Ballistic Electron Transport in Nanoscale Three-Branch Junctions

by

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The author was born in Shizuoka, Japan, on Aug 9th, 1979. He attended the University of Tokyo in Japan from 1998 to 2005, graduating with a Bachelor of Engineering in 2003 and a Master of Engineering in 2005, both in Materials Engineering. In the summer of 2005, he enrolled in the Ph.D program in the Department of Electrical and Computer Engineering at the University of Rochester. He began his research in the Superconducting Digital Electronics Group under the late Professor Marc J. Feldman, where he developed an unshunted Josephson junction comparator comprised of the superconducting Single-Flux Quantum circuitry. After the death of Professor Feldman, he joined Professor Roman Sobolewski’s Group at the Laboratory for Laser Energetics to conduct the thesis research on development of room temperature ballistic devices. His research interests include high-speed electron devices made of two-dimensional electron gas and quantum information processing in superconducting electronics.
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Abstract

Presented here is an experimental study on a novel electron device utilizing ballistic electron transport. This device is a three-terminal structure comprised of lithographically defined Y-shaped two-dimensional electron gas (2DEG) in a compound semiconductor heterostructure. Ballistic electron transport causes a nonlinear input-output transfer curve, which can be exploited for signal rectification, frequency multiplication, and logic gate function. Device fabrication technique using electron beam lithography and a carbon-hard-mask was developed, in order to reliably fabricate ~100-nm-wide 2DEG wires. Direct current measurements while changing the device length and operating temperature revealed the role of ballistic transport in the nonlinear behavior and elucidated that the intervalley transfer mechanism took over the nonlinear behavior under the high electric field. Electrical response at terahertz frequencies was studied by ultrafast time-domain analysis based on an electro-optic sampling and sub-picosecond pulse generation using a photoconductive switch. The terahertz measurement validated the superior high-speed performance of this device as a consequence of small internal capacitance. This research opens up a way to realize various ballistic devices with room-temperature operation and terahertz-bandwidth performance.
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Foreword

The author performed all experimental procedures in this thesis. Other contributions from colleagues are listed below:

Chapter 3: The carbon-hard-mask process was developed in collaboration with Dr. Quentin Diduck (Department of Electrical and Computer Engineering, University of Rochester).

Chapter 4: I am the primary author of this chapter, which is based on the material published in the Applied Physics Letters. Data analysis and discussion have been performed with co-authors of the paper: Roman Sobolewski, late Marc J. Feldman, and Martin Margala (Department of Electrical and Computer Engineering, University of Massachusetts Lowell).

Chapter 5: I am the primary author of this chapter. Results of a double TBJ rectifier in Section 5.3 have been published in the Journal of Applied Physics. Data analysis and discussion have been performed with co-author of the paper: Roman Sobolewski.
1. Introduction

1.1 Ballistic electron transport at room temperature

When the device dimensions become smaller than the electron mean free path $l_e$, electrons travels unimpeded in those devices, or the electron’s motion is governed by ballistic transport.\(^1\) Recent advancements in nanofabrication and high-quality semiconductor growth techniques have enabled researchers to observe such ballistic transport even at room temperature.\(^2\),\(^3\) This section overviews the two technological advancements as well as past research on room temperature ballistic transport.

Nanofabrication techniques have been developed for faster and denser electronic integrated circuits. Among various lithography techniques, the electron beam lithography (EBL)\(^4\),\(^5\) is generally accepted to have the highest practical resolution capability to date. EBL scans a focused electron beam across a surface covered by a resist to draw patterns, and the patterns are transferred to the resist by selectively removing either exposed or non-exposed regions of the resist. The main advantage of EBL is that the minimum feature size is not limited by the diffraction limit as with photolithography. The spot size of a focused electron beam can approach to a sub-nm by electron beam optics.\(^4\) The ultimate resolution of EBL, however, is not set by the resolution of electron optical system but by the resolution of electron scattering processes in the resist and/or by
the subsequent fabrication steps, e.g., etching. The minimum line spacing is typically limited to $\sim 20$ nm for polymethyl methacrylate (PMMA) positive resists, but smaller features can be patterned for sparse objects as small as 3 nm.

Another key element for room temperature ballistic transport is long $l_e$. Two-dimensional electron gas (2DEG) formed in III-V compound semiconductor heterostructures offers a long $l_e$. The $l_e$ is defined as the average traveling distance of electrons between successive scattering events and can be written as the product of the Fermi velocity $v_f$ and mean relaxation time $\tau$. For the 2D system, $v_f$ and $\tau$ can be written as $v_f = \frac{h k_f}{2 \pi m^*} = \frac{h}{m^* \sqrt{n/2\pi}}$ and $\tau = \frac{m^* \mu}{e}$, where $h$ is Planck's constant, $k_f$ is the Fermi wave vector, $m^*$ is the effective mass, $n$ is the electron concentration, and $\mu$ and $e$ represent the electron mobility and elementary charge, respectively. Thus, $l_e$ can be described as:

$$l_e = \frac{h \mu}{e} \sqrt{\frac{n}{2\pi}},$$

(1.1)
saying that $l_e$ is proportional to $\mu$ and square root of $n$. The invention of molecular beam epitaxy (MBE) has enabled us to grow high-quality crystal growth for two-dimensional electron gas, which dramatically improved both $\mu$ and $n$. The large $\mu$ is achieved primarily by separating the conductive 2DEG from ionized impurities used for doping. Other techniques such as material selection with lighter $m^*$, band alignment engineering, and mechanical strain by a lattice mismatch have been incorporated to enhance both $\mu$ and $n$, thanks to the flexibility of the MBE growth process. The 2DEG in InGaAs/InAlAs heterostructures, one of the most common material systems
used for commercial high-speed electronics, has a $l_e$ of $\sim$370 nm at room temperature.\textsuperscript{9} Continuous efforts for a further enhancement of $l_e$ are made by seeking a new compound semiconductor system or looking into exotic materials such as graphene sheets.\textsuperscript{10}

From the overview above, it is evident that we can fabricate structures with dimensions smaller than $l_e$ at room temperature, where the ballistic transport governs the electron transport. The first demonstration of room temperature ballistic transport was done using four-terminal cross-junction devices made of InGaAs-based 2DEG.\textsuperscript{2} The authors investigated a temperature dependence of a representative ballistic effect, so-called negative bend resistance, in a wide range of temperatures from 1.5 K to 290 K. The results showed not only that the ballistic effect can survive at room temperature but also that thermal broadening of electron energy enhances the ballistic nature of the system at high temperatures as compared to the estimate based on average electron energy.\textsuperscript{2} Those findings clearly demonstrated the robustness of the ballistic transport over a wide range of temperatures.

### 1.2 Novel electron devices utilizing ballistic transport

The demonstrations of ballistic transport at room temperature have naturally driven some attempts to create novel electron devices utilizing the ballistic effect for better performance and/or new functionalities.\textsuperscript{3,11} Thanks to the unique nature of ballistic transport, in which the electron travels unimpeded, it is expected that a ballistic device could possess a novel principle of operation and exhibits some advantages over conven-
tional electron devices; e.g., CMOS. As an example of such ballistic devices, the ballistic rectifier was proposed by Song et al.\textsuperscript{12} The ballistic rectifier exploits the fact that the electron travels straight until reaching a geometrical border where it undergoes a specular reflection. Figure 1.1(a) presents the device structure. Rectification of electron flow is carried out by a downward deflection at the triangular barrier. In addition to such a unique working principle, it was claimed that the ballistic rectifier can work at unprecedented high-frequencies of $\sim 1$ THz\textsuperscript{13} with low power consumption as a consequence of ballistic electron transport. Following Song’s work, our team at the University of Rochester proposed the ballistic deflection transistor (BDT).\textsuperscript{14} BDT has the same geometry as the ballistic rectifier except for two additional gates on both sides of an electron injection channel, as can be seen in Fig. 1.1(b). Depending on the direction of the electric field created between the gate electrodes, the gates actively switch the path of electron flow to either right or left drains. This structure adds an additional feature such as gain, while maintaining the high-speed and low-power characteristics as with the ballistic rectifier. So far, transistor operation with a reasonable gain has been demonstrated by direct current (DC) measurements.\textsuperscript{15}

Realization of ballistic devices that can operate in a THz-frequency band is expected to have a significant impact, because there is a wide variety of emerging applications using the electromagnetic spectrum from 0.3 to 20 THz.\textsuperscript{16–18} The first example of such applications is field-effect transistors (FETs) with a cutoff frequency exceeding 1 THz to sustain the evolutionary development of the integrated circuit. A data transmis-
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(a) Ballistic rectifier

(b) Ballistic deflection transistor

(c) Three-branch junction

Fig. 1.1: Examples of ballistic devices (a) ballistic rectifier, (b) ballistic deflection transistor (BDT), and (c) three-branch junction (TBJ).
sion system using optical carriers requires various kinds of digital/analog electronics with THz bandwidth, e.g., amplifiers, multiplexers, digital-to-analog converters, and analog-to-digital converters).\textsuperscript{19} Although the most advanced FETs have reached a cutoff frequency of $\sim 1$ THz,\textsuperscript{20} further drastic enhancement of their speed seems to be practically and fundamentally difficult.\textsuperscript{21} Another important application is a so-called free-space THz technology.\textsuperscript{17} The free-space THz electromagnetic wave, considered as an optical wave rather than electronic signal, has been intensely studied in the area of novel chemical detection schemes, medical imaging, radio astronomy, etc. The key technology for this application is the development of compact solid-state THz detectors and emitters. Novel electron devices, such as plasmon-resonance transistors\textsuperscript{22} and self switching-diodes,\textsuperscript{23} have been considered, yet satisfactory performance has not been achieved. In all the cases above, the THz frequency operation of ballistic devices is expected to be the solution for those emerging applications.

There are many problems to be solved regarding the room temperature ballistic devices. Especially, the theoretical framework of both the ballistic device itself and the circuit consisting of ballistic devices is immature due to its novelty. Despite the vast and detailed studies of ballistic transport in the linear regime\textsuperscript{1} in which the system is at low temperature and under low applied bias, there have been only several works devoted to ballistic transport in the nonlinear regime, where the system is at high temperature and under high bias\textsuperscript{23,24} (e.g., at room temperature and under $\sim 1$ V operating voltage). Because of the above operating conditions, any practical device would be operated in
the nonlinear regime. Thus, understanding ballistic transport in the nonlinear regime is indispensable in terms of realizing room temperature ballistic devices. Also, since the ballistic device would be based on a novel principle of operation, the method of integrating them to make a large circuit should be reinvestigated to receive the full benefit of the ballistic effect.

1.3 Scope of the thesis

This thesis focuses on the three-branch junction (TBJ)\textsuperscript{25–38} which is one of the most studied ballistic devices [Fig. 1.1(c)]. The TBJ is a three-terminal device with a Y-shape conductor made of 2DEG. The TBJ exhibits a nonlinear input-output transfer function originated from ballistic transport.\textsuperscript{25} Thanks to the nonlinear response, it can perform signal rectification,\textsuperscript{26,29,30} frequency multiplication,\textsuperscript{36} and even signal amplification like a transistor.\textsuperscript{37} The details of the device structure and operation modes will be discussed in the next chapter. Here, some unique features of TBJs as a room temperature ballistic device are mentioned: first and the most important, the intrinsic capacitance of the device is extremely small. This feature implies that the operating speed is not limited by resistance-capacitance (RC) time delay, which usually determines the cut-off frequency of most electron devices. Monte Carlo simulation predicts that the TBJ is able to work at frequencies above 1 THz,\textsuperscript{38} although there is no experimental verification yet. Second, the nonlinear response persists in a wide range of temperatures.\textsuperscript{27,31} Several
groups have reported its room temperature operation. For the last, high density of integration is possible because of the inherently small area and simple structure.

Apart from such nice features, the physical mechanism of the nonlinear response is not well understood. So far, some models applicable only in the ballistic limit exist,\textsuperscript{25} which still leave the question why the TBJ is robust over a wide range of temperature. Thus, it is of importance to investigate the nonlinear response when the transport mode becomes quasi-ballistic or diffusive, considering the practical operating condition (e.g., at room temperature and under considerable bias). Since the device functionalities are based on the nonlinear response, elucidating its physical origin is important in terms of modeling the electrical behavior, optimizing its device structure, and assessing the ultimate limit of performance. Also, as mentioned above, the physics of nonlinear ballistic transport still remains an unexplored subject compared to that of linear ballistic transport. Investigating nonlinear ballistic transport is interesting in its own right.

The specific scope of this thesis can be categorized into two parts: first, the mechanism of nonlinear response of the TBJ is explored. This part will give clear insights into the TBJ’s device physics as well as fundamental aspects of nonlinear ballistic transport in 2DEG. Second, the ultrafast phenomena in the TBJ are studied to verify that the TBJ is indeed able to operate in the THz-frequency regime. Since the THz-frequency lies above the frequency range that conventional electronics cover, developing a test system with a THz bandwidth must be the first problem to solve. The knowledge gained
through this research should carry forward not only the development of novel TBJ devices, but also the realization of room temperature ballistic devices.

1.4 Thesis overview

The thesis material is divided into 6 chapters.

Chapter 2 discusses the device physics of the TBJ. First, typical device geometry and operation mode of the TBJ are described along with its advantages over the conventional electron devices. Next, theoretical modeling is performed, based on the Landauer–Büttiker formalism and intervalley electron transfer mechanism.

Chapter 3 describes the experimental techniques used in the following chapters. Those include the fabrication recipe developed for patterning with <100 nm resolution, the apparatuses for DC electrical tests, and the principle of electro-optic (EO) sampling technique for THz electrical tests.

Chapter 4 summarizes the experimental results on the physical origins of TBJ’s nonlinear response. Systematic experiments using various device lengths and electron mean free paths investigated the role of ballistic transport in the nonlinear response. The study reveals that there are two independent origins of the nonlinear behavior depending on the external applied bias. It also demonstrates the robustness of the nonlinear response over a wide range of operating temperatures and device sizes, which promises the TBJ to be a practical room-temperature ballistic device.

Chapter 5 describes the experimental study on the THz electrical response of the
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TBJ. This chapter starts with illustrating the experimental implementation of an ultra-fast EO sampler that allows for measuring TBJ’s THz response. Using the EO sampling technique, the temporal response of the TBJ is characterized and its operation at sub-THz frequencies was demonstrated.

Finally, chapter 6 summarizes the works done in this thesis, and lists the most prospective future directions of research.
2. Device Physics of the Three-Branch Junction

This chapter describes the device physics of the TBJ. The first part introduces the device structure, operating modes and applications using TBJ’s nonlinear response. Theoretical analysis on the nonlinear response is then carried out based on the two existing models: the Landauer–Büttiker (L-B) formalism and the intervalley electron transfer model.

2.1 Device structure, operation modes, and applications

Device structure of the TBJ is very simple. The TBJ is a three-terminal device consisting of a T-shape (or Y-shape) conductor with three ohmic contacts at the end of each branch. The conductor is typically made of a 2DEG formed in either modulation-doped GaAs/AlGaAs heterostructures or modulation-doped InGaAs/InP quantum-well wafers. Employing electron-beam lithography and dry etching, the 2DEG wafer is patterned into a T-shape structure with a dimension comparable to \( l_e \). Figure 2.1 shows a scanning electron microscopy (SEM) image of a typical device structure. We see that the semiconductor materials in the area other than the T-shape structure are etched away, which leaves a T-shape mesa structure that possesses 2DEG layer. We also see three fan-out structures connecting the T-shape structure with the ohmic contacts.
and metal interconnects. The metal interconnects are connected to contact pads for DC probing, or are a part of planar transmission lines for high-frequency signals.

There are other ways to realize the T-shape structure, such as branched semiconductor nanowires$^{39}$ or carbon nanotubes.$^{40}$ However, in this thesis, we restrict ourselves to the lithographically defined T-shape 2DEG structure as the TBJ hereafter.

TBJ’s operation can be categorized into two modes: two-input passive mode$^{25}$ and three-terminal active mode.$^{35}$ For the sake of the explanation of the operation modes, a schematic of the TBJ is illustrated in Fig. 2.2(a), in which the three branches are marked L (left), R (right), and C (center) along with corresponding voltages.

In the two-input passive mode, input voltages are applied to the L– and R–terminals, while the C– terminal is used as the output terminal. Figure 2.2(b) presents a simulated input-output transfer curve,$^{26}$ where it is assumed that the two inputs are biased in a push-fix fashion (i.e., $V_L = V_{IN}$ and $V_R = 0$) and the output voltage $V_C$ is calculated

Fig. 2.1: A typical device structure of the TBJ.
Fig. 2.2: (a) Schematic illustration of the TBJ. (b) Simulated diode-like behavior of a single TBJ (adopted from Ref. 27).

as a function of $V_{IN}$. Detail description of the model will be discussed in the following sections. A simple analytical form of the transfer curve can be expressed as follows:

$$V_C(V_{IN}) \approx \begin{cases} -\frac{\alpha}{2} V_{IN}^2 + \frac{1}{2} V_{IN} & (V_{IN} \sim 0) \\ -\frac{1}{2} |V_{IN}| + V_{off} + \frac{1}{2} V_{IN} & (|V_{IN}| \gg 0) \end{cases} \quad (2.1)$$

where $\alpha$ and $V_{off}$ represent the curvature around zero bias and the offset voltage, respectively.

The striking feature of this transfer curve is that $V_C$ is always smaller than the average of the two inputs ($V_{IN}/2$). If one assumes that the T-shape conductor is made of a resistor with a linear current-voltage relationship, the output $V_C$ should be exactly
the average of the two inputs because of the geometrical symmetry. However in the TBJ, the output voltage $V_C$ shifts to the lower voltage side regardless of the polarity of the applied voltage. Equation (2.1) says that the deviation from the average is parabolic around $V_{IN} = 0$ with the curvature of $\alpha$. On the other hand, when $V_{IN}$ becomes sufficiently high, $V_C$ is completely pinned to either $V_L$ or $V_R$ with the finite offset voltage $V_{off}$, depending on the polarity of $V_{IN}$. This diode-like nonlinear response is the heart of the TBJ operation.

The nonlinear response can be exploited for various applications. An example is a signal rectification\textsuperscript{29} and second-harmonic generation.\textsuperscript{36} If we assume that an sinusoidal signal $V_{IN} = A \sin(\omega t)$ is applied to L-terminos while keeping $V_R = 0$, the output voltage $V_C$ becomes:

$$V_C(t) = -\frac{\alpha A^2}{4} + \frac{\alpha A^2}{4} \cos(2\omega t) + \frac{A}{2} \sin(\omega t)$$  \hspace{1cm} (2.2)$$

where $A$ and $\omega$ represent the amplitude and the angular frequency of the sinusoidal input, respectively. Here, the magnitude of $A$ is assumed sufficiently small, so the nonlinear response has a parabolic dependence as seen in Eq. (2.1). The first and second term in Eq. (2.2) correspond to the DC rectified signal and the second harmonic signal, respectively. The third term is a linear response that can be eliminated by feeding a signal of $A \sin(\omega t + \pi)$ to R-terminal, if necessary. So far, successful rectification has been experimentally confirmed in a frequency range up to 110 GHz with a sensitivity as high as 150 V/W under ideal impedance matching.\textsuperscript{29} The frequency doubling was demonstrated for frequencies between 20 MHz and 1 GHz.\textsuperscript{36} Thanks to the quadratic
response of the TBJ, significant contrast between the second harmonic signal and higher harmonics was observed.

Another example of applications is a logic gate for digital electronics. Since the diode-like response [Eq. (2.1)] can be simplified to $V_C = \min(V_R, V_L)$, the single TBJ can be considered as AND logic gate by itself. If we take negative voltage as the binary value of 1, this operation can also be considered as OR logic function. Furthermore, it was reported that a triple TBJ structure consisting of three TBJs can work as a NAND or XNOR logic gate.\textsuperscript{42} Those features open up a possibility of realizing ultrasmall logic gates comprising of multiple TBJs.

As the second operation mode, three-terminal active mode using all terminals as inputs and simultaneously as outputs is discussed. Figure 2.3 shows the biasing configuration and the TBJ response in this mode.\textsuperscript{35} In an analogy to a field-effect transistor, the L– and R–terminals are used as a drain and source, respectively, and the C–terminal as a gate. The response presented in Fig. 2.3(b) demonstrates that the drain current saturates at a certain drain voltage and the saturation current increases with the increase of the gate voltage. Figure 2.3(c) presents the drain current vs. gate voltage, which shows a clear cutoff of drain current below the threshold gate voltage. The overall presented behavior is similar to the operation of field effect transistors, meaning that we can use a TBJ as a conventional transistor. Various digital/analog electronics using TBJs in this operation mode such as logic NOR gate,\textsuperscript{35} half-adder circuit,\textsuperscript{43} amplifier with power gain up to 1 GHz\textsuperscript{37} have been demonstrated. Compared to the two-terminal passive
Fig. 2.3: (a) Biasing scheme for three-terminal active mode. (b) Drain current as a function of the drain voltage. (c) Drain current as a function of the gate voltage (adopted from Ref. 36).

mode, this mode brings flexibility in TBJ’s functionality. However, the switching mechanism of this transistor action is still in discussion. Previous studies suggest that the switching has nothing to do with the ballistic transport, but the field effect through the fringing and quantum capacitances in the TBJ structure causes the response. Therefore, we will focus on the two-input passive mode in the following chapters. Despite those facts, this three-terminal active mode shares many advantages with ballistic two-input passive mode, and it would be useful to combine those two modes of operation to extend the functionality of TBJ circuits.
Finally, TBJ’s unique advantages over the conventional electron devices are discussed. The most striking advantage is a high-frequency operation capability due to its small capacitance. The small capacitance implies that the cutoff frequency ($f_T$) is not limited by the RC time delay that usually limits $f_T$ in conventional electron devices. Three main reasons explain the TBJ’s small capacitance: (i) there is no junction-capacitance associated with a highly doped p-n junction, (ii) there is no gate-capacitance as with MOS structures, and (iii) the intrinsic capacitance is just the fringing capacitance between adjacent 2DEGs, which is extremely small. Therefore, the high-frequency performance of TBJ is expected to be free from the RC time delay, which is quite different from any conventional devices. Another important advantage is that the TBJ structure is compatible with the high electron mobility transistors (HEMTs), because those two types of devices are fabricated from the same compound semiconductor heterostructure. Combination of TBJs with a well-established HEMT circuitry should bring versatility for many practical applications.

2.2 Landauer–Büttiker formalism for low-field transport

In this section, theoretical analysis based on the Landauer-Büttiker formalism is performed to provide insights of physical mechanism of the TBJ’s nonlinear response. The L–B formalism is the most frequently used model to describe the electron transport in mesoscopic systems in which inelastic scattering is negligible and electrons travel in ballistic fashion. Contrary to the conventional drift-diffusion theory, the electrical
conduction is considered as a transmission of an electron wave in the L–B formalism. In
the following, the L–B formalism is applied to the TBJ structure, which demonstrates
that the TBJ’s parabolic response under small input-biases can be successfully modeled
by taking into account the energy dependence of the transmivitity.

We start with modeling the TBJ as a T-shape ballistic conductor with three reflectionless contacts at the end of each branch. Each contact is connected to a reservoir
with a chemical potential of $\mu_\alpha$, where $\alpha$ (L, R, or C) represents the name of the
electrode. According to the L–B formalism, the current flowing into the electrode $\alpha$ in a
certain energy window $E$ can be written as:

\begin{equation}
  i_\alpha(E) = i_\alpha^{\text{out}}(E) + i_\alpha^{\text{in}}(E) = \sum_{\beta \neq \alpha} \frac{e}{h} f(E - \mu_\alpha) \cdot T_{\beta\alpha} - \sum_{\beta \neq \alpha} \frac{e}{h} f(E - \mu_\beta) \cdot T_{\alpha\beta},
\end{equation}

where $f(E)$ and $T_{ji}$ represent the Fermi-Dirac distribution function and the transmissivity from $i$-th to $j$-th electrode, respectively. Due to the geometrical and time reversal
symmetry under no magnetic field, we can simplify the $T_{ji}$ as follows:

\begin{equation}
  T_{LC} = T_{CL} = T_{RC} = T_{CR} = T_1
  
  T_{LR} = T_{RL} = T_2.
\end{equation}

To make the problem even simpler, we employ two assumptions: $T = 0$ K and the
so-called linear response assumption that $T_{ji}$ is constant over the entire energy range
that we are interested in. The total current through each electrode is then calculated by
integrating Eq. (2.3) over the entire energy range, which gives the following relations:

\[
\begin{align*}
I_L &= \frac{e}{\hbar} \left\{ -(\mu_C - \mu_L)T_1 - (\mu_R - \mu_L)T_2 \right\}, \\
I_C &= \frac{e}{\hbar} \left\{ (2\mu_C - \mu_L - \mu_R)T_1 \right\}, \quad (2.5) \\
I_R &= \frac{e}{\hbar} \left\{ (\mu_R - \mu_C)T_1 + (\mu_R - \mu_L)T_2 \right\}.
\end{align*}
\]

Since we connect the high impedance voltmeter to the lower stem C to measure the voltage \(V_C\), \(I_C\) is set to zero. The condition gives

\[
\mu_C = \frac{\mu_L + \mu_R}{2}. \quad (2.6)
\]

Therefore, it is proved that the voltage appearing at the lower stem C is equal to the average of the two input voltages according to the L-B formalism under the linear response assumption. Interestingly, the result is the same as what we expect for the strictly diffusive channel. This analysis tells us that the TBJ’s nonlinear response is not a direct consequence of ballistic transport.

Next, we extend the L-B approach to the nonlinear regime by following the model proposed by Xu,\textsuperscript{25} in which the energy dependence of the transmissivity \(T_{ji}(E)\) is explicitly taken into account rather than using the linear response assumption, while keeping the no inelastic scattering assumption valid. For the sake of simplicity, the TBJ’s structure is modeled by three quantum point contacts (QPCs) connected via a ballistic cavity with adiabatic boundaries. The three QPCs are assumed to have a saddle-like potential,\textsuperscript{47} so the transmittivity of the QPC can be simplified by a linear dependence on the
energy:

\[ T(E) = \frac{E - \mu \Theta(E - \mu)}{\Delta}, \]  

(2.7)

where \( \Delta \) is the energy spacing of each transverse mode, \( \Theta(E) \) is the Heaviside step function, and \( \mu \) is the electrostatic potential at the saddle point, respectively. As with Eq. (2.3), we can write the current flowing into lower stem \( C \) in an integral form as:

\[ I_C = \frac{e}{h} \int \{ f(E - \mu_C) \cdot [T_{LC}(E) + T_{RC}(E)] - f(E - \mu_L) \cdot T_{CL}(E) - f(E - \mu_R) \cdot T_{CR}(E) \} dE, \]  

(2.8)

where the energy dependence of each transmissivity is explicitly expressed. Push-pull input voltages are applied to the side branches, changing the electrochemical potential at each terminal as \( \mu_L = \mu_F + eV, \mu_R = \mu_F - eV \), where \( \mu_F \) and \( V \) are the Fermi potential and the variable input voltage, respectively. Given \( T = 0 \) K and \( I_C = 0 \), Eq. (2.8) becomes

\[ \int_{\mu_C}^{\mu_L} T_{CL}(E) dE = \int_{\mu_R}^{\mu_C} T_{CR}(E) dE. \]  

(2.9)

Finally, the \( \mu_C \) is determined to satisfy Eq. (2.9), and \( V_C \) is calculated as \((\mu_C - \mu_F)/e\). A graphical approach illustrated in Fig. 2.4 is helpful to intuitively understand the solution of Eq. (2.9). First, the left side and right side of Eq. (2.9) are viewed as the incoming and outgoing current through lower stem \( C \), respectively. Then, Eq. (2.9) becomes a balance equation between the incoming and outgoing currents to make the total current \( I_C \) equal to zero. In Fig. 2.4, the two shaded areas correspond to the two current flows. One can determine \( \mu_C \) so that the two regions have the same area.
Because the $T_{ji}(E)$ has a linear dependence on energy [Eq. (2.7)] and incoming and outgoing currents use different energy windows, the $\mu_C$ becomes larger than the average of $\mu_R$ and $\mu_L$ to make the two areas equal. Therefore, $\mu_C$ shifts toward the higher potential side than the average of $\mu_R$ and $\mu_L$.

In another approach, one can obtain an analytical solution of $\mu_C^{25}$ from Eq. (2.8). This is done by taking the Taylor series expansion of Eq. (2.8) around $E = \mu_F$ with assumption that the applied voltage $V$ is small. The final form of $V_C$ has a quadratic dependence on the applied push-pull voltage $V$ as follows:

$$V_C = -\frac{1}{2} \alpha V^2,$$

(2.10)

where

$$\alpha = e \left( \frac{\int G_C(E) \frac{\partial^2 f(E - \mu_F, T)}{\partial E^2} dE}{\int G_C(E) \frac{\partial f(E - \mu_F, T)}{\partial E} dE} \right)$$

$$= e \frac{G_C'(\mu_F)}{G_C(\mu_F)},$$

(2.11)
and

\[ G_C = \frac{e^2}{\hbar}(T_{LC} + T_{RC}). \quad (2.12) \]

Note that the analytical solution above can be applied to the system at arbitrary temperature, because we did not use the \( T = 0 \) K assumption. As a conclusion, the parabolic output voltage can be successfully explained by this extended L–B formulation. The reason of this nonlinearity is caused by the energy dependent transmissivity in narrow channels.

It is now worth to point out that the \( V_C \) shows nonlinear dependence even when the applied voltage is small. In other words, the TBJ can be easily driven away from the linear response regime and there is no intrinsic threshold for the nonlinear regime. This feature brings a unique advantage over the conventional rectifiers such as Schottky diodes; namely, no threshold rectification. While the Schottky diode has a threshold voltage to turn on the nonlinear response, the TBJ rectifier shows the nonlinear dependence even around the zero-bias point, which enables rectification of small signals without any external circuitry to feed DC offset biases. In addition, the quadratic dependence of the output implies that the TBJ can be used as an efficient second harmonic generator as mentioned in the previous section.
2.3 Intervalley transfer model for high-field transport

Another approach to the TBJ’s nonlinear response is based on the intervalley transfer mechanism. Contrary to the previous models, the intervalley transfer inherently involves inelastic scattering, and is therefore regarded as a non-ballistic process.

Figure 2.5(a) shows the energy band structure of the conduction band in InGaAs. The conduction band has a multi-valley structure in $k$-space. Under an equilibrium condition, all electrons occupy the lowest energy valley, the $\Gamma$ valley. When the electron system obtains excess energy by an external perturbation (e.g., electric field or optical excitation), some electrons start populating the high energy levels in the $\Gamma$ valley. Once electrons gain kinetic energy higher than the energy separation between the $\Gamma$ valley and the second-lowest $L$ valley ($\epsilon_{\Gamma L}$), those electrons are transferred into the $L$ valley along with optical phonon scattering. Since the effective mass of electrons in the $L$ valley is heavier than that of the $\Gamma$ valley, the intervalley transfer reduces the group velocity. In other words, an additional voltage induces a slower group velocity once the intervalley transfer takes place, resulting in a negative differential conductance with respect to the applied voltage. Since the additional voltage is absorbed in this confined area, the region is called the high-field domain. This intervalley transfer and resultant formation of the high-field domain are well known in bulk III-V semiconductors operating under the high electric field.

In TBJs, the high-field domain is created in the nano-channel near the higher voltage side where the electrons have the highest kinetic energy. As mentioned above, it is
Fig. 2.5: (a) Energy band diagram of the conduction band in InGaAs, (b) Nonuniform carrier and electric field distributions, calculated by Monte-Carlo simulation (adopted from Ref. 34).

expected that the high-field domain can induce a nonuniform distribution of the electric field and the carrier density. To account for this effect quantitatively, place Monte Carlo studies were done by Mateos et al.\textsuperscript{33} Figure 2.5(b) shows the actual distribution of the electrostatic potential and the carrier density as a function of the applied push-pull voltage. Looking at the data for the high applied push-pull voltages, there are an electron accumulation and a larger potential gradient near the high voltage side, which are the clear indications of the high-field domain formation. The nonuniform potential gradient results in a negative voltage at the center of the channel ($x/L = 0.5$) or at the lower stem C. Under small applied voltages, on the other hand, the potential gradient is uniform, and the center voltage remains zero as we expect from a diffusive channel.

One might think that this non-ballistic intervalley transfer mechanism harms the
high-speed operation of the TBJ. In fact, the time constant of the high-field domain formation is usually fast enough (< 1ps)\(^{49}\) so that it does not limit the TBJ frequency response at least in the sub-THz frequency regime. Moreover, no requirement of ballistic transport in this process even becomes an advantage in terms of high-temperature operation.

As we have discussed in this chapter, the nonlinear response in the TBJ takes place with or without the ballistic effect. Thus, we should be careful about the fact that the nonlinear response does not mean the manifestation of ballistic transport. In fact, the intervalley transfer mechanism should be more robust than the ballistic mechanism in a condition that the applied voltage is much larger than optical phonon energy where the inelastic scattering by optical phonons takes place. In addition, the electric field is inevitably intense inside any nanodevices because of its inherently small size. The intense electric field enhances the disturbance of the system, thus it is expected that the high-field effect becomes pronounced in nanodevices fabricated from III-V materials.

2.4 Summary

Device physics of the TBJ is summarized in this chapter. The unique structure and nonlinear response of the TBJ are introduced along with several applications exploiting TBJ’s nonlinear response. The nonlinear mechanism for the low-field regime has been modeled by the extended L–B formalism that takes into account for the electron’s
energy dependence of the transmissivity. The intervalley transfer model was introduced to account for the nonlinear response for the high-field regime.
3. Experimental

This chapter describes the experimental techniques used in the following chapters. Properties of 2DEG studied as well as device fabrication steps are discussed in detail. The measurement system for DC electrical characterization is then described. The last part is devoted for time-domain EO sampling system allowing for THz electrical characterization of the TBJ.

3.1 Device fabrication

3.1.1 2DEG in III-V semiconductor heterostructures

Two kinds of 2DEGs formed in two types of heterostructure wafers were used in this thesis. The first one (referred to as wafer A) is an In$_{0.53}$Ga$_{0.47}$As/In$_{0.52}$Al$_{0.48}$As-based lattice matched heterostructure grown by IQE, inc.$^{50}$ Figure 3.1 presents the thickness and material of each layer in wafer A. The heterostructure was grown by the MBE technique on a $\sim$600-µm-thick semi-insulating InP substrate with undoped In$_{0.52}$Al$_{0.48}$As buffer (500 nm), undoped In$_{0.53}$Ga$_{0.47}$As quantum well (40 nm), undoped In$_{0.52}$Al$_{0.48}$As spacer (20 nm), a $2 \times 10^{12}$ cm$^{-2}$ silicon δ-doping, undoped In$_{0.52}$Al$_{0.48}$As barrier (30 nm), and an undoped In$_{0.53}$Ga$_{0.47}$As cap layer (10 nm). The MBE growth enabled high-quality single crystalline structure and atomically flat interfaces. As can
3. Experimental

<table>
<thead>
<tr>
<th>Layer</th>
<th>Thickness</th>
<th>Composition</th>
</tr>
</thead>
<tbody>
<tr>
<td>Cap</td>
<td>100Å</td>
<td>In$<em>{0.53}$Ga$</em>{0.47}$As</td>
</tr>
<tr>
<td>Barrier</td>
<td>300Å</td>
<td>In$<em>{0.52}$Al$</em>{0.48}$As</td>
</tr>
<tr>
<td>Spacer</td>
<td>200Å</td>
<td>In$<em>{0.52}$Al$</em>{0.48}$As</td>
</tr>
<tr>
<td>Channel</td>
<td>500Å</td>
<td>In$<em>{0.55}$Ga$</em>{0.45}$As</td>
</tr>
<tr>
<td>Buffer</td>
<td>4500Å</td>
<td>In$<em>{0.43}$Al$</em>{0.55}$As</td>
</tr>
<tr>
<td>Substrate</td>
<td>635µm</td>
<td>S.I. InP</td>
</tr>
</tbody>
</table>

**Fig. 3.1:** Physical layer diagram of wafer A (left) and its energy band diagram of conduction band (right).

be seen in Fig. 3.1, the 2DEG layer is formed at the interface between the InGaAs channel layer and InAlAs spacer layer, where the conduction band forms a triangular shape quantum well due to the band discontinuity and dopant induced band bending. A $\delta$-doping was employed in a layer that is separated by 20-nm-thick spacer from 2DEG, which suppresses the impurity scattering and enhances the electron mobility. Table 3.1 shows the sheet resistance $R_{\text{sheet}}$, carrier density $n$ and electron mobility $\mu$, evaluated by the Hall measurement in the dark. To minimize the extrinsic effects such as process induced damage, van der Pauw technique$^{51}$ was employed so that any lithography steps were not required for sample preparation. We can also calculate the elastic mean free path $l_e$ from $\mu$ and $n$ as follows;

$$l_e = \frac{h\mu}{e} \sqrt{\frac{n}{2\pi}},$$

(3.1)

where $h$ and $e$ represent Plank's constant and elementary charge, respectively. Table
3. Experimental

3.1 shows that the $l_e$ exceeds 100 nm at 295 K. Thus, it is substantially longer than the minimum feature size achievable by the electron beam lithography as discussed in Ch. 1. We also note that that $l_e$ is enhanced at 77 K due to a suppression of phonon scattering that limits $\mu$ at 295 K.\(^{52}\)

The second wafer (referred to as wafer B) is the second-generation heterostructure designed to realize a low value of $R_{\text{sheet}}$, which turned out to be important for THz experiments. Figure 3.2 presents the thickness and material of each layer in wafer B. As compared to wafer A, wafer B has i) a higher In content in the channel layer, ii) thinner spacer layer, and iii) higher doping concentration in the $\delta$-doing layer. The higher In content enhances $\mu$ through a change of the effective mass.\(^{7}\) Both the thinner spacer layer and higher doping concentration are aimed for increasing the carrier density in 2DEG.\(^{53}\) In Table 3.1, the 2DEG properties of wafer B are listed in addition to those of wafer A, which clearly shows a reduction of $R_{\text{sheet}}$ of wafer B compared to that of wafer A.

![Physical layer diagram of wafer B.](image)
3. Experimental

<table>
<thead>
<tr>
<th></th>
<th>Wafer A (295 K)</th>
<th>Wafer A (77 K)</th>
<th>Wafer B (295 K)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Sheet resistance: $R_{\text{sheet}}$ (Ω/□)</td>
<td>530</td>
<td>103</td>
<td>220</td>
</tr>
<tr>
<td>Carrier density: $n$ (1/cm$^2$)</td>
<td>$9.8 \times 10^{11}$</td>
<td>$9.3 \times 10^{11}$</td>
<td>$2.1 \times 10^{12}$</td>
</tr>
<tr>
<td>Electron mobility: $\mu$ (cm$^2$/Vs)</td>
<td>$1.2 \times 10^4$</td>
<td>$6.5 \times 10^4$</td>
<td>$1.4 \times 10^4$</td>
</tr>
<tr>
<td>Mean-free-path: $l_e$ (nm)</td>
<td>196</td>
<td>1030</td>
<td>323</td>
</tr>
</tbody>
</table>

3.1.2 Fabrication flow

Device fabrication of all TBJs presented in this thesis was performed at the Cornell NanoScale Science and Technology Facility (CNF). The fabrication process can be divided into four steps: preparation, 2DEG patterning, ohmic contact formation, and interconnect formation. In the following, each step is described in the order corresponding to the actual fabrication flow.

The preparation step starts with dicing a 3” heterostructure wafer into small chips with a ~1-cm-wide rectangular shape. The wafer is scratched by a wafer scribe and cleaved along the scratch mark by gently pressing on both sides of the mark. On each piece, alignment marks for the overlay EBL steps are fabricated by the EBL and lift-off technique. Each alignment mark is a 10 µm x 10 µm square pattern made of a 300-nm-thick Au film.

The next step is patterning the 2DEG layer. A process flow named carbon-hard-mask technique has been developed in this study for this step. Figure 3.3 illustrates the process flow of carbon-hard-mask technique. It is started by depositing a 500-Å carbon layer on the entire surface of the chip using an electron beam (EB) evaporator.
Fig. 3.3: Process flow of the carbon-hard-mask technique.
3. Experimental

[Fig. 3.3(a)]. Then, the chip is coated by a positive EB resist named poly methyl methacrylate (PMMA) using a resist spinner. Very thin layer of PMMA (∼100 nm) is selected to achieve a sub-100-nm resolution. The chip is then set in an EB writer (Leica VB6) and the area of device patterns are exposed to a focused electron beam. The exposed chip is dipped in a developing solution, a mixture of methyl isobutyl ketone (MIBK) and isopropyl alcohol (IPA), to transfer the pattern on the PMMA [Fig. 3.3(b)]. On top of the patterned PMMA, a 60-Å Cr layer is deposited by EB evaporator, and residual PMMA and Cr layers on the PMMA are removed by lift-off technique in a methylene chloride solution [Fig. 3.3(c)]. Now, the device patterns are transferred on the Cr layer. The chip is then exposed to O₂ plasma in a reactive ion etcher (RIE) chamber to transfer the patterns on the carbon layer using the Cr layer as an etch-mask [Fig. 3.3(d)]. The remaining carbon layer acted as an etch-mask in the subsequent substrate etching. Ion mill etching by Ar ions is employed for the substrate etching, which removes the area left uncovered by the carbon hard mask. Typically, the etching depth was chosen ∼130 nm so that the channel InGaAs layer is completely etched away. Finally, the remaining carbon mask is removed by O₂ plasma in RIE [Fig. 3.3(e)].

This carbon-hard-mask technique was crucial for fabricating the TBJ devices studied in the following chapters, because the experiments require devices with a channel length close to $l_e$ at room temperature (∼100 nm). To demonstrate the patterning resolution, a diagnostic pattern was fabricated, which consists of multiple wire structures with several wire-widths and wire separation distances. Figure 3.4 shows a scanning electron
Fig. 3.4: An SEM image of a diagnostic pattern. Wire width and separation distance was varied from 40 to 90 nm and from 40 nm to 160 nm, respectively. The actual dimension for each structure is summarized in the inset table.

microscope (SEM) image of the diagnostic pattern. The minimum trench width and minimum wire width extracted from the SEM image are approximately 40 nm and 50 nm, respectively. This patterning resolution of the carbon-hard-mask technique satisfies the prerequisite of the following experiments. Moreover, it is confirmed that this technique gives a good uniformity over the entire chip and high reproducibility for every run.

The next fabrication step is ohmic contact formation. As with the previous step, the EBL and lift-off are employed to deposit Ni/Ge/Au (150 Å/350 Å/1000 Å) multilayers on the contact areas. For this step, however, a relatively thick (~500 nm) resist, named methyl methacrylate (MMA) copolymer, are used because of the thick ohmic
metals. The metal deposition is followed by a rapid thermal annealing at 430 °C for 40 seconds in forming gas (Ar/H mixture), through which the Ge and Au formed an alloy diffusing into the substrate. To obtain a reasonable value of ohmic contact resistance, it is important to remove the native oxide by dipping the chip into a buffered oxide etchant right before loading it in an EB evaporator. The metal layer thickness and the annealing temperature have been optimized successfully, resulting in an acceptably low contact resistance (∼0.5 Ωmm) at every run.

In the last step, an interconnect layer is formed by the EBL and lift-off technique in a full analogy to the previous step. This layer provides electrical interface to device electrodes through contact pads for either wire-bonding or probing. It is also used as a transmission line for high-frequency signals. Typically a bilayer of Ti/Au (200 Å/2000 Å) is used. Since this layer forms a Schottky junction on the substrate, the leakage current flowing through the substrate is negligibly small.

Figure 3.5 shows a multi-scale view of a complete device chip for DC electrical characterization. Typically, a chip area is 7 mm × 7mm, and approximately 100 devices are on a single chip. The device structure for the THz experiment will be discussed in Ch. 5.

### 3.2 DC electrical characterization system

Most of the DC electrical characterizations presented in this thesis were performed in a probe station located at the Center for Nanoscale System at Cornell University.
Fig. 3.5: Multi-scale view of a typical TBJ chip for DC electrical characterization.
This probe station has both DC and RF probing capability between 0–40 GHz. Sample sizes up to a full 100 mm (4”) diameter can be probed with any of six DC/RF probes. The system is turbo-pumped to enable cryogenic sample temperatures. Continuous liquid helium flow is used to cool down the sample stage and a temperature controller measures/controls the sample and probe temperatures. The temperature is tunable between 4.2 and 375 K with excellent temperature stability. The probe station also has a high resolution microscope with CCD camera, which facilitates probe placement with 3 µm ultimate resolution.

DC characterization by a home-made dipping probe for liquid-He/N was also performed in some cases. The dipping probe consists of a sample mount for a dual-in-line ceramic package (DIP) and a breakout box for BNC connectors. A sample under test is wire-bonded on a DIP, then mounted on the sample mount. The dipping probe can fit both the liquid-He and liquid-N containers, thus low temperature experiments can be performed by dipping the sample directly in cryogens.

For the measurement electronics, I assembled a computer controlled system consisting of commercial DC voltage/current sources and voltmeters for various purposes of experiments. Actual measurement configurations will be presented in the following chapters.
3.3 Time-domain electro-optic sampling system for THz tests

To investigate the device response at THz frequencies, I implemented a time-domain electro-optic (EO) sampling technique. The EO sampling system may be the most readily understood as an ultrafast sampling oscilloscope with subpicosecond temporal resolution. This system achieves this exceptional performance by using a modelocked Ti:Sapphire laser generating 100-fs-wide optical pulses, which essentially defines the temporal resolution. The optical pulse excites a photoconductive switch inducing sub-picosecond electrical transient, which provides a trigger signal for the sampling oscilloscope as well as an input signal to the device under test (DUT). The electric field component of the transient signal interacts with an EO transducer affecting the polarization of a probing optical pulse, allowing us to measure the intensity of the electric field. Sampling the intensity of the electric field as a function of the time delay between arrival times of the excitation and sampling optical pulses, the temporal response of the DUT can be obtained. Thus, the EO sampling system is an oscilloscope with the optical input and output.

In the following sections, each building block of the EO sampling system is discussed in detail. We start with the description of the photoconductive switch as an optic-electric transducer and the EO transducer made of LiTaO$_3$. Then, actual implementation of the entire system is presented.
3.3.1 Photoconductive switch

The basic operation of a photoconductive switch is as follows: two electrodes are deposited on an semi-insulating semiconductor, and a voltage is applied between the electrodes. Exposing the semiconductor to light with the photon energy larger than the semiconductor’s bandgap generates electron and hole pairs. The initial response time of carrier generation is almost instantaneous $\sim$fs, and is limited by the speed at which the optical energy is delivered to the semiconductor (temporal shape of the optical pulse). This fast generation of photoexcited carriers leads to a fast electrical transient at the electrodes. On the other hand, the falling edge is determined by the carrier lifetime.\textsuperscript{54} In typical MBE grown III-V compound semiconductors like the substrate used in this thesis, the carrier lifetime is much longer than 10 ps. Thus, additional efforts to shorten the duration time are required.

To overcome the limit set by a long carrier lifetime, a technique called a nonuniform edge-illumination is employed in this thesis.\textsuperscript{55–58} The technique enables generating subpicosecond electrical pulses in any photoconductive semiconductor substrate. In this technique, the excitation optical pulse is focused at the edge of the negatively biased electrode, where the electric field is confined because of the reversely biased Schottky junction. Upon arrival of the excitation optical pulse, the built-in electric field is screened by the photoexcited carriers, followed by redistribution of the electric field in the electrodes’ gap.\textsuperscript{58} Those processes successfully excite a subpicosecond electrical pulse in the transmission line in which the switch is embedded. The biggest advantage
of the nonuniform edge-illumination technique is that it does not rely on the material properties of the photoconductive substrate, such as carrier recombination rate and drift velocity.\textsuperscript{58} In addition, the wide electrode-gap relaxes the complexity of the fabrication and reduces the dark current of the switch. The subpicosecond pulse generation has been observed on coplanar waveguide structures made on GaAs\textsuperscript{55–57} as well as on Si and InP,\textsuperscript{59} thus establishing the material independent nature of this subpicosecond pulse generation technique.

### 3.3.2 LiTaO\textsubscript{3} EO sampler

The EO effect, or Pockels effect is the change of refractive index of a material, linearly proportional to the applied electric field.\textsuperscript{60} This effect is generally different in ordinary and extraordinary axes of the material; thus it induces extra birefringence in the material. When light propagates through a birefringent crystal such as LiTaO\textsubscript{3}, polarization of the output light is altered. The LiTaO\textsubscript{3} EO sampler is an electro-optic transducer using the EO effect. In the following, the basic theory of the EO effect as well as the EO sampler is introduced, based on Ref. 61.

The properties of an optical crystal are generally described by the impermeability tensor $\eta_{ij}$:

$$
\eta_{ij} = \frac{\varepsilon_0}{\varepsilon_{ij}} = \frac{1}{n_{ij}^2} \quad (i, j = 1, 2, 3)
$$

(3.2)

where $\varepsilon_0$ is the permittivity of vacuum, $\varepsilon_{ij}$ is an element of the dielectric tensor, and $n_{ij}$ is an element of the refractive index tensor. The indices $i$ and $j$ correspond to the three Cartesian coordinates. The propagation of light through such optical medium is
governed by the index ellipsoid:

$$\eta_{ij} x_i x_j = 1.$$  \hfill (3.3)

Equation (3.3) and henceforward, the convention of summation over repeated indices is assumed. The Pockels effect induces a refractive index change as a function of the applied electric field $\vec{E}$. If the function varies slightly with $\vec{E}$, $\eta$ can be series expanded:

$$\eta_i(\vec{E}) = \eta_i(0) + r_{ik} E_k + \cdots,$$  \hfill (3.4)

where a reduced tensor with $i = 1–6$ and $k = 1–3$ is used to avoid the complex tensor calculation. The $r_{ij}$ are called the linear or Pockels EO coefficients. Specific crystal symmetry can further reduce the independent elements of the Pockels tensor $r$. For trigonal $3m$ crystals such as LiTaO$_3$, there are only 4 independent nonzero Pockels coefficients, as shown below:

$$
\begin{pmatrix}
0 & -r_{22} & r_{13} \\
0 & r_{22} & r_{13} \\
0 & 0 & r_{33} \\
0 & r_{51} & 0 \\
r_{51} & 0 & 0 \\
-r_{22} & 0 & 0 \\
\end{pmatrix}.
$$  \hfill (3.5)

Now, we consider specific example illustrated in Fig. 3.6 where a cross section of a LiTaO$_3$ crystal with electrodes applied in microstrip configuration is shown. The application of a voltage $V$ on the top electrode produces an electric field in $z$-direction with a magnitude $E_z = V/d$. The crystal is cut so that the $c$-axis (optical axis) is parallel to the $z$-axis.
Combining Eq. (3.3) and Eq. (3.4), we can get the propagation function of light through the crystal as a function of $E_z$:

$$
\left( \frac{1}{n_0^2} + r_{13}E_z \right) x^2 + \left( \frac{1}{n_0^2} + r_{13}E_z \right) y^2 + \left( \frac{1}{n_e^2} + r_{33}E_z \right) z^2 = 1,
$$

(3.6)

where $n_0$ and $n_e$ are the ordinary and extraordinary indices of refraction. Thus, the principal axes are unchanged under the external electric field, but the indices of refraction are modified by $E_z$ as shown below:

$$
n_x = n_0 - \frac{1}{2} n_0^3 r_{13} E_z, \\
n_y = n_0 - \frac{1}{2} n_0^3 r_{13} E_z, \\
n_z = n_e - \frac{1}{2} n_e^3 r_{33} E_z.
$$

(3.7)

where we have assumed that $n^2 r_{ij} E_z \ll 1$ and ignored higher order terms of $E_z$. When light propagates along the $y$-axis and with the polarization at $45^\circ$ to the $x$ and $z$ axes, the phase retardation $\Gamma$ from the LiTaO$_3$ crystal can be expressed as:

$$
\Gamma = k_0 |n_x - n_z| \\
= \frac{2\pi}{\lambda} (n_e - n_0) L_1 - \frac{\pi}{\lambda} (n_e^3 r_{33} - n_0^3 r_{13}) \frac{L_2}{d} V,
$$

(3.8)
where \( L_1 \) and \( L_2 \) are the width of the crystal and the region of electric field existing, respectively. \( d \) is the length along \( z \)-axis where the voltage is applied [Fig. 3.6]. The second term in Eq. (3.8) is the consequence of EO effect and a linear function of the voltage signal. The first term is caused by the static birefringence of the crystal. This static retardation can be compensated using an optical compensator, thus the first term will be ignored hereafter.

An intensity modulator can be built based on the EO effect, by putting two cross-polarized polarizers at two ends of the optical path through the EO crystal. The optical axis of the first polarizer is oriented at \( 45^\circ \) relative to the optical axis of the crystal. Thus, the intensity transfer function is:

\[
I = I_0 \sin^2 \left( \frac{\Gamma}{2} \right). \tag{3.9}
\]

This intensity transfer function is illustrated in Fig. 3.7. When \( \Gamma = \pi \), all the incoming light can be transmitted. The voltage applied to the EO crystal that produces \( \pi \) retardation is called the half-wave voltage \( V_\pi \). Equation (3.9) can be rewritten in terms of the voltage induced retardation as:

\[
I = I_0 \sin^2 \left( \frac{\pi V}{2 V_\pi} \right), \tag{3.10}
\]

The \( V_\pi \) can be derived from the second term of Eq. (3.8) as follows:

\[
V_\pi = \frac{\lambda d}{L_2 (n_3^3 r_{33} - n_0^3 r_{13})}. \tag{3.11}
\]

Typically, \( V_\pi \) is on the order of 10 kV in EO crystals used in EO sampling system, while voltage signals that are under investigation are in general comparable or smaller.
than 1 V. The fact that $V/V_\pi \ll 1$ indicates that the modulator works at point A, very close to the zero point in Fig. 3.7. In order to maximize the sensitivity and the linearity of intensity modulation, it is usually optically biased at point B in Fig. 3.7, where $\Gamma_0 = \pi/2$, by adding a quarter-wave plate between the two polarizers as shown in Fig. 3.8. Transmission function including the static optical bias becomes as follows:

$$I = I_0 \sin^2 \left( \frac{\pi}{4} + \frac{\pi V}{2 V_\pi} \right),$$

(3.12)

Because $V/V_\pi \ll 1$, the differential change in transmission intensity can be expanded. Neglecting high order ($n \geq 2$) terms, the change of light intensity is directly proportional to the amplitude of the voltage signal as follows:

$$\frac{\Delta I}{I} = \frac{\pi V}{V_\pi}.$$  

(3.13)
3.3.3 Time-domain EO sampling system

Schematics of the EO sampling system used in this thesis are illustrated in Fig. 3.9. The device chip consists of a photoconductive semiconductor substrate and a transmission line. A DUT and a voltage biased gap as a photoconductive switch are embedded in the transmission line. Both excitation and sampling optical pulses are generated by a commercial Ti:Sapphire mode-locked laser system. The center wavelength is set at 780 nm with a FWHM of \( \sim 3 \) nm in frequency domain, corresponding to \( \sim 100\)-fs pulse width in time domain. This extraordinarily short duration of the laser pulse enables the temporal resolution of EO sampling to be in a subpicosecond time scale.

The excitation beam is focused onto the gap embedded in the transmission line. As discussed in the previous section, the nonuniform edge-illumination technique is used to overcome the limit set by the relatively long carrier life time in the MBE grown InGaAs...
3. Experimental

Fig. 3.9: Schematic diagram of the EO sampling system.

A photoconductive switch is used for the incident beam. The beam is focused on the edge of negatively biased electrode. The electrode gap is typically 20–30 µm, wider than the diameter of the focused beam spot on the sample (~15 µm). This wide electrode gap greatly eases the fabrication, reduces the dark current, and allows a high bias voltage.

The EO intensity modulator is constructed as described in the previous section, by adding a pair of cross-polarized polarizers, a quarter-wave plate, and a LiTaO₃ crystal as an EO transducer in the sampling beam path. The lateral dimension of the crystal is approximately 6 mm × 3.5 mm. The thickness is approximately 0.5 mm. The crystal is cut so that the optical axis is parallel to its long edge and placed on the sample so that the electric field is parallel to the long edge of the crystal. The crystal is clamped down on the sample to ensure that it is in close contact with the transmission line, thus the electric field propagating in the transmission line can efficiently penetrate into
the crystal. A high-reflection (HR) coating is applied to the bottom surface of the EO crystal so that the optical pulse is able to go through the region with electric field and is reflected back by the HR coating.

Lock-in detection technique is employed to enhance the signal-to-noise (SN) ratio. The intensity of excitation pulse focused on the photoconductive switch is modulated by an acousto-optic modulator with a 91 kHz modulation frequency from the internal oscillator in a lock-in amplifier. The lock-in amplifier measures only the frequency component of the output signal at the modulation frequency, thus unwanted fluctuations such as $1/f$ noise of the laser intensity and the auxiliary electronics can be minimized.

Differential signal detection scheme using a balanced photodetector is also employed to further improve the SN ratio. The balanced detector is a pair of two photodetectors that works in a differential detection mode. By using a polarized beam splitter as an analyzer, not only the transmission but also the reflection at the analyzer can be measured. The reflection beam is the orthogonal component to the transmission beam, thus the transfer function can be written in a full analogy of the transmission beam [Eq. (3.12)]:

$$I = I_0 \sin^2 \left( \frac{3\pi}{4} + \frac{\pi V}{2V_\pi} \right),$$

(3.14)

This equation can be series expanded as was done in the previous section. Neglecting high order ($n \geq 2$) terms, the change of light intensity is:

$$\frac{\Delta I}{I} = -\frac{\pi V}{V_\pi},$$

(3.15)

Therefore, the reflection beam gives an output signal with the same magnitude but
the opposite sign compared to the transmission beam [Eq. (3.13)]. Measuring differential
signal of the two components, the total output signal is doubled. Furthermore, this
scheme reduces the noise from a laser intensity fluctuation, because the differential
detection cancels out the static response, i.e., $I_0$ in Eqs. (3.12) and (3.14).

The arrival timing of the optical sampling pulse is controlled by a computer-controlled
translational stage. Varying the time delay between switch excitation and signal sam-
pling pulses, a temporal shape of the electrical pulse generated by the photoconductive
switch can be measured in time domain. If we put a DUT between the photoconductive
switch and the sampling point, the response of DUT can also be evaluated.

The temporal resolution of the system is affected by several parameters. First is the
laser pulse width $\tau_P$, which in our case is $\sim$100 fs. Thus, the electric field in the LiTaO$_3$
crystal is averaged over the 100 fs period. Next is the spatial interaction between the
electric field and the crystal. This is broken into two situations. In the reflection mode,
the sampling beam must propagate through electric field region twice. The time for the
light to propagate can be calculate from the interaction length $L_2$ and speed of light in
the crystal $n/c$, as $\tau_L = nL_2/c$. For a typical case, $\tau_L \sim$140 fs with $L_2$ of $\sim$20 um and
$n$ of 2.1. The other factor is the spread of sampling beam in lateral dimension. The
EO effect is smeared by $\tau_A = \sqrt{\varepsilon d/c}$, where $d$ is the diameter of probe beam. For a
typical case, $\tau_A \sim$110 fs with $\sqrt{\varepsilon}$ of $\sim$3.3 and $d$ of 10 $\mu$m. All of those effects combine
to provide a temporal resolution given by:

$$\tau = \sqrt{\tau_P^2 + \tau_L^2 + \tau_A^2} \sim 200\text{fs}$$

(3.16)
Therefore, the EO sampling system offers subpicosecond temporal resolution, which is a significant advantage over all electronic sampling techniques.

3.4 Summary

This chapter described experimental techniques used in the following chapters. First, device fabrication technique was overviewed. Fabrication technique based on EBL and carbon-hard-mask technique have been developed successfully, which offers patterning capabilities of <100 nm resolution. Combining MBE grown 2DEG heterostructures having $l_e > 100$ nm at room temperature, it is now possible to fabricate structures smaller than the mean free path. The explanations of the EO effect and EO sampling system are presented. With the help of a femtosecond pulsed laser as an optical excitation and sampling source, both generating subpicosecond electrical pulses and measuring the electrical transient signal with $\sim 200$ fs temporal resolution can be realized. This technique allows examining the ultrafast (THz) response of TBJ devices.
4. Ballistic vs. Diffusion Transport in
Three-Branch Junction

TBJ’s nonlinear response in both the diffusive and ballistic transport mode has been
investigated by systematically varying the device size \( L \) and operating temperature \( T \).
The operating temperature alters \( l_e \) through the temperature-dependent \( \mu \) and \( n \) in
2DEG. Since the degree of ballistic transport is determined by the ratio of \( L \) and \( l_e \),
this approach allowed us to study the role of ballistic transport on the TBJ’s nonlinear
characteristics, as well as to determine the involved physical mechanisms.

4.1 Experimental

TBJs with different sizes were fabricated. Figure 4.1 presents the parameters that
define the device geometry. A set of six devices with different channel lengths \( L \) (\( L = 160, 200, 300, 500, 1000, 2000 \text{ nm} \)) were prepared, while the other parameters were
fixed: channel width \( W = 140 \text{ nm} \), probe length \( L_p = 190 \text{ nm} \), and probe width \( W_p \)
= 120 nm. The left and right branches of the TBJs are connected to two independent
ohmic contacts at each side through a fan-out region made of 2DEG. The two contacts
at each branch allowed for a four-point measurement to eliminate the contact resis-
tances, which will be revisited later. In Fig. 4.2, SEM images of the six devices used
for DC characterizations are presented. In actual fabrication, five copies for each TBJ were fabricated for a statistical analysis. No significant shape variations in SEM images among the copies were observed, showing a good reproducibility of the fabrication process. The trapezoidal structures seen just above the T-shape structure in long-channel devices \((L \geq 500 \text{ nm})\) are dummy structures. In the EB pattern writing process, accelerated electrons affect not only the region covered by incident electron beam spot but also the surrounding area, which is called the proximity effect. The dummy structures were used to average out the proximity effect in the narrow wire region.

To control the \(l_e\) of 2DEG, \(T\) was varied from room temperature to 4.5 K. Here, temperature dependence of \(l_e\) is discussed. Figure 4.3 presents the temperature dependence of \(\mu\), \(n\), and \(l_e\), measured in a Hall bar device\(^{62}\) with a relatively large size \((L = 100 \text{ \mu m}, W = 20 \text{ \mu m})\). Single conduction layer was assumed in the Hall measurement analysis. For comparison, Fig. 4.3 also shows the unprocessed wafer’s data that were presented in Table 3.1.

Looking at the \(n\) in Fig. 4.3, we should note two unusual behaviors: first, there is a
Fig. 4.2: SEM images of TBJ devices with different nanochannel lengths.
significant reduction of \(n\) after the device fabrication, especially at 295 K, as compared to the unprocessed wafer’s data. The second observation is that \(n\) increases with decreasing temperature in the Hall bar device, which is an opposite behavior to the unprocessed wafer and previous studies.\(^8,\,^52\) This unusual behavior of \(n\) is most likely caused by crystal defects formed during fabrication steps, although the real cause is not clear at this moment. More studies are needed to clarify the reasons as well as to find solutions to the problems.

Next, we see in Fig. 4.3 that \(\mu\) increases as the temperature decreases and saturates at \(\sim 9 \times 10^4\) cm\(^2\)/Vs at \(T < 40\) K. This temperature dependence of \(\mu\) is typical for the 2DEG formed in InGaAs/InAlAs heterostructures.\(^8,\,^52\) Previous studies\(^8,\,^52\) revealed that scattering by acoustic and/or polar-optical phonons determines \(\mu\) in the high temperature regime \((T > 60\) K), resulting in that \(\mu\) has a \(T^{-1}\) dependence. In low temperature regime \((T < 60\) K), \(\mu\) is limited by the remote impurity scattering and alloy-disorder scattering, which leads to a constant \(\mu\) over the temperature range studied.

From the values of \(\mu\) and \(n\), we can calculate \(l_e\), [see, Eq. (3.1)] at each \(T\) as shown in Fig. 4.3. We see that \(l_e\) changes from 146 nm to 1.28 \(\mu\)m when \(T\) changes from 295 K to 20 K. The temperature dependence of \(l_e\) is primarily governed by that of \(\mu\), because of the stronger temperature dependence of \(\mu\) than that of \(n\).

Now, we discuss the electrical characterization technique employed for TBJ devices. To accurately test electrical properties of TBJs, measurements with a careful subtraction
Fig. 4.3: Temperature dependence of the 2DEG carrier density, mobility, and mean free path measured in a Hall bar device (square points) and unprocessed wafer (triangular points).
of parasitic resistances were performed. Figure 4.4 illustrates the measurement configuration, in which the constant bias current \( I_{LR} \) was applied between side branches and voltages at all three electrodes were simultaneously measured. This four-point-probe configuration for the side branches allowed us to eliminate the ohmic contact resistances, because the voltages measured by voltage probes \( (V_L \) and \( V_R) \) did not include the voltage drops at ohmic contacts for the current path. Since the configuration by itself eliminates the contact resistances at every bias point, the effect of the contact resistance is removed including its possible nonlinear bias dependence. We should also consider that the fan-out regions (transitional regions between the nanochannel and the ohmic contacts) give a considerable parasitic resistance, referred to as the access resistance \( R_{access} \) hereafter. The value of \( R_{access} \) was evaluated by extrapolating the TBJ’s resistance \( R_{tot} = (V_R - V_L)/I_{LR} \) versus \( L \) dependence to \( L = 0 \). The value of \( R_{access} \) was obtained as 7.5 kΩ at 295 K, 3.0 kΩ at 200 K, and 1.2 kΩ at 4.5 K. Then, the residual voltage drop due to \( R_{access} \) was subtracted from the input voltages with an assumption that \( R_{access} \) was independent of bias current.

Hereafter, for the sake of convenience, we introduce a virtual zero potential reference \( V_0 \), defined as \( V_0 = (V_L + V_R)/2 \), and redefine the voltages as the relevant potentials with respect to the \( V_0 \) (see Fig. 4.4):

\[
V_{pp} = (V_R - V_L - I_{LR}R_{access})/2,
\]
\[
V_{C0} = V_C - V_0,
\]

(4.1)

where \( V_{pp} \) and \( V_{C0} \) are referred to as push-pull voltages applied to the nanochannel
Fig. 4.4: Schematic illustration of measurement configuration employed. The variable dc current $I_{LR}$ is applied between side stems and the voltages at three voltage probes—$V_L$, $V_R$, and $V_C$—are measured by high impedance voltmeters.

and the center probe voltage, both with respect to $V_0$. This way, we could neglect the effects of two parasitic resistances and view $V_{pp}$ being applied in a push-pull fashion to the TBJ structure. Under the push-pull voltage input, TBJ’s linear response gives zero output. Thus, any output voltage deviation from zero would indicate the TBJ’s nonlinear response.

4.2 Nonlinear response in ballistic and diffusive transport regimes

Typical electrical responses of TBJs are shown in Fig. 4.5, with Figs. 4.5(a) and (b) corresponding to the $V_{C0}$ vs. $V_{pp}$ and $I_{LR}$ vs. $V_{pp}$ dependencies, respectively. The two presented examples in Fig. 4.5 represent the two extreme cases: one for the ballistic
Fig. 4.5: (a) The central probe voltage $V_{C0}$ and (b) the device current $I_{LR}$ versus the push–pull bias voltage $V_{pp}$ for the mostly ballistic (solid line) and mostly diffusive (dashed line) modes of the TBJ operation.

mode ($L = 160 \text{ nm}; \ T = 4.5 \ K; \text{ solid line}$) in which $l_c \gg L$ and the other for the diffusive mode ($L = 2000 \ \text{nm}; \ T = 295 \ K; \text{ dashed line}$) in which $L \gg l_c$. Interestingly, the two curves in Fig. 4.5 exhibit very similar behavior despite the difference of the transport mode. At small $V_{pp}$, $V_{C0}$ has a quadratic dependence and $I_{LR}$ increases linearly. On the other hand for large $V_{pp}$, $V_{C0}$ becomes linear with nearly a unity slope while $I_{LR}$ is saturated. The only obvious difference between the data for the diffusive and ballistic modes is the value of the transition voltage separating the two regimes.

For a more quantitative analysis of the data presented in Fig. 4.5(a), we took a derivative of $V_{C0}$ with respect of $V_{pp}$ and plotted it as a function of $V_{pp}$, as shown in Fig. 4.6. Figures 4.6(a) and 4.6(b) present the $dV_{C0}/dV_{pp}$ dependences measured at different temperatures for the longest and shortest TBJs, respectively, while Figs.
4. Ballistic vs. Diffusion Transport

4.6(c) and 4.6(d) illustrate the $L$ dependences at 295 K and 4.5 K, respectively. All the dependences shown in Fig. 4.6 share two main common features: first, the derivative has a linear dependence in the small $V_{pp}$ regime, which means that $V_{C0}$ vs. $V_{pp}$ has a quadratic dependence; second, all derivatives reach the ±1 value at large push-pull voltages, which indicates that in this regime $V_{C0}$ is completely pinned to the voltage of the side branch with negative voltage. In addition, we note that for short TBJs or the structures measured at low temperatures, the $dV_{C0}/dV_{pp}$ characteristics are step-like functions with constant slopes between the two saturation regions, while for the intermediate/high $L$ and $T$ ranges, the curves broaden and we observe an additional bend at a certain transition voltage. In the following, we will discuss each point addressed above, with the emphasis on the physics involved.

In the low $V_{pp}$ regime, $V_{C0}$ has been found to be represented by

$$V_{C0} = -\frac{\alpha}{2} |V_{pp}|^2,$$  \hspace{1cm} (4.2)

where $\alpha$ is a fitting parameter that represents the curvature in Fig. 4.5(a) and can be extracted from the curves in Fig. 4.6 as the value of the slope around the origin. The quadratic behavior of $V_{C0}(V_{pp})$ confirms that the nonlinear ballistic effect explained in Ch. 2 is responsible for the electron transport in this regime. This conclusion is supported by the fact that the value of $\alpha$ decreases as the channel becomes more diffusive as seen in Fig. 4.6, since the perfectly diffusive channel should give $V_{C0} = 0$, or, equivalently, $\alpha = 0$. For the opposite, i.e., ballistic case, the extended Landauer-Büttiker model predicts that $\alpha_0 = e/2\mu_F$ at $T = 0$ K under a hard-wall potential shape simpli-
Fig. 4.6: The derivative of $V_{C0}$ with respect to the push–pull voltage. The temperature dependence of TBJs with (a) $L = 2000$ nm and (b) $L = 160$ nm. The channel-length dependence at (c) $T = 295$ K and (d) $T = 4.5$ K.
4. Ballistic vs. Diffusion Transport

where $\mu_F = \pi \hbar n/m^*$ is the Fermi level of the 2DEG. Since $\mu_F$ in our 2DEG system is approximately 0.02 eV, $\alpha_0$ is expected to be 25 V$^{-1}$.

In Fig. 4.7(a), $\alpha/\alpha_0$, namely the ballisticity factor, for all our tested devices as a function of the nanochannel length at several operating temperatures is plotted. One notes that for the TBJ with $L = 160$ nm at 4.5 K, which has the largest $l_c$ to $L$ ratio, $\alpha/\alpha_0$ $\sim$27%. That infers that the TBJ is partly ballistic even in the best case, suggesting that a proper modeling should include a finite scattering effect. Simultaneously, at the worst-case, i.e., for the 2000-nm-long device operated at 295 K, the $\alpha/\alpha_0$ factor is only $\sim$6.5%. This low value is expected, because the transport mode should be predominantly diffusive in this case. However, it is interesting to note that the $\alpha/\alpha_0$ is still nonzero. This nonzero $\alpha/\alpha_0$ seems to implicate that the nonlinear ballistic effect does not completely vanish even for long TBJs operated at high temperatures, although the overall behavior is strongly dominated by the scattering. This unexpected robustness of the TBJ’s nonlinear response is favorable in terms of realizing room temperature ballistic devices. Overall, as expected, $\alpha/\alpha_0$ is a function of temperature (increases with decreasing $T$) and sharply decreases with the $L$ increase, which reflects the degree of ballistic transport in the nanochannel. All those experimental findings support that the parabolic nonlinear response is a consequence of ballistic transport.

In the high $V_{pp}$ regime, on the other hand, the $V_{C0}(V_{pp})$ dependence can be approximated by the following formula:

$$V_{C0} = -|V_{pp}| + V_{onset},$$

(4.3)
Fig. 4.7: The channel-length dependence of (a) the ballisticity factor $\alpha/\alpha_0$ and (b) the onset voltage $V_{\text{onset}}$. Points represent our experimental data extracted from Figs. 4.2 and 4.3, while the star is a datum of the Monte Carlo simulation published in Ref. 34. The lines are just guides for the eye.
where \( V_{\text{onset}} \) is the extrapolation to \( V_{C0} = 0 \). This linear dependence of \( V_{C0} \) for large \( V_{pp} \) can be explained by the intervalley transfer model explained in Ch. 2 and Ref. 34. This is supported by the facts that the slope of \( V_{C0} \) is exactly equal to \( \pm 1 \) as well as that the current saturates in this voltage region, which are the consequences of the high-field domain formation as discussed in Ch. 2.

Next, the effects of \( L \) on the nonlinear response caused by the intervalley transfer process are discussed. In Fig. 4.7(b), we plot \( V_{\text{onset}} \) for all our tested devices as a function of \( L \) at several operating temperatures. The behavior of \( V_{\text{onset}} \) in Fig. 4.7(b) can be understood in relationship to the transport mode. In the diffusive mode, i.e., at 295 K and at 200 K for \( L > \sim 300 \) nm, the \( V_{\text{onset}} \) is proportional to \( L \). This indicates that the intervalley transfer is triggered at a constant \( E_{th} \). This fact is consistent with the conventional intervalley transfer model for long channel devices, in which the electron temperature \( T_e \) representing the kinetic energy that electron can gain is determined by the electric field applied. It has been considered that \( E_{th} \) at which the intervalley transfer takes place cannot be modeled by the simple analytical formula, thus other methods such as experiments and Monte Carlo simulations have been used to obtain \( E_{th} \). Previous experiments\textsuperscript{63,64} report that \( E_{th} \) is in the range from 2 kV/cm to 5 kV/cm for the InGaAs-based long-channel \((L \gg l_e)\) devices at room temperature, which is consistent with the 2.0-kV/cm value obtained from the slope of the data shown in Fig. 4.7(b). Furthermore, according to the Monte Carlo simulations in Ref. 35, \( V_{\text{onset}} \) should be about 0.15 V at 300 K for a TBJ with \( L = 750 \) nm, which is again consistent
with our experimental data. Those findings further support the intervalley transfer model in this high bias regime.

In ballistic cases, i.e., at 4.5 K for all \( L \) and at 200 K for \( L < \sim 300 \) nm, Fig. 4.7(b) shows that \( V_{\text{onset}} \) remains constant and independent of \( L \). This behavior is very different from the diffusive channel case discussed above in which \( V_{\text{onset}} \) is proportional to \( L \), but can be understood by considering how electrons acquire their kinetic energy from the external bias. As an extreme example, we consider a ballistic nanochannel with the length \( L \) and biased by \( V_{\text{in}} \). The kinetic energy that an electron can acquire during traveling along the channel is determined by \( V_{\text{in}} \), because the electron does not lose its energy by scattering. Thus, not the electric field \( V_{\text{in}}/L \) but the applied voltage \( V_{\text{in}} \) determines the kinetic energy in the ballistic channel. Since the electron has to acquire a sufficient amount of kinetic energy \( \sim \Delta \varepsilon_{\Gamma L}/e \) to make a transition to the upper valley, \( V_{\text{in}} \sim \Delta \varepsilon_{\Gamma L}/e \) is required for the intervalley transfer at the very least. In other words, \( V_{\text{onset}} \) is determined by \( \Delta \varepsilon_{\Gamma L}/e \) and should be independent of \( L \) in short channel devices, when \( E_{\text{th}} L \) is smaller than \( \Delta \varepsilon_{\Gamma L}/e \). We can define the crossover channel length \( l_{i} \) at which \( \Delta \varepsilon_{\Gamma L}/e = E_{\text{th}} l_{i} \). The \( l_{i} \) obtained from Fig. 4.7(b) is \( \sim 160 \) nm at 295 K, \( \sim 300 \) nm at 200 K, and >2000 nm at 4.5 K. We expect \( l_{i} \) to correlate with \( l_{c} \), because this transition is related to the transport mode. The experimentally obtained \( l_{i} \) shows similar temperature dependence as \( l_{c} \), however \( l_{i} \) is longer than \( l_{c} \) at all temperatures. To model \( l_{i} \) quantitatively, further study such as Monte Carlo simulations are necessary.
it is irrelevant to the ballistic effect. However, the process by which electrons gain their kinetic energy is affected by the actual transport mode (ballistic vs. diffusive), giving rise to a difference as discussed above.

4.3 Summary

The main result of this chapter is that two distinct mechanisms are involved in the electron transport and nonlinear response in nanostructured TBJs. First, under small applied push-pull voltages, a nonlinear ballistic effect is clearly observed at all temperatures and results in a quadratic $V_{C0}(V_{pp})$ dependence with a curvature that is sensitive to the amount of the transport “ballisticity.” Second, the $\Gamma$–$L$ intervalley electron transfer dominates the nonlinear response when the applied $V_{pp}$ exceeds the onset voltage $V_{onset}$ that is related to the critical electric field value or voltage needed for the intervalley transfer to emerge. Both effects are present in all tested TBJs and the strong nonlinear response was observed even in the longest devices operated at room temperature. The robustness of the TBJ’s nonlinear response is a unique advantage in terms of realizing a room-temperature integrated circuitry using TBJs.
5. THz Electrical Response of the Three-Branch Junction rectifier

The most prominent feature of the TBJ is its high-frequency characteristics. To demonstrate the high-speed (THz) operation of the TBJ, an experimental setup based on the time-domain EO sampling has been developed. TBJ’s ultrafast behavior is characterized in time-domain by exciting the TBJ with sub-picosecond electrical pulses and measuring the transient response with sub-picosecond temporal resolution. In frequency-domain, this corresponds to a broadband characterization in a frequency range up to THz, because the sub-picosecond electrical pulse has a frequency spectrum extending into the THz frequency range. This chapter is devoted to describe the THz test system as well as the results on TBJ’s response in THz frequencies.

5.1 Experimental

First, the structure of device chip for EO sampling characterization is described, using a representative example of a double-TBJ rectifier, one of the three kinds of devices tested. Device designs of the rest of TBJ devices are illustrated in Sec. 5.4 along with their THz responses. Figure 5.1 presents optical microscope images of a double-TBJ rectifier chip. Figure 5.1(a) shows the overall device structure. The chip consists of
5. THz Electrical Response

Fig. 5.1: (a) Optical microscope image of overall device structure. (b) Enlarged view of the photoconductive switch. (c) & (d) Enlarged views of the double-TBJ rectifier.

A photoconductive switch and DUT both embedded in a coplanar waveguide (CPW). The CPW is made of 200-nm-thick Ti/Au film deposited in an EB metal evaporator. The thickness of the Au was chosen 200 nm so that the thickness is larger than the skin depth at frequencies around 1 THz. The line width and separation of the CPW are both 30 µm. The total length of the CPW is 6 mm, and the DUT is positioned roughly at the center to avoid unwanted reflections from the ends of CPW. Figure 5.1(b) shows an enlarged view of the photoconductive switch. The photoconductive switch is comprised of two 30-µm-wide electrodes with a voltage biased gap of 25 µm. The photoconductive substrate underneath the electrodes is the InAlAs buffer layer that was brought to surface when the mesa etching for the TBJ structure was performed.
Since the photoconductive switch is a part of the CPW as seen in Fig. 5.1(b), excited electrical pulses propagate to the DUT along the CPW. The DUT in this particular chip is a double-TBJ rectifier,\textsuperscript{29,30} a combination of 2 TBJs in parallel as is shown in Figs. 5.1(c) and 5.1(d). The purpose of having 2 TBJs is to reduce the input impedance of the DUT as well as to make the geometry as symmetrical as possible. To minimize the parasitic access resistances, small ohmic contacts as well as small fanout 2DEGs are employed as seen in Fig. 5.1(d).

To operate the TBJ rectifier, three branches are used as input, ground, and output, as is shown in Fig. 5.2(a). This configuration is known as push-fix biasing scheme,\textsuperscript{26} because one of the inputs is always tied to the ground. Figure 5.2(b) shows the DC nonlinear transfer curve measured in the biasing configuration shown in Fig. 5.2(a). The transfer curve has a quadratic dependence in the low-bias regime ($-0.4 \text{ V} < V_{\text{in}} < 0.4 \text{ V}$) and a linear dependence in the high-bias regime ($|V_{\text{in}}| > 0.4 \text{ V}$), which can be approximated by

$$V_C(V_{\text{IN}}) \approx \begin{cases} -\frac{\alpha}{2} V_{\text{IN}}^2 + \frac{1}{2} V_{\text{IN}} & (V_{\text{IN}} \sim 0) \\ -\frac{1}{2} |V_{\text{IN}}| + V_{\text{off}} + \frac{1}{2} V_{\text{IN}} & (|V_{\text{IN}}| \gg 0) \end{cases},$$

(5.1)

where $\alpha$ and $V_{\text{off}}$ are the curvature around the zero bias and the offset voltage, respectively. As discussed in the previous chapters, the quadratic dependence in the low-bias regime originates from the nonlinear ballistic electron transport. On the other hand, in the high-bias regime, the transfer curve has a linear dependence with different slopes depending on the polarity of the input voltage, which can be explained by the intervalley...
Fig. 5.2: (a) Electrical test configuration for the dc transfer-curve measurement; (b) DC transfer curve at room temperature.

transfer model. (See Ch. 4 for the discussion on the carrier transport in the TBJ and the transfer characteristics.)

Figure 5.3 presents a schematic illustration of the complete EO sampling system, which mainly consists of generation and detection components for sub-picosecond electrical pulses. For the generation part, a photoconductive switch excited with 100-fs-wide optical pulses from a mode-locked Ti:Sapphire laser (Coherent Mira 900-F) is used. Due to a long life time of photoexcited carriers in the InAlAs layer, the InAlAs-base PC switch generally has a relatively slow response on the order of 100 ps. To overcome the limit and generate sub-picosecond pulses, the nonuniform edge-illumination technique discussed in Ch. 3 is employed. In this technique, the excitation optical pulse is focused at the edge of the negatively biased electrode, where the electric field is concentrated.
because of the reversely biased Schottky junction. Upon arrival of the excitation optical pulse, the built-in electric field is screened by the photoexcited carriers, followed by subsequent redistribution of the electric field in the electrodes’ gap. Those processes successfully excite a subpicosecond electrical pulse in the CPW line.

For detection of electrical pulses, an ultrafast EO modulator consisting of an EO transducer (a LiTaO$_3$ crystal overlaying the CPW line) and the optics system as a polarization analyzer are implemented as discussed in Ch. 3. The EO transducer converts the electrical field signal, which travels in the CPW and couples to the EO crystal, into the change of polarization state of the optical sensing pulses through the Pockels
effects. The optical pulse is focused on the gap between CPW lines through the EO transducer, so that it penetrates the crystal and is reflected at the dielectric mirror at the bottom of the EO crystal, where the fringing electric field of the signal exists. By using a linearly polarized incident light and two polarizers that are cross polarized each other, the EO sampler acts as an optical intensity modulator for the sensing beam. A quarter-wave plate in front of the second polarizer (analyzer) is used to maximize the sensitivity of the modulator as well as to linearize the intensity modulation with respect to the change of the electric field applied. To enhance the signal-to-noise ratio, the differential signal detection by using a balanced photodetector and the lock-in detection at the modulation frequency of 91 kHz are used.

To obtain the temporal waveform of the electrical transient signals, the optical pulses from the Ti:Sapphire laser are split into two paths, creating both the excitation and sensing pulses. Using a computer-controlled translational stage, EO sampling is performed at each delay $\tau$ that is defined as the difference of arriving times of the two pulses at the device chip. Thus, transient signal can be recovered as a function of $\tau$. In addition, the sensing position can be changed by moving either the EO transducer or the focusing spot of sampling pulse along the CPW line, thus sensing at an arbitrary position, i.e., both the incident and transmitted electrical pulses of the TBJ rectifier, is possible.
5.2 Characterization of the photoconductive switch

In the first phase of the studies, electrical pulses from the photoconductive switch were characterized. For this purpose, a structure identical to the one shown in Fig. 5.3 but without the rectifier was used. Figure 5.4 presents a transient voltage signal from the photoconductive switch that was illuminated with an optical power of 2 mW and biased at 9 V. The spatial separation between the PC switch and the sampling point was 1 mm, corresponding to the distance between the PC switch and the TBJ rectifier in the complete test chip. Thus, this electrical pulse will be later considered as the incident pulse to the TBJ rectifier. The main pulse in Fig. 5.4 has a FWHM of 1.75 ps, and is followed by a broader and smaller peak. The second peak turned out to be ringing due to the modal dispersion effect, which will be discussed shortly. Changing the applied bias and/or the excitation optical power, the pulse height of the main peak was tunable between 0 and \( \sim 2 \) V without noticeable change of its temporal characteristics.

Next, to study the effects of pulse propagation in the CPW line, a series of transient signals at different positions along CPW were collected. Figure 5.5(a) summarizes the position dependence of the FWHM and the rise time of the voltage pulse as a function of a distance from the photoconductive switch. A significant change of the temporal waveform was observed, i.e., the FWHM increased from 0.92 ps to 1.63 ps, and the rise time increased from 0.61 ps to 1.33 ps, while the electrical pulse propagated from 0.2 mm to 0.8 mm away from the switch. This result indicates that the pulse shape seen
Fig. 5.4: Temporal waveform of a voltage pulse generated by the photoconductive switch. The spatial separation between the switch and sampling point was 1 mm. The inset shows a waveform sampled at 0.22 mm away from the switch.

In Fig. 5.4 (or the incident pulse in the complete test chip) was not determined by the intrinsic switch response but rather by the propagation properties of the CPW.

For more quantitative understanding on the propagation properties, a frequency-spectrum analysis was performed. Generally, voltage pulses propagating along the CPW can be modeled in the frequency-domain as:

\[ V(f, z) = V(f, 0)e^{-(\alpha(f)+j\beta(f))z}, \]  

(5.2)

where \( V(f, z) \) is the Fourier transform of the time-domain waveform at a distance \( z \) along the waveguide, \( \alpha(f) \) is the frequency dependent attenuation, and \( \beta(f) \) is the frequency dependent phase. Using multiple data set of the voltage waveforms taken at different \( z \), \( \alpha(f) \) and \( \beta(f) \) can be extracted from Eq. (5.2).
Fig. 5.5: (a) The points represent experimental data on the sampling-position dependence of the FWHM and rise time of the electrical pulse. The solid line represents simulated dependences, as described in the text. (b) Propagation constants calculated from the measured pulse shape and model. Small oscillations visible in both $\alpha(f)$ and $v_{ph}$ curves are artifacts caused by truncation of the Fourier transform.
Figure 5.5(b) presents $\alpha(f)$ and the phase velocity $v_{ph}(f) = 2\pi f/\beta(f)$ obtained in that way. It is seen that $\alpha(f)$ stays negligibly small below $\sim 0.3$ THz, and subsequently rises up above $\sim 0.3$ THz. On the other hand, $v_{ph}$ remains almost constant (slightly decreasing with $f$) over the entire frequency range. The frequency-spectrum analysis indicates that $v_{ph}$ has much weaker frequency dependence than $\alpha$. We can therefore, conclude that the pulse broadening observed in the Fig. 5.5(a) was primarily caused by attenuation of the high-frequency spectrum rather than a modal dispersion. As another approach, using the experimental $\alpha(f)$ and $\beta(f)$ data shown in Fig. 5.5(b), we can simulate temporal waveforms at different positions along the CPW line. The solid lines in Fig. 5.5(a) correspond to the FWHM and rise time of simulated pulses with the input waveform sampled at 0.23 mm away from the photoconductive switch. One can see that the solid lines successfully reproduce the trends of our experimental data. In addition, the simulated time-domain waveforms in Fig. 5.6 succeeded in reproducing the post-pulse ringing observed in Fig. 5.4, and allowed us to conclude that it was caused by the frequency dependent $v_{ph}$ as seen in Fig. 5.5(b). Furthermore, one should note that the measured waveform [Fig. 5.4 inset] near the photoconductive switch did not show ringing, which provides evidence that the post-pulse ringing is caused by the pulse propagation.
Fig. 5.6: Simulated waveform at different positions from the photoconductive switch. The numbers represent the distances from photoconductive switch. The arrow shows the post-pulse ringing due to the modal dispersion.

5.3 Picosecond time-domain response of the double-TBJ rectifier

In the complete test chip, including the double-TBJ rectifier [Fig. 5.3], two transient waveforms were sampled before and after the TBJ rectifier. Figures 5.7(a) and 5.7(b) show these two waveforms and their sampling positions, while Fig. 5.7(c) is an enlarged view of Figs. 5.7(a) and 5.7(b) showing the early parts of the pulses. The waveform taken at the position before the TBJ rectifier exhibits an initial sharp double peak, followed by broadened double-peak structures repeated every 20 ps. The origin of the initial
double peak is a combination of the incident pulse and its reflection at the interface between the CPW and TBJ rectifier. This observation was confirmed by measuring the time separation $\Delta t$ between these two peaks while moving the sensing positions with respect to the DUT. The measurement showed that $\Delta t$ increased linearly with the distance $\Delta x$ of the sensing position from the TBJ rectifier. The propagation velocity $v_g$ calculated from the slope of the $\Delta x$ versus $\Delta t$ dependence was $\sim 1.0 \times 10^8$ m/s, and was consistent with the phase velocity data shown in Fig. 5.6(b). The subsequent double-peak structures, repeated every 20 ps, were multiple reflections of the original voltage pulse inside a 1-mm-long cavity that is formed by the CPW terminated by the TBJ and photoconductive switch at both ends. This idea was confirmed by the repetition interval (20 ps) and the calculated $v_g$ ($1.0 \times 10^8$ m/s).

The other waveform shown in Fig. 5.7(b) was taken at a position after the TBJ rectifier, and referred to as transmitted pulse hereafter. It shows similar time dependence to the incident waveform. The main difference is the negative dip and oscillatory behavior after the main pulse with positive peak, which is caused by parasitic capacitive coupling between the input and output CPW lines. The effects of parasitic capacitances will be discussed later.

Another important observation from Fig. 5.7 is that the amplitude of the transmitted pulse was significantly smaller than the incident. Likewise, the amplitude of the reflected pulse in the Fig. 5.7(a) was comparable to the incident pulse. Those facts indicate that a large portion of the incident signal was reflected at the CPW/TBJ interface, rather than
Fig. 5.7: (a) Temporal shape of the incident and reflected signal sampled before the TBJ rectifier. The inset shows the sampling position. (b) Temporal shape of the transmitted signal after the TBJ rectifier. The inset shows the sampling position. (c) Enlarged view of the main pulses in Figs. 5.7(a) and 5.7(b). In this particular measurement, the photoconductive switch was excited with an optical power of 4 mW and biased at 5 V, although the overall temporal dependence of recorded transients did not change significantly with the excitation conditions used.
being coupled into the TBJ. The small transmission and corresponding large reflection were a consequence of a significant impedance mismatch at the CPW/TBJ plane. To estimate the voltage pulse height of the transmission, the transmission coefficient $s_{21}$ in the scattering matrix was computed based on a two-port network analysis:\textsuperscript{65} 

$$s_{21} = \frac{2z_{21}Z_{\text{CPW}}}{(z_{11} + Z_{\text{CPW}})(z_{22} + Z_{\text{CPW}}) - z_{12}z_{21}}, \quad (5.3)$$

where $z_{ij}$ and $Z_{\text{CPW}}$ are the impedance matrix elements of the TBJ rectifier and the characteristic impedance of the CPW line, respectively. In our case, $Z_{\text{CPW}}$ is calculated as $\sim 62 \, \Omega$ by quasi-static analysis,\textsuperscript{66} while $z_{ij}$s are $z_{11} = z_{22} = 2z_{12} = 2z_{21} = 10 \, k\Omega$ that were estimated from the DC resistance measurements under an assumption that three branches have the same resistance and the TBJ is modeled by the T-network shown in Fig. 5.8. The calculated $s_{21}$ from Eq. (5.3) was only $\sim 0.008$, despite the fact that the measured amplitude ratio between the incidence and transmission pulses in Fig. 5.7 is $\sim 0.051$. This discrepancy implies that there must exist another mechanism that

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**Fig. 5.8:** Equivalent circuit that is used to estimate transmission and reflection at the CPW/TBJ interface. $z_{ij}$s are matrix elements of the impedance matrix of the TBJ rectifier.
transmitted the pulse through the TBJ rectifier. One possible mechanism is capacitive coupling between the input and output electrodes of the CPW. If one assumes that a capacitive element \( C_e \) bridges the input and output port rather than the rectifier, the output voltage can be computed as follows:

\[
V_{\text{out}}(t) = Z_{\text{CPW}} \cdot C_e \cdot \frac{dV_{\text{in}}(t)}{dt}.
\]  

(5.4)

Figure 5.9 presents the calculated waveform from Eq. (5.4) with \( C_e = 0.818 \) fF, obtained by solving Poisson’s equation using the 3-dimensional finite-element method with boundary conditions reflecting the actual geometry. The overall time-dependence of the calculated waveform in Fig. 5.9 is similar to the experimental transmission data including the negative dip observed in Fig. 5.7. This analysis indicates that the capacitive coupling dominates the transmitted signal, determining the overall transient shape. However,
looking at the calculated waveform with great care, one notes that the positive pulse height of 3.1 % is still less than the experimental value (5.1 %) and the negative peak is more pronounced in the calculated waveform than the experimental data. Those two observations suggest that, although the capacitive coupling is the dominating component of the transmitted signal, a finite portion of the transmitted waveform comes from the response of the TBJ rectifier.

Next, the nonlinear behavior of the double-TBJ rectifier at THz frequencies was studied by varying the pulse height of the incidence. Figure 5.10 presents the dependence of the transmitted-pulse height on the incident one; namely sub-THz transfer curve. In the Fig. 5.10, the points show the experimental data, while the solid line is the expected transfer curve derived from the DC transfer curve in Fig.5.2 (b). We observe an excellent agreement, although in the tested range the sub-THz transfer curve has an almost linear dependence. The reason for this linear behavior is the small effective power coupling into the TBJ rectifier. In other words, effective voltage of the incident pulse was only a few % of the actual pulse height. Since the DC transfer curve had a low rectifying efficiency and was almost linear around the zero bias, the nonlinear response should be miniscule under the small amplitude modulation in the sub-THz frequency range.

Finally, frequency spectrum analysis was performed on the incident and transmitted signals to examine the TBJ’s response in frequency-domain. Figure 5.11 presents the Fourier transform spectra of the incident and transmitted pulse signals. To remove the influence of the reflection pulse, the waveform in Fig. 5.4 was used as the incident
Fig. 5.10: Incident pulse height dependence of the transmitted pulse height. The points with error bars show experimental data. The solid line represents a scaled dc transfer curve [Fig. 5.2(b)] with a scaling factor of 0.102 to account for the effective coupling of incident voltage in the sub-THz frequencies.

pulse. Looking at the frequency spectrum of the incident pulse, we note that it has usable spectrum up to 0.8 THz at which point the signal was buried in the background noise. Thus, it is confirmed that the system has a near 1 THz bandwidth. Considering that the presented bandwidth was limited by the propagation properties of the CPW rather than the intrinsic photoconductive switch response, further improvement is possible by reducing the separation between the photoconductive switch and the DUT. As for the spectrum of the transmitted pulse, one can note that it has a frequency-dependence similar to the incident pulse up to \(~0.5\) THz at which point the signal was buried in the background noise. To show the similarity of the frequency-dependences, the inset
5. THz Electrical Response

of Fig. 5.11 presents the amplitude ratio of the transmitted and incident spectra. For the sake of simplicity, the ratio was normalized by the values at low frequencies. As is seen in the figure, the normalized ratio has constant value within the bandwidth up to 0.5 THz. This similarity of frequency-dependences suggests that the TBJ rectifier does not degrade the frequency response in the tested frequency range, and indirectly demonstrate the superiority of the high-frequency response of the TBJ rectifier. However, further extension of the bandwidth is necessary to study the electrical response in the frequencies exceeding 1 THz and obtain the cutoff frequency of the TBJ rectifier.

5.4 Single-TBJ rectifier with dc-bias offset

To clearly observe the nonlinear response in THz frequencies (sub-picosecond response in time-domain), refinements of device structure have been performed. As discussed in the previous section, the main problems suppressing the nonlinear response are: i) a severe impedance mismatch between the transmission line and DUT, ii) a subtle nonlinearity around zero bias, and iii) a large feedthrough mediated by the capacitive coupling between input and output ports of the transmission line. In the followings, an improved design intended to solve those problems is presented along with their time-domain responses to picosecond electrical excitation pulses.

The device is named a common ground single-TBJ rectifier. The overall device structure and close-up views of the photoconductive switch and the TBJ rectifier are presented in Fig. 5.12. As shown in Fig. 5.12(b), the photoconductive switch is identical
Fig. 5.11: Fourier-transformed spectra of the incident and transmitted voltage transient signals. The temporal waveform in Fig. 5.4 was used for the incident spectrum, and that in Fig. 5.7(b) was used for the transmitted spectrum. The inset shows the normalized amplitude ratio of the transmitted spectrum to the incident one.

to as the one in previous setup, except it has an additional meander line connected to the output electrode. The meander line acts as an inductor, thus application of a DC offset superimposed on sub-picosecond electrical pulses is possible. The meander biasing structure can be understood as analogous to a microwave bias-tee component. The line width and length of the inductive line is 1 μm and 160 μm with 20 turns, respectively. The transmission line is modified to coplanar strip (CPS) structure instead of CPW, because the inductive line geometry does not fit into CPW structure.

Another modification is for mitigating the impedance mismatch by decreasing the
DUT resistance and increasing the CPS characteristic impedance \( Z_{\text{CPS}} \). Firstly, the wafer B that had a lower sheet resistance than wafer A [see Chapter 3] was used for the TBJ fabrication process. The high indium content in channel InGaAs layer of wafer B reduces the edge depletion layer, resulting in a wider electrical branch width under the same physical branch width. Because of those two changes, an approximately 4-folds reduction of input resistance was achieved, compared to the TBJ of the double-TBJ rectifier in the previous section. However, the CPS structure does not allow having two TBJs as with the double-TBJ rectifier, thus the improvement in the input impedance was reduced to approximately 2-fold.

Secondary, increasing the \( Z_{\text{CPS}} \) was done along with DUT resistance reduction, by
changing the ratio of the line width $w$ to the line separation $s$. Since the EO-sampling setup does not require using 50-$\Omega$ matched electronics, $Z_{\text{CPS}}$ can be chosen arbitrarily. The actual CPS in the presented device had a $w/s$ of 0.5, while CPW in the previous section had a $w/s$ of 1.0. This change brought a $\sim 22\%$ increase of the characteristic impedance, and the resulting value of $Z_{\text{CPS}}$ was calculated as $\sim 84 \Omega$ according to the quasi-static analysis.\

For the last, reduction of the capacitive coupling was achieved by shrinking the lateral size of CPS. Because the area of input/output electrodes determines the capacitance, smaller size is better for minimizing the capacitive coupling. As seen in Fig. 5.12(a), the reduction of lateral dimension of CPS is done with a bend of metal lines. The $w/s$ is kept constant around the bend so that $Z_{\text{CPS}}$ is maintained during the transition from wide to narrow CPS region. The actual $s$ and $w$ of narrow (wide) CPS region were 20 $\mu$m (40 $\mu$m) and 10 $\mu$m (20 $\mu$m), respectively. The lower limit of the lateral size was determined by the sensitivity of EO-sampling. Small lateral dimensions reduce the sensitivity through the interaction length $L_2$ in Eq. (3.11). Also, small $L_2$ makes the system more susceptible to the physical contact of LiTaO$_3$ crystal clamped on the transmission line, giving rise an uncertainty in the experiment. The lateral dimensions above are chosen by taking into account those practical limitations.

Figure 5.13(a) shows the DC transfer curve of the TBJ at room temperature. Overall, an ideal transfer curve that can be described by the Eq. (5.1) was obtained. Compared to the previous double-TBJ device, the transfer curve in Fig. 5.13(a) is more
Fig. 5.13: (a) DC transfer curve and (b) IV curve of the presented TBJ at room temperature.

linear for the low-field regime, while clear nonlinear response is observed for the high-field regime. In Fig. 5.13(b), IV curve is presented from which the input resistance was obtained as 6.5 kΩ.

Figure 5.14 exhibits the temporal response of the incident pulse shape. The sampling point was \( \sim 200 \mu \text{m} \) before the TBJ. As with the incident signal of the previous double-TBJ experiment [Fig. 5.7], the temporal response has a double peak structure, which is caused by the reflection at the CPS/TBJ interface. The FWHM and rise time of the incident pulse are 1.48 ps and 1.30 ps, respectively: both numbers are consistent with the data in Fig. 5.5. Comparing the pulse shape in Fig. 5.14 with that of the previous double-TBJ device in Fig. 5.7, we note that the amplitude of the reflection in Fig. 5.14 is significantly smaller than that earlier observed in Fig. 5.7.
Figure 5.15(a) shows the transmitted pulse response sampled at 100 µm after the TBJ. Two curves represent the temporal responses at the different DC offset superimposed on the picosecond excitation pulse. For the sake of later discussion, the data were normalized by the incident pulse height that was measured separately. We note that the transmitted pulse height was \( \sim 40\% \) of the incident, which is significantly larger than the previous double-TBJ device (\( \sim 5.1\% \)). Since the \( s_{21} \) calculated from the \( z_{ij} \) of the single-TBJ device and \( Z_{CPS} \) is only 1.7 % and capacitive coupling between input and output ports of CPS should be smaller than the previous design, this large transmission is not caused by either the improved impedance matching or capacitive coupling, but by other effects. The cause of this large feedthrough has not been identified in this study.

As for the DC offset dependence, a clear dependence is not observed in Fig. 5.15(a).
Fig. 5.15: (a) Temporal response of the transmitted voltage pulse at dc bias offsets of 0 V and −1.2 V. (b) Voltage difference between the two signals in (a) normalized by transmitted signal. The error bar is the standard deviation of 30 measurements.
Since the slope of the DC transfer curve is quite different at two biasing points (0 V and -1.2 V), a significant difference in small signal response is expected. If we assume that the transmitted voltage is proportional to the slope of the DC transfer curve, the voltage difference between the two cases are estimated as approximately $s_{21}V_i(t)$, where $V_i(t)$ is the time-dependent incident voltage. Considering that the $s_{21}$ is only 0.017, the expected difference is much smaller than the background feedthrough signal. Therefore, extra care is needed to resolve the TBJ’s nonlinear response buried in the background.

In Fig. 5.15(b), the difference of the two signals in Fig. 5.15(a) is plotted, where the normalized voltage difference $\Delta V(t)$ is used so that we can evaluate the proportion of the voltage change in the transmitted pulse signal normalized by the incident pulse height:

$$\Delta V(t) = \frac{V_{-1.2}(t) - V_0(t)}{V_0(t)} \cdot \frac{V_{t,\text{peak}}}{V_{i,\text{peak}}}$$  \hspace{1cm} (5.5)

where $V_0(t)$ and $V_{-1.2}(t)$ is the time-domain response of transmitted voltage with DC offset of 0 V and -1.2 V, respectively, and $V_{t,\text{peak}}$ and $V_{i,\text{peak}}$ are the pulse heights of transmitted and incident pulses, respectively. The error bars in Fig. 5.15(b) represent the standard deviation of the 30 measurements at each data point. The dash line indicates the magnitude of expected signal amplitude from the $s_{21}$ analysis. Since we do not know the temporal shape of the ideal response, the dash line only represents the magnitude of the expected difference. The data in Fig. 5.15(b) indicate that there is a slight difference between the two biasing points. As is clear from the comparison between the expected value and the error bar, however, the presented experimental
setup could not reliably resolve the TBJ’s response. In future experiments, elimination of the feedthrough signal or significant improvement of the impedance matching is indispensable to observe the clear bias dependence and to demonstrate THz nonlinear response.

5.5 Summary

A THz electrical response of the ballistic TBJ rectifier was measured by picosecond electrical pulse excitation with a PC switch and the time-domain EO sampling technique. From the Fourier transform of the time-domain signals, frequency spectra of the incident and transmitted signals up to $\sim0.5$ THz were successfully obtained. The TBJ rectifier did not degrade the frequency spectrum of the transmitted electrical pulses within the frequency range studied, which indirectly confirmed its THz operation. Meanwhile, the limiting factor of the system bandwidth was studied, which proved that the wave propagation properties of the CPW, rather than the intrinsic response of the photoconductive switch, limited the system bandwidth. In fact, the intrinsic response of the PC switch had subpicosecond FWHM and rise time, which promises the extension of the bandwidth $>1$ THz by reducing the dispersion effect during the wave propagation. As for the nonlinear response in THz frequencies, because of the small power coupling and the low rectifying efficiency of the TBJ rectifier in the tested bias range, a clear nonlinear response was not observed in this study. In the small-signal regime, however, the THz data points followed the nonlinear DC transfer curve. Two approaches for
enhancing the THz rectification have been performed: superimposing a DC bias voltage through an inductive line and mitigating the impedance mismatch at the transmission line/DUT interface. Unfortunately, those approaches raised other problems preventing the clear nonlinear behaviors in THz frequencies. Further refinement of the device structure is needed for practical application of the THz TBJ rectifier.
6. Conclusions

This thesis studied the TBJ device, a novel high-speed electron device utilizing ballistic transport at room temperature. The achievements throughout this research can be categorized into the three parts:

The first was development of the fabrication technique. Room temperature ballistic devices require structures with a dimension smaller than the electron mean free path at room temperature. Since the typical mean free path at 300 K of high-mobility 2DEG in III-V semiconductor heterostructures is $\sim100$ nm, a fabrication technique with a resolution less than 100 nm is imperative. This work has developed a fabrication technique using electron beam lithography with a carbon-hard-mask, which allowed us to pattern a 2DEG layer with sub-100 nm resolution.

The second was clarification of the nonlinear mechanism involved in the input-output transfer curve of the TBJ. It was found that two distinct mechanisms were involved in the nonlinear response depending on the magnitude of the external bias voltage. Under small voltages, the nonlinear ballistic effect was responsible, resulting in a quadratic dependence with a curvature that was determined by the amount of the transport "ballisticity". In high voltage regime, the intervalley electron transfer mechanism took over the nonlinear response, when the applied voltage exceeded the onset voltage that was related to the critical voltage needed for the intervalley transfer to emerge. The
robustness of the nonlinear response over a wide range of operating temperature and channel length was demonstrated experimentally, and was explained by the non-ballistic nature of the intervalley transfer mechanism. This robustness of the TBJ’s nonlinear response is a unique advantage in terms of realizing a room-temperature integrated circuitry using TBJs.

The last element was the demonstration of THz operation of the TBJ. An experimental setup based on the time-domain EO sampling has been developed to observe ultrafast response of the TBJ rectifier. The obtained temporal response to picosecond excitation pulse was analyzed by the two-port network analysis, which revealed a significant impedance mismatch and a signal feedthrough by the extrinsic capacitive coupling. Frequency spectrum of the time-domain signals showed that the TBJ rectifier did not degrade the frequency spectrum within the system bandwidth of $\sim0.5$ THz, indicating superior high-frequency performance in THz frequencies.

In terms of the future direction of the presented research, the most promising application of TBJs seems to be as THz detectors. From the theoretical analysis, the TBJ rectifier gives a parabolic transfer curve with curvature $\alpha_0 = e/2\mu_F$, corresponding to the responsivity of $\sim600$ V/W under perfect impedance matching. The responsivity of current state-of-the-art Schottky diodes are typically in the range of 4,000 V/W at 100 GHz to 400 V/W at 900 GHz. Thus, the responsivity of the TBJ detector can approach Schottky diode’s performance around 1 THz, and could get better performance at even higher frequencies. Another important measure of THz detector is the
6. Conclusions

noise performance. The noise performance of the TBJ has never been studied. However, due to the unique noise properties of the nanoscale ballistic channel,\textsuperscript{68} systematic noise studies of the TBJ would be interesting from the basic physics and device development point of view. In order to realize the practical TBJ THz rectifier, there remain practical problems to be solved, such as impedance matching and THz antenna design. Further studies on both fundamental and practical aspects of the ballistic TBJ device are necessary to realize THz detectors using TBJs.

Another interesting aspect to explore in TBJs and ballistic nanodevices is the sub-picosecond temporal characteristics of the electron motion. In a time scale shorter than the mean scattering time $\tau_e$, one would expect that new physical phenomena, such as the kinetic-inductive effect,\textsuperscript{69} should be observed. Kinetic inductance is negligible in conventional nonsuperconducting electron devices, because the operating frequency is typically much lower than $\tau_e^{-1}$. Since the $\tau_e$ in high-mobility 2DEG is of the order of 1 ps at room temperature, it is likely that the sub-picosecond temporal characteristics of ballistic devices are affected by such unconventional effects. The EO sampling system presented in this thesis has a unique capability to directly observe such ultrafast phenomena. Thus, EO sampling study should give important insights into the THz electrical response of the room temperature ballistic devices.


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