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An Intelligent ISFET Sensory System With Temperature and Drift Compensation for Long-Term Monitoring

D. Y. Chen and P. K. Chan

Abstract—This paper presents a new intelligent ISFET sensory system dedicated to a precision pH sensory function, as well as a long-term monitoring capability without being jeopardized by temperature and drift fluctuations in the water-quality monitoring environment. It includes the novel compensation technique for counteracting a simultaneous change of temperature and drift, the design of the sensory system incorporating hardware and software co-design for enhancing the performance stability of a standard ISFET device and a new programmable current source with reduced dynamic current error for a dedicated temperature compensation process. The proposed ISFET sensory system, with about a 0.01 pH resolution in system design at a 3.3 V supply, has been validated by the experiments, exhibiting a maximum accuracy error of 0.02 pH at 23 °C and 0.05 pH at 40 °C with dual compensation. The compensation results show a maximum time drift of 0.003 pH/hour (0.166 mV/hour) at 23 °C and an average temperature drift of 0.00049 pH/hour/°C (0.0245 mV/hour/°C) for a reference temperature increase from 23 °C to 40 °C, with the value of the pH solution ranging from 4 to 9 in six-hour measurements. These measured results outperform those of the reported drift reduction techniques, suggesting that the ISFET sensory system using novel compensation can provide significant immunity against temperature change, time drift, and temperature drift, which are favorable towards robust measurements in environmental monitoring applications.

Index Terms—Intelligent sensor, ISFET drift, ISFET interface circuit, ISFET temperature compensation, long-term monitoring, microcontroller sensor system, precision current mirror, water quality monitoring.

I. INTRODUCTION

THE REGULAR monitoring of the pH value of water resources and sewage discharge for potential water pollution is a necessary task for an overdeveloped environment. The normal pH range for surface water systems is between 6.5 and 8.5 and is between 6 and 8.5 for ground water systems. Water

with a pH lower than 6.5 is acidic, soft or corrosive and contains elevated levels of toxic metals, thus introducing a health risk. Although water with a pH higher than 8.5, sometimes referred to as hard water, does not pose a health risk, it can cause aesthetic problems [1].

The invention of ISFETs (Ion Sensitive FETs) [2] has allowed for the construction of relatively cheap portable pH measurement instruments [3]. With further advancement in semiconductor technology, ISFET sensors have emerged as a standard. Despite the fact that the ISFET sensor has been developed for over 30 years [4], several effects of ISFET sensors remain that are not ideal. The usual phenomena are fluctuation with time and fluctuation with temperature. The first phenomenon is a result of the drift mechanism [5], as well as slow pH response [6]–[8], which poses challenges in maintaining the long-term stability of standard ISFET devices. The second phenomenon is a result of the pH dependent temperature coefficient [9], [11]–[13] and nonlinear temperature dependent mobility in MOS part of ISFET [17], [24], which increase the difficulty of thermal compensation. Furthermore, it has been reported that the ISFET drift rate has an exponential incremental tendency, with increasing pH value, as well as temperature value [25]. All these undesirable properties severely affect the accuracy of the ISFET sensor in its long-term monitoring applications. With reference to the fundamental studies [6] as well as taking into account the ISFET temperature dependence, the ISFET output signal variation, ΔV can be generalized in this form

$$\Delta V = \frac{\partial V}{\partial \text{pH}} \Delta \text{pH} \left[1 - \sum_{i=1}^n \varepsilon_i \exp\left(-\frac{t}{\tau_i}\right) \right] + \frac{\partial V}{\partial t} dt + \frac{\partial V}{\partial T} dT \quad (1)$$

where $\partial V/\partial \text{pH}$ is the pH sensitivity of the ISFET. ΔpH is the amplitude of the pH-step signal introduced. ε_i and τ_i are the normalized amplitude and time constant of the corresponding exponential term i . From (1), it can be observed that the time-variant variables come from: (i) the bracket term that represents the slow pH response of the ISFET sensor and (ii) the drift term $(\partial V/\partial t) \times dt$ that represents the change of output with time. Turning to the time-invariant term $(\partial V/\partial T) \times dT$, this denotes the change in output with temperature.

Very little research has been conducted on drift compensation techniques such as eliminating drift as a common mode signal by using an ISFET/REFET pair [14], neglecting the drift over a short-time interval of several seconds due to a relatively longer time constant of drift in the order of several hours [15] and employment of negative feedback in the electronic path [16]. Unfortunately, all these methods require a thermo-stable

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condition to avoid the temperature dependence of ISFET. For a broad range of pH value, the temperature compensation becomes more difficult because the temperature coefficient ($T.C.$) of the electrochemical component is a function of pH value. As a result, the composite time-invariant and time-variant effects greatly limit the performance of ISFET sensors, leading to unacceptable results in critical measurements. It is even more challenge for ISFET to operate in a wide pH range environment. Therefore, the main objective of this paper is to propose an ISFET-based pH measurement system with dedicated hardware design and software algorithms that not only provides a precise pH sensing function, but that also performs smart compensation simultaneously in terms of time and temperature so as to allow for long-term monitoring applications in a low-cost solution.

The smart compensation is processed by two approaches: (i) It makes use of recent work [17] on a novel dynamic biasing current temperature compensation technique to provide optimum biasing current for ISFET dynamically, thus resulting in an almost zero temperature coefficient ($Z.T.C.$) for nonlinear biasing point at different pH values. (ii) It makes use of a proposed time-variant compensation algorithm to control the hardware that periodically samples the output voltage of ISFET readout circuit in intervals of several seconds such that the relatively smaller value of the time-variant components can be distinguished from the pH signal component, $(\partial V/\partial \text{pH}) \times d\text{pH}$. The value of the time-variant component is then recorded and stored in the memory, and will be subsequently utilized to adjust the foundation calibration data in the time domain, facilitating the time-variant compensation in long-term measurements.

Following the introduction, this paper is organized as follows. In Section II, the ISFET sensor characteristics pertaining to temperature-based time-invariant parameters and time-variant parameters are presented. In Section III, the microcontroller-based ISFET sensory system together with its interface integrated circuit (IC) involving a new programmable current mirror with reduced dynamic error is described in detail. Section IV deals with the algorithms relating to the operation of the ISFET sensory system in a smart sensor environment. In Section V, the measurement results pertaining to the proposed current mirror structure as well as the proposed intelligent ISFET sensory system are discussed.

II. ISFET SENSOR CHARACTERISTICS

The ISFET sensor begins with a unified temperature-dependent drain current expression in saturation region, which is derived from the electrochemical theory and semiconductor theory of MOS transistor, with an indication of the key process model parameters and their definitions. Following this, the time-invariant thermal characteristic of ISFET and the time-variant behaviors, including drift and slow pH response, are described to provide useful insight as well as to understand their impact on the ISFET sensing device. This also reveals the design challenge for smart compensation in a portable instrument system [3].

A. ISFET in Saturation Region

The operation mechanism of ISFET has been studied in [4], [18]–[21]. Like the standard MOSFET, for a long-channel

ISFET operating in a saturation region, the drain current can be expressed as

$$I_{ds} = \frac{\mu C_{OX} W}{2L} (V_{gs} - V_{th(\text{ISFET})})^2 \quad (2)$$

where μC_{OX} is the process transconductance parameter, W/L is the aspect ratio and the ISFET threshold voltage [4] combined with the MOS transistor counterpart which includes body effect in a four-terminal device [22] can be unified as

$$V_{th(\text{ISFET})} = V_{th_chem} + V_{th_mos} \quad (3)$$

or

$$V_{th(\text{ISFET})} = E_{\text{Ref}} + \Delta\phi^{lj} - \Psi_{eo} + \chi^{sol} + \left[\frac{-\Phi_{Si}}{q} - \frac{Q_{OX} + Q_{SS} + Q_B}{C_{OX}} + 2\varphi_f + \gamma(\sqrt{2\varphi_f + V_{SB}} - \sqrt{2\varphi_f}) \right] \quad (4)$$

with

$$V_{th_chem} = E_{\text{Ref}} + \Delta\phi^{lj} - \Psi_{eo} + \chi^{sol} \quad (5)$$

and

$$V_{th_mos} = \frac{-\Phi_{Si}}{q} - \frac{Q_{OX} + Q_{SS} + Q_B}{C_{OX}} + 2\varphi_f + \gamma(\sqrt{2\varphi_f + V_{SB}} - \sqrt{2\varphi_f}) \quad (6)$$

where V_{th_chem} denotes the chemical threshold voltage of ISFET. E_{ref} is the potential of the Ag/AgCl reference electrode. $\Delta\phi^{lj}$ is the potential drop between the reference electrode and the solution, which typically has a value of 3 mV [9]. As this potential is pH-independent, it can be viewed as a common-mode input signal for an ISFET interface circuit in any pH buffer solution and can be nullified during system calibration and measurement procedures. χ^{sol} , with a typical value of 50 mV [8], is the surface dipole potential of the solvent being independent of pH. ψ_{eo} is the only chemical input parameter that is responsible for ISFET pH sensitivity, which has already been explained by Hal and Eijkel's theory [21] and is defined as the change of ψ_{eo} with respect to a change in the pH value of the solution as follows:

$$\frac{\partial \psi_{eo}}{\partial \text{pH}} = -2.303 \frac{kT}{q} \alpha \quad (7)$$

where α is a dimensionless sensitivity parameter with a value between 0 and 1, k is the Boltzmann constant, T is the absolute temperature and q is the magnitude of the electron charge. Regarding the ISFET-based MOS threshold voltage V_{th_mos} , the terms in (6) are similar to those of the standard MOSFET threshold voltage, except for the absence of the gate metal work function. In (6), the first three terms represent the threshold voltage at zero body-to-source bias, while the fourth denotes the presence of the body effect if $V_{SB} \neq 0$. It should also be mentioned that the n-type ISFET is mostly used for its low drift and high mobility properties [9], hence, in the application of the integrated microsystem where the ISFET sensor and interface

circuit are fabricated in the same die, the source-to-bulk potential of ISFET plays a crucial role in circuit performance when referencing the apparent body effect item in (4) and (6).

B. ISFET Temperature Characteristic

Several investigations [9], [11]–[13] have demonstrated that the ISFET exhibits thermal instability. The temperature characteristic of ISFET is a complex function pertaining to the reference electrode, electrolyte-insulator potential, and MOS transistor. It can be expressed as

$$T.C_{\text{total}} = T.C.R. + T.C.F. + T.C.I. \quad (8)$$

where $T.C_{\text{total}}$ is the total temperature coefficient of ISFET. $T.C.R.$ is the temperature coefficient of E_{ref} , which has been reported by [9] to have a typical value of 0.14 mV/°C. $T.C.F.$ is the temperature coefficient of the MOS structure of ISFET. Based on the device physics [22], [23], there are two main parameters that are responsible for the temperature characteristics of the MOS transistor: mobility and threshold voltage. For a given temperature, T_1 , they can be expressed as follows:

$$\mu(T)|_{T=T_1} = \mu(T_0)(T_1/T_0)^m \quad (9)$$

$$V_T(T)|_{T=T_1} = V_T(T_0) + K_{VT}(T_1 - T_0) \quad (10)$$

where T_0 is the reference temperature (note that room temperature is taken as the reference) and m is a constant, with its typical value ranging between -1.9 and -2.2 [24], depending on the doping concentrations of silicon. K_{VT} is the temperature coefficient of the threshold voltage, usually between -0.5 mV/°C and -3.0 mV/°C [22]. The third item in (8), $T.C.I.$, is the temperature coefficient of the electrolyte-insulator interface potential. Previous studies have shown that $T.C.I.$ is also a function of pH value [9], [11]–[13]. It may vary from 0.54 mV/°C to 1.1 mV/°C as the pH increases from 4 to 10 [10]. As described in an earlier section, $\Delta\phi^{lj}$, the junction potential between the reference electrode and solution is about 3 mV. Taking into account the temperature-dependent effect, the junction potential can be expressed as [11]

$$\Delta\phi^{lj}(T) = \Delta\phi^{lj}(T_0) + 10^{-5}(T - T_0). \quad (11)$$

For instance, the experimental data estimates that the contribution of the temperature effect to $\Delta\phi^{lj}$ is about 0.17 mV when the temperature increases from room temperature 23 °C to 40 °C, which gives an equivalent accuracy error of 0.0034 pH in a pH sensor having a sensitivity of about 50 mV/pH.

C. ISFET Time-Variant Effects

Apart from time-invariant thermal instability, the ISFET also suffers from time-variant effects, which include drift and slow pH response. Several investigations [5]–[8] have been carried out to characterize and model the behaviors of these two effects, providing useful insight into their mechanisms. The drift in Si_3N_4 -gate ISFET is characterized by a relatively slow, monotonic temporal change in the threshold voltage, which occurs over a period of several hours and is postulated to be associated with relatively slow conversion of the silicon nitride surface to a hydrated SiO_2 or oxynitride layer. This hydration causes the variation of effective insulator capacitance

due to different dielectric constants of the hydrated and nitride layers. According to Jamasb's physical model [4] on the drift characteristics, with reference to (4), the expression for the drift of ISFET threshold voltage can be expressed as

$$\Delta V_{th(\text{ISFET-Drift})}(t) = -(Q_{OX} + Q_{SS} + Q_B) \times \left(\frac{\varepsilon_n - \varepsilon_{HL}}{\varepsilon_n \varepsilon_{HL}} \right) x_{HL}(t) \quad (12)$$

where ε_n and ε_{HL} are the dielectric constants of the nitride and hydrated layers and x_{HL} is the thickness of the hydrated layer. It can be observed from (12) that the drift is directly proportional to the thickness of the hydrated layer. Although the value of drift is independent from the ISFET biasing condition [5], the drift rate of the ISFET device has an exponentially increasing tendency with increasing pH value or with increasing temperature value [25]. This further complicates the ISFET sensor compensation in terms of both temperature and drift influence at the same time.

Besides ISFET drift, the ISFET slow pH response [6], which is about 3%–7% of the total pH response, is another time-variant parameter. It is postulated that this slow response behavior can be interpreted in terms of a surface model [7], in which a portion of the surface sites located under the surface react very slowly for the diffusion of ions in the insulator to the site. This can be very slow, even for small distances. The behavior of this slow response effect has been modeled on the first bracket term of (1). Contrary to the drift effect, the amplitude of ISFET slow pH response as well as its polarity is also determined by the amplitude and polarity of input pH-step signal, ΔpH [6], [16]. Although these two time-variant nonideals of ISFET are triggered by different mechanisms, the variation on the ISFET output caused by these two effects overlaps in the time domain. For instance, if there is a negative step in the input pH signal, the variation on ISFET output contributed by a slow pH response is negative, which will counteract with that contributed by the drift effect. Using the ISFET readout circuit described in Section III, the output signal that is essentially equivalent to gate-to-source voltage, V_{GS} of ISFET, would be the appropriate signal to demonstrate the drift and slow pH response of ISFET.

III. ISFET SENSORY SYSTEM

This section provides an overall description of the proposed microcontroller-based ISFET sensory system. This includes the sensory system architecture, the readout IC with novel wide dynamic current biasing circuit, and the microcontroller system.

A. Sensory System Architecture

Dedicated hardware on the basis of the flexible programming feature of the microcontroller serves as an automation platform for the signal translation of ISFET interface IC, numerical computation, data storage, implementation of the smart compensation algorithms and communication with peripherals for smart sensor signal processing. The system block diagram is depicted in Fig. 1, which consists of an improved ISFET interface IC incorporating a recently published ISFET readout circuit [17] with a new programmable wide dynamic biasing current circuit, a 12-bit Digital-to-Analog Converter IC (TLV5616), a 16-bit

Fig. 1. Block diagram of the microcontroller-based pH measurement system.

Fig. 2. Schematic diagram of ISFET interface circuit (R_d and R_s are ISFET intrinsic drain and source diffusion resistors).

microcontroller (MSP430F149) with built-in RS232 facilities and an LCD display module. Note that the ISFET interface IC has been fabricated using TSMC 0.25 μm CMOS process technology. The whole system is powered with a single 3.3 V supply.

B. ISFET Readout Circuit With Wide Dynamic Biasing Current Circuit

The ISFET sensor usually operates with an analog readout circuit to translate the pH physical variable to the voltage domain for sensor signal processing. It is desirable for the circuit to exhibit good sensitivity, good linearity, insignificant bulk modulation, low circuit temperature dependence, and ease of compensation. A tremendous number of interface circuits [17], [26]–[32] have been reported in the literature. The ISFET sensory system design strategies follow the desirable goals and extend the recent interface circuit [17] to include the replacement of a conventional current source with a new programmable wide dynamic biasing current circuit (dotted box) shown in Fig. 2. Such an arrangement supports the long-term monitoring tasks required to meet the precision objective.

The ISFET readout circuit aims at detecting the change of the ISFET threshold voltage that reflects the variation of the pH value in the solution. The ISFET working in the region of saturation is biased by I_{B2} . According to (2) and (4) and taking into

account the ISFET intrinsic drain and source diffusion resistors R_d and R_s , as shown in Fig. 2, the output of the interface circuit can be derived as

$$V_{\text{out}} = V_{gs} = V_{th(\text{ISFET})} + \sqrt{\frac{2I_{B2}}{\mu_n C_{ox}(W/L)}} + I_{B2}R_s \quad (13)$$

where the second term is defined as a quiescent dc biasing voltage arising from the gate-overdrive bias and the third term is the voltage drop across the source diffusion resistor, R_s . The voltage across R_s will only introduce a DC offset in the output voltage of the ISFET readout circuit, which is regarded as a common-mode signal and which will be nullified during calibration. Therefore, the output V_{out} becomes the only function of the threshold voltage of ISFET, $V_{th(\text{ISFET})}$, which is linearly proportional to the pH value of the solution. In grounding ISFET's source and bulk terminals, the body effect dc does not alter sensitivity but the temperature-dependent body effect contributes to the total $T.C.$ of temperature uncompensated output by a few percentage points [17], which results in an estimated accuracy error of about 0.005 pH for a temperature change from a room temperature of 23 $^{\circ}\text{C}$ –40 $^{\circ}\text{C}$ using the typical process and design parameters in calculations. In water quality monitoring applications, this temperature-based error is small and tolerable.

Based on the output expression of ISFET readout circuit, another design consideration of the ISFET interface circuit involves recognizing a robust biasing circuit that provides necessary ISFET biasing current I_{B2} to support the dynamic current temperature compensation algorithm in a precise manner. Hence, the circuit technique for improving the precision of the current mirror under the dynamic biasing condition is required.

With reference to the simple CMOS current mirror, it has the advantage of wide operational headroom but displays in low-resolution due to potential poor output impedance (dependent upon the choice of channel length) and contributes significant systematic error arising from different V_{DS} values across the input and output transistor of the current mirror. One way to increase the output impedance of the current mirror is to employ the cascode structure. However, the conventional cascode current mirror utilizes a minimum input voltage of $2V_{GS}$, which hinders its application in the proposed ISFET interface circuit. Although one diode-connected transistor in the input branch can be eliminated to attain the minimum input voltage of one V_{GS} , the constant gate biasing voltage applied to the output cascode transistor can only support high precision for a small range of operating current. Another method for achieving output impedance enhancement is through the regulated-cascode circuit [33] or active-input regulated-cascode circuit [34]. However, the V_{DS} across the output transistor will deviate significantly from the V_{DS} across the input transistor under dynamic biasing operation, introducing dynamic systematic error in a wide input current range. Besides, the active-input regulated-cascode scheme [34] requires two amplifiers that incur extra complexity, power and silicon area.

In order to meet precision requirements, a new V_{DS} -tracked regulated-cascode current mirror topology, consisting of an operational transconductance amplifier (OTA1) [35], a cascode

transistor M3 and a current mirror transistor pair M4–M5, is proposed in the biasing circuit (dotted box), as shown in Fig. 2. The biasing circuit also includes a voltage-to-current (VI) converter comprising an OTA2, a regulated transistor M2 and an external reference resistor R_B . The biasing current generated by the VI converter is copied to the ISFET device by the V_{DS} -tracked regulated-cascode current mirror. The input transistor M5 is diode-configured, while the OTA1 is responsible for regulating the gate voltage of M3 to track the V_{DS} of M4 and M5. Turning to the other part of the circuit design, the OTA2 in association with M2 clamps the voltage across R_B by the output voltage of the digital-to-analog converter (DAC), V_{DA} , which is programmed by the microcontroller. The output of the OTA2 regulates the gate voltage of M2 such that the biasing current is defined as follows:

$$I_{B1} = \frac{V_{DA}}{R_B}. \quad (14)$$

However, there are unavoidable static offsets generated by the dc offset of OTA2 (typically, a few mV in value) and tolerance of external reference resistor R_B (typical of 1%). These random errors are regarded as static systematic errors that will not be harmful to the overall system accuracy because they can be nullified during system calibration and measurement procedures. It turns out that the dynamic systematic error arising from a wide programmable biasing current becomes one of the most critical design issues. Another dynamic error source coming from the DAC chip can also affect the precision. An accurate DAC chip is therefore required in the design. Finally, the proposed current mirror structure solves the problem arising from the dynamic systematic error.

C. Microcontroller System

The microcontroller in association with its peripherals plays an important role in the intelligent ISFET sensory system. The microcontroller is not only responsible for sampling and converting the sensor output signal to the digital domain for further digital signal processing but is also responsible for implementing the proposed drift and temperature compensation algorithms in the ISFET sensor. The microcontroller used in the system is based on the Texas Instruments MSP430 series, which is an ultra-low-power 16-bit microcontroller family that comprises several devices featuring different sets of modules targeted at various applications [36]. It is designed to be battery operated for use in extended-time applications. The MSP430F149 microcontroller consists of two built-in 16-bit timers, a fast 12-bit A/D converter, two universal serial synchronous/asynchronous communication interfaces (USART), 16×16 bits hardware multiplier and 48 I/O pins. One of the major advantages of the MSP430F149 is that it has 16 registers to reduce execution time. This reduces the register-to-register operation time to one cycle of the processor frequency. Another major advantage is that the 16×16 bit hardware multiplier is not integrated into the CPU, thus requiring no special instructions and operating independent of the CPU, which enhances the performance and offers a broad code and hardware-compatible family solution. As shown in Fig. 1, the output signal of the ISFET readout circuit is sampled by

the built-in 12-bit A/D converter and converted to the digital domain. Then, the digitalized voltage signal is processed by the microcontroller and converted into the corresponding pH value, which is transmitted to TBM12864-64A 128×64 dot matrix LCD for a display of the results. Meanwhile, the measurement data can be transmitted to a personal computer (PC) via the RS232 port. As will be described in next section, during the implementation of the ISFET dynamic current temperature compensation technique [17], the biasing current of the ISFET readout circuit is digitally controlled by the microcontroller via a 12-bit Digital-to-Analog Converter (DAC), TLV5616. The TLV5616 is a voltage output DAC [37] with a flexible four-wire serial interface, which allows for a glue-less interface with MSP430F149's serial peripheral interface (SPI).

IV. MULTIPARAMETERS COMPENSATION

For reliable monitoring, the essential feature of the pH measurement system is to support multiparameters compensation simultaneously. A smart compensation algorithm is proposed in this section to achieve the objective.

A. Software Algorithm of Proposed pH Measurement System

Dedicated software has been developed for the microcontroller hardware system depicted in Fig. 1 to control the whole procedure of pH measurement, including calibration of the ISFET sensor in standard pH buffer solutions, nonlinear ISFET temperature compensation at different pH values, ISFET drift and slow pH response compensation, transmission of the measurement results to an LCD display module, and via RS232 port to a PC. Both the C and assemble language are employed in the software programming process.

Fig. 3 depicts a flow chart of the application-specific software programming for the pH measurement system. As described in the flow chart, the software design allows the user to select the option of calibration of the ISFET sensor in the standard pH4, pH7, and pH9 buffer solutions at room temperature followed by an update of the calibration data stored in the memory or skip the calibration procedure but utilize the previous calibration data read from the memory. During the measurement procedure, regarding the temperature and drift compensation, there are four options for ISFET pH measurement: (i) pH measurement with only temperature compensation; (ii) pH measurement with both temperature and drift compensation; (iii) pH measurement with only drift compensation; and (iv) pH measurement without temperature and drift compensation. For the option involving both temperature and drift compensation, the temperature compensation is carried out prior to drift compensation, during which the ISFET sensor is biased by the optimum biasing current that forces the ISFET to operate at the athermal point. After temperature compensation, the timer in the microcontroller is set to 2 s, meaning that the output of the ISFET readout circuit will be sampled by the built-in ADC every 2 s and the difference of every two successive sampled output voltages will be compared with a detection threshold. The criteria for choosing an optimum detection threshold is based on the fundamental error sources arising from a practical experimental environment, given static parameters such as resolution, differential nonlinearity (DNL) and integral nonlinearity (INL) in the embedded ADC of the

Fig. 3. Flow chart of software programming.

microcontroller, static errors of the DAC chip, and the noise plus other potential fluctuations in the ISFET interface output voltage. If the differential value of the two successive sampled output voltages calculated by the controller's computation resource is more than the defined threshold, it can be regarded as a change in solution pH value, and hence the ISFET dynamic current temperature compensation will be carried out again. On the contrary, if the differential voltage is less than the threshold, it is treated as a random fluctuation caused by ISFET drift and slow pH response. Such differential value is then accumulated with the old $\Delta V_{th_time}(t)$ data stored in the memory for the following compensation. At this juncture, the pH value can be converted from the measured ISFET output voltage by employing the drift-compensated equations that realize time-variant compensation algorithm.

B. Revisiting the ISFET Dynamic Current Temperature Compensation Technique

As suggested in (1), the accuracy of ISFET can be significantly impeded by time-invariant temperature parameters as well as other time-variant parameters at the same time. In order to effectively compensate for the composite effect, the measurement strategy begins with the dynamic current temperature compensation [17] prior to time-variant compensation. This is mainly because the temperature compensation only requires

milliseconds to accomplish the task when compared with the drift phenomenon, which occurs in the order of minutes.

The temperature compensation technique stems from the physical theory of mutual compensation of mobility and threshold voltage temperature variation to achieve the *Z.T.C.* biasing point in MOSFET by choosing an appropriate value of the biasing current. Therefore, for a given pH buffer solution, the optimum biasing current that forces ISFET to operate at the athermal point can be derived as

$$I_{DS(pH)} = \left(K_{VT(pH7)} + 2.303 \frac{k}{T} \alpha \times \Delta pH \right)^2 T_0^2 \frac{2}{m^2} \times \left(\frac{T_1}{T_0} \right)^{m+2} \frac{\beta_0}{1 + \theta [V_{GS(pH)}(T_0) - V_{th(ISFET)}(T_0)]} \quad (15)$$

where α is a dimensionless sensitivity parameter, which has been explained in (7). $\Delta pH = pH - pH7$ is the pH difference between a given pH buffer solution and reference pH7 buffer solution, and the other symbols are defined and explained in [17]. To overcome the obstacle that the value of ΔpH cannot be obtained directly in a practical pH measurement, a numerical iteration technique has been proposed in [17]. In this iteration method, the $I_{DS(pH7)}$ derived in [17] is treated as an initial biasing current and the measured output voltage of the ISFET interface circuit in Fig. 2 is converted to the corresponding pH value, pH_C by a microcontroller using a computational equation as follows:

$$pH_C = pH4 + \frac{V_{out} - V_{Calib(pH4)}}{(V_{Calib(pH7)} - V_{Calib(pH4)}) / (pH7 - pH4)} \quad (16)$$

if $V_{out} < V_{Calib(pH7)}$ or else

$$pH_C = pH7 + \frac{V_{out} - V_{Calib(pH7)}}{(V_{Calib(pH9)} - V_{Calib(pH7)}) / (pH9 - pH7)} \quad (17)$$

where pH_C is the pH value being converted from the measured output voltage of ISFET. $V_{Calib(pH4)}$, $V_{Calib(pH7)}$, and $V_{Calib(pH9)}$ are the corresponding output voltages of ISFET at pH4, pH7, and pH9 input standard buffer solutions, respectively. Then, ΔpH is calculated from $\Delta pH = pH_C - pH7$, which is substituted into (15) again to calculate another new biasing current. The above process is repeated until the convergence of pH_C from successive iterations occurs. By using this iterative method, the optimum biasing current to achieve *Z.T.C.* of ISFET at different pH values can then be nonlinearly found. When biased by the optimum biasing current, the ISFET operating in the *Z.T.C.* point is almost insensitive to temperature variation. Therefore, (1) becomes

$$\Delta V = \frac{\partial V}{\partial pH} \Delta pH \left[1 - \sum_{i=1}^n \varepsilon_i \exp\left(-\frac{t}{\tau_i}\right) \right] + \frac{\partial V}{\partial T} dt + \left(\frac{\partial V}{\partial T} dT - \Delta V'_{temp} \right). \quad (18)$$

It can be seen that the differential of the ISFET output signal, ΔV , is now contributed by input pH signal variation, ISFET slow pH response, and drift because the temperature compensated term $\Delta V'_{temp}$ generated by the dynamic biasing current technique tends to counteract the temperature dependent term

$(\partial V/\partial T)dT$. The time-variant compensation is the focus of Sections IV-C and IV-D.

C. Review of ISFET Time-Variant Compensation Techniques

Much research effort has been expended on various ISFET drift and slow pH response compensation techniques [14]–[16]. In earlier work [15], the relatively small instantaneous drift rate in ISFETs was used to compensate for the drift signal superimposed on the sensor output. This method is based on the fact that the time constant characterizing drift is in the order of several hours, which implies that over an individual sampling time of several seconds, the variation in the ISFET output signal contributed by drift can be negligible when compared to that contributed by a change in the pH input signal. However, the main limitation of this method is that of the thermostable measurement setup required to avoid the influence of ambient temperature variation due to the strong temperature dependence of the ISFET device. Other research [16] has employed negative feedback in the electronic path to overcome the long-term drifts. This method is based on the fact that if the drain-to-source current (I_{DS}) of ISFET is held constant, an increase in drain-to-source voltage (V_{DS}) produces a decrease in gate-to-source voltage (V_{GS}) over a constant pH value. Thus, the variation of V_{GS} of ISFET caused by a drift in the time domain can be counteracted by regulating the V_{DS} , which is controlled by the electronic conditioning circuit. However, this method can only be effective when the temperature is kept constant. Since the pH sensitivity of the ISFET device ($\partial V/\partial \text{pH}$) is strongly temperature-dependent, the relationship between input (pH value) and output (voltage) of the ISFET sensor may not be accurate across different temperatures.

D. Proposed ISFET Time-Variant Compensation Technique

As described in Section II, both ISFET drift and slow pH response are time-variant parameters. Although arising from independent physicochemical processes, they both cause instability in the dc operating point of ISFET as a time-dependent ISFET threshold voltage, which can be quantitatively represented as $\Delta V_{th(\text{ISFET_Drift})}(t)$ and $\Delta V_{th(\text{ISFET_Slow})}(t)$, respectively. These two effects on time-dependent ISFET threshold voltage overlap in the time domain can be functionally merged and represented by one lumped parameter $\Delta V_{th_time}(t)$

$$\Delta V_{th_time}(t) = \Delta V_{th(\text{ISFET_Drift})}(t) + \Delta V_{th(\text{ISFET_Slow})}(t). \quad (19)$$

The first term in (19), which has been postulated to be associated with a slow and temporal change in the overall gate insulator capacitance, will always occur in the positive direction [4], and the second term would depend on the positive or negative pH step produced [5], [6], [16]. Therefore, the overall time-variant component, $\Delta V_{th_time}(t)$, could be in a positive or negative direction.

The basic concept of the proposed ISFET time-variant compensation technique is to keep the relationship of the ISFET input signal (pH value)-output signal (voltage) at a certain level of accuracy by correcting the calibration data stored in the memory by the parameter, $\Delta V_{th_time}(t)$ on the basis of the predicted result. In case of long-term pH monitoring, the output voltage of the readout circuit that is the gate-source voltage of ISFET will deviate by an amount of $\Delta V_{th_time}(t)$ caused by the deviation of ISFET threshold voltage through ISFET drift and slow pH response. Therefore, the output voltage of the readout circuit cannot be converted to pH value precisely if referenced to the original calibration data. One way to correct this error is to recalibrate the ISFET in standard pH buffer solutions to update the relationship between the input pH signal and output voltage value. However, frequent calibration of the ISFET sensor is not only costly but also unpractical for long-term monitoring purposes. In this proposed ISFET time-variant compensation technique, the calibrated data is automatically updated in real time using the microcontroller hardware and software resource in the process of data acquisition and signal processing in the digital domain. Therefore, formula (16) and (17) can be modified, as shown in (20) and (21) at the bottom of page, where $V_{out}(t)$ is the measured output voltage of the ISFET readout circuit in the time-domain. It can be observed that the deviation of $V_{out}(t)$ in the time-domain by the amount of $\Delta V_{th_time}(t)$ can be compensated by updated calibration data, which indicates that the output voltage is able to be converted to pH value, $\text{pH}_C(t)$ precisely.

The key point of achieving this goal is to precisely detect the value of the time-variant component, $\Delta V_{th_time}(t)$. According to the models [5], [6] for time-variant nonideals, the time constants characterizing drift and slow pH response are in the order of several hours and several minutes, respectively, while the intrinsic pH response time for an ISFET is in the order of milliseconds [5]. Therefore, it is easy to distinguish the time-variation component, $\Delta V_{th_time}(t)$ from the solution pH-variation component, $(\partial V/\partial \text{pH}) \times \Delta \text{pH}$ by periodically sampling the output voltage of the readout circuit over a time interval of a few sec-

$$\text{pH}_C(t) = \text{pH4} + \frac{V_{out}(t) - (V_{Calib(\text{pH4})} + \Delta V_{th_time}(t))}{[(V_{Calib(\text{pH7})} + \Delta V_{th_time}(t)) - (V_{Calib(\text{pH4})} + \Delta V_{th_time}(t))] / (\text{pH7} - \text{pH4})} \quad (20)$$

$$\text{pH}_C(t) = \text{pH7} + \frac{V_{out}(t) - (V_{Calib(\text{pH7})} + \Delta V_{th_time}(t))}{[(V_{Calib(\text{pH9})} + \Delta V_{th_time}(t)) - (V_{Calib(\text{pH7})} + \Delta V_{th_time}(t))] / (\text{pH9} - \text{pH7})} \quad (21)$$

onds and comparing two successive samples. Considering the embedded 12-bit ADC with ± 1 LSB differential (DNL) and ± 1 LSB integral nonlinearity (INL) [36], the worst nonlinearity could lead to ± 2 LSB, which is equivalent to ± 1 mV with an ADC reference voltage of 1.8 V. Therefore, maximum fluctuation on the converter results due to DNL and INL could lead to 2 mVpp (peak-to-peak). Further considering other potential noise coupling from the digital blocks of the microcontroller, the threshold detection voltage is chosen as 2.7 mV under careful design of the printed circuit board (PCB), which is equivalent to a pH of 0.05. If the differential voltage between every two successive samples is less than the detection threshold, it is treated as a time-variant component and will be accumulated with the old time-invariant component stored in the memory in the time-domain. Therefore, the time-variant component defined in (19) is approximately evaluated by this accumulated value, $\Delta V_{th_time}^*(t)$, which can be expressed as follows:

$$\Delta V_{th_time}(t) \approx \Delta V_{th_time}^*(t) = \sum_{n=1}^{\infty} V_{out}[(n+1)t] - V_{out}(nt) \quad (22)$$

Alternatively, if a differential voltage between two successive samples is more than the detection threshold, it can be viewed as a change in the pH value of the solution, which will trigger the ISFET dynamic current temperature compensation as shown in the flow chart of software programming in Fig. 3. Therefore, the integral summary can precisely reflect the actual time-variation components, $\Delta V_{th_time}(t)$, and is thus employed to correct the initial calibration data when converting the output voltage of ISFET to pH value based on the (20) and (21).

As a final remark, the proposed time-variant compensation technique can operate in a temperature fluctuation environment when incorporating nonlinear temperature compensation for wide pH sensing. This overcomes the conventional long-term drift compensation techniques that rely on constant temperature constraint.

V. MEASURED RESULTS AND DISCUSSIONS

This section is concerned with evaluating the accuracy of the proposed current mirror structure using the fabricated interface IC, test equipment setup, testing the functionality of the microcontroller-based ISFET pH measurement system and validating the effectiveness of a novel compensation technique for ISFET drift and slow pH response at two test temperatures. Finally, a comparison with previously published ISFET drift reduction techniques is conducted and discussed.

A. Measured Results of V_{DS} -Tracked Regulated-Cascode Current Mirror

The ISFET interface circuit has been fabricated in TSMC CMOS 0.25 μm process technology, with the micrograph shown in Fig. 4. Prior to verifying the functionality of the entire pH measurement system, the proposed V_{DS} -tracked regulated-cascode current mirror on the basis of Fig. 2 was tested. The experimental dc characteristic of the current mirror were measured by sweeping its input current I_{B1} for two decades, ranging from 10 μA to 1 mA. The error of the current mirror within the measured current range is illustrated in Fig. 5. Based on the ISFET

Fig. 4. Micrograph of the integrated ISFET interface circuit.

Fig. 5. Error of the proposed current mirror structure at different input currents.

dynamic current temperature compensation technique in this design, the ISFET operating current range is approximately between 100 and 400 μA . The measured results in Fig. 5 have shown that the error of the current mirror is less than 0.5 μA . The results are reasonably good because the channel length employed for the current mirror transistor pair is only 2 μm . A better matching characteristic can be obtained if the channel length is further increased at the expense of the increased silicon area.

In terms of the ISFET temperature compensation methodology, some of the important parameters in (15) have been extracted in [17], and the values of those parameters are repeated here: $V_{th(\text{pH}4)} = -244.413$ mV, $V_{th(\text{pH}7)} = -87.6536$ mV, $V_{th(\text{pH}9)} = -21.8782$ mV, $\beta_0 = 0.001357$ A/V², $\theta = 0.0821$ V⁻¹, and $m = -2.2$. With reference to the ISFET gate-source voltage expression [17]

$$V_{GS(\text{pH})}(T)|_{T=T_1} \approx V_{th(\text{pH})}(T_0) + K_{VT(\text{pH}7)} \times (T_1 - T_0) + \sqrt{\frac{2I_{DS}}{1 + \theta \left[\frac{\beta_0}{V_{GS(\text{pH})}(T_1) - V_{th(\text{pH}7)}(T_1)} \right] \left(\frac{T_1}{T_0} \right)^m}} \quad (23)$$

Fig. 6. Experiment setup for the microcontroller-based ISFET pH measurement system.

In the presence of biasing current error ΔI_{DS} , the ISFET gate-source voltage can be shown as

$$V_{GS(pH)}^*(T)|_{T=T_1} \approx V_{th(pH)}(T_0) + K_{VT(pH7)} \times (T_1 - T_0) + \sqrt{\frac{2(I_{DS} + \Delta I_{DS})}{\frac{\beta_0}{1 + \theta[V_{GS(pH)}(T_1) - V_{th(pH7)}(T_1)]} \left(\frac{T_1}{T_0}\right)^m}}. \quad (24)$$

As can be observed from Fig. 5, in the ISFET operating current ranging from 100 to 400 μA , the largest current mismatch appears at a biasing current of 400 μA . Hence, the ΔV_{GS} error can be calculated by setting the worst case condition as $I_{DS} = 400 \mu A$, $\Delta I_{DS} = 0.5 \mu A$, and $T_1 = (50 + 273)K$. Therefore, we have

$$\Delta V_{GS} = V_{GS(pH)}^*(T)|_{T=T_1} - V_{GS(pH)}(T)|_{T=T_1} \approx 0.56 \text{ mV}. \quad (25)$$

This value is approximately equal to a pH of 0.01 unit, which has negligible impact on the long-term monitoring applications, suggesting that the dynamic accuracy of the proposed current mirror is satisfactory.

B. pH Measurement Setup

The Si_3N_4 -gate pH-sensitive ISFET sensor is commercially available from D+T Microelectrónica, A.I.E (CNM), Spain. Fig. 6 shows the experimental setup that includes the evaluation board where the microcontroller-based pH measurement system is built. The whole system works with a single 3.3 V supply, having adequate headroom for the ISFET device to operate in the region of saturation. Meanwhile, a FLUKE 87 III TRUE RMS handheld multimeter was employed to monitor the ISFET drain-source biasing current. The meter value was compared with the expected biasing current, which is calculated by the microcontroller in the process of running the ISFET dynamic current temperature compensation algorithm and displayed on the LCD module. Other measurement results, such as the pH value and the amount of $\Delta V_{th_time}(t)$ in (22),

TABLE I
LIST OF CURRENT CONSUMPTION OF EACH COMPONENT IN THE EVALUATION BOARD OF pH MEASUREMENT SYSTEM

Components	ISFET Interface Circuit Chip	REF3318 Voltage Reference	MSP430F149 Microcontroller	TLV5616 DAC	TBM12864-64A LCD
Current Consumption	780 μA	5 $\mu A \times 2$	340 μA	700 μA	2.5 mA

are also displayed on the LCD. At this juncture, the measured results were also transmitted via RS232 port to a PC and stored as a text file for further result analysis. The buffer solution where the ISFET was inserted was kept in a thermal buffer so as to control its temperature during the measurement process. Although the thermal buffer supported the measurement at room temperature or above, the method of ISFET drift and slow pH response compensation at various temperatures was still verified. The pH value of the solution was monitored by the *CyberScan 1000* automatic temperature compensation (ATC) pH meter, with a resolution of 0.01 pH. Table I lists the current consumption of each module in Fig. 1, the system is estimated to operate 739 hours continuously for two AA batteries having 3200 mA hr capacity. This is quite a sustainable long operation time when compared to existing instruments.

Although the LCD module employed in this prototype consumes significant power consumption, a low-power LCD module can be employed to save the whole system power substantially.

C. Measured Results of Proposed pH Measurement System at Room Temperature

To demonstrate the effectiveness of the proposed ISFET time-variant (ISFET drift and slow pH response) compensation technique at room temperature at thermal buffer, an experiment pertaining to long-term pH monitoring was carried out when the pH value of the solutions was changed in a step sequence of pH7-pH5.65-pH4-pH5.65-pH7-pH9 at 1 h intervals. The pH values with the symbol "X" denote the readings of the *CyberScan 1000* pH meter at defined time points in the pH measurement experiments. The *CyberScan 1000* pH meter measured each pH value twice, once at the beginning of the one hour interval and again at the end to monitor the stability of the solution across each time period being monitoring. Fig. 7 illustrates the comparison of the measured results based on an uncompensated scheme and various compensated schemes. It can be observed that both the ISFET drift and slow pH response can be compensated with long-term monitoring. It is interesting to observe that there exists some minute compensation when the solution's pH value changes from pH7 to pH4 but not when it moves from pH4 onwards. This confirms that the positive slope of V_{GS} of ISFET caused by the drift mechanism is intrinsically counteracted by the negative slope of V_{GS} caused by ISFET slow pH response in negative pH step [6], [16]. Meanwhile, it can also be observed that the measurement results with standalone ISFET drift compensation are almost identical to dual compensation. This is due to the fact that both the measurement and calibration are conducted under the same room temperature (23 $^{\circ}C$). Thus, the temperature sensitive parameters, such as $V_{th(ISFET)}$

TABLE II
COMPARISON OF MEASURED VALUES BETWEEN THE PROPOSED pH MEASUREMENT SYSTEM WITH
DUAL COMPENSATION AND THE *CyberScan 1000* pH METER AT 40 °C

Time Interval (Hours)	Reading of Proposed pH measurement System (pH unit)	Reading of <i>CyberScan 1000</i> pH meter (pH unit)	Difference (pH unit)
0 to 1	7.00	6.95	0.05
1 to 2	5.70	5.67	0.03
2 to 3	4.04	4.01	0.03
3 to 4	5.68	5.67	0.01
4 to 5	7.00	6.95	0.05
5 to 6	8.90	8.85	0.05

Fig. 7. Comparison of measured results with and without dual compensation at room temperature (23 °C).

Fig. 8. Comparison of measured results with and without dual compensation at a temperature of 40 °C.

and μ_n in measurement would not deviate from those in calibration.

D. Measured Results of Proposed pH Measurement System at Other Temperature

In order to further demonstrate the effectiveness of ISFET drift and the slow pH response compensation technique in the case of other temperature, the experiment was conducted at 40 °C in a thermal buffer depicted in Fig. 6. The pH values of the measured buffer solutions at 40 °C were pH6.95, pH5.67, pH4.01, and pH8.85, which were recorded by the *CyberScan 1000* pH meter with automatic temperature compensation. It is evident from Fig. 8 that the dual compensation scheme is far more superior to the ISFET sensing system without any compensation, and significantly better than that of the standalone drift compensation or that of the standalone temperature

compensation. Of the measurement data being jointed at periodic sample time points of 2 s under the smart compensation algorithm, almost straight lines are formed at different pH values across each duration of the monitoring period. The readings of the *CyberScan 1000* pH meter have been indicated with the symbol X, while the readings of the proposed pH measurement system, illustrated as the corresponding solid line in Fig. 8, are pH7.00, pH5.70, pH4.04, pH5.68, and pH8.90. It should be noted that the commercial *CyberScan 1000* pH meter only measures at any instant time point (to avoid the drift effect), with the purpose of providing accurate solution pH value for reference. A comparison of the readings between the proposed pH measurement system having dual compensation and the *CyberScan 1000* pH meter is summarized in Table II. Using the measured data in Figs. 7 and 8, it can be calculated that after dual compensation, the output of the proposed measurement system only produces a temperature drift of 0.00049 pH/hour/°C over a pH range of 4–9 at a temperature difference of (40 °C–23 °C). This shows that the proposed sensory system achieves respectful precision for long term monitoring condition even at different temperatures. This suggests that not only does the sensory system compensate the drift against time, but it also demonstrates good immunity against the temperature-dependent drift which is regarded as the technical merit of this work. The worst case accuracy of 0.05 pH unit meets the minimum requirement in water quality monitoring [40].

E. Comparison of Drift Performance With Prior-Art Works

The overall drift performance of the prior-art works and proposed work, in terms of time drift, temperature drift, pH range, incorporation of initial drift rate, and incorporation of slow pH response, are compared in Table III alongside a standard uncompensated ISFET device as a reference. The proposed sensory system improves the time drift by at least one order of magnitude with respect to an uncompensated ISFET device and about twice the reduction with respect to the analytical drift reduction technique. Besides, the temperature drift of this work is also one order magnitude lower than the instrumentation system drift reduction technique. More importantly, the computation of measured drift data also involves initial drift and slow response. This suggests that the sensory system can offer robust measurements because it takes a full range of drift data and handles slow response in view of a potential larger step change in pH value. It has validated that the proposed smart compensation technique

TABLE III
COMPARISON OF PROPOSED ISFET DUAL COMPENSATION TECHNIQUE WITH
PREVIOUSLY REPORTED DRIFT COMPENSATION TECHNIQUES

	Conventional ISFET [38]	Analytical Technique [15]	Instrumentation System [39]	Proposed ISFET Sensory System
Time Drift at room temperature	0.06 pH/hour	0.005 pH/hour	N.A.	0.003 pH/hour
Temperature Drift	0.024pH/hour/°C	N.A.	0.0033pH/hour/°C	0.00049pH/hour/°C
pH range	pH1 - pH7	pH3.5 - pH10.1	pH3.34 - pH7	pH4 - pH9
Temperature	15 °C - 45 °C	Room Temperature	20 °C - 26 °C	23 °C, 40 °C
Consider initial drift	No	No	Yes	Yes
Consider slow pH response	No	No	Yes	Yes

has the key advantages of low time drift and low-temperature drift.

VI. CONCLUSION

A microcontroller-based ISFET sensory system featuring a smart compensation algorithm has been realized in both hardware and software to alleviate the problems associated with the composite temperature-dependence and time-variant effect in a basic ISFET sensing element. A new programmable current mirror that supports a wide range of biasing current with reduced dynamic error in the ISFET interface IC has been presented. Incorporating the earlier-established nonlinear temperature compensation technique, the proposed time-invariant compensation technique enhances the stability performance of a standard ISFET sensor in the context of time drift and temperature drift for the long-term water quality monitoring environment in a low-cost solution. The experimental results have validated the circuit, system, and algorithms. The comparison has shown that the measured results are better than those of the prior art works. The accuracy and drift performance are adequate for environmental monitoring applications. The proposed compensation technique will be found useful for the improvement of commercial ISFET technology sensor products for long-term immersion. The proposed sensory system can be extended easily to detect different types of chemical ions in other chemical sensing systems. Therefore, the sensory system is particularly suitable for smart sensor.

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