Application of Analog Adaptive Filters for Dynamic Sensor Compensation

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***Abstract*—This paper investigates the application of analog adaptive techniques to the area of dynamic sensor compensation, of which there is little reported work in the literature. The case is illustrated by showing how the response of a load cell can be improved to speed up the process of measurement. The load cell is a sensor with an oscillatory response in which the measurand contributes to the response parameters. Thus, a compensation filter needs to track variation in measurand, whereas a simple fixed filter is only valid at one specific load value. To facilitate this investigation, computer models for the load cell and the adaptive compensation filter have been developed. To allow a practical implementation of the adaptive techniques, a novel piecewise linearization technique is proposed in order to vary a floating voltage-controlled resistor in a linear manner over a wide range. Simulation and practical results are presented, thus, demonstrating the effectiveness of the proposed techniques.**

***Index Terms*—Analog adaptive filter, dynamic sensor, load cell, response compensation.**

I. Introduction

LOAD CELLS are used in a variety of industrial weighing applications such as vending machines and checkweighing systems. Since information processing and control systems cannot function correctly if they receive inaccurate input data, compensation of the imperfections of sensors is one of the most important aspects of sensor research. Influence of unwanted signals, nonideal frequency response, parameter drift, nonlinearity, and cross sensitivity are the five major defects in primary sensors [1]. In the new generation of sensors, called intelligent or smart sensors, the influence of these imperfections has been dramatically reduced by using signal processing techniques, which have resulted from advances in the field of digital systems.

Some sensors, such as load cells, have an oscillatory response which needs time to settle down. Dynamic measurement refers to the ascertainment of the final value of a sensor signal while its output is still in oscillation. It is, therefore, necessary to determine the value of the measurand in the fastest time possible to speed up the process of measurement, which is of particular importance in some applications including vending machines and checkweighing systems. One example of processing that can be done on the sensor output signal is filtering to achieve response correction. Several methods have been reported addressing this problem. Software techniques for sensor compensation are reviewed in [2]. Digital adaptive techniques have been used in [3] for load cell response correction. An artificial neural network has been proposed for dynamic measurement which needs a learning phase [4]. Other methods, such as employing a Kalman filter [5] and estimation with a recursive least square (RLS) procedure [6], have also been applied for dynamic weighing systems. Almost all the above reported methods are based on digital signal processing techniques which need analog-to-digital convertors and powerful signal processors. Although digital techniques have been used efficiently, the aim of this paper is to investigate the possibility of using analog adaptive techniques for load cell response correction. The potential benefits of analog adaptive techniques compared to digital methods include higher signal processing speeds, lower power dissipations, and smaller integrated circuit areas. It should be noted that most applications of analog adaptive techniques have focused on communications and digital magnetic storage [7], and there has been little or no work on application of analog adaptive techniques to intelligent sensors which is the main focus of this paper.

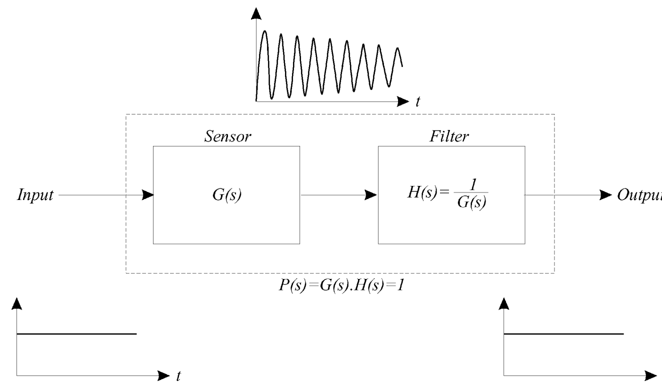

Fig. 1. General principle of load cell response correction.

II. Load Cell Response Compensation

The primary sensor is considered as a system with transfer function $G(s)$. The general principle for eliminating the transient time is shown in Fig. 1. A filter having the reciprocal characteristic of the sensor is cascaded with it. Therefore, the transfer function of the whole system is *"unity,"* which means that any changes in the input transfer to the output without any distortion. The response of a load cell can change for different measurands. For example, the characteristic of a load cell changes when a load is applied to it because the mass of the load contributes to the inertial parameters of the system. Therefore, the transfer function of the filter should change

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accordingly. In other words, a fixed filter can be used only for one specific load value.

Previous work has shown that the load cell can be modeled as a second-order system [3]

$$(m+m_0)\cdot\frac{d^2y(t)}{dt^2}+c\cdot\frac{dy(t)}{dt}+k\cdot y(t)=F(t) \quad (1)$$

where m is the mass being weighed, m_0 is the effective mass of the sensor, c is the damping factor, k is the spring constant, and $F(t)$ is the force function. The Laplace transfer function of this sensor is

$$G(s)=\frac{Y(s)}{F(s)}=\frac{\frac{1}{m+m_0}}{s^2+\frac{c}{m+m_0}s+\frac{k}{m+m_0}}=\frac{A}{s^2+\frac{\omega_0}{Q}s+\omega_0^2}. \quad (2)$$

This shows that m affects all inertial parameters of the sensor such as gain factor A, quality factor Q, and natural frequency ω_0.

Equation (2) yields a pair of complex conjugate poles $a \pm jb$ where

$$a=-\frac{c}{2(m+m_0)} \quad (3)$$

and

$$b=\sqrt{\frac{k}{(m+m_0)}-\frac{c^2}{4(m+m_0)^2}}. \quad (4)$$

Thus, the zeros of the adaptive filter, which are the poles of the sensor, can be obtained.

In general, assume w is defined as a vector that contains all of the parameters of adaptive filter, i.e.,

$$\boldsymbol{w}=[w_1\ w_2\ w_3\ \ldots]^T. \quad (5)$$

The elements of $\boldsymbol{w}$ can be calculated for different values of the measurand. To emphasise that $\boldsymbol{w}$ depends on m, it can be written as $\boldsymbol{w}(m)$. m is unknown in the first instance when a new measurement begins. Therefore, the parameters of the adaptive filter cannot be set to appropriate values in order that the filter behaves as an inverse system. Hence, an adaptive rule is required to modify the parameters of the adaptive filter according to the value of measurand. This rule is a crucial element, but there is not a straightforward solution for it. Usually, in classic adaptive techniques, an adaptive algorithm, such as a least mean squares (LMS) method, updates $\boldsymbol{w}$ to minimize a cost function. However, (2) shows that, for a load cell, the suitable filter has a pair of conjugate zeros $z_{1,2}=a\pm jb$, which a and b can be considered as the parameters of adaptive filter and the relationship between them and load can be modeled as in (3) and (4). The real-time measurement operation is shown in Fig. 2. In this block diagram, m has been substituted with y, the output of the whole system, which is proportional to m. Initially, the zeros of the filter are set to arbitrary values. Then, the output y is calculated. This new value of y is used to calculate the zeros of the filter once again. Repeating these steps results in a rapid approach to obtain the steady state value of y.

So far, the zeros of the second-order compensation filter have been examined. In order that the analog filter can be realized, it is necessary to add at least two poles to the filter. The values of these poles can be determined practically. For simulation purposes, these poles are selected by trial and error so that the output of the filter quickly reaches its steady-state value with minimum oscillation. The transfer function of the compensation filter is

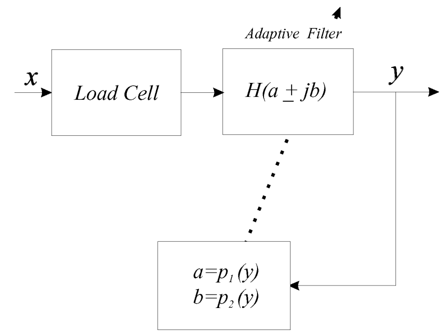

Fig. 2. Block diagram of adaptive load cell response correction.

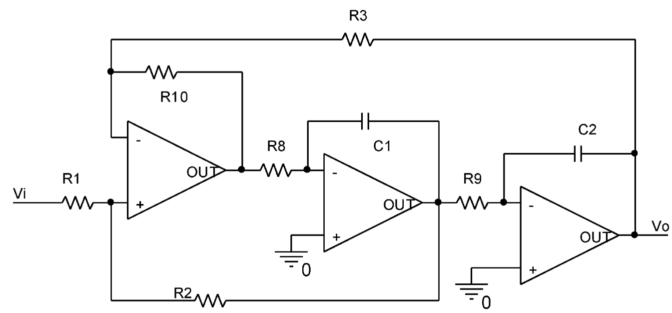

Fig. 3. State-variable low-pass filter.

$$H(s)=\frac{(m+m_0)}{10^{-5}}\cdot\frac{s^2+\frac{c}{m+m_0}s+\frac{k}{m+m_0}}{s^2+600s+10^5}. \quad (6)$$

The transfer functions of the load cell (2) and its compensation filter (6) are biquadratic functions. There now exists a wealth of theoretical and experimental information on the design of fixed or nonadaptive analog biquads [8]. The problem is how to make a biquad adaptive, and it is necessary to have only one filter component to track changes in m without any influence on the other parameters such as damping factor c and the spring coefficient k.

III. Proposed Load Cell Model

Amongst the various biquad structures, the state-variable lowpass filter [8], shown in Fig. 3, can be used to model the behavior of the load cell. The state-variable filter transfer function is

$$T_{sv}(s)=K1\frac{\frac{1}{R_8R_9C_1C_2}}{s^2+K1\left(\frac{R_1}{R_2R_8C_1}\right)s+\frac{R_{10}}{R_3R_8R_9C_1C_2}} \quad (7)$$

where

$$K1=\frac{R_2(R_3+R_{10})}{R_3(R_1+R_2)}. \quad (8)$$

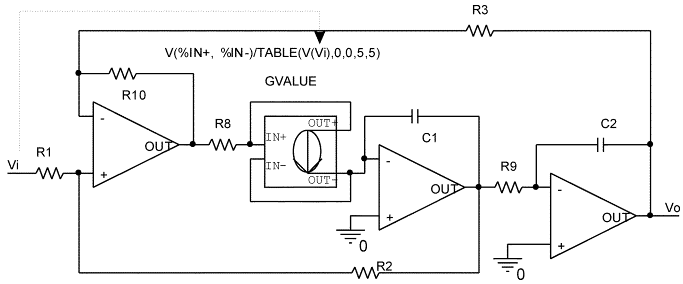

Fig. 4. Load cell model.

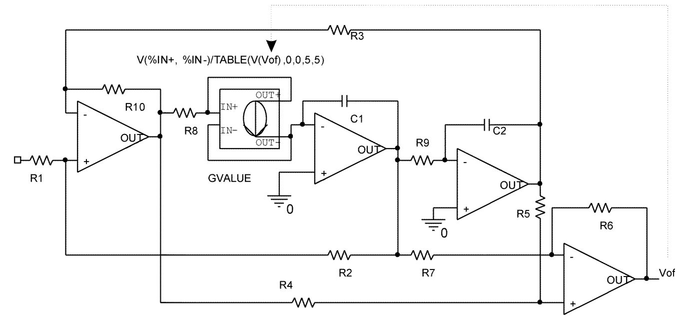

Fig. 5. Adaptive compensation filter model.

Comparing this transfer function with the load cell transfer function, (2), shows that R_8 can model $(m + m_0)$. R_8 has to be split into a fixed resistor equal to m_0 and an additional resistor proportional to m. Since m is the mass being weighed, and in the model it is equivalent to stimulating voltage (V_i), the resistor has to be a voltage-controlled device whose resistance should be directly proportional to V_i. The practical implementation of such a resistor will be discussed in Section VI. For simulation purposes, the analog behavioral modeling facility in PSPICE can be used. This is achieved by using the G component (a voltage-controlled current source) and "TABLE," which allows the user to enter different resistors for different voltages. Using this voltage-controlled resistor in the lowpass filter (Fig. 3) produces an analog biquadratic filter which can model the behavior of the load cell. The complete model is depicted in Fig. 4. From experimental data for a particular load cell [4], the damping factor c, spring constant k, and the effective mass of the load cell m_0 are 3.5, 2700 Pa, and 0.5 kg, respectively. These numbers are used to determine the values for resistors and capacitors in Fig. 4.

For step excitation, the input voltage of the model is a step function whose amplitude is proportional to m. The simulation results for two different values of m are shown in Figs. 6 and 7, which indicate that changing the input (similar to the practical case) varies all inertial parameters of the output waveform such as the steady state value, resonant frequency, and damping factor.

IV. Proposed Adaptive Compensation Filter Model

Since the transfer function of the compensation filter (6) is a biquadratic function, different scaled outputs in the state variable filter, shown in Fig. 3, need to be added to form a complete biquad. To make this biquad adaptive, as described in the block diagram of Fig. 2, the filter's zeros have to be changed by the output of the biquad. Similar to the sensor model approach, it is possible to use a voltage-controlled resistor in the compensation filter. The filter output voltage is used to control this resistance. The complete adaptive biquad is shown in Fig. 5. The transfer function of this filter is

$$H(s) = K\frac{s^2 + \frac{R_6(R_4+R_5)}{R_5R_8C_1(R_6+R_7)}s + \frac{R_4}{R_5R_8R_9C_1C_2}}{s^2 + K1\left(\frac{R_1}{R_2R_8C_1}\right)s + \frac{R_{10}}{R_3R_8R_9C_1C_2}} \tag{9}$$

where

$$K = K1\frac{R_5(R_6+R_7)}{R_7(R_4+R_5)} \tag{10}$$

and $K1$ was previously defined in (8). Similar to the sensor model, R_8 consists of a fixed resistor and a voltage-controlled resistor whose resistance is controlled by the filter's output voltage. In other words, R_8 models $(m + m_0)$ in (6).

The adaptation sequence will now be described in detail. Before stimulating the load cell, the filter output voltage is zero, and the initial transfer function of the filter will be

$$H_0(s) = A\cdot\frac{(s-a_0-jb_0)(s-a_0+jb_0)}{(s-d-je)(s-d+je)}. \tag{11}$$

Where A is the gain factor of the filter, a and b are the real and imaginary parts of filter's zeros respectively, d and e are real and imaginary parts of filter's poles, respectively, and the subscript ($_0$) denotes the initial values. The zeros of the filter need to cancel the poles of the sensor, i.e., a and b are the same as (3) and (4). Since m (the output of the filter) is unknown at first, the values of a and b, which depend on m, cannot be fixed. The initial values for a and b are

$$a_0 = -\frac{c}{2(0+m_0)} \text{ and } b_0 = \sqrt{\frac{k}{(0+m_0)} - \frac{c^2}{4(0+m_0)^2}}.$$

When the input is applied to the filter with initial transfer function of $H_0(s)$, it produces an output, say m_1. Since the zeros of the filter change with the output voltage, the new values for a and b will be a_1 and b_1, and then the transfer function of the filter changes to

$$H_1(s) = A\cdot\frac{(s-a_1-jb_1)(s-a_1+jb_1)}{(s-d-je)(s-d+je)}. \tag{12}$$

With this new transfer function, the filter produces a new output that changes the filter's zeros again and this procedure continues until a and b converge to their final values.

It should be noted that the poles of this compensation filter (9) vary as m varies, which is not the case with the filter model (6). However, detailed analysis has shown [9] that the filter remains stable for all values of m.

V. Simulation Results

In this section, the load cell and the adaptive compensation filter models will be used to examine how analog adaptive techniques can be used for load cell response correction. Fig. 6 shows the load cell output and the compensation filter output for $m = 0.1$ kg. To illustrate the capability in tracking changes in m, Fig. 7 shows the results when $m = 1$ kg. Clearly, the simulation results show that the analog adaptive biquad filter (shown

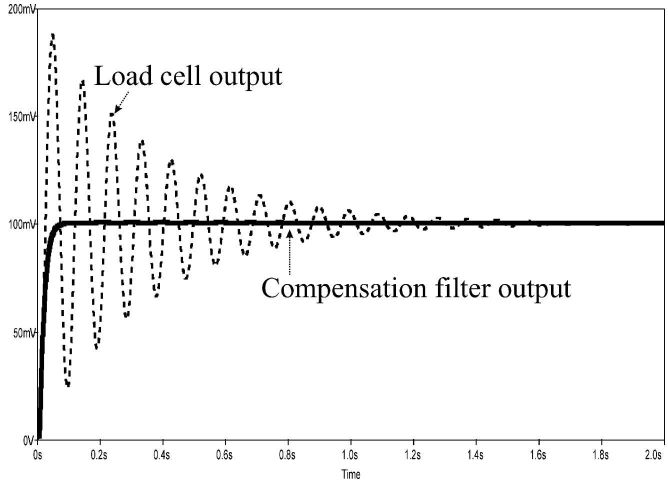

Fig. 6. Simulation result of adaptive compensation for $m = 0.1$ kg.

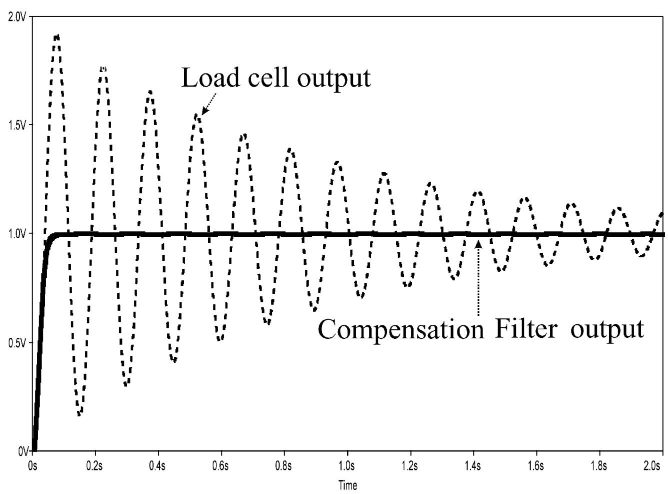

Fig. 7. Simulation result of adaptive compensation for $m = 1$ kg.

in Fig. 5) can be used to correct the response of a second-order sensor system.

VI. Practical Implementation

In the sensor and compensation models (Figs. 4 and 5), there is a floating linear voltage-controlled resistor (VCR) which must be directly proportional to the controlling voltage over a wide range because this resistor is equivalent to $(m_0 + m)$, as described in Sections III and IV. In PSPICE simulation, this VCR was modeled using a voltage-controlled current source whose voltage values and resistors were set by a theoretical table format, hence, there was no restrictions. In practice, however, such component does not exist. The challenging part of the practical implementation of the sensor and compensation filter models is the realization of such a VCR. The properties of this VCR can be summarized as follows.

- The resistor should be floating (none of its terminals is connected to ground or power supply).
- The resistor should be linear (linear relationship between voltage and current).
- The relationship between resistance and controlling voltage should be linear over a wide range, enhancing its practical application.

A. Voltage-Controlled Resistor

Junction field-effect transistors (JFETs) are used as voltage-controlled resistors [10]–[13] where the channel resistance can be controlled by applying a variable voltage between gate and source. There are, however, two major problems: 1) voltage across the channel should be small, i.e., for large $V_{\rm DS}$, the channel is a nonlinear resistor and 2) channel resistance is inversely proportional to gate voltage. There have been some papers that address linearizing the current and voltage relationship (the first problem) [10], [14], [13], and [11]. However, an extensive literature search has shown that there is very little or no work in the area of producing a resistor that varies with input voltage in a linear manner. In this section, these two problems are addressed and a novel linearization technique is proposed to solve the second one.

The relationship between drain current, I_D, and drain–source voltage, $V_{\rm DS}$, of a JFET in the triode or ohmic region [$V_{\rm DS} < (V_{\rm GS} - V_{\rm GS(off)})$] is

$$I_D = \frac{2I_{\rm DSS}}{V_P^2}[(V_P - V_{\rm GS})V_{\rm DS} - 0.5V_{\rm DS}^2] \tag{13}$$

which shows a nonlinear relationship between I_D and $V_{\rm DS}$ (a nonlinear resistance). For small values of $V_{\rm DS}$, the square term in (13) can be ignored and in this case the drain–source resistance can be considered as a linear resistance, which in practice limits the voltage across the resistance to several hundred millivolts. It is possible to have a linear resistance over a wider range [10], [14], [13], and [11]. However, since in the analog adaptive filter model, the voltage across VCR is very small; $(V_{\rm DS})^2$ in (13) can be ignored without any significant impact. In this case, for $V_{\rm GS}$ less than $V_{\rm GS(off)}$, the channel resistance becomes

$$R_{\rm DS} = \frac{R_{\rm DS(on)}}{\left(1 - \frac{V_{\rm GS}}{V_P}\right)} \tag{14}$$

where $R_{\rm DS(on)} = V_P/2I_{\rm DSS}$ is the minimum resistance for $V_{\rm GS} = 0$. Equation (14) shows that $R_{\rm DS}$ is inversely proportional to controlling voltage $V_{\rm GS}$. Equation (14) can be rewritten as

$$R_{\rm DS} = R_{\rm DS(on)} \cdot \left[1 + \left(\frac{V_{\rm GS}}{V_P}\right) - \frac{1}{2} \cdot \left(\frac{V_{\rm GS}}{V_P}\right)^2 + \frac{1}{6} \cdot \left(\frac{V_{\rm GS}}{V_P}\right)^3 + \cdots\right]. \tag{15}$$

Equation (15) can be approximated as a linear equation if $1/2 \cdot (V_{\rm GS}/V_P)^2 \ll (V_{\rm GS}/V_P)$. This restriction is equivalent to $m \ll m_0$, which means that the mass being weighed should be much less than the effective mass of the sensor. However, in practice, typically, m can be several order of magnitude bigger than m_0. To increase the capability of the compensation filter over a wider range, we propose to introduce a nonlinear amplifier between the output of the filter, which is controlling voltage (V_C), and the gate of JFET (Fig. 8). The nonlinear relation between output Y and input X of this amplifier can be

Fig. 8. Nonlinear amplifier which provides the gate voltage from V_C.

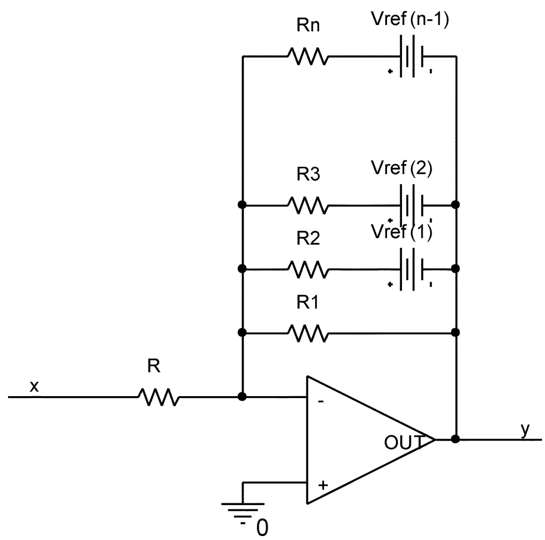

Fig. 9. Implementation of the proposed piecewise linear approximation technique.

obtained as follows. The VCR needs to change linearly with V_C, i.e.,

$$R_{\mathrm{VCR}} = R_0(1 + \alpha \cdot V_C) = R_0(1 + \alpha \cdot X) \tag{16}$$

where α is a proportional constant. The relationship between R_{DS} of the JFET with its V_{GS} is given in (14), and it is required that this resistance changes linearly with V_C. In other words, (14) and (16) should be equal

$$\frac{R_{\mathrm{DS(on)}}}{\left(1 - \frac{Y}{V_P}\right)} = R_0(1 + \alpha \cdot X). \tag{17}$$

The value of R_0 is not important, and it is assumed $R_0 = R_{\mathrm{DS(on)}}$. By manipulating (17), the input-output relationship of the nonlinear amplifier is

$$Y = \frac{V_P \cdot \alpha \cdot X}{1 + \alpha \cdot X}. \tag{18}$$

One possible implementation of (18) is shown in Fig. 9, which realize piecewise linear approximation of this equation. This circuit has n different gains for different intervals ($V_{ref(i-1)} < |Y| < V_{ref(i)}$)

$$A_{v(i)} = -\frac{R_1 \| R_2 \ldots \| R_i}{R}, \qquad \text{for} \quad i = 1, 2, \ldots n. \tag{19}$$

To calculate the gains, (18) can be approximated by n straight lines, as shown in Fig. 10, for $n = 3$. The breaking points of this curve are $V_{GS(i)} = Y_{(i)} = V_{ref(i)}$ (for $i = 1, 2, \ldots n$). Corresponding inputs of the nonlinear amplifier $X_{(i)}$ can be calculated from (18). This approximation is equivalent to n piece of straight lines and, hence, needs an amplifier with n different gains. Slopes of the piecewise lines $k_{(i)}$ are equal to the gains of amplifie $A_{v(i)}$

$$k_{(i)} = \frac{\Delta Y_{(i)}}{\Delta X_{(i)}} = \frac{Y_{(i)} - Y_{(i-1)}}{X_{(i)} - X_{(i-1)}} = A_{v(i)}, \text{ m for } i = 1, 2, \ldots n \tag{20}$$

where $Y_{(0)} = X_{(0)} = 0$.

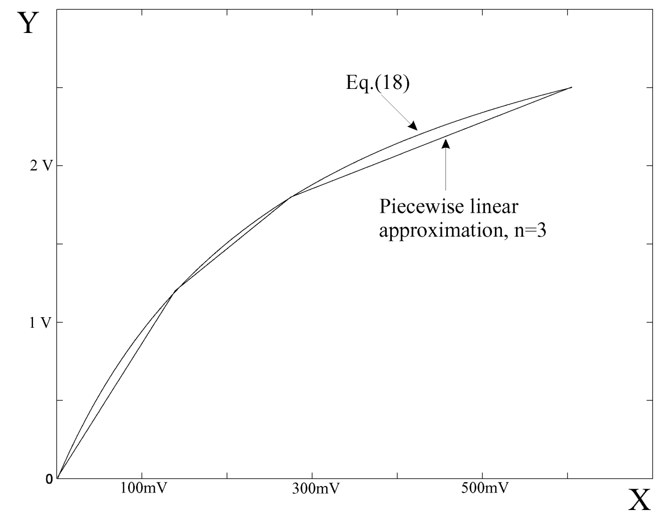

Fig. 10. Piecewise linear approximation of (18).

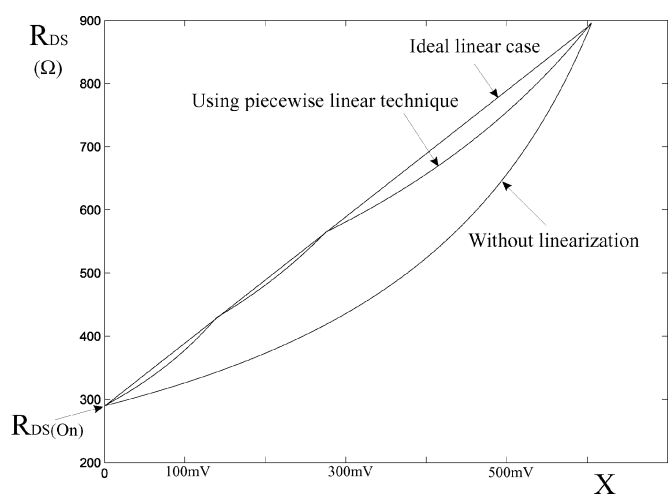

Fig. 11. Linearization of R_{DS} using the proposed piecewise linear technique.

Having used this piecewise linear amplifier to provide the gate voltage, Fig. 11 shows R_{DS} of the JFET with changing V_C. To indicate the effectiveness of the proposed technique, the ideal linear case and nonlinear case are also depicted. This figure shows that R_{DS} can be controlled linearly over a wide range (more than three times $R_{\mathrm{DS(on)}}$). In other words, the adaptive compensation filter can be used for the values of m as large as $3m_0$.

B. Experimental Results

The complete analog adaptive compensation filter using a JFET as the VCR and piecewise linear amplifier to provide gate voltage is shown in Fig. 12. The voltage references in Fig. 9 have

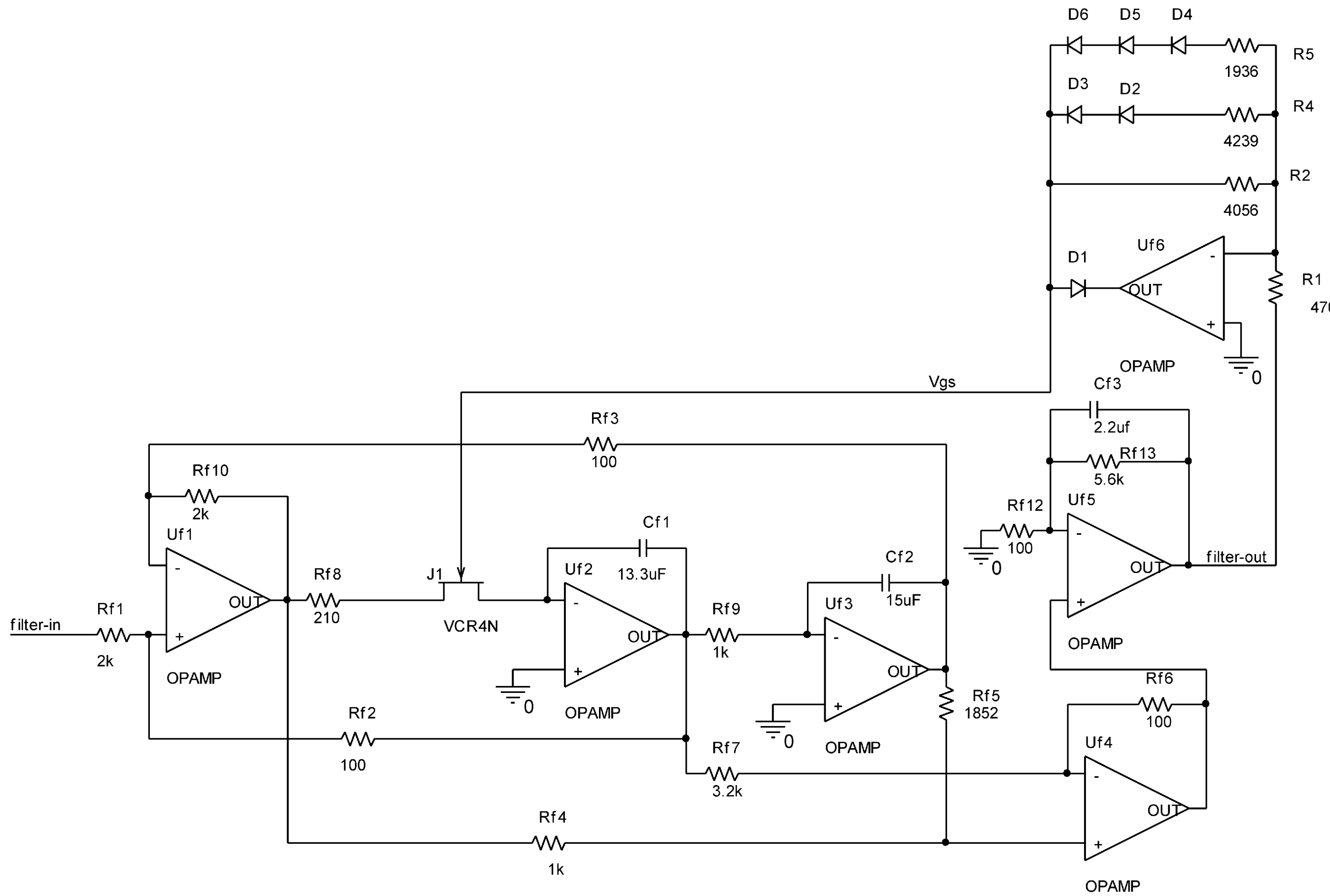

Fig. 12. Complete practical analog adaptive compensation filter.

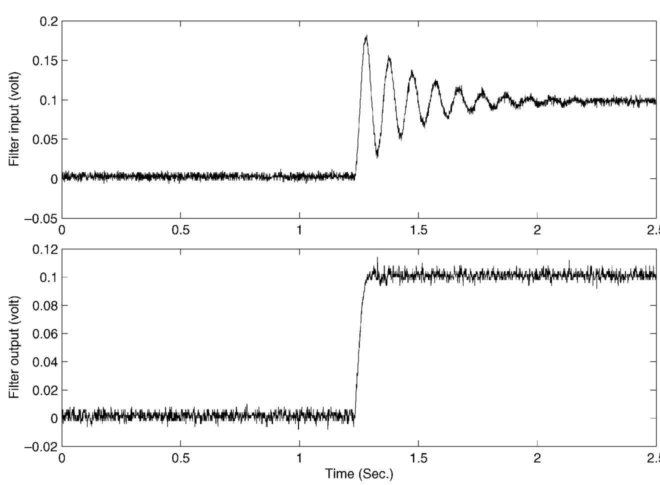

Fig. 13. Input and output of the adaptive filter with $m = 0.1$ kg.

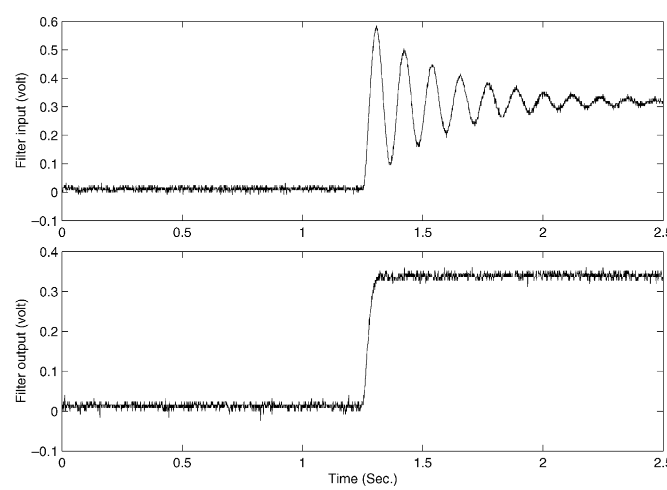

Fig. 14. Input and output of the adaptive filter with $m = 0.33$ kg.

been implemented with a cascade of single diodes. For practical testing, a square wave excitation voltage was applied to the sensor circuit, and according to the amplitude of the excitation voltage, the sensor VCR (R_8) was changed manually. Figs. 13 and 14 show two sample inputs and outputs of the adaptive filter. The input amplitudes are 100 and 330 mV, which are equivalent to $m = 0.1$ and 0.33 kg. Clearly, the practical results show that the analog adaptive biquad filter (Fig. 12) can be used to correct the response of the load cell. It should be noted that there is good correlation between simulation and experimental results (Figs. 6 and 13). To indicate the effectiveness of using an adaptive filter, a fixed filter was also used for compensation. When the excitation voltage is 330 mV, the gate of JFET is connected to ground (i.e., only fixed $R_{\mathrm{DS(on)}}$ is in the filter circuit) and the input and output of the filter are depicted in Fig. 15, which shows that the fixed filter is unable to perform the response correction. The accuracy of the proposed analog adaptive compensation filter depends on the accuracy of the resistors and capacitors (9), and in

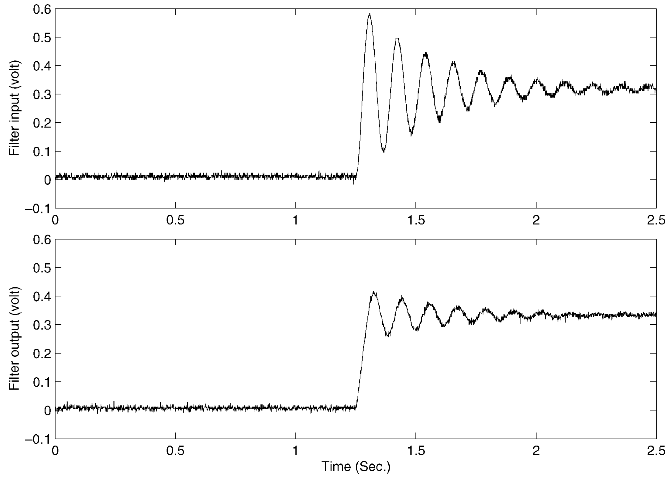

Fig. 15. Input and output of the nonadaptive filter with $m = 0.33$ kg.

practice, careful selection is needed to achieve good correction performance.

VII. Concluding Remarks

This paper has shown that it is possible to perform effective response compensation of dynamic sensors using analog adaptive filter techniques. This has been demonstrated with a reference to a load cell sensor. It has been shown that the state-variable biquadratic filter provides an accurate and flexible sensor and adaptive compensation filter models. Simulation and experimental results, showing the viability of the proposed technique, were presented. The practical prototype consisted of a novel piecewise linearization technique for floating voltage-controlled resistor. We are currently developing circuit techniques that allow the integrated fabrication of the analog adaptive filter.

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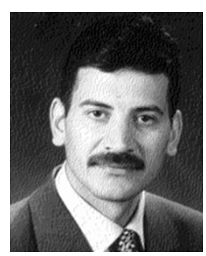

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