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Silicon Temperature Sensors

E. Habekotté

Silicon temperature sensors exploit the inherent temperature sensitivity of silicon. Yet, the possibilities to measure the temperature accurately are limited to bipolar devices and MOS devices in weak inversion with their exponential temperature behaviour. The three basic principles to measure the temperature and some simplified circuit examples are presented. Some special cases of sensors in bipolar as well as in CMOS technology show the variety of possibilities in this field of sensors.

Les capteurs de température au silicium utilisent la sensibilité inhérente du silicium à la température. Toutefois les possibilités de mesurer la température de façon précise sont limitées aux transistors bipolaires et aux transistors MOS à inversion faible, grâce à leur comportement exponentiel. L'article présente les trois principes de base permettant de mesurer la température ainsi que des exemples de circuits simplifiés. Quelques cas spéciaux de capteurs bipolaires et en technique CMOS font preuve d'une grande variété de possibilités.

Silizium-Temperatursensoren basieren auf der inhärenten Temperaturempfindlichkeit von Silizium. Genaue Messungen sind jedoch nur mit Bipolartransistoren sowie mit MOS-Transistoren in schwacher Inversion möglich, die ein exponentielles Temperaturverhalten aufweisen. Die drei Messprinzipien sowie verschiedene vereinfachte Schaltungen werden gezeigt. Einige spezielle Sensoren in Bipolar- sowie in CMOS-Technologie veranschaulichen die Vielfalt von Möglichkeiten.

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1. Introduction

Silicon is in most cases more sensitive to temperature than one would like it to be. In search for silicon temperature sensors it is not only the sensitivity to temperature one is looking for but also the predictability, the reproducibility and the linearity of this behaviour.

For example, one could think of using a diffused or implanted resistor (R) as a temperature-measuring device; its temperature dependency follows the one of the mobility μ ($R \approx 1/\mu\alpha T^{2.5}$) [1]. However, the large nonlinear relationship between the resistor value and the temperature dictates the necessity of linearization even for small temperature ranges (10 to 20 °C). Further, the absolute output value (voltage or current) of this type of sensor is difficult to control as the absolute value of the resistor can vary somewhere between 20 and 50% around its mean value. This makes the necessary calibration rather complicated as it should allow for at least 50% adjustment of the output value.

Nevertheless, the diffused resistor can very well be used to compensate for similar temperature behaviour ($\alpha \cdot 1/\mu$) elsewhere on the same chip, as in that case the matching properties in temperature behaviour are of more interest than the absolute accuracy.

But the resistor is not very useful as monolithic temperature sensor. Thus, the possibilities to measure accurately the temperature with a silicon sensor are in fact limited to bipolar transistors and to MOS transistors in weak inversion which have a similar temperature behaviour as bipolar devices. This similarity means that in many cases it is possible to copy, partially, bipolar sensors one to one in CMOS technologies.

On the other hand, there are two bipolar transistors available in every CMOS technology: a vertical one and a lateral one as shown in figure 1 [2]. They allow to realize certain bipolar temperature sensors directly in CMOS technologies however, with the following two restrictions:

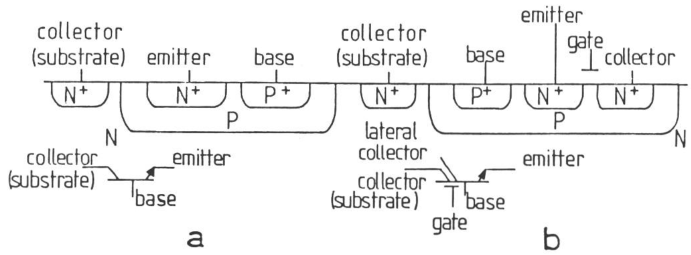
- the collector of the vertical transistor is not available as it is common with the substrate and
- the emitter of the lateral transistor is not available either because the emitter current is the sum of two emitter currents, one coming from the parasitic vertical transistor and the other one coming from the lateral transistor.

Therefore, it is sufficient for most temperature sensors to give only the bipolar circuit schematics as an example of the three basic principles to measure the temperature.

2. Bipolar temperature sensors

Silicon is very sensitive to temperature, especially the base-emitter vol-

Fig. 1
Bipolar transistors available in a *p*-well CMOS Technology [2]
a Vertical bipolar transistor
b Lateral bipolar transistor



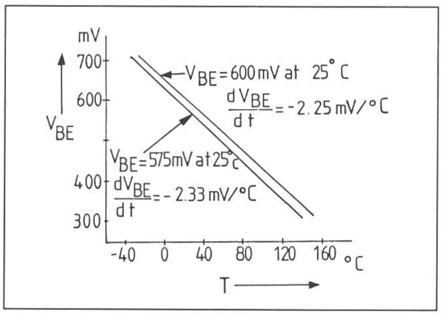


Fig. 2 Base-emitter voltage V_{BE} of two bipolar transistors of the same type as function of the temperature T

tage of a bipolar transistor. In figure 2 is shown how the base-emitter voltage V_{BE} depends on the absolute temperature; it is almost a straight line with a negative temperature coefficient (approximately -2.3 mV/K). A very important point concerning the base-emitter voltage is that its temperature behaviour is very well described and predictable. However, the absolute value for V_{BE} depends on much less predictable variations in the production process, so that it is necessary to calibrate the absolute output voltage (current) of such a sensor at a certain temperature. The inclination depends on the current density at which the bipolar transistor will be operated, and can thus be adjusted by choosing the correct current density (fig. 3).

The inherent non-linearity (curvature) is shown in figure 4 as a temperature error for a large temperature range of almost 200 °C around a calibration point (no temperature error at this point). The total curvature is relatively small so that the base-emitter junction of a bipolar transistor can directly be used to measure the absolute temperature if no high accuracy (not better than 2 to 3%) is required for large temperature ranges. However, for small temperature ranges around the calibration point a relatively accurate temperature measurement stays feasible with only one bipolar transistor as measuring device.

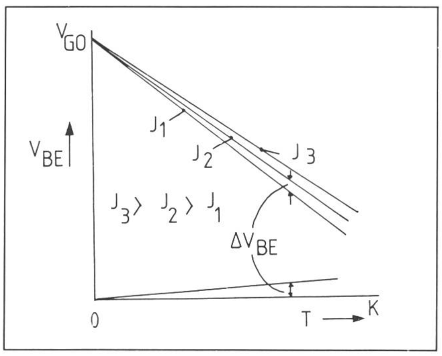


Fig. 3 V_{BE} and ΔV_{BE} for different current densities J as $F(T)$

It is possible to compensate the nonlinearity by introducing a current which depends on the absolute temperature to a certain power as shown in figure 5 [4]. This compensation technique becomes evident by looking at the mathematical description of the ideal collector current I_C for a bipolar transistor:

$$I_C = I_S \exp[V_{BE}/(kT/q)]$$

$$= \{\alpha\, T^r \exp[-V_{G0}/(kT/q)]\}$$

$$\exp[V_{BE}/(kT/q)] \qquad (1)$$

where

- I_s reverse current of an ideal diode
- V_{BE} is assumed to be a few times larger than (kT/q)
- α constant, which depends on parameters such as the base width; it may thus differ from one transistor to another, even for transistors of the same type. It is proportional to the effective emitter surface A_E
- T absolute temperature
- r constant, which mainly depends on the temperature dependency of the mobility of minority carriers in the base region ($2 < r < 4$)
- V_{G0} extrapolated bandgap voltage of silicon (approximately 1.2 V)
- q electron charge ($1.6\ 10^{-19}$C)
- k Boltzman's constant ($k/q = 86.1706\ \mu V/K$)

Equation (1) does not only show the well-known exponential impact of V_{BE} on the collector current but it exhibits as well how the temperature influences the current.

Rewriting this expression reveals how V_{BE} depends on the extrapolated bandgap voltage V_{G0} and a term where the absolute temperature appears through the thermal voltage (kT/q) and through a logarithmic dependency:

$$V_{BE} = V_{G0} - (kT/q)\ln(\alpha\, T^r/I_C) \qquad (2)$$

It is this logarithmic dependency which provokes the nonlinear behaviour. By introducing a collector current, which is not a constant but which is proportional to the absolute temperature T or even to T^r, it is possible to reduce to a large extent the negative influence of the curvature as indicated in figure 5.

It is also possible to cancel out the nonlinearity by using the excellent

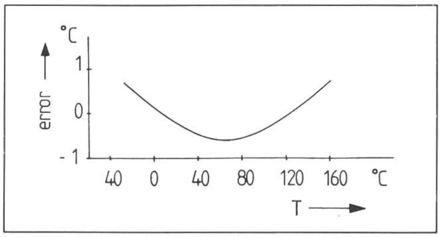


Fig. 4 Inherent nonlinearity of a bipolar transistor as $F(T)$

matching properties between similar devices on the same chip. In that case no complex compensation is needed for high accuracy. Note that I_S in equation (1) is proportional to the effective emitter surface A_E through α. If one takes the difference between two base-emitter voltages of two bipolar transistors operated at different current densities J_1 and J_2, the following relationship can be found:

$$\Delta V_{BE} = V_{BE1} - V_{BE2} = (kT/q)\ln(J_2/J_1) \qquad (3)$$

which is directly proportional to the absolute temperature T. As J_1 and J_2 are easily realized by introducing the same collector current in both transistors with different emitter surfaces A_{E1} and A_{E2} respectively, equation (3) will simply become

$$\Delta V_{BE} = (kT/q)\ln(A_{E2}/A_{E1}) \qquad (4)$$

This of course holds only if good matching can be realized between the two transistors, which is the case when both transistors are close together on the same chip and if possible with one center of gravity. The best way to exploit this is using several times (4 to 24) the emitter surface of the smaller transistor to create the larger transistor by putting them in parallel and symmetrically around the smaller one. The emitter-surface ratio defined in such a way can be very accurate (better then 1%) as it is only one mask which defines the wanted ratio.

It is due to this almost perfect measure of the absolute temperature that many temperature sensors do exploit

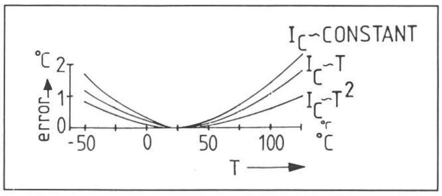


Fig. 5 Compensation of the inherent nonlinearity of a bipolar transistor

ΔV_{BE}. Note that the ΔV_{BE} has a positive temperature coefficient and, due to the In (ratio), presents a small voltage of only a few times kT/q (26 mV at 27 °C), while V_{BE} itself has a negative temperature coefficient and presents a relatively large voltage (fig. 2). The voltage difference ΔV_{BE} is usually called V_{PTAT}, which stands for Voltage Proportional To the Absolute Temperature.

One important disadvantage of the so far presented methods to measure the absolute temperature is the necessary high resolution for further signal processing; the required resolution in order to detect a temperature change of e.g. 0.1 K at 300 K with V_{BE} = 600 mV and dV_{BE}/dT = -2 mV/K comes to 3000:1, which corresponds to 12 bits. It is therefore better to convert the information on the absolute temperature into a more handy temperature scale (°C). This can be done by shifting the absolute temperature scale with a reference voltage to a celsius temperature scale. The introduction of a separate reference voltage will, however, reduce the precision of the temperature sensor due to the very small but still existing erroneous-temperature dependence of the reference. Furthermore, this voltage reference is a relatively expensive component.

A way to avoid this problem is an intrinsically referenced temperature sensor. Here a combination of V_{BE} and ΔV_{BE} is used as a measure for the temperature; by forming the difference between V_{BE} with negative and ΔV_{BE} with positive temperature coefficient it is possible to introduce the desired scale conversion while the two temperature coefficients are added to each other with the same sign [5; 6].

As a conclusion of this section there are three basic methods of measuring the temperature [5]:

- the base-emitter voltage of one bipolar transistor as measure for the absolute temperature with negative temperature coefficient;
- the difference between two base-emitter voltages of bipolar transistors operated at different current densities as a measure of the absolute temperature with positive temperature coefficient;
- a combination of a base-emitter voltage and the difference between two base-emitter voltages with either positive or negative temperature coefficient presenting a measure of the temperature in any desired scale.

3. MOS temperature sensors

The MOS transistor in weak inversion exhibits a similar exponential relationship between the control (gate) voltage V_G and the channel current I_D as the one between V_{BE} and the collector current I_C for bipolar transistors [7]:

$$I_D = I_{D0} \exp[V_G/(n\,U_T)]\,[\exp(-V_S/U_T) - \exp(-V_D/U_T)] \qquad (5)$$

where

$I_{D0} \approx S$	$S \exp[-V_T/(n\,U_T)]$, which is very sensitive to process parameters
S	W/L
W	width of the channel
L	length of the channel
V_T	threshold voltage of the MOS transistor
U_T	thermal voltage (kT/q)
n	slope factor $1.2 < n < 2$ depending on technology and transistor type, it is fairly controllable but depends slightly on the temperature
V_G	gate voltage referred to the substrate potential
V_S	source voltage referred to the substrate potential
V_D	drain voltage referred to the substrate potential

Equation (5) simplifies in case V_S = 0 and V_D is a few times larger than U_T to

$$I_D = I_{D0} \exp[V_G/(n\,U_T)] \qquad (6)$$

Equation (6) put into logarithmic form shows how good the temperature behaviour of MOS transistors in weak inversion resembles the one of bipolar tranistors (equ. 2). Therefore bipolar temperature sensors can almost directly be copied in a MOS technology by replacing the actual temperature measuring bipolar transistors with MOS transistors in the weak inversion mode. As MOS transistors do not have a noticeable input (gate) current, the usual base-current compensation in bipolar temperature sensors is not necessary in the MOS equivalent of the sensor.

However, the absolute output level of an all-MOS sensor will be less well controlled than it is the case with a bipolar sensor due to the relatively large mismatch in the threshold voltage. Further, there is a small but non-neglectable temperature dependency of the slope factor n which has to be taken into account if high accuracy is wanted. And finally, the current levels

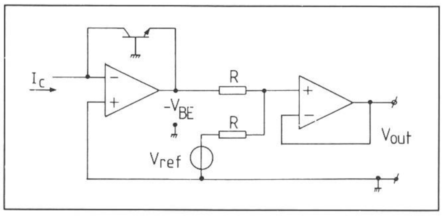


Fig. 6 One transistor temperature sensor [4]

for the weak inversion mode are below several µA, so that already at relatively low temperatures (70–80 °C) the temperature behaviour of the leakage currents overwhelms the one of the MOS transistors [8].

4. Examples of the three basic principles

Figure 6 shows a simplified circuit, where the first temperature measurement method is implemented. The temperature-measuring transistor is taken in the feedback loop of an operational amplifier. In this way the collectorbase junction is almost short-circuited so that it hardly introduces an error. The leakage and recombination-generation currents in the total emitter current do not flow in the collector circuit; they are delivered by the output of the operational amplifier. So, they will not interfere, either. The output of this circuit is $-V_{BE}$. In case another temperature scale than kelvin is needed the conversion is possible by adding a reference voltage V_{ref} to $-V_{BE}$. A good compensation of the non-linearity in V_{BE} can be realized by introducing a collector current I_c which is not constant but is proportional to the square of the temperature as already suggested. In [4] a deviation from linearity within ±0.1 °C throughout a measuring temperature range of -50 to 125 °C has been reported. This was for a discrete circuit realization, where the collector current was made proportional to the square of the absolute temperature.

The second method to measure the temperature is depicted in the figures 7

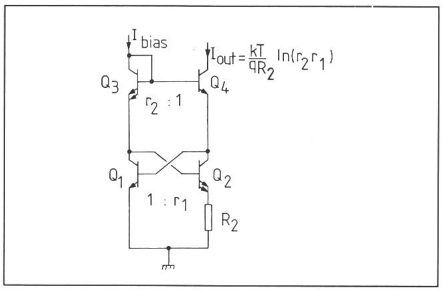


Fig. 7 I_{PTAT}-sink [5]

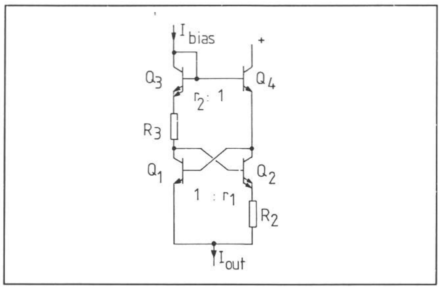


Fig. 8 I_{PTAT}**-source [5]**

and 8. Here, two possibilities to realize relatively large V_{PTAT}s are given [5]. Going through the circuit in figure 7 starting at the base of Q_3 one recognizes the sum (V_{BE3} + V_{BE2}), plus a voltage drop across resistor R_2 and this total has to be equal to (V_{BE1} + V_{BE4}) going back from the emitter of Q_1 to the base of Q_3. So, the voltage drop across resistor R_2 can be written as

$$V_{R2} = V_{BE1} - V_{BE2} + V_{BE4} - V_{BE3} \quad (7)$$

Rewriting (7) and using the logarithmic relationship results in a V_{PTAT}

$$V_{R2} = (kT/q)\ln(r_1 \cdot r_2) \quad (8)$$

Due to the product r_1 times r_2 it is possible to realize relatively large V_{PTAT}s with small emitter-area ratios. This voltage is converted by resistor R_2 into a current, which is proportional to the absolute temperature.

The circuit in figure 8 converts in a similar way two V_{PTAT} voltages across R_3 and R_2 into a total current proportional to the absolute temperature which in case R_3 equals R_2 comes to the same result as found for the circuit in figure 7. The circuit in figure 7, however, sinks the PTAT current, while in figure 8 it is a PTAT-current source.

Note that the incoming bias current value does not influence the found relationship. This special feature comes from the crosscoupling of transistors Q_1 and Q_2; it is much better than realizing the same ratio product with a PNP current mirror; PNP bipolar transistors are lateral devices and show due to the parasitic, vertical PNP transistor rather poor matching properties in comparison with the vertical NPN bipolar transistors especially when relatively large emitter-area ratios are needed.

The actual realization of V_{PTAT} sources is much more complicated than indicated here because the negative influence of the base currents on the temperature behaviour have to be compensated in order to achieve high accuracy [5; 9]. Further, bipolar sensors using current mirror configurations need start-up circuitry which will make the sensor more complex than actually needed.

In the shown circuits (fig. 7, 8) one voltage with a positive temperature coefficient (ΔV_{BE}) and one with a negative temperature coefficient V_{BE} are available. So, the difference between these two voltages will again result in a temperature-dependent voltage but now in any desired temperature scale. This type of intrinsically referenced temperature sensor presents the third method to measure the temperature. Figure 9 gives an example of such a sensor with current output [5]. The netto output current is the difference between the PTAT current and the current flowing through resistor R_1 imposed by the base-emitter voltage V_{BE4} of Q_4 neglecting the effects of the base currents:

$$I_{out} = V_{PTAT}/R_2 - V_{BE4}/R_1$$

$$= (1/R_1)\,[V_{PTAT}(R_1/R_2) - V_{BE4}] \quad (9)$$

The nonlinearity in V_{BE4} is partly compensated as a PTAT current is forced to flow into the collector of Q_4 by the PNP-current mirror (1:1). Therefore, V_{BE4} will show an almost linear decrease with the temperature:

$$V_{BE} = V_{G0} - CT \quad (10)$$

where C is a constant. In case R_1 and R_2 do match in temperature behaviour the temperature dependence of their ratio may be neglected. Using (10) and assuming that the output current is zero at $T = T_Z$, equation (9) can be rewritten as

$$I_{out} = (V_{G0}/R_1)\;(T - T_Z)/T_Z \quad (11)$$

R_1 can be used to adjust the desired sensitivity and R_2 [in $(R_1/R_2)\,V_{PTAT}$] al-

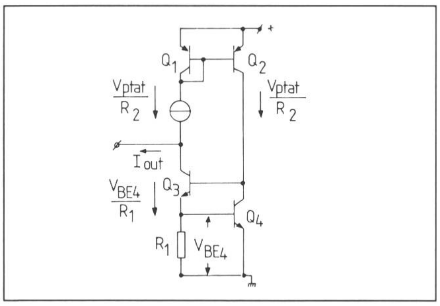


Fig. 9 Intrinsically referenced current source [5]

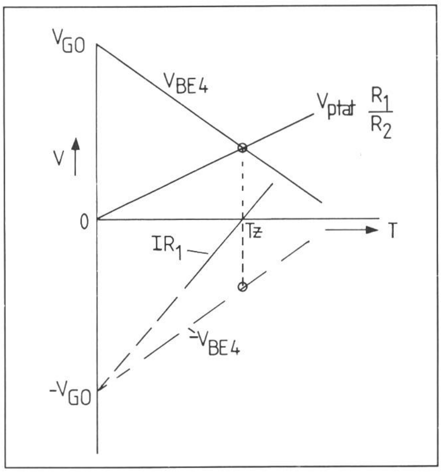


Fig. 10 Calibration of the temperature sensor from figure 9

lows to adjust the output current to zero at a given T_Z as indicated in figure 10. Note that the resistors should have a very small temperature dependency. In [5] it is shown that thin-film resistors with a temperature coefficient of approximately 92 ppm/°C will compensate the still existing nonlinearity due to V_{BE4}.

All temperature sensors realized in a bipolar technology will look similar to the given examples following the three basic principles for temperature measurement. This is also true for most CMOS temperature sensors [10] using either the available bipolar transistors (fig. 1), with their restrictions, or MOS transistors in the weak inversion mode as temperature-measuring device. However, in a CMOS process there are two other approaches possible.

5. Special CMOS approaches

One totally different approach possible to measure the temperature in a CMOS process exploits the temperature behaviour of the threshold voltage of a MOS transistor. The temperature coefficient is approximately −2 to −3 mV/°C. This type of temperature sensor will suffer a lot from the large variation in the absolute value of the threshold voltage (±100 mV) of a MOS transistor.

The second typical CMOS approach would be the implementation of switched-capacitor techniques [8]. It is still the base-emitter voltage or the difference between two of them, which will be used as a measure of the temperature. However, the temperature-dependent voltage or voltage differ-

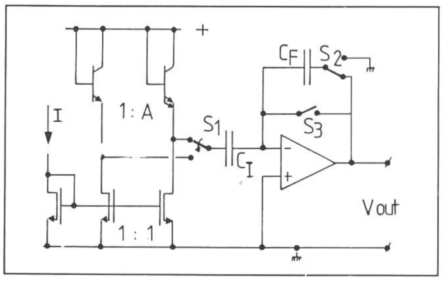


Fig. 11 Switched capacitor approach of a V_{PTAT}-source [8]

ence is now introduced into the input capacitor of a differential amplifier by switching between the two terminals of the diodes (switch S_1 in fig. 11). This means that the output voltage has to be sampled and held as well. Switch S_3 takes care of the DC biasing of the amplifier by resetting the amplifier between every two measurements. Switch S_3 in combination with S_2 reduces offset and $1/f$ noise by double correlated sampling. Figure 11 shows an example of a switched-capacitor V_{PTAT} source using the vertical bipolar transistor with collector common with substrate, available in every CMOS process. The output voltage of the amplifier is simply the capacitor ratio (closed loop gain) times the voltage difference between the two base-emitter junctions operated at different current densities (emitter-area ratio A)

$$V_{out} = (C_I / C_F)\ (kT/q) \ln A \qquad (12)$$

6. Bipolar micropower temperature sensor

This section gives an example of an integrated micropower low-voltage temperature sensor with *frequency output* realized in a bipolar process. It shows that micropower circuits are not the exclusive domain for CMOS technologies. This temperature-controlled oscillator is one of the latest developments of temperature sensors done at the University of Delft [3]. It has been implemented in a standard bipolar IC technology and thick-film technology. The blockdiagram of the temperature-to-frequency converter is given in figure 12. First the information on the absolute temperature is converted into a current I_{PTAT}. This current is used in

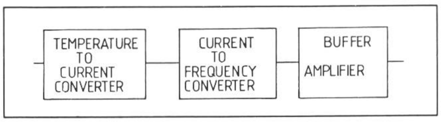


Fig. 12 Temperature-to-frequency converter [3]

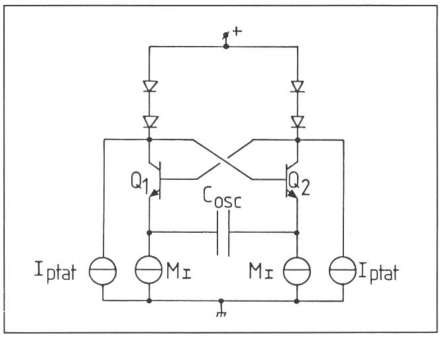


Fig. 13 Current controlled single-capacitor oscillator

an oscillator whose oscillation frequency depends linearly on the output current of the temperature-current converter. The output frequency is meant to be 100 Hz/°C so that the temperature can then be directly displayed on a frequency counter. In order to minimize the problems with the relatively large dimensions of non-integrated capacitors, the chosen oscillator type uses only one capacitor. The principle is shown in figure 13.

The frequency is directly proportional to the current in the emitter branch and inversely proportional to the product of the emitter-coupled capacitor C_{osc} and the peak-to-peak voltage swing on this capacitor. By making the current proportional to the absolute temperature, the frequency of this oscillator will also be proportional to the absolute temperature. However, by using an intrinsically referenced temperature-dependent current source it is possible to calibrate the whole circuitry to have a more appropriate temperature scale. The basic part of such a temperature-dependent current source is indicated in figure 14; the cross-coupled pair forces a V_{PTAT} across resistor R_1, so that the current through R_1 is PTAT as well. There is further an extra current flowing through R_2 and its temperature behaviour is totally defined by the difference between the V_{PTAT} across R_1 and the base-emitter

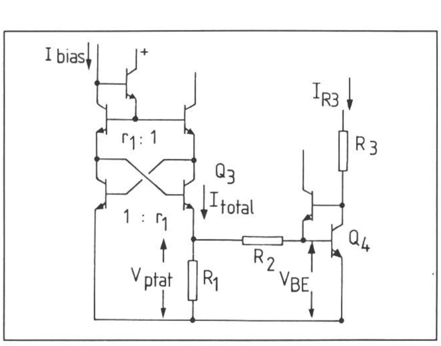


Fig. 14 Temperature-to-current converter

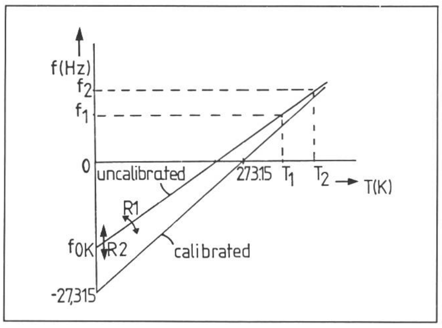


Fig. 15 Calibration of the temperature controlled oscillator

voltage of Q_3. The sum of both currents flows through Q_4:

$$I_{tot} = I_{Em\,Q4} =$$

$$[V_{PTAT}(R_1 + R_2)/R_1 - V_{BE4}](1/R_2) \qquad (13)$$

The temperature sensitivity of the total current can be adjusted by changing R_1. Changing R_2 will influence the absolute level of the total current. By introducing this current in the oscillator circuit it is possible to calibrate the absolute level as well as the temperature coefficient of the output frequency as shown in figure 15.

The circuit can measure body temperatures within the range of 32 to 42 °C with an accuracy of ±0.1 °C. The temperature coefficient can be calibrated to 100 Hz/°C. The power consumption is less than 40 μW for 2.7 V supply voltage.

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