Tunable Interface Electronics for HEMT based Sensor

Kaushal Kishore[1, 2, a)] and S.A. Akbar[1, 2, b)]

[1)]*CSIR-Central Electronics Engineering Research Institute, Pilani-333031, India*
[2)]*Academy of Scientific and Innovative Research (AcSIR), Ghaziabad, 201002, India*

[a)]*Corresponding author: kaushal@ceeri.res.in*
[b)]*Electronic mail: saakbar@ceeri.res.in*

Abstract. The paper reports a tunable interface electronics for High Electron Mobility Transistors (HEMTs) based Field Effect Transistor (FET) sensor. The approach presented in the work can estimate both the 'on resistance' and drain current of the FET sensor while maintaining a constant drain-source voltage (V_{DS}). The interface electronics consist of a full Wilson current mirror with a floating voltage-controlled resistance which is controlled by the output of an integrator forming a closed-loop feedback control. The bias point for the sensor is set by the external voltage, V_{bias} and the difference between the sensor drain-source voltage, V_{DS} and V_{bias} is fed to the integrator as an error signal. The output of the integrator, V_c varies the tunable resistance which in turn varies the current in the current mirror branch. This current is mirrored to the sensor branch and continues to vary until the drain-source voltage, V_{DS} matches the applied external bias voltage, V_{bias}. At this point, the circuit is at the balanced state and the corresponding control voltage, V_c represents the sensor signal. The circuit is evaluated with SPICE simulation and experimentally verified using a prototype PCB. Experimental testing revealed that the circuit can work for a range of 200Ω to 2000Ω with an absolute relative error of less than $\pm 1\%$. The circuit can also be tuned for the desired working range and allows the user to apply external bias potential based on the sensor's requirement. The proposed circuit has great potential as a bio-chemical interface circuit for FET based sensors.

INTRODUCTION

Interest in Field Effect Transistor (FET) based devices as bio-chemical sensors was propelled with the introduction of open-gate ion-selective field-effect transistor by Bergveld [1]. The device combined the effect of a chemically sensitive membrane at the open gate area and the conduction channel between drain and source terminal of the FET underneath the gate. The resulting structure give rise to modulation of source-drain current (I_{DS}) of the FET. Though it was a breakthrough, extensive study of such devices revealed shortcomings such as thermal dependency, drift and hysteresis. Overcoming these issues, GaN/AlGaN high electron mobility transistors (HEMTs) based FET have been extremely popular for number of reasons. These devices are bio-compatible, chemically inert, stable at high temperature, inherently sensitive to ambient. Unlike conventional semiconductor FET, the channel formation is not dependent on the doping concentration but rely on two-dimensional electron gas (2DEG) located at the interface of the AlGaN and GaN hetero-structure [2, 3]. This 2DEG is formed due to the piezoelectric polarization of the strained AlGaN layer and the difference in the spontaneous polarization between AlGaN and GaN. Also, this high electron sheet carrier concentration channel is sensitive to the ambient [4, 5]. Due to these advantages, group III – nitride including GaN, AlN, InN and their alloys are increasingly used for next generation chemical and biological sensing platforms. The inherent property of the heterostructure device to form a polarization – induced 2DEG is an appealing feature to detect ions, impurities in liquid, bio molecules and polar liquids.

Seeing the promising possibility of these devices, a great deal of research has also been made to model the Al-GaN/GaN devices to comprehend its behavior [6]. Close study of the device reveals that the drain-source current (I_{DS}) is a coupled function of gate-source voltage (V_{GS}) and drain-source voltage (V_{DS}) represented mathematically as:

$$I_{DS}(V_{GS}, V_{DS}) = f(V_{GS})g(V_{DS}) \tag{1}$$

While FET devices are typically characterized by 'on resistance' or drain current (IDS) variation ratio with respect to the analyte concentration, the working of the device is ensured by its interface circuit to implement a precise biasing. For ultra-low analyte concentration, such as proteins, bio-markers, impurities and heavy metals in liquid, a precise and stable interface circuit is highly desirable.

As the sensor design field progresses, the need for novel interface circuits exclusive to HEMT based sensors has also increased. Although, HEMT based sensors have clear advantages, system design based on these sensors are still not reported. One of the contributing factors is the fact that these sensors have a high base/offset current with a response <10% of the base current [7, 8, 14]. Also, biasing for the sensor has to be controlled as the response is dependent on both drain – source voltage (V_{DS}) and gate – source voltage (V_{GS}) as reveled in Eq. (1). Circuits based on Null-balancing techniques are commonly used for FET based devices [9] and are preferred for sensors with low offset current. With increasing base current, the null-balancer saturates due to high offset. This effect is also true for the conventional FET biasing technique based on constant current source [10, 11]. Also, the drain-source voltage (V_{DS}) is externally applied and is not ensured to be constant by any control loop. This may affect the sensor output as HEMT based sensors are highly sensitive to bias potential and ambient conditions. For circuits based on voltage clamping technique, a reference FET is used along with the primary sensor [12, 13, 15]. This decreases the substrate utilization as reference sensor is not functionalized for any analyte detection.

This work presents a novel approach to estimate both the drain current and the 'on resistance' of an AlGaN/GaN HEMT sensor at a constant (V_{DS}). This is accomplished by an auto-balancing loop implemented using a floating voltage-controlled resistance in a full Wilson current mirror. The technique provides linearized output and automatic bias voltage adjustment which can be controlled by applying an external voltage. Since the drain-source voltage is constant, the drain current in a direct function of gate potential, which is capacitively coupled with the analyte concentration. The interface design also enhances the CMRR as the output from the sensor is taken in differential mode and also ensures a constant sensitivity for the entire working range of the sensor. The circuit is also tested with an in-house designed and fabricated HEMT sensor for heavy metal sensing in liquid.

SIGNAL CONDITIONING CIRCUIT

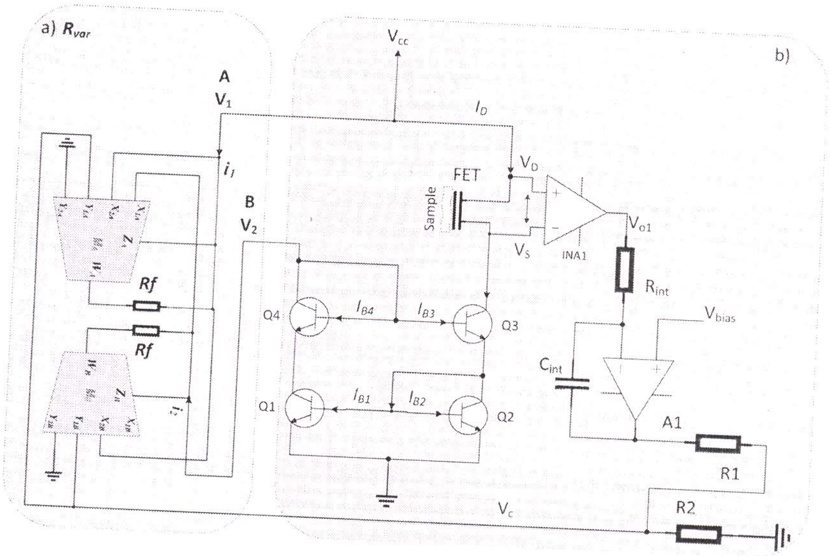

FIGURE 1. Proposed interface circuit, a) floating voltage-controlled resistance, b) auto-balancing full Wilson current mirror

The proposed interface can be introduced in three different sections namely, a full Wilson current mirror, voltage-controlled resistor and auto balancing loop. Transistors Q1 – Q4 form a full Wilson mirror with better current accuracy []. The auto-balancing loop designed using A1 ensures that the drain-source voltage (V_{DS}) is always constant and is equal to the applied voltage, V_{bias}. To perform this task, the control voltage generated by the integrator is applied to the floating voltage-controlled resistor, R_{var} which in turns changes the current, IR of the Wilson current mirror. An equal current, I_D is mirrored due to the feedback loop which causes a constant (V_{DS}) across the FET sensor. The following sections are dedicated for detailed description of the circuit.

Full-Wilson current mirror

Transistors Q1 – Q4 are arranged to form a current mirror, equalizing the collector voltages of Q1 and Q2 by lowering the collector voltage of Q1 by V_{BE4}. The incoming branch current IR is mirrored as IL which acts as the driving current for the FET sensor. The output current equation for a full-Wilson current mirror is given as:

$$I_L = I_R = \frac{(V_{cc} - V_{BE2} - V_{BE3})}{R_{var}} \tag{2}$$

Considering $V_{BE2} = V_{BE3} = V_{BE}$,

$$I_D = I_R = \frac{(V_{cc} - 2V_{BE})}{R_{var}} \tag{3}$$

From Eq. 3, it can be noted that the load current, I_D is tunable with inverse proportionality to the resistance Rvar.

Floating voltage-controlled resistance

As discussed, the output current of the current mirror is tunable by the variable resistance R_{var}. We have considered a floating voltage-controlled resistance between terminal A and B as shown in Fig. 2 which consist of two analog voltage multiplier M1 and M2.

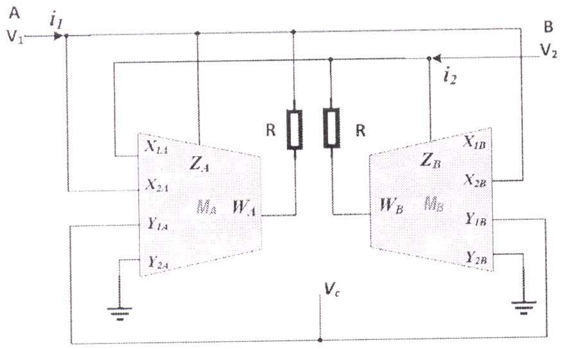

FIGURE 2. Floating voltage-controlled resistance

Considering properties of AD633 as voltage multiplier, the transfer function can be expressed as:

$$V_{W_A} = \frac{(X_{1A} - X_{2A})(Y_{1A} - Y_{2A})}{10V} + Z_A \tag{4}$$

Using the terminal voltage shown in Fig. 2, Eq. (2) can be expressed in terms of voltages as:

$$\begin{aligned} V_{W_A} &= \frac{(V_2 - V_1)(V_c)}{10V} + V_1 \\ V_{W_B} &= \frac{(V_1 - V_2)(V_c)}{10V} + V_2 \end{aligned} \tag{5}$$

The input current from terminal A and B can be expressed as:

$$\begin{aligned} i_1 &= V_1 - \left(V_1 - \left((V_1 - V_2)\frac{V_c}{10V}\right)\right)\frac{1}{R} = \frac{\Delta V V_c}{R} \\ i_1 &= V_1 - (V_1 - \left(\left(V_1 - V_2\frac{V_c}{10V}\right)\right)\frac{1}{R} = \frac{\Delta V V_c}{R} \end{aligned} \tag{6}$$

The equivalent resistance between terminal A and B can be expressed as:

$$R_{var}(V_c) = \frac{V_1 - V_2}{i} \tag{7}$$

Where, $i = |i_1| = |-i_2|$ Consider $R_A(V_c)$ and $R_B(V_c)$ as voltage-controlled resistors formed by multiplier MA and MB respectively. The voltage difference between terminal A and B can be expressed as:

$$V_1 - V_2 = \Delta V = i_1 R_A(V_c) - i_2 R_B(V_c) \tag{8}$$

$$= \frac{\Delta V V_c/10}{R}[R_A(V_A) + R_B(V_c)] \tag{9}$$

Thus, the voltage-controlled resistor, $R_{var}(V_C)$ between floating terminal A and B is expressed as:

$$R_{var}(V_c) = R_A(V_c) + R_B(V_c) = \frac{R_f}{V_c/10V} \tag{10}$$

The practical limit of V_c is around [0.05, 9.8] V. Using the above Eq. 8, the range of the tunable resistance can be estimated to be:

$$R_{var}(V_c) = \begin{cases} 200R_f & ; V_c = 0.05V \\ 1.02R_f & ; V_c = 9.8V \end{cases} \tag{11}$$

For a given power supply, the resistance R_f is selected such that the Wilson current mirror can deliver that maximum possible current so as to maximize the range of operation of the interface circuit. However, the circuit can be designed for suitable range by selecting the excitation voltage, V_{cc} and the resistance, R_f.

Auto-balancing feedback circuit

The input to the auto-balancing feedback circuit is provided by the instrumentation amplifier INA1 that calculates the differential voltage, V_{DS} across the drain and source of FET sensor. The circuit maintains a high CMRR as the output voltage is taken in differential mode. As shown in Fig.1 the amplifier A1 with R and C forms an integrator that generates a control voltage, V_c in order to nullify the error between the external applied voltage, V_{bias} and the output voltage, V_{o1} of INA1. The generated control voltage is scaled by resistance divider R1 and R2 so that the voltage swing is always below 10V. The voltage at the output of INA1 is given as:

$$V_{o1} = V_D - V_S = R_{on} I_D \tag{12}$$

From Eq. 3, to express the current I_D,

$$V_{o1} = \frac{(V_{cc} - 2V_{BE})}{R_{var}} R_{on}$$
$$V_{o1} = \frac{(V_{cc} - 2V_{BE})V_c)}{10R_f} R_{on} \tag{13}$$

Under balanced condition, $V_{o1} = V_{bias}$,

$$R_{on} = \frac{(10R_f V_{bias})}{V_c(V_{cc} - 2V_{BE})} \tag{14}$$

From Eq. (14), it can be noticed that the 'on' resistance of HEMT (R_{on}) is directly proportional to the control voltage. V_c. Response of the FET sensors are often expressed in term of the drain-source current, I_{DS} which can be expressed in the interface circuit design as:

$$I_{DS} = \frac{V_{DS}}{R_{on}} = \frac{V_{bias}}{R_{on}} = \frac{(V_{cc} - 2V_{BE})}{V_c R_f} \tag{15}$$

PROTOTYPING AND EXPERIMENTAL RESULTS

The proposed interface circuit in 1 has been experimentally verified by choosing commercial off-the shelf components. BC548 is used as transistors Q1-Q4, INA114 as precision instrumentation amplifier INA1 and multiplier, Mo, M1 and M2 are implemented by AD633. The feedback loop has OPA177 as amplifier A1, RINT = 128kΩ, CINT = 1000pF. The amplifier is supplied with ±15V and so the saturation voltage may exceed the control voltage limit of multiplier AD633. For this reason, a resistance divider is formed with R1 = 50kΩ and R2 = 105kΩ scaling the output voltage of the integrator within the range of input control voltage of the multiplier. The floating voltage-controlled resistance has base resistance R with a value of 214Ω. All passive components are verified for tolerance within ±1% for resistive element and ±10% for capacitive elements. For testing purpose, the bias voltage of the sensor (V_{bias}) is set as 3.3V and can be changes as per the sensor's working voltage. The prototype PCB is tested with a resistive decade box,

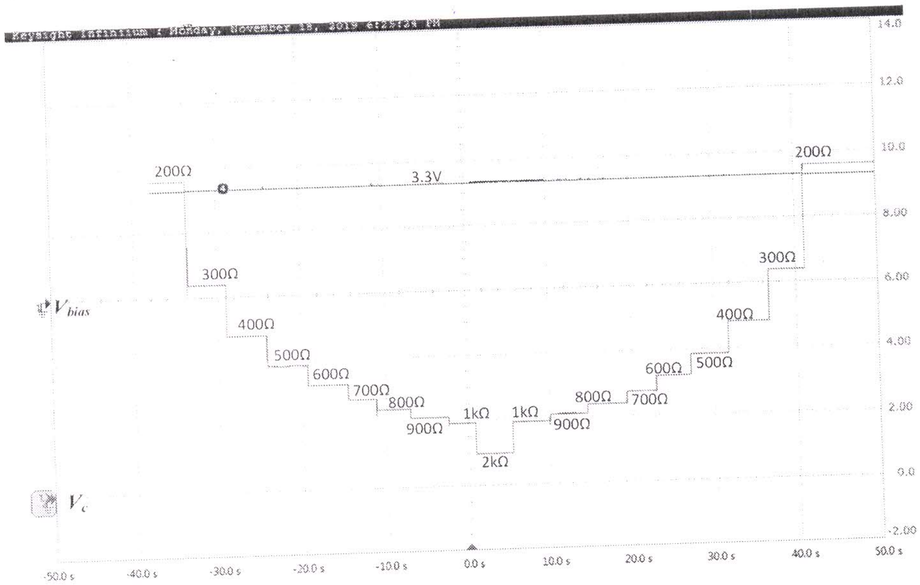

FIGURE 3. Control voltage variation with different sensor values. Note that the clamping voltage is fixed at 3.3 V for the full range variation in the sensor value.

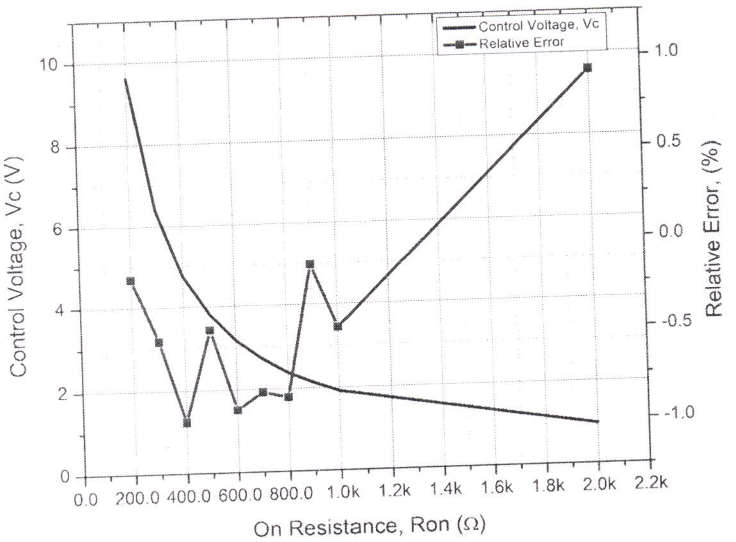

FIGURE 4. Control voltage response with varying 'on resistance' of sensor along with the relative error

emulating the 'on resistance' of the FET sensor. For interface circuit response is recorded in an oscilloscope while changing the sensor resistance. The resulting plot is shown in 3. It can be noted that the bias voltage is maintained at 3.3V by the auto-balancing loop for the entire range of operation. The data from the experiment is used for calculating the relative error and the resulting plot is shown in 4. Note that the change in 'on resistance' of a typical FET sensor is of the order of 10's of Ω. The designer can choose the operating point and tune the circuit parameters to operate the circuit in the operational region of the sensor. The current experiment exhibits a control output for the 'on resistance' of 200Ω to 2000Ω with a relative error less that 1%.

CONCLUSION

A new interface technique utilizing an auto balancing current biasing approach is presented in this paper. By integrating a feedback loop with a full wilson current mirror, the circuit maintains a constant drain-source voltage and thus the changes in the drain current and equivalently in the 'on resistance' is measured in a differential mode. The clamping voltage for the sensor is tunable and can be adjusted as per the sensors working voltage. Experimental evaluation of the prototype PCB is also presented showing the working of the interface circuit from 200Ω to 2000Ω. The analog interface exhibits high CMRR with a low relative error ($< 1\%$). We believe that the proposed interface is suitable for FET based chemical and biological sensors and is also customizable as per the sensor's range and sensitivity.

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