DEVELOPMENT AND INVESTIGATION OF A SWITCHED-CAPACITOR INSTRUMENTATION AMPLIFIER WITH DYNAMIC-ELEMENT-MATCHING FEEDBACK

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This paper shows the implementation and measurement result of a recently designed switched-capacitor dynamic- element- matching amplifier. The main advantage of this amplifier concerns it immunity against input common-mode voltages. As compared with a previous design, the performance of the circuit has been improved by optimizing the design of the switches and by extending the auto-calibration procedure to two different offset measurements. For an experimental evaluation, a test chip has been designed and realized in 0.7µm CMOS technology. ***.*** *The experimental result shows 1.4µV input referred noise for a measurement time of 40 ms, which is close to the calculated value. A very high immunity against input common-mode voltages has been found. However, the non-linearity over the full input-voltage range is only 0.27%.*

Keywords: Dynamic-Element-Matching, Resolution, Instrumentation Amplifier, Common-Mode Rejection

1. INTRODUCTION

In conventional measurement systems, major systematic-error sources are related to offset and parameters drift by temperature and aging. However, the use of advanced techniques such as chopping and three-signal auto-calibration will eliminate the effects of any systematic additive (offset) and multiplicative (gain) uncertainties [1]. To achieve this, the core processing circuit should be designed with a high degree of linearity for the relevant input range. Therefore, before applying the input signal to the main circuit, it is necessary to scale its size to an appropriate range. Depending on the signal levels, a voltage divider or an amplifier will be needed. Because these pre-scaling circuits are outside of the auto-calibration path, any inaccuracy in the scaling factor will affect the accuracy of the whole system. Because the offset of a pre-scaling circuit can still be eliminated, the inaccuracy of the scaling factor is the main concern.

Nowadays, the most accurate amplifiers are realized by applying negative feedback (by means of passive element) around an active gain stage. Then, the accuracy of the transfer function is limited by the matching properties of the feedback elements.

The best method for designing amplifier with accurate gain is using Dynamic-Element-Matching Feedback [2]. In the following sections a short summary of these measurement techniques will be presented. Afterwards the details and experimental result of the optimized design will be presented.

2. INSTRUMENTATION AMPLIFIER WITH DYNAMIC-ELEMENT-MATCHING FEEDBACK

In [2], [3] an instrumentation amplifier with resistive dynamic-element-matching (DEM) feedback has been presented. The main problem of this circuit concerns its limited common-mode range. However, in [4] a Switched-Capacitor - DEM Amplifier (Figure 1) has been presented, that can handle rail-to-rail common mode voltage. On each cycle, N-1 equal capacitors (equal in layout, but in reality showing some mismatches) are connected to the input and the remaining capacitor is connected to the output. After a complete cycle of interchanging the capacitors, the average gain $\overline{G}$ over N clock cycles equals N-1.

In first order this gain is independent of mismatching [5] and equals:

$$\overline{G} = \frac{1}{N}\sum_{j=1}^{N}\frac{\sum_{i=1}^{N}C_i - C_j}{C_j} \approx N-1+\frac{1}{N}\sum_{j=1}^{N}\delta_j^2 . \tag{1}$$

In this equation δ_j is the relative mismatching between the capacitors with respect to the average value. There is some leakage current at the inverting input of the amplifier, so we need a reset to prevent excessive drift at the output. The core processing system that we use converts the voltages to periods and measures the periods with a micro-controller. Therefore, it can do averaging without any extra needs. The main problem of this circuit is that switch-charge injection causes noise and inaccuracy. The performance of the circuit has been improved by optimizing the switch geometries for a minimum charge injection and an acceptable ON resistance R_{on}.

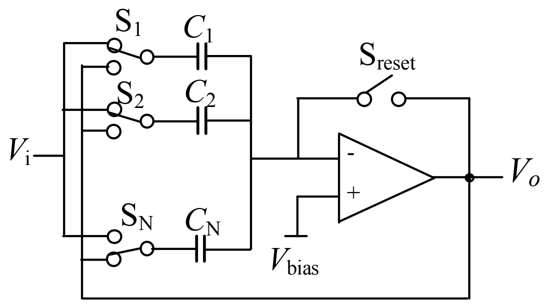

Fig. 1. Switched-Capacitor Dynamic-Element-Matching Amplifier (SC-DEM Amplifier).

3. EXTENDED AUTO CALIBRATION

In the original three-signal measurement [1], only a single offset measurement is performed. However, due to voltage dependency of the capacitor and the use of two different choppers, which were needed for some practical reasons, the offset voltages slightly depend on whether or not an amplifier is used. Also this offset depends on the common-mode voltage. Therefore, it is better to apply an offset measurement for each of the applied configurations. This can be explained with the help of Figure 2, which shows the applied extended auto calibration.

The four different signals are measured in the following way:

In phase 1, the switches S_1 and S_3 are ON and the other switches are OFF. In this case we convert V_x to T_x, where:

$$T_1 = T_x = KAV_x + T_{off1}. \tag{2}$$

In this equation K and T_{off1} represent the gain and the offset of voltage-to-period converter, and A is the gain of our amplifier.

In phase 2, the switches S_1 and S_4 are ON and the other switches are OFF. Then it holds that

$$T_2 = T_{off1}. \tag{3}$$

In phase 3, the switches S_2 and S_5 are ON and the other switches are OFF, which yields that

$$T_3 = T_{ref} = KV_{ref} + T_{off2}. \tag{4}$$

In Phase 4, the switches S_2 and S_6 are ON and the other switches are OFF. So in this phase we can measure T_{off2}:

$$T_4 = T_{off2}. \tag{5}$$

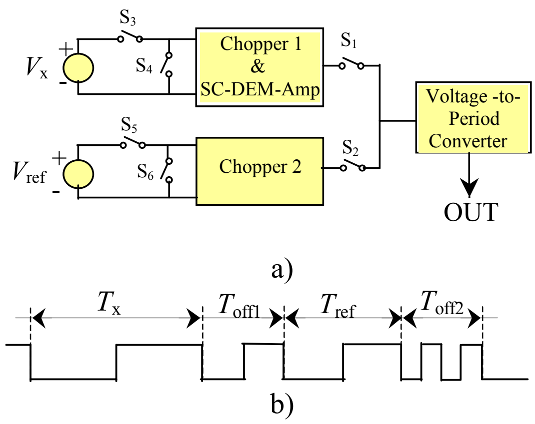

Fig. 2. (a) The configuration for auto-calibration and (b) the output signal, which consists of four concatenated periods.

Therefore we can calculate M with equation:

$$M = \frac{T_x - T_{off1}}{T_{ref} - T_{off2}} = \frac{AV_x}{V_{ref}}. \tag{6}$$

When knowing V_{ref} and after measuring M we can calculate V_x according to the equation:

$$V_x = \frac{MV_{ref}}{A}. \tag{7}$$

In this equation we suppose that the gain is 7.

During testing, when applying known voltages V_{ref} and V_x and measuring M, we can calculate the amplification factor with the equation:

$$A = \frac{MV_{ref}}{V_x}. \tag{8}$$

As it can be seen the measured M and so V_x and A are independent of additive and multiplicative errors of the voltage- to-period converter.

4. Measurement Results

The improved design has been realized in 0.7µm CMOS technology of AMIS (Fig. 3).

We measured the different periods of the chip output using a micro-controller (for identification purposes the frequency of T_{off2} is twice that of the rest). The data is read via a serial port (RS232) and transferred to a PC, where the data is analyzed with a Labview program.

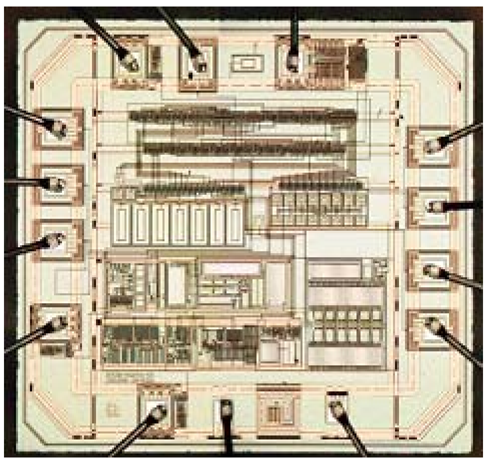

Fig. 3. Photograph of the 2000µm ×1800µm chip.

From that, we calculate the voltage-to-time transfer functions H_x and H_{ref} with the following equations:

$$H_x = \frac{T_x - T_{off1}}{V_x}, \tag{9}$$

$$H_{ref} = \frac{T_{ref} - T_{off2}}{V_{ref}}. \tag{10}$$

In our setup it holds that H_x = 434 µs/mV and H_{ref} = 62 µs/mV. For the standard deviation of these periods we found:

$$\sigma_{(T_x - T_{off1})} = 0.6\mu s, \tag{11}$$

$$\sigma_{(T_{ref} - T_{off2})} = 0.2\mu s. \tag{12}$$

We can transform these jitters to input referred noise according to the equations:

$$V_{n,x} = \frac{\sigma_{(T_x - T_{off1})}}{H_x}, \tag{13}$$

$$V_{n,ref} = \frac{\sigma_{(Tref - T_{off2})}}{H_{ref}}. \tag{14}$$

The results are 1.4 µV and 3.2 µV for $V_{n,x}$ and $V_{n,ref}$, respectively. Comparing the values of $V_{n,x}$ and $V_{n,ref}$ it can be concluded, that the amplifier improves the VPC resolution with more than one bit. At the cost of more power consumption it would be possible to increase the resolution even more.

Figure 4 shows the measured voltage $V_{x,measured}$ for V_x = 4.088mV, V_{ref} = 46mV, and a measurement time of 40 ms. The average measured value is 4.093mV, corresponding to A= 7.008. The standard deviation amounts to 1.4 µV.

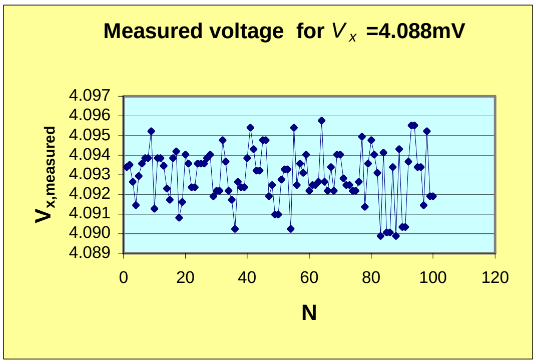

Fig. 4. Measurement result for V_x = 4.088 mV

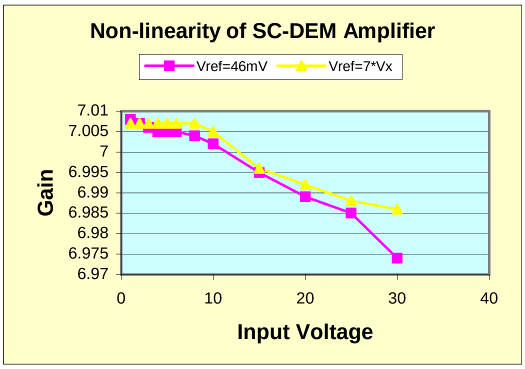

Fig. 5. Gain versus input voltage

In order to measure the gain of the amplifier with respect to the input voltage, we performed two measurements. In the first measurement, we applied a constant voltage as V_{ref}. In the second measurement, we changed the value of V_{ref} in such a way that always V_{ref} = 7V_x, so that at the input of the VPC, for the V_x and the V_{ref} measurements, the voltages are equal. So, in the second measurement, the non-linearity of the voltage-to-period converter (VPC) is eliminated by auto-calibration and the measured non-linearity is due to the amplifier only. In the first measurement the non-linearity is related to the whole system.

The results are shown in the Fig. 5.

For input voltages V_x<25mV, both measurements give similar results, which shows that the amplifier non-linearity is dominant in the overall system non-linearity. However, for V_x>25mV, between the two curves a significant difference is found. This is due to saturation of the integrator in the VPC, which causes an increasing system non-linearity, which is measured in the first measurement.

Therefore, we use 25mV as the maximum value of the dynamic range. If we define the worst-case non-linearity as:

$$\lambda = \frac{G_{\max} - G_{\min}}{G_{\text{avr}}}, \qquad (14)$$

Then, for a full-range of 25mV, the non-linearity λ amounts to 0.27%. Usually, for low-sensitivity thermocouples, such as the types B, R, and S, such non-linearity is too large. At this moment, the reasons for this non-linearity are not yet fully understood and are under investigation.

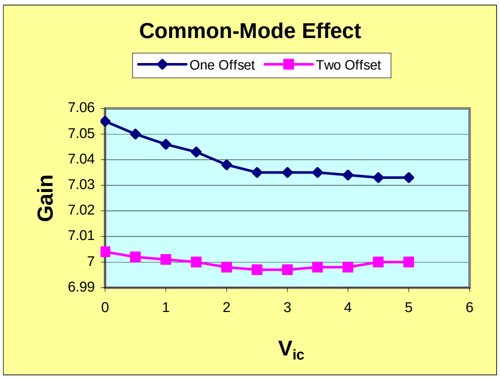

Fig. 6. Measurement gain variations for V_x = 4.088 mV and varying common-mode voltage V_{ic}

To test the effect of the input common-mode voltage V_{ic} on the amplification factor, we changed the input common-mode voltage from 0V to 5V, with a step of 0.5V, and measured the gain of the amplifier for V_x=4.088mV, with one and two offset measurements, respectively (Fig. 6).

From Fig. 6 we can see two distinctive advantages of the two-offset measurement in comparison with one-offset measurement. Firstly, the effect of the input-common-mode voltage on the amplification factor is much less and secondly, the gain is much closer to the designed value of 7. The first advantage is independent of the input voltage V_x, however, the second advantage is only valid for V_x<10mV (for signals to be amplified, this is the most important part the input voltage range).

5. CONCLUSION

A SC-DEM amplifier as well as a voltage-to-period converter has been optimized, which resulted in a significant reduction of the noise level of the interface circuit. To improve the common-mode rejection of the system, an extension of the auto-calibration technique with an additional offset measurement has been proposed. The amplifier has been designed and realized in a 0.7 µm CMOS process. Measurements show 1.4 µV input referred noise for a measurement time of 40 ms, which for a thermocouple interface is rather good. The relative gain variations due to the input common-mode voltage are less than 0.01%. However, the 0.27% non-linearity of the gain over full dynamic range of the desired input-signal is too high to be tolerated. The reasons for this non-linearity are not yet fully understood and are under investigation.

6. REFERENCES

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