Smart voltage measurement systems and bandgap references

Gerard C.M. Meijer,
Delft University of Technology, Department of Electrical Engineering
Mekelweg 4, 2628 CD Delft, (Netherlands)

August 21, 1995

Abstract

This paper gives a review of the concepts and basic problems of voltage measurement systems and bandgap references. It deals with the principles of a continuous autocalibration technique. By applying this technique all of the linear errors of the processing circuits are eliminated. This enables the use of high-offset amplifiers, such as CMOS amplifiers, to implement the circuits. The basic signals are converted to the time domain and decoded by a microcontroller. The microcontroller performs the processing of all of the basic signal components, including the composing voltages V_{BE} and V_{PTAT} of bandgap voltage references. The paper discusses the temperature dependence of the $I_C(V_{BE})$ characteristics of bipolar transistors, and methods to compensate for this temperature dependence for application in accurate bandgap references. Methods and technologies to calibrate bandgap references are discussed. As case studies, two types of bandgap references are discussed. One of them is implemented in bipolar technology, the other one in CMOS technology. A novel type of bandgap reference and voltage processor, the so-called dynamic voltage processor, is also presented. This voltage processor is particularly useful for CMOS measurement systems, where it can improve the accuracy of the system. Some basic problems for the accuracy of bandgap references are discussed.

1 Introduction

A reference signal S_{ref} is often used to determine the magnitude of an unknown signal S_x, by measuring the ratio S_x / S_{ref}. For this application the reference signal should be of the same type as the unknown signal. For instance, reference resistors are used to measure unknown resistors, reference resonators are used to measure frequencies and reference voltages are used to measure voltages.

As an example, Fig.1 shows a typical measurement system for voltages: An unknown voltage V_x is measured by comparing its value to that of a reference voltage V_{ref}. The processing circuit generates an output signal S_{out}, which is linearly related to the input voltage. In the three positions 1,2 and 3 of the multiplexer switch it holds that, respectively,

$$S_{out,x} = aV_x + b \tag{1}$$

$$S_{out,ref} = aV_{ref} + b \tag{2}$$

$$S_{out,0} = b \tag{3}$$

After measurement of these three output signals the unknown voltage V_x can be calculated from the equation:

$$V_x = V_{ref} \frac{S_{out,x} - S_{out,0}}{S_{out,ref} - S_{out,0}} \tag{4}$$

From equation (4) it can be concluded that the transfer parameters a and b of the processing circuit do not affect the measurement result, provided that these parameters are constant during the measurement and that the system is really linear. However, inaccuracy of the reference voltage directly limits the accuracy of the measurement. Therefore voltage references should have a low temperature coefficient and an exellent long and short- term stability. They also have to be insensitive to thermal and mechanical shocks and, in general, insensitive to changes of any external or internal physical condition.

Another application of references is the generation and control of signals. As an example Fig.2(a) shows a voltage regulator whose output voltage V_0 is directly related to a reference voltage V_{ref}, according to the equation

$$V_0 = V_{ref}(1 + \frac{R_2}{R_1}) \tag{5}$$

Figure 2(b) shows a voltage-to-current converter and a current copier (current mirror). Neglecting the non-idealities of the applied components it is found that

$$I_0 = I_0' = I_0'' = \frac{V_{ref}}{R_s} \tag{6}$$

Circuits of the type shown in Fig.2 can be used to generate the currents and voltages which are needed to bias electronic and other non-linear electrical components.

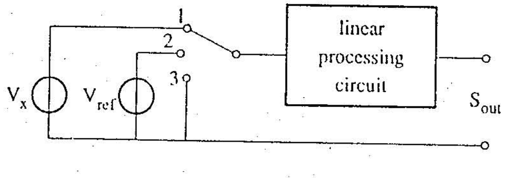

Figure 1: *A system setup for the measurement of an unknown voltage V_x.*

This paper deals with smart voltage measurement systems and bandgap voltage references. In smart measurement systems a microcontroller or computer makes part of an overall design. We will show that the memory and calculation capability of the microcontroller can help to solve some traditional problems of CMOS bandgap references.

In bandgap references the reference voltage is generated by adding a correction voltage $V_C(T)$ to the base-emitter voltage $V_{BE}(T)$ of a bipolar transistor (the reference transistor)

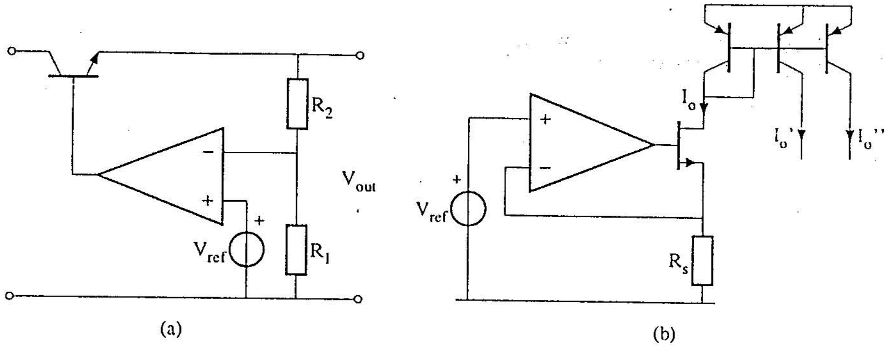

Figure 2: *Voltage references applied to control voltages and currents (a) a voltage regulator (b) a current source and a current copier.*

in order to cancel out its temperature dependence.
Since the introduction of this principle in 1964 and 1971 by, respectively, Hilbiber [1] and Widlar [2] 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 b[illegible]ences concern their low supply voltage and low power dissipation. A suprisi[illegible]cision can be achieved with bandgap references. The temperature coefficient [illegible]ase-emitter voltage is rather high: circa 3000 ppm/°C. The TC of the outpu[illegible] the best bandgap references is less than 2 ppm /°C. This means that the r[illegible]C 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 ICs. 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 behaviour which is accurate enough to be used for this purpose is the difference ΔV_{BE} of the base-emitter voltages of a matched pair of transistors operated at unequal collector-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. Such a voltage is suited for compensation of the first-order temperature dependence of V_{BE}. To compensate for the higher-order temperature dependence the ratio p can be made temperature dependent.
In this paper we will give a review of the basic principles and problems of bandgap references, starting with a discussion of the current and voltage dependence of base-emitter voltages. Various methods to compensate for the first-and higher-order temperature coefficients will be presented. After a discussion of calibration methods and techniques, some typical circuits implemented in bipolar and CMOS technology will be discussed. Some of the typical problems of CMOS bandgap references can be solved by the novel dynamic

bandgap references, presented in this paper. Some major non-idealities which limit the accuracy of bandgap references are discussed briefly.

2 The $V_{BE}(I_C, T)$ characteristics

The main properties of the forward-biased bipolar (npn) transistors applied for bandgap references are revealed by the following equation[3], [5]:

$$I_C = I_S \exp\left(\frac{qV_{BE}}{kT}\right) = CT^{\eta}\exp\frac{(qV_{BE} - V_{g0})}{kT}. \tag{7}$$

where

I_S	=	the saturation current ($I_S = I_S(T)$)
T	=	the absolute temperature,
V_{BE}	=	the base-emitter voltage,
q	=	the electron charge,
k	=	the Boltzmann constant,
V_{g0}	=	the extrapolated bandgap voltage of silicon for $T = 0$ K,
η	=	a semi-empirical parameter which expresses the nonlinearity of the $V_{BE}(T)$ characteristic.

The third pa[illegible] ation is originally found as an approximation of the physical characteristic [illegible] shown that with empirical constants this equation is suited to describe th[illegible] racteristic with a precision which is good enough for practical applications. [illegible] ound that the values for V_{g0} and η satisfy the empirical equation

$$V_{g0} = 1268 \text{ mV} - \eta\frac{kT_r}{q}, \tag{8}$$

where $T_r = 323$ K.

For practical reasons the collector current is made often proportional to some power of T.

$$I_C \propto T^m \tag{9}$$

As will become clear in the following sections 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.3). We obtain from (7) and (9) that:

$$V_{BE}(T) = \{V_{go} + (\eta - m)\frac{kT_r}{q}\} - \lambda T + (\eta - m)\frac{k}{q}(T - T_r - T\ln\frac{T}{T_r}), \tag{10}$$

where

$$\lambda = \frac{V_{g0} + \frac{kT_r}{q}(\eta - m) - V_{BE}(T_r)}{T_r}. \tag{11}$$

To obtain an impression of the magnitude of the different terms of (10) and (11) we substitute $V_{g0} = 1156$ mV, $\eta = 4$, and for example $T_r = 300$ K and $V_{BE}(T_r) = 700$ mV; while for the collector current the following three practical cases are considered:

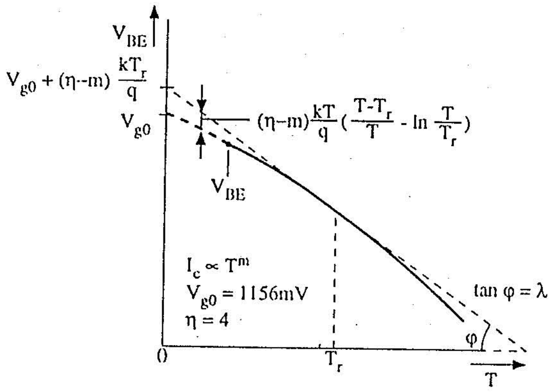

Figure 3: *The base-emitter voltage V_{BE} versus the temperature T. The curvature is exaggerated in order to indicate the characteristic points clearly.*

- $I_C = V_{PTAT}/$
- $I_C = V_{ref}/R$
- $I_C = V_{BE}/R$

To get an impression of the effects of the various parameters we suppose for a moment that R is constant. For this simplified case Fig. 4(a) shows the extrapolated $V_{BE}(T)$ curves over the full 0 - 500 K range. However, the reader should be aware of the fact that the mathematical models used in this paper only represent the physical behaviour over a limited temperature range of about -50°C to + 150°C. The tangents to the three curves at $T = 300$ K intersect the vertical axis (at $T = 0$ K) at, respectively, 1.234 V, 1.260 V and 1.286 V. So that the slope $\lambda = -\mathrm{d}\ V_{BE}/\mathrm{d}T$ of the tangents respectively amount to 1.780 mV/K, 1.867 mV/K and 1.953 mV/K.

To get a better impression of the curvature of the $V_{BE}(T)$ characteristics we also plotted the difference of the three curves with respect to their tangents (Fig.4(b)). From this figure it can be concluded that the curvature of the characteristic is significantly affected by the temperature dependence of the collector current I_C. This property is used in many of the so-called curvature-corrected bandgap references (see Section 3). When the resistor R is also temperature dependent, then this will affect the curvature. For resistors with a negative temperature coefficient the curvature will be smaller and for resistors with a positive temperature coefficient the curvature will be larger. Moreover, the temperature dependence of the resistors also slightly effects the optimal value for the output voltage of a bandgap reference. However, it does not effect the methods and principles discussed in this paper. Therefore, for simplicity reasons, in the rest of this paper, we will R consider as to be constant.

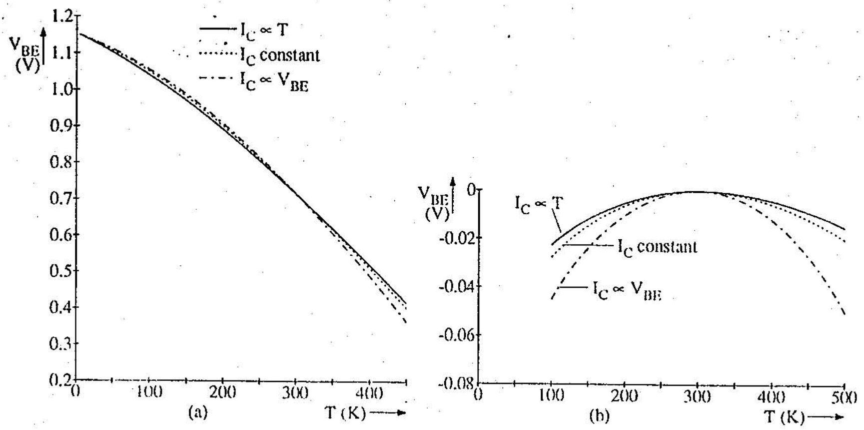

Figure 4: *The effect of the temperature dependence of the collector current on the $V_{BE}(T)$ characteristics (a) The extrapolated $V_{BE}(T)$ characteristics, (b) the non-linearity of the $V_{BE}(T)$ characteristics.*

3 Compensation of the temperature dependence of the base-emitter voltage

3.1 Principles

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. The compensating voltage is in the same order of magnitude as the base-emitter voltage itself. Therefore, the demands with respect to accuracy, reliability and temperature range are as stringent for both of the voltages V_{BE} and V_{PTAT}. Generally, calibration of the voltage references is necessary. As economic considerations also play role in the choise of compensation technique, the simplicity of calibration has to be taken into account. With respect to these requirements very good results are obtained 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 difference $\Delta V_{BE} = (kT/q)\ln p$ of the base-emitter voltages of two transistors operated at unequal collector-current densities with ratio p. In this way an output voltage V_{ref} is obtained for which it holds that:

$$V_{ref} = V_{BE}(T) + V_C(T) = V_{BE}(T) + A \triangle V_{BE}(T), \tag{12}$$

where A denotes an amplification factor. A simple implementation is shown in Fig. 5. The action of the current mirror (Q_3, Q_4) ensures that the ratio of the collector currents of Q_1 and Q_2 remains constant. Neglecting the influence of the base currents and base-width modulation, for the voltage $\triangle V_{BE}$ across R_2 we find

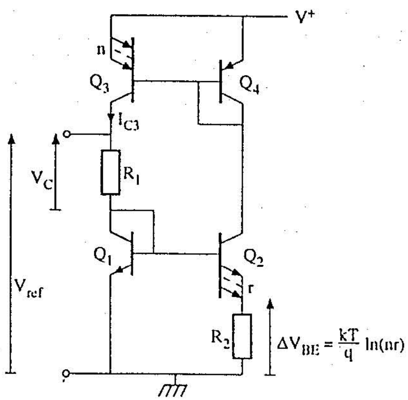

Figure 5: *A simple bandgap-reference circuit.*

$$\Delta V_{BE} = (kT/q) \ln (nr), \tag{13}$$

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

$$I_{C3} = nI_{C2} = n\frac{kT}{qR_2}\ln(nr). \tag{14}$$

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

$$V_{ref} = V_{BE1} + V_c = V_{BE1} + n\frac{R_1}{R_2}\frac{kT}{q}\ln(nr). \tag{15}$$

We assume that the resistors are temperature independent so that I_{C3} is PTAT. The base-emitter voltage V_{BE1} of Q_1, biased with a PTAT current, is given by Eq. (10) with $m = 1$.

With proper calibration the PTAT correction voltage represented by the second term in the right-hand side of (15) compensates for the PTAT component λT in V_{BE1}, resulting in an output voltage V_{ref}:

$$V_{ref} = V_{g0} + (\eta - 1)\frac{kT_r}{q} + (\eta - 1)\frac{k}{q}(T - T_r - T\ln\frac{T}{T_r}). \tag{16}$$

The non-linear 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(b).

The accuracy of the simple circuit of Fig. 5 is poor because of the influence 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. Further, the output voltage will depend on the load current and supply voltage.

Another problem is that circuits of the type of Fig. 5 have a stable state at zero current flow even when the power supply is non-zero. A separate start-up circuit is required to prevent the circuit from remaining in this undesired state. These problems can easily be overcome, as has been shown in many well-designed circuits presented over the last two decades [2],[7]-[14]. Many of these papers also deal with the so-called curvature correction, which is a method for compensating for the non-linearity of $V_{BE}(T)$ curve. This compensation technique will be discussed in the next section.

3.2 Curvature correction

There are many ways to compensate for the non-linearity of the $V_{BE}(T)$ characteristics. In [12] Song and Gray present a method in which a correction voltage proportional to the square of the absolute temperature ($PTAT^2$) is generated. In [13] Lee et al present a method which exploits the temperature dependence of the current-gain factor β of a bipolar transistor. In [9] Widlar presents a method which exploits the difference in the curvature of the $V_{BE}(T)$ characteristics for transistors biased at a constant and a PTAT current, respectively (Fig.4(b)). The latter method has been applied in many of the curvature-corrected bandgap references presented in various papers [10],[11] and [14]. An attractive feature of this method is the direct relation of the correction voltage to the basic signals V_{BE} and V_{PTAT}. This makes it easy to compensate in the same time for both the first-order and the higher-order temperature dependence of V_{BE}, without using additional components. The principle of the method is shown in Fig.6: The curvature of the $V_{BE2}(T)$ curve for a transistor biased at a constant current ($m = 0$) is larger than the curvature of $V_{BE1}(T)$ of a similar transistor biased at a PTAT current ($m = 1$), where the ratio of the corresponding non-linearities is about a factor $\eta/(\eta - 1)$. This ratio is independent of the current levels. To achieve full curvature correction for $\eta = 3.5$ the base-emitter voltage V_{BE2} is multiplied by a factor $(\eta - 1)/\eta \simeq 0.7143$ (Fig.6). This factor is independent of the applied current densities and current density ratio [15], so that it is possible to obtain first-order as well as higher-order temperature compensation (Fig.6) [9], [10] and [14].

In [10] it is shown that the same principle can be applied using series-connected base-emitter voltages. A benefit of this method is that the large magnitude of the basic signals improves the accuracy. On the other hand for such a circuit a relatively large supply voltage is required. The circuits presented in [9],[11] and [14] concern low-power, low-voltage implementation where an output voltage in the order of a few hundred mV is generated. In [15] it is shown that even with a 1 V supply voltage there are many degrees of freedom which can be used to optimize, for instance, the noise performance, the minimum supply voltage or the supply current.

4 Calibration methods and techniques

To minimize the temperature coefficient of a bandgap reference it is necessary to calibrate the device. Without calibration the spread in V_{BE} and V_{PTAT} will cause a corresponding spread in both V_{ref} and its temperature coefficient.

Example 1: The consequences of a deviation in V_{BE} for the reference voltage $V_{ref} = V_{BE} + AV_{PTAT}$ are shown in Fig.7. A worst-case deviation in V_{BE} and V_{PTAT} of 20 mV and 0.5 mV, respectively, and an amplification factor A =

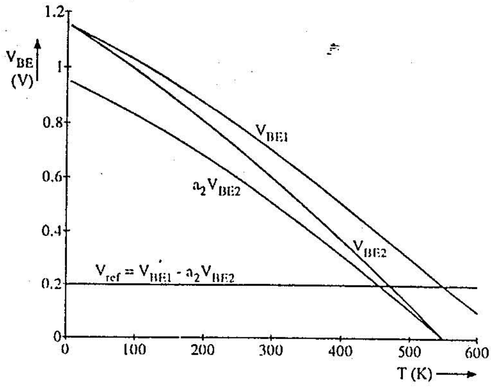

Figure 6: *Principle of making a curvature-corrected reference voltage V_{ref} with a linear combination of two base-emitter voltages.*

15 result in a worst-case deviation of the bandgap-reference voltage of about 27.5 mV and a worst-case temperature coefficient of 74 ppm/K. To reduce this large deviation, calibration is required.

For simplicity we discuss the calibration methods using linearly extrapolated approximations of the temperature characteristic. To trim a linear (temperature) characteristic, two points have to be fixed. Fortunately in the case of bandgap references, one point is already fixed by physical constants: at 0 K the extrapolated curves of V_{PTAT}, V_{BE} and V_{ref} intersect the vertical axis at the values of 0, V_{BE0} and V_{BE0}, respectively, where V_{BE0} is the linearly extrapolated value of V_{BE} at $T = 0$ K (Fig. 7). In the case of a scaled bandgap reference the output voltage amounts to

$$V_{ref} = A_2(V_{BE} + A_1 \triangle V_{BE}). \tag{17}$$

In this case the extrapolated reference voltage at $T = 0$ K is fixed within certain limits, to the value $A_2 V_{BE0}$. The factor A_2 can be realized using well-matched resistors. The worst-case deviation of such a resistor cluster depends on the geometry (the matching improves with larger chip areas). Typical values for the mismatch are in the order of 0.2%. Exploitation of the above-mentioned property enables a single step trimming at an arbitrary temperature. The trimming has to be performed in such a way that the V_{ref} curve rotates around the "high point" at 0 K. For an unscaled bandgap reference, according to Eq.(13), an output voltage $V_{ref} = V_{BE0}$ at a certain temperature automatically involves a zero temperature coefficient. When designing the trim procedure care has to be taken to maintain the advantages of this remarkable property.

Example 2: When for calibration purposes the constant A_2 of Eq.(17) would be changed, then the position of the fixed point at the 0 K axis would also be changed and the attractive possibility of a one-step trimming would be lost. Good ways to trim the scaled bandgap references of Eq.(17) are to change the

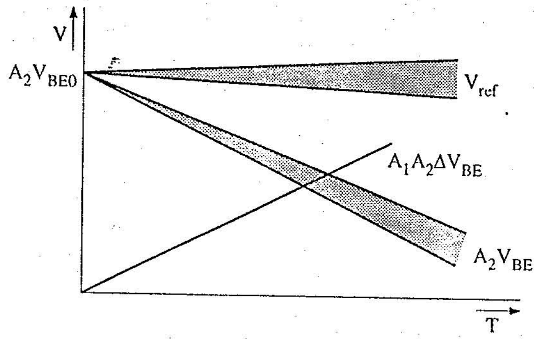

Figure 7: *The effect of spreading in V_{BE} at the output voltage V_{ref} of an uncalibrated bandgap reference, (the curves are shown as their linear extrapolated approximations.*

amplification factor A_1 or to change V_{BE} (not A_2!) by changing the current density of the transistor.

Example 3: When a bandgap reference has been calibrated at 300 K for the nominal output voltage $A_2 V_{BE0}$, then a deviation in A_2 of 0.2% (2000 ppm) will cause a temperature coefficient of 2000ppm/300 K $\simeq$ 7 ppm/K. Therefore, for very accurate bandgap references it is better to use unscaled bandgap references, where $A_2 = 1$ or another integer.

Various trimming techniques can be applied. In common use are [3]:

- fusible links, to blow-up connections,
- zener zapping, to short-circuit connections,
- laser-trimming, to adjust resistors

5 Practical Realizations

Since the introduction of the first integrated bandgap reference, by Widlar, many high-performance circuits have been developed. The first improvements of the early circuits concerned:

- the use of cascoding transistors or internal regulators to eliminate the effect of changes of the collector-voltage of the supply-voltage,
- the use of compensation methods to reduce the effect of base currents.

Other improvements concerned the elimination of the need for low-performance components, such as:

- the early lateral pnp-transistors,

- the (voltage-dependent) diffusion resistors.

As an example Fig. 8(a) shows an "all-npn" PTAT current source [4]. The output voltage V_{ref} of this circuit is insensitive to the performance of the pnp current mirror. Figure 8(b) shows an alternative way to get "all-npn" performance. In modern IC processes the properties of the applied components can significantly differ from those of earlier processes. This change can partly explain the continuous stream of new modified circuits. For another part new or modified circuits are developed to meet special requirements, such as the ability to operate at low supply voltages or at very high temperatures, or to withstand large voltage transients [16].

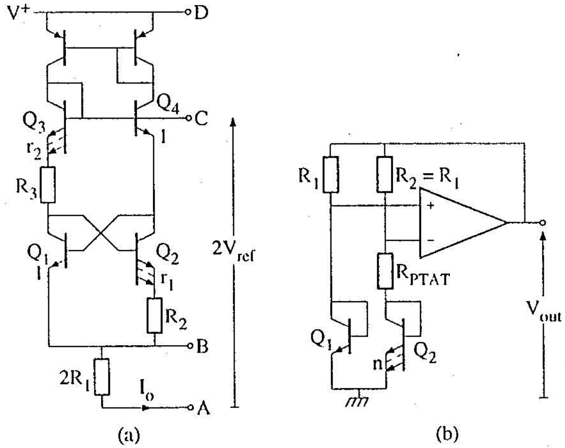

Figure 8: *(a) Simple implementation of a bandgap reference, where accuracy is hardly affected by the pnp transistors (b) an alternative way to make an all-npn bandgap reference.*

5.1 Bipolar implementations

Figure 9 shows the basic principle of the low-voltage curvature-corrected bandgap reference of Widlar [9]. For the output voltage V_{ref} it holds that

$$V_{ref} = V_{BE72} - V_{BE71} + I_{Vbe}R_{62} - (I_{const71} + I_{PTAT})R_{61} \simeq V_{PTAT} + AV_{BE72} \qquad (18)$$

A more recent version of this type of circuit, designed by van Staveren et al[14] is shown in Fig. 10. A simple start-up circuit is formed by the transistors Q_{start1}, Q_{start2} and the pinch resistor R_{start}. Start-up circuits are required to guarantee that the circuit is properly biased. Usually, self-biasing current sources, such as PTAT sources, do have more than one stable state of biasing [17].
Start-up and anti-latch-up circuits [17] are required to eliminate the undesired states. Simple start-up circuits as shown in Fig. 10 are suited for a limited range of the supply voltage. At large values of the supply voltage the start-up current can affect the accuracy of the circuit. Figure 11 shows an alternative start-up circuit where a current I_{st}, via D_{st},

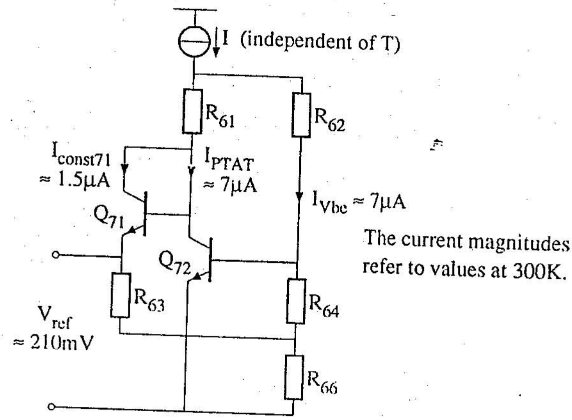

Figure 9: *The basic principle of Widlar's curvature-corrected bandgap reference.*

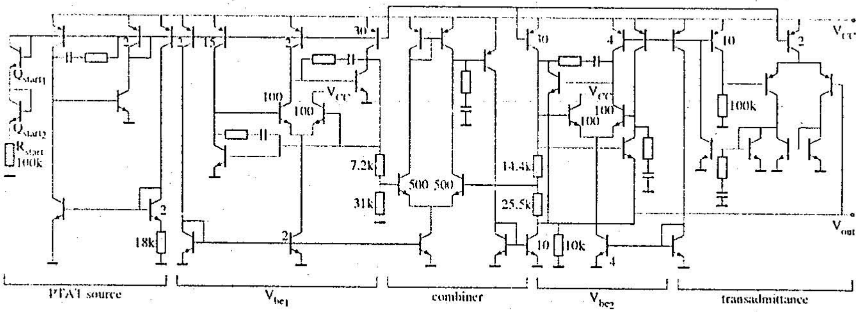

Figure 10: *A complete version of a low-voltage curvature-corrected bandgap reference [14].*

is injected in the PTAT source in the starting phase only. Sometimes dynamic starters are applied. With this type of circuits, directly after power switch-on, a capacitor charge is injected into the current source to start-up the circuit. However, the use of this type of starter has to be discouraged, because proper start-up cannot be guaranteed in worst-case circumstances. Typical specifications of the circuit of Fig. 10 have been listed in Table 1.

TABLE 1: Typical performance of the circuit of Fig. 10.

Minimum supply voltage	1 V
Supply current at 27°C	100 μA
Output voltage	194.2 mV
Temperature dependence over the 0°C - 100°C range	150 ppm/100°C
Output impedance	40 Ω
Noise	< 80 nV/$\sqrt{\text{Hz}}$
Line regulation	-80 dB
Chip size	3 mm^2

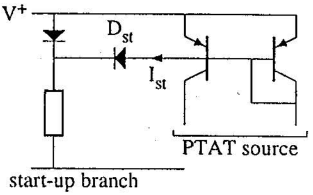

Figure 11: *A simple start-up circuit which only affects the PTAT current source during starting.*

5.2 CMOS Implementations

The principles of CMOS bandgap references are similar to those of bipolar ones. At least one bipolar transistor is required to make a bandgap reference [18]. For this application lateral (Fig. 12) or substrate transistors can be used. In the lateral pnp transistor shown in Fig. 12 a gate is used to obtain a thin oxide layer which makes it easy to etch the holes for the emitter and collector diffusions. As compared to the lateral pnp-transistors made in a bipolar process, the CMOS versions do have two main drawbacks:

- There is no buried layer, which gives rise to a relatively large substrate current I_{sub}.
- The $I_C(V_{BE})$ characteristic deviates from the idealized exponential one. This non-ideality has to be expected especially in transistors made in an n-well process; because of the low surface doping. Consequently, high-level effects occur even at rather low current levels.

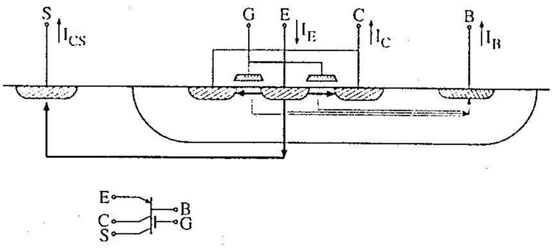

Figure 12: *A cross section of a lateral pnp-transistor made in an n-well CMOS process.*

As an example, Fig. 13 shows the $I_C(V_{BE})$ characteristics for a lateral pnp-transistor made in a 1.2 μm CMOS process of Mietec. In a 0.7 μm process the non-ideality of the characteristics are even worse [19]. The (vertical) substrate transistors show a much better performance with respect to the ideality of the $I_C(V_{BE}, T)$ characteristics. Therefore, this type of transistors is preferred for generating the V_{BE} and V_{PTAT} voltages in bandgap references. Because all of the collectors of the substrate transistors are connected to the common substrate, special amplifier configurations are required to amplify the PTAT voltage. Figure 14(a) shows a basic configuration for such a bandgap reference. The generated output voltage amounts to

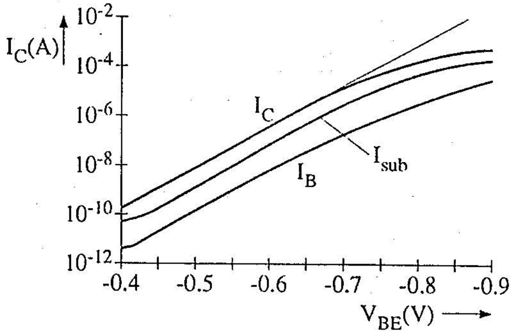

Figure 13: *The $I(V_{BE})$ charateristics of a lateral pnp transistor fabricated in an n-well CMOS process (courtesy of Mietec).*

$$V_{out} \simeq V_{EB2} + (1 + \frac{R_2}{R_{PTAT}})(V_{PTAT} + V_{OS}), \tag{19}$$

where $V_{PTAT} = (kT/q)\ln n = (V_{EB1} - V_{EB2})$ and V_{OS} is the offset voltage of the amplifier A. In the configuration of Fig. 15(a) the current density ratio of Q_1 and Q_2 equals their emitter-area ratio. To obtain an accurate CMOS bandgap reference it is important to minimize the amplifier offset voltage V_{OS}. When the supply voltage is sufficiently high, the relative influence of the offset can be reduced by stacking the base-emitter voltages of several substrate transistors [16]. For lower supply voltages other solutions have to be found.

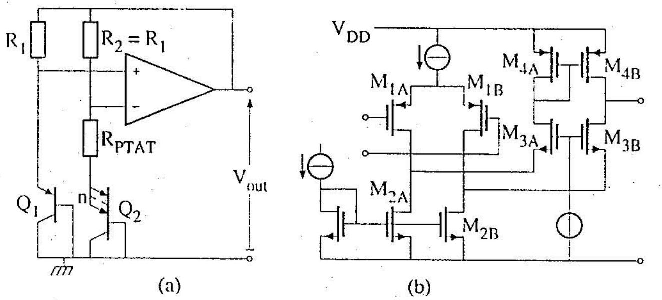

Figure 14: *Principles of CMOS bandgap-reference circuits (a) a basic circuit configuration, (b) setup of the amplifier-input stage.*

Figure 14(b) shows the input stage of a CMOS amplifier recently designed by Mietec. The input transistors M_{1A} and M_{1B} are buffered by the folded-cascode transistors M_{3A} and M_{3B}, which are biased with the current sources M_{2A} and M_{2B}. The main causes for the offset voltages of this input stage are the V_{GS} mismatch of the input transistors and the drain-current mismatch of the current sources M_{2A}, M_{2B} and the current mirror M_{4A}, M_{4B}. In both CMOS and bipolar input stages the offset can be minimized by using quad transistor configurations and large geometries. In addition to these general methods, the offset of CMOS input stages can be further reduced by optimizing the current density [20].

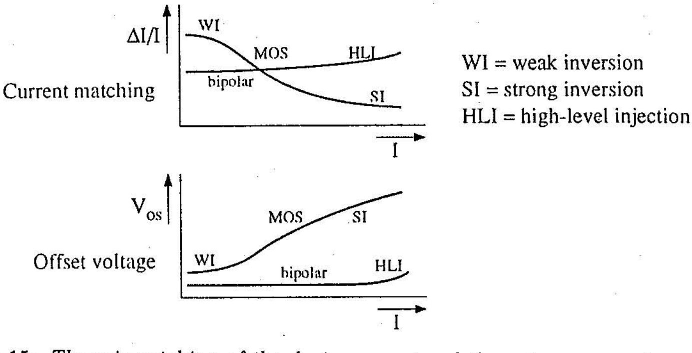

Figure 15: *The mismatching of the drain current and the gate-source voltage for bipolar and CMOS transistors as a function of the biasing current.*

Figure 15 shows typical characteristics of the mismatching versus the current density for bipolar as well as CMOS transistors [21]. To minimize the V_{GS} mismatching, it is advantageous to minimize the current denisity. Therefore the input stage of Fig.14(b) has been designed for the transistor M_{1A} and M_{1B} to operate in weak inversion region. On the other hand, to minimize the effect of current mismatching the transistor pairs M_2 and M_4 are operated in strong inversion, where $(V_{GS} - V_T)$ is as large as possible.

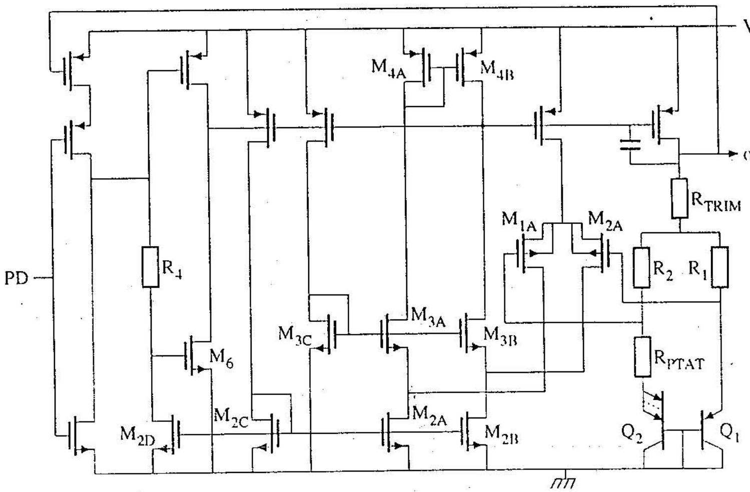

Figure 16: *A complete circuit diagram of a CMOS bandgap reference. (Courtesy of Mietec).*

Figure 16 shows a complete circuit diagram of the CMOS bandgap reference. The circuit is self-starting and is provided with a power-down (PD) function. Four digital inputs (not shown in the figure) allow the output voltage to be trimmed over a range of +/- 45 mV. The main specifications of the circuit are listed in Table 2.

Table 2 The main specifications of the circuit of Fig. 16.

Parameter	min	typ	max	units
Supply voltage	2.7		5.5	V
Supply current at 25°C		17	23	μA
Output voltage	1.18	1.20	1.23	V
Output current	-2		10	μA
Line regulation	60			dB
Temperature dependence over the -50°C to +150°C range		4000		ppm/200°

6 Dynamic bandgap references for voltage processors

For some applications there is an interesting new method for solving the problem of amplifier offset discussed in Section 5.2. This method, which is based on classical principles, can be used when the bandgap reference is a part of a measurement system for voltages (Fig. 1). In this particular case the basic voltages V_x, V_{BE}, V_{PTAT} and V_{OS} can be directly converted to, for instance, the time domain. In the system setup of Fig.17(a) the voltage V_x, V_{BE}, V_{PTAT} and V_{OS} are converted to the time domain by a linear voltage-to-period convertor. When the output periods during the successive measurements are T_x, T_{BE}, T_{PTAT} and T_{OS}, respectively, then the final measurement result M_3 amounts to

$$M_3 = \frac{T_x - T_{OS}}{T_{BE} + AT_{PTAT} - (A+1)T_{OS}} = \frac{V_x}{V_{ref}} \tag{20}$$

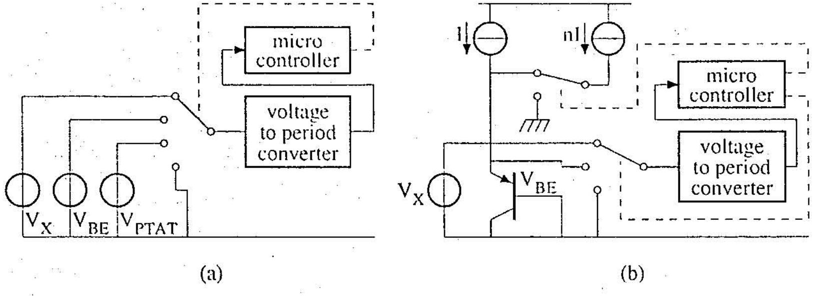

Figure 17: *Two different measurement systems implemented with dynamic bandgap references.*

In the alternative measurement setup of Fig. 17(b) resp. V_x, V_{BE1}, V_{BE2} and V_{OS} are converted to the time domain. The voltages V_{BE1} and V_{BE2} are obtained with biasing currents $(1+n)I$ and I, respectively. The systems shown in Fig. 17 are continuously autocalibrated for gain and offset errors of the converter and therefore very suited for CMOS implementations.

7 Accuracy limits of bandgap references

In well-designed bandgap references the accuracy is only limited by that of the base-emitter voltages of the applied bipolar reference transistors. To the best of the author's knowledge no results of systematical research on the fundamental limits to this accuracy have been published. Therefore the discussions, ideas and suggestions in this section are based on information fragments, speculations and some experimental work of the research group of the present author. According to our experiments mechanical stress is the main cause of the inaccuracy of bandgap references operated in the temperature range of -50°C to +150°C. Mechanical stress affects the base-emitter voltage by the so-called piezo-junction effect [23],[24]. The macro effects of mechanical stress will cause a change in the value of V_{BE}, while micro effects, which can be different from transistor to transistor, also affect the V_{BE} matching and consequently V_{PTAT}. These effects can result in the following non-idealities:

- After packaging, the output voltage of a calibrated bandgap-reference chip differs from the initial value. The difference is more or less unpredictable and causes additional temperature dependence of the output voltage.
- Some of the bandgap references, even when they have the correct output voltage at ambient temperatures, do have a relatively large temperature coefficient. Such an effect can be caused by stress-induced mismatching of a PTAT pair of transistors.
- The output voltage of bandgap references shows hysteresis during thermal cycling

In our experiments bandgap references trimmed within 300 ppm of the nominal value showed a deviation up to 2000 ppm after packaging in a metal can and up to 6000 ppm after packaging in (low-stress) epoxy. After removal of the epoxy by chemical etching accuracy recovered to the initial value. The use of a silicone blob top was found to be helpful in reducing the effect of epoxy-induced mechanical stress.
To minimize the relative effects of stress it is advantageous to maximize the magnitude of V_{PTAT}. However, very large values of V_{PTAT} will introduce new errors due to high-level and self-heating effects in the transistor operated at a high current density, and leakage-current, noise and interference effects in the transistor operated at a low current density. A completely different way to cope with the stress problems is to use micro-machined structures to eliminate the stress in the most critical parts of the chip.

8 Conclusions

Bandgap references are applied as a voltage-reference element in voltage and current regulators and in voltage-measurement systems. To obtain the best result at the lowest price, it is often advantageous to make an overall system design. It has been shown that in voltage-measurement systems the linear errors of the processor can be eliminated by using an autocalibration technique. This technique is applied in the novel dynamic bandgap references presented in this paper. The dynamic bandgap references are very suited for application in CMOS technology, because they eliminate the need for low-offset amplifiers.

In bandgap references, compensation can be made for the linear as well as the nonlinear temperature dependence. Bandgap references can be realised in bipolar as well as in CMOS technology. Usually, at least two transistors are required to generate the basic voltages V_{BE} and V_{PTAT}. However, in the dynamic bandgap references a single bipolar transistor can be sufficient.

Bandgap references can be calibrated with a single-step trimming at, for instance, ambient temperatures; but certain procedures have to be followed to get the full advantage of this attractive feature.

Many well-designed bandgap references have been presented in literature. The accuracy of these bandgap references is limited by the effect of mechanical stress in the reference transistor. In this paper various ways to reduce this problem are suggested. More systematic research on this topic is required to achieve a major breakthrough with respect to accuracy.

9 Acknowledgement

The author would like to thank Herman Casier and Stefaan Blieck of Mietec for their contributions in the field of CMOS bandgap references and Arie van Staveren of TU Delft for discussions and assistance concerning the curvature-corrected bandgap references.

10 References

[1] *D.F. Hilbiber*, "A new semiconductor voltage standard", Dig. Techn. Papers, ISSCC, pp. 32-33, Feb. 1964.

[2] *R.J. Widlar*, "New developments in IC voltage regulators", IEEE J. Solid-State Circuits, vol. SC-6, pp. 2-7, Feb. 1971.

[3] *G.C.M. Meijer*, "Bandgap references", Proc. Workshop Advances in Analog Circuit Design, 26-28 april, 1995, Austria.

[4] *G.C.M. Meijer*, "Integrated circuits and components for bandgap references and temperature transducers, PhD thesis, TU Delft, The Netherlands, 1982.

[5] *J.W. Slotboom and H.C. de Graaff*, "Measurements of bandgap narrowing in Si bipolar transistors", Solid-State Electron., vol. 19, pp.857-862, Oct. 1976.

[6] *Y.P. Tsividis*,"Accurate analysis of the temperature effects in $I_C - V_{BE}$ characteristics with application to bandgap reference sources", IEEE J. Solid-State Circuits, vol. SC-15, pp. 1076-1084, Dec. 1980.

[7] *A.P. Brokaw*,"A simple three-terminal IC bandgap reference", IEEE J. Solid-State Circuits, vol. SC-9, pp. 388-393, Dec 1974.

[8] *G.C.M. Meijer and J.B. Verhoeff*, "An integrated bandgap reference", IEEE J. Solid-State Circuits, vol.-11, pp. 403-406, June 1976.

[9] *R.J. Widlar*,"Low-Volage Techniques", IEEE J. Solid-State Circuits, vol. SC-13, pp. 838-846, Dec. 1978.

[10] *G.C.M. Meijer, P.C. Schmale and K. van Zalinge*, "A new curvature-corrected bandgap reference", IEEE J. Solid-State Circuits, vol. SC-17, pp. 1139- 1143, Dec. 1982.

[11] *M. Gunawan, G.C.M. Meijer, J. Fonderie and J.H. Huysing*, "A curvature-corrected low-voltage bandgap reference", IEEE J. Solid-State Circuits, vol. 28, pp. 667-670, June 1993.

[12] *B. Song and P.R. Gray*, "A precision curvature-compensated CMOS bandgap reference", IEEE J. Solid-State Circuits, vol SC-18, pp. 634-643, Dec. 1983.

[13] *I. Lee, G. Kim and W. Kim*, "Exponential curvature-compensated BICMOS bandgap references", IEEE J. Solid State-Circuits, vol. 29, pp. 1396-1403, Nov.1994.

[14] *A. van Staveren, J. van Velzen, C.J.M. Verhoeven and A.H.M. van Roermund*, "An integratable second-order compensated bandgap reference for 1 V supply", to be published in: Analog Integrated Circuits and Signal processing, July 1995.

[15] *A. van Staveren*, "A systematic design method for bandgap references", Research Report, Delft University of Technology, March 1995.

[16] *M. Martins*, "CMOS shunt regulator with bandgap references for automotive environment", IEE Proc. Circuits Devices and Systems, vol. 141, pp. 157-161, June 1994.

[17] *A.G. van Lienden, G.C.M. Meijer and J. van Drecht*, "Latch-up in bipolar low-voltage current sources", IEEE J. Solid-State Circuits, vol. SC-22, pp. 1139-1143, Dec. 1987.

[18] *G. Tzanateas, C.A.T. Salama and Y.D. Tsividis*, "A CMOS bandgap voltage reference", IEEE J. Solid-State Circuits, SC-14, pp. 655-657, June 1979.

[19] *H. Casier*, Private communication, March 1995.

[20] *M.J.M. Pelgrom, A.C.J. Duinmaijer and A.P.G. Welbers*, "Matching properties of MOS transistors, IEEE J. Solid-State Circuits, vol. 24, pp. 1433-1440, October 1989.

[21] *H. Casier*, "BICMOS analog circuit design techniques", Course on practical aspects in analog IC-design, EPFL-Lausanne, June 1992.

[22] *G.C.M. Meijer and A.W. van Herwaarden*, "Thermal Sensors", IOP publ., Bristol, 1994.

[23] *B. Puers and W.M.C. Sansen*, "New mechanical sensors in silicon by micromachining piezojunction transistors", Proc. Transducers '87, Tokyo, pp. 324-327, 1987.

[24] *Y. Kanda*, "Effect of compressive stress on silicon bipolar devices", J. Appl. Phys., vol. 44, pp. 389-394, 1973.