An Auto-balancing Capacitance-to-Pulse-Width Converter for Capacitive Sensors

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Abstract—A novel auto-balancing capacitance-to-pulse-width converter (CPC) that uses sinusoidal excitation, and operates in a closed-loop configuration, is presented in this paper. Unlike most of the existing CPCs, the proposed interface circuit is compatible with both single-element and differential capacitive sensors. In addition, it provides a pulse-width modulated (PWM) signal which can easily be digitized using a counter. From this PWM signal, a ratio output is derived when a single-element sensor is interfaced, and a ratiometric output is obtained for a differential sensor. The final digital output is independent of the nominal capacitance of the sensor and has a linear characteristic irrespective of the sensor characteristic being linear or inverse. The CPC is designed such that the PWM output depends on the change in the sensor capacitance alone. It is insensitive to parasitic capacitance and has very low sensitivity to the non-idealities of the components and ICs used. The effects due to some of the non-idealities are automatically corrected by the negative feedback based auto-balancing employed. The effect of component mismatch is significantly reduced by a one-time correction mechanism. These benefits are achieved without the use of any complex or expensive analog building blocks. The prototype exhibits a maximum non-linearity error of less than 0.7%, a resolution of 13.02 effective number of bits (ENOB), a signal-to-noise ratio (SNR) of 80.12 dB, and a rise time of 5 ms. Thus, the proposed simple, yet effective, low-power, low cost, auto-balancing CPC can be used to interface a wide range of existing and new capacitive sensors to digital systems.

Index Terms—Capacitive sensor, differential capacitive sensor, pulse-width converter, duty cycle, sine wave, auto-balancing, digitizer, linearization, PWM, linear output, parasitic capacitance.

I. Introduction

SENSORS with efficient interfacing circuits that are simple and reliable are in high demand in the world today. They play a key role in a variety of domains ranging from consumer electronics, agriculture, healthcare, to automation, industry, and scientific research [1]. The demand is a constantly growing one, driven by the popular technologies of data science and artificial intelligence, which are heavily dependent upon enormous amounts of useful sensor data. Among various types of sensors that are available, capacitive sensors are used in numerous applications, and have a noticeable share in the sensor market, due to their advantages such as ease of fabrication, high sensitivity, high resolution, low power dissipation, broad bandwidth, and low cost [1].

Depending on the application and requirement, measurement engineers opt for (a) single-element capacitive sensors, which are designed using two electrodes, (b) differential capacitive sensors (DCS), designed using three electrodes, or (c) sensors that use more electrodes, which can be considered as a combination of multiple single-element sensors [1]–[5]. The differential capacitive sensor has a wide range of applications in automation and control. Its output can be processed in a way that the effect due to most of the common parameters that are

of no interest can be eliminated or reduced [1]. It is used not only to sense mechanical parameters such as position [2], displacement [3], acceleration [4], force [5], and shear stress [6], but also to realize proximity sensing [7], and non-destructive estimation [8]. These are also used in areas like gesture recognition [9], bacterial growth monitoring [10], robotic and upper limb prosthesis [6], pressure sensing in harsh environments [11], and accurate detection of a sound source [12]. The DCS consists of two sensing capacitances, C_1 and C_2 , with a common electrode, as shown in Fig. 1. The input-output characteristic of C_1 and C_2 can be linear, as in (1), or inverse, as shown in (2), where k is a transformation constant and C_0 is the nominal capacitance of the sensor [13]. In the case of the ubiquitous parallel plate arrangement of the DCS, (1) holds true if the change in capacitance is caused by a change in the area of overlap between the plates, and (2) if the change is caused by the change in the distance between the plates. In the case of single-element capacitive sensors, C_1 in (1) can be used to represent the sensor with a linear characteristic, and similarly, C_2 in (2) can be used to represent the one following inverse characteristic.

$$C_1 = C_0(1 \pm kx) \text{ and } C_2 = C_0(1 \mp kx) \quad (1)$$

Fig. 1. Simplified circuit diagram of the proposed capacitance-to-pulse-width converter.

$$C_1 = \frac{C_0}{(1 \mp kx)} \text{ and } C_2 = \frac{C_0}{(1 \pm kx)} \quad (2)$$

The conversion of the sensor capacitance to a form suitable for analog or digital signal processing can be done in many ways. It can be converted to an equivalent voltage [14], [15] or current [16]. In such a case, different techniques for compensating the effects of parasitic capacitances have been proposed, for instance: the use of a negative capacitance placed in parallel with the parasitic one [17], even including an automatic compensation circuit [18]. Capacitance-to-Digital converters (CDCs) based on voltage conversion, followed by an analog-to-digital converter, are generally complex in design, prone to noise and consume more power, compared to other conversion paradigms [19]. Interfacing schemes based on quasi-digital signals, where the information is given by either the frequency or pulse duration, have emerged as possible alternatives owing to their simplicity, the possibility of direct transmission over comparatively noisy or non-linear channels and the facility to digitize these signals using counters [19]-[22]. Another approach is to interface the capacitive sensor directly to a digital system such as a microcontroller, but the accuracy that can be achieved is limited even after employing special calibration [23]. Also, in this case, it gets relatively cumbersome if a sensor needs to be excited at a particular frequency. Capacitance-to-frequency converters (CFCs) are comparatively simpler and less power demanding circuits. However, their practical realizations are often fraught with high sensitivity to temperature and process parameters when used in an open-loop configuration [19]. Most of the CFCs require a voltage-to-frequency converter (VFC), which increases the cost and complexity of the interfacing circuit. Similar to the CFCs, the output of the capacitance-to-pulse-width converters (CPCs) can be easily interfaced to a microcontroller/digital system, but without the additional cost and complexity of analog blocks like the VFC. This is because the frequency of the PWM signal can be synchronized to that of the inbuilt clock of the

microcontroller [22]. Also, the PWM signals, being quasi-digital, possess all the advantages of a frequency output [19]. In spite of these attractive features, CPCs have not been widely explored so far. Although there are some CPCs reported for capacitive sensors excited from DC sources, there are no attempts of CPCs reported for differential capacitive sensors excited using a sinusoidal source. A basic idea of such a scheme for a single element capacitive sensor is presented in [24].

An important factor to consider in the design of the signal conditioning circuitry, for any capacitive sensor, is the type of excitation signal used. The sinusoidal AC signal is one such excitation signal. A sinusoidal signal is one of the best choices to evaluate a given system. Also, in the case of some of the capacitive sensors which use the change in the dielectric property with frequency [25], [26] the use of a sinusoidal excitation signal, within a specified frequency range, is unavoidable. This AC excitation signal has been used to successfully realize many sensing applications such as flow rate measurement [27], proximity sensing [28] and humidity measurement [29]. The use of this AC signal also places lesser demand on the opamps, used in the circuit, for required bandwidth and slew rate, in comparison with a square-wave signal. The use of such an excitation signal also enables the easy implementation of a synchronous demodulator to achieve interference insensitivity and high signal-to-noise ratio (SNR) [30].

In this paper, a novel auto-balancing CPC, with a sinusoidal excitation, which employs a new closed-loop pulse-width conversion technique, is presented. Its PWM output facilitates easy digitization of the interfaced sensor capacitance. This new scheme can be used to interface single-element as well as differential capacitive sensors. The sensors can possess either a linear characteristic as in (1), or an inverse one as in (2). The final output from the proposed CPC is linear. Thus, this scheme combines the use of a sinusoidal AC excitation, and provides a PWM signal, and so has several advantages, as mentioned above. Additionally, the proposed scheme is insensitive to

parasitic capacitance. It also offers a solution to the degradation in performance of CDCs due to offset capacitance. Generally, the capacitive transducers used in practice produce very small changes due to the measurand. At the same time, they exhibit a relatively large offset capacitance. If not canceled, the offset also gets converted, and thus degrades the resolution of the capacitance measurement. The pulse-width converter presented in this paper processes the output corresponding to the change in capacitance alone, not the offset capacitance. In addition, the ratio output provided in the case of the single element sensor, and the ratiometric output obtained for the differential capacitive sensor are independent of the nominal capacitance C_0 . These features are achieved without complex signal processing, which otherwise adds to the cost and power consumed.

II. CAPACITANCE-TO-PULSE-WIDTH CONVERTER

The circuit diagram of the proposed CPC is shown in Fig. 1. Here, the variable capacitors C_1 and C_2 represent the differential capacitive sensor (DCS). The circuit is also suitable for single-element capacitive sensors, which is discussed at the end of this section. The circuit uses a sinusoidal excitation signal $v_s = \sqrt{2}V_s \sin \omega t$, where V_s is the root mean square (rms) voltage, and $\omega = 2\pi f_s$, where f_s is the frequency of the signal. Its time period $T_s = 1/f_s$. The maximum frequency at which C_1 and C_2 change is assumed to be much less compared to f_s . The opamps OA₁ and OA₂, feedback capacitors C_{F1} and C_{F2} , single-pole, double throw (SPDT) switches S₁, S₂, S₃ and S₄, and resistors R_1 and R_2 , form two branches of the proposed circuit, beginning at C_1 and at C_2 , respectively, and converging at the inverting terminal of opamp OA₃. OA₁, C_{F1} , and C_1 constitute a charge amplifier. So do OA₂, C_{F2} , and C_2 . The outputs v_{oi1} and v_{oi2} of these two charge amplifiers are expressed as, $v_{oi1} = -\frac{C_1}{C_{F1}}\sqrt{2}V_s \sin \omega t$ and $v_{oi2} = -\frac{C_2}{C_{F2}}\sqrt{2}V_s \sin \omega t$.

As shown in Fig. 1, v_{oi1} and v_{oi2} are processed using S₁ and S₂, respectively, and two unidirectional signals of opposite polarity, v_{o1} and v_{o2} , respectively, are obtained. Both S₁ and S₂ are controlled by the output v_{oc1} of the comparator OC₁. The inputs of OC₁ are v_s at the non-inverting terminal, and ground potential at the inverting terminal. Thus, v_{oc1} is a square wave of frequency f_s . When v_{oc1} is low, S₁ and S₂ are at position-0 leading to $v_{o1} = v_{oi1}$ and $v_{o2} = -v_{oi2}$. Alternatively, when v_{oc1} is high, both S₁ and S₂ are at position-1, and so v_{o1} and v_{o2} are $-v_{oi1}$ and v_{oi2} , respectively. A pictorial representation of these signals is included in Fig. 1. These signals are fed into S₃ and S₄, respectively. Both S₃ and S₄ are controlled by v_{oc2} , a pulse signal generated by the comparator OC₂. The signal at the output of S₃, $v_{o3} = v_{o1}$ when v_{oc2} is high, and zero (ground) otherwise. The opposite occurs at the output of S₄, with $v_{o4} = 0$ when v_{oc2} is high, and $v_{o4} = v_{o2}$ otherwise. v_{o3} and v_{o4} are input via R_1 and R_2 , respectively, into the inverting terminal of OA₃. R_1 , R_2 , and OA₃, along with the feedback capacitor C_{F3} constitute an inverting summing integrator. Its output, v_{oi3} , is given to the non-inverting terminal of OC₂. A reference triangular signal v_{tri} , with peak value V_p and time period $T = 1/f_{tri}$, is given to its inverting terminal.

The pulse-width of v_{oc2} depends on v_{oi3} and v_{tri} . v_{oc2} is high while $v_{tri} < v_{oi3}$ and low otherwise. In the proffered circuit f_{tri}

is set such that $f_{tri} \gg f_s$ (explained in section II.A.). Hence, the change in v_{oi3} within a cycle of v_{tri} is very low compared to that in v_{tri} . When $v_{oi3} = 0$, the pulse-width of v_{oc2} is $T/2$, and the duty cycle $D = \frac{T_{ON}}{T}$ is 0.5. Here, T_{ON} is the duration for which v_{oc2} is high. T_{OFF} is the duration for which v_{oc2} is low and thus, $T = T_{ON} + T_{OFF}$. As v_{oi3} increases in the positive direction, T_{ON} increases, and T_{OFF} decreases. The opposite occurs (T_{ON} decreases and T_{OFF} increases) if v_{oi3} decreases from zero and goes negative. The control of S₃ and S₄ by v_{oc2} , whose pulse-width is a function of v_{oi3} , realizes a negative feedback based the auto-balancing mechanism in the proposed circuit. The operation of the proposed CPC when interfaced with (a) differential, and (b) single-element capacitive sensors, respectively, is discussed below.

A. Differential Capacitive Sensor

In the proposed circuit, $C_{F1} = C_{F2} = C_F$ and $R_1 = R_2 = R$. The output voltages v_{o1} and v_{o2} of the switches S₁ and S₂, respectively, can be expressed as $v_{o1} = \left| \frac{C_1}{C_F} \sqrt{2}V_s \sin \omega t \right|$ and $v_{o2} = -\left| \frac{C_2}{C_F} \sqrt{2}V_s \sin \omega t \right|$. As S₃ is controlled by v_{oc2} , $v_{o3} = v_{o1}$ during $T_{ON} (= DT)$, but zero for T_{OFF} . The average value, $\overline{v_{o1}}$, of current i_{C1} flowing through R_1 , as shown in Fig. 1, for one cycle of v_s is expressed as in (3). In (3), (T_s/T) is a positive integer.

$$\overline{v_{o1}} = \frac{1}{T_s} \sum_{j=0}^{(T_s/T)-1} \int_{jT}^{DT+jT} \left| \frac{C_1}{C_F} \sqrt{2}V_s \sin \omega t \right| dt \quad (3)$$

Similarly, the average value, $\overline{v_{o2}}$, of current i_{C2} flowing through R_2 , for one cycle of v_s is expressed as in (4). This is written based on $v_{o4} = v_{o2}$ during T_{OFF} , i.e. $(1-D)T$, and zero otherwise.

$$\overline{v_{o2}} = \frac{1}{T_s} \sum_{j=0}^{(T_s/T)-1} \int_{DT+jT}^{T+jT} \left| \frac{C_2}{C_F} \sqrt{2}V_s \sin \omega t \right| dt \quad (4)$$

From (3) and (4), $\overline{v_{o1}}$ and $\overline{v_{o2}}$ are directly proportional to C_1 and C_2 , respectively. Consider the case $D = 50\%$. $v_{o3} = v_{o1}$ when v_{oc2} is high, and zero otherwise. Similarly, $v_{o4} = v_{o2}$ when v_{oc2} is low, and zero otherwise. These are illustrated in Fig. 2. If the top portions of each non-zero segment of v_{o3} and v_{o4} are assumed to be flat, the segments become rectangular blocks. As

Fig. 2. Signals v_{o3} and v_{o4} when $D = 0.5$. This is illustrated for $f_{tri} = 20f_s$. The sum of the average of these signals was found to be negligible (0.006% of $\overline{v_{o4}}$). The computation was done using Scilab.

long as $T_s \gg T$, this approximation will introduce only a negligible error and the average value of the signal can be computed using these rectangular blocks. Under this condition, the average value of v_{o3} will vary proportional to D , and that of v_{o4} proportional to $(1 - D)$. Thus, the values of $\bar{i}_{C1}$ computed using (3) and $\bar{i}_{C2}$ computed using (4) are directly proportional to D and $(1 - D)$, respectively. The relationship between $\bar{i}_{C1}$ and D is shown in Fig. 3, which is computed by varying D while maintaining C_1 at a fixed value. It can be seen that the curve becomes linear once the condition $T_s \gg T$ is satisfied. The nonlinearity error becomes negligible once f_{tri} is $20f_s$ or more. The same can be concluded in case of $\bar{i}_{C2}$ and $(1 - D)$.

The integrator output voltage v_{oi3} can be expressed as in (5).

$$v_{oi3} = -\frac{1}{C_{F3}} \int_0^{T_s} (i_{C1} + i_{C2}) dt \quad (5)$$

When $C_1 = C_2$, $\bar{i}_{C1} = -\bar{i}_{C2}$ and from (5), $v_{oi3} = 0$. Then, $T_{ON} = T_{OFF}$ and $D = 50\%$. In case $C_1 < C_2$, initially $|\bar{i}_{C1}| < |\bar{i}_{C2}|$, as per (3) and (4). As $\bar{i}_{C2}$ is a negative quantity, v_{oi3} will increase in the positive direction. This causes T_{ON} to increase and T_{OFF} to decrease, increasing D . This will increase $\bar{i}_{C1}$ and decrease $\bar{i}_{C2}$. This process continues until v_{oi3} reaches the steady state, i.e., $\bar{i}_{C1} = -\bar{i}_{C2}$. The opposite happens (i.e., $|\bar{i}_{C1}| > |\bar{i}_{C2}|$ and v_{oi3} will increase in the negative direction) when $C_1 > C_2$.

In the steady state condition, i.e., when $\frac{d(v_{oi3})}{dt} = 0$, i_{C1} and i_{C2} are related as in (6).

$$\bar{i}_{C1} = -\bar{i}_{C2} \quad (6)$$

Considering $\bar{i}_{C1}$ is directly proportional to C_1 and D , and $\bar{i}_{C2}$ to C_2 and $(1 - D)$, (6) can be rewritten as in (7).

$$C_1 \times D = C_2 \times (1 - D) \quad (7)$$

Rearranging C_1 and C_2 and expressing (7) in terms of T_{ON} and T_{OFF} , gives (8).

$$\frac{C_1}{C_2} = \frac{T_{OFF}}{T_{ON}} \quad (8)$$

Further, in ratiometric form, (8) becomes

$$\frac{C_1 - C_2}{C_1 + C_2} = \frac{T_{OFF} - T_{ON}}{T_{OFF} + T_{ON}} \quad (9)$$

By substituting for C_1 and C_2 as in either (1) or (2), i.e., sensors with either linear or inverse characteristic, respectively, (9) becomes (10). As can be seen from (10), the final output of the proposed scheme is linear, irrespective of the sensor characteristic being linear or inverse.

$$\pm kx = \frac{T_{OFF} - T_{ON}}{T_{OFF} + T_{ON}} \quad (10)$$

B. Single-Element Capacitive Sensor

In addition to being compatible with differential capacitive transducers, the proffered CPC is also suitable for single-element capacitive sensors. To interface a single-element capacitive sensor, either C_1 or C_2 , shown in Fig. 1, is set as the variable capacitance of sensor, C_x , while the other acts as a fixed reference capacitance, $C_R = C_0$. The choice of either C_1 or C_2 as C_x depends upon the sensor characteristic of C_x : (a) if

Fig. 3. The average value of the signal i_{C1} , $\bar{i}_{C1}$, for different values of D , from 0 to 1. This was computed using Scilab, for $(f_{tri}/f_s) = 1, 2, 3, 4, 5, 10$, and 20 . The percentage non-linearity error (%NLE) for all the corresponding ratios of (f_{tri}/f_s) are shown in the bottom inset. The maximum of the %NLE, for the same ratios are given in the top left inset. It also gives the %NLE for $(f_{tri}/f_s) = 21$. It can be seen that the %NLE is very low as long as the ratio is high, say 20 .

C_x possesses a linear characteristic, i.e. $C_x = C_0(1 \pm kx)$, then C_1 is set as C_x and C_2 is set as C_0 , and (b) if C_x follows an inverse characteristic, i.e., $C_x = C_0/(1 \pm kx)$, then C_2 is set as C_x and $C_1 = C_0$. Substituting either of the choices (a) or (b) in (8) leads to (11).

$$\pm kx = \frac{(T_{OFF} - T_{ON})}{T_{ON}} \quad (11)$$

According to (11), when a single-element capacitive sensor C_x is interfaced to the proposed CPC, the final output is linear irrespective of its sensor characteristic.

III. ANALYSIS: EFFECT OF NON-IDEALITIES

A. Component Mismatch - One time Correction

The derivation of (10) has been done assuming $C_{F1} = C_{F2} = C_F$, and $R_1 = R_2 = R$. In a practical implementation of the proposed scheme, the match in the aforementioned capacitors and resistors will be limited by the tolerance of these components. For example, let $C_{F2} = C_{F1}(1 + \epsilon_F)$, where $\epsilon_F = \frac{\Delta C_{F1}}{C_{F1}}$, and $\Delta C_{F1} = C_{F2} - C_{F1}$. Then, in the steady state condition $|\bar{i}_{C1}| = -|\bar{i}_{C2}|$, (7) gets modified as in (12).

$$\left(\frac{1}{C_{F1}}\right) C_1 \times D = \left(\frac{1}{C_{F2}}\right) C_2 \times (1 - D) \quad (12)$$

Substituting $C_{F2} = C_{F1}(1 + \epsilon_F)$ and $D = T_{ON}/(T_{ON} + T_{OFF})$, and rearranging, gives (13), where $k_F = 1/(1 + \epsilon_F)$.

$$\frac{C_1}{C_2} = k_F \times \frac{T_{OFF}}{T_{ON}} \quad (13)$$

The impact of k_F in the final output can be compensated for by performing a simple one-time calibration. For this, C_1 and C_2 are set to their nominal value C_0 . In the presence of mismatch T_{ON} and T_{OFF} will not be equal. Then, the value of k_F can be obtained as $k_F = T_{ON}/T_{OFF}$. From then onwards, the measured $T_{OFF(m)}$ can be divided by k_F to get the gain corrected T_{OFF} . This can be used in the final ratiometric output in (10) or ratio output in (11) to ensure negligible sensitivity to capacitor mismatch. A mismatch between R_1 and R_2 will result in a similar gain error. The k_F obtained through the one-time calibration, is one that can be used for the gain correction of the signal paths (a) C_{F1} -S₃-S₁-R₁ and (b) C_{F2} -S₂-S₄-R₂ of the circuit in Fig. 1.

B. Parasitic Capacitances of the Transducer

The parasitic capacitances C_{P1} , C_{P2} and C_{P3} of the differential capacitive transducer are shown in Fig. 4. C_{P1} and C_{P2} remain inactive because in both of them one end is connected to the virtual ground while the other is grounded. C_{P3} has no impact on the output of the proposed circuit as long as it does not load the excitation source due to the current I_{P3} from v_s to ground, through C_{P3} .

Fig. 4. Sensor with parasitic capacitances in the proposed CPC.

C. Input Offset Voltage of Comparators

Let V_{OSC1} and V_{OSC2} be the input offset voltages of OC₁ and OC₂, respectively. If the polarity of V_{OSC1} is such that the output v_{OC1} is high when the input signal to OC₁ is zero, then, the duration for which v_{OC1} is high within a cycle of v_s increases, and that for which it is low decreases. This will affect the switching of S₁ and S₂. In v_{o1} , there will be small portions of negative voltage near each of the zero-crossing points of v_{oi1} . The average value of this negative voltage within a cycle T_s is $\cong \frac{V_{OSC1}^2}{8V_s^2} v_{oi1(max)}$, when v_s and v_{oi1} are approximated to triangular waveforms (considering V_{OSC1} is very low compared to the peaks of the signals v_s and v_{oi1}), instead of sinewaves. $v_{oi1(max)} = \sqrt{2}V_s \frac{C_1}{C_F} v_{oi1(max)}$ can be designed such that it is

close to $V_{SAT} - 0.5$, where V_{SAT} is the saturation voltage of OA₁. For a comparator with a significant input offset voltage of 5 mV, $V_s = 2.5/\sqrt{2}$ V, and $V_{SAT} = 8$ V, and $v_{oi1(max)} = 0.5(V_{SAT} - 0.5)$, the resulting average value is 3.75 μ V, which is negligible compared to the average value of $v_{o1} \cong 2(V_{SAT} - 0.5)/\pi = 4.7$ V at full scale. The same logic can be extended to quantify the effect of V_{OSC1} on the output of switch S₂, and arrive at the conclusion that it can be neglected.

To understand the effect of V_{OSC2} , consider first the case $C_1 = C_2$. In this condition, ideally, i.e. $V_{OSC2} = 0$, v_{oi3} should be zero and $T_{ON} = T_{OFF}$. However, due to V_{OSC2} , when $v_{oi3} = 0$, v_{OC2} will not have transitions at the erstwhile zero crossing points of v_{tri} . Instead, the transitions occur when the value of v_{tri} crosses V_{OSC2} . If the polarity of V_{OSC2} is such that v_{OC2} is high when $v_{oi3} = v_{tri} = 0$, then $T_{ON} > T_{OFF}$ and $|\overline{i_{C1}}| > |\overline{i_{C2}}|$. This will start driving v_{oi3} in the negative direction and the difference between T_{ON} and T_{OFF} will keep reducing with time. This process will stop when v_{oi3} has the same magnitude of that of V_{OSC2} but opposite polarity, ensuring $T_{ON} = T_{OFF}$. The impact of V_{OSC2} with opposite polarity can be analyzed in the similar manner. Thus, thanks to the negative feedback approach adopted in the proposed auto-balancing pulse-width converter circuit, V_{OSC2} has no effect in the final output.

D. Bias Currents and Input Offset Voltage of Opamps

Let I_{B1} , I_{B2} and I_{B3} be the bias currents at the inverting terminals of OA₁, OA₂ and OA₃, respectively. I_{B1} and I_{B2} charge C_{F1} and C_{F2} , respectively, in a continuous manner. As in the case of charge amplifiers, to avoid driving the opamp into saturation, appropriately chosen large resistors are connected in parallel to C_{F1} and C_{F2} , respectively, to provide a discharging path for each. The value of bias current is very low for commercially available opamps. Thus, the effect can be neglected, once the integrating effect is compensated using suitable parallel resistors.

OA₃ is in the negative feedback loop of the auto-balancing pulse width converter. It does not require a parallel resistor as the feedback will regulate the integrating effect due to I_{B3} . In the presence of I_{B3} , v_{oi3} will have an integrating term $\frac{1}{C_{F3}} \int I_{B3} dt$. Due to this, to reach steady state, an additional current (equal and opposite to I_{B3}) will have to be provided by $\overline{i_{C1}}$ and $\overline{i_{C2}}$. At steady state, (6) will get modified to become (14), instead of being $\overline{i_{C1}} + \overline{i_{C2}} = 0$.

$$\overline{i_{C1}} + \overline{i_{C2}} = I_{B3} \quad (14)$$

If we consider the condition $C_1 = C_2$, T_{ON} and T_{OFF} will get modified as $T_{ON} \mp \Delta T_\epsilon$ and $T_{OFF} \pm \Delta T_\epsilon$ to contribute the opposite current that amounts to I_{B3} , through C_{F3} . ΔT_ϵ is the change in the ON and OFF periods due to I_{B3} . This is equivalent to having an ideal opamp OA₃ but with C_1 a bit greater than C_2 . The effect of I_{B3} can be treated as an offset in the final output, and (10) will get modified as (15). It can be corrected digitally in the measured ON and OFF periods of v_{OC2} . To keep the term $\frac{2\Delta T_\epsilon}{T_{OFF} + T_{ON}}$ very low, an opamp with very low bias current compared to i_{C1} and i_{C2} is preferred.

$$\pm kx = \frac{T_{OFF}-T_{ON}}{T_{OFF}+T_{ON}} \mp \frac{2\Delta T_e}{T_{OFF}+T_{ON}} \quad (15)$$

Let V_{OS1} , V_{OS2} , and V_{OS3} be the input offset voltages of OA₁, OA₂ and OA₃, respectively. V_{OS1} becomes present at the output of OA₁ and, therefore, $v_{o1} = (|v_{oi1}| \mp V_{OS1})$ during the first half cycle, and $v_{o1} = (|v_{oi1}| \pm V_{OS1})$ in the second; this is assuming that the circuit is not in the initial transient after being supplied and, hence, the offset is not affected by the charge amplifier gain. Thus, the average value of the signal v_{o1} over one full cycle of v_s will not have any effect due to V_{OS1} . Hence, the final output will not be affected due to V_{OS1} . Similarly, it can be seen that the effect of V_{OS2} on v_{o2} , and thereby the final output is negligible.

Due to V_{OS3} there will be a fixed term V_{OS3} in v_{oi3} . The effect of this will get cancelled as in the case of V_{OS2} , thanks to the negative feedback in the auto-balancing approach. In addition, there will be a current V_{OS3}/R_E flowing through C_{F3} , introducing an additional integrating term $\frac{1}{C_{F3}} \int \left(\frac{V_{OS3}}{R_E}\right) dt$ in v_{oi3} , where $R_E = R/2$. This is equivalent to the case of I_{B3} . The effect can be analyzed by replacing I_{B3} in (14) with $\frac{V_{OS3}}{R_E}$. The final output will modify as in (15). Again, to keep the term $2\Delta T_e$ in (15) extremely low compared to $T_{ON} + T_{OFF}$, an opamp with very low input offset voltage that drives negligible current $\frac{V_{OS3}}{R_E}$ compared to i_{C1} and i_{C2} needs to be selected and used.

E. Non-idealities of the Inverting Amplifiers

The above explanation assumed ideal inverting amplifiers, with gain -1 , feeding to the switches S_1 and S_2 as in Fig. 1. The inverting amplifier processing v_{oi1} will provide an output $-v_{oi1}(1 + \epsilon_{R1})$, where ϵ_{R1} is a gain error due to the mismatch between the feedback and input resistors. Similarly, the one processing v_{oi2} will give $-v_{oi2}(1 + \epsilon_{R2})$, where ϵ_{R2} is again a gain error due to the resistors mismatch. These gain errors will be present throughout a half cycle of v_{o1} and v_{o2} . The effect, in the final output, will automatically get corrected once the one-time correction for the mismatch, between the signal paths, is implemented as described in section III.A.

Also, the inverting amplifier will have a very small dc output that is twice the input offset voltage of the opamp. An opamp with very low input offset voltage (75 μ V) is selected such that this value is negligible compared to average values of the switch outputs v_{o1} and v_{o2} corresponding to the lowest range of the capacitive sensor. Additionally, there will be a dc part in the inverting amplifier output due to the bias current. To keep this value very low, the appropriate value of the resistor was introduced in the non-inverting terminal of the opamp [30].

F. Non-idealities of the Switches

Two main factors of the switch that need to be analyzed are charge injection, and ON resistance, r_{ON} . Switches with zero charge injection are available, e.g., MAX4709, hence the first factor can easily be taken care of. (6), and then (10) and (11) have been derived assuming that the equivalent resistance seen from the input of OA₃ is same for both the signal paths (i) with R_1 , S_3 and S_1 , and (ii) with R_2 , S_4 , and S_2 . In the circuit, $R_1 = R_2 = R$. Hence switches with identical ON resistance is preferred. Since the durations for which S_3 and S_4 are

respectively connected to the corresponding input signals are not the same, when $C_1 \neq C_2$ the effective ON resistance of the switches, for those two channels (i) and (ii), will be a function of T_{ON} and T_{OFF} . Thus, the best way to limit the effect of r_{ON} is by selecting switches that have $r_{ON} \ll R$.

Each switch takes a finite delay time t_d for every transition in its output. For the switch used in the prototype, t_d is a few hundreds of nanoseconds which is a factor of 1000 times less than the time period T of v_{tri} . It is important that $t_d \ll T$. Since the same delay is present for both low-to-high and high-to-low transitions, the width of the v_{o3} and v_{o4} waveforms (refer Fig. 2) remains the same except that there will be a finite delay in each transition. In addition, switches S_3 and S_4 are controlled by the same signal and, hence, there will be no error as long as the mismatch between t_d for both the switches is negligible. As both S_3 and S_4 have the same value ($R_1 = R_2$) of the resistor connected at the common terminal, the mismatch in t_d is expected to be negligible. On the other hand, the delay time will have no noticeable error for S_1 and S_2 as it is much smaller than T_s ($= 2$ ms). Also, for S_1 and S_2 the value of the signal during the transition is very close to zero. The delay time due to comparator OC₁ will have the same effect as that due to t_d of S_1 and S_2 . Similarly, the delay time of OC₂ can be treated as an additional delay time in the transitions of S_3 and S_4 .

Another limitation of the switches is the OFF resistance of the non-selected input. When S_1 is at position-0, as voltages at node-0 and node-1 are in phase opposition, there will be a small current between these nodes through the OFF resistance. This current will not go to the next stage through the common terminal of the switch. The same logic applies to the switch S_2 . On the other hand, when S_3 is at position-1, there will be a small current flowing from node-1 to node-0. This value will be very low as $R_1 = 100$ k Ω is much less than the OFF resistance. Due to this, there will be a small gain error in this signal path. A similar effect is present in the case of S_4 . The mismatch between these gain errors will be corrected in the one-time mismatch correction described in section III.A.

G. Saturation of Opamps

Let $\pm V_{SAT}$ be the positive/negative saturation voltage of both OA₁ and OA₂. In order to avoid saturation problems, the following condition must be satisfied: $-V_{SAT} < v_{oij} < +V_{SAT}$, where j can be either 1 or 2. Assuming the expressions of v_{oij} given in Section II, the previous condition implies that $C_j < V_{SAT} C_{Fj} / \sqrt{2} V_s$. Accordingly, for a sensor with a linear and inverse characteristic, the range of kx is limited as indicated in (16) and (17), respectively.

$$|kx| < \frac{V_{SAT} C_{Fj}}{\sqrt{2} V_s C_0} - 1 \quad (16)$$

$$|kx| < 1 - \frac{\sqrt{2} V_s C_0}{V_{SAT} C_{Fj}} \quad (17)$$

Considering the value of the components employed later in Section IV, the proposed circuit provides a full range operation (i.e., $-1 \leq kx \leq 1$) for the linear case, but not completely for the inverse case, to be precise: $-0.92 \leq kx \leq 0.92$.

H. Summary

As a brief summary of the analysis, the important parameters that are considered in the analysis and their effect or the condition to keep the associated error within the limit are tabulated and presented in Table I.

TABLE I
CIRCUIT PARAMETERS AND THEIR EFFECT

Source	Effect/remedy	Source	Effect/remedy
Component mismatch	One-time correction.	$r_{ON, OFF}$ resistance	One-time correction.
Parasitic capacitance	Not affected	Charge injection	Zero charge injection switches
V_{OS1}	negligible		
$V_{OS2}, V_{OS1}, V_{OS2}$	Not affected	Delay of switches	No effect if no mismatch, and $t_d \ll T$
V_{OS3}	$\frac{V_{OS3}}{R_E} \ll i_{C1}, i_{C2}$	Delay of OC ₁ and OC ₂	No effect if no mismatch
I_{B3}	$I_{B3} \ll i_{C1}, i_{C2}$		

IV. EXPERIMENTAL SETUP AND RESULTS

The prototype of the proposed CPC was realized using the components specified in Table II, according to the circuit diagram in Fig. 1, powered from a ± 10 V supply, and tested using the experimental setup shown in Fig. 5. As a proof-of-concept, the prototype was implemented with discrete components and ICs and, hence, has not been optimised for low power consumption. The power consumption of the opamps, comparators, and switches were $P_{OA} \sim 27$ mW, $P_{OC} = 102$ mW, and $P_{SW} = 14.8$ mW, respectively. This leads to a total power consumption < 150 mW. The sinusoidal AC signal and the triangular reference signal in the prototype were generated by a function generator. In a real application, these signals could be generated through either a microcontroller with an embedded DAC or a waveform generator chip, for instance, the AD9837 from Analog Devices. This is a low-cost (1\$), low-power (4 mA) and programmable solution that can provide an output signal (sinusoidal or triangular) with a very low thermal sensitivity and very low harmonic distortion (-68 dB).

To test the prototype for known capacitance values, two identical standard variable capacitance boxes, with negligible leakage conductance were selected. These are used to represent C_1 and C_2 . To correct for the effect of bias current of opamps, a 10 M Ω resistor was introduced across C_{F1} and C_{F2} . To smoothen output of OA₃, a Butterworth low pass filter with a corner frequency of 100 Hz was introduced between the output

TABLE II
PROTOTYPE

Component	Part/Value	Component	Part/Value
v_S	5 V _{pp} , Using AFG3022B	f_S	500 Hz
v_{tri}	From AFG3022B	f_{tri}	10 kHz
$S_1 - S_4$	MAX4709	V_p	5 V
C_1, C_2	250 pF ± 50 pF	OA ₁ to OA ₃	OP07
OC ₁ , OC ₂	LM311	$C_{F1} = C_{F2}$	1 nF
R_1, R_2	100 k Ω	C_{F3}	1 nF

Fig. 5. Experimental setup to characterize the performance of the proposed CPC.

Fig. 6. Important waveforms from the prototype captured using the oscilloscope. The waveforms of v_{o3} , v_{o4} , v_{o13} and v_{oc2} , correspond to $C_1 = 300$ pF and $C_2 = 200$ pF.

of OA₃ and non-inverting terminal of OC₂. During the experiments, the important waveforms from the prototype were observed using a digital oscilloscope. A screenshot of the same, corresponding to $C_1 = 300$ pF and $C_2 = 200$ pF, is shown in Fig. 6. The signals v_{o3} , v_{o4} , v_{o13} and v_{oc2} are indicated in Fig. 6. Since $C_1 > C_2$, the amplitude of the v_{o3} is higher than that of v_{o4} . This causes v_{o13} to be negative and thus, $T_{ON} < T_{OFF}$.

A. Testing the Linearity

In order to test the linearity of the proposed scheme, C_1 and C_2 were initially set to the nominal capacitance value of 250 pF. Then, C_1 was varied for a range of ± 50 pF, in steps of 10 pF, while C_2 was varied by the same amount in opposite direction. For each set of values of C_1 and C_2 , T_{ON} and T_{OFF} were measured using the 24-bit timer module of the ATSAM3X8E microcontroller, from Atmel, with a timing resolution of 12 ns. This was also simultaneously measured and compared using an oscilloscope. The duty cycle, D , in each case was determined. The plot of $\frac{(T_{OFF} - T_{ON})}{(T_{OFF} + T_{ON})}$ and $\frac{(C_1 - C_2)}{(C_1 + C_2)}$ is given in Fig. 7. The relationship between these two, observed experimentally, is highly linear, with a maximum non-linearity error (NLE) of 0.64% FS, as can be seen in Fig. 7. In addition, the prototype circuit has a sensitivity, defined as $(T_{OFF} - T_{ON}) / (C_1 - C_2)$, of $0.2 \mu\text{s/pF}$, which can also be obtained theoretically from (9) as $T/2C_0$. The same test was conducted for sensors possessing inverse characteristic as in (2). It was also found to be linear as expected.

Fig. 7. Results from the prototype. The ratiometric output and the corresponding percentage non-linearity error (NLE %) are shown.

B. Testing the SNR, Resolution, and Repeatability

To ascertain the signal-to-noise ratio (SNR), the resolution in terms of effective number of bits (ENOB) and the repeatability, first C_1 and C_2 were set to $C_1 = 200$ pF and $C_2 = 300$ pF. Then, the output from the prototype was recorded for 100 consecutive measurements, all the while keeping C_1 and C_2 unchanged. The performance parameters were calculated from the recorded values using the formulae [31] given in Table III. The results obtained are presented in the same table. As can be seen, an effective resolution higher than 13 bits was obtained, which is high enough for many industrial and laboratory applications. Taking into account that the sensitivity is $0.2 \mu\text{s/pF}$ and that the main uncertainty source in the time-interval measurements is the timer quantization (12 ns), the resolution of the circuit prototype is 60 fF.

C. Rise Time

In order to determine the rise time of the proposed CPC, an additional capacitor, $C_{new} = 20$ pF, was introduced. An SPDT switch S_5 was used to switch C_{new} between C_1 and C_2 . One of the ends of C_{new} was connected to node q of the circuit in Fig. 1, while the other end to the common terminal of S_5 . The position-0 of S_5 connected C_{new} to node p , whereas position-1 to node r . S_5 was controlled using a switching signal v_{sw} , which was a 5 Vpp square pulse signal at a frequency of 5 Hz. For the test, C_2 and C_1 were set to 210 pF and 270 pF, respectively. At position-0 of S_5 , the resulting capacitances were $C_2 = 210$ pF and $C_1 = 290$ pF, whereas at position-1, $C_2 = (210 + 20)$ pF and $C_1 = (290 - 20)$ pF, introducing a $\Delta C = 20$ pF in both sensing capacitors. This causes a rise in v_{oi3} . The screenshot of the

Fig. 8. Rise time test of the prototype capacitance-to-pulse-width converter.

TABLE III
PROTOTYPE

Parameter	Formula	Value
Signal-to-noise ratio (SNR)	$SNR = 10 \log \frac{\sum_{i=1}^M X(i)^2}{\sum_{i=1}^M [X(i) - \bar{X}]^2}$	80.12 dB
Resolution (ENOB), n bits	$n = \frac{(SNR - 1.76)}{6.02}$	13.02 bits
Repeatability, d	$d = \frac{\Delta_{rmax}}{kx_u - kx_l} \times 100\%$	0.027%

$X(i) = i^{\text{th}}$ measurement, $\bar{X}$ = average value of the measured data
 M = total number of measurements recorded for the repeatability study,
 kx_u = upper limit of measurement range of CPC, kx_l = lower limit of measurement range of CPC, and Δ_{rmax} = maximum difference between repeated measurements

oscilloscope indicating v_{sw} and v_{oi3} is given in Fig. 8. The time taken for v_{oi3} to rise from 10% to 90% of its final value was found to be approximately 5 ms. The rise time could be improved by decreasing the value of C_{F3} .

V. DISCUSSION

The proposed CPC provides a PWM signal corresponding to the change in sensor capacitance, with the measurand. The quasi-digital PWM signal is easily digitized using a counter. This approach gives a high update rate at relatively low complexity, cost, and power. The sensor is excited by a sinusoidal signal, which is inevitable in applications where measurement at particular frequency/frequencies is required to ensure functionality [25], [26], and helps the overall measurement system deliver improved performance [30]. The final linear output, derived from the PWM signal, is independent of the sensor characteristic (linear or inverse) and C_0 . The output is insensitive to parasitic capacitance and has very low sensitivity to the non-idealities of the components and ICs used. The impact of component mismatch is minimized by a one-time correction. The prototype has an NLE $< 0.7\%$, a resolution of 13.02 bits (ENOB), an SNR of 80.12 dB and a rise time of 5 ms. These features widen the range of sensors that can be interfaced using the CPC. For instance, the proposed CPC can be easily applied to measure liquid-level [15] and relative humidity [32] capacitive sensors that are in the range of hundreds of pF. The design of the proposed circuit is also suitable for IC fabrication, although it requires four resistances (R_1 , R_2 , and those two in parallel with C_{F1} and C_{F2}) of high value that could occupy a significant layout area; if this was critical, those resistances could be placed off-chip. A comparison of the performance of the proposed circuit vis-à-vis existing interfacing circuits for capacitive sensors was conducted focusing on similar schemes and the results are presented in Table IV.

As can be seen in Table IV, there is only one scheme [36] that accepts single-element and differential capacitive sensors and provides a quasi-digital output, like the proposed CPC. But the circuit in [36] requires a relatively expensive multiplier IC for its implementation. Such ICs have limited accuracy and hence limits the overall performance. The final output depends upon the nominal capacitance, which is another demerit. Since

TABLE IV
COMPARISON

Work	Method	Suitable to interface	Parasitic capacitance	Excitation	Conversion time	Resolution	NLE (%)	Number of components	Type of Output	Output sensitive to C_0
[33]	Dual-slope ^(a)	Single element	Not sensitive	Sine wave	24 ms	NS	0.27	M	Time-to-digital	Not sensitive
[34]	Single-slope	Single element	Sensitive	Square wave	15 μ s	13 bits	--	L	Duty Cycle	Sensitive
[35]	Switch-Bridge	DCS	Sensitive	Square wave	NM	NS	NM	M	analog	Sensitive
[36]	Relaxation Oscillator and multiplier IC	Single element, and DCS	Sensitive	Square wave	~15 ms	0.1 fF	0.999 ^(b)	M/H	quasi-digital ^(c)	Sensitive
[37]	Self-balancing Bridge ^(d)	DCS	Sensitive	Sine wave	NM	0.05 ^(e) , NM	0.8	H	analog	Not Sensitive
[24]	Capacitance-to-Pulse-width	Single-element	Not Sensitive	Sine wave	13 ms	13.31 bits ^(g)	0.06 ^(h)	M	quasi-digital	Not Sensitive
This work	Capacitance-to-Pulse-width	Single-element, and DCS	Not Sensitive	Sine wave	~5 ms	13.02 bits ^(g)	0.64	M	quasi-digital	Not Sensitive

Notes: NS- Not studied. NM-Not measured and reported. DCS-Differential capacitive sensor. M-Medium. L-Low. H-High.

^(a)Requires a special clock and associated circuit. ^(b)Linearity coefficient R^2 . ^(c)Neither ratio nor ratiometric. ^(d)Using two multiplier ICs,

^(e)Standard deviation of kx , ^(f)Without the power consumption of sinewave and triangular wave generators, ^(g)ENOB, ^(h)Range of ± 5 pF.

the output is not ratiometric, it does not have the attractive feature of being able to interface either a linear or an inverse DCS, and still deliver a linear output. Also, it is a relaxation oscillator and hence measurement at a particular frequency of interest is not feasible. Another limitation of this scheme is its sensitivity to parasitic capacitance. Although the methods in [33] and [37] use sine wave excitation, [33] is suitable only for single-element capacitive sensors. Also, it is difficult to improve the conversion time further as it relies on the dual-slope technique. In addition, [33] requires a very special clock to achieve good linearity, adding to its complexity, cost, and power consumption. The method in [37] needs two multiplier ICs, owing to which its overall performance is limited. This method is suitable only for differential capacitive sensors, and the final output is analog. So, it does not possess the advantages of a quasi-digital output that the proffered circuit has. The output in [37] is sensitive to parasitic capacitance, which is another major concern. The schemes in [34] and [35] use square wave excitation. The outputs in both these schemes are sensitive to parasitic capacitance and the final outputs sensitive to the nominal value of the capacitance. [34] is useful only for single-element capacitive sensors, while [35] is suitable for sensing differential capacitance alone. Additionally, they provide only an analog output and hence need additional circuitry to be interfaced with a digital system, which will add to the cost and amount of power consumed. The circuits in [34]-[37] do not have insensitivity to parasitic capacitance. The proffered converter achieves parasitic insensitivity through its simple design. The circuit in [24] is a CPC where an initial study about its feasibility for a single-element capacitive sensor is reported. It is compatible with single-element sensors with linear characteristic alone, and not suitable for DCS.

VI. CONCLUSION

The conception and realization of a new capacitance-to-pulse-width converter that is compatible with both single-element and differential capacitive sensors has been presented in this paper. It uses sinusoidal excitation and provides a linear output. The output from the converter is independent of the nominal capacitance, and sensor characteristic; the sensor characteristic can be either linear or inverse. Further, it is insensitive to parasitic capacitance and has only negligible impact due to non-idealities such as offset voltage, bias current, charge injection and ON resistance, provided the components and ICs are selected as per the guidelines given in the paper. In the proposed scheme, the pulse-width converter part of the circuit processes the signals due to the change in the sensor capacitance alone. Hence its full capacity is used to process the most useful information; not the unwanted signal due to the offset capacitance. The prototype of the proposed CPC exhibited a maximum non-linearity error of less than 0.7 %, a resolution of 13.02 bits (ENOB), and an SNR of 80.12 dB. It has a fast response indicated by a rise time of 5 ms. All these features make the proposed CPC suitable for a wide range of applications. It is noteworthy that many capacitive-based applications employ sinusoidal AC excitation, and the proposed first of its kind CPC, with its quasi-digital PWM output, offers an effective solution to realize the digitization of the capacitive sensor output in these applications.

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