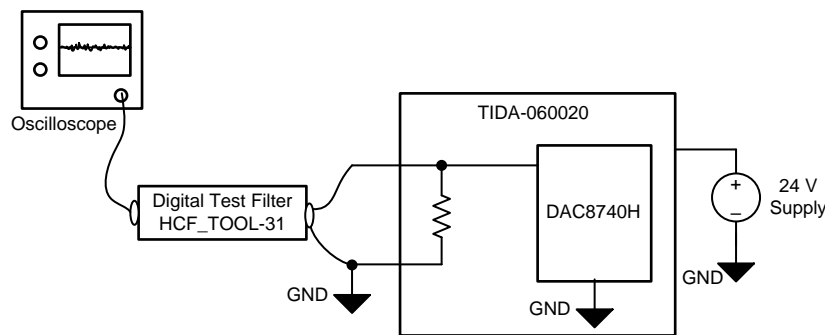


Figure 22. Output Noise During Silence Test Setup

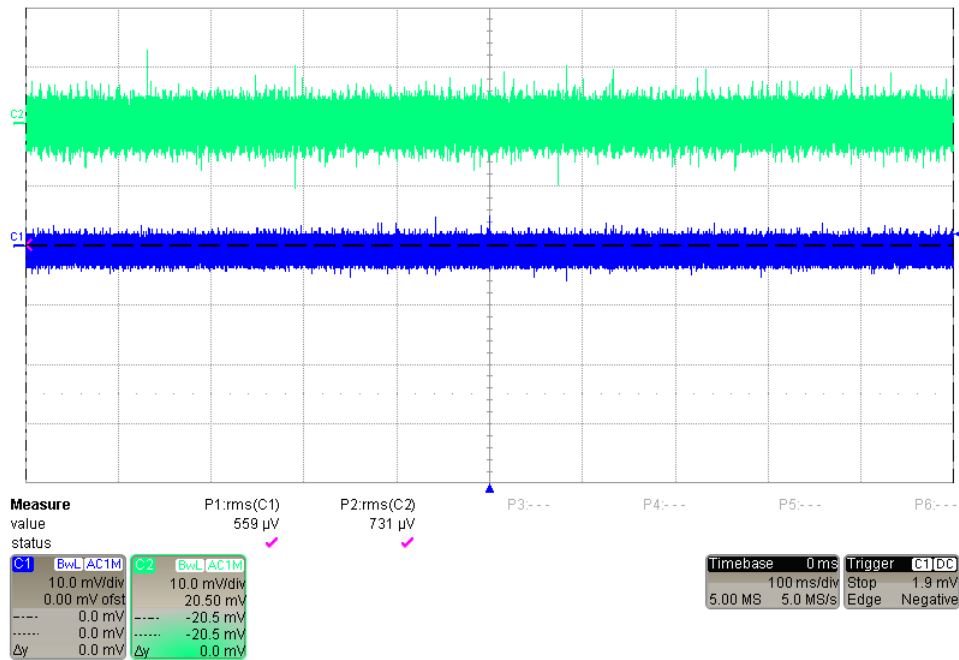


Figure 23. Output Noise During Silence Scope Capture

Table 3. Output Noise During Silence Test Results

TEST	MEASURED	PASS CRITERIA
Broadband noise (no filter)	559 μV_{RMS}	138 mV_{RMS} max
In-band noise (500 Hz to 10 kHz) digital test filter (10x gain)	731 μV_{RMS}	22 mV_{RMS} max

3.2.2.5 HART Receive Impedance Test Results

The receive impedance is the input impedance of the tested device. The impedance affects the HART signal integrity and can reduce the distance over which HART can be used. Figure 24 shows the test setup for receive impedance. A sinusoidal AC source generates a voltage and the amplitude is measured twice: first across a 250- Ω test load (V_a) and again across the input of the TIDA-060020 (V_b). The impedance is calculated by $Z = (250 \Omega / V_a) \times V_b$. Table 4 lists the measured values of V_a , V_b , and the calculation of Z at the required frequencies. Figure 25 illustrates a plot of the receive impedance. A parallel combination of $R = 261 \Omega$ and $C = 13 \text{ nF}$ approximates the receive impedance of TIDA-060020.

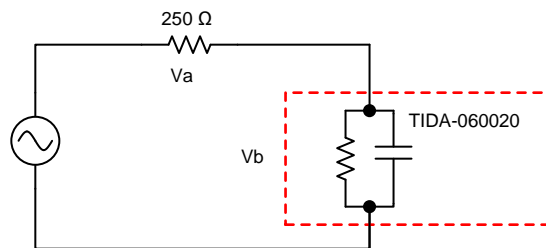
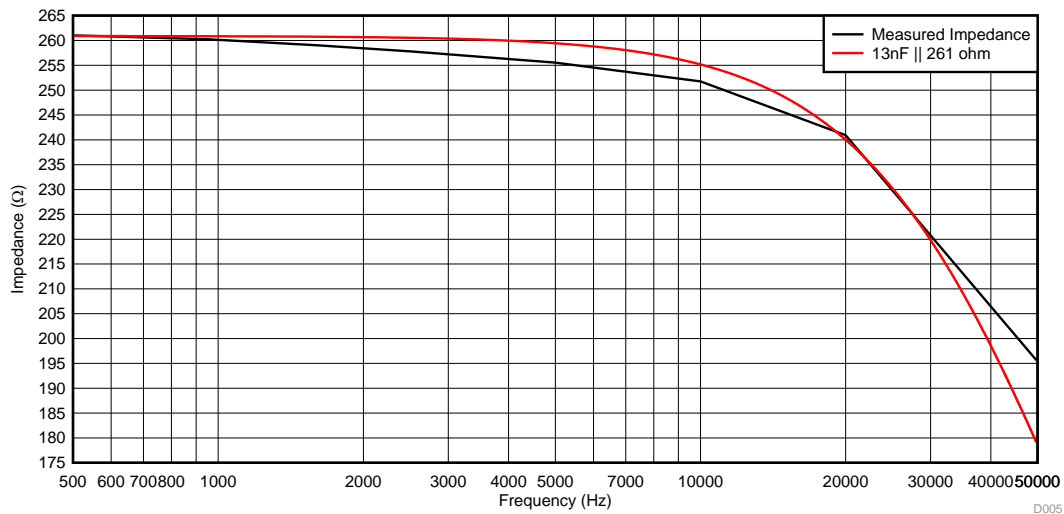


Figure 24. Receive Impedance Test Setup

Table 4. Receive Impedance Measurements

FREQUENCY (Hz)	Va (V RMS)	Vb (V RMS)	Z (Ω)
500	0.3529	0.3710	261.0
950	0.3534	0.3704	260.2
1.6k	0.3542	0.3695	259.0
2.5k	0.3550	0.3686	257.8
5k	0.3567	0.3671	255.5
10k	0.3600	0.3560	251.7
20k	0.3704	0.3594	240.9
50k	0.4217	0.3318	195.4


Figure 25. Receive Impedance vs Frequency

The receive impedance must be between 230 Ω and 600 Ω between 950 Hz and 2500 Hz. The receive impedance over the extended frequency band of 500 Hz to 10 kHz must have a difference between the maximum and minimum value difference less than 6 dB (factor of 2). The receive impedance of the TIDA-060020 decreases by 9.3 Ω over the extended frequency band.

3.2.2.6 Send Impedance Test Results

Figure 26 shows the send impedance test setup. The output of the TIDA-060020 is connected to an external test load resistor and the 1200-Hz and 2200-Hz waveforms are produced by the DAC8740H. The test is repeated for two load resistance values. V1 is the voltage measured across a 10-kΩ test load and V2 is the voltage measured across a 1-kΩ test load. Equation 8 calculates the send impedance as the change in voltage across the test load (ΔV) divided by the change in current (ΔI). Table 5 shows the results for the 1200-Hz and 2200-Hz tests. The send impedance is required to be less than 80 percent of the receive impedance.

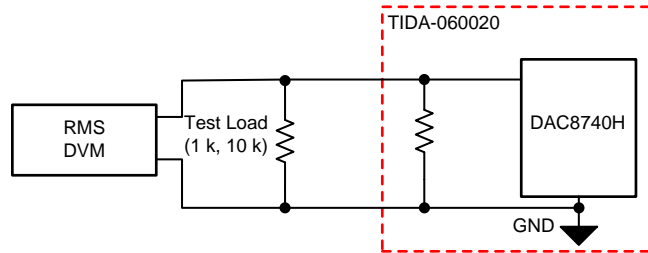


Figure 26. Send Impedance Test Setup

$$Z_{OUT} = \frac{\Delta V}{\Delta I} = \frac{V1 - V2}{\left(\frac{V2}{1\text{ k}\Omega}\right) - \left(\frac{V1}{10\text{ k}\Omega}\right)} \tag{8}$$

Table 5. Send Impedance Measurements

TEST	RMS VOLTAGE (V _{RMS})	FREQUENCY (Hz)	Z _{OUT} (Ω)
10 kΩ, 1200 Hz (V1)	0.1573	1200	32.5
10 kΩ, 2200 Hz (V1)	0.1624		
1 kΩ, 1200 Hz (V2)	0.1528	2200	18.9
1 kΩ, 2200 Hz (V2)	0.1597		

3.2.2.7 Noise Sensitivity Test Results

Figure 27 shows the noise sensitivity test setup. This test is designed to ensure HART communication can still occur with noise added to the loop. Noise of different frequencies is injected by a signal generator and adjusted so the specified amplitude is detected across the load resistor of the TIDA-060020. A HART test modem is then used with a waveform amplitude of 175 mV_{PP} to generate 100 consecutive response request exchanges with the noise on the loop. No bit errors occurred in this testing of the TIDA-060020. Table 6 lists the noise sensitivity levels. In the case of the 29-Hz and 63-Hz frequency levels, the amplitude is restricted to 0.9 V so as not to exceed the minimum HART-enabled operating current of 3 mA. Figure 28 illustrates an example of setting the sinusoidal noise level across the load resistor before generating the response and request sequence.

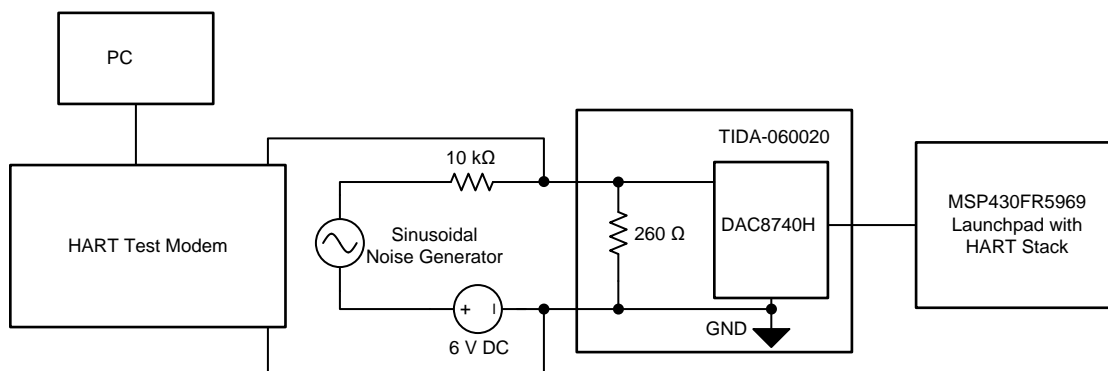


Figure 27. Noise Sensitivity Test Setup

Table 6. Noise Sensitivity Levels

AMPLITUDE	FREQUENCY
55 mV _{PP}	1.7 kHz
220 mV _{PP}	250 Hz
880 mV _{PP}	125 Hz
3.52 V _{PP} (limited to 0.9 V _{PP})	63 Hz
16 V _{PP} (limited to 0.9 V _{PP})	29 Hz

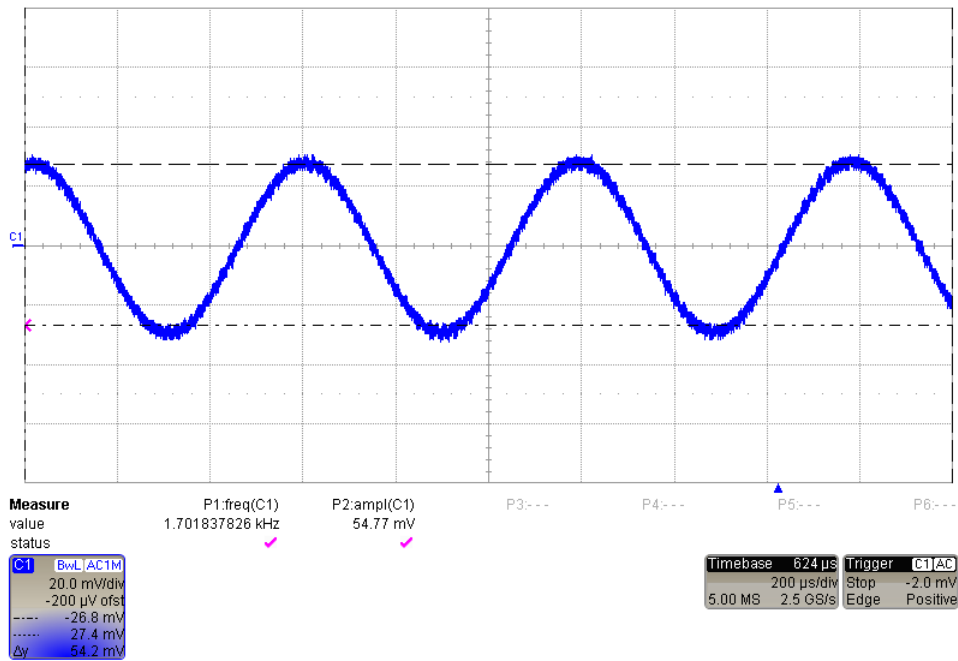
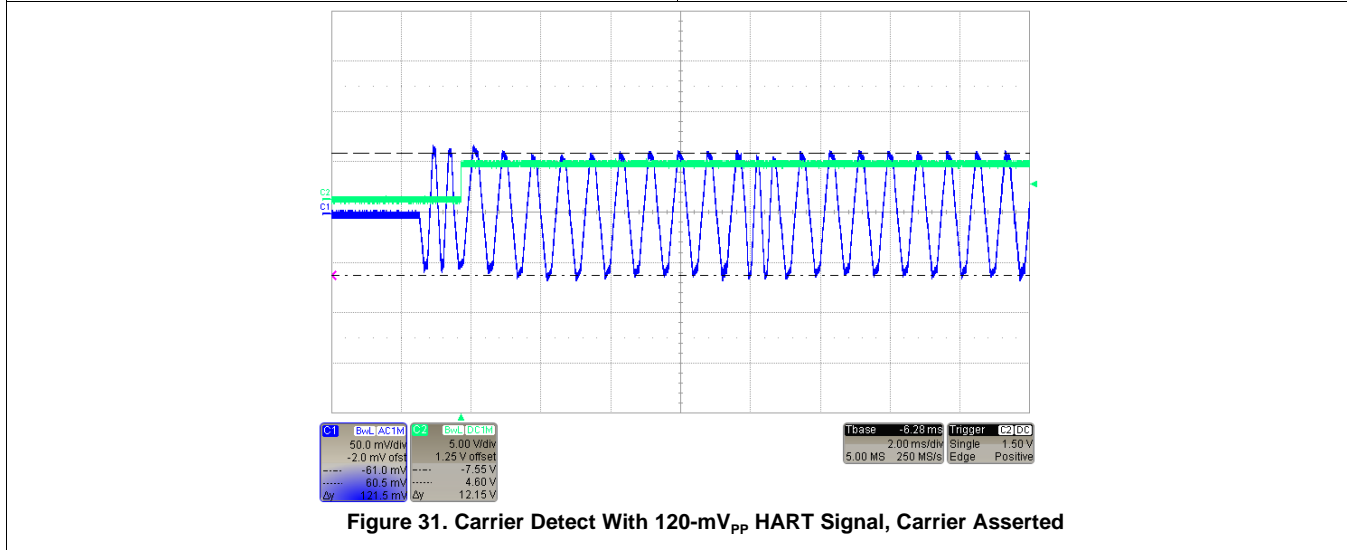
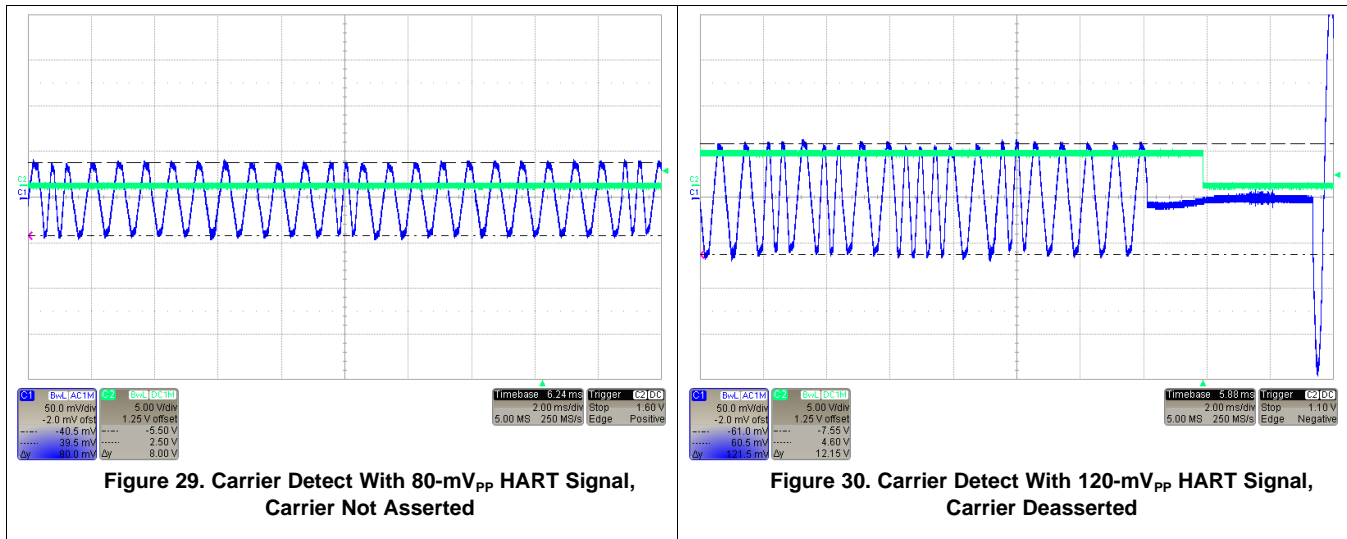


Figure 28. Example of Setting the Noise Level Across R_{LOAD}

3.2.2.8 Carrier Detect Tests

The carrier detect tests ensure that the modem does not attempt to demodulate a HART signal below 80 mV_{PP} and always demodulates HART signals above 120 mV_{PP}. To perform this test, the setup from the noise sensitivity test setup is used and the test modem is set to an amplitude of 80 mV_{PP} and 120 mV_{PP}. The carrier detect pin is monitored and is not asserted in the 80-mV_{PP} case but is asserted in the 120-mV_{PP} case. The time from the carrier on to the carrier detect assertion and the time from the carrier off to the carrier detect deassertion must be no more than 6 bit times.

Figure 29 shows that the carrier detect is not asserted when the test is run with a HART signal amplitude of 80 mV_{PP}. Figure 31 shows that the carrier detect is asserted after applying a HART signal amplitude of 120 mV_{PP}. Figure 30 shows that the carrier detect is deasserted after same 120 mV_{PP} carrier is turned off. Both Figure 31 and Figure 30 demonstrate approximately a 1 bit time delay.



3.2.2.9 DC Current Measurement Test Setup (ADC)

To measure the accuracy of the ADC DC loop current measurement, the test setup shown in Section 3.2.1.2 is used. The DAC8760 sets the DC loop current in steps of 500 μA . The voltage across the sense resistor is measured by both the 8.5 digit DMM and the ADC. The TUE is calculated based on Equation 9, where $I_{\text{LOOP,ADC}}$ is the current calculated from the ADC voltage reading and $I_{\text{LOOP,DMM}}$ is the current calculated from the voltage measured by the 8.5 digit DMM. Figure 32 plots the TUE over the input current range of the design. The maximum error is 0.007% of the full-scale range (FSR) at room temperature.

$$\text{TUE}(\% \text{FSR}) = \frac{I_{\text{LOOP,ADC}} - I_{\text{LOOP,DMM}}}{23.5 \text{ mA} - 3.5 \text{ mA}} \times 100 \tag{9}$$

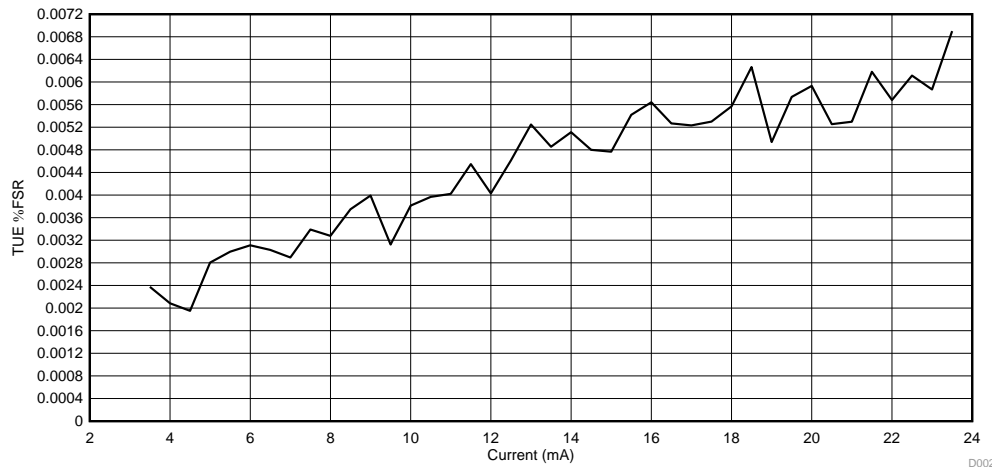
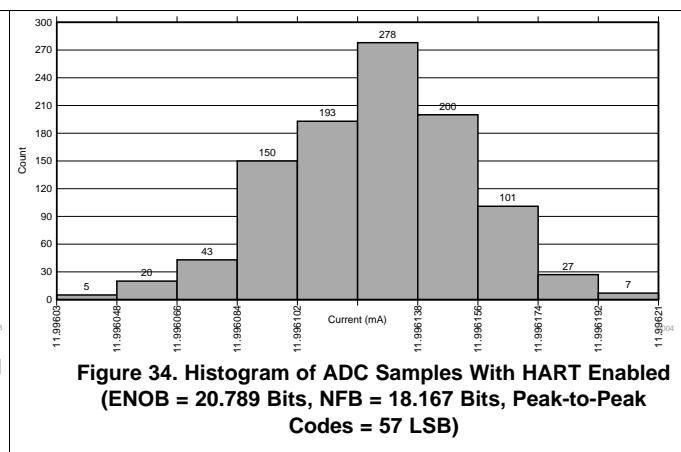
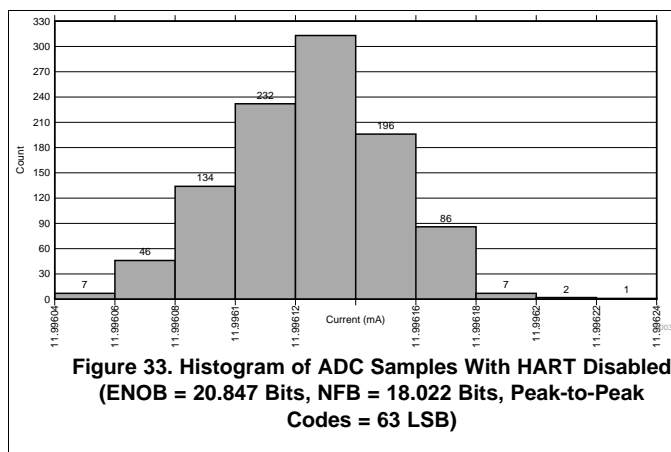


Figure 32. Current TUE vs Loop Current

The ADC performance was evaluated by taking 1024 samples of the voltage across the sense resistor with the loop current set to mid-scale (i.e. 12 mA). This evaluation is first done with HART disabled and then repeated with the HART enabled.. This process ensures that the HART signal which is superimposed on the loop current is being filtered out effectively and is not affecting the DC current measurement. Figure 33 and Figure 34 plot a histogram of the DC measurements taken with HART disabled and with HART enabled, respectively. No significant difference is observed when the HART is enabled, confirming that the combination of analog and digital filtering sufficiently attenuates the HART signal from the loop current.



4 Design Files

4.1 Schematics

To download the schematics, see the design files at [TIDA-060020](#)

4.2 Bill of Materials

To download the bill of materials (BOM), see the design files at [TIDA-060020](#).

4.3 PCB Layout Recommendations

4.3.1 Layout Prints

To download the layer plots, see the design files at [TIDA-060020](#).

4.4 Altium Project

To download the Altium Designer® project files, see the design files at [TIDA-060020](#).

4.5 Gerber Files

To download the Gerber files, see the design files at [TIDA-060020](#).

4.6 Assembly Drawings

To download the assembly drawings, see the design files at [TIDA-060020](#).

5 Software Files

To download the software files, see the design files at [TIDA-060020](#).

6 Related Documentation

1. [TI Designs: TIDA-01504 Highly-Accurate, Loop-Powered, 4mA to 20mA Field Transmitter with HART Modem Reference Design](#)

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