Compact and Low-Power Under-Voltage Lockout and Thermal-Shutdown Protection Circuits Using a Novel Low- I_Q All-in-One Bandgap Comparator

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Abstract—This paper presents a novel and compact bandgap comparator (BGRCOMP) for under-voltage lockout (UVLO) and thermal shutdown (TSD) protection circuits. The proposed BGRCOMP is self-referenced and combines the advantages of both a high-accuracy bandgap reference and a comparator into one single circuit. A latch-controlled biasing technique is also presented, which reduces static power consumption of the proposed BGRCOMP. The proposed BGRCOMP is used for the design of compact and low power UVLO and TSD circuits. The post-layout simulation results using a 0.18 μm BCD-on-SOI technology prove the attractive performance of the UVLO and TSD with a static current (I_Q) of 7.76 μA and 5.4 μA from a 5 V supply, respectively. The deviations of UVLO thresholds are less than 3 mV in the temperature range of -40 – 85 $^{\circ}\text{C}$.

Keywords—Bandgap comparator (BGRCOMP), under-voltage lock-out (UVLO), thermal shutdown (TSD), low power, protection circuits.

I. INTRODUCTION

Nowadays, electronic systems integrate various functions on the same die to achieve compactness, improve performance and reduce cost. To enhance the reliability requirement of these systems, under-voltage lockout (UVLO) and thermal shutdown (TSD) circuits are frequently needed for most integrated circuit (IC) systems as shown in Fig. 1(a) [1-2]. The UVLO circuit is employed to prevent the operation of circuits in case of a supply voltage failure. The TSD circuit senses and detects high temperature within the IC to turn off the source of heat when the temperature is higher than a certain preset value.

Fig. 1(b) shows conventional UVLO and TSD circuits consisting of three basic blocks: a sensor, a voltage reference and a hysteresis comparator. The temperature sensor of the TSD senses the temperature within the IC and converts the temperature into temperature-dependent voltage, $V_s(T)$. $V_s(T)$ is then compared with a voltage reference V_{REF} to detect the critical temperature condition and output a TSD signal. In the same way, the voltage sensor in the UVLO senses the value of the bias supply VDD and compares this sensing value $V_s(V)$ with V_{REF} to detect the under-voltage condition and output a UVLO signal. These conventional UVLO and TSD structures have been widely applied and reported in the literature [3-10]. However, several issues related to the design of UVLO/TSD circuits remain. First, a dedicated V_{REF} is needed, so the power consumption is high as in the UVLO designs reported by Song [3], Cho [4], and the design of the TSD reported by Ding [8]. A second very important issue for the design of TSD and UVLO is reducing the static quiescent current of the utilized

comparator, especially for the applications where the power supply is limited. The reason for this is that except UVLO/TSD circuits continue to operate when UVLO, TSD conditions occur, the other blocks of the IC are off. A third issue is that the accuracy of the trip points (voltage thresholds of UVLO and temperature threshold of TSD) for the UVLO and TSD circuits is low [3,5,7]. In order to simplify the design and reduce power consumption, a non-bandgap structure has been proposed [11]. However, the drifts of the trip points are even worse and not well defined. Therefore, the design of low power UVLO/TSD with precise trip points remains a challenge.

Fig. 1. (a) IC system with the integration protection functions, (b) Conventional structures and (c) new structure of the UVLO and TSD.

This paper proposes new solutions for the design of low-power, size-compact UVLO and TSD using a novel bandgap comparator as shown in Fig. 1(c). In this structure, a dedicated bandgap V_{REF} is no longer needed since the proposed bandgap comparator (BGRCOMP) can work as a comparator and provide a high-precision internal reference voltage. The two main contributions of this paper are: (1) a novel BGRCOMP with low quiescent current and (2) new UVLO and TSD circuits using the proposed BGRCOMP.

The rest of the paper is organized as follows: Section II presents the proposed BGRCOMP circuit. In Section III, the application of BGRCOMP in designing UVLO and TSD circuits is presented. Section IV provides circuit design and post-layout simulation results of the UVLO and TSD. Finally, Section V draws the main conclusions of this works.

II. PROPOSED BANDGAP COMPARATOR

The proposed BGRCOMP in Fig. 2 consists of 4 stages. The first stage includes Q1, Q2, R1, R2 and a latch load MP1-MP4 that constitutes the core of the BGRCOMP. Q1, Q2, R1, and R2 compose the comparator utilizing the bandgap reference principle, so an internal reference voltage exists. The second stage (MN1-MN2, MP5-MP6) with an amplification current factor of M is used to increase the gain and the resolution of the comparator. Two latches (MN3-MN8, MP7-

MP12) help to provide a rail-to-rail output. The two extra switches M_{SW1} and M_{SW2} help to reduce quiescent current.

2.1 Bandgap comparator operation principle

Fig. 3(a) shows the core of the proposed BGRCOMP. As will be explained below, this structure can self-generate an internal reference and compare the input voltage with this internal reference value. For the ease of analysis, assume that Q1 and Q2 are in the forward-active region. As the emitter area of Q2 is N times that of Q1, $I_{S2} = N \times I_{S1}$ where I_{S1} and I_{S2} are the saturation currents of Q1 and Q2, respectively. The large-signal collector currents of Q1 and Q2 can be expressed as:

$$i_{C1} = I_{S1} \times e^{\frac{v_{IN} - (i_{C1} + i_{C2})R_2}{U_T}}; i_{C2} = N \times I_{S1} \times e^{\frac{v_{IN} - (i_{C1} + i_{C2})R_2 - i_{C2}R_1}{U_T}} \quad (1)$$

where $U_T = kT/q$ is the thermal voltage, q is the electronic charge, k is the Boltzmann constant, and T is the temperature. The current ratio between i_{C1} and i_{C2} can be formulated as:

$$\frac{i_{C1}}{i_{C2}} = \frac{1}{N} \times e^{\frac{R_1 i_{C2}}{U_T}} \quad (2)$$

We define a balancing condition at which $i_{C1} = i_{C2} = i_{CB}$ and $v_{IN} = V_{INREF}$ where i_{CB} and V_{INREF} can be expressed as:

$$\frac{i_{C1}}{i_{C2}} = \frac{1}{N} \times e^{\frac{R_1 i_{CB}}{U_T}} = 1; i_{C1} = i_{C2} = i_{CB} = \frac{1}{R_1} \times \frac{kT}{q} \ln(N) \quad (3)$$

$$V_{INREF} = V_{BE1} + 2 \times \frac{R_2}{R_1} \frac{kT}{q} \ln(N) \quad (4)$$

Fundamentally, V_{INREF} in (4) is a function of temperature. It is well known that the V_{BE1} is complementary to absolute temperature (CTAT) while kT/q is proportional to absolute temperature (PTAT). With proper values of design parameters R_1 , R_2 , N , V_{INREF} can be designed to be insensitive w.r.t temperature variations. Under this balancing condition, as $i_{C1} = i_{C2}$, $V_A = V_B$.

Fig. 2. The proposed bandgap comparator consisting of internal reference voltage and latch-based control for low I_Q .

From (1), it can be seen that i_{C1} and i_{C2} are functions of v_{IN} . When $v_{IN} \neq V_{INREF}$, $i_{C1} \neq i_{C2}$, the BGRCOMP is out of the balancing condition. According to the value of v_{IN} , there are two scenarios depicted in Fig. 3(b) as follows:

- When $v_{IN} < V_{INREF}$, both of i_{C1} and i_{C2} are lower than i_{CB} . As $i_{C2} < i_{CB}$, from (2) and (3) it can be verified that $i_{C1}/i_{C2} < 1$, i.e. $i_{C1} < i_{C2}$. The current difference between i_{C1} and i_{C2} triggers the latch load MP1-MP4. Voltage V_B at node B is pushed down and V_A at node A is pulled up to VDD.
- When $v_{IN} > V_{INREF}$, both of i_{C1} and i_{C2} are higher than i_{CB} . As $i_{C2} > i_{CB}$, from (2) and (3) it can be verified that $i_{C1}/i_{C2} > 1$,

i.e. $i_{C1} > i_{C2}$. The current difference between i_{C1} and i_{C2} triggers the latch load. This current difference triggers the latch load MP1-MP4. Voltage V_B at node B is pulled up to VDD and V_A is pushed down.

Through the above analysis, when the value of $v_{IN} = V_{INREF}$, the comparator is in balancing state and helps to define an internal temperature-insensitive reference voltage. The operation of the bandgap comparator can now be modeled as a comparator with its internal bandgap reference as shown in Fig. 1(c) in which the input voltage v_{IN} to its internal temperature-insensitive voltage reference V_{INREF} .

Fig. 3. (a) Core of the bandgap comparator. (b) Variations of the collector currents i_{C1} and i_{C2} versus v_{IN} .

It should be noted that the proposed BGRCOMP in Fig. 2 utilizes BJT devices that might not be available with the standard CMOS process [12-13]. Another low-cost technique and highly compatible for digital CMOS technologies is to use a MOSFET pair in sub-threshold (weak inversion) replacing the Q1-Q2 pair. In sub-threshold regime, when the gate-source voltage is lower than the threshold voltage ($V_{GS} < V_T$), MOSFET works as a lateral BJT device [14-15].

Fig. 4. I_Q reduction with latch-controlled technique [16].

2.2 Low- I_Q BGRCOMP with latch-controlled method

To reduce power consumption, a latch-controlled biasing technique is also implemented with switches M_{SW1} and M_{SW2} that are controlled by the pre-latch (MN3-MN4, MP7-MP10) [16]. Fig. 4 shows the operation of the comparator (without the post-latch for the sake of simplicity) for the pre-latch state: $V_{OUT(-)Pre} = \text{HIGH}$, $V_{OUT(+Pre)} = \text{LOW}$. When $v_{IN} > V_{INREF}$, $i_{C1} > i_{C2}$ and the latch load (MP1-MP4) is triggered so that the drain current through MP1 increases rapidly and the drain current of MP2 decreases rapidly to zero. The generated current through MP1 is copied and amplified ($\times M$) by the current mirrors MP1-MP5/MN1-MN3 and pulls $V_{OUT(-Pre)}$ from HIGH to LOW.

Therefore, the pre-latch flips its internal states. When $V_{IN} < V_{INREF}$, the current through MP1 decreases to zero while the current through MP2 increases rapidly. However, since the output state of the pre-latch keeps M_{SW2} off, the triggering current from the current mirror MP2-MP6 cannot be copied to MN2-MN4. In this case, the internal logic of the pre-latch holds. The operation of the comparator using latch-controlled technique for the pre-latch state $V_{OUT(-)(Pre)} = \text{LOW}$, $V_{OUT(+)(Pre)} = \text{HIGH}$ is similar.

As a result of the latch-controlled biasing technique, the triggering currents $M \times I_{MP1}$ and $M \times I_{MP2}$ are generated only when the pre-latch changes its output state, i.e. the static current of the overall BGRCOMP is only in the first stage. This helps to minimize the quiescent current (I_Q) of the proposed BGRCOMP. Besides, by using a high value of R_1 and R_2 , I_Q could be designed in a nA range.

III. DESIGN OF UVLO AND TSD PROTECTION CIRCUITS

3.1 Design of BGRCOMP-based Under-Voltage Lockout

Fig. 5(a) shows the proposed UVLO, which is composed of the proposed BGRCOMP and a voltage sensor that is simply a voltage divider network (R_1 , R_2 , R_H). The voltage sensor R_1 - R_H senses the VDD values $V_S(\text{VDD})$ and compares the internal V_{INREF} of the BGRCOMP. The output of the comparator is under-voltage signal (UVLO), depending on the relative value between the V_S and V_{INREF} . The operation of the proposed UVLO can be briefly summarized as following: when VDD is low, $V_S < V_{INREF}$, meaning that the system supply VDD is in under-voltage state, the BGRCOMP compares V_S and V_{INREF} and outputs a signal at low logic (UVLO = '0'). When VDD increases, meaning that the system supply VDD is in a safe range, V_S increases and surpasses V_{INREF} of the BGRCOMP, the BGRCOMP outputs a signal at high logic (UVLO = '1'). The UVLO output signal of the UVLO is then used to allow the ON/OFF operation of the IC as shown in Fig. 1(a). To avoid an output oscillation when V_S fluctuates around V_{INREF} , a hysteresis is added and controlled by the switch SW_{HYS} which bypasses R_H . As the on-resistance of SW_{HYS} is small to bypass R_H , the size of SW_{HYS} is large; therefore, SW_{HYS} is driven by a buffer that consists of 2 inverters connected in series.

Fig. 5. Schematic of: (a) BGRCOMP-based UVLO, (b) BGR-based TSD.

3.2 Design of BGRCOMP-based Thermal Shut-Down

A simplified schematic of the proposed TSD circuit is constructed using a BGRCOMP, a PTAT temperature sensor, a start-up circuit, and a switch SW_{HYS} driven by buffers BUFF (Fig. 5(b)). The temperature sensor circuit generates a voltage-

dependent $V_{TS}(T)$, which is then compared to the internal reference V_{INREF} of the BGRCOMP. The PTAT temperature sensor is the current mirror BGR. Compared to PTAT temperature sensor based on [8], there is no need for the use of an op-amp, thus making area and power consumption lower. As the current mirror temperature sensor is self-biased, the startup circuit is included to avoid them from off state.

For the current mirror temperature sensor, the emitter areas of Q_{TA} and Q_{TB} are set as 1:N. By setting up the same size ratio for the current mirror MNT1-MNT2 and MPT1-MPT4, the collector currents of Q_{TA} and Q_{TB} are equal and the following relationship can be derived:

$$V_{TS}(T) = K \times I_{PTAT}(T) \times R_{TS2} = K \times \frac{R_{TS2}}{R_{ST}} \times \frac{kT}{q} \ln(N) \quad (4)$$

The $V_{TS}(T)$ is proportional to the absolute temperature T and depends on the design parameters K , R_{TS1} , R_{TS2} , N_S and physical constants k , T , q . In particular, $V_{TS}(T)$ depends on the relative ratio of R_{TS1} and R_{TS2} , not their absolute values; thus, the $V_{TS}(T)$ tracks the temperature T with a very high accuracy despite process variation. By letting $V_{TS}(T) = V_{INREF}$, the preset critical temperature $T = T_{CRIT}$ can be found. The operation of the proposed BGRCOMP based TSD can be briefly summarized as follows: when the temperature T is low, $V_{TS}(T)$ is lower than V_{INREF} of the BGR comparator, thus making the output of the TSD at low logic (TSD = '0'), meaning that the IC's temperature is low and in a safe range for the IC's operation. When the temperature increase, $V_{TS}(T)$ increases and surpasses V_{INREF} of the BGRCOMP, the BGRCOMP outputs a signal at high logic (TSD = '1'), meaning that the IC's temperature is higher than a safe level. The TSD output signal of the TSD circuit is then used to allow the ON/OFF operation of the IC as shown in Fig. 1(a).

IV. CIRCUIT DESIGN AND POST-LAYOUT SIMULATIONS

4.1 UVLO and TSD design specifications

The proposed UVLO/TSD circuits based on BGRCOMP are targeted for the use in the design of gate drivers, such as in [17-18]. The trip-point specifications of the UVLO and TSD circuits are set at 4.4 V/4.2 V (hysteresis window $\Delta V_{TH} = 200$ mV) and 90°C/70°C (hysteresis window $\Delta T = 20^\circ\text{C}$), respectively.

4.2 Layout and post-layout simulation results

The proposed BGRCOMP, UVLO and TSD circuits have been designed and simulated in a 0.18 μm BCD-on-SOI process with VDD = 5 V. For the BGRCOMP, the key design parameters were calculated and tuned as follows: $Q_1 = 3 \mu\text{m}/2 \mu\text{m}$, $N = 4$. The balancing condition of $I_{C1} = I_{C2} = 448$ nA was selected which is determined by the value of biasing resistor $R_1 = 78$ k Ω , $R_2 = 8 \times 78$ k Ω . Figure 6(a) shows the testbench to determine the V_{INREF} of the BGRCOMP and Fig. 6(b) shows the pre-layout simulation results. These results show that the BGRCOMP has a very small internal reference variation of 240 μV (2.68 ppm/ $^\circ\text{C}$) over a temperature range of -40°C to 100°C , thanks to the high amplification factor (β) of the utilized BJTs and temperature coefficient of the resistors R_1 , R_2 . Figure 6(c) shows the propagation delay of the

BGRCOMP at different input voltage ($V_{IN} - V_{INREF}$) to verify that the BJT BGRCOMP is fast enough for UVLO and TSD designs with a delay lower than 1 μ s.

Fig. 6. (a) Testbench to determine V_{INREF} of the BGRCOMP, (b) Pre-simulation results of V_{INREF} , (c) Pre-layout delay the BGRCOMP.

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