

Simulation And Analysis of Various Bandgap Reference Circuits

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Abstract. In analog integrated circuits, it is essential to provide accurate and undisturbed circuits. Bandgap reference circuits could use a Bipolar transistor to generate positive temperature coefficient voltage and negative temperature coefficient voltage, which could reduce the influence of temperature. Therefore, it is used widely in analog integrated circuits. This paper first introduces the basic notion of a bandgap reference circuit, including its characteristics that are not affected by voltage, temperature, and process, so that the circuit works stably in a stable state. Then, the circuit structure of several mainstream bandgap reference sources is deduced and analyzed and their performance is verified by simulation experiments. Finally, a circuit with a higher-order temperature coefficient is proposed analyzed, and simulated.

Keywords: Bandgap reference; zero-temperature coefficient; higher-order temperature coefficient; simulation.

1. Introduction

With the development of electrical technology, the performance and reliability requirements of integrated circuits are constantly increasing. The bandgap reference circuit could provide stable reference voltage, and it is not affected by temperature, process, or voltage. Thus, it plays a significant role in electronic systems, especially in ADC, DAC, voltage stabilizers, and other fields.

However, designing a stable bandgap reference circuit requires a deep understanding of the characteristics of the relevant analog integrated circuit components. Moreover, precise control of each parameter in the circuit and comprehensive consideration of the impact of the process, temperature, and voltage-related parameters. Therefore, this paper will start from the point of view of the principal introduction of the reference circuit and the formula derivation of the bandgap reference circuit. Next, analyze several kinds of mainstream bandgap reference structures and verify the simulation circuit. Finally, the influence of the higher-order temperature coefficient of the circuit is explained, and a new reference voltage source using the higher-order temperature coefficient to compensate for the voltage is proposed.

2. Bandgap Reference Circuit

The integrated circuit requires voltage or current reference. In actual operation, every circuit needs a reference to stabilize the output voltage, ensuring the circuit works stably in an ideal state. For instance, in ADC and DAC conversions, the reference voltage must be precisely compared to the input voltage to obtain the correct result. The ideal reference circuit should meet the requirements of PVT: voltage-independent, temperature and process-independent.

In practical work, temperature has a significant impact on the reference of circuits, and new designs are needed to compensate for this deficiency. BJT transistor has a negative temperature coefficient voltage V_{BE} , which will decrease as the temperature increases, and a positive temperature coefficient ΔV_{BE} , which is caused by the voltage difference. By combining the two, temperature-independent reference voltage or current can be obtained, which is the bandgap reference.

In the design process, many typical bandgaps reference circuits need to consider various factors, each with its advantages and disadvantages. In the following, it will be elaborated on one by one.

3. Formula Derivation of the Bandgap Reference Circuit

The purpose of the formula derivation of the bandgap reference circuit is to find a V_{ref} where the weighted sum of the temperature coefficients voltage is 0. BJT transistors are the proper target because they can produce positive and negative temperature coefficient voltages. Through the related design of BJT transistors, an ideal circuit with zero temperature coefficient voltage or current can be realized.

3.1. Negative Temperature Coefficient Voltage

For BJT transistor, its PN junction diode has a positive voltage V_{BE} with a negative temperature coefficient, $I_C = I_S \exp\left(\frac{V_{BE}}{V_T}\right)$, and $V_T = \frac{kT}{q}$, $I_S = bT^{4+m} \exp\left(\frac{-E_g}{kT}\right)$, k is the Boltzmann constant, equal to 1.38×10^{23} J/K. E_g is the bandgap energy of silicon, equal to about 1.12eV. m is the temperature index of minority carrier mobility, equal to about -1.5. It can be concluded that.

$$V_{BE} = V_T \ln\left(\frac{I_C}{I_S}\right) \tag{1}$$

Take the derivative of this.

$$\frac{\partial V_{BE}}{\partial T} = \frac{\partial V_T}{\partial T} \ln\frac{I_C}{I_S} - \frac{V_T}{I_S} \frac{\partial I_S}{\partial T} \tag{2}$$

$$\frac{\partial V_{BE}}{\partial T} = \frac{\partial V_T}{\partial T} \ln\frac{I_C}{I_S} - \frac{V_T}{I_S} \frac{\partial I_S}{\partial T} \tag{3}$$

As a result:

$$\frac{\partial V_{BE}}{\partial T} = \frac{V_{BE} - (4+m)V_T - E_g/q}{T} \tag{4}$$

At room temperature, generally $V_{BE} = 750mV$, $\frac{\partial V_{BE}}{\partial T} = 1.5mV/K$ [2].

3.2. Positive Temperature Coefficient

If two BJT transistors operate at different current densities, their difference in V_{BE} is proportional to temperature. The specific derivation is as follows:

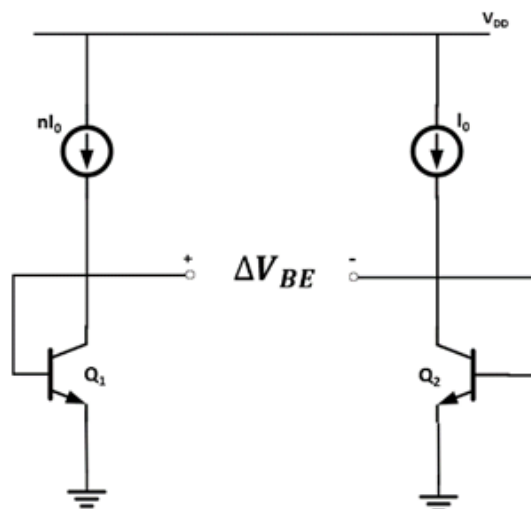


Figure. 1 Positive temperature coefficient

As shown in Figure 1, the current density of Q_1 is n times that of Q_2 .

$$\begin{aligned}
 \Delta V_{BE} &= V_{BE1} - V_{BE2} \\
 &= V_T \ln \frac{n I_0}{I_{S1}} - V_T \ln \frac{I_0}{I_{S2}} \\
 &= V_T \ln n \\
 &= V_T \ln n \tag{5}
 \end{aligned}$$

Therefore, ΔV_{BE} can represent a positive temperature coefficient.

3.3. Bandgap Reference

By the weighted sum of the positive temperature coefficient voltage and the negative temperature coefficient voltage: $V_{REF} = a V_{BE} + b V_T \ln n$.

Take the derivative of this and get $\frac{\partial V_{REF}}{\partial T} = a \frac{\partial V_{BE}}{\partial T} + b \frac{\partial V_T \ln n}{\partial T}$. When the first partial derivative of V_{REF} with respect to temperature is 0, the relationship between a and b can be obtained. At room temperature, generally $V_{BE} = 750mV$, $\frac{\partial V_{BE}}{\partial T} = -1.5mV/K$. If a=1, then b=17.2, that is: $V_{ref} \approx V_{BE} + 17.2V_T = 1.2V$ [1].

In this way, by superimposing the positive temperature coefficient voltage and the negative temperature coefficient voltage of the BJT transistor, a zero-temperature coefficient voltage relationship can be obtained. Using this relationship, more and more perfect reference and circuits can be designed in the future.

4. Structural Analysis of Several Mainstream Bandgap Reference Sources

4.1. Bandgap Reference Voltage Generation Circuit

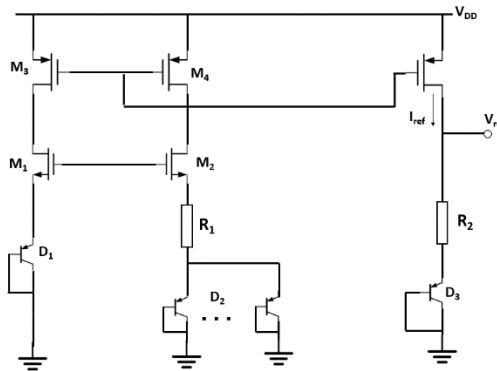


Figure. 2 Bandgap reference voltage generation circuit

In Figure 2, D_2 expands the emission area by connecting multiple bipolar transistors in parallel [1]. The reference voltage is $V_{REF} = V_{D3} + I_{REF}R_2$, the voltage of negative temperature coefficient is V_{D3} , and the voltage of positive temperature coefficient is $I_{REF}R_2$, $I_{REF} = \frac{V_{D1}-V_{D2}}{R_1} = V_T \times \frac{R_2}{R_1}$. Positive temperature coefficient voltage is equal to $I_{REF}R_2 = \ln N \times V_T$. At normal temperature, $\frac{R_2}{R_1} \ln N = 17.2$. The reference voltage is $V_{REF} = V_{D3} + 17.2V_T$.

Therefore, only if $\frac{R_2}{R_1} \ln N = 17.2$ is satisfied, the positive temperature coefficient and the negative temperature coefficient can cancel each other, so as to achieve a stable bandgap voltage reference. In the process design process, $\frac{R_2}{R_1} = 5$, $n=31$ is generally selected to make this circuit.

4.2. Generating Circuit Using Bandgap Reference Voltage of Operational Amplifier

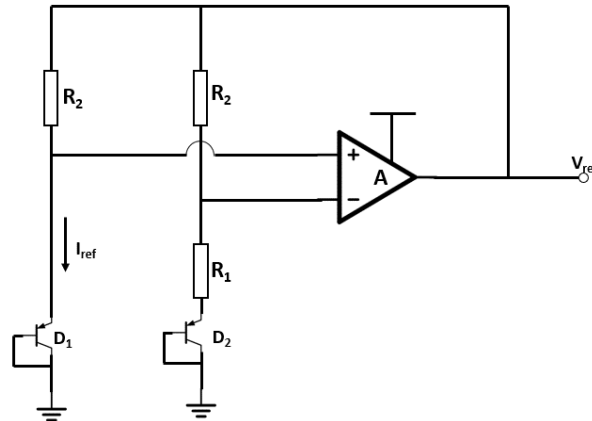


Figure. 3 Generating circuit using bandgap reference voltage of operational amplifier.

As shown in Figure 3, D_2 is also multiple bipolar transistors in parallel to increase the emission area, and $V_A = V_B$ is achieved through virtual short and negative feedback of the operational amplifier [3]. In this circuit, the reference voltage $V_{REF} = V_{D1} + I_{REF2} = V_{D2} + I_{REF}(R_2 + R_1)$, where $I_{REF} = V_T \times \ln n \times \frac{R_2}{R_1}$. After simplification, $V_{REF} = D_2 + V_T \times \ln n \times (1 + \frac{R_2}{R_1})$.

4.3. Low Pressure Bandgap Reference

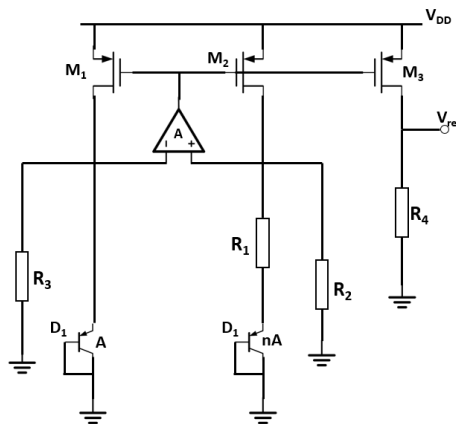


Figure. 4 Low pressure bandgap reference

As shown in Figure 4, an arbitrary voltage value with zero temperature coefficient can be converted by adding the current with positive temperature coefficient and the current with negative temperature coefficient [2]. Due to the negative feedback characteristics of the amplifier, $V_X = V_Y$, $V_{R3} = V_{Q2} + I_{Q2}R_1$. Similarly, it still exists $V_{Q1} - V_{Q2} = V_T \ln n$, $I_1 = \frac{V_T \ln n}{R_1}$, $I_4 = I_1 + I_2$, and $I_2 = \frac{V_{Q1}}{R_2}$. The combination is $I_{R4} = \frac{V_T \ln n}{R_1} + \frac{V_{Q1}}{R_2}$. After simplification, we can get $V_{BG} = \frac{R_4}{R_2} \left(V_{BE1} + \frac{R_2}{R_1} V_T \ln n \right)$ [2].

As before, the zero-temperature coefficient voltage can be reached if $\frac{R_2}{R_1} V_T \ln n = 17.2$ at room temperature. In this case, due to the existence of $\frac{R_4}{R_2}$, it can be lowered below the traditional limit of 1.2V [2].

5. Simulation Circuit Verification

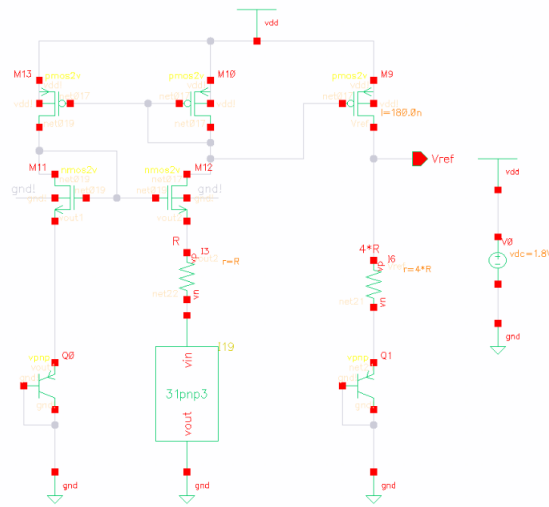


Figure. 5 first-order temperature coefficient bandgap reference circuit diagram

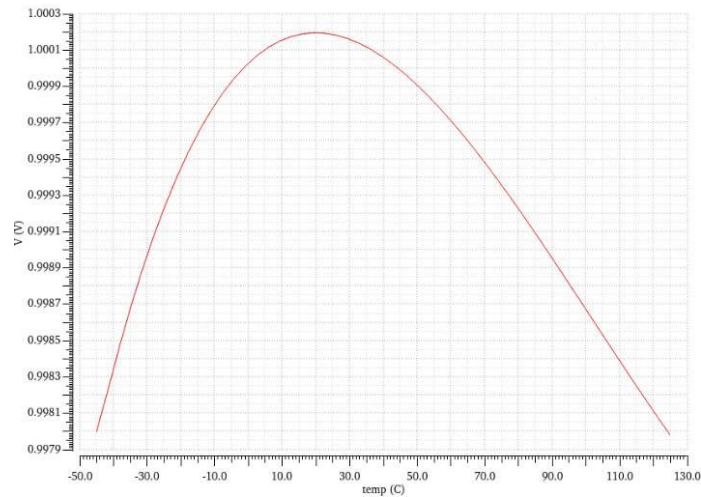


Figure. 6 the result obtained by using DC temperature.

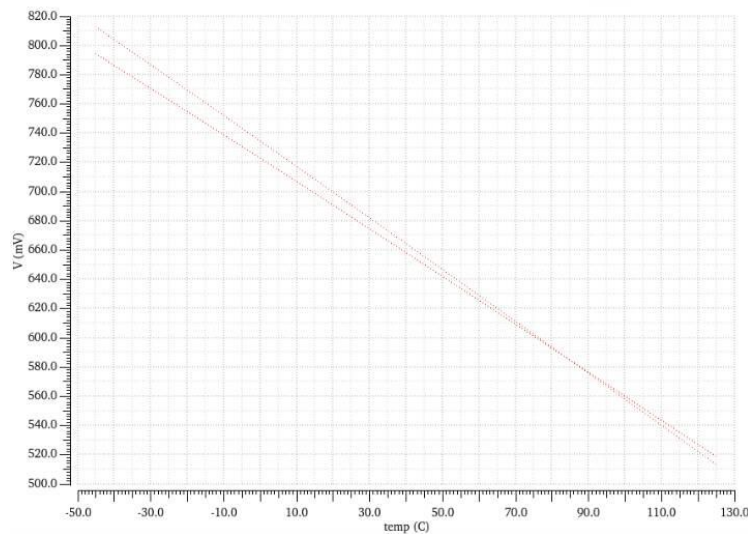


Figure. 7 The electrical mismatch results

The simulation circuit in this paper adopts the CMOS technology of TSMC 0.18 μm . Figure 5 is a first-order temperature coefficient bandgap reference circuit diagram, and Figure 6 is the result

obtained by using DC temperature scanning (-45, 125). Through the relevant calculation, we can get the temperature coefficient of this circuit $TC=31\text{ppm/K}$, is a more qualified bandgap reference source.

There are many areas for improvement in this simulation. The first is due to the use of current mirrors to design bandgap reference voltage sources. The electrical mismatch results in a certain gap between the voltage values of M_6 and M_3 in Figure 7. The difference between M_6 and M_3 is also amplified, leading to a peak voltage of only around 1V. After calculation, it is found that reducing the difference between M_6 and M_3 significantly decreases the error. The second problem is that the previous analysis and calculation are based on the general $V_{BE} = 750\text{mV/K}$, $\frac{\partial V_{BE}}{\partial T} = -1.5\text{mV/K}$. However, as the temperature changes, this value will also change. The effect of this parameter should be further considered through simulations or higher-order models. The third thing that can be improved is that the effect of temperature on the circuit is not simply $V_T = \frac{kT}{q}$. There are also higher-order temperature coefficients that need to be taken into account, and these related problems will be analyzed and improved in the following paper.

6. Analysis and Design of Higher-order Temperature Coefficient Circuit

In the previous formula derivation, the first-order temperature compensation of BJT transistor is based on the positive temperature characteristic of thermal voltage. However, the existence of higher-order temperature coefficient voltage is not taken into account, so the temperature coefficient is too high. With the deepening of the research, it is found that there is a nonlinear higher-order temperature coefficient. It is possible to compensate for the nonlinear term of higher-order temperature coefficients and more accurately reduce the temperature coefficient of the output voltage. The following is the expression of the higher-order temperature coefficient of V_{BE} :

$$V_{BE}(T) = V_G(T) + [V_{BE}(T_0) - V_G(T_0)] \frac{T}{T_0} + (\eta - \alpha)V_T \ln \frac{T}{T_0} \quad (6)$$

$V_G(T)$ refers to the bandgap voltage of silicon material at temperature T , and $V_G(T_0)$ represents the bandgap voltage of silicon at same reference temperature T_0 . $V_{BE}(T_0)$ is the voltage at the emitter of the transistor at the reference temperature T_0 . The parameter η is the coefficient of the process properties of the semiconductor material, and its value is usually between 3.6 and 4. α describes the behavior of the current flowing through the transistor as a function of temperature: When the bias current is proportional to the temperature, $\alpha = 1$; If the bias current is not significantly related to temperature, $\alpha=0$. When expanded according to Taylor's formula, we get [4]:

$$V_{BE}(T) = a_0 + a_1T + a_2T^2 + a_3T^3 + \dots \quad (7)$$

a_0, a_1, a_2, a_3 represent constants. It can be found that $V_{BE}(T)$ is composed of multi-order temperature coefficients, which is not a simple linear relationship with temperature.

In related research, there are some higher-order temperature coefficient compensation methods. Such as PTAT2 compensation, resistance proportional compensation, exponential curvature compensation, piecewise curvature compensation, subthreshold mos compensation and other methods [5]. The higher-order temperature coefficient voltage is cleverly weakened to obtain a more stable bandgap reference circuit [5].

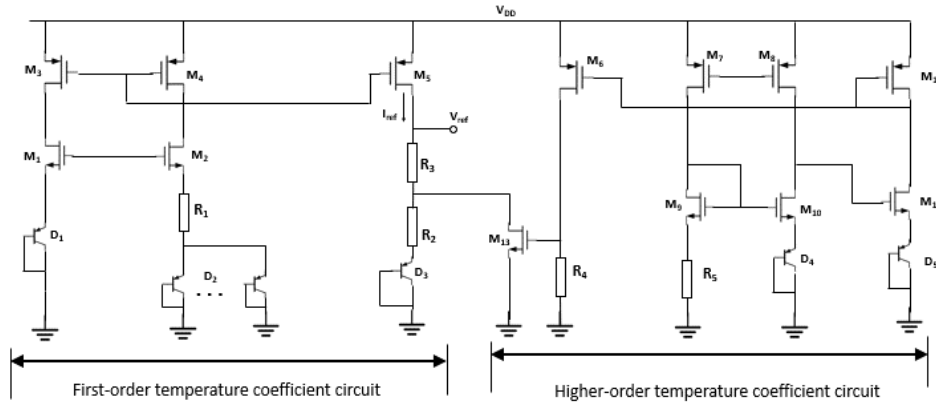


Figure. 8 The design of a higher-order temperature compensation circuit

Figure 8 shows the design of a higher-order temperature compensation circuit using the current characteristics of the subthreshold region to compensate. On the left is the circuit with the first-order temperature coefficient simulated above, and on the right is the circuit using exponential curvature to compensate for higher order temperature [6]. M_6 and M_{11} constitute the structure of the current mirror, with a width to length ratio $\frac{W_6}{L_6} = M(\frac{W_6}{L_6})$, so $I_6 = M \times I_{11}$. The current flowing through R_4 makes up the Complementary to absolute temperature (CTAT) current. By selecting the appropriate values for M , R_4 , R_5 , the M_{13} transistor is placed in the subthreshold region. At this time, the current I through M_{13} is [7]:

$$I = \mu_n C_{OX}(n - 1) \left(\frac{W_7}{L_7}\right) V_{TH}^2 \exp\left(\frac{V_{GS} - V_{TH}}{nV_T}\right) \quad (8)$$

μ_n is the electron mobility of the NMOS tube, $\mu_n = \mu_0(T/T_0)^{-2.3}$. C_{OX} is the gate oxide capacitance per unit area [8]. The threshold voltage V_{TH} is a negative temperature coefficient, which decreases linearly with the increase of temperature.

$$V_{TH} = V_{TH0} - \delta(T - T_0) \quad (9)$$

Where, $\delta \approx (1-4) \text{ mV}/^\circ\text{C}$; V_{TH0} is the threshold voltage when the temperature equals T_0 . Since the V_{GS} of M_{13} is V_4 , it can be approximated.

$$V_{GS7} = V_{GS0} - \beta(T - T_0) \quad (10)$$

Substituting the equation gives:

$$I_{D7} = AT^{-0.3} \exp\left[\frac{V_{GS0} - V_{TH0} + (\beta - \delta)T_0}{nV_T}\right] q \quad (11)$$

Among them, $A = \mu_0 T_0^{2.3} C_{OX}(n - 1) \left(\frac{W_7}{L_7}\right)$, is a constant and independent of the temperature. Due to the TMS18 process, the M_{13} transistor is in the subthreshold region, so $V_{GS0} - V_{TH0} \approx 50\text{mV}$; $V_{TH0} \approx 420\text{mV}$, $V_{BE0} \approx 750\text{mV}$; $\frac{R_5}{R_4} \approx 0.5$. Therefore, it can continue to be simplified to:

$$I_{D7} = AT^{-0.3} \exp\left(\frac{m}{nT}\right) \quad (12)$$

$m = V_{GS0} - V_{TH0} + (\beta - \delta)T_0 < 0$. The partial derivative of the temperature can be obtained:

$$\frac{\partial I_{13}}{\partial T} = A \frac{e^{-\frac{m}{nT}} (CT^{1.3} - 0.3T^{2.3})}{T^{3.6}} \quad (13)$$

Where $C = -\frac{m}{n}$; $n = 1 + \frac{C_d}{C_{OX}}$, C_d is the drain capacitance. The zero point of the equation is $T_1 = \frac{C}{0.3}$. When $T \leq T_1$, I_{D2} will increase with the increase of temperature, and when $T \geq T_1$, I_{D2} will decrease with the increase of temperature. The variation in the image of the bandgap reference with the first-

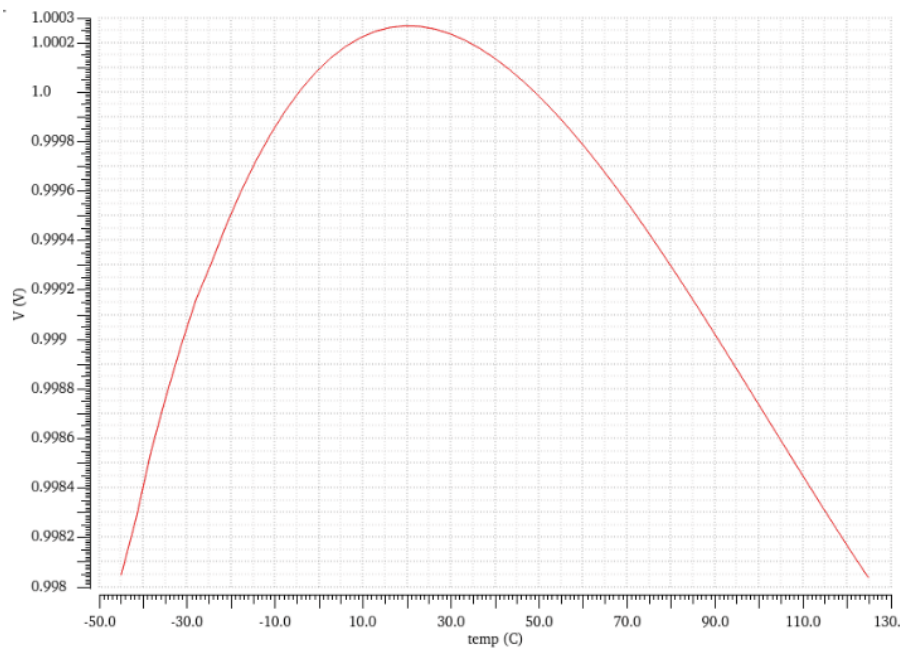


Figure. 11 the result obtained by using DC temperature scanning.

Figure 9 is the circuit diagram of higher-order temperature simulation. Figure 10 is the image of I_{13} changing with temperature, and Figure 11 is the result obtained by using DC temperature scanning (-45-125).

It can be seen from Figure 10 that the image of I_{13} changing with temperature is consistent with the trend of voltage change in Figure 6. Figure 11 shows that the higher-order temperature coefficient TC of voltage is 20 ppm/K. Compared with the voltage temperature coefficient of the first-order temperature coefficient, the temperature coefficient is reduced by about 33%. It can be seen that the design of higher-order temperature coefficient circuit can significantly reduce the temperature coefficient and improve the temperature stability of the reference voltage.

7. Conclusion

In this paper, the structures of several mainstream bandgap reference sources are analyzed, and the simulation circuit is verified by $0.18\mu\text{m}$ CMOS process. Then, the effect of higher-order temperature coefficient on the circuit is discussed, and the method of exponential curvature compensation is proposed. Finally, a circuit with higher-order temperature coefficient characteristics is designed and simulated. The simulation results show that the temperature coefficient TC reaches 20ppm/K, which is 33% lower than that of the first-order temperature coefficient simulation, and the circuit stability is improved. In integrated circuit design, it can be used as a reference voltage source for AD/DA conversion circuit, ADC acquisition circuit, etc.

However, due to the use of current mirror to design the bandgap reference voltage source, the mismatch of the first-order temperature coefficient voltage does not reach the ideal 1.2V. Therefore, the simulation of the higher-order temperature coefficient does not reach the ideal voltage. In the subsequent design, the virtual short characteristic of the amplifier can be adopted to design the bandgap reference circuit, so as to avoid this situation.

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