PAPER

A New Method for the Determination of the Extrinsic Resistances of MESFETs and HEMTs from the Measured *S*-Parameters under Active Bias

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SUMMARY A new method is proposed for determining the parasitic extrinsic resistances of MESFETs and HEMTs from the measured S-parameters under active bias. The proposed method is based on the fact that the difference between drain resistance (R_d) and source resistance (R_s) can be found from the measured S-parameters under zero bias condition. It is possible to define the new internal device including intrinsic device and three extrinsic resistances by eliminating the parasitic imaginary terms. Three resistances can be calculated easily via the presented explicit three equations, which are induced from the fact that 1) the real parts of $Y_{int,11}$ and $Y_{int,12}$ of intrinsic Y-parameters are very small or almost zero, 2) the transformation relations between S-, Z-, and Y-matrices. The modelled S-parameters calculated by the obtained resistances and all the other equivalent circuit parameters are in good agreement with the measured S-parameters up to 40 GHz.

key words: *small signal model, extrinsic resistances, MESFETs, HEMTs*

1. Introduction

GaAs MESFETs and HEMTs have been widely used for amplifiers at microwave and millimeter wave frequency range due to their low noise and high power performances. Especially, HEMTs have ultra high cut off and maximum oscillation frequencies up to a few hundred GHz because of their short gate length and high electron mobility. The large signal equivalent circuit model of a device is required for the prediction of performances of nonlinear circuits. Prior to nonlinear circuit model, a small signal equivalent circuit model is also important for the prediction of linear operation, and, especially, for the development of accurate large signal model.

Many papers have been presented for finding the parasitic extrinsic elements in the equivalent circuit [1]–[9]. As for extrinsic resistances, the methods used in previous works require additional DC measurements in addition to S-parameter measurement [1]–[4], [7], and optimization or iteration in order to obtain acceptable solutions [10], [11]. However, the additional DC measurements require forward gate bias which, sometimes, can bring about serious problem such as damage to device during or after measurement [8], [9]. The optimization or iteration method usually results in the equivalent circuit parameters which may be meaningless value in practical world. Also, the optimization technique always entails the initial value-dependence problem and the local minimum problem [12]. The modelling technique using the device physics and fabrication processes may be extremely difficult and complex to circuit designer.

In order to escape from all the problems mentioned above, Sommer et al. have presented a method to obtain the parasitic resistances from the measured S-parameters under active bias [9]. However, in their work, three resistnaces (R_d, R_s, and R_g) are not fixed to unique value, but the ratios between them ($\alpha_1 = R_g/R_s$ and $\alpha_2 = R_d/R_s$) are assumed to have some initial values. After that, optimization or iteration procedure is required to choose the proper α_1 and α_2. In [9], R_s has three solutions; one real root, which is to be chosen as R_s, and two complex roots which are discarded. R_g and R_d are calculated from the assumed relations, α_1 and α_2, after R_s is chosen.

We propose a new method for finding the three parasitic resistances, as a unique solution, from the measured S-parameters under active bias in order to overcome all the problems mentioned above. In this method, there is no need for additional DC measurement and optimization or iteration. There are two basic and important facts from which the key idea of the proposed method is stemmed. The first is that there exists a finite difference between R_s and R_d and this difference can be calculated from the measured S-parameters under zero bias condition. The second is that the real parts of $Y_{int,11}$ and $Y_{int,12}$ of intrinsic Y-parameters are theoretically zero or extremely small to be approximated as zero. The simple transformations between S-, Z-, Y-matrices and two basic facts mentioned above will produce the explicit formulae for the unique solution of three parasitic resistances. To our knowledge, the explicit formulae which show three extrinsic resistances uniquely are proposed in this work firstly.

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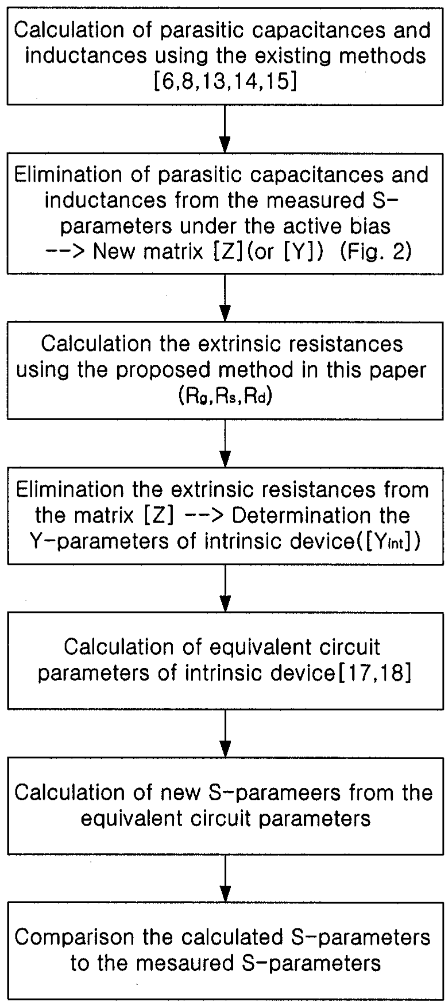


Fig. 1 Flow chart for the verification of the proposed method.

Figure 1 shows the procedure to verify the validity of the proposed method. For the beginning, the parasitic capacitances and inductances will be extracted by existing methods [6], [8], [13]–[16]. Then, the dotted-line box shown in Fig. 2 including intrinsic device and extrinsic resistances can be defined as a new $[Z]$ (or $[Y]$) matrix. In the next step, the extrinsic resistances will be calculated by the proposed method. Now Y_{int}, the Y-parameters of intrinsic device, can be obtained by subtracting the extrinsic resistances. Finally, intrinsic equivalent circuit parameters are calculated by the Berroth et al. [17], [18].

The modelled S-parameters will be calculated using the obtained equivalent circuit parameters, and compared with the measured S-parameters. We apply the proposed method to the S-parameters with various bias conditions, and verify its validity up to 40 GHz.

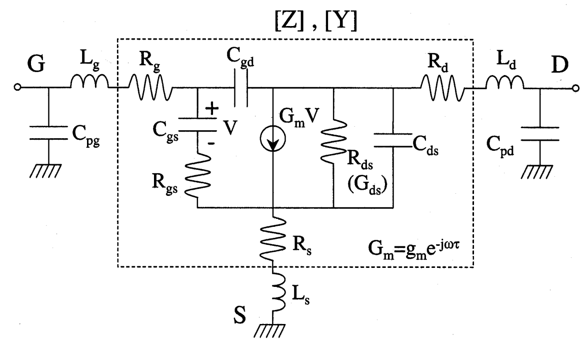


Fig. 2 Widely used small signal equivalent circuit model of GaAs MESFETs and HEMTs.

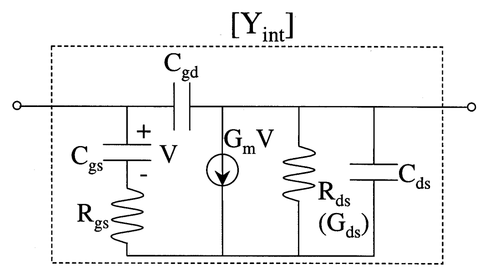


Fig. 3 Equivalent circuit of intrinsic part of GaAs MESFETs and HEMTs with 7 ECPs.

2. Small Signal Equivalent Circuit Model and Parasitic Extrinsic Parameters

Figure 2 shows the equivalent circuit of MESFETs and HEMTs with 7 intrinsic Equivalent Circuit Parameters (ECPs) which has been used widely. Dambrine et al. presented the method for determining the ECPs using the transformations between S-, Z-, and Y-parameters [13]. Berroth et al. proposed another method for finding the 7 or 10 ECPs using the device symmetry at low drain voltage [17], [18]. Figure 3 shows the intrinsic equivalent circuits with 7 ECPs in [17]. They extended the valid frequency limit of equivalent circuit modelling up to 60 GHz, and refined the accuracy between the modelled S-parameters and the measured ones.

It is important to extract the correct parasitic elements since the intrinsic part is deduced from the whole equivalent circuit model by subtracting the extrinsic parts. It is generally accepted that there are eight extrinsic parasitic elements (C_{pg},C_{pd},L_g,L_s,L_d,R_g,R_s,R_d) as shown in Fig. 2, and these elements are known not to depend on bias conditions. Strictly speaking, it is just an assumption that extrinsic parameters do not depend on bias conditions. However, their dependence on bias conditions is not significant, so this assumption has been widely and generally accepted for the convenience of modelling.

Since parasitic capacitances and inductances can be extracted from the measured S-parameters under zero bias or pinched-off bias [3], or from the curve-fitting method, no damage can occur [6], [13], [15], [16]. However, parasitic resistances can not be obtained using the same measurement because these resistances are caused by the conductor resistivity of terminals such as gate, drain, and source. Hence, the method using DC current for measurement of the parasitic resistances has been used conventionally [1]–[4], [7], [13]. However, these DC current methods sometimes incur a serious problem that the device under test may have a damage because of forward gate current [8], [9].

3. New Method for the Determination of the Extrinsic Resistancesfrom the Measured S-Parameters under Active Bias

A new method for the determination of the extrinsic resistances will be described in detail in this section. Prior to entering the main process, the parasitic imaginary elements should be determined using the existing method in [6], [8], [13]–[16]. At $V_{ds} = 0\,\text{V}$, i.e. in the cold modelling, a FET and HEMT can be considered as a symmetric device. The measured S-parameters on this bias condition can be transformed into Z-parameters. By the way, at $V_{ds} = 0\text{V}$, the intrinsic section of FET can be expressed by the sum of channel resistance and equivalent impedance of Schottky barrier. If the extrinsic resistances (R_g,R_s,R_d) and inductances (L_g,L_s,L_d) are added to the intrinsic section, the resultant Z-parameters can be expressed as follows [13], [17], [18];

$$Z_{11} = R_s + R_g + \frac{R_c}{3} + \frac{nkT}{qI_g} + j\omega(L_s + L_g) \tag{1}$$

$$Z_{12} = Z_{21} = R_s + \frac{R_c}{2} + j\omega L_s \tag{2}$$

$$Z_{22} = R_s + R_d + R_c + j\omega(L_s + L_d) \tag{3}$$

where R_c is the channel resistance under given gate bias, nkT/qI_g is the equivalent resistance of Schottky barrier, n is the ideality factor, k is the Boltzmann's constant, T is the temperature, and I_g is the DC gate current.

As can be seen in (1)–(3), the imaginary parts increase with frequency, while the real parts of Z-parameters are not dependent on the frequency theoretically. There are four unknowns (R_c, R_g, R_s, and R_d) in three real parts. However, (4), the relation between R_d and R_s, can be obtained from (1)–(3) regardless of R_c. Therefore a unique solution for three resistances can exist.

$$\text{Re}(Z_{22}) - 2\text{Re}(Z_{12}) = R_d - R_s = \Delta R_{ds} \tag{4}$$

where ΔR_{ds} means the difference between R_d and R_s. Two cases are possible according to the sign of ΔR_{ds}. If $\Delta R_{ds} > 0$, then $R_d > R_s$, i.e. $R_d = R_s + \Delta R_{ds}$, and vice versa.

In summary, the difference between R_d and R_s, $|\text{Re}(Z_{22}) - 2\text{Re}(Z_{12})| = |R_d - R_s| = |\Delta R_{ds}|$ can be calculated from the Z-parameters transformed from the S-parameters measured under cold bias. It should be noted that either R_d or R_s gives the other one provided ΔR_{ds} is known. This means that the number of unknowns has been reduced from four to three because R_d and R_s can be considered as one due to (4). Therefore the unique solution for three extrinsic resistances can be obtained through the explicit formulae, (25)–(27), which will be explained later in detail.

The Y-parameters for the intrinsic part with 7 ECPs, shown in Fig. 3, can be expressed like (5)–(8) after being simplified since $\omega^2 C_{gs}^2 R_{gs}^2 \ll 1$ and $\omega t \ll 1$ [9], [13], [18]. [18] has dealt with the Y-parameters for the intrinsic part of equivalent circuit with 10 ECPs in more detail.

$$Y_{int,11} \approx \omega^2 C_{gs}^2 R_{gs} + j\omega(C_{gd} + C_{gs}) \tag{5}$$

$$Y_{int,12} = -j\omega C_{gd} \tag{6}$$

$$Y_{int,21} \approx g_m - j\omega\{g_m(\tau + R_{gs}C_{gs}) + C_{gd}\} \tag{7}$$

$$Y_{int,22} \approx G_{ds} + j\omega(C_{gd} + C_{ds}) \tag{8}$$

Those parameters can be represented in a simple matrix form equation.

$$[Y_{int}] = [G_{int}] + j[B_{int}] \tag{9}$$

where

$$[G_{int}] = \begin{bmatrix} \omega^2 C_{gs}^2 R_{gs} & 0 \\ g_m & G_{ds} \end{bmatrix} \tag{10}$$

$$[B_{int}] = \omega[C_{int}] \tag{11}$$

$$[C_{int}] = \begin{bmatrix} C_{gd} + C_{gs} & -C_{gd} \\ -\{g_m(\tau + R_{gs}C_{gs}) + C_{gd}\} & C_{gd} + C_{ds} \end{bmatrix} \tag{12}$$

In Fig. 2, the Y- (or Z-)parameters for the dotted box including intrinsic device and external resistances are defined as $[Y]$ (or $[Z]$). $[Y]$ can be obtained by eliminating the parasitic capacitances and inductances from $[Y_{mea}]$, the transformed Y-parameters from the measured S-parameters. Equation (13) shows the relation between $[Y]$ and $[Y_{mea}]$.

$$[Y] = \left[\{[Y_{mea}] - j\omega[C_{ext}]\}^{-1} - j\omega[L_{ext}]\right]^{-1} \tag{13}$$

where

$$[C_{ext}] = \begin{bmatrix} C_{pg} & 0 \\ 0 & C_{pd} \end{bmatrix} \tag{14}$$

$$[L_{ext}] = \begin{bmatrix} L_g + L_s & L_s \\ L_s & L_d + L_s \end{bmatrix} \tag{15}$$

Since $[Y]$ is the combination of $[Y_{int}]$, and the external resistances, the following relationship is derived.

$$[Y]=[Z]^{-1}=\{[R_{ext}]+[Y_{int}]^{-1}\}^{-1} \quad (16)$$

where

$$[R_{ext}]=\begin{bmatrix} R_g+R_s & R_s \\ R_s & R_d+R_s \end{bmatrix} \quad (17)$$

Therefore,

$$[Y_{int}]\{[Z]-[R_{ext}]\}=[I_2] \quad (18)$$

Equation (19) is obtained by inserting (9) and (17) into (18).

$$\begin{bmatrix} G_{int,11}+jB_{int,11} & G_{int,12}+jB_{int,12} \\ G_{int,21}+jB_{int,21} & G_{int,22}+jB_{int,22} \end{bmatrix}\begin{bmatrix} T_{11} & T_{12} \\ T_{21} & T_{22} \end{bmatrix}$$
$$=[I_2]=\begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}$$
$$=\begin{bmatrix} 1+j0 & 0+j0 \\ 0+j0 & 1+j0 \end{bmatrix}=\begin{bmatrix} I_{11} & I_{12} \\ I_{21} & I_{22} \end{bmatrix} \quad (19)$$

where

$T_{11}=\{\mathrm{Re}(Z_{11})-(R_g+R_s)\}+j\mathrm{Im}(Z_{11})$,
$T_{12}=\{\mathrm{Re}(Z_{12})-R_s\}+j\mathrm{Im}(Z_{12})$,
$T_{21}=\{\mathrm{Re}(Z_{21})-R_s\}+j\mathrm{Im}(Z_{21})$, and
$T_{22}=\{\mathrm{Re}(Z_{22})-(R_d+R_s)\}+j\mathrm{Im}(Z_{22})$.

Here, $\mathrm{Re}(Z_{ij})$ represents the real part of Z_{ij}, and $\mathrm{Im}(Z_{ij})$ the imaginary part, where $i,j=1,2$. $\mathrm{Re}(Z_{ij})$ and $\mathrm{Im}(Z_{ij})$ were already known to us through the elimination of the parasitic capacitances and inductances from the measured S-parameters. It can be seen that $G_{int,12}=0$ theoretically from (10). This means the feedback in device is purely capacitive.

The matrix expression of (19) contains 4 equations, which can be turned into 8 equations by dividing the real and imaginary parts. The real numbers 1 and 0, four elements of $[I_2]$, can be expressed in complex numbers whose imaginary parts are zero. Now, it is true that $\mathrm{Re}(I_{11})=\mathrm{Re}(I_{22})=1$ and $\mathrm{Re}(I_{12})=\mathrm{Re}(I_{21})=\mathrm{Im}(I_{11})=\mathrm{Im}(I_{12})=\mathrm{Im}(I_{21})=\mathrm{Im}(I_{22})=0$. Only four equations out of the eight are required for determining the parasitic resistances, as expressed in (20)–(23).

$$\mathrm{Re}(I_{11})=-B_{int,11}\mathrm{Im}(Z_{11})-B_{int,12}\mathrm{Im}(Z_{21})=1 \quad (20)$$

$$\mathrm{Im}(I_{11})=B_{int,11}\{\mathrm{Re}(Z_{11})-(R_g+R_s)\}+B_{int,12}\{\mathrm{Re}(Z_{21})-R_s)\}=0 \quad (21)$$

$$\mathrm{Re}(I_{12})=-B_{int,11}\mathrm{Im}(Z_{12})-B_{int,12}\mathrm{Im}(Z_{22})=0 \quad (22)$$

$$\mathrm{Im}(I_{12})=B_{int,11}\{\mathrm{Re}(Z_{12})-R_s\}+B_{int,12}\{\mathrm{Re}(Z_{22})-(R_d+R_s)\}=0 \quad (23)$$

It should be noted that the time constant of $R_{gs}C_{gs}$ and $G_{int,11}(=\omega^2C_{gs}^2R_{gs})$ are extremely small. The typical value of $G_{int,11}$ is order of $10^{-4}-10^{-5}$ over a few GHz to a few decade of GHz. The more important fact is that the first terms including the multiplication by $G_{int,11}$ in the expansion of I_{11} and I_{12} in (19), currently not shown in (20)–(23), can be ignored because these are much smaller than the other terms by a factor of ten or more. Appendix will verify that this approximation is quite reasonable and tolerable by using this widely used approximation, and that the effects of the first terms multiplied by $\omega^2C_{gs}^2R_{gs}$ are so weak even though it is not ignored. In practice, this approximation has been discussed already in previous works [9], [13].

Because $B_{int,11}$ and $B_{int,12}$ can be expressed as functions of $\mathrm{Im}(Z_{ij})$ in (20) and (22), they are produced by a simple matrix relation as follows :

$$\begin{bmatrix} B_{int,11} \\ B_{int,12} \end{bmatrix}=\frac{1}{D}\begin{bmatrix} -\mathrm{Im}(Z_{22}) \\ \mathrm{Im}(Z_{12}) \end{bmatrix} \quad (24)$$

where D means the determinant, i.e., $\mathrm{Im}(Z_{11})\mathrm{Im}(Z_{22})-\mathrm{Im}(Z_{12})\mathrm{Im}(Z_{21})$.

R_s and R_d are calculated from (25) or (26) by inserting $B_{int,11}$ and $B_{int,12}$ into (23). Finally, R_g is obtained from (27) by inserting $B_{int,11}$, $B_{int,12}$, R_s, and R_d into (21). Equations (25)–(27) show that R_s, R_d, and R_g are determined as a unique solution set by simple explicit formulae.

i) $\Delta R_{ds}>0 \Longrightarrow R_d>R_s \Longrightarrow R_d=R_s+\Delta R_{ds}$

$$R_s=\frac{\left[\begin{array}{l}\mathrm{Re}(Z_{12})B_{int,11} \\ \quad +\{\mathrm{Re}(Z_{22})-\Delta R_{ds}\}B_{int,12}\end{array}\right]}{B_{int,11}+2B_{int,12}} \quad (25)$$

ii) $\Delta R_{ds}<0 \Longrightarrow R_d<R_s \Longrightarrow R_d=R_s+\Delta R_{ds}$ or $R_s=R_d+|\Delta R_{ds}|$

$$R_d=\frac{\left[\begin{array}{l}\{\mathrm{Re}(Z_{12})-\Delta R_{ds}\}B_{int,11} \\ \quad +\{\mathrm{Re}(Z_{22})-\Delta R_{ds}\}B_{int,12}\end{array}\right]}{B_{int,11}+2B_{int,12}} \quad (26)$$

iii) R_g

$$R_g=\frac{\left[\begin{array}{l}\{\mathrm{Re}(Z_{11})-R_s\}B_{int,11} \\ \quad +\{\mathrm{Re}(Z_{21})-R_s\}B_{int,12}\end{array}\right]}{B_{int,11}} \quad (27)$$

4. Verification of the Proposed Method; Comparison of S-Parameters

A new method for determining the extrinsic resistances from the measured S-parameters under active bias has been described so far. Since $[Y_{int}]$ can be found by subtracting the extrinsic resistances from the Y matrix using (16), the equivalent circuit parameters (ECPs) for the intrinsic device can be determined by applying the widely used conventional method [13], [17], [18].

For the verification of the proposed method, a HEMT device with $0.2\,\mu$m of gate length and $100\,\mu$m of gate width was investigated. For the beginning step, parasitic capacitances and inductances were found using the previous method [6], [8], [13]–[16]. Then, we calculated 3 extrinsic resistances by the proposed method. Finally, we calculated the ECPs of intrinsic device. The flow chart shown in Fig. 1 has been processed for the verification.

Figure 4 shows the frequency characteristics of R_s, R_d, R_g, and ΔR_{ds}, which have been obtained by the proposed method. It can be seen that the R_s, R_d, and R_g converge to fixed values above 5 GHz. At low frequency, as widely known, the resolution of the measurement system is limited due to so small real part of the feedback admittance and almost open characteristics of S_{11} of the device, which is predominated by a small dc-current. Figure 4 is a very typical plot and the final values of R_s, R_d, and R_g determined by averaging over 5–40 GHz are $4.22\,\Omega$, $6.82\,\Omega$, and $3.13\,\Omega$, respectively.

Table 1 shows the extrinsic and intrinsic equivalent circuit parameters of this device under the bias that $V_{gs} = -0.4$ V and $V_{ds} = 2$ V. Figure 5 shows the modelled S-parameters agree well with the measured S-parameters up to 40 GHz. This agreement implies that the proposed method for determining the extrinsic resistances and successive step for small signal modelling are well applicable and suitable up to millimeter wave range. In order to verify whether the proposed method is acceptable for various bias conditions, we extracted the small signal ECPs, calculated the S-parameters, and compared them, in Fig. 6, with each S-parameters measured under various bias conditions. The calculated and measured S-parameters, under $V_{gs} = -1$ V and 0 V are shown in Fig. 6. Good agreement is observed for all bias conditions.

It should be noted that maximum available gain (G_{max} if $K \geq 1$) or most stable gain (MSG if $K < 1$) and stability factor (K) are practically important parameters for the microwave circuit designers. Figure 7 illustrates the values of G_{max}(or MSG) and K from the modelled and measured S-parameters up to 40 GHz, and good agreement between them under all bias conditions. Therefore, it can be said that the verification of the proposed method has been achieved successfully through the above agreements.

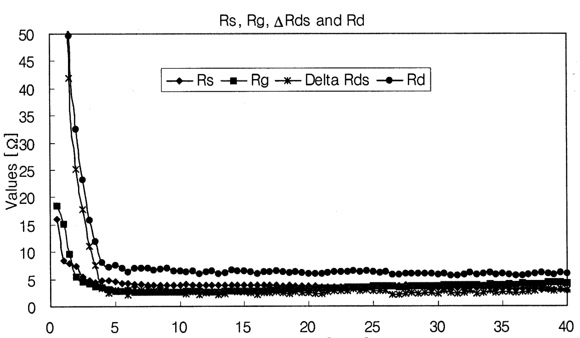


Fig. 4 Frequency Characteristics of R_s, R_d, R_g, and ΔR_{ds}.

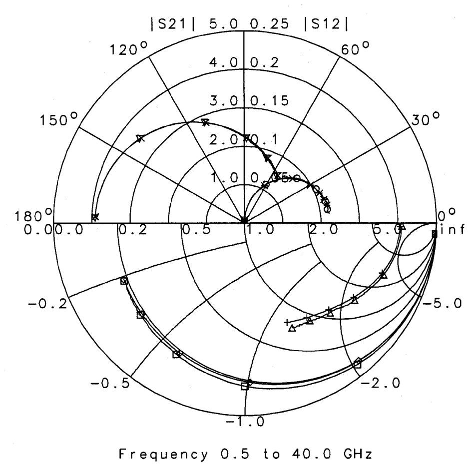
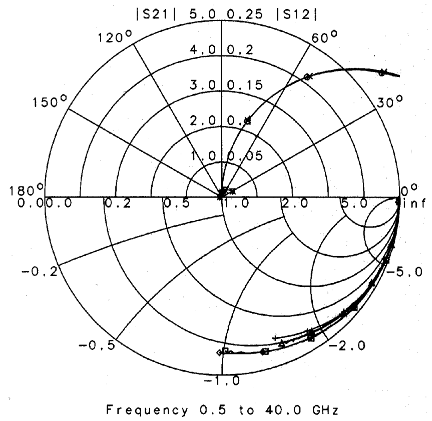


Fig. 5 Comparison between the calculated and measured S-parameters under $V_{gs} = -0.4$ V and $V_{ds} = 2$ V. Good agreement shows the validity of the proposed method for the determination of the extrinsic resistances. (□: S11(mea), ○: S21(mea), ▽: S12(mea), △: S22(mea), ◇: S11(model), ∗: S21(model), ×: S12(model), +: S22(model)).

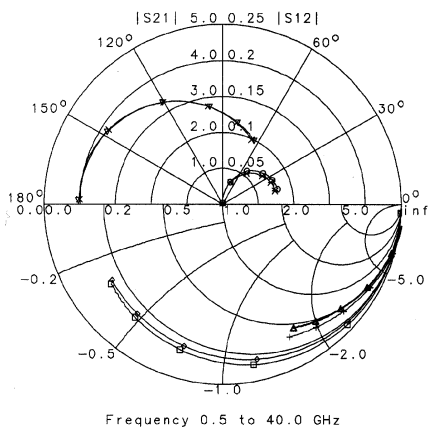
Table 1 The extrinsic and intrinsic equivalent circuit circuit parameters of a HEMT device with $0.2\,\mu$m of gate length and $100\,\mu$m of gate width under the bias that $V_{gs} = -0.4$V and $V_{ds} = 2$V.

Element	L_g (pH)	L_d (pH)	L_s (pH)	C_{pg} (fF)	C_{pd} (fF)	R_g (Ω)	R_s (Ω)
Value	46	8	35.9	7.7	4.5	3.13	4.22

R_d (Ω)	C_{gd} (fF)	C_{gs} (fF)	R_{gs} (Ω)	g_m (mS)	τ (psec)	C_{ds} (fF)	R_{ds} (Ω)
6.82	11.7	152.8	4.83	54.4	0.5226	30.5	367



(a) $V_{gs} = -1\,\mathrm{V}$



(b) $V_{gs} = 0\,\mathrm{V}$

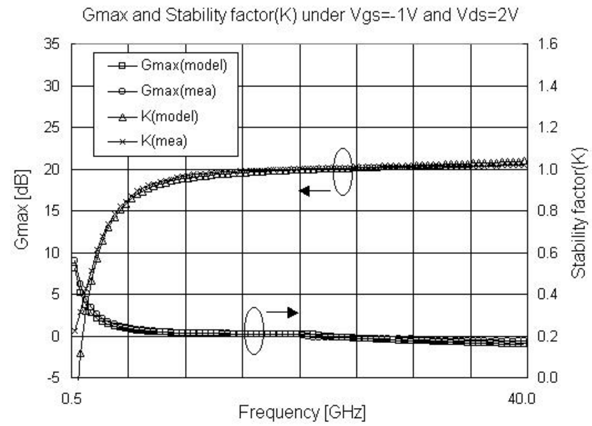
Fig. 6 Comparison between the calculated and measured S-parameters under $V_{ds} = 2\,\mathrm{V}$ and various gate voltages.

5. Conclusion

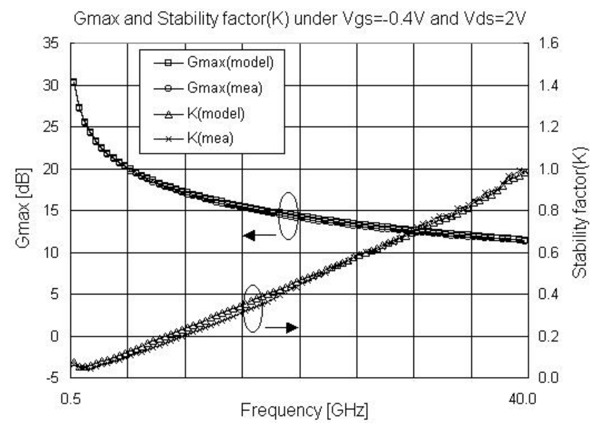
We proposed a new method for determining the extrinsic resistances as explicit formulae from the predetermined parasitic capacitances and inductances, and the measured S-parameters under active bias. The ΔR_{ds}, difference between R_d and R_s, has been found from the measured Z-parameters under cold bias. This ΔR_{ds} reduced the number of unknowns and made it possible to solve the unique solution for extrinsic resistances as explicit formulae.

The advantages of the proposed method are summarized as follows. First, no additional DC measurement for the device under test is required. Second, no requirement for iteration or optimization is needed. Third, no assumptions for the ratio between the resistances are made, and no need for optimization or iteration for these ratios is required. Finally, there is no problem of complex solutions for R_s. The unique solution set for R_s, R_d, and R_g are calculated by simple explicit formulae. All parasitic resistances are extracted from the S-parameters measured under normal active bias.

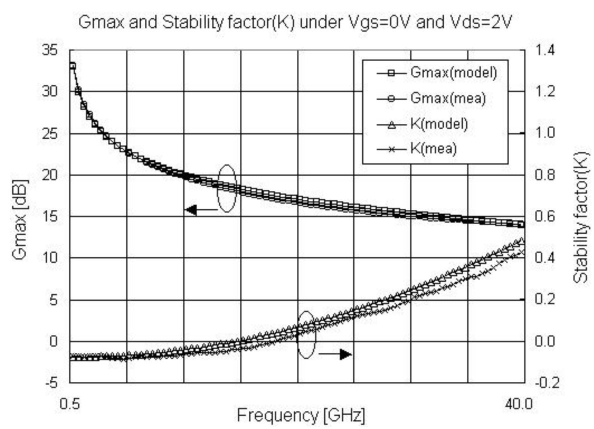
By applying the proposed method to a HEMT device of $0.2\,\mu\mathrm{m} \times 100\,\mu\mathrm{m}$ under various bias conditions, we found the extrinsic resistances and all the other equivalent circuit parameters. We calculated the S-parameters, maximum available gain, and stability fac-



(a) $V_{gs} = -1\,\mathrm{V}$



(b) $V_{gs} = -0.4\,\mathrm{V}$



(c) $V_{gs} = 0\,\mathrm{V}$

Fig. 7 G_{max} and stability factor (K) from the modelled and measured S-parameters under $V_{ds} = 2\,\mathrm{V}$ and various gate voltages. G_{max} means maximum available gain if $K \geq 1$, otherwise most stable gain (MSG).

tor, compared these to the measured ones, and verified the agreement between them. The validity of the proposed method has been extended up to 40 GHz. This implies that the proposed method has the possibility of application up to high millimeter wave range beyond 40 GHz.

Acknowledgements

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Appendix: Verification of the Very Weak Effects of $G_{int,11}$ in (20)–(23)

By expanding (19) for $\mathrm{Im}(Z_{11})$ and $\mathrm{Im}(Z_{12})$, dividing them into real and imaginary parts, and using the fact $G_{int,12}$ is zero from (10), (A·1)–(A·4) are obtained. If the effects of the first terms are so weak compared to the other terms, (A·1)–(A·4) can be rewritten as (20)–(23). In this Appendix, the magnitudes of the first terms will be compared to the other terms in order to verify that there are little differences in the resulting sums in (A·1)–(A·4) irrespective of ignoring the first terms.

$$G_{int,11}\{\mathrm{Re}(Z_{11})-(R_g+R_s)\} \\ -B_{int,11}\mathrm{Im}(Z_{11}) - B_{int,12}\mathrm{Im}(Z_{21}) = 1 \quad (\mathrm{A}\cdot 1)$$

$$G_{int,11}\mathrm{Im}(Z_{11}) + B_{int,11}\{\mathrm{Re}(Z_{11}) \\ -(R_g+R_s)\} + B_{int,12}\{\mathrm{Re}(Z_{21})-R_s\} = 0 \quad (\mathrm{A}\cdot 2)$$

$$G_{int,11}\{\mathrm{Re}(Z_{12})-R_s\} - B_{int,11}\mathrm{Im}(Z_{12}) \\ - B_{int,12}\mathrm{Im}(Z_{22}) = 0 \quad (\mathrm{A}\cdot 3)$$

$$G_{int,11}\mathrm{Im}(Z_{12})+B_{int,11}\{\mathrm{Re}(Z_{12})-R_s\} \\ +B_{int,12}\{\mathrm{Re}(Z_{22})-(R_d-R_s)\} = 0 \quad (\mathrm{A}\cdot 4)$$

Provided that the first terms are smaller than other terms by one tenth or less, they can be ignored in the engineering sense and (20)–(23) are acceptable. The ratios between the first terms and other terms were calculated up to 40 GHz using (A·1)–(A·4) and the R_s, R_d, and R_g shown in Table 1. The results are illustrated in Fig. A·1. From this plot, this approximation is well acceptable in (A·1) and (A·3) up to 40 GHz, and acceptable up to 32 GHz and 35 GHz for (A·2) and (A·4), respectively. Therefore, in the strictest sense,

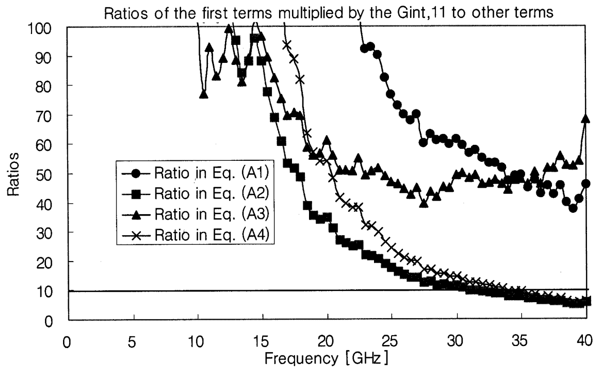

Fig. A· 1 Ratios between the first terms multiplied by $G_{int,11}(=\omega^2 C_{gs}^2 R_{gs})$ and other terms in (A· 1)–(A· 4).

the extrinsic resistances should be calculated by averaging the values shown in Fig. 4 over 5–32 GHz. However, as can be seen in Fig. 4, the extrinsic resistances at each frequency go almost constant over 5 GHz up to 40 GHz, and the finally resulting extrinsic resistances are so similar for any averaging range over 5–32 GHz or 5–40 GHz. Therefore it can be said that the proposed method in this paper is well applicable up to 40 GHz.

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