Tao Hong

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List of symbols

α_{eff}	effective ionization coefficient	$1/cm$
$\alpha_{n,p}$	ionization coefficient of electrons, holes	$1/cm$
ϵ	relative permittivity	—
ϵ_0	vacuum permittivity	As/Vm
η	efficiency coefficient of the RF-power conversion	—
η_1	injection dependent efficiency coefficient	—
η_2	transit time dependent efficiency coefficient	—
λ	proportionality between N_{h_nett} and I_{CE}	—
$\bar{v}_{dn,dp}$	effective velocity of electrons, holes	cm/s
Θ_D	transit angle	—
Θ_M	phase angle of the injection	—
Θ_W	injection angle	—
$v_{dn,dp}$	drift velocity of electrons, holes	cm/s
v_Q	velocity of charge carrier	cm/s

A	chip area	cm^2
C_{DC}	DC-link capacitor	F
C_{diff}	effective capacitance in RF-range	F
C_{Diode}	differential capacitance of SCR in diode	F
C_{IGBT}	differential capacitance of SCR in IGBT	F
E_C	critical E-field strength	V/cm
E_{max}	maximal E-field in the space charge region	V/cm
I	current	A
I_L	load current, trun-off current of IGBT	A
I_N	norm current	A
I_{wb}	load current in the wire bonds	A
J_{DC}	induced average current density	A/cm^2
J_{ind}	induced current density on the terminal	A/cm^2
l	coil length	mm
L_a	inductance of the avalanche region	H
$L_{B1,B2}$	inductance of wire bonding	H
$L_{lc,sc}$	inductance of long, short coil	H
L_P	load inductance	H
L_{wb}	inductance of wire bondings	H
N_A	density of the acceptors	$1/cm^3$
N_D	density of donors	$1/cm^3$

N_{e_Plasma}	electron-density in plasma	$1/cm^3$
N_e	electron-density	$1/cm^3$
N_{h_nett}	net hole-density	$1/cm^3$
N_{h_Plasma}	hole-density in plasma	$1/cm^3$
N_h	hole-density	$1/cm^3$
P_{DC}	power consumption of DC voltage source	W
P_{RF}	radio frequency power generation	W
Q	injection charge	Q
r_c	radius of coils	mm
$R_{B1,B2}$	resistance of wire bonding	Ω
R_g	gate resistance	Ω
$RC_{1,2,3}$	resonant circuits	—
t	time	s
T_D	transit time	s
T_i	injection time	s
t_{on}	on time	s
V_{DC}	DC-link voltage	V
v_{e_Plasma}	electron-velocity in Plasma	cm/s
V_{GE}	gate-emitter voltage	V
v_{h_Plasma}	hole-velocity in Plasma	cm/s
V_{RF}	amplitude of the AC voltage	V

V_T	terminal voltage on the transit time structure	V
W_A	width of the avalanche region	cm
W_c	field energy of coils	W
W_{SCR}	width of the space charge area	cm
W_{wb}	field energy of emitter wire bondings	W

Zusammenfassung

Radio Frequenz Oszillationen bis zu mehreren hundert MHz wurden während des Abschaltens von Hochstrom leitenden IGBTs beobachtet. Sie werden in dieser Arbeit als Transient Avalanche Oscillationen beschrieben. Anders als PETT Oszillationen, die in der Tail-Strom Phase vorkommen, erscheint diese Oszillation während des Anstiegs der Kollektor-Emitter-Spannung V_{CE} und des Fallens des Kollektor-Emitter Stroms I_{CE} . In dieser Arbeit werden die Abschaltvorgänge von IGBTs mit Hilfe von Messungen, Schaltungs- bzw. Device-Simulationen untersucht, um die Wirkungsmechanismen der transient erscheinenden hochfrequenten Oszillationen zu erklären. Es ist erstmals gelungen solche Transient Avalanche Oszillationen mit der Simulation zu reproduzieren und anschließend die Einflüsse einzelner Parameter zu untersuchen. Die Beteiligung des IMPATT Mechanismus an der Transient Avalanche Oszillation wurde nachgewiesen. Die Interaktion von IMPATT und PETT während Aufbau der Raumladungszone im IGBT wurde untersucht. Maßnahmen zur Unterdrückung der Oszillation werden auf den Ebenen von Chip-Design, Modulauslegung und der Treiberschaltung diskutiert. Das Zuführen des Elektrostromes während der kritischen Phase wird als eine effective Methode zur Vermeidung der TA-Oscillation nachgewiesen.

IGBT, Abschalten, Oszillation, Transient Avalanche, Resonanz, Raumladung, IMPATT, PETT, TA-Oszillation

Abstract

Radio frequency oscillations up to several hundreds of MHz were observed during turn-off the high-current conducting IGBTs. They are described in this work as Transient Avalanche oscillations. Other than PETT oscillations that happen in the tail current phase, this oscillation appears during the rise of collector-emitter voltage V_{CE} and during the fall of the collector-emitter current I_{CE} . The turn-off process of IGBTs are investigated with the help of measurements, the circuit- and device-simulations to reveal the working mechanisms of transient high-frequency oscillations. For the first time the Transient Avalanche Oscillation is successfully reproduced with computer simulation, which allows the further investigation on the influences of individual parameters. The participation of IMPATT mechanism in Transient Avalanche oscillation is demonstrated. The interaction between IMPATT mechanism und PETT mechanism during the setting up of depletion region in IGBTs was investigated. Measures to suppress such oscillation are discussed on the levels of chip design, module design and the driver concept. Supplying electron current during critical phase is found to be an effective method to avoid the dynamic avalanche and the TA-oscillation.

IGBT, turn-off, Oscillation, Transient Avalanche, Resonance, Space charge, depletion region, IMPATT, PETT, TA-oscillation

Chapter 1

Introduction

In addition to the reliability, higher blocking capability, lower conduction losses, lower switching losses and higher switching frequency are the basic requirements for designing power semiconductors. The approach to improve such properties by combining the advantages of MOSFET (metal-oxide-semiconductor field-effect transistor) and BJT (bipolar transistor) was reported in the late seventies [Bal79]. In 1980 the U.S. Patent Office granted a U.S. patent “Power MOSFET with an anode region” with patent No. 4,364,073 to Hans W. Becke and C. Frank Jr. Wheathley [BW80]. This patent is regarded as the invention of the insulated gate bipolar transistor, briefly called IGBT, which merged BJT and MOSFET technically.

After the development over three decades, more and more approaches have been developed by advantageous combination of bipolar and unipolar elements. Up to date, IGBT finds wider and wider application fields. Parallel to the industrial motor control and inverters of renewable energy, IGBTs are also widely used in traction applications, auto mobiles and consumer electronics. The application area of IGBTs extends even to the typical application area of thyristors with the development of high power

IGBTs. The first 3.3kV IGBT Module was developed by Eupec GmbH in 1995. The first 6.5kV IGBT module was developed also by Eupec GmbH in 1999 [ABG⁺99]. The development of 8kV IGBT and diode chips was reported in 2002 [RKK⁺02]. Up to date the 6.5kV IGBT modules are still the commercially available Modules with maximal blocking capability.

One important aspect of the IGBT development focuses on the reduction of conduction and switching losses and to increase the current density on silicon. To fulfil increasing demands on the electrical and thermal performance in high power applications, IGBTs and diodes are paralleled and integrated in modules. In the modules, IGBT chips work in a complicated parasitic environment which consists of IGBT capacitance, diode capacitance, parasitic inductances and capacitances as a consequence of the module construction. Such parasitic components build up resonant circuits. In the dynamic process, such as turn-on and turn-off, the resonant circuits can be excited which cause undesirable oscillations. Especially the high-frequency oscillations can cause electromagnetic interference for the complete system.

Under a special design of a module, high frequency oscillations of up to several hundreds of megahertz are observed during turn-off of the IGBT. This oscillation phenomena has been found in the tail current phase of NPT IGBTs in flat package [TYK⁺97] and was explained by the the mechanism of plasma extraction transit time (PETT) oscillation [GMS01] [GMS02] [SLNM03]. The PETT oscillations were usually observed during the tail current. After the IGBT has taken the reverse voltage, there is still electron-hole plasma remaining close to the collector side. This residual plasma is extracted out of IGBT by the tail current. The holes drift through space charge region (*SCR*) under the influence of the strong electric field to the emitter side and recombine with electrons. The holes need a certain time to be injected in to the space charge region and a transit time to drift through the space charge zone. Such injection time and the

transit time together induce a phase shift between terminal voltage and the terminal current in the RF-frequency range. For a certain frequency range the phase shift lies in the interval between $1\pi/2$ and $3\pi/2$. This generates therefore power and works in this frequency range as an amplifier. This BARITT-diode similar mechanisms (barrier injection transit-time diode) [Roe77] [Sze81] was analysed in details in [Gut06] and [SMNL06]. RF-frequency oscillations has also been found during the turn-off of fast recovery diode. The temporarily charged K-defects, radiation induced deep levels, leads to the temporary reduction of the avalanche breakdown voltage and lead to the dynamic IMPATT (Impact Ionization Avalanche Transit Time) oscillation during diode turn-off [LSG98] [Lut00] [SLH04]. Both IMPATT and PETT oscillations lead to deterioration of EMC and should be avoided. consequently [SLNM03].

Recently a new type of high-frequency oscillation was found during the turn-off of IGBTs. It happens in the dynamic avalanche [SN89] [DLS03] phase during the rise of the collector-emitter voltage and the fall of the collector-emitter current. At the moment of oscillation, the load current or a large portion of the load current is still flowing. Since this oscillation is observed only transiently and only during the dynamic avalanche process, it is called Transient Avalanche oscillation (TA-oscillation) [HPB⁺14]. The mechanisms of this oscillation seems to be closely related to the process found in IMPATT diode [EH98]. The target of this work is to reveal its mechanism and to figure out the countermeasures to prevent or suppress such TA-oscillations.

In Chapter 2, the construction and characterizations of IGBTs are shown. The IGBT turn-off process is described according to the internal semiconductor states. Finally, the state of art mechanisms of PETT oscillation and the dynamic IMPATT oscillation is represented.

In Chapter 3, the theoretical analysis for the transit-time effect and mechanisms of IMPATT- and BARITT-diode are summarized. The anal-

ysis of the efficiency of the transit-time effect in the space charge zone of the IGBTs results in the basic of theoretical explanation of the TA oscillation. Whereas efficiency means the efficiency to amplify RF-oscillations by the TA-effect.

The detailed investigation of the TA-oscillation by means of measurements, resonant circuit analysis and simulations are presented and followed by the theory of the TA-oscillation in Chapter 4.

In chapter 5, the countermeasures of TA-oscillation are investigated, and a new driver concept is introduced to suppress TA-oscillation.

The summary in chapter 6 provides an overview of the HF-frequency oscillations in power modules, that is followed by the outlook on the further possibilities to suppress such oscillation by methods of module design and chip designs.

Chapter 2

Functionality, turn-off and phenomenons of RF-oscillations in IGBT modules

To analyse the TA-oscillation mentioned in Chapter 1, it is necessary to understand the functionality of IGBT and to describe the turn-off process with the internal semiconductor states. A hypothesis on the mechanisms of TA-oscillation can be set up based on the results of the investigation on the relationships between IGBT-, diode-chips and parasitic components of the IGBT modules constructions.

Figure 2.1: Cell structure of a modern trench IGBT with field stop on the collector side [Hon14].

2.1 Functionality of modern IGBTs

An IGBT works essentially like a MOS-channel controlled bipolar Transistor. The cell structure of a modern trench IGBT with field stop on the back side is illustrated schematically on the left side of Fig. 2.1. The cathode (IGBT emitter) is on the top of the chip, and the anode (IGBT collector) is on the bottom. The electrical contacts to the outside circuit are usually realized by the metallizations on the silicon body. A highly doped poly-silicon layer serves as a gate electrode. The MOS-channel is controlled by the voltage at the gate. Due to the positive voltage between collector and emitter, the junction J_1 (n^-p) on the emitter side is polarized in the reverse direction. The junction J_2 (p^+n^+) on the collector side is polarized in the forward direction. In the blocking state, the gate-emitter potential is generally set to a voltage between 0V and -15V to keep the MOS-channel off. A space charge region is built around J_1 and extends far into the low doped n^- type base region. With the increasing reverse voltage, the space charge region reaches the n^+ layer of J_2 , where the heavy doping makes a high electric field gradient. Thus a trapezoidal E-field distribution is formed as shown on the right side of Fig. 2.1. Currently, depending on

the thickness of the base region IGBTs can be commercially designed for a blocking capability up to 6.5 kV.

In forward operation, the gate-emitter voltage is usually switched to +15V. The MOS-channel is turned on. Electrons flow through the MOS channel and n^- base region to junction J_2 on the collector side. Similar to a bipolar transistor, holes from p^+ are injected into the n^- base. Due to the wide n^- base of the PNP-transistor (see Fig. 2.1), the current gain of the transistor is less than one [Sil05]. That is, its base current is greater than the collector current. As the PNP transistor gets into the on-state, holes get from the p^+ layer (collector) in the n^- base. The base is then flooded with an electron-holes plasma and gets in the state of high injection. This results in a good conductivity in the forward operation of the IGBT. The hole current flow is suppressed by the deep trench gate geometry. The injection efficiency of holes near to J_1 at the emitter side is therefore enhanced. A plasma distribution similar to a thyristor can be realized due to this injection enhancement effect [KOH⁺93] [OOSO97]. Compared to the planar and trench IGBT, as shown in Fig. 2.2, the combination of field stop layer and trench cell design results in the optimized carrier distribution, the minimum forward voltage drops and the minimum switching losses [LMP⁺00]. Further analysis in this work will be carried out using this cell structure.

2.2 Basic design of IGBT modules

Packaging is generally an important topic of semiconductor industry. On one side, the package offers mechanical protection to the chips. On other side, the chips are thereby electrically and thermally connected to the outside. In the applications, a power circuit always consists of several components to satisfy the electrical and thermal requirements. Integration of IGBTs and diodes into power modules facilitate the design of power

Figure 2.2: Plasma distributions of Planar-IGBT, Trench-IGBT and Trench FS-IGBT [LMP⁺00] [R.B10]

electronics.

DCB-substrates (direct copper bounding) are frequently used as the circuit substrate in the IGBT modules where two copper layers are bonded on each side of a ceramic sheet. Fig. 2.3a. shows the construction of such a substrate equipped with IGBT and diode chips. IGBT- and diode-chips are bonded on the top side of a DCB. The top side copper layer of the DCB serves as the circuit carrier. The ceramic and the bottom copper layer of the DCB ensure the electrical insulation and heat conduction to the heat sink. Wire-bonds are applied to realize the electrical connection between the top side of the chips and electrical terminals on the DCB. Fig. 2.3b. shows an half bridge circuit. The circuit consists of two IGBTs connected in series and their anti-parallel free-wheeling diodes.

Besides the desired electrical and thermal connections, the module structure results in inevitable parasitic effects. The parasitic electrical resistances and inductances of the bond wires, parasitic capacitance within

the substrate are some main parasitic components of a module structure. The equivalent circuit of an IGBT module, as can be seen in Fig. 2.3c, is therefore much more complex than the circuit in Fig. 2.3b. The circuits for parallel connected dies are even more complex than Fig. 2.3c. The parasitics of the module structure together with the semiconductor chips may constitute resonant circuits which are excited during switching processes resulting in oscillations. This leads to significant deviation from the ideal clean switching [BD10]. The switching performances is worsened by such oscillations. Further investigation of the resonant circuit will be addressed in Chap.4.2.

2.3 IGBT turn-off

To bring the IGBT from the conducting state to the blocking state, the plasma must be extracted out of the base region. This extraction process determines the dynamic behaviour of the IGBT. Fig. 2.4 shows the circuit setup applied for the dynamic characterization of the IGBT modules. During the dynamic test of the bottom IGBT, the top IGBT of the half bridge is kept off by a gate-emitter voltage V_{GE_off} between 0V and -15V. The anti-parallel diode in the upper leg of the half bridge acts as the free-wheeling diode. The load inductance is connected parallel to the Top IGBT. A capacitor bank (so-called DC-link C_{DC}) is connected across the half bridge. C_{DC} acts as an ideal voltage source for a short time. When the DUT is open (V_{GE} of the DUT is driven to V_{GE_off} , the DUT is therefore blocking). The terminal voltage V_{CE} on the DUT is equal to V_{DC} . And there is no current in the load inductance L_P . After the DUT is turned on, V_{GE} is switched to V_{GE_on} (typically +15V), the current in the L_P and in the DUT increases. The load current I_L can be adjusted by load inductance L_P and the time t_{on} according to Eq.(2.2). In order to bring the IGBT to a stationary flooded condition, t_{on} is generally adjusted to

a. Construction of a substrate in IGBT modules

b. Ideal circuit of a half bridge

c. Schematic circuit of a real IGBT switch

Figure 2.3: a. ideal diagram of a IGBT module, b. Substrates with DCB structure and bonding wire, c. Equivalent circuit of a real IGBT module

Figure 2.4: Circuit setup of the dynamic test

more than $100\mu\text{s}$.

$$\frac{dI_L}{dt} = \frac{V_{DC}}{L_P} \quad (2.1)$$

$$I_L = t_{on} \frac{V_{DC}}{L_P} \quad (2.2)$$

With a forward current up to two times of the rated current ($2 \times I_N$), the IGBT is still in the IGBT-saturation¹. The base region is flooded with a plasma of electrons and holes and the basic region is electrically neutral. Since the charge carrier density in the plasma is several orders of magnitude over the base doping, a weak E-field is sufficient to drive the load current passing through the base region. Fig. 2.5 presents the charge distribution in the base region schematically. Due to the in the weak E-field in the plasma, the mobility of electrons is about three times higher than that of holes. The ratio of electron-current to the hole-current should be roughly 3/1. On the emitter-side only holes are deduced over J_1 . The total electron-current is delivered by MOS channel.

Figure 2.5: Schematic of plasma and carrier-flows in the stationary on state of an IGBT

¹IGBT-saturation describes a state of the IGBT in which the charge carriers in the IGBT base region are ample and the forward current increase strongly with the forward voltage.

Figure 2.6: Turn-off diagram of a 3.3kV Trench Field Stop IGBT (simulation)

Fig. 2.6 represents a turn-off diagram of a 3.3 kV Trench Field Stop IGBT. The turn-off process begins with the switching off the driver voltage. The driver voltage is switched from +15V to 0V at the time t_0 . Similarly to a MOSFET, the discharging of the gate capacitance drives the gate voltage first down to the Miller-plateau. In such period, the conductivity of the MOS-channel and V_{CE} is kept nearly unchanged. On the Miller plateau, from t_1 to t_2 , the MOS channel is in the MOS-saturation². Due to the reduction of the channel conductivity, V_{CE} increases by a few volts. The enlarged part below in Fig. 2.6 shows a linear increase of about 10V. The major portion of this voltage drops on the MOS channel. The reduction of the channel conductance can be thereby compensated and the channel

²MOS-saturation describes a state of the MOS-channel, in which the conductivity is sensitively dependent on the gate voltage.

current is maintained. Because of the Miller capacitance, the feedback of V_{CE} on the gate is so strong that the V_{GE} remains almost constant in this period. Such behaviour can also be found on the Miller-plateau during turn-off of a MOSFET (Fig. 2.7). The channel current vanished with further discharging the gate capacity.

From time t_2 , the IGBT passes to the active-state³. The channel current begins to decrease with the gate voltage. The sinking of channel current means less incoming electrons from the emitter side to the base region. Because V_{CE} is still below the V_{DC} , the load current can not be commutated to the free-wheeling diode. The total load current must be carried by the IGBT. On the emitter side the missing electron-current must be taken over by holes. The electrical neutrality under the J_1 is broken. A layer of space charge with a net holes density of N_{h_net} appears in the vicinity of J_1 . The space charge region SCR is formed and expands with the time to the collector. Poisson-equation (2.3) describes the dependence between the electric field and charge density in the SCR . Since the charge carrier density is much higher than the base doping, in one dimensional-case Eq.(2.3) can be expressed to Eq.(2.4).

$$\frac{\partial E}{\partial x} = \frac{q}{\epsilon\epsilon_0} (N_D - N_A + N_h - N_e) \quad (2.3)$$

with $N_{h_net} = N_h - N_e \gg N_D - N_A$

$$\frac{dE}{dx} = \frac{q}{\epsilon\epsilon_0} N_{h_net} \quad (2.4)$$

With the modern computer technology, the essential semiconductor equations can be numerically solved by the Device Simulators to represent the

³In the active-state of a IGBT, the channel current is linearly dependent on the gate voltage.

Figure 2.7: Turn-off process of a MOSFET with inductive load (Fig.9.17 [LSSD11]). At the time t_{rv} the MOSFET is on the Miller-plateau.

dynamic processes in semiconductors. Nevertheless the following analytical assessment is often useful to better understand the processes.

During the turning-off of IGBTs, the building-up of *SCR* can be analytically derived under the following assumptions and simplifications approximately.

- Homogeneous charge distribution and neutrality in the plasma with the charge density $N_{Plasma} = N_{h_Plasma} = N_{e_Plasma}$
- Homogeneous charge distribution and triangular E-field distribution in main part of *SCR*
- In the plasma the electron velocity v_{e_Plasma} is three times higher than that of the holes v_{h_Plasma} .

$$I_{CE} = AqN_{Plasma} (v_{h_Plasma} + v_{e_Plasma}) \quad (2.5)$$

A: chip area

With $v_{e_Plasma} = 3v_{h_Plasma}$ the velocity of the electrons in the plasma can be represented as a function of load current and plasma density. The *SCR* extends to the collector with the velocity v_{e_Plasma} . After the turn-off of the channel current the neutral plasma front drifts toward the collector with the velocity of v_{e_Plasma} . For the same IGBT, higher load current would lead to higher electron velocity in the plasma during the on-state and therefore leads to the faster extension of *SCR* during turning-off.

$$v_{e_Plasma} = \frac{3}{4} \frac{I_{CE}}{AqN_{Plasma}} \quad (2.6)$$

$$W_{SCR} = v_{e_Plasma} (t - t_2) \quad (2.7)$$

The space charge, Q_{SCR} in Eq. (2.8) and the maximum field strength, E_{max}

in Eq.(2.10), increases first linearly with W_{SCR} . Accordingly, the collector-emitter voltage V_{CE} increases proportionally to the square of W_{SCR} . In the analysis of high-frequency oscillations, the instantaneous chip capacitor of SCR and it's effect of the resonant circuit are important factors. A differential capacitance C_{diff} can be defined as Eq.(2.9). C_{diff} acts as the instantaneous chip capacitor. It is inversely proportional to W_{SCR} . This can also be applied to the reverse-biased diodes. Generally, larger C_{diff} leads to lower resonant frequency in the parasitic resonant circuit.

$$Q_{SCR} = qAN_{h_net}W_{SCR} \quad (2.8)$$

$$C_{diff} = \frac{\frac{dQ_{SCR}}{dt}}{\frac{dV_{CE}}{dt}} = \frac{A\epsilon\epsilon_0}{W_{SCR}} \quad (2.9)$$

Under higher current, E_{max} raises to the critical field strength E_C in short time after the MOS-channel turn-off. E_C is about $2 \times 10^5 V/cm$ for Silicon. The gate resistor limits the discharge current of the gate circuit and thereby slows down the constriction of the MOS-channel. For the building-up of SCR and ramping-up of V_{CE} , the larger R_g acts in the same direction as a smaller I_{CE} does.

$$V_{CE} = \frac{1}{2} \frac{qN_{h_net}}{\epsilon\epsilon_0} W_{SCR}^2$$

$$E_{max} = \frac{qN_{h_net}}{\epsilon\epsilon_0} W_{SCR} \quad (2.10)$$

$$= \sqrt{\frac{2q}{\epsilon\epsilon_0} V_{CE} N_{h_net}} \quad (2.11)$$

With a proportionality λ between the nett hole density and the collector current as defined in Eq.(2.12), Eq.(2.11) can be eliminated to Eq.(2.13)

which makes the relationship between the E_{max} in the SCR and the V_{CE} as well as I_{CE} visible. For the ideal abrupt channel turn-off λ can be taken as $\frac{1}{Aqv_{h_Plasma}}$.

$$N_{h_net} = \lambda I_{CE} \quad (2.12)$$

$$E_{max} = \sqrt{\frac{2\lambda q}{\epsilon\epsilon_0} V_{CE} I_{CE}} \quad (2.13)$$

During the forward conduction of IGBT, the anti-parallel diode is of reverse bias, so that during the raise of V_{CE} , no plasma in the diode need to be extracted. The space charge density in the diode is then equal to the doping concentration N_D in the base region. The W_{SCR} in diode can be calculated according to Eq.(2.14) to estimate the C_{diff} of diode.

$$W_{SCR} = \sqrt{\frac{2\epsilon\epsilon_0 V_{CE}}{qN_D}} \quad (2.14)$$

In the IGBT, during the high current turn-off, the E_{max} reaches the critical value E_C at a V_{CE} which is significant below the blocking capability of the IGBT due to the high hole-density in *SCR*. The further increase of the E-field is then limited by the impact ionization. The generated holes are evacuated over J_1 . The electrons drift through the *SCR* to J_2 . This leads to a reduction of the net hole density and dE/dx in *SCR*. From t_3 , the resulted slew rate reduction on V_{CE} can be observed in the turn-off diagram. After V_{CE} exceeds V_{DC} at t_4 the load current begins to commutate to the free-wheeling diode. The I_{CE} reduction leads to the further dE/dx reduction in *SCR* and the reduction of the *SCR* extension velocity. However, this process is limited by the parasitic inductance in the load current circuit. Up to t_5 the current has dropped so far that effect of the dE/dx reduction compensates the effect of *SCR* extension. V_{CE} reaches its maximum at t_5 . After t_5 , the slew rate of V_{CE} can not

remain positive. From t_5 to t_6 V_{CE} decreases rapidly to V_{DC} . Up to t_7 , the residual plasma is evacuated by the so-called tail current. Just by t_7 , SCR reaches the field stop layer. The tail current gets rapidly to zero and causes a second peak on the V_{CE} .

2.4 Possible RF-oscillations during IGBT switching

In the turn-off process of an IGBT, there is always temporary states in which a SCR is already built up and at least part of the load current is still flowing. Because of the high field strength in SCR , the charge carriers are accelerated up to the saturation velocities. In case of a superimposed RF-voltage, the charge carrier transit time and the charge injection time would induce a phase shift between current and voltage on the device terminal. At certain frequencies, such phase shifts are between $\pi/2$ und $3\pi/2$. At such frequencies, the power consumptions on the device are negative. The device works therefore as an amplifier. In Chapter 3, the mechanisms and the efficiency of the transit time effect are discussed in detail.

As mentioned previously in 2.2, the module constructions are always accompanied with parasitic resonant circuits. The noise signal can be raised to disturbing HF-oscillations, if the eigenfrequencies of the resonant circuit match the frequency band of the transit time effect.

2.4.1 Tail oscillation PETT

One type of high-frequency oscillation in IGBT modules is reported as PETT(plasma extraction transit time) oscillation in 2002 [GMS02]. Since the oscillation is usually observed in the tail-current phase, it is also referred as tail oscillation. Fig.2.8 shows the diagrams of the PETT oscilla-

Figure 2.8: PETT oscillation of IGBT [Gut06]

tion. The oscillation signal is sensed by a wire loop antenna through the electromagnetic coupling. The Fourier spectrum in Fig.2.8 shows that the frequency of oscillation is around 246 MHz. It lies well above the oscillations frequencies caused by negative gate capacitances(0.4 to 2.5 MHz is shown in [OFO99]). The mechanisms of PETT oscillation was explained to be similar to the mechanisms of a BARITT(Barrier Injection Transit Time) diode. The work mechanism of PETT can be seen in Fig.2.9. The superposition of a small AC-signal on the collector-emitter voltage will lead to a pulse-like holes extraction on the plasma boundary to *SCR*. The AC-signal would be amplified by the transit-time effect periodically, when the semiconductor chip and the module structures build up the appropriate resonant circuits. Finally, the oscillation signal can be perceived as the HF-disturbance. The PETT mechanisms was also found in the turn-off process of fast recovery diodes [SMNL06].

Figure 2.9: Mechanisms of PETT oscillation [Gut06]

There are generally two methods to suppress oscillations. The first one is to integrate power consumers into the resonant circuit. The power consumer may be resistance in the line, which affects all frequencies. It may also be the resistive coating on the conductor-line and is especially effective for high-frequency signals. Energy of the HF-signal can be extracted also by a resistive conductor adjacent to the resonant circuit. The HF-power is converted to heat by eddy current in such conductor [SGMM03] [SGMM02]. The second method of the oscillation suppression is to mismatch the eigenfrequencies between the amplifier and the resonant circuits.

The emitter cross wire-bonding between parallel IGBTs has already been patented in 1995 as a countermeasure to HF-oscillations during IGBT turn-off as shown here in Fig.2.10. During switching off of modules with paralleled IGBT chips (3,4), high frequency oscillation can cause the electromagnetic interference and the destruction of the chips can still not be excluded. These oscillations can be prevented by the direct emitter wire-bonding (14) between the adjacent chips. The cross wire-bondings reduce the inductance between the adjacent IGBTs and thus the resonant circuits are out of tune. There are also proposed solutions by increasing the inductance of the resonant circuit to suppress PETT-oscillation [TKY⁺98]. This usually means the increasing of inductance in the commutation path of the load current. Such inductance is generally unfavourable for high-performance power modules. Additional parasitic capacitances and inductance can also be integrated on the DCB-layout to mismatch the eigenfrequencies [Mou01a]. Countermeasures to PETT-oscillation are summarized in chapter 5 of [Gut06].

Although the PETT-oscillation in the tail current phase is a dynamic process, it shows certain quasi-stationary characteristics. The IGBT has taken over the entire DC-link voltage before PETT-oscillation appears. The collector-emitter voltage remains almost unchanged during PETT-oscillation. The *SCR* capacitance of IGBT- and diode-chip as well as the

Figure 2.10: Cross wire-bonding to suppress the high-frequency turn-off oscillation in IGBT modules - EUPEC patent Zimmermann 1995 [ZS95]

eigenfrequencies of the resonance circuits in the module-structure vary only within a narrow range. For example, tail oscillations were found in the range between 74MHz and 78.2 MHz on a 4500V IGBT module [FHK09]. Therefore, the mismatching between the resonant circuit and the transit time effect can be relatively simply realized.

2.4.2 Dynamic Impact-Ionization Transit-Time oscillations of freewheeling diodes

It is already known that avalanche generation may accrue in the reverse recovery phase even for a reverse bias voltage below the steady-state breakdown voltage [BS67]. This phenomenon is similar to the dynamic avalanche during the turning off of GTO-thyristors and IGBTs [SN89] and of the fast recovery Si diodes [DLS03]. High-frequency oscillations have been found during the turning-off of fast recovery diode in the dynamic avalanche phase [LSG98] [Lut06]. Fig.2.11 shows the circuit of the measurement

setup to test the reverse recovery behaviour of the freewheeling diode in the IGBT Module by double pulse mode. The top side IGBT of the half bridge is actively switched with double pulse during the bottom IGBT is kept turned-off. The load current in L_p is adjusted by the V_{DC} and the first pulse width. After the first pulse, the top IGBT is in off-state and the bottom freewheeling diode (DUT) is loaded with the current in L_p . As the top IGBT is turned-on again, the reverse recovery characteristic of the freewheeling diode occurs. The dynamic oscillation is observed after the recovery current peak and during the tail current phase of the diode.

Fig.2.12 shows the measurement on the 1200V diode at 0°C [LSG98]. After the Reverse recovery current peak (time point t_1) the reverse current decreases. Between t_1 and t_2 , the voltage climbs up close to the value of the DC-link voltage, while the tail current flows in the diode. After t_2 a current hump grows from the tail current. This hump occurs only if the DC-link voltage is above 910V. The further increase in the DC-link voltage to 930V effects that a high reverse current suddenly shoots up from the current hump whose amplitude is a multiple of the reverse recovery current peak. A high frequency oscillation in the range >500 MHz is superimposed. After some 100ns(t_4), the oscillation is finished. A reduction of the voltage by only 1 or 2 V, or an increase of the temperature by 1 or 2°C, removes the effect [LSG98].

The dynamic IMPATT oscillation is caused by the K-centre, which is created at irradiation of the semiconductor with high-energy particles [LSG98] for example in the Controlled Axial Lifetime (CAL) technology [LS94] [SL04]. The K-center has the characteristic of a temporary donor and is positively charged during the forward conduction. After the voltage has changed its polarity the K-centre need to be discharged. The time constant of the discharge process is temperature dependent. It is ca. 100ns at 400K and is in the order of some microseconds at temperatures below 300K [LSSD11]. Especially for low temperature, when the IGBT applies

the DC-link voltage to the diode in a very short time after the the current zero-crossing point of the diode, the break down voltage of the diode is temporarily reduced due to the positively charged K-center and impact ionization sets on [LSSD11]. Fig.2.12 shows the measurement on a 1200V

Figure 2.11: Test circuit of the reverse recovery behaviour

diode at 0°C . The work mechanism is shown in Fig.2.13. The electron packets generated by the avalanche run through the SCR to the right side. During the dynamic avalanche the device has the similar property as the IMPATT diode [Sze81] and is able to raise HF-oscillations. To avoid these oscillations, it is necessary to restrict the number of the generated donor states. The irradiation parameter as well as annealing temperature and time have to be chosen carefully [SSL02b] [SLH04]. Early works [SSL02a] [SL04] show the possibility to optimize the irradiated devices by means of simulations. The dynamic IMPATT oscillation can be surely avoided by the the proper adjustment of the K-centre concentration [LSSD11]

Figure 2.12: Dynamic IMPATT oscillation measured at 0°C on 1200V diode [LSG98]

Figure 2.13: Process in the device at dynamic IMPATT oscillation [LSSD11]

2.4.3 Transient Avalanche oscillations

Meanwhile, a new type of HF-oscillation was found during the IGBT high current turn-off. Fig.2.14 shows TA-oscillation [HPB⁺14] during high current turn-off of 3.3 kV IGBT. The oscillation appears only during the rise of V_{CE} and fall of the I_{CE} for several hundreds *ns*. And it was observed only during the dynamic avalanche process. TA-oscillations occur during the fast expanding the SCR. Therefore, in contrast to the PETT-oscillation, TA-oscillation shows even more dynamic characteristics. During the TA-oscillation, high current is still flowing through the IGBT. The typical frequency of TA-oscillations is found between 100MHz and 700MHz. This indicates the relevance of TA-oscillation to the transit time effect. The understanding of the mechanism of TA-oscillation by analysis of measurements and simulations is presented in Chap.4.

Figure 2.14: Measurement diagram of TA-oscillation, adapted from [HPB⁺14]

Chapter 3

Mechanisms of Transit time effect and RF-oscillations in IGBT Modules

When a point charge is moving in a conductor system, the terminal current of the conductor system is determined by the movement of the point charge but not the charge carriers arriving to the electrodes [Sho38]. This is known as “Ramo-Shockley theorem”. Using this theorem the transit time effect of charge carriers through the space charge region can be explained. IMPATT diodes and BARITT diodes are the devices based on the similar principle corresponding to the transit time effect. In the diodes, a space charge region (SCR) is formed at a reverse biased pn-junction by a DC-voltage. The charge injection into the *SCR* can be induced by a AC voltage supplied additionally to the DC voltage. In a IMPATT diode, the E_{max} is set to just below the E_C so that the additional AC-voltage would drive the E_{max} to exceed E_C . The charge injection is caused by the avalanche generation. In the BARITT diodes, there is a forward biased pn-junction

Figure 3.1: Structure and work mechanisms of IMPATT diode

opposite to the blocking pn-junction J_2 . Under the DC-voltage, the J_2 stands just below the threshold voltage. The additional AC-voltage to the DC-voltage would drive the J_2 into the forward conduction. The forward current of J_2 induces the charge injection into the SCR. The injection current needs a certain injection time T_{inj} to reach its maximum before it vanishes in the negative half period of the AC-voltage. The injected charge carriers need a transit time T_D to drift through the SCR. T_{inj} and T_D together induce a phase shift between the AC terminal voltage and the AC terminal current of the diodes. In a certain frequency range, in which the phase shift is between $1\pi/2$ and $3\pi/2$, the diode acts as amplifier.

3.1 Amplification process

The IMPATT diode is an avalanche transit time structure. The simplest variant of the avalanche transit time structure is shown in Fig. 3.1. Under the reverse bias voltage V_{DC} , the SCR is built up around p^+n junction

J_1 . On the p^+ side, the E-field decreases steeply to zero. Due to the weak doping level of the n region, the *SCR* stretches wide into it. At the operation point, the V_{DC} is adjusted just below the breakdown voltage. The maximum field strength E_{max} directly at the position J_1 is just below the critical field strength E_c . In operation, an AC voltage V_{AC} is superimposed on V_{DC} . In the positive half cycle of V_{AC} , the structure is driven into avalanche. Electron-hole pairs are generated near J_1 . The holes are extracted through p^+ to anode during the electrons have to drift through the transit-time region to the cathode side. The E-field in the drift region is adjusted high enough so that the drift velocity of the charge carriers reach the saturation velocities v_s . As already described in Chap.2, a similar avalanche transit-time structure is formed from t_3 (Abb. 2.6) during turn-off of the IGBT.

3.1.1 Avalanche breakdown

Under the strong E-field, electrons / holes can be so strongly accelerated that they gather sufficient energy on their free-moving length to generate new electron-hole pairs through the collision with the next atomic lattice. The ionization coefficient $\alpha_{n,p}$ indicates the number of electron-hole pairs generated by an electron / hole on the distance of 1cm. The structure comes into breakdown, when the charge carrier multiplication in the *SCR* leads to a infinity current. The avalanche breakdown criterion is given by equation (3.1) in [Sze81] S.100.

$$\int_0^W \alpha_n \exp \left(- \int_x^W (\alpha_n - \alpha_p) dx' \right) dx = 1 \quad (3.1)$$

The equation (3.1) can be simplified to (3.3) using the definition of the effective ionization coefficient α_{eff} in equation (3.2) [LSSD11] P.62. For the E-field strength over 1×10^5 , the ionization rate increases exponentially. Fig. 3.2(a) shows the field dependence of the ionization coefficient

Figure 3.2: (a) The ionization coefficient as a function of E-field at 300K, similar to Fig. in [LSSD11] P.65. (b) The temperature dependence of the ionization coefficient under the E-field of 2.5×10^5 V/cm

for silicon at 300K [LSSD11].

With increasing temperature, the free-moving length and thus the ionization coefficient decreases. The temperature dependence of the ionization coefficient between -20°C and 150°C for silicon can be described by equations (3.4) to (3.6) with good approximation [LSSD11] P.64. Fig. 3.2(b) shows the temperature dependence of the ionization coefficients for silicon under the E-field strength of 2.5×10^5 V/cm. From 300K to 450K the ionization coefficients decrease by roughly 50%.

$$\alpha_{eff} = \frac{\alpha_n - \alpha_p}{\ln(\alpha_n/\alpha_p)} \quad [\text{LSSD11}] \text{ P.61.} \quad (3.2)$$

$$\int \alpha_{eff} dx = 1 \quad (3.3)$$

$$\alpha_n = 1.1 \times 10^6 \cdot \exp\left(-\frac{1.46 \times 10^6 + 1100(T - 300K)}{E}\right) cm^{-1} \quad (3.4)$$

$$\alpha_p = 2.1 \times 10^6 \cdot \exp\left(-\frac{2.2 \times 10^6 + 1100(T - 300K)}{E}\right) cm^{-1} \quad (3.5)$$

$$\alpha_{eff} \cong 1.06 \times 10^6 \cdot \exp\left(-\frac{1.68 \times 10^6 + 1100(T - 300K)}{E}\right) cm^{-1} \quad (3.6)$$

The avalanche region is generally defined as eq. (3.7), where the integral of the ionisation coefficient along the avalanche region reach 0.95 [Sze81] P.573.

$$\int_0^{W_A} \alpha_{eff} dx = 0.95 \quad (3.7)$$

3.1.2 Amplification efficiency

Fig. 3.3 shows the diagram of terminal voltage, charge injection and the induced current of the transit time structure. The terminal voltage is a result of the AC-voltage override $V_{RF} \sin(\omega t)$ to the V_{DC} .

$$V_T = V_{DC} + V_{RF} \sin \omega t \quad (3.8)$$

According to the "Ramo-Shockley theorem", the diode current is equal to the average convection current through out the drift region ([Pau81] P.315). When a charge packet of Q is injected into the SCR at the position of W_A and drifts with the velocity of v_Q , as shown in Fig.3.1, the inducing current J_{ind} on the terminal can be described by Eq.3.9 [EH98] P.354.

$$J_{ind} = \frac{Q}{W_{SCR}} \left(v_Q - \frac{W_A}{W_{SCR}} \frac{dW_{SCR}}{dt} \right) \quad (3.9)$$

Figure 3.3: Diagram of voltage and current in a transit time structure with the simplified charge injection, Fig. from [EH98], extended

In the positive half period of V_{RF} , electrons are injected into the drift region in the injection time T_i with a injection angle Θ_W ($=T_i \times \omega$). The injected electrons need a transit time of T_D ($=W/v_s$) to drift through the SCR. The transit angle is calculated as $\Theta_D = T_D \times \omega$. Assume that a charge pulse is injected into the drift region at the phase angle Θ_M . The diagram of V_T , V_{RF} , charge injection and the induced terminal current J_{ind} are shown in Fig.3.3 (generalization on the base figure on [EH98] P.355.). The ability to generate RF-power by the transit time structure can be described using the efficiency coefficient η as in Eq.(3.13). The

efficiency coefficient is the ratio of the RF-power generation to the total power consumption which is drawn from the DC-voltage source V_{DC} . The introduction steps are involved in Appendix A

$$J_{DC} = \frac{1}{2\pi} \int_{\Theta_M - \Theta_W/2}^{\Theta_D + \Theta_M + \Theta_W/2} J_{ind}(\omega t) d(\omega t) \quad (3.10)$$

$$P_{DC} = J_{DC} V_{DC} \quad (3.11)$$

$$P_{RF} = -\frac{A}{2\pi} \int_{\Theta_M - \Theta_W/2}^{\Theta_D + \Theta_M + \Theta_W/2} J_{ind}(\omega t) V_{RF} \sin(\omega t) d(\omega t) \quad (3.12)$$

$$\eta = \frac{P_{RF}}{P_{DC}} = \frac{V_{RF}}{V_{DC}} \frac{\sin(\frac{\Theta_W}{2})}{\frac{\Theta_W}{2}} \frac{\cos(\Theta_M + \Theta_D) - \cos \Theta_M}{\Theta_D} \quad (3.13)$$

The injection dependent efficiency coefficient η_1 (3.14) and the transit time dependent efficiency coefficient η_2 (3.15) can be eliminated out of η . η_1 gets its maximum as the Θ_W goes to Zero. The efficient coefficient favoured from the short injection pulse. In an IMPATT diode, the charge injection reach its maximum roughly at $\Theta_M = \pi$. For the BARITT diode, the maximal charge injection happens at $\Theta_M = \pi/2$. The optimal width of drift region and the corresponding transit angle can therefore be estimated for the most effective Rf-power conversion. For IMPATT diode, the optimal transit angle $\Theta_D \approx 3\pi/4$. And for BARITT diode the optimal transit

angle is found to be $\Theta_D \approx 3\pi/2$ [EH98] P.355.

$$\eta_1(\Theta_W) = \frac{\sin(\frac{\Theta_W}{2})}{\frac{\Theta_W}{2}} \quad (3.14)$$

$$\eta_2(\Theta_D) = \frac{\cos(\Theta_M + \Theta_D) - \cos \Theta_M}{\Theta_D} \quad (3.15)$$

As Eq.(3.13) is derived without the interval limitation of 2π , it can be applied to the case of wide drift regions as in IGBT ($\Theta_M + \Theta_D \geq 2\pi$). In the frequency range that the transit angle lies in $(2n-1)\pi < \Theta_D(2n-1)\pi$, the BARITT and the corresponding PETT mechanisms have the ability to generate RF-power. The IMPATT mechanism can theoretically generate Rf-power in the whole frequency range. As can be seen in Fig.3.4, η_2 has at least a second and a third peak value in addition to its maximum. The peak value of η_2 reduces with the increase of the transit angle, but the second peak value of IMPATT mechanism is still as high as the maximum of BARITT structure. The PETT and BARITT mechanism shows the damping effect as long as the frequencies are low [Sil14]. The drift velocity of electrons and holes (v_{dn} , v_{dp}) can be derived on the basis of the measurement of electron and hole drift velocity in Silicon [CMMO75]. The equations Eq.(3.16) and Eq.(3.17) describe the dependencies on temperature and E-field [Bal79] P.10.

$$v_{dn} = \frac{1.42 \times 10^9 T^{-2.42} E}{[1 + (E/1.01 T^{1.55})^{2.57 \times 10^{-2} T^{0.66}}]^{(2.57 \times 10^{-2} T^{0.66})^{-1}}} \quad (3.16)$$

$$v_{dp} = \frac{1.31 \times 10^9 T^{-2.2} E}{[1 + (E/1.24 T^{1.68})^{0.46 \times 10^{-2} T^{0.17}}]^{(0.46 \times 10^{-2} T^{0.17})^{-1}}} \quad (3.17)$$

Under the assumption of a triangle E-field distribution and the maximal E-field of $2 \times 10^5 V/cm$, the effective velocities of electrons and holes ($\bar{v}_{dn}$,

Figure 3.4: η_2 of IMPATT und BARITT/PETT structures according to the transit angle

$\bar{v}_{dp}$) in the SCR are estimated. This results under 25°C $\bar{v}_{dn} \approx 8.93 \times 10^6$ and $\bar{v}_{dp} \approx 6.44 \times 10^6$. As shown in Fig.3.5, the eigenfrequencies of the transit structures decrease according to the expansion of the SCR. The frequency F_{1BA} , at which the BARITT structure is most effective, is about $3/2$ of F_{1IM} . The IMPATT structure at F_{2IM} is still as effective as BARITT structure at F_{1BA} . The transit time frequency F_{2IM} is almost 4 times higher as F_{1IM} . The internal states during the IGBT turning-off are shown in Fig. 4.35. It is generally possible that the IMPATT and PETT mechanisms exist temporally parallel to each other. The SCR acts as the transit region of electrons for the IMPATT mechanism and also as the transit region of holes for the PETT mechanism. The IGBT chips are physically constructed so that they can generally act as RF-generator during the turn-off process. If an adequate resonant circuit exists and it is not well damped, then the thermal noise would be raised to oscillations.

Figure 3.5: Transit angle dependent efficiencies of RF-power Generation η_2 in IMPATT und BARITT/PETT structure using $\bar{v}_{dn} \approx 8.93 \times 10^6$ and $\bar{v}_{dp} \approx 6.44 \times 10^6$

3.2 Resonant circuits in IGBT modules

There are always rich number of resonant circuits in the IGBT applications. They may be constructed of parasitics of the module setups, diode chips and even IGBT chip itself. For the analysis of RF-oscillation, the resonant circuit near to the IGBT chips are of special interest. Fig.2.3 shows for example that the parasitic capacitance between DCB areas and the parasitic inductance can be found near the IGBT- and diode chips. The three resonant circuits in this construction are shown in Fig.4.19. The chip capacitance for the RF-analysis can be estimated according to Eq.(2.9). As the chip capacitances are inversely proportional to the width of SCR, the eigenfrequencies of the resonant circuits increase with the expanding of SCR. At the moment that the eigen frequency matches the transit time frequencies as shown in Fig.3.5, RF-oscillations can be raised. The quantitative analysis are presented according to the sample setups in Chap.4.

Chapter 4

Analysis of Transient Avalanche oscillation

Under high current density, high frequency oscillation signals are found on the trace of gate-emitter voltage during the turning off of 3.3 kV IGBTs as shown in Fig. 2.14. This leads to the presumption that gate capacitance and the Miller-feedback could be responsible for the this oscillation. However this presumption is disproved by the measurements and simulations introduced in this chapter. Moreover the RF-signals are found to be emitted by the emitter wire bonds. As this oscillation occurs only transiently during the dynamic avalanche of the IGBT, it is called TA-oscillation in this work. The resonant circuits of TA-oscillations can be localized nearby the IGBT chips. The parasitics extraction of setups on DCB and the analytical estimations of IGBT and diode chip-capacitance allow the quantitative analysis of the resonant circuits according to the eigenfrequencies. The mechanisms of TA-oscillation are illustrated corresponding to the transit time effect and are verified by device simulations.

4.1 Measurement of TA-oscillations

Systematic measurements are accomplished on 3.3kV IGBT modules and substrates setups to find out the source and the relevant parameters of TA-oscillations.

4.1.1 Measurement of TA-oscillation on IGBT modules

TA-oscillations are first observed on an 3.3kV IGBT module under development, in which $24 \times$ IGBT chips are parallel arranged and act as a single switch. Two modules are series connected to a half-bridge for the dynamic characterization as shown in Fig.2.4. The hardware configuration of the dynamic characterization is represented in Fig.4.1. Because of the high voltage, the dynamic characterizations are generally carried out in a grounded security box. For the conventional measurements three channels of oscilloscope are necessary to record V_{CE} , V_{GE} and I_{CE} . A $1000 \times$ voltage probe is applied for V_{CE} . A $10 \times$ voltage probe is applied for V_{GE} . I_{CE} is converted to voltage signal using a Pearson current transformer. Additionally to the conventional dynamic module characterization, for the oscillation analysis, a wire loop is placed near the DUT to capture the RF-disturbance. Care must be taken for the parasitics in the ground loop in order to distinguish the real RF-signal and the interferences, see appendix B.

Similarly to the PETT oscillation, the oscillation frequencies of TA-oscillation are in the range of several hundreds of MHz. But unlike the PETT oscillation, TA-oscillation does not appear in the phase of the tail current. TA-oscillations are found during the dynamic avalanche of the high current turn-off of the IGBT. As TA-oscillation happens, the DUT is under high voltage and high current density.

Figure 4.1: Test setup for the dynamic characterization of IGBT modules

Effect of load current I_L and DC-link voltage V_{DC}

Three measurement series, MS1, MS2 and MS3, are processed to verify dependences on I_L and V_{DC} . The DUT and roadmap of the measurements is shown in Fig. 4.2. In MS1, the V_{DC} is kept at 2100V during the varying of I_L by T_{on} . In MS2, the I_L is kept at 3200A during the varying of V_{DC} and T_{on} . In MS3, the L_P and T_{on} are kept constant. According to Eq.(2.2), the variation of V_{DC} leads simultaneously to a proportional change of I_L .

TA-oscillations are found under various combinations of I_L and V_{DC} . Fourier transform is applied on the measured signals to evaluate the oscillation in the frequency domain. Fig.4.3 shows the discrete Fourier transform (FFT) of the antenna signals measured in MS1. The RF-oscillation of 625MHz is found under the condition of V_{DC} =2100V and I_L =3200A, right in Fig.4.3. The amplitude of the oscillation decreases first with the reduction of turn-off current I_L . The oscillation vanishes until I_L =2800A and appears again at I_L =2500A with a lower frequency of about 300MHz, left in Fig.4.3. Further reduction of I_L leads to the final oscillation-free turn-off. In MS2, I_L is kept at 3200A during the varying of V_{DC} . The

Figure 4.2: Roadmap of measurement series to find out the dependences of TA-oscillations on I_L and V_{DC} .

Figure 4.3: Discrete Fourier transform (FFT) of antenna signal of MS1: variation of load current I_{CE}

varying of DC-link voltage shows the influence on the TA-oscillation in the same manner as the varying of I_L . As shown in Fig.4.4, after the 625MHz oscillation is found at 3200A I_L and 2800V V_{DC} the V_{DC} is reduced. The oscillation almost disappears as V_{DC} decreases from 2100V to 1800V and appears again at V_{DC} =1650V with the lower frequency of about 300MHz. For the DC-link voltage below 1600V, no oscillation is found during the turn-off process. Fig.4.5 shows that MS3 has same tendency as MS2 and

Figure 4.4: Discrete Fourier transform (FFT) of antenna signal of MS2: variation of DC-link voltage V_{DC}

MS1. In the measurement serie MS3, the 300Mh oscillation is found at 2720A I_L and 1890V V_{DC} . The Product of V_{DC} and I_L is found to be a key parameter for the TA-oscillations. The frequency of the oscillations under the hard condition ($I_L \times V_{DC} = 6.7 \times 10^6 VA$) can identified at about 625MHz, and the frequency of the oscillation under soft conditions

Figure 4.5: Discrete Fourier transform (FFT) of antenna signal of MS3: variation of DC-link voltage and load current

Figure 4.6: Product of V_{DC} and I_L as the TA-oscillations occur

($I_L \times V_{DC} = 5.3 \times 10^6 VA$), lower current and (or) lower voltage, is identified to be about 300MHz. As shown in Fig.4.6 the product of I_L and V_{DC} is found to be a decisive parameter for the oscillations in all the three measurement series.

The analytical estimation in chap.2 helps to understand the effect of $V_{DC} \times I_L$. As Eq.(2.13) represents, the maximal e-field E_{max} of the SCR is proportional to $\sqrt{I_L \times V_{DC}}$. This indicates the relevance between the avalanche generation and the TA-oscillation. To reach the critical e-field, the W_{SCR} is roughly inversely proportional to the current of the IGBT. Therefore the lower current of IGBT means wider SCR and longer transit time of the SCR and accordingly the lower frequency of the transit time effect.

Further measurements show that under certain conditions more partial frequencies are even found in a single turn-off process of the IGBT module. Fig.4.7 shows for example a finely resolved diagram of the TA-oscillation. Beside the frequencies of 300MHz and 625MHz, a partial frequency of 430Mhz is even found to be the dominant frequency of the oscillation. The analysis in the Chap.3 shows that, for the IMPATT mechanism, RF-power generation on at least two frequencies is possible. Together with

Figure 4.7: Multi frequency TA-oscillations during the Module turn-off

the possible hole transit time effect, the frequencies population of TA-oscillation can be more complicated.

Effect of temperature

The dynamic Avalanche oscillation was reported in the similar frequency range of several hundreds of MHz as TA-oscillation [LSG98]. As discussed in chap.2.4.2, such oscillations are quite sensitive to the temperature. The following measurements are processed to verify, if TA-oscillation is affected by the temperature similarly as the dynamic IMPATT oscillations. In this connection, the junction temperature is varied from 25°C to 125°C. MS3 as shown in Fig.4.2 is processed at each temperature. The oscillation of 625MHz becomes weaker due to the raise in temperature. In order to reproduce the 625MHz oscillation at 80°C, the product of I_L and V_{DC} must be adjusted 30% higher than the test at 40°C. The condition to reproduce the 300MHz oscillation increased by about 10% when the temperature rises from 40°C to 80°C. For the temperature above 80°C, no 625Mhz oscillation can be detected. Due to the limitation of over voltage by the IGBT blocking capability, we have to dispense with the further increase of the current. The 300MHz oscillations are still found at 125°C. Fig.4.8 shows the conditions of the oscillation accordingly to the temperature variation. The

Figure 4.8: Start conditions of the oscillation corresponding to the temperature

dynamic avalanche oscillation is caused by the K-centre and can be suppressed by the temperature variation of a few degrees. No such strong temperature sensitivity can be demonstrated for TA-oscillation. The causes of the two oscillation types are not the same.

The discussion in Chap.3 shows that the coefficient of the impact ionization decreases about 1/4 when the temperature rises from 40°C to 80°C, that can be read out Fig.3.2. This agrees with the temperature sensitivity of the 625MHz oscillations. The 300MHz oscillations do not show such temperature sensitivity.

4.1.2 Measurement of TA-oscillation on a substrate and single chip

The early investigations show that the PETT oscillations are found on the parallel switching of IGBTs and diodes [MGS02] [GMS02]. To investigate the effects of paralleling chips on TA-oscillations, additionally to the

engineering IGBT module with $24\times$ paralleling IGBTs, measurements on substrate setups with $4\times$ IGBTs, $2\times$ IGBTs and also on setups with single IGBT chip are conducted. The TA-oscillations are even found on the single IGBT setups. The paralleling of chips is not absolutely necessary for TA-oscillation.

The following measurements on the substrate setups are carried out to find out the critical parameter, the amplifier, the resonant circuits of TA-oscillation.

Effect of R_G on substrate

To eliminate the effect of the asymmetry in the gate circuit of the IGBT Modules, measurements are processed on the substrate setups with much simpler gate circuit than the whole engineering module. Fig.4.9 shows a substrate setup with $4\times$ IGBT- and $4\times$ diode-chips arranged on a DCB. Each IGBT chip is connected to an anti-parallel diode via emitter wire bonds. This substrate setup works as a single switch.

During the varying of gate resistor from 6.8Ω to 15Ω , all the other conditions are kept unchanged. Fig.4.10 shows the traces of the gate-emitter voltage. The amplitude of TA-oscillation increases with R_G till $R_G=10\Omega$ and then reduces with the further increase of R_G . No oscillation is found in the range of $R_G \leq 6.8\Omega$ and $R_G \geq 15\Omega$.

Smaller R_G leads to the faster turn-off of the MOS channel. The resulted higher net hole density at the emitter side leads to a steep raise of V_{CE} . Therefore the varying of R_G has a similar effect as the varying of the I_L in the reverse direction.

Effect of cross wire bonding between IGBT chips on a substrate

According to Zimmermann patent, shown in Fig.2.10, and [Gut06], the Cross wire bonds between the adjacent IGBTs were proved to be an effective manner to suppress the PETT oscillations by mismatching the reso-

Figure 4.9: Substrate setup with 4×IGBT- and 4×diode-Chips

Figure 4.10: RF-oscillations dependence on R_G

nant circuit frequency and the PETT frequency. As the TA-oscillations are found in the similar frequency range as PETT oscillations, the cross wire bonds were also expected to suppress TA-oscillations. Fig.4.11 shows the measurements on substrates with different cross wire bond configurations.

A 30MHz oscillation is found on the substrate which can be suppressed by the cross wire bond 1 and 2 between the adjacent IGBTs as shown in Fig.4.9. The 600MHz oscillation is however not obviously suppressed by the cross wire bonds. As discussed in Chap.??, during the time the PETT normally happens, the SCR in IGBT and diode vary only within a narrow range. The eigenfrequencies of the resonant circuits can be shifted out of the transit time frequency of PETT mechanism by the cross wire bond. But the TA-oscillation occurs during the raise of V_{CE} and fall of the I_{CE} . The SCR is expanding much more dynamically than during PETT oscillation. During TA-oscillation, because of the expanding of SCR, both the transit time frequency and the resonant frequencies are changing within a wide range. This makes it difficult to mismatch the resonant circuit to the transit time frequency by cross wire bonds.

Effect of chip thickness

In case of the PETT oscillation, the oscillation occurs at the end of the tail current phase. The SCR has reached almost the field stop layer on the back side of the IGBT. Therefore the transit time frequency should be inversely proportional to the width of the base region of IGBT. To vary the base region IGBT chip types with identical front side and back side structures but different thickness are processed. Substrates with 4×IGBTs, as shown in Fig.4.9, were setup for each IGBT types. Fig.4.12 shows the discrete Fourier transform (DFT) of the TA-oscillation signals measured on the substrate with different IGBT chips. Oscillations of 475MHz are found at all the four chip types. No influence of the chip thickness can be identified. This indicates that, as the TA-oscillations occur, the SCRs do not reach

Figure 4.11: a. TA-oscillations on substrate without cross wire bond. b. TA-oscillations on substrate with two cross wire bonds

the field stop of all chips.

Source of the oscillation signal

As the TA-oscillation was first observed on the gate-emitter voltage of the IGBT module, the active participation of gate was assumed. To find out if the the gate emitter voltage actively participates in the TA-oscillation, it is necessary to measure the chip internal gate emitter voltage during TA-oscillation. As a gate resistor is already integrated on the IGBT chip, a special sense contact as shown in Fig.4.13 was constructed to pick up the potential between the gate- and the emitter-metallization. At the end of a BNC cable two feather needles are contacted. The distance between the needles is adjusted so that one of them contacts on the gate and the other on the emitter-metallization. The loop area between the two needles is kept as small as possible to avoid the EMI coupling. For the single chip

Figure 4.12: TA-oscillations are measured at $I_L = 500A$, $V_{CD} = 2500V$ under $25^\circ C$

Figure 4.13: Sense contact to measure the The internal gate-emitter voltage V_{GEIN} is directly on the IGBT chip with needle sense contact, oscillation on the antenna and no oscillation on the gate is observed.

measurements, just one IGBT and one diode on the substrate setup are connected. The signal of antenna and the internal gate voltage V_{GEIN} of the IGBT are shown in Fig.4.13. As compared to the oscillation on the antenna, no high frequency disturbances can be found on the V_{GEIN} . As TA-oscillation happens, the internal gate-emitter potential is driven to under the threshold voltage of the IGBT. Therefore the previously detected oscillation signal on V_{GE} is just passive raised by the EMI coupling. The active participation of V_{GE} in TA-oscillation can not be demonstrated.

Furthermore varying the relative position and direction of the antenna to the emitter wire bonds cause the strong influence on amplitude of the measured oscillation signals. The oscillation signal received by the antenna parallel to the wire bonds (Fig.4.14) is more than three times stronger than that received by the perpendicular antenna (Fig.4.15). This indicates that emitter wire bonds are part of the resonant circuit of TA-oscillation. The wire bonds act as antenna and emitted the RF-oscillation signals.

Figure 4.14: Measurement of the TA-oscillation with the antenna parallel to the emitter wire bonds

Figure 4.15: Measurement of the TA-oscillation with the antenna perpendicular to the emitter wire bonds

Effect of the wire bond configuration

After the emitter wire bonds are found to be part of the resonant circuit of TA-oscillation, measurements are processed on different wire bond configurations, as shown in Fig.4.16. Fig.4.16 shows the test setups with different configurations of wire bonds. On DUT1 in Fig.4.16(a), the IGBT and the anti parallel diode are directly connected by the emitter wire bonds. On DUT2 in Fig.4.16(b), the IGBT and the anti parallel diode are connected via a metal island of the DCB top layer. Both DUTs are turned off at a current density of $130Acm^2$, which is out of the safe operating area. Fig. 4.18 shows the measurement on DUT1. Oscillations are found at the begin of the dynamic avalanche process. Fig.4.17 shows the result of the measurement on DUT2. Oscillations are found at the end of the dynamic avalanche process. The frequencies are determined as 600MHz on DUT1 and 155MHz on DUT2. The strong influence of the emitter wire bonds is demonstrated.

Typical characters of TA-oscillation

After all the measurements on the 3.3kV IGBT setups, the typical characters of TA-oscillations can be summarized as the following:

- TA-oscillations are found during the IGBT turn-off under high load current and high DC-link voltage and during the dynamic avalanche.
- The gate resistance has to be in a certain range.
- Large stray inductance in the current commutation circuit leads to high over voltage and thus has the similar effect on TA-oscillation as the high DC-link voltage.
- Longer emitter wire bonds leads to lower frequency of TA-oscillations
- TA-oscillations are found at the junction temperature from 25°C to 125°C.

Figure 4.16: (a) Direct wire bonding between IGBT and diode (DUT1)
 (b) Indirect wire bonding between IGBT and diode (DUT2)

Figure 4.17: Measurement on DUT2 with long emitter wire bonds. HF-oscillation at end of the dynamic avalanche, oscillation frequency is about 150MHz

Figure 4.18: Measurement on DUT1 with direct emitter wire bonds. HF-oscillation at the begin of dynamic avalanche, oscillation frequency is about 600MHz

- TA-oscillations are found on all chip types with the thickness from $350\mu m$ to $475\mu m$, and the frequencies of TA-oscillations are found to be the same.
- Paralleling of IGBTs is not necessary for the TA-oscillations. TA-oscillations are found also on setups with a single IGBT and a single diode.
- The frequencies of TA-oscillations are found in a frequency band from 100 to 800MHz for 3.3kv IGBTs. Multi frequencies are found in a single turn-off process.
- The cross wire bond between the adjacent IGBTs is effective for the suppression of the low frequency oscillation but not effective for the high frequency oscillation.
- The oscillation signals on V_{GE} are proved to be EMI coupling. As TA-oscillation happens, the internal gate emitter voltage has already dropped to below the threshold voltage. The gate did not participate actively in TA-oscillation.
- The emission source of the oscillation signals are proven to be the emitter wire bonds.

4.2 Analysis of the resonant circuits

Resonant circuit is necessary to raise the TA-oscillation. Due to the high frequency of the oscillation, the resonant circuit must consist of small inductors and capacitors. And they must be located close to the chips, and the capacitance of the chip itself is expected to participate in the oscillation. Fig.4.19 shows a schematic illustration of the resonance circuits near to the IGBT chip with the participation of emitter wire bondings. Three

Figure 4.19: Resonance circuits ($RC1$, $RC2$ and $RC3$) of the DCB setups shown in Fig.4.16(a)

resonance circuits are constituted of IGBT and diode capacitance, capacitance between DCB areas and the inductance of bond-wires. Resistances in the circuit act as damping factor and have just minimal impact on the resonance frequencies. For the quantitative analysis of complex structures, such as substrate and even the total IGBT modules, simulation tools such as Q3D and HFSS are developed by ANSYS to assist the design analysis using FEM.

4.2.1 Inductance of wire bonds

Due to its reliability and flexibility, wire bondings are the most popular method to realize the internal electrical connections of power modules. The parasitic inductance, electric resistance as well as the thermal resistance are critical parameters that limit the design and application of power modules. The inductance of wire bonding is one of the key parameter for the analysis of the resonant circuit of the TA-oscillations. For the resonant circuit analysis of setup with just one IGBT and one diode, the parasitics

Figure 4.20: Analytical estimating of the wire bonding inductance

can be analytically estimated. Although the analytical estimation is not so precise as the FEM simulations, but it is a practice way to find a start point of the frequency analyse by means of network simulations. The wire bondings in the power modules are always arranged close above a conductor surface such as the DCB-copper layer. One of the emitter wire bonding configuration is shown for example in Fig.4.9. Theoretically for high frequency analysis, the copper layer can be taken als ideal conductor. The electric and the magnet-field caused by a wire bonding over a copper layer is equivalent to the field caused by two wire bondings arranged antisymmetric to the copper surface.(Mirror principle theoretical electro technique) Fig.4.20 shows the method to estimate the inductance of the emitter wire bondings analogous to the air coil.

$$W_{wb} = \frac{1}{2}W_c \quad (4.1)$$

$$W_c = \frac{1}{2}L_{sc}I_c^2 \quad (4.2)$$

$$W_{wb} = \frac{1}{2}L_{wb}I_{wb}^2 \quad (4.3)$$

$$I_c = \frac{1}{N}I_{wb} \quad (4.4)$$

A emitter wire bonding can be firstly approximated by a half circle wire conductor, whose electrical magnet field energy is equal to half of a wire ring. The electro-magnet field energy of the parallel arrangement of the wire bonding W_{wb} is equivalent to the half electro-magnet field energy of the coil W_c , Eq.(4.1) (4.2) and (4.3), with the current ration $1/N$ as expressed in Eq.(4.4). Here N is the winding number of a coil and the number of the parallel wire bondings. I_c is the coil current, this means the current in a single bond wire. And I_{wb} is the total current of the parallel wire bondings.

For long coil with $l \gg r$, as in case of the emitter wire bonding on the substrate as in Fig.4.9, the inductance L_{lc} is calculated as in Eq.(4.5) [MRP05]. For the short coils with $l > 0.6r$, as in case of the wire bonding of a single chip, the inductance L_{sc} is given by Eq.(4.6) [MRP05].

$$L_{lc} = N^2 \frac{\mu_0 \mu_r \pi r^2}{l} \quad (4.5)$$

$$L_{sc} = N^2 \frac{\mu_0 \mu_r \pi r^2}{\sqrt{l^2 + 4r^2}} \quad (4.6)$$

The inductance of the emitter wire bonding can be therefore estimated

Table 4.1: Parasitics extracted for the resonant circuit analysis

parasitics	value	unit
L_B	2.65	nH
C_{EB}	140	pF
C_{CB}	10	pF

analogues to a coil by Eq.(4.7), which is eliminated out of the equations of Eq.(4.1) and (4.6).

$$L_{wb} = \frac{1}{2} \frac{\mu_0 \mu_r \pi r^2}{\sqrt{l^2 + 4r^2}} \quad (4.7)$$

The inductance of the emitter wire bondings on the example substrate setups as shown in Fig. 4.16, is estimated as approximately 2.7nH, which agrees well with the result of the Q3D simulation. Although in the real applications the magnet coupling to the adjacent layout structures must be considered, this analogous inductance estimation is still an effective method to find out a start point for the network analysis of the resonant circuit in IGBT module.

The capacitances of the DCB copper areas can be calculated as conventional plate capacitor according to Eq.(4.8). For the further analysis, the parasitics of the resonant circuits as shown in Fig.4.19 are estimated and listed in Tab.4.1.

$$C_{DCB} = \frac{\epsilon_0 \epsilon_r A}{D} \quad (4.8)$$

4.2.2 Differential capacitance of IGBTs and diodes

As mentioned in Chap.2, the differential capacitance corresponding to the actual SCR can be defined as Eq.(2.9) for the resonant circuit analysis.

The chip capacitances in the resonant circuits as shown in Fig.4.19 are estimated according to the actual width W_{SCR} of the space charge regions in the respective chips.

Although IGBTs are so called voltage controlled semiconductor switches, But their switching processes are indeed current controlled. For the IGBT, the expanding of the SCR and the ramp up of V_{CE} is mainly limited by the net hole density, which is determined by the load current and the ratio between the electron current and hole current in SCR . For the different turn-off conditions, such as different R_G and different I_L the expansion of SCR in the IGBT and the diode can totally differ from each other. By the assumption that the W_{SCR} in the diode is γ times of the W_{SCR} in the IGBT, the chip capacitances and the corresponding resonance frequencies can be calculated as shown in Fig.4.21 for the analytical understanding. Due to the absence of plasma in the diode, γ is generally larger than one. Thereby it is taken into account that chip capacitance reaches its minimum after the SCR reaches the field stop layer, and the active area of the diode is roughly 50% of the IGBT. Higher I_L and smaller R_G lead to faster switching process. The faster is the switching process of the IGBT the larger is γ .

Fig.4.22 shows the diagram of the eigen frequencies of the three resonant circuits, estimated according to the W_{SCRs} in IGBT at $\gamma = 2$, together with the transit time frequencies of the corresponding SCR . The transit time frequencies decrease with the expanding of SCR . The resonance frequencies increase due to the decrease of the chip capacitances given by (2.9). The intersections of group A and group B in the figure represent the frequencies, at which the oscillations can be effectively amplified by the transit time mechanisms. In the frequency range above 1GHz oscillations is theoretically possible, but such oscillations are not yet observed in the real measurements. During the switching off of the channel current, the ratio between the electron current and hole current in the IGBT and

Figure 4.21: Chip capacitance of IGBT and diode according to W_{SCR}

the corresponding γ is not constant. So that the analytical estimation of the resonant frequencies just helps to the understanding of the oscillation mechanisms. The device simulations are indispensable for the fine quantitative analysis of the dynamic IGBT turn off process.

4.3 Device simulations

Compared to measurements, simulations gives more flexible way to analyse the oscillation mechanisms and allow the access to the internal process of the IGBT chips. Similarly to the measurement a 3.3kV field stop IGBT cell with trench gate structure on emitter side is created for the device simulation [Pfi12]. To accelerate the simulation process, just half of cross section of a IGBT cell is constructed as 2D model for the cylinder symmetrical simulation. Doping profiles are adjusted so that the blocking capability, forward bias and short circuit characteristic fit to the measurements on

Figure 4.22: Intersection between eigenfrequencies of the transit time mechanism and the resonant frequencies during the expanding of W_{SCR}

Figure 4.23: Circuit of Device-Simulation, a 2D model of half IGBT cell is integrated in the circuit

the IGBT modules.

4.3.1 Circuit

The rapid development of the computer techniques makes it now possible to solve the semiconductor device model in a spice-similar circuit environment [syn11]. For the oscillation analysis the resonant circuit is just as important as the proper device model of the IGBT chip, therefore the realistic estimation of parasitic components in the module and substrate setup is necessary to find the start point for the TA-oscillation analysis. To reproduce the resonant circuits near the IGBT chips, the parasitics on the DCB setups are extracted analytically and verified by Q3D simulations.

The total circuit of the system simulation is shown in Fig.4.23. The device IGBT model builds first together with the parasitics of the substrate

setups a single IGBT switch (DUT), which consists of three resonant circuits as shown in Fig.4.19. $Ls3$ and $R3$ together with $Ls5$ and $R5$ represent the wire bonding between the IGBT and the diode, and the $Ls6$ and $R6$ represent the wire bonding between the diode and the copper area on the top of the DCB. The series switching of C_{CB} and C_{EB} is simplified as C_{DB1} . And C_{DB} represents the junction capacitance of the diode DB . The DUT is again integrated in the test circuit corresponding to labour setup for the measurement. The entire test process is solved by Sentera device simulator. In contrast to the previous HF-oscillation simulations, the turn-off process is simulated without artificial HF-stimulations. The maximal time step during the IGBT turning-off is limited to 0.2ns, which is found to be a good compromise between the simulation time and the convergence. The time derivative of the current signal on R5 represents the antenna signal in the measurement.

Oscillations are observed on the current of R5 during the whole ramp up process of V_{CE} . The frequencies of the weak oscillation decrease according to the extension of the SCR from 560MHz to 200MHz before the strong oscillation occurs. At the time of MOS-channel turned-off, the strong oscillation occurs with the frequency of around 600Mhz. Oscillations with similar frequency trend can also be observed on the fine resolved V_{CE} and I_{CE} .

4.3.2 Effect of resonant circuit's eigenfrequencies

The varying of C_{DB1} shows a significant effect on the oscillation signals. The frequencies of the oscillations reduce according to the increase of C_{DB1} . The maximal amplitude is found at 610MHz and $C_{DB1}=22\text{pF}$ as shown in Fig.4.25(a). The circuit configurations of the device simulation are analysed by network simulation parallel to the device simulation. As the network simulator does not solve the semiconductor equations, the parameters such capacitance and inductance of the semiconduc-

Figure 4.24: Oscillations are found during the turn-off simulation.

Figure 4.25: Compare the device simulations with the network simulations at $C_{DB1}=10\text{pF}$, 22pF and 40pF

tor chips must be extracted manually according to the actual SCR. For the time point of 600MHz oscillation, at $111.2\mu\text{s}$ in Fig.4.24, the combination of $C_{IGBT} = 100\text{pF}$ and $L_{IGBT} = 1\text{nH}$ is found as the proper configuration by comparing the Simetrix simulation and the device simulations. Three eigenfrequency groups are calculated using SIMetrix as shown in Fig.4.25(b). Group 2, which lies around 600MHz, matches well with the oscillation frequencies of the device simulations.

Further simulations with diverse configurations of C_{DB1} and C_{DB} are processed both in device simulations and in circuit simulations. The diagram in Fig.4.26 shows the device simulations and SIMetrix simulations with various resonant circuit configurations. The amplitude and the frequencies of device simulation show the same tendency as the network simulations. The simulation time and the expense of the network simulation using SIMetrix is significantly lower compared to device simulation. The network simulation is therefore attractive for the module designer to

Figure 4.26: Device simulations show the same trend as the network simulations with the same C_{DB1} and C_{DB} , $C_{IGBT} = 100pF$ and $L_{IGBT} = 1nH$ are used for IGBT circuit model

optimize the module layout in respect to the oscillations as the chip capacitance and the chip inductance are dependent on the SCR in the chip. The combination of $C_{IGBT} = 100pF$ and $L_{IGBT} = 1nH$ is not valid for universally. For the module layout optimization, the parameters of the equivalent chip model need to be extracted for the different work conditions and different time points.

4.3.3 Effect of load current

The effect of load current has been theoretically analysed in Chap.2. The higher load current means the higher carrier densities in the IGBT and smaller R_G leads to faster turn off of the MOS channel. The resulted higher hole current and therefore higher hole density in the space charge region drive the the maximal e-field E_{max} to the critical value E_c at a smaller W_{SCR} . This leads to a small transit time and a higher transit time frequency. The possible oscillations may be raised when the transit time frequencies cross the eigenfrequencies of the resonant circuits, for example near the intersections as shown in Fig.4.22. To verify this theory,

the load current I_L is varied from $18 A/cm^2$ to $181 A/cm^2$.

The oscillation peak is first found at $56 A/cm^2$ with the frequency of 400MHz, then reduces with the increasing I_L and then reaches its second peak amplitude at $109 A/cm^2$ with the frequency of 650MHz. Fig.4.28 shows the expansion of SCR in the IGBT device model at the moment of the oscillations. The oscillation frequencies of the device simulation match well with the measurements. For $I_L = 56 A$, $W_{SCR} = 155 \mu m$ as the 400MHz oscillation happens. For $I_L = 109 A$, $W_{SCR} = 205 \mu m$ as the 650MHz oscillation happens. The eigenfrequencies of the resonant circuit can be read out in Fig.4.22 with indications of work points group A for 400MHz and work points group B for 650MHz oscillations according to the current W_{SCR} . During the turn-off, the expanding of SCR leads to the decrease of transit time frequencies of IGBT and increase of the resonance frequencies on the substrate. The transit time frequencies match the resonance frequencies first time at around 400MHz near the work points group A in Fig.4.22. As the dynamic avalanche does not occur, the 400MHz oscillation is induced mainly by PETF mechanisms. The further expanding of the SCR leads temporarily to the mismatch between the transit time frequencies and the resonance frequencies and decay of the oscillation. For $I_L = 109 A$, as the SCR expands to about $200 \mu m$, the dynamic avalanche happens. The transit time frequencies match the resonance frequencies at 650MHz near the work points group B in Fig.4.22. The frequencies of the simulated oscillations are lower than the analytical estimated frequencies. The possible reason for the deviations between the oscillation frequencies and the estimated resonant frequencies is the time-delay of the avalanche region in IGBT, which is not involved in the simplified resonant circuit model in Fig.4.19. The equivalent circuit of the avalanche region for microwave diode has been derived analytically by Gilden and Hines [GH66]. Although the avalanche region can be described by the parallel switching of an inductance L_a and a capacitance C_a , but they are strong

Figure 4.27: Simulation of the dependency on load current

Figure 4.28: W_{SCR} in the IGBT at the moment of the oscillations of 400MHz and 650MHz

Table 4.2: Simulated oscillations by $I_L = 109A$ with varying of R_G

R_G (Ω)	F_O (MHz)	$IR5_{AC}$ (A)
5	600	9
11	590	0.5
17	705	80
30	680	3

non-linear dependent on the E_{max} . Due to the dynamic changing of E_{max} , it is difficult to find the proper equivalent circuit of the avalanche region analytically for each time during the turn off of IGBTs.

4.3.4 Effect of gate resistance

The gate resistor limits the current in the gate circuit so that the switching off of the MOS channel can be delayed by larger gate resistors. Using a smaller gate resistor the electron current would be turned off faster and would lead to earlier and higher hole current on the emitter side during the IGBT turning-off. Simulation results of varying R_G from 5Ω to 40Ω in Tab.4.2 show that the amplitudes of oscillation are sensitive to R_G during the frequencies are moderately shifted by the different R_G configurations. This phenomenon is also found by the measurements. Under certain conditions, when the transit time frequencies match well with the resonance frequencies, the simulations can not run till the end due the divergence. The maximal amplitude is found as 80A at 705Mhz and $R_G = 17\Omega$, which would be strong enough to destroy the device. In this case, the E-field in IGBT shows a strong deformation compared to a triangle and indicates a TRAPATT (trapped plasma avalanche-triggered transit) similar process.

Figure 4.29: Impact ionization and charge injection during TA-oscillation

4.3.5 Internal process of IGBT chips

Device simulations make it possible to get a view on the internal state of the IGBT chip. Fig.4.29 shows the impact ionization process within one period of the TA-oscillation. During the TA-oscillation, 10 time points of the current in R5 are shown on the top of the figure. The corresponding internal states of IGBT chip are shown in the middle of Fig.4.29, in which the periodical change of the impact ionization can be observed. The integration of the impact ionization over the SCR during TA-oscillation is shown on the bottom of Fig.4.29, which indicates a periodical current injection during the avalanche process.

The e-field distributions in the SCR, corresponding to the time points during the TA-oscillation, are shown in Fig.4.30. The sinusoidal overlapping can be observed on the triangle e-field distributions. The differences of the e-field between the individual time points and the average value of

one oscillation period are defined as Eq.(4.9). In Fig.4.31 a propagating wave of $\Delta E(x, n)$ is clearly visible. As shown in Fig.4.32 and Fig.4.33, the propagating waves are also observed on the electron distribution and hole distribution in the SCR. The wave is induced by the drift of electrons packet from emitter side to collector side with the velocities of 8.56cm/s . And the wave of holes spreads in the reverse direction with the velocity of 7.75cm/s . The velocities comply with the effective saturation velocities of electrons packet and holes packet in the SCR respectively. The variation of the eDensity in the SCR is much stronger than that of the hDensity, which indicates the dominant roll of the electron transit time effect in the TA-oscillation.

$$\Delta E(x, n) = E(x, n) - \overline{E(x, n)}; n = 1, 2, \dots, 11 \quad (4.9)$$

$$\Delta eDensity(x, n) = eDensity(x, n) - \overline{eDensity(x, n)}; n = 1, 2, \dots, 11 \quad (4.10)$$

$$\Delta hDensity(x, n) = hDensity(x, n) - \overline{hDensity(x, n)}; n = 1, 2, \dots, 11 \quad (4.11)$$

4.3.6 Effect of impact ionization

The flexibility of device Simulation allows to answer the question, if the TA-oscillation is solely caused by the avalanche. After that the oscillation of 600Mhz is reproduced by the reference simulation, the simulation is repeated with deactivated impact ionization. Although it is not possible to turn off the impact ionization in real IGBT chips, this comparison is the simple method to verify, if the TA-oscillation is solely caused by the impact ionization. Fig.4.34 shows two simulations with identical parameter settings except of the impact Ionization. On the top left is the turn-

Figure 4.30: e-field distribution in IGBT during TA-oscillation

Figure 4.31: E-field variation during TA-oscillation according to Eq.(4.9)

Figure 4.32: eDensity variation in the SCR of IGBT during TA-oscillation according to Eq.(4.10)

Figure 4.33: hDensity variation in the SCR of IGBT during TA-oscillation according to Eq.(4.11)

Figure 4.34: Simulation of the effect of impact ionization on TA-oscillations

off diagram of the reference simulation, in which the impact ionization is allowed. On the left bottom side is the diagram of the simulation, in which the impact ionization is deactivated. The simulation with deactivated impact ionization shows a 455MHz oscillation with a half of the amplitude of the 600MHz oscillation. The 455MHz oscillation occurs 60ns earlier and its duration is three time longer than that of the 600MHz oscillation. All of these indicate that the impact ionization can not be the single cause of TA-oscillations. Additionally to the IMPATT, the PETT mechanism is also involved in the TA-oscillation because of the remaining plasma behind the SCR. Due to the poor efficiency of the PETT mechanism, without IMPATT it need more time to raise the oscillation.

Figure 4.35: Mechanism of TA-oscillation mechanisms

4.4 Summarized explanation of TA-oscillations

Fig.4.35 represents the one-dimensional transit time structure during IGBT turn-off. Under forward conduction of the IGBT, the base region is flooded with electrons and holes. After turning-off the MOS-channel, most of the current must be carried by the holes instead of electrons. At that moment, the electron-density decreases and the net hole-density near to the emitter side increases according to the current-share of holes. This results in a net positive charge density (hole-density minus electron-density) near the MOS channel and increases the gradient of electrical field according to the Poisson equation. Due to the high current, the net hole-density is much higher than the doping level. So that, at a V_{CE} far below the block capability, E_{max} has reached the critical value E_C and begins to generate electron-hole pairs at the emitter-side. The so called dynamic avalanche occurs. The electrons generated by avalanche drift through the SCR under the influence of the strong electric field and reach to the anode side.

Similar to the IMPATT mechanism the transit-time effect hereby gives an oscillation trend to the IGBT chip. The injection-time together with the transit-time of the holes induces a phase shift between the terminal voltage and terminal current in RF-range. At certain frequencies, the IGBT can act as an amplifier. When the addressable resonant circuits are present, oscillations would be raised. Due to the high current, the plasma in the base region is quickly extracted and the *SCR* is expanding fast to the collector. The expanding of the SCR during the turn-off process disperses the corresponding frequencies of the transit time structure from single frequencies to a continuous frequency band depending on the actual width of the depletion region (W_{SCR}).

The parasitic components in IGBT modules, for example stray inductance of bond wires and junction capacitance of the free wheel diode can together with the IGBT chip build up resonance circuits. The extension of *SCR* leads to on one hand the longer transit time of the carriers, on the other hand leads to the reduction of the chip capacitance and therefore the higher eigenfrequencies of the according resonant circuits. If the eigenfrequencies of such resonance circuits match the frequency of the transient-time effect, the Transient Avalanche Oscillation would be raised. Oscillations would occur around the intersections of the eigenfrequencies of the resonant circuit and the efficiency of the amplifier being high.

TA-oscillations are successfully reproduced by device simulations using a two dimension IGBT model which is integrated in a resonant circuit similar to the module setups. Eigenfrequency of the resonant circuit fits the second maximum of IMPATT efficiency resulting in an oscillation around 600MHz. The internal semiconductor states show that the e-field and carriers distributions in the space charge region are overlapped by space waves. The velocities of the waves are appropriate to the corresponding carrier velocities. The IMPATT mechanism is the key player in the TA-oscillation. And additionally, the remaining plasma behind the SCR allows also for

the plasma extraction transition time (BARITT/PETT) mechanism.

Although TA-oscillations can be raised theoretically at each voltage class of IGBTs. TA-oscillations are until now only observed by IGBTs with over 3.3kV blocking voltage. The device simulations in this work are based on a 2D half IGBT cell. The filament effects due to the current commutation between the adjacent cell and the paralleled chips can therefore not be investigated.

Chapter 5

Countermeasures to Transient Avalanche oscillations

5.1 Essential measures to suppress TA-oscillations

According to the measurements and analysis in the previous chapters, the dynamic avalanche generation and the transit time effect of the avalanche electrons are identified as the root cause of the TA-oscillation. The high load current density and the switching off of the electron current from the MOS-channel are the key character of the turn off process, in which TA-oscillation occurs. The corresponding general possible measures to avoid the TA-oscillations can be summarised as following:

- To apply limitation to the load current and the DC-link voltage so that the dynamic avalanche does not occur.

- To keep the gate resistor out of the range, in which the TA-oscillation would be raised.
- To avoid the high over voltage by low inductive module design so as to suppress the dynamic avalanche process. This is because that the large stray inductance in the current commutation circuit leads to high over voltage and thus has the similar effect on TA-oscillation as the high DC-link voltage.
- To increase the stray inductance in the resonant circuit for pushing down the resonant frequencies so that they can not be addressed by the transit time effect of the IGBT. Simply, longer emitter wire bonds lead to lower frequency of TA-oscillations. But the high stray inductance would lead to the high over voltage, that affects the module performance negatively.
- For the investigated 3.3kV IGBT chip, the TA-oscillations can be avoided by setting up the resonant circuits, so that the resonant frequencies are out of the frequency range from 100MHz to 800MHz. As the eigenfrequencies and the resonant frequencies change dynamically during the turning off, it is difficult to mismatch the frequencies all the turn-off time. The cross wire bonding between the adjacent IGBTs is effective for the suppression of the low frequency oscillation but not effective for the high frequency TA-oscillation.
- To apply additional electron current from the channel, during the time that TA-oscillation normally happens, is proved to be an effective method against the TA-oscillation. The absence of electrons leads to the high e-field peak near to the emitter side. When the additional electron current is applied from the MOS channel to the SCR, near to the emitter side the derivative dE/dX of the E-field could be reduced and therefore the peak value of the e-field and accordingly the avalanche generation is suppressed. The timing of

the electron injection should be shortly before the begin of the TA-oscillation. The injection should be short enough that the IGBT can be still surely turned off. This method has been realized on a driver concept, which I introduce as next.

5.2 Driver concept to suppress TA-oscillations - pace maker

As already mentioned in Chap.4.4, the high total current density and the fast switching off of the channel current lead to the dynamic avalanche during the turning off of the IGBT. On the system level, the load current and the gate resistance can be limited in a certain range, so that the TA-oscillation does not happen in the SOA (Safe Operation Area). However, methods to suppress and to avoid TA-oscillation with out restriction of the applications are desired. In this work a driver concept - pace maker [Hon13] is developed to get oscillation free switching performance of IGBT out of the conventional SOA. The basic idea of this driver concept is to apply additional electron current from MOS channel to the SCR during the time that TA-oscillation were found. The additional electrons reduce the net hole density and the gradient of the e-field in the SCR according to Poisson's equation. The maximal e-field and the corresponding avalanche process near to the emitter would be suppressed by the additional channel current. The additional channel current pulse during the tail current has been introduced to suppress the PETT oscillation [Mou01b].

For the conventional IGBT turn off, the driver voltage is pulled down from +15V to -15V, and the gate current flows out of the gate to the driver. The gate is therefore discharged. Fig.5.1 shows the pace maker driver concept. After the normal turn off and short time before the dynamic avalanche occurs, an additional positive gate pulse for a few hundreds of *ns* is applied to the IGBT so that the gate current is forced to change

its direction. And the gate is charged again for the short time before it is finally pulled down to -15V . A short electron current pulse is injected from the MOS channel to the SCR. As the pulse is short enough, it is not able to turn-on the IGBT completely, but just opening the channel to some extent that dynamic avalanche does not occur. The start condition of the additional pulse can be defined as Eq.5.1. The V_{CE} is compared with the DC-link voltage during the ramp up of V_{CE} . The additional gate pulse is triggered as the V_{CE} exceeds the certain ratio of V_{DC} . α and β are adjusted by the voltage dividers as shown in Fig.5.2.

Figure 5.1: Driver concept pace maker

$$\alpha V_{CE} \geq \beta V_{DC} \quad (5.1)$$

$$\alpha = \frac{R4}{R3 + R4}$$

$$\beta = \frac{R2}{R1 + R2}$$

Figure 5.2: Voltage divider on pace maker to define the condition of the additional gate pulse

The additional channel current leads to the reduction of the E_{max} and causes a reduction of the miller capacitance. The reduction of miller capacitance leads to a negative gate current, which tends to turn off the

Figure 5.3: Driver concept pace maker

channel again. This so called self turn off effect [Boel14] [Bas14] must be compensated by the pace maker in order to establish the channel current. Measurements to verify the driver concept are processed with the driver board as shown in Fig.5.3. The reference measurement with the additional pulse is shown on the left side of Fig.5.4, on which the TA-oscillation can be observed in the antenna signal. The measurement is repeated with the additional gate pulse and the result can be seen on the right side of Fig.5.4. The TA-oscillation is totally suppressed by the pace maker.

Table 5.1: comparison of the turn off energy E_{off} of the conventional driver with pace maker at 25°C, $I_L = 1500A$ and $V_{CE} = 1800V$ with varying of R_G

R_G (Ω)	E_{off} (mJ) conventional driver	E_{off} (mJ) pace maker
2.7	2144	2240
1.5	not allowed	1907
1.3	not allowed	1848

Figure 5.4: Improvement using driver concept pace maker

In the practice, the essential asymmetry and inhomogeneity in the module- and chip-layout is inevitable. Especially during the dynamic process, such as IGBT turn off, such asymmetry and inhomogeneity have generally negative effect on the module performance. Fig.5.5 shows schematically the possible asymmetry within an IGBT module and also within a IGBT chip. During turn off, the gate current flows from gate to the driver. Due to the asymmetry of the parasitic resistance and inductance in the gate circuits, dynamically the gate potential would be different between the chips in a module or between the cells within a chip. This would lead to the asymmetry of the current distribution in the IGBT module and even within a IGBT chip. The pacemaker driver forces a current direction change in the gate circuit. This compensates partially the possible asymmetrical current distribution due to any existing asymmetrical module and chip construction. Besides the suppression of TA-oscillation, the pacemaker concept makes the IGBT even more robust. Generally IGBT is just allowed to be turned off up to two times of the nominal current. If the I_{CE} exceeds this limit, short circuit should be imposed before the IGBT is turned off to avoid the destruction [HB97]. Using the pace maker driver, the 3.3kV 1500A test module can be successful switched off at 5000A. And

no limitation of gate resistor is needed any more. Compared to the conventional driver concept the switching energy can also be reduced. For the test module, to avoid the TA-oscillation 2.7Ω is defined as the lower limit of the SOA. As shown in Tab.5.1, the turn off energy at 2.7Ω and of the pace maker driver is $2240mJ$, which is 4.5% slightly higher than the $2144mJ$ of the conventional driver. But with the pace maker oscillation free turn off can be also realized at $R_G = 1.3\Omega$. The E_{off} of $1848mJ$ is then 13.8% lower than standard.

Figure 5.5: Asymmetry of IGBT modules and chips

Chapter 6

Summary and outlook

6.1 Summary

Additionally to the PETT oscillation [GMS02] and dynamic IMPATT [LSG98] oscillation, TA-oscillation is found to be a new type HF-oscillation in IGBT modules. TA-oscillations are observed during the raising of the collector voltage and the falling of the collector current, as the dynamic avalanche is ongoing. TA-oscillation is much more dynamic compared to the PETT oscillation and dynamic IMPATT oscillation in diode. The measurements in this thesis show that the gate resistance, the DC-link voltage and the load current are the essential parameter for the TA-oscillation. Although the TA-oscillation is first observed on the gate voltage trace, no active participation of gate can be demonstrated. The HF-oscillation signal on the gate trace is caused by HF-current in the emitter wire bonding through electromagnetic coupling. The TA-oscillation appears just in a certain range of gate resistor while the product of load current and DC-link voltage is significant of set of TA-oscillation, that indicates the correlation of the TA-oscillation to the maximal e-field of the space charge region according to Eq.2.13. And the frequency of the TA-oscillation is found

between 100 and 800MHz, that indicates the effect from carrier transit time.

The explanation of the TA-oscillation can be expressed as follows:

Under forward conduction of the IGBT, the base region is flooded with electrons and holes. Due to turn off of the channel, the total load current must be carried by holes. A space charge layer is building up on the emitter side. Due to the high current, the net hole-density is much higher than the doping level. Thus, dynamic avalanche occurs at collector voltage far below the static blocking voltage of the IGBT. Electrons generated by avalanche drift through the space charge region to the collector side similar to the IMPATT diode. The transit time effect causes the chip to act as a high-frequency amplifier. Additionally, the remaining plasma behind the collector side allows also for the plasma extraction transition time (BARITT/PETT) mechanism.

The carrier injection time and the transit time lead to the phase shift between the terminal HF-voltage and the HF-current. According to the conventional consideration, for the IMPATT diode the maximal HF-power generation occurs at the transit angle of π for IMPATT diode and $3/2\pi$ for the BARITT diode [Sze81] [EH98]. Till now such consideration is also applied to the frequency estimation of PETT oscillation and dynamic IMPATT oscillations [GMS02] [LSG98] [MD10] [Hee14]. And no consideration are taken for the transit angle above 2π . With this theory, for a $200\mu\text{m}$ SCR the transit time frequency can be estimated as around 170MHz for IMPATT and 240MHz for PETT/ BARITT mechanisms. This is much lower than the measured oscillation frequencies in this thesis, i.e. 650MHz. As the the power semiconductor chip thickness is not optimized for the RF-power generation, the none ideal situation should also be considered. For the understanding of the amplifier mechanism, the analytical estimation of the HF-energy conversion efficiency of transit time diode of [EH98] is extended in this thesis. The analysis in Chap.3 shows that additional to

the optimal transit angle, RF-power generation is also quite possible at the transit angle about 2.9π for IMPATT mechanism. Compared with the conventional consideration that corresponds to three times higher IMPATT frequency. The frequency for maximal power generation efficiency is reversely proportional to the width of the space charge region. The expanding of the SCR during turning off of IGBT leads to decreasing of the transit time frequencies. Not only the amplification but also the resonant circuit varies with the expanding of the space charge region. In the IGBT module the resonant circuit is formed out of chip and substrate capacitance as well as the inductance of the wire bonds from IGBT to the diode and diode to the substrate. The diode capacitance is also part of the resonant circuit. The chip capacitance is reversely proportional to the width of the space charge region. During the turning off, the eigenfrequencies of the resonant circuit rise with the expansion of the space charge region. Oscillations would occur around the intersections of the eigenfrequencies of the resonant circuit and the transit time frequencies, that the efficiency of the amplifier being high. Eigenfrequency of the resonant circuit fits with the second maximum of IMPATT efficiency and results in an oscillation around 600MHz.

To reproduce the TA-oscillation, A two dimensional device model of 3.3kV trench field stop IGBT was setup and integrated in the resonant circuit similar to a IGBT module. The device model and the resonant circuit was again integrated in the test circuit similar to the measurement setup. In contrast to the PETT simulation, TA-oscillations are successfully reproduced without artificial stimulations. The influence of the parameters, such as load current, gate resistance and emitter wire bonds, are verified by the device simulation. The impact ionization shows a periodical process, according to the sine wave of the oscillation. The electrical field has a sinusoidal shape overlapping a triangle, determined by the space charge. Similar wave forms are also found on the electron-distribution and hole-

distribution in space Charge Region. IMPATT-like mechanism plays the key role in the TA-oscillation. The simulation with deactivated impact ionization shows that in addition to the IMPATT, the PETT mechanism can be also involved in the TA-oscillation.

All the analysis shows that the temporary absence of the electron on the emitter side causes the high E-field peak on the emitter side and leads to TA-oscillation. In this thesis, a driver concept "pace maker" is developed to open the channel again during the time the TA-oscillation normally happens. As the pulse is short enough, it is not able to turn-on the IGBT completely, but just opening the channel to some extent that dynamic avalanche does not occur. Supplying electron current during critical period of time is verified to be an effective method to suppression of dynamic avalanche and the TA-oscillation. The pace maker makes the IGBT even more robust because of the compensation of the possible asymmetry in the IGBT module.

6.2 Outlook

The one dimension model is used to analyse the RF-power generation efficiency of IMPATT structures analytically. And a narrow avalanche region assumed to inject charge carriers into the SCR with a delay of π . During the turn off process of IGBT, the E-field distribution in the SCR changes quickly and this assumption is not always in accordance with the reality and leads to deviation between the frequency estimation and the measurement. Theoretical work can be done to develop a more suitable analytical model of the avalanche region during the dynamic changing of E-field distribution. Although the avalanche generation and the electron transit time are dominant in TA-oscillation, transit time effect of holes is also found. Further work can be done to investigate the interaction between electrons and holes during TA-oscillations.

Further research work can be done on the module and chip design of IGBT to improve the dynamic performance. Generally, there are always inevitable asymmetry and inhomogeneities in the IGBT Module, IGBT chip and also within a single cell of the chip. In the conventional design, the IGBT cells are constructed to get a uniform characteristic. Accordingly, the current should be distributed homogeneously. But in reality, a small non-homogeneity and unbalance is unavoidable. And it is generally known that, the slight unavoidable non-homogeneity between and within the IGBT cells may lead to current filaments during the turn off processes of IGBTs [MD10]. This would happen especially under the avalanche condition. The kink on the V_{CE} trace in Fig.4.14 is a sign of current filaments. In the current filaments the local current density can reach up to ten times the average value. Such filaments therefore enhance the transient avalanche dramatically, especially in IGBT for high voltage and high current application, such transient avalanche in IGBTs can become strong enough, so that TA-oscillation takes place and raises EMI problem. Reduction of the density of the current filaments would lead to improved utilization of the IGBT volume. This leads to the idea to reduce the current filament by integration of controlled inhomogeneities into the IGBT chip and into the single cell designs. Further research work can be suggested as follows [Hon14]:

- To investigate the IGBT chip design by arrangement of different (more than two) types of cells within an IGBT chip.

Generally, on the IGBT chip there are large number of IGBT cells with the same cell type implemented. Due to the unavoidable manufacturing tolerances and the asymmetry of the cell arrangement on the chip the switching behaviour of the cells may slightly differ from each other. Since the distribution of such differences cannot be controlled, such minimal difference can lead to strong current filaments. In an IGBT chip, cells can be implemented with differ-

ent channel widths; different channel contacts, different cell shapes and sizes as well as different doping levels. To generate the non-homogeneity on IGBT chips by regular or irregular arrangement of various of cell types. Non-homogeneity of the order of regular arrays of various cells of irregular shapes. The target is to suppress the current filament by chip design from the uncontrolled non-homogeneity to a controlled non-homogeneity.

- To investigate the IGBT cell design to integrate inhomogeneity to the gate channel within a IGBT cell.

The lateral non-homogeneity can be realized by deformation of trench gate oxide. Due to the inhomogeneous field distribution in the channel region, during turn-off and before complete channel pinch-off, the channel current can be divided to several sub-streams. The limitation of the current commutation between the adjacent region with in the cell would suppress the stronger filaments and thus the followed avalanche at channel pinch-off.

The trench bottom can also be implemented as wave form or step form, so that the charge commutation from the p-float to the channel region can be homogeneously divided. This would help to suppress TA-oscillations. And there must be other constructions, such as jagged structure, can be introduced to get such effects.

Appendix A

Derivation of transit time efficiency

Here with is the derivation of amplifier efficiency of transit time structures (IMPATT and BARITT). The injection current is consist of three linear functions, which is indeed afflicted with theoretical error. But in case of short injection time this error trend to zero.

$$P_{DC} = J_{DC}V_{DC} = \frac{A}{2\pi}J_{max}\Theta_D V_{DC}$$

$$P_{RF} = -\frac{A}{2\pi} \int_{\Theta_M - \Theta_W/2}^{\Theta_D + \Theta_M + \Theta_W/2} J_{ind}(\omega t) V_{RF} \sin(\omega t) d(\omega t)$$

$$= -\frac{A}{2\pi} (P1 + P2 + P3)$$

Figure A.1: Diagram of voltage and current in a transit time structure with the simplified charge injection

$$\begin{aligned}
P1 &= \int_{\Theta_M - \Theta_W/2}^{\Theta_M + \Theta_W/2} J_{ind}(\omega t) V_{RF} \sin(\omega t) d(\omega t) \\
&= -J_{max} V_{RF} \cos(\Theta_M + \Theta_W/2) + 2 \frac{J_{max} V_{RF}}{\Theta_W} \cos(\Theta_M) \sin\left(\frac{\Theta_W}{2}\right)
\end{aligned}$$

$$\begin{aligned}
P2 &= \int_{\Theta_M + \Theta_W/2}^{\Theta_M + \Theta_D - \Theta_W/2} J_{ind}(\omega t) V_{RF} \sin(\omega t) d(\omega t) \\
&= J_{max} V_{RF} (\cos(\Theta_M + \Theta_W/2) - \cos(\Theta_M + \Theta_D - \Theta_W/2))
\end{aligned}$$

$$\begin{aligned}
P3 &= \int_{\Theta_M + \Theta_D - \Theta_W/2}^{\Theta_M + \Theta_D + \Theta_W/2} J_{ind}(\omega t) V_{RF} \sin(\omega t) d(\omega t) \\
&= J_{max} V_{RF} \cos(\Theta_M + \Theta_D - \Theta_W/2) \\
&\quad + \frac{J_{max} V_{RF}}{\Theta_W} (\sin(\Theta_M + \Theta_D + \Theta_W/2) - \sin(\Theta_M + \Theta_D - \Theta_W/2))
\end{aligned}$$

$$P_{RF} = -\frac{A}{2\pi} J_{max} V_{RF} \frac{\sin(\Theta_W/2)}{\Theta_W/2} (\cos(\Theta_M + \Theta_D) - \cos(\Theta_M))$$

$$\eta = \frac{P_{RF}}{P_{DC}} = -\frac{V_{RF}}{V_{DC}} \frac{\sin\left(\frac{\Theta_W}{2}\right)}{\frac{\Theta_W}{2}} \frac{\cos(\Theta_M + \Theta_D) - \cos \Theta_M}{\Theta_D}$$

$$\eta_1(\Theta_W) = \frac{\sin(\frac{\Theta_W}{2})}{\frac{\Theta_W}{2}}$$

$$\eta_2(\Theta_D) = \frac{\cos(\Theta_M + \Theta_D) - \cos \Theta_M}{\Theta_D}$$

Appendix B

Test box and the RF-noise problems

At begin of the investigation on TA-oscillation, HF-disturbance at about 600MHz is some time observed on the trace of load current, as shown in Fig.B.1. But the load current is measured by a Person coil, whose response frequency is under 30MHz. The capacitive coupling in this case is responsible for disturbance. Fig.B.2 shows the schematic of the test setup. The impedance of the GND conductor between the test box and the oscilloscope, the impedance of the GND conductor of the BNC cabel between Person current sense and the coupling capacitance of Persen current sense to the test box built a closed loop, that capture the HF-oscillation signal similarly to an antenna. The HF-signal measured at the oscilloscope is the voltage drop on the BNC cable of the Person current sensor according to the voltage divider as shown in Fig.B.2, but not the current signal. This effect can be eliminated by enlarging the impedance of the GND conductor between the test box and the oscilloscope.

Figure B.1: oscillation signal observed on the load current

C_P : capacitive coupling between Person current sensor and test box

Z_{GND} : impedance of the GND conductor between test box and oscilloscope

Z_{BNC} : impedance of the BNC cable from Person current sensor to oscilloscope

Figure B.2: Schematic of capacitive coupling and the voltage divider in the test setup

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