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A High-Resolution, Linear Resistance-to-Frequency Converter

Kouji Mochizuki and Kenzo Watanabe

***Abstract*—A resistance-to-frequency converter consisting of a Wheatstone bridge followed by an integrator and a comparator is described. In concept the circuit represents a relaxation oscillator whose frequency changes linearly with a resistance change detected by the bridge. Analyses show that a resolution better than 0.05% is possible with the simple configuration, and an excellent linearity is maintained over the wide resistance change by using a simple compensation method. The converter is therefore suited as a signal conditioner of a resistive sensor. Experimental results are included to demonstrate its performance.**

I. Introduction

Detection of a small resistance change is often needed in industrial and process control systems and medical instrumentation [1], [2]. In such systems, microprocessors are used exclusively for digital signal processing, and thus input data in a digital format is highly desired. A conventional way of detecting the resistance in the digital form is to use a relaxation oscillator whose frequency is determined by the resistance under measurement and a known capacitor [3], [4]. Though simple, this method suffers from nonlinearity. In addition, high sensitivity cannot be expected since the oscillation frequency is inversely proportional to the total resistance.

The most sensitive means of detecting the resistance change is a Wheatstone bridge. Combining a Wheatstone bridge with a relaxation oscillator appears to be a promising approach to the high-sensitivity resistance-to-frequency (R-to-F) converter. Such a scheme has been proposed by Johnson and Richeh [5]. Their circuit is a resistance-to-duty-ratio converter, and the feedback circuit, consisting of an integrator and a comparator, maintains the balance of the bridge by controlling the switched resistors incorporated in the arms of the bridge. This paper describes a true R-to-F converter designed based on the above-mentioned approach. Different from the earlier circuit [5], the Wheatstone bridge is used merely to detect the resistance deviation from the offset value. Since no balance operation is involved, the circuit configuration is much simpler. The sensitivity is high enough, because of the differential integration involved, to be compared to that of the resistance-to-duty-ratio converter [5]. Therefore, the R-to-F converter described in detail in the following sections will find wide applicability as the signal conditioner of platinum RTD's for temperature measurements, strain gauges for electronic balances, piezoresistors for pressure detection, and the other resistive sensors for instrumentation and measurements.

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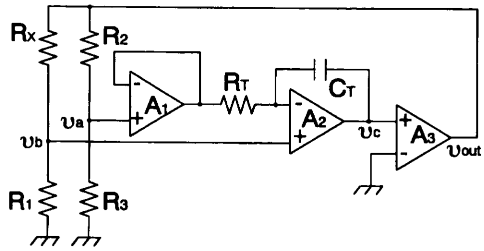


Fig. 1. A linear resistance-to-frequency converter: Basic configuration.

II. Circuit Description

Fig. 1 shows the circuit diagram of the linear resistance-to-frequency (R-to-F) converter. It is essentially a relaxation oscillator consisting of a Wheatstone bridge, an integrator, and a zero-crossing detector. The resistor R_x whose resistance change is to be detected forms one arm of the bridge. The unbalance voltage due to the resistance change is integrated, and its polarity is fed back to the bridge as the bias voltage, to sustain the oscillation.

When op-amps involved are ideal, node voltages are given by

$$v_a = \alpha v_{\text{out}} \tag{1}$$

$$v_b = \beta v_{\text{out}} \tag{2}$$

$$v_c = v_b - \frac{1}{\tau}\int (v_a - v_b)dt \tag{3}$$

where

$$\alpha = \frac{R_3}{R_2 + R_3} \tag{4}$$

$$\beta = \frac{R_1}{R_1 + R_x} \tag{5}$$

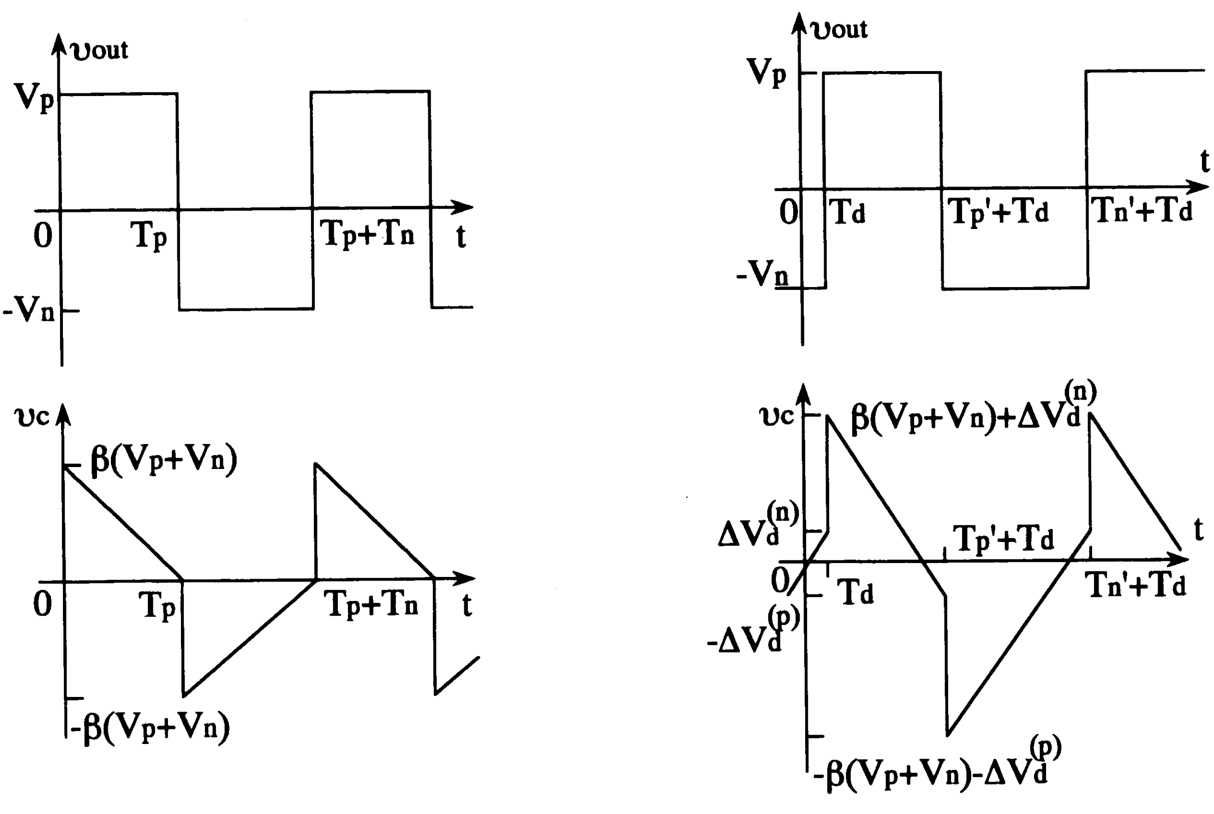
and

$$\tau = C_T R_T. \tag{6}$$

Output voltage waveforms of the zero-crossing detector and the integrator are shown in Fig. 2(a). Here,

$$v_{\text{out}} = \begin{Bmatrix} V_P, & \text{if } v_c \geq 0 \\ -V_n, & \text{if } v_c < 0 \end{Bmatrix}. \tag{7}$$

is assumed, and T_p and T_n are the periods during which v_{out} assumes V_p and $-V_n$, respectively. Referring to these waveforms, one can



(a) (b)

Fig. 2. Voltage waveforms in Fig. 1, (a) when op-amps are ideal and (b) when the zero-crossing detector has the response delay T_d.

derive the following expressions for T_p and T_n

$$T_p = \frac{\beta}{\alpha - \beta}(1+\gamma)\tau \tag{8}$$

$$T_n = \frac{\beta}{\alpha - \beta}\left(1 + \frac{1}{\gamma}\right)\tau \tag{9}$$

where

$$\gamma = |V_n/V_p|. \tag{10}$$

The oscillation frequency is thus given by

$$f = \frac{\alpha - \beta}{\beta}\frac{\gamma}{(1+\gamma)^2\tau} = f_0 + \Delta f \tag{11}$$

where

$$f_0 = \frac{R_0R_3 - R_1R_2}{R_1(R_2+R_3)}f_\tau \tag{12}$$

is the offset frequency corresponding to the bridge offset,

$$\Delta f = \frac{R_3\Delta R}{R_1(R_2+R_3)}f_\tau \tag{13}$$

is the frequency shift due to the resistance change ΔR in R_x, and

$$f_\tau = \frac{\gamma}{(1+\gamma)^2}\frac{1}{\tau} \tag{14}$$

$$R_x = R_0 + \Delta R. \tag{15}$$

Equation (13) indicates that the resistance change is linearly converted into the frequency.

One of the error sources which degrades the linearity is the response delay of the zero-crossing detector. Waveforms when the delay time is T_d are shown in Fig. 2(b). Here,

$$\Delta V_d^{(p/n)} = \frac{T_d}{\tau}(\alpha - \beta)v_{\text{out}} = \frac{T_d}{\tau}(\alpha - \beta)V_{(p/n)} \tag{16}$$

is the voltage change in the integrator output during the delay time T_d. Referring to Fig. 2(b), T_p' and T_n' are derived as

$$T_p' = T_p + (1+\gamma)T_d \tag{17}$$

$$T_n' = T_n + \left(1 + \frac{1}{\gamma}\right)T_d. \tag{18}$$

T_p and T_n given by (8) and (9), respectively, change inversely with ΔR, and thus the nonlinearity due to T_d increases with ΔR.

Waveforms in Fig. 2(b) suggest the compensation method for the response delay: The effect of the delay time upon the oscillation frequency could be eliminated by replacing the zero-crossing detector by the comparator which compares the integrator output with $\Delta V_d^{(p/n)}$. Substituting the comparator threshold voltage $\Delta V_d^{(p/n)}$ given by (16) into (3) explicitly, one can get the design equation which describes the required operations

$$v_c - \beta v_{\text{out}} - \frac{T_d}{\tau}(\alpha - \beta)v_{\text{out}} = -\frac{1}{\tau}\int(\alpha - \beta)v_{\text{out}}dt. \tag{19}$$

The right-hand side indicates the differential integration while the left-hand side the comparison. The threshold in the comparison is βv_{out}, shifted by $\Delta V_d^{(p/n)}$ for the delay compensation.

The delay-compensated R-to-F converter thus designed is shown in Fig. 3(a). It should be noted that the voltage at the inverting input terminal Ⓐ of op-amp A_1 is βv_{out}; and R_3 is split into two parts, δR_3 and $(1-\delta)R_3$, to shift the threshold voltage. Fig. 3(b) shows waveforms of the integrator output v_c and the threshold voltage v_{TH}, which is given by

$$v_{\text{TH}} = \left\{\{\beta(1-\delta)\} - \frac{\delta(\alpha-\beta)}{1-\alpha}\right\}v_{\text{out}}. \tag{20}$$

Referring to these waveforms, one obtains

$$T_p = \frac{\beta(1-\delta)(1+\gamma)(1-\alpha)}{\alpha - \beta}\tau + (1+\gamma)(T_d - \delta\tau) \tag{21}$$

$$T_n = \frac{\beta(1-\delta)(1+\gamma^{-1})(1-\alpha)}{\alpha - \beta}\tau + \left(1 + \frac{1}{\gamma}\right)(T_d - \delta\tau). \tag{22}$$

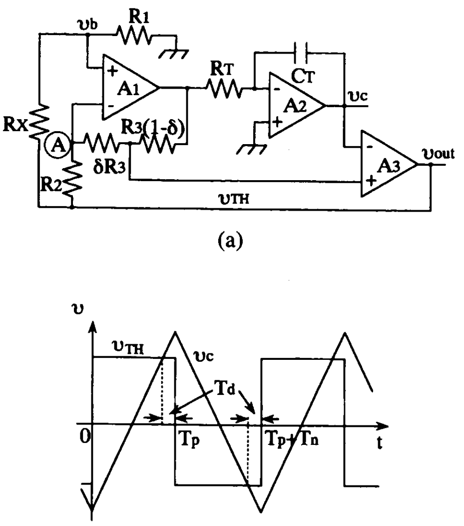


Fig. 3. (a) The circuit diagram of the delay-compensated resistance-to-frequency converter, (b) Voltage waveforms at the input terminals of comparator A_3.

Therefore, if δ is chosen such that

$$\delta\tau = T_d \tag{23}$$

then the oscillation frequency changes linearly with ΔR

$$\begin{aligned} f &= f_0 + \Delta f \\ &= \frac{1}{(1-\delta)R_1R_2}\{(R_3R_0 - R_1R_2) + R_3\Delta R\}f_\tau. \end{aligned} \tag{24}$$

III. Resolution

Another error source that influences the converter operation shown in Fig. 3(a) is the offset voltage of an op-amp. Let V_{os1} and V_{os2} be the offset voltages of op-amps A_1 and A_2, respectively. These offset voltages produce the constant error voltage. The converter is driven by its output voltage v_{out} which assumes the constant value V_p or V_n. Therefore, the offset voltage is equivalent to the change in v_{out} and scales T_p and T_n as

$$T_p' = \frac{T_p}{1-\Delta} \tag{25}$$

$$T_n' = \frac{T_n}{1+\Delta/\gamma} \tag{26}$$

where

$$\Delta = \frac{1-\alpha}{\alpha-\beta}\left(\frac{\varepsilon_1}{(1-\alpha)} - \varepsilon_2\right) \tag{27}$$

and

$$\varepsilon_1 = V_{\text{os1}}/V_p \tag{28}$$

$$\varepsilon_2 = V_{\text{os2}}/V_p. \tag{29}$$

The oscillation frequency f' is then given by

$$f' = \frac{1}{T_p' + T_n'} = f(1-\Delta)(1+\Delta/\gamma) \tag{30}$$

where f is the oscillation frequency of the delay-compensated R-to-F converter given by (24).

It is clear from (27) that if the amplified offset voltage of op-amp A_1 cancels the offset voltage of A_2, then no offset error appears. This condition does not always hold true because of the temperature drifts in the offset voltages. The deviation Δf from the offset frequency f_0 then consists of two parts; one is Δf_R due to the resistance change

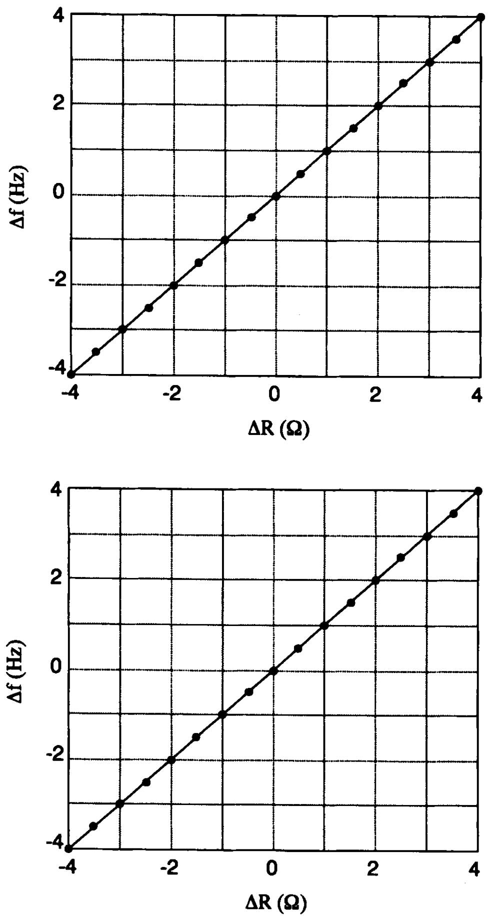


Fig. 4. The oscillation frequency change Δf of the prototype converter for the resistance change ΔR.

ΔR, and the other is Δf_ε due to the offset voltage. Δf_ε is given, to the first order, by

$$\begin{aligned} \Delta f_\varepsilon &= f_0\Delta\left(1-\frac{1}{\gamma}\right) \\ &= \frac{1}{1-\delta}\cdot\left(1+\frac{R_0}{R_1}\right)\left(\frac{\varepsilon_1}{1-\alpha} - \varepsilon_2\right)\left(1-\frac{1}{\gamma}\right)f_\tau. \end{aligned} \tag{31}$$

For the resistance change ΔR to be detected unambiguously, Δf_R should be larger than Δf_ε. The resolution limited by this condition is given by

$$\frac{\Delta R_{\min}}{R_0} = \left(1+\frac{R_1}{R_0}\right)\frac{1}{\alpha}\left(1-\frac{1}{\gamma}\right)\{\varepsilon_1 - (1-\alpha)\varepsilon_2\} \tag{32}$$

where $\Delta R_{\min}$ denotes the minimum detectable resistance change.

The offset voltage of comparator A_3 also affects the converter operation. Its effect on the oscillation frequency is second-order, because it influences only the delay compensation through shifting the threshold voltage. Another factor that limits the resolution is the temperature drift Δf_T of the offset frequency. Δf_T can be evaluated by

$$\frac{\Delta f_T}{f_0} = \left(\frac{\partial f_0}{f_0}/\partial T\right)\Delta T \tag{33}$$

where ΔT represents the operating temperature range of the converter. The frequency stability $(\partial f_0/f_0)/\partial T$ of a conventional relaxation oscillator built using resistors and capacitors with low temperature coefficients is typically $10^{-5}/°\text{C}$. Assuming $\Delta T = 40°\text{C}$, the resolution limited by the temperature drift is

$$\frac{\Delta R_{\min}}{R_0} = 4\times10^{-4}\left(1-\frac{R_1R_2}{R_0R_3}\right). \tag{34}$$

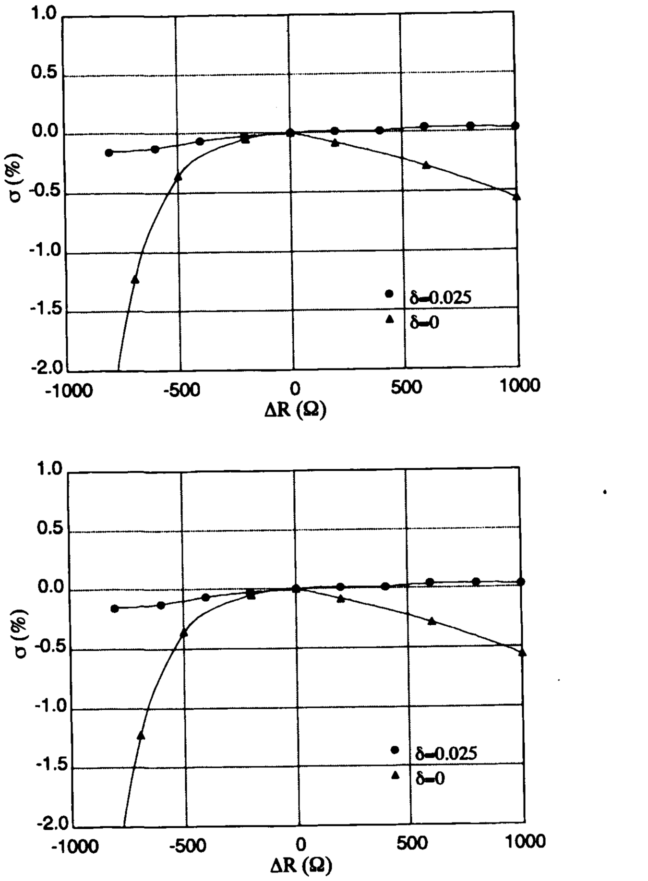


Fig. 5. The nonlinear error σ of the prototype converter with ($\delta = 0.025$) and without ($\delta = 0$) the delay compensation.

Assuming $V_{os1} = -V_{os2} = 2$ mV, $V_p = 5$ V, $\gamma = 0.95, \alpha = 1/2$, $R_0 \approx R_1$, one can estimate the offset-limited resolution given by (32) to be

$$\frac{\Delta R_{\min}}{R_0} = 10^{-4}.$$

Summarizing the above estimation, it is concluded that the resolution limited by the offset voltages is comparable to that limited by the temperature drift of the offset frequency, and the overall resolution is of the order of 0.05%.

IV. Experimental Results

To confirm the principle of operation, the R-to-F converter shown in Fig. 3(a) was breadboarded using off-the-shelf components. The op-amps used are LF411. Its offset voltage is less than 2 mV, and the delay time is about 2.7 μs. No adjustment was made for the offset compensation. The power supplies are ± 12 V. The other relevant parameters are $R_1 = R_3 = 1.6$ kΩ, $R_2 = 2.4$ kΩ. The conversion sensitivity was adjusted to be 1 Hz/Ω. The integration time constant $\tau = C_T R_T$ was then about 110 μs. The oscillation frequency was measured by the frequency counter with 0.1 Hz resolution.

Fig. 4 shows the measured frequency changes when R_x was changed in 0.5 Ω steps from its offset value of $R_0 = 3.4$ kΩ. The results indicate that the resolution is higher than 0.015%. Fig. 5 shows the nonlinear error σ of the prototype converter with ($\delta = 0.025$) and without ($\delta = 0$) the delay compensation. The nonlinear error σ is defined here by

$$\sigma = \frac{\Delta f \text{ measured} - \Delta f \text{ given by (24)}}{\Delta f \text{ given by (24)}}. \tag{35}$$

The results shown in Fig. 5 indicate that a resolution higher than 0.1% is achievable over the wide resistance range with the delay compensation, thus confirming the performance evaluation in the previous sections.

V. Conclusion

A novel circuit has been presented which provides the frequency equivalent of a resistance change. The component requirement is minimum. Besides the simple configuration, it features a high resolution and an excellent linearity over the wide resistance range. These advantages make the R-to-F converter proposed herein especially useful for signal processing of such transducers as strain gauges, RTD's, and piezo resistive devices.

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A Simple Measurement Method to Determine the Burst Acquisition Time of Digital Systems

Curtis Hatamoto, Huajing Fu, and Kamilo Feher

Abstract—**A cost-efficient solution for measuring fast and ultrafast synchronization-caused bit errors is disclosed. This invention enables the user to modify conventional test instruments and to eliminate the need to obtain very expensive test equipment. Our new instrumentation implementation concept comprises a gated switch added to a pseudorandom bit sequence generator and analyzer. The synchronization time of a couple of circuits is tested.**

I. Introduction

In certain applications, such as in the wireless field, synchronization time is crucial to the transmission of data. For instance, in frequency-hopping spread spectrum (FH-SS), the carrier frequency is constantly

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