Calibration-less direct capacitor-to-microcontroller interface

O. Lopez-Lapeña, E. Serrano-Finetti, and O. Casas

A new technique to measure a capacitor or a capacitive sensor by means of a direct sensor-to-microcontroller interface circuit that does not need a calibration capacitor is proposed. Basically, the measurement process consists of three consecutive steps of charge, discharge and charge of the capacitor under test. A non-linear equation is obtained and solved that is dependent only on known circuit parameters. Experimental results show that it is possible to measure a wide range of capacitor values with a maximum deviation of 2 % from the reference value, and that temperature changes from 18 °C to 70 °C yield relative errors below 0.1 %. For the lowest measured capacitor range (33 pF to 4.7 nF) the uncertainty holds below 1 pF which enables measurement of commercially available capacitive sensors. The main advantage of the proposed technique is cost and space reduction of the final design.

Introduction: Capacitive sensors are used in a growing number of applications. In order to reduce the cost, complexity, space and power consumption of its interface circuits, direct capacitive sensor-to-microcontroller interfaces have been proposed. There, the sensor is connected directly to the microcontroller (µC) without using either a signal conditioner or an analogue-to-digital converter. These circuits are an evolution of the conditioning circuits found in digital capacitance meters based on measuring the charging/discharging time of an RC circuit that were proposed before µCs became available [1][2] and that needed several digital and analogue discrete components to measure the desired time interval.

Fig. 1*a* shows the simplest direct capacitor-to-microcontroller interface [3]. In this circuit, the unknown capacitor C_m is charged and discharged consecutively by changing the settings of two digital ports, P_A and P_B. The first stage ensures that C_m is fully charged by setting P_A and P_B as outputs. P_A outputs a high level (V_{OH}, "1") while P_B outputs in low level (V_{OL}, "0".) Afterwards, C_m is discharged towards V_{OL} through R by setting P_A as an input hence behaving as a Schmitt trigger in high impedance state (HZ) while P_B remains the same. Simultaneously, the µC embedded timer starts counting until V_{PA} reaches V_{TL}, the port's lower threshold voltage. At this point, P_A is triggered and the timer stops yielding $T_m \approx N_m T_{CLK}$ where T_{CLK} is the clock period and N_m is the timer register count. Fig. 1*b* shows the waveform at P_A that follows the voltage drop across C_m. Then, C_m can be estimated by

$$C_m = \frac{T_m}{R\ln\left[\left(V_{OH}-V_{OL}\right)/\left(V_{TL}-V_{OL}\right)\right]}\,. \tag{1}$$

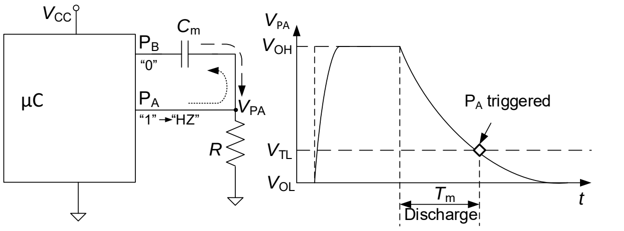

Fig. 1 *Simplest direct capacitor-to-microcontroller interface.*
a Circuit detail
b Voltage across C_m during the charge and discharge stages

One major drawback of using (1) to estimate C_m is that V_{TL} is not accurately known because of its usually large tolerance. A common solution to this is to include one or more known calibration capacitors C_{ref} connected to other µC digital ports [3][4], as shown in Fig. 2*a*, or to use a more complex circuit as in [5]. With one C_{ref}, a sequence of two charge/discharge cycles allows to measure two time lengths, T_{ref} for C_{ref} and T_m for C_m. In this way, C_m can now be estimated by

$$C_m = \frac{T_m}{T_{ref}} C_{ref} \tag{2}$$

which is independent of V_{TL}, V_{OH} and V_{OL}. However, it is now essential to select a low-tolerance and low-drift capacitor for C_{ref}, being the consequence an increased cost for the final design. In this work we present a new low-cost direct capacitor-to-microcontroller interface that no further uses any C_{ref} but is still able to accurately measure a capacitor.

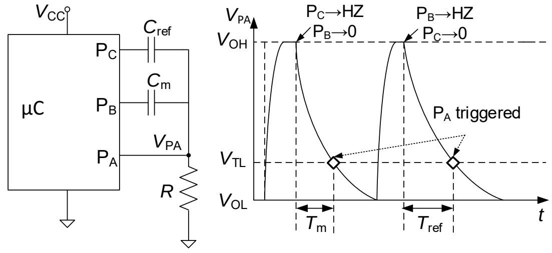

Fig. 2 *Direct interface using a reference (calibration) capacitor*
a Circuit detail
b Waveform at P_A, showing two time measurements: T_m for C_m, the sensor, and T_{ref} for C_{ref}, the calibration capacitor

Proposed measurement technique: Instead of using a reference capacitor to achieve an equation that eliminates the dependence on V_{TL}, V_{OH} and V_{OL}, we propose to use another relationship resulting from the measurement of the time length (T_{CH}) needed to charge C_m from V_{OL} to V_{TL} (Fig. 3*b*.) Unfortunately, all input ports use different trigger thresholds for decreasing signals (V_{TL}) than for rising signals (V_{TH}.) Thus, the use of a simple charge cycle would introduce another unknown, V_{TH}, not solving the problem. To overcome this, we propose to estimate T_{CH} from T_1 (time length to charge from V_{OL} to V_{TH}) and T_2 (time length to charge from V_{TL} to V_{TH}) as is shown in Fig. 3*b*. By noting that the time between points 1 and 2 is the same as T_2 we have $T_{CH} = T_1 - T_2$. The start and end points of T_1 and T_2 are set by toggling P_B and also by the input port thresholds, hence they can be measured using the circuit of Fig. 3*a*. P_A is always configured as an input port (HZ) that compares V_{PA} with V_{TL} or V_{TH} and P_B is configured as an output port. In a first stage, P_B outputs a "1" and C_m, initially discharged, charges towards V_{OH} through R. Referring to Fig. 3*b*, V_{PA} eventually reaches V_{TH} (point 2) where T_1 is measured. The capacitor is further charged until it can be considered fully-charged. Following, P_B outputs a "0" (point 3) and C_m discharges towards V_{OL} until it reaches V_{TL} (point 4) where T_{DS} is measured. In a third and last stage, P_B is set back to "1" and a second charging stage begins until V_{TH} (point 5) is reached again and T_2 is measured.

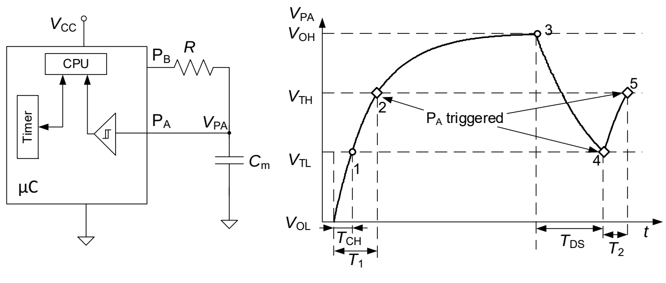

Fig. 3 *Proposed direct interface measure to measure C_m.*
a Circuit detail
b Voltage across C_m during the different charge and discharge stages

Taking into account that V_{PA} equals V_{TL} at points 1 and 4, it follows that

$$V_{TL} = V_{OH} + (V_{OL} - V_{OH})e^{\frac{-T_{CH}}{RC_m}} = V_{OL} + (V_{OH} - V_{OL})e^{\frac{-T_{DS}}{RC_m}} \tag{3}$$

which leads to an equation where C_m is the only unknown, hence overcoming the problems of the interface in Fig. 1*a*. To solve (3), we first define

$$f(t) = 0.5 - e^{\frac{-t}{RC_m}} \tag{4}$$

and rewrite (3) to

$$-f(T_{DS}) = f(T_{CH}) \tag{5}$$

To solve (5) we approximate $f(t)$ by its Taylor second order polynomial around some T_0 such that $f(T_0) = 0$. Hence (5) becomes

$$A(RC_m)^2 + B(RC_m) + C \approx 0 \tag{6}$$

where $A = -[2+\ln(2)]\cdot\ln(2)$, $B = [1+\ln(2)]\cdot(T_{CH}+T_{DS})$, $C = -0.5\cdot[(T_{CH})^2+(T_{DS})^2]$. Finally, the capacitance is estimated as

$$C_m \approx -\frac{1}{R}\frac{B+\sqrt{B^2-4AC}}{2A} \tag{7}$$

With typical V_{TL} values, T_{CH} and T_{DS} are not so far from T_0 as to introduce significant errors when using this approximation. In fact, using the proposed Taylor polynomial yields a theoretical residual error of less than 0.18 % when solving for C_m. The main advantage of this approach compared to that of Fig. 1 is that (6) is now independent of V_{TL}, V_{OH} and V_{OL} as well as its associated tolerances and thermal drifts and does not need extra ports nor extra capacitors. On the other hand, the dynamic range of this technique will now depend on the timer specs i.e. register length, clock frequency and clock stability.

Experimental results: The µC chosen to test the proposed method was the MSP430F1471 (Texas Instruments Inc.) A low-power, low-noise 3.3 V LDO (NCP702, ON Semiconductor) was selected to supply the µC that helps in reducing the noise of the input port trigger levels. We chose two general-purpose digital ports to implement the method described in Fig. 3, P_A = P1.2 and is P_B = P2.3, and whose measured V_{TL} and V_{TH} were 1.2 V and 1.8 V respectively. We used the µC's 16-bit Timer A with its clock sourced by an 8 MHz quartz crystal (30 ppm/°C.) A resistor R of 200 kΩ (1 % tolerance, 20 ppm/°C) was used for the smaller capacitor range (33 pF up to 4.7 nF) and was changed to 20 kΩ for values up to 220 nF to avoid timer overload. For each C_m, 200 consecutive measurements were performed and the mean value and standard deviation were computed. Fig. 4*a* shows a comparative between the mean value of C_m estimated with the proposed technique and its reference value, C_r, measured with an LCR meter (U1733C, Agilent). As seen in Fig. 4*b*, choosing an optimal R value according to the capacitor range (200 kΩ for values under 4.7 nF and 20 kΩ for the rest) yields a difference below 2 % between C_m and C_r.

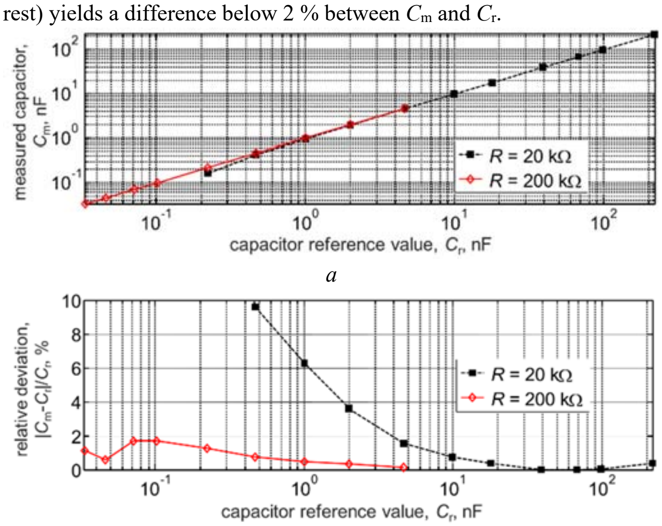

Fig. 4 *Performance of the proposed technique.*
a Linearity between C_m and the reference value C_r
b Relative deviation between the measurement and the reference

The maximum standard deviation obtained was below 1 pF for R = 200 kΩ and below 15 pF for R = 20 kΩ. To assess the robustness of the proposed technique against the thermal drift of V_{TL}, we have tested a 68 nF capacitor for a circuit temperature ranging from 18 °C to 70 °C. As can be seen in Fig. 5*b*, the maximum measured variation in V_{TL} was around 4 % resulting also in an approximately 4 % error in C_m when using (1). The proposed technique showed an error of less than 0.1 % in all the tested temperature range as shown in Fig. 5*a*. Because using (1) yields a measurement deviation in C_m that is proportional to any deviation in V_{TL}, a larger tolerance of the latter would increase the measurement error. For example, the µC manufacturer specs state that V_{TL} ranges from 0.9 V to 1.3 V (3 V supply voltage) hence the maximum expected error would be 25 %. However, other µCs might have different thermal drift and tolerance specs therefore these results shall differ accordingly.

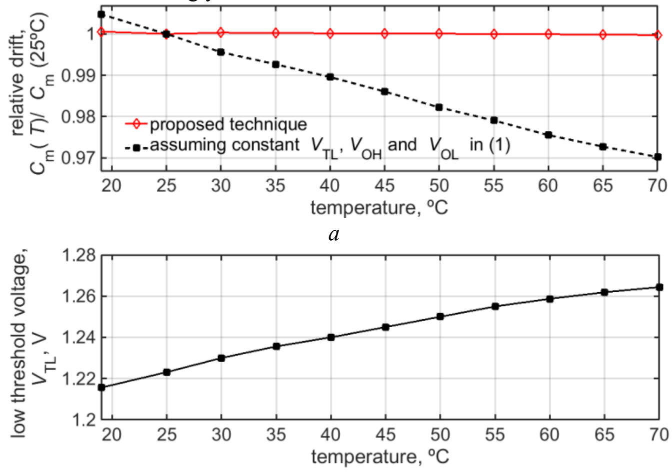

Fig. 5 *Performance of the proposed technique when air temperature changes between 18 °C and 70 °C.*
a Drift in C_m induced by the temperature change when using (1) (discontinuous trace) and with the new technique (continuous trace)
b Measured changes in V_{TL} for different temperatures

Conclusions: A new direct capacitor-to-microcontroller interface circuit has been proposed that does not use a calibration capacitor and reduces the number of ports used by the µC. Instead of using the discharge time of a reference capacitor, both the charge and discharge times of the measured capacitor are used to improve accuracy. Experimental results show that it is feasible to measure capacitors (or capacitive sensors) down to the tens of picofarad range with errors below 2 % and negligible thermal drifts.

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