

A Resistorless Switched Bandgap Reference Topology

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Abstract—This work presents a novel Switched-capacitor Bandgap Reference circuit using no resistors at all. The negative drift V_{EB} and the positive drift V_T voltages are generated by the same PNP bipolar transistor. A switched-capacitor circuit stores and processes these voltages, generating the bandgap reference voltage in only 4 clock phases. The proposed topology was simulated in CMOS 130nm process, presenting better performance than traditional band-gap topologies. Monte Carlo simulation shows lower variability, that results from the lower variability of capacitor ratios instead of resistor ratios and from the use of just one bipolar transistor.

I. INTRODUCTION

Voltage references, besides operational amplifiers, are fundamental blocks that are present today in every analog, mixed-signal, and even digital design. They are essential for bias, data-conversion, power regulation and management, and almost any other electronic circuit implementation. Its most important characteristic is to offer a voltage that is very insensitive to temperature variations, presenting a good thermal stability. Also, this voltage must be almost the same in any fabricated unit, which means that the circuit topology has to be very insensitive to uncontrollable factors in process fabrication that cause device mismatch.

Bandgap voltage reference is one of the most relevant circuits that successfully achieve these requirements, but as CMOS technology continues to shrink its size, resulting in larger variation in device parameters, it becomes more and more difficult to design precise bandgap references. As a consequence, traditional topologies such as in [1] are becoming unable to handle this challenge.

A lot of work has been done in order to improve bandgap reference's performance, such as in [2] based on weighted ΔV_{GS} , using curvature compensation of higher order temperature terms [3], [4], using only one BJT and MOSFETs in weak inversion [5] and using resistor trimmers to adjust process and mismatch errors [6]. Most of them use the same technique proposed originally by Robert Widlar in 1971's LM113 [7], the first commercial bandgap reference, where the difference of the V_{BE} voltages of two bipolar transistors, working under

different junction current densities, is applied over a resistor to generate a proportional-to-absolute-temperature (PTAT) current. Then, this current is applied to another resistor which value is calculated to make its voltage drop counterbalance the almost linear negative voltage drift of a bipolar junction under forward bias, around $-2\text{mV}/^\circ\text{C}$.

The work in [8] describes the process and mismatch variation dependency of bandgap references and proposes a process-mismatch design-oriented strategy. It presents the dependency of the bandgap voltage on resistors and current mirrors mismatch. Furthermore, [9] proposes a bandgap reference with intrinsic compensation, in order to solve the current mirror mismatch problem and [10] proposes a bandgap reference without resistors. However, [9] depends on resistors good matching, and [10] on many current mirrors matching.

In this work, we propose a resistorless switched bandgap reference topology, which presents better voltage standard deviation due to device mismatch when comparing with a traditional bandgap reference using resistors. The topology presented uses a single current mirror, a single BJT, along with switched capacitors to implement the bandgap PTAT plus V_{EB} voltage computing with matching improved. The circuit gain depends on capacitors matching, which is far more accurate than resistors in current CMOS processes. Moreover, lower thermal noise is produced.

For comparison purposes, we present simulations from our topology compared to the traditional one in CMOS 130 nm process. Thermal behavior and Monte Carlo simulations using Cadence Virtuoso are shown.

This paper is organized as follows: section II introduces the concept and characteristics of a switched-capacitor bandgap reference circuit; section III presents the results of nominal voltage variation with temperature, control signals and Monte Carlo simulations, as well as Monte Carlo comparisons; and section IV presents the conclusions.

II. THE SWITCHED BANDGAP CIRCUIT

In our topology, the bandgap reference is based on the cancellation of the almost linear negative voltage drift of a bipolar

junction, around $-2\text{mV}/^\circ\text{C}$, by the proportional-to-absolute-temperature (PTAT) difference of the junction voltages of the same bipolar transistor, working under two different junction current densities. The switched bandgap reference proposed in this work is shown in Fig. 1. The op-amp is initially considered ideal for a simpler explanation.

The equation for the V_{EB} voltage as a function of its current I is approximately given by

$$V_{EB} = V_T \cdot \ln \frac{I}{I_S} \quad (1)$$

being V_T the thermal potential and I_S the junction reverse saturation current. Subtracting two V_{EB} voltages generated by different currents, i.e., I and $2I$, the following equation holds

$$V_{EB1} - V_{EB2} = V_T \cdot \ln \frac{2I}{I_S} - V_T \cdot \ln \frac{I}{I_S} \quad (2)$$

$$\Delta V_{EB} = V_T \cdot \ln 2 = \frac{kT}{q} \cdot \ln 2 \quad (3)$$

where k is the boltzman constant, T the temperature and q the electron charge. One can see in (3) that the voltage ΔV_{EB} is proportional to temperature. Also, the junction voltage V_{EB} shows a negative temperature drift, which to first order can be approximated by

$$V_{EB} = V_{EB}(T_0) - \alpha(T - T_0) \quad (4)$$

where α is around $2\text{mV}/^\circ\text{C}$. If one sums eqs. (3) and (4) with the correct gain in the first one, the bandgap reference is then obtained.

The switched bandgap reference proposed here does exactly as described above using switched capacitors, as shown in Fig 1. A single capacitor could be used instead of two (C_a and C_b), however the circuit would be more susceptible to current mismatches from the mirror (I_a and I_b should be equal). Let's suppose that I_a and I_b are almost equal, and its difference results from MOSFETs mismatch.

The behavior of this circuit is divided into 4 phases (assume the switches that are not closed are openned) for switches control:

- Phase 1) Switches S_0 , S_1 , S_3 and S_5 are closed and the capacitor C_a is charged to $V_{EB}(I_a)$;
- Phase 2) Switches S_0 , S_2 , S_3 and S_6 are closed and the capacitor C_b is charged to $V_{EB}(I_b)$;
- Phase 3) Switches S_1 , S_2 , S_3 , S_5 and S_6 are closed causing:
 - Average of V_{C_a} and V_{C_b} to $V_{EB}(I_{ab})$;
 - The capacitor C_1 is charged to the difference between $V_{EB}(I_a + I_b)$ and $V_{EB}(I_{ab})$;
- Phase 4) Switches S_0 , S_1 , S_2 , S_4 , S_5 and S_6 are closed and the output V_{REF} goes to its final value;

In phase 3, charges in C_a and C_b are averaged, making the voltage $V_{EB}(I_{ab})$ an intermediate result from currents I_a and I_b . Since both currents are summed in this phase to generate

$V_{EB}(I_a + I_b)$, this average reduces the impact of the mismatch between the mirror currents, resulting

$$\Delta V_{EB} = V_{EB}(I_a + I_b) - V_{EB}(I_{ab}) \approx \frac{kT}{q} \cdot \ln 2 \quad (5)$$

In this phase the charge stored in C_1 is given by

$$\Delta Q_1 = \Delta V_{EB} \cdot C_1 \quad (6)$$

as in phase 4 the charge in C_1 is transferred to C_2 , it yields

$$V_{C2} = \frac{\Delta V_{EB} \cdot C_1}{C_2} \quad (7)$$

Finally, the output voltage of the circuit is given by

$$V_{REF} = V_{C2} + V_{EB} \quad (8)$$

$$V_{REF} = \frac{C_1 k}{C_2 q} \cdot \ln 2 \cdot T + V_{EB}(T_0) - \alpha(T - T_0) \quad (9)$$

The thermal dependency is canceled if

$$\frac{C_1}{C_2} = \frac{\alpha \cdot q}{k \cdot \ln 2} \quad (10)$$

which is the amplifier gain. One can easily adjust the amplifier gain by simulations, using typical parameters and canceling thermal dependency for 300K. In our case, we obtained the optimum value of 25.6.

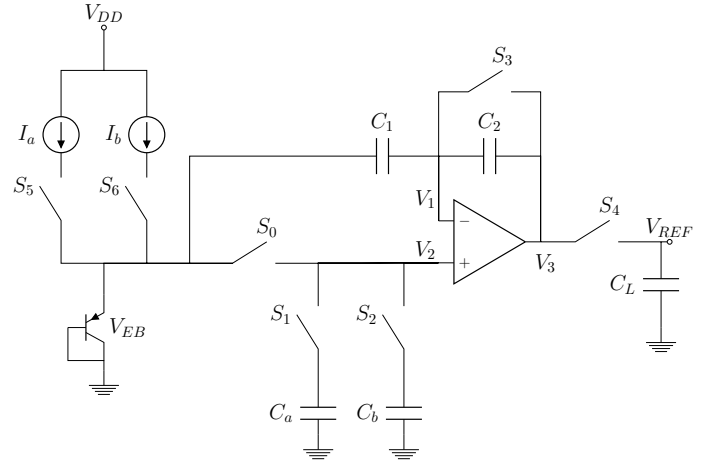


Fig. 1. Switched bandgap reference circuit.

III. SIMULATION RESULTS

The switched bandgap reference circuit was simulated with Cadence Virtuoso Spectre, using IBM 130 nm technology PDK. Electrical simulations used the foundry-supplied parameters for MOSFET model BSIM4. The ideal current sources were replaced by a mirror, controlled by the current bias source proposed in [11], which is also shown in Fig. 2. This current source uses no resistors and can be designed to have a good matching and acceptable temperature variation. For

this simulation we designed the mirror for 10uA, but lower currents could be used.

The traditional bandgap topology was designed to work with the same currents in the bipolar transistors. Since the process used offers many options for resistors, it was designed using OPRP (precision poly resistor), which would result in a large area taken for their layout. This circuit is shown in Fig. 3, being the base of most bandgap references. The start-up circuit is not shown. In this simulation we used $R1 = R2$ to keep the same current in both arms, and $n = 2$, meaning $Q2$ works with half the current density of $Q1$.

Both circuits were designed to operate in similar conditions, and the operational amplifier was modeled as ideal in both circuits, to focus on the circuit topology comparison.

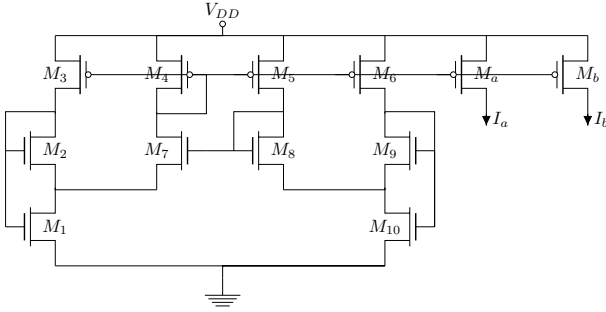


Fig. 2. Current bias source used for simulations.

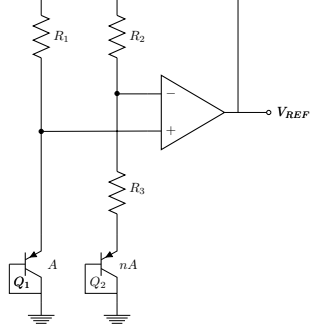


Fig. 3. Traditional bandgap reference circuit.

The V_{REF} equation for this traditional circuit is:

$$V_{REF} = V_{EB1} + V_T \cdot \ln(n) \frac{R_2}{R_3} \quad (11)$$

Fig. 4 shows V_{REF} vs temperature simulation results from both circuits, using typical parameters. The switched bandgap voltage has a 5 mV variation around 1.24 V, while the traditional bandgap voltage has the same variation around 1.21 V. From the temperature drift point of view, both circuits behave similarly.

The time behavior of voltages V_1 , V_2 and V_3 of Fig. 1 are shown in Fig. 5. The control signals of switches $S0$ to $S6$ are shown in Fig. 6. They follow the description of the 4 phases in the previous section.

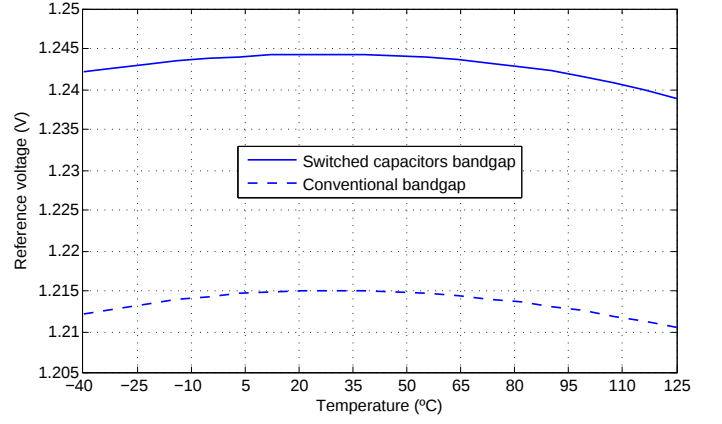


Fig. 4. V_{REF} vs Temperature curves.

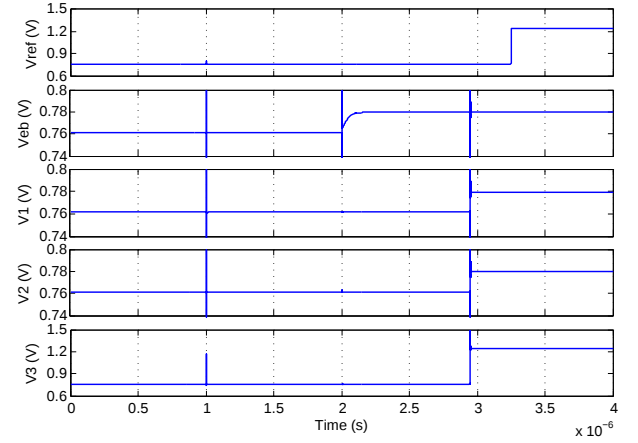


Fig. 5. Intermediate voltage signals.

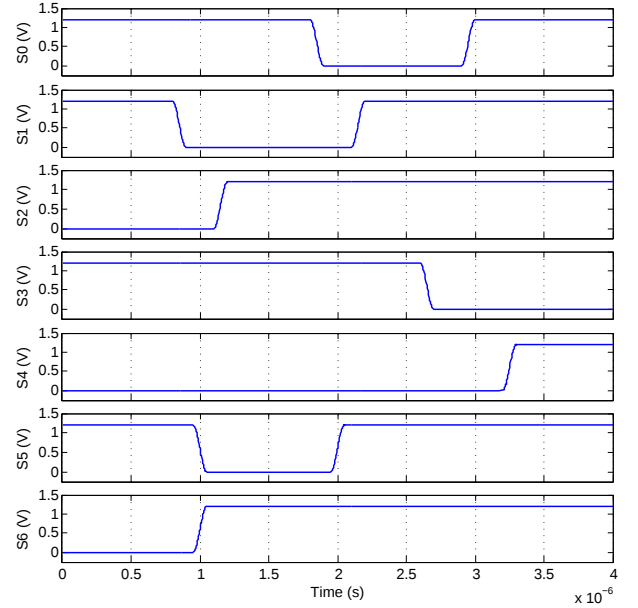


Fig. 6. Control signals for the switches.

Monte Carlo simulation of device mismatch impact in

V_{REF} , for 200 samples and 27 °C, is presented in Fig. 7 and Fig. 8, showing that the switched bandgap circuit presents a standard deviation in V_{REF} of 2.7 mV against the 4.8 mV presented by the traditional topology. This improvement results mainly from the use of capacitors instead of resistors in this topology. The use of only one bipolar transistor, and the mirror mismatch effect reduction by charge averaging, also contributed for the better performance. In simpler and lower cost processes, where precision resistors are not an option, this performance difference would increase. Also, the area taken by the critical capacitors (C_1 and C_2) is lower than the area used by the resistors. In Fig. 9 one can see the Temperature Coefficient (TC) variation of the switched bandgap presenting a mean value of 26.7 ppm/°C and a maximum of 30.3 ppm/°C for 3σ .

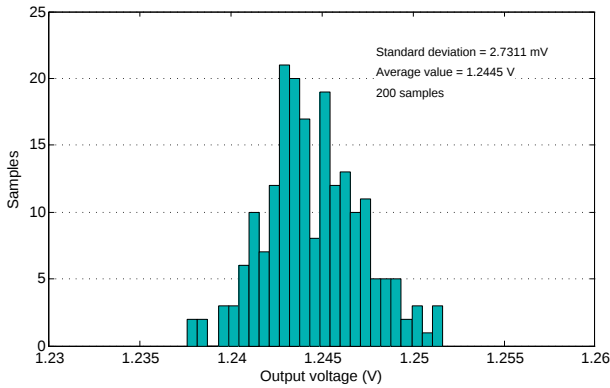


Fig. 7. Monte Carlo simulation for the switched bandgap.

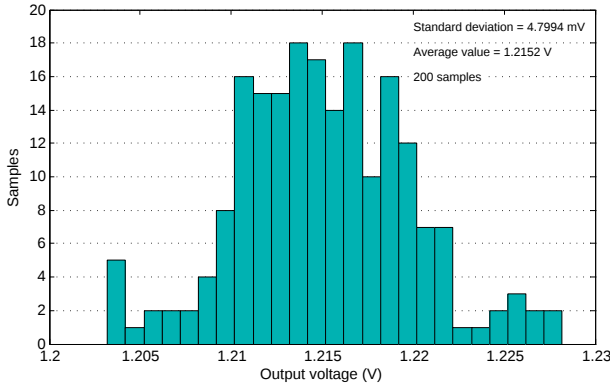


Fig. 8. Monte Carlo simulation for the traditional bandgap.

IV. CONCLUSION

We have introduced a new resistorless bandgap reference topology concept, without resistors, which offers lower variability based on better matching properties of capacitors in commercial processes. Also, bipolar current density variability, related to current mirror mismatch, is compensated by charge averaging in different phases. In simulations using IBM 130nm PDK, the reference voltage of the proposed circuit presented half the standard deviation of a traditional topology, working

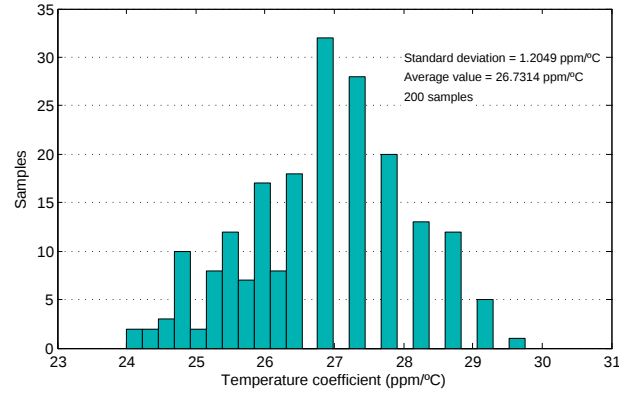


Fig. 9. Monte Carlo thermal drift simulation for the switched bandgap.

under similar conditions and using precision resistors. Also, a average TC of 26.7 ppm/°C enables this circuit to be used without calibration in many applications.

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