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TI Designs — Precision: Verified Design

Analog Linearized 3-Wire PT1000 RTD to 2-Wire 4-20mA Current Loop Transmitter Reference Design

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Design Resources

TIPD202	All Design files
XTR112	Product Folder

Circuit Description

This 2-wire (loop-powered) 4-20mA sensor transmitter uses analog linearization methods to provide a simple single-chip solution to interface 3-wire PT1000 resistive temperature detectors (RTD) to 4-20mA current loops. The design is configured for a -50°C to 250°C RTD temperature, but other ranges can be achieved using the equations in the Component Selection Section. The design operates over a wide supply range of +7.5 V – 40 V and includes circuit protection to pass the IEC61000-4 tests.

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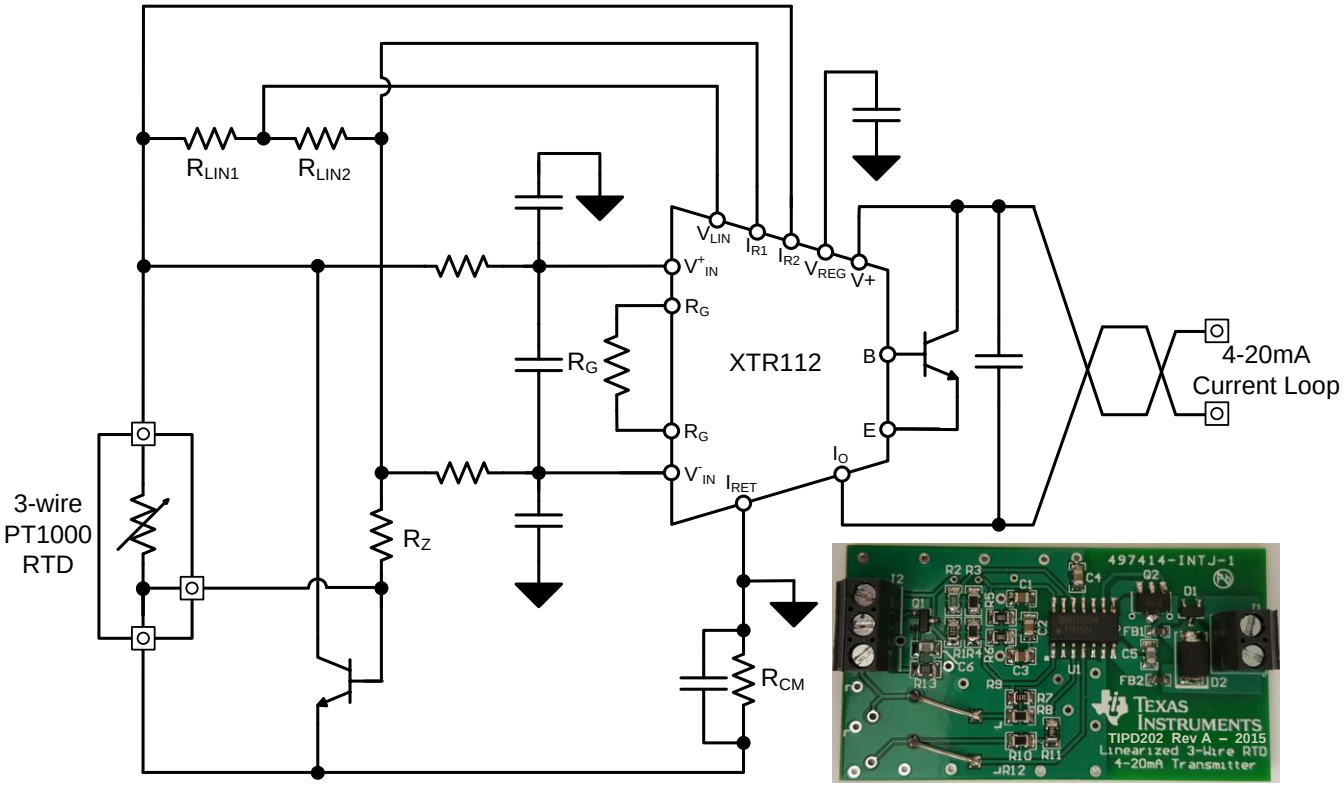

1 Design Summary

The design requirements are as follows:

- Loop Supply Voltage: +7.5 – 40V
- Input: 3-Wire PT1000 RTD
- RTD Temperature Range: -50 -250°C (300°C span)
- Output: 4-20mA 2-Wire Current Loop
  - Ideal Output Slope of 53.33µA/°C
- Circuit Protection: Protected against IEC61000-4 Transients and EMI/EMC

The design goals and performance for the PT1000 RTD sensor transmitter are summarized in Table 1. Figure 1 and Figure 2 depict the unadjusted and calibrated temperature accuracy results for the design.

Table 1: Comparison of Design Goals, Simulated, and Measured Performance

Specification	Goals	Calculated	Measured
Total Unadjusted Error (TUE)	±0.3°C, (±0.1 %FSR)	±0.315°C, (±0.105%FSR)	0.146°C, (0.049%FSR)
Calibrated Output Error	±0.1°C, (±0.033%FSR)	±0.102°C, (±0.034%FSR)	±0.11°C, (±0.037 %FSR)

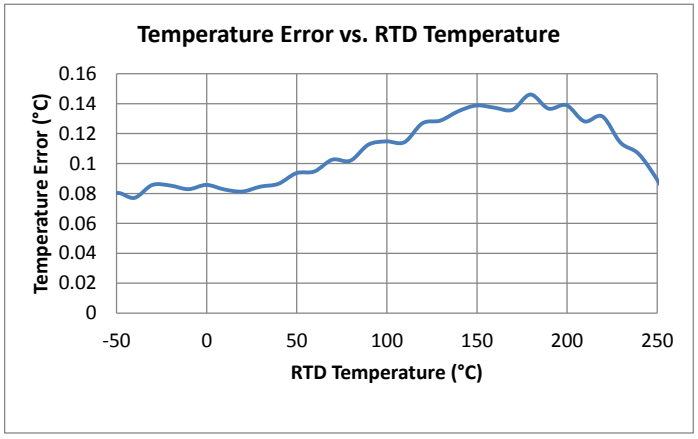

Figure 1: Measured Unadjusted Temperature Error

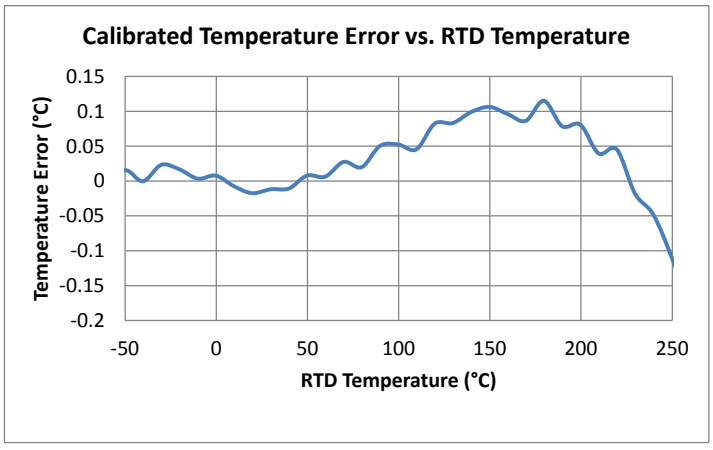

Figure 2: Calibrated Temperature Error

2 Theory of Operation

PT1000 RTD sensors require excitation sources, amplification, and linearization circuits to convert the sensor resistance into a linear output representing temperature [1]. In this application, the linearized RTD output will be transmitted over a standard 2-wire 4-20mA current loop to a 2-wire analog input module for use in process/control applications. The complete RTD sensor conditioning and 2-wire transmitter circuitry for this design is featured in a single integrated component, the XTR112, as shown in Figure 3. The output current transfer function of the transmitter is shown in Equation 1.

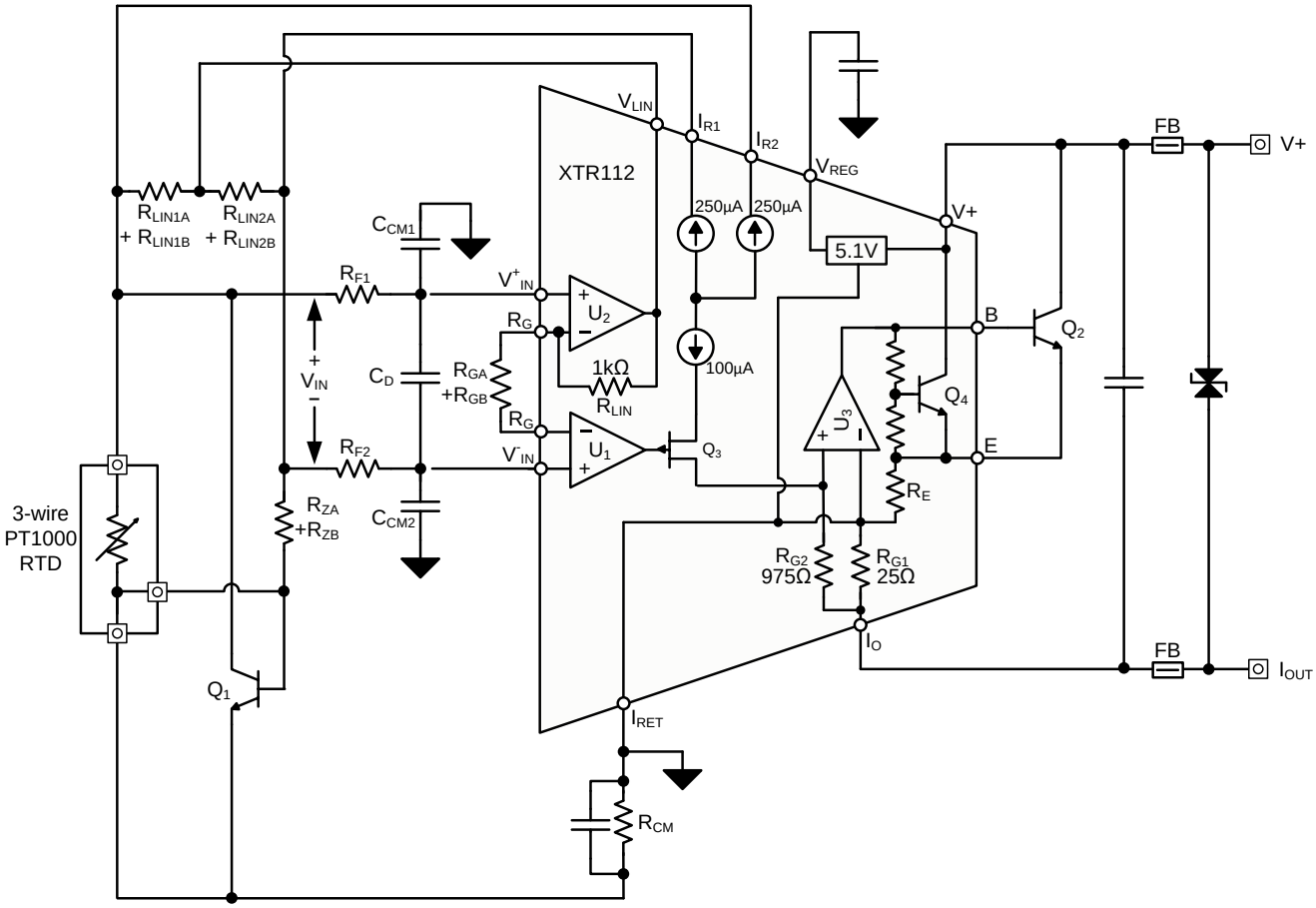

Figure 3: Detailed Schematic

$$I_{OUT}(mA) = \left(V^{+}_{IN} - V^{-}_{IN}\right) * \left(\frac{40000}{R_G}\right) + 4 \quad (1)$$

2.1 Input Amplification and Linearization Stage

The input amplification and linearization stage of the XTR112 is shown in Figure 4. It is comprised of two excitation current sources, I_{R1} and I_{R2}, several passive components, and two op amps, U1 and U2, that work together to amplify and linearize the RTD resistance. The two 250µA current sources, I_{R1} and I_{R2}, are used to create voltage drops across the RTD and zero-scale resistor, R_Z. Using two current sources allows for the lead resistances of the 3-wire RTD to be cancelled [1]. The difference in voltage drops across the RTD and R_Z resistors creates the differential input voltage, V_{IN}, to the XTR112 input stage. Differential and common-mode input filters are formed with R_{F1}, R_{F2}, C_D, C_{CM1}, and C_{CM2} which help reject noise that may couple into the RTD lead wires.

The gain of the input stage, formed by U1 and U2, is set by the ratio of the external resistor, R_G, and internal resistor, R_{LIN}. The U1 output controls the amount of current the Q_3 JFET sources to the output stage. The JFET current, I_{Q3}, is the sum of a 100uA offsetting current source and a current proportional to the differential input voltage, V_{IN}, and the ratio of the R_{LIN1} and R_G resistors. The 100uA offsetting current source is included so a zero-volt differential input signal results in a transmitter output of 4mA after the output stage gain. The transfer function for the output current of the input stage is shown in Equation 2.

The U2 amplifier also provides a controlled amount of positive feedback that increases the excitation current to the RTD over the span of the measurement. The increase in excitation current is designed to correct for the decrease in RTD sensitivity (α) over the span of the temperature measurement [1] [2]. The amount of excitation current increase through the RTD is set by R_{LIN1}. R_{LIN2} balances the circuit, increasing the effectiveness of the lead resistance cancellation.

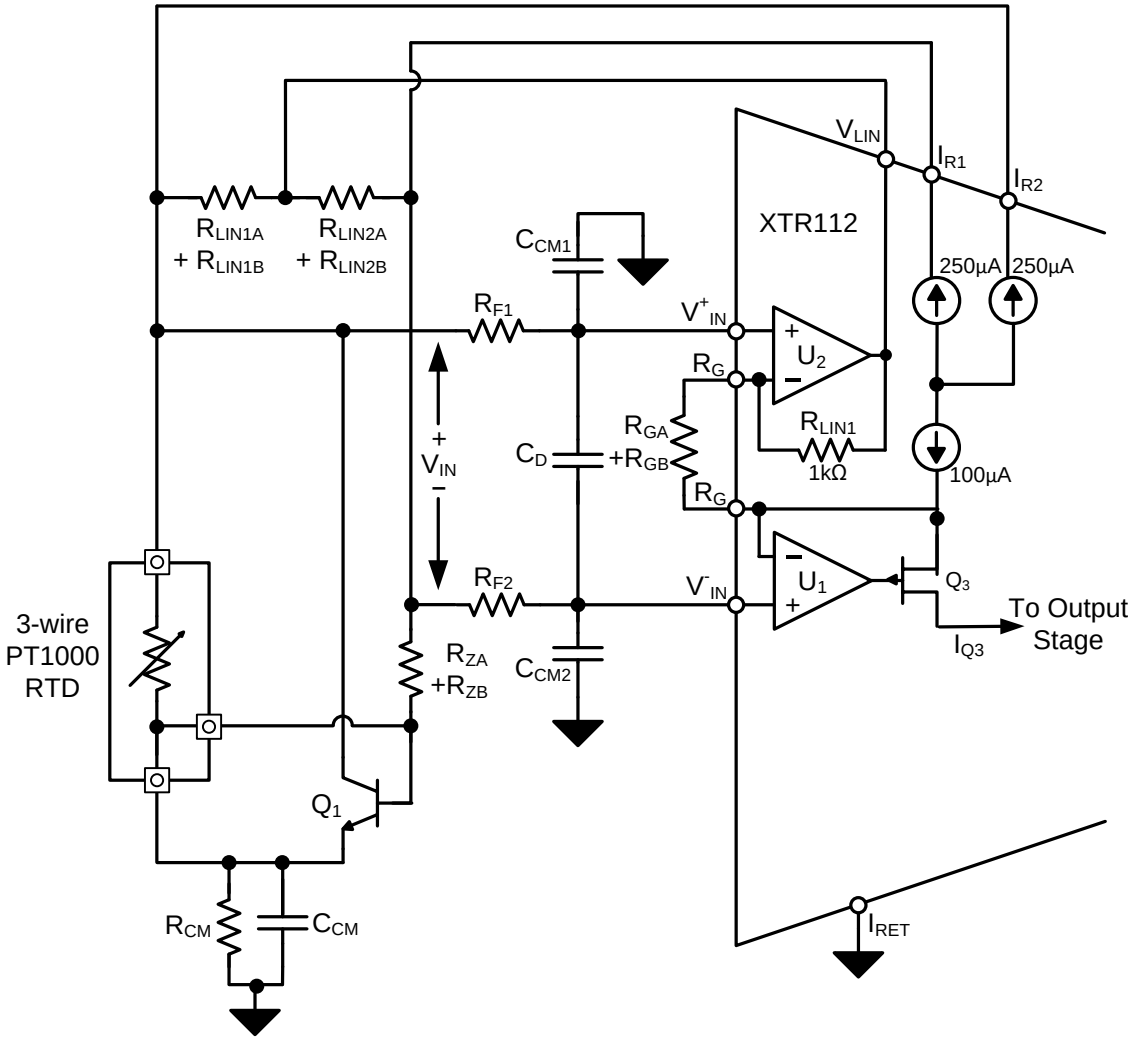

Figure 4: XTR112 Input Excitation, Linearization, and Amplification Stage

$$I_{Q3}(\mu A) = \left(V^{+}_{IN} - V^{-}_{IN}\right) * \left(\frac{R_{LIN1}}{R_G}\right) + 100\mu A = V_{IN} * \left(\frac{1000}{R_G}\right) + 100\mu A \quad (2)$$

2.2 Output Current Loop Driver Stage

The output current loop driver stage controls the output current in the 2-wire loop based on the output current from the input stage, I_{Q3}. The I_{Q3} current passes through the R_{G2} resistor creating a voltage at the non-inverting input of U3. U3 controls the current flowing through the external transistor, Q2, so the voltage drop across R_{G1} is equal to the drop across R_{G2}. The sensor excitation currents as well as the current for circuitry powered from the V_{REG} output (I_{RET}) also flow through R_{G1} and the op amp controls the Q2 output to provide the remaining required current [3]. The transfer function for the output stage is shown in Equation 3.

$$I_O(mA) = (I_{Q3}) * \left(1 + \frac{R_{G2}}{R_{G1}}\right) \quad (3)$$

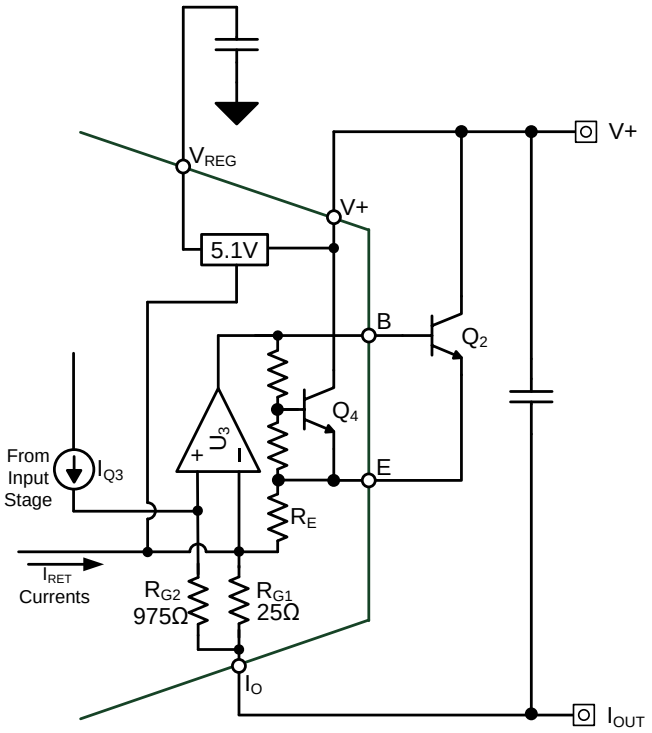

Figure 5: XTR112 Current Loop Transmitter Circuitry

3 Component Selection

3.1 XTR112

The XTR112 is a single-chip 2-wire current loop sensor transmitter designed specifically for 2 or 3-wire PT1000 RTD sensors. The XTR112 integrates the excitation sources, amplification, linearization, and current loop drive circuitry described in Section 2. The linearization engine in the XTR112 results in roughly 40:1 improvement in the linearity of the final output current over the measured temperature span versus a non-linearized RTD sensor as shown in Section 4.2.

The XTR112 features strong dc specifications with a maximum 100 µV initial offset and 1.5 µV/°C maximum offset drift. The maximum span error is 0.2% with 25 ppm/°C of span drift. The internal 4 mA zero-scale offsetting circuitry has a maximum error of 25 µA. The internal 250 µA excitation sources have a maximum error of 0.2% with worst-case matching of 0.1%. The maximum non-linearity of 0.01% enables strong post-calibration accuracy.

3.2 Passive Component Selection

Four external resistors are required to configure the gain, offset, and linearization circuitry for the XTR112 based on the desired 3-wire RTD temperature measurement range. These resistors are R_Z, R_G, R_{LIN1}, and R_{LIN2}. The R_Z resistor is selected to produce a zero-volt differential input signal when the RTD is at the minimum resistance ($R_{RTD}(T_{MIN})$) as shown in Equation 4.

$$R_Z = R_{RTD}(T_{MIN}) \tag{4}$$

R_G sets the gain of the circuit and is based on the change of the RTD over the measurement temperature span. The complete equation to calculate the R_G value is shown in Equation 5 where $R_{RTD}(T_{MAX})$ is the maximum RTD resistance over the measurement temperature and $R_{RTD}(T_{MID})$ is equal to $R_{RTD}(T_{MAX}) – R_{RTD}(T_{MIN}) / 2$.

$$R_G = \frac{0.625 * (R_{MAX} - R_Z)(R_{MAX} - R_Z)}{R_{MAX} - R_{MID}}$$

Where

$$R_{MAX} = R_{RTD}(T_{MAX}) \tag{5}$$

$$R_{MID} = R_{RTD}(T_{MID})$$

$$T_{MID} = \frac{T_{MAX} - T_{MIN}}{2}$$

R_{LIN1} and R_{LIN2} are calculated using the same variables found in the R_G equation as shown in Equation 6.

$$R_{LIN1} = \frac{1.6 * R_{LIN}(R_{MAX} - R_{MID})}{(2R_{MID} - R_{MAX} - R_Z)}$$

$$R_{LIN2} = \frac{1.6 * (R_{LIN} + R_G)(R_{MAX} - R_{MID})}{(2R_{MID} - R_{MAX} - R_Z)} \tag{6}$$

Where

$$R_{LIN} = 1000\Omega$$

The min, mid, and max values of a PT1000 RTD operating from -50C - 250C are shown in Table 2.

Table 2: PT1000 Resistance Values for -50°C to 250°C Span

	Temperature	Resistance
RTD(T_{MIN})	-50°C	803.1 Ω
RTD(T_{MID})	100°C	1385 Ω
RTD(T_{MAX})	250°C	1941 Ω

Substituting the RTD resistance values from Table 2 into Equations 4 – 6 results in the external passive component values shown in Table 3. The closest 0.1% values are also shown along with the error resulting from the discrepancy from the ideal value.

Table 3: XTR112 Zero, Gain, and Linearity Resistor Selections with Error

	Calculated Value	Closest 0.1% Value	Error Between Calculated and Available Values
RZ	803.06 Ω	806 Ω	0.37%
RG	744.55 Ω	750 Ω	0.73%
RLIN1	34120 Ω	34000 Ω	0.35%
RLIN2	59530 Ω	59000 Ω	0.89%

The initial gain and offset errors caused by the deviation between the closest 0.1% values and the ideal calculated value are outside of the unadjusted accuracy goals so the combination of two resistors will be used for these components as shown in Table 4.

Table 4: Updated Resistor Selections with Error

	Calculated Value	Resistor A	Resistor B	Total	Error
R_Z	803.06 Ω	787 Ω	16 Ω	803 Ω	0.007%
R_G	744.55 Ω	715 Ω	29.4 Ω	744.4 Ω	0.02%
R_{LIN1}	34120 Ω	34000 Ω	120 Ω	34120 Ω	0.0%
R_{LIN2}	59530 Ω	59000 Ω	536 Ω	59536 Ω	0.01%

The two 250 µA excitation currents flow through the R_{CM} resistor to bias the differential input signal to within the XTR112 input common-mode voltage range of 1.25 V to 3.5 V. The minimum resistance is calculated in Equation 7.

$$R_{CM\_MIN} = \frac{1.25V}{2 \cdot 250\mu A} = 2500\Omega \tag{7}$$

R_{CM} was selected to be a standard value of 3.3 kΩ, resulting in a minimum common-mode input voltage of 1.65 V.

Most temperature measurements do not require high bandwidths because the change in temperature is not very fast. In this design, the inputs to the XTR112 have been filtered to reduce interference from extrinsic noise sources. R_{F1} and R_{F2} combine with C_D, C_{CM1} and C_{CM2} to form common-mode and differential input filters. The common-mode filter cutoff frequency is set roughly forty times greater than the differential filter to prevent component tolerances from degrading the common-mode rejection of the amplifier. The cutoff frequencies for these filters are calculated in Equation 8.

$$f_{-3dB\_DIFF} = \frac{1}{2 \cdot \pi \cdot (2 * 10k\Omega) \cdot 0.22\mu F} = 36Hz$$
$$f_{-3dB\_CM} = \frac{1}{2 \cdot \pi \cdot 10k\Omega \cdot 0.01\mu F} = 1591Hz \tag{8}$$

3.3 Protection Component Selection

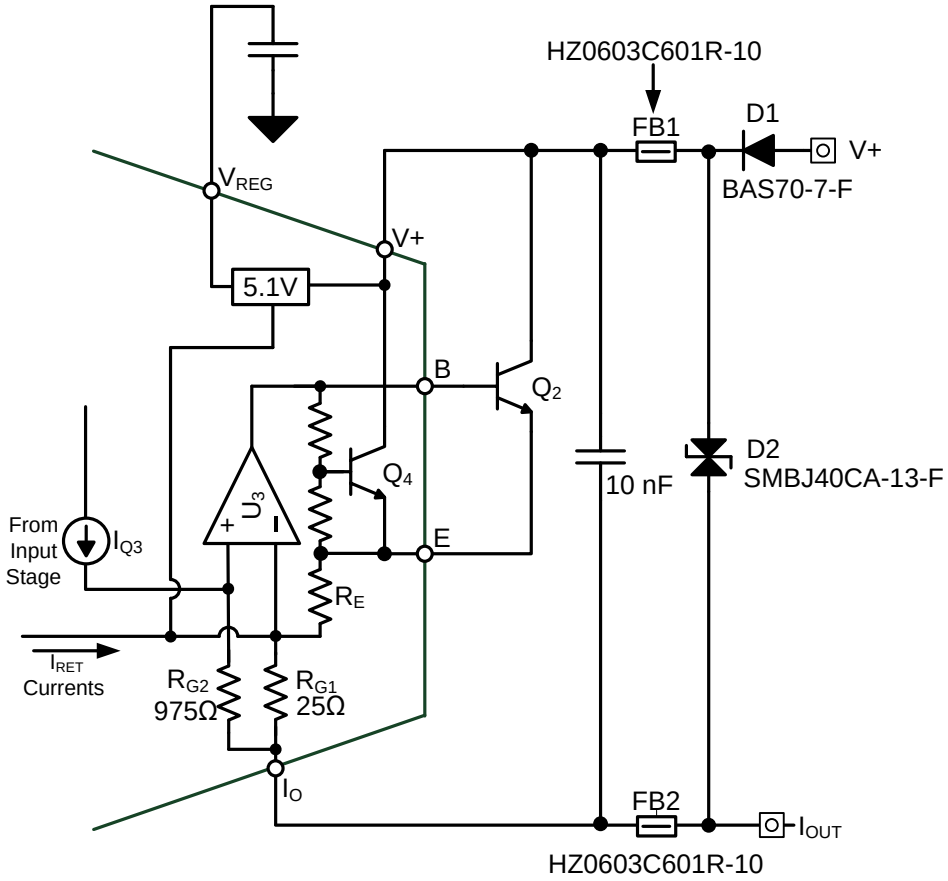

Figure 6: Simplified Schematic Showing Output Protection Components

The protection scheme in this design, shown in Figure 6, is intended to provide immunity to EMI/RFI interference and transient voltage disturbances as described in the IEC61000-4 tests[5][6][7][8][9][10]. While this design was not taken to the external test house, TIPD161 which features the XTR105 and effectively the same output stage was tested with the same components and passed.

The IEC61000-4 transients have two components: a high frequency component and a high energy component. Therefore the protection strategy focuses on attenuating the high frequency transients and diverting the energy from the high energy components away from the sensitive circuitry[4].

Attenuation is achieved through resistors and capacitors that attenuate high-frequency transients and also limit series current. Ferrite beads are used to maintain dc accuracy while still limiting current from high frequency transients. Capacitors placed between the output terminals help attenuate transient energy from high frequency tests such as electrostatic discharge (ESD) and electrically fast transients (EFT).

Diverting the large energy transients is accomplished using transient voltage suppressor (TVS) diodes. The protection scheme must limit the level of the voltage transients to a level less than the absolute maximum rating for the output of the XTR112. More information on the IEC61000-4 tests and the protection circuitry requirements can be obtained in TIPD153 [4].

3.3.1 Reverse Protection Diode

A Schottky diode rectifier is used in this design to prevent damage to the components in the design if the supply and output wires are connected improperly. Selection of this diode should be based on low reverse leakage current and low forward voltage. The BAS70 was chosen for this design because it features low forward voltages of 0.41 V, a maximum reverse breakdown voltage of 70 V and forward current rating of 70 mA continuous. A diode bridge could have been used to maintain proper transmitter operation regardless of the wiring connections at the cost of an extra diode drop in the loop.

3.3.2 TVS Diode

A bidirectional TVS diode is used to divert energy from high voltage transients away from the XTR112 and sensor. Select the TVS diode based on the working voltage, breakdown voltage, leakage current and power rating. The working voltage specification defines the largest reverse voltage that the diode is meant to be operated at continuously without it conducting. This is the voltage at the “knee” of the reverse breakdown curve where the diode begins to break down and exhibits some small leakage current. As the voltage increases above the working voltage, more current will begin to flow through the diode. The breakdown voltage defines the reverse voltage at which the diode is fully allowing current to flow. It is important to keep in mind that if large currents flow through the diode, the breakdown voltage will rise.

The diode breakdown voltage should be low enough to protect all components connected to the output terminals and to provide headroom to continue providing this protection as the breakdown voltage rises with large currents. In this design the working voltage of the TVS diode should be at or above the upper limit of the allowed supply voltages since any higher voltage would cause leakage through the diode. In this case the SMBJ40CA diode was selected with a working voltage of 40 V, breakdown voltage of 44 V, and power rating of 600 W. An additional parameter to consider for TVS diode selection is leakage current. At the working voltage, when the diode is not operating in its breakdown region, some current will flow through the diode and can affect system accuracy. The diode selected for this design features 1 µA maximum leakage current at the working voltage.

3.3.3 Passive Comonents

Series ferrite beads and a parallel capacitor are used to attenuate transient signals that may remain after passing across the TVS diode. The ferrites are chosen based on their current rating, impedance at dc, and impedance at high frequency. In this design the chosen ferrites feature 42 mΩ max impedance at dc, 600 Ω impedance at 100 MHz, and 3 A current rating. The capacitor chosen has a voltage rating of 100 V.

4 Circuit Performance Calculations

The circuit performance is based on the specifications of the XTR112 and the accuracy of the external R_Z, R_G, R_{LIN1}, and R_{LIN2} resistances. The final temperature accuracy depends also on the effectiveness of the XTR112 analog RTD non-linearity correction methods.

4.1 Circuit Accuracy

A simplified error calculation using only the dominant error terms is shown in this section. A full error calculation is shown in the datasheet for the XTR112 that includes all error sources in the XTR112. The specifications that dominate the XTR112 performance are listed in Table 5.

Table 5: XTR112 Performance Specifications

XTR112 Specification	Typical Value
Offset (mV)	0.05
Zero Current Error (µA)	5
Excitation Current Error (%)	0.05
Gain Error (%)	0.05
Nonlinearity (%)	0.003

The input span for the measurement range featured in this design is calculated in Equation 9 and used in the error calculations.

$$V_{IN\_SPAN}(V) = \left(RTD(T_{MAX}) - RTD(T_{MIN})\right) * I_{EXC}$$
$$V_{IN\_SPAN}(V) = \left(1941\Omega - 803.1\Omega\right) * 250\mu A = 0.2845V \quad (9)$$

The total unadjusted error, TUE, of the XTR112 (TUE_{XTR112}) can be roughly calculated as shown in Equation 10.

$$TUE_{XTR112}(\%FSR) = \left(\sqrt{\left(\frac{V_{OS}}{V_{IN\_SPAN}}\right)^2 + \left(\frac{Zero}{I_{OUT\_SPAN}}\right)^2 + \left(\frac{GainError\%}{100}\right)^2 + \left(\frac{ExcitationError\%}{100}\right)^2 + \left(\frac{Nonlinearity}{100}\right)^2}\right) \times 100$$
$$TUE_{XTR112}(\%FSR) = \left(\sqrt{\left(\frac{0.05mV}{0.2845V}\right)^2 + \left(\frac{0.005mA}{16mA}\right)^2 + \left(\frac{0.05}{100}\right)^2 + \left(\frac{0.05}{100}\right)^2 + \left(\frac{0.003}{100}\right)^2}\right) \times 100 = 0.085\%FSR \quad (10)$$

R_Z resistor tolerance and deviation from the ideal calculated value introduce offset errors into the design. The errors caused by tolerance (R_{Z_TOL}) deviation from the actual value (R_{Z_DEV}) of R_Z are calculated in Equation 11.

$$R_{Z\_TOL} = 0.1\%FSR$$
$$R_{Z\_DEV}(\%FSR) = \left(\frac{\left(R_Z - R_{Z\_IDEAL}\right) * 250\mu A}{V_{IN\_SPAN}}\right) \times 100$$
$$R_{Z\_DEV}(\%FSR) = \left(\frac{\left(803.1\Omega - (787\Omega + 16\Omega)\right) * 250\mu A}{0.2845V}\right) \times 100 = 0.01\%FSR \quad (11)$$

The R_G resistor tolerance and deviation from the ideal calculated value introduce gain errors into the design. The errors caused by the tolerance (R_{G_TOL}) deviation from the actual value (R_{G_DEV}) of R_G are calculated in Equation 12.

$$R_{G\_TOL} = 0.1\%FSR$$
$$R_{G\_DEV}(\%FSR) = \left(\frac{\left(R_G - R_{G\_IDEAL}\right)}{R_{G\_IDEAL}}\right) \times 100$$
$$R_{G\_DEV}(\%FSR) = \left(\frac{744.55\Omega - \left(715\Omega + 29.4\Omega\right)}{744.55\Omega}\right) \times 100 = 0.02\%FSR \quad (12)$$

The XTR112 errors can be combined with the errors from the external components to calculate the total error of the circuit as shown in Equation 13. Note that the resistor tolerances were divided by three to obtain the typical error for the resistors to match the typical values used in the rest of the calculations.

$$I_{OUT\_ERROR}(\%FSR) = \left(\sqrt{\left(TUE_{XTR105}\right)^2 + \left(\frac{R_{Z\_TOL}}{3}\right)^2 + \left(R_{Z\_DEV}\right)^2 + \left(\frac{R_{G\_TOL}}{3}\right)^2 + \left(R_{G\_DEV}\right)^2}\right) \times 100$$
$$I_{OUT\_ERROR}(\%FSR) = \left(\sqrt{(0.085\%)^2 + (0.033\%)^2 + (0.01\%)^2 + (0.033\%)^2 + (0.02\%)^2}\right) = 0.099\%FSR \quad (13)$$

4.2 RTD Sensor Linearization Circuitry Accuracy

As shown in Figure 7 a PT1000 RTD measurement will have roughly 1.14% nonlinearity over the -50°C – 250°C span if a simple endpoint fit is used. The linearization circuit in the XTR112 can ideally correct a standard RTD nonlinearity by over forty times compared with an end-point fit. Figure 7 displays the nonlinearity of a PT1000 compared with the ideal output using the XTR112. As shown, the XTR112 linearization circuitry can ideally reduce the output nonlinearity to roughly ±0.008%FSR.

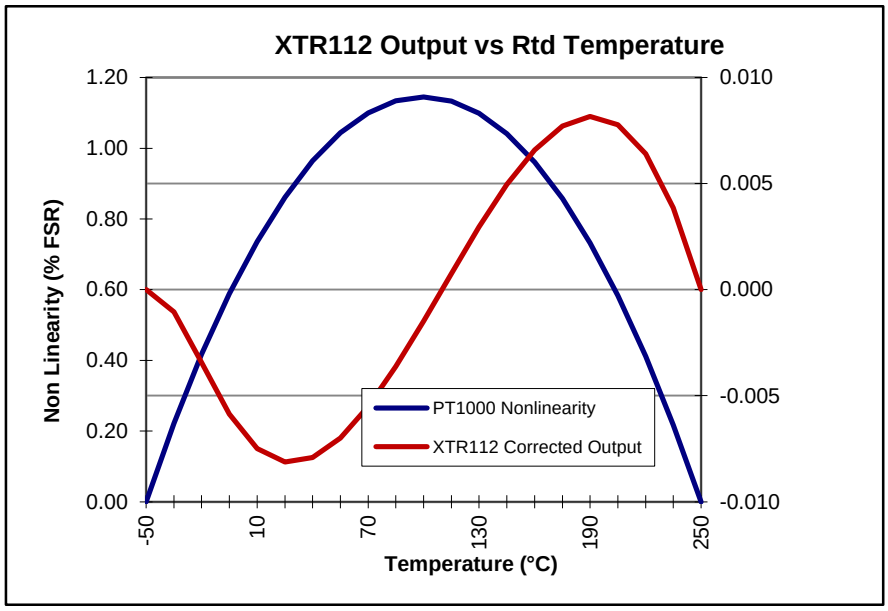

Figure 7: XTR112 Linearization Circuitry Ideal Improvements

Component tolerance and deviation from the ideal calculated value for RLIN1 will reduce the effectiveness of the linearization engine. The tolerance and deviation of RLIN2 will have minimal impact on the transfer function but will impact the effectiveness of the 3-wire RTD lead resistance cancellation. The accuracy of the linearization engine and the effects of RLIN1 are combined with the error calculations to predict the final transmitter accuracy as shown in Equation 14.

$$I_{OUT\_ERROR}(\%FSR) = \left(\sqrt{(TUE_{TOTAL})^2 + (Linearization_{ERROR})^2 + \left(\frac{R_{LIN1\_TOL}}{3}\right)^2 + (R_{LIN1\_DEV})^2} \right) \times 100$$
$$I_{OUT\_ERROR}(\%FSR) = \left(\sqrt{(0.099\%)^2 + (0.008\%)^2 + (0.033\%)^2 + (0\%)^2} \right) = 0.105\%FSR \quad (14)$$

4.3 Calibrated Output Accuracy

The gain and offset calibration circuit shown and explained in Appendix A.3 can be used to remove the gain and offset errors of the circuit leaving only the linearity errors of the XTR112 gain stage and linearization circuitry. These terms are combined in Equation 15 which predicts the calibrated circuit accuracy.

$$I_{OUT\_CAL}(\%FSR) = \left(\sqrt{(XTR105_{NONLINEARITY})^2 + (Linearization_{ERROR})^2 + (R_{LIN1\_TOL})^2 + (R_{LIN1\_ACTUAL})^2} \right) \times 100$$
$$I_{OUT\_CAL}(\%FSR) = \left(\sqrt{0.003^2 + 0.008^2 + 0.033^2 + 0.00^2} \right) \times 100 = 0.034\%FSR \quad (15)$$

5 PCB Design

The PCB schematic and bill of materials can be found in Appendix A.1.

5.1 PCB Layout

For optimal performance of this design follow standard precision PCB layout guidelines, including proper decoupling close to all mixed signal integrated circuits and providing adequate power and GND connections with large copper pours.

Additional considerations must be made for providing robust EMC/EMI immunity. All protection elements should be placed as close to the output connectors as possible to provide a controlled return path for transient currents that does not cross sensitive components. Optimal dissipation of the transient energy occurs with wide, low-impedance, low-inductance traces along the output signal path and protection elements. When possible, copper pours are used in place of traces. Stitching the GND pours provides an effective return path around the PCB and helps reduce the impact of radiated interference.

The layout for the TIPD202 design is shown in Figure 8.

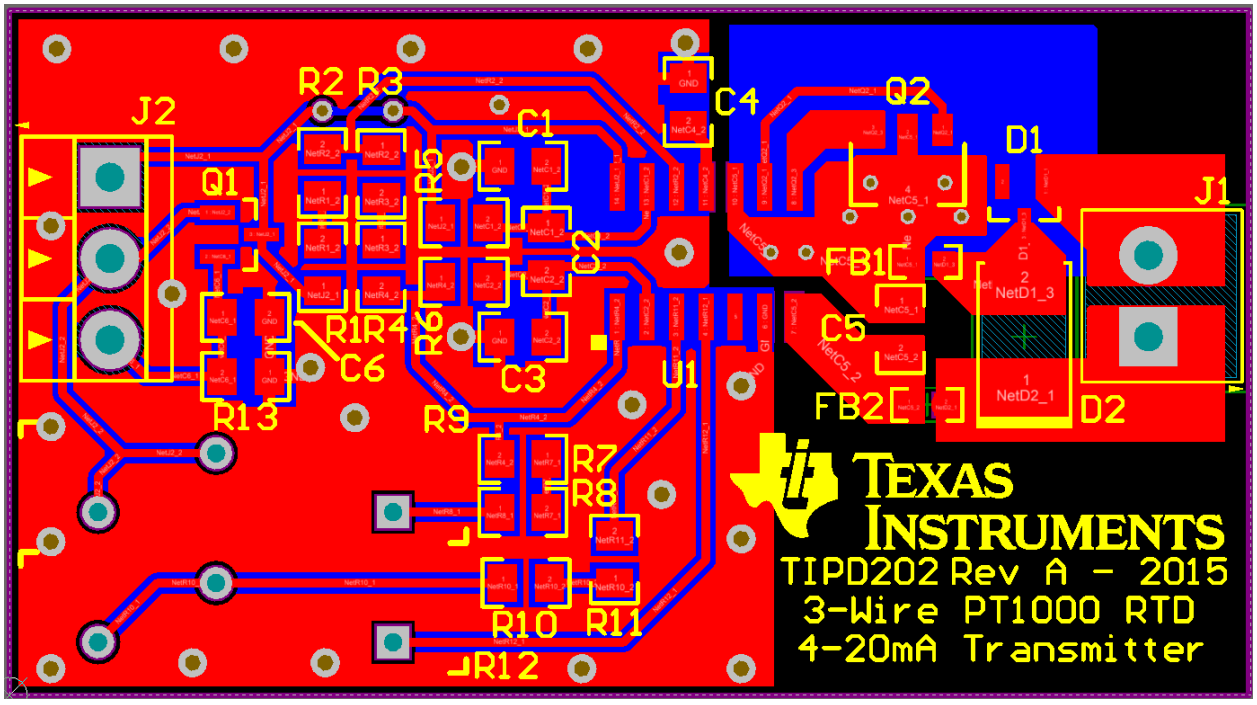

Figure 8: Altium PCB Layout

6 Verification and Measured Performance

6.1 Measured Transfer Function with Precision Resistor Input

To test the accuracy of the acquisition circuit, the CL3001 lab calibrator from Omega Engineering was used as the input to the system. Figure 9 displays the output current of the system over an input temperature range of -50°C to 250°C, or input resistance of roughly 803 Ω to 1941 Ω. The design performs as expected and outputs a linear 4-20 mA over the input RTD temperature range. The power supply voltage was +24 V for the measured results in this section.

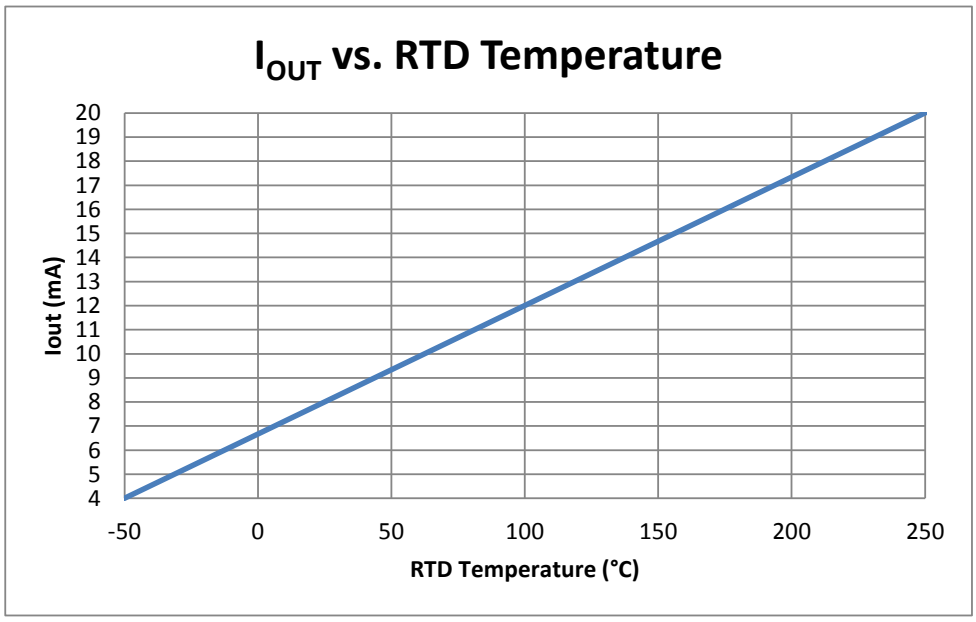

Figure 9: Output Current (I_{OUT}) vs. RTD Temperature

The output current error is displayed in Figure 10. The maximum output current error was less than 8 µA with around 5 µA of offset error and gain and linearity errors less than ±3 uA.

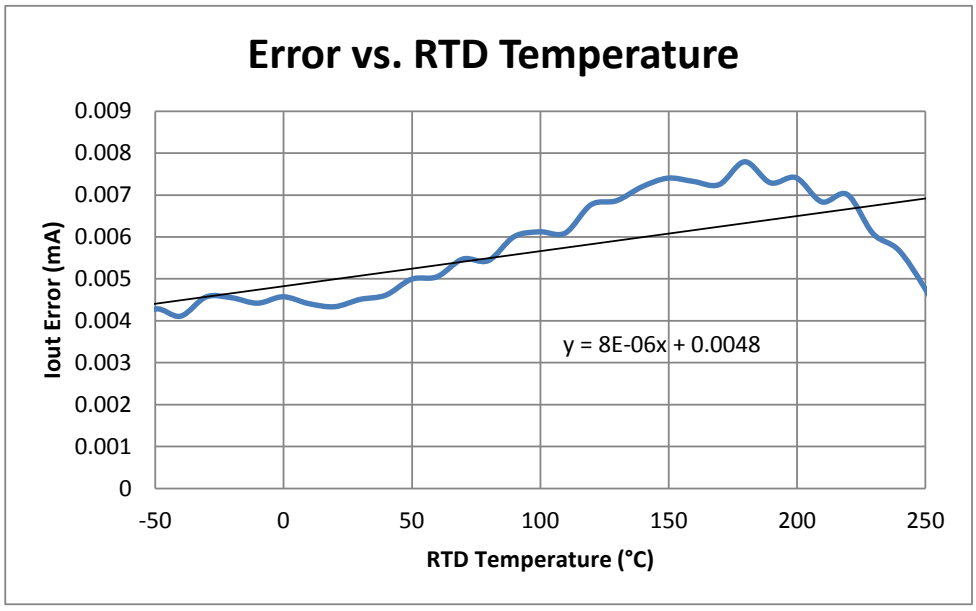

Figure 10: Output Current Error vs. RTD Temperature

The output current error can be converted to output temperature error based on the ideal output slope of 0.0533mA/°C or 18.75°C/mA. The output temperature error over the measurement range is under 0.15°C as shown in Figure 11.

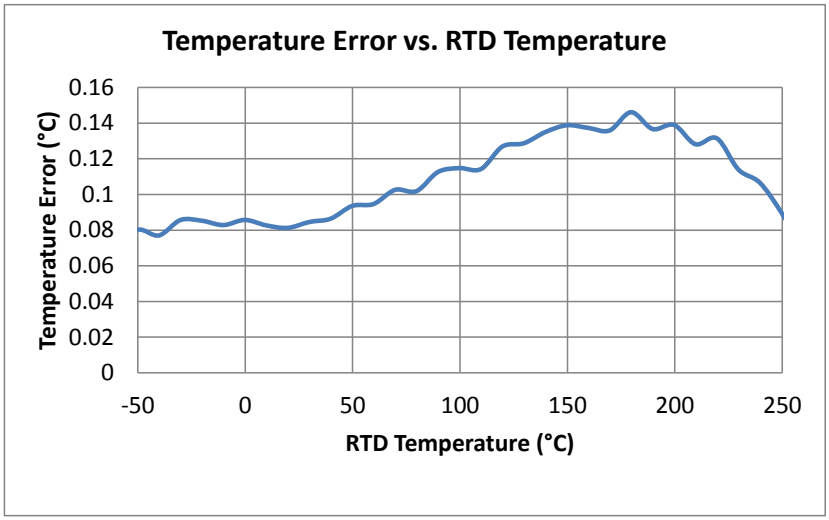

Figure 11: Output Temperature Error vs. RTD Temperature

While the initial accuracy is likely suitable for many applications using this design, the accuracy can be further improved by removing the gain and offset errors with the manual calibration described in Appendix A.3. Using this method, the results in Figure 12 were achieved. The final calibrated error is about ±0.11°C (0.037% FSR) which matches very well with the calculated value of 0.102°C (0.034% FSR).

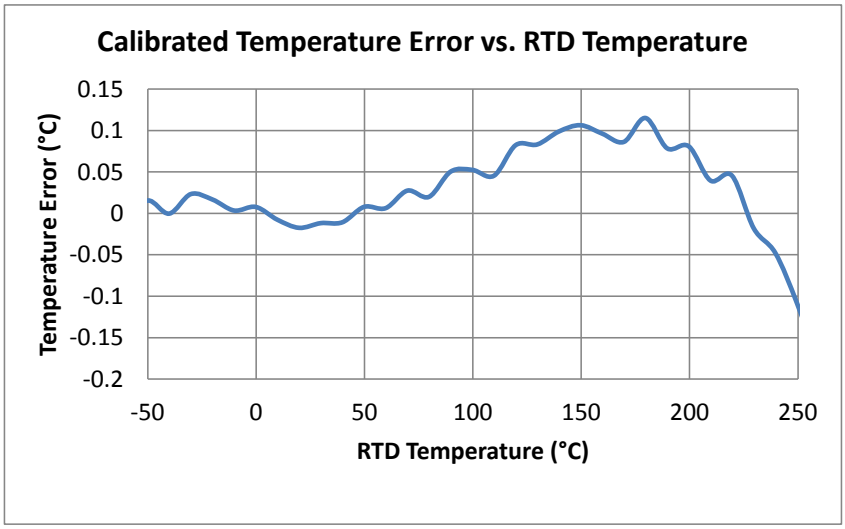

Figure 12: Calibrated Output Current Error vs. RTD Temperature

6.2 Measured Result Summary

The measured performance is summarized and compared with the goals and calculated values in Table 6.

Table 6: Comparison of Design Goals, Calculated, and Measured Performance

Specification	Goals	Calculated	Measured
Total Unadjusted Error (TUE)	±0.3°C, (±0.1 %FSR)	±0.315°C, (±0.105%FSR)	0.146°C, (0.049%FSR)
Calibrated Output Error	±0.1°C, (±0.033%FSR)	±0.102°C, (±0.034%FSR)	±0.11°C, (±0.037 %FSR)

7 Modifications

The XTR112 is the only integrated solution available from Texas Instruments that convert a PT1000 RTD to a linear 4-20mA output without any external digital or active circuitry. A discrete solution could be used to create a similar system using an ADC, MCU, and DAC such as the design featured in http://www.ti.com/tool/TIDA-00095.

8 About the Authors

Collin Wells is an applications engineer in the Precision Linear group at Texas Instruments where he supports industrial products and applications. Collin received his BSEE from the University of Texas, Dallas.

9 Acknowledgements & References

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Appendix A.

A.1 Electrical Schematic

The Altium electrical schematic for this design can be seen in Figure 13.

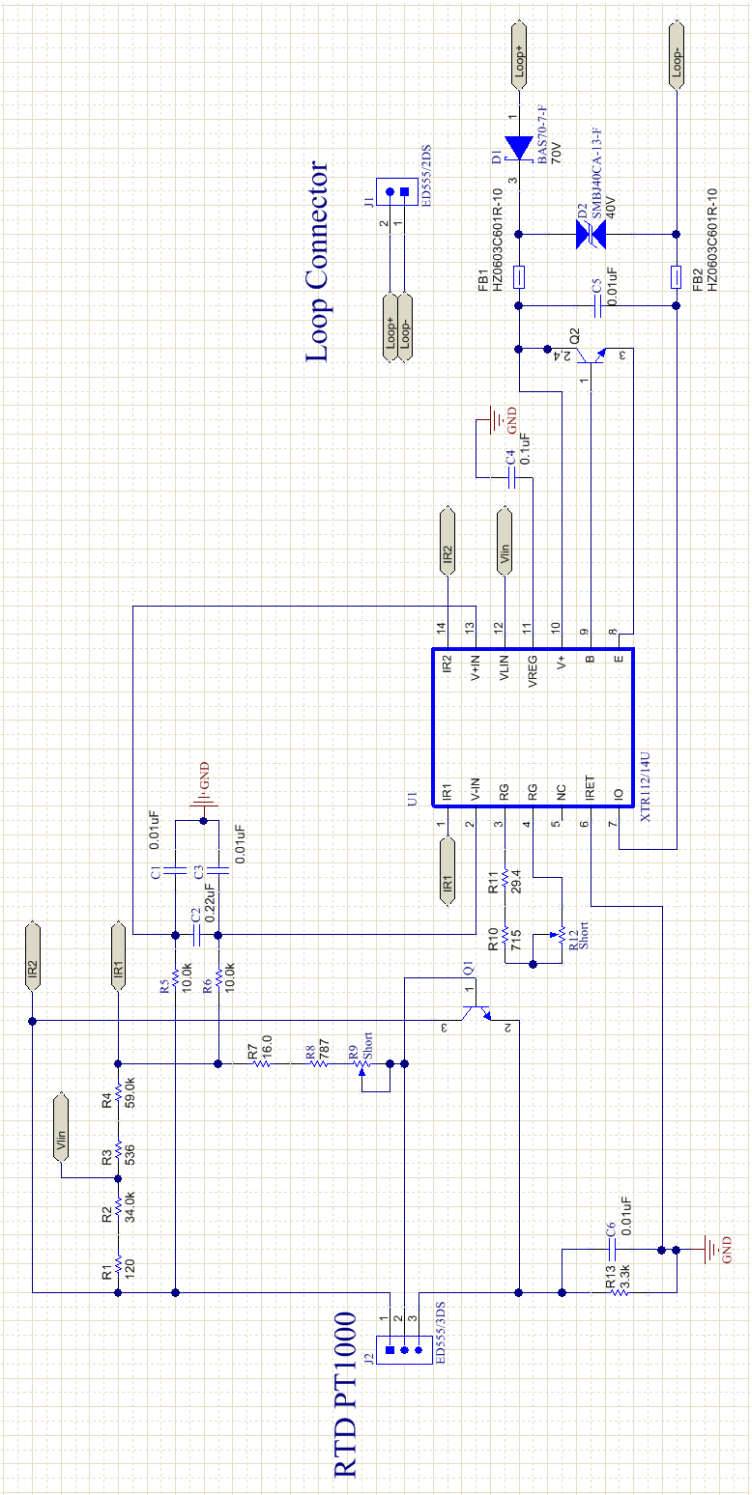

Figure 13: Altium Schematic

A.2 Bill of Materials

The bill of materials for this circuit can be seen in Figure 14.

Texas Instruments

Bill of Materials

TI DESIGNS

TIPD202: Analog Linearized 3-Wire PT1000 RTD to 4-20mA Current Loop Transmitter Reference Design

Item #	Quantity	Designator	Value	Description	Manufacturer	Part Number
1	2	C1, C3	0.01uF	CAP, CERM, 0.01uF, 50V, +/-5%, C0G/NP0, 0805	TDK	C2012C0G1H103J
2	1	C2	0.22uF	CAP, CERM, 0.22uF, 50V, +/-10%, X7R, 0805	MuRata	GRM21BR71H224KA01L
3	1	C4	0.1uF	CAP, CERM, 0.1uF, 50V, +/-10%, X7R, 0805	AVX	08055C104KAT2A
4	2	C5, C6	0.01uF	CAP, CERM, 0.01uF, 50V, +/-10%, X7R, 0805	MuRata	GRM216R71H103KA01D
5	1	D1		Diode, Schottky, 70V, 0.07A, SOT-23	Diodes Inc.	BAS70-7-F
6	1	D2		Diode, TVS, Bi, 40V, 600W, SMB	Diodes Inc.	SMBJ40CA-13-F
7	2	FB1, FB2		300mA Ferrite Bead, 600 ohm @ 100MHz	Used in BOM report	Used in BOM report
8	1	J1		Terminal Block, 6A, 3.5mm Pitch, 2-Pos, TH	On-Shore Technology	ED555/2DS
9	1	J2		Terminal Block, 6A, 3.5mm Pitch, 3-Pos, TH	On-Shore Technology	ED555/3DS
10	1	Q1		Transistor, NPN, 45V, 0.1A, SOT-23	ON Semiconductor	BC847CLT1G
11	1	Q2		Transistor, NPN, 45V, 1A, SOT-89	Diodes Inc.	FCX690BTA
12	1	R1	120	RES SMD 120 OHM 0.1% 1/8W 0805	Panasonic Electronics	ERA-6AEB121V
13	1	R2	34.0k	RES SMD 34K OHM 0.1% 1/8W 0805	Vishay-Dale	TNPW080534K0BEEA
14	1	R3	536	RES, 536, 0.1%, 0.125 W, 0805	Panasonic Electronics	RT0805BRD07536RL
15	1	R4	59.0k	RES SMD 59K OHM 0.1% 1/8W 0805	Panasonic Electronics	ERA-6AEB5902V
16	2	R5, R6	10.0k	RES, 10.0k ohm, 1%, 0.125W, 0805	Vishay-Dale	CRCW080510K0FKEA
17	1	R7	16	RES SMD 16 OHM 0.1% 1/10W 0805	TE Connectivity	2-1614884-2
18	1	R8	787	RES SMD 787 OHM 0.1% 1/8W 0805	Panasonic Electronics	ERA-6AEB7870V
19	2	R9, R12	50	TRIMMER, 10K, 0.75W, TH	Bourns	3006P-1-500LF
20	1	R10	715	RES SMD 715 OHM 0.1% 1/8W 0805	Panasonic Electronics	ERA-6AEB7150V
21	1	R11	29.4	RES SMD 29.4K OHM 0.1% 1/8W 0805	Panasonic Electronics	ERA-6AEB2942V
22	1	R13	3.30k	RES, 3.3 k, 5%, 0.125 W, 0805	Vishay-Dale	CRCW08053K30JNEA
23	1	U1		IC 4-20MA TRANSMITTER 14-SOIC	Texas Instruments	XTR112U

Figure 14: Bill of Materials

A.3 Manual Calibration Using Trim Potentiometers

The calibrated circuit results shown in Figure 12 were achieved using trim potentiometers to set the gain resistor, R_G, and zero resistor, R_Z, values as shown in Figure 15. The range of values for R_G and R_Z are shown in Table 7.

Table 7: Comparison of Design Goals, Simulated, and Measured Performance

Calibration Element	Ideal Value	Calibration Range	Calibration % of Span
R_Z	803.06 Ω	787 Ω + 50 Ω	-2% to +4.23%
R_G	744.55 Ω	715 Ω +50 Ω	-3.97% to +2.75%

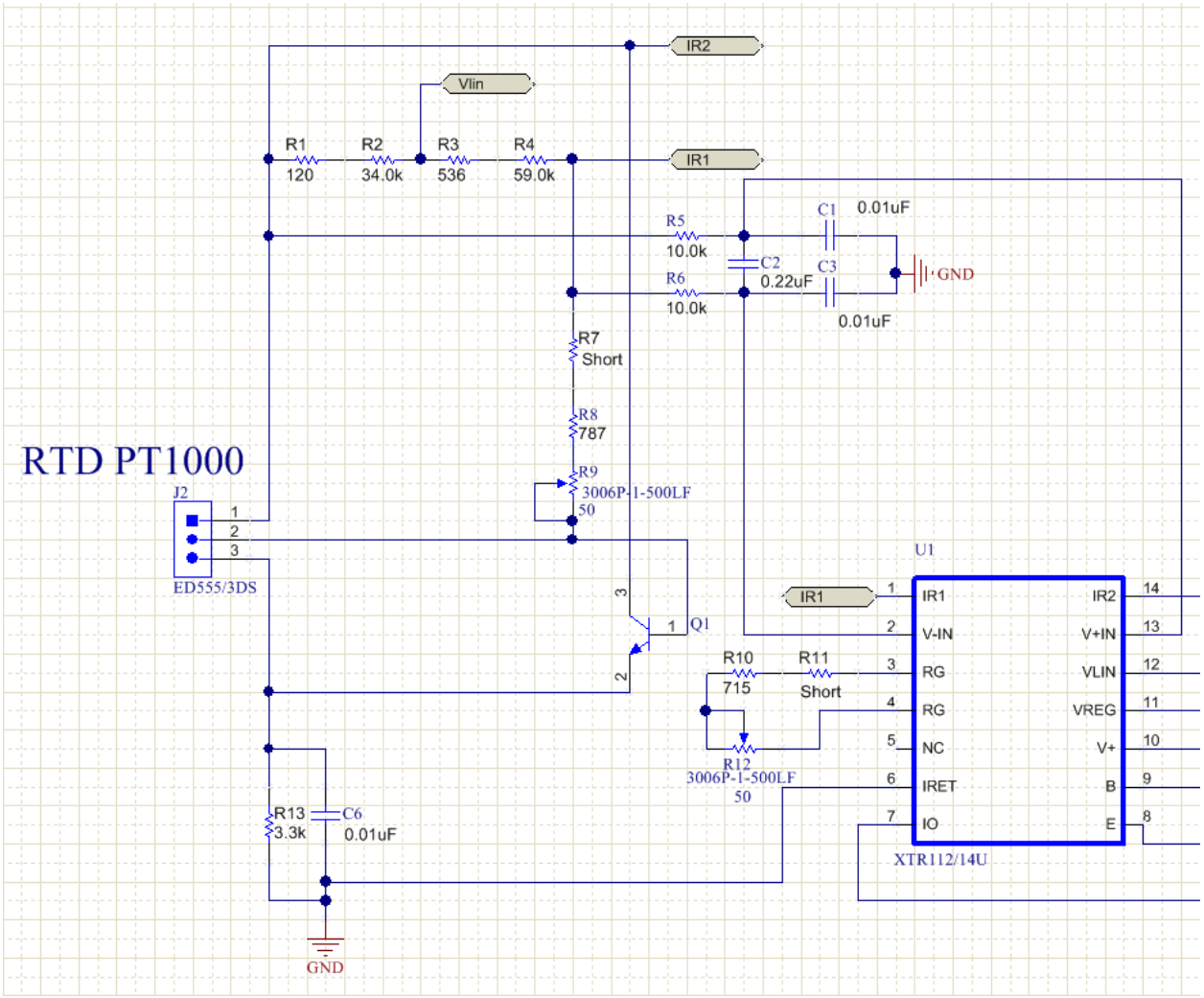

Figure 15: Schematic with Calibration Components Shown

The results from Figure 16 are shown in terms of output current (I_{OUT}) error to help explain the calibration process. The first step is to set the R_Z trim potentiometer with the RTD set to the PT1000 resistance at 0°C (1000 Ω). Adjust the output current to a 6.657mA so at 0°C the output will have effectively zero error as shown in the figure. Then adjust the R_G trim resistor so the output is 12.001mA with the RTD set to the mid temperature resistance value (1385.06 Ω). Next go back and check the output level with the input set again to 1000 Ω and adjust the output back to 6.657mA if it was affected by the gain adjustment. Repeat the gain adjustment one more time with the input set the 1385.06 Ω and the results below should be achieved.

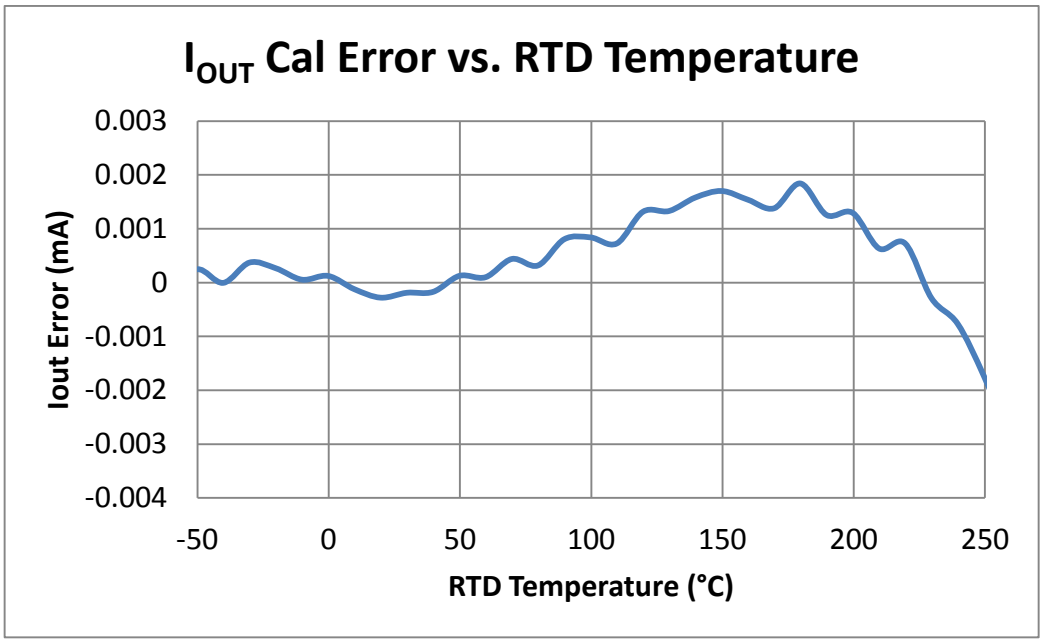

Figure 16: Calibrated Output Current Error vs. RTD Temperature

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