A Self-Adaptive Front-End for Grounded Conductivity Sensors in Liquid-Monitoring Applications

Qi Jia, Xiujun Li, and Gerard C.M. Meijer

Abstract— **In this paper a self-adaptive front-end for measuring liquid conductivity has been proposed. To address the chemically active environment that the sensor electrodes will face, the DC component of the excitation signal is set to be zero. Moreover the final system will be functional with a single power supply even when one terminal of the sensor electrode is grounded. The use of a reference conductance will make sure the full dynamic range of the system can be always used, therefore the best signal-to-noise ratio can be achieved. The design has been implemented and submitted for fabrication in a 0.35µm CMOS technology.**

Index Terms— **conductivity measurement, adaptive signal processing**

I. INTRODUCTION

CONDUCTIVITY sensors are widely used in agricultural and food industries for monitoring and controlling specific biological/chemical features and qualities of the liquid.

The conductivity of such liquid is usually measured by means of detecting the conductance/resistance of a specific structure, which is made of the very same kind of liquid and the sensor electrodes.

However, because of the direct exposure of the electrode to the liquid, the chemically active environment will cause special problems for the sensor electrodes. One of those problems is that a DC component in the excitation signal will cause undesired chemical reactions, thus accelerating the corrosions, on the electrodes. So the proposed design should be able to protect the electrodes from the corrosions or at least not to accelerate it.

Another problem is that in many applications one electrode of a conductivity sensor is grounded to earth physically, which is due to specific restrictions of the application. The result is that, for symmetrical excitation, the other electrode will see both a positive and a negative voltage, which could require an extra negative power supply for the sensor system. For today's single- supply-microcontroller centered sensor systems, this would be rather undesirable. To address this problem, in this paper a combination of a current-driving scheme with the capacitor-based galvanic isolation is proposed.

One common problem of the current excitation of sensor probes is that it usually limits the measurement range. For a predefined current value, the response voltage on the sensor is proportional to the equivalent resistance R_x of the sensor. However the modern electronics system has a fairly small power supply voltage, which greatly limits the system's dynamic range in voltage domain. Thus only some specific applications, which suit the measurement range of the data-acquisition system, can be measured. An application independent solution has been given in this paper and detailed discussion is given in section II.

The presented work is a subsystem of a novel multi-functional sensor interface called Universal Sensor Interface, where different types of sensors can be interfaced with one chip by configuring the chip in one of the different operation modes.

II. DESIGN CONSIDERATIONS

The simplified system diagram is shown in Figure 1. The sensor R_x is galvanic-isolated from the measurement system with capacitors C_1 and C_2. Two operational floating amplifiers (OFA) generate a current flowing through the C_1, R_x and C_2. The level of this current is defined by V_{ex} and R_{ref}. The resulting voltage is then sensed by an instrumentation amplifier. Capacitors C_5 and C_6 again isolate the charge carriers between the sensor and the measurement system. To obtain the value of R_x, we measure the voltage Vref across the reference resistor R_{ref} and the V_x from the sensor. With three-signal autocalibration method{Meijer, 1992 171 /id} we can calculate the value of R_x by

$$R_x = R_{ref} \frac{V_x - V_{off}}{V_{ref} - V_{off}} \tag{1}$$

where V_{off} is the system offset referred to the input of the instrumentation amplifier A3 and is obtained by taking the output result while short circuit the input of A3.

Detailed explanations for each design consideration will be given in this section.

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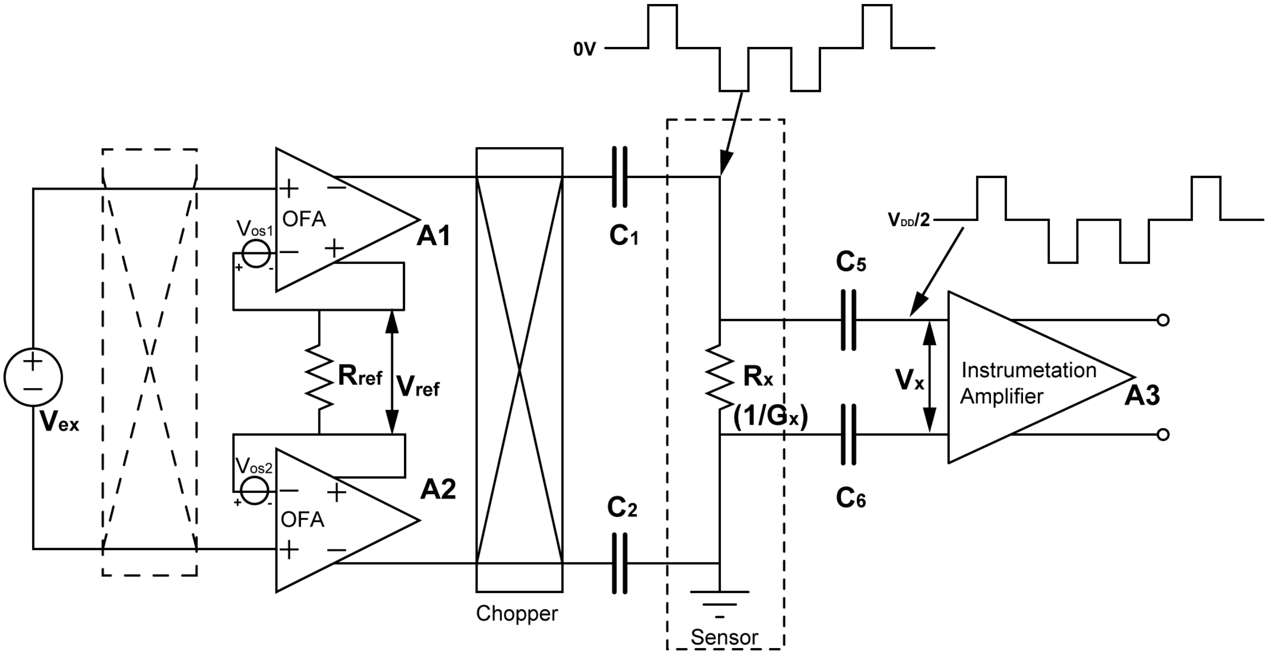


Figure 1. Simplified system diagram

A. *Current Excitation with Capacitive Isolation*

For a grounded conductivity sensor, a symmetrical excitation signal means a positive and negative voltage on that non-grounded terminal of the sensor. In this case both voltage and current excitation can be applied to have this negative voltage. However directly applying a negative voltage signal on the sensor would require a double-supply amplifier and a negative power supply. For nowadays' microcontroller centered data acquisition system a negative power supply is certainly not preferred. On the other hand with a capacitive isolation the bipolar current excitation is compatible with both the single power supply on the electronics side and the negative voltage on the sensor side. Therefore the combination of a current excitation with a capacitive isolation has been used.

To have a bi-directional current on the sensor, a chopper has been applied. Generally there are two places to put the chopper. One is between the voltage input of OFAs and V_{ex}. Here the voltage V_{ex} is chopped, so a $\pm V_{ex}$ will become the input of the OFAs. Another place to insert the chopper is at the output terminals of the OFAs, where the current generated by the OFAs is chopped.

In practice even with great carefulness during the layout design, the OFAs will still have some offset due to the random mismatch. In the phase where V_{ex} is the input voltage, the output current is

$$I_{out} = (V_{ex} + V_{os1} - V_{os2}) / R_{ref} \tag{2}$$

While in the other phase, where $-V_{ex}$ is the input voltage, the output current would be

$$I_{out} = (-V_{ex} + V_{os1} - V_{os2}) / R_{ref} \tag{3}$$

Compare (2) and (3), we can see that they not equal in amplitude. Therefore a long-term DC component will reside on the sensor, thus accelerate the chemical reactions on the electrodes. However the chopper at the output of OFAs would not observe this problem, since the same current is used and only the direction is reversed in two phases. So the current position of the chopper is decided.

B. *Adaptive Current Generation*

As introduced in section I, the current driving compared with voltage driving suffers a problem of limited dynamic range of the system, thus the system is only able to interface a limited input range.

A conventional solution to this problem is to divide the input range into several sub ranges. For each sub range an excitation current is used. In this way a total large input range can be implemented. However the sub ranges are usually predefined during the design phase of the measurement system. The real world sensors may not exactly vary in any sub range of the system. Suppose that we have a measurement system, which has sub ranges of 100Ω and 1000Ω, and a sensor, which varies between 50Ω and 200Ω. We would have to use the 1000Ω range of the measurement system, but then we lose 80% of the dynamic range of the system.

To address this problem we introduce a reference resistance R_{ref}, which can be chosen in such a way that R_{ref} equals to the maximum value of the R_x. The current to excite the R_x is defined as

$$I_{ex} = V_{ex} / R_{ref} \tag{4}$$

where V_{ex} is defined by the system. In this way the maximum response voltage V_x on R_x would be V_{ex}, we can design the gain of the instrumentation amplifier to make full use of the system's dynamic range. And because the user can always select R_{ref} to be the maximum possible value of interest, the performance will be application independent.

C. *Differential Driving*

For the grounded conductivity sensor, it would be enough at the circuit level to have a single ended driving and detecting. However in practice the ground connection may not be very explicit. The so called ground can easily be the earth of the whole sensor system. And if we inject the current into the sensor's floating terminal, the return path of this current is undefined and can be rather large and complex. Therefore this setup would be very bad for EMC and vulnerable to all kinds of EMI. To have a good EMC property, we have to define the current return path explicitly. Therefore a differential driving method has been applied. In Figure 1 with two OFAs we provide a complete current loop for the sensor.

III. DETAILED CIRCUIT FOR IMPORTANT BUILDING BLOCKS

Based on the considerations in section II and Figure 1, we have a system that can perform all the requirements to be an adaptive front-end for a wide range of conductivity sensors. However to successfully implement all the building blocks will need in depth analysis of the requirements and the trade-offs we have to take.

A. *Implementation of OFAs*

The operational floating amplifiers are different from other types of operational amplifiers because of the unique feature that the two output terminals of an OFA have exactly the same amount of current flowing in/out of that terminal. This is particularly important in this design because in order to know the value of R_x by V_x, V_{ref} and R_{ref} through equation (1), it has to be true that the current flowing through R_x is exactly the same to the current flowing through R_{ref}. Otherwise the final result will be erroneous. Equation (1) can be written as

$$\begin{aligned} R_x &= R_{ref}\frac{V_x - V_{off}}{V_{ref} - V_{off}} \\ &= R_{ref}\frac{(I_{ex} + \Delta I_{ex})R_x}{I_{ex}R_{ref}} \end{aligned} \tag{5}$$

where the ΔI_{ex} is the mismatched current between the R_{ref} and the R_x. It is clear that any mismatch is directly translated into the final error without any attenuation. Therefore the design of the two OFAs has to insure the perfect balance between the output currents.

Because the output current is defined by the V_{ex} and the R_{ref}, which is chosen by the user according to the application, the level of this current can be very different. Suppose a V_{ex} of 100 mV, for R_{ref} of 100 Ω and 100 kΩ the resulted current will vary from 1 μA to 1 mA. In this case, to be able to provide the necessary amount of current for the 100 Ω R_{ref}, a class A output stage would have a quiescent current of at least 1.2 mA to avoid serious distortion. However when R_{ref} is in the range of 100 kΩ, this will be a big waste for the power consumption. For high power efficiency, the output stage of the OFA should be class AB or other type of circuits with a similar functionality.

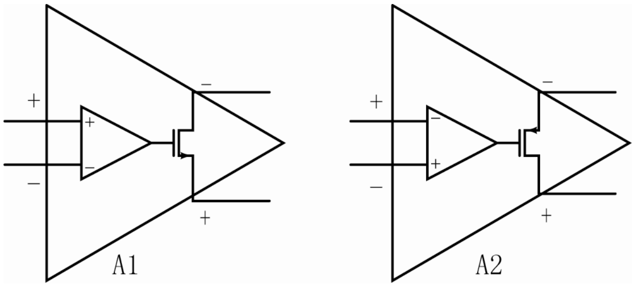


Figure 2. The circuit implementation of OFAs

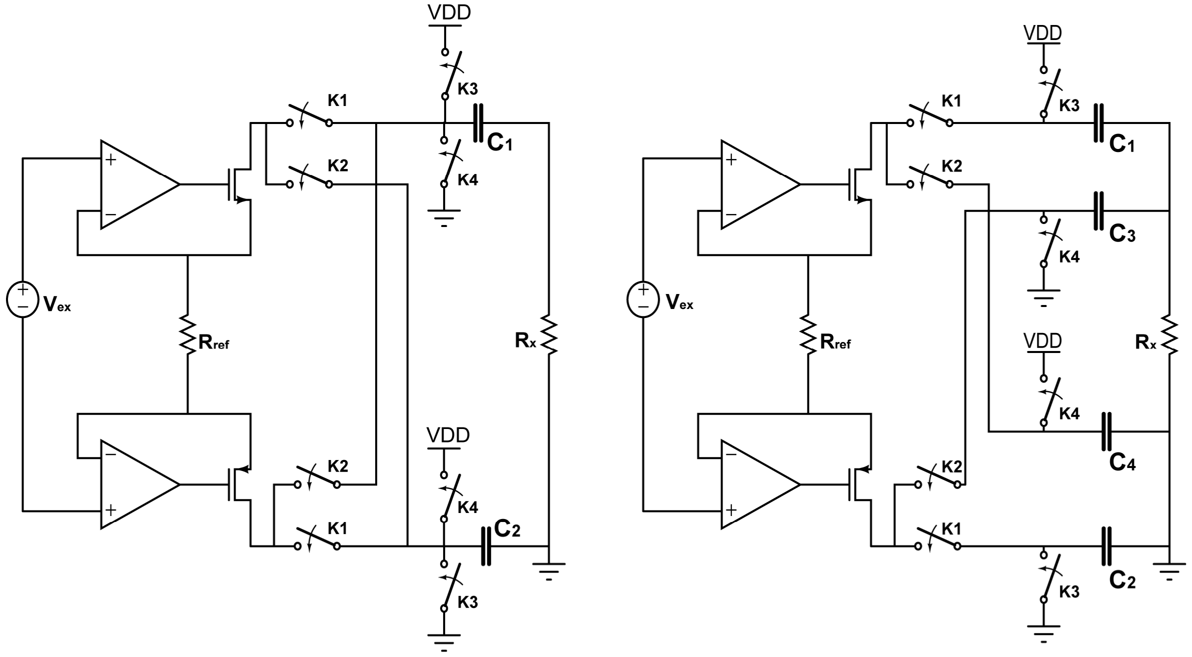
There are several ways to implement an operational floating amplifier. In {Sedra, 1989 320 /id;Haslett, 1979 321 /id;John H.Huijsing, 2001 322 /id} different types of OFA designs have been demonstrated. In [2][3] the matching of the output currents directly depends on the matching of the transistors. And in silicon implementation, this matching factor would be 0.5% at best for a reasonable area. And therefore the final system accuracy can not be better than this. So this is not acceptable for a high accuracy measurement system. In [3] the output current of one terminal is measured by a monitor resistor in series with the output and accurately copied into another current. In this case the matching of the output current is depending on the matching of two resistors, which is much better than the matching between transistors. But since the current level is changing significantly according to the R_{ref} value, it is very difficult to select the value of the monitoring resistor and impossible to cover a large range of R_{ref} with a single value monitoring resistor.

Figure 2 shows the implementation of the OFAs A1 and A2. With one opamp and one pmos/nmos transistor, we now have an operational amplifier that fulfills all the requirements. The two output terminals are source and drain of the same transistor; therefore they are always equal to each other.

The problem of such an amplifier is that it has no driving ability, which means we have to initiate a current that flows in the OFAs, and then the negative feedback configuration will regulate the current level. The isolation capacitors C_1, C_2 have been assigned another role here as the energy storage. During the reset phase, left pins of C_1 and C_2 are reset to either *VDD* or *GND* respectively depending on the current direction. See Figure 3. During the driving phase, the charge stored in C_1 and C_2 will provide a current flowing through the OFAs with the relation of (4). The reset phase is necessary here to perform two functions: one is to define the voltage on the high impedance nodes at the left pins of the isolation capacitors C_1 and C_2; the other is to recharge the capacitors C_1 and C_2 so that next phase there will be current flowing in the circuit. And the reset circuit is discussed in next section.

B. *Reset Circuits*

Two reset circuits are shown side by side in Figure 3. The circuit is working in the sequence of K3, K1, K4 and K2. The



Reset Circuit A

Reset Circuit B

Figure 3. Two types of reset circuits

circuit A uses two capacitors for the isolation whereas circuit B uses four capacitors. However the circuit A consumes much more power than circuit B, since it is frequently resetting capacitors between VDD and GND. The two extra capacitors do not need to be accurate and hence do not increase the cost and board area very much. Therefore the circuit B has been chosen for the final design.

IV. IMPLEMENTATION

The proposed design has been designed for manufacture with AMI Semiconductor 0.35µm CMOS-A technology. And the active area measures 530 µm × 640 µm.

V. CONCLUSION

An application-adaptive front-end circuit for grounded conductivity sensor has been designed. The proposed system can operate with a single power supply while providing an accurate symmetrical excitation on the sensor. The special way of implementation of the system's building blocks gives very high power efficiency. The proposed system can be applied in a very wide range of conductance/resistance measurement applications without any modification.

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