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Determination of PHEMT's Microwave Noise Parameters Only by Means of the Small-Signal Equivalent Circuit and Experimental Comparisons

D. Gasquet ⁽¹⁾, F. Barberousse ⁽¹⁾, M. de Murcia ⁽¹⁾, W. de Raedt ⁽²⁾
and C. Claeys ⁽²⁾

⁽¹⁾ Centre d'Electronique de Montpellier, UM2, 34095 Montpellier cedex 05, France

⁽²⁾ IMEC, Kapeldreef 75, B-3001 Leuven, Belgium

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Résumé. — Le but de cet article est de donner une méthode analytique de modélisation des paramètres de bruit de transistors à effet de champ et plus particulièrement de HEMT pseudomorphiques. L'originalité de cette méthode tient dans le fait qu'elle ne comprend pas de paramètres ajustables et qu'elle ne nécessite que la connaissance des éléments du schéma équivalent petit signal. Elle prend en compte toutes les sources de bruit thermique associées aux résistances, le bruit du canal étant décrit par un générateur de courant de bruit de densité spectrale $4kT_0(g_m + g_{ds})$. Les résultats théoriques sont comparés à nos mesures mais aussi à une autre technique de mesure faisant intervenir une part de modélisation (Dambrine [7]). Nous montrons que l'accord est excellent pour les quatre paramètres de bruit aussi bien en fonction de la fréquence qu'en fonction de la polarisation.

Abstract. — The purpose of this paper is to develop an analytical method in order to model the noise parameters of pseudomorphic HEMT's. The principal interest of this method is that it does not need adjustable parameters and it only needs the knowledge of the components of the small signal equivalent circuit. It takes into account all the thermal noise sources associated to the resistors and the noise of the channel is described by a noise current source with a spectral density $4kT_0(g_m + g_{ds})$. The theoretical results are compared to our measurements but also to other measurement technique involving a part of modelization (Dambrine [7]). We show that the agreement is excellent for the four noise parameters as well *versus* the frequency as *versus* the bias.

1. Introduction

Many papers have been written on noise modeling GaAs FET's [1–7]. Unfortunately, in order to predict the noise behavior of a device, all these models either need the knowledge of noise measurements results [7] and/or the use of two or three adjustable parameters [1–6] in addition to the knowledge of the small signal equivalent circuit. We do not speak here about the

physical microscopic models taking into account interactions and transport phenomena but about electrical models only. Our model is only based on analytical electrical calculations on condition that we know exactly the component values of the equivalent circuit (Fig. 1). So we will briefly explain here the method used to determine this equivalent circuit. Then the theoretical noise calculations will be compared to noise measurements performed on pseudomorphic HEMT's processed by IMEC. Noise measurement has been performed in the range 8 GHz - 18 GHz.

2. HEMT structure

The layer sequence for PHEMT devices has been grown by using molecular beam epitaxy (MBE). A $1\text{ }\mu\text{m}$ thick undoped GaAs buffer is grown on a semi-insulating GaAs substrate followed by the 13 nm-thick undoped InGaAs (In% = 20%) active layer. The 30 nm-thick GaAlAs (Al% = 22%) supply layer is separated from the active layer by a 5 nm-thick undoped GaAlAs ($5 \times 10^{14}\text{ cm}^{-3}$) spacer in order to provide a high quality interface. To reduce parasitic resistances the 40 nm cap layer is highly doped. The carrier concentration in the channel is $1.6 \times 10^{12}\text{ cm}^{-2}$ and the mobility at 300 K is $5800\text{ cm}^2/\text{Vs}$.

The devices are processed by using standard technology. The gate pattern is defined by e-beam lithography resulting in a gate length of $0.2\text{ }\mu\text{m}$. The surface is passivated by using SiN. Multi-gate-finger structure 4 fingers with a total length of $150\text{ }\mu\text{m}$ has been used and the source pads are interconnected by air bridges.

The outer pad configuration has been designed for on-wafer testing with G-S-G probe tips. This structure exhibits high gain values and very low noise figure typically $g_m = 800\text{ mS/mm}$ and $\text{NF} = 0.48\text{ dB}$ at 8 GHz.

3. Extraction of the equivalent circuit

As we said in the introduction, in order to determine the noise parameters, it is necessary to obtain with a good accuracy the component values of the equivalent circuit. This equivalent circuit (Fig. 1) can be divided into two parts, the intrinsic one and the extrinsic one. The extrinsic components including the access inductances and resistances and the pad capacitors are assumed to be bias independent. In the intrinsic part the transconductance G_m is taken frequency dependent varying as $g_m e^{-j\omega\tau}$, τ is the transit time under the gate. These values are determined using the method proposed by Cappy [8] and Dambrine [9]. This method includes

Fig. 1. — Equivalent circuit of the PHEMT. The dot rectangle define the intrinsic device.

Fig. 2

Fig. 3

Fig. 4

Fig. 5

Fig. 2. — Drain source capacitance *versus* V_{gs} . (●) $V_{ds} = 1$ V, (▽) $V_{ds} = 2$ V.

Fig. 3. — Gate drain capacitance *versus* V_{gs} . (●) $V_{ds} = 1$ V, (▽) $V_{ds} = 2$ V.

Fig. 4. — Gate source capacitance *versus* V_{gs} . (●) $V_{ds} = 1$ V, (▽) $V_{ds} = 2$ V.

Fig. 5. — Transconductance *versus* V_{gs} . (●) $V_{ds} = 1$ V, (▽) $V_{ds} = 2$ V.

two steps. In the first one we measure the scattering parameters of the device under special bias conditions [10] ($V_{ds} = 0$ V, $V_{gs} > 0$, and $V_{ds} = 0$ under the pinchoff) in order to determine the parasitic elements. In the second step we measure the scattering parameters of the device under normal bias conditions and then by matrix transformation and subtractions of the parasitic

Fig. 6

Fig. 7

Fig. 8

Fig. 6. — Drain source resistance *versus* V_{gs} . (●) $V_{ds} = 1$ V, (▽) $V_{ds} = 2$ V.

Fig. 7. — Input resistance *versus* V_{gs} . (●) $V_{ds} = 1$ V, (▽) $V_{ds} = 2$ V.

Fig. 8. — Transit time under the gate *versus* V_{gs} . (●) $V_{ds} = 1$ V, (▽) $V_{ds} = 2$ V.

elements we obtain the admittance matrix of the intrinsic device. The values of the components are obtained by identification [9,11]. In Table I we give the extrinsic element values. We can notice the very low values found for the pad capacitances ($C_{pg} \approx 0$ and $C_{pd} \approx 30$ fF). The measurements were performed in range 6-18 GHz directly on wafer. This fact explains the low inductance values found. Figures 2 to 8 give the behavior of C_{ds} , C_{gd} , C_{gs} , G_m , R_{ds} , R_i , and τ *versus* the bias conditions. From these results it is possible to evaluate the extrinsic transition

noise sources which are uncorrelated, we only have neglected the pad capacitors. In order to determine the noise parameters of the device we have to transform this equivalent circuit in the one given on Figure 9b therefrom, if we know i and e , it will be easy to calculate the noise parameters as:

$$F_{\min} = 1 + \frac{1}{2kT\Delta f} \left\{ \operatorname{Re}(\langle ei^* \rangle) + \sqrt{\langle ee^* \rangle \langle ii^* \rangle - [\operatorname{Im}(\langle ei^* \rangle)]^2} \right\} \quad (1)$$

$$R_n = \frac{\langle ee^* \rangle}{4kT\Delta f} \quad (2)$$

$$G_{\text{opt}} = \sqrt{\frac{\langle ii^* \rangle}{\langle ee^* \rangle} - \left[\frac{\operatorname{Im}(\langle ei^* \rangle)}{\langle ee^* \rangle} \right]^2} \quad (3)$$

$$B_{\text{opt}} = \frac{\operatorname{Im}(\langle ei^* \rangle)}{\langle ee^* \rangle} \quad (4)$$

The Z -matrix of the circuit (Fig. 9b) can be written as follow, if Z_{ij} are the Z coefficients of the noiseless two ports:

$$V_1 = Z_{11}I_1 + Z_{12}I_2 - Z_{11}i - e \quad (5)$$

$$V_2 = Z_{21}I_1 + Z_{22}I_2 - Z_{21}i \quad (6)$$

In the first step, considering the noise generators in the device circuit (Fig. 9a) as standard generators, we calculate the Z -matrix. This calculation is not very complicated but very tedious because we do not neglect any element except the pad capacitors. Then it is possible, by identification with (5) and (6), to obtain the expression of the generators e and i . These expressions are quite complicated and are given here:

$$i = \frac{i_m(a_1 + ja_2) + e_{ds}(b_1 + jb_2) + e_s(c_1 + jc_2) + e_d(d_1 + jd_2) + e_i(e_1 + je_2) + e_{gd}(f_1 + jf_2)}{\alpha + j\beta} \cdot C_{gs}\omega \quad (7)$$

$$e = \frac{e_g(g_1 + jg_2) + e_s(h_1 + jh_2) + e_i(i_1 + ji_2) + e_{ds}(j_1 + jj_2) + e_{gd}(k_1 + jk_2) + e_d(l_1 + jl_2) + i_m(m_1 + jm_2)}{\alpha + j\beta} \quad (8)$$

The values of the variables α , β and a_i to m_i are given in Appendix A. These variables only depend on the circuit elements and frequency.

Now in the second step we calculate $\langle ee^* \rangle$, $\langle ii^* \rangle$ and $\langle ei^* \rangle$, where $*$ denotes the complex conjugate and $\langle \rangle$ denotes the mean value. To perform these calculations we consider that the noise sources e_x are uncorrelated thus the terms $\langle e_x e_y^* \rangle$ and $\langle e_x i_y^* \rangle$ vanish. The terms $\langle e_x e_x^* \rangle$ are equal to the thermal noise of the resistance R_x ($4kR_x T_0 \Delta f$) and as we said earlier the term $\langle i_m i_m^* \rangle$ is equal to $4kg_m T_0 \Delta f$. Under these assumptions we can express the noise parameters from equations (1), (2), (3) and (4) as follows:

$$F_{\min} = 1 + \frac{2Q}{\alpha^2 + \beta^2} \cdot C_{gs}\omega + \frac{2C_{gs}\omega}{\alpha^2 + \beta^2} \sqrt{ST - R^2} \quad (9)$$

$$R_n = \frac{S}{\alpha^2 + \beta^2} \quad (10)$$

$$|Y_{\text{opt}}|^2 = \frac{T}{S}(C_{\text{gs}}\omega)^2 \quad (11)$$

$$B_{\text{opt}} = -\frac{R}{S}C_{\text{gs}}\omega \quad (12)$$

The values of the variables Q , R , S and T are expressed in Appendix A. They only depend on the values of the elements of the small-signal equivalent circuit, so that under these assumptions it is possible to evaluate the noise performances of a given device only by means of the measurement of its equivalent circuit. The value of the optimal conductance G_{opt} may be evaluated from the equations (11) and (12).

In order to calculate the noise parameters of the equivalent circuit it would be possible to use the correlation matrix approach [21]. This method avoids tedious calculations and leads directly to the results. Nevertheless it does not permit to obtain an analytical formulation of the results. Then, it is not easily possible to try to link noise parameters with technological parameters.

As we said in the introduction the literature proposes a lot of electrical noise modeling techniques. The most commonly used one is the Pospieszalski's [1] approach. In this model, which uses only two noise temperatures T_g and T_d , the noisy device is assumed to be represented by a noiseless device associated with an input noise voltage generator $\langle e_g^2 \rangle$ and an output noise current generator $\langle i_d^2 \rangle$. The expressions of these uncorrelated generators are:

$$\langle e_g^2 \rangle = 4kT_g R_i \Delta f$$

$$\langle i_d^2 \rangle = 4kT_d R_{ds} \Delta f$$

From our model, using the transformation allowing us to obtain the Pospieszalski's representation, it is possible to express the temperatures T_g and T_d :

$$T_g = \frac{C_{\text{gs}}^2}{(C_{\text{gs}} + C_{\text{gd}})^2} T_0$$

and,

$$T_d \approx \left(1 + \frac{g_m}{g_{ds}}\right) T_0$$

5. Noise measurement

The noise measurements have been carried out on wafer by means of a probe station "Alessi" coupled with a noise figure meter HP 8970 B, a noise figure test set HP 8971 B and a synthesized CW generator HP 8673 G. At the input of the noise apparatus we have placed a low-noise amplifier and an isolator in order to diminish its noise figure and to keep its input impedance constant. The LNA has a noise factor lower than 4 dB and a gain higher than 25 dB in the range 1 -18 GHz. In order to determine the noise parameters we use the multiple impedances method [16], [17]. These impedances are realized by means of an automatic mechanical tuner. This tuner is a home made one, concerning the magnitude of the reflection coefficient its repeatability is better than 2% and the uncertainty on its phase is in the worst case (at 18 GHz) 2 degrees. In order to calibrate the tuner and deembed [18] the noise of the DUT we use a network analyzer HP 8720. This calibration step is very important and it must be performed with a great carefulness to avoid erroneous measurements.

Fig. 10

Fig. 11

Fig. 12

Fig. 13

Fig. 10. — Noise factor under $50\ \Omega$. Condition *versus* f^2 . $V_{ds} = 1\text{ V}$, $I_d = 26\text{ mA}$. (■) experimental results, (▽) results of the calculation from equation (13) using $Y_s = 20\text{ mS}$.

Fig. 11. — Minimum noise figure *versus* frequency. $V_{ds} = 1\text{ V}$, $I_d = 9\text{ mA}$. (●) Experimental results, (▽) Results of the calculation from equation (9), (▼) Results obtained from Dambrine's method.

Fig. 12. — Phase of the optimum reflection coefficient *versus* frequency. $V_{ds} = 1\text{ V}$, $I_d = 9\text{ mA}$. (●) Experimental results, (▽) Results of the calculation from equations (11) and (12), (▼) Results obtained from Dambrine's method.

Fig. 13. — Magnitude of the optimum reflection coefficient *versus* frequency. $V_{ds} = 1\text{ V}$, $I_d = 9\text{ mA}$. (●) Experimental results, (▽) Results of the calculation from equation (11) and (12), (▼) Results obtained from Dambrine's method.

Fig. 14

Fig. 15

Fig. 14. — Equivalent noise resistance *versus* frequency. $V_{ds} = 1$ V, $I_d = 9$ mA. (●) Experimental results, (▽) Results of the calculation from equation (10), (▼) Results obtained from Dambrine's method.

Fig. 15. — Minimum noise factor *versus* drain current. $f = 8$ GHz (●) Experimental results, (▽) Results of the calculation from equation (9), (▼) Results obtained from Dambrine's method.

In Figures 10 to 15 we give the noise results. We have plotted the experimental ones, the calculated ones using equations (9) to (12) and the results obtained using the Dambrine's method [7].

In this method the authors use two assumptions:

- The first one is that the noise resistance R_n is (in the case of on wafer measurement) frequency independent. This fact is exact if we consider the intrinsic device, but we will show that it is not true when we take into account the gate, the source and the drain inductances even if they are very low.

- The second one concerns the correlation coefficient between the gate and the drain which is considered to be purely imaginary.

From this two considerations, it is then possible to obtain the noise parameters from a simple 50Ω noise measurement knowing the small-signal equivalent signal.

In Figure 10, we give the noise factor measurement (black squares) performed under 50Ω conditions *versus* the square of the frequency. We can see that the variation is linear and by means of the Dambrine's method we determine noise parameters which will be used further on. The white triangles are the results calculated from our noise parameters by means of the classical formula:

$$F = F_{\min} + \frac{R_n}{G_s} \cdot [(G_s - G_{\text{opt}})^2 + (B_s - B_{\text{opt}})^2] \quad \text{with } B_s = 0 \text{ and } G_s = 20 \text{ mS}$$

The agreement is quite correct.

In Figure 11 we have plotted the minimum noise factor *versus* the frequency. The agreement is better than 0.05 dB over the whole frequency range.

Figure 12 represents the phase of the optimal reflection coefficient *versus* the frequency, Dambrine's values are lightly lower than our results.

In Figure 13 where we have plotted the magnitude of the optimal reflection coefficient the disagreement is much more important and Dambrine's values are between 5 and 10% lower. Concerning the noise resistance (Fig. 14) Dambrine's value is constant of course and it seems that our results give a light dependence on the frequency. The calculated R_n decreases when frequency increases but the experimental parameter which is the most difficult to obtain with a good accuracy seems to increase. Nevertheless the difference between the three curves is lower than 2%.

Then in Figure 15 we have plotted the minimum noise factor *versus* the drain current and the agreement is always very good. The minimum of the noise factor is obtained for a drain current of 9 mA. For a higher current the noise factor increases relatively slowly. However when the drain current tends to zero the noise factor increases dramatically.

6. Conclusions

This work shows that it is possible with a good accuracy to predict, only by means of, the equivalent circuit determination, the noise behaviour of a PHEMT, assuming that the noise is only due to thermal noise. This fact suggests that the contribution of the hot carriers to the noise of the device may be neglected as predicted by Van der Ziel [14]. In this study we show that the Dambrine's method gives correct results excepted for the determination of the optimal reflection coefficient where the disagreement is about 6%. This fact may be due to the initial hypothesis which neglects the extrinsic elements. Nevertheless this method needs an additional noise measurement to provide noise parameters. The experimental results given in this paper are in agreement with those found in literature [19, 20].

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Appendix A

$$\alpha = g_m R_{ds} \sin(\omega\tau) + \omega \{ R_s (C_{gs} + C_{gd}) + R_{ds} C_{gd} + g_m R_{ds} \cos(\omega\tau) [C_{gd} (R_s - R_{gd}) - R_{ds} C_{ds}] \} + \omega^2 g_m R_{ds} L_s C_{gd} \sin(\omega\tau)$$

$$\beta = g_m R_{ds} \cos(\omega\tau) - \omega g_m R_{ds} \sin(\omega\tau) [C_{gd} (R_s - R_{gd}) - R_{ds} C_{ds}] + \omega^2 L_s \{ C_{gs} + C_{gd} + g_m R_{ds} C_{gd} \cos(\omega\tau) \}$$

$$a_1 = R_{ds} \left(1 + \frac{C_{gd}}{C_{gs}} \right)$$

$$a_2 = \omega R_{ds} \left\{ C_{gd} (R_i + R_{gd}) + R_{ds} C_{ds} \left(1 + \frac{C_{gd}}{C_{gs}} \right) \right\}$$

$$b_1 = - \left(1 + \frac{C_{gd}}{C_{gs}} \right)$$

$$b_2 = -\omega \left\{ C_{gd} (R_i + R_{gd}) + R_{ds} C_{ds} \left(1 + \frac{C_{gd}}{C_{gs}} \right) \right\}$$

$$\begin{aligned}
c_1 &= \left\{ 1 + \frac{C_{gd}}{C_{gs}} [1 + g_m R_{ds} \cos(\omega\tau)] \right\} - \omega g_m R_{ds}^2 C_{ds} \frac{C_{gd}}{C_{gs}} \sin(\omega\tau) \\
c_2 &= g_m R_{ds} \frac{C_{gd}}{C_{gs}} \sin(\omega\tau) - \omega \left\{ C_{gd} (R_i + R_{gd} + R_{ds}) + R_{ds} C_{ds} \left[2 + 2 \frac{C_{gd}}{C_{gs}} + g_m R_{ds} \frac{C_{gd}}{C_{gs}} \cos(\omega\tau) \right] \right\} \\
d_1 &= c_1 \\
d_2 &= c_2 \\
e_1 &= -g_m R_{ds} \frac{C_{gd}}{C_{gs}} \cos(\omega\tau) - \omega g_m R_{ds}^2 C_{ds} \frac{C_{gd}}{C_{gs}} \sin(\omega\tau) \\
e_2 &= g_m R_{ds} \frac{C_{gd}}{C_{gs}} \sin(\omega\tau) - \omega R_{ds} \left\{ C_{gd} + g_m R_{ds} C_{ds} \frac{C_{gd}}{C_{gs}} \cos(\omega\tau) \right\} \\
f_1 &= -e_1 \\
f_2 &= -e_2 \\
g_1 &= -g_m R_{ds} \sin(\omega\tau) - \omega \{ C_{gd} (R_{ds} + R_s) + C_{gs} R_s + g_m R_{ds} \cos(\omega\tau) [C_{gd} (R_s - R_{gd}) - C_{ds} R_{ds}] \} \\
&\quad - \omega^2 g_m R_{ds} C_{gd} L_s \sin(\omega\tau) \\
g_2 &= -g_m R_{ds} \cos(\omega\tau) + \omega \{ g_m R_{ds} \sin(\omega\tau) \cdot (R_s C_{gd} - R_{ds} C_{ds} - R_{gd} C_{gd}) \} \\
&\quad - \omega^2 L_s \{ C_{gs} + C_{gd} (1 + g_m R_{ds} \cos(\omega\tau)) \} \\
h_1 &= -g_m R_{ds} \sin(\omega\tau) + \omega \{ C_{gs} (R_g + R_i) + C_{gd} (R_g + R_{gd}) + 2 C_{ds} R_{ds} + g_m R_{ds} \cos(\omega\tau) \\
&\quad \times [C_{ds} R_{ds} + C_{gd} (R_{ds} + R_{gd})] \} + \omega^2 g_m R_{ds} L_g C_{gd} \sin(\omega\tau) \\
h_2 &= -\{ 1 + g_m R_{ds} \cos(\omega\tau) \} - \omega g_m R_{ds} \sin(\omega\tau) (R_g C_{gd} + R_{ds} C_{ds} + R_{gd} C_{gd}) \\
&\quad - \omega^2 \{ L_g (C_{gs} + C_{gd}) + g_m R_{ds} C_{gd} \cos(\omega\tau) \} \\
i_1 &= -g_m R_{ds} \sin(\omega\tau) - \omega \{ R_s C_{gs} + g_m R_{ds} \cos(\omega\tau) [R_s C_{gd} - R_{ds} C_{ds} - R_{gd} C_{gd} - (R_g + R_s) C_{gs}] \} \\
&\quad - \omega^2 \{ g_m R_{ds} \sin(\omega\tau) [L_s C_{gd} - (L_g + L_s) C_{gs}] \} \\
i_2 &= -g_m R_{ds} \cos(\omega\tau) + \omega \{ g_m R_{ds} \sin(\omega\tau) [C_{gd} (R_s - R_{gd}) - R_{ds} C_{ds} - (R_g + R_s) C_{gs}] \} \\
&\quad - \omega^2 \{ L_s (C_{gs} + C_{gd}) + g_m R_{ds} \cos(\omega\tau) [L_s C_{gd} - (L_s + L_g) C_{gs}] \} \\
j_1 &= \omega \{ (R_g + R_s + R_i) C_{gs} + (R_{gd} + R_g) C_{gd} + R_{ds} C_{ds} + g_m R_{ds} \cos(\omega\tau) (R_g + R_s) (C_{gd} - C_{gs}) \} \\
&\quad + \omega^2 g_m R_{ds} \sin(\omega\tau) (L_g + L_s) (C_{gd} - C_{gs}) \\
j_2 &= -1 - \omega g_m R_{ds} \sin(\omega\tau) (R_g + R_s) (C_{gd} - C_{gs}) \\
&\quad + \omega^2 \{ C_{gs} (L_g + L_s) + L_g C_{gd} + (L_g + L_s) g_m R_{ds} \cos(\omega\tau) (C_{gd} - C_{gs}) \} \\
k_1 &= -\omega \{ C_{gd} (R_{ds} + R_s) - g_m R_{ds} (R_g + R_s) C_{gs} \cos(\omega\tau) \} + \omega^2 (L_g + L_s) g_m R_{ds} C_{gs} \sin(\omega\tau) \\
k_2 &= -\omega g_m R_{ds} (R_g + R_s) C_{gs} \sin(\omega\tau) - \omega^2 \{ L_s C_{gd} - (L_g + L_s) g_m R_{ds} C_{gs} \cos(\omega\tau) \} \\
l_1 &= \omega \{ C_{gs} (R_g + R_s + R_i) + C_{gd} (R_g + R_s + R_{ds} + R_{gd}) + 2 C_{ds} R_{ds} \\
&\quad + (R_g + R_s) g_m R_{gd} C_{gd} \cos(\omega\tau) \} + \omega^2 (L_g + L_s) g_m R_{ds} C_{gd} \sin(\omega\tau)
\end{aligned}$$

$$l_2 = -1 - \omega(R_g + R_s)g_m R_{ds} C_{gd} \sin(\omega\tau) + \omega^2 \{(L_g + L_s)C_{gs} + (L_g + L_s)g_m R_{ds} C_{gd} \cos(\omega\tau)\}$$

$$m_1 = -\omega R_{ds} \{C_{gs}(R_g + R_s + R_i) + C_{gd}(R_g + R_{gd}) + C_{ds}R_{ds}\}$$

$$m_2 = R_{ds} - \omega^2 R_{ds} \{C_{gs}(L_g + L_s) + C_{gd}L_g\}$$

$$Q = 0$$

$$R = g_m(a_1 m_2 - a_2 m_1) + R_{ds}(b_1 j_2 - b_2 j_1) + R_s(c_1 h_2 - c_2 h_1) \\ + R_d(d_1 l_2 - d_2 l_1) + R_i(e_1 i_2 - e_2 i_1) + R_{gd}(f_1 k_2 - f_2 k_1)$$

$$S = R_g(g_1^2 + g_2^2) + R_s(h_1^2 + h_2^2) + R_i(i_1^2 + i_2^2) + R_{ds}(j_1^2 + j_2^2) + R_d(l_1^2 + l_2^2) + g_m(m_1^2 + m_2^2)$$

$$T = g_m(a_1^2 + a_2^2) + R_{ds}(b_1^2 + b_2^2) + R_s(c_1^2 + c_2^2) + R_d(d_1^2 + d_2^2) + R_i(e_1^2 + e_2^2) + R_{gd}(f_1^2 + f_2^2)$$

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