

MEASUREMENT OF RESISTANCE IN CHARGE BALANCING CIRCUIT

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Abstract: This paper presents a method of the measurement of resistance that can be used in a simple microcontroller systems. A direct analog-to-digital conversion (without the measurement of any additional values like a voltage or current) was based on the charge balancing scheme. The whole measurement is divided into two stages. The final result is achieved by computing the difference between them. It makes it possible to eliminate theoretically influences of many factors evoking errors such as long time instability and accuracy of a voltage reference source, influences of supply voltage or offset voltages of a comparator or integrator etc.

Keywords: measurement of resistance, charge balancing circuit, two stages measurement

1 INTRODUCTION

The method of measurement based on the charge balancing circuit is commonly used to voltage signals, especially audio signals [1]. It makes it possible to simplify an analog part of a converter expense of expending a digital part. The main advantages of the method are good linearity of conversion, low sensitivity to tolerances of components and the ability to achieve a high resolution and accuracy of the analog to digital conversion [2][3].

The idea of charge balancing circuit can be easily applied to measurements of resistance and the main goal of this article is to find properties of such direct „resistance-to-digital” conversion. It is worth noticing that the measurement of resistance based on a charge balancing scheme is not realised indirectly by measurements of any other physical values like voltages and currents.

In a test circuit a single, cheap 8-bit microcontroller to control the measurement and furthermore to realise a simple processing of the achieved results has been used. In the simplest form, a complete measurement system could consist of a controller, an operational amplifier working as Miller’s integrator and a few other passive components only. Currently many producers of microcontrollers present applications of their chips to measurements of resistance of sensors using the dual slope scheme. The method of charge balancing is very similar to the dual slope method as the circuit is regarded, but it has many advantages and it seems it could be used alternatively. Moreover due to a simple analog part it could be easily applied to constructing monolithic circuits.

2 DESCRIPTION AND PROPERTIES OF METHOD

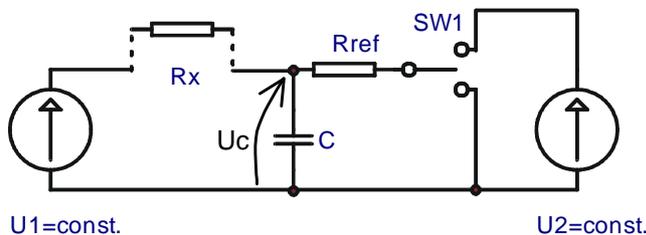


Figure 1. The simplest charge balancing circuit for the measurements of resistance.

Figure 1 shows the simplest circuit using charge-balancing scheme. The R_x is variable and unknown, values of voltages in the circuit: U_1 and U_2 are constant and known. The value of R_x can be evaluated from equation:

$$R_x = \frac{(U_1 - U_c) \cdot R_{ref}}{U_c - U_2 \cdot h} \quad (1)$$

$$\text{where: } h = \frac{T_H}{T_{total}} \quad (2)$$

- T_H is the time when the switch SW1 was connected to U_2 source,
- T_{total} is the time of one measurement.

The SW1 is switched by the controller to keep the constant value of the charge stored in a capacitor C. It is already equivalent to keeping the constant value of the voltage U_c . In fact the

equation (1) is only true when changes of the voltage on pins of the capacitor C are very small: $\Delta U_c \rightarrow 0$. But the most significant is that the final result depends on many factors as reference voltage sources, a reference resistor or assumed voltage on capacitor C. The equation (1) has been achieved by making a great simplification and in reality there are a lot of other factors that can influence the final result and eventually evoke errors of the measurement. In the ideal case the value of Rx could be expressed as a function of two factors: reference resistor and coefficient η obtained as the result of the charge balancing process:

$$R_x = f(R_{ref}, h) \quad (3)$$

It would be theoretically possible if the "auto zero" method, known from processing voltages [4], were applied. In that case two measurements have to be done and the final result is the difference between them. The first measurement is realized when the value of U_1 is equal to zero and the second one when U_1 is the same as U_2 , so the obtained values of the coefficient η are:

$$\text{for } U_1=0: h_1 = \frac{U_c}{U_2} \cdot \left(1 + \frac{R_{ref}}{R_x}\right) \quad (4)$$

$$\text{for } U_1=U_2: h_2 = -\frac{R_{ref}}{R_x} + \frac{U_c}{U_2} \cdot \left(1 + \frac{R_{ref}}{R_x}\right) \quad (5)$$

$$\text{and the difference is: } h_d = h_1 - h_2 = \frac{R_{ref}}{R_x} \quad (6)$$

$$\text{hence: } R_x = \frac{R_{ref}}{h_d} \quad (7)$$

The equation (6) shows that the processing function $h_d = f(R_x)$ is non linear. It is reverse proportional to the measured resistance. However, it would be possible to exchange resistors Rx with Rref at Fig. (1) and obtain a linear dependence, but when the Rx were switched with high frequency, a non-resistant part of its impedance, for instance: capacity of a connection wire, could affect the achieved result.

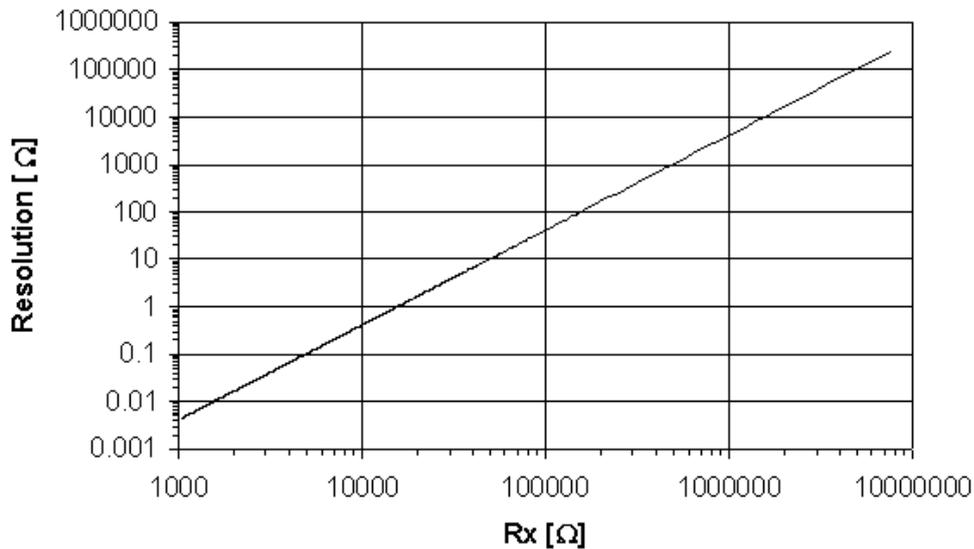


Figure 2. The resolution versus measured resistance

In the circuit the analog value is converted to its digital representation and hence additional process called quantization of results must be considered. The available number of results is finite and (6) should be rewritten as:

$$n_d = \text{ant} \left\{ \frac{R_{ref}}{R_x} \cdot N \right\}, \text{ where} \quad (8)$$

n_d - is a binary result of the measurement,
N - a count of available quantization levels.

The main effect of the non-linear characteristic of the conversion is different resolution of the measurement depending on the value of resistance. Smaller values of the R_x are measured with a far better resolution than bigger ones. Figure 2 shows hypothetical characteristic of these changes versus measured resistance where the following values: $N=230400$ and $R_{ref}=1050 \Omega$ were assumed. It is a good feature, because the most of real sensors used in technology, which convert physical values to the resistance for e.g.: thermistors, photoresistors, have non-linear functions of changes of the resistance versus a measured value. Such characteristic of conversion makes it possible to achieve the better final resolution not only in its particular part, but in a whole measurement range. It leads to relatively constant resolution of the measurement with regard to the measured physical value.

3 INFLUENCE OF RESISTANCES OF SWITCHES

In this chapter, a bit more complicated circuit than that shown at Fig. 1 and the influence of more important factors evoking errors on accuracy of the method, will be considered. The new circuit has been presented in Fig. 3a. A single capacitor C storing unbalanced portion of the charge has been replaced with the Miller's integrator circuit. In this way values of the injected charge through the R_x and R_{ref} became independent of temporary voltage changes on the capacitor C . This voltage changes in real circuit cannot tend toward zero due to the limited sensitivity of the comparator. Additionally another switch $SW2$ which controls measurement stages has been added. The switch $SW1$ is used to balance the charge. It should be noticed that the $SW2$ works with a much slower frequency than the $SW1$. If an ordinary binary counter were used as a lowpass filter and decimator (Fig. 3b), their ratio would be equal to the length of the counter. The reference source U_2 has been replaced by V_{cc} . It has become possible since this voltage has been eliminated from the final equation (7) as the result of using two stage measurement. The precise and stable voltage source has become unnecessary.

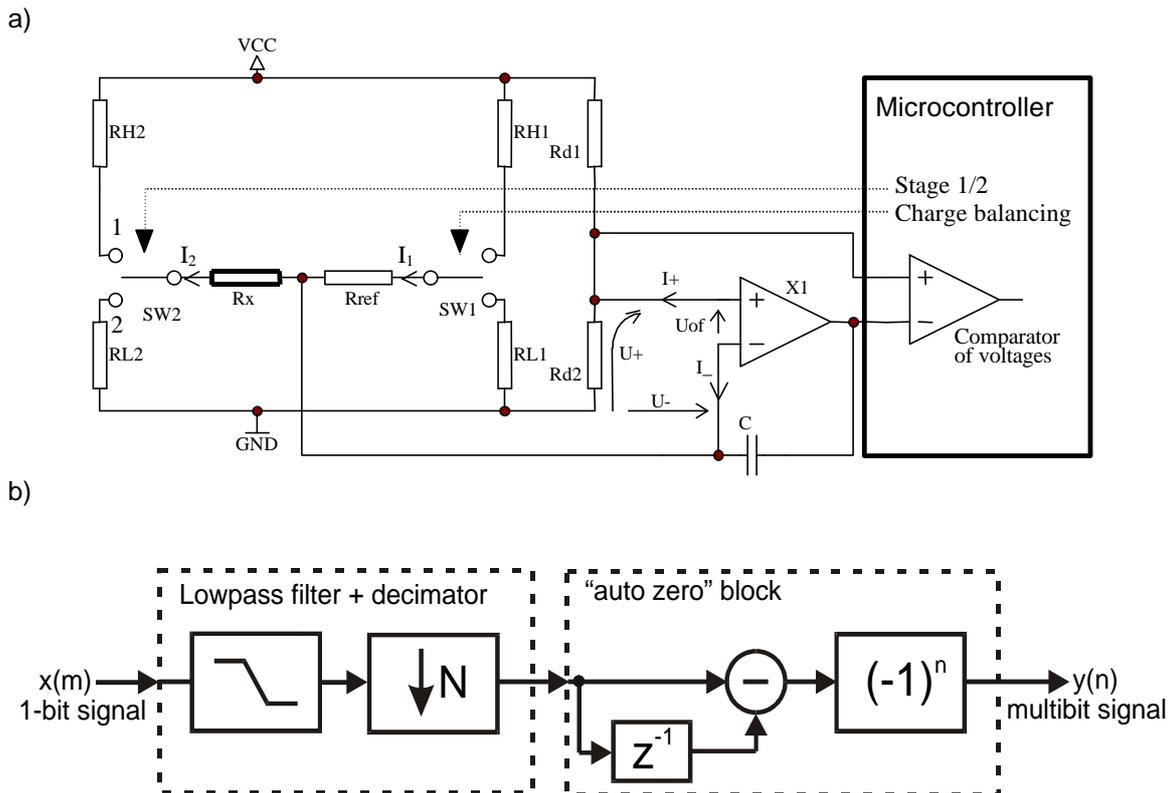


Figure 3. The circuit for two stages measurement: a) analog part, b) digital part

Real switches based on a MOSFET transistors have finite on and off values of resistance. The most important are on values. They are represented in Fig. 3a by R_{H1} and R_{L1} resistors for the $SW1$ switch and R_{H2} and R_{L2} for the $SW2$ respectively. An operational amplifier X1 working as an integrator has an offset voltage U_{of} and can provide some currents: I_+ , I_- to the circuit from its inputs independently of currents provided by R_x and R_{ref} . It will also be regarded in further considerations.

Every factor mentioned above can cause measurement error, but due to uses the "auto zero" method, their influence can be reduced significantly. Next equation shows the result of computing the

difference between two stages of the measurement where these factors have been taken into consideration:

$$\mathbf{h}_d = \mathbf{h}_1 - \mathbf{h}_2 = \frac{\frac{V_{cc} - U_-}{R_x + R_{H2}} + \frac{U_-}{R_x + R_{L2}}}{\frac{U_-}{R_{ref} + R_{L1}} + \frac{(V_{cc} - U_-)}{R_{ref} + R_{H1}}} \quad (9)$$

The voltage U_- can be expressed as a part of V_{cc} , because its value is almost the same as the U_+ voltage obtained by dividing V_{cc} in R_{d1} , R_{d2} divider:

$$U_- = \mathbf{b} \cdot V_{cc} \quad (10)$$

In this case a value of coefficient β will be the function the following factors:

$$\mathbf{b} = f(R_{d1}, R_{d2}, I_+, U_{of}, V_{cc}) \quad (11)$$

Substituting equation (10) for (9) it is possible to "hide" all voltages in the variable β :

$$\mathbf{h}_d = \frac{\frac{1 - \mathbf{b}}{R_x + R_{H2}} + \frac{\mathbf{b}}{R_x + R_{L2}}}{\frac{\mathbf{b}}{R_{ref} + R_{L1}} + \frac{1 - \mathbf{b}}{R_{ref} + R_{H1}}} = \frac{(R_{ref} + R_{H1}) \cdot (R_{ref} + R_{L1}) \cdot [\mathbf{b} \cdot (R_{H2} - R_{L2}) + R_x + R_{L2}]}{(R_x + R_{H2}) \cdot (R_x + R_{L2}) \cdot [\mathbf{b} \cdot (R_{H1} - R_{L1}) + R_{ref} + R_{L1}]} \quad (12)$$

Equation (12) shows that the full elimination of undesirable factors is only theoretical. But if on-resistances of switches in high and low states were equal, it would be possible to remove the variable β . As it has been specified in (11), this variable is the function of the most undesired factors and it amounts to their rejection.

$$\text{If: } R_{H2} = R_{L2} = R_2, \quad R_{H1} = R_{L1} = R_1 \quad (13)$$

equation (12) can be simplified to:

$$\mathbf{h}_d = \frac{R_{ref} + R_1}{R_x + R_2} \quad (14)$$

It means that non-zero values of on-resistance will always influence the result and this method is unable to eliminate it. The best way to reduce these errors is to try fulfil conditions when:

$$R_{ref} \gg (R_{H1}, R_{L1}) \quad \text{and} \quad R_x \gg (R_{H2}, R_{L2}) \quad (15)$$

Another method of solving this problem could be to correct the final result. It is already possible if on-resistances of switches are known and do not change in time importantly. The transformation of equation (12) leads to:

$$\mathbf{h}_d = \frac{R_{ref}}{1} \cdot \left[\frac{R_{ref} + R_{H1} + R_{L1} + \frac{R_{H1} \cdot R_{L1}}{R_{ref}}}{R_{ref} + \mathbf{b} \cdot R_{H1} + (1 - \mathbf{b}) \cdot R_{L1}} \right] \cdot \frac{1}{R_x} \cdot \left[\frac{R_x + \mathbf{b} \cdot R_{H2} + (1 - \mathbf{b}) \cdot R_{L2}}{R_x + R_{H2} + R_{L2} + \frac{R_{H2} \cdot R_{L2}}{R_x}} \right] \quad (16)$$

The factor in the first square bracket is a constant value and does not depend on the measured resistance. Hence it can be easily replaced by a correction constant for the reference resistor R_{ref} :

$$\mathbf{a}_1 = \frac{R_{ref} + R_{H1} + R_{L1} + \frac{R_{H1} \cdot R_{L1}}{R_{ref}}}{R_{ref} + \mathbf{b} \cdot R_{H1} + (1 - \mathbf{b}) \cdot R_{L1}} \quad (17)$$

In this case the corrected value of R_x is equal to:

$$R_x = \frac{\mathbf{a}_1 R_{ref} - (R_{H2} - R_{L2}) \cdot \mathbf{h}_d + \sqrt{\mathbf{a}_1^2 R_{ref}^2 + 2\mathbf{a}_1 R_{ref} \cdot (2\mathbf{b} - 1) \cdot (R_{H2} - R_{L2}) \cdot \mathbf{h}_d + (R_{H2} - R_{L2})^2 \cdot \mathbf{h}_d^2}}{2\mathbf{h}_d} \quad (18)$$

If the value of β is exactly equal to 0.5, the voltage on inverting input of the amplifier X1 is precisely equal to half of V_{cc} , the Eq. (12) can be additionally simplified:

$$R_x = \frac{\alpha_1 R_{ref} - (R_{H2} - R_{L2}) \cdot h_d + \sqrt{\alpha_1^2 R_{ref}^2 + (R_{H2} - R_{L2})^2 \cdot h_d^2}}{2h_d} \quad (19)$$

Unfortunately it does not mean that the whole elimination of influence of Vcc and other undesired factors on the measurement result will be achieved, because parameter β is included in constant α_1 and there is not any other way to remove it as to fulfil the condition (15). It worth noticing that in β the offset voltage of X1 and the current of its non-inverting input are included and these factors strongly depend on temperature and differ one from another depending on a component. Moreover it neither possible nor purposeful to create the resistant divider of Vcc to achieve exactly half of Vcc voltage. First of all on-resistances of switches should be minimized and the condition of their equality fulfilled. Nowadays it is easier to minimize on-resistance due to technology progress. Presently one can find MOSFET transistors with the value of this resistance of about 10 m Ω . The second condition is a bit more difficult because transistors with n-channel still have lower values of on resistance than with p-channel. This problem is especially important in the case of direct use of microcontroller port lines. Usually there is strong asymmetry between resistance of line in high and low state. Figure 4 shows the error of measurement for direct switching by a microcontroller line and the efficiency of the rejection of influences of circuit voltages. The value of voltage U+ has been changed by -5% to check it. It is already equivalent to the change of the value β . The parameters of the microcontroller line were: about 120 Ω when the line was forced high and about 20 Ω when forced low. In this case measurement error is only acceptable when the values of the measured resistance are close to the value of reference resistor. By increasing Rref the expected measurement error can be decreased, because the ratio Rref and Rx to resistances of switches is increased. But the condition of balancing charge forces that measured resistance cannot be lower than the value of the Rref, so by changing the Rref one simultaneously changes the measurement range. If the Rref is increased a minimal measured value of the resistance will be also higher, it is a disadvantage. High limit theoretically does not exist, but while the resistance goes to infinity the quantization error goes to infinity too and in fact a minimal acceptable resolution points out a maximal measured value of resistance.

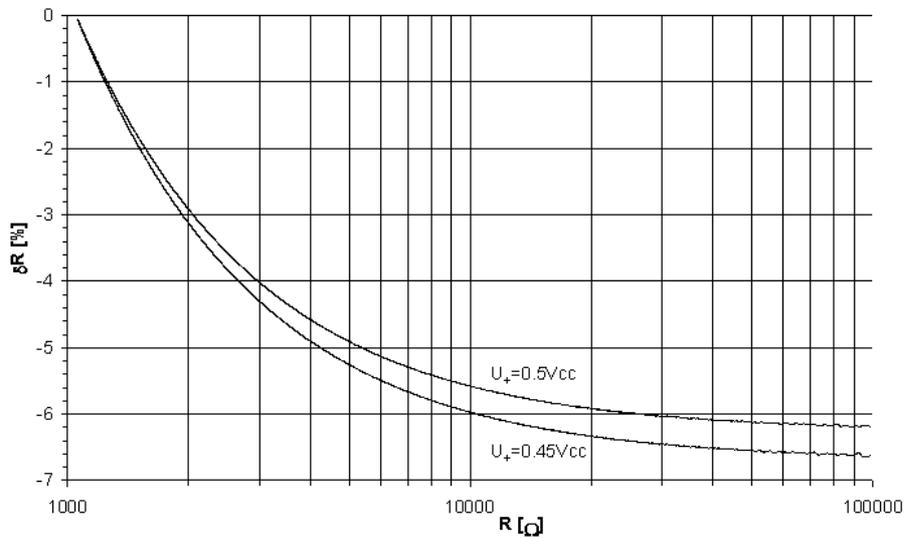


Figure 4. The relative error of measurement versus measured resistance $\delta R=f(R)$ for Rref=1050 Ω and direct switching by the microcontroller line.

Figure 5 presents results of simulation that has been done for the case of very small values of on-resistances: 10 m Ω in low state and 30 m Ω in high state. As previously, the characteristic has been drawn for two values of U₁ but in this case both curves have overlapped. It means that the influence of voltages in the circuit on the accuracy of measurement radically decreased. The value of error evoked by non-zero on-resistance of switches is similar or less than the quantization error. Thus the values of on-resistances are no longer important and the accuracy seems to become perfect. But in practical realization of such circuit an additional problem may appear: transistors with very low on-resistances usually have a big capacitance of gate that limits the speed of switching them and could cause additional dynamic errors which have not been considered in this paper.

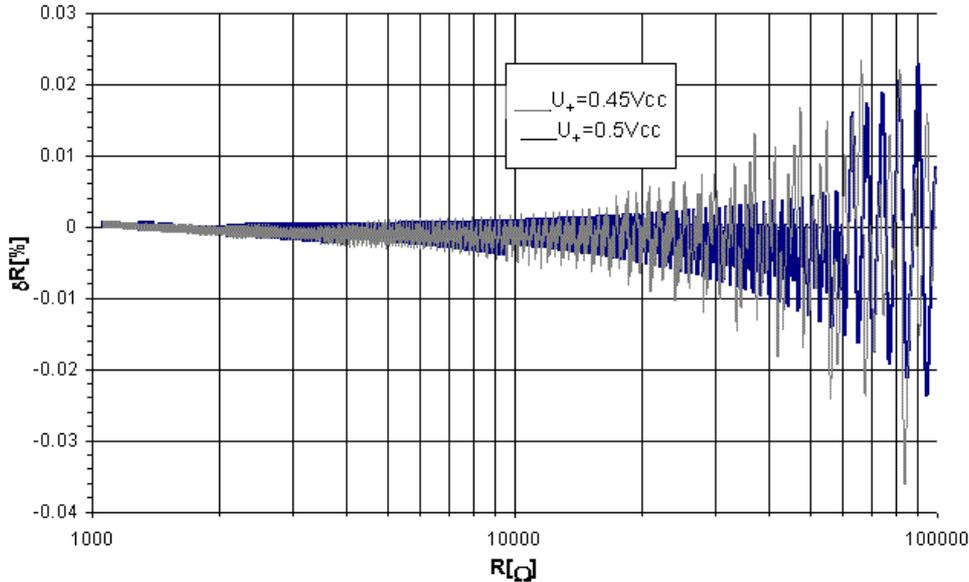


Figure 5. The result of a simulation of relative error versus measured resistance $\delta R=f(R)$ for $R_{ref}=1050 \Omega$ and switching by MOSFET transistors with low on-resistance.

4 CONCLUSION

In the article an application of charge balancing circuit to the measurement of resistance has been presented. The “auto zero” method has been used which enabled the reduction of influences not only offset voltages of operational amplifiers, but also other voltages in the circuit, especially V_{cc} . Precise and high stable voltage have become redundant. It is enough that V_{cc} is stable during both stages of the measurement.

This method can be an alternative to the similar but simpler “dual-slope” method. Although it requires fast switching but it does not have limitation in reaching higher values of resolution, which in “dual-slope” converter is connected with a problem of accurate and fast detection of the equality of two voltages, one of which is slowly rising or falling saw. Additionally, parameters of components are less critical than in “dual-slope” method, especially the capacitor C . By the right choice of a conversion time or use of a digital filter with an adequately shaped frequency characteristic, a perfect rejection of signals of some undesired frequencies or whole bands of frequencies, for example: line frequency and its higher harmonics can be obtained.

When the high ratio of the measured and reference resistances to the on-resistances of switches is kept, the processing function depends only on the former so no process of calibration of this converter is required. Also the characteristic of conversion depends on this ratio. Unfortunately full overlapping of the theoretical – Eq. (6) and real characteristics is not possible to obtain when the switches have non-zero resistances. But the progress in technology of MOSFET transistors with extremely low on-resistance that has been made in recent years and will be done in future can make this method more attractive to use.

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