integrated circuits and components for

bandgap references and temperature transducers

gerard c.m. meijer
Integrated Circuits and Components for Bandgap References and Temperature Transducers

Gerard C.M. Meijer

Electronics Research Laboratory
(Dissertation Delft 18 March 1982),
Department of Electrical Engineering,
Delft University of Technology,
Delft, The Netherlands.

Postal address: P.O. Box 5031, 2600 GA Delft, The Netherlands.
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References

Samenvatting
Voltage references are applied in data acquisition systems, voltage regulators and a large variety of measurement equipment. With all of these applications the magnitude of an unknown voltage is determined by measuring its ratio with respect to a reference voltage.

The magnitude of the reference voltage has to be known every time and under all circumstances in which the measurement has to be performed. Uncertainty about its value directly limits the accuracy of the measurement. Therefore, a voltage reference should supply a voltage with a very low temperature coefficient, a great long- and short-term stability, a low internal resistance and an insensitivity to thermal and mechanical shocks.

Nowadays, generally bandgap or zener references are applied.

In bandgap references the output voltage is obtained by adding a correction voltage $V_{c}(T)$ to the base-emitter voltage $V_{BE}(T)$ of a bipolar transistor (the reference transistor) in order to cancel its temperature dependence.

Since the introduction of this principle in 1971 by Widlar [1.1] a lot of material has been published about bandgap references and a large number of commercial products in which these references have been applied have become available. Bandgap references have to compete with zener references. The main advantages of bandgap references concern their low supply voltage, low power dissipation and good long-term stability. A surprisingly high precision can be achieved with bandgap references. The temperature coefficient (TC) of a base-emitter voltage is rather high: circa 3000 ppm/$^\circ$C. The TC of the output voltage of the best bandgap references is less than 2 ppm/$^\circ$C. This means that the reduction in TC is over a factor of 1000!

This high order of compensation has to be maintained over a period of many months and under various external circumstances without recalibration. The calibration has to be performed during production in a simple, inexpensive way.

In fact, the realization of accurate bandgap references clearly demonstrates the amazing rate of precision achieved with analog IC's. The correction voltage $V_{c}(T)$ needed to compensate for the TC of the base-emitter voltage of the reference transistor is in the same order of magnitude as $V_{BE}$ itself. Therefore, demands with respect to accuracy are as stringent for the correction voltage as for the $V_{BE}$ of the reference transistor. The only voltage with the desired thermal behavior which is accurate enough to be used for this purpose is the difference $\Delta V_{BE}$ of the base-emitter voltages of a matched pair of transistors operated at unequal emitter-current densities. This differential voltage is proportional to the absolute temperature (PTAT) when the ratio $p$ ($p \neq 1$) of the current densities is constant.

In bandgap references signals with a relatively high TC are generated and processed in a highly accurate way. The same type of component and circuits can be used for the implementation of IC temperature transducers. For the last few years these temperature transducers have provided a welcome solution for many temperature measurement problems. They deliver an accurate electrical output signal in the temperature range from $-55^\circ$C to $+150^\circ$C.

During the course of the work presented in this dissertation a number of new, accurate circuits for bandgap references and temperature transducers have been developed. The limitations with respect to accuracy have been investigated.

Much emphasis has had to be put on the thermal behavior of components and circuits, and many second-order effects also had to be taken into account.

To achieve high accuracy not only the components and circuits but also the layout had to be carefully designed in order to optimize the matching of components and to minimize the effects of thermal gradients and mechanical stress.

A brief outline of the material found in this dissertation is given below.

The considerations of and the investigations into the properties of the applied components and the influence of thermal-electrical interaction and mechanical stress have been clustered in Chaps. 2. and 3.

The reader who is mainly interested in the circuit designs may omit these chapters and simply adopt
the main results derived from these considerations. In Chapter 2 the properties of the components used for the generation and amplification of the basic signals in bandgap references and temperature transducers are discussed. In Chapter 3 the temperature rise of the chip due to internal power dissipation is considered. With the results of these considerations values are found for the maximum admissible power dissipation of temperature transducers. Furthermore, both empirical and calculated data are presented about the temperature gradients caused by internal dissipation in the chip. These data are important for an optimal design of the layout of the chip. This chapter also includes the influence of mechanical stress on circuit performance. It will be argued that this stress is the dominant source of inaccuracy in many well-designed circuits. In Chapter 4 accurate bandgap-reference circuits developed during the course of this work are presented. In Chapter 5 various methods of using bipolar transistors to sense the temperature are discussed. Much attention has been paid to accuracy, circuit simplicity and minimizing the effort needed to calibrate the devices. There is a large variety of temperature-transducer applications. The demands for specific applications can better be met by special circuit designs. Examples of such special designs are described. In Chapter 6 the main conclusions of this work are summarized.
2. COMPONENTS FOR IC BANDGAP REFERENCES AND TEMPERATURE TRANSDUCERS

2.1 Introduction

The favorable properties of IC bandgap references and temperature transducers are due to the highly predictable and time-independent way in which the base-emitter voltage of a bipolar transistor is related to the temperature.

For this reason bipolar technology has been selected to fabricate the devices. Bandgap references fabricated in MOS technology [2.1] are less accurate and therefore left out of consideration in this work.

In this chapter firstly the properties of the most important component - the bipolar npn transistor - are dealt with.

The considerations about the current-voltage characteristic and the temperature dependence of the base-emitter voltage of the bipolar transistor presented in Sections 2.2.1 and 2.2.2 provide the theoretical basis for the circuit designs dealt with in this dissertation. For high accuracy second-order effects, such as base-width modulation and temperature dependence of the current gain which are discussed in Sections 2.2.3 and 2.2.4, also have to be taken into account.

Mismatch of npn transistors and nonidealities of PTAT voltages, as dealt with in Sections 2.2.5 and 2.2.6, are decisive for the accuracy of the devices.

In Section 2.3 the properties of lateral pnp transistors are discussed. It is shown that the nonidealities in their properties are so large that they cannot be used in the basic reference and sensor cells. However, in the supporting electronics lateral pnp's are very useful, especially for level shifting.

Resistors are used for voltage-to-current conversion and for setting the amplifier gain. Their properties have been dealt with in Section 2.4. Finally, some attention is paid to the Seebeck effect. It is shown that in the design of the resistor layout this effect has to be taken into account.

2.2 The bipolar npn transistor in the forward active mode

In bandgap references and temperature transducers bipolar npn transistors are used to generate and amplify the basic signals. In this section the d.c. behavior of these transistors is considered, with emphasis on the temperature dependence of their properties. In our applications the transistors are usually operated at low voltages and low currents in order to obtain high accuracy and to minimize power dissipation.

Therefore, no attention has been paid to effects such as high-level injection and avalanche multiplication. The sign conventions used for transistor voltages and currents are shown in Fig. 2.1.

2.2.1 The $I_C (V_{BE})$ characteristic

The main properties of bipolar transistors applied for bandgap references and temperature transducers are revealed by the well-known equation for the collector current $I_C [2.2]$

$$I_C = I_S \exp \left( \frac{qV_{BE}}{kT} \right), \quad (2.1)$$

where

$T = \text{the absolute temperature},$

$V_{BE} = \text{the base-emitter voltage},$

$q = \text{the electron charge},$

$k = \text{the Boltzmann constant},$

and $I_S$ is given by

$$I_S = \frac{q^2 \mu A_E \bar{D}}{2 A_E B}, \quad (2.2)$$

with

$A_E = \text{the emitter-junction area},$

$n_i = \text{the intrinsic carrier concentration in the base},$

$\bar{D} = \text{the effective minority-carrier diffusion constant in the base},$
\[ Q_B = \text{the charge represented by the net number of} \]
\[ \text{doping atoms in the neutral base per unit area.} \]

The charge \( Q_B \) can be written as
\[ Q_B = q \int_{x_E}^{x_C} N_B \, d \xi, \quad (2.3) \]
where \( N_B \) is the net base-doping density and \( x_E \) and \( x_C \) represent the boundaries of the neutral base region on the emitter and the collector side, respectively. These boundaries depend on the junction voltages. This causes the base-widening effect which will be discussed in Section 2.2.4.

The base current is at low current levels dominated by mechanisms such as surface recombination and recombination in the depletion regions. At this current level the base current is proportional to
\[ \exp \left( \frac{qV_{BE}}{mkT} \right), \]
where the nonideality factor \( m \) is between 1 and 2 [2.3]. Due to the nonideality of the base current the temperature dependence of the \( I_C(V_{BE}) \) characteristic can be predicted with greater precision than that of the \( I_E(V_{BE}) \) characteristic.

At moderate current levels the base current \( I_B \) is mainly due to injection of holes from the base into the emitter. Then \( m = 1 \) and it holds that [2.4]
\[ I_B = \frac{q n_{IE}^2 A E \nu_B}{G_E} \exp \left( \frac{qV_{BE}}{kT} \right), \quad (2.4) \]
where
- \( n_{IE} \) = the intrinsic carrier concentration in the emitter,
- \( G_E \) = a constant which is called the effective Gummel number for the emitter,
- \( \nu_B \) = the effective minority-carrier diffusion constant in the emitter.

### 2.2.2 The temperature dependence of the base-emitter voltage

To calculate the temperature dependence of the base-emitter voltage \( V_{BE} \) we consider the temperature dependence of the terms of (2.1) and (2.2).

From [2.5] we have
\[ n_{IE}^2 = T^3 \exp \left( -\frac{qV_g}{kT} \right), \quad (2.5) \]
\[ \nu_B = \left( \frac{kT}{q} \right) \tau_B, \quad (2.6) \]
where
- \( \nu_B \) = the effective value of the mobility of electrons in the base,
- \( V_g \) = the bandgap voltage of the base material.

The base charge \( Q_B \) also depends somewhat on the temperature because of the temperature dependence of the boundaries \( x_E \) and \( x_C \). However, calculations showed that this effect is negligible in all practical cases dealt with in this dissertation.

The mobility \( \nu_B \) and the bandgap voltage \( V_g \) are related to the temperature in a nonlinear way. When we make the common approximations [2.5]:
\[ \nu_B = T^{-n}, \quad (2.7) \]
\[ V_g = V_{g0} - \alpha T, \quad (2.8) \]
where \( n \) and \( \alpha \) are constants and \( V_{g0} \) is the extrapolated bandgap voltage at 0 K, and then substitute (2.5) and (2.6) in (2.1) and (2.2) and neglecting the Early effect it is found that:
\[ I_C = C \tau^n \exp \left( \frac{qV_{BE} - V_{g0}}{kT} \right), \quad (2.9) \]
where
- \( C \) = a constant,
- \( \tau = 4-n \).

An accurate description of the temperature dependence of the \( I_C(V_{BE}) \) characteristic is very important for the designers of bandgap references and temperature transducers.

In applying approximations (2.7) and (2.8) we found it necessary to check the validity of (2.9). For this purpose we performed the measurements described in Appendix A. The empirical values \( V_{g0} = 1166 \text{ mV} \) and \( \eta = 3.72 \) were found from these measurements. These values differ considerably from those expected on the basis of physical considerations. Tsividis [2.6] showed that this is mainly due to the poor approximation (2.8) for the \( V_g(T) \) dependence and presented a more accurate, physically based analysis. However, even with this analysis the resulting equations found are not as accurate as is desired in bandgap reference applications.

In Appendix B it is shown that equation (2.9) with empirical values for \( n \) and \( V_{g0} \) can provide the
Fig. 2.2 The base-emitter voltage $V_{BE}$ versus the temperature $T$. The curvature is exaggerated in order to indicate the characteristic points clearly.

required accuracy. For this reason and also because of its simplicity we will use this equation to describe the temperature dependence of the $I_C(V_{BE})$ characteristic.

To examine spread in $V_0$ and $n$ a number of measurements have been performed on transistors located on different chips on one wafer and having a large range of bias currents. The small spread found was within the range determined by the measurement accuracy, which amounted to 0.03 K.

These measurement results confirm the accuracy of (2.9) with the best fitting values for $V_0$ and $n$. To develop the equations for $V_{BE}(T)$ we consider two temperatures: an arbitrary temperature $T'$ and a specified reference temperature $T_r$. Applying (2.9) for each temperature, one can derive the following equation:

$$V_{BE}(T) = V_0 + (n-m) - \frac{kT}{q} \ln \frac{T}{T_r} + \frac{kT}{q} \ln \frac{I_C(T)}{I_C(T_r)}.$$  

(2.10)

For practical reasons in many of the circuits dealt with in this dissertation the collector current is made proportional to some power of $T$:

$$I_C = T^m.$$  

(2.11)

Substituting this in (2.10) gives:

$$V_{BE}(T) = V_{BE}(T) + \frac{kT}{q} \ln \frac{T}{T_r} + (n-m) - \frac{kT}{q} \ln \frac{T}{T_r}.$$  

(2.12)

As will become clear in the following chapters it is convenient to express $V_{BE}(T)$ as the sum of a constant term, a term proportional to $T$, and higher-order terms in such a way that the linear terms represent the tangent to the $V_{BE}(T)$ curve for $T = T_r$ (Fig. 2.2). We obtain from (2.12):

$$V_{BE}(T) = V_{BE}(T) + \frac{kT}{q} \ln \frac{T}{T_r}.$$  

(2.13)

where

$$\lambda = \frac{kT}{q} \ln \frac{T}{T_r} - (n-m).$$  

(2.14)

To obtain an impression of the magnitude of the different terms of (2.13) and (2.14) we substitute $V_0 = 1166 \text{ mV}, n = 3.72$, and for example $m = 0$ ($I_C$ is constant), $T_r = 323 \text{ K}$ and $V_{BE}(T_r) = 630 \text{ mV}$; then we find for the constant:

$$V_{BE}(T) = V_{BE}(T) + \frac{kT}{q} \ln \frac{T}{T_r} = 1269.5 \text{ mV}.$$  

and for the linear term:

$$\lambda = 1.98 \text{ mV/K}.$$  

The nonlinearity in $V_{BE}(T)$, which is represented by the last term in (2.13), is plotted in Fig. 2.3 against the temperature $t$ in °C for various values of $(n-m)$.

For relatively small temperature changes $\Delta T = T - T_r < T_r$ (2.13) can be approximated well by the first three terms of its Taylor expansion, which results in

$$V_{BE}(T) = V_{BE}(T) + \frac{kT}{q} \ln \frac{T}{T_r} - (n-m).$$  

(2.15)
2.2.3 The temperature dependence of the current gain $h_{FE}$

The intrinsic carrier concentration $n_{iE}$ in the emitter is exponentially related to the bandgap voltage in the emitter in a way similar to that given by (2.5) for the intrinsic carrier concentration $n_i$ in the base. Because of the heavy emitter doping the bandgap voltage in the emitter is an amount $AV$ lower than that in the base [2.5]. The temperature dependence of the common-emitter-current gain factor $h_{FE}$ is, at moderate current levels, mainly due to this effect [2.8]. When this is taken into account it can be calculated from (2.1), (2.4) and (2.5) that

$$h_{FE} = I_C/I_B = (n_i/n_{iE})^2 = \exp\left(-\frac{qAV}{kT}\right).$$

(2.16)

If a Taylor expansion at $T = T_r$ is performed on (2.16), then it follows for $\Delta T = T - T_r << kT/(qAV)$ that:

$$h_{FE} = h_{FE}(T_r)\left(1 + \frac{qAV\Delta T}{kT}\right).$$

(2.17)

Figure 2.4 shows typical characteristics of $h_{FE}$ versus the temperature for integrated npn transistors fabricated in a conventional IC process. These curves fit well for those predicted by (2.16) for $AV = 40$ mV. For a number of transistors fabricated on different wafers, we found values for $(h_{FE})^{-1}d h_{FE}/dT$ within the range of 0.005/°C up to 0.006/°C.

2.2.4 Base-width modulation effects

The base charge $Q_B$ given by (2.3) depends on the boundaries $x_E$ and $x_C$ of the neutral base region. These boundaries in turn depend on the junction voltages. An increase in $V_{CB}$ or a decrease in $V_{BE}$ will cause an increase in the corresponding depletion-layer widths. As a consequence of this base-width modulation effect the depletion-layer charge increases and the base charge $Q_B$ decreases. In [2.9] Jespers presents a calculation of this effect for changes of $V_{CB}$. When we extend this calculations to include changes of $V_{BE}$ as well [2.10] we find for the base charge $Q_B(V_{CB}, V_{BE})$.

For small changes in $V_{BE}$ and $V_{CB}$ of the junction voltages when $V_{cb} = V_{CB} + V_{CB}$ and $V_{be} = V_{BE} + V_{BE}$ it is found from (2.18) for the base charge

$$Q_B(V_{CB}, V_{BE}) = Q_{BO}\left\{1 - \frac{V_{CB}}{V_{A,C,0}}\frac{C_{CO}(V_{CB})}{C_{CO}}dV_{CB}
+ \frac{1}{V_{A,E,0}}\frac{V_{BE}}{C_E(V_{BE})}dV_{BE}\right\},$$

(2.18)

where

$$Q_{BO} = Q_{B} \text{ for } V_{CB} = V_{BE} = 0 \text{ V},$$

$$C_{C}(V_{CB}) = \text{ the collector junction capacity for a bias voltage } V_{CB},$$

$$C_{E}(V_{BE}) = \text{ the emitter junction capacity for a bias voltage } V_{BE},$$

$$C_{CO} = C_{C} \text{ for } V_{CB} = 0 \text{ V},$$

$$C_{EO} = C_{E} \text{ for } V_{BE} = 0 \text{ V},$$

$$V_{A,C} = Q_{BO}/C_{CO} \text{ which is called the Early voltage,}$$

$$V_{A,E} = Q_{BO}/C_{EO}. $$

For small changes in $V_{BE}$ and $V_{CB}$ of the junction voltages when $V_{cb} = V_{CB} + V_{CB}$ and $V_{be} = V_{BE} + V_{BE}$ it is found from (2.19) for the base charge $Q_B(V_{cb}, V_{be})$ that:

$$Q_B(V_{cb}, V_{be}) = Q_{BO}\left(V_{cb}, V_{be}\right)\left(1 - \frac{V_{CB}}{V_{I,C}} + \frac{V_{BE}}{V_{I,E}}\right),$$

(2.19)

where

$$V_{I,C} = V_{A,C,0}\frac{C_{CO}}{C_{C}(V_{CB})},$$

$$V_{I,E} = V_{A,E,0}\frac{C_{EO}}{C_E(V_{BE})}.$$

(2.20)

Taking into account that the last two terms of the right-hand side of (2.19) are always small compared to unity, we derive from (2.1), (2.2) and (2.19) for the collector current $I_C(V_{cb}, V_{be})$ that

$$I_C(V_{cb}, V_{be}) = I_C(V_{cb}, V_{be})\left(1 + \frac{V_{CB}}{V_{I,C}} - \frac{V_{BE}}{V_{I,E}}\right).$$

(2.21)
I_(V ) characteristics. For transistor fabricated with today's technology the last terms in (2.18) and (2.19) can be neglected in all practical situations dealt with in this dissertation.

Matching of npn transistors operated at equal voltages

Due to the occurrence of mismatches the collector-current ratio \( p = \frac{I_{C1}}{I_{C2}} \) of two supposedly identical transistors operated at equal voltages deviates slightly from unity. This non-unity limits the attainable performance of many basic integrated circuits such as current mirrors and differential amplifiers. The collector-current ratio \( p \) is directly related to the input offset voltage \( V_{\text{offset}} \), where

\[
V_{\text{offset}} = k_\text{T} \ln p.
\]

In npn transistors mismatches are mainly caused by:

- Variations in the doping profile, which cause base-load mismatches as well as bulk-resistance mismatches.
- Variations in the transistor geometry as a consequence of the limited resolution of the photolithographic process. This causes emitter-area mismatches.
- Temperature gradients and variations in mechanical stress. These effects will be discussed in Chap. 3.
- Variations in leakage currents.

The matching of components gets better the more closely together the components are positioned or the larger they are. We performed a large number of measurements to obtain empirical data about the magnitude and the temperature coefficients of the collector-current mismatches.

Firstly, Straver [2.13] tested a wafer containing pairs of transistors with rectangular and circular emitters having areas of 20 \( \mu m \times 25 \mu m \), 20 \( \mu m \times 50 \mu m \) and 20 \( \mu m \times 100 \mu m \) and diameters of 25.2 \( \mu m \), 35.7 \( \mu m \) and 50.5 \( \mu m \), respectively. A microphotograph of a chip is shown in Fig. 2.7. The measurement set-up is pictured in Fig. 2.8. A total of 67 chips have been tested. By deleting the results of the 10% worst cases, the standard deviation \( \sigma(\Delta I_C/I_C) \) of the collector-current mismatch of the remaining 90% best cases is as
Fig. 2.7 Microphotograph of a chip containing transistor pairs with various geometries.

Fig. 2.8 Measurement set-up for the determination of collector-current mismatches shown in Fig. 2.9 for various emitter areas.

In addition, transistor pairs formed from cross-connected segments of a quad of transistors have been tested (Fig. 2.10).

Such pairs are often applied to reduce the influence of thermal gradients (see Chap. 3).

However, with respect to collector-current matching the observed improvement indicated in Fig. 2.9 is not better than was expected from the large total emitter area.

The collector-current ratio $p$ of a pair of transistors operated at equal bias voltages is also temperature dependent. At large current levels ($I_C > 200 \mu A$) this is mainly due to mismatches of the emitter and base bulk resistances. At low current levels ($I_C < 100 \, \text{nA}$) unequal leakage currents are responsible for this phenomenon. From measurements taken at high and low current levels, respectively, the influence of these effects at moderate current levels can be calculated. However, it has been found that at moderate current levels ($1 \, \mu A < I_C < 100 \, \mu A$) the measured temperature dependence is much larger than can be explained on the basis of the effects of bulk resistances and leakage currents.

The measurements have been performed with the set-up shown in Fig. 2.8. Typical measurement results are depicted in Fig. 2.11, which represents the relative change of the collector-current ratio $p$ with respect to its value at $t = 40^\circ C$ versus the temperature $t(\, ^\circ C)$ for six chips.

A remarkable phenomenon observed during these measurements is the occurrence of a hysteresis effect for successive heating and cooling cycles, shown in Fig. 2.12. The measurement results depicted in this figure have been found for a pair of transistors afflicted with a strong hysteresis effect. At a lower level this effect has also been found for the other devices. The hysteresis effect can possibly be explained by the occurrence of mechanical stress at the Al-SiO₂-Si interfaces [Sect. 3.4].
2.2.6 Non-ideality of the PTAT voltage

When two transistors are operated at a constant ratio $r$ ($r \neq 1$) of their emitter-current densities, the difference $AV_{BE}$ in their base-emitter voltages is proportional to the absolute temperature (PTAT). This PTAT voltage is a basic signal for bandgap references and temperature transducers. Non-idealities of this signal limit the accuracy of well-designed devices.

When we assume that $n$ and $D_B$ are equal for both transistors, then with (2.1) it is found that:

$$AV_{BE} = V_{BE1} - V_{BE2} = \frac{kT}{q} \ln \frac{I_{C1}Q_{BE1}^{E2}}{I_{C2}Q_{BE2}^{E1}}.$$  (2.23)

In this equation the numerical subscripts correspond to those of the transistors.

For transistors with identical diffusion profiles and operated at equal collector-base voltages we find from (2.3) with $x_{C1} = x_{C2} = x_C$ that

$$Q_{BE1} = 1 + \frac{x_{E2}^{E1} N_B(x) dx}{x_{E1}^{E2} N_B(x) dx} = 1.$$  (2.24)

Even though in the approximation in (2.24) the influence of base-width modulation by the (unequal) base-emitter voltages is neglected, the error made by this approximation is small when compared to the observed non-idealities in $AV_{BE}$ discussed in this section.

With (2.23) and (2.24) we have

$$AV_{BE} = \frac{kT}{q} \ln \frac{I_{C1}Q_{BE1}^{E2}}{I_{C2}Q_{BE2}^{E1}}.$$  (2.25)

Empirically, the validity of (2.25) has been tested with the measurement set-up described in Appendix A for npn transistor pairs with an emitter-area ratio of $(20 \mu m \times 60 \mu m) : (8 \times 20 \mu m \times 60 \mu m)$.

A microphotograph of these transistors is shown in Fig. 2.13. The collector currents were adjusted until they were equal and ranged from 10 $\mu A$ up to 100 $\mu A$. The deviation $\Delta T$ of $AV_{BE}(T)$ from pure proportionality has been calculated from measured values of $AV_{BE}$ and $T$ by means of the equation:

$$\Delta T = \frac{AV_{BE}(T)}{AV_{BE}(T_r)} T_r - T.$$  (2.26)

where $T_r$ denotes a reference temperature.

Typical results of these measurements are shown in Fig. 2.14 versus the temperature $t$ in °C for $T_r = 40^\circ C$. The current dependency of this deviation was found to be small. The deviation $\Delta T$ shown in Fig. 2.14 turned out to be larger than what can be
explained by the influence of bulk resistances, leakage current, base-width modulation or internal power dissipation. Possibly, mechanical stress is responsible for the deviation [Sec. 3.4].

2.3 The lateral pnp transistor

Lateral pnp transistors have a bad reputation because of their low current gain $h_{FE}$ and low cutoff frequency $f_T$. Little known is that also the $I_C(V_{BE})$ characteristics are also far from ideal. Figure 2.15 shows these characteristics as measured for a G312 breadboard component of Philips and a CA3096 array transistor of RCA, and for comparison, those for an npn transistor of the type CA3046 (RCA).

The CA3096 transistor consists of seven transistors connected in parallel. This and the difference in doping profiles explains why the two lateral pnp's behave differently. The deviations $\Delta V_{BE}$ of the $I_C(V_{BE})$ characteristics from the ideal can be described by

$$\Delta V_{BE} = V_{BE} - V_{BE,ref} = \frac{kT}{q} \ln \frac{I_C}{I_{C,ref}}, \quad (2.27)$$

where $V_{BE,ref}$ denotes the base-emitter voltage for a certain reference value $I_{C,ref}$ of the collector current (Fig. 2.15).

These non-idealities of the $I_C(V_{BE})$ characteristics are mainly due to a high-level injection mechanism and to the influence of the emitter-bulk resistance. We can model this effect by a current-dependent

emitter-series resistor $r = \frac{d(\Delta V_{BE})}{dI_C}$ whose values are shown in Fig. 2.16.

The non-idealities in the $I_C(V_{BE})$ characteristics are so large that lateral pnp transistors cannot be applied to generate accurate PTAT signals. In one respect the lateral pnp has an advantage: Due to the low emitter doping and the consequent little bandgap narrowing its current-gain factor $h_{FE}$ is less temperature-dependent than that of the npn transistors. Typical values for the temperature coefficient of $h_{FE}$ as measured for the G 312 breadboard components (Philips) are listed in Table 2.1.

The base-width modulation effect (see Section 2.2.4) can be characterized by a single parameter $V_{1,C}$ whose magnitude as measured for a G 312 is shown

![Fig. 2.15 The $I_C(V_{BE})$ characteristics of two different types of lateral pnp transistors and an npn transistor.](image-url)
### Table 2.1

<table>
<thead>
<tr>
<th>$I_E$ (μA)</th>
<th>$h_{FE}$ at 25°C</th>
<th>$\frac{\partial h_{FE}}{\partial T}$ (°C⁻¹)</th>
</tr>
</thead>
<tbody>
<tr>
<td>10</td>
<td>32</td>
<td>$4 \times 10^{-4}$</td>
</tr>
<tr>
<td>100</td>
<td>34</td>
<td>$11 \times 10^{-4}$</td>
</tr>
<tr>
<td>1</td>
<td>9</td>
<td>$16 \times 10^{-4}$</td>
</tr>
</tbody>
</table>

![Fig. 2.17 The base-width-modulation parameter $V_{IC}$ of a lateral pnp transistor versus the collector-base voltage.](image)

For a number of wafers measurements were made on the collector-current matching of pairs of pnp transistors operated at equal bias voltages. We observed considerable differences in the mismatches for different wafers. However, for the best wafers the mismatches were comparable to those of the npn transistors [2.14].

### 2.4 Resistors

#### 2.4.1 Properties of resistors

In bandgap references and temperature transducers special attention has to be paid to the quality of the resistors, which perform important tasks in the circuitry. They are applied for voltage-to-current conversion and for setting the amplifier gain. Furthermore, by trimming the resistors, the devices can be calibrated.

The main types of resistors applied for IC bandgap references and temperature transducers are:

1) **Thick-film resistors.** These resistors, which are fabricated on a ceramic substrate, are applied because they have a low temperature coefficient, excellent matching, a wide range of available sheet resistance (from 100 Ω/□ to 1 MΩ/□), and lend themselves to laser trimming. Their main disadvantage is the relatively high cost of fabricating and testing the hybrid circuits.

2) **Thin-film-on-silicon resistors.** The production cost of devices made with this technology is much less than that of hybrids because trimming and testing can be performed on the wafer. As compared to thick-film resistors a disadvantage is the limited range of sheet-resistance values (100 Ω/□ to 2 kΩ/□).

3) **Base-diffused resistors.** These resistors are fabricated with standard IC technology, which makes them very attractive in view of the low production costs. An excellent long-term stability is reported [2.15]. Their main disadvantages concern the high temperature and voltage dependence, and the limited trimming possibilities, which will be discussed in detail in this section and in Section 2.4.2, respectively. The sheet resistance cannot be freely chosen but is determined according to the IC process applied.

In order to acquire numerical data concerning their properties the various types of resistors, fabricated with standard technology, have been tested. The main measurement results are listed in Table 2.2. The matching data in this table concern the properties of nominally identical resistors located as closely to each other as possible. Note that diffused resistors match almost as well as thick- and thin-film resistors. Short diffused resistors match less well than long ones, due to the relatively larger influence of the Si-Al contact.

The temperature coefficient $\delta_T = R^{-1} \frac{dR}{dT}$ of diffused resistors strongly depends on the temperature as shown in Fig. 2.18. The temperature dependence of thin-film resistors has been found to be linear in $T$.

Diffused resistors are voltage-dependent because of the influence of depletion-layer-width modulation. Small-signal values $R_{sp}$ of base-diffused (sp) resistors have been measured as a function of the bias voltage applied to the (epitaxial) island. The voltage coefficient $\delta_V = R_{sp}^{-1} \frac{dR_{sp}}{dV_{epi-sp}}$ of the resistor for bias voltage changes has been presented in Fig. 2.19.
### Table 2.2
Summary of resistor properties measured for various types of IC resistors

<table>
<thead>
<tr>
<th>Resistor type</th>
<th>Size: length width (μm x μm)</th>
<th>Sheet resistance (Ω/μ)</th>
<th>Absolute tolerance (%)</th>
<th>Matching tolerance (%)</th>
<th>Temperature coefficient (TC) (ppm/°C)</th>
<th>Difference in TC's of matched resistors (ppm/°C)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Base diffused</td>
<td>40 x 80</td>
<td>200</td>
<td>± 20</td>
<td>± 1.2</td>
<td>See Fig.2.18</td>
<td>± 30</td>
</tr>
<tr>
<td></td>
<td>100 x 20</td>
<td>200</td>
<td>± 20</td>
<td>± 0.7</td>
<td>&quot; &quot;</td>
<td>&quot; &quot;</td>
</tr>
<tr>
<td></td>
<td>500 x 20</td>
<td>200</td>
<td>± 20</td>
<td>± 0.2</td>
<td>&quot; &quot;</td>
<td>&quot; &quot;</td>
</tr>
<tr>
<td>Base pinch</td>
<td>6000</td>
<td></td>
<td>± 50</td>
<td></td>
<td>3500</td>
<td></td>
</tr>
<tr>
<td>Nichrome on silicon</td>
<td>100 x 20</td>
<td>200</td>
<td>± 20</td>
<td>± 0.7</td>
<td>50 to 100</td>
<td>± 2</td>
</tr>
<tr>
<td></td>
<td>750 x 15</td>
<td>200</td>
<td>± 20</td>
<td>± 0.7</td>
<td>50 to 100</td>
<td>± 2</td>
</tr>
<tr>
<td></td>
<td>30 x 60</td>
<td>200</td>
<td>± 20</td>
<td>± 0.7</td>
<td>50 to 100</td>
<td>± 2</td>
</tr>
<tr>
<td>Thick film</td>
<td>1000 x 1000</td>
<td>1000</td>
<td>± 10</td>
<td>± 1.5</td>
<td>30</td>
<td>± 2</td>
</tr>
<tr>
<td>Dupont comp. 1731</td>
<td>500 x 500</td>
<td>1000</td>
<td>± 10</td>
<td>± 2.0</td>
<td>30</td>
<td>± 2</td>
</tr>
</tbody>
</table>

![Fig. 2.18](image1.png)

Fig. 2.18 The temperature coefficient \( \delta_T = K^2 \frac{dR}{dT} \) of diffused resistors versus the temperature.

versus the bias voltage \( V_{epi-sp} \) for three values of the temperature. An important item not mentioned in Table 2.2 is the long-term stability of matching properties. To measure this one needs to observe a large number of components over a long period of time, under prescribed conditions. Unfortunately, we had no opportunity of perform these measurements.

A few data about the long-term stability of resistors are reported in literature. In [2.16] for monolithic thin-film resistor networks it is mentioned that the absolute drift is less than 1000 ppm per 1000 hrs (at 125°C) and that the drift of the ratio of two resistors is less than 100 ppm per 1000 hours (at 125°C). For thick-film resistors in [2.17] an absolute drift of typically 500 ppm (2000 hrs at 70°C) is reported. With respect to diffused resistors it is argued in [2.15] that their long-term stability is better than that of thin-film resistors. However, no data have been given.

![Fig. 2.19](image2.png)

Fig. 2.19. The voltage coefficient \( \delta_V = R^2 \frac{dR}{dV} \) of base-diffused resistors as a function of the bias voltage \( V_{epi-sp} \) between the resistor and the resistor island.

2.4.2 Trimining of resistors

Calibration of bandgap references and temperature transducers is performed by trimming one or more resistors. Thick- and thin-film resistors can be laser-trimmed, and can thus achieve an absolute accuracy of \( 1^\circ \) of the resistor value. A microphotograph of a laser-trimmed thick-film resistor is shown in Fig. 2.20. Diffused resistors can be trimmed by short-circuiting E-B zener diodes (zener-zap trimming) or...
Fig. 2.20 A laser-trimmed thick-film resistor.

Fig. 2.21 Adjustment of resistors is achieved by
a) blowing-up interconnections
b) short-circuiting zener diodes.

by blowing-up interconnections (fusable-link trimming) (Fig. 2.21).

The zener-zap method is preferable because it does not have the disadvantages mentioned by Erdi [2.18] for the fusable-link method, namely that metal regrowth due to electromigration and thermal expansion may occur, that wafer test probes deteriorate quickly and that blown metal interconnections are unsightly and therefore often unacceptable.

In the zener-zap method a large current is passed through an emitter-base diode in the avalanche mode, which destroys the junction and produces a highly reliable short circuit [2.18]. Some precautions have to be taken to avoid damaging the sp-epi junction, as will be explained now.

The zener-diode bulk resistance is rather high. For emitter-base diodes with a 20 μm × 20 μm emitter size the zener series resistor amounts to about 150 Ω. This series resistor is an order of magnitude larger than that of a forward-biased diode because of the different current-flow paths (Fig. 2.22). The emitter-base zener diodes make part of (parasitic) npn transistors as shown in Fig. 2.23 for two resistors and two diodes. The collectors have been connected to a positive voltage in order to prevent the collector-base junctions of going into conduction.

Because of the high series resistor $R_B$, the zener voltage $V_{EB}$ of $Q_1$ at large current levels may exceed the breakdown voltage $V_{(BR)CER}$ of $Q_2$ (Fig. 2.23). In this case the collector-base junction of $Q_2$ will be destroyed.

This undesirable effect can be avoided by designing the zener diodes with a smaller series resistor or by short-circuiting the unused B-E junctions during zener-zap trimming. In the latter case the parasitic breakdown occurs at a voltage $V_{(BR)CES} > V_{(BR)CER}$.

2.5 The Seebeck effect

In the layout design of the precision circuits discussed in this dissertation the Seebeck effect has to be taken into account, especially when temperature gradients in the chip are to be expected. The Seebeck effect occurs when the junctions a and b of two materials (for instance
silicon and aluminum) are at different temperatures $T_1$ and $T_2$ (Fig. 2.24). In this case a voltage $V_S$ which is proportional to the temperature difference $T_1 - T_2$ is generated. Therefore one can write:

$$V_S = a(T_1 - T_2),$$

(2.28)

where $a$ denotes the Seebeck coefficient.

In Si-Al couples the value of $a$ depends on the doping concentration of the silicon and can be as large as 1.4 mV/K, which is in the same order of magnitude as the temperature coefficients of a base-emitter voltage.

A chip containing integrated thermocouples of Al and all different types of Si was designed by Kerkhoff [2.19]. For these thermocouples the Seebeck coefficients have been measured and are listed in Table 2.3.

Especially with long resistors temperature gradients can induce undesired Seebeck voltages. This will be illustrated with the structure of Fig. 2.25.

One of the Al-Si junctions of a resistor is located close to a power-dissipating component (for instance a transistor).

As will be dealt with in Section 3.3 the temperature difference $T_1 - T_2$ between the resistor contacts amounts to

$$T_1 - T_2 = \frac{P_h}{2\kappa k} \left( \frac{1}{r_1} - \frac{1}{r_2} \right),$$

(2.29)

where

- $k$ = the thermal conductivity of silicon,
- $P_h$ = the dissipated power.

If, for instance, $r_1 = 70 \mu m$, $r_2 = 1000 \mu m$ and $P_h = 10$ mW, then with $k = 140$ W/mK it is found that $T_1 - T_2 = 0.14^\circ C$.

Substitution of this value in (2.28) with $a = 950$ µV/K gives for the Seebeck voltage:

$$V_S = 0.134 \text{ mV}.$$

In a sensitive part of the circuitry such a voltage may be unacceptable. This effect can be minimized by carefully designing the layout, for instance by placing the resistor contacts close to each other.

### Table 2.3

<table>
<thead>
<tr>
<th>Type of silicon</th>
<th>Sheet resistance (Ω/□)</th>
<th>Layer thickness (µm)</th>
<th>$a$ (µV/K)</th>
</tr>
</thead>
<tbody>
<tr>
<td>sn</td>
<td>10</td>
<td>1.9</td>
<td>-210</td>
</tr>
<tr>
<td>sp</td>
<td>200</td>
<td>2.6</td>
<td>+950</td>
</tr>
<tr>
<td>dn</td>
<td>8</td>
<td>4</td>
<td>-340</td>
</tr>
<tr>
<td>dp</td>
<td>5</td>
<td>12</td>
<td>+610</td>
</tr>
<tr>
<td>n-epi</td>
<td>1250</td>
<td>8</td>
<td>-1420</td>
</tr>
</tbody>
</table>

$s$ = shallow, $d$ = deep, $n$ = n-type silicon, $p$ = p-type silicon.
3. INTERACTION WITH REGARD TO POWER DISSIPATION AND MECHANICAL STRESS

3.1. Introduction

Power dissipation in the chip gives rise to a temperature increase in the chip. Such a temperature rise contributes directly to the absolute error of a temperature sensor. Bandgap references are insensitive to uniform temperature changes in the chip. However, these devices are, like temperature transducers, very sensitive to thermal gradients, which can easily be caused by internal power dissipation.

This self-heating effect can be characterized by a thermal model. A simple model is shown in Fig. 3.1, in which $\theta_{jB}$ is the thermal resistance between the emitter junction and the bulk of the chip, $\theta_{BC}$ is the thermal resistance between the bulk of the chip and the case, $\theta_{CA}$ is the thermal resistance between the case and its surroundings, $C_{CH}$ is the thermal capacity of the chip and $C_{C}$ is the thermal capacity of the case. The power source is represented by the current source $P$ while the ambient temperature with respect to a certain reference is represented by the voltage source $T_A$.

The validity of this model will be discussed in Section 3.2. Although it will appear to be useful for qualitative analysis of self-heating effects, the model is not suitable for predicting temperature gradients in the chip. In Section 3.3 a simple manner of calculating these gradients from empirical and theoretical data is shown.

Not only thermal-electric interaction but also mechanical-electric interaction can cause deviations in the behavior of transistors. The magnitude of mechanical stress and its influence on base-emitter voltages will be discussed in Section 3.4.

3.2 The effect of power dissipation; thermal response time

We can get a good idea of the thermal behavior of IC's by measuring the thermal step response of chips mounted in different types of packages. The temperature changes in the chip can be determined from the values of the base-emitter voltages. A typical result, observed for a CA 3046 npn-transistor array in a DIL package is shown in Fig. 3.2(a).

The layout of the array is given schematically in Fig. 3.2(b). The upper solid curve of Fig. 3.2(a) represents the relative temperature change $\Delta T_3$ of

![Fig. 3.1](image_url) Simple thermal model.

![Fig. 3.2](image_url) (a) Temperature rise $\Delta T$ of the base-emitter junctions as caused by a power step $\Delta P_j$ in the collector dissipation of $Q_j$. (b) Layout of the CA 3046 array.
the emitter junction of Q₃ resulting from a power step ΔP₃ in the collector dissipation of that transistor. Roughly, the temperature rise can be broken down into two components, viz., a fast one and a slow one. The fast effect is concerned with the transistor geometry. The slow effect is dependent on the thermal properties of the package. The lower curve of Fig. 3.2(a) represents the temperature change ΔT₄ of Q₄ caused by a power step ΔP₃ in the dissipation of Q₃. Note that this curve parallels the upper curve for time >1 ms and represents the slow effect mentioned above. In plastic packages, between chip and ambient the heat is mainly transported by thermal conductivity of the metal leads. The broken curves in Fig. 3.2(a) represent measurement results for a similar chip mounted in an outwardly identical DIL package but having a different metal construction internally. A considerable difference in thermal impedances is found. For times longer than 10 s the temperature rise depends on the thermal resistance between the package and its surroundings, which is influenced by the presence of assembling material, cooling bodies, air velocity, etc. The time response shown in Fig. 3.2 deviates considerably from the exponential behavior to be expected from the model of Fig. 3.1. However, for our applications it is not interesting to improve the accuracy of the model at the price of larger complexity. From Section 3.1 it follows that we mainly need the model for two purposes:

a) To predict the rise of the chip temperature in the stationary case. For this purpose we can omit the capacitors in the model of Fig. 3.1

b) To predict the time response of fast temperature transducers. Fast temperature transducers are made by mounting the chips in special packages with low thermal resistances and capacitances. In this particular case useful results are obtained with the simple model of Fig. 3.1.

<table>
<thead>
<tr>
<th>Type of package</th>
<th>Thermal resistance (°C/W)</th>
<th>Table 3.1 Thermal resistances for various types of packages in still air.</th>
</tr>
</thead>
<tbody>
<tr>
<td>RCA transistor array</td>
<td>CA 3046</td>
<td>CA 3045</td>
</tr>
<tr>
<td>14-pens DIL (plastic)</td>
<td>62</td>
<td>62</td>
</tr>
<tr>
<td>14-pens DIL (ceramic)</td>
<td>72</td>
<td>50</td>
</tr>
<tr>
<td>16-pens DIL (plastic)</td>
<td>18</td>
<td>14</td>
</tr>
<tr>
<td>0₂BA</td>
<td>152</td>
<td>126</td>
</tr>
</tbody>
</table>

A detailed study of the thermal properties of diebond, bonding wires and the construction of the package has been performed by F.A. Dietz [3.1]. A summary of his measurement results for different types of packages is given in Table 3.1. To study the limitations to the response time of fast temperature sensors Duyverman [3.2] investigated the thermal properties of unpackaged chips placed in an air stream with a specified velocity. For these measurements a temperature-sensor chip which measures 2 mm x 1 mm x 0.2 mm has been glued on a grooved printed-circuit board (Fig. 3.3). For air velocities > 1 m/s the thermal response to a step change in the power dissipation on the chip agreed well with those predicted by the simple model of Fig. 3.1 with Cₐ = 0 Ws/°C and 0₂BC = 0°C/W. The observed magnitudes of the thermal conductance h₂BA between the chip and ambient and the thermal time constant τ = 0₂BA x CₐCH are shown in Fig. 3.4 versus

![Fig. 3.4](image-url)
the air velocity \( v \). The small response time means that the IC temperature sensor is applicable for very fast temperature measurements.

The time constant is only slightly dependent on the chip area because with increasing area the thermal conductance as well as the thermal capacitance increases. For a low response time the chip is etched as thin as is permissible from the point of view of mechanical strength. Generally, temperature-sensor chips are mounted in special types of packages to obtain mechanical and chemical protection together with good thermal properties.

In [3.3] Analog Devices specifies the thermal properties of temperature sensors in two types of packages in various thermal environments. The time constants of these devices are about 30 times larger than those of the unpackaged chips so that further improvement of these sensors with respect to response time seems to be possible. An interesting device can be obtained by gluing a piece of silicon on the sensor chip (Fig. 3.5) in order to obtain some mechanical protection of the chip. The thermal time constant of such a device is only 2 times that of the single unprotected chip.

### 3.3 Thermal gradients in the chip

The accuracy of bandgap references as well as of temperature transducers can easily be spoiled by temperature gradients over the chip. Temperature gradients caused by internal power dissipation have been experimentally investigated by measuring the temperature difference of two transistors caused by a power step \( \Delta P \) in a third transistor. An example of such a step response has been shown in Fig. 3.6 for a CA 3046 transistor array.

It has been found that these thermal gradients are hardly influenced by the thermal properties of the package and surroundings. The magnitude of this effect can be calculated in a relatively simple way. For the case in which the heat source is half spherical and located at the surface of a half space (Fig. 3.7) and the heat is exclusively transported via conduction in this half space, a thermal differential resistance \( \theta_d(r_1, r_2) \) is found for two points located at distances \( r_1 \) and \( r_2 \) from the center of the source for which it holds that in the stationary case [3.4]:

\[
\theta_d(r_1, r_2) = \frac{\Delta(T(r_1) - T(r_2))}{\Delta P} = \frac{1}{2\pi k'} \left( \frac{1}{r_1} - \frac{1}{r_2} \right),
\]

where \( k' = \) the thermal conductivity of silicon \((\approx 140 \text{ W/mK at } T = 300 \text{ K})\).

For transistor arrays the thermal differential resistances calculated with (3.1) differ less than 25% from those found empirically [3.5]. When the layout is designed Eq. 3.1 can be used to make a rough estimation of the effect of dissipating components on the most sensitive circuit parts. When this effect is too large, the layout can be optimized by choosing a more symmetrical position.
of the dissipating component with respect to the sensitive circuit part or by increasing their mutual distance.

3.4 Mechanical stress

As discussed in Section 2.2.5 for the collector-current ratio \( p \) of npn transistors operated at equal voltages a hysteresis effect has been observed for successive heating and cooling cycles. Such a hysteresis effect has also been found for the output signal of bandgap references and temperature transducers and may be due to the occurrence of mechanical stress. Mechanical stress changes the base-emitter voltage of a transistor [3.6]. In IC's this stress can be caused by unequal thermal expansion of silicon, silicon dioxide, aluminum interconnections or mounting base. The thermal-expansion coefficients at 25°C are listed in Table 3.2

Observe the strong deviation between the expansion coefficients of Al and Si. The temperature dependence of stress in aluminum films on oxidized silicon substrates has been examined by Sinha and Sheng [3.9]. During successive heating and cooling cycles they observed a hysteresis effect for the mechanical stress, which they attributed to diffusion creep, grain growth and dislocation slip in the aluminum. In the temperature range \( 0°C < t < 500°C \) the stress amounts to from \( 0.5 \times 10^8 \) to \( 10^9 \) N/m².

In our laboratory the influence of stress on the base-emitter voltage of transistors has been investigated by P.C. Schmale by means of the following measurement method. A strip of silicon wafer has been mounted in a test set-up as indicated in Fig. 3.8. In this set-up the surface stress of the tested chip amounted to \( 0.23 \times 10^8 \) N/m². This stress caused a change in \( V_{BE} \) which came to \( +0.5 \) mV.

This value is so large that it can be concluded that mechanical stress may be the dominating factor limiting the accuracy of well-designed bandgap references and temperature transducers. Further investigations are necessary to find a solution to this problem. Mechanical stress induced by packaging of the chip can also cause serious problems with respect to the accuracy of high-precision analog IC's [3.11], [3.12]. However,

<table>
<thead>
<tr>
<th>Material</th>
<th>( \alpha ) at 25°C ( \times 10^{-6} ) K(^{-1} )</th>
<th>reference</th>
</tr>
</thead>
<tbody>
<tr>
<td>Al</td>
<td>23</td>
<td>[3.7]</td>
</tr>
<tr>
<td>Thermal SiO(_2)</td>
<td>0.6</td>
<td>[3.7]</td>
</tr>
<tr>
<td>&lt; 100 &gt; Si</td>
<td>2.5</td>
<td>[3.7]</td>
</tr>
<tr>
<td>Kovar</td>
<td>4.7</td>
<td>[3.8]</td>
</tr>
</tbody>
</table>

Fig. 3.8 Test set-up to determine the influence of surface stress on the base-emitter voltage of bipolar transistors.
4. Bandgap References

4.1 Introduction

In bandgap references the reference voltage is obtained by compensating the base-emitter voltage of a bipolar transistor for its temperature dependence. There are many ways to achieve this compensation. Because the temperature dependence of base-emitter voltages is rather high the compensation technique chosen has to be very accurate and reliable (Chap. 1) over a temperature range as wide as possible. Generally, calibration of the voltage references is necessary. As economic considerations play also a role in choosing the compensation technique, the simplicity of calibration has to be taken into account.

With respect to these requirements very good results are met with the compensation technique commonly applied in bandgap references: A correction voltage \( V_c(T) \) is added to \( V_{BE}(T) \) to compensate for at least the first-order temperature dependence of \( V_{BE}(T) \). This correction voltage is obtained by amplifying the differences \( \Delta V_{BE} = (kT/q)\ln p \) of the base-emitter voltages of two transistors operated at unequal emitter-current densities with ratio \( p \).

In this way an output voltage \( V_{ref} \) is obtained for which it holds that:

\[
V_{ref} = V_{BE}(T) + V_c(T) = V_{BE}(T) + A\Delta V_{BE}(T),
\]

where \( A \) denotes an amplification factor. A simple implementation is shown in Fig. 4.1. The action of the current mirror \((Q_3, Q_4)\) ensures that the ratio of the emitter currents of \( Q_1 \) and \( Q_2 \) remains constant. Neglecting the influence of base currents and base-width modulation, for the voltage \( \Delta V_{BE} \) across \( R_2 \) we find

\[
\Delta V_{BE} = (kT/q)\ln (nr),
\]

where \( r \) and \( n \) denote the saturation-current ratios \( I_{S2}/I_{S1} \) and \( I_{S3}/I_{S4} \) of \((Q_2, Q_1)\) and \((Q_3, Q_4)\), respectively. For the collector current \( I_{C3} \) of \( Q_3 \) we find

\[
I_{C3} = nI_{C2} = n \frac{kT}{qR_2} \ln (nr).
\]

For the output voltage \( V_{ref} \) it holds that:

\[
V_{ref} = V_{BE1} + V_c = V_{BE1} + n \frac{kT}{qR_2} \ln (nr).
\]

We assume that the resistors are temperature independent so that \( I_{BE1} \) is PTAT. The base-emitter voltage \( V_{BE1} \) of \( Q_1 \), biased with a PTAT current, is given by Eq. 2.13 with \( m = 1 \):

\[
V_{BE1} = V_0 + (n-1) \frac{kT}{q} - \lambda T + (n-1) \frac{k}{q} (T-T_r) \ln \frac{T}{T_r}.
\]

The nonlinear portion of the temperature dependence of \( V_{BE1} \) is not compensated for and causes the typical temperature characteristic of the output voltage shown in Fig. 4.2. Note that the top of the curve occurs for \( T = T_r \).

The accuracy of the simple circuit of Fig. 4.1 is bad because of the influences of the base currents, base-width modulation and load current. The uncertainty about the magnitudes and temperature
coefficients of these effects makes it impossible to calibrate the circuit in a simple way. Furthermore, the output voltage will depend on the load current and supply voltage.

Another problem is that circuits of the type of Fig. 4.1 have a stable state at zero current flow even when the power supply is nonzero. A separate start-up circuit is required to prevent the circuit from remaining in this undesired state. These problems have been overcome by the circuit presented in [4.1] - [4.3]. One of these circuits will be discussed in Section 4.2.

The main disadvantage of the circuit of the type of Fig. 4.1 concerns the use of lateral pnp transistors because of their poor performance, viz. a low current gain, large bulk resistances and the occurrence of high-level injection at a comparatively low current level (Section 2.3). This problem has been solved by designing the high-performance all-npn configuration [4.4] described in Section 4.3.

Further improvement of bandgap references is obtained by compensation of the thermal nonlinearity shown in Fig. 4.2, [4.5], [4.6]. For these circuits extremely high performance is reported. One of these circuits will be discussed later in Section 4.4.

With the best circuits the precision of bandgap references is limited by imperfection in the components used to generate and process the basic signals. These limitations will be discussed in Section 4.5.

4.2 An accurate, straightforward implementation of the basic circuit

4.2.1 Design considerations

In an initial attempt to realize an accurate bandgap reference the basic configuration of Fig. 4.1 has been improved. The main imperfections in the basic circuits are due to:

a) The error in the current gain \( n \) of the pnp current mirror which depends on temperature as well as on the supply voltage.

b) The inequality and supply-voltage dependence of the collector currents of \( Q_1 \) and \( Q_2 \), which influence the basic PTAT current.

c) The influence of the temperature dependent base currents of \( Q_1 \) and \( Q_2 \).

d) The influence of load currents.

As will be shown, all these errors can be minimized by careful design and by extending the basic circuit with additional elements. First of all, we will pay special attention to the behavior of pnp current mirrors, because these are basic circuits frequently used in many of the circuits dealt with in this dissertation. The sensitivity of the output voltage of the circuit of Fig. 4.1 to changes in the current transfer factor \( n \) is rather high. For the sensitivity factor \( S_n^{V_{\text{ref}}} \) we find from (4.4) that
With the approximation in (4.7) we neglect the influence of the change in \( V_{BE1} \) caused by a change in \( n \), which results in an insignificant error when \( V_c \gg \frac{kT}{q} \).

For \( R_1/R_2 \) we find from (4.4) that:

\[
\frac{R_1}{R_2} = \frac{V_c}{kT \ln(nr)}
\]

(4.8)

Substituting (4.8) in (4.7) yields:

\[
S_{n, \text{ref}} = \frac{V_c}{V_{\text{ref}}} \left( 1 + \frac{1}{\ln(nr)} \right)
\]

(4.9)

Common values for \( V_{BE1} \) at room temperature amount to \( V_{BE1} = \frac{1}{2} V_{\text{ref}} \) so that

\[
S_{n, \text{ref}} = \frac{1}{2} \left( 1 + \frac{1}{\ln(nr)} \right)
\]

(4.10)

Substituting, for example, \( nr = 5 \) results in

\[
S_{n, \text{ref}} = 0.81,
\]

which means that a 1% change in \( nr \) causes a 0.81% change in \( V_{\text{ref}} \).

### 4.2.2. Temperature and voltage dependence of the current transfer factor of pnp current mirrors

Figure 4.3(a) shows a simple pnp current mirror with a multi-emitter output transistor. The current transfer factor

\[
n = \frac{p}{1 + (p+1)/h_{FE}}
\]

(4.11)

where \( p \) denotes the effective emitter-area ratio of \( Q_2 \) and \( Q_1 \), and \( h_{FE} \) is the common-emitter current-gain factor of \( Q_1 \) and \( Q_2 \).

The temperature dependence of \( h_{FE} \) creates a relative temperature coefficient of the transfer factor amounting to:

\[
\frac{\delta n}{n \delta T} = \frac{\delta h_{FE}}{h_{FE} \delta T} = \frac{p + 1}{h_{FE}(h_{FE} + p + 1) \delta T}
\]

(4.12)

A large error can be caused by inequalities in the effects of high-level injection and particularly bulk resistances. As shown in Section 2.3, these effects cause considerable errors in the \( I_C(V_{BE}) \) characteristics of the transistors. This problem is overcome by applying well-matched emitter resistors. Application of these resistors is also advantageous in reducing the influence of base-width modulation. To be effective these resistors \( R \) have to be so large that \( R \gg r_e \) holds, where \( r_e = (kT)/(qI) \).

Due to the limited accuracy of the fabrication...
process there will be some mismatch between the emitter-area ratio and the emitter-series resistor ratio, which causes an additional temperature dependence of the transfer factor. To demonstrate this, let us assume that $Q_1$ and $Q_2$ are nominally equal transistors with a deviation $\Delta p$ from unity in their saturation-current ratio $I_{S1}/I_{S2}$ (Fig. 4.4(a)), or a difference $\Delta R$ in the emitter series-resistor values (Fig. 4.4(b)), respectively. Then, when $\Delta p/p << 1$ and $\Delta R/R << 1$ we can calculate for the transfer factor $n$ of the circuit of Fig. 4.4(a) that

$$n = 1 - \frac{R_e}{R + r_e} \Delta p,$$

and for the circuit of Fig. 4.4(b) that

$$n = 1 - \frac{\Delta R}{R + r_e},$$

where $r_e$ denotes $(kT)/(qI)$. The temperature dependence of $r_e$ causes a temperature dependence of the transfer factor $n$, where for the circuit of Fig. 4.4(a) it holds that

$$\frac{1}{n} \frac{\delta n}{\delta T} = -\frac{x}{T(1 + x)^2} \Delta p,$$

while for the circuit of Fig. 4.4(b) it is found that

$$\frac{1}{n} \frac{\delta n}{\delta T} = -\frac{x}{T(1 + x)^2} \frac{\Delta R}{R},$$

where $x = R/r_e$. Particular attention has to be paid to the term $x/(1 + x)^2$, which has been depicted in Fig. 4.5.

When the current mirror is operated at low current levels, then for practical reasons it is often impossible to choose $R >> r_e$. Insertion of resistors $R = r_e (x \neq 1)$ would maximize the effect of the term $x/(1 + x)^2$. In this case it would be preferable to omit the emitter-series resistors completely. As an example we shall substitute $\Delta p = 1.4\%$ and $\Delta R/R = 1\%$ (typical data for our standard IC process). These mismatches are uncorrelated and we can calculate their effective influence on the temperature coefficient of $n$ by taking the square root of the summed squares of their separate contributions. In this way we find with (4.16(a) and (b)) for $x = 1$ and $T = 300$ K that

$$\frac{1}{n} \frac{\delta n}{\delta T} = 14 \text{ ppm/k.}$$

The transfer factor of the simple current mirrors of Fig. 4.3 is very sensitive to changes in the output voltage. Using (2.21) we find

$$\frac{1}{n} \frac{dV_{CB2}}{V_{CB2}} = \frac{1}{I_{out}} \frac{\delta I_{out}}{I_{out}} = \frac{1}{I_{out}} \frac{\delta V}{V} \frac{1}{I_{C}},$$

where $V_{I,C}$ ranges from 35 to 60 V (Sect. 2.3). A well-known remedy for reducing this voltage sensitivity is either to apply emitter-series resistors ($R >> r_e$) or to use cascode transistors. In the latter case the voltage dependence is reduced by a factor of $h_{FE}/2 [4.7]$. A problem with cascoded current mirrors is the fluctuation of the output current by low-frequency (1/f) noise which can be attributed to fluctuations in the base currents. The cascoding transistors contribute even more to this noise than the mirror transistors $Q_1$ and $Q_2$. Using high-gain combinations for the cascoding transistor (Fig. 4.6) considerably reduces their base currents and thus their 1/f noise contribution. A much better method for reducing the influence of 1/f noise in a bandgap reference is to avoid the use of lateral pnp transistors. Circuits using this approach will be discussed in Section 4.3.
We will not deal further with the 1/f noise aspects because of their minor importance for these improved circuits.

### 4.2.3 Practical realization

A circuit with good performance with respect to the temperature dependence and low-frequency noise is shown in Fig. 4.7. In comparison to the circuit of Fig. 4.1 the permissible output current is increased by a factor $h_{FE21}$, where $h_{FE21}$ is the common-emitter forward-current gain of the Darlington pair $Q_{21}^{-} Q_{11}$. The output voltage is two times as large.

The influence of the base currents of $Q_1$ and $Q_2$ is compensated for by the base current of $Q_{16}$, which is supplied via the current mirror $Q_{18}^{-} Q_{19}$ to the bases of $Q_1$ and $Q_2$. The effect of collector-base voltage variations of $Q_2$ is combated by cascoding $Q_2$ with the Darlington pair $Q_{10}^{-} Q_{11}$. For the current mirror the circuit of Fig. 4.6 has been used.

The transistors $Q_{22}$, $Q_{23}$ and $Q_{24}$ form the start-up circuit. When the power supply is switched on, a small current flows via the 100 K resistor into the base of $Q_{22}$. The collector current of $Q_{22}$ provides the start-up current for the reference circuit which drives itself toward the desired stable state. In this state $Q_{23}$ is switched 'on' by $Q_{24}$, causing $Q_{22}$ to be switched 'off' so that the start-up circuit does not interfere with the reference circuit once the desired operating point has been reached.

A breadboard model of the circuit has been tested.

#### Table 4.1

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Measured Values</th>
<th>Units</th>
</tr>
</thead>
<tbody>
<tr>
<td>Output-voltage change 0°C to 85°C</td>
<td>150</td>
<td>ppm Vout</td>
</tr>
<tr>
<td>Line regulation</td>
<td>500</td>
<td>ppm Vout/V*</td>
</tr>
<tr>
<td>Output impedance</td>
<td>3</td>
<td>Ω</td>
</tr>
<tr>
<td>Supply current</td>
<td>700</td>
<td>μA</td>
</tr>
<tr>
<td>Supply-voltage range</td>
<td>5.5 - 15</td>
<td>V</td>
</tr>
</tbody>
</table>

The temperature dependence of the output voltage is very low. Not only is the first-order temperature dependence of $V_{BE}(T)$ compensated for, but also the nonlinearity of $V_{BE}(T)$ with $T$. However, there is little reason to be satisfied with this compensation, which has been found to be mainly due to a coinciding of nonidealities in the current-mirror transfer and changes in the collector-base voltage of $Q_{10}$. Due to the large spread in these effects this higher-order
compensation is unreliable and consequently undesirable. Because of this disadvantage we did not fabricate IC's of the type dealt with in this section, but instead developed circuits with intrinsic higher precision. These will be discussed in the coming sections.

4.3 An all-npn configuration

4.3.1 Design considerations and principle

The main drawback of the circuit discussed in the previous section is the great influence of the pnp current mirror on the circuit performance. Imperfections in this current mirror, due to the poor quality of lateral pnp transistors, limit the precision of the bandgap references. Higher accuracy is achieved when the pnp current mirror is replaced by two resistors [4.1], [4.8] (Fig. 4.8). By sensing the difference between the voltages across the resistors and by applying negative feedback to the circuit, a constant collector-current ratio for Q1 and Q2 is achieved. A disadvantage of this circuit is that it requires additional, perfectly matched, resistors and a high-gain feedback amplifier.

In this section an alternative, basically all-npn configuration is described which features a generated PTAT voltage that is highly insensitive with respect to the bias-current levels and ratios. With this circuit a quite modest number of components is needed. Furthermore a number of secondary advantages contribute significantly to the high precision of the circuit.

The basic principle is shown in Fig. 4.9(a). Characteristic of this circuit are the cross-connected bases of Q1 and Q2, the unequal emitter areas with ratios r1 and r2 for the transistor pairs (Q2, Q1) and (Q3, Q4) and the emitter resistances R2 and R3. We can now write the following equation:

\[ V_{BE1} + V_{BE4} - V_{BE2} = V_{BE3} = I_{E2}R_2 + I_{E3}R_3, \]

(4.18)

where the numerical subscripts correspond to those of the components. Initially neglecting the base currents and the base-widening effect, we find, with \( V_{BE} = (kT/q) \ln \left( I_C/I_O \right) \), that

\[ V_{BE1} + V_{BE4} - V_{BE2} = \frac{kT}{q} \ln \left( I_{C1}/I_O \right) = I_{E2}R_2 + I_{E3}R_3, \]

(4.19)

where the numerical subscripts correspond to those of the components. Initially neglecting the base currents and the base-widening effect, we find, with \( V_{BE} = (kT/q) \ln \left( I_C/I_O \right) \), that

From (4.20) some important conclusions may be drawn:

- The output current \( I_O \) is PTAT, assuming that \( R \) is temperature-independent.
- The output current \( I_O \) is independent of the bias current \( I_{bias} \). This remarkable feature is due to the cross connection of the bases of Q1 and Q2.
- The level of the PTAT voltage \( V_{PTAT} \) is high thanks to the multiplication of \( r_1 \) and \( r_2 \). For instance, let \( r_1 = r_2 = 4 \); then, for \( T = 300 K \) this voltage amounts to 72 mV. In the configuration of Fig. 4.1 with \( n = 1 \), an impractical emitter-area ratio \( r = 16 \) is needed to generate this voltage. A large value for \( n \) is disadvantageous in view of the larger number of lateral PNP's needed. Thanks to the high level of the PTAT voltage, the performance of the PTAT current source only
little depends on that of the remaining part of the bandgap reference circuit.

If the influence of the non-zero base currents is taken into account it is found that

\[ \frac{kT}{q} \ln \left( \frac{I_{E2} - I_{B1} - I_{B4}}{I_{E1} - I_{B1}} \right) \left( \frac{I_{E1} - I_{B1}}{I_{E2} - I_{B2}} \right) I_{B1} R_1 \]

\[ = I_{E1} R_2 + (I_{E1} - I_{B1} + I_{B2}) R_3. \]  

(4.21)

When \( I_{B1} = I_{B2} = I_{B3} = I_{B4} \), Eqs. (4.19) and (4.21) are identical. This means that in this case full compensation of the base currents is obtained. If there is a difference in the left-hand and right-hand branch currents \((I_{C3}, I_{C1})\) and \((I_{C4}, I_{C2})\), respectively, then the base-current compensation is not complete. This effect causes the output current \( I_o \) to be weakly dependent on the bias current. This is shown in Fig. 4.10, which represents the normalized output current versus the normalized bias current, where \( I_{opt} \) denotes the optimum bias-current level \((kT/2qR) \ln (r_1r_2)\), for which all collector currents are equal, and \( B \) denotes the common-emitter current gain. From this curve it is found that the sensitivity of the output current to changes in the bias current amounts to only 2.5%.

Another interesting circuit is obtained if \( R_3 = 0 \) (Fig. 4.9(b)). Equation (4.19) shows that in this case the right-hand branch current \( I_{E1} \) and therefore also \( I_{C4} \) are PTAT.

If a current sink is needed instead of a current source, the circuit of Fig. 4.9(b) may be preferable. An advantage of this circuit is that only one resistor is required. However, for many applications the circuit of Fig. 4.9(a) will be preferred, because with this circuit the total current is converted into a PTAT current, thus saving power consumption. Moreover, we found that with this circuit implementation of bandgap references turned out to be simpler.

To achieve full base-current compensation, the collector-current ratio \( I_{C3}/I_{C4} \) has to approximate unity. This is accomplished by a pnp current mirror (Fig. 4.11). The output current \( I_o \) only weakly depends on the current-mirror gain.

So, in contrast to the circuit of Fig. 4.1, in this design the bad performance of the lateral pnp transistors does not spoil the performance of the complete circuit.
to changes in the current-mirror gain is increased. However, for well-matched resistors \( \Delta p < 2\% \) no serious problems are caused by this effect. For a bandgap-voltage reference, analogous to the circuit of Fig. 4.1, a series resistor \( Z_R \) is connected to the circuit (Fig. 4.11). Its value is twice as high, because double voltage is generated.

### 4.3.2 Realization

The current \( I_0 \) in the circuit of Fig. 4.11 is rather sensitive to supply-voltage changes. An increase in the supply voltage \( V^+ \) and consequently in \( V_{CB4} \) causes a decrease in the base-emitter voltage of \( Q_4 \), for which it is found from (2.1) and (2.21) that

\[
\frac{\delta V_{BE4}}{\delta V_{CB4}} \frac{1}{I_c} = -\frac{kT}{qV_{I,C}}.
\]

When we extend equations (4.18) - (4.20) to include this effect of base-width modulation we find that

\[
\frac{1}{I_0} \frac{\delta I_0}{\delta V^+} = \frac{1}{I_0} \frac{\delta I_0}{\delta V_{CB4}} = -\frac{1}{V_{I,C} \ln(r_1 r_2)}.
\]

The base-width-modulation parameter \( V_{I,C} \) amounts to about 100 V (Sect. 2.2.4). It is a remarkable feature of this circuit that \( I_0 \) decreases with increasing supply voltage. By cascoding \( Q_3 \) and \( Q_4 \) the supply-voltage dependence of the output signal is reduced to an insignificant level.

Figure 4.12 shows a complete schematic of a bandgap-reference circuit incorporating this improvement.

The string of diodes ensures the start-up of the circuit when the power supply is switched on. Such a start-up method can only be used for a limited range of the supply voltage. Compared to the start-up circuit discussed in Section 4.2.3, this circuit has the advantage of being very simple and having minimal power consumption in the steady state of the bandgap reference. If a wide range of the supply voltage is desired, a different type of start-up circuit, for instance one of those described in [4.2], [4.3] and [4.5], has to be applied.

A MOS capacitor of 30 pF prevents parasitic oscillations, which can easily occur due to the internal feedback in the circuit. Empirically, we determined the optimum value and the best location for this capacitor and found them to be dependent on the bias-current level and load conditions. For a circuit with a lower bias current than that of

---

**Fig. 4.12** Complete schematic of an accurate all-npn bandgap reference.

**Fig. 4.13** Microphotograph of the chip.
gradients and strain is cancelled in the first-order by the crosscoupled configuration of the transistors Q1 - Q4 in quad layout [4.9]. Devices have been made by using conventional IC technology and discrete metal-film resistors. Figure 4.13 shows a microphotograph of the chip.

### 4.3.3 Performance

Bandgap references of the type of Fig. 4.12 have been tested between -20°C and 100°C. Figure 4.14 shows the normalized output voltage versus the temperature. This measurement result fits very well with that predicted by Eq. (4.6) for $V_{g0} = 1172$ mV and $\eta = 3.5$. Clearly, the temperature characteristic of this bandgap reference is dominated by the behavior of the npn transistors, while the remaining circuitry is so good that it hardly contributes to the imperfections. Further specifications of this bandgap reference are listed in Table 4.2.

**Table 4.2**

Performance of the bandgap reference of Fig. 4.12.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Measured Values</th>
<th>Units</th>
</tr>
</thead>
<tbody>
<tr>
<td>Output-voltage change</td>
<td>see fig.4.14</td>
<td></td>
</tr>
<tr>
<td>0°C to 85°C</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Line regulation</td>
<td>50</td>
<td>ppm $V_{out}/V^+$</td>
</tr>
<tr>
<td>Output impedance</td>
<td>2.9</td>
<td>Ω</td>
</tr>
<tr>
<td>Supply current</td>
<td>200</td>
<td>µA</td>
</tr>
<tr>
<td>Supply-voltage range</td>
<td>4.5 - 6</td>
<td>V</td>
</tr>
</tbody>
</table>

The precision of this bandgap-reference circuit is so high that it is worthwhile to further improve accuracy by compensating the higher-order temperature dependence of $V_{BE}(T)$. This will be described in the next section.

### 4.4 Curvature-corrected bandgap references

#### 4.4.1 Design considerations and principle

For a number of years bandgap-reference designers have been trying to cancel the thermal nonlinearity of the base-emitter voltage to obtain a 'curvature-corrected' reference voltage with a lower temperature dependence. Sometimes this correction is accomplished by connecting a pinch resistor in series with the base of a transistor of the PTAT circuit [4.8]. However, there is large spread in the magnitude and temperature coefficient of these resistors. Furthermore, the effect of such a base resistor is proportional to $V_{BE}^{-1}$, a quantity which is also afflicted with large spread. Due to these effects first-order compensation of the temperature dependence of $V_{BE}$ is troublesome. Better results can be expected when an accurate resistor with a well-known temperature coefficient is connected in series with the emitter of a transistor in the PTAT circuit. In this way the $h_{FE}$ dependence is eliminated. This method will be discussed in section 4.4.2.

A quite different and more accurate method for curvature correction of bandgap references has been presented by Palmer and Dobkin [4.5]. In their design a nonlinearity in the correction voltage is effected which approximately compensates for the $V_{BE}$ nonlinearity. An alternative method with very high accuracy is presented in Section 4.4.3.

#### 4.4.2 Curvature correction by applying temperature-dependent resistors

The principle of this method is shown in Fig. 4.15. The resistors $R_c$ with a high positive temperature coefficient are inserted in series with the emitters of $Q_1$ and $Q_4$. The resistors $R$ and $2R_c$ are temperature independent. In this way a small thermal nonlinearity of the correction voltage $V_c(T)$ is effected which can cancel the second-order tempera-
Curvature correction of the reference voltage by applying temperature-dependent resistors $R_c$.

![Fig. 4.15](image)

The curvature-corrected reference voltage and its components for the circuit of Fig. 4.15 plotted versus the temperature. The broken lines represent the tangents to the curves at $T = T_r$.

The curvature is exaggerated in order to indicate the characteristic points clearly.

Diffused resistors for $R_c$ are an obvious choice, because of their large temperature coefficient. However, these resistors are non-linearly related to temperature, while their absolute tolerance is rather high (see Section 2.4.1). An uncommon but rather interesting alternative is the use of aluminum resistors [3.10]. Aluminum resistors have a very large temperature coefficient $\alpha$ of approximately $4 \times 10^{-3}/^\circ C$ at $t = 40^\circ C$, while their change with temperature is linear over the entire range of interest to us. Inserting these resistors causes the correction voltage $V_{c}(T)$ to deviate from pure PTAT behavior and shows a considerable second-order temperature dependence (Fig. 4.16). Straightforward calculations show that the second-order temperature dependence of $V_{c}$ is just compensated by that of $V_{BE}$ when:

$$V_{c0} = -2(V_{ref} - V_{BE}(T_r)) R_c / R \alpha T_r \quad (4.28)$$

Substituting the values given in the example above and $T_r = 300 K$ in (4.28) yields $V_{c0} = -73.44 mV$.

The circuit of Fig. 4.12 modified in the way indicated in Fig. 4.15 has been tested.

For the resistors $2R$ and $R$ thin-film resistors with a low temperature coefficient were applied, while for $R_c$ aluminum resistors were used. The chip area $C$ for the two Al resistors amounted to 0.8mm$^2$. The observed temperature dependence of the output voltage ($\approx 2.4 V$) has been plotted in Fig. 4.17 versus the temperature. In comparison with conventional bandgap references the reduction in temperature dependence amounts to about a factor of 10. An advantage of this method of curvature correction is that existing circuit designs can be easily modified in this way. However, there are also important disadvantages. Especially for...
circuits operated at low current levels a large chip area is needed to realize the Al resistors. Another disadvantage is that the ratio of the resistors $R_c$ and $R$ made with different technologies, will show a large spreading. As is seen from (4.27) and (4.28) this spreading results in a deviation $\Delta V_{\text{ref}}$ of the predetermined (calculated) value of $V_{\text{ref}}$ from its optimum value (with the lowest temperature coefficient). With today's technology we can expect a standard deviation $\sigma(R_c/R)$ in the resistor ratio, which will amount to about 10%. With (4.27) and (4.28) it can be calculated that this results in a standard deviation $\sigma(\Delta V_{\text{ref}}/V_{\text{ref}})$, amounting to 3000 ppm. When the bandgap references are calibrated for the predetermined value of their output voltage this deviation causes a first-order temperature dependence (see Sect. 4.5) of about $T^{-1} \sigma(\Delta V_{\text{ref}}/V_{\text{ref}}) = 10 \text{ ppm/K}$, which is unacceptably high. So, successful application of this method requires a higher absolute accuracy of the resistors, or a (more expensive) calibration method with higher precision (Sect. 4.5). An alternative method for curvature correction which does not rely on the high precision of resistors is presented in the next section.

4.4.3 Curvature correction by linearizing $V_{\text{BE}}(T)$

The large temperature dependence of the conventional bandgap references discussed in Sections 4.1 - 4.3 is caused by the thermal nonlinearity of $V_{\text{BE}}(T)$. We have developed a new accurate bandgap reference with a very low temperature coefficient in which a voltage equal to a base-emitter voltage but having a linear temperature dependence is used [4.6]. The principle of this circuit is shown in Fig. 4.18(b). For comparison the basic circuit of a conventional bandgap reference (Fig. 4.18(a)) has also been depicted. The four series-connected base-emitter junctions of $Q_1 - Q_4$ are biased at a PTAT current $I_{\text{PTAT}}$, while the three series connected base-emitter junctions of $Q_{12} - Q_{14}$ are biased at a temperature-independent current $I_{\text{ref}}$. For a transistor operated at PTAT current the thermal nonlinearity in $V_{\text{BE}}$ is about 25% less (depending on $\eta$) than that of a transistor biased at a constant current. Subtracting the three base-emitter voltages with higher nonlinearity from the four with lower nonlinearity results in a voltage $V_{\text{BE}}$, which is compensated for thermal nonlinearity. The nonlinearity of $V_{\text{BE}}(T)$ is somewhat dependent on the bias current, so that the compensation can be optimized by properly choosing this current. The linear portion of the temperature dependence of $V_{\text{BE}}$ is conventionally canceled by connecting a series resistor $R_1$ in the path of the PTAT current, which makes a reference voltage available at the emitter of $Q_{14}$. The temperature-independent current $I_{\text{ref}}$ is automatically obtained when a temperature-independent resistor is employed for $R_2$. The output voltage $V_{\text{out}}$ is regulated by the error amplifier and the series regulator transistor $Q_{23}$ in such a way that it equals $V_{\text{ref}}$. With this circuit a large reduction in the temperature dependence of the output voltage has been realized as shown in Fig. 4.19. This figure represents typical measurement results for the curvature-corrected bandgap references whose practical implementation is discussed in the next section. For comparison, also the temperature characteristic of a conventional bandgap reference is depicted.

Second-order effects as caused by base-width modulation, base currents, leakage current, internal power dissipation and resistor temperature dependence have been minimized by proper circuit design, good layout and employment of thin-film resistors. The offset voltage of the error amplifier is PTAT and is added to that of the correction voltage $V_{\text{PTAT}}$ without decreasing accuracy.
An accurate analysis of the thermal behavior will be given now.

For the voltage $V'_{BE}$ it holds that

$$V'_{BE} = \frac{4}{12} V_{BE1} - \frac{14}{T} V_{BEj}$$

From (2.13) and (2.14), with a substitution of $m = 1$ for the transistors biased with PTAT current and $m = 0$ for those biased with a constant reference current, we find

$$V'_{BE} = V_{g0} + (\eta-4) \frac{kT}{q} - \lambda' T$$

where

$$\lambda' = \frac{V_{g0}}{q} + \frac{kT}{q} (\eta-4) - \frac{V'_B T}{T}$$

When $\eta = 4$ (see App. A),

$$V'_{BE} = V_{g0} - \lambda' T.$$

The first-order temperature dependence in $V'_{BE}$ is compensated by the correction voltage $V_{PTAT}$.

Perfect compensation is achieved when

$$V_{out} = V_{g0}.$$

An offset voltage of the error amplifier, in so far it is PTAT, does not change this calibration condition. The optimal number of transistors with series-connected base-emitter junctions in the

circuit of Fig. 4.18(b) depends on the value of $\eta$. It can easily be shown that the number $n$ of transistors in the left-hand branch has to approximate $\eta$ as closely as possible, while in the right-hand branch it must equal $n - 1$. The value of $\eta$ can differ for transistors made by using different IC processes [2.7]. The number of transistors in the branches of the circuit has to be chosen in agreement with the value of $\eta$.

The magnitude of $\eta$ only slightly depends on process tolerances and on the bias current (App. A). For a given process it is possible to optimize the compensation by properly choosing the bias-current level.

### 4.4.4 Realization

Figure 4.20 is a complete schematic of the curvature-corrected bandgap reference discussed in the previous section. This reference has been fabricated by means of standard bipolar processing and thin-film-resistor technology.

Figure 4.21 shows a photograph of the chip. We will now turn to a detailed discussion of the circuit. The PTAT current is generated by the accurate source $Q_3 - Q_{10}$ which has been described in Section 4.3. The components in Fig. 4.20 have been given the same numerical subscripts as in the basic circuit of Fig. 4.18(b) for easy recognition. The left-hand
branch of Figure 4.18(b) has been interwoven with the PTAT current source in order to minimize the supply voltage. The unbuffered reference voltage $V_{\text{ref}}$ across $R_2$ is compared with the output voltage $V_{\text{out}}$ at the emitter inputs of the differential amplifier $Q_{12} - Q_{20}$. This differential stage makes up the first stage of the error amplifier, while the second stage is formed by the high gain amplifier $Q_{21} - Q_{23}$. The supply voltage for the PTAT current source and for the reference cell has been stabilized in a very effective way by having the series regulator transistor $Q_{26}$ be controlled by the collector current of $Q_{11}$ in such a way that the voltage difference between the bases of $(Q_{18}, Q_{20})$ and $(Q_{12}, Q_{15})$ equals $650 \text{ mV} (=V_{\text{BE1}})$. In this way the transistors $Q_4, Q_{12}$ and $Q_{15}$ of the basic reference cell are biased at $V_{\text{cb}} = 0 \text{ V}$ over the entire range of supply voltages and temperatures, resulting in a very low sensitivity of the output voltage to changes in the supply voltage $V^+$, amounting to only 20 ppm/$V^+$ (this figure is usually referred to as line regulation).

The method of supply-voltage stabilization chosen is to be preferred over cascoding transistors $Q_4, Q_6, Q_{12}$ and $Q_{15}$, because the base currents of the cascoding transistors would cause a decrease in accuracy of the basic reference cell, while the use of Darlington pairs would imply an increase in the minimal supply voltage. Biasing of $Q_{11}$ is provided by current mirror $Q_{24}, Q_{25}$ and $Q_{22}$. The circuit is started by means of pinch resistor $R_5$ which injects current via $D_1$ into the base of $Q_7$. Once started, $Q_{23}$ delivers current to the output and causes a voltage drop across $R_5$ which drives $D_2$ into conduction and blocks diode $D_1$. Frequency compensation is obtained by three small capacitors $C_1 - C_3$. To maintain proper biasing of the output stage the output terminals are shunted with a load resistor of less than 8 kΩ. The load current is delivered by $Q_{23}$ via $D_2$ and $Q_7$. The load regulation is 600 ppm/mA. The supply voltage has a conveniently low minimum value of 5.5 V.

Special attention has been paid to the chip layout. The transistors $Q_3 - Q_4$ have been placed in a quad configuration to reduce the influence of linear thermal gradients on the PTAT current source. Furthermore, $Q_1, Q_2, Q_{16}$ and $Q_{17}$ have also been placed in a quad configuration, while $Q_{15}$ is positioned in the geometric center of $Q_3$ and $Q_4$. In this way the difference $V_{\text{BE}}$ in the base-emitter voltages of $Q_1 - Q_4$ and $Q_{15} - Q_{17}$ are compensated for the influence of linear thermal gradients. The transistors with the highest power dissipation have been located far from the sensitive parts of the basic reference cell. The main specifications of this bandgap reference calibrated for optimum thermal behavior have been summarized in Table 4.3.

**Table 4.3**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Typical values</th>
<th>Units</th>
</tr>
</thead>
<tbody>
<tr>
<td>Output-voltage change</td>
<td>50</td>
<td>ppm $V_{\text{out}}$</td>
</tr>
<tr>
<td>-25°C to 85°C</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Line regulation</td>
<td>20</td>
<td>ppm $V_{\text{out}} / V^+$</td>
</tr>
<tr>
<td>Load regulation</td>
<td>600</td>
<td>ppm $V_{\text{out}} / mA$</td>
</tr>
<tr>
<td>Supply current (minimal)</td>
<td>400</td>
<td>μA</td>
</tr>
<tr>
<td>Supply-voltage range</td>
<td>5.5 - 15</td>
<td>V</td>
</tr>
<tr>
<td>$V_{\text{out}}$</td>
<td>1153</td>
<td>mV</td>
</tr>
</tbody>
</table>

The temperature dependence of the reference voltage is very low. The line regulation is excellent. Only the load regulation is considered to be insufficient.

When a voltage reference is very accurate with regard to temperature dependence, other specifications such as load regulation must, of course, comply with this accuracy. This pinpoints some problems which were irrelevant with conventional bandgap references. One of these problems is stability after thermal cycling. Measurements performed on the devices discussed in this section as well as on commercially available conventional devices have shown variations in the reference voltages after thermal cycling which amount to about 100 ppm for $-20°C < t < 100°C$. These variations are probably caused by the PTAT voltage: Their magnitude agrees with
Fig. 4.22 Improved version of the circuit of Fig. 4.20.

well with that derived from the results of measurements performed on transistors pairs. We hope that with an improved fabrication process this problem can be solved.

For the problem of load regulation a solution has already been found. In the circuit of Fig. 4.20 this problem is created by the base current of \( Q_7 \), which depends on the load current. Figure 4.22 shows a schematic of the most recent circuit which has been tested as a breadboard model and is currently being fabricated as an IC. In this circuit the load current is directly supplied from the positive supply voltage and has no significant influence on the output voltage. With the improved output stage it was necessary to change the start-up circuit. It is now made with a separate current source (shown in the left-hand side of the figure), which starts the circuit by injecting a current in the base of the series regulator transistor. This new start-up circuit also prevents some latch-up problems which happen to occur with the circuit of Fig. 4.20 under certain load conditions.

4.5 Precision and calibration

Even with the best circuit the precision of bandgap references is limited by nonidealities of the components used to generate and process the basic signals. As can be seen from the equation

\[
V_{\text{ref}} = V_{BE}(T) + \Delta V_{BE}
\]

(4.1) these components are

- the pair of transistors used to generate the PTAT voltage \( \Delta V_{BE} \),
- the pair of resistors used to adjust the amplification factor \( A \),
- the reference transistor by which the voltage \( V_{BE}(T) \) is generated,
- the components whose values determine the bias current for the reference transistor.

We can characterize the sensitivity of the reference voltage \( V_{\text{ref}} \) for a change in a parameter \( x \) by the sensitivity factor:

\[
\frac{\delta V_{\text{ref}}}{\delta x} = \frac{\Delta V_{\text{ref}}}{\Delta x}
\]

(4.34)

These factors have been calculated from (4.1) for changes in \( \Delta V_{BE} \), \( A \), \( V_{BE} \) and \( I_C \) respectively. Their values, assuming that \( V_{BE} = V_{\text{ref}}/2 \), have been listed in Table 4.4.

This table also includes the standard deviations \( \sigma(\Delta x/x) \) in the error distribution of the parameters over the temperature range \(-20^\circ C < T < 100^\circ C\).

This figure can be considered as the typical relative deviation of a parameter \( x \) at the extremes of the indicated temperature range as compared to what is to be expected when the value of \( x \) at \( T = T_r \) is known. The effect of these deviations on the reference voltage depends on the method of calibration. For instance, this calibration can be performed by successively measuring the thermal behavior of the individual devices and trimming each device for optimal thermal behavior.
However, it is a very complex, difficult and expensive procedure to perform such a calibration with the required precision. Therefore, we assume that the calibration is performed in the usual, much simpler, way which will be explained with the aid of Fig. 4.23. This figure plots the linear approximations of the reference voltage \( V_{\text{ref}} \) and its components \( V_{\text{BE}} \) and \( \Delta V_{\text{BE}} \) versus the temperature. The voltage reference is calibrated by adjusting the amplification factor \( A \). The curve for the reference voltage \( V_{\text{ref}} \) intersects the vertical axis at a voltage \( V_{\text{ref}}(0) \), independent of process parameters, bias-current levels or the magnitude of \( A \). The calibration is performed by trimming one of the resistors whose values determine the amplification factor \( A \) for the desired value \( V_{\text{cal}} = V_{\text{ref}}(0) \) of the reference voltage. With this simple adjustment the reference voltage automatically obtains its lowest temperature coefficient. For conventional bandgap references, as those discussed in the Sections 4.1 - 4.3, the voltage \( V_{\text{cal}} \) amounts to

\[
V_{\text{cal}} = V_{g0} + (T_{\text{cal}} - T_0) \frac{kT}{q}, \tag{4.35}
\]

where \( T_{\text{cal}} \) is the temperature at which the calibration is performed. For the curvature-corrected bandgap reference discussed in Sect. 4.4 this voltage amounts to \( V_{\text{cal}} = V_{g0} \).

With this calibration method the deviations in the temperature dependence of the various parameters \( x \) mentioned in Table 4.4 from their nominal behavior cause a temperature dependence of the reference voltage.

Table 4.4

<table>
<thead>
<tr>
<th>Parameter ( x )</th>
<th>( S_{V_{\text{ref}}} ) when ( V_{\text{BE}} = \frac{V_{\text{ref}}}{2} )</th>
<th>( \Delta V_{\text{BE}} )</th>
<th>( \Delta A )</th>
</tr>
</thead>
<tbody>
<tr>
<td>( V_{\text{BE}} )</td>
<td>1/2</td>
<td>550 ppm</td>
<td>225 ppm</td>
</tr>
<tr>
<td>( A )</td>
<td>1/2</td>
<td>240 ppm</td>
<td>120 ppm</td>
</tr>
<tr>
<td>( V_{\text{BE}} )</td>
<td>1/2</td>
<td>&lt; 100 ppm</td>
<td>&lt; 50 ppm</td>
</tr>
<tr>
<td>Bias current ( I_C )</td>
<td>( kT(qV_{\text{ref}})^{-1} )</td>
<td>2500 ppm</td>
<td>50 ppm</td>
</tr>
</tbody>
</table>

As an example, the influence of a deviation \( \delta V_C \) in the correction voltage \( V_C = \Delta V_{\text{BE}} \) from being accurately proportional to the absolute temperature is illustrated in Fig. 4.24 (only first-order temperature dependence is taken into account). When the reference voltage is calibrated for the nominal value \( V_{\text{cal}} \) at \( T = T_{\text{cal}} \) and \( \Delta V_{\text{BE}} \) is not exactly PTAT, the deviation \( \delta V_C \) in the correction voltage causes an equal deviation of the reference voltage for \( T \neq T_{\text{cal}} \). The same type of errors are caused by deviations in the temperature dependence of \( V_{\text{BE}} \) and the bias current \( I_C \).

The standard deviations \( \sigma(\delta V_{\text{ref}}/V_{\text{ref}}) \) of the relative errors in \( V_{\text{ref}} \) at the extremes of the temperature range \(-20^\circ \text{C} < t < 100^\circ \text{C} \) caused by these deviations equal the product of \( \sigma(A/x) \) and \( S_{V/x} \) and are listed in Table 4.4. From this table it can
be seen that the nonideality in the temperature
dependence of the PTAT voltage $\Delta V_{BE}$ dominates the
others. It should be clear that easy calibration of
very accurate bandgap references such as those
discussed in Sect. 4.4 necessitates a reduction in
the nonideality in $\Delta V_{BE}$. For this reason, further
investigations into improving the transistor
properties with respect to their thermal behavior
are necessary.

For the accurate bandgap references presented in
[4.5] an output-voltage change of only 160 ppm over
the $-55^\circ C$ to $125^\circ C$ temperature range has been
reported. From this it can be concluded that
National Semiconductor can make transistors with
even less spreading in their thermal behavior than
those whose data are given in Table 4.4.
5. MONOLITHIC TEMPERATURE TRANSDUCERS

5.1 Introduction

Transistors are well suited to use as temperature sensors for the temperature range \(-50^\circ C < t < 150^\circ C\), especially when low costs, high accuracy, good long-term stability and high sensitivity are required. As discussed in the previous chapters the favorable properties of transistors for this purpose are due to the highly predictable and time-independent way in which the base-emitter voltage is related to the temperature.

There are several methods for determining the temperature from the base-emitter voltage \(V_{BE}\). These can be concluded from Fig. 5.1 which shows the base-emitter voltages of two transistors operated at different levels of their emitter-current densities versus the temperature. According to the method applied, we can distinguish three types of sensors:

- the single-transistor temperature sensors, in which the base-emitter voltage \(V_{BE}\) of a single transistor is used as a measure for the temperature,
- the PTAT temperature sensors, in which the difference \(\Delta V_{BE}\) between the base-emitter voltages of two transistors is used as a measure for the temperature,
- the temperature sensors with intrinsic reference, in which a combination of \(V_{BE}\) and \(\Delta V_{BE}\) is used as a measure for the temperature.

These sensors are discussed in Sections 5.2 - 5.4, respectively. A comparison of the long-term stability and the precision of these different types of sensors is made in Section 5.5.

A number of practical circuits for IC temperature transducers have been implemented. These transducers have been designed for optimum performance of specific features, some of which are:

- accuracy and long-term stability,
- linearity,
- temperature range,
- power dissipation,
- supply-voltage range,
- type of output signal
- simplicity of calibration,
- costs (including supporting electronics).

The specific circuits dealt with in Sections 5.6 - 5.9 are:

- a general-purpose temperature transducer with an output current which is calibrated for both sensitivity and 'zero',
- a low-power easy-to-calibrate temperature transducer with a voltage output,
- an accurate small-range temperature transducer,
- a temperature transducer with a logic output and an adjustable hysteresis.

5.2 A single-transistor temperature sensor

The value of the base-emitter voltage of a transistor operated at a constant collector current is an accurate measure for the emitter-junction temperature.

A temperature meter using this feature can be implemented as shown in Fig. 5.2. The transistor is biased at a current \(I_C\) by means of the current source and the feedback amplifier A. From Eq. (2.12) it follows that for constant collector current \((m = 0)\) it holds that:

\[
\frac{T}{T_c} V_{BE}(T) - T V_{BE}(T_c) = (T_c - T) V_{BE} - q \frac{k T_c}{q} \ln \frac{T}{T_c},
\]

(5.1)
or

\[ T = T_r \left( \frac{V_{g0} - V_{BE}(T_r)}{V_{g0} - V_{BE}(T_r)} \right)^{-1} \times \left\{ (V_{g0} - V_{BE}(T)) + \frac{kT}{q} \ln \frac{T}{T_r} \right\} . \]  

(5.2)

With a microprocessor, using Eq. (5.2), the temperature \( T \) of the sensor is calculated from the digitalized value of the base-emitter voltage. The base-emitter voltage \( V_{BE} \) is not very suitable for direct use as the input signal for the A/D converter because such a use requires a high-resolution converter. To detect a temperature change \( \Delta T \) of 0.1 K at, for instance, \( T = 300 \) K with \( V_{BE} = 600 \) mV and \( \Delta V_{BE}/\Delta T = 0.2 \) mV/K the required resolution is 3000:1. So, a 12-bit converter would be needed. Therefore, in the system of Fig. 5.2 the difference of \( V_{BE} \) and a well-chosen reference voltage is used as input signal for the A/D converter. In this case the required resolution amounts to \( (T_{\text{max}} - T_{\text{min}}) : 0.1 \) K, where \( (T_{\text{max}} - T_{\text{min}}) \) is the temperature range of interest. Especially for small-range temperature transducers a much smaller resolution is needed.

The system is calibrated by measuring the base-emitter voltage \( V_{BE}(T_r) \) at an arbitrary reference temperature \( T_r \) and storing these values and the value of the reference voltage in the computer memory. The values of \( V_{g0} \) and \( \eta \) are constants for transistors made in the same production process (Apps. A and B).

The measurement system can be simplified when it is calibrated by adjusting the collector current for a specified value \( V_{BE}(T_r) \) at the measured temperature \( T_r \). In this case the values of \( V_{BE}(T) \) can be stored in a PROM and a much simpler processing unit can be used instead of a microcomputer. Presently, transistors can be manufactured with a \( V_{BE} \) tolerance of \( \pm 3 \) mV [5.1]. To adjust the base-emitter voltage of these transistors a collector-current range of \( \pm 12 \% \) is sufficient. With the single-transistor sensor an accuracy of 0.05°C for \( -20°C \leq t \leq 100°C \) can be achieved. The long-term stability of this system is mainly determined by the reference voltage \( V_{\text{ref}} \) (Sect. 5.5).

The main advantages of the single-transistor sensor are high accuracy and low sensor price. However, the processing unit is complicated and expensive. In many applications a simpler system will be preferred. Furthermore, it is often desirable that the sensors be interchangeable, but the processing unit should be the same and there should be no custom-performed calibration. These wishes are met by the IC sensors dealt with in the next section.

### 5.3 PTAT temperature sensors

#### 5.3.1 An accurate PTAT current source

In PTAT temperature sensors the basic signal is the difference \( \Delta V_{BE} \) between the base-emitter voltages of two transistors operated at a constant ratio \( r \) (\( r \neq 1 \)) of their emitter-current densities. The voltage is nominally a PTAT voltage [Sect. 2.2.6].

In temperature-sensor IC's like those manufactured by Analog Devices [5.2] and National Semiconductor [5.3] this basic voltage is amplified and buffered. The amplifier gain is adjusted at the wafer so that interchangeable calibrated devices are obtained. An accurate PTAT current source can be implemented with the configuration discussed in Section 4.3 (Fig. 5.3). This circuit implemented with thick-film resistors, has been tested. With the terminals B and D only, the device is a two-terminal current source whose main specifications have been listed in Table 5.1.

A more accurate measure for the temperature is the voltage \( V_{AB} \) between the terminals A and B. This
Table 5.1

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Measured values</th>
<th>Units</th>
</tr>
</thead>
<tbody>
<tr>
<td>Nominal supply voltage</td>
<td>4</td>
<td>V</td>
</tr>
<tr>
<td>Temperature range</td>
<td>-50 to +125</td>
<td>°C</td>
</tr>
<tr>
<td>Power dissipation</td>
<td>1.2</td>
<td>mW</td>
</tr>
<tr>
<td>Temperature coefficient of the output current</td>
<td>1.0</td>
<td>μA/K</td>
</tr>
<tr>
<td>Output impedance</td>
<td>150</td>
<td>MΩ</td>
</tr>
<tr>
<td>Absolute error</td>
<td>± 0.5</td>
<td>K</td>
</tr>
<tr>
<td>Time constant of the hybrid in stirred oil</td>
<td>2</td>
<td>s</td>
</tr>
</tbody>
</table>

The absolute error equals the product of the basic PTAT voltage and a resistor ratio and is therefore not dependent on the temperature coefficient of the resistors.

Figure 5.4 shows the absolute error of the PTAT voltage, which is within ± 0.2°C between -20°C and +100°C. This small inaccuracy is due to the imperfections of the basic PTAT voltage discussed in Section 2.2.6.

5.3.2 A voltage or current output

Some of the PTAT temperature sensors are implemented as current sources ([5.2] and Sect. 5.3.1), others as voltage sources [5.3]. Let us compare these two methods of implementation.

An advantage of the current output is that the output signal is insensitive to a voltage drop over (long) wires. However, the voltage output has the inherent advantage that the output signal is directly related to the basic physical PTAT signal kT/q, which is a voltage. Hence, to derive a current from this signal requires a resistor whose non-ideality introduces an additional inaccuracy of the output signal. To derive a voltage from this signal requires only a constant ratio of the resistors determining the amplifier gain. Therefore, in this case resistors fabricated with simpler technology (for instance base-diffused resistors) can be used.

An important difference between transducers with a voltage and a current output concerns their sensitivity to disturbing signals which are electrically or magnetically coupled to the output. This is illustrated by Fig. 5.5. For the system with the current-output sensor shown in Fig. 5.5(a) the relative influence of a capacitively coupled disturbing sinusoidal voltage \( V_{\text{dist}} \) with frequency \( \omega \) at the detected sensor current \( I_o \) amounts to:

\[
\frac{\frac{1}{V_o} \frac{\delta I_o}{\delta V_{\text{dist}}}}{\frac{\delta V_o}{V_o}} = \frac{\omega C}{1/C}.
\]

A system with a voltage-output sensor (Fig. 5.5(b)) is insensitive to this type of disturbance provided that the output impedance of the sensor is low. Actually, the network shown in Fig. 5.5(b) is the dual of that shown in Fig. 5.5(a), which means that in this case the sensor signal is sensitive for magnetically coupled disturbing currents, where it holds that:

\[
\frac{\frac{1}{V_o} \frac{\delta V_o}{\delta I_{\text{dist}}}}{\frac{\delta V_{\text{dist}}}{V_{\text{dist}}}} = \frac{\omega M}{V_o}.
\]
twisting the wires or using filters, can be applied. The choice of either a current or a voltage output may depend on what type of disturbance can be expected in the environment in which the sensor is to be used.

5.4 Temperature sensors with an intrinsic reference

Already for a number of years IC temperature transducers with a PTAT output signal have provided a welcome solution to many temperature measurement problems. An important feature of these transducers ([5.2], [5.3] and Sect. 5.3) is not only the linearity of the output signal with temperature changes, but also the possibility to calibrate them economically by trimming a resistor at the chip. Commonly, the output signal of a temperature transducer serves as the input signal for an A/D converter or another type of data-processing system. Like the single-transistor sensor (Sect. 5.2) PTAT temperature sensors also have the main drawback of requiring a high-resolution measurement system to detect small temperature changes. For instance, to detect a 0.1 K temperature change at a temperature of 300 K the required resolution is 3000 : 1. With an output signal at a °C scale the required resolution would amount to only 270 : 1. This example shows that when the temperature range of interest is small it is advantageous to have a temperature transducer with its 'zero' at a temperature in or close to the range of interest. A signal at a °C, °F or another scale can be obtained with the system shown in Fig. 5.6, which has been implemented with a PTAT temperature sensor, a voltage reference and a differential amplifier. Such a system is expensive because of the large number of accurate components needed and the many calibration steps performed during the production of the IC's or during manufacture of the system. For just these applications an IC temperature transducer which is almost as simple as a single PTAT temperature transducer has been designed. By application of the intrinsic bandgap voltage as reference high accuracy and good long-term stability are achieved. An important advantage of this transducer concerns the simplicity of the calibration method [5.4], [5.5]. The basic circuit of the sensor with intrinsic reference is shown in Fig. 5.7. The current source generates a current \( I_{\text{PTAT}}/R_2 \), where \( V_{\text{PTAT}} \) is a voltage proportional to the absolute temperature (PTAT) and \( R_2 \) is an internal resistor in the current source. The bias current for transistor \( Q_7 \) is provided by means of the pnp-current mirror. The high-gain feedback amplifier \( A_1 \) forces the collector current of \( Q_7 \) to equal the output current of the current mirror and by the output shunt feedback lowers the output impedance. The output voltage \( V_0 \) amounts to

\[
V_0 = V_{BE7} - V_{\text{PTAT}} \frac{R_1}{R_2}.
\]

The first term on the right-hand side of (5.5) shows an almost linear decrease with temperature:

\[
V_{BE} = V_{g0} - C_T,
\]

where \( V_{g0} \) is the linearly extrapolated bandgap voltage at 0 K. Hence, when the output voltage \( V_0 \) is zero at a temperature \( T_Z \) (Fig. 5.8), we find for the output voltage

\[
V_0 = -\frac{T - T_Z}{T_Z} V_{g0}.
\]

The curve for the output voltage intersects the
vertical axis at the voltage $V_{0}$ independent of process parameters. Calibration of the transducer is performed by trimming a resistor (for instance $R_{1}$) for a desired value $V_{0}(T_{\text{cal}})$ of the output voltage. With this single adjustment the output voltage is calibrated over a wide temperature range.

Because it is our aim to realize accurate temperature transducers, the thermal nonlinearity in $V_{BE}(T)$ and the effect of the temperature dependence of the resistors also have to be taken into account.

But, as can be shown, these effects do not complicate the calibration. Firstly, we deal with the thermal nonlinearity of $V_{BE}(T)$, assuming that the resistors $R_{1}$ and $R_{2}$ are temperature-independent.

In Section 5.7.2 we will discuss the influence of the temperature dependence of the resistors.

With $V_{PTAT}(T) = (T/T_{Z}) V_{PTAT}(T_{Z})$ it can be found from (5.5) and from $V_{o}(T_{Z}) = 0$ that

$$V_{o} = \frac{-T V_{BE7}(T_{Z}) + T_{Z} V_{BE7}(T)}{T_{Z}}.$$  (5.8)

Substitution of Eq. (2.13) in (5.8) gives with $m = 1$:

$$V_{o} = \frac{V}{T_{Z}} \left( \frac{R_{0}}{T_{Z}} + \frac{g - 1}{q} \right) \text{linear term} + \frac{g - 1}{q} \frac{K}{T_{Z}} \text{higher-order term}.$$  (5.9)

The linear term in the right-hand side of (5.9), almost equals the approximated one of (5.7), and represents the tangent to the $V_{o}(T)$ curve for $T = T_{Z}$.

Approximation of $V_{o}$ by only this linear term causes a small error which can be represented as $\Delta T$ (see Fig. 5.9(a)). This error is shown in Fig. 5.9(b) for $T_{Z} = 273.15 \text{ K}$. The sensitivity $\delta V_{o}/\delta T$ is concatenated with $T_{Z}$ and amounts for $T = T_{Z}$ to:

$$\frac{\delta V_{o}}{\delta T}\bigg|_{T = T_{Z}} = -\frac{V_{0}}{T_{Z}} (g - 1) \frac{K}{q},$$  (5.10)

which yields $-4.51 \text{ mV/°C}$ for $V_{0} = 1172 \text{ mV}$, $g = 3.5$ and $T_{Z} = 273.15 \text{ °C}$.

Calibration of the device is necessary because of spreading in the process parameters and can be performed at an arbitrary temperature $T_{\text{cal}}$ by trimming $R_{1}$ in such a way that the output voltage has the correct value, i.e. the one calculated with (5.9) by substituting $T = T_{Z}$.

Note that for this calibration it is not necessary to know the actual values of $V_{PTAT}$ and $V_{BE7}$. For this reason, the spreading in the transistor parameters does not affect the output current of the calibrated transducer. Figure 5.10 shows the 'zero' $T_{Z}$ in °C versus the magnitude of $R_{1}$ which is normalized with respect to a value $R_{1,\text{ref}}$ for which $T_{Z} = 0 \text{ °C}$. The curve shown has been calculated by
When $V_{BE}$ is equal for both the bandgap reference and the temperature sensor at equal temperatures, and bearing in mind that $V_o(T_z) = 0$ for $T = T_z$, we can calculate from (5.11) and (5.12) that

$$A_1 = \frac{V_{ref} - V_{BE}(T_z)}{\Delta V_{BE}(T_z)} = A_2 \frac{V_{ref} - V_{BE}(T_z)}{V_{BE}(T_z)}.$$  \hspace{2cm} (5.13)

With $V_{BE}(T_z) = \frac{1}{2}V_{ref}$ it is found that

$$A_1 = A_2.$$ \hspace{2cm} (5.14)

Assuming that the long-term changes in $V_{BE}$ and $A_1 \Delta V_{BE}$ are uncorrelated (in view of their different origins this seems to be justified), it can be concluded that the effective values of the long-term drift in $V_{ref}$ and in $V_o$ are approximately equal. If the relative change $\Delta V_{ref}/V_{ref}$ of a bandgap reference is known, the corresponding change $\Delta T$ of a temperature sensor with intrinsic reference and a voltage output is found from

$$\Delta T = (\Delta V_{ref}/V_{ref})T_z.$$ \hspace{2cm} (5.15)

Practical values for the long-term stability of bandgap voltage references amount to $10^{-4}$ [5.6], which corresponds to $\Delta T = 0.027^\circ C$ for the temperature transducer.

For the temperature transducer with a current output, to be discussed in Section 5.6, an additional scale-factor error due to the long-term drift of the resistor values has to be accounted for. For thick-film resistors, the long-term stability amounts to $5 \times 10^{-4}$, which results in an additional error $\Delta T = 5 \times 10^{-4} \times (T - T_z)$.

In Section 4.5 we showed that the nonidealities in bandgap references are mainly due to those of the PTAT voltage $A_1 \Delta V_{BE}$. Under the assumption that this also holds true for the long-term stability we can compare the various types of temperature transducers with respect to their long-term stability.

When the assumption just made holds true, then a single-transistor temperature sensor has a better long-term stability than a PTAT sensor. However, with the circuit of Fig. 5.2 the applied reference voltage can be a cause of stability problems. A relative drift $\Delta V_{ref}$ of the reference voltage would cause an error $\Delta T$ in the temperature measurement which amounts to:
Fig. 5.11 A very accurate and simple system with single-transistor sensors for the detection of small temperature differences. For this special application no voltage reference is needed.

$$\Delta T = \frac{\Delta V_{\text{ref}}}{V_{\text{ref}}} \frac{\delta V_{\text{BE}}}{V_{\text{BE}}} - 1.$$  \hfill (5.16)

When a bandgap reference is used (because of its good long-term stability), with $V_{\text{BE}} = 600 \text{ mV}$ and $\delta V_{\text{BE}}/\delta T = -2 \text{ mV/K}$ at 300 K, this error is just as large as given in (5.15) for the sensor with intrinsic reference. So a better long-term stability for a single-transistor sensor is only obtained when the environmental conditions for the reference-voltage source are better than those for the sensor.

In some applications it is not necessary to use a voltage reference. Then it is often advantageous to use single-transistor sensors in view of their high precision and superior long-term stability. For instance, to detect small temperature differences the circuit shown in Fig. 5.11 would be very appropriate.

The PTAT temperature sensors are the least accurate ones. An error $\Delta V_{\text{PTAT}}$ in the basic PTAT voltage causes an error $\Delta T$ in the temperature measurement which amounts to:

$$\Delta T = \frac{\Delta V_{\text{PTAT}}}{V_{\text{PTAT}}} T.$$  \hfill (5.17)

In temperature sensors with an intrinsic reference the basic PTAT voltage contributes to only about half the sensitivity of the output signal (the other half is contributed by the base-emitter voltage). Consequently, in the temperature sensor with intrinsic reference an error in $V_{\text{PTAT}}$ causes an error in the output signal which is only half that of a PTAT sensor. When a PTAT sensor is used in combination with a voltage-reference source, to reduce the requirements with respect to resolution, long-term stability will be worse, due to drift of the reference source.

The influence of nonidealities of the components on the precision of the various types of temperature transducers can be calculated from the data presented in Table 4.4 for bandgap references. The spread in the basic PTAT voltage, which amounts to 550 ppm for $-20^\circ \text{C} < t < 100^\circ \text{C}$ corresponds to an error of 0.15°C of the PTAT signal. For the temperature transducer with intrinsic reference this error is two times lower, while for the single-transistor sensor an error $< 30 \text{ mK}$ is found. For the PTAT sensor and the single-transistor sensor the nonidealities of an applied voltage reference also have to be taken into account.

5.6 A general-purpose temperature transducer with a fully calibrated output current

5.6.1 Principle

In the previous sections it has been shown that temperature sensors with intrinsic reference have important advantages with regard to resolution aspects, simplicity of calibration, and precision, as compared to the PTAT- and single-transistor sensors. In the remaining part of this chapter we will discuss a number of practical implementations of this type of temperature sensor.

In an initial attempt to realize such an implementation we designed a temperature transducer with a current output whose 'zero' and sensitivity can be adjusted for the desired values by laser trimming two resistors. This temperature transducer is intended to operate over the $-50^\circ \text{C}$ to $+125^\circ \text{C}$ temperature range. The basic circuit configuration is shown in Fig. 5.12. The current source generates a current $V_{\text{PTAT}}/R_2$, where $V_{\text{PTAT}}$ is a PTAT voltage and $R_2$ is an internal resistor in the current source. The bias current for transistor $Q_9$ is provided by means of the pnp current mirror and the feedback amplifier $Q_8$. When we neglect the influence of the base currents and base-width modulation it is found for the output current $I_0$ that

$$I_0 = V_{\text{PTAT}}/R_2 - V_{\text{BE}}/R_1 = \frac{1}{R_1}(V_{\text{PTAT}}R_2 - V_{\text{BE}}).$$  \hfill (5.18)

This equation is almost identical to Eq. (5.5) which describes the output voltage of a temperature
The sensitivity \( \delta \frac{I}{\delta T} \) can freely be chosen within a range limited by the maximum tolerable power dissipation on the upper side and the influence of leakage currents and interfering signals on the lower side. When a low value of \( \delta \frac{I}{\delta T} \) is chosen, for instance in order to minimize power dissipation, this may present the problem of requiring high-valued resistors to convert the basic voltages into currents.

\section{The influence of base currents and temperature dependence of resistors}

The main deviations of the actual circuit behavior from the idealized case discussed in the previous section are due to the base currents of the transistors \( Q_7 \) and \( Q_8 \), and the temperature dependence of the resistors. We will successively calculate their influence.

Taking account of the finite current gain \( h_{FE} \) of \( Q_7 \) and \( Q_8 \), and assuming that \( h_{FE7} = h_{FE8} = h_{FE} \), we find for the output current \( I_o \) that

\[ I_o = \frac{h_{FE}^2 + 1}{h_{FE}^2 + h_{FE} + 1} \left[ \frac{V_{BE7}}{R_I} \right] \]

(5.21)

With \( I_{PTAT}(T) = (T/T_Z)I_{PTAT}(T_Z) \) and \( I_o(T_Z) = 0 \AA \) it can be found from (5.21) that

\[ I_o = \frac{h_{FE}^2}{h_{FE}^2 + h_{FE} + 1} \left[ \frac{T}{R_I} \right] \]

(5.22)

The temperature dependence of \( h_{FE} \) can be represented well by the equation (Sect. 2.2.3)

\[ h_{FE}(T) = h_{FE}(T_Z) \left[ 1 + p(T - T_Z) \right] \]

(5.23)

From (5.22) and (5.23) it can be calculated that

\[ \left( \frac{\delta I}{\delta T} \right) \bigg|_{T=T_Z} = \frac{1}{R_I} \left( \frac{h_{FE}(T_Z)^2}{h_{FE}(T_Z)^2 + h_{FE}(T_Z) + 1} \right) \left[ \frac{V_{BE7} + (\eta-1)kT_Z}{T_Z} \right] \]

(5.24)

Comparing (5.24) and (5.20), after substituting the values of \( V_{BE7} \) and \( I_{PTAT} \).
typical values $h_{FE} = 100$ and $V_{BE} = 700$ mV for $t_z = 0{\text{C}}$, shows that the effect of finite current gain $h_{FE}$ lowers the sensitivity $\delta I / \delta T$ about 1%. With the adjustment of $R_1$ this effect has to be counteracted by trimming $R_1$ for a correspondingly smaller value. For instance, when the desired sensitivity $\delta I / \delta T$ amounts to $1$ $\mu$A/$^\circ$C for $t_z = 0{\text{C}}$, it is found with $V_{g0} = 1172$ mV, $n = 3.5$ and $h_{FE} = 100$ that $R_1 = 4461$ $\Omega$, while with infinite current gain it is found that $R_1 = 4506$ $\Omega$.

The nonlinearity in $I_0(T)$ is hardly affected by the influence of $h^\infty_{FE}$. Next, we will account for the temperature dependence of the resistors. For transducers with a current output we need rather accurate resistors, because the basic signals $V_{g0}$ and $kT/q$ are voltages. Thin-film resistors are well suited to our purpose. These resistors vary linearly with $T$ (Sect. 2.4.1) so it holds that

$$R_1 = R_{1,0} \left(1 + a(T - T_z)\right), \quad (5.25)$$

where $R_{1,0}$ is a temperature-independent resistance and $a$ is the temperature coefficient. Substitution of (5.25) in (5.19) gives a good approximation:

$$I_0 = I_{0,1} - \frac{a(T - T_z)^2}{R_{1,0} \cdot T_z} \cdot V_{g0}, \quad (5.26)$$

where the first term in the right-hand side of (5.26) denotes the output current for the case in which $a = 0$ $\mu$A/$^\circ$C.

Equation (5.26) shows that a first-order temperature dependence of $R_1$ effects a second-order temperature dependence of $I_0$ (represented by the last term in (5.26)). We can use this effect to compensate for the second-order temperature dependence in $I_0 | R_1 = R_{1,0}$, which, as the approximation of (5.19) by the first three terms of its Taylor expansion shows, amounts to

$$\frac{(n-1)}{2R_{1,0}} \cdot \frac{(T-T_z)^2}{T_z} \cdot \frac{kT}{q}.$$  

With (5.26) it is found that this term is fully compensated for when $a$ has a value $a_{\text{comp}}$ for which it holds that

$$a_{\text{comp}} = \frac{k \cdot n-1 \cdot V_{g0}}{q \cdot 2V_{g0}}, \quad (5.27)$$

With $V_{g0} = 1172$ mV and $n = 3.5$ this yields $a_{\text{comp}} = 92$ ppm/$^\circ$C, which is a realistic value for a thin-film-resistor temperature coefficient. So we can conclude that the temperature transducers can be curvature-corrected by applying thin-film resistors with a proper composition of their alloy.

### 5.6.3 Experimental results

Figure 5.13 shows a complete schematic of a general-purpose Celsius temperature transducer conforming to the principle discussed in the previous sections. For the PTAT current source, the configuration presented in Section 4.3.2 has been chosen because of its high accuracy, the low sensitivity of the PTAT output current to leakage currents flowing into the substrate, and the high output impedance. Devices have been fabricated using laser-trimmed thick-film resistors and thin-film-on-silicon resistors, respectively. Photographs of these devices are shown in Figs. 5.14 and 5.15. The thick-film resistors have been calibrated at ambient temperatures in the way just discussed in the previous sections. The sensitivity amounts to $1$ $\mu$A/$^\circ$C, the 'zero' to $0^\circ$C. The thin-film devices have not been calibrated. Figure 5.16 shows the measured nonlinearity of the devices. The difference in nonlinearity was found to be due to different temperature coefficients (TC) of the resistors. For the thick-film resistors a TC of $20$ ppm/$^\circ$C was measured, while for the thin-film...
resistors this figure amounted to 180 ppm/°C. The high TC of the thin-film resistors causes over-compensation of the nonlinearity, which agrees with the analysis given in Section 5.6.2. Further specification of the devices as measured for the hybrid circuit are listed in Table 5.2. Note the excellent power-supply rejection and the high output impedance.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>measured values</th>
<th>Units</th>
</tr>
</thead>
<tbody>
<tr>
<td>Nominal supply voltage</td>
<td>+5 and -3</td>
<td>V</td>
</tr>
<tr>
<td>Rated performance temperature</td>
<td>-50 to +125</td>
<td>°C</td>
</tr>
<tr>
<td>Power dissipation at 300 K</td>
<td>1.5</td>
<td>μW</td>
</tr>
<tr>
<td>Temperature coefficient of the output current</td>
<td>1</td>
<td>μA/°C</td>
</tr>
<tr>
<td>Zero-output-signal temperature</td>
<td>0</td>
<td>°C</td>
</tr>
<tr>
<td>Output impedance at 300 K</td>
<td>150</td>
<td>MΩ</td>
</tr>
<tr>
<td>Absolute error</td>
<td>±0.5</td>
<td>°C</td>
</tr>
<tr>
<td>Error due to supply-voltage variations</td>
<td>0.02</td>
<td>°C/V</td>
</tr>
<tr>
<td>Time constant of the hybrid device in stirred oil</td>
<td>2</td>
<td>s</td>
</tr>
</tbody>
</table>

5.7 A micropower easy-to-calibrate temperature transducer implemented with diffused resistors

5.7.1 Design considerations

Temperature transducers with a current output as discussed in Section 5.6, are afflicted with two main drawbacks: Their implementation requires accurate resistors to convert the basic signals
$V_{g0}$ and $kT/q$ into currents, and at least two calibration steps are needed to calibrate for the desired 'zero' and sensitivity. These drawbacks are overcome by temperature transducers with a voltage output conforming to the principle discussed in Sect. 5.4. With these circuits the output voltage only weakly depends on the resistor magnitudes. Therefore, even diffused resistors, which have a large nonlinear temperature coefficient, can be used. In addition to process simplicity an important advantage of diffused resistors is their good long-term stability.

Once the 'zero' $T_z$ of these transducers has been chosen, as has been shown in Sect. 5.4 by Eq. (5.10), the sensitivity is also determined. So, in a single calibration step both the desired 'zero' and the sensitivity are calibrated. The simplicity of this calibration is an important advantage of the device because of the associated reduction of complexity and costs of fabrication.

As will be shown in Sect. 5.7.4 in a consideration of thermocouple cold-junction compensation, in some specific applications calibration is done even easier by using the offset control of an op amp.

Compared to the transducer presented in Sect. 5.6, this device has been designed for a much lower power dissipation (200 $\mu$W). This has been achieved by stabilizing the internal supply voltage at a minimum value and using supply currents instead of voltages. Due to the reduced level of power dissipation only a little self-heating occurs, enabling the sensor to be used in applications where the thermal resistance between chip and ambient is relatively high (Chap. 3). Furthermore, the low value of the power dissipation is attractive when a battery is used for the energy supply.

5.7.2 Influence of temperature dependence of the resistors and finite current gain

The main deviation of the actual circuit behavior from the idealized one, discussed in Sect. 5.4, is due to the large temperature dependence of the diffused resistors (Sect. 2.4.1). This temperature dependence can be approximated well by the second-order relationship:

$$R(T) = R(T_r)(1 + a_1(T - T_r) + a_2(T - T_r)^2)$$

(5.28)

For pairs of resistors with a carefully designed layout (Sect. 2.4) the observed mismatch in temperature coefficient lies in the range of 2 to 10 ppm/°C. This mismatch is so small that the temperature dependence of the resistor ratio $R_1/R_2$ in (5.5) can be neglected. The collector current of $Q_7$ amounts to $V_{BE7}/R_2$. Substituting this value in $V_{BE7} = (kT/q) \ln (I_{C7}/I_{S7})$ yields with (5.28)

$$V_{BE7} = V_{BE7}R_2$$

is constant

$$-\frac{kT}{q} \ln (1 + a_1(T - T_r) + a_2(T - T_r)^2),$$

(5.29)

where $V_{BE7}R_2$ is constant is the base-emitter voltage for the case in which $a_1 = 0K^{-1}$ and $a_2 = 0K^{-2}$.

With this equation we find from (5.5):

$$V_o = V_{BE7}$$

is constant

$$-\frac{kT}{q} \ln (1 + a_1(T - T_r) + a_2(T - T_r)^2).$$

(5.30)

For the temperature range -50°C < $t$ < +125°C one finds, as a good approximation when $T = T_r$, with the approximation $\ln(1+x) = 1 + x - \frac{x^2}{2}$ that:

$$V_o = - (T - T_z) \left( \frac{V_{g0}}{T_z} + \frac{k}{q} (\eta - 1 + a_1 T_z) \right)$$

$$- (T - T_z)^2 \frac{k}{q} \frac{\eta - 1}{2 T_z^2} + a_1 + \frac{(2a_2 - a_1^2) T_z}{2}.$$  

(5.31)

With $T_z = T_2 = 273.15K = 0°C$, $V_{g0} = 1172 mV$, $\eta = 3.5$ and $a_1 = 1700 ppm/°C, a_2 = 5.5 ppm/(°C)^2$ (experimental values) for the sensitivity $\delta V_o/\delta T$ of the transducer at $t = 40°C$ (midrange temperature) we find

$$\frac{\delta V_o}{\delta T} \bigg|_{t = 40°C} = -4.60 mV/°C.$$  

Comparison of (5.31) with (5.9) shows that the temperature dependence of the resistors causes the nonlinearity to increase by about 60% and amounts to ± 0.6°C from -50°C to 125°C. The base current $I_{BE7}$ of $Q_7$ has only a little influence on the performance. With the typical values $h_{FE7} = 100$ and $\delta h_{FE7}/\delta T = 0.005/°C$ it can be calculated that the decrease in sensitivity amounts to only 0.44%.
5.7.3 Practical Realization

Figure 5.17 is a complete schematic of the general-purpose temperature transducer conforming to the principle discussed in Section 5.4. For the current source \( Q_1 - Q_6 \) the configuration discussed in Section 4.3 has been chosen because of its accuracy and simplicity. Two junction capacitors of 30 pF prevent parasitic oscillations, which can easily occur due to the internal feedback in the circuit. Empirically, we determined the optimum values and locations for these capacitors.

The supply voltage is stabilized by the diodes \( D_1 - D_4 \). This method of stabilization has a number of advantages: To begin with, the collector-base voltages of \( Q_1 - Q_6 \) are almost temperature independent, which eliminates base-width modulation effects. Furthermore, the supply voltage has its minimum value over the entire temperature range, which results in a low power dissipation. Finally, a starting circuit which works perfectly over the entire temperature range can be realized with only two diodes (D5 and D6). The nominal supply currents amount to about +80 μA and -45 μA.

Transistor \( Q_9 \) operates as a current amplifier and lowers the output impedance. Transistor \( Q_{10} \) prevents latch-up problems. Trimming of the circuit is obtained by short-circuiting some parts of the resistor \( R_1 \) by zener zapping (Sect. 2.4.2). In our device the resistor parts amount to \( R_n^2 \) (n = 0, 1, ..., 4 and \( R_0 = 150 \Omega \)). In this way the 'zero' is trimmed to within ±0.3°C over a range of ±10°C.

For the zeros of uncalibrated devices we found a spread of only ±5°C, which easily falls within the trimming range.

A microphotograph of the chip (type EL-227) is shown in Fig. 5.18.

The measured values of sensitivity (-4.62 mV/°C) at 40°C as well as of the nonlinearity (Fig. 5.19) are in good agreement with those predicted by theory (see Sect. 5.7.2 and Fig. 5.19). Essential specifications of the device are listed in Table 5.3. Note the excellent power-supply rejection. We found that the sensitivity of the output signal to changes

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**Fig. 5.17** Complete schematic of the micropower easy-to-calibrate temperature transducer. For \( V^+ = 5 \) V and \( V^- = -6 \) V choose \( R^+ = 27 \) kΩ and \( R^- = 120 \) kΩ.

**Fig. 5.18** Microphotograph of the 1100 μm x 1900 μm chip (type EL-227).

**Fig. 5.19** Measured and calculated nonlinearity of the 6°C temperature transducer (made zero for midrange temperature) plotted versus the temperature.
in the negative supply current is mainly due to the
influence of the base current of $Q_9$, which affects
the collector current of $Q_7$.

### Table 5.2

Performance of the temperature transducer with a
voltage output of the type El-227.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Measured values</th>
<th>Units</th>
</tr>
</thead>
<tbody>
<tr>
<td>Nominal supply current</td>
<td>+80 and -45</td>
<td>µA</td>
</tr>
<tr>
<td>Rated performance temperature range</td>
<td>-50 to +125</td>
<td>°C</td>
</tr>
<tr>
<td>Power dissipation at $40°C$</td>
<td>200</td>
<td>µW</td>
</tr>
<tr>
<td>Temperature coefficient of the output voltage at $0°C$</td>
<td>-4.6</td>
<td>mV/°C</td>
</tr>
<tr>
<td>Zero-output-signal temperature</td>
<td>0</td>
<td>°C</td>
</tr>
<tr>
<td>Nonlinearity</td>
<td>See Fig. 5.5</td>
<td></td>
</tr>
<tr>
<td>Output impedance at $40°C$</td>
<td>40</td>
<td>Ω</td>
</tr>
<tr>
<td>Sensitivity to supply-current variations</td>
<td>$1.6 \times 10^{-3}$</td>
<td>°C per % change in $I$</td>
</tr>
<tr>
<td></td>
<td>$4 \times 10^{-3}$</td>
<td>°C per % change in $I$</td>
</tr>
</tbody>
</table>

The power dissipation has a pretty low value which
keeps self-heating effects small. To calculate the
power to be supplied by the batteries the
dissipation in the external resistors also has to
be accounted for. With $V^+ = 5$ V and $V^- = -5$ V the
total supplied power is about 600 µW.

#### 5.7.4 Thermocouple cold-junction compensation

An interesting application for the temperature
transducer El-227 discussed in the preceding
sections is that of cold-junction compensation of
thermocouple signals. As will be shown in this case
it is possible to use even an uncalibrated El-227
as a cold-junction temperature sensor.

Figure 5.20 shows a new thermocouple system with
cold-junction compensation for a °C-thermometer.
The temperature $T_m = T_m - 273.15 °$K of the measuring
junction of the thermocouple has to be depicted on
a display.

In this measuring system the El-227 is held
isothermally to the thermocouple reference (cold)
junction as well as to the op amp. Let us assume
that the El-227 is uncalibrated; then the objectives
of the calibration procedure of the system are to
adjust
- the sensitivity of the amplified thermocouple signal;
- the sensitivity of the amplified signal of the
  El-227;
- the compensation for the offset voltage of the op
  amp;
- the zero (0°C) of the output signal.

Next we will show that all of these adjustments can
be performed in only two steps by using the offset
control of the op amp and the gain control of the
panel meter.

When we assume that the offset voltage is PTAT
(which holds true for a well-designed op amp) and
amounts to $V_{ offset} = \gamma T_{ ref}$, then for the value $V_D$
displayed at the panel meter it holds that:

$$V_D = A_1 \lambda_1 (T_m - T_{ ref}^*) + A_1 \gamma T_{ ref} - A_2 \lambda_2 (T_{ ref} - T_z),$$

where
- $A_1$ and $A_2$ = the gain factors for the thermocouple
  signal and the El-227 signal, respectively;
- $\lambda_1$ and $\lambda_2$ = the temperature coefficient of the
  thermocouple signal and the El-227
  signal, respectively;
- $T_m$ = the temperature of the thermocouple
  measuring junction;
- $T_{ ref}$ = the temperature of the thermocouple
  reference junction;
- $T_z$ = the zero of the uncalibrated El-227.

The resistors $R_2$ and $R_1$ are chosen in such a way
that

$$\frac{R_2}{R_1} = \frac{-\lambda_2}{\lambda_1}.$$

(5.33)
By substituting, for example, \( \lambda_1 = 40 \mu W/°C \) and 
\( \lambda_2 = -4.6 mV/°C \), we find \( r_2/r_1 = 115 \). This ratio, which determines the gain of the EL-227 signal, is not very critical and can be made with 5% resistors without adjustment. With the first calibration step the gain \( A_1 \) for the thermocouple signal is adjusted by means of the gain control of the panel meter or of \( R_3 \). For this purpose, the thermocouple is replaced by a calibrated voltage source.

The final calibration step is performed by disconnecting the thermocouple while shorting the thermocouple inputs and adjusting the offset control of the op amp for a displayed value \( V_D \) which corresponds to the measured value of \( T_{ref} \). It can be found from (5.32) and (5.33) that with these two adjustments it holds that:

\[
V_D = A_1 \lambda_1 T_{ref}'
\]  
(5.34)

Note that with the last calibration step we obtain compensation of the offset voltage as well as adjustment of the zero of the system. This combined adjustment is only possible if the offset voltage is PTAT and can be adjusted in a range of at least \( \pm 0.4 mV \). The OP-05 of Precision Monolithics [5.7] is highly suitable for this purpose. The long-term stability of the offset voltage of this op amp amounts typically to \( 0.2 \mu V/month \) which would cause a drift in the calibrated temperature meter of only \( 0.005 °C/month \).

We measured the temperature dependence of the offset voltage of the OP-05 for six different devices and found that the deviation from being PTAT is about 1.5% over the -20°C to 100°C temperature range. The effect of such a deviation in the °C thermometer of Fig. 5.20 is proportional to the magnitude of the adjusted offset voltage and the width of the temperature range over which the op amp temperature \( T_{ref} \) changes. For the maximum offset voltage of 0.4 mV mentioned above, the deviation of the offset voltage from being PTAT would cause an error of 0.15°C of the thermometer for a 100°C temperature change in \( T_{ref} \) (in many applications \( T_{ref} \) will change only over a small temperature range resulting in a corresponding lower value of the error).

5.8 An accurate small-range temperature transducer

5.8.1 Design considerations and basic circuit

In this section a special-purpose thermometer especially intended for clinical use [5.8] is discussed. In comparison to the general-purpose transducers discussed in the previous sections, the one presented here has the following features:

- higher accuracy (0.1°C) over a smaller temperature range (from 20°C to 50°C),
- low dissipation (250 \mu W) and consequently little self heating,
- only one power supply (9 V),
- a buffered output voltage with a large temperature coefficient (0.1 V/°C).

The specific features of this temperature transducer made it necessary to design a special circuit whose basic configuration differs considerably from the previously discussed ones. The temperature coefficient of the output voltage has been given a large value to enable readout with a simple digital voltmeter. Consequently, the output voltage has a large swing over the temperature range of the transducer, and a relatively large value of the supply voltage is needed.

To keep power dissipation low it was necessary to reduce the supply current. We performed this by making the current in the basic sensor cell proportional to the temperature in °C instead of K as is the case with PTAT current sources. This enables the circuit to be biased at a current which is small but nevertheless has a large temperature coefficient which keeps the influence of non-idealities such as leakage currents, etc., small.

The power dissipation has further been reduced by interweaving the sensor cell with the buffer amplifier (see Sect. 5.8.2).

The principle of the circuit will be explained on the basis of Fig. 5.21. The collector current of \( Q_1 \) is kept equal to that of \( Q_2 \), for instance, by means of a pnp-current mirror. Initially, neglecting the base currents and the base-widening effects, we find that the difference \( \Delta V_{BE} \) in the base-emitter voltages of \( Q_1 \) and \( Q_2 \) amounts to

\[
\Delta V_{BE} = \frac{kT}{q} \ln (r),
\]  
(5.35)
where \( r \) denotes the emitter-area ratio of \( Q_2 \) and \( Q_1 \).

The current source \( I \) provides the bias current for \( Q_7 \) by means of the feedback amplifier \( Q_8 \). The value of this current \( I \) is not critical and is allowed to depend on the temperature, as indicated in the figure.

For the collector current \( I_{C2} \) of \( Q_2 \) it is found that

\[
I_{C2} = I_{E2} = \frac{1}{R_2} \left( \frac{R_1 + R_2}{R_1} \Delta V_{BE} - V_{BE7} \right). \tag{5.36}
\]

Neglecting the offset voltage of the voltage follower for the output voltage \( V_o(T) \) it holds that \( V_o(T) = I_{C2} R_3 \). Note that Eq. (5.36) is almost identical to Eq. (5.5). However, for the circuit of Fig. 5.21 the collector current of \( Q_7 \) is not PTAT but proportional to \( T^m \). Therefore, \( V_{BE7} \) is found directly from Eq. (2.13) without having to make a substitution for \( m \). Taking this into account the calculation of the output voltage \( V_o \) is similar to that presented in Sect. 5.4 and results in

\[
V_o(T) = \frac{R_3}{R_2} \left[ (T - T_z) \left\{ \frac{V^{BE}_0}{T_z} + \frac{(n-m) k}{q} \right\} + \frac{(n-m) k}{q} (T - T_z - T \ln \frac{T}{T_z}) \right]. \tag{5.37}
\]

In Sect. 5.4 it is shown that such an output voltage is almost a linear function of the temperature. As will be discussed in Section 5.8.3 calibration is obtained by trimming the resistor ratio \( R_3/R_2 \) for the desired scale factor and then trimming \( R_1 \) for the right value of \( T_z \). Choosing \( T_z = 273.15 \, K \) makes \( V_o(T) \) a °C signal. Note from Eq. (5.37) that the output signal of the calibrated sensor is not affected by spread in transistor parameters.

Equations (5.36) and (5.37) show that when the output voltage is proportional to the temperature

\[
\text{in °C the bias currents of } Q_1 \text{ and } Q_2 \text{ are as well.}
\]

### 5.8.2 Practical realization

The complete schematic of the clinical thermometer is shown in Fig. 5.22. A very low supply current is obtained by interweaving the separate circuit parts. The transistor pair \( Q_1, Q_2 \) of the basic circuit has been replaced by the quad of transistors \( Q_1 - Q_4 \) through which a PTAT voltage \( \Delta V_{BE} = kT/q \ln (r_{12} r_2) \) across \( R_1 \) is developed, which is independent of bias-current levels (Sect. 4.3).

Cascoding \( Q_3 \) and \( Q_4 \) by \( Q_5 \) and \( Q_6 \) reduces the influence of base-widening effects. The current \( I_3 \) through \( R_3 \) is by good approximation equal to the collector current \( I_{C4} \) of \( Q_4 \). For the voltage across \( R_3 \), an expression identical to (5.37) holds. The transistors \( Q_{11} - Q_{16} \) form a unity-gain buffer amplifier. The combination of transistors \( Q_{17} \) and \( Q_{18} \) together with the 100 kΩ base-diffused emitter resistors operate as a pnp-current mirror and provide a bias current \( I_1 \) for the left-hand branch which equals \( I_2 \). This equality is necessary to compensate for the influence of base currents (Sect. 4.3). The emitters of \( Q_{17} \) and \( Q_{18} \) are not tied together as is usual but are kept at the same potential by the action of the buffer...
amplifier \( Q_{11} - Q_{16} \). In this way it is possible to use \( I_1 \) to bias the output transistor \( Q_{13} \). By connecting the collector of \( Q_{10} \) to the emitter of \( Q_{18} \), full base-current compensation is obtained (assuming equal base currents for \( Q_1 - Q_6 \)). The current through \( R_2 \) via \( Q_8 \) is used as the tail current for the buffer amplifier. The bias current of \( Q_7 \), which is led via \( Q_9 \) through \( Q_{13} \), is almost temperature independent, as calculations can show. This causes the parameter \( m \) in Eq. (5.37) for the circuit to approximately zero. The circuit is started by a current injected into the base of \( Q_9 \) via diodes \( D_1 - D_6 \). Once started, the voltage rise at the base of \( Q_9 \) blocks the diodes \( D_4 - D_6 \), preventing the start-up circuit from influencing the circuit properties. A resistor of 1 kΩ in the negative supply rail protects the circuit in case the output terminals short-circuit.

The circuit has been realized as a hybrid with dimensions 3.5 mm × 14.5 mm (Fig. 5.23). For demonstration purposes, this hybrid has been mounted in a probe (Fig. 5.24), designed for oral measurement. The 'tip' of the probe is of ABS plastic and is plated with a thin gold film to make it easy to clean.

### 5.8.3 Calibration and results

Let us assume that the temperature transducer has a rated range from \( T_1 \) to \( T_2 \). We will call the temperature \( (T_1 + T_2)/2 \) the 'midrange temperature' and the output signal at this temperature the 'midrange signal'. The circuit has to be calibrated for correct values of the midrange signal and of \( dV_o/dT \). As (5.37) shows, the sensitivity \( dV_o/dT \) is proportional to \( R_3/R_2 \). This ratio can be trimmed at ambient temperatures. Next, \( R_4 \) is trimmed for the desired midrange signal. It is not necessary to perform this at the midrange temperature. With (5.37) it is possible to calculate the output signal at any temperature \( (t > 0°C) \), assuming a correct output signal at the midrange temperature. It can be shown that trimming of \( R_3/R_2 \) can be omitted if this ratio deviates less than 2.5 percent from the correct value. In that case only adjustment of \( R_4 \) at ambient temperature remains. Fine adjustment of the output signal can be obtained by trimming \( R^4 \). This slightly affects the output signal in a direction opposite to that caused by trimming \( R_4 \).

In the device of Fig. 5.23 the actual adjustment has been performed by laser trimming the thick-film resistors \( R_1 \) and \( R_4 \). The main specifications of this calibrated device are listed in Table 5.4.

### Table 5.4

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Measured values</th>
<th>Units</th>
</tr>
</thead>
<tbody>
<tr>
<td>Nominal supply voltage</td>
<td>9</td>
<td>V</td>
</tr>
<tr>
<td>Rated performance temperature range</td>
<td>+20 to +50°C</td>
<td></td>
</tr>
<tr>
<td>Power dissipation at 300K</td>
<td>0.25 mW</td>
<td></td>
</tr>
<tr>
<td>Temperature coefficient of the output signal</td>
<td>100 mV/°C</td>
<td></td>
</tr>
<tr>
<td>Extrapolated value of the 'zero'</td>
<td>0°C</td>
<td></td>
</tr>
<tr>
<td>Output impedance</td>
<td>70 Ω</td>
<td></td>
</tr>
<tr>
<td>Absolute error</td>
<td>± 0.1°C</td>
<td></td>
</tr>
<tr>
<td>Error due to supply-voltage variations</td>
<td>0.004°C/V</td>
<td></td>
</tr>
<tr>
<td>Time constant of the hybrid in stirred oil</td>
<td>2 s</td>
<td></td>
</tr>
</tbody>
</table>
5.9 A temperature sensor with binary output, adjustable trip point and trimmable hysteresis

5.9.1 Design considerations and basic circuit

The temperature sensor described in this section delivers a binary output signal as shown in Fig. 5.25. The trip temperature $T_t$ and the hysteresis width $\Delta T$ can be adjusted. This type of temperature sensor is very useful in the design of temperature-control systems. The hysteresis is often desired in view of the stability of the control system where the optimum value of $\Delta T$ depends on the thermal properties of the medium to be controlled and the distance between heater and sensor. The sensor has been designed in such a way that the trip temperature can be controlled with a potentiometer, while the hysteresis width is only trimmable at the chip in order to limit the number of interconnection wires. The trip temperature has a maximum range of $-50^\circ\text{C}$ to $+125^\circ\text{C}$, while the hysteresis amounts to from $0.1^\circ\text{C}$ to $1.5^\circ\text{C}$. A sensor with properties as described above could be made by combining the circuit of Fig. 5.7 with a Schmitt trigger and an output amplifier, where the resistor $R_1$ can be used to adjust the trip temperature. A major disadvantage of such a sensor concerns the influence of the contact resistance of the variable resistor which limits the accuracy and reliability of the device. This drawback has been overcome with the modified sensor circuit whose basic principle is shown in Fig. 5.26. The PTAT current $I_{\text{PTAT}}$ effects a voltage $V_{\text{BE}}(T_{t})$ at the positive input terminal of the sensor amplifier and provides the biasing of $Q_7$ by means of the current mirror. The trip temperature $T_t$ is found from

$$pR_1 I_{\text{PTAT}}(T_t) = V_{\text{BE}}(T_t). \tag{5.38}$$

When the sensor is calibrated for a trip temperature $T_{t,\text{ref}}$ at a reference position $p_{\text{ref}}$ of the potentiometer, it follows from (5.38) that:

$$\frac{p}{p_{\text{ref}}} = \frac{T_{t,\text{ref}}}{V_{\text{BE}}(T_t) V_{\text{BE}}(T_{t,\text{ref}})}. \tag{5.39}$$

The dependence of $T_t$ on $p$ is similar to that shown in Fig. 5.10 when the substitution $p/p_{\text{ref}} = R/R_{\text{ref}}$ is made. To obtain in a simple manner a well-controlled hysteresis in the output signal a small current $nI_{\text{PTAT}}$ is added to $I_{\text{PTAT}}$ when the output signal is 'high' (Fig. 5.27). The hysteresis width $\Delta T$ can be calculated from (5.38) to be

$$pR_1(1+n) I_{\text{PTAT}}(T_t - \Delta T) = V_{\text{BE}}(T_t - \Delta T). \tag{5.40}$$

When as first approximation for $V_{\text{BE}}(T)$ it is written that $V_{\text{BE}}(T) = V_{\text{GO}} - C T$, then with (5.38) and (5.40) it is found that

$$\Delta T = T_t \frac{n(V_{\text{GO}} - CT)}{V_{\text{GO}} - CN T}. \tag{5.41}$$
A more accurate solution is found when for the values found from Eq. (2.13) are subtracted. The result of such a calculation is depicted in Fig. 5.28, which shows the hysteresis width \( \Delta T \) versus the trip temperature for \( V_{BE} = 1172 \text{ mV}; \; n = 3.5; \; n = 0.01; \; V_{BE}(40^\circ \text{C}) = 540 \text{ mV} \) (so \( C = 2.02 \text{ mV/K} \)).

Note that the hysteresis width hardly depends on the trip temperature. In [5.9] it is shown that this is an advantage of the specific configuration of Fig. 5.27 over a number of alternative configurations.

2.9.2 Practical realization

A problem with the circuit of Fig. 5.27 is to generate the small current \( nI_{PTAT} \) without making use of very large-valued (impractical) resistors. The transfer factor \( n \) is very small and amounts to \( 6.7 \times 10^{-4} \) to \( 10^{-2} \) when \( \Delta T \) is from 0.1 K to 1.5 K. The configuration of the so-called 'peaking current source' [5.10] is very suitable for application as a current-gain cell for our purpose because it enables the realization of rather small transfer ratios with pretty low values of the resistors [5.11], [5.9]. When for the input current \( I_{PTAT} \) it holds \( I_{PTAT} = \frac{kT}{qR_2} \ln r \), then for this circuit (Fig. 5.29) a transfer factor \( n \) is found amounting to

\[
n = \exp \left( \frac{R_2}{R_2} \ln r \right),
\]

which does not depend on temperature.

The complete schematic of the temperature sensor has been shown in Fig. 5.30. For the PTAT current source \( (Q_1 - Q_8) \) the configuration discussed in Section 4.3 has been chosen for its accuracy and simplicity. By means of the current mirror \( (Q_9, Q_{10}) \) and current-gain cell \( (Q_{11} - Q_{14}) \) a small bias current \( I_{C14} \) is generated which is used to bias the sensor amplifier \( (Q_{22} - Q_{23}) \). The bias current \( I_{C14} \) has been deliberately chosen small in order to minimize the slide current of the potentiometer \( R_1 \) which equals \( I_{C14}/2I_{FE24} \). A second output transistor \( Q_{15} \) of the current-gain cell has an even smaller output current which is trimmable with \( R_2 \) and is used to provoke hysteresis. Depending on the state of the switch \( (Q_{26}, Q_{27}) \) this current is transferred via current mirror \( (Q_{28}, Q_{30}) \) and gain cell \( (Q_{17}, Q_{18}) \), where finally a small PTAT current \( I_{C17} \) causes a decrease of the current through \( R_1 \) and effects the hysteresis effect in the way discussed in Sect. 5.9.1.

The collector current \( I_{C29} \) is used as input signal for the 'totem-pole' output amplifier \( (Q_{33} - Q_{35}) \). The output stage is capable of delivering sufficient output current to drive a number of TTL input stages. The circuit of Fig. 5.30 has been built by using breadboard components. Preliminary specifications have been listed in Table 5.5.

### Table 5.5

**Performance of the sensor with binary output**

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Measured values</th>
<th>Units</th>
</tr>
</thead>
<tbody>
<tr>
<td>Trip-temperature range</td>
<td>-50 to +125</td>
<td>°C</td>
</tr>
<tr>
<td>Hysteresis width</td>
<td>0.1 to 1.5</td>
<td>°C</td>
</tr>
<tr>
<td>Accuracy</td>
<td>0.1</td>
<td>°C</td>
</tr>
<tr>
<td>Output impedance</td>
<td>1500 in '1' state</td>
<td>Ω</td>
</tr>
<tr>
<td></td>
<td>1000 in '0' state</td>
<td>Ω</td>
</tr>
<tr>
<td>Max. output current</td>
<td>100</td>
<td>μA</td>
</tr>
<tr>
<td>Supply voltage</td>
<td>5</td>
<td>V</td>
</tr>
<tr>
<td>Supply current (no load)</td>
<td>100</td>
<td>μA</td>
</tr>
<tr>
<td>Change in trip temperature due to supply-voltage variations</td>
<td>0.01</td>
<td>°C/V</td>
</tr>
</tbody>
</table>

Using a standard potentiometer, calibration is
easily obtained by trimming the resistor $R_2$ or $R_3$ in the PTAT current source for the desired trip temperature in a given position of the potentiometer.

A disadvantage of the sensor is the relatively high sensitivity to disturbing signals, which is due to the high impedance of the potentiometer. Therefore, a shielded cable is necessary to connect the potentiometer to the sensor. This disadvantage can be overcome when a modified circuit configuration is used. This is still under investigation at the time of writing.
Integrated bandgap references are designed to deliver a reference voltage with a low temperature coefficient and a good long-term stability. The reference voltage is obtained by compensating the base-emitter voltage for its temperature dependence. This accuracy of these references is due to the highly predictable and time-independent way in which the base-emitter voltage $V_{BE}$ of integrated bipolar npn transistors is related to the temperature and to the possibility of generating an accurate compensating voltage, by using the difference $\Delta V_{BE}$ between the base-emitter voltages of two transistors operated at a well-controlled ratio of their emitter-current densities.

Integrated temperature sensors are designed to deliver an electrical output signal which is an accurate measure for the temperature. Just as in the bandgap references in temperature transducers here also the voltages $V_r$ and $\Delta V_{BE}$ are used as the basic signals. There is a lot of similarity between bandgap-reference circuits and temperature-sensor circuits.

To make use of the favorable transistor properties the circuits for amplifying and processing the basic signals have to be carefully designed to avoid having nonidealities spoil the inherent high precision. Even when using a proper circuit design, performance can be deteriorated by thermal- and mechanical-electrical interactions. These effects can be combated by a careful layout design and the use of special technologies, respectively. We will summarize the main conclusions about each of these aspects.

**Components**

The temperature dependence of the $I_C(V_{BE})$ relationship of bipolar transistors can accurately be characterized by two parameters $\eta$ and $V_{BO}$. Knowledge of these parameters is important for the designers of IC bandgap references and temperature transducers. With physically based analysis a good estimation of the values of these parameters can be made. However, at this moment, the required (very high) precision can only be obtained by making measurements on transistors fabricated in the IC process of interest and applying curve-fitting methods. Accurate measurement of $V_{BO}$ and $\eta$ requires high precision of the equipment, set-up and method applied. Some methods for these measurements have been proposed, discussed and compared in this dissertation. A low spread in the $V_{BE}(T)$ characteristics is very important for simple calibration of bandgap references and temperature transducers. For transistors fabricated on a single wafer over the temperature range $-20^\circ C < t < 100^\circ C$ only a small spread has been found. This spread was within the range determined by the temperature measurement accuracy of 0.03 K.

In bandgap references and temperature transducers with intrinsic reference the demands with respect to the precision of the PTAT voltage $\Delta V_{BE}$ are as stringent as for $V_{BE}$ itself. However, as has been shown, the precision of $\Delta V_{BE}$ is less than that of $V_{BE}$ and therefore decisive for the accuracy of the devices.

In this dissertation it is argued that mechanical-electrical interaction is a possible cause of the nonidealities in the PTAT voltage. Further investigations would be necessary to confirm this suspicion and to improve processing should it hold true.

**Interaction with regard to power dissipation**

Power dissipation in the chip effects self-heating which in turn causes an error in IC temperature sensors. The empirical data presented in Chap. 3 enable circuit designers to calculate the magnitude of this effect. Measurement of thermal time constants of unpackaged chips shows that IC temperature sensors can be very fast. However, in practice the package needed for mechanical and chemical protection of the chips will be decisive for the response time. Thermal gradients in the chip can cause serious errors in the bandgap references as well as in temperature transducers. In this dissertation a simple manner to calculate these gradients using empirical or theoretical data has been shown. By properly designing the layout the influence of thermal gradients can be minimized.
Bandgap references

In an initial attempt to realize an accurate bandgap reference a circuit has been designed whose precision is mainly limited by the nonidealties in a pnp current mirror. An improved circuit has been made by using a basically all-npn configuration. In this bandgap reference the first-order temperature dependence of $V_{BE}(T)$ is accurately compensated for. The precision of this circuit was found to be so high that it was considered worthwhile to further improve the accuracy by compensating for the higher-order temperature dependence of $V_{BE}(T)$ as well. This was done in the curvature-corrected bandgap references. With such a reference an extremely low temperature coefficient was achieved. Further by using a unique circuit for supply voltage stabilization an excellent line regulation was also reached.

Although an amazing rate of precision in bandgap references has thus far been achieved, we do not expect that this will curtail the development of newer and better devices. On the contrary, improvement in the stability of the reference voltage after thermal cycling and in the spreading in the optimal value of the output voltage for which the temperature coefficient has its lowest value is desirable. These improvements would require an even higher precision of the PTAT voltage, which can potentially be achieved through a better fabrication process. Furthermore, improvement of the circuit with respect to its sensitivity to disturbing signals would also be desirable.

Temperature transducers

There are several methods for using transistors as temperature sensors with $V_{BE}$ or $\Delta V_{BE}$ as basic signals. To begin with, the base-emitter voltage $V_{BE}$ of a single transistor can be directly used as a measure for the temperature. The intrinsic precision of such a single-transistor sensor is high. However, for an accurate temperature measurement the nonlinearity of the signal has to be accounted for, which requires a complex measurement system. To solve resolution problems the use of a reference voltage is required. However, this increases the system complexity, and the inaccuracy of the reference voltage contributes to that of the system and can easily dominate the sensor's own low inaccuracy. In some special applications, for instance in the detection of small temperature differences, these drawbacks do not play an important role. For these applications single-transistor sensors are very suitable because of their simplicity and high precision. Some of the disadvantages of single-transistor sensors are overcome by PTAT sensors, in which the difference $\Delta V_{BE}$ between the base-emitter voltages of two transistors is used as a measure for their temperature. Some important advantages of these sensors are the linearity of the output signal with regard to temperature changes and the fact that calibration can be performed at the chip, which reduces the overall costs of the measurement system.

The resolution problems of PTAT and single-transistor sensors are solved in an attractive way in the temperature sensors with intrinsic reference. With these sensors a 'zero' is obtained in or close to the temperature range of interest. So, an output signal at °F, °C or an arbitrary scale can be generated. In these sensors both $V_{BE}$ as well as $\Delta V_{BE}$ are used as basic signals, which results in a higher intrinsic sensitivity. The nonlinearity of the sensor signal is two times as low as that of single-transistor sensors, its precision being about two times as high as that of PTAT sensors. An important advantage of the temperature with intrinsic reference transducers is that they can be calibrated at the wafer in a rather simple way. A number of practical implementations of these transducers have been made. In an initial attempt to realize such a transducer, we made a circuit with a current output whose sensitivity and 'zero' can be calibrated by laser trimming two resistors. Some attractive features of this device are the high output resistance, the simplicity of calibration and the possibility to linearize the output signal by applying thin-film resistors with a properly chosen temperature coefficient.

A serious drawback of this transducer is that at least two resistors have to be trimmed to calibrate it, which is basically due to the fact that the sensitivity and the 'zero' of the transducer can be chosen independently of each other. Another disadvantage is that the resistors which are needed to convert the basic voltages $kT/q$ and $V_{g0}$ into currents have to be very accurate. For this reason, base-diffused resistors cannot be applied.
This is a pity because other properties such as the low costs and the good long-term stability make their application attractive.

To reduce the number of calibration steps we designed an easy-to-calibrate temperature transducer whose 'zero' and sensitivity are adjusted in a single step. The 'zero' of this transducer can freely be chosen, while the sensitivity automatically obtains a predetermined value, which depends only on the bandgap voltage of silicon and the 'zero' chosen. The choice of a voltage output enables diffused resistors to be used. The high nonlinear temperature coefficient and large absolute error of these resistors do not deteriorate the circuit performance (as they do in the case of a current output) because, to amplify the basic signals $kT/q$ and $V_{g0}$, only perfect resistor matching is required. The internal power dissipation of this transducer is kept low by stabilizing the supply voltage at a value of only $4 \times V_{BE}$. An interesting application of this temperature transducer is in cold-junction compensation of thermocouple measurement systems. It has been shown that in this application separate calibration of the IC temperature sensor can be entirely omitted.

With biomedical applications in mind, we designed a temperature transducer which in comparison to general-purpose transducers, features a higher accuracy over a relatively small temperature range. This transducer is implemented with an output amplifier delivering a large output signal which allows readout with a simple digital voltmeter. To keep the internal power dissipation small a modified circuit of the temperature transducer with intrinsic reference has been used. In this circuit the transistors are biased at a small current which nevertheless has a large temperature coefficient, which keeps the influence of non-idealities small. Further reduction in the power dissipation has been achieved by interweaving the sensor and buffer amplifier circuitry. This transducer, which is implemented with thick-film resistors, can be calibrated in a single step by laser trimming the resistors.

For application in temperature-control systems a temperature sensor with a binary output, adjustable trip point and trimmable hysteresis has been designed. The trip point of this device is controlled by means of a potentiometer. The hysteresis width can be adjusted by trimming a resistor at the chip.

The totem-pole output amplifier is capable of driving both CMOS and TTL input stages.
APPENDIX A

Measurement of the Temperature Dependence of the $I_C(V_{be})$ Characteristics of Integrated Bipolar Transistors

GERARD C. M. MEIJER and Kees Vingerling

Abstract—The temperature dependence of the $I_C(V_{be})$ relationship of bipolar transistors can be characterized by two parameters $\eta$ and $V_{go}$. In this correspondence a new method for the determination of these parameters is discussed. With this method there is no need for accurate temperature measurements. It is shown that the results fit very well with bandgap-reference temperature characteristics.

An analytical method for the calculation of $V_{go}$ and $\eta$ from values of the base emitter voltage or the bandgap reference voltage at different temperatures is presented.

INTRODUCTION

The relationship between the collector current $I_c$, the base emitter voltage $V_{be}$, and the absolute temperature $T$ can be accurately described at low current levels [1] by the well-known equation

$$I_c = CT^\eta \exp \left( \frac{q}{kT} (V_{be} - V_{go}) \right),$$

(1)

where

$V_{go} =$ the extrapolated bandgap voltage at $T = 0$ K;

$\eta =$ a constant, related to the temperature dependence of the mobility of minority carriers in the base region;

$C =$ a constant.

Knowledge of the parameters $\eta$ and $V_{go}$ is important, especially for the designers of bandgap references and integrated transducers such as temperature transducers [2].

For well-designed bandgap references the spread in $\eta$ is one of the dominating factors limiting the accuracy. Ohte and Yamagata have measured $V_{go}$ and $\eta$ for various types of transistors [3], and found values ranging from 1185 to 1120 mV and 3 to 5, respectively. There is still no satisfactory explanation available for the deviation of these values from theoretical ones [31, 4]. For this reason $V_{go}$ and $\eta$ will be considered in this correspondence as “the best fitting” parameters. $V_{go}$ and $\eta$ can be determined for a fixed collector current [1] from measured values of $V_{be}(T)$. This method involves a long measurement time because, as will be shown, accurate measurement is necessary and therefore perfect temperature matching between the temperature sensor and the device under test has to be ensured. For this reason this method is not very practical for investigating spreads in $\eta$ and $V_{go}$ for a large number of transistors. In this correspondence a more suitable method, in which the temperature is derived from the difference ($\Delta V_{be}$) in the base-emitter voltages of two matched transistors, is presented.

The advantage of this method derives from the excellent temperature matching between the sensor (the base-emitter junctions of a pair of transistors) and the device under test (one of the transistors) by which measuring time can be considerably shortened. The accuracy of this method and the usability of the results for bandgap references will be described.

METHOD

Let us first consider how to determine $V_{go}$ and $\eta$ for a constant collector current $I_c$ from measured values of $V_{be}(T)$. Equation (1) contains three independent parameters: $V_{go}$, $\eta$, and $C$; therefore $V_{be}$ has to be measured at at least three different temperatures $T_1$, $T_2$, and $T_3$. From (1) it can be derived that the following holds:

$$T_2 V_{be}(T_1) - T_1 V_{be}(T_2) = (T_2 - T_1) V_{go} + \frac{k T_1 T_2}{q} \ln \frac{T_2}{T_1},$$

(2)

$$T_3 V_{be}(T_2) - T_2 V_{be}(T_3) = (T_3 - T_2) V_{go} + \frac{k T_2 T_3}{q} \ln \frac{T_3}{T_2},$$

(3)

By substituting measured values of $T_1$, $T_2$, and $T_3$ and $V_{be}(T_1)$, $V_{be}(T_2)$, $V_{be}(T_3)$ with (2) and (3) $V_{go}$ and $\eta$ can be calculated analytically. The accuracy of this method is mainly limited by the measurement accuracy of the temperature. An error $\Delta T$ in the measured value of $T_2$ results in errors $\Delta V_{go}$ and $\Delta \eta$ as large as

$$\Delta V_{go} \approx \Delta T \left( -\frac{2T_1 T_3 (V_{be}(T_1) - V_{be}(T_3))}{(T_2 - T_1)(T_2 - T_3)(T_3 - T_1)} \right),$$

(4)

and

$$\Delta \eta \approx \frac{\Delta V_{go}}{q} \left( \frac{k T_2}{q} \right)^{-1}.$$

(5)

Similarly, the influence of errors in $T_1$ and $T_3$ can be calculated, which appears to be approximately half as large as those due to an error $\Delta T_2$. For typical values of $V_{be}(T)$ (725.27, 627.18, and 526.59 mV for $T = 273$ K, 323 K, and 373 K, respectively) values of $V_{go}$ and $\eta$ as calculated with (2) and (3) are represented in Fig. 1 with the error $\Delta T_2$ as parameter. As can be seen in this figure values of $V_{go}$ and $\eta$ are located on a straight line obeying the equation

$$V_{go} = 1.268 V - \frac{k T_2}{q} \eta.$$

(6)

Note that this equation is very similar to that found by Ohte and Yamagata for the correlation between the spread in $V_{go}$ and $\eta$ for various transistor types [3]. Fig. 1 shows that an error of $\pm 0.05$ K in $T_2$ results in an error of $\pm 0.3$ in $\eta$ and $\pm 78$ mV in $V_{go}$. In our new method the requirement of accurate temperature measurement is avoided by the measurement of the difference $\Delta V_{be}$ of the base-emitter voltages of a pair of transistors on the same chip. Fig. 2 shows the principle of this method. If the transistors $Q_1$ and $Q_2$ are operated at a constant ratio $p$ ($p \neq 1$) of their collector current densities and as--
Fig. 1. Values of $V_{be}$ and $\eta$ calculated with (2) and (3) with the error $\Delta T_2$ in the value of $T_2$ as a parameter.

![Fig. 2. Measurement setup.](image)

Fig. 2. Measurement setup.

Experimental Results

The results for $T_1 = 273$ K, $T_2 = 323$ K, and $T_3 = 373$ K are listed in the table below.

<table>
<thead>
<tr>
<th>$V_{be}$</th>
<th>$\eta$</th>
<th>$T$ determined from</th>
</tr>
</thead>
<tbody>
<tr>
<td>1166 mV</td>
<td>3.72</td>
<td>direct measurement</td>
</tr>
<tr>
<td>1158 mV</td>
<td>3.97</td>
<td>$\Delta V_{be}(T)$ values</td>
</tr>
</tbody>
</table>

It can be concluded that the results of both methods are in good agreement. With direct temperature measurement at intermediate values of $T$ it has been found that there is no significant temperature dependence of the parameters $V_{be}$ and $\eta$. Next, a number of chips fabricated on the same wafer were tested with the indirect temperature measurement. The parameters $V_{be}$ and $\eta$ as calculated from the $V_{be}$ and $\Delta V_{be}$ values at $T_1 \approx 273$ K, $T_2 \approx 323$ K, and $T_3 \approx 373$ K are plotted as crosses in Fig. 3. The collector currents amounted to 100 $\mu$A. The influence of bulk resistances and the small internal heat dissipation depends on the magnitude of the collector current. Their effect has been examined by performing measurements at various current levels. A decrease of the current from 100 to $10 \mu$A leads to a decrease of 10 mV in $V_{be}$ and an increase of 0.35 in $\eta$. These changes are as large as those shown in Fig. 1 for an equivalent error $\Delta T_2 = 0.05$ K in $T_2$. Discussion of these small differences is beyond the scope of this text.

As is shown in Fig. 3, the relation between $V_{be}$ and $\eta$ is linear and can be approximated well by a straight line obeying (6). Because (6) represents the correlation between $V_{be}$ and $\eta$ for inaccuracy of the temperature measurements the suspicion arises that this is related to the nature of the indirect method of temperature measurements.

In order to check this suspicion, the proportionality of $\Delta V_{be}$ with $T$ was tested. When the measurement results of the first experiment mentioned in this section are used, the deviation $\Delta T$ of the $T$ versus $\Delta V_{be}$ characteristic from the ideal one has been calculated and is represented in Fig. 4 for two values of the collector currents. This measurement was accurate within 0.02 $\Delta T$.

It is seen that within the temperature range 270 K < $T$ < 370 K $\Delta T$ is less than 0.2 $\Delta T$. However, this deviation is so large that it can limit the accuracy of temperature transducers [2] and, as will be discussed in the next section, also that of bandgap references. To account for this effect (7) can be modified into

$$I_C = CT^n \exp \left( \frac{q}{kT} (mV_{be} - V_{be}) \right).$$

As has been shown by Gummel and Poon [5], due to base-width modulation by the forward biased e-b junction, the emission coefficient $m$ deviates from unity.

This coefficient depends slightly on temperature and process tolerances. Calculations revealed that by this effect for $m \approx 0.996$ a nonlinearity in $V_{be}(T)$ corresponding to 0.15 K over the temperature range 270 K < $T$ < 370 K can be expected. So the nonidealities shown in Fig. 4 as well as the spread in the results shown in Fig. 3 can be reasonably explained by this effect. It can be calculated that, if $V_{be}$ and $\eta$ are determined from $V_{be}(T)$ with direct measurement of $T$, the influence of the nonunity of $m$ is negligible.

From the foregoing it has to be concluded that by indirect
temperature measurement (from $\Delta V_{be}(T)$) values for $V_{g0}$ and $\eta$ are found which deviate slightly from the actual ones. However, if the intention is to apply the measurement results for bandgap references, this is hardly a disadvantage because, as will be discussed in the next section, $V_{g0}$ and $\eta$ found with the indirect method fit very well with the bandgap-reference temperature characteristic.

The measurements described have also been performed for some devices fabricated elsewhere (Philips breadboard components). The results are very similar to those described above.

**APPLICATION TO BANDGAP REFERENCES**

For the output voltage $V_{ref}(T)$ of a bandgap reference when $\delta V_{ref}/\delta T = 0$ at $T = T_0$, it holds [4]

$$V_{ref}(T) = V_{g0} + \frac{kT}{q} \left( \eta - 1 \right) \left( 1 - \ln \frac{T}{T_0} \right). \quad (10)$$

If $V_{ref}(T)$ is measured at three temperatures, $T_1$, $T_2$, and $T_3$, values for $V_{g0}$ and $\eta$ can be found from:

$$T_2 V_{ref}(T_1) - T_1 V_{ref}(T_2) = V_{g0}(T_2 - T_1)$$

$$+ \frac{kT_1 T_2}{q} \left( \eta - 1 \right) \ln \frac{T_2}{T_1}. \quad (11)$$

$$T_3 V_{ref}(T_2) - T_2 V_{ref}(T_3) = V_{g0}(T_3 - T_2)$$

$$+ \frac{kT_2 T_3}{q} \left( \eta - 1 \right) \ln \frac{T_3}{T_2}. \quad (12)$$

Although these equations are very similar to (2) and (3), there is no need for accurate temperature measurements. This can be seen from (4). When $V_{ref}(T)$ is substituted for $V_{be}(T)$, the quotient $\Delta V_{g0}/\Delta T_2$ is much smaller because of the small temperature dependency of $V_{ref}(T)$.

A number of accurate bandgap references of the type presented in [2] have been tested. From the measurement results $V_{g0}$ and $\eta$ were calculated and are plotted as circles in Fig. 3.

Notice that the values of $V_{g0}$ and $\eta$ fit very well with those found with the new measurement method and can be approximated by (6). This experimental result is understandable, considering that $V_{ref}(T) = V_{be}(T) + C \Delta V_{be}(T)$. Due to the non-linearity of $\Delta V_{be}(T)$ discussed in the preceding section, values for $V_{g0}$ and $\eta$ found from (11) and (12) deviate slightly from the actual ones. Straightforward calculation shows that these deviations are nearly identical to those found with the new measurement method.

This shows that $V_{g0}$ and $\eta$, as found from the $V_{be}$ versus $\Delta V_{be}$ measurements, fit very well with bandgap-reference temperature characteristics. Substituting (6) with $T_2 = T_0$ into (10) gives

$$V_{ref}(T) = 1.268 V - \frac{kT}{q} T_0 \left( \frac{\eta - 1}{\eta} \right) \left( 1 - \ln \frac{T}{T_0} \right). \quad (13)$$

$V_{ref}(T)$ is calculated with (13) for $T_0 = 318.15 K (45^\circ C)$ and is represented in Fig. 5 with $\eta$ as parameter. Notice that due to the correlation between the best fitting parameters $\eta$ and $V_{g0}$ the maximum value of $V_{ref}$ is independent of their spreading.

As will become clear, measurement of the bandgap-reference temperature dependence provides a third method for the determination of $V_{g0}$ and $\eta$. However, a complete bandgap-reference circuit has to be available, while the method discussed in the preceding section can be performed for each single pair of transistors.

**CONCLUSIONS AND DISCUSSION**

The transistor parameters $V_{g0}$ and $\eta$ can be determined from $V_{be}(T)$ values at three different temperatures. An analytical solution for the calculation of $V_{g0}$ and $\eta$ has been presented.

Accurate temperature measurements for finding $V_{g0}$ and $\eta$ can be avoided by the measurement of $V_{be}(T)$ and $\Delta V_{be}(T)$ for a matched pair of transistors. In this case, because of the tight thermal coupling between the emitter junctions of the transistors, measuring time can be considerably shortened.

The values for $V_{g0}$ and $\eta$ found in this way fit very well with the temperature characteristics of bandgap references. The spread in the best-fitting parameters $V_{g0}$ and $\eta$ are correlated and can be described by (6). This spread can be explained by the small nonlinearity in the $\Delta V_{be}(T)$ characteristic. This nonlinearity which also limits the accuracy of integrated temperature transducers is probably due to the small spreading in the emission coefficient $m$, whose influence can be accounted for by the more accurate (9).

Measurement results confirming those presented in previous papers [3] and [4] show a high value of $\eta$ and a corresponding low value of $V_{g0}$. These values, which are also found when $T$ is measured directly, differ considerably from those found by others [1]. Until now no satisfactory explanation has been found.
ACKNOWLEDGMENT

The authors wish to thank Prof. J. Davidse for many helpful and stimulating discussions in the course of this work.

REFERENCES

APPENDIX B ACCURATE DESCRIPTION OF TEMPERATURE EFFECTS IN $I_C(V_{BE})$ CHARACTERISTICS

The temperature dependence of the $I_C(V_{BE})$ characteristics can be described by the equation derived in Section 2.2.2 for the case $I_C = T^r$:

$$V_{BE}(T) = V_{g0} \left( 1 - \frac{T}{T_r} \right) + \frac{T_r}{T} V_{BE}(T_r) - (\eta - m) \frac{kT}{q} \ln \frac{T}{T_r}.$$  \hspace{1cm} (B 1)

From measurements of the energy gap $V_g$ and the mobility $\mu_B$ it is found that $\eta = 2.2$ and $V_{g0} = 1205$ mV \[4.1\]. However, with these values the accuracy of (B 1) is very poor. A large discrepancy between theory and experiment is found \[2.6\]. The accuracy of (B 1) can be improved very much by substituting for the parameters $V_{g0}$ and $\eta$ the empirical values mentioned in Appendix A.

Tsividis \[2.6\] showed that the disagreement between these parameter values is mainly due to the non-linearity of bandgap-voltage variations with temperature changes. To account for this effect he proposes the equations:

$$V_{BE}(T) = V_g(T) - \frac{T}{T_r} V_{g0}(T) + \frac{T_r}{T} V_{BE}(T_r) - (\eta - m) \frac{kT}{q} \ln \frac{T}{T_r},$$  \hspace{1cm} (B 2)

where $V_g(T)$ follows the empirical relation:

$$V_g(T) = a-bT-cT^2,$$  \hspace{1cm} (B 3)

with

$$a = 1.1785V$$
$$b = 9.025 \times 10^{-5} V/K \quad \text{for} \quad 150 \text{K} < T < 300 \text{K};$$
$$c = 3.05 \times 10^{-5} V/(K)^2$$
$$a = 1.20595 V$$
$$b = 2.7325 \times 10^{-6} V/K \quad \text{for} \quad 300 \text{K} < T < 400 \text{K};$$
$$c = 0 V/(K)^2$$

There is no way to accurately calculate the value of $\eta$ in (B 2) from process data; hence for this parameter an empirical value determined from $V_{BE}(T)$ measurements is substituted.

To check and compare the accuracies of (B 1) and (B 2) we performed a number of measurements of $V_{BE}(T)$ on npn transistors fabricated in a conventional IC process (using a 9 $\mu$m thick 1 $\Omega$ cm n-on-p-epitaxial layer, a 2.6 $\mu$m deep 200 $\Omega/sq$ diffusion, a 1.9 $\mu$m deep emitter and a SOT package).

For this experiment a platinum resistor thermometer was mounted together with six different components to be tested in a copper thermal buffer which was placed in a thermostat. With this arrangement the temperature measurement was accurate within 0.025°C. This inaccuracy of the temperature measurement together with a 10 $\mu$V inaccuracy of the voltage measurement result in a total inaccuracy of the $V_{BE}(T)$ measurement amounting to 60 $\mu$V.

We measured the base-emitter voltages for $T = -20, 10, 40, 70$ and 100°C, respectively, and collector currents of 10, 50 and 100 mA, respectively, with a collector-base voltage of 0 V. The measured $V_{BE}(T)$ values have been compared with those calculated with (B 1) and (B 2), respectively. With these calculations we substituted for $V_{BE}(T_r)$ the values measured at $T = 40^\circ C$. Furthermore, with (B 1) the best fitting values $\eta = 3.618$ and $V_{g0} = 1.1707 V$ have been substituted for $\eta$ and $V_{g0}$ while, with (B 2) for $\eta$ the value $\eta = 2.303$, calculated from $V_g(T)$ data at $T = -20^\circ C$ and $40^\circ C$ \[2.6\], has been substituted.

The deviations of calculated values from measured ones are shown in Fig. (B 1) for a typical transistor at two current levels. The advantage of (B 1) is apparent. For the other transistors similar results have been found. At the three levels of bias current for all transistors the errors found with (B 1) were less than 100 $\mu$V. To predict the behavior of accurate bandgap references such a high precision of the equation for $V_{BE}(T)$ is necessary.

It is possible to improve the accuracy of (B 2) by substituting a best fitting value for parameter $a$ in (B 3), but then the advantage of the close relationship to physical considerations is lost.

In conclusion, it can be said that at this moment even the most accurate physically based analysis, contained in (B 2), is not accurate enough to predict the $V_{BE}(T)$ curves with the required precision. This makes it necessary to apply empirical parameters found by curve fitting.

There is need for at least two of these parameters for instance those used in (B 1).

Equation (B 1) is simple and accurate although not unique. Other equations can provide the same accuracy. We preferred (B 1) because of its common
Fig. B 1 Discrepancy between theory and experiment; the calculations have been performed with (B 1) and (B 2), respectively.

use in existing literature. Very good agreement between this equation and measurements has been demonstrated. For a number of transistors and a large range of current levels the errors observed are of the order of measurement inaccuracy amounting to 60 μV. For those who prefer to use (B 2) the following conversion equations have been derived:

\[ n_1 - n_2 = \frac{-c(T-T_R)^2}{k(T-T_R-T\ln\frac{T}{T_R})} = \frac{2cT_R}{k/q}, \]  

\[ V_{g0} - (a+cT_R)^2 = (n_2-n_1) - \frac{kT_R}{q}, \]  

where \( n_1 \) is the value of \( n \) in (B 1) and \( n_2 \) is the value of \( n \) in (B 2)
REFERENCES


[3.4] O. Mueller, 'Der Einfluss des Mitlaufens der Collectorperrschichttemperatur auf das Wechselstromverhalten des Transistors (Mitlaufefeekt) (Internal thermal feedback


Dit proefschrift beschrijft het ontwerp en de eigenschappen van geïntegreerde spanningsreferentiebronnen en temperatuurtransducenten. Na een uiteenzetting over de toepassing, de eigenschappen en de basisprincipes van deze produkten volgt een behandeling van de eigenschappen van de gebruikte componenten. De hoge precisie van de schakelingen is vooral te danken aan de hoge mate van voorspelbaarheid waarmee de basis-emitter spanning van bipolaire transistoren van de temperatuur afhangt en de mogelijkheid om met deze transistoren nauwkeurige spanningen met een tegenovergestelde temperatuurcoëfficiënt op te wekken. Uitgebreid wordt behandeld hoe het thermisch gedrag van basis-emitterspanningen het best gekarakteriseerd kan worden en hoe nauwkeurig deze beschrijving is. Meetmethoden en meetresultaten worden hierbij uitvoerig besproken.

Een aantal verschillende bandgap-spanningsreferenties zijn ontworpen en vervaardigd als IC. De meest geavanceerde schakeling levert bij optimale afregeling een uitgangsspanning van ongeveer 1.2 V met een temperatuurafhankelijkheid van slechts 50 ppm over het temperatuurbereik van -20°C tot +100°C. Dezezelfde fundamentele transistoreigenschappen waarop bandgap-referenties hun nauwkeurigheid te danken hebben maken de realisatie van nauwkeurige temperatuursensoren mogelijk. De benodigde schakelingen vertonen veel overeenkomst met die van bandgap-referenties. Er blijken verschillende manieren te zijn om transistoren als temperatuursensor te gebruiken. De voor- en nadelen van deze methoden worden besproken. De meest belovende methode is die welke toegepast wordt in de z.g. temperatuursensoren met intrinsieke referentie. Met deze sensoren wordt een uitgangssignaal verkregen op basis van een °C, °F of andere gunstig te kiezen schaal.

Opmerkelijk is dat de gevoeligheid en het nulpunt in een enkele stap gecalibreerd kunnen worden, hetgeen grote voordelen biedt bij het fabriceren van grote aantallen nauwkeurige goedkope temperatuursensoren. Een aantal transducenten, waarin van deze methode gebruik wordt gemaakt, zijn als IC gerealiseerd. De meest opmerkelijke uitvoering heeft als belangrijkste kenmerken: eenvoudige calibratie (in één enkele stap), een hoge gevoeligheid over een groot temperatuurbereik (-4,6 mV/°C; van -50°C tot +125°C), nulpunt vrij te kiezen, lage uitgangsimpedantie (40 Ω) en lage interne dissipatie (200 μW). Deze transducent is vervaardigd in een conventioneel IC-proces en gecalibreerd met behulp van de zener-zap methode.
About the author

Gerard C.M. Meijer was born in Wateringen, The Netherlands, on June 28, 1945. He received the 'ingenieurs' (M.S.) degree in electrical engineering from the Delft University of Technology, The Netherlands, in 1972.

In the course of his study at the university he investigated and improved ESR spectrometers with regard to the influence of FM noise.

Since 1972 he has been part of the Laboratory of Electronics at the Delft University of Technology, where he is engaged in research and teaching on analog IC's.
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