Improving Accuracy and Linearity of Low-Cost Flow Meters

M. Dias Pereira[1,2], O. Postolache[1,2], P. Silva Girão[2]

[1]*ESTSetúbal-LabIM/IPS, Rua do Vale de Chaves, Estefanilha, 2910-761 Setúbal, Portugal*
Email: joseper@est.ips.pt,
[2]*Instituto de Telecomunicações/DEEC, Av. Rovisco Pais, 1049-001, Lisboa, Portugal*
Emails: poctav@alfa.ist.utl.pt, psgirao@ist.utl.pt

Abstract - Two of the most important static transducer characteristics, in any measurement system, are linearity and accuracy. In the present paper, the calibration and linearization of a low-cost orifice flow meter is considered. Accuracy improvement is based on calibration coefficients that are evaluated with the measurements obtained from a reference flow meter. Linearity improvement is based on a non-linear A/D conversion whose transfer characteristic is the inverse of the transducer's characteristic. The linearity can be strongly improved using a non-linear A/D conversion performed by a low-cost ADC, using only single-bit input and output ports. The proposed method, already described by the authors in previous papers, is applied to an orifice-based flow meter, simplifying the typical conditioning circuit that requires an additional square root extractor circuit. The paper includes simulation and experimental results that validate theoretical expectations.

I. Introduction

Flow measurements are crucial whenever the performance of industrial processes is at stake. The effectiveness and performance of an industrial process is strongly affected by accurate flow meter measurements. Flow calibration is essential to obtain an accurate measurement of the volume or mass that crosses a transversal section of an industrial pipe. The accuracies of flow meters are generally worst that the accuracies provided by measuring devices of other quantities like temperature, pressure, mass or electrical voltages or currents. The uncertainties of gas flow or liquid flow measuring devices range between 200-2000 p.p.m. and 800-1000 p.p.m., respectively [1].

In the present paper the focus in on the linearization of an orifice based flow meter. Generally, flow meter calibration requires expensive flow calibration devices that implement a prover weight tank or PVTt (**P**ressure-**V**olume-**T**emperature-**t**ime) technique [2]. The calibration procedure used in our work is based on a set of two reference flow meters, one of the electromagnetic type [3] and the other of the Coriolis type [4], installed in series with the orifice based flow meter. In this flow meter, implemented in our laboratory facilities, the fluid is measured through the pressure drop across an obstruction (orifice plate) that narrows the pipe and forces the flowing fluid to constrict. Figure 1 depicts an orifice sensing element and the pressure drop between upstream (HP) and downstream (LP) sides.
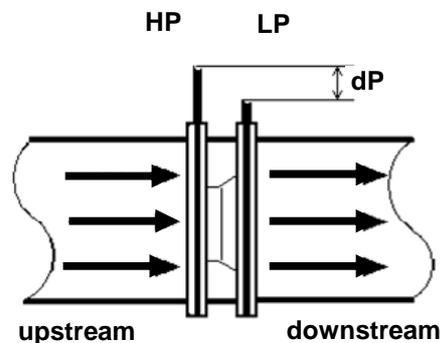

Figure 1. Pressure drop (dP) between upstream (HP) and downstream (LP) sides of an orifice sensing element

The orifice sensing elements are simple, cheap and can be delivered for almost any application in any material. Assuming stationary conditions, the flow is measured by the pressure drop whose value is obtained from Bernoulli's equation [5, 6],

$$\left(\frac{v^2}{2}+\frac{p}{\rho}+g\cdot h\right)=0 \tag{1}$$

where v represents the flow speed, p the fluid's static pressure, ρ the fluid's density, g the gravity's acceleration and h the fluid's height above a reference plane. The quadratic relationship between the pressure drop and flow speed is highly non linear and requires the usage of square root extractors to linearize flow measurements.

II. System description

Figure 2 represents the piping and instrumentation (P&I) diagram we used for flow meters' experimental tests. It includes the following set of equipments: an electro-pump (CPT45 Pentax) controlled by a three phase compact general purpose inverter (VS mini J7 Omron), a tank (14 gallon capacity), a manual valve, a pressure indicator (PI), an electromagnetic flow meter (Rosemount 8712C), a Coriolis flow meter (RFT9712) and an orifice flow meter under test (*in-house made*).
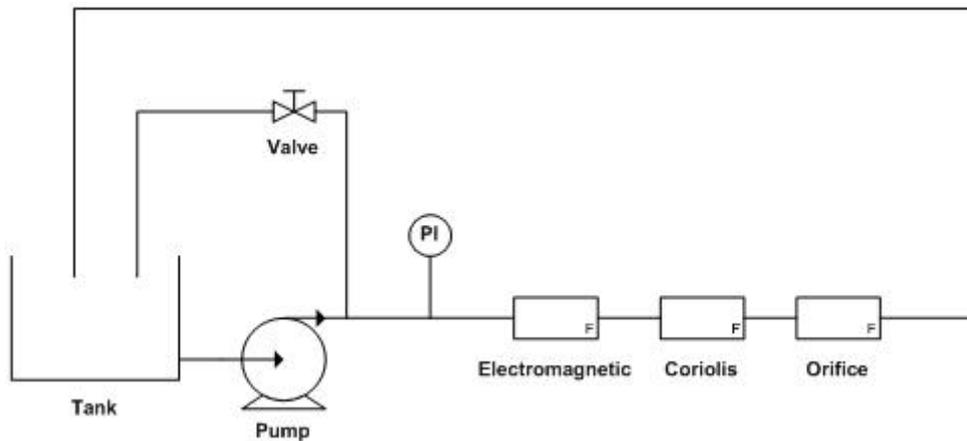

Figure 2. P&I diagram used for flow meters' experimental tests

Figure 3 represents the basic hardware of each flow transmitter (F1, F2 and F3) measuring channel, which includes the current output of each flow meter (4-20 mA), the current to voltage converter (RCV420 integrated circuit), the conditioning circuit (CC) and the PWM-ADC. Obviously, the difference circuit, implemented with an operational amplifier (OA), can be replaced by an instrumentation amplifier (e.g. AD620) but the proposed solution is cheaper and works properly as long as the output current delivered by RCV420 integrated circuit is lower than 49 mA.
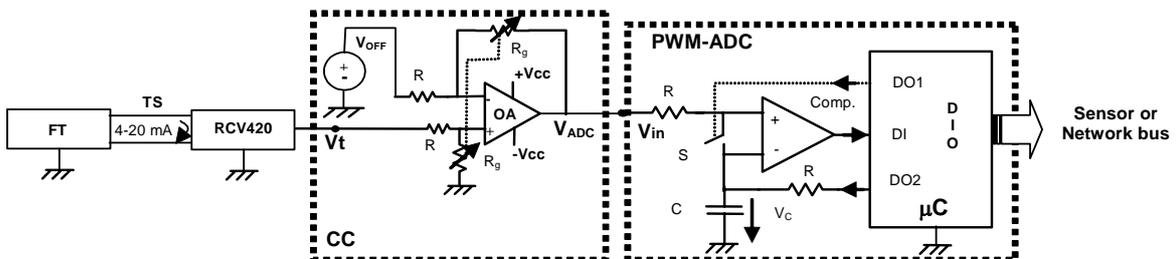

Figure 3. Basic hardware of each flow meter measuring channel: flow transmitter (FT), transmission support (TS), current to voltage converter (RC420), operational amplifier (OA); analog-to-digital converter based on pulse width modulation (PWM-ADC), comparator (Comp.), digital input (DI), digital output (DO), digital input-output ports (DIO), analog switch (S), microcontroller (μC)

The output voltage of the conditioning circuit is given by,

$$V_{ADC} = \frac{R_g}{R}\left(V_T - V_{OFF}\right) \tag{2}$$

In this way, it is possible, after calibration, to adjust offset and gain errors. Based on the PWM signal generated by the microcontroller [7-8], the number of high and low pulses is evaluated to obtain a linearized digital representation of the input signal amplitude (V_{ADC}).

Assuming ideal components, it can be demonstrated (Appendix) that the relation between the number of PWM (high and low pulses) and the input voltage amplitude (V_{in}) is given by:

$$V_{ADC} = \frac{m}{m + (2^n - m)k_{HL}^{-1}}V_H + \frac{2^n - m}{mk_{HL} + (2^n - m)}V_L \tag{3}$$

where V_H and V_L represent the amplitude of high and low pulses of the PWM signal, m is the number of high pulses, n is the converter's number of bits and k_{HL} is a coefficient that can be used for non-linearity correction and whose value is defined by,

$$k_{HL} = (1 - e^{-T_H/\tau})\big/(1 - e^{-T_L/\tau}) \tag{4}$$

being $\tau = RC$ the time-constant of the low-pass filter. The coefficients T_H and T_L represent the time duration of high and low pulses of the PWM signal, respectively. Choosing an adequate value for k_{HL} it is possible to linearize quadratic profiles of any measurement channel's characteristic [9].

III. Simulation Results

MATLAB routines were developed to evaluate the best value of the time constant (τ) of the low pass filter included in the PWM-ADC. A correct dimensioning of this parameter is essential in terms of A/D accuracy and resolution. A very low value of τ generates an excessive quantization error. On the other hand, a very high value of τ lowers the quantization error amplitude but makes the ADC very sensitive to noise. Thus, it is crucial to obtain the appropriate τ value in order to improve ADC performance. Figure 4 represents the amplitude variation of the average voltage modulus, across capacitor's terminals, as a function of the pulse width (T_P), during a conversion cycle (2^n pulses generated by the microcontroller). Input voltage range is normalized relatively to full-scale amplitude and the amplitude of the voltage variation (vertical axis) is normalized to ADC step amplitude (Q).

The simulation parameters were: $V_H = -V_L = 0.5$, n=5 bits, and the ratio between pulse width (T_P) and RC time constant (τ) assumes three different values $\left(1/2^{n-2}, 1/2^n, 1/2^{n+2}\right)$.
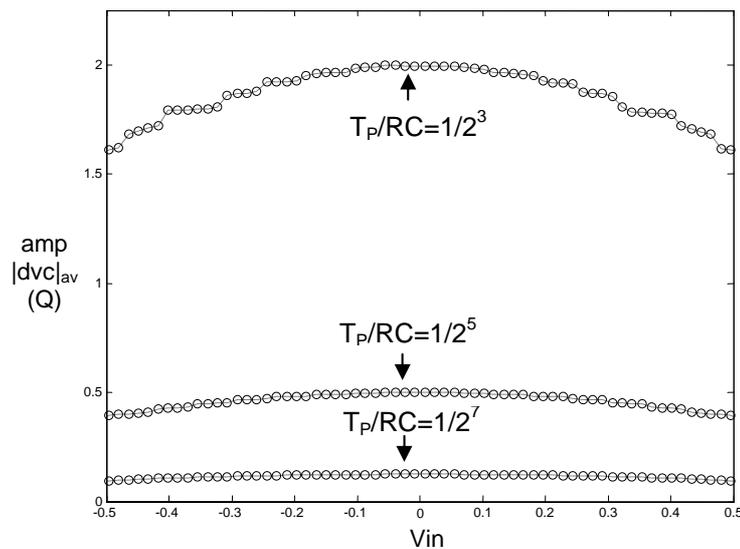

Figure 4. Modulus of the amplitude voltage across capacitor's terminal for three different values of the ratio between T_P and τ ($1/2^{n-2}$, $1/2^n$, $1/2^{n+2}$)

As expected [10], the amplitude variation of the voltage modulus between capacitor terminals is maximum in the mean value of the conversion range (Vin=0), being that amplitude higher when the coefficient T_P/RC decreases.

IV. Experimental Results

Using the A/D converter depicted in Figure 3, and adjusting a suitable value for the time duration ratio (k_{HL}), it is possible to linearize the characteristic of the flow measurement system without any need of a square root extractor, which is a conditioning circuit that introduces complexity and inaccuracy to the measuring channel.

Figures 5(a) represents the graphical representation of the output voltage (Vt) versus flow intensity, and Figure 5(b) represents the linearized characteristic of the measuring channel using the proposed PWM-ADC and associated conditioning circuit (V_{OFF}=1 V; R_g/R=1.25; k_{HL}=1.42). In this example, the linearization error is reduced from almost 20% to about 1 % without additional circuitry (root square extractor) or complex linearization algorithms requirements.
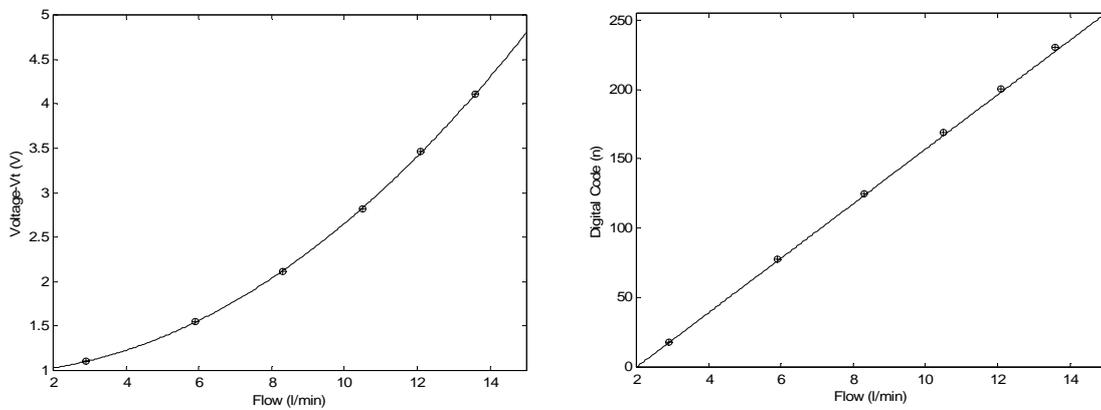

Figure 5. Graphical representations: (a) experimental (o) and interpolated (-) measurements (E_{LIN}=21,2%) and (b) linearization error after ADC-PWM A/D conversion (E_{LIN}=0.96% for p>5 p.s.i. and k_{HL} =1,42)

V. Conclusion

In the present paper a modified version of the traditional A/D PWM is presented. Linearity improvements can be reached by the converter whose transfer characteristic is the inverse of the transducer's characteristic. The conversion parameter K_{HL}, related with the ratio between high and low pulse durations, is used to adjust the nonlinear characteristic of the A/D converter. The proposed method is applied to an orifice-based flow meter, simplifying the conditioning circuit that requires an additional square root extractor circuit module or complex linearization algorithms. The paper includes simulation and experimental results that validate theoretical expectations.

Appendix

Considering the capacitor voltage variation in a RC series circuit we can write the following relationship:

$$V_C(t) = V_F + (V_{C0} - V_F) \cdot e^{-\frac{t}{\tau}} \tag{A1}$$

where V_F represents the charging voltage and V_{C0} the initial capacitor's voltage (t=0). For a time constant τ equal to RC the capacitor's voltage variation during a time period equal to (Δt) is given by:

$$\Delta V_C(\Delta t) = (V_F - V_{in}) \cdot (1 - e^{-\frac{\Delta t}{\tau}}) \tag{A2}$$

where Vin represents the initial capacitor's voltage. This voltage is imposed by the analogue switch S (Fig.3) before starting any A/D conversion cycle.

The negative feedback loop that includes the comparator, the microcontroller and the low-pass filter, imposes that the average capacitor's voltage, during each conversion cycle, is equal to V_{in}. If during the conversion cycle of the input voltage (V_{in}) there are m positive pulses, with duration T_H, and p negative

pulses, with duration T_L, the final capacitor's voltage, obtained in the end of the conversion cycle (T_C) is given by: The incremental voltage variation during each conversion pulse is given by:

$$V_C(T_C) = V_{C0} + m \cdot \Delta V_{CH} + p \cdot \Delta V_{CL} \tag{A3}$$

where $\{\Delta V_{CH}, \Delta V_{LH}\}$ represents the increment and decrement of capacitor's voltage for each positive and negative pulse, respectively.

From (A2) it is possible to obtain the average values of voltages $\{\Delta V_{CH}, \Delta V_{LH}\}$ that are given by:

$$\begin{cases} (\Delta V_{CH})_{av} = (V_H - V_{in}) \cdot \left(1 - e^{-T_H/\tau}\right) \\ (\Delta V_{CL})_{av} = (V_L - V_{in}) \cdot \left(1 - e^{-T_L/\tau}\right) \end{cases} \tag{A4}$$

In the conversion's end the average value of voltage $V_C(T_C)$ is equal to V_{in} with an error lower that $Q/2$ where Q represents the ADC's quantization step.

From (A3) and (A4) it is possible to obtain the digital representation of the input voltage:

$$V_{ADC} = \frac{mV_H \cdot (1 - e^{-T_H/\tau}) + pV_L \cdot (1 - e^{-T_L/\tau})}{m \cdot (1 - e^{-T_H/\tau}) + p \cdot (1 - e^{-T_L/\tau})} \tag{A5}$$

where V_{ADC} represents the digital voltage value associated with the input voltage V_{in}.

Finally, if we define the coefficient (k) as:

$$k_{HL} = \frac{(1 - e^{-T_H/\tau})}{(1 - e^{-T_L/\tau})} \tag{A6}$$

Expression (A5) can be written as:

$$V_{in} = \frac{m}{m + (2^n - m)k_{HL}^{-1}} \cdot V_H + \frac{(2^n - m)}{mk_{HL} + (2^n - m)} \cdot V_L \tag{A7}$$

Where n represents the converter's number of bits, being $m + p = 2^n$.

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