Abstract—This paper presents a novel, on-wafer de-embedding technique for the accurate small-signal equivalent circuit modelling of resonant tunneling diodes (RTDs). The approach is applicable to stabilised RTDs, and so enables the modelling of the negative differential resistance (NDR) region of the device’s current-voltage (I-V) characteristics. Further, a novel quasi-analytical procedure to determine all the equivalent circuit elements from the de-embedded S-parameter data is developed. Extraction results for a $10 \times 10 \mu m^2$ stabilised, low-current density RTD at different bias points show excellent fits between modelled and measured S-parameters up to 110 GHz.

Index Terms—Bias oscillations, experimental modelling, parameter extraction, Resonant tunneling diode (RTD), small-signal equivalent circuit.

I. INTRODUCTION

Terahertz (THz) radiation, which has an electromagnetic spectrum that lies between millimetre-waves and infrared light, has become, over the last decade, a primary research interest in the fields of medical diagnostics, security imaging and wireless communications [1]. The resonant tunnelling diode (RTD) is a quantum-well structure, which exhibits negative differential resistance (NDR) that extends into the THz range, thus making it one of the target candidates for such applications. There is intense on-going research on this device technology to realise compact and coherent THz sources [2]-[5]. Fundamental frequency oscillations at around 2 THz have recently been obtained from RTD oscillators [6]. Compared to other electronic device technologies that may be used in the 0.1 – 1 THz band such as transistors, Gunn diodes, etc., the RTD exhibits the largest bandwidth (for a given device size), requires simple circuitry, is compact, and consumes low power [7].

Recently, high performance RTD based THz short range multi-gigabit wireless links [8]-[10] and imaging applications have been demonstrated [11]. Thus, the accurate modelling of RTDs to support reliable THz circuit design is of paramount importance.

The key device operation region of interest, which gives the RTD its performance advantages over competing technologies (for the THz band), is its negative differential resistance (NDR) region. Without stabilisation circuitry, the device bursts into oscillations when biased in the NDR, rendering the characterisation of this region difficult [12], [13]. The common approach to achieve bias stability is to use a suitable shunt resistance connected across the RTD, chosen such that the combined conductance is positive. Using this approach, the device I-V characteristic within the NDR can be determined indirectly [14], [15]. This approach has also been used once for the RF characterisation of the NDR region, but no details about the de-embedding of the stabilising resistor were provided [16]. The increased complexity of the resistor model at high frequency to account for parasitic effects such as self-inductance together with the RTD makes this approach difficult to implement at millimeter-waves.

Another approach to characterise the NDR region is to use physically small devices, usually sub-micron device dimensions [17], since for such devices the negative differential conductance is also small making them stable in a conventional measurement setup, e.g. when characterised by a vector network analyser in the typical 50-Ω system impedance. This approach is, however, only applicable to low peak current density RTDs, so less than about 100 kA/cm$^2$ [18]. For high current density designs (>300 kA/cm$^2$), even small submicron devices remain unstable when biased in the NDR region and so must employ a stabilising resistance [19]. Therefore, the characterisation of the NDR region of such devices is usually not possible. Presently, RF characterisation of an RTD in its positive differential resistance (PDR) region is used to estimate its equivalent circuit elements in the NDR region [20], [21]. Even for a key parameter such as the device self-capacitance, its extraction is done at only a single frequency, 10 GHz [20], and so there is limited scope to validate the accuracy of this approach.

In this paper, we report a new approach to characterise the NDR regions of (stabilised) RTDs without limitations to device
sizing or frequency. It uses a universal on-wafer bond-pad and shunt resistor de-embedding technique for reliable high frequency characterisation. Further, a quasi-analytical procedure to determine the RTD equivalent circuit elements is also developed. The new de-embedding and extraction procedure is applied to a 10 x 10 µm² AlAs–InGaAs–AlAs device stabilised with a 20 Ω shunt stabilisation resistor at different bias points.

The paper is organised as follows: Section II describes the procedure for de-embedding the shunt stabilisation resistor from S-parameter measurements, while section III provides details of the RTD epitaxial structure, its manufacturing including that of the shunt resistor and the measured RTD I-V characteristics. RF device and test-fixture characterisation and validation is described in Section IV, while the new parameter extraction of the RTD equivalent circuit elements is described in Section V. A comparison of the extracted element values with those determined using alternate methods to estimate the device parameters, in particular, those found on the basis of device physics, is given in Section VI. Conclusions are given in Section VII.

II. RF ANALYSIS OF A STABILISED RTD

Fig. 1 shows an RTD with a shunt resistor for bias stabilisation connected across it. The stabilising resistor including the interconnections can be considered as a standard 2-port network as shown in Fig. 1. If the RTD has a reflection coefficient, \( I_{\text{RTD}} \), and the stabilising network is described by its \( S \)-parameter matrix, then the input reflection coefficient \( I_{\text{in}} \) will be given by [22]:

\[
I_{\text{in}} = S_{11} + \frac{S_{12}S_{21}}{1-S_{22}} I_{\text{RTD}}.
\]  

(1)

![Fig. 1. RTD biased through a shunt resistor \( R_s \), modelled as a 2-port network.](image)

Re-arranging (1) gives the reflection coefficient of the RTD, \( I_{\text{RTD}} \):

\[
I_{\text{RTD}} = \frac{r_{\text{in}}-S_{11}}{r_{\text{in}}S_{22}-S_{11}S_{22}+S_{12}S_{21}}.
\]  

(2)

where \( S_{11}, S_{12}, S_{21}, \) and \( S_{22} \) are the S-parameters of the 2-port stabilising network. These S-parameters can be obtained by a two-port measurement of a fabricated auxiliary test structure, identical to the stabilised RTD but without the device under test (DUT).

III. RTD DEVICE AND TEST-FIXTURE FABRICATION

The RTD epitaxial wafer used in this work was grown by molecular beam epitaxy (MBE) by IQE Ltd on a semi-insulating InP substrate. The epitaxial layer structure consists of a 4.7 nm InGaAs quantum well (\( E_g = 0.75 \) eV) sandwiched between 2.5 nm thick AlAs barriers (\( E_g = 2.16 \) eV), forming a double barrier quantum well structure (DBQW). The structure is completed by spacer layers on either side of the DBQW, a drift layer on the collector side and contact layers on both sides as detailed in Table I. It was chosen in order to enhance the device I-V characteristics by maximising peak-to-valley voltage and current differences as proposed in [23].

![Table 1](image)

<table>
<thead>
<tr>
<th>Layer</th>
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RTD devices with 10 x 10 µm² top contact mesa were fabricated using optical lithography. Chemical wet etching (\(\text{H}_3\text{PO}_4 : \text{H}_2\text{O}_2 : \text{H}_2\text{O} = 1 : 1 : 38\)) was used to define the RTD mesa. This recipe has an etching rate of around 100 nm/min. Polyimide PI-2545 was used for device passivation. The Ohmic contacts comprised of Ti/Pd/Au (20/30/150 nm) metallisation. The device contact resistance was characterized through TLM (transmission line method) measurements and found to be 2.6 Ω. The 20-Ω stabilising resistor was realised from a thin film NiCr (60:40) which has a sheet resistance of 50 Ω/square. It was realised across the coplanar wave-guide (CPW with 50 Ω characteristic impedance) input line and was 60 µm wide and 220 µm long. On the same sample, auxiliary test structures were fabricated with identical metal pads (CPW with length \( L = 220 \) µm) and resistor geometry (Fig. 2 bottom). A micrograph of the fabricated stabilized RTD and the auxiliary test structure is shown in Fig. 2.

![Fig. 2. Micrograph of a fabricated RTD with a 20 Ω stabilizing resistor (top) and de-embedding auxiliary test structure of pads and resistor (bottom).](image)
The DC characterisation of the RTD was done using a Keysight B1500A device parameter analyser. The measured I-V characteristic of the stabilised device is presented in Fig. 3. The device exhibits a peak-valley bias voltage difference ($\Delta V$) of around 2 V and peak-valley current difference ($\Delta I$) of around 16 mA.

As expected, significant difference in acquired data can be observed across the frequencies for the stabilized device (blue curve), due to the presence of the shunt resistor. The proposed de-embedding method was applied in both cases using two-port measurements from their corresponding test structure (metal pads & shunt resistor – stabilised device; metal pads – unstabilised device). Fig. 5, shows the S-parameters of the pads and shunt resistor test structure used for de-embedding the stabilised device. The de-embedded data is shown in Fig. 6. (blue trace) alongside that of the un-stabilised device (red trace). Good agreement can be observed between the two measurements under the same bias conditions, which validates the applicability of this procedure. There are some minor discrepancies, which can be attributed to the fabrication process not yielding two perfectly identical devices. Overall, this result shows that the proposed de-embedding method is not limited by a specific device external circuitry, and so is suitable to accommodate different bond-pad and stabilizing network configurations.

IV. RF DEVICE & TEST-FIXTURE CHARACTERISATION

For S-parameter measurements, a Keysight E8361A vector network analyser (VNA) was used. The calibration was done using the short-open-load-through (SOLT) technique with a port power of -17 dBm. The frequency range was 10 MHz to 110 GHz.

To establish the applicability of the proposed de-embedding procedure, measurements of both a stabilized and an unstabilized device RTD device at identical bias points in the positive differential resistance (PDR) region were initially taken. The measured $S_{11}$ parameters of a stabilised and unstabilised RTD at a bias of at 2.9V (close to the peak region) are shown in Fig. 4.

As expected, significant difference in acquired data can be observed across the frequencies for the stabilized device (blue curve), due to the presence of the shunt resistor. The proposed de-embedding method was applied in both cases using two-port measurements from their corresponding test structure (metal pads & shunt resistor – stabilised device; metal pads – unstabilised device). Fig. 5, shows the S-parameters of the pads and shunt resistor test structure used for de-embedding the stabilised device. The de-embedded data is shown in Fig. 6. (blue trace) alongside that of the un-stabilised device (red trace). Good agreement can be observed between the two measurements under the same bias conditions, which validates the applicability of this procedure. There are some minor discrepancies, which can be attributed to the fabrication process not yielding two perfectly identical devices. Overall, this result shows that the proposed de-embedding method is not limited by a specific device external circuitry, and so is suitable to accommodate different bond-pad and stabilizing network configurations.
V. SMALL-SIGNAL EQUIVALENT CIRCUIT MODELLING

Fig. 7 shows the small-signal equivalent circuit of an RTD [24]. It consists of a contact and access resistance $R_s$, in series with the parallel combination of the device self-capacitance $C_n$ together with the device conductance $G_n$ which models the intrinsic current-voltage characteristic, and the quantum well inductance $L_{qw}$ which models the charging and discharging effect of the quantum well.

An inspection of the circuit in Fig. 7 reveals that at high enough frequencies, $C_n$ would provide a short-circuit path which would effectively mask any contributions to the input impedance from $G_n$ and $L_{qw}$. This is clearer to see with a quasi-static model for the RTD, i.e. one which neglects $L_{qw}$. In this case, the impedance of the device can be written as:

$$Z_{RTD} = R_s + \frac{1}{G_n+j\omega C_n} = R_s + \frac{G_n-j\omega C_n}{G_n^2+\omega^2 C_n^2} \quad (3)$$

From (3), we can infer that the contact resistance $R_s$ (which will typically be a few Ohms) dominates the real part of the device impedance at high frequencies, since $G_n$ is typically a few milli-Siemens, $C_n$ is tens of femtofarads or larger, and therefore the contribution of the intrinsic device to the device resistance can be negligible at millimeter-waves. Therefore, to determine the small-signal equivalent circuit elements of Fig. 7, the following procedure is proposed:

a) The de-embedded S-parameter data of the device ($Y_{RTD}$) is first converted into Z-parameters which provides $Z_{RTD}$. A plot of the real part of $Z_{RTD}$ with frequency at high frequency provides an estimate of $R_s$, i.e. $R_s \approx \text{Re}(Z_{RTD})_{f \rightarrow \infty}$. Fig. 8, provides such plots for one bias point in the PDR and another in the NDR, respectively. As expected, the real parts of the input impedance become frequency independent at high frequencies, with $R_s \approx 2.5 \ \Omega$ at $V_{bias} = 1 \ \text{V}$ and $R_s \approx 3.5 \ \Omega$ at $V_{bias} = 3.1 \ \text{V}$. These values of $R_s$ are initial estimates at the respective bias voltages. The final value of $R_s$ at each bias point is determined in conjunction with the simultaneous determination of the other intrinsic parameters, namely $C_n$, $R_s$, and $L_{qw}$. Here, the basis of the parameter extraction is that each of these lumped elements is independent of frequency.

b) We assume that $R_s$ is known and de-embed it from $Z_{RTD}$. The resulting data then should represent $C_n$ in parallel with $G_n$ and $L_{qw}$ and so can be concisely expressed using its admittance as follows:

$$Y_{RTD} = j\omega C_n + \frac{1}{R_n+j\omega L_{qw}} = \frac{R_n}{R_n^2+\omega^2 L_{qw}^2} + j\omega \left(C_n - \frac{L_{qw}}{R_n^2+\omega^2 L_{qw}^2}\right) \quad (4)$$

with $R_n = 1/ G_n$.

c) From (4), it should be clear that a correct value of $R_s$ would provide an imaginary part which varies linearly with frequency (at low frequencies). The value of $R_s$ may be adjusted at this stage to achieve this. We can then estimate $C_n$ using $C_n \approx \text{Im}(Y_{RTD})/\omega$ and de-embed it from the data. The resulting data then represents the series circuit of $G_n$ and $L_{qw}$. As such, we can convert this data into an impedance which can be expressed as:

$$Z_{RTD1} = R_n + j\omega L_{qw} \quad (5)$$

d) Using (5), a correct $R_s$ (and $C_n$) should provide frequency independent values of $R_s$ and $L_{qw}$. Thus, $R_s$ and can be adjusted further to achieve this. In specific cases, for which the magnitude $L_{qw}$ is large (i.e. in the NDR region), its effects can be observed more dominant at lower frequencies in the susceptance of the circuit ($\text{Im}(Y_{RTD})$). Using the estimation presented in (c) would provide an overcompensated value of $C_n$, which needs further adjustment in order to achieve linearity in the real and imaginary part of $Z_{RTD1}$.

In summary, starting with the estimate of $R_s$ as described in step (a) above, the extraction of the other intrinsic RTD elements from measured data proceeds as described above. The
initial estimate of $R_s$ may be varied within ±10% to achieve the expected frequency independence of $C_n$, $R_n$ and $L_{qw}$. Example extraction results from the PDR and NDR regions of the RTD are shown in Fig. 9. and Fig. 10. respectively.

Fig. 9. Imaginary part of Y-parameters in the PDR region with de-embedded $R_s$ – linear region shown in inset used to estimate $C_n = 93$ fF (a). Real part of Z-parameters in the PDR region with de-embedded $R_s$ and $C_n$ – linear region shown in inset used to estimate $R_n = 284$ Ω (b). Imaginary part of Z-parameters in the PDR region with de-embedded $R_s$ and $C_n$ – linear region shown in inset used to estimate $L_{qw} = 0.37$ nH (c).

Fig. 10. Imaginary part of Y-parameters in the NDR region with de-embedded $R_s$ – linear region shown in inset used to estimate $C_n = 110$ fF (a). Real part of Z-parameters in the NDR region with de-embedded $R_s$ and $C_n$ – linear region shown in inset used to estimate $R_n = 136$ Ω. Initial value of $C_n$ adjusted to 94 fF to achieve linearity (b). Imaginary part of Z-parameters in the NDR region with de-embedded $R_s$ and $C_n$ – linear region shown in inset used to estimate $L_{qw} = -0.48$ nH (c).

As may be seen from these graphs, the values of the intrinsic elements are largely independent of frequency at lower frequencies, up to around 40 GHz. Insets for the frequency bands used for the actual extraction are shown in each figure.
The somewhat random behaviour at higher frequencies may be attributed to reduced measurement accuracy at these frequencies.

The extracted parameters were used to compute simulated device input impedance and reflection coefficient. The results were compared with the measured data and are shown in Fig. 11, for the input impedance at two different bias points, one in the PDR and the other in the NDR. Fig. 12. shows the measured and simulated reflection coefficient for the NDR bias point on a Smith chart. Excellent agreement between measurement and simulated RTD data can be observed over the complete measurement frequency range, 10 MHz to 110 GHz, demonstrating the accuracy of the proposed procedure.

Table II shows the extracted small-signal equivalent circuit elements at various bias points, throughout the entire bias range. In order to validate the accuracy of the extracted parameters a standard optimization process was further used, in order to obtain an error range for each coefficient (tabulated in brackets), with 95% confidence bounds.

The results of the complete extraction procedure indicate a relatively small linear bias dependence of the contact and access resistance $R_s$ (1.3 Ω over a range of 4V) This variation was possible to observe due to the high frequency nature of the S-parameter measurements (between 80 – 110 GHz), as it is generally compensated at lower frequencies by errors in the extracted values of device conductance $G_n$. We think that this phenomenon could be related to the electric field dependence of the carrier mobility [25], and will be investigated further.

**VI. OTHER ESTIMATION TECHNIQUES FOR RTD ELEMENTS**

This section reviews previously used methods to estimate the RTD small-signal equivalent circuit elements and compares
these with the new approach described in the preceding section. In particular, physics-based approaches for estimating $C_n$ and $L_{qw}$ are described, but first the alternative techniques of determining $R_s$ and $G_n$ are described.

A. Device contact resistance and differential conductance

The metal-semiconductor contact resistance can be determined experimentally by using standard transmission line model (TLM) measurements. For the devices described in this paper, the estimated $R_s$ is $-2.6$ $\Omega$ for the $10 \mu m \times 10 \mu m$ RTDs. This contact resistance can be de-embedded from the measured I-V characteristics (Fig. 3) to yield the intrinsic device I-V characteristics. The differential conductance $G_n$ can be computed directly from these characteristics. Good agreement can be seen in Fig. 13, between the variation of $G_n$ with bias and extracted $G_n$ values from S-parameter data.

The basis of the charge variation has been derived by [28] and validated in [16], and can be described by the change in quantum-well – collector current density ($J_c$) as a function of electron escape $v_c$ ($s^{-1}$), and thus $C_{qw}$ is also expressed as:

$$C_{qw} = A \frac{\Delta Q_c}{\Delta v} = \frac{C_n}{v_c}$$

where $\Delta Q_c$ represents the variation of charge in the collector, $\Delta Q_c \approx -\Delta Q_{qw}$ assuming no contribution from the electrons tunneling back from the quantum-well into the emitter. Therefore using (6)-(8), i.e. $C_n$ and extracted $G_n$, $v_c$ can be determined. It can be seen that quantum well-collector escape rate (assumed bias independent) is $1/v_c = 0.55$ ps. From this value and the differential conductance in Fig. 13, the modelled capacitance variation with bias is plotted in Fig. 14, alongside the extracted device capacitance from the S-parameter data.

B. Device capacitance

The RTD self-capacitance ($C_0$) comprises the device geometric capacitance ($C_0$) and the quantum-well capacitance ($C_{qw}$), which arises from the electron density change in the quantum well as a function of applied bias, and so is given by:

$$C_n = C_0 + C_{qw}.$$  

(6)

Considering the DBQW structure as a standard parallel-plate capacitor (an undoped region confined between a highly doped collector and emitter), $C_0$ can be approximated by:

$$C_0 = \frac{\mathcal{A}}{\varepsilon_{qw} \varepsilon_0 \varepsilon_{qb} \varepsilon_{ed}}.$$  

(7)

where $L_{on}$, $L_b$, $L_d$ are the widths of the quantum well, barrier and depletion region (4.7 nm 2.5 nm, 120 nm) and $\varepsilon_{qw}$, $\varepsilon_b$, $\varepsilon_{ed}$ are the corresponding material dielectric constants (13.1, 13.1 [26], 10.1 [27]). This equates to 88 fF for the presented epi-layer structure. The extracted device capacitance from S-parameters, which corresponds to $C_n$, indicates that the static capacitance $C_0$ is slightly higher (~90 fF). Thus, $C_{qw}$ can be determined from the total capacitance variation using (6).

The RTD device capacitance $C_n$ computed from the intrinsic I-V characteristics (black trace) and extracted (red dots).

Fig. 13. Differential device conductance $G_n$ computed from the intrinsic I-V characteristics (black trace) and extracted (red dots).

Fig. 14. RTD Device capacitance $C_n$ simulated (black trace) and extracted (red dots).

C. Quantum well inductance

As earlier described, the quantum-well inductance ($L_{qw}$) is attributed to the charging and discharging effect of the quantum-well, and is given by [24]:

$$L_{qw} = \frac{\tau_{dwell}}{\varepsilon_n}.$$  

(9)

where $\tau_{dwell}$ is the electron quasi-bound-state lifetime in the quantum-well [29] and can be estimated from:

$$\tau_{dwell} = \frac{\hbar}{\Delta E_n}.$$  

(10)

Here, $\Delta E_n$ is the energy full-width of the transmission probability function through the resonant state which can be obtained by the Wentzel–Kramers–Brillouin (WKB) approximation method as follows:

$$\Delta E_n = E_n \exp \left[ -2L_b \sqrt{\frac{2m_b(U_0-E_n)}{\hbar^2}} \right].$$  

(11)

$E_n$ is the $n$th resonance level, $L_b$ is the width of the barrier (2.5 nm), $U_0$ is the barrier energy level (1.322 eV) and $m_b$ is the
effective electron mass in the barrier. Fig. 15 shows the transmission probability through the RTDs DBQW structure using WinGreen software, which is a nonequilibrium Green function based 1D quantum transport simulator [30]. The estimation for the carrier lifetime is generally calculated for the first resonance energy level \((E_0 = 0.17 \text{ eV for this structure})\), however, the transmission probability plot suggests that the primary resonant current occurs at a higher energy level \((E_1 = 0.73 \text{ eV})\).

Using (10) and (11), the computed value for \(\tau_{\text{dwell}}\) was calculated to be 1.86 ps, and then using (9) and the intrinsic \(G_n\) (Fig. 7), the variation of \(L_{qw}\) with bias was calculated and is shown in Fig. 16. (black solid trace). As expected, the computed inductance becomes negative in the NDR region following the nature of the differential conductance.

From Fig. 16, it can be seen that the largest discrepancy between the simulated and experimentally extracted electron life time can be observed around the centre of the NDR region. This effect can be simply explained by the variations of the electron escape rates through the barriers, which were assumed bias independent throughout simulations.

VII. CONCLUSION

A universal on-wafer bond-pad and shunt resistor de-embedding technique for stabilised RTDs was proposed and demonstrated up to 110 GHz for a 10 x 10 \(\mu\text{m}^2\) InP RTD device. The accuracy of the method relies principally on measured data from one test structure. Further, a new simple and robust small-signal equivalent circuit parameter extraction procedure for RTDs which yielded physically relevant parameters and provided an excellent fit between the model and measured S-parameter data up to 110 GHz was described. It is expected that these results will accelerate the development of RTD technology for THz applications by providing the foundation to develop compact CAD models for the device.

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