A 2.93μW 8-Bit Capacitance-to-RF Converter for Movable Laboratory Mice Blood Pressure Monitoring

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Abstract—**An ultra-low-power capacitance-to-RF (C/RF) converter for laboratory mice blood pressure monitoring is proposed. Unlike the conventional design involving capacitance-to-analog (C/A) conversion followed by analog-to-digital (A/D) conversion, the proposed front-end is a direct capacitance-to-digital (C/D) converter that can simplify the hardware while saving both power and area. The C/D converter also features an automatic capacitance-range finder mapping the input capacitance range to the full scale of the digitization core. The generated digital data is compressed before driving the back-end RF transmitter, which is based on a power-ON/OFF VCO with direct FSK modulation operated at the 915-MHz ISM band. Optimized in 65-nm CMOS, the simulated 8-bit 6.4-kSa/s C/RF converter exhibits a 7.5 effective number of bit (ENOB) and a ~0.7 V_{pp} output swing at the RF transmitter output, while drawing 2.93 μW of power. The DNL and INL are ±0.125 and ±0.188 LSB, respectively. The attained capacitance sensing resolution is equivalent to 1.25 fF/LSB.**

I. INTRODUCTION

Treatment tests on animals are essential parts in pharmacology. The responses from the animals imply indicative consequences on developing new treatments and medicine. Among the available kinds of animals used in the research stage, mice and rats are the most popular. Several reasons for this phenomenon are: 1) they are easy to handle and maintain; 2) they reproduce fast which is critical for animal research as impacts on several generations of mice are entailed to obtain the desirable strains with targeted genetic sequences; and 3) they share a high degree of homology with humans, having, as well, a completely sequenced genetic generation. Among all the biological information, blood pressure is one of the most important vital signals [1]. Many physiological factors can affect the blood pressure such as blood volume, blood vessel resistance (related to length and diameter of it) and blood viscosity. Also, the heart beat rate of the mouse can be known if the blood pressure waveform can be obtained.

Traditional ways to measure blood pressure include invasive catheter-tip [2] or tail-cuff [3]. Regrettably, it either requires complex procedure or causes stress-induced signal distortion. In addition, the mouse cannot move freely due to wired electronics. Recently, a batteryless blood pressure monitoring microsystem with wireless data telemetry was proposed in [4]. It exploits an immersed MEMS capacitive pressure sensor to enhance the level of integration, capturing the biological information from a free roaming laboratory mouse. The MEMS sensor signal was first converted from capacitance-to-analog (C/A), and then digitized by an analog-to-digital (A/D) converter. Such two conversions draw totally 36 μW of power which are still too high for implantable applications. Also, the RF transmitter (TX) is operated at a low data rate and was always ON [4], dominating the system power (i.e., 240 out of 300 μW).

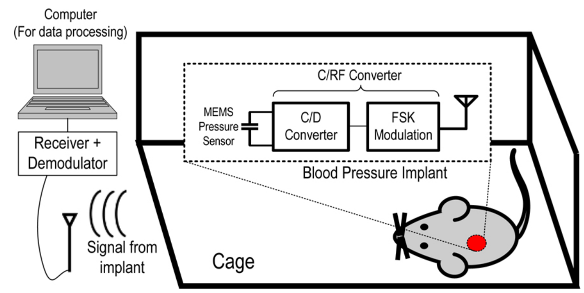

Fig. 1 The wireless mouse arterial blood pressure measuring system.

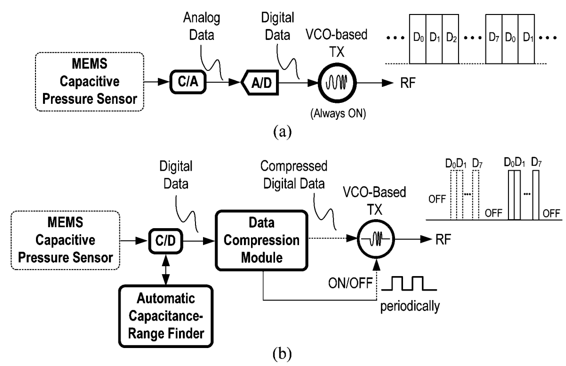

Fig. 2 Block diagram of C/RF converters: (a) [4]. (b) Proposed.

This paper proposes a novel ultra-low-power capacitance-to-RF (C/RF) converter composed of an 8-bit capacitance-to-digital (C/D) converter and a FSK modulation block for a movable laboratory mice blood pressure monitoring system, as illustrated in Fig. 1. As depicted in Fig. 2, instead of using the C/A followed by the A/D approach proposed in [4] (Fig. 2(a)), the input signal is directly digitized using a C/D converter, which also features an automatic capacitance-range finder to match the range of the blood pressure to the full scale of the C/D converter (Fig. 2(b)). A data compression module lowers significantly the power-ON time of the TX, which is based on a power-ON/OFF voltage-controlled oscillator (VCO) with FSK modulation. Although the resolution of the C/D converter is just 8 bits ([4] is 11 bits), the system power is reduced by around 100x (i.e., 2.93 μW) when comparing with [4], rendering it as a promising candidate for real-time long-term monitoring.

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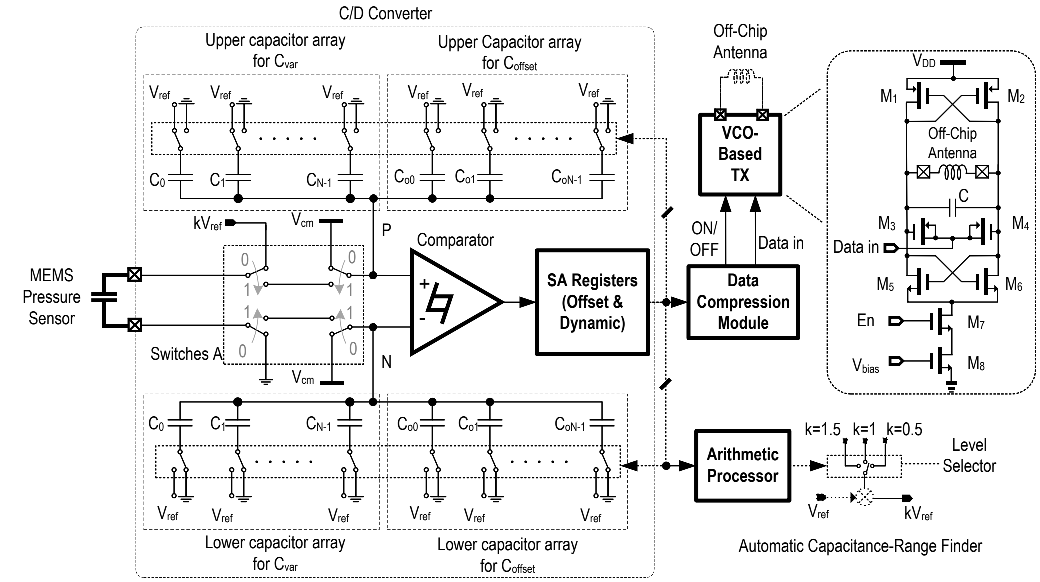

Fig. 3 Schematic of the proposed C/RF converter.

II. SYSTEM ARCHITECTURE

The Schematic of the proposed C/RF converter is shown in Fig. 3. It includes a C/D converter that interfaces with the off-chip MEMS capacitive pressure sensor, a data compression module, a power-ON/OFF VCO-based TX and an automatic capacitance-range finder. The details are described as follows:

A. C/D Converter

The C/D converter is responsible to directly digitize the capacitance change from the MEMS sensor C_{MEMS} to avoid using explicit C/A followed by A/D converter topology. Moreover, the C/A converter in [4] was generally built by an operational amplifier that consumes static power even if the system can be powered down. The C/D converter is therefore a more promising alternative. A differential architecture is employed to avoid the variation of C_{MEMS} during digitization, thus, not affecting the conversion results [5].

C_{MEMS} can be partitioned into two parts: static offset and dynamic capacitances. Only the dynamic will reflect the blood pressure value while the offset part will only degrade the useful resolution if no offset cancellation is employed. An extra offset capacitor array C_{offset} is used, together with the capacitor array C_{var}, to compensate the effect of significant MEMS sensor offset capacitance and improve the usable resolution for blood pressure A/D conversion. The C/D converter will initially determine C_{offset} of the MEMS sensor through calibration. In the sampling phase, all of the switches, in the set of switches *A*, will be set to position 0, and all the bottom plates of the capacitors in the upper and lower capacitor arrays will be connected to V_{ref} and ground, respectively. *k* is a scaling factor to match the range of the capacitor array to the pressure sensor and will be described later.

The calibration of C_{offset} is initiated by charging the pressure sensor and capacitor arrays in the sampling phase. In addition, the comparator offset is also sampled by connecting both inputs to V_{cm}. The C/D converter will enter the conversion phase by turning switches A to position 1. The MSB of the upper (lower) offset capacitor array $C_{\text{oN-1}}$ will then be switched to ground (V_{ref}), where *N* is the number of bits of the C/D converter. The comparator determines whether kC_{offset} is larger or smaller than $C_{\text{oN-1}}$ by comparing the voltage difference between *P* and *N*. If $kC_{\text{offset}} > C_{\text{oN-1}}$, $V_{\text{p}} > V_{\text{n}}$ and the SA register will connect capacitor $C_{\text{oN-1}}$ in the upper (lower) offset capacitor array to ground (V_{ref}). Otherwise, the SA register will switch $C_{\text{oN-1}}$ of the upper (lower) offset capacitor array back to V_{ref} (ground). Subsequent conversion steps will be carried out until all of the bits are obtained. The SA registers for offset calibration will store the result for future conversion.

After C_{offset} has been calculated, the C/D converter is able to digitize C_{var} without deterioration caused by the offset capacitance. The operation is similar to the case for offset capacitance calibration with the difference that the offset SA register will reload the offset capacitance result and flip the corresponding switches during the conversion phase to cancel the MEMS sensor offset capacitance. In our targeted system, the sensor mean offset capacitance is 2 pF and the sensitivity is 1 fF/mmHg [4]. C_{offset} can be set within a range from 1.6 pF to 2.4 pF to cover a ±20% variation in the MEMS sensor offset capacitance, with a resolution of 3.125 fF/LSB. C_{var} is set to 320 fF (for *k=1*) with a resolution of 8 bits (1.25mmHg/LSB) to account for margins of hypertension of the object under observation, as well as residue offsets due to resolution mismatch between C_{var} and C_{offset}. Both capacitor arrays are implemented using metal plate capacitors to achieve fF resolution. A two-stage weighted capacitor network (not shown) is employed to reduce the required total capacitance, and capacitor mismatch can be resolved by using a calibration technique similar to [6].

B. Power-ON/OFF VCO-Based TX

When the comparison results are obtained, they should be delivered to the external receiver. In [4] the TX stays ON even if no data transmission is required. This unnecessary overhead heavily reduces the system power efficiency as the TX is the most power-hungry block.

In this system, FSK modulation is employed for its simplicity. The carrier frequency is chosen to be 915 MHz, which is an ISM band. The total available bandwidth in this band is 26 MHz. If the sampling rate of the blood pressure sensing system is set to 6.4 kSa/s with 8 bits, then the original

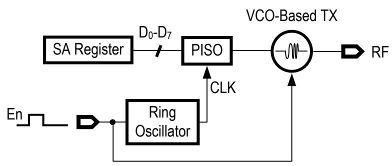

Fig. 4 Architecture of the proposed VCO-based TX.

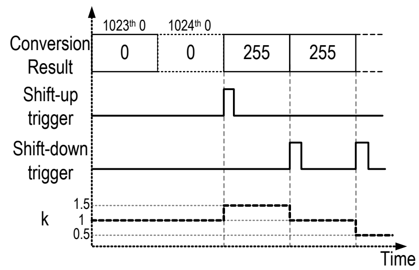

Fig. 5 Timing diagram of the automatic capacitance-range finder.

data rate will be 48 kbps. However, this is much lower than the bandwidth of the carrier frequency and it wastes much RF resources. If the data is compressed in time (i.e., higher data rate) and emitted only once in a while, the TX power can be better managed and multiple channel transmission is possible. In addition, since the data can be sent within a shorter period of time, when the next batch of data (conversion results) is not available yet, the TX can be turned OFF to save power. The structure of the VCO is shown in Fig. 4. At the end of the conversion, an enabling signal *En* will be triggered and sent to a ring oscillator (RO) and the VCO to start oscillation. The RO will generate a *CLK* signal to the parallel-in serial-out registers (PISO). The conversion results from the SA register will then be loaded to the PISO before the data are wiped. The data will be sent to the VCO at a higher data rate which depends on the frequency of *CLK*. The VCO, with its structure highlighted in Fig. 3 already, is built by two pairs of cross coupling transistors M_1-M_2 and M_5-M_6 which provide the necessary negative resistance ($-1/g_m$) to compensate for the parasitic resistances of the inductor and capacitor, exhibiting oscillation at:

$$f_{osc} = \frac{1}{2\pi\sqrt{L(C + C_{MOS})}} \tag{1}$$

where C_{MOS} is the series capacitance of two varactors M_3-M_4. It will be changed according to the voltage applied to *Data In* to achieve FSK modulation. M_7 is used to control the state of the TX. If *En*=0, the TX is OFF. The external inductor *L*, which can give a higher quality factor than its on-chip counterpart [7], is used as the loop inductor. It also acts as a loop antenna [4] to emit the data. In addition, the inductance can be trimmed to match the frequency in ISM band even after fabrication. The inductance is set around 82 nH. This is chosen as the self-resonant frequency of the commercial available inductor will drop below a few GHz when the inductance is larger than this value [8]. An initialization signal should be sent before the end of the conversion to notify the receiver to accept the data. This can be achieved by a preamble right before the conversion results. This preamble can provide a reference frequency for synchronization as well as data retrieval and storage.

The duration of *En* must ensure that all of the data will be sent to the receiver before the TX was turned off. A counter can be used to count the number of *CLK* cycles to turn off *En* appropriately. In addition, since the VCO needs time to start oscillation after *En* triggered from "0" to "1", a delay for few hundred ns is required between the start of VCO oscillation and the data retrieval from the PISO. This delay depends on the settling time of the VCO and can be estimated through simulation (section III).

C. Automatic Capacitance-Range Finder

As proposed in [5], a scaling factor *k* is used to control the sensing range of the C/D converter to avoid signal distortion as well as to increase the corresponding resolution as C_{VAR} varies. Here, an automatic capacitance-range controller to find the suitable value of *k* for system performance optimization is proposed. To reduce circuit complexity, continuous modeling for *k* is not implemented. In addition, the receiver should be able to recognize the value of *k* from the conversion result so it can correctly read out the capacitance value. Thus, 3 discrete values of $k = 0.5$, 1 and 1.5 are implemented as shown in Fig. 3. Originally, *k* is set to 1 when the system is powered up. The arithmetic processor will keep observing the conversion results from the C/D converter. If the result is too large (all of the bits are "1"), the arithmetic processor will shift down the value of *k* by one level (from 1.5 to 1 or from 1 to 0.5). The algorithm for the shift up process is more complex than the shift down case. Since the blood pressure can vary in a cycle, it is possible that the diastolic pressure is below the threshold of the capacitance-range finder but is actually working properly within the range. To address this issue, a counter was employed in this design to count the number of conversion results that are under the threshold. The normal heart beat rate of a laboratory mouse is around 600 beats per minute, which is equivalent to a beat interval of 0.1 second. For a converter with sampling rate of 6400 samples per second, the number of samples between beat intervals is 640. Based on this observation, it can be concluded that if the output codes are below the critical point for a cycle of blood pressure, the variable *k* should be shifted up (from 0.5 to 1 or from 1 to 1.5) to increase the dynamic range of the converter. For simplicity, the number of counts to trigger the shift up action is set to 1024. The threshold code of the shift down process was set to 96 so that if the output code of 1024 consecutive conversion results is below 96, *k* will be shifted down by one unit. These sets of operation are illustrated in Fig. 5. Similarly, the processor on the receiver can also decide the value of *k* by using the same algorithm stated above. By doing it, the value of *k* can be obtained in the receiver even without direct transmitting through the antenna.

III. Simulation Results

The C/RF converter was simulated in 65-nm CMOS. The power consumption of the VCO-based TX is reduced from 18.7 to 0.91 μW with the power-ON/OFF and data compression techniques at $V_{DD} = 1.5$ V and 3.33 Mbps. This corresponds to a 95% power reduction.

The standalone C/D converter is assessed under the conditions of $C_{offset} = 1.6$ pF, $V_{ref} = 1$ V, $V_{cm} = 0.5$ V, and $C_{var} = 320$ fF. The effective number of bits (ENOB) was evaluated by using a sinusoidal input capacitance variation created in Verilog-A. Fig. 6 shows the ENOB of the C/D converter with three values of *k*. The ENOBs stay around 7.5 bits up to the Nyquist frequency of 3.2 kHz. This implies that the differential architecture maintains its accuracy even when the capacitance changes at a rate close to the Nyquist frequency. In addition, as

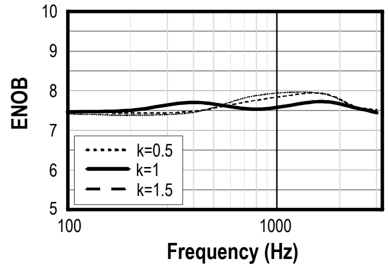

Fig. 6 ENOB of the C/D converter with different values of k.

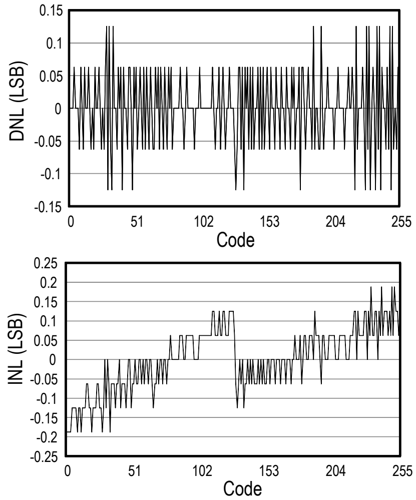

Fig. 7 DNL and INL of the C/D converter when k = 1.

TABLE 1 COMPARISON OF THE CAPACITANCE-TO-DIGITAL CONVERTER

	Bits	Freq.	C/D power	FOM
This work (Simulation)	8	6.4 kS/s	1.92 µW	1.17 pJ/step
[4] (Measurement)	11	2 kS/s	36 µW	8.79 pJ/step
[9] (Simulation)	9	62.5 kS/s	158.3µW	4.95 pJ/step

the three values of k yield a similar ENOB, the C/D converter will be insensitive to the variation of k. The DNL and INL are ±0.125 and ±0.1875 LSB, respectively, as shown in Fig. 7. The standalone C/D converter dissipates 1.92 µW. If the same MEMS sensor as [4] is employed, which features a sensitivity level of 1 fF/mmHg, the C/D converter equivalently achieves a pressure sensing resolution of 1.25 mmHg/LSB. Table 1 shows the comparison of the C/D converter presented here, with the structures from [4] and [9]. It could be seen that this work achieves the lowest figure of merit (FOM) when compared with [4] and [9], since it eliminates the C/A and A/D converters, rendering it much more favorable for long term monitoring.

The overall C/RF converter draws 2.93 µW, including the TX operating at a data rate of 6.4 kSa/s. The TX delivers an output swing of ~0.7 V_{pp} with 100-ns settling time (Fig. 8). The FSK frequency separation is 20 MHz (at 902 and 922 MHz) as depicted in Fig. 9. Comparing with [4], the achievable power consumption is reduced by ~100x, with a 3-bit lower resolution target in the C/D converter. The data compression and power-ON/OFF TX techniques are generally applicable to other micro-systems.

IV. CONCLUSIONS

This paper has reported a novel ultra-low-power C/RF converter for long-term laboratory mice blood pressure monitoring. Direct C/D conversion minimizes the hardware,

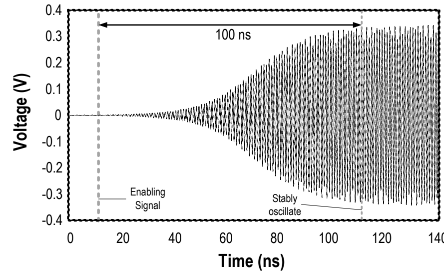

Fig. 8 TX output transient behavior.

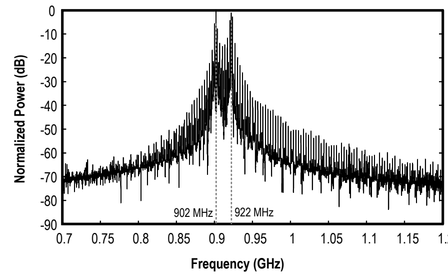

Fig. 9 TX output spectrum with FSK modulation.

resulting in better area and power efficiencies. The data compression module reduces significantly the required power-ON time of the TX. An automatic capacitance-range finder matches the internal capacitance range to the MEMS pressure sensor, avoiding distortion and deterioration of the useful resolution. The simulated ENOB is 7.5 up to the Nyquist frequency of 3.2 kHz. The DNL and INL are ±0.125 and ±0.1875 LSB, respectively. The power consumption is 2.93 µW at 6.4 kSa/s TX output rate. For a 1-fF/mmHg MEMS capacitive pressure sensor, the achieved pressure sensing resolution is equivalent to 1.25 mmHg/LSB.

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