# Noise tunability in planar junction diodes: theory, experiment and additional support by SPICE

J. I. IZPURA[1] and J. MALO[2]

*Grupo de Microsistemas y Materiales Electrónicos, Universidad Politécnica de Madrid. Spain.*
*[1]Departamento de Ingeniería Electrónica, ETSI Telecomunicación.*
*[2]Departamento de Sistemas Electrónicos y de Control, EUIT Telecomunicación.*

*This document deals with a rather unknown feature of planar junction diodes: their bias-dependent dynamical resistance making them Lorentzian noise generators whose lifetime can be tuned electrically. This feature, useful to produce electrically coloured noise, also is the basis of a 1/f noise synthesis process taking place in the channel of resistors while their resistance noise is measured. From the above, the main question this paper answers is: Do we teach our engineers the proper way to measure resistance noise in resistors? And the surprising answer this paper gives is: Not at all.*

## 1. Introduction

The excellent work of H. Nyquist to explain thermal noise in resistances [1], applied to the noise measured by J. B. Johnson in different resistors [2], has led to a misconception making the magnitude resistance equivalent to the resistor device. Although resistance R in general is the real part of a frequency-dependent impedance function $Z(j\omega)=R(\omega)+jX(\omega)$, we will focus on its dc value $R_{dc}=R(\omega\rightarrow 0)$ defined by Ohm´s Law: $R_{dc}=V/I$, where V is the dc voltage applied to the resistor *device* and I is the current flowing throught it. This will be so because the $R_{dc}$ of an homogeneous resistor and the resistivity $\rho$ ($\Omega\times$cm) of its inner material are linearly related by geometrical parameters that define the way the electrical field of V is applied to such *device*. The word *device* has been emphasised because it plays a key role that is the origin of several effects of great interest concerning resistance noise, overseen in the course of the years.

To keep a close link with the teaching of Electronics we have chosen a simple device (a Schottky diode), whose voltage-dependent capacitance is a known feature easily predicted by differentiating the formula that gives the width of the Space-Charge Region (SCR) of their one-sided junction as a function of its bias voltage [3]. A less known feature of Schottky diodes is their dynamical resistance $r_d(V_Q)$ that depends exponentially on the quiescent voltage of the diode $V_Q$. This results from differentiating their exponential current-voltage (i-v) characteristic [3], also present in other type of junctions whose experimental handling not always is easy. The very high values of $r_d(V_Q)$ in p-n junction diodes (well over the G$\Omega$) makes difficult such handling. These high values come from the high built-in voltage $V_{bi}$ of Silicon or GaAs-based junction diodes [4], but when low-$V_{bi}$ Schottky diodes are used the above resistances fall below the M$\Omega$ and they can be handled easily by low cost electronics.

Using commercial Silicon Schottky diodes we have measured the effects of a dc bias voltage in the kT/C noise of their junction capacitor, mainly its tunability, that also has been confirmed by SPICE with the models available from the semiconductor industry. The proper understanding of this tunable noise allows to explain electrically other Low-frequency noises as the 1/f electrical noise of Field Effect Transistors (FETs) and planar resistors and this new view of the *excess noise* found in conductors reveals that resistance noise measurement performed in the past have been creating their own 1/f electrical noise from a backgating noise overseen in the course of the years. This noise due to interfaces limiting conductors is Lorentzian in thermal equilibrium (e.g. while it is not measured) but becomes a 1/f noise as soon as a bias current or voltage is applied to the conducting channel to convert resistance noise into voltage noise able to excite an spectrum analyzer. The main lesson derived from the above is that we have to warn our students (and ourselves first) against the innocent use of Ohm´s Law to convert resistance noise into voltage noise able to excite a spectrum analyzer. The voltage disturbing the sample during the experiment must be taken into account because it will bias any device existing in the sample, usually one or more Field-Effect transistors with the gate left floating (FGFET).

## 2. The kT/C noise of a junction diode: a tunable Lorentzian spectrum

Although it is known long time ago that *the thermal noise of a capacitance C shunted by any resistance always gives the same mean squared voltage noise* $\langle v_n(t)^2\rangle = kT/C$ [5] (k is the Boltzmann constant) their consequences have not been fully exploited yet. In a Sample and Hold system storing voltage samples in a capacitor C=1pF, the power $\langle v_n(t)^2\rangle = kT/C$ ($V^2$) sets a root mean squared voltage noise $v_{rms}=64\mu V$ with zero mean (pure ac) superimposed to any dc voltage value stored in the capacitor. For this capacitor holding a dc voltage sample of 20μV, this added kT/C noise is a true problem because the voltage read after the sampling action will be any value around 20μV dc with a variance of 64μV. Thus, kT/C noise becomes a problem in small capacitances as those present in Charge Coupled Devices [6]. Since actual devices and integrated circuits are plenty of planar capacitors (both intentional as well as parasitic ones), the proper knowledge of thermal noise in planar junction capacitors is of capital interest. For small-signals as noise, a reverse biased Schottky diode is a parallel RC circuit whose capacitance C comes from the junction SCR and whose shunting resistance R will contain any leakage resistance $R_{leak}$ of the diode in parallel with its dynamical resistance $r_d(V_Q)$. Using this ideal i-v characteristic (ideality factors other than unity give similar results):

$$i = I_{sat} \times \left( \exp\left(\frac{v}{V_T}\right) - 1 \right) = I_{sat}[\exp(u) - 1] \qquad (1)$$

the dynamical resistance of the diode is:

$$r_d(v) = \left(\frac{\partial i}{\partial v}\right)^{-1} = \frac{V_T}{I_{sat}} \exp(-u) = r_{d0} \times \exp(-u) \qquad (2)$$

where u is the bias voltage v measured in thermal voltage units $V_T=kT/q$ ($V_T \approx 26mV$ at room T).

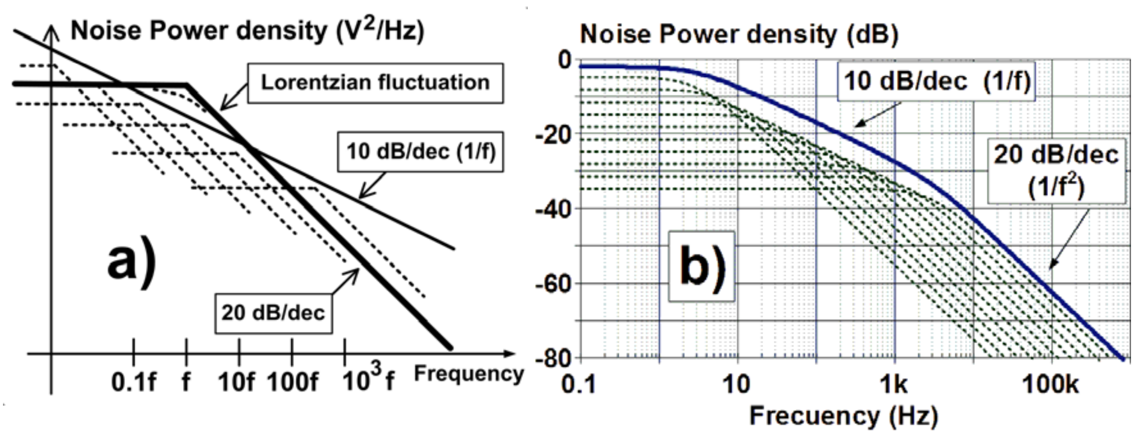


**Figure 1.** a) Theoretical noise of a Schottky diode. b) Set of Lorentzians synthesizing 1/f noise.

From equation (2), a small reverse bias v=-60mV will set the dynamical resistance of the diode to a ten times higher value ($10r_{d0}$) than the Thermal Equilibrium (TE) $r_d$ with no bias applied ($r_{d0}$), while its capacitance value will remain roughly the same (actually C will decrease slightly as it happens in varactor diodes). Considering a high $R_{leak}$ in parallel with a much lower $r_d(V_Q)$ (hence the reason to choose a low-$V_{bi}$ Schottky junction), the RC parallel circuit of our Schottky diode will have a kT/C noise power of: $\langle v_n(t)^2\rangle = kT/C$ ($V^2$) whose spectral distribution will be *shaped* by C and $r_d(V_Q)$ in parallel. Therefore, the power density $S_v(f)$ ($V^2$/Hz) of this noise measured between terminals of the Schottky diode will be this Lorentzian spectrum:

$$S_v(f) = \frac{4kTr_d(v)}{1+\left(2\pi fCr_d(v)\right)^2} = \frac{4kTr_d(v)}{1+\left(f / f_C(v)\right)^2} = \frac{4kT}{2\pi C} \times \frac{f_C}{f_C^2+f^2} \quad (3)$$

This spectrum is the Bode plot depicted in Fig. 1-a by a thick solid line coming from this first order, low-pass, R-C filter. The flat part of this spectrum below its cut-off frequency $f_C$ *must be numerically equal to the thermal noise of the shaping resistance R that shunts C for small-signal no matter if R includes noiseless resistances as $r_d(v)=(\partial i/\partial v)^{-1}$*. Otherwise, the kT/C noise power of a capacitance C shunted by no matter which resistance would not be conserved and *this noise power conservation law* for a given C [5, 6] is the reason why the kT/C noise of a Schottky or junction diode *must show a tunable Lorentzian spectrum* whose flat power density below $f_C$ is inversely proportional to $f_C$, as stated by equation (3). This has been depicted in the Bode plots drawn by dashed lines in Fig. 1, that form a kind of ladder whose steps are aligned parallel to the 1/f thin line (10dB/decade slope) of Fig. 1. This is so because equation (3) is a *power density*, that is converted to dB by $10\log(S_v)$ and not by $20\log(X)$ that is used when X is voltage or current. This feature is very interesting to synthesize 1/f noise from a set of these Lorentzian noise terms, since *the graphical requirement to achieve such synthesis is to have a constant step height in the ladder* of a Bode plot. For a discrete set of such Lorentzian terms this means that the cut-off frequencies of each Lorentzian must be spaced logarithmically in the band where the 1/f region of noise must be synthesized, a kind of spacing occurring naturally in the conducting channel of actual resistors far from pinch-off conditions as we will show.

### 3. Experimental tunability of the noise in a BAT85 Schottky diode

An experimental demonstration of the electrical tunability of the noise spectrum of a junction in general and that of a metal-semiconductor junction in particular, was obtained in commercial Si-Schottky diodes some time ago [7]. From the forward voltage $V_f$=235mV and dc current $I_f$=1.23mA measured in a BAT85 diode at room T ($V_T$=25.9mV), equations (1) and (2) give: $I_{sat}=1.4\times10^{-7}$A and $r_{d0}$=185KΩ, two useful data to design the circuit of Fig. 2. For C=10pF (from BAT85 datasheets) equation (3) gives: $f_0=f_C(v=0)$=86KHz, a high *native* $f_0$ that was lowered to $f_{01}$=183Hz by a fixed capacitor of 4700pF in parallel with the BAT85. This new $f_{01}$ is better suited to the input characteristics of our Low-Noise Amplifier (LNA) and below 25KHz, the cut-off frequency of the anti-aliasing filter of our noise analyzer described in [8]. The high value of this added capacitor also makes negligible any small change of C with $V_Q$, a beneficial feature we used in [4] to measure comfortably the thermal activation of $r_d(V_Q)$ in GaAs Shottky junctions with an activation energy $E_T \approx qV_{bi}$ (q is the electronic charge).

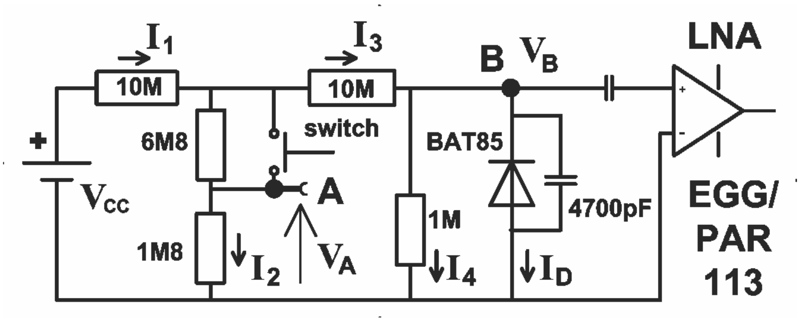


**Figure 2.** Electrical circuit used to bias the Schottky diode and to measure its noise.

Fig. 2 shows the circuit used to measure voltage noise in the BAT85 biased by a small battery ($V_{CC}$) enclosed in the shielded test fixture. It offers a rather high and constant resistance $R_L \approx 1$MΩ in parallel with the diode, thus producing a rather low loading effect on $r_{d0}$=185KΩ of this "C-enhanced" diode. The switch allows to select two reverse bias in the diode: $V_Q \approx -V_T$ and $V_Q \approx -7V_T$ as it can be seen in Table I, where also are shown two voltages $\mathbf{V_A}$ and $\mathbf{V_B}$ *measured* on the circuit with a Keithley-195A voltmeter (1GΩ load resistance *for input voltages below 2V*, hence the circuit used) that allow to *calculate* the currents $I_1$-$I_4$ and $I_D$. These $V_Q$ values together with $V_Q$=0 ($V_{CC}$=0) will show the bias dependence of $r_d(V_Q)$, whose values for $I_{sat}$≈0.208μA (see $I_D$ for the diode under high reverse bias) are: $r_{d0}$=125KΩ, $r_d(-V_T)$≈340KΩ and $r_d(-7V_T)$≈158MΩ. This would produce clear shifts in the $f_C(V_Q)$ of the BAT85 noise spectra from $f_{01}$=271Hz to $f_{02}$=100Hz and $f_{03}$=0.21Hz if the loading effects of $R_L$=1MΩ on $r_d(V_Q)$ were low. This is quite true for $r_{d0}$=125KΩ, but not for $r_d$=158MΩ. In this case a noise power spectrum shaped by 1MΩ in parallel with 4700pF would result with $f_{03true}$≈158×0.21=33Hz, as well as a 158 times lower flat noise spectral density corresponding to 1MΩ and not to 158MΩ (the $R_{in}$=100MΩ of the LNA in parallel with $R_L$=1MΩ has been neglected).

| **$V_{CC}$=12.38V** | **$V_A$ (V)** | **$V_B$ (V)** | $I_1$ (μA) | $I_2$ (μA) | $I_3$ (μA) | $I_4$ (μA) | $I_D$ (μA) |
|---|---|---|---|---|---|---|---|
| **Swich ON** | **1.630** | **0.025** | 1.075 | 0.906 | 0.169 | 0.025 | 0.144 |
| **Switch OFF** | **0.818** | **0.185** | 0.847 | 0.454 | 0.393 | 0.185 | 0.208 |

**Table I.** Measured voltages (bold) and calculated currents in the circuit of Fig. 2.

The measured noise spectra shown in Fig. 3 confirm the above predictions, since curve a) is the noise spectrum measured in the circuit of Fig. 3 with the input of the LNA shorted, thus being a reference about the lowest power spectral density (in $V^2$/Hz) that our system can handle due to the equivalent input noise $(e_n)^2$ of the old LNA used. Curve b) is the Lorentzian spectrum measured for $V_Q$=0 ($V_{CC}$=0) whose $f_{01exp}$≈300Hz is not far from $f_{01}$. Curve c) is the spectrum measured for $V_Q$=-$V_T$ (switch ON), whose $f_{02exp}$≈95Hz agrees with $f_{02true}$=133Hz obtained by considering that $R_L$=1MΩ in parallel with $r_d(-V_T)$=340KΩ gives 254KΩ as the shaping resistor for this noise. Curve d) is the noise spectrum measured for $V_Q$=-7$V_T$ (switch OFF), whose $f_{03exp}$≈35Hz agrees well with the theoretical $f_{03true}$≈33Hz.

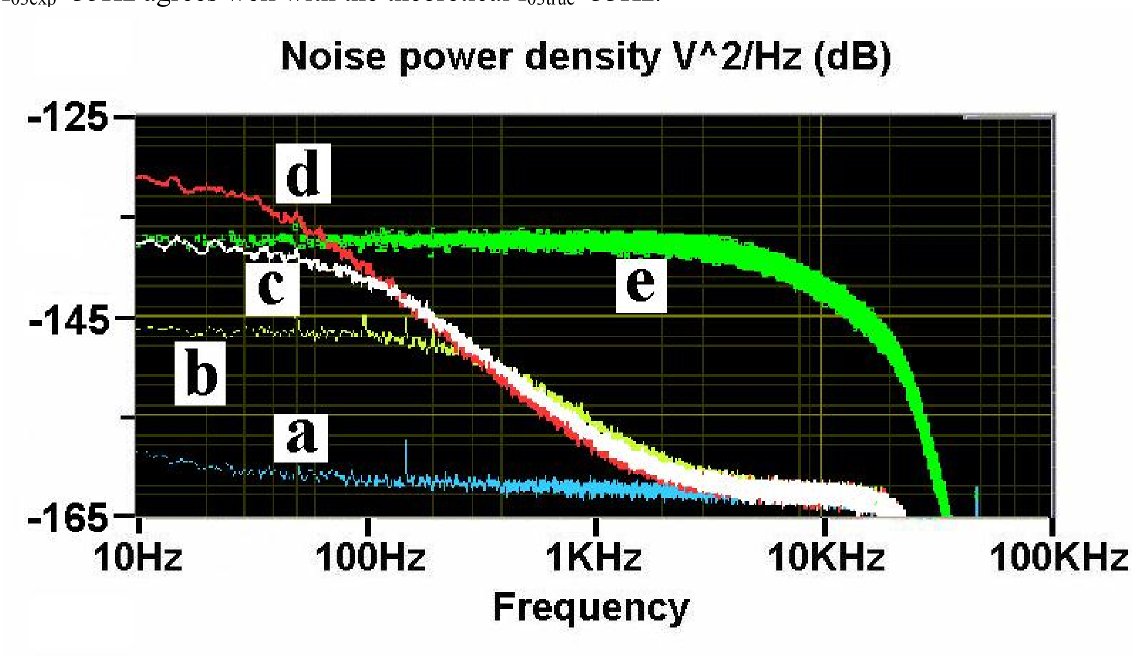


**Figure 3.** Noise spectra measured in the circuit of Fig. 2 under different conditions (see text).

Curve e) in Fig. 3 is a final check that revealed *an added contribution* to the kT/C noise of the diode. It shows the noise measured in the circuit of Fig. 2 (Switch OFF) *with the BAT85 and the 4700pF capacitor removed*. This was done to check this: since $r_d(-7V_T)$=158MΩ in parallel with $R_L$≈1MΩ *vanishes*, the flat part of curve d) would have to be equal to that of curve e). But the flat part of curve d) ($S_v$≈-131dB, 7.94×$10^{-14}$$V^2$/Hz) is *over*, the familiar thermal noise of 1MΩ resistor (-137.8dB, 1.66×$10^{-14}$ $V^2$/Hz) often used to check the system and whose drop around $f_{drop}$≈6KHz is due to the parasitic capacitance ($C_{in}$≈25pF) of the input of the LNA.

Thus, there is *more noise* in curve d) *than the expected one from the above argument.* The reason is that although $r_d(-7V_T)=158M\Omega$ *vanishes* in parallel with $R_L \approx 1M\Omega$, *the shot noise of the diode current* $I_D(-7V_T)=0.208\mu A$ (Table I) *does not vanishes at all*, and moreover, its power spectral density: $2\times q\times I_D=6.7\times10^{-26}A^2/Hz$, will be converted to a voltage power density through $|Z(j\omega)|^2=(R_L)^2$ in this case. This gives: $6.7\times10^{-14}V^2/Hz$, that will add in power with the thermal noise of $R_L=1M\Omega$, thus giving $S_n(f<34Hz)=8.3\times10^{-14}V^2/Hz$ (-130.8dB) for curve d) that explains perfectly the extra noise found in the reverse biased diode. Therefore, Schottky diodes are Lorentzian noise generators, easily *tuned by small bias voltages* $V_Q \approx V_T$, as equations (2) and (3) predict, but this noise *not only includes the kT/C noise* of the diode capacitance existing in TE ($V_Q=0$), but *also includes the shot noise* of any current existing in the diode for $V_Q \neq 0V$.

Hence, we have to be very cautious with this tuning feature because any bias voltage $V_Q$ applied to the diode will take it out of TE, thus setting some bias current $I_Q$ (forward or reverse), whose *shot noise will add to the Johnson noise* of any *leakage resistance* $R_{leak}$ assumed to *justify* its kT/C noise. This *noisy* $R_{leak}$ always can be assumed in parallel with $r_d(v)$ (*noiseless*) to obtain a noise power in C: $\langle v_n^2\rangle=kT/C$ ($V^2$) *no matter the R value used*. Taking $R_{leak}=10^{40}\Omega$ that means $R_{leak}\rightarrow\infty$ from a practical viewpoint, we *justify electrically* the kT/C noise in the diode under any small reverse bias, *besides* the shot noise due to any $I_Q$. This shot noise will be a broadband noise (due to the short transit time of carriers across the junction) that will add in power (uncorrelated noises) to the Johnson noise of $R_{leak}$ filtered by C (the kT/C noise of C). The final result is *a unique Lorentzian noise spectrum shaped by C and $r_d(v)$* comprising both the $4kTr_d$ noise ($V^2$/Hz) and the contribution of the shot noise $2qI_Q$ ($A^2$/Hz), converted to voltage noise ($V^2$/Hz) through $[r_d(v)]^2$. Readers familiar with *degrees of freedom* to store energy in the system (e.g. electrostatic energy in the RC cell) may not need to assume $R_{leak}\rightarrow\infty$, but just a capacitance C *in thermal equilibrium at temperature T* to admit its kT/C noise as a result of the equipartition theorem [6]. The carriers in thermal contact with C to keep it at temperature T could be seen as the origin of $R_{leak}\rightarrow\infty$ if desired, but what matters for the filtering of the noise is $r_d(v) << R_{leak}$.

## 4. Computer simulation by SPICE of the noise in a BAT85 Schottky diode

Let us show how the electrical simulator SPICE handles this electrical tunability for the noise of a Schottky diode BAT85. This will give some added support to the above theory that despite its simplicity, seems hard to be understood [9], probably due to its radical departure [10] from today's views on low-frequency electrical noise [11]. Using the BAT85 SPICE model provided by Philips [12] (see the Appendix) we have simulated the noise spectra of the circuit of Fig. 2 for the same experimental conditions we used in Fig. 3. It is worth noting that this model uses $I_{sat}=2.117\times10^{-7}A$, very close to the value $I_D=0.208\mu A$ shown in Table I for the diode under clear reverse bias (switch OFF). This value so close to $I_D$ and not far from the initial $I_{sat}=1.4\times10^{-7}A$ obtained from equation (1) applied to one point of the I-V curve of the diode, gives an added support to the experimental data of previous Section. Table II shows the voltages and currents obtained by SPICE in the circuit of Fig. 2 for $V_{CC}=12.38V$. The good agreement with Table I indicates that the SPICE model used for the BAT85 diode is very accurate for dc.

| **$V_{CC}$=12.38V** | **$V_A$ (V)** | **$V_B$ (V)** | $I_1$ (μA) | $I_2$ (μA) | $I_3$ (μA) | $I_4$ (μA) | $I_D$ (μA) |
|---|---|---|---|---|---|---|---|
| **Swich ON** | 1.642 | 0.026 | 1.074 | 0.912 | 0.162 | 0.026 | 0.135 |
| **Switch OFF** | 0.830 | 0.164 | 0.841 | 0.461 | 0.380 | 0.164 | 0.216 |

**Table II.** Voltages and currents calculated by SPICE in the circuit of Fig. 2 for $V_{CC}=12.38V$.

From the noise viewpoint, Fig. 4 shows the noise spectra given by SPICE for the same bias conditions of curves b), c), d) and e) of Fig. 3. The first feature that appears in Fig. 4 is that *SPICE also shows the tunability* already predicted. The flat curve e') corresponds to curve e) in Fig. 3, whose drop above 6KHz was explained in previous Section. Both curves e) and e') have a -138dB flat region. Since curve e') corresponds to the circuit of Fig 2 *with the BAT85 and the 4700pF capacitor removed*, we observe that SPICE calculates well noise in resistors.

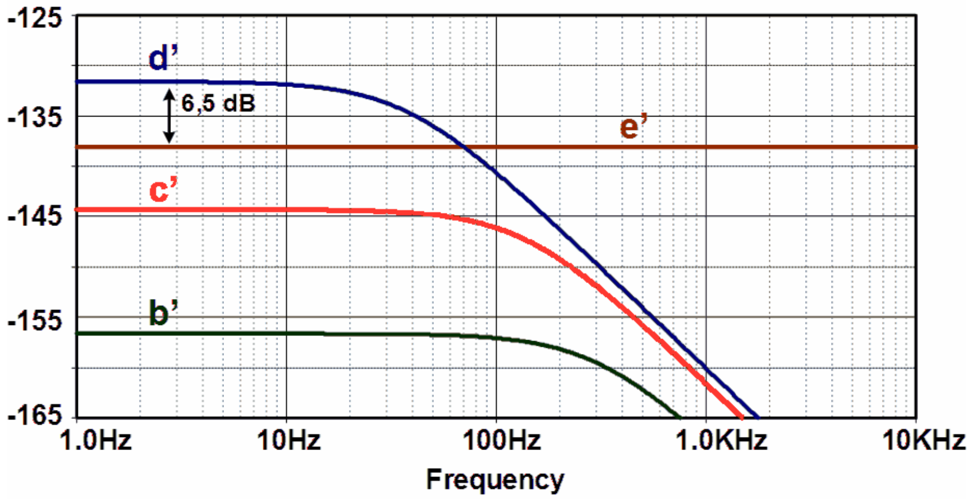


**Figure 4.** Noise spectra obtained by SPICE for the circuit of Fig. 2 under different conditions (see text).

Curve d') from SPICE and curve d) in Fig. 3 (measurement) look similar. The values given by SPICE: $f_{03SIM}$=37.7Hz and $S_{vSIM}$=-131.58dB (flat region) agree very well with the measured ones: $f_{03exp}\approx$35Hz and $S_v$=-131dB, that we handled in previous Section to show the contribution due to the shot noise. From this agreement we observe that SPICE simulates well thermal noise and shot noise in the circuit, *when the diode has a rather high reverse bias* (seven thermal units $V_T$ in this case). But when we approach thermal equilibrium (v=0V), things are different. Curve c') in Fig, 4 corresponds to curve c) in Fig. 3 and its $f_{02SIM}$=137.2Hz agrees with $f_{02true}$=133Hz obtained by considering that $R_L$=1MΩ in parallel with $r_d(-V_T)$=340KΩ gives 254KΩ as the shaping resistor for this noise. What differs is the spectral power density given by SPICE: $S_{vSIM}$=-144.3dB: flat region of curve c') in Fig. 4 and $S_{vexp}\approx$-138dB, flat region of curve c) in Fig. 3. This difference (roughly 6dB) would indicate that SPICE is giving *the right cut-off frequency but a lower noise power than the actual one*. This is somewhat surprising after the good agreement between curves d') (SPICE) and d) (experiment) and hence, we have studied the reason for this departure. A clue for this study comes from curve b') obtained by SPICE for $V_{CC}$=0V whose $f_{01SIM}$=310Hz agrees with $f_{01exp}\approx$300Hz of the previous Section ($f_{01}$=307Hz, if we consider, as SPICE does, that the loading effect of the circuit on the diode is not $R_L\approx$1MΩ but $R_L$=936KΩ). Again, SPICE is giving *the right cut-off frequency but a lower noise power than the measured one*, because $S_{vSIM}$=-156.6dB for the flat region of curve b') roughly is 9dB lower than $S_{vexp}\approx$-147dB, flat region of curve b) in Fig. 3.

**5. Discussion on the suitability of SPICE for noise calculations**

These low noise values given by SPICE correspond to the thermal noise of a noisy or dissipative resistance $R_L$ given by the Nyquist formula [1], connected in parallel with a noiseless resistance $r_d(v)$ with no thermal noise at all, as it can be checked easily by hand. In this way the noise of the noisy resistance is attenuated by the noiseless resistance, that in turn, does not add thermal noise at all. This produces *the right shaping resistance for the noise spectrum* ($R_L$ and $r_d(v)$ in parallel), but *the noise magnitude of the diode for low bias voltages is underestimated by SPICE*. And the reason for the above is the simple noise model used by SPICE, summarized in the noisy and noiseless resistances just commented, that also is the most extended view among researchers and teachers. From this viewpoint, as the reverse bias in the diode is reduced, $r_d(v)$ decreases markedly and attenuates strongly any thermal noise of noisy resistances in parallel, thus giving a low noise level.

If the above view about noise in circuits was true, the next question is: Why the noise given by SPICE is lower than the experimental noise?. And the answer involves the way SPICE handles shot noise, that only is accurately when the diode is under forward bias or under high reverse bias as we are going to show. To obtain the shot noise of the diode, SPICE uses *the net* *current in the diode* $I_D$ to calculate the current noise density $2qI_D$ ($A^2$/Hz) that is converted to a voltage noise density through the impedance of the circuit. *But this approach fails for the diode biased with v=0V* because it gives a null shot noise for the diode having a null $I_D$ on average, *achieved by two equal and opposed currents $I_{sat}$ each having its own shot noise*. Therefore, the shot noise of the diode at v=0 is: $2\times2qI_{sat}$ ($A^2$/Hz) no matter its null $I_D$, a result that is coherent with equation (3), that predicts a noise power density in the diode that will be a Lorentzian of amplitude $4kTr_{d0}$ ($V^2$/Hz) for v=0V. Converting this shot noise that SPICE does not consider: $4qI_{sat}$ ($A^2$/Hz), into a voltage noise power density through $(r_{d0})^2$ we have: $4qI_{sat}(V_T/I_{sat})^2=4kTr_{d0}$. In this way the noise of a junction diode in TE can be seen as the kT/C noise of its capacitance or the shot noise of its two equal and opposed $I_{sat}$ added in power (uncorrelated), but this is a view falling out of SPICE's capabilities. This fact is somewhat surprising because the *diode noise model* of Van der Ziel cited in [6] uses a shot noise term $2q(2I_{sat}+I_D)$ that points towards a careful study of the inner currents in the device from the noise viewpoint. With today's electronics moving towards low-voltage circuits, perhaps the time has come to improve the simulator SPICE paying attention to this mishandled noise term.

Adding manually this shot noise: $2q(2I_{sat}+I_D)\times[r_{d0}(v)]^2$ to curves b') and c') in Fig. 4, their new flat regions appear at -147dB and -141.6dB respectively, thus fitting well the experimental curves b) and c) of Fig. 3. It is easy to understand that under *clear forward bias*, we have: $I_D>>2I_{sat}\Rightarrow(2I_{sat}+I_D)\approx I_D$ and the approach used by SPICE is quite valid. Under *high reverse bias* the net current in the diode is: $I_D=-I_{sat}$, making $(2I_{sat}+I_D)=I_{sat}$ that validates the net current approach used by SPICE to calculate shot noise as we observed with curves d') and d). But *for a few thermal voltage units $V_T$ (forward or reverse) biasing the diode, its shot noise will be underestimated*, being this a drawback for the use of SPICE in electrical noise research.

## 6. Application to the measurement of resistance noise: electrical origin of 1/f noise.

This section consider a very basic device that exists in most planar resistors where Low-frequency noise is measured and where an *excess noise* is found, usually having a 1/f spectrum rather than the Lorentzian one we have shown to exist in Schottky diodes. The same kind of noise can be predicted for other planar junctions as p-n or p-i-n homojunctions, heterojunctions as that of the Shottky diode and so on. Such is the case of the junction existing between any semiconductor epilayer and the underlying substrate (or layer) having different conductivity to confine the carrier flow in the upper epilayer. Let us think about an n-type epilayer onto a p-type substrate, although a semi-insulating substrate is equally valid. Let us consider too that the substrate doping is high, thus allowing to consider it as an equipotential region similarly to a metallic layer. Therefore, the junction existing between the substrate and the epilayer will have and electrical field orthogonal to the epilayer and a SCR will exist at the substrate-epilayer interface. Let us suppose that we have fabricated (e.g. by mesa etching) a rib in the epilayer to make a strip resistor in the onto the p-type substrate. It is quite apparent that *we don't have a pure resistor but a conducting channel coupled with a FET-like gate that is the underlying substrate*. And *let us also leave floating this gate* to measure electrical noise in the resistor strip, no matter that a FET gate left floating is not very advisable for such measurements.

The reason to consider this Floating-Gate Field-Effect Transistor device (FGFET) is because *most researchers use it inadvertently as soon as they forget to connect properly the underlying substrate to the upper resistors*. In this way, *a FGFET device is found in nearly all the resistors we can make*. Even a rod of conducting material not requiring an underlying substrate due to their own rigidity *will have a surface and a surface SCR* that converts it into a FGFET with a weak gate layer on the surface [8] able to produce 1/f resistance noise in these non-planar resistors. Fig. 5 shows the FGFET device that researchers use to have inadvertently while they focus their attention in the upper epilayer where the strip resistor or devices have been made. The gate G is thus the underlying substrate or layer, usually overseen. The strip resistor of thickness H, length L and width W has two ohmic contacts at its ends that we will call Drain (D) and Source (S). *As soon as we have defined the transversal area $A=W\times L$ of the resistor on top we have defined the area of the disturbing capacitance C existing between the* *resistor channel of thickness H and the floating gate* G. Once this capacitance is defined, it will have a kT/C voltage noise power with a Lorentzian spectrum as we have shown. But each small voltage fluctuation taking place in C requires a small modulation of the channel thickness H (FET action) and this *means that the resistance of the strip resistor will not be strictly constant*. Therefore, actual resistors will have a *mean resistance* $R_{ch}$ *together with a small resistance noise* that, with no bias applied, will have a Lorentzian spectrum. *No bias applied* means $V_{DS}$=0V or $I_d$=0, a condition that allows to consider as an equipotential region the resistor body (channel) as it also is considered the floating gate. This means that the junction between the gate and the channel has the same bias voltage (e.g. v=0V) at each position along the x axis in Fig. 5. Connecting Drain and Source together as the cathode and using the floating gate G (p-type substrate) as the anode of the junction diode, we can apply some bias voltage $V_Q$ to the junction under the strip resistor to tune its Lorentzian noise.

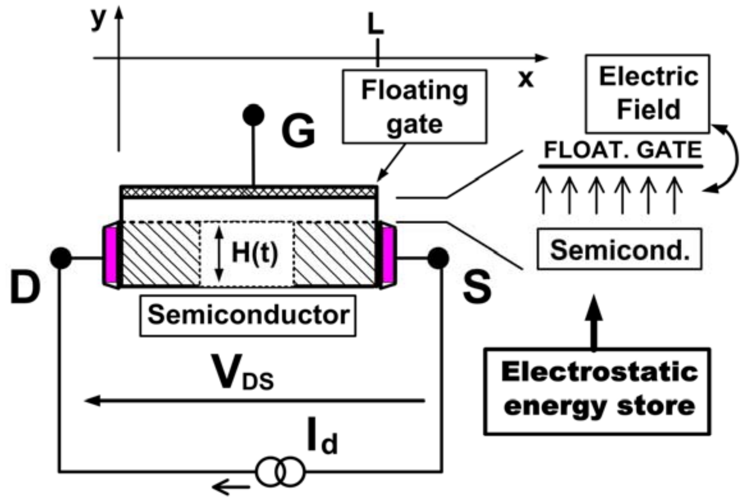


**Figure 5.** Floating-Gate FET device commonly used in noise measurements (see text).

Once this tunability has been accepted for the noise of the interface between the strip resistor sketched in Fig. 5 and the floating gate due to its underlying substrate, let us go check the *unsound statement on actual resistors having a non-constant resistance*, or in other words: having a small thickness modulation or an Interface-Induced Thermal noise (*IIT noise* in what follows). If this is so and such a Lorentzian resistance noise exists in the resistor of our example, we can try to measure it. But a spectrum analyzer connected between D and S terminals only will sense the thermal (Johnson) noise of the mean resistance $R_{ch}$ of our strip resistor, because it is a *voltage fluctuation* that such spectrum analyzer can sense. A small resistance noise as IIT noise does not produce any voltage fluctuation on its own and the resistance noise we want to measure won't be sensed by the above spectrum analyzer, except for a very sophisticated analyzer able to measure *the noise of the thermal noise* [10]. This is a difficult task due to the statistical nature of thermal noise that requires some *time averaging that attenuates this noise of the Johnson noise* due to small resistance fluctuations around the mean resistance $R_{ch}$.

Hence, *we need to convert the above IIT noise into a voltage noise emerging over the Johnson noise* existing between D and S terminals, and the traditional method used long time ago for this conversion has been to inject a dc current $I_d$ through the resistor (see Fig. 5) to convert any resistance fluctuation $\Delta R_{ch}$ into a voltage fluctuation $\Delta V=\Delta R_{ch}\times I_d$ by the well known Ohm's Law. But Ohm´s Law can not be applied in this way to achieve the expected conversion, because we don't have a resistance but a resistor (device) that is a FET. The reason is that as soon as $I_d$ is injected, a voltage drop $V_{DS}$ will appear along the channel. Since the gate region is equipotential, the planar junction along the channel becomes differently biased as we move along the x axis in Fig. 5. As it has been shown in [9, 10] the floating gate G will acquire a floating voltage $V_{float}$ between 0V and $V_{DS}$ in order make null the net current between the equipotential gate and the non-equipotential channel. This means that the junction aside the Source terminal (S) becomes forward biased by $V_{float}$ while it becomes reverse biased aside the Drain terminal D where the Drain-Gate bias voltage will be ($V_{float}$-$V_{DS}$).

The above situation means that *there is not a unique Lorentzian noise in the junction as we move along the channel* of the resistor strip. Viewing the channel as a set of slices connected in series, slices close to the Drain will have low cut-off frequencies in their planar junction due to their reverse bias while slices near the Source will have high cut-off frequencies due to their forward bias, as equation (2) and (3) state. Then, a continuous set of Lorentzian IIT noises appear along the channel and this set of resistance noise terms is viewed as (or synthesizes a) 1/f spectrum of resistance noise in the channel while its resistance noise is being measured [9, 10]. In summary: *Ohm's Law applied to the whole channel of actual resistors does not give the right conversion factor for their resistance noise because they are FGFET devices where the own current $I_d$ or voltage $V_{DS}$ applied to convert their IIT noise into voltage or current noise, affects the result obtained* because it bias differently each position of the resistor *device*. When this distributed bias and the backgating effect of the kT/C noise of disturbing capacitors are taken into account, *excess noise in resistors appears naturally, as well as its 1/f spectrum for $V_{DS}$ values of a few thermal voltages $V_T$* [9, 10]. To our knowledge, this is the first electrical explanation for 1/f electrical noise since 1925, when it appeared in noise measurements carried out by Johnson [14] in vacuum tubes (Audions). Our new theory also applies straightforwardly to Johnson's results, where the voltage drop along the filament to heat it would play the role of $V_{DS}$ to modify the cathode-grid potential barrier for thermoionic emission from the cathode of those Audions. The mathematical work behind the 1/f noise synthesis in the resistor channel will be published elsewhere [10] and here we will replace it by the Bode plot obtained by summing several Lorentzian noises *logarithmically spaced in the desired band* where the 1/f noise is synthesized, *all obeying equation (3)*. This kind of spacing appears naturally for each slab in the channel due to the exponential dependence of $r_d(v)$ given by equation (2) and the linear drop of $V_{DS}$ along the channel. This sum for ten discrete Lorentzian spectra with cut-off frequencies of 3, 6.5, 13.9, 30, 64.6, 139.2, 300, 646.3, 1392,5 and 3000Hz, is shown in Fig. 1-b to synthesize three decades of 1/f noise that mimics the action of $V_{DS}=7V_T\approx 180$mV at room T.

**Conclusions**

Present knowledge on electrical noise in junction diodes has not been integrated in the simulator SPICE yet. This is a major drawback for the use of this excellent circuit simulator as a CAD tool in electrical noise research. This situation perhaps reveals our incomplete knowledge about noise in active devices (junction and heterojunction diodes, Bipolar and FET transistors, etc.) and in those devices thought to be passive. Among these, actual resistors use to be a kind of Field Effect Transistors with one or more gates left floating (FGFETs) that behave in a non-linear way for the resistance noise they have. This noise is seen as an excess noise at low frequencies, usually with a 1/f spectral distribution over several decades, although its original spectrum under thermal equilibrium is a Lorentzian one coming from an R-C relaxating cell.

Considering these FGFET devices that exist in most resistors and the backgating effect of the kT/C noise of boundary Space Charge Regions surrounding or embedded in such devices, a Low-frequency excess noise is predicted for such resistors whose 1/f spectrum reflects the disturbance of the dc bias applied to the resistor while its resistance noise is measured.

From the considerations made in the above paragraph, that have been overseen in the past, the answer to the main question in the abstract is: *No, we are not teaching our engineers the proper way to measure resistance noise in resistors because we have to learn first how to do it*. Teachers in Electronics can find interesting our theory that allows to explain electrically a lot of 1/f electrical noise, but researchers defending other theories for the origin of 1/f noise (e.g. carrier traps) can find our electrical explanation quite disillusioning, but they would have to learn that *electrical measurements always are done on devices* whose behaviour can affect quite a lot the *measured data* that such theories try to explain.

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**Appendix**

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.SUBCKT BAT85 1 2
* The Resistor R1 does not reflect a physical device.  Instead
it improves modeling in the reverse mode of operation.
     *
     R1 1 2 3.6E+07
     D1 1 2 BAT85-1
     .ENDS
     *
     .MODEL BAT85-1 D(
     +    IS = 2.117E-07
     +    N = 1.016
     +    BV = 36
     +    IBV = 1.196E-06
     +    RS = 2.637
     +    CJO = 1.114E-11
     +    VJ = 0.2013
     +    M = 0.3868
     +    FC = 0
     +    TT = 0
     +    EG = 0.69
     +    XTI = 2)
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