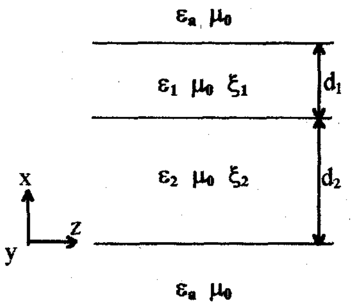

Fig. 4. A two-layer open chirowaveguide.

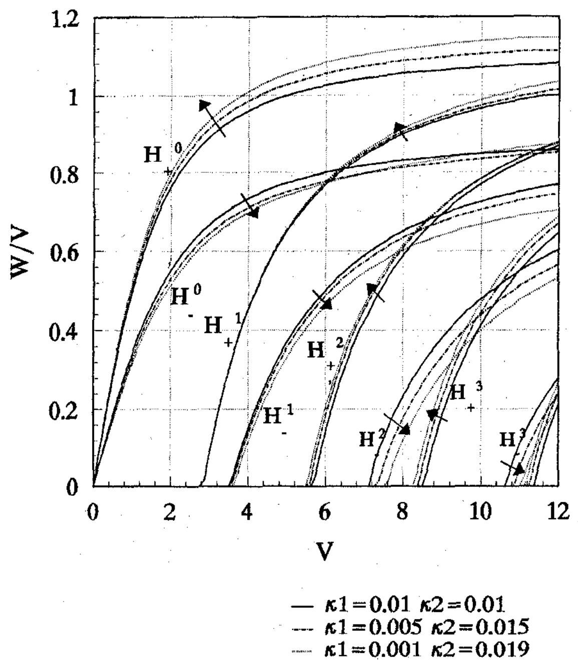

Fig. 5. Dispersion property for the open structure shown in Fig. 4. $v = k_o D\sqrt{\epsilon_1 - \epsilon_a}, w/V = \sqrt{\beta^2 - \epsilon_a}/\sqrt{\epsilon_1 - \epsilon_a}$.

To verify the effectiveness of the network method developed in this paper, we have calculated a two-layer open structure as shown in Fig. 4. The value of media parameter are assigned as

$$\epsilon_1 = \epsilon_2 = 4.41, \quad \epsilon_a = 4.0, \quad \xi_1 = -j2.1\kappa_1,$$
$$\xi_2 = -j2.l\kappa_2, \quad d_1 = D/3, \quad d_2 = 2D/3.$$

When κ_1 and κ_2 are chosen to be equal, we get exactly the same results as in [2]. When κ_1 and κ_2 deviate from the initial value of 0.01 ($\kappa_{2,1} = 0.01 \pm 0.005$, or $\kappa_{2,1} = 0.01 \pm 0.009$), the change of eigenvalue is shown in Fig. 5 by the arrows. Because of $d_2 > d_1$, it is expected that κ_2 contributes more to the change than κ_1 does, i.e. as κ_2 increases, the pairs of curves originally having the same cut off in the chiral limit will more and more separate from each other [2]. This trend can be observed in Fig. 5. Judging from this, the effectiveness and accuracy of the present approach are thus verified, though no comparison is given in Fig. 5 between our results and others because of no data available for a two-layer open chirowaveguide in the literature.

III. Conclusion

This paper presents an equivalent network method for the dispersion analysis of general multilayered planar chirowaveguides. Use has been made of the concepts of multimode network method for planar dielectric waveguides. Different kinds of impedance transformation relations are given, including the transformation of input impedance matrix within a homogeneous layer, the impedance matrix transformation at the interface of two media and the input impedance matrix for a layer terminated with an open or short circuit. Also, the transverse resonance technique is extended to treat the chirowaveguides.

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Fast and Efficient Extraction of HBT Model Parameters Using Multibias S-Parameter Sets

Seonghearn Lee

***Abstract*—Accurate parameter extraction technique has been presented for a small-signal equivalent circuit model of AlGaAs/GaAs HBT's. This technique makes use of multibias data optimization regarding two sets of S-parameters in the active mode and one in the cut-off mode, under the physics-based constrain that current-dependent elements in two active bias circuits are linked each other by the ratio of their currents. This multibias optimization as well as the constrain imposed on intrinsic parameters may reduce the degree of freedom of circuit variables and increase the probability of finding a global minimum result. As a result of this extraction, good agreement is seen between the circuit models and their measured S-parameters in the frequency range of 0.045 to 26.5 GHz.**

I. Introduction

For the development of microwave circuit applications using heterojunction bipolar transistors (HBT's), it is essential to use an accurate HBT equivalent circuit model for simulating monolithic microwave integrated circuit (MMIC). Although physical and analytical HBT models have been reported previously [1], [2], an empirical HBT model requiring circuit parameters extracted from measurements has been generally used. In order to provide precise parameter values, reliable and efficient extraction method should be established. In a typical approach, a small-signal equivalent circuit model is optimized

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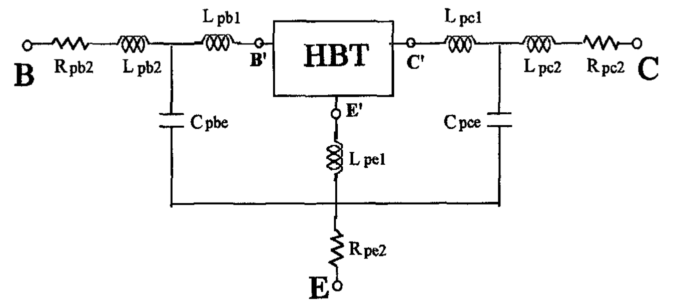

Fig. 1. A small-signal equivalent circuit model for RF probe pads and interconnections.

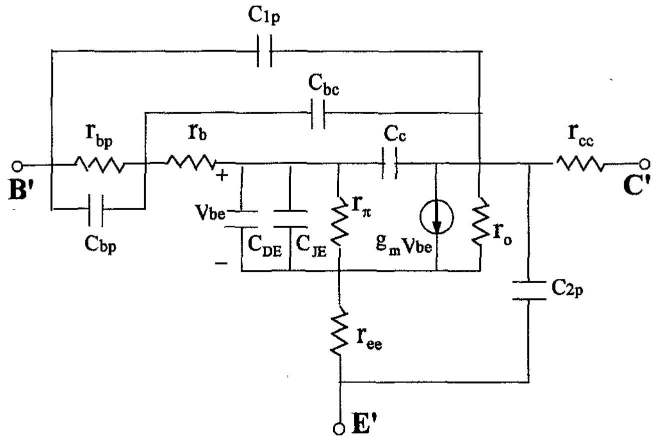

Fig. 2. A small-signal equivalent circuit model for AlGaAs/GaAs HBT biased to the active mode without probe pads and interconnections.

to fit the measured S-parameters as close as possible. However, it suffers the problem such as nonuniqueness of extracted parameters [3]. As a method to overcome this problem, it was reported that an approach using multiple sets of measurement contributes to enhancing the degree of unique identification of parameters [4], [5]. Such a merit of this approach was verified by attempting to achieve the best possible match between each multiple sets of S-parameters and their equivalent circuits simultaneously [4], [5], but the number of unknown variables increases rapidly because of the addition of perturbed circuit parameters differing from original circuit. Therefore, this large number may produce another uncertainties in finding an unique solution, depending on the initial values during the optimization. To solve this problem, it is essential to reduce the number of unknown parameters by imposing physics-based constrains on some unknowns [3] or using independent measurements [6].

In this paper, the uncertainties are minimized by using three different sets of measured S-parameters where the first set is obtained under the bias condition of cut-off mode, and the second and third are measured after applying two different currents in the active mode. Using three data sets, multibias data finding is performed while constraining current-dependent intrinsic elements in the second circuit to vary with those in the third circuit multiplied by the ratio of their currents.

TABLE I
EXTRACTED SMALL-SIGNAL EQUIVALENT CIRCUIT MODEL PARAMETERS OF AlGaAs/GaAs HBT INCLUDING RF PROBE PADS AND INTERCONNECTIONS FOR DATA SET A

L_{pb1}	40.0 pH	r_{ee}	9.0 Ω
L_{pc1}	23.0 pH	r_{cc}	8.4 Ω
L_{pe1}	15.0 pH	r_o	3.6×10^7 Ω
L_{pb2}	27.0 pH	C_{DE}	572 fF
L_{pc2}	27.0 pH	C_{JE}	120 fF
R_{pb2}	9.0 Ω	C_c	28.1 fF
R_{pc2}	4.4 Ω	C_{bc}	68.3 fF
R_{pe2}	2.3 Ω	C_{bp}	297 fF
C_{pbe}	15.0 fF	C_{1p}	12.7 fF
C_{pce}	20.8 fF	C_{2p}	35.6 fF
r_π	520 Ω	g_{mo}	95 mS
r_b	136 Ω	τ_d	1.8 ps
r_{bp}	360 Ω		

II. PARAMETER EXTRACTION TECHNIQUE

The novel parameter extraction technique was applied to extract the parameters of Al graded ($x = 0.1$–0) base n-p-n AlGaAs/GaAs HBT's (0.2 μm thick, $1 \times 10^{19}\,\text{cm}^{-3}$ doped base, emitter area of $5.5 \times 5.5\,\mu\text{m}^2$) with a non self-aligned structure. S-parameter measurements for HBT's biased at two different currents ($I_c = 2.51$ mA and $V_{CB} = 1.6$ V for data set A, $I_C = 0.35$ mA and $V_{CB} = 1.6$ V for data set B) were carried out using on-wafer RF probes from 0.045 to 26.5 GHz. Since the probe pads and interconnections for on-wafer probing add large parasitics to the equivalent circuit, these have precisely been modeled as Fig. 1, and the accuracy of this model was demonstrated previously in the frequency range of 0.045 to 26.5 GHz [7].

An extended hybrid-π small-signal equivalent circuit in Fig. 2 is developed to model an AlGaAs/GaAs HBT accurately in this work. In this model, C_{DE} is the emitter-base diffusion capacitance, C_{JE} is the emitter-base junction capacitance, r_{bp} is the base contact resistance, C_{bp} is the base contact capacitance [7], capacitances (C_{1p} and C_{2p}) are associated with SiO_2 between interconnect metals and n^+ collector contact layer, and the transconductance is expressed as [6]: $g_m = g_{mo}\exp(-j\omega\tau_d)$, where g_{mo} is its dc value and τ_d is its phase delay.

S-parameter measurements are repeated on the same HBT biased to the cut-off mode at $V_{CB} = 1.6$ V and $V_{BE} = -1$ V (data set C). In order to modify Fig. 2 into a cut-off mode circuit, $C_{DE}, r_\pi, g_m V_{be}$, and r_o are omitted and C_{JE} is replaced as C due to cut-off biasing [7]. Since this cut-off circuit is a kind of passive open-circuit, capacitors may be more sensitive to the error function for optimization than resistances and inductances. Therefore, the virtual number of unknown parameters in optimization is likely to be reduced, because the cut-off mode circuit is essentially simpler than the active mode circuit and the error function is mainly governed by capacitances in the frequency of measurement. Thus, if this cut-off circuit is included in the multicircuit optimization, it may increase the chance of uniquely identifying unknown variables.

The two active mode circuits and a cut-off mode circuit are simultaneously optimized to fit their measured S-parameters using the Touchstone [8] until no further change in the parameter values is seen. Since pad and interconnection parameters, capacitances ($C_c, C_{bc}, C_{bp}, C_{1p}, C_{2p}$) and resistances ($r_b, r_{cc}, r_{ee}, r_{bp}$) are assumed to have same values in three circuit models, these parameters are assigned as common variables in this optimization. This assumption seems to be valid, since the reduction of r_b due to current crowding can be ignored in AlGaAs/GaAs HBT, and current dependence of C_c and C_{bc} due to widening of base-collector depletion region [9] can be neglected in a given current range below base pushout. However, some of intrinsic parameters ($C_{DE}, r_\pi, g_{mo}, r_o$) for one of the active circuit are connected to those for the other by current-ratio equation. The current dependences of these intrinsic

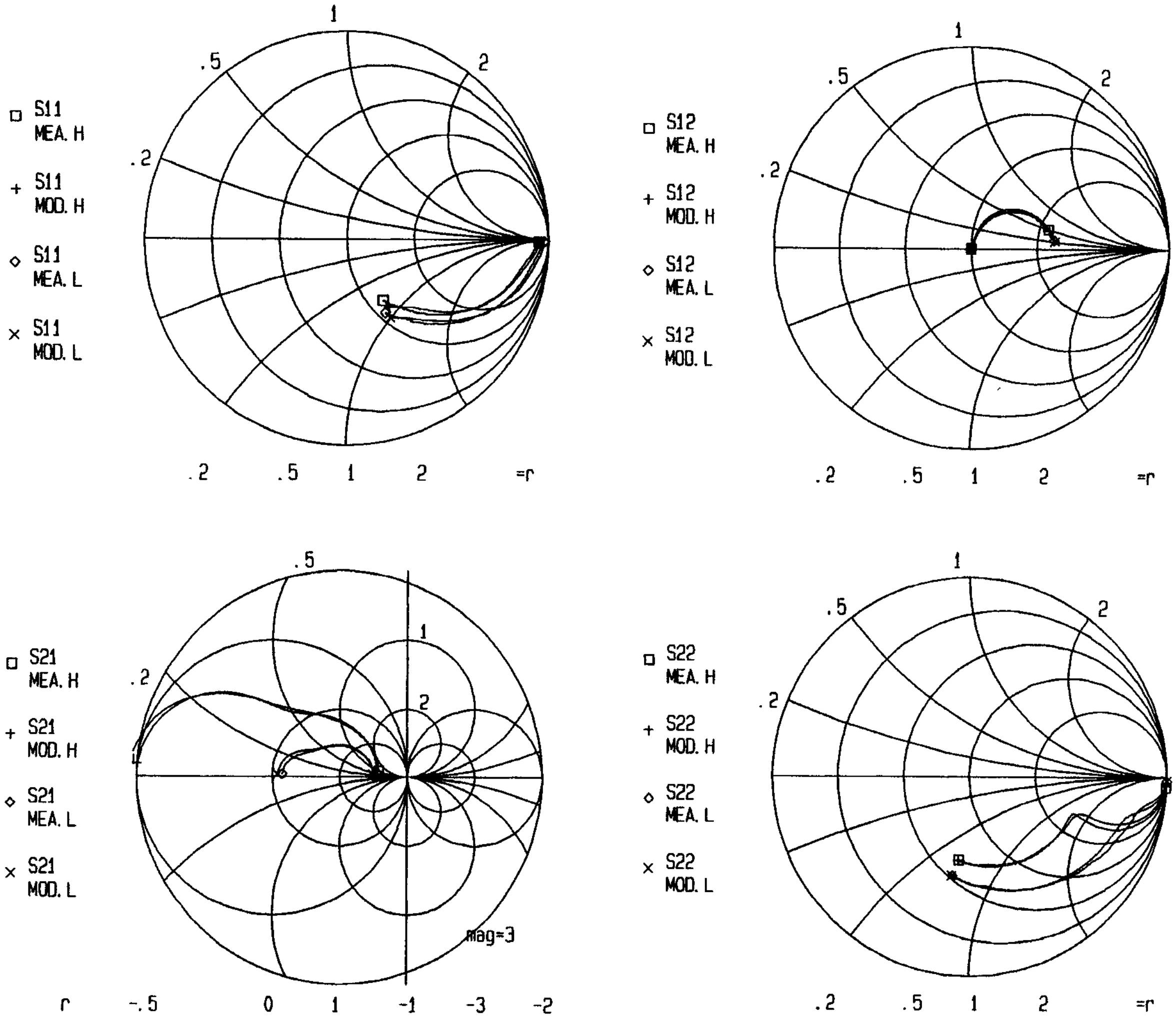

Fig. 3. Comparison between measured (MEA) and modeled (MOD) S-parameters from 0.045 to 26.5 GHz at different currents. The extension name of .H and .L represents data set A and data set B, respectively.

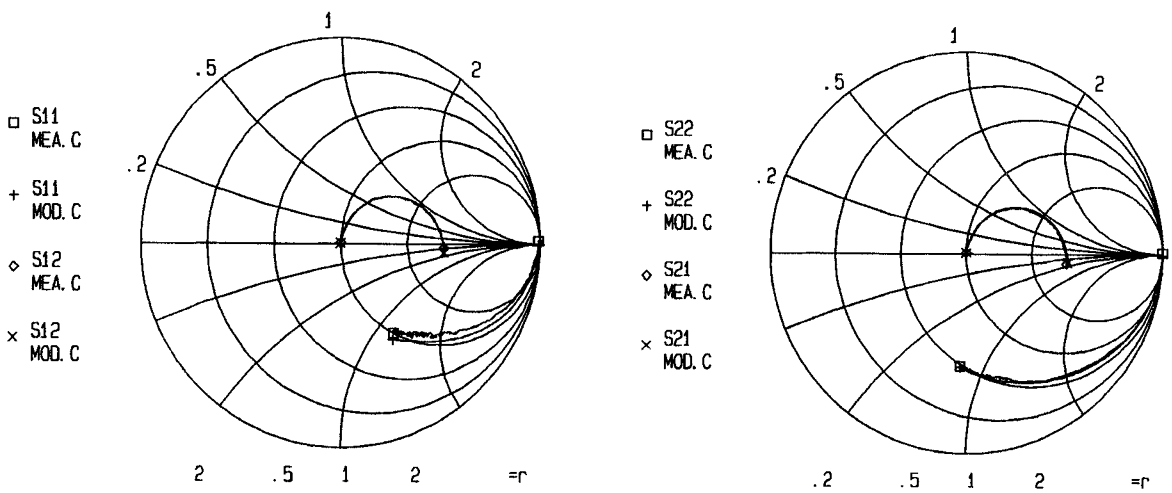

Fig. 4. Comparison between measured (MEA.C) and modeled (MOD.C) S-parameters from 0.045 to 26.5 GHz for data set C in the cut-off mode.

parameters are described as

$$g_{mo} = \frac{qI_c}{n_c kT}, \quad r_\pi = \frac{n_b kT}{qI_b},$$
$$r_o = \frac{|V_A|}{I_c}, \quad C_{DE} = \frac{q\tau_F I_c}{n_c KT}$$

where n_c is the ideality factor of I_c, n_b is the ideality factor of I_b, V_A is the Early voltage, and τ_F is ideal forward transit time.

Using the above equations and assuming constant n_c and n_b versus current over a given current range, the current-dependent elements of one of active mode circuit can be expressed by those of the other multiplied by the ratio of both currents as follows:

$$g_{mo2} = \frac{I_{c2}}{I_{c1}} g_{mo1}, \quad r_{\pi 2} = \frac{I_{b1}}{I_{b2}} r_{\pi 1},$$
$$r_{o2} = \frac{I_{c1}}{I_{c2}} r_{o1}, \quad C_{DE2} = \frac{I_{c2}}{I_{c1}} C_{DE1}.$$

The two bias currents must be chosen within the current range where n_c and n_b remain unchanged. These parameters in one of active mode circuit are constrained by the above formulations using EQN block in Touchstone, while three circuits are optimized simultaneously. These equations force the above intrinsic parameters to comply with a physics-based model. In addition to these constrains, an accurate optimization for obtaining the physical values of device parasitics is also important because the performance is generally influenced by parasitics. For this accurate optimization, parasitics were independently measured (or calculated), and these values are then used as initial values with narrow bounds. The resistances are obtained from dc I-V [10] or ac de-embedded Z-parameters [6], those of capacitances are from ac de-embedded Y-parameters in the cut-off mode, and rest of parameters were calculated with theoretical methods [11]. In the ac extraction method, some dispersion effects between low and high frequency extracted data may exist because base-collector distributed components and external capacitances ignored at low frequencies become important factors in the high frequency region. These effects become important for this HBT, because of large extrinsic capacitances. Thus, great care must be taken to extract the parameters.

The values of C_c and C_{bc} are extracted using the result of $C_c + C_{bc} = 89$ fF obtained from $\text{Im}[Y_{12}]$ of HBT biased to cut-off mode at low frequencies. For 0.7 μm thick $SiO_2, C_{1p} = 11$ fF and $C_{2p} = 18$ fF. The term of C_{JE} is extracted to be 0.12 pF because the slope of total emitter-collector transit time $[\tau_{ec} = 1/(2\pi f_T)]$ vs. $1/I_c$ plot is $C_{JE} + C_c + C_{bc} + C_{1p}$. Using $g_{mo} = qI_c/n_c kT, g_{mo} = 95.5$ mS. The value of $r_{ee} \approx 12\,\Omega$ is extraced by using $r_{ee} \approx \text{Re}(Z_{12}) - 1/g_{mo}$ at low frequencies [6]. These initial values result in very rapid optimization convergence.

The extracted parameters of equivalent circuit for data A are listed in Table I. The modeled S-parameters of three circuits including probe pads and interconnections are compared with their measured S-parameters in Figs. 3 and 4, and show excellent agreement from 0.045 to 26.5 GHz. Comparing these extracted values with the initial values, good agreement is obtained. This good agreement demonstrates that these parameters extracted from the new technique are physically reliable. However, the fact that extracted C_{2p} is larger than the calculated value may be due to the fringing capacitance caused by the proximity of n^+ collector contact layer to emitter pad. Since the number of unknown variables is virtually reduced by both cut-off biasing and current-ratio related constrains, this new method may provide a higher degree of confidence in the validity of extracted parameters, compared with the traditional method of direct fitting. As a different approach, direct extraction technique employing analytical expressions derived from HBT equivalent circuit have been reported [12], but this technique is not suitable for the HBT under this study because the complexity of Fig. 2 makes analytical derivations difficult.

III. Conclusion

Accurate extraction technique using multibias optimization has been described for determining small-signal model parameters of AlGaAs/GaAs HBT's. Three equivalent circuits for a cut-off biasing and two active biasing at different currents are optimized simultaneously by minimizing the error function, under the physics-based constrain that intrinsic elements for one of active mode circuit are connected to those for the other multiplied by the ratio of two currents. The cut-off mode circuit and the constrain give the advantage of speeding up unique identification of parameters, because the variable space dimension of equivalent circuit is reduced by biasing to cut-off and constraining current-dependent variables. Using this method, small-signal equivalent circuits at different bias currents are determined simultaneously by only one optimization process. After this optimization, three sets of modeled S-parameters agree well with their measured S-parameters.

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