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Using the Non-linear Behavior of the Brokaw Bandgap Voltage Reference Cell to Linearize Resistance Temperature Detectors (RTD)

[1, 2] Pedro Carvalhaes-Dias, [2] Ivan P. Ferreira, [1, 2] Luís F. C. Duarte, [3] Flávio J. O. Morais and [2] J. A. Siqueira Dias

[1] Paraná Federal University of Technology, Av. A. Carazzai, Cornélio Procópio, 1640, PR, Brazil
[2] Department of Semiconductors, Instrumentation and Photonics, School of Electrical and Computer Engineering - Unicamp, Av. Albert Einstein, 400, Campinas, SP, Brazil
[3] Faculty of Sciences and Engineering, UNESP, R. D. Costa Lopes, 780, Tupã, SP, Brazil
[2] Tel.: +55 19 3521-4901, fax: + 55 19 3521-4896
[2] E-mail: siqueira@demic.fee.unicamp.br

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Abstract: Resistance temperature detectors (RTD) present second and third order non-linearities, and a linear signal processing circuit which converts the voltage on a Pt-100 RTD to an output voltage with 10 mV/ºC presents a maximum non-linearity error of 1.07 ºC (10.7 mV) in the 0 to 85 ºC temperature range. These non-linearities can be corrected digitally, but there are applications where a simple analog linearization can be used with advantages, as in the case of a direct interface of an RTD sensor with a 4-20 mA current loop transducer. The intrinsic curvature of the Brokaw bandgap voltage reference cell, caused by the non-linear variation of the bipolar transistor´s VBE with temperature, can be used to create a compensation voltage that can be used to reduce the non-linearity of the signal processing circuit of resistance temperature. A discrete Brokaw bandgap reference cell using a bipolar analog array LM3046 and a high precision op-amp was designed, and the calculated values of the final circuit indicate that the non-linearity of the signal conditioning circuit of a Pt-100 RTD is reduced by one order of magnitude (down to approximately 0.14 ºC) in the 0-85 ºC temperature range.

Keywords: RTD, Bandgap curvature, Brokaw cell, Linearization, Temperature sensors.

1. Resistance Temperature Detectors

Resistance Temperature Detectors (RTD) are one of the most used type of temperature sensors in industrial plants [1]. These sensors are usually fabricated using two techniques: (a) by wrapping a coil of a fine platinum wire around a ceramic substrate; (b) depositing a thin film of platinum on an insulating substrate. The resistance of the platinum wire has a positive temperature coefficient (TC), and these variations of temperature are used to correlate the values of resistance with temperature.

Although more linear than thermocouples, RTDs present important second and third order non-linearities, and depending on the accuracy required by the temperature measurement system, cannot be read with a simple linear amplifier.

These non-linearities can be corrected digitally but, in many applications, it is desired to have a pure analog signal processing circuit [2]. The 4-20 mA current loop transducer widely used in process control in industrial plants [3] is a typical example where a low-cost linearizing analog RTD signal processing circuit could be used with advantages.

In this paper we present an analog linearization technique that uses the intrinsic curvature of the output voltage of a conventional Brokaw bandgap cell to compensate for the non-linearities of the RTD [4].

2. The Platinum RTD

RTDs can operate in a wide temperature range, and for temperatures in the range 0 ºC < T < 661 ºC a platinum Pt-100 RTD is described by the Callendar-Van Dusen equation [3] which, in its simpler form, (also known as the Callendar equation), is given by:

$$R_T = R_0[1 + A\,T + B\,T^2], \tag{1}$$

where A=3.9083×10^{-3} and B=-5,775×10^{-7}, and for a Pt -100 RTD at T=0 ºC, R_T=100 Ω (that is, R_0 = 100 Ω).

Although in many applications the non-linearity of the RTD is disregarded and a simple signal conditioning circuit with a linear amplifier is used. However, in applications where high precision is required, the quadratic term of Eq. (1) cannot be neglected, even in limited temperature ranges.

For example, in the 0 ºC ≤ T ≤ 85 ºC temperature range, the non-linearity of the RTD cannot be neglected, as can be observed in Fig. 1, where the non-linearity of R_T is plotted as a function of the temperature.

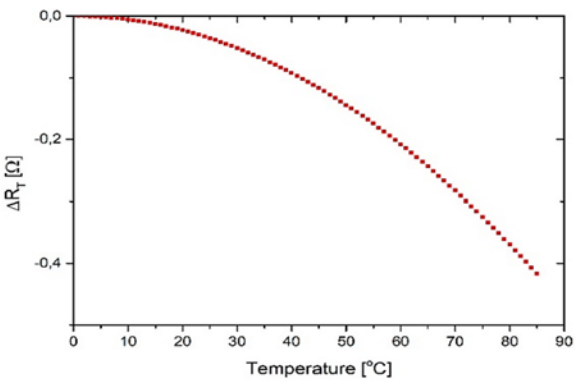

Fig. 1. Non-linearity error in R_T due to the quadratic term of the Callendar-Van Dusen equation.

The calculated value of R_T at T = 100 ºC using Eq. (1) is R_T = 132.80 Ω, while the non-linear part of Eq. (1) is ΔR_T = 0.417 Ω.

3. Amplifying the RTD´s Voltage Signal

There are several techniques to read RTD sensors [5] (as 2-wire, 4-wire, ratiometric), but the most common circuit technique used to read RTD sensors is to force a constant and temperature independent current source (I_0 = 1 mA) through the RTD, and then amplify the voltage difference between the RTD terminals with an instrumentation amplifier.

If the instrumentation amplifier has a high input impedance (low input currents in the inputs), the resistance of the wires that connect the RTD to the instrumentation do not add errors to the measurement, and a simple circuit with a single instrumentation amplifier can be used.

The schematic diagram of a circuit that provides a linear amplification of the voltage across an RTD is presented in Fig. 2. If the instrumentation amplifier has a gain G_v, the output voltage V_O is given by:

$$V_o = G_v.[I_0.R_T] + V_{REF}, \tag{2}$$

where V_{REF} is an external voltage that is summed to the output of the amplifier.

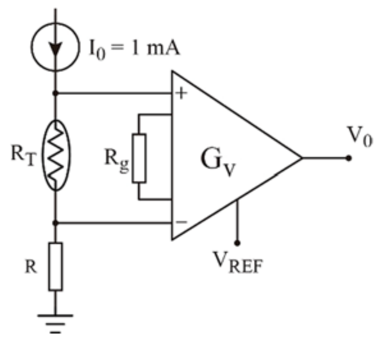

Fig. 2. Basic RTD circuit using an instrumentation amplifier.

If we consider only the linear term of Eq. (1), a temperature change of ΔT = 1 ºC in the RTD results in a R_T change of ΔR_T = 100 Ω×3.9083×10^{-3}= 0.39083 Ω. A common value of full-scale input voltages in 4-20 mA current loop transducers is Vin = 1 V, and if we want to produce an output voltage V_O = 10 mV/ºC (so that at T = 100 ºC we would have V_O = 1.0 V), we need to apply a gain G_v = 10/(1 mA×0.39083 Ω) = 25.58.

However, according to Eq. (2), if we apply this gain with V_{REF} = 0, the output of the amplifier for T=0 ºC (R_T=100 Ω) would be V_O=100 mV×25.58=2.558 V.

Therefore, since we need to have V_O = 0 V at the output of the instrumentation amplifier for T = 0 ºC, it is necessary to have V_{ref} = -2.558 V. In Fig. 3 we present a basic schematic of the linear amplifier described.

The output of a high precision (100 µA ± 0.5 %) and temperature stable (± 25 ppm/ºC) current source REF200 (from Texas Instruments) is amplified by op-amp A1 and resistors R1, R2 (R1 = 10 R2) and a current I_0 = 1 mA is established to drive the RTD. A differential amplifier with gain G_v = 25.58 amplifies the voltage at the RTD terminals, and a voltage V_{ref} = -2.558 V, created by a 2.5 V voltage reference, op-amp A2 and resistors R4, R5.

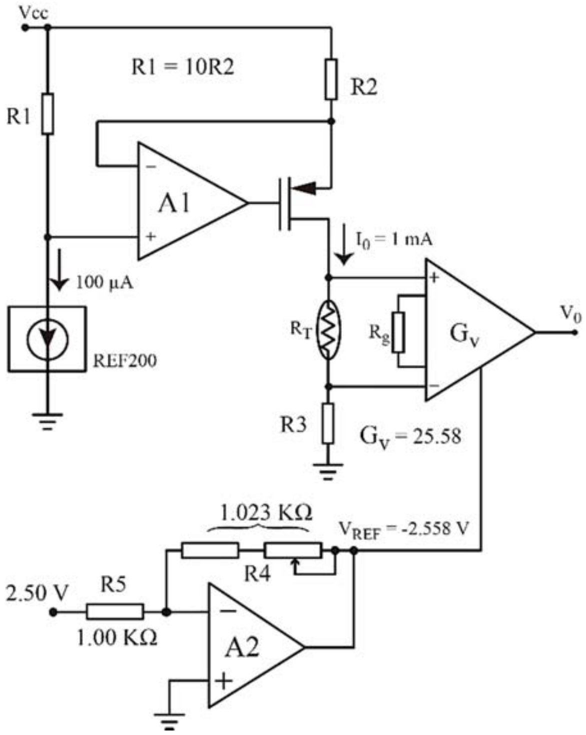

Fig. 3. Simplified schematic of RTD amplifier using an instrumentation amplifier.

The error due to the non-linearity of the output voltage V_O in the circuit of Fig. 3 is plotted as a function of the temperature in Fig. 4. Observing the data in this plot and comparing it to a simple linear output voltage, we see that the error can reach up to 1.26 % at T = 85 ºC.

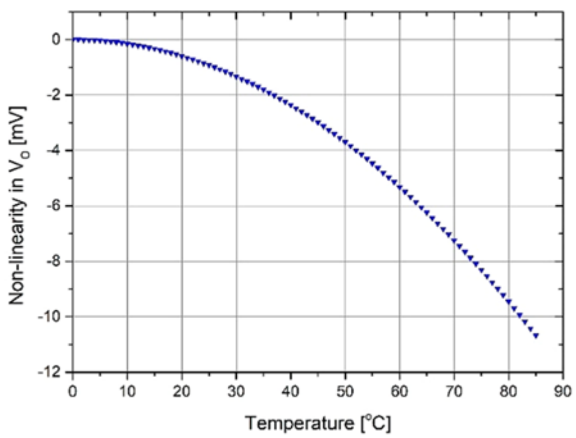

Fig. 4. Non-linearity at the output of the instrumentation amplifier with gain G_v = 25.58.

4. The Brokaw Bandgap Reference Cell

The Brokaw bandgap cell [4] is probably the most used building block in voltage reference ICs. In its discrete version, the Brokaw cell can be implemented with a circuit based on a LM3046 array of bipolar transistors, an op-amp, and some resistors as shown in Fig. 5.

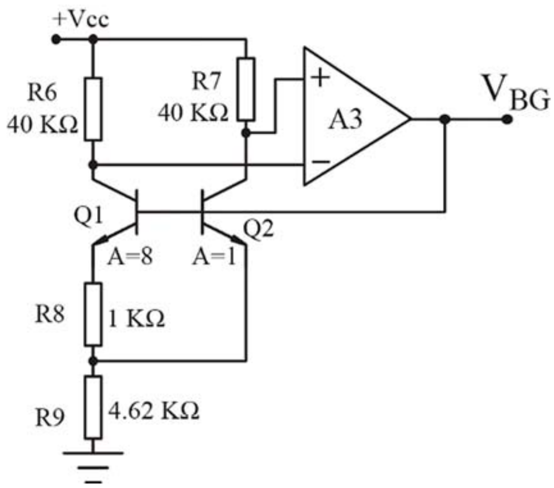

Fig. 5. Discrete implementation of the Brokaw bandgap reference voltage cell.

The op-amp A3 forces the voltage on resistors R6, R7 to be equal, by adjusting the voltage at the connected bases of Q1, Q2. Neglecting the input current of the op-amp and making R6=R7, then the collector currents in Q1 and Q2 are equal.

For these two transistors conducting equal currents, if Q1 has emitter area A1 = 8 and Q2 has emitter area A2 = 1, the difference in their V_{BE} voltages is written as:

$$V_{BE1} - V_{BE2} = \left[\left(\frac{k}{q}\right)\ln(8)\right]T, \tag{3}$$

where k is the Boltzmann constant and q is the charge of the electron.

Therefore, this ΔVBE voltage is proportional to the absolute temperature T (PTAT), and since this voltage appears on resistor R8, we can calculate the current on R8 (which is the same as in Q1) as:

$$I_{R6} = \left[\left(\frac{k}{q\ R8}\right)\ln(8)\right]T \tag{4}$$

Since the currents in Q1 and Q2 are equal and the sum of these currents passes through R9, the voltage drop in R9 is given by:

$$V_{R9} = 2\left[\left(\frac{R9}{R8}\ \frac{k}{q}\right)\ln(8)\right]T \tag{5}$$

The output of the bandgap cell (V_{GB}) is taken at the connected bases of Q1 and Q2, and can be calculated as

$$V_{BG} = V_{R9} + V_{BE2} \tag{6}$$

The variation of V_{BE} with temperature is very well known, and for a transistor biased with a PTAT collector (which is the case of Q2), it can be conveniently expressed by the sum of a constant term, a linear term (that decreases with temperature), and a term that is non-linear with temperature [10]:

$$V_{BE}(T) = [V_{G0} + (\eta - 1)\frac{k}{q}T_r] - \lambda T + \varphi(T), \quad (7)$$

where V_{G0} is the is the extrapolated band-gap voltage at 0 K; η is a constant (dependent on the fabrication process), and T_r is a reference temperature (in K).

The value of the linear term λ is given by:

$$\lambda = \left(\frac{[V_{G0} + (\eta - 1)\frac{k}{q}T_r] - V_{BE}(T_r)}{T_r}\right) \quad (8)$$

And the non-linear term is written as:

$$\varphi = (\eta - 1)\frac{k}{q}\,[T - T_r - T ln\left(\frac{T}{T_r}\right)] \quad (9)$$

Looking into Eq. (6), where a PTAT voltage is summed to a V_{BE} voltage, if we adjust the values of R6, R7 to make the PTAT term of Eq. (5) equal to the linear term λ of Eq. (8), these two terms are cancelled and the value of V_{BG} is described simply by:

$$V_{BG} = V_{G0} + (\eta - 1)\frac{k}{q}T_r + (\eta - 1)\left(\frac{k}{q}\right)[T - T_r - T\ln\left(\frac{T}{T_r}\right)] \quad (10)$$

The first two terms of Eq. (10) are constants, and the remaining term is non-linear with temperature. This non-linear term is responsible for the well-known curvature of the bandgap voltage reference, and if we plot this non-linear term as a function of the temperature, with Tr = 273.15 K (T = 0 ºC), we obtain the curve presented in Fig. 6.

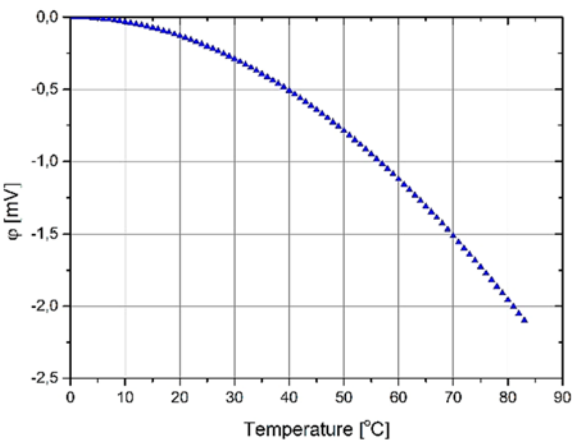

Fig. 6. Non-linear term of Eq. (10) (output voltage of a bandgap reference), as a function of the temperature.

5. Using the Curvature of the Brokaw Bandgap Reference Cell to Compensate for the Non-linearities of RDTs

If we apply a gain to the non-linear curve of Fig. 6 in order to make the values of the non-linear term at T = 85 ºC to be equal to the value of the quadratic non-linearity of V_o in Fig. 4 (also at T = 85 ºC), we obtain the plot shown in Fig. 7.

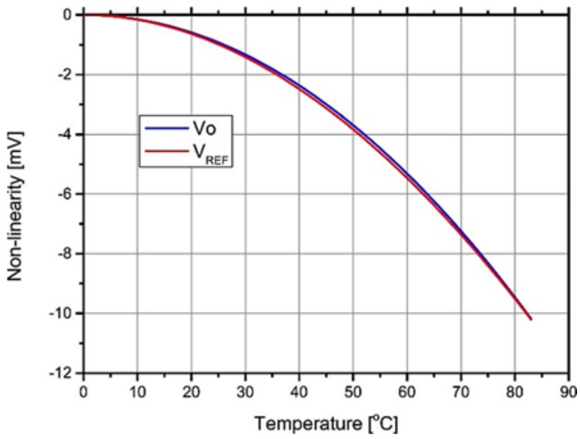

Fig. 7. Comparison of the non-linearity of V_O and V_{REF}.

As we can observe in Fig. 7, both curves almost match perfectly, and if we invert the bandgap curve and sum it to the RTD curve, we will obtain an almost perfect linearization of the response n the output of the amplifier shown in Fig. 3.

From Fig. 4 we can obtain the non-linearity error in V_O for T = 85 ºC as V_{OE} = 10.6758 mV. The value of φ for T = 85 ºC is 2.1070 mV, and if we want to provide a zero error at T = 85 ºC, the compensation signal φ must be amplified by G_{BG} = 10.6758 mV/2.1070 mV = 5.0668.

The circuit which can apply an inverting gain of G_{BG} = - 5.0668 is presented in Fig. 8.

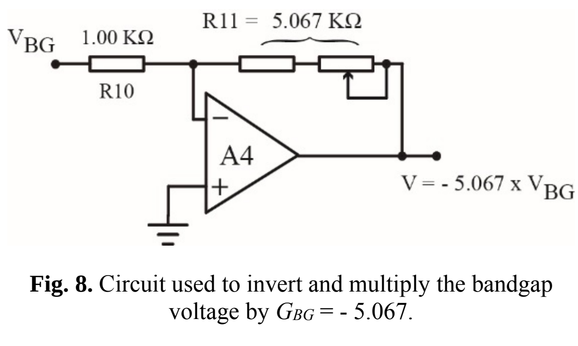

Fig. 8. Circuit used to invert and multiply the bandgap voltage by G_{BG} = - 5.067.

However, if we amplify the value of the bandgap voltage by G_{BG} = - 5.0668, we will multiply both the constant and non-linear terms of Eq. (10).

Using the value of V_{GO} = 1177 mV (as obtained in [10]), using Eq. (10) (with η = 3.12) we calculate the value of V_{BG}, for T = T_r, as V_{BG} = 1229 mV.

However, we need only the non-linear term to create the compensation voltage V_{REF}, and to eliminate the voltage created by the constant term of the VBG, the circuit shown in Fig. 9 was developed.

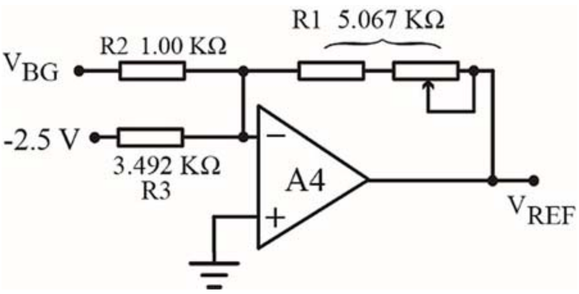

Fig. 9. Circuit used to create the inverted voltage of the non-linear part of Eq. (10) multiplied by G_{BG} = - 5.067.

By applying a negative voltage (-2.5 V) at the R3 resistor, we sum a positive voltage ate the output of the op-amp (that eliminates the dc voltage of the V_{BG} multiplied by G_{BG} = - 5.067). Thus, the resulting V_{REF} voltage consists only of the non-linear part of Eq. (10) multiplied by the same gain, G_{BG} = - 5.067.

The complete circuit used to generate V_{REF} includes another op-amp (A5), connected as a unity gain inverter, and a reference voltage V = 2.5 V, as presented in Fig. 10.

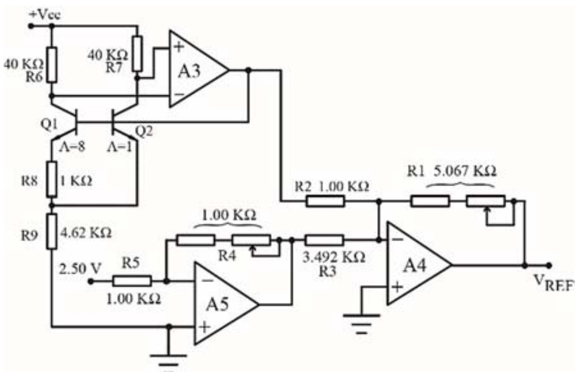

Fig. 10. Circuit used to generate V_{REF}.

6. Results

The non-linearity in the output voltage of the amplifier V_O calculated with V_{REF} obtained from the circuit of Fig. 10 is presented in Fig. 11. When compared to the graph of Fig. 3, where the non-linearity of a non-compensated circuit is presented, we observe that a reduction from 10.27 mV to 143 µV, a two order of magnitude reduction.

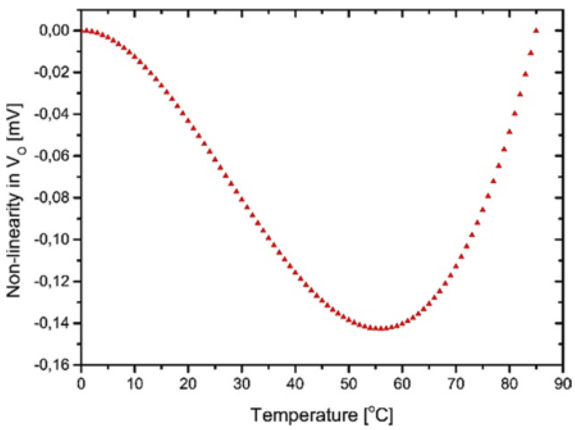

Fig. 11. Non-linearity in V_O calculated with V_{REF} obtained from the circuit of Fig. 10, in the 0 to 85 °C temperature range.

The same principle can be applied to a wider temperature range, and in Fig. 11 we present the plot of the non-linearity in V_O when the same technique is applied in the 0 to 125 °C.

Although the technique is not so efficient as in the case of a limited temperature range, it is still possible to observe a reduction of more than 70 % in the non-linearity, from 23 mV to 6.4 mV.

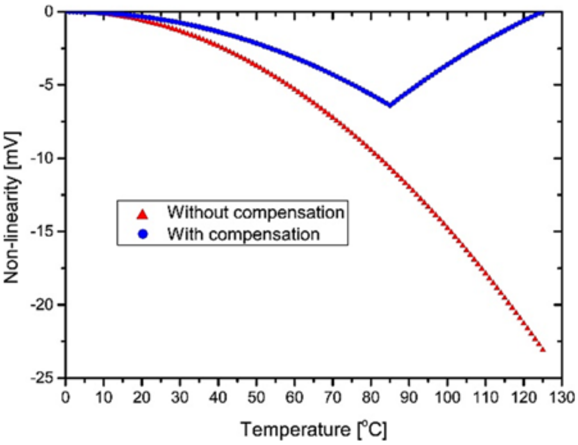

Fig. 12. Comparison of the non-linearities errors of a linear circuit and the proposed technique, in the 0 to 125 °C temperature range.

Even though we are dealing with errors in the order of a few microvolts, we decided to perform an electrical simulation using the software LTSPICE to check if the technique is robust. As the complete models of the transistors in the LM3046 array are not available, transistors Q1-Q2 were simulated using the available models of bipolar transistors from a conventional bipolar process used to fabricate the 741 op-amp [11], since these transistors are fabricated using the same type of process of the LM3046.

The first simulation was realized just to check how the parameters of the transistors used in the

741 op-amp would affect the behavior of the bandgap voltage reference.

We noticed that a small difference between the non-linear term φ obtained in the simulation and calculated using the NPN parameters the values given in [10]. In Fig. 13 it is shown the comparison of the calculated and the simulated curvature of the bandgap reference circuit presented in Fig. 5. The observed differences are due to the differences in the values of η and V_{G0} in Eq. (10).

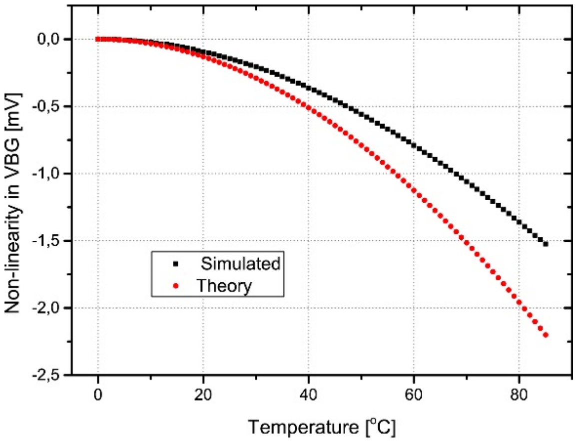

Fig. 13. Difference in the non-linearity of V_{BG} in the circuit of Fig. 5, simulated and calculated with the theoretical equation of the V_{BE} of the transistors, in the 0 °C to 85 °C temperature range.

In Fig. 14 we present a comparison of the value of the quadratic non-linearity of V_o calculated from the circuit of Fig. 3 and the result obtained in the simulation of the circuit presented in Fig. 10. Due to the variations of the parameters of the NPN transistors, minor adjustments in the values of resistors R1 (R1 = 7.006 kΩ), R3 (R3 = 3.184 kΩ) and R9 (R9 = 4.7756 kΩ) were necessary to make, at T= 85 °C, the simulated values of the non-linear term to be equal to the value of the quadratic non-linearity of V_o in Fig. 4.

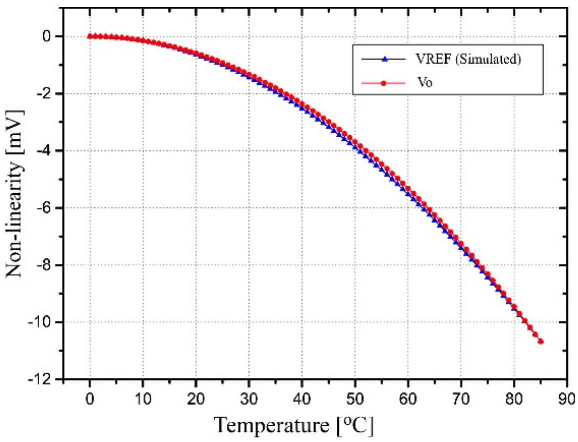

Fig. 14. Comparison of the non-linearity of V_O (from the circuit of Fig. 3) and in V_{REF} from the simulated circuit.

The non-linearity in the output voltage of the amplifier V_O calculated with V_{REF} obtained from the electrical stimulation of the circuit presented in Fig. 10 is shown in Fig. 15.

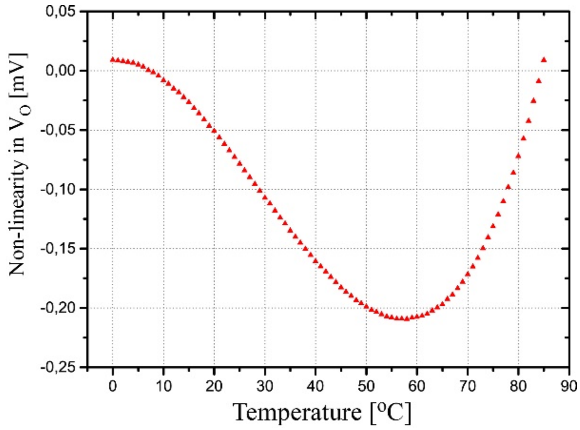

Fig. 15. Non-linearity in V_O with V_{REF} obtained from a simulation of the the circuit of Fig. 10, in the 0 °C to 85 °C temperature range.

Observing the results shown in Fig. 15 we see that the shape of the error curve is similar to the plot shown in Fig. 11, but performance of the linearization obtained with the simulated circuit is not as good as those calculated with the theoretical bandgap circuit.

It was interesting to notice that, due to the limited resolution of the LTSPICE algorithms, we could not "trim" the resistors in the program to obtain the desired values. An example of such limitation of the electrical simulation can be seen in the error at T= 0 °C (Fig. 15), where we could not find a value of the resistor R3 that would make the value of VREF exactly equal to -2.558 V.

Even with these limitations, we see that the maximum error shown in Fig. 15 was limited to 209 µV. Thus, when a simulation of the circuit is performed in LTSPICE, a reduction in the error from 10.27 mV to only 209 µV was obtained with the proposed technique.

7. Conclusions

A simple and extremely efficient technique for the analog linearization of an RTD signal processing circuit using the curvature of a Brokaw bandgap reference cell was presented. A signal processing circuit with a discrete bandgap Brokaw cell generates a curve that almost matches the inverse curvature of the non-linearity of a RTD, and the sum of the curves implement an excellent linearization technique.

The technique can reduce, in the 0 to 85 °C temperature range, the non-linearities errors of the signal processing circuit by two orders of magnitude, from 10.27 mV (calculated with a linear signal processing circuit, that is, with VREF constant), to only 143 µV.

An electrical simulation (LTSPICE) of the proposed circuit was performed using the model of typical NPN transistors found in the 741 op-amp and, after a few adjustments of the resistors, the circuit showed that a reduction of the non-linearities errors of the signal processing circuit from 10.27 mV to only 209 µV can be achieved.

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