Analysis of GaN-based HEMTs operating as RF detectors over a wide temperature range

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Abstract— This paper presents an analysis of detection in the microwave region with AlGaN/GaN high-electron mobility transistors in terms of the key figures of merit: responsivity (both with voltage- and current-mode detection schemes) and noise equivalent power, in the temperature range between 20 and 400 K. Transistors with three different gate lengths (75, 150 and 250 nm) have been measured and favorably compared with a simple quasi-static model extracted from the DC curves, which is able to reproduce the zero-bias experiments at 1 GHz in the entire gate-bias sweep and operating temperature range. Such model allows to explain the detection experiments performed both with the devices operating in zero-current or zero-voltage conditions, and demonstrates that the bowing coefficient of the $I_D-V_{DS}$ curves is the parameter which determines the performances of the devices as RF detectors. The voltage responsivity (in V/W) increases when the gate bias approaches the threshold voltage, but shows different behaviors in subthreshold conditions. Depending on whether the zero-current bias point is near zero drain voltage (for leaky devices, when the channel is not well pinched-off due to the short gate or to the conduction through the buffer at low temperature) or shifted to the point where third quadrant conduction starts (for well-behaved devices), the responsivity presents a decrease or an approximately constant value, respectively. On the other hand, when the detector is biased, the curves of current responsivity (in A/W) show a characteristic bell-shape with a maximum which is approximately the same for all tested devices and hardly depends on the gate bias (it is only shifted to different drain bias points).

Index Terms— RF detection, zero-bias detector, GaN HEMTs, quasi-static model.

I. INTRODUCTION

The recent advances in transistor technology using narrow-bandgap high-mobility semiconductors have made possible to reach the milestone of THz amplification with a 25-nm InP high electron mobility transistor (HEMT) [1]. On the other hand, widebandgap-based transistors such as GaN HEMTs have desirable electrical characteristics for high-power high-temperature operation at moderately high frequencies [2].

In the last years different types of Field Effect Transistors (FETs) have been used as RF and THz detectors. Indeed, HEMTs are able to detect RF signals with frequencies much above their cutoff for amplification (described by their commonly used figures of merit $f_t$ and $f_{max}$). Even if there is still controversy about the physical origin of such detection mechanism (plasmonic or thermoelectric effects can play a role [3]–[16]), practical THz detection is feasible with GaN HEMTs despite their lower cutoff frequencies with respect to faster InP-based devices. Wide bandgap semiconductor detectors have the advantage of improved robustness and the possibility for high-power high-temperature operation, which may allow for applications such as ultrahigh-power RF detection or wireless power transmission [17].

In FET-based RF detectors operating in the voltage-mode scheme (with open-circuited output), the best noise equivalent power (NEP) is found when $V_{GS}$ is just above the threshold bias value ($V_{th}$) [18], [19], but the maximum value of responsivity appears in the subthreshold region [19]–[21]. This operation region has attracted some attention in the literature and models have been proposed, some based on self-mixing square law detectors [20], [22]–[25] and plasma wave theory [21], [26]. Interestingly, in the current-mode detection scheme (with short-circuited output, best suited for fast operation), the results presented in the literature show that the responsivity vanishes for $V_{GS}<V_{th}$ [16], [22]. In practical cases (which are neither voltage- nor current-mode operation), the impedance of the readout system has also an effect in the value of the responsivity, diminishing as the load resistance decreases [18], [21], [23].

Even if zero-bias operation leads to optimum low noise operation and allows one to get rid of self-heating effects, in this work we present a complete study of both zero-bias and biased RF detection with AlGaN/GaN HEMTs in a wide range of temperature (20-400 K). Drain-coupled RF injection in both current- and voltage-mode detection schemes will be employed in our experiments and the results will be interpreted in terms of the non-linearity of the $I_D-V_{DS}$ curves (low-frequency resistive mixing mechanism) [27]. For that sake, a simple quasi static (QS) model [27], [28] will be used, able to capture all the features of the RF responsivity in all of the operation conditions just by analyzing the static $I_D-V_{DS}$ characteristics at the corresponding operating point. In Section II the devices under test (DUT) and the experimental set-up are described. The QS model is explained in Section III and afterwards, in Section III,
the experimental and modeled results are presented and the origin of the behavior of the RF detection in the different operation modes is discussed. Finally, the main conclusions of our work are drawn in Section V.

II. DEVICE UNDER TEST, EXPERIMENTAL SETUP

In this work, we focus our study on AlGaN/GaN HEMTs grown on a high resistivity silicon substrate whose details are provided in [29]–[31]. The characterized devices have a drain-to-source distance $L_{DS}=2.5 \mu$m, a width of $W=2.5 \times 2.5 \mu$m and three different gate lengths $L_{G}=75$ nm, 150 nm and 250 nm. For the measurements, see Figure 1, the DUTs are placed inside a LakeShore CRX-VF cryogenic probe station, and are on-wafer connected by means of GSG RF probes to a VNA Keysight N5244A, used as RF power source (which is injected to the drain contact). A two-channel SMU B2902A allows both biasing the device and record the output voltage/current detected at the drain terminal. One channel of the SMU is connected to the gate and the other to the drain terminal, in this case through the internal bias-tee of the VNA, making possible combining the RF input and the DC bias. The selected operating temperatures are 20, 100, 200, 300 and 400 K.

Two different zero-bias detection schemes can be employed, zero-current voltage-mode (ZC), where drain terminal is open-circuited and its voltage is measured, or zero-voltage current-mode (ZV) when drain terminal is short-circuited and its current is measured. The responsivity in ZV conditions is the ratio $\beta^{ZV}=\Delta V_{DS}/P_{in}$, where $\Delta V_{DS}$ is the DC drain voltage shift caused by the RF excitation.

The most important figure of merit for detectors is the noise equivalent power (NEP), which indicates their sensitivity for detecting low-power signals and is defined as the relation between the noise amplitude and the responsivity (both under current- or voltage-mode operation):

$$NEP = \frac{V_n}{\beta_V} = \frac{I_n}{\beta_I}, \text{ in } W/\sqrt{\text{Hz}} \quad (1)$$

being $V_n$ and $I_n$ the voltage- and current-noise amplitudes, respectively. Under ZC or ZV conditions, we can assume that $V_n$ and $I_n$ take the equilibrium values given by the Nyquist theorem, so that the value of the NEP is:

$$NEP^{ZC} = \sqrt{(4k_BT/R_D^{ZC})/\beta_V^{ZC}}, \quad (2)$$

$$NEP^{ZV} = \sqrt{(4k_BT/R_D^{ZV})/\beta_I^{ZV}}, \quad (3)$$

being $R_D^{ZC}$ and $R_D^{ZV}$ the drain resistance at $I_D=0$ and $V_{DS}=0$, respectively, $k_B$ the Boltzmann constant and $T$ the lattice temperature. The only possible difference between $NEP^{ZC}$ and $NEP^{ZV}$ is just associated to the fact that ZC and ZV bias points may not be the same, as we will see later in the results section.

When the detector is biased (and neither ZC nor ZV conditions apply), excess noise is generated, so that the noise power should be measured in order to determine the exact value of the NEP.

III. QUASI-STATIC MODEL

Based on the static $I_D$-$V_{DS}$ curves, an analytical quasi-static (QS) model is able to predict the RF responsivity ($\beta_{QS}$) at low frequency of FETs based on the differential drain-source resistance, $R_D$, and the bowing coefficient $\gamma$ which can be extracted from the $I_D$($V_{DS}$) dependence as:

$$R_D = \left(\frac{\partial V_{DS}}{\partial I_D}\right)^{-1}, \quad (4)$$

$$\gamma = \frac{\partial^2 I_D}{\partial V_{DS}^2} = \frac{\partial^2 I_D}{\partial V_{BS}^2} R_D (\text{in } V^{-1} \text{ units}) \quad (5)$$

We have to remark that non-quasi-static mechanisms such as self-heating or trapping effects [32]–[34] may lead to a low-frequency dispersion of the responsivity of the devices (as happens in nanometric-width diodes [35]), and originate discrepancies between the predictions of the QS model and the experimental results at GHz frequencies. In the case of the transistors studied here, first, self-heating is almost negligible (mainly in zero-bias conditions, but also when biasing the devices, since the dissipated power is very low). Self-heating effects in the same transistors were previously studied by means of Monte-Carlo simulations and pulsed $I_D$-$V_{DS}$ characterization in [31]. Moreover, low frequency dispersion was observed in virgin devices [36] but disappeared after biasing the transistors. Therefore, an almost perfect coincidence between the between the QS model predictions taking as a base the DC $I_D$-$V_{DS}$ curves and the responsivities measured at 1 GHz is found.

In case trapping or heating effects were significant in the operation of the transistors as RF detectors, the QS model would account for them all since they could be evidenced by
means of pulsed $I_D-V_{DS}$ measurements. By measuring the $I_D-V_{DS}$ curves with different pulse widths, the frequency dependence of $R_D$ and $\gamma$ would be obtained, and by applying the proposed QS model (which is still valid for linking the so obtained $I_D-V_{DS}$ and the RF detection) the possible frequency dispersion of the detection would be straightforwardly obtained.

III.1 Current-mode detection

The QS model [27], [28], [37] is based on the second order Taylor series expansion of the current as a function of the voltage,

$$I_D \approx I_{D0} + \Delta V_{DS} \frac{\partial I_D}{\partial V_{DS}} + \frac{1}{2} \Delta V_{DS}^2 \frac{\partial^2 I_D}{\partial V_{DS}^2}, \tag{6}$$

where $I_{D0} = I_D(V_{DS0})$ is the DC bias term, $\Delta V_{DS} = V_{DS} - V_{DS0}$ with $V_{DS}(t) = V_{DS0} + V_{RF} \cos(2\pi f t)$ the harmonic RF excitation with frequency $f$ and amplitude $V_{RF}$ sufficiently small so that higher order terms can be neglected. The derivatives have to be evaluated around the bias point $(I_{D0}, V_{DS0})$. Because of the harmonic signal excitation, the time-dependent current can be obtained from eq. (6) as:

$$I_D(t) \approx I_{D0} + V_{RF} \cos(2\pi f t) \frac{\partial I_D}{\partial V_{DS}} + \frac{1}{4} V_{RF}^2 [1 + \cos(4 \pi f t)] \frac{\partial^2 I_D}{\partial V_{DS}^2}. \tag{7}$$

Therefore, the detected current $\Delta I_D$, obtained as the difference between the average value of $I_D(t) = \overline{I_D(t)}$ and the drain current without RF excitation $I_{D0}$ is:

$$\Delta I_D = \overline{I_D(t)} - I_{D0} \approx \frac{1}{4} V_{RF}^2 \frac{\partial^2 I_D}{\partial V_{DS}^2}. \tag{8}$$

Since the RF responsivity, $\beta$, is defined as the ratio between the output signal and the input power, in order to evaluate the value of the current-mode responsivity $\beta_I = \Delta I_D/P_{in}$ (in A/W units), one must first take into account that the RF power dissipated by the detector is:

$$P_{in} = \frac{1}{2} T \int [I_D(t) - I_{D0}][V_{DS}(t) - V_{DS0}] dt = \frac{1}{2} V_{RF}^2. \tag{9}$$

Using eqs. (2), (5) and (6), the optimum value of $\beta_I$, without considering the impedance mismatch between the RF source and the detector, can be obtained with this QS model as:

$$\beta_{I,opt} = \frac{1}{2} \frac{\partial^2 I_D}{\partial V_{DS}^2} R_D = \frac{1}{2} \gamma. \tag{10}$$

III.2 Voltage-mode detection

The same reasoning can be applied to the voltage-mode detection if a RF harmonic current excitation around the bias point $(I_{D0}, V_{DS0})$ is considered, $I_D(t) = I_{D0} + I_{RF} \cos(2\pi f t)$. In this case the output signal obtained is the drain bias shift $\Delta V_{DS} = \overline{V_{DS}(t)} - V_{DS0}$, and the voltage-mode responsivity is obtained as $\beta_V = \Delta V_{DS}/P_{in}$ (in V/W units). For that sake one must start with the Taylor expansion of the reciprocal expression of the $I_D-V_{DS}$ curves $V_{DS}(I_D)$, so that an expression similar to eq. (8) is obtained:

$$\Delta V_{DS} = \overline{V_{DS}(t)} - V_{DS0} \approx \frac{1}{4} I_{RF} \frac{\partial^2 V_{DS}}{\partial I_D^2}. \tag{11}$$

and knowing that in this case the RF power can be calculated as $P_{in} = \frac{1}{2} \frac{I_{RF}^2 R_D}{\gamma}$. The optimum voltage responsivity $\beta_{V,opt}^{QS}$ can be computed as:

$$\beta_{V,opt}^{QS} = \frac{\Delta V_{DS}}{P_{in}} = \frac{1}{2} \frac{\partial^2 V_{DS}}{\partial I_D^2} \frac{1}{2} \frac{\partial^2 V_{DS}}{\partial I_D^2} \frac{R_D}{\gamma} = \frac{1}{2} \frac{\partial^2 V_{DS}}{\partial I_D^2} R_D. \tag{12}$$

and since

$$\frac{\partial^2 V_{DS}}{\partial I_D^2} = \frac{\partial}{\partial I_D} \frac{\partial V_{DS}}{\partial I_D} = \frac{\partial}{\partial I_D} \left( \frac{\partial V_{DS}}{\partial I_D} \right)^{-1} \frac{\partial}{\partial I_D} \left( \frac{\partial V_{DS}}{\partial I_D} \right)^{-1} = - \frac{\partial^2 V_{DS}}{\partial I_D^2}, \tag{13}$$

one can easily obtain the value of the voltage responsivity using the definition of $\gamma$ given in eq. (5), as:

$$\beta_{V,opt} = \frac{1}{2} \frac{\partial^2 V_{DS}}{\partial I_D^2} R_D = - \frac{1}{2} \gamma R_D = - \beta_{I,opt}^{QS} R_D. \tag{14}$$

Therefore, an often overlooked property of RF detectors is demonstrated, (i) current- and voltage-mode responsivities have opposite sign and, (ii) as expected, the rate of their modules is $R_D$.

III.3 Impedance Mismatch

In order to obtain the practical value of the responsivities of a given detector, the mismatch between the impedance of the device ($Z_d$, which can be assumed to be equal to $R_D$ at low
presented the values obtained for devices operate for ZC or ZV detection. The inset of Figure 2

and the detector, being $Z_s = 50 \Omega$. So that, the channel resistance, bowing coefficient and the current- and voltage-mode responsivities (also and the NEP, if in which happens

In the strongly mismatched case, when $Z_s \ll R_D$ (as happens in our devices when $V_{GS}$ approaches $V_{th}$) $|\Gamma| \approx 1$, and $(1 - |\Gamma|^2) \approx 4Z_s/R_D$, so that $\beta_{Q50}$ vanishes and the voltage responsivity is independent of the device impedance taking a value of

$$\beta_{Q50} \approx -2Z_s\gamma = -100\gamma. (17)$$

IV. RESULTS AND DISCUSSION

Figure 2 shows the $I_P - V_{DS}$ characteristics for the GaN HEMTs with gate lengths $L_g = 75$, 150 and 250 nm, including the negative $V_{DS}$ region, which plays a key role for drain-injected RF power detection. Indeed, the third quadrant conduction (TQC) region (typical of FETs) is clearly observed. The TQC region appears as a consequence of the symmetry of the device [37–39], so that the conduction condition for the gate voltage when a drain bias is applied ($V_{GD} = V_{GS} - V_{DS} > V_{th}$) is fulfilled when $V_{DS}$ takes negative values below $V_{GS} - V_{th}$. For FETs operating as RF detectors, TQC has an important influence, particularly in subthreshold conditions ($V_{GS} < V_{th}$) as we will show in this study. The $I_P - V_{DS}$ curves at low temperatures do not show any kink or other non-ideal behavior in the drain-bias range under analysis (see the results shown in [37]).

The value of $V_{th}$ is calculated as the $V_{GS}$ value where the extrapolation of the linear region of the curve $\sqrt{V_{DS} - V_{GS}}$ intercept the $V_{GS}$ axis for $V_{DS} = 0.1$ V (approximately where the devices operate for ZC or ZV detection). The inset of Figure 2 (b) presents the values obtained for $V_{th}$ as a function of temperature. As expected, $V_{th}$ is more negative as the gate length is reduced due to short channel effects. Moreover, regardless of the gate length, it is constant with temperature below 100 K and then slightly decreases with increasing temperature.

In order to understand correctly the detection results that will be shown later, a key point to bear in mind is the location the static operation point (Q-point) of the HEMTs, since it will determine their performances. By applying the QS model to the measured DC characteristics around the Q-point we will have a precise estimation of the RF detection figures of merit of the transistor at low frequency. Figure 3 shows two selected $I_P - V_{DS}$ curves from Figure 2(c) for $V_{GS} = -4.1$ V and -3.1 V, corresponding to subthreshold and above-threshold conditions, respectively ($V_{th} \approx -3.7$ V for this 250-nm-gate HEMT at 300 K). The parameters of the QS model are obtained at each Q-point of the measured DC curves through a polynomial fitting in a window of $\Delta V_{DS} = 0.4$ V (see Figure 3) centered at a specific point in the curve fixed by its $I_P$ or $V_{DS}$ value depending on whether voltage- or current-mode detection is used. In that way, the channel resistance, bowing coefficient and the current- and voltage-mode responsivities (also the NEP, if in equilibrium conditions) can be obtained as a function of the bias. In Figure 3 the Q-points represented correspond to the zero-$I_P$ bias point, which, surprisingly shifts to negative values of $V_{DS}$ when $V_{GS}$ enters into the subthreshold region. This is so because there is always a small positive $I_P$ leakage current which flows out of the drain terminal even if $V_{DS} = 0$, due to the non-zero value of the gate leakage current (with $I_P \approx |I_G|/2$, since half of the gate flows out to the source terminal, see [37]). As a consequence, a negative $V_{DS}$ is created when the bias condition $I_P = 0$ is maintained. This is especially evident well in subthreshold conditions ($V_{GS} < V_{th}$), because the Q-point is significantly shifted to the (negative) $V_{DS}$ value corresponding approximately to the onset of the TQC, $V_{DS} \approx V_{GS} - V_{th}$ (in the case of no significant buffer leakage [37]). Note the quite

![Fig. 3. $I_P - V_{DS}$ output characteristics for $V_{GS} = -4.1$ V and -3.1 V for $L_g = 250$ nm at 300 K. At each Q-point a polynomial fitting is made in a window of $\Delta V_{DS} = 0.4$ V.](image)

![Fig. 4. (a) Bowing coefficient, $\gamma$, and (b) $R_D$ in ZC (solid lines) and ZV (symbols, dashed lines) conditions for the HEMT with $L_g = 250$ nm. The values are obtained by applying the QS model to the DC measurements of the $I_P - V_{DS}$ curves of the device in the temperature range 20-400 K. The vertical dashed lines indicate the value of $V_{th}$ for each $T$.](image)
different shape of the $I_D - V_{DS}$ curve around the Q-point when this is shifted to negative values of $V_{GS}$.

In order to evidence the effect of the Q-point shift on the detection performances, Figure 4 shows the values of $\gamma$ and $R_D$ for $L_g = 250$ nm, both at zero-$I_D$ (ZC) and zero-$V_{DS}$ (ZV) conditions, obtained using the previously explained QS model. For $T \leq 100$ K, the value of $\gamma$ is similar in ZC and ZV conditions, with a maximum value for $V_{GS}$ around (or slightly below) $V_{th}$, dropping to zero both in open channel and subthreshold regions. However, for higher $T$, the curves start to diverge when $V_{GS} < V_{th}$: $\gamma$ goes to zero in ZV but remains nearly constant at its maximum value in ZC. This happens because of the previously explained shift of the Q-point in ZC with respect to ZV conditions, which takes place only at $T > 100$ K. As explained in [37], at low $T$, the ionization level of the buffer acceptors decreases (C atoms are present in our samples in a concentration of about $10^{17}$ cm$^{-3}$ leading to a semi-insulating GaN buffer), so that a significant drain leakage current flows through the buffer. In those conditions, a small drain voltage induces a significant negative drain leakage current, which compensates the positive gate contribution (without the need of TQC), and therefore the ZC and ZV Q-points are both located at the origin of the graph (the zero-$I_D$ point is nearly at $V_{DS} = 0$).

On the other hand, $R_D$ increases when decreasing $V_{GS}$ and saturates when entering the subthreshold region, Figure 4(b), with a value which strongly depends on $T$. Indeed, the value of $R_D$ in subthreshold conditions first increases when lowering $T$ and then sharply decreases as a consequence of the drain current leakage through the buffer (associated to non-ionized acceptors) previously discussed. And, since the values of $R_D$ when $V_{GS} < V_{th}$ are lowered by the shift of the Q-point in ZC conditions at $T > 100$ K, their maximum values (≈ $\text{M}\Omega$) are observed at 200 K in ZC and 300 K in ZV conditions, a similar behavior to that of $\gamma$.

IV.1 Zero-Bias detection

Figure 5 shows the excellent agreement found between the RF detection measurements made at 1 GHz (with a small input RF power of $P_{in} = -15$ dBm) and the results obtained with the QS model for the three devices with $L_g = 75$, 150 and 250 nm (in a wide temperature range of 20-400 K). Remarkably, only a weak dependence on $L_g$ is observed.

Regarding current-mode ZV detection, the responsivity always exhibits a bell-shaped behavior, vanishing in subthreshold conditions, as expected from eq. (15), due to the strong impedance mismatch. This also makes the maximum of $\beta_{V_{th}}$ to be found for $V_{GS}$ around 0.2-0.3 V above $V_{th}$ in all the cases, shifted to the right with respect to the maximum of $\gamma$, Figure 4(a), due to the influence of the $(1 - |\Gamma|^2)$ factor which approaches zero when $V_{GS}$ approaches $V_{th}$.

The maximum value of $\beta_{V_{th}}^{ZV}$ increases as $T$ decreases due to the enhancement of $\gamma$ observed in Figure 4(a), but only down to a certain $T$ when the pinch-off of the device starts to be degraded, as observed in Figure 6(b), where the values of the main detection figures of merit for the optimum $V_{GS}$ are presented (in absolute value). However, even if the maximum value of $\gamma$ is obtained at $T = 200$ K, the best current responsivity is found at the lowest $T$. This happens because as $T$ decreases a lower $\gamma$ is compensated with a better impedance matching ($R_D$ is lower, below the kΩ range, due to a higher electron mobility) at around $V_{GS} = V_{th} + 0.2$ V, where the maximum $\beta_{V_{th}}^{ZV}$ is obtained, see Figure 4(b).

Regarding ZC detection, Figures 5(d)-(f), the values of $\beta_{V_{th}}^{ZC}$ follow the two different trends (depending on the temperature range) observed for $\gamma$ in Figure 4(a): the responsivity reaches its maximum near $V_{GS} = V_{th}$ and then drops in subthreshold conditions for $T \leq 100$ K, while for $T > 100$ K $\beta_{V_{th}}^{ZC}$ remains constant at around its maximum. In this case, contrary to what
found for $\beta_{Z,50}^{ZC}$, the maximum value of $\beta_{V,50}^{ZC}$ decreases when lowering the temperature, which is best in the 200-300 K range (when the pinch-off behavior of the transistors is optimum), since it is also degraded at high $T$ [19]. As observed in Figure 6(b), this happens because $\beta_{V,50}^{ZC}$ can be approximated by the value given by eq. (17), -100 $\gamma$ (its maximum is always found in conditions of strong mismatch), so that its dependence with $T$ is much the same as that of $\gamma$.

Regarding the noise equivalent power $NEP_{ZC}$ calculated with eq. (2), Figs. 5(g), (h) and (i), the optimum values are obtained, regardless of $L_G$, approximately for the same $V_{GD}$ as the maximum of $\beta_{I,50}^{ZC}$ (around $V_{th}+0.2$ V), again because the higher resistance degrades the $NEP$. We have chosen to represent the value of $NEP_{ZC}$, but the values of $NEP_{ZV}$ are nearly the same above threshold, so that the optimum values of the $NEP_{ZC}$ and $NEP_{ZV}$ coincide. Figure 6(a) shows that the $NEP$ of our devices increases with $T$, taking very competitive values [19], around 1 pW/√Hz at 20 K rising to around 5 pW/√Hz at 400 K, and are (slightly) better for $L_G=250$ nm.

Only a slight dependence of the zero-bias detection figures of merit on $L_G$ is observed in Fig. 6. Since a better pinch-off behavior leads to an enhanced non-linearity, a slight degradation when reducing $L_G$ is observed due to short-channel effects, thus decreasing $\gamma$ and consequently $\beta_{V,50}^{ZC}$, $\beta_{I,50}^{ZC}$ and $NEP$ for the shortest $L_G=75$ nm. The other side of the coin is that, even if a longer gate improves the low-frequency detection, it reduces the cutoff frequency of the resistive mixing detection mechanism, so that a tradeoff between both of them must be chosen.

IV.2 Biased detection

Even though zero-bias operation is preferable in terms of $NEP$ (no excess noise is generated), biased detection allows to improve the responsivity by purposely locating the Q-point where the $I_D-V_{DS}$ curves exhibit their optimum non-linear behavior [40]-[42]. Figure 7 shows the different figures of merit calculated with the QS model for $L_G=250$ nm at $T =300$ K as a function of $V_{DS}$ for different $V_{GS}$. Figure 7(a) shows the same $I_D-V_{DS}$ curves of Figure 2(c), but this time representing $|I_D|$ in log scale, so that the ZC point ($I_D=0$) is
showed that the presence of a drain leakage current and its competition with TQC fixes the zero-crossing of allowing to strongly enhance the current-mode responsivity width of the transistors, which decreases their resistance, thus with the 50\textsuperscript{Ω} [43]. However, the voltage-mode responsivity is not improved because the lower input losses are compensated by the smaller resistance of the transistor, see eq. (16). As a consequence, the maximum voltage responsivity is always near the constant value predicted by eq. (17), corresponding to the strongly-mismatched case, since it is obtained near pinch-off.

Regarding γ, Figure 7(d), in subthreshold conditions its maximum value is attained at the onset of the TQC, for the (negative) $V_{DS}$ value corresponding to the ZC point. The dependence of γ on $V_{DS}$ is similar while $V_{GS} < V_{th}$, reaching a similar minimum of about -30 V\textsuperscript{ℓ} with just a shift of $V_{DS}$ parallel to the $V_{GS}$ increase. On the other hand, for $V_{GS} > V_{th}$ the maximum of γ decreases and appears for positive $V_{DS}$ values (at the knee of the $I_D - V_{DS}$ curve, where the increase of $R_D$ is less pronounced than for TQC), always shifting to higher $V_{DS}$ as $V_{GS}$ increases. This dependence of increase of γ allows to straightforwardly explain the values of the voltage responsivity (γ $\beta_{V50}$) at the onset of the TQC, where $R_D$ is in the kΩ range, Fig. 7(b). Second, for $V_{GS} > V_{th}$, γ and $\beta_{V50}$ ($I_D$) rapidly decrease, even if the mismatch is not as large as in subthreshold conditions.

On the other hand, the values of $\beta_{1.50}$ ($V_{DS}$) obtained with eq. (15), shown in Figure 7(f), always exhibit a bell shape for each given $V_{GS}$. However, even though γ sharply decreases for $V_{GS} > V_{th}$, in this case its decrease is completely compensated by the lower mismatch, so that the optimum $\beta_{1.50}^{QS}$ is around 1.5 A/W almost independently of $V_{GS}$, and even slightly increases just above $V_{th}$. The optimum $V_{DS}$ bias point is indicated with circular symbols in Figures 7(b), (c) and (f), showing that in subthreshold the peak value for $\beta_{1.50}$ ($V_{DS}$) takes place around $R_D \approx 284$ Ω (corresponding to a mismatch of about 50%). The reason why $\beta_{1.50}^{QS}$ slightly improves for $V_{GS} > V_{th}$ is the enhanced power matching (reaching about 72%). These dependences on $V_{DS}$ and $V_{GS}$ obtained by the QS model have been confirmed by the experiments shown in Figure 8, where the theoretical current responsivities in biased conditions are quite favorably compared with the measured values at 1 GHz for all the three different gate lengths at $T=300$ K. Remarkably, the maximum value of $\beta_{1.50}$ hardly depends on $L_G$, taking values of around -1.6 A/W.

It must be noted that the impedance mismatch of the devices with the 50 Ω source can be improved by increasing the gate width of the transistors, which decreases their resistance, thus allowing to strongly enhance their current-mode responsivity [43]. However, the voltage-mode responsivity is not improved because the lower input losses are compensated by the smaller resistance of the transistor, see eq. (16). As a consequence, the maximum voltage responsivity is always near the constant clearly visible. The drain voltage corresponding to the zero-current point (indicated by the vertical dashed lines for each $V_{GS}$), shifts to the left in subthreshold conditions ($V_{GS} < V_{th}$, with $V_{th} \approx -3.7$ V in this case). In our previous work [37] we showed that the presence of a drain leakage current and its competition with TQC fixes the zero-crossing of $I_D$ in subthreshold. The lower the drain leakage the stronger the phenomenon is, so that short gate lengths (giving rise to short-channel effects) and low temperatures (buffer acceptor ionization level decreases) significantly suppress this voltage shift of the ZC point.

Figure 7(b) shows the values of the differential resistance, $R_D$, taking a minimum value of 72 Ω in open channel and TQC conditions, sharply increasing with $V_{DS}$, reaching the MΩ range when $V_{GS}$ is in subthreshold or up to some kΩ when saturation in the 1\textsuperscript{st} quadrant is attained. Conversely, the mismatch factor $\left(1 - |\Gamma|^2\right)$ calculated from the previous values of $R_D$, Figure 7(c), goes, when $V_{DS}$ is increased, from almost perfect matching (97%), to almost 0 in subthreshold or to around 20% for high $V_{GS}$ (open channel), with the curves shifted to higher $V_{DS}$ when $V_{GS}$ is raised.

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qualitative analysis and conclusions extracted here for the HEMT with $L_G=250$ nm are general and valid also for the devices with $L_G=150$ nm and $L_G=75$ nm. It is also interesting to remark that the current responsivity $\beta_{1.5}$ has much smaller variability due to the fact that it goes to zero when approaching subthreshold conditions, so that its values are much less sensitive to the dissimilarities on the pinch-off behaviour of the devices.

The main result of this section is that the RF detection performances of the HEMTs cannot be improved by biasing the devices, neither in voltage-mode nor in current-mode operation. In voltage mode, the optimum $V_{DS}$ is obtained by the spontaneous self-bias at the zero-current operation point (which provides the best non-linearity), while in current mode the maximum responsivity (which coincides with the optimum NEP) can be obtained with $V_{DS}=0$ by adequately choosing the $V_{GS}$ value to around $V_{th}+0.3$ V.

V. CONCLUSIONS

The QS-model employed in this work is a simplistic and easy way to understand the RF detection performances at low frequencies (while the resistive mixing mechanism based on the non-linearity of the devices holds). In this work it has been applied to obtain the responsivity and NEP of GaN HEMTs operating as RF detectors both in current- and voltage-mode and in zero-bias and biased operation in a wide temperature range, correctly reproducing the experimental results.

There are marked differences between the two detection schemes. When the voltage responsivity is measured, the best response is presented at the ZC point. Under this detection mode scheme, if the transistor has a good pinch-off, the bias point is shifted to the $V_{DS}$ corresponding to the onset of TQC, thus maintaining a high responsivity in subthreshold. On the other hand, leaky devices (due to short channel effects or non-ionized buffer acceptors at low $T$) have low voltage responsivity in subthreshold regime due to their linearity around the ZC point. Regarding the dependence with $T$, the maximum voltage responsivity is attained at intermediate temperatures, associated to an optimum non-linear behavior of the transistors. Conversely, when operating in current-mode, the responsivity is null in subthreshold conditions. A maximum value around 1.5-1.6 A/W is obtained for $V_{GS} \approx V_{th}+0.3$ V at $T=300$ K, increasing to about 2 A/W at low $T$, almost independently of the gate length.

Finally, no improvement of the RF detection performances can be obtained by means of drain biasing, the optimum zero-bias detection conditions can be achieved just by choosing the adequate value of $V_{GS}$.

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