THE OPERATION OF AN ISFET AS AN ELECTRONIC DEVICE

P BERGVELD

Afdeling der Electrotechniek, Technische Hogeschool Twente, Enschede, P O B 217
(The Netherlands)

1 Introduction

The conventional ion sensitive sensors, such as glass membrane electrodes, coated wire electrodes, metal–metal oxide electrodes, etc., convert a non-electrical quantity into an electrical quantity, which can then be measured with appropriate electronic circuitry connected in series. Unpredictable deviations in transducer parameters cannot be controlled in such a measuring system. Because the ISFET also has an electrical input besides the ion-sensitive input, this results in unique application possibilities, with consequences for the development of adequate electronic circuitry. In order to obtain an insight into this matter, it is useful first to understand the electrical behaviour of a MOSFET, from which the ISFET behaviour can be derived. Based on this theory some electronic design starting points can be developed for an appropriate ISFET application.

2. The MOSFET and ISFET small-signal behaviour

In first‐order MOS transistor theory, the equation for the drain current, \( I_d \), in the unsaturated region \( (V_d < V_g - V_t) \) is

\[
I_d = \beta [(V_g - V_t)V_d - \frac{1}{2} V_d^2]
\]

where \( \beta = \mu C_{ox} W/L \) is a geometry constant, \( V_g \) and \( V_d \) are the d.c. gate to source and drain to source voltages, respectively, and \( V_t \) is the threshold voltage which represents all the effects of substrate depletion charge, work function of gate metal, interface states, and fixed charges in the oxide.

The equation for the a.c. drain current, \( i_d \), is given by

\[
i_d = S V_g + \frac{1}{R_{ch}} V_d
\]

where \( V_g \) and \( V_d \) are the a.c. gate to source and drain to source voltages, respectively, \( S \) is the mutual conductance and \( R_{ch} \) is the differential channel resistance. Expressions for \( S \) and \( R_{ch} \) follow from differentiations

\[
S = \frac{dI_d}{dV_g|_{V_d=\text{const}}} = \beta V_d
\]
\[ \frac{1}{R_{ch}} = \left. \frac{dI_d}{dV_d} \right|_{V_g = \text{const}} = \beta [V_g - V_t - V_d] \quad (4) \]

The question arises whether these equations also hold for ISFETs or whether they must be modified. For an ISFET, \( V_g \) is kept constant and \( V_t \) contains the variable input signal, while in the case of a MOSFET, \( V_t \) is assumed to be constant and \( V_g \) is the variable.

Seen electronically, this difference in basis conception will give no complications, because the term \( V_g - V_t \) can be seen in both cases as the actual input variable.

Considering further the difference between a MOSFET and an ISFET, besides the change in input variable, the method of contacting the actual source and drain is also necessarily quite different. With a MOSFET the source and drain regions can, in principle, be completely evaporated with aluminum, which makes a very low resistive contact after alloying. By contrast, with ISFETs this contact method is impossible because the gate has to be contacted by an electrolyte, which means that no metal contacts can exist in the vicinity of this area. Depending on the necessary length of the insulating lacquers over the oxidized source and drain regions, the contact places are usually some millimeters away from the actual source- and drain-to-gate interfaces. The most common ISFET configuration is shown in Fig 1.

Using a donor concentration \( N_D = 5 \times 10^{19} \text{ cm}^{-3} \) for the source and drain regions, resulting in a square resistance \( R = 40 \Omega \), internal source and drain resistances are created with a value \( 40 \times l/w \Omega \), in which \( l/w \) is the length/width ratio of the diffusion region. In the case of the geometry as shown in Fig 1, the series resistance of source and drain will thus be \( 4 \times 40 \Omega = 160 \Omega \). In practical cases of needle-shaped ISFETs, this value is even larger (Esashi et al., 1978). This internal resistance of source and drain, of which the value depends, of course, on the actual geometry of the device, will never be zero, and gives rise to serious problems with regard to the device parameters as derived in eqns (3) and (4). This can easily be seen from the following calculations, based on the model given in Fig 2.

\[ V_d = V_{d's'} + I_d(R_s + R_d) \quad (5) \]
\[ V_g = V_{gs'} + I_dR_s = V_{gs'} + \frac{V_d - V_{d's'}}{2} \quad \text{(for } R_d = R_s) \quad (6) \]

Substituting for \( V_g \) and \( V_d \) of eqn (1), \( V_{gs'} \) (eqn (6)) and \( V_{d's'} \) (eqn (5)) respectively, gives

\[ I_d = \beta [(V_g - \frac{1}{2} V_d + \frac{1}{2} V_{d's'} - V_t) V_{d's'} - \frac{1}{2} V_{d's'}^2] \]
\[ = \beta [(V_g - \frac{1}{2} V_d - V_t)(V_d - I_d(R_d + R_s))] \quad (7) \]

From eqn (7) the equation can be derived for the drain current of a transistor with internal resistances

\[ I_d = \beta V_d \frac{V_g - \frac{1}{2} V_d - V_t}{1 + \beta (R_d + R_s)(V_g - \frac{1}{2} V_d - V_t)} \quad (8) \]
The expressions for the mutual conductance as well as the differential channel resistance, follow from differentiations

\[ S = \left. \frac{dI_d}{dV_g} \right|_{V_d=\text{const}} = \beta V_d \left[ 1 + \beta (R_d + R_s) (V_g - \frac{1}{2} V_d - V_t) \right]^{-2} \]  
(9)

\[ \frac{1}{R_{ch}} = \left. \frac{dI_d}{dV_d} \right|_{V_g=\text{const}} = \frac{\beta (V_g - \frac{1}{2} V_d - V_t)}{1 + \beta (R_d + R_s) (V_g - \frac{1}{2} V_d - V_t)} \]  
(10)

For \( R_s = R_d = 0 \), eqns (9) and (10) change into eqns (3) and (4). The influence of \( R_d \) and \( R_s \) on the channel resistance gives no electronic problems, because it only increases \( R_{ch} \). A decrease of \( 1/R_{ch} \) decreases the influence of \( V_d \) on \( I_d \) (see eqn (2)), which is an advantage. Problems, however, do arise from the influence of \( R_s \) and \( R_d \) on the mutual conductance, or, in other words, the sensitivity of the device, which decreases drastically for real values of \( R_s \) and \( R_d \). This effect is shown in Fig. 3 for a MOSFET with additional \( R_s \) and \( R_d \). In accordance with eqn (9) the sensitivity decreases more for larger values of \( (V_g - V_t) \), as can also be seen in Fig. 3.

The shape of the \( I_d-V_g \) curves published from ISFETs [1 - 3] can be fully explained as being the result of the internal source and drain resistances, the effect of which is very pronounced, especially for small values of \( V_d \) as shown in Fig. 4, and in accordance with eqn (9).
It can be concluded that, electronically seen, a MOSFET is a device in which the sensitivity depends only on the device parameter, $\beta$, and the applied voltage, $V_d$ (see eqn (3)), while an ISFET is a device in which the sensitivity is dependent, in addition, on the input signal $V_i = f(\text{pH} + \Delta \text{pH})$ and its bias $V_g$ (see eqn (9)). There are, in principle, two ways of solving this problem, namely, technologically and electronically. A technological way of
approaching the problem is to shorten the length of the source and drain diffusions by contacting the source and drain from the liquid-free side of the device. However, this implies new technologies because this technique is very unusual for standard transistor devices. The most promising solution is described by Cline et al. [4], while Zemel [5] shows that it can be used for gated diodes. It is, however, much easier to solve the problem of decreased sensitivity by means of an electronic circuit, which is insensitive to series resistors due to the application of the feedback principle.

3. Electronic circuit design adapted to ISFETs

From the curves of Fig 4 it can be seen that biasing an ISFET at a constant $I_d$, at the same time maintaining a constant $V_d$, can only be obtained by (automatic) control of $V_g$, compensating for a change in pH. This control action is mentioned in the previous Section by the proposed feedback system. This means, with regard to the small signal condition (eqn (2)), that $v_d$ is kept zero, making the measurement insensitive to the value of $R_{ch}$, while $v_d$ is also kept at zero by effectively controlling $v_g$ to zero, which makes the measurement independent of $S$ and thus of $R_d$ and $R_s$. The required condition can be obtained in two ways. The first possibility is control of $i_d = 0$ in a feedback loop due to automatic adjustment of the potential of the reference electrode, and thus of the liquid, in respect of the source and drain potentials. The second possibility is an automatic adjustment of the source and drain potentials with regard to a constant liquid potential, usually the ground potential. An example of the first possibility is the automatically balanced bridge circuit as shown in Fig 5.

A change in $V_t$ due to a pH variation is compensated by a change in $V_g$ via the reference electrode. A disadvantage of this system is that the liquid may not be grounded, which is sometimes required for the benefit of certain measuring conditions. Also, the possibility of accidental grounding has to be prevented. This problem can be solved by the use of isolation amplifiers and an isolated power supply. A further disadvantage of this bridge circuit is the asymmetrical impedance of the source and drain leads, making the system also under floating conditions, sensitive to interference from external electric fields and static electricity.

An example of the second circuit design, as mentioned above, is the source and drain follower concept, which is shown in Fig 6. In contrast to the bridge circuit shown in Fig 5, the liquid is now connected to the ground of the circuit by means of the reference electrode. The system consists essentially of a power supply (current source and adjustable reference voltage, $V_{ref}$), an instrumentation amplifier system, and an operational amplifier. The ISFET is connected to the leads of the instrumentation amplifier which are normally used to connect a resistor that determines the amplification of the amplifier. The usual inputs of the amplifier are, in this case, connected to a fixed voltage $IR_1$, provided by the current source. The output voltage of the
The difference between the output voltage of the instrumentation amplifier and an adjustable reference voltage, $V_{\text{ref}}$, is amplified by the operational amplifier, from which the output ‘injects’ a feedback current $I_f$ into $R_2$, thus controlling the source and drain voltages which are equal to $V_{R_1}$ and $V_{(R_3+R_4)}$, respectively. This control action results in a constant $I_d$ at a constant $V_d = IR_1$ or, in other words, $i_d = 0$, $v_d = 0$, while the feedback causes $v_g = 0$. The feedback current is measured via the adjustable resistor $R_9$. If the open-loop amplification of the system, determined by amplification of the combination of ISFET and instrumentation amplifier (approximately $R_8/R_7 \times S \times (R_3 + R_4)$), and the open-loop amplification of the operational amplifier is high enough, the source and drain potential with regard to ground follow a change in the effective input voltage $v_i$ of the
ISFET  At the same time, the amplified input voltage is available across $R_9$ according to

$$v_0 = \frac{R_9}{R_2} v_i$$

(11)

Note that, due to the fact that the potential of the source and drain leads follows the input potential with regard to earth, no capacitive loads exist, which gives the system a maximum of frequency response, independent of the length of the connecting leads. Further, the system includes a very simple calibration possibility, which corresponds to the usual calibration facilities of pH measurements, namely, the starting point that pH = 7 corresponds to a voltage of 0 V.

The procedure of a pH measurement with this system is as follows: If the ISFET is placed in a buffer of pH = 7, the reference voltage is adjusted in such a way that $I_t = 0$. The output voltage across $R_9$ is thus also zero, independent of the value of $R_9$. If the ISFET is then placed in another buffer, e.g., pH = 4, the value of $R_9$ can be adjusted in such a way that an appropriate voltage is measured, e.g., 3 V if the desired sensitivity of the system should be 1 V/pH. Of course the output voltage can be directly calibrated in pH units.

As already mentioned in another paper [8], the reference point pH = 7, corresponding to $V_0 = 0$ V, should be independent of temperature to facilitate absolute measurement. With the development of glass membrane electrodes, this problem is solved in a technological way by the choice of the inner buffer solution. The question arises whether a similar technological solution can also be found for an ISFET. Therefore, we have to focus our attention on the basic equation of the ISFET and determine it in view of the temperature sensitivity of all terms.

4. The temperature sensitivity of ISFETs

The equation for the dc current, $I_d$, of a pH-sensitive ISFET is given by [8]

$$I_d = \beta \left[ \left( V_g - E_{ref} + \Delta \phi_1 + \phi_0 + \frac{RT}{F} \ln a_{H^+} + \Phi_{S1} + \frac{Q_{ss} + Q_{ox} + Q_B}{C_{ox}} - 2\Phi_f \right) V_d \cdot \frac{1}{2} V_d^2 \right]$$

(12)

where $\beta = \mu C_{ox} W/L$ is a geometric constant.

$V_g$ is the potential applied to the reference electrode with regard to the source,

$E_{ref} - \Delta \phi_1$ is the voltage of the reference electrode including the liquid junction potential,

$\phi_0 + \frac{RT}{F} \ln a_{H^+}$ is the voltage across the electrolyte–oxide interface,
\( \Phi_{Si} \) is the silicon work function,
\( Q_{ss} \) is the charge of the interface states,
\( Q_{ox} \) is the charge in the oxide,
\( Q_B \) is the depletion charge in the bulk,
\( \Phi_f \) is the bulk Fermi potential,
\( V_d \) is the dc drain to source voltage,
\( C_{ox} \) is the oxide capacity per unit area.

In the electronic circuit as given in Fig 6, \( V_d \) is kept constant, while \( I_d \) can be set to a desired value, which is also kept constant due to the feedback. For \( pH = 7 \) this value is chosen in such a way that the feedback current \( I_f \) is zero \((V_o = 0)\), resulting in a certain value of \( V_g \) which can be derived from eqn (12)

\[
V_g = \frac{\text{const}}{\beta} + E_{\text{ref}} - \Delta \phi_i - \phi_0 - \frac{RT}{F} \ln a_{H^+} - \Phi_{Si} - \frac{Q_{ss} + Q_{ox} + Q_B}{C_{ox}} + 2\Phi_f
\]  

(13)

For \( pH = 7 \) it yields
\[
\frac{d}{dT} \left( -\frac{RT}{F} \ln a_{H^+} \right) = 1.39 \text{ mV/°C},
\]  

(14)

which means that by maintaining \( I_f = 0 \), independent of temperature, the condition given in the next equation should be fulfilled

\[
\frac{d}{dT} \left( \frac{\text{const.}}{\beta} + E_{\text{ref}} - \Delta \phi_i - \phi_0 - \Phi_{Si} - \frac{(Q_{ss} + Q_{ox} + Q_B)}{C_{ox}} + 2\Phi_f \right) = -1.39 \text{ mV/°C}
\]  

(15)

Of course the same equation has to be obeyed if the bridge circuit as given in Fig 5 is used.

The requirement given in eqn (15) differs from the corresponding one for glass membrane electrodes [8], in the first place due to the fact that now solid-state parameters are part of the equation. Further, the reference electrode voltage and the standard potential of the electrolyte-oxide interface are now also part of the equation.

It is unrealistic to assume that the ISFET process technology should be so accurate that the requirement as given in eqn (15) can be met by a technological process control as is the case for glass membrane electrodes [8]. Fortunately, we have, in contrast to glass membrane electrodes, electronic possibilities to tackle the problem, which will be further discussed in this Section.

In general, it can be stated that a compensation for temperature drift, which requires adjustment for each individual ISFET, is not convenient with regard to the desired interchangeability of the devices. A necessary calibration for the input variable, e.g., the pH, may already be less desirable, this cannot in any way be accepted with regard to an interference signal, as is the temperature in this case.
A usual approach in electronics to compensate for temperature drift in solid-state devices is to create a differential pair on one chip from which one device is the active input device and the other is used for temperature compensation, assuming that the temperature characteristics of both devices are equal. As can be seen from Fig 7, which shows a generalized $I_d-V_g$ curve of a MOSFET or an ISFET as a function of temperature, the requirement $\frac{dI_{d_1}}{dT} = \frac{dI_{d_2}}{dT}$ for a pair of devices having the same characteristics, can only be maintained if, in addition, the electrical bias of both devices is kept equal ($I_{d_1} = I_{d_2}$ and $V_{g_1} = V_{g_2}$).

Both requirements can be met reasonably for a pair of MOSFETs with today's MOSFET technology and the application of electronic feedback (see Fig 8(a)). It is, however, not realistic to use this approach for a pair consisting of an ISFET and a MOSFET on the same chip [9], due to the explicit existence of differences in $V_t$, resulting in a bias difference (see Fig. 8(b))

This is the reason why this system needs adjustment for each individual probe, calibrated by means of a known temperature variation, as reported by McKinley [10]. A principally better approach to solve the problem of automatic compensation for temperature interference is the construction of a differential ISFET pair, one for the measurement of the pH and one with a

![Fig 7 Generalized $I_d-V_g$ curve of a MOSFET or ISFET as function of temperature](image)

![Fig 8 Principle diagrams for differential amplifier circuit with feedback for (a) a pair of MOSFETs, (b) an ISFET and a MOSFET, (c) a pair of ISFETs](image)
separate compartment on top of the gate, filled with a buffered agarose, which is in contact with the solution to be measured via a liquid junction, as described by Janata and Huber [11] and shown in Fig 9

A practical problem with this construction is that the reference ISFET cannot be made completely by a technology which is compatible with IC-technology. Further, the construction introduces a difference in electrical bias due to the additional liquid junction potential for the reference ISFET. This will, of course, also be the case for a difference between the pH of the solution to be measured and the buffer solution of the reference ISFET, if this is not compensated by a feedback system. Then, however, the possibility must be present to contact the liquid gate of the reference ISFET (the buffer solution) separately (see Fig 8(c)). Therefore, this system has to be further investigated electronically as well as technologically.

The conclusion is that the approach of a differential pair construction on one chip to prevent temperature drift, as commonly in use for MOSFETs, cannot be applied directly to ISFETs.

An unusual approach in conventional electronics is the simultaneous detection of those parameters of the measuring device which are responsible for the temperature drift, and to use this signal for compensation.

The function $I_d(T)$ or, in the feedback circuit of Fig 6, the corresponding function $V_q(T)$ (multiplied by a constant factor) is unknown, in such a way that the theory concerning the temperature sensitivity of the terms $\beta$ and $\Phi_f$ contains some empirical coefficients:

$$\beta = C_{ox} \frac{W}{L} T^a \quad -1 < a < -1.5 \quad [12]$$

$$\Phi_f = \frac{kT}{q} \ln \frac{N_A}{CT^{-3/2} \exp(-W_qS_1/T)} \quad [13]$$

Therefore, the simultaneous measurement of the temperature with a separate sensor cannot be used for compensation of temperature drift in $V_q(T)$ for $I_d = \text{constant}$. Instead of this, we have continuously to measure the unknown function for each individual ISFET connected to the amplifier during operation. With this measure, the set value of $I_d$ (see Fig 6) can be controlled in such a way that $V_q = \text{constant}$. The same signal can be used to

![Fig 9 Construction of differential pair of a pH ISFET and a reference ISFET (after Janata and Huber [10])](image-url)
adjust the amplification of the measured output signal as a function of pH, in agreement with the slope correction of glass membrane electrodes.

It has already been shown that for MOSFETs the substrate or bulk, which up to now in this paper has been assumed to be shortened with the source, can be used as an additional signal input. In this case we have to extend eqn (1) with terms which reflect the influence of the bulk to source voltage $V_b$, resulting in eqn (16)

$$I_d = \beta \left[ (V_g - V_t)V_d - \frac{1}{2} V_d^2 - \frac{2}{3} \alpha \{(V_d - V_b + 2\Phi_t)^{3/2} - (V_b + 2\Phi_t)^{3/2}\} \right]$$

where $\alpha = (2\varepsilon_0\varepsilon_s q N_A)^{1/2}/C_{ox}$ and the further symbols being already mentioned in the preceding text.

For the small-signal behaviour of the bulk, differentiation of eqn (16) gives the mutual conductance of the bulk

$$S_b = \left. \frac{dI_d}{dV_b} \right|_{V_d=\text{const}} = \alpha \beta \left[ (V_d - V_b + 2\Phi_t)^{1/2} - (-V_b + 2\Phi_t)^{1/2} \right]$$

As can be seen from eqns (16) and (17), $S_b$ and $I_d$ contain the same temperature dependent terms, $\beta$ and $\Phi_t$. It can be shown [7] that $\Delta I_d/\Delta S_b$ is a constant independent of $V_g$, which means that a simultaneous measurement of $\Delta S_b(T)$ reflects the temperature sensitivity of $I_d$ ($\Delta I_d(T)$). This means for an ISFET that, so far as it concerns the temperature sensitive solid-state parameters, an additional signal can be withdrawn from the device, $\Delta S_b(T)$, independent of the input signal (pH), which can be used for readjustment of $I_d$ via the reference potential $V_{\text{ref}}$ in the circuit of Fig 6 during operation. The result will be that the ISFET, adjusted in a certain bias condition for the reference point pH = 7, will maintain this condition automatically.

The realisation of this system is quite simple as is shown in Fig 10. In order to create a sensitivity independent of $\beta$, $R_s$, and $R_d$, the feedback system of Fig 6 is used. In this system, an additional sinusoidal signal is injected by means of a transformer that is connected between the source and the bulk.

Because the whole feedback system is limited to a frequency of 3 kHz, the frequency of the bulk signal is chosen as 30 kHz, and is thus not affected by the feedback loop. The amplitude of the 30 kHz signal is measured at the output of the instrumentation amplifier by means of a lock-in amplifier, and appears to be proportional to $\Delta S_b$. This signal is used to readjust the set reference voltage $V_{\text{ref}}$.

Considering eqn (15), it will be obvious that the temperature compensation mentioned above only yields for the terms $\text{const}/\beta(T)$ and $\Phi_t(T)$ does not correct changes in the voltage of the reference electrode ($E_{\text{ref}} - \Delta \Phi_t$) and the electrolyte-oxide standard potential, $\phi_0$, as a function of the temperature. If an external reference electrode is used it is clear that this reference electrode has to be kept at a constant temperature. It remains to be seen whether an over compensation of the terms $\text{const}/\beta(T)$ and $\Phi_t(T)$
may also include the effect of $\phi_0(T)$ This may possibly also yield for a reference electrode integrated with the ISFET, and thus subject to the same temperature variations In this respect, eqn (15) can possibly be obeyed As the equation does not give a direct indication for the best value of over compensation, this has to be determined empirically

5 Conclusion

It can be concluded that ISFETs can be seen as a new class of ion sensitive devices which greatly differs from the conventional potentiometric pH-sensor in the sense of being an electronic component which can be controlled by electronic means such as feedback, simultaneous parameter measurement with corresponding compensation, etc On the other hand the ISFET differs from the original electronic component, the MOSFET from which the ISFET is derived, in the sense of having additional parameters of non-physical origin, which cannot be directly influenced by electronic means It is useful to investigate whether a total feedback system including the actual pH unit may solve this problem Such an approach requires, however, the development of an electronic pH actuator This development would lead to far-reaching progress in the application of the devices

References

3 P Bergveld and N F de Rooij, The single electrode operation of an ion-sensitive field


11 J Janata and R J Huber, Ion sensitive field effect transistors, *Ion Select Electrode Rev.*, 1 (1979) 31 - 79
