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Universal Asynchronous Sensor Interfaces

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Abstract

Sensor interfaces should provide an output signal that can be read out by a microcontroller. This paper shows, that besides digital output signals also time-modulated signals are suited to be read out by microcontrollers. It is shown, that the attractive features of time-modulated signals concern the simplicity of the circuits and the flexibility of the signal processing. It is shown that precision interface chips can be implemented with low-cost CMOS technology. To make the interface chips universal, specific physical problems for typical applications and measurements have been taken into account. Interface chips in which dedicated measurement techniques have been applied, including autocalibration, two-port measurement and advanced chopping, can be very users friendly. Moreover, these techniques enable selective detection of the measurand with a high immunity for parasitic effects of the sensing elements, and for the effects of the connecting wires. As case studies, interface systems have been presented for capacitive sensors, resistive-bridge sensors.

Keywords: Smart Sensor Systems, Sensor Interfaces, Smart Sensors, Asynchronous Interfaces.

1. Introduction

Universal sensor interfaces form the "signal bridge" between common sensing elements, which convert physical signals into electrical ones and the digital world. Sometimes, the functions of physical or chemical sensors, electronic interfaces, and microcontrollers are combined in overall designs. These functions include sensing, signal conditioning, analogue-to-digital conversion, bus interfacing and data processing. There are cases where these functions can even be implemented in a single chip. For instance, with acceleration sensors, it is quite well possible to integrate the sensing element into a micromachined chip, together with the required electronics for signal interfacing. Packaging the chip will not degrade the performance of this smart sensor. For other types of sensors there can be many reasons to implement the various parts of a sensor system separately, using different components. Usually, these reasons stem from physical or economical considerations, such as:

- o Harsh temperature conditions and corrosion disable the use of electronic circuits in an environment in which the sensing elements are operated.
- o The product volumes are often far too low to make full integration economically feasible.
- o The sensing elements often cannot be manufactured in a process compatible with electronics.
- o Using off-the-shelf products speeds up system design.

For designs in which the sensing elements have to be separate from the interface electronics, the hardware configuration of figure 1 could offer a good solution for sensor systems targeted for a medium-volume market. The use of a clock line is optionally and will be discussed in section 3. The modifier consists of electronic circuits that provide multiplexing, conditioning and conversion of the sensor signals, including A/D conversion. In this setup, each of the system parts can be selected or designed for optimum system performance. This often requires a variety of different technologies, such as:

- o Packaging technologies that yield a good compromise between interaction with the physical environment and immunity against corrosion,
- o Technologies for the modifier hardware, to optimise precision and speed of the analog and digital *signal* processing,
- o Technologies for the microcontroller hardware, to optimise *data* processing for the best performance-to-price ratio.

For such a setup, the use of a universal interfaces could reduce the burden of separately redesigning the electronics for each type of sensing element. As explained elsewhere [1], the use of two-port measurement helps to overcome the problems posed by the parasitic impedances of connecting wires or cables. In many cases the cost of wiring can make up a considerable part of the total system costs. To reduce wiring costs, high-frequency signals transported via the wiring should be minimized. For this purpose, a modifier can be selected that does not apply high-frequency excitation signals or high-frequency clock signals. In sections 3 and 4, it will be shown that for this purpose modifiers with an internal asynchronous converter can be used.
In sensor system design, the appreciation of universal interfaces is rapidly increasing. In the next subsection, a number of desirable features and the design criteria of such interfaces will be discussed, while in section 3 the typical features of asynchronous converters will be discussed. For any interface circuit the quality of the front-end design is very important for the overall performance of the whole sensor system. Such front-ends will be introduced in section 4. Finally, as a case study, section 5 discusses the designs and features of asynchronous signal processors for capacitive sensors and resistive bridges.

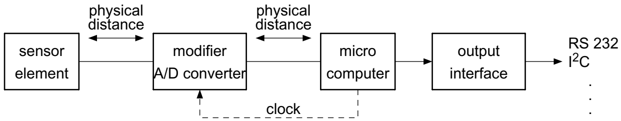


Fig. 1. Description Possible hardware configuration for a smart sensor system.

2. Universal sensor systems

Universal sensor interfaces are designed for low-cost rapid prototyping. The universal properties allow users to re-use their tools and knowledge in alternative applications. Paradoxically, the front-end circuitries in these interfaces are not universal at all, but have been designed to offer optimal signal-processing quality for a range of specific types of sensing elements. These dedicated front-ends should assist users in applying the most appropriate excitation signal, measurement technique, and measurement configuration.

Generally, sensor systems lack standardization. For communication between microcontrollers and computers, standard bus protocols can be applied. However, due to physical constraints of the sensing elements, no standards are available. Sensing elements generate many different types of electrical signals, e.g. changes of resistance, capacitance, voltage, current, charge, bridge imbalance, power, etc. Some sensing elements, such as pH sensors, require input amplifiers with very high input impedance, while other types of sensing elements, such as photocurrent sensors, require input amplifiers with very low input impedance. These examples show that for the design of a universal sensor interface, conflicting demands must be met. Therefore, it is not possible to implement a universal sensor interface just with only one type of input amplifier stage. A variety of possible input amplifier stages is required, with each of the stages optimized for a certain category of applications.

During the last decade, many interface chips were introduced to the market. Some of these interface chips mainly consist of a microcontroller or DSP core with only a few front-end circuits for some specific applications. Other interface chips, [1] and [2], mainly consist of an A/D converter with a programmable gain-amplifier implemented with front-end circuits for a few types of sensors. Interface chips that are more universal are presented in [3], [4] and [5].

Besides economic reasons, important considerations in selecting a certain type of interface include: the types of supported sensing elements, the applied measurement techniques, the specifications and the supporting tools.

Types of supported sensing elements: The input stages of sensor interfaces must be optimized for the type and the dynamic range of the input signals and for the source impedances. Therefore, in the various interface modes, dedicated front-end circuits with a special configuration are applied to specific classes of sensing elements, such as:

- o Low-impedance resistive elements, such as Pt100, etc.
- o High-impedance resistive elements, such as conductivity sensors.
- o Thermistors.
- o Capacitive elements.
- o Resistive bridges for strain gauges, load cells, etc.
- o Voltage-generating elements, including thermopiles, thermocouples etc.
- o Current-generating elements, including photo-cells, PSDs etc.

Specifications and features: There is a wide range of special features to be considered by the user. For instance:

- o Accuracy and dynamic range.
- o Measurement time. The effects of random errors can be reduced by filtering (for instance averaging) over a longer measurement-time interval. In some sensor interfaces, the measurement time is programmable, so that it can be changed for the best performance for each specific measurement.
- o Temperature range.
- o Supply voltage and current.
- o The need for clock signals, reference voltages, or external components (see next sections).
- o The type of output signal.
- o Built-in peripherals for temperature compensation and voltage references.

The applied measurement techniques: In addition to the specifications of a sensor interface, it is important for the user to know and understand the applied measurement techniques, because these are important for the reliability, long-term stability, and immunity against cross effects and undesired signals. Important items concern:

- o Autocalibration. Autocalibration [6] compensates for the effects of inaccuracy, drift, and temperature and voltage coefficients of the interface transfer parameters.
- o Two-port measurements. The application of this technique [6] reduces or eliminates the parasitic effect of connecting wires.
- o The frequency, wave shape, and magnitude of the excitation signals generated by the interface. A wrong choice can lead to corrosion, nonlinearity, and increased influence of parasitics.
- o The noise and EMC properties.
- o Chopping. The use of this modulation/demodulation technique reduces the effects of $1/f$ noise and offset.
- o Input biasing conditions. Special attention is needed for grounding of the sensing elements and the power supply.
- o Dynamic element matching (DEM). Application of this technique yields better reliability and long-term stability.
- o The type of A/D conversion (see the next subsections).

3. Asyncronous converters

Sensor systems are equiped with analog-to-digital converters (ADCs), which require very high (absolute) accuracy, linearity and resolution. Often, it is important that the power consumption of these converters is low and that their clock signals do not cause interference for the sensitive front-end circuits. On the other hand, the bandwidth of the sensor signals to be converted is often low.

The A/D converter requirements can be met with both incremental sigma-delta A/D converters [7] and asynchronous converters. In this paper, we will mainly focus on asynchronous converters, which have the advantages of simplicity, flexibility, and do not require a clock line. For a review of principles and architectures of A/D converters, the reader is referred to [8].

The output signals of asynchronous converters are square-wave signals, which have been modulated in the time domain. Various types of modulation can be applied, such as duty-cycle modulation, period modulation, and pulse-width modulation. The trade-offs originate from the unavoidable distance between the various system parts.

> ***Example 1*** Let us assume that for the system configuration of Fig. 1, the sensing element has to sense a low-frequency physical signal while being situated 4 meters away from the microcomputer hardware. For economical reasons it could be appealing to implement the modifier at the same PCB board as the microcontroller, so that only a single board is required and the modifier and the microcontroller can share the same clock line for timing. However, this yields a rather long distance between the sensing element and the modifier. The wiring is a long antenna for interfering signals, or when shielded, cause a large parasitic capacitance, which can result in too long time constants for the excitation signal. In that case, it could be preferable to make the distance between the modifier and the sensing element as small as possible, for instance 50 cm or less and to accept a distance of about 3.5 m between the modifier and the microcomputer.
>
> In such a case, the modifier and the sensing element could be assembled in, for instance, a single probe. Because of the long distance between the modifier and microcontroller, it would be an advantage when the clock line could be skipped, because high-frequency signals can cause interference, or, when shielded, can be attenuated due to the time constants of the connection. The use of coaxial cable would have the drawbacks of increased costs and size of assembly material and handling. Using asynchronous converters can solve such problems. With these converters, no clock line is required. The frequency of the output signal can be limited to a bandwidth of a few kHz. The use of period-modulated output signals allows rather large time constants, up to about 10% of the period time.

This example shows why it could be advantageous to use asynchronous converters. On the other hand, many manufacturers of sensor interface circuits, [1] and [2], have decided to use on-board sigma-delta converters. In section 3.5, a comparison is made between the features of these two types of converters.

3.1. Conversion of sensor signals to the time domain

In asynchronous converters, the conversion of sensor signals to the time domain is accomplished in two steps.

Firstly, an integrator capacitor C_i is charged with a charge Q_x that represents the sensor signal. In the so-called *switched-capacitor* (SC) converters, for voltage-independent capacitor, the charge Q_x equals the product C_sV_x of the sampling-capacitor value and the capacitor voltage V_x. Often, the voltage V_x represents the sensor signal. These front-end circuits consist of selectors, choppers and amplifiers. In capacitive sensors, the capacitor C_s could also represent the measurand. However, for accuracy reasons, often the sensing element and the converter have been separated by a buffer amplifier.

Secondly, in the charge balancing concept, the charge Q_x of a sampling capacitor C_s is compared with a well-known reference charge. The reference charge is changed up to the level that the both charges are equal (balanced). In a frequently used implementation, the reference charge is obtained by integrating a precision reference current I_{ref} over a certain time interval t_x. This charge removes the charge Q_x of the sampling capacitor till the capacitor voltage equals the original value, which was there before sampling the charge Q_x. Because no charge has been lost, it holds that $I_{ref}\, t_x = Q_x$, so that:

$$t_x = \frac{Q_x}{I_{ref}} = \frac{C_sV_x}{I_{ref}} \,. \tag{1}$$

In this way a signal represented by charge, voltage or capacitance can be converted into the time domain.

Figure 2 shows a simple circuit in which this principle has been implemented. This circuit is suited for capacitive sensors as well as voltage-generating sensors. In the latter case, after closing the switches S_1 and S_3, the sensor voltage V_x is sampled. Next, upon a control signal of the comparator output, the pair of switches (S_1, S_3) are opened and (S_2, S_4) are closed, successively.

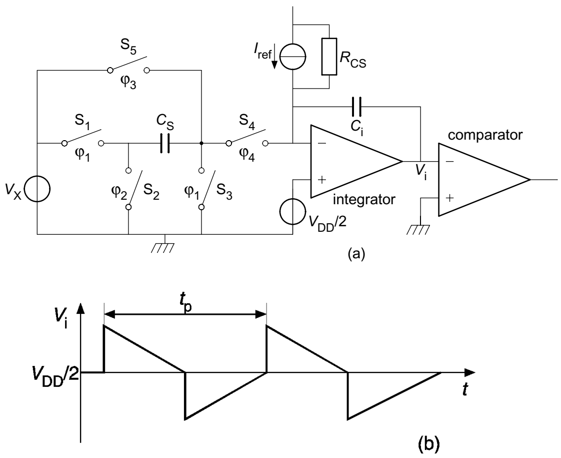


Figure 2 Basic circuit configuration for conversion of sensor signals to time intervals; (a) circuit diagram; (b) the integrator output voltage V_I. Charge added as a result of the switched-capacitor action is – on average - removed by the reference current source I_{ref}. The switches are controlled by the comparator output voltage.

By this action, the charge $Q_x = V_xC_s$ in the capacitor C_s is transferred to the integrator capacitor C_i, which causes a jump in the output voltage. The integrator consists of an opamp with a feedback capacitor C_i. Supposed, that the opamp is ideal, with an infinite slew rate, then the opamp is operated in its linear mode, so that the input voltage is "zero" and the inverting input terminal is virtually at ground potential. In case of linear operation, the effects of the various excitation voltages and currents can be found by applying the superposition theorem, which says that the total response upon the various excitations equals the sum the responses for each excitation, separately. When we suppose that the internal resistance R_{cs} of the reference-current source is infinitely high, then the current I_{ref} will flow into capacitor C_i, causing a linear decrease of the integrator output voltage. Once, the integrator voltage crosses the threshold voltage $V_{DD}/2$ at the non-inverting input terminal of the comparator, the comparator output voltage changes its state, which generates the command signal for controlling the switches, as mentioned before.

To eliminate the effect of $1/f$ noise, offset and other low-frequency nonidealities of the opamp and comparator, the whole process is repeated with the opposite signs of the relevant voltages and currents. This is accomplished by inverting of direction of the current I_{ref}. Inversion of the charge Q_x is obtained by replacing the switch pair (S_1, S_3) by the pair (S_5, S_2). In this way, DC sensor signals are converted to AC signals, which can selectively be processed, while removing the DC disturbances of the electronic circuitry.

From this simple circuit, the basic requirement to obtain an accurate conversion can easily be understood: Let us suppose that autocalibration [1] is applied. Autocalbration will eliminate the effects of multiplicative and additive parameters, including the absolute values of C_s and I_{ref}. However, these values should be stable during the conversion time and therefore also be independent of the sensor signal. Furthermore, undesired charge injection, as caused by clock feedthrough and channel-charge injection of the switches, should be as low as possible and/or compensated by applying autocalibration. For the opamp, a high slew rate is not a first requirement. Only in combination with the finite value of the internal impedance R_{cs} of this current source a finite slew rate will cause a problem. The effects of a finite loop-gain are partly reduced by autocalibration. The effects of transients in the integrator output signal are only relevant, when there is a residual effect at the moment of crossing the threshold value of the comparator. Therefore, the bandwidth and slew-rate requirements for the opamp are rather moderate [9]. In table 1 the basic requirements of the components of the converter circuit (Fig. 2(a)) have been summarized.

Table 1 Basic requirements of the components of the charge-to-time converter.

Component	Requirements
Switches	Very low clock feedthrough and channel-charge injection
Capacitors C_s and C_i	Very low leakage current and voltage coefficient
Opamp	Low-noise, low input-bias current
Comparator	Low-noise voltage
Current source I_{ref}	Stable, low-noise, high output impedance

3.2. Output signals

The output signal of an asynchronous converter is a square-wave signal (Fig. 3), which is modulated in the time domain. Various types of modulation can be applied, such as duty-cycle modulation, period modulation and pulse-width modulation.

In **period modulated signals** (Fig. 3(b)), the sensor information is modulated in the length of one or more periods. In this case, the time intervals are measured using only one type transient: the rising or the falling edge. In case of a large time constant, there will be some time

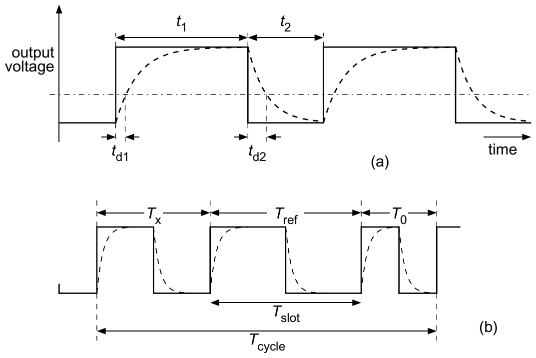


Figure 3 Various types of output signals (a) duty-cycle modulation (b) period modulation.

delay in the detection moment. These delays are equal for all detected transients, so that their effect on the measured time differences (intervals) is compensated. In **duty-cycle-modulated** signals (Fig. 3(a)) both the time delays of the rising and the falling edges affect the duty-cycle measurement. This has the drawback, that any change in the threshold voltage will influence the measured duty cycle. On the other hand duty-cycle modulated signals have the advantage that the avarage value is proportional to the modulating signal. So, in addition to information in the time-domain, in each period there is also information in the amplitude domain, which can easily be acquired with a simple low-pass filter.

3.3. Quantization noise of sampled time-modulated signals

With a microcontroller the various time intervals, t_1, t_2 and/or the period time t_p (= t_1 + t_1) of a time-modulated signal (Fig. 4) can easily be digitized. For this purpose, the output-voltage line of the asynchronous converter is connected to a timer input of the microcontroller. At the first sampling moment, after an up-going or down-going edge of the signal, the contents of a fast-running timer are copied to a special register (the capture register). In this way, time moments and time intervals are converted into integer numbers of counts. In between two sampling moments, within the sampling period t_s exact information about the precise moment of the transient is lost.

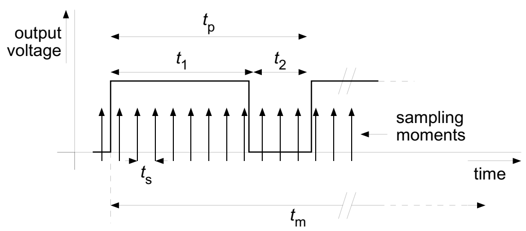


Figure 4 Sampling moments with time intervals t_s during a total measurement time t_m of a time-modulated output signal.

This loss of information is similar to the loss of information in any other type of A/D converter and gives rise to a conversion error, which is called *sampling error* or *quantization error*. Averaging the result of the measurement over a large number of periods will significantly reduce this type of noise. For period modulation, this averaging process is more effective than for duty-cycle modulation. This can be explained as follows:

a) With period modulation, after a series of concatenated periods, only the very beginning and the very end of the counting interval will introduce uncertainties. Therefore, the relative effect of the quantization error will go down proportionally to the total number N of periods.
b) With duty-cycle modulation, any up-going and down-going edge will introduce uncertainty. When these edges are not synchronized with the sampling moments, the quantization noise contributions of the various transients are stochastically independent. Then, in the average, the quantization noise will go down with the square root of the number of counts. Care should be taken, to avoid that sampling pulses of the microcontroller do not cause undesired synchronization of the sensor signal. Undesired synchronization would change the stochastic errors into systematic ones, which will exactly be repeated, each period. In that case, increasing the number of periods will not help to reduce the relative error. Injection of noise or pseudo noise can help to avoid undesired locking (de Jong *et al.*, 1996).

Averaging over N concatenated periods will yield a total

$$t_m = Nt_p = N(t_1 + t_2) \tag{2}$$

measurement time t_m which equals
The quantization noise has a uniform distribution function. In case of duty-cycle modulation, this noise causes a relative error with a

$$\sigma_{q,dcm} = \frac{1}{\sqrt{6N}} \frac{t_s}{t_p} = t_s \sqrt{\frac{1}{6 t_m t_p}} \tag{3}$$

standard deviation that equals:
Often resolution is expressed in term of bits. In that case, the resolution R_{bits} is related to the standard deviation σ as:

$$R_{bits} = -{}^2\log \sigma = \frac{-\ln \sigma}{\ln 2} \; . \tag{4}$$

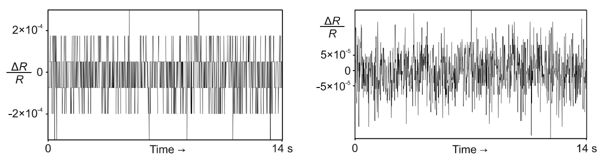
Often, for ADC converters the resolution is specified for 3 or 6 times the standard deviation. This will yield resolution figures being 1 or 2 bits less. For comparison with other publications, in this chapter we

$$\sigma_{q,pm} = \frac{1}{\sqrt{6}} \frac{t_s}{Nt_p} = \frac{1}{\sqrt{6}} \frac{t_s}{t_m} \; , \tag{5}$$

will use resolution figures as defined in Eq. (4).
In case of period modulation, quantization noise causes a relative error with a standard deviation $\sigma_{q,pm}$ that equals:
For both types of modulation, using a fast microcontroller will significantly reduce the quantization noise.

Example 2: Quantization noise for a sampled period-modulated signal. Suppose that the asynchronous converter would generate a period-modulated output signal with period time t_p = 0.3 ms and that the available measurement time t_m = 30 ms. Furthermore, let us suppose that the sampling time t_s = 0.3 μs, which represents a typical value for microcontrollers of, for instance, the type 8051. According to Eq. (5), it is found that $\sigma_{q,pm} = 4.1 \times 10^{-6}$, which corresponds to a resolution of $-({}^2\log \sigma_{q,pm})$ = 18 bits.
Nowadays, microcontrollers can be much faster. For instance, with microcontrollers of the type LPC2101, running at 70 MHz, the internal sampling frequency is as fast as 70 MHz, so that t_s = 14.3 ns. With such a microcontroller, for the same measurement conditions, the quantization noise would be reduced to only $\sigma_{q,pm} = 0.2\times 10^{-6}$, which corresponds to more than 22 bits. Such a resolution is close to that mentioned in [7] for a third-order incremental sigma-delta converter.

Example 3: Quantization noise and thermal (white) noise. Figure 5 shows the noise in the measurement for a sensor resistor R (= 100 Ω), as measured with the universal interface [5]. The two graphs represent the results for various values of the sampling time t_s. In figure 5(a) the effect of quantization noise can easily be recognized. When the sampling time is reduced to 14.2 ns (figure 5(b)), the graphs show that there is no significant quantization noise. Note that systems in which thermal noise dominates over quantization noise can not be improved by using an A/D converter with a higher resolution. For this reason a sensor systems with a 24-bits A/D converter seldom shows a 24-bits resolution. The values of the standard deviations of the measured noise signals shown in Fig. 5 have been listed in Table 2.



(a) (b)
(b)

Figure 5 Noise of an interface system for a measurement time t_m =14 ms and various sampling periods t_s: (a) t_s = 200 ns; (b) t_s = 14.2 ns.

Table 2 Standard deviation σ_R in the value of ($\Delta R/R$)

	Standard deviation	
t_m in ms	for t_s = 200 ns	for t_s = 14.2 ns
14	1×10^{-4}	0.58×10^{-4}

3.4. A comparison between synchronous converters and sigma-delta converters

Both asynchronous and sigma-delta converters belong to the group of indirect A/D converters, with as key characteristic the use of an intermediate time-domain signals in A/D conversion processes for signals in the low- and intermediate frequency range. The attractive features of indirect converters concern their high resolution, low power consumption and simplicity. Therefore, indirect converters are highly suited for smart sensor systems. The basic block diagrams of asynchronous and sigma-delta converters show a strong similarity (Fig. 6). In asynchronous converters, the microcontroller (Fig. 6(a)) takes care for digitizing, using a clocked gate and a fast-running timer.

To reduce power dissipation and emission of disturbing signals, it is advised to limit the application of high-frequency signals to a limited space/area. Because the timer clock is used only within a small area at the microcontroller chip, without external wiring or connections and far away from the sensitive input amplifiers, a very high timer frequency can be applied. As shown in example 2, a high sampling frequency results in (very) low quantization noise. In sigma-delta converters, the clocked gate is within the feedback loop. The advantage is that the output signal is already digitized (Fig. 6(b)), which usually will be appreciated by the user. Furthermore, when applying higher-order converters, the quantization noise and/or the clock frequency can highly be reduced.

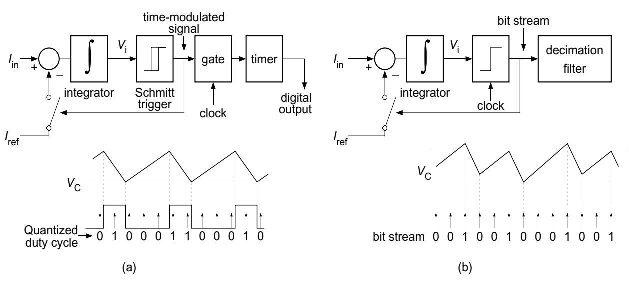


Figure 6 Block diagrams indirect A/D converters; (a) asynchronous converters, (b) synchronous (sigma-delta) converters.

However, with such converters also a number of problems will arise:

- o The occurrence of *dead zones*. Practical integrators will show some leakage. Dependent of the amount of leakage, at specific DC values of the sensor signal, the resolution can drop significantly. Consequently, the worst-case resolution will drop. This problem is important for the large group of sensing elements that generate low-frequency or DC output signals.
- o *Multiplexed sensing elements.* To perform autocalibration or to select elements from a sensor array, at the input of the sensor system the sensing elements are multiplexed. This caused the problem that at the beginning of each conversion the memory elements, including all of the capacitors, have to be reset. This would have the drawback of an increased acquisition time, thus slowing down the operation speed of the sensor system. The use of special filter topologies and initialization circuits (feed-forward paths) [10], can solve this problem and speed up the conversion rate.

In [10], many examples can be found of sigma-delta converters that have been designed for smart temperature sensors. It has been shown that such circuits can be combined with other circuits, which are needed for, for instance calibration of reference purposes. In [7], a third-order incremental converter is presented, implemented in 0.6 μm CMOS technology. The ADC chip contains also an internal programmable oscillator, a decimation filter and a bus interface. With a clock frequency of only 30 kHz, a resolution of 22 bits is obtained for a conversion time of 67 ms.
When using a microcontroller with a fast timer (see example 2), with asynchronous converters a similar performance can be obtained. In this case, the excellent properties of the asynchronous converters are due to the very high sampling frequency in the microcontroller timer. These high frequencies are only found locally in a limited area and volume of the micrcocontroller chip. In this part of the chip, power consumption can be limited by minimizing the parasitic capacitances and by using a reduced supply voltage. Related to this, also possible undesired effects of interference and cross-talk can be limited.

The simplicity of asynchronous converters is due to the fact that it concerns a first-order system and that the quantization process is performed in the microcontroller. Furthermore, the very limited amount of internal memory cells allows rapid adaptation of the converter to changes in the front-end circuit, or to multiplexing of sensing elements, or to change the acquisition time. It is possible to make asynchronous converters with a built-in quantizer, so that a digital output signal is achieved. However, then the advantage, that no counter, external clock line or internal clock is needed, is lost.

In summary, it can be concluded that asynchronous converters have the attractive features of simplicity and flexibility for rapid changes in the front-end circuitry or in the signal processing. These features make these converters attractive for universal converters with many front-ends and many users' options. Because of the small chip size, integrated asynchronous converters can easily be combined with sensing elements, as a system in a package. Moreover, because of their simplicity, asynchronous converters are rather suited for low- and medium volume applications. In such applications, a universal interface can perform multiplexing, chopping and precision analog processing of the sensor signal, while a low-cost microcontroller performs quantization and digital signal processing.

On the other hand, sigma-delta converters offer the advantage of delivering a digital output. For high-volume applications the complexity of sigma-delta converts is not a big problem, which makes these converters rather suited for dedicated designs of smart sensors and systems-on-a-chip (SOCs) with embedded microcontrollers.

4. Front-end circuits

4.1. Cross effects and interaction

The front-end circuits are very important, because they have to act as converters with, at the input side, optimum adaptation to the sensing element and, at the output side, with the A/D converter. Often the sensing elements acquire physical information under difficult physical circumstances. Therefore, for selective signal processing also the nonidealities of the sensing elements have to be taken into account. These nonidealities will include physical cross effects, noise, interference, electrostatic discharge (ESD), impedance levels, etc. To cope with this problem, even in a universal sensor interface, the front-ends are not universal at all. Instead of this, front-end circuits are tailored to have optimal features for specific tasks. This is accomplished by applying the best measurement techniques and, in addition to this, by using high-quality components. In front-end circuits, the electronic properties of the interface interact with those of the environment. This can cause undesired physical, chemical and electrical effects in both parts of the system.

Example 3: Undesired Interaction
Chemical effects: A DC biasing input current or excitation voltage generated by the interface can cause chemical problems, including corrosion, of the sensing element. Also the use of a switched-capacitor (SC) input of the interface will cause some DC current, which flows through the sensing element.
Physical effects: An excitation current for a resistive temperature sensor can cause undesired heating of the sensitive element.
Electrical effects: The decision to connect or not to connect sensing elements to ground can easily affect the biasing conditions of an input amplifier in a front end.

In addition to these measures, the properties of the interface circuit can be improved by minimizing the systematic and random errors of the interface and the connecting wires, as will be explained in the next sub-section.

4.2. Interference

In high-resolution sensor systems, noise and interference should be kept as low as possible, to minimize the stochastic errors. As a first step, thermal noise can be reduced by optimizing filter characteristics for the best signal-to-noise ratio of the front-end circuit. As a next step, the resolution can be improved with an increase of the measurement time. As shown before, quantization noise reduces inversely proportionally to the measurement time, while thermal noise reduces inversely proportionally to the square root of the measurement time.
In addition to noise reduction, it is very important to design for a high immunity for interfering signals. Shielding of connecting wires at the interface input should reduce the interfering signals as much as possible. However, to function properly, sensing elements have to interact with their environment, which limits the possibilities of taking proper measures against electro-magnetic interference. Therefore, it is advised where possible to use interference filters in the connecting wires. For low-frequency disturbing signals, filtering can be realized by auto-calibration and chopping.
To suppress high-frequency disturbing signals, good low-pass filtering is required. Interference of the microcontroller clock can cause undesired locking of the oscillator signal to the microcontroller clock frequency. In that case, the quantization errors are repeated in an exact way over a range of periods, so that a reduction of quantization noise can not be obtained by averaging over a number of periods. To disable locking, in addition to filtering of clock-feedthrough signals, dithering techniques can be applied [9] and [11].

5. Case studies

5.1. Front-end circuits for capacitive sensors

As a first case study we will compare and discuss the architectures and properties of the various interfaces for capacitive sensors, which have been presented in section 3 of this chapter. In [12] an interface circuit (Fig. 7) has been presented, that generates a square-wave output signal which is proportional to the sensor capacitance C_s.

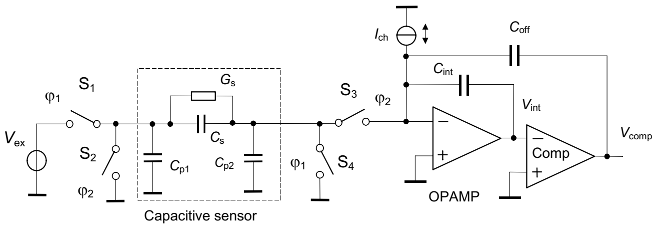


Figure 7 Direct use of the charge-to-time converter as interface by using the capacitive sensing element C_s as sampling capacitor.

It appears that the circuit diagram is equal to that of the charge-to-time converter presented in Fig. 2. The sensor capacitor is used as sampling capacitor. The circuit has the attractive feature that when the sampling capacitor is discharged rapidly, the effect of a shunting conductance G_s is reduced. This makes this configuration rather suited for sensing elements with leaky capacitors, such as capacitive humidity sensors. However, when the sensing capacitors have a small value, then the signals can be too small to take full advantage of the large dynamic range of the charge-to-time converter. Especially in universal interfaces the charge-to-time converters will have a relatively large range, to support a wide variety of applications. Therefore, in the universal transducer interface (UTI) presented in [5] and [13], the charge-to-period converter is preceded by a front-end amplifier (Fig. 8). Because this amplifier has been designed for just one type of sensing element, its noise performance can be optimized for this application. In fact this design consists of a universal charge-to-period converter and a dedicated pre-amplifier. This explains that, even for sensor capacitors with a full-range value of only 2 pF, still a very high resolution/repeatability of 50 aF can be obtained. However, the user should realize that with an excitation frequency in the range of 50 kHz to 100 kHz the sensor impedance is very high, so that even a small shunting conductance will have a large impact for the resolution.

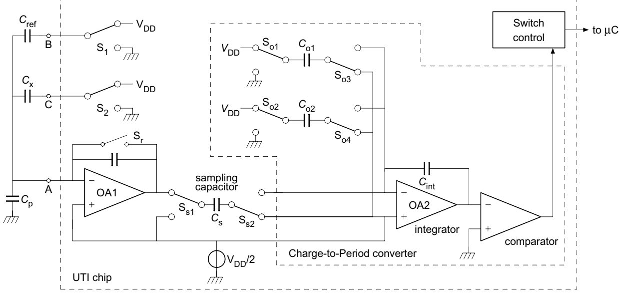


Figure 8 Principle of the UTI system for capacitive-sensor modes.

Example 4: It can be shown, that for the interface circuit of Fig. 8, a shunting leakage conductance G_{leak} of sensor capacitor C_x will cause a relative error ε_{sh} in the measurement, which equals:

$$\varepsilon_{sh} = \frac{(t_2 - t_1)G_{\text{leak}}}{2C_{\text{x}}} \tag{6}$$

From this equation it can be calculated, that for $C_{\text{x}} = 2$ pF, and $(t_2 - t_1) = 10\mu\text{s}$ and an error smaller than 500 aF ($\varepsilon_{\text{sh}} = 0.25 \times 10^{-3}$), it should hold that: $G_{\text{leak}} < 2 \times 10^{-10}$ S!

This example shows that for high-precision capacitive sensors the occurrence of shunting conductance should be avoided. Therefore, such sensors should be applied in a clean environment. When this is not possible, because of its lower sensitivity for shunting conductance, application of the circuit of Fig. 8 would be a better option. In both circuits, the use of the advanced chopping (+, -, -, +) technique suppresses interfering signals at the input of the sensing element. In fact the interface circuit, which generates the excitation signal and detects the responding signal of the sensing element, acts as a synchronous detector (Fig. 9), [13].

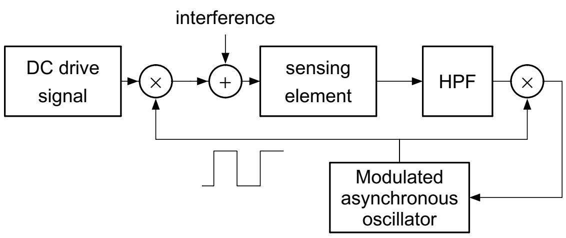


Figure 9 Principle of the interface system for capacitive-sensor modes.

The high-pass filter (HPF) represents the action of the advanced chopping technique. The filter transfer function has a second-order characteristic and provides filtering for low frequencies of offset $1/f$ noise and interference. In an experimental setup for capacitive sensors, [13] test results on the suppression of low-frequency interfering signals have been reported. In an experimental setup a test voltage, modeling the interference, was capacitively coupled to the input A (the common electrode in Fig. 8). The coupling capacitor was as large as the sensor capacitance C_{x}. The relative interfering amplitude equals the ratio of the amplitude of the interference and the amplitude of the signal on the transmitting electrode ($V_{\text{DD}}/2$). Figure 10 depicts the experimental results together with the calculated ones which appear to be close to each other.

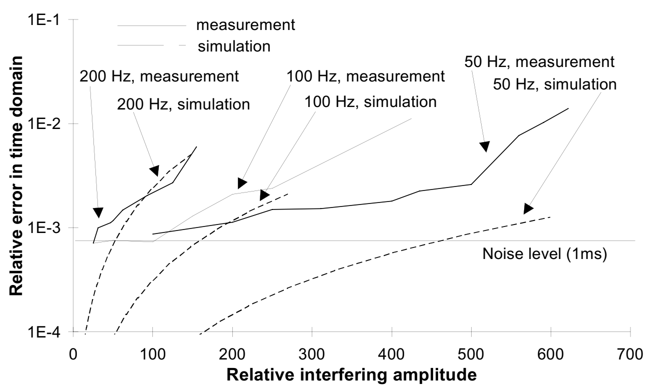


Figure 10 Suppression of low-frequency interfering signals for capacitive measurements with the system of Fig. 8.

5.2. Front-end circuits for resistive bridges

The second case study concerns a front-end circuit for resistive-bridge sensors that are used in many mechanical sensors for measuring force, pressure, acceleration etc. In many traditional interface circuits for resistive bridges, reference voltages are applied as DC excitation source. The same or another DC reference source is used to perform precise measurement of the bridge-output voltage V_{o} (Fig. 11), where it holds that

$$V_o = \varepsilon_{bridge} V_s \, , \tag{6}$$

where ε_{bridge} is the relative bridge imbalance, R_{bridge} is the bridge resistance, and V_{s} and I_{s} the bridge supply voltage and current, respectively.

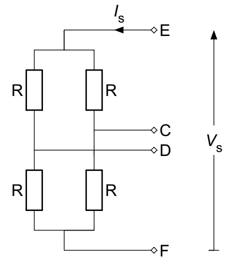


Figure 11 Resistive bridges with voltage supply or current supply.

From equation (6) it can be concluded, that it is not necessary to use (expensive) references sources. Because the relative bridge imbalance ε_{bridge} represents the measurand, a ratiometric measurement of V_{o} and V_{s} will be more precise. In that case, for the supply voltage only stability is required over the short measurement-time interval, which is in the order of tens of ms.
Another drawback of the traditional setup is that it uses a DC supply voltage. This is often not allowed, because it can cause corrosion, thus reducing the life time of the sensor gauge. Moreover, many signal nonidealities, such as offset, parasitic Seebeck voltages, and $1/f$ noise, are also in the low-frequency range. Therefore, often AC supply voltages are preferred.
These problems are solved in the universal sensor interface presented in [13]. Figure 12 shows the system setup.

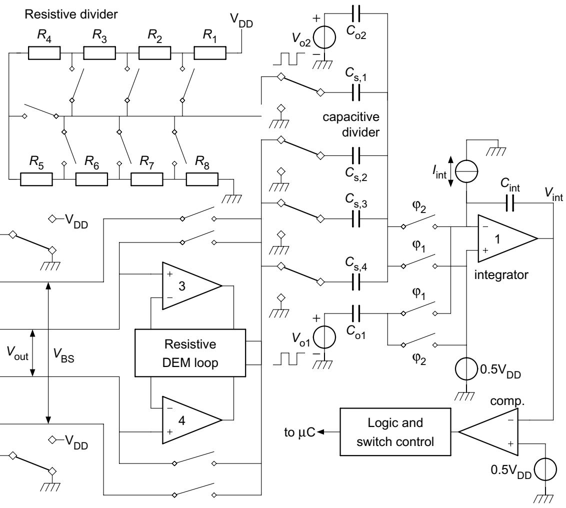


Figure 12 Setup of a bridge interface circuit.

The interface circuit consists of:

- an excitation part which generated an AC bridge-supply voltage/current,
- a divider part to measure the bridge-supply voltage/current,
- a selector, which selects the various voltages to be measured,
- a DEM amplifier to amplify small output voltages,

- o a voltage-sampling circuit,
- o a universal charge-to-time converter.

The excitation signal is derived from the supply voltage using a pair of chopper switches S_1 and S_2, which can invert the supply voltage, thus generating a square-wave bridge-supply voltage with amplitude V_{DD} (and peak-to-peak value of $2V_{DD}$!).

The switches are controlled by the relaxation oscillator in the charge-to-period converter. This is an interesting feature: Although the interface is asynchronous with respect to an *external* clock in the quantizer, *internally* it is synchronized by the oscillator signal, which period also represents the measurand. The importance of this feature is that in this way it is possible to apply synchronous detection as shown before in Fig. 9, and enabling filtering of low-frequency interference as described for the capacitive sensors.

For the output voltage V_{out} optionally a DEM amplifier can be used. In case that no amplifier is used, the bridge output voltage is sampled with the four sampling capacitors $C_{s,1}$ - $C_{s,4}$ connected in parallel. Next, the sampled charge is transferred to the charge-to-period converter. To do an accurate ratiometric measurement of V_o/V_s, the same sampling capacitors should be used to sample the bridge supply voltage V_s. Because of the high value of the supply voltage, this voltage is split up into smaller parts, which are more close to the size of the output voltage, but do not have to be exactly equal. This action is performed with the resistive divider R_1 – R_8 and the corresponding switches. The voltage across each of the resistors is sampled by one of the capacitors $C_{s,1}$ - $C_{s,4}$. After a complete cycle of 32 two phases, each of the 8 voltage parts have been sampled by each of the four capacitors. The results of all of these sub-measurements are added in the microcontroller. In this way, after a complete cycle, the sampled supply voltage is measured part-by-part with the same sampling components as the output voltage. This equalization of the size of the charge samples, results in much more relaxed requirements for the dynamic range of the charge-to-period converter.

When an amplifier is used, the ratiometric measurement is applied for AV_o/V_s. Therefore, the amplification factor A should be known very precisely. This can be achieved by applying an amplifier with dynamic element matching (DEM) [14].

To complete the method of three-signal autocalibration in each measurement phase also offset measurements are performed. During the offset measurement the sensor signal is zeroed. In case of the resistive bridge measurement, this is performed by sampling with the sampling capacitors of zero-voltage difference, for instance, at the CM level of $V_{DD}/2$. In the microcontroller it is easy to process the series of data and to calculate the measurand.
In figure 12 two offset capacitors C_{01} and C_{02} have been indicated. The offset capacitor C_{off1} is used to realize time intervals for capacitors $C_{s,1}$ - $C_{s,4}$ to sample the voltage to be measured. Together with the comparator and conrolled switches a charge-controlled relaxation oscillator is formed which linearly converts the voltages into periods of the oscillator output signal. The second offset capacitor C_{off2} is needed to enable negative values of the thermocouple voltages. More details concerning the applied signal processing techniques in this circuit can be found elsewhere [13].

6. Conclusions

Especially, for medium and low-volume markets, universal sensor interfaces enable rapid prototyping of low-cost, high-performance sensor systems. For easy readout by a microcontroller these interfaces should provide output signal which are digital or a time-modulated. For use in universal interfaces time-modulated output signals have the attractive features of simplicity, flexibility for the user and low-power consumption. The typical drawbacks of CMOS technology, such as $1/f$ noise and component mismatching can be overcome by applying auto-calibration and dynamic element matching (DEM). The application of dedicated measurement techniques, including autocalibration, two-port measurement and advanced chopping, results in a high precision and reliability of the sensor interfaces. Applying synchronous detection, autocalibration and advanced chopping also yields high immunity for interfering signals, $1/f$ noise and parameter drift.

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