

An Improved-Accuracy Approach for Readout of Large-Array Resistive Sensors

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Abstract—Large-array resistive sensors have found a variety of applications from industrial instrumentation to medical wearable applications. In such large-array sensors, several non-idealities limit the accuracy of the readout circuit. In this paper, after analyzing the crosstalk error caused by the input offset voltage and input bias current of the operational amplifier, a novel double-sampling technique is proposed to considerably reduce the error. Spice simulation results of a 20×20 array confirm that for a 1-M Ω resistance, the proposed technique can decrease the maximum measurable relative error from 20% to 0.089%. Furthermore, the measurement results of a 20×20 array confirm that for a resistance value of 19 M Ω being measured, the proposed technique can reduce the error from 84% to 0.69%.

Index Terms—Large-array sensors, resistive pressure sensors, readout circuits, operational amplifier non-idealities, crosstalk.

I. INTRODUCTION

SENSOR arrays are nowadays employed in a variety of applications including biometric sensing, gas detection, tactile sensing, temperature sensing, electronic nose, thermal imaging based on infrared (IR) sensor, and physical or physiological monitoring [1]–[20]. In such arrays, different sensors including pressure sensors [1]–[14], Hall-effect sensors [15], [16], etc. are employed. Among different types of pressure sensors, i.e. resistive, capacitive and piezo-electric sensors, resistive sensors are more commonly used due to their lower cost, higher robustness, faster response, and simpler read-out circuitry [6]. For simpler readout of large-array resistive sensors, in most recent large-array sensors, there is only access to terminals of the rows and columns (by sharing the rows and columns [3]).

The schematics of a small fraction of a large-array resistive sensor and the corresponding readout circuit are shown in Fig.1. In order to measure the resistance of an individual resistor, e.g. R_{23} in Fig.1, the switch of the corresponding

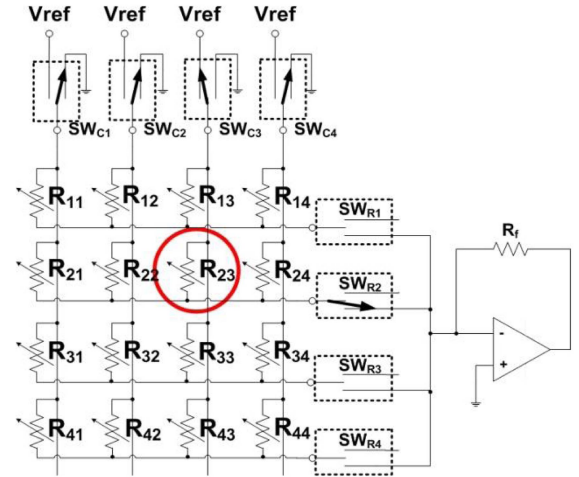


Fig. 1. The schematic of a part of the sensor array and the readout circuit.

column, i.e. SW_{C3} , is connected to V_{ref} and all other column switches are connected to ground. On the other hand, the switch of the corresponding row, i.e. SW_{R2} , is connected to the operational amplifier (Opamp) input and all other row switches are left open. In this way, neglecting all non-ideal effects, the current flowing through R_f will be equal to V_{ref}/R_{23} and the output voltage is thus:

$$V_o = -\frac{V_{ref}}{R_{23}} \cdot R_f \quad (1)$$

In such arrays, due to several non-idealities, the resistance being read out is affected by the resistances of other sensors. This effect is called crosstalk. For instance, a non-zero input offset voltage of the operational amplifier (Opamp) leads to a leakage current being added to the main current of interest (flowing through R_f). The value of this error current is affected by the offset voltage, the array size and the values of the “non-measured” resistances. Therefore, the crosstalk effect becomes more severe in larger arrays or when measuring very large resistances. In [1], an Improved Isolated Drive Feedback Circuit (IIDFC) using one operational amplifier (op-amp) has been designed to reduce the crosstalk caused by the adjacent column elements in resistive sensor array. In [2], it has been proved that the solution of crosstalk and snapshot capability cannot be achieved simultaneously in networked resistive sensor arrays. Furthermore, several techniques have been already reported in order to reduce the crosstalk effect [3]–[9]. For instance, in [3], for each row, an individual Opamp has been used. While the cross-talk effect has been reduced, the technique suffers from increased

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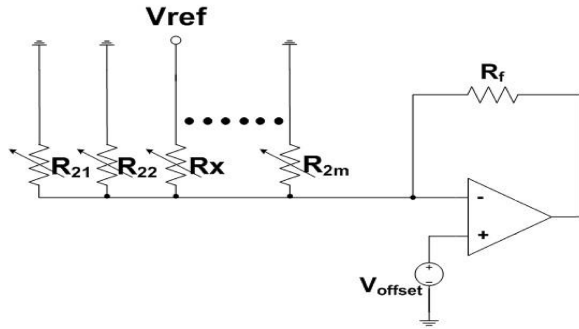


Fig. 2. Equivalent circuit with the Opamp offset modeled R_x .

complexity, power consumption and cost. Although the effects of finite Opamp gain and input impedance are analyzed in [4], nothing has been presented for very-large-array sensors. Another technique is proposed in [5] that reduces the effect of non-zero “on” resistances of the switches. It can be observed that none of the above has either analyzed the effect of the Opamp offset voltage and input bias current or proposed a technique for reducing this effect in large-array sensors. In this paper, however, the effect of two Opamp non-idealities, i.e. its offset voltage and input bias current, will be analyzed and a double-sampling technique to reduce the effect will be proposed. The proposed technique does not impose any additional hardware overhead to the conventional circuit. Since for resistive pressure sensors, the resistance is usually inversely proportional to the pressure, the proposed technique is indeed increasing the pressure *sensitivity* of the sensor. It should be noted that the analysis/proposed technique of this paper can be employed in any kind of large-array resistive sensors.

The rest of the paper is organized as follows. In section II, the effect of non-zero offset voltage and input bias current of the Opamp will be analyzed. It will be shown that this effect is much more pronounced for large-array sensors or the cases where the resistance being measured is much larger than the others. Section III, proposes a novel double-sampling technique effectively reducing the errors caused by the non-idealities mentioned earlier. In Section IV, simulation results of a 20×20 resistive array are presented followed by the measurement results of the array in section V. Finally, section VI concludes the paper.

II. ERRORS DUE TO OPAMP OFFSET VOLTAGE AND INPUT CURRENT

As mentioned above, in practice there are several non-idealities that reduce the accuracy of the aforementioned readout technique, especially for high values of resistors; including (but not limited to) non-zero on-resistances of the switches as well as the non-idealities of the Opamp. Among various non-idealities of the Opamp, i.e. finite DC gain, finite bandwidth and slew-rate, non-zero input offset voltage, and non-zero input bias current, in this paper the effects of the last two will be analyzed.

Fig.2 shows the equivalent circuit when measuring the resistance of R_x with the Opamp input offset voltage modeled as a voltage source, V_{offset} . It can be observed that a non-zero offset voltage for the Opamp causes the voltages across other

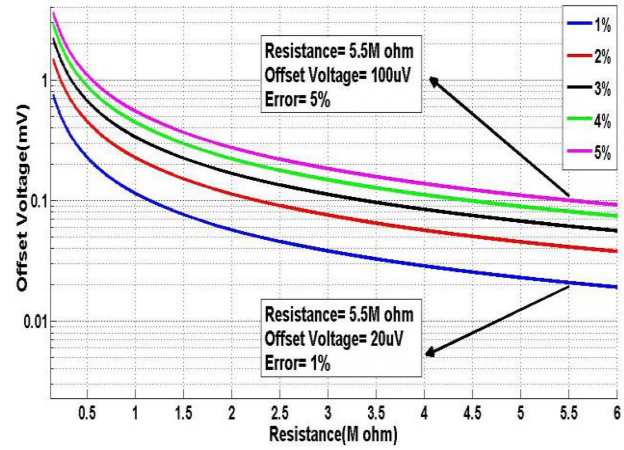


Fig. 3. Required value for the Opamp's offset voltage as a function of R_x for different values of relative error.

resistors and therefore their currents to be non-zero. Therefore, the values of other resistors affect the readout of R_x through a mechanism called crosstalk.

Assuming a very large gain for the Opamp, the output voltage of the circuit shown in Fig. 2 is obtained from

$$V_o = R_f \left(\frac{-V_{ref}}{R_x} + V_{offset} \left(\frac{1}{R_{21}} + \dots + \frac{1}{R_x} + \dots + \frac{1}{R_{2m}} \right) \right) + V_{offset} \quad (2)$$

Therefore the error voltage, V_{err} , will be equal to

$$V_{err} = V_{offset} \times R_f \left(\frac{1}{R_{21}} + \dots + \frac{1}{R_x} + \dots + \frac{1}{R_{2m}} + \frac{1}{R_f} \right)$$

It can be observed that the value of the error voltage due to the Opamp's offset can be considerable for large arrays of resistive sensors or for the case where the value of R_x is very large.

If ϵ is the value of the relative error in the measured value of R_x , then it can be shown that the offset voltage of the Opamp should satisfy

$$V_{offset} \leq \frac{V_{ref}}{R_x \left(\frac{1}{\epsilon} + 1 \right) \left(\sum \frac{1}{R_i} + \frac{1}{R_f} \right)} \quad (3)$$

where R_i 's are the other resistances of the row.

Fig.3 shows the maximum acceptable offset voltage as a function of the value of R_x for different values of ϵ . It has been assumed that $R_f = 24 \text{ k}\Omega$, $V_{ref} = 5 \text{ V}$, $m=40$ and all other resistors are $100 \text{ k}\Omega$. It can be observed that, for instance, for errors of less than 5% and 1% in measuring a resistance of $5.5 \text{ M}\Omega$, the offset voltage should be smaller than $100 \mu\text{V}$ and $20 \mu\text{V}$, respectively.

It should be noted that the value of the offset voltage is not constant and it varies with temperature. Therefore, foreground calibration methods cannot be effectively employed. Even if the value of the offset voltage was known, there are numerous unknowns in the equation expressed in (2) and solving it is troublesome.

It is obvious that any factor changing the value of the current flowing in R_f affects the readout accuracy. Therefore, another Opamp non-ideality reducing the accuracy of the

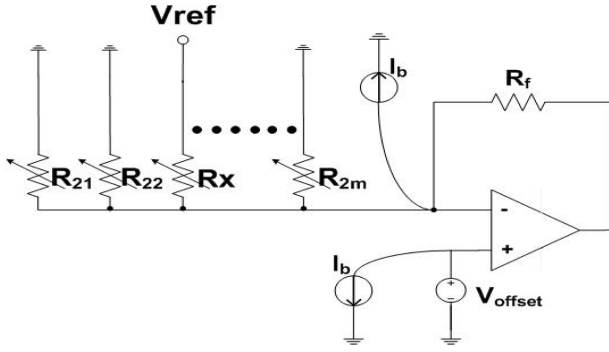


Fig. 4. Equivalent circuit with both the OpAmp's offset voltage and input bias current modeled.

readout technique is the non-zero input bias current of the Opamp. Although this value is negligibly small for most CMOS Opamps, it is not zero for some conventional Opamps used in discrete circuits. It can be shown that the output voltage of the circuit with only the finite input bias current of the opamp as the non-ideality source, is obtained from

$$V_o = R_f \left(\frac{-V_{ref}}{R_x} + I_b \right) \quad (4)$$

If ϵ is the value of the relative error in the measured value of R_x and considering only the effect of the input bias current of the Opamp, then it can be shown that the input bias current of the Opamp should satisfy

$$I_b < \frac{V_{ref}}{R_x(1 + \frac{1}{\epsilon})} \quad (5)$$

It can be observed that in spite of the case for the offset voltage, the effect of the non-zero input bias current of the Opamp does not depend on the size of the array. However, similar to the offset voltage case, for larger values of the measured resistor, the required value for the input bias current of the Opamp is smaller. For instance, an Opamp with an input bias current of 30nA, can measure a resistance of up to 7.94 M Ω with an error of 5% ignoring all other non-ideal factors.

Fig.4 shows the equivalent circuit including both the offset voltage and the input bias current in the model. The output voltage can be calculated from

$$V_o = R_f \left(\frac{-V_{ref}}{R_x} + V_{offset} \left(\sum \frac{1}{R_i} + \frac{1}{R_f} \right) + I_b \right) \quad (6)$$

It can be observed that the errors caused by the offset voltage of the Opamp and its input bias current are “added” to the ideal output. Note that these two terms are repeated in the output voltage if another input rather than V_{ref} is applied as the input. Therefore, in the following section, a double-sampling technique is proposed where the constant term of $V_{offset} \left(\sum \frac{1}{R_i} + \frac{1}{R_f} \right) + I_b$ is subtracted in the digital domain.

III. PROPOSED DOUBLE-SAMPLING READOUT TECHNIQUE

A. Proposed Technique

In the proposed readout technique, once an input voltage of V_{ref} (Fig.5-a) and then another DC voltage (GND or 0V in our

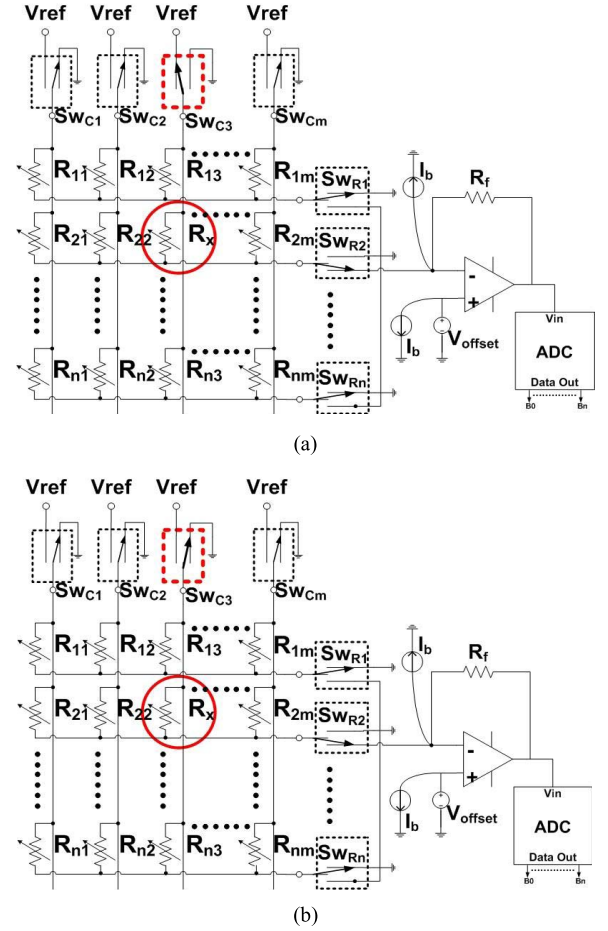


Fig. 5. Proposed double-sampling circuit with the Opamp offset voltage and input bias current modeled: (a) phase 1, (b) phase 2.

proposed system) are applied to the readout circuit (Fig.5-b). The latter is simply done by connecting SW_{C3} to ground.

Then, the digital versions of the output voltages are subtracted in the digital domain. Hence, the result of the subtraction in the digital domain will become:

$$V_o = R_f \cdot \left[\frac{-V_{ref}}{R_x} + V_{offset} \left(\sum \frac{1}{R_i} + \frac{1}{R_f} \right) + I_b \right] - R_f \left[V_{offset} \left(\sum \frac{1}{R_i} + \frac{1}{R_f} \right) + I_b \right] = -R_f \left(\frac{V_{ref}}{R_x} \right) \quad (7)$$

It can be observed that the effects of both the offset voltage and the input bias current are cancelled. It should be noted that the main limitation imposed by the approach is the sampling speed being halved due to the double-sampling nature of the technique. Furthermore, the only complexity (hardware) overhead of the approach is the required resolution of the ADC that might be higher than that of the conventional approach.

B. Design Considerations

In this section most important design considerations will be briefly discussed.

In order for the technique to be effective, the resolution of the employed analog-to-digital converter (ADC) should be high enough to be able to digitize the error voltage with meaningful number of bits. In fact the resolution of the ADC must be high enough to be able to measure the Opamp output voltage when the SWC3 switch is connected to GND and the output voltage is equal to: $-R_f \left(V_{offset} \left(\sum \frac{1}{R_i} + \frac{1}{R_f} \right) + I_b \right)$. In the other word, the main parameter which determines the ADC resolution is the Opamp's output voltage at this phase.

The proposed technique can effectively reduce the effects of the Opamp's offset voltage and input bias current. However, other specifications of the Opamp must satisfy the requirements imposed by the circuit. For instance, the DC gain of the Opamp should be higher than 90dB readily achievable by popular Opamps. As for the required bandwidth and slew rate, for an array of 400 sensors and for a frame rate of 200 frame/sec, a unity-gain bandwidth of at least 125MHz and a slew rate of 180V/ μ s are needed. It should be emphasized, however, that the proposed technique only reduces the circuit speed by a factor of two without tightening the required Opamp specifications.

Since the resistances of the switches are in series with the main resistance, R_x , another non-ideal parameter is the on-resistance of the switches. In this work, we considered a 2 Ω resistance for switches negligibly small compared to the resistances being measured.

Finally, as can be seen in (7), the value of V_{ref} directly affects the measurement accuracy. As a result, before starting the reading procedure, the value of V_{ref} is read by the ADC and stored in the memory to be employed for accurate measurement of resistances.

IV. SIMULATION RESULTS

In order to illustrate the effectiveness of the proposed technique, a 20-row 20-column resistive array has been simulated using Spice. Based on the practical values of an industrial pressure sensor array, the minimum and maximum values of the resistors to be measured are 100k Ω and 1000M Ω , respectively. The objective is to measure the resistance of the sensor located on the 10th row and 10th column while all the other resistors are at their minimum value, i.e. 100k Ω (leading to maximum possible crosstalk). In this simulation, AD8042 has been used as the OpAmp with an offset voltage and input bias current about 2mV and 1.53 μ A, respectively. It should be noted that the values of the "on" resistances of all switches are assumed to be 2 Ω (realizable using off-the-shelf multiplexers).

Fig. 6 shows the effect of the values of other resistors of the same row, denoted by R_r (i.e. all other resistors of the same row not being measured when measuring R_x) on the read-out accuracy. It can be observed that by increasing the value of R_r , the cross-talk issue is degraded and the error became smaller. Fig. 6 compares the simulated values of the error as a function of R_x for both the conventional single-sampling (dashed curve) and the proposed double-sampling (solid curve) techniques assuming an offset voltage of 2mV for the OpAmp. It can be observed that the proposed technique can dramatically

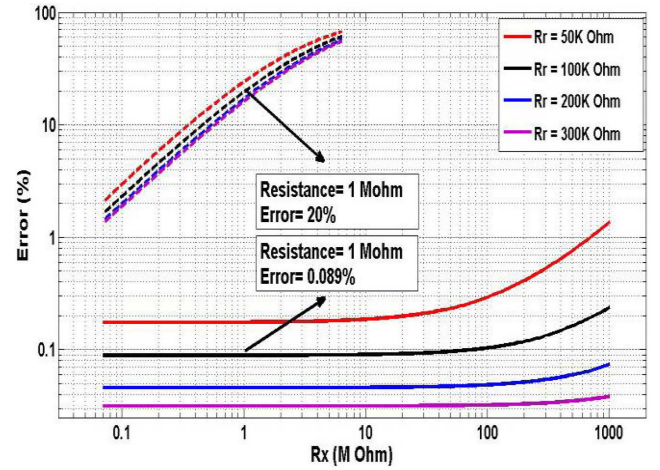


Fig. 6. Simulated values of the read-out error for both single-sampling and double-sampling techniques.

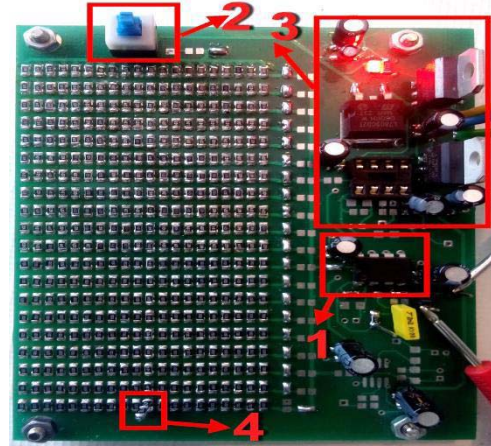


Fig. 7. Fabricated PCB for evaluating the proposed technique.

increase the read-out accuracy. For instance, when measuring a 1M Ω resistor and when R_r 's are all 100k Ω , by employing the double-sampling approach, the measurement error can be reduced from 20% to less than 0.089%. Furthermore, by increasing the value of R_x (for a constant R_r), the error is increased to unacceptably large values for the conventional technique. However, by using the proposed technique, a large resistance of 1G Ω can be measured with less than 0.24% of error.

V. MEASUREMENT RESULTS

Finally, a 20 \times 20 resistive sensor array, the board-level implementation of which is shown in Fig. 7, has been realized. The board includes the sensor array located in the left part, the AD8042 Opamp (marked as 1), the state-alternating switch that switches between 0V and 5V (marked as 2) and the supply/regulator part (marked as 3). The values of the input offset voltage and the input bias current of the Opamp are 9.8 mV and 4.8 μ A, respectively. Also the DC gain is 90dB, the unity gain bandwidth is 150MHz and the slew rate is 200V/ μ s. The values of all the resistors are 100k Ω while the value of the resistance being measured,

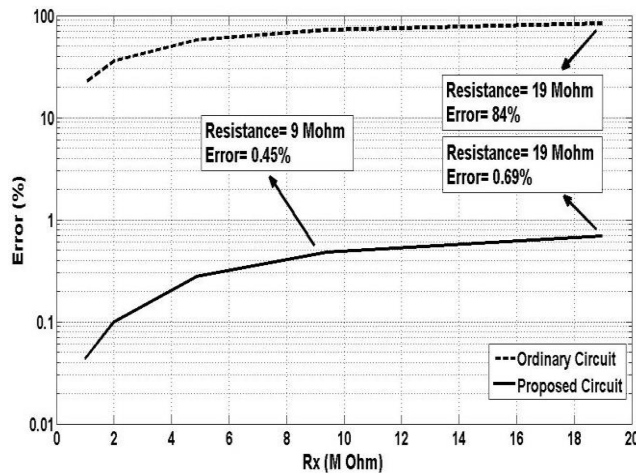


Fig. 8. Measurement results for single-sampling and proposed double sampling technique.

i.e. R_x (and marked as 4), is changed from $1\text{M}\Omega$ to $20\text{M}\Omega$. In addition, the output voltage of Opamp is measured and digitized by AD7656 ADC. Fig. 8 shows the measurement error as a function of R_x for both the conventional single-sampling and the proposed double-sampling approaches for R_x values between $1\text{M}\Omega$ and $19\text{M}\Omega$. It can be observed that using the proposed approach, R_x values as high as $19\text{M}\Omega$ can be measured with errors of 0.69% and 84% for the proposed and conventional approaches, respectively.

VI. CONCLUSION

In this paper, the effects of two main non-ideality sources of operational amplifier, i.e. the input offset voltage and the input bias current have been analyzed. It was observed that those two non-idealities considerably reduce the accuracy of resistive readout for large arrays and large resistance values. A double-sampling technique was therefore proposed to reduce the effect of those non-idealities. Both simulation and measurement results confirm the effectiveness of the proposed technique in increasing the measurement accuracy for large arrays and/or large ratios of R_{\max}/R_{\min} . Although, the proposed approach reduces the operational speed by a factor of 2, one can increase the speed of the readout circuit by employing additional Opamps in parallel.

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