An RF-Powered Temperature Sensor Designed for Biomedical Applications

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ABSTRACT

Continuous patient monitoring enables early detection and treatment of diseases. Small and wireless devices are more comfortable to the patient and simplify the implantation procedure. This paper presents the design and partial testing of an integrated RF-powered temperature sensor suitable for the human body temperature range, from 35 to 42°C. The device was designed using a 130-nm standard CMOS technology and it receives energy from a 900-MHz signal that has power as low as −10 dBm. A calibration method was designed to obtain conversion errors smaller than 0.2°C.

Index Terms: Biomedical Electronics, RF Power Transfer, CMOS.

I. INTRODUCTION

A general application of biomedical electronic devices is continuous patient monitoring, which can enable early detection of diseases or complications, permitting a more effective treatment. To perform this task in a comfortable manner to the patient, the employed device must be small and wireless. One way to reduce size is to remove the device’s battery and use an energy harvesting technique or transfer power from an external source [1]. The latter method permits higher power levels at the device. There are several ways of transmitting power and one way of doing it is through electromagnetic waves, in which the transmitted power levels are mainly limited by tissue damaging [2].

There are several kinds of biomedical-signal sensors, and a simple one that can be readily implemented in CMOS technology is the temperature sensor. It is an important sensor for biomedical applications because temperature is a relevant parameter to diagnose many diseases.

One example of a well-known device that implements power transfer and communication to an external device is the fully-passive RFID tag. In this scheme, illustrated in Fig. 1, the external device, called reader, sends power to the tag and the information is sent back from the tag through the reflected signal modulation [3].

This work presents an RF-powered temperature sensor suitable to measure human body temperature (35 to 42°C), developed in a standard 130-nm CMOS technology and designed and simulated with the Cadence environment. Based on the RFID scheme, power and information are transferred between reader and sensor with a 900-MHz signal. The system architecture is presented in the next section, the design of each block and some simulation results are presented in Sec. III, a full-system simulation is discussed in Sec. IV, partial measurement results are presented in Sec. V and, finally, the work is concluded in the last section.

II. SYSTEM ARCHITECTURE

Fig. 2 shows the system block diagram. The RF front end, formed by the impedance matching, rectifier and backscattering device, is responsible for receiving and rectifying the input signal and send information about the measured temperature back to the reader. The remaining blocks do the power management and sense the temperature.
The impedance matching block was designed to provide a 50-Ω input impedance to match our 50-Ω based measurement setup. Next to this block, a rectifier is used to generate a DC supply voltage $V_{dc}$ while charging the $C_s$ capacitor, which is responsible to provide energy autonomy to the system. Since the input signal power can vary, $V_{dc}$ can increase and damage the circuits, so a voltage limiter is employed to limit its value. A mode selector is used to monitor $V_{dc}$ and, based on its value, switch the system between standby and active mode. A voltage regulator is employed to generate the supply voltage $V_{dd}$ of the sensor core. This core is formed by a reference source, which is the most temperature sensitive circuit in the system, and an oscillator. The oscillator is biased with the reference current $I_{ref}$ generated by the source, thus providing a signal with temperature dependent frequency. This signal controls the backscattering device, which modulates the input impedance and, consequently, the reflected signal. Through this reflected signal, the reader device obtains information about the measured temperature.

III. CIRCUIT DETAILS

In this section, the design of each block along with its simulation is presented. The reader device was not implemented in this work, so we considered the specifications of the reader presented in [4] to design the sensor device.

A. Backscattering Device

The backscattering device is responsible for modulating the input impedance, thus modulating the reflected signal amplitude and phase [3]. This technique is also known as Load Shift Keying (LSK) modulation.

The backscattering device circuit is presented in Fig. 3, along with other elements. It is composed by transistor $M_1$, which works as a switch, ideally commuting the input impedance $Z_{in}$ between 50 and 0 Ω. When $Z_{in} = 50$ Ω, there is no reflected signal, and when $Z_{in} = 0$ Ω, the reflected power is at its maximum. This scheme was chosen to increase the maximum distance between reader and sensor, because the reflected signal is the maximum possible.

Transistor $M_1$ has width $W = 50 \mu m$ and length $L = 120$ nm, minimum channel length of the used technology. Fig. 4 shows the S11 simulation results. Considering the reader specifications [4], the Friis transmission equation [5], the losses on the backscattering device and that our system uses a typical RFID antenna of 2 dBi gain [3], the minimum input signal power is approximately $-10$ dBm and the maximum distance from reader to sensor is 15 m.
B. Rectifier and Impedance Matching

The rectifier topology is presented in Fig. 5 and it is composed of 16 stages that are based on the Greinacher voltage doubler [6]. The transistors in diode configuration are native transistors, enabling the rectifier operation with a lower input signal power. The rectifier was partially designed using the method presented in [7], for an output voltage $V_{dc} = 1$ V, load current $I_{dc} = 10 \ \mu A$ and input power $P_{av} = -10$ dBm. We obtained 10% of power conversion efficiency (PCE), which is defined by:

$$\text{PCE} = \frac{P_{dc}}{P_{av}}, \quad (1)$$

where $P_{dc}$ is the output power and $P_{av}$ is the available power at the input, i.e., the input power if the input impedance is perfectly matched to the source impedance.

The L-match network, composed by $L_M$ and $C_M$ in Fig. 3, was designed to match the input impedance of the rectifier, $219 - 291 \ \Omega$, to 50 $\Omega$. This network was designed considering the parasites extracted from layout and the chosen values of $L_M$ and $C_M$ were 12 nH and 1.68 pF. With input and output conditions mentioned before, the simulated nominal $S_{11}$ at 900 MHz is $-35$ dB. A Monte Carlo simulation with 1000 samples was performed, resulting in a worst case $S_{11}$ equal of $-15$ dB.

C. Voltage Limiter

The voltage limiter circuit schematic is presented in Fig. 6 and it is based on the limiter proposed in [8]. This circuit consumes the excess current generated by the rectifier, preventing $V_{dc}$ from increasing to high values. It was designed to limit $V_{dc}$ to 1.6 V, which is the maximum supply voltage permitted by the employed technology.

For $V_{dc}$ increasing, and when $V_{dc}$ is high enough for transistors $M_1$-$M_4$ to conduct current, the voltage drop on $R_1$ increases. When the gate voltage of $M_5$ is low enough, the current that flows through $M_5$ increases, thus increasing the gate voltage of transistor $M_6$. In this final stage, the current that flows through $M_6$ is higher than the current of the other branches, limiting the $V_{dc}$ at higher input power levels.

The values of resistors $R_1$ and $R_2$ are 20 and 400 k$\Omega$, respectively. The transistor dimensions for all transistors used in this circuit are $W = 10 \ \mu m$ and $L = 1 \ \mu m$. Simulation results of the voltage limiter connected to the rectifier (with impedance matching) are presented in Fig. 7. The circuit limits $V_{dc}$ to values not greater than 1.6 V for a wide input power range.
D. Mode Selector

The voltage and current reference source starts up only when its supply voltage $V_{dd}$ is greater than a certain value $V_{stable}$, i.e., the temperature sensor works correctly only when $V_{dd} > V_{stable}$. So it is interesting to have a block that monitors $V_{dc}$ and turns the sensor on and off at $V_{dc}$ values greater than $V_{stable}$ to not produce an incorrect output. To perform the task, the comparator with internal hysteresis [9] presented in Fig. 8 is employed to compare a reference voltage $V_{ref}$ to a voltage $V_{cres}$ that increases with $V_{dc}$. For an increasing $V_{dc}$, when $V_{cres}$ is greater than $V_{ref}$ by certain value, the system turns on (starts active mode) at $V_{dc} = V_{on}$. After the system is in active mode and $V_{dc}$ is decreasing, when $V_{cres}$ is lower than $V_{ref}$ by certain value, the system turns off, which happens at $V_{dc} = V_{off}$. When the system is in active mode, the mode selector output $V_{ctr}$ is high (follows $V_{dc}$), otherwise (standby mode) $V_{ctr}$ is low. This behavior is shown in Fig. 9 along with the input voltages curves.

Transistor $M_3$ is biased in sub-threshold region, providing a low bias current for the comparator. The transistors dimensions used in this circuit are shown in table I. The value of the current mirror ratio B, which controls the comparator hysteresis, is 15. We chose B to set the values of $V_{on}$ and $V_{off}$ to 1.05 and 0.75 V, respectively. A Monte Carlo simulation was also performed and, among 1000 samples, not a single sample presented $V_{off}$ lower than $V_{stable}$, which is 0.6 V, as can be seen in Fig. 9(a). The current consumption of this circuit is 34 nA at $V_{dc} = 1 V$.

The increasing voltage $V_{cres}$ generator circuit is presented in Fig. 10. To relate $V_{cres}$ to $V_{ref}$, the inputs $V_{b0}$ and $V_{b1}$ come from the reference source in order to mirror its current. The transistors $M_{14}$-$M_{16}$ are connected as diodes to prevent current from flowing through R at the beginning and keeping $V_{cres}$ initially lower than $V_{ref}$. This circuit's transistors dimensions are shown in table I. The resistance of R is 512 kΩ.

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Table I. Mode selector and increasing voltage generator’s transistors dimensions

<table>
<thead>
<tr>
<th>Transistor</th>
<th>W (μm)</th>
<th>L (μm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>$M_1$, $M_2$, $M_8$, $M_9$</td>
<td>120</td>
<td>0.42</td>
</tr>
<tr>
<td>$M_3$</td>
<td>80</td>
<td>0.5</td>
</tr>
<tr>
<td>$M_4$, $M_5$</td>
<td>1</td>
<td>5</td>
</tr>
<tr>
<td>$M_5$, $M_6$</td>
<td>15</td>
<td>5</td>
</tr>
<tr>
<td>$M_{10}$, $M_{11}$</td>
<td>2</td>
<td>0.5</td>
</tr>
<tr>
<td>$M_{12}$, $M_{13}$</td>
<td>18</td>
<td>1.2</td>
</tr>
<tr>
<td>$M_{14}$, $M_{15}$</td>
<td>8</td>
<td>5</td>
</tr>
</tbody>
</table>

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Figure 8. Mode selector schematic

Figure 9. Mode selector simulated (a) inputs and (b) output for $V_{dc}$ increasing and decreasing, direction indicated by the arrows

Figure 10. Increasing voltage generator circuit
E. Voltage Regulator

A voltage regulator is necessary to suppress the wide $V_{dc}$ swings influence on the sensor output. We chose to use the Low Dropout (LDO) regulator topology shown in Fig. 11 due to its low voltage drop from the input $V_{dc}$ to the output $V_{dd}$ [10]. The regulator is switched on/off by the mode selector. Since the reference voltage $V_{ref}$ is obtained from the reference source, powered by the regulator output, it is important that $V_{dd}$ follows $V_{dc}$ when the regulator is off to correctly initialize the reference source.

The operational amplifier employed in the regulator is presented in Fig. 12 and it is switched on/off by $V_{ctr}$. When it is off ($V_{ctr}$ is low), its output voltage is low and current flows through transistor $M_P$. When $V_{ctr}$ is high, the circuit works as an amplifier. This amplifier does not have internal compensation and the regulator is stable due to its feedback network and load capacitor $C_O$. The regulator total power consumption is 2.6 $\mu$W when turned on. The transistor parameters used in this circuit are presented in table II. The capacitor $C_O$ has 50 pF and the resistors $R_1$ and $R_2$ have 3 and 0.5 M$\Omega$, respectively.

F. Temperature Sensor

The threshold-referenced current source presented in Fig. 13 [11] is the most temperature-sensitive circuit in the system and it biases the current-starved ring oscillator, presented in Fig. 14. These two blocks form a temperature sensor which output is a signal that has temperature-dependent frequency.

Table II. Parameters of components used in the LDO regulator

<table>
<thead>
<tr>
<th>Transistor</th>
<th>W (μm)</th>
<th>L (μm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>$M_1$, $M_2$</td>
<td>15</td>
<td>0.8</td>
</tr>
<tr>
<td>$M_3$, $M_4$, $M_5$</td>
<td>3</td>
<td>1.2</td>
</tr>
<tr>
<td>$M_6$</td>
<td>0.16</td>
<td>3</td>
</tr>
<tr>
<td>$M_8$</td>
<td>0.32</td>
<td>3</td>
</tr>
<tr>
<td>$M_9$, $M_{10}$</td>
<td>6</td>
<td>1.2</td>
</tr>
<tr>
<td>$M_{11}$</td>
<td>2</td>
<td>0.5</td>
</tr>
<tr>
<td>$M_{12}$, $M_{13}$</td>
<td>16</td>
<td>1.2</td>
</tr>
<tr>
<td>$M_{14}$</td>
<td>4</td>
<td>0.5</td>
</tr>
</tbody>
</table>

Figure 11. LDO voltage regulator topology

Figure 12. Operational amplifier circuit used in the LDO regulator

Figure 13. Threshold-referenced current source

Figure 14. Current-starved ring oscillator with level shifter
The reference source nominal current and voltage outputs with respect to temperature and supply variations are presented in Fig. 15. The reference current temperature coefficient is \( T_{C_i} = -1.22 \, \%/{ }^\circ C \) and its value at \( T = 38.5^\circ C \) (middle of the temperature range) is 613 nA. The source starts working with \( V_{dd} = V_{stable} = 0.6 \, V \).

At 38.5\(^\circ\)C, the oscillator frequency is 660 kHz and its temperature coefficient is \(-0.78 \, \%/{ }^\circ C\). The transistor \( M_{10} \) switches the oscillator on or off. Transistors \( M_{13}-M_{16} \) increase the output peak-to-peak voltage to \( V_{dc} \), allowing a greater power reflection of the backscattering device when \( V_{bks} \) is high.

The transistor dimensions used in the oscillator and reference source are presented in Table III. The capacitor \( C_1 \) used in the startup circuit has 1 pF capacitance and the resistors \( R_1 \) and \( R_2 \) have nominal values of 270 and 512 k\( \Omega \), respectively. The power consumption of these two circuits combined is 2.8 \( \mu W \).

The variable resistor is composed by 8 series resistors, of which 7 can be short circuited through switches, as shown in Fig. 16(a). Each variable

I. Calibration

Due to fabrication process variations, the sensor output will change. To match the output frequency versus temperature curve to the nominal one, we designed a two-stage calibration method.

The first stage consists in calibrating the TC through the variation of resistor \( R_1 \), since \( T_{C_i} \) can be approximated by [11]:

\[
T_{C_i} = \frac{1}{V_{f_1}} \frac{\partial V_{f_1}}{\partial T} - \frac{1}{R_1} \frac{\partial R_1}{\partial T} = T_{C_{V_{f_1}}} - T_{C_{R_1}} \quad (2)
\]

The second stage consists in calibrating the absolute value of the frequency at \( T = 38.5^\circ C \) through the variation of transistors \( M_{11}-M_{12} \) aspect ratios, i.e., a modification of the current mirror ratio. Ideally, this stage does not modify \( T_{C_i} \) and the resulting \( I_{bias} \) curve will be fitted to the nominal one. The calibration of \( I_{bias} \) current calibrates the output frequency, because the frequency is approximately proportional to the bias current [12].

The variable resistor is composed by 8 series resistors, of which 7 can be short circuited through switches, as shown in Fig. 16(a). Each variable

Table III. Dimensions of transistors used in the reference source and the oscillator

<table>
<thead>
<tr>
<th>Transistor</th>
<th>W (( \mu m ))</th>
<th>L (( \mu m ))</th>
</tr>
</thead>
<tbody>
<tr>
<td>( M_1 - M_3 )</td>
<td>0.5</td>
<td>0.5</td>
</tr>
<tr>
<td>( M_{10} - M_{12}, M_{13} - M_{18} )</td>
<td>18</td>
<td>1.2</td>
</tr>
<tr>
<td>( M_{10} )</td>
<td>0.5</td>
<td>0.12</td>
</tr>
<tr>
<td>( M_{13} )</td>
<td>1</td>
<td>0.12</td>
</tr>
<tr>
<td>( M_{14} )</td>
<td>2</td>
<td>0.12</td>
</tr>
<tr>
<td>( M_{15} )</td>
<td>1.5</td>
<td>0.12</td>
</tr>
<tr>
<td>( M_{16} )</td>
<td>0.5</td>
<td>0.12</td>
</tr>
<tr>
<td>( M_{17} - M_{21} )</td>
<td>8</td>
<td>5</td>
</tr>
<tr>
<td>MP</td>
<td>3</td>
<td>12</td>
</tr>
<tr>
<td>MN</td>
<td>1</td>
<td>12</td>
</tr>
</tbody>
</table>

Figure 15. Reference source outputs for (a) temperature and (b) supply voltage variations

Figure 16. Variable (a) resistor and (b) current mirror used to calibrate the sensor
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transistor is composed by 13 shunt transistors as shown in Fig. 16(b), forming current mirrors. These shunt mirrors can be activated by the series switches, modifying the ratio of the equivalent current mirror. The R and $R_0$ values are 4 and 50 KΩ, respectively, and the aspect ratio $S$ and $S_0$ values are $7 \times 10^{-4}$ and 0.3, respectively.

We did a 100-sample Monte Carlo simulation with extracted parasitics and calibrated each sample. Due to time limitations, these simulations were performed considering only the reference current. The results are presented in Fig. 17. A total of 96% of the samples presented less than 0.2°C of conversion error.

IV. FULL-SYSTEM SIMULATION

We did full-system transient simulations with extracted parasitics from layout. The results are presented in Figs. 18 and 19.

The waveforms of the oscillator output $V_{bks}$, the rectifier output $V_{dc}$ and the voltage regulator output $V_{dd}$ are presented in Fig. 18. These waveforms were obtained at $T = 38.5°C$ and with an input-signal source that had 900-MHz frequency and 50-Ω impedance. After a start-up time of 35 μs, the oscillator produces a 627.3 kHz signal. The ripple of $V_{dc}$ after 35 μs is due to the backscattering device operation, i.e., there is a $V_{dc}$ drop because the input power is reflected when $V_{bks}$ is high.
Fig. 19 presents the frequency of $V_{\text{bks}}$ in simulations at different temperatures and with several input power levels. Due to time limitations, in these simulations the rectifier with impedance matching was modeled as an ideal current source. For $P_{\text{av}} > -9.2$ dBm, the frequency variations produce negligible conversion errors.

V. MEASUREMENTS RESULTS

Some of the circuits described in this paper were sent to fabrication and tested. In this section, the measurement results are presented.

A. Rectifier

The rectifier with impedance matching circuit and the $C_S$ capacitor are presented in Fig. 20. The RF input probe is on the left side of the photograph and the DC output probe is on the right side.

A 100-kΩ resistor is placed at the rectifier output to act as a 10 μA load current at 1 V output voltage. As expected, the rectifier’s PCE is approximately 10% at −10 dBm input signal power. In Fig. 21, the measured S11 is presented. At 900 MHz, we measured S11 equal to −27.4 dB. There is a difference between the S11 obtained through measurements and nominal simulation, although this difference is inside the limits predicted by the Monte Carlo simulation.

B. Voltage Limiter

The measurement results for the voltage limiter circuit are presented in Figs. 22(a) and 22(b) in linear and logarithmic scales, respectively. The results are close to the simulated and at $V_{\text{DD}} = 1$ V the current consumption is 2 μA.

C. Backscattering Device

To test the backscattering device separately from the other circuits, we placed a 50-Ω on-chip resistor in parallel to it. Fig. 23 presents the S11 curve for a sweep of the control voltage $V_{\text{bks}}$. The measured S11 when $V_{\text{bks}} = 0$ V is −24.4 dB, which is far from simulated...
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VI. CONCLUSION

The design and partial testing of an integrated RF-powered temperature sensor was discussed. The developed system operates in two modes: standby, in which the system consumes 4.9 μA, and active, in which the current consumption is 8.5 μA. With the minimum input signal specification, the rectifier works with 10% efficiency, providing 10 μA output current. The active area of the combined layouts of all blocks is approximately 0.34 mm². The largest layout is the rectifier block one (0.26 mm²), which includes an inductor (from the impedance matching circuit) and the storage capacitor. The sensor has a temperature coefficient of −0.78 %/°C and maximum conversion error, after calibration, of 0.2°C.

In table IV we compare our work to recent RF-powered temperature sensors, developed with similar technology and input signal frequency. We managed to develop a system that occupies a small area, if compared to the others, while also containing circuits for sensor calibration.

Table IV. Comparison with recent publications

<table>
<thead>
<tr>
<th>Reference</th>
<th>Technology (nm)</th>
<th>Input Frequency (MHz)</th>
<th>Area (mm²)</th>
<th>Standby power consumption (μW)</th>
<th>Active power consumption (μW)</th>
<th>RF-DC efficiency (%)</th>
<th>Minimum input power (dBm)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>250</td>
<td>450</td>
<td>1.2</td>
<td>5</td>
<td>1500</td>
<td>-</td>
<td>-12.5</td>
</tr>
<tr>
<td></td>
<td>130</td>
<td>900</td>
<td>0.95</td>
<td>-</td>
<td>9</td>
<td>30*</td>
<td>-12</td>
</tr>
<tr>
<td></td>
<td>130</td>
<td>868</td>
<td>3.96</td>
<td>-</td>
<td>-</td>
<td>8.6</td>
<td>-10.3</td>
</tr>
<tr>
<td></td>
<td>180</td>
<td>910</td>
<td>1.2</td>
<td>-</td>
<td>-</td>
<td>35*</td>
<td>-5</td>
</tr>
</tbody>
</table>

REFERENCES


