S-Parameter Measurements of the Impedance of mm-Wave IMPATT Diodes in Dependency on the Current Density

Claus J. Schoellhorn, M. Morschbach, H. Xu, W. Zhao and E. Kasper, Associate Member, IEEE

Abstract

S-parameter measurements were performed to characterize IMPATT diodes integrated in coplanar waveguides, up to a maximum frequency of 40GHz. With a de-embedding procedure the real- and imaginary part of the impedance of the inner diodes were calculated. Above the avalanche frequency the diodes showed the expected negative real- and imaginary parts of the impedance. Due to theory, with increasing current density the avalanche frequency shifted to higher values. To manufacture the diodes and oscillators a complete monolithically integrated process was used. This process avoids the use of heatsinks and the difficult and time consuming bonding process.

I. INTRODUCTION

S-parameter measurements are a well known method to determine the small signal behavior of circuits for RF and microwave applications [1],[2]. The determination of the impedance of IMPATT diodes by means of S-parameter measurements is performed only very scarcely because IMPATT diodes tend to oscillate easily in appropriate resonator structures which may also be provided by the measurement setup.

To allow on-wafer measurements we integrated IMPATT diodes in coplanar waveguides (CPWs). These structures are processed in monolithic technique with a characteristic impedance of 50Ω. On the same wafer, open- and short-circuits were fabricated to execute the de-embedding procedure. The S-parameter measurements were performed up to frequencies of 40GHz.

II. INTEGRATION TECHNOLOGY

Because IMPATT diodes are operated in the breakdown they are often near their thermal limit. With a smaller resistance of the heat flow the thermal exposure of the devices can be reduced.
In most of the earlier work the diodes were mounted on special diamond or copper heat sinks [3], [4]. Because this is a time-consuming and costly process, requiring a lot of manual work, it is not possible to process IMPATT oscillators cost-effective and in very large numbers. In contrast to this we used a complete monolithical process to manufacture the IMPATT diodes.

The active layers, in this case a $p^+n^-n^+$-stack for a single-drift (SD) IMPATT diode with a $n$-doped avalanche/drift region, are grown with molecular beam epitaxy (MBE). This method facilitates the growth of high-quality layers in corporation with the abrupt change of the doping level, more than four orders of magnitude within a few nanometers, that is needed at $p^+n^-$-junction of the device. The process consists of two etch steps, one to define the mesas of the active devices, the second to remove the high doped buried layer beneath the passive parts (CPWs and resonators). To isolate and passivate the devices a low temperature PECVD silicon oxide is used. Finally the aluminum metallization layer is applied. Fig. 1 shows a 3D-scheme of an integrated IMPATT diode with the double-mesa and the metallization.

![Fig. 1. Scheme of a monolithically integrated IMPATT diode with the double mesa under the passivating silicon oxide and the aluminum coplanar waveguides.](image)

### III. Theory

The IMPATT diode was first proposed in 1958 by Read [5]. IMPATTs are operated in the reverse direction at breakdown. The occurrence of a negative differential resistance is caused by two main effects:

The first one is the avalanche breakdown caused by the very high electric field. This impact ionization leads to a multiplication process of electron-hole pairs. The second effect is the drift of the charge carriers to the $n^+$-contact. While the holes reach the $p^+$-contact immediately, the electrons drift through the drift region towards the $n^+$-contact with the saturation velocity $v_s$.

The avalanche process causes an inductive phase shift of 90°. Taking the drift process into account, an additional phase shift can be achieved. The amount of this phase shift can be determined by the length of the drift region. Considering both effects a phase shift between voltage and current at the $n^+$- and $p^+$-contact of 180° can be achieved. This leads to a negative differential resistance [6].

A small signal analysis, in a first step performed for a Read diode [7], later proved to be valid for general IMPATT structures [8], leads to the following equation for the avalanche frequency $f_a$:

\[ f_a = \frac{v_s}{2L} \]
where \( J_0 \) is the current density, \( \alpha' \) the differential of the ionization coefficient with respect to the electric field, \( v_s \) the saturation velocity and \( \varepsilon \) the semiconductor permittivity. At the avalanche frequency the imaginary part of the impedance changes its sign from inductive to capacitive. The zero crossing of the imaginary part is used to determine the exact avalanche frequency. For IMPATT diodes \( f_a \) is a very significant property as oscillator operations are possible above the avalanche frequency. As optimum operating frequency Culshaw suggested a frequency of \( f_{Osc} \sim 1.5 f_a \) [9]. According to equation (1) the avalanche frequency is proportional to the square root of the DC current density. In a fixed oscillator this effect is used to push the avalanche frequency to an optimized value.

\[
f_a = \frac{1}{2\pi} \sqrt{\frac{2\alpha' v_s \cdot J_0}{\varepsilon}} \quad (1)
\]

\[
f_a \sim \sqrt{J_0} \quad (2)
\]

IV. DE-EMBEDDING

Because the inner diode cannot be measured directly a de-embedding procedure has to be executed. With open- and short structures which are designed as close as possible to the structure of the measured diode, the impedance of the parallel and series parasitics are estimated and subtracted from the quantified impedance of the diode. The impedance of the measured diode with its coplanar surrounding is indicated as \( Z_{DUT} \) (DUT : device under test) in contrast to the inner diode, in our case called \( Z_{IMP} \). The procedure used in this paper was first published by Koolen et al. in 1991 [10]. In contrast to earlier published de-embedding methods it takes not only the parallel parasitics into account but also the series parasitics. Short and open structures are realized with the same dimensions as the active diode. In case of the open structure the contact holes in the passivation layer are not opened, while in the short structure the signal line is elongated to the ground metallization. Fig. 2 shows the equivalent circuit of an IMPATT diode integrated into a coplanar waveguide. The CPW is assumed to consist of a series impedance \( Z_s \) and a parallel admittance \( Y_p \). In this case it is not necessary to look at the parasitics in more detail, as they are described in [11] for example, because they are handled only as complex impedance and admittance respectively.
In a first step the open structure is used to determine the parallel admittance $Y_P$ of the contact structure: $Y_{\text{Open}} = Y_P$. In the second step the admittance of the short circuit $Y_{\text{Short}}$ is measured. In this case the admittance $Y_P$ and the reciprocal of the series impedance $Z_S$, $Y_S = Z_S^{-1}$ represent a parallel circuit. With equation (3) the series admittance can be easily calculated:

$$Y_S = Y_{\text{Short}} - Y_P \quad (3)$$

With knowledge of the admittance of the measured device it is possible to calculate the impedance of the inner diode by using equation (4):

$$Z_{\text{IMP}} = (Y_{\text{DUT}} - Y_{\text{Open}})^{-1} - (Y_{\text{Short}} - Y_{\text{Open}})^{-1} \quad (4)$$

All results in the following chapters are corrected with the de-embedding procedure.

V. RF MEASUREMENTS

A. Measurement Setup

To carry out measurements of the scattering parameters the diodes are integrated in coplanar waveguides. Fig. 3 shows the scheme of the S-parameter measurement. The diode is parallel to signal and ground of the CPW. Because only port 1 (1-2) is accessible ((3-4) marks the inner diode) solely the reflection coefficient $S_{11}$ of the whole structure can be measured. The impedance of the IMPATT diode $Z_{\text{IMP}}$ has to be calculated by using the de-embedding procedure presented in chapter IV.
Fig. 3. S-parameter measurement of an RF diode integrated in a coplanar waveguide.

The appropriate photo of the on-wafer measurement setup with the GSG-(ground-signal-ground) prober head can be seen in fig. 4. In contrast to the very small IMPATT diode, marked with the circle, the coplanar structure is quite large to fit the pitch of the prober heads. Fig. 5 shows the active device in more detail. The active area has a length of $l=30\mu m$ and a width of $w=10\mu m$. The mesa of the buried layer, which is used instead of a backside contact, is much larger to provide a good contact and small series resistances. The coplanar waveguides are lying on the passivating oxide which covers the high resistivity silicon substrate (Float Zone, FZ, $\rho>1000\Omega\cdot cm$). The buried layer is etched below the signal line (compare fig. 5 with fig. 1) because highly doped silicon layers beneath the CPWs would increase the attenuation of the coplanar waveguides and thus the losses of the oscillator.

Fig. 4. On-wafer measurement of an integrated IMPATT diode. The coplanar waveguide is adapted to the pitch of the GSG-prober head.
Fig. 5: IMPATT diode with an active area of $l \times w = 30 \times 10 \mu m^2$ and the metallization of the signal and ground contact.

**B. Smith Chart**

The Smith chart shows the reflection coefficient as a function of the frequency. The reflection coefficient of an IMPATT diode at its operating frequency is outside of the Smith Chart because of its ability to emit RF power. This happens for the reflection coefficient $r > 1$, as to say the reflected power gets larger than the input power. Fig. 6 shows the Smith Chart of an IMPATT diode for different current densities and a maximum frequency of 40GHz. With increasing current density the curve runs further outside indicating the negative differential resistance above the avalanche frequency. For very high frequencies the curve returns into the Smith Chart. This is caused by the larger series resistance of the diode. The negative resistance of the IMPATT decreases above the avalanche frequency and the total resistance is the sum of both.
Fig. 6. Measured reflection coefficient of a 30µm×10µm IMPATT diode for different current densities. The maximum frequency is 40GHz and the maximum current density is 0.133mA/µm².

C. Impedance of the IMPATT Diodes

To get the diode impedance from the S-parameter measurement, it is necessary to insert the reflection coefficient into equation (5)

\[ Z = \frac{1 + r}{1 - r} \cdot Z_0 \]  

(5)

with \( Z \) the impedance of the device, \( r \) the measured reflection coefficient and \( Z_0 \) the characteristic impedance (50Ω). The curves for the real- and the imaginary part of the diode impedance are shown in fig. 7. The upper picture shows the real part and the lower one the imaginary part. The current is increased from 15mA to 40mA. The size of the diode is 300µm², so the current density increases from 0.05mA/µm² to 0.133mA/µm² in steps of 0.0167mA/µm².
Fig. 7. IMPATT impedance: real- and imaginary part after the transformation of the $S_{11}$-parameter. The increase of the avalanche frequency is roughly proportional to the square root of the current density.

As expected the avalanche frequency increases with increasing current density. After equation (2) the ratio of $f_a/\sqrt{J_0}$ has to be constant. Table 1 compares the quotients for the different current densities.
As it can be seen in table 1 the ratio decreases slightly instead of being constant. This is caused by the increase of the temperature of the diode with increasing current density. According to equation (1) the avalanche frequency depends on the square root of the differential of the ionization coefficient $\alpha'$. But for increasing temperature the ionization coefficient $\alpha$ decreases, for example by the factor of $\sim 2$ for a rise of the temperature from 20°C to 200°C. Additionally the saturation drift velocity decreases too. For an increase of the temperature from 20° to 200° it decreases from $1\times 10^7$ cm/s to $0.85\times 10^6$ cm/s [12].

### D. Negative Resistance of the integrated IMPATT Diodes

IMPATT diodes are able to oscillate if they are embedded in a resonator. The impedance of the resonator has to be adapted to the impedance of the diodes. A mandatory condition for an operating oscillator is equation (6a,b):

\[
\text{Im}_{\text{Osc}} = -\text{Im}_{\text{IMPATT}} \quad \text{and} \quad \text{Re}_{\text{Osc}} = -\text{Re}_{\text{IMPATT}} \quad (6a)
\]

or:

\[
Z_{\text{Res}} = -Z_{\text{IMPATT}} \quad (6b)
\]

An important parameter is the negative real part of the IMPATT which has to balance the resistance of the resonator. Because of this a high negative real part of the ideal IMPATT is necessary to overcome the series resistance of the diode. With our MBE made diodes very high values of the negative resistance Re (Z) are obtained. Fig. 8 shows the real part of the diode impedance near the avalanche frequency for a current of 35mA. Especially the region above the avalanche frequency is important, because in this range an oscillator is operated. In this case we obtain negative resistances of $-957 \Omega$ at a frequency of 30GHz, $-106 \Omega$ at 32GHz and $-27 \Omega$ at 34GHz. At 36GHz the negative real part of the impedance is still $-12 \Omega$.

The curve of the real part has a maximum negative resistance at the avalanche frequency. For frequencies above the avalanche frequency the negative resistance decreases. Basic theory requires a negative Re (Z) for all frequencies above the avalanche frequency, but the series resistance limits the usable frequency range for an oscillator application. A high series resistance shrinks the operation regime of the diodes. In our case DC measurements resulted in a series resistance of $2.8 \Omega$. Fig. 9 shows the ratio of the negative resistance to the maximum
negative value in dependency of the scaled frequency $f/f_a$. The exact knowledge of the impedance is necessary to design and optimize the resonator.

Fig. 8. Impedance of an IMPATT diode showing the high values of the negative resistance ($A=300\mu m^2$, $I=35 mA$, $f_a=29.5 GHz$).

Fig. 9. Ratio of the negative real part to the maximal negative resistance in dependency of the frequency (same diode parameters as in fig.8)
E. Ratio $\text{Im}(Z)/\text{Re}(Z)$

To simplify the resonator design a well defined ratio between imaginary part and real part of the impedance is favorable. Fig. 10 shows the ratio for a frequency above the avalanche frequency, because the IMPATT oscillators are operated at frequencies $f > f_a$. For an improved depiction the curve was smoothed.

![Graph showing the ratio of imaginary and real part of the impedance for frequencies $f > f_a$.](image)

Fig. 10. Ratio of imaginary and real part of the impedance for frequencies $f > f_a$ (same parameters as in fig. 8).

VI. DESIGN AND SIMULATION OF THE RESONATOR

If the IMPATT diode is embedded in a adapted resonator, oscillations above the avalanche frequency are possible. To achieve this, the conditions described in equation (6a) and (6b) have to be fulfilled around the oscillating frequency. Fig. 11 shows the layout of a planar resonator which is designed for frequency measurements with GSG prober heads of the S-parameter measurement setup.
The impedance of the resonator is determined by an open which is transformed by the two coplanar waveguides CPW1 and CPW2. Both waveguides have a signal line width of 120µm, the gap between signal and ground is 72µm. The lengths are 700µm for CPW1 and 480µm for CPW2. The GSG prober head is placed between the two waveguides. The prober head is used to apply the reverse breakdown voltage and to measure the oscillation frequency. The 50Ω impedance of the measurement setup has to be considered in the simulation of the resonator. The short circuit on the right side is transformed by the third coplanar waveguide CPW3. It has a length of 275µm, the signal line width is 60µm and the gap is 36µm. The impedance of CPW3 is in series with the IMPATT diode. Line AA’ marks the position in the signal line for which the impedances are simulated. The resulting impedance has to be adapted to the resonator impedance. To apply the DC ground to the diode on the right side a short circuit has to be used.

For designing the resonator the impedance of the IMPATT diode at 38.9GHz was selected (Fig.12). The ratio between oscillating frequency and avalanche frequency was chosen to be around 1.3 due to the fact that the decrease of the real part above the avalanche frequency is very steep. At the oscillating frequency the diode has a negative real part of -1.4Ω and a imaginary part of -96.3Ω. Smaller current densities will lead to smaller negative values of the real part so that the compensation of the series resistance of the diode and the resistance of the resonator might not be possible. Higher current densities will increase the thermal load of the diode. Fig. 12 shows real and imaginary part of the impedance around the oscillating frequency. The characteristic impedance $Z_0$ is 50Ω.
The simulations of the resonator were performed with a commercial design tool [13]. As substrate we supposed a silicon wafer with a thickness of 525µm. Fig. 13 shows the simulated resonator consisting of CPW1, CPW2 and the GSG.prober head.

Fig. 12: Real- and imaginary part of the impedance around the oscillation frequency (A=300µm²; I=35mA).

Fig. 13: Simulation of the resonator for a frequency range from 45MHz to 40GHz.
Fig. 14 shows the simulation of the serial connection of the measured $S_{11}$ parameter of the diode and the coplanar waveguide CPW3. Simultaneously the mirrored curve is depicted too. The mirroring was performed by building the reciprocal value of the reflection coefficient

$$r' = -\frac{1}{r}$$  \hspace{1cm} (7)$$

The maximum negative resistance is reached when the curve intersects the real axis. In this case the imaginary part is zero. Simultaneously this marks the avalanche frequency of the diode. The negative resistance at the avalanche frequency is $\sim 1125 \Omega$.

Fig. 14 : Simulated and mirrored curve of the $S_{11}$ parameter of the IMPATT diode and coplanar waveguide CPW3. The frequency range is from 45MHz to 40GHz.

Fig. 15 shows the condition for the oscillator. The intersection of the mirrored curve and the simulated curve of the resonator marks the oscillation frequency of 38.90GHz. At the oscillating frequency the impedance of the resonator is $Z_{\text{Res}} = (1.2 + j 61.0) \Omega$. The impedance of the mirrored diode with the coplanar waveguide is $Z_{\text{Diode}} = (1.15 + j 61.3) \Omega$. Within the precision of the S-parameter measurement the absolute values of the impedances can be denoted as equal. The feasibility of the coplanar oscillator concept was proven even for higher mm-wave frequencies [14].
VII. Conclusion

The achieved results demonstrated the possibility to measure the scattering parameters of monolithically integrated IMPATT diodes in the millimeter wave region. The measurements were performed on-wafer with a network analyzer. Undesired oscillations which disturb the measurement are suppressed by an accurate design of the 50Ω coplanar test structures. From the measured S11-parameter the impedance Z of the devices can be easily calculated. Real- and imaginary part of the impedance characterize very accurate and explicit the behavior of the IMPATT diode at the avalanche frequency and above. This is necessary to design and optimize the oscillators. The results of the analysis pave the way to IMPATT oscillators using a completely monolithically integrated process. The time consuming manual bonding processes can be avoided.

ACKNOWLEDGMENT

The authors would like to thank M. Oehme and Dr. C. Parry for growth of the MBE layers and J. Hasch and H. Irion, Robert Bosch GmbH, Central Research and Development, FV/FLO for helpful discussions and the basic design of the resonator.

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