Extraction and verification of the small-signal model for InP DHBTs in the 0.2–325 GHz frequency range

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Abstract: This study presents the extraction and verification of a small-signal model suitable for characterizing THz InP double heterojunction bipolar transistors (DHBTs). The π-type topology is adopted in the intrinsic model. Capacitances $C_{ce}$ and $C_{cc}$ are used to characterize the capacitive parasitics caused by the routing line connecting the collector terminal, base terminal and emitter terminal, respectively. The inductive parasitics introduced by the routing line are also considered. The initial values of the model parameters are extracted using a direct extraction method. The model and extraction method for the model parameters are verified by adopting an InP DHBT with 1 emitter finger and an emitter size of 0.5 µm × 5 µm. The simulation results correspond with the measured results in the frequency range from 200 MHz to 325 GHz.

Keywords: InP DHBTs, small-signal model, extraction, verification

Classification: Microwave and millimeter-wave devices, circuits, and modules

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1 Introduction

Compared with GaAs heterojunction bipolar transistors (HBTs), InP HBTs are advantageous in terms of their DC gain, high frequency performance, power consumption and 1/f noise due to the inherent characteristics of the material in the system. After decades of development as a typical InP HBT device, InP DHBTs have gradually approached application frequency bands in the THz [1, 2] range and are considered one of the most promising solutions for THz technology [3], making such transistors a topic of interest in the fields of high-speed, low-power millimeter wave, submillimeter and ultrafast devices and related integrated circuit technologies.

The small signal model is not only the basis of the large signal model but also an important support means to guide structure optimization and process improvement. Small signal models that can be used for InP HBT/DHBT modeling have been extensively reported [4, 5, 6, 7, 8, 9, 10, 11] and mainly verified below 100 GHz, with a few reported results reaching 200 GHz. [5] Considering that the characteristic frequency of the InP DHBT device has been approaching the THz range, few studies have verified the small signal model for an InP DHBT device beyond 100 GHz and the model parameter extraction technology. Based on traditional modeling techniques, this paper first attempts to extend the small-signal model modeling technology of InP DHBTs to 325 GHz and verify the validity of the model using measured data.

The $\pi$-type equivalent circuit model was adopted for the topology of this model in the intrinsic region to make it more in line with the physical structure of InP DHBT devices. Fifteen elements are used in the model to achieve a compromise between the simplicity and accuracy of the model topology. Authors use an
analytical extraction method to extract the initial values of the circuit elements and curve-fit them to the measured S parameters to obtain the final optimized results.

2 Model topology

The proposed small-signal model for InP DHBTs is shown in Fig. 1. $R_b$, $R_c$ and $R_e$ are the base, collector and emitter parasitic resistances, respectively. $C_{cx}$ and $C_{ce}$ were introduced for characterization considering the capacitive parasitics caused by the routing line connecting the collector terminal to the base terminal and emitter terminal, respectively. The inductive parasitics introduced by the routing line connecting the base terminal, collector terminal and emitter terminal were characterized using inductances $L_b$, $L_c$ and $L_e$, respectively. $R_{bc}$ and $R_{be}$ are the intrinsic base-collector and base-emitter resistances, and $C_{bc}$ and $C_{be}$ are the intrinsic base-collector and base-emitter capacitances, respectively. $R_{bi}$ is the intrinsic base distributed resistance, and $R_{ce}$ is the output resistance. $G_m$ and $G_{m0}$ are the small-signal and DC transconductances, respectively. $\tau$ is the delay time. To facilitate the following analysis and discussion, the portion enclosed in a dashed box in Fig. 1 is labeled Inner, and the corresponding Y and Z parameters are identified as $Y_{in}$ and $Z_{in}$, respectively.

3 Transistor small-signal equivalent circuit extraction

The majority of studies in the literature have reported on the extraction of small signal model parameters for III–V or SiGe HBT, which often introduced small parasitic inductances at the port leads (typically at several pH levels) and encountered greater accuracy problems with the extraction of the intrinsic base resistance. The frequency response of several pH-level lead parasitic inductances in the network parameters at HBT device magnification typically occurs beyond 100 GHz, which is the main reason for the traditional derivation of parasitic inductance extraction methods, which tend to be less effective at low frequencies. When the port lead resistance and inductance can be accurately determined and stripped from the total measured data, the intrinsic transistor elements can be extracted at a lower frequency, allowing for a model parameter extraction strategy at low frequency that can be applied to the extraction at 300 GHz.

In this letter, the port parasitic resistance and inductance of InP DHBT devices are extracted from the perspective of engineering applications, and the accuracy of
the model’s parameter extraction is determined using the method reported in [11]. Considering the topological structure of the model, as shown in Fig. 1, when derived as a Z-parameter, \( Y_m \) has a negligible influence and can thus be neglected. The calculation for port lead parasitics in [11] is as follows: 
\[
R_b = \text{real}(Z_{11} - Z_{12}), \quad R_c = \text{real}(Z_{22} - Z_{12}), \quad R_e = \text{real}(Z_{12}), \quad L_b = \text{imag}(Z_{11} - Z_{12})/\omega, \quad L_c = \text{imag}(Z_{22} - Z_{12})/\omega, \quad R_e = \text{imag}(Z_{12})/\omega.
\]

Using the above method, the port parasitic resistances and inductances of the InP DHBT device with bias conditions of \( I_{bc} = 400 \mu A \) and \( V_{ce} = 1.5 \) V are extracted, and the results are shown in Figs. 2(a) and (b). Resistances \( R_b \) and \( R_c \) tend to stabilize at frequencies beyond 20 GHz, \( R_e \) tends to be constant at frequencies over 40 GHz, and the parasitic inductances \( L_b, L_c, \) and \( L_e \) tend to stabilize at frequencies beyond 80 GHz (see Fig. 2). The results extracted by this algorithm exhibit good frequency stability and can be used to reliably extract port lead parasitics.

Once the port lead parasitics are determined, the impedance matrix \( Z_m \) of the inner part can be stripped from the total two-port Z-parameters by matrix subtraction, as shown below.

\[
Z_m = Z_t \left( \begin{array}{cc} R_b + R_c + j\omega (L_b + L_c) & R_c + j\omega L_c \\ R_c + j\omega L_c & R_c + R_e + j\omega (L_c + L_e) \end{array} \right)
\]

(1)

\( Z_t \) is the two-port measured Z parameters of the transistor. \( Y_{in} = Z_m^{-1} \). With reference to the dashed box portion of Fig. 1, \( Y_{in} \) can also be expressed as

\[
Y_{in} = \begin{pmatrix} Y_{in11} & Y_{in12} \\ Y_{in21} & Y_{in22} \end{pmatrix} = \begin{pmatrix} \frac{Z_{bc} + j\omega C_{cs}}{N} & -\frac{Z_{bc} - j\omega C_{cs}}{N} \\ \frac{Z_{bc} G_m - \frac{Z_{bc}}{N} - j\omega C_{cs}}{N} & \frac{Z_{bc} G_m + \frac{Z_{bc}}{N} + j\omega C_{cs} + \frac{1}{Z_{ce}}} {N} \end{pmatrix}
\]

(2)

where

\[
G_m = G_{m0} e^{-j\omega \tau}
\]

(3)

\[
Z_{bc} = R_{bc}/(1 + j\omega R_{bc} C_{bc})
\]

(4)

\[
Z_{ce} = R_{ce}/(1 + j\omega R_{ce} C_{ce})
\]

(5)

\[
Z_{be} = R_{be}/(1 + j\omega R_{be} C_{be})
\]

(6)

\[
N = Z_{be} Z_{be} + Z_{bc} R_{bi} + Z_{he} R_{bi}
\]

(7)

At this point, the circuit parameters are determined analytically below.

When an HBT is operated under low frequencies, the impedance of the intrinsic base resistance \( R_{bi} \) is sufficiently small [7] that it can be ignored. The admittance is expressed as follows:

\[
Y_{in11} + Y_{in12} = R_{bc}^{-1} + j\omega C_{be}
\]

(8)

\( R_{bc} \) and \( C_{be} \) can be extracted from the real and imaginary parts of Eq. (8) as follows: \( R_{bc} = \text{real}(Y_{in11} + Y_{in12})^{-1} \) and \( C_{be} = \text{imag}(Y_{in11} + Y_{in12})/\omega. \)

From (2) and (6), one can deduce

\[
\frac{Y_{in11} + Y_{in12}}{Y_{in21} - Y_{in22}} = \frac{1}{G_m Z_{be}}
\]

(9)

or, equivalently,
\[ G_m = G_{m0}e^{-j\omega \tau} = \frac{(Y_{in21} - Y_{in12})(1 + j\omega R_{bc}C_{be})}{(Y_{in11} + Y_{in12})R_{be}} \] (10)

By substituting the extracted \( R_{bc} \) and \( C_{be} \) into (10), \( G_{m0} \) can be obtained from the magnitude of (10). \( G_{m0} = \text{Mag}[(Y_{in21} - Y_{in12})(1 + j\omega R_{bc}C_{be})/R_{be}(Y_{in11} + Y_{in12})] \).

In this work, \( \tau \) is calculated as \( \tau = 1/(2\pi f_1) \), where \( f_1 \) is the cutoff frequency of the device.

Using (2), (4) and (7), \( N/Z_{bc} \) can be written as

\[
\frac{1}{Y_{in11} + Y_{in12}} = \frac{N}{Z_{bc}} = \frac{R + j\omega X}{1 + j\omega T} \] (11)

with \( R = R_{bi} + R_{bc} + R_{bi}R_{bc}/R_{bc} \), \( X = R_{bi}R_{bc}(C_{bc} + C_{be}) \), and \( T = R_{bc}C_{be} \).

Equation (11) can be written as a linear system as follows:

\[
(1 + j\omega T)/(Y_{in11} + Y_{in12}) = R + j\omega X
\] (12)

or, equivalently,

\[
\text{real}[(1 + j\omega T)/(Y_{in11} + Y_{in12})] = R = R_{bi} + R_{bc} + R_{bi}R_{bc}/R_{bc}
\] (13)

\[
\text{imag}[(1 + j\omega T)/(Y_{in11} + Y_{in12})]/\omega = X = R_{bi}R_{bc}(C_{bc} + C_{be})
\] (14)

In practical applications, the base distributed resistances \( R_{bi} \) and \( R_{bc} \) are considerably smaller than \( R_{bc} \). Equation (13) can be simplified as follows:

\[
\text{real}[(1 + j\omega T)/(Y_{in11} + Y_{in12})] = R = R_{bi} + R_{bc}
\] (15)

From Eqs. (14) and (15), \( R_{bi} \) and \( C_{bc} \) can be determined as \( R_{bi} = \text{real}[(1 + j\omega T)/(Y_{in11} + Y_{in12})] - R_{bc} \) and \( C_{bc} = \text{imag}[(1 + j\omega T)/(Y_{in11} + Y_{in12})]/(\omega R_{bi}R_{bc}) - C_{be} \).

From (2), we obtain

\[ Y_{in12} = -Z_{bc}/N - j\omega C_{cx} \] (16)

By substituting Equations (6) and (7) into Equation (16), \( R_{bc} \) can be determined by the real part of Eq. (16) (i.e., \( \text{real}(Y_{in12}) = \text{real}(-Z_{bc}/N) \)) once \( R_{bc}, C_{be}, R_{bi} \) and \( C_{bc} \) have been extracted. Thus, the imaginary part of Eq. (16) yields \( C_{cx} = -\text{imag}(Y_{in12} + Z_{bc}/N)/\omega \).

The same methodology is used to extract \( R_{ce} \) and \( C_{ce} \). The real and imaginary parts of \( Y_{in22} \) yield \( R_{ce} = \{\text{real}(Y_{in22}) - \text{real}[(R_{bi}Z_{bc}G_m + Z_{be} + R_{bi})/N]\}^{-1} \) and \( C_{ce} = \text{imag}(Y_{in22}) - \text{imag}[(R_{bi}Z_{bc}G_m + Z_{be} + R_{bi})/N - j\omega C_{cx}]/\omega \), respectively.

### 4 Verification of the model parameter extraction

For verification, a single-finger emitter InP DHBT device with an emitter size of \( 0.5 \times 5 \mu m^2 \) featuring an \( f_1 \) of more than 550 GHz was fabricated at the Nanjing Electronic Devices Institute in China. The S-parameters were measured using an Agilent E8461A network analyzer (200 MHz–66 GHz), a vector network analyzer frequency extenders from Farran Technology FEV-10-TR (75–110 GHz), FEV-06-TR (110–170 GHz), FEV-05 (140–220 GHz), FEV-03 (220–325 GHz) and a probe station from Cascade Microtech (Summit 11000B-S). On-wafer test structures for the InP HBT device with a traditional open/short de-embedding method were used in the 0.2–66 GHz frequency band. For the measured data with frequencies above 75 GHz, the DUT was designed as a TRL measurement connection structure, and...
the system was calibrated through the corresponding TRL calibration kits to synchronize calibration and de-embedding of the test structure parasitics. After the test system was calibrated, the DC power was supplied using the HP 4156C Precision Semiconductor Parameter Analyzer. The small-signal S-parameters of the device from 200 MHz to 325 GHz were analyzed with biases of Bias 1 ($V_{ce} = 1.0 \text{ V}, I_{be} = 200 \mu\text{A}$), Bias 2 ($V_{ce} = 1.5 \text{ V}, I_{be} = 200 \mu\text{A}$), Bias 3 ($V_{ce} = 1.0 \text{ V}, I_{be} = 400 \mu\text{A}$) and Bias 4 ($V_{ce} = 1.5 \text{ V}, I_{be} = 400 \mu\text{A}$).

In this paper, the measured S-parameters at the bias conditions of $V_{ce} = 1.5 \text{ V}$ and $I_{be} = 400 \mu\text{A}$ were selected to illustrate the parameter extraction process. The extraction results are shown in Figs. 2–5. $R_b$, $R_c$ and $R_e$ are 3.2 $\Omega$, 3.8 $\Omega$ and 7.26 $\Omega$ (see Fig. 2(a)), respectively, and $L_b$, $L_c$ and $L_e$ are 2.6 pH, 4.4 pH and 2.7 pH, respectively (see Fig. 2(b)). The experimental results of $R_{bc}$ and $R_{be}$ are shown in

![Fig. 2. Relationship between the frequency and extracted resistances ($R_b$, $R_c$ and $R_e$) (a) and inductances ($L_b$, $L_c$ and $L_e$) (b)](image)

![Fig. 3. Extraction result of $R_{bc}$, $R_{be}$ is determined as when the real part of $-Z_{bc}/D$ is equal to the real part of $Y_{in12}$.)](image)

![Fig. 4. Extraction results of the capacitances ($C_{bc}$, $C_{bc}$, $C_{cx}$ and $C_{ce}$) and resistance $R_{ce}$)](image)
Fig. 3. The average value of the resistance in the frequency range of 200 MHz to 66 GHz is considered for the resistance: $R_{be}$ is 54 $\Omega$. Fig. 5 shows the extraction of $C_{be}$, $C_{ce}$, $C_{bc}$ and $C_{cx}$. The right ordinate in Fig. 4 is the algorithm used to extract $R_{ce}$. The value of $R_{ce}$ is $2.9 \times 10^3$ $\Omega$, and was obtained in the frequency range where the resistance is relatively stable. The calculation results of the intrinsic base resistance $R_{bi}$ and $G_m^0$ are shown in Fig. 5.

After the parameters were extracted, the obtained model parameters were fine-tuned using Keysight ICCAP software to achieve a better fitting accuracy between the simulated results and measured data. Because the external parasitic resistances and inductances are independent of the bias conditions, the article only extracts them under Bias 4. The analysis results are provided in Table I. The initial extraction and optimization of DHBT device under Bias 4 conditions and the extraction results under other biases are shown in Table II.

### Table I. Initially extracted and optimized values of the model port parasitics at $V_{ce} = 1.5$ V, $I_{be} = 0.4$ $\mu$A. Error = $\mid$Extracted value − Optimized value$\mid$/Extracted value $\times$ 100%

| Parameters | Extracted value | Optimized value | Error (%) |
|------------|-----------------|-----------------|-----------|
| $R_b$ ($\Omega$) | 3.2 | 2.7 | 15.9 |
| $R_c$ ($\Omega$) | 3.8 | 3.4 | 10.5 |
| $R_{ce}$ ($\Omega$) | 7.26 | 6.0 | 17.4 |
| $L_b$ (pH) | 2.6 | 2.3 | 11.5 |
| $L_c$ (pH) | 4.37 | 4.66 | 6.2 |
| $L_e$ (pH) | 2.7 | 2.88 | 6.2 |

### Table II. The extraction results under four sets of bias conditions, and the optimal value and error are given when the bias condition is $V_{ce} = 1.5$ V, $I_{be} = 0.4$ $\mu$A

| Parameters | $V_{ce}$ = 1.5 V, $I_{be}$ = 0.4 $\mu$A | Bias 1 | Bias 2 | Bias 3 |
|------------|---------------------------------|--------|--------|--------|
| $R_{bi}$ ($\Omega$) | Extracted | Optimized | Error (%) | Extracted | Optimized | Error (%) | Extracted | Optimized | Error (%) |
| $R_{be}$ ($\Omega$) | 23.6 | 27.6 | 17.4 | 21.0 | 28.6 | 27.0 |
| $R_{bc}$ (K$\Omega$) | 11.0 | 9.2 | 12.7 | 13.7 | 18.0 | 4.4 |
| $R_{ce}$ (K$\Omega$) | 2.9 | 6.9 | 57.9 | 4.78 | 9.9 | 3.6 |
| $C_{bc}$ (fF) | 5.4 | 4.5 | 16.7 | 6.0 | 4.5 | 10.0 |
| $C_{ce}$ (fF) | 4.6 | 4.5 | 15.1 | 4.0 | 4.48 | 4.2 |
| $C_{cx}$ (fF) | 55.0 | 259 | 78.7 | 90.0 | 100.0 | 350.0 |
| $G_{m0}$ (mS) | 32.0 | 13.0 | 59.3 | 9.0 | 11.0 | 9.6 |
| $\tau$ (fs) | 608.0 | 590.0 | 3.0 | 280.0 | 240.0 | 600.0 |
| $\tau$ (fs) | 279.0 | 270.0 | 3.2 | 315.0 | 310.0 | 290.0 |
5 Model verification and discussion

The model, as shown in Fig. 1, was further simulated directly in the Keysight Advanced Design System (ADS) based on the extracted results. The results are compared in Fig. 6(a), (b), (c) and (d), including both the real and imaginary parts of \( S_{11} \), \( S_{12} \), \( S_{21} \) and \( S_{22} \). The gap between these S-parameters was enlarged by adding a delta variable \( \Delta \), as shown in Fig. 6, to observe the fitting result under each bias condition more clearly. The simulation results obtained by the extraction model and measured data are in good agreement.

The model parameter extraction results presented in this paper, except for the parasitic inductances, were extracted in the 66 GHz frequency band. When applied to the model simulation, the model shows excellent agreement between the measured and simulated S-parameters up to 325 GHz. In other words, the intrinsic model parameters of InP DHBT devices can be reliably extracted in the lower frequency band.

The simulation results both with and without consideration of the port parasitic inductance model are presented in Fig. 7. The simulation results of the two models exhibit notable differences when the frequency is higher than 80 GHz. In the frequency band below 80 GHz, the port parasitic inductance has only a slight effect on the S-parameters. This result illustrates that the parasitic inductance will become the main source of model error when the operating frequency of the transistor is...
high, which directly affects the fitting accuracy of the S parameter, particularly the imaginary part. The errors of the imaginary parts of $S_{11}$ and $S_{22}$ under different bias conditions in different frequency bands are provided in Table III. W.L. and N.L characterize the error analysis results with and without consideration of the parasitic inductance model, respectively. The influence of inductive parasitics becomes larger with increases in the working frequency.

![Fig. 7. Comparison of the model simulated (solid lines) and measured (symbols) S-parameters at different biases, including the imaginary of $S_{11}$ and $S_{22}$. Of these, the suffix L is the simulated result with inductive parasitic.](image)

| Table III. Under four sets of bias in different frequency bands, the error between measured data and model simulated result with inductive parasitic |
|---------------------------------|-----------------|-----------------|-----------------|
| Frequency Band                 | 0.2~66 GHz      | 0.2~220 GHz     | 0.2~325 GHz     |
| Bias                           | Imag ($S_{11}$) | Imag ($S_{22}$) | Imag ($S_{11}$) | Imag ($S_{22}$) |
| $V_{ce} = 1.0 \, V$, $I_{be} = 0.2 \, \mu A$ | 9.2          | 4.3              | 45.3            | 20.6            | 65.9            | 36.3            |
| $V_{ce} = 1.0 \, V$, $I_{be} = 0.4 \, \mu A$ | 21.7         | 16.6             | 30.3            | 19.9            | 33.7            | 17.6            |
| $V_{ce} = 1.5 \, V$, $I_{be} = 0.2 \, \mu A$ | 6.6          | 2.9              | 11.9            | 4.0             | 14.3            | 4.3             |
| $V_{ce} = 1.5 \, V$, $I_{be} = 0.4 \, \mu A$ | 8.5          | 3.9              | 40              | 34.3            | 77.9            | 44.4            |
| $V_{ce} = 1.0 \, V$, $I_{be} = 0.2 \, \mu A$ | 5.4          | 1.5              | 20.7            | 9.5             | 30.1            | 9.2             |
| $V_{ce} = 1.0 \, V$, $I_{be} = 0.4 \, \mu A$ | 30.2         | 23.7             | 74.9            | 40.2            | 80.3            | 48.4            |
| $V_{ce} = 1.5 \, V$, $I_{be} = 0.2 \, \mu A$ | 9.3          | 5.7              | 21.1            | 4.1             | 27.6            | 4.3             |
| $V_{ce} = 1.5 \, V$, $I_{be} = 0.4 \, \mu A$ | 4.4          | 2.1              | 14.3            | 4.5             | 22.4            | 4.7             |

6 Conclusion

This paper presented a direct parameter extraction method for determining the small signal equivalent circuit model for frequencies of up to 325 GHz. The intrinsic parameters of the model were extracted in the frequency band below 66 GHz, and the port lead parasitic inductance was extracted from the S-parameters measured in the frequency band above 80 GHz. The method was validated by using the data measured under different bias conditions to extract the model parameters. The simulation results of the model and the measured data are in good agreement.