DESIGN AND OPERATION OF LOW POWER TEMPERATURE SENSOR - BANDGAP REFERENCE CIRCUIT IN SUBMICRON TECHNOLOGY

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ABSTRACT: A low power CMOS temperature sensor and a voltage reference circuit is presented. The circuit was designed and produced in submicron 0.25µm technology. It can work in temperature range from at least -20^0C to $+120^0$C. The minimum measured accuracy of the calibrated temperature sensor is of about 0.1^0C (rms). The bandgap reference circuit generates *Vref*=1160mV with 1.6mV chip to chip spread. The power consumption is about 75µW and the chip can work for bias voltage down to 1.9V.

INTRODUCTION

The demand for the integrated low power high performance temperature sensors is increasing in a wide range of applications [1]. These sensors are used for temperature measurements, to control the power dissipation or to compensate the temperature dependence of other devices. Among many fields they find applications in the High Energy Physics experiments where complex detector systems, with millions of channels, must operate without direct human intervention for months and years. In such systems temperature sensors on chip are requested for different diagnostic purposes.

In this paper we describe the design and measurements of a prototype temperature sensor chip designed for on chip diagnostics for the silicon drift detector (SDD) system within the ALICE experiment at CERN. Since the requested specifications left a large margin of freedom for the design, the circuit architecture was chosen in order to include also the voltage reference circuit. This was done since both temperature sensor and voltage reference often use the same "core" and so it is straightforward to design them together.

In the following we briefly remind the theory of operation of temperature sensors and voltage reference circuits. Then the design and simulations of the present circuit are described. Finally we present the measurements and the results obtained for both the temperature sensor and the voltage reference circuit.

PRINCIPLE OF OPERATION

Temperature Sensor. The temperature sensors implemented in the VLSI technology commonly use the I-V characteristics of a forward biased *p-n* junction [2]. For a forward biased diode the current voltage relation reads as:

$$I_C = I_s \exp\left[\frac{V_{BE}}{V_{th}}\right] \qquad (1)$$

The voltage difference between two diodes drawing the same currents but working at different current densities may be simply derived as:

$$\Delta V = V_{th}\ln[n] = \frac{kT}{q}\ln[n] \qquad (2)$$

where n is the current density ratio.

In practice this condition may be satisfied drawing the same current into two diodes of different area. In this case the ratio of saturation currents will become the ratio of diodes areas. As seen from the above equation the resulting voltage difference is proportional to the absolute temperature (PTAT). A whole class of PTAT sensors uses this relation for practical applications.

Voltage Reference. The forward biased *p-n* junctions are also used to build voltage references. At a constant current the junction voltage drop may be expressed as:

$$V_{BE} = V_{th}\ln\frac{I_C}{I_s} = \frac{kT}{q}\ln\frac{I_C}{I_s} \qquad (3)$$

Its temperature derivative is:

$$\frac{\partial V_{BE}}{\partial T} = \frac{\partial V_{th}}{\partial T}\ln\frac{I_C}{I_s} - \frac{V_{th}}{I_s}\frac{\partial I_s}{\partial T} \qquad (4)$$

Taking the saturation current as:

$$I_s \propto bT^{4+m}\exp\left[-\frac{E_g}{kT}\right] \qquad (5)$$

where m ≅ -1.5, one obtains:

$$\frac{\partial V_{BE}}{\partial T} = \frac{V_{BE} - (4+m)V_{th} - E_g/q}{T} \qquad (6)$$

This expression is characterized by a negative temperature coefficient (TC), which in first approximation may be assumed constant. The idea of a voltage reference is to add a positive TC voltage to a negative TC junction voltage in order to compensate the temperature dependence. The ideal candidate for the positive TC is the PTAT voltage. Adding the PTAT voltage to the V_{BE} with the weight which assures the cancellation of temperature dependence (in a first approximation) of the sum, a voltage reference source may be built. Adding (2) and (3) and using equations (4-6) the expression for the voltage reference *Vref* may be found:

$$V_{ref} = \frac{E_g}{q} + (4+m)V_{th} \qquad (7)$$

A quantitatively dominating term in this expression comes from the semiconductor energy bandgap E_g and only very small temperature dependent term reminds. For this reason a whole class of voltage references using this principle are called the bandgap voltage references. As seen from the above expression the bandgap voltage has precise and almost constant absolute value.

CIRCUIT DESIGN

Overall Architecture. Few considerations were done in order to choose the circuit architecture [2,3]. At first, the circuit was aimed mainly to measure the temperature but a simple implementation of a bandgap reference was also foreseen. At second, the availability of forward biased junction devices was limited to a substrate grounded diodes. With this in mind a simple structure shown in fig. 1 was chosen. The first two branches constitute a typical PTAT core with diodes D1, D2 working as the temperature sensing devices. The voltage difference between the D1 and D2, proportional to an absolute temperature, is seen on the resistor R2. This is true if the amplifier's inputs are at the same potentials. To satisfy this condition a negative feedback is applied which sets the proper currents of D1 and D2. The currents, equal in all branches with the exception of M3, were chosen to ensure the "ideal" current-voltage characteristics of D1 and D2. This was done knowing the right current density range for used technology and resulted in a current value of about 2.3µA at room temperature.

The PTAT voltage drop on the resistor R2 is mirrored on the resistor R3 with a multiplication factor $k_I \cdot k_R$. The *Vtemp* is the output of the temperature sensor. Taking into account the amplifier offset voltage an expression for *Vtemp* may be derived as:

$$V_{temp} = k_I \cdot k_R \cdot \left(V_{th} \ln(k_D) - V_{OS}\right) \qquad (8)$$

where k_I, k_R and k_D are the ratios of the mirrored currents, resistances and diodes areas respectively, while V_{OS} is the amplifier input offset voltage.

Also current PTAT signal is available as *Itemp* in fourth branch. If not used it should be grounded.

The bandgap reference voltage *Vref* is build in the fifth branch as the sum of the D5 voltage drop (with negative TC) and the voltage drop on R5 (with positive TC). The D5 diode is the same as D1. The cancellation of temperature dependence is obtained ensuring the right multiplication factor $\ln(k_D) \cdot k_R$. The final expression for *Vref* is:

$$V_{ref} = V_{D5} + k_R\left(V_{th} \ln(k_D) - V_{OS}\right) \qquad (9)$$

To ensure that the circuit enters the stable operating point a diode connected transistor M0 was added as a simple starter. The nominal power consumption of the circuit without the amplifier is $2.5V \cdot (4+k_I) \cdot 2.3\mu A$ which gives about 40µW.

Amplifier Design. A proper functioning of the circuit is ensured by the operational amplifier which establishes the critical potentials. Some circuit aspects were taken into account to chose the amplifier topology. The amplifier sees only a capacitive load and so the high current capacity was not necessary.

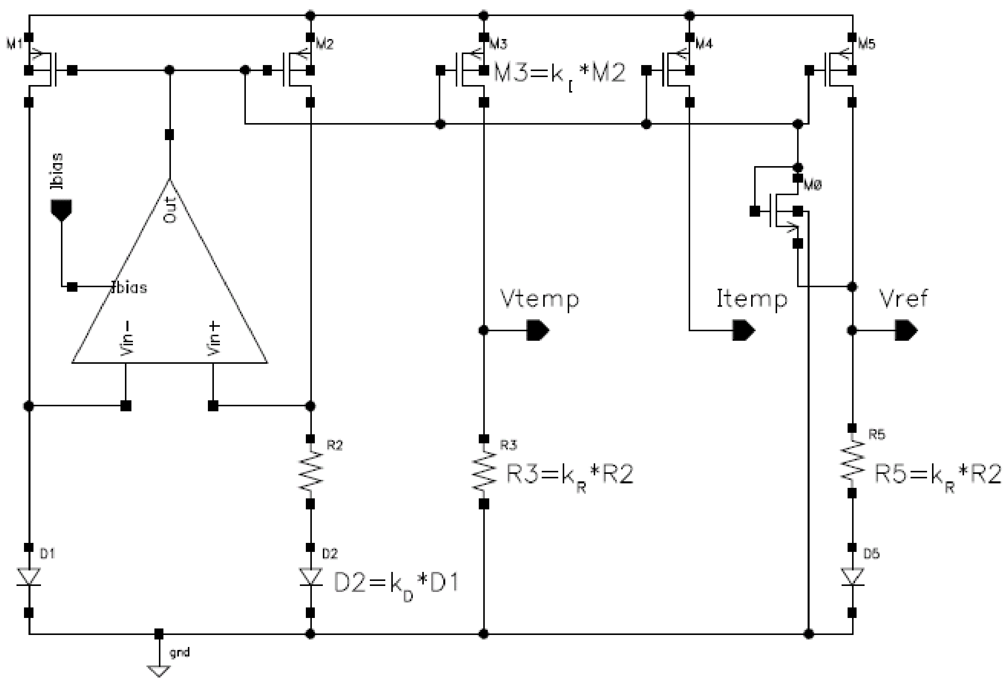

Fig. 1. Schematic of temperature sensor and bandgap reference circuit

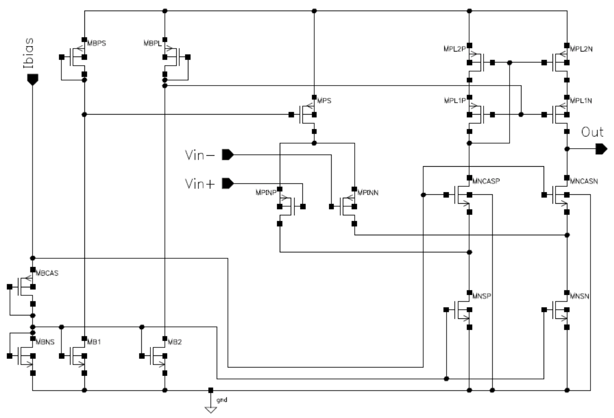

Fig. 2. Schematic of amplifier

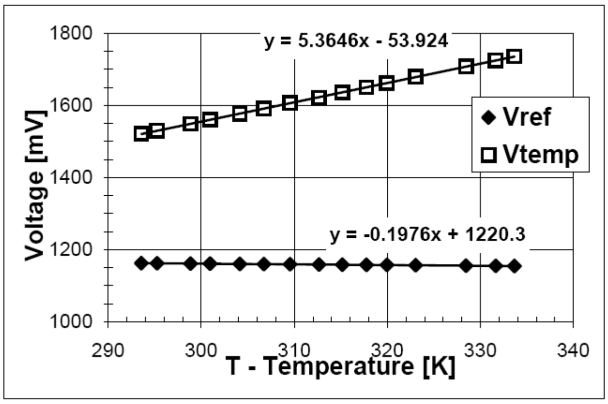

Fig. 3. Vtemp and Vref outputs vs temperature for typical chip

To obtain a high voltage gain of above 80dB either a one stage high gain configuration could be used or a standard two stage configuration. Here the first solution was chosen. One of important reasons was the fact that in a single stage configuration the stability is improved by increasing the output capacitance. In present circuit all current sources M1-M5 are long length devices and so their input capacitance could be used for the amplifier compensation. To obtain a high voltage gain a folded cascode topology with a cascode current load was chosen. This configuration is also preferred due to its good PSRR. The overall amplifier scheme is shown in fig. 2. Another important factor in the amplifier design was a low input voltage offset requested by the application. For this reason large transistors in the differential input pair and in the cascode load were used. The amplifier power consumption is about 35µW. For a test purpose an additional copy of the amplifier was also included on the prototype chip.

Simulations and Layout. The whole circuit was designed and simulated with hspice simulator. The simulations were performed for the temperature range from -50^{0}C to +100^{0}C showing very good sensor linearity and stable (within ~2mV) bandgap voltage output in the whole range. The total power consumption is about 75µW.

During the layout process all critical components, diodes D1, D2 and D5, resistors R1, R2 and R5, transistors M1-M5 were built as the arrays of elements using common centroid geometry and dummy devices.

MEASUREMENTS AND RESULTS

The first qualitative measurements of the circuit were done in the temperature range from -20^{0}C to +120^{0}C showing the expected behavior of the temperature sensor and bandgap voltage reference. The detailed measurements discussed in this paper were done in a temperature range foreseen for the application from +20^{0}C to +60^{0}C.

The measurements were done on a sample of 22 temperature sensors and 22 bandgap references. In the presented plots descriptions in legends like "C1D2" regard the chip number and device (temperature sensor or bandgap reference) number. Few separate amplifiers were also available and the dedicated tests (mainly of the input voltage offset) were performed.

The temperature dependence tests were done using a simple setup including: the chip bonded on a chip carrier, the aluminium box containing the chip carrier and the cabling, the heater built of high power ceramic resistors placed below the aluminium box and the isolating polystyrene box containing the whole setup contents. The temperature was measured by the platinum thermometer of 0.1^{0}C precision which was in thermal contact with the chip carrier. Temperature measurements were done for raising and falling temperatures in order to account the inertia effects.

The qualitative results for a temperature sensor and bandgap voltage reference are shown in fig. 3 for a typical chip. As expected the temperature output voltage *Vtemp* increases linearly with temperature while the bandgap reference voltage *Vref* is practically constant. It was verified that also the power consumption agrees with simulations. Below we discuss more in detail both subcircuits separately. We will also discuss the effect of amplifier non-ideality on the circuit behavior.

Temperature Sensor

As seen from fig. 3 the *Vtemp* sensitivity is of about 5mV/K. The precision of temperature measurement depends on a sensor linearity and a knowledge of its linear fit parameters, the slope and offset. Depending on requested accuracy one can use the sensor with or without calibration. We have studied three possible types of applications: without any calibration, with single temperature calibration (probably at room temperature) and with precise calibration.

Not Calibrated Sensor. To show the sensor behavior without calibration in fig. 4 we plotted the sensor output voltage *Vtemp* versus temperature for all measured circuits. In order to determine the average precision we have done a linear fit to all measurements. In this way we obtained the averaged slope and offset for all sensors:

$$V_{temp} = 5.26\frac{mV}{K}\cdot T - 22.74mV \qquad (10)$$

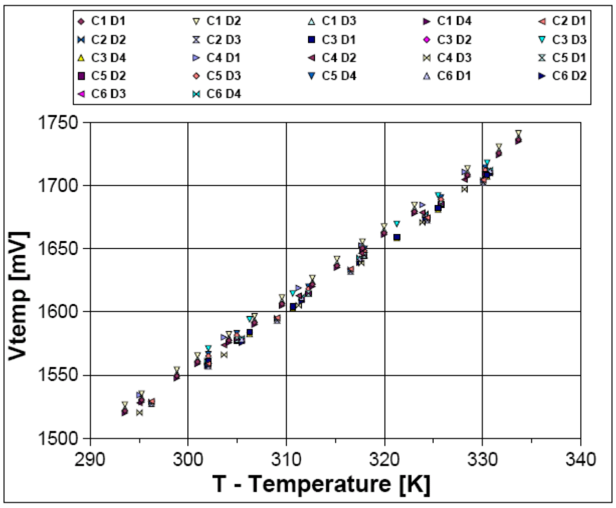

Fig. 4. Not calibrated Vtemp versus temperature for all sensors

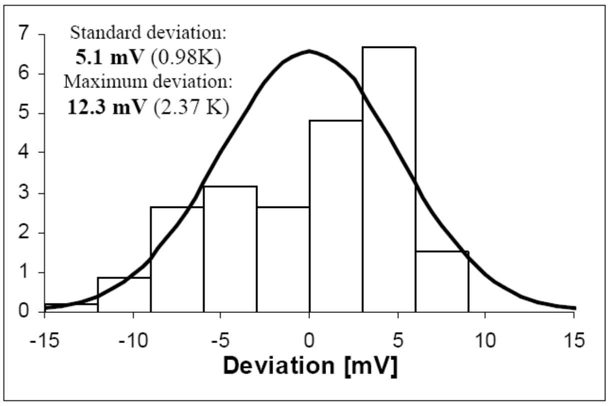

Fig. 5. Histogram Vtemp deviations for not calibrated sensor

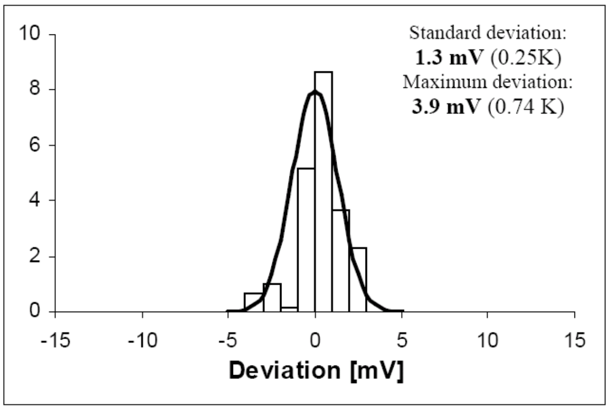

Fig. 6. Histogram of Vtemp deviations for one point calibration

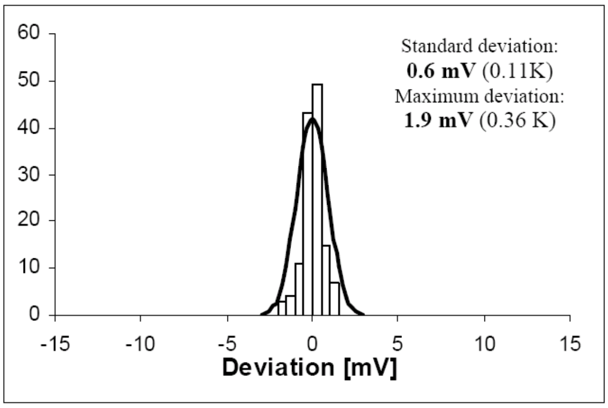

Fig. 7. Histogram of Vtemp deviations for multipoint calibration

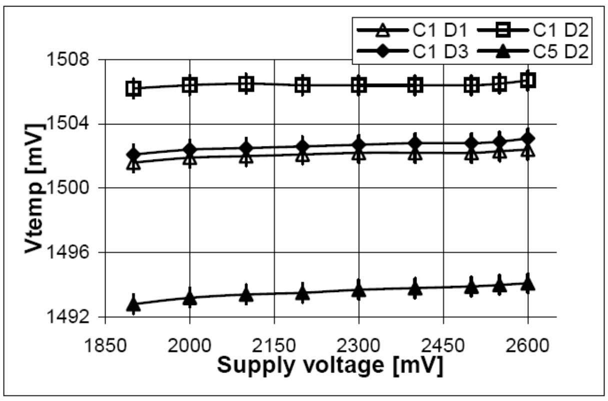

Fig. 8. Vtemp versus bias voltage

To get an idea about the accuracy possible to obtain using the averaged fit in fig. 5 we plot the histogram of deviations between the measurements and the fit. The standard deviation obtained from this histogram is 5.1mV which corresponds to about 1^0C and it may be used as a precision estimation of a randomly chosen not calibrated sensor.

Single Point Calibration. An intermediate accuracy measurement could be done using the sensor after a calibration at one chosen temperature (the simplest at room temperature). Practically we make this calibration taking the slope from previously found averaged fit (eq.10) and finding the optimal offset parameter for each sensor separately. To see the measurement accuracy after such calibration we create the histogram of deviations between the measurements and the linear fit for each sensor. The histogram of deviations summed over all measured sensors is plotted in fig. 6. Using standard deviation to estimate the measurement precision an improvement from 1^0C (no calibration) to 0.25^0C (one point calibration) is found.

Multipoint Calibration. The most accurate, but also the most time consuming, measurement could be done calibrating every sensor in different temperatures. This was done making the linear fit to the measured values for each sensor separately. The histogram of a fit to measurement deviations summed over all sensors is shown in fig. 7. The obtained standard deviation of 0.6mV corresponds to 0.11^0C. However we can not consider this value as an estimation of the sensor accuracy since the uncertainties connected to the setup are of the same order. The main two are the resolution of the thermometer of 0.1^0C and the uncertainty connected to the setup thermal inertia together with a quality of the thermal contact between the sensor and the thermometer. So we expect the sensor precision to be at least not worse than the estimated value but due to the setup limitations we were not able to verify it.

Bias Voltage Dependence. We have also studied the effect of the bias voltage value on sensor behavior. The typical curves of *Vtemp* versus the bias voltage are shown in fig. 8. The sensor works properly for the bias voltage range from 1.9V to 2.6V. No significant dependence of *Vref* on bias voltage is seen, although in case of multipoint calibration temperature measurements a small correction may be needed.

Summary. To summarize the temperature sensor discussion it is interesting to compare the measured results to hspice simulated model and to a pure theoretical model. As it was shown before from pure theoretical *p-n* junction considerations one gets (eq. 8):

$$V_{temp}^{theory} \cong 5.259\frac{mV}{K}\cdot T - 26.25\cdot V_{OS} \qquad (11)$$

for present application parameters k_I, k_R and k_D.
The fit obtained from hspice simulations using the models appropriate for the used technology gives:

$$V_{temp}^{hspice} = 5.1914\frac{mV}{K}\cdot T + 3.1008mV \qquad (12)$$

The measurement averaged over all sensors (eq. 10) gives:

$$V_{temp}^{meas} = 5.26\frac{mV}{K}\cdot T - 22.74mV$$

A good agreement of the slope of the order of ~1% is seen between the theory, simulations and measurements.

Bandgap voltage reference

Overall Bandgap Performance. The main feature requested from the reference voltage is its precision and independence from external factors. In fig.9 we plot the results of *Vref* measurements versus temperature for all bandgap circuits. All circuits give almost the same output reference voltage with few milivolts precision. To quantify it better the histogram of *Vref* measured at the same temperature for all bandgap reference circuits is shown in fig. 10. A small standard deviation of 1.6mV is found with a maximum spread of 4.1mV.
The average *Vref* value from all measurements, taken at room temperature is:

$$V_{ref}^{meas} = 1160mV \qquad (13)$$

It agrees very well with the results of hspice simulations where the same value was found:

$$V_{ref}^{hspice} = 1160mV \qquad (14)$$

Temperature and Bias Voltage Dependence. From the *Vref* temperature dependence shown in fig. 9, taking a first order linear fit, one can find that:

$$\frac{\partial V_{ref}}{\partial T} \cong -0.18\,{}^{mV}\!/_{K} \qquad (15)$$

This value, although acceptable in many applications, is few (~5) times higher than one could expect from the simulations or theoretical calculations. As seen from fig.9 the temperature dependence is a systematic feature of all bandgaps and so can not be explained neither by the statistical variations of technology parameters nor by the measurement uncertainties. This dependence can not come from the PTAT component of the *Vref*, since the results obtained for the temperature sensor were in good agreement with the simulations and theory. To understand the reason a study of junction voltage dependence on temperature for a separate diode would be very helpful. Unfortunately we did not have an additional diode on the chip, for the test purpose. Reference voltage was also studied versus bias voltage value and the results measured for a few typical bandgap circuits are shown in fig. 11. The negligible

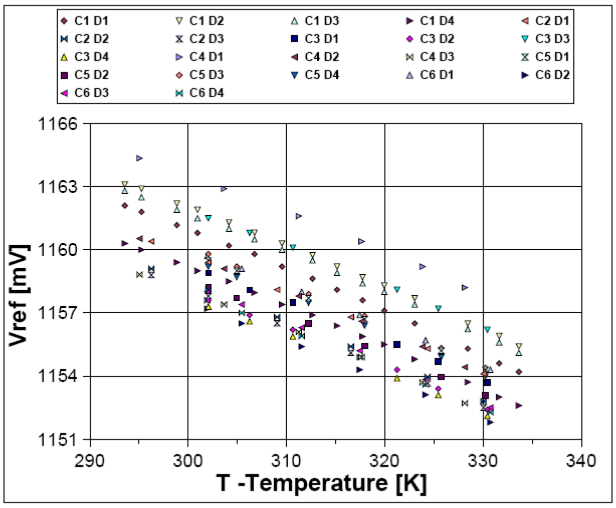

Fig. 9. Vref temperature dependence for all bandgaps

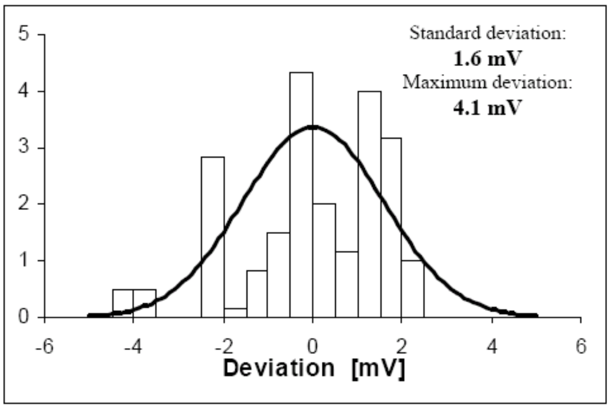

Fig. 10. Histogram of bandgap voltage deviations

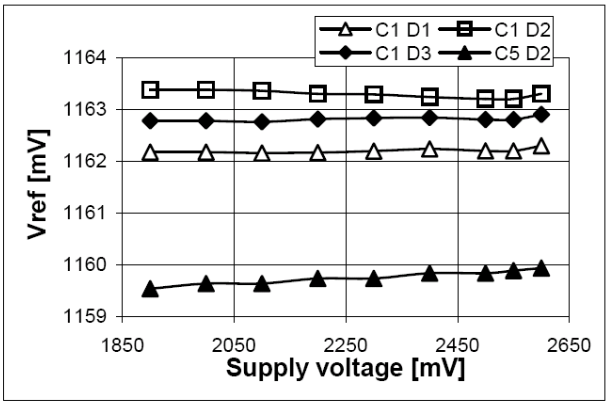

Fig. 11. Vref versus bias voltage

changes of *Vref* allow to conclude that *Vref* does not depend on bias voltage within the measured bias range.

Amplifier measurements

The main aim of the amplifier is to set the branch currents so as to have the same potentials at its inputs. Any non zero input voltage would add directly to the D1, D2 diode voltage difference and would deteriorate the circuit operation. For this reason one of the main requirements in the amplifier design was to minimize its input voltage offset.
We have measured the amplifier input offset for a few (six) available amplifiers. The qualitative results showed a small amplifier average offset of the order of ~250µV. We have also studied the temperature dependence and bias voltage dependence of amplifier offset. In both cases no change of the offset voltage was observed.

CONCLUSIONS

A low power temperature sensor and bandgap voltage reference circuit was designed and produced in submicron 0.25µm technology. A good operation of the temperature sensor and bandgap voltage reference was verified in a wide temperature range (-20^0C, $+120^0$C) and the detailed measurements were done in the interesting range ($+20^0$C, $+60^0$C). The temperature sensor was tested for different type of applications showing that the temperature accuracy in the range from about 1^0C to 0.1^0C may be obtained. The bandgap voltage reference produces an output voltage *Vref*=1160mV with chip to chip spread below 2mV.

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