MODELING OF TUNNEL OXIDE TRANSISTORS

By

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Abstract

Two improvements to a comprehensive analytic model which describes the steady-state current in a metal-insulator-semiconductor tunnel junction are reported. The first modification replaces the conventional two-band representation of the thin oxide band structure with a one-band model. In this approach the electron barrier height for tunneling is always less than the hole barrier height by an amount equal to the semiconductor band gap. The second improvement enables the energy dependence of the electron and hole tunneling probability factors to be taken into account. This is accomplished by expressing the tunneling probabilities as short-term series expressions. The capability of the model to accurately predict the electrical characteristics of metal-insulator-semiconductor (MIS) tunnel junctions is demonstrated by simulating the d.c. and a.c. performance of three major types of transistor with tunnel oxide emitters, namely the tunnel emitter transistor (TETRAN), MIS-emitter transistor (MISET) and polysilicon emitter transistor (PET). Experimental data for the d.c. characteristics of all these devices are available and are found to be well-
described by the predictions of the models. No experimental data for the limits of high frequency operation of the TETRAN, MISET and pnp PET have yet been reported. The models presented here suggest what those limits can be expected to be.
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Chapter 1

Introduction

1.1 General Background

In the recent years, the metal-insulator-semiconductor (MIS) tunnel junction has played an important role in the development of new device structures. One very promising application of MIS tunneling devices is in solar cells. It has been shown that this structure gives a higher open-circuit photovoltage than that of the Schottky barrier solar cell, while maintaining the attractive features that make the Schottky junction potentially more suited to very large area, terrestrial, solar cell applications [1], than the p-n junction.

Another application of tunnel oxide structures is found in negative resistance devices such as the MIS-switch [2]. This device has a metal-tunnel oxide-n-p+ structure. The p-n junction is biased in the forward direction. In the high impedance state there is a wide depletion layer at the oxide-semiconductor in-
terface, and the current in the device is low. In the low impedance state there is
a significant inversion layer charge formed by holes at the oxide-semiconductor
interface, leading to a high field strength in the oxide, which in turn enhances
the electron current tunneling from the metal into the conduction band of the
semiconductor. This device has potential as a static memory element since the
two distinct states allow it to perform the memory storage function which nor-
manly requires a bistable circuit. However, the switching times are calculated
as being too long for high speed usage [3].

Tunnel oxide structures have also found application in the design of bipolar
current amplifiers. Recently three such devices, namely, the tunnel emitter
transistor (TETRAN), MIS-emitter transistor (MISET) and polysilicon emitter
transistor (PET) have been reported. The structure of a TETRAN [3] is shown
in Fig. 1.1. Without an excess supply of minority carriers a deep depletion
layer develops at the oxide-semiconductor interface because the holes generated
thermally in this layer can tunnel through the oxide. Most of the voltage
is dropped across the depletion layer while the field strength in the oxide is
low, and there is a small flow of electrons tunneling from the metal into the
semiconductor conduction band. Increasing the supply of holes by injection
through the $p^+$ diffusion region results in the formation of a hole inversion layer
at the interface, which in turn increases the voltage drop across the oxide and
the electron current tunneling from the metal into the semiconductor. Modest
Figure 1.1: Structure of a TETRAN
gains of around 120 have been measured in operational devices. Of greater interest is the prediction of an intrinsic cut-off frequency as high as 600 GHz [4].

The structure of the MISET is shown in Fig. 1.2. The electron to hole injection ratio of the MIS junction forming the emitter can be quite high, if the metal has a low work function, as for example in the case for aluminium. The electrons, injected by the metal emitter, are collected with the aid of the reversed biased p-n junction forming the collector. Green and Godfrey [5] have recently reported a MISET device with common emitter current gain approaching 25000, which is extremely high for a bipolar device. Another advantage of this device is the simple structure, thus it has small lateral and vertical extensions as well as fewer preparation steps than conventional BJTs. However, other aspects of this device, such as the degradation of the ultra-thin tunnel oxide (~20Å) [6] and the high frequency performance need to be studied further before the practicality of this device can be assessed.

The replacement of the metal emitter in the MISET by a polysilicon emitter can result in a more stable structure, because now the thin tunnel oxide is sandwiched between silicon layers. This arrangement is likely to lead to less degradation and create lower surface states than when contacted with metal on one side. A typical polysilicon emitter transistor structure is shown
Figure 1.2: Structure of a MISET
in Fig. 1.3. The thickness of the tunnel oxide can greatly affect the electrical characteristics of the devices [7]. There are two type of device: one with an intentional, chemically-grown interface oxide (≈20-30Å) [8], and one with a “clean” interface [9,10] at which there is an unavoidable native oxide (≈5Å). The first type of device has a higher current gain and a lower (or even negative) temperature coefficient for the current gain than the second type [11]. This behavior can be roughly predicted by using tunneling theory [8] for the first device and minority carrier transport theory [10,12] in polycrystalline silicon for the second device. Eltoukhy and Roulston [13] later proposed a unified theory which could apply to any polysilicon transistor. For the high frequency performance, the second type of device would seem to be superior to the first type. The use of an intentionally grown oxide significantly degrades the high frequency performance capability by increasing the emitter resistance by an order of magnitude [14].

The polysilicon layer can be deposited after the monocrystalline emitter is formed by conventional ion implantation techniques [10], or the polysilicon can itself be used as a source for the emitter diffusion [8,9]. For the device with a “clean” interface, the high temperature polysilicon anneal (e.g. 1000°C/30 min.) degrades the integrity of the native oxide layer and destroys its blocking action [14], thus increasing the base current and reducing the gain of the device. The best alternative for avoiding this problem is the use of in-situ doped polysilicon.
Figure 1.3: Structure of a npn PET
Rowlandson and Tarr [15] report a “true” polysilicon emitter transistor, formed by depositing in-situ doped polysilicon at a temperature (627°C) which is low enough to prevent diffusion into the substrate, and by not subjecting the device to any additional annealing. Therefore, in this structure the emitter-base junction coincides with the interface between polycrystalline and monocrystalline material. A common emitter current gain as high as 10000 has been reported for this device.

The major application of MIS and semiconductor-insulator-semiconductor (SIS) junctions has already been mentioned above. However, there are some other interesting applications. Moravvej-Farshi and Green [16] have replaced the normally implanted source and drain regions of NMOS transistors by SIS diodes to produce devices with near-zero junction depth, which is a critical parameter in determining the onset of short-channel effects in MOSFETs. Unfortunately, a self-aligned structure has not yet been fabricated.

Fossum and Barker [17] have attempted to use a MIS junction to serve as a charge packet threshold detector. A charge-coupled device (CCD) input structure was designed to inject a metered packet of holes into the MIS junction. The MIS junction may switch to a “high” current state or stay at a “low” state, depending on the amount of the charge injected. However, the slow switching speed (10-100 ms) of this device may prevent it from being very useful.
Because of the interest in the MIS tunnel junction, several models to aid in device analysis and design have been formulated by different researchers. The first general model was that of Green and Shewchun [44]. It has been shown that the results of this comprehensive numerical model can be duplicated for a wide variety of operating conditions by a much simpler analytical model [22]. This latter model has been incorporated in the popular semiconductor device analysis program SEDAN III [55]. One of the simplifying features of the analytic model in [22] is that the electron and hole tunnel currents can be derived assuming constant tunneling probability factors. This approximation is valid so long as the semiconductor surface is not strongly inverted and the metal Fermi level lies within the range of energies defined by the semiconductor band gap at the semiconductor surface. These conditions may not hold in some tunnel junction applications. Another feature of the original analytical model [22] is its utilization of a two-band representation of the thin $SiO_x$ insulator. Recent work [30], which compares experimental data on electron and hole barrier heights [24] with theoretical predictions, suggests that a one-band model of the oxide is more appropriate for analysis of tunneling phenomena in MIS junctions. However, until the work described in this thesis, this representation of the oxide had not been incorporated in models to explain the characteristics of MIS or SIS devices.
1.2 Objective

Many interesting applications of MIS and SIS tunnel junctions incorporating thin oxide layers on silicon have been proposed and, in some case, demonstrated. The particular area that is presently attracting much attention is the application of MIS and SIS junctions in the design of small-sized, easily-fabricated and high speed silicon bipolar transistors. Of special interest is the modeling and characterization of the TETRAN [4], MISET [5] and PET [9,15].

The TETRAN and its gallium arsenide counterpart, the inversion-base bipolar transistor (IBT) [64], represent a completely new type of device. Unlike the conventional bipolar transistor there is no doped neutral base region. The base is actually formed by a very thin inversion layer of mobile holes induced by the emitter; consequently, all of the problems associated with the diffusion of minority electrons in a neutral base region are eliminated. Furthermore, the device is expected to operate at extremely high frequencies because the base transit time is essentially zero. This is on account of the very high field in the induced base which assists the passage of electrons to the collector.

The MISET can exhibit very high common emitter current gain; values up to 25000 have already been reported [5]. This is because the MIS contact behaves as a heterojunction, which causes the tunneling of the conduction-
band electrons between the metal and silicon to be greatly favored over that of valence-band holes. In this respect the silicon MISET is similar to the GaAs heterojunction emitter transistor [56]. Another interesting property of heterojunction transistors is the removal of the necessity to trade-off the transconductance ($g_m$) against the base resistance ($R_b$) to obtain a high maximum oscillation frequency ($f_{max}$). This property still needs to be demonstrated in the MISET.

The npn PET can be optimized to obtain high d.c. gain (~10000) [15] or high cutoff frequency (~16 GHz) [58], depending on the processing treatment of the emitter-base contact. The gain-speed trade-off for npn PETs with different tunnel oxide thicknesses, base-emitter junction depths, base and emitter doping concentrations has been studied [59] by numerical simulations and experimental measurements. Recently, Maritan and Tarr [70] have fabricated some pnp PETs and shown that these devices can exhibit reasonable values of gain and emitter series resistance. This is an extraordinary result as one would expect that the pnp device performance would be very poor, due to the fact that the probability of holes tunneling through the thin oxide is considerably less than that of electrons.

For the design, understanding and development of MIS and SIS tunnel junction devices, an accurate model of the structure is needed. The primary goal of this work is to refine and improve the analytical MIS model of [22]
by representing the oxide via a one-band model, and by allowing for energy-dependent tunneling probability factors for electrons and holes. The new model is used to predict the d.c. characteristics of TETRANs, MISETs and PETs, and to make comparisons with the published experimental results for these devices. Some important a.c. model parameters are extracted in order to evaluate the high frequency performance of these devices. The interesting gain-speed trade-off in MISETs and the surprisingly good performance of pnp PETs are investigated by computer simulations.

1.3 Thesis Outline

Chapter One introduces the background to, and current work on, MIS and SIS devices. The necessity of developing an improved tunneling model is emphasized.

The first two sections of Chapter Two summarize the mathematical formulation of two major tunneling mechanisms in MIS tunnel junctions, namely band-to-band direct tunneling and surface-state tunneling. A key parameter in modeling tunnel junctions is the tunneling barrier height. The last section surveys the different experimental methods of determining the tunneling barrier height.

Chapter Three details the improvement of an earlier analytical model of
the MIS tunnel junction. The improvements include: (1) replacement of the
conventional two-band formulation of the oxide band structure with a one-band
representation, (2) making allowance for the energy dependence of the tunneling
probabilities for electrons and holes, (3) extending the previous model so that it
can deal with three terminal transistor structures, rather than only two-terminal
diode structures.

Chapter Four is devoted to the use of the new model to predict both
the d.c. and the small-signal a.c. performance of the TETRAN. The predicted
characteristics are compared with the published experimental data. The inap-
propriateness of using the previous model to fit the experimental data is
demonstrated.

In Chapter Five, the d.c. and high frequency performance of MISETs
is evaluated and compared to the previous experimental results. The tradeoff between the base conductance and cut-off frequency is studied in order to
optimize the device for high maximum oscillation frequency.

Chapter Six describes the modeling of pnp polysilicon emitter transistors,
and treats both annealed and unannealed devices. The fact that these transis-
tors exhibit reasonable gain and low values of emitter series resistance is shown
to be consistent with the new model proposed. The potential for high frequency
operation of both annealed and unannealed devices is assessed.
Conclusions are drawn and suggestions for further work are made in the last chapter.
Chapter 2

Basic Theory of Thin Oxide Tunnel Junctions

2.1 Existing Models of Band-To-Band Direct Tunneling

In metal-insulator-metal (MIM), MIS or SIS systems, electrons and holes can tunnel through the thin insulator by three different mechanisms, namely, direct band tunneling, surface state tunneling and trap-assisted tunneling. We will consider the direct band tunneling in detail, since it is generally the dominant mechanism.

For a symmetrical MIM structure (see Fig. 2.1(a)) at 0 K, the electron tunnel current through the insulator is [18, p.553]

\[ J_n = \frac{q}{2\pi \hbar d^2} \left[ \chi_1 \exp(-\gamma \sqrt{\chi_1}) - \chi_2 \exp(-\gamma \sqrt{\chi_2}) \right] \]  

(2.1)
Figure 2.1: Energy band diagrams of MIM junctions. (a) Symmetrical structure. (b) Non-symmetrical structure.
where

\[ \chi_1 = \chi_m - \frac{q\psi_f}{2} \]
\[ \chi_2 = \chi_m + \frac{q\psi_f}{2} \]
\[ \gamma = \frac{4\pi d\sqrt{2m^*_e}}{h} \]

and \( m^*_e \) is the effective mass of electrons in the conduction band of the insulator.

For large applied voltages, \( \psi_f > \chi_m/q \), (2.1) can be simplified to the well-known Fowler-Nordheim equation.

Extending the MIM theory to non-symmetrical structures and non-zero temperatures, Stratton [19] has derived an equation for the electron tunnel current (see Fig. 2.1(b))

\[ J_n = \frac{q^4 m^*_e kT}{h^3} \int_{-\infty}^{\infty} \theta_e(\varepsilon_e) S(\varepsilon_e) d\varepsilon_e \]  

(2.2)

where \( m^*_e \) is the effective transverse mass of electrons in the metal, \( \varepsilon_e \) is defined in Fig. 2.1(b) and \( \theta_e(\varepsilon_e) \) is the tunneling probability, which is energy dependent, and can be expanded using the Taylor's series

\[ \ln[\theta_e(\varepsilon_e)] = -(b_1 + c_1 \varepsilon_e + f_1 \varepsilon_e^2 + \cdots) \]  

(2.3)

where coefficients \( b_1, c_1, f_1, \cdots \) are dependent on the shape of the tunnel barrier. The first two terms are enough if \( c_1 < 1/kT \).

The supply function \( S(\varepsilon_e) \) represents the difference between the number of electrons per second per unit area, having (x-directed) energy in the range
\( E_z \) to \( E_z + dE_z \), incident on opposite sides of the barrier.

\[
S(\varepsilon_z) = \int_0^\infty (f_1 - f_2) dE_t
\]

\[
= \int_0^\infty \left[ \frac{1}{1 + \exp(\frac{E_t - \varepsilon_z}{kT})} - \frac{1}{1 + \exp(\frac{E_t - \varepsilon_z + qV}{kT})} \right] dE_t
\]

\[
= kT \ln \left[ \frac{1 + \exp(\frac{\varepsilon_z}{kT})}{1 + \exp(\frac{\varepsilon_z - qV}{kT})} \right]
\]

(2.4)

Truncating the expression for \( \theta_c \), i.e. (2.3), to its first two terms and putting this into (2.2), gives after much mathematical treatment (and use of (2.4)),

\[
J_n = q \frac{4\pi m^* kT}{c_1 h^3} \exp(-b_1) \left\{ \frac{\pi}{\sin(\pi c_1 kT)} [1 - \exp(-c_1 qV)] - R_1 - R_2 \right\}
\]

(2.5)

where

\[
R_1 = \exp(-c_1 qV) \cdot \ln\left[ \frac{1 + \exp(\frac{\varepsilon_z}{kT})}{1 + \exp(\frac{\varepsilon_z - qV}{kT})} \right]
\]

\[
R_2 = \int_0^{\exp(-c_1 kT)} x^{c_1 kT - 1} \cdot \left[ \frac{1}{1 + x} - \frac{1}{1 + x \exp(qV/kT)} \right] dx
\]

In the case of a metal, \( \zeta \) is usually very large and \( R_1 \) and \( R_2 \) are negligible. \( J_n \) can therefore be approximated by the first term of (2.5). Although this approximation is often used in the case of heavily doped semiconductors [20] (where \( \zeta \) is only several \( kT \)), or even in non-degenerate semiconductors [21] (where \( \zeta < 0 \)), the error caused by this approximation may be significant.

For electron tunneling at non-degenerate semiconductor surfaces [13] (where \( \zeta \ll 0 \)), we could use Maxwell-Boltzmann statistics rather than Fermi-Dirac
statistics in deriving the supply function in (2.4). From (2.2) we obtain

\[ J_n = q \frac{4\pi m^*_e (kT)^2}{h^3} \frac{\exp(-b_1)}{1 - c_1 kT} \left[ \exp\left(\frac{\xi}{kT}\right) - \exp\left(\frac{\xi - qV}{kT}\right) \right] \] (2.6)

The same form of expression is also used to describe minority carrier tunneling between semiconductor surfaces and metals [8,20], where Maxwell-Boltzmann statistics can apply.

To model the tunneling current in a MIS system without borrowing from the MIM theory, Tarr et al. [22] derived an expression for the electron tunnel current (see Fig. 2.2(a))

\[ J_n = q \frac{4\pi m^*_e k^2 T^2}{h^3} \theta_e \left[ \mathcal{F}_1 \left( \frac{E_{Fno} - E_{co}}{kT} \right) - \mathcal{F}_1 \left( \frac{E_{Fm} - E_{co}}{kT} \right) \right] \] (2.7)

where \( m^*_e \) is the electron transverse effective mass in the semiconductor. The tunneling probability \( \theta_e \) is assumed to be constant. Except for large tunnel barrier heights \( (\chi_e) \) and small oxide thickness \( (d) \), and the case when the Fermi-levels are well below the conduction band edge, this expression can cause very large errors. For \( E_{Fno} - E_{co} \ll 0 \) and \( E_{Fm} - E_{co} \ll 0 \), the Fermi-Dirac integral in (2.7) can be replaced by the exponential function. After a few mathematical steps, the electron tunnel current can be approximated by [23,24]

\[ J_n = A^* T^2 \theta_e \exp\left( -\frac{q\phi_b}{kT} \right) \left[ \exp\left( \frac{qV}{nkT} \right) - 1 \right] \] (2.8)

where \( A^* \) is the effective Richardson constant \( (= 4\pi q m^*_e k^2 / h^3) \) and \( \phi_b \) is the Schottky barrier height. The ideality factor \( n \) (usually \( \sim 1 - 1.5 \)) is defined as
Figure 2.2: Energy band diagrams for MIS junctions in which (a) $E_{Fm} < E_{co}$ and (b) $E_{Fm} > E_{co}$. 
$-V/ \Delta \psi_s$, where $\Delta \psi_s$ is the change in semiconductor surface potential due to the bias.

In a MIS system where the metal Fermi-level rises above the conduction band edge (Fig. 2.2(b)), Simmons and Taylor [3] have suggested an approximate expression for the electron tunnel current which takes the (parabolic) energy dependence of the tunneling probability into account. This expression is in the following form:

$$J_n = A^* T^2 \{(K_1 + K_2) \exp[-\gamma_e (\chi_m)^{3/4}] - K_3 \exp[-\gamma_e (\chi_e)^{3/4}]\} \quad (2.9)$$

where

$$\gamma_e = \frac{4\pi d}{\hbar} (2m^*_e)^{1/4}$$

$$K_1 = \frac{1}{1 + \frac{\gamma_e kT}{2(\chi_e)^{1/4}}}$$

$$K_2 = \frac{2(E_{Fm} - E_{co})(\chi_m)^{3/4}}{3k^2 T^2 \gamma_e}$$

$$K_3 = \frac{2(E_{Fm} - E_{co})(\chi_e)^{3/4}}{3k^2 T^2 \gamma_e}$$

and $m^*_e$ is the effective mass of electrons in the conduction band of the oxide.

The expressions presented so far allow the electron tunnel current to be calculated if the tunneling probability $\theta$ is known. However, there is controversy over the tunneling band structure appropriate to thin silicon oxides, and on the different forms of $\theta_e$. The tunneling probability, using the WKB approximation,
is given by \[ (2.10) \]
\[ \theta = \exp[-2 \int_{x_m}^{x_s} |k_{IX}| dx] \]
where \( x_s, x_m \) are essentially the positions of the oxide interfaces with the semiconductor and metal respectively, and \( k_{IX} \) is the x-component of the complex electron wavevector in the oxide.

In order to evaluate (2.10), knowledge of the band structure in the forbidden gap of the oxide is required. The complex electron wavevector \( k_I \) in the oxide is usually taken to obey the Franz dispersion relationship [26]

\[ \frac{1}{k_i^2} = \frac{\hbar^2}{2m_{ci}^* (E - E_{ci})} + \frac{\hbar^2}{2m_{vi}^* (E_{vi} - E)} \]  
where \( m_{ci}^*, m_{vi}^* \) are the effective masses associated with the conduction and valence bands in the oxide respectively, and \( E_{ci}, E_{vi} \) are the energies of the band edges in the oxide. Once the transverse wavevector \( k_t \) is known, the x-component of wavevector in the oxide is given by

\[ k_{IX} = \sqrt{k_i^2 - k_t^2} \]  
(2.12)

Note that (2.11) can be rewritten as

\[ k_I = \left( \frac{2m_{ci}^*}{\hbar} \right)^{\frac{1}{2}} (E - E_{ci})^{\frac{1}{2}} \left( 1 - \frac{E_{ci} - E}{E_{vi}} \right)^{\frac{1}{2}} \left[ 1 - \left( 1 - \frac{m_{ci}^*}{m_{vi}^*} \right) \frac{E_{ci} - E}{E_{vi}} \right]^{-\frac{1}{2}} \]  
(2.13)

where \( E_{vi} \) is the oxide bandgap \((= E_{ci} - E_{vi})\).

In the past, a two-band model (or Franz-single mass model) has generally been accepted. It assumes that \( m_{ci}^* \) and \( m_{vi}^* \) are equal, and renders (2.13) to
However, recent calculations of the band structure for SiO$_2$ in the $\alpha$-quartz form indicate that $m^*_c \sim 0.5m_e$ ($m_e$ is the electron rest mass) and $m^*_{vi} \sim (5 - 10)m_e$ [28]. Also, Weinberg [29] observes that the conduction band and the top of the valence band originate from very different electronic orbitals, as indicated in Fig. 1 of [28]. It is possible that the conduction band should be connected (in the complex domain) to one of the deeper lying valence bands, rather than to the top of the valence band. This implies that the appropriate value of $E_{pi}$ in (2.13) could be as high as 18 eV. A small ratio of $m^*_c/m^*_{vi}$ or a large $E_{pi}$ both reduce (2.13) to

$$k_I = \frac{(2m^*_c)^{\frac{1}{2}}}{\hbar} (E - E_c)^{\frac{1}{2}} \left(1 - \frac{E_{c} - E}{E_{pi}}\right)^{\frac{1}{4}}$$  \hspace{1cm} (2.14)$$

This is the so-called parabolic or one-band model. O'Neill [30] has shown that the one-band model gives better agreement between estimates of tunnel probability and some experimental results than the two-band model. Eq. (2.15) also suggests that the effective barrier height for holes is higher than that for electrons and that the energy difference is equivalent to one silicon band gap. Fig. 2.3 compares the dispersion curves of the one-band and two-band models. Note that $k_I$ represents the damping factor of electron waves in the oxide. The smaller $k_I$ is, the higher will be the tunneling probability.
Figure 2.3: Dispersion curves for the one-band and two-band representations of the oxide
Combining (2.15) and (2.12), we obtain

\[ k_{lx} = \left(\frac{2m_e^*}{\hbar}\right)^{\frac{1}{2}} (E_x - E_{ei})^{\frac{1}{2}} \]  

(2.16)

where \( E_x \) is the energy associated with the electron momentum in the \( x \)-direction. Putting (2.16) into (2.10), the tunneling probability can be found. Since \( E_{ei} \) can be a function of \( x \), the shape of the oxide potential barrier can affect \( \theta \). Although many forms of barrier, such as parabolic or triangular shape [19], have been considered, it is commonly assumed that the barrier is either rectangular or trapezoidal (see Fig. 2.4). From (2.10), it can be shown that:

for a rectangular barrier,

\[ \theta(E_x) = \exp\left[-\frac{4\pi d}{\hbar} (2m_e^*)^{\frac{1}{2}} (E_{ei} - E_x)^{\frac{1}{2}}\right] \]  

(2.17)

for a trapezoidal barrier,

\[ \theta(E_x) = \exp\left(-\frac{4\pi d}{\hbar} (2m_e^*)^{\frac{1}{2}} \left(\frac{2}{3q\psi_f}\right)(E_{ei} - E_x)^{\frac{1}{2}} - (E_{ei} - q\psi_f - E_x)^{\frac{3}{2}}\right) \]  

(2.18)

for a triangular barrier,

\[ \theta(E_x) = \exp\left[-\frac{4\pi d}{\hbar} (2m_e^*)^{\frac{1}{2}} \left(\frac{2}{3q\psi_f}\right)(E_{ei} - E_x)^{\frac{3}{2}}\right] \]  

(2.19)

Once \( \theta(E_x) \) is known, \( J_n \) in (2.7) can be determined by setting \( \theta_e = \theta(E_{co}) \). Moreover, the coefficients of the Taylor's series in (2.3) can be fixed.

Another factor affecting the tunneling probability \( \theta \) is the effective electron barrier height \( \chi_e = E_{ei} - E_{co} \) (see Fig. 2.2). The values for \( \chi_e \) (or \( \chi_e \)) reported
Figure 2.4: Different types of tunnel barrier. (a) Rectangular. (b) Trapezoidal. (c) Triangular.
in the literature vary widely, e.g. from 0.25 to 3.3 eV for a 25 Å thick oxide [40]. The barrier height also appears to depend upon oxide thickness, being smaller for thinner oxides [32]. This barrier lowering effect is not fully understood at present and may be due to a number of factors, including image forces, surface effects, fixed oxide charge, and the presence of amorphous silicon oxide.

2.2 Surface-State Tunneling

In a MIS system electrons can tunnel through the thin oxide to surface states lying within the forbidden gap at the semiconductor-oxide interface, and then communicate with the bulk by recombination processes [33,34] as shown in Fig. 2.5. The capture ($c_n, c_p$) and thermal emission ($e_n, e_p$) rates for electrons and holes per active trap at the semiconductor surface are given by

\[
\begin{align*}
    c_n &= \frac{1}{\tau_{cn}} = v_{th}\sigma_bn_s \\
    c_p &= \frac{1}{\tau_{cp}} = v_{th}\sigma_pp_s \\
    e_n &= \frac{1}{\tau_{en}} = v_{th}\sigma_n n_1 \\
    e_p &= \frac{1}{\tau_{ep}} = v_{th}\sigma_p p_1
\end{align*}
\]

(2.20) \hspace{1cm} (2.21) \hspace{1cm} (2.22) \hspace{1cm} (2.23)

where $n_s, p_s$ are the electron and hole concentrations at the surface, $\sigma_n$ and $\sigma_p$ are the thermal capture cross sections for electrons and holes, $n_1$ and $p_1$ are defined as $N_e \exp\left(\frac{E_t-E_c}{kT}\right)$ and $N_v \exp\left(\frac{E_v-E_w}{kT}\right)$ respectively, and $v_{th}$ is the carrier thermal velocity. Neglecting trap photoemission [35], the various currents that
Figure 2.5: Surface-state tunneling in a MIS system
enter and leave the trap can be written as

\[ J_{ct} = -qN_t[c_n(1 - f_t) - e_n f_t] \]  \hspace{1cm} (2.24)

\[ J_{vt} = qN_t[c_p f_t - e_n(1 - f_t)] \]  \hspace{1cm} (2.25)

\[ J_{mt} = qN_t \nu_m (f_t - f_m) \]  \hspace{1cm} (2.26)

where \( N_t \) is the density of states at trap energy level \( E_{tr} \), \( f_t \) is the occupancy probability at \( E_t \), \( f_m \) is the occupancy of \( E_t \) in equilibrium with the metal, and \( \nu_m (=1/\tau_t) \) is the tunneling rate.

The occupancy of the interface states is determined by a competition between tunneling transitions to the metal and the capture of carriers from the conduction and valence bands of the semiconductor surface. Under the steady state condition \( J_{ct} + J_{vt} + J_{mt} = 0 \), the surface state occupancy \( f_t \) has a value between \( f_m \) and \( f_s \) and is given by

\[ f_t = \frac{\tau_t f_s + \tau_s f_m}{\tau_t + \tau_s} \]  \hspace{1cm} (2.27)

where

\[ \tau_s = \frac{1}{v_{th} [\sigma_n (n_s + n_1) + \sigma_p (p_s + p_1)]]} \]  \hspace{1cm} (2.28)

and

\[ f_s = \frac{n_s \sigma_n + p_1 \sigma_p}{(n_s + n_1) \sigma_n + (p_s + p_1) \sigma_p} \]  \hspace{1cm} (2.29)

\[ \approx \frac{1}{1 + \exp\left(\frac{E_{tr} - E_{ir}}{kT}\right)} \text{ if } p_s \gg n_s \]

\[ \approx \frac{1}{1 + \exp\left(\frac{E_{ir} - E_{ir}}{kT}\right)} \text{ if } n_s \gg p_s \]
Putting (2.27) into (2.26), the surface state tunneling current via a single level trap is given by

\[ J_{mt} = qN_{t}\left(\frac{f_s - f_m}{\tau_s + \tau_t}\right) \]  

(2.30)

For \( \tau_s \gg \tau_t \), e.g. in the case of oxide thickness < 15\AA, (2.27) simplifies to \( f_t \approx f_m \). Interface states and metal are in equilibrium, the metal Fermi level is pinned to the energy level of the surface states, and \( J_{mt} \) is controlled by the interface recombination time \( \tau_s \). We call this the interface recombination controlled case.

For \( \tau_t \gg \tau_s \), e.g. in the case of oxide thickness > 25\AA, (2.27) reduces to \( f_t \approx f_s \), the majority carrier quasi-Fermi level at the semiconductor surface is pinned to the states, and \( J_{mt} \) is controlled by the time constant for tunneling through the oxide \( \tau_t \). We call this the oxide tunneling controlled case.

The tunneling rate can be defined as [33,36]

\[ \nu_m = \frac{1}{\tau_t} = \nu_m(E_{tr})\theta(E_{tr}) \]  

(2.31)

where \( \nu_m(E_{tr}) \) is the attempt-to-escape frequency at the trap energy level \( E_{tr} \), and \( \theta(E_{tr}) \) is the tunneling probability at \( E_{tr} \). A variety of experimental methods have yielded values of \( \nu_m \) in the range from \( 10^3 \) to \( 10^{13} \) Hz. Recently, Jain and Dahlke [36], by measuring the photo- and dark capacitance transients in Cr-SiO\(_2\)-nSi MIS tunnel diodes and assuming a two-band model, found that \( \nu_m \) decreases from about \( 10^{18} \) Hz at midgap to \( 10^{10} \) Hz at the silicon conduction band edge.
The surface state density distribution $N_s(E)$ within the semiconductor bandgap and the carrier capture cross section areas $\sigma_n, \sigma_p$ can be estimated by small signal capacitance and conductance techniques [37,38]. Consider a metal-SiO$_2$-pSi tunnel diode as shown in Fig. 2.6, the applied bias $V_B$ is varied such that the hole Fermi level $E_{Fp}$ can scan across the whole Si bandgap. With each biasing point, the a.c. capacitance $C_m$ and conductance $G_m$ of the diode are measured at different frequencies. From the capacitance and conductance dispersion curve, useful surface state parameters can be extracted.

The small-signal equivalent circuit for a MIS diode is shown in Fig. 2.7(a). For oxide thickness $> 20\text{Å}$, the recombination currents are much larger than the tunneling current $J_{mt}$ under the application of an a.c. signal. The circuit
Figure 2.7: Small-signal equivalent circuits of a MIS diode. Steps (a)-(d) represent different stages of simplification.
in Fig. 2.7(a) can be reduced to (b), and subsequently to (c) and (d). Some typical measured capacitance and conductance \( (C_m(\psi, \omega), G_m(\psi, \omega)) \) curves [37] are shown in Fig. 2.8.

Once \( C_m \) and \( G_m \) are known, \( C_d \) and \( G_d \) can be deduced from Fig. 2.7(c) and (d) providing the bulk resistance \( R_B \) is given. The surface state capacitance \( C_s \) is obtained by subtracting the calculated space charge capacitance \( C_{sc} \) from \( C_d \) at equilibrium (or very low) frequency \( \omega_{eq} \):

\[
C_s(\psi_s) = C_d(\psi, \omega_{eq}) - C_{sc}(\psi_s)
\]

Then, the interface state density, \( N_t = C_s/q \), is found from \( C_s \) and its energy position in the Si bandgap. From the conductance, the distribution \( N_t(E) \) can be independently determined by setting

\[
N_t = \frac{2}{q} \left[ G_d(\psi, \omega)/\omega \right]_{\text{max}}
\]

The interface recombination constant \( r_s \) can be determined by the fact that the maxima occurs at \( \omega r_s = 1 \). From (2.28), the majority carrier capture area \( (\sigma_p \text{ or } \sigma_n) \) can also be found.

The surface state distribution depends on the type of metal used and the processing steps. For example, magnesium exhibits one peak in the center of the bandgap, while gold exhibits a large peak near the conduction band and a small peak near the valence band [39]. An annealing step also tends to increase
Figure 2.8: Typical capacitance (a) and conductance (b) curves of a MIS tunnel diode [37].
the surface state density. This arises most probably from thermally stimulated migration of the metal into the oxide. The typical values of \( N_t \) range from \( 10^{11} \) to \( 10^{13} \) cm\(^{-2}\) eV\(^{-1}\), and are greatly affected by metal diffusion through the oxide film. The upper limit of \( N_t \) can possibly be due to a limited solubility of metals in the oxide film. The surface states are usually found to be acceptor-like. The typical values of \( \sigma_n, \sigma_p \) are found to be \( 10^{13} \) cm\(^2\), although a spread of several orders of magnitude is possible [36,37].

### 2.3 The Determination of Barrier Height

There are several methods to determine the tunnel barrier energies. Dressendorfer and Barker [40] have measured the Si-SiO\(_2\) and Al-SiO\(_2\) electron barrier heights on oxides of thickness \( \sim 50 \) Å by photoemission measurements. Since the current across the oxide consists of a background tunneling current and a much smaller photoemissive current, a slow light-on/light-off cycle is needed to resolve these two components. The currents through the device under illuminated and dark conditions are time averaged and subtracted to find the photoemissive component. The typical relation between the square root of the photoyield and photon energy is shown in Fig. 2.9.

There are three distinct sections of the curves: the portion from \( \phi_{T1} \) to \( \phi_{T2} \) caused by photoemission from the conduction band of the silicon; the linear
Figure 2.9: Typical result of photoemission measurements on a MIS diode [40].
section from $\phi_{T2}$ onwards; and the steepest section at higher energies. These latter two portions of the curves are caused by indirect and direct absorption processes from the silicon valence band. The $\phi_{T2}$, which is the barrier between the Si valence band and SiO$_2$ conduction band, is obtained to be about 3.6 eV for a slightly forward biased MIS junction with degenerate n-substrate and oxide thickness about 56 Å. This implies the average electron barrier $\bar{X}_e$ on the Si side (see Fig. 2.2 for definition) is about 2.5 eV, assuming a Si bandgap of 1.1 eV.

For the same device under reverse bias, $\phi_{T1}$ is found to be about 2.6 eV. This value can be understood as the average electron barrier $\bar{X}_m$ on the metal side. The difference from $\chi_m(\sim 3.2$ eV) and $\chi_e(\sim 3.1$ eV) measured with much thicker oxide (>300 Å) suggests that a barrier lowering correction of about 0.5 eV must be included. This technique cannot be used to determine the barrier heights of thin oxide <40 Å, because of the difficulty of resolving the very small photoemissive current from the large background tunneling current.

Kasprzak et al. [32] have deduced the energy barriers of ultrathin (10-30 Å) SiO$_2$ layers between aluminium and degenerate silicon. The MIM tunneling theory was used to interpret the result (see also Fig. 2.1(b)). The $J_n$-vs-$V$ and $[\partial(\ln J_n)/\partial V]$-vs-$V$ curves from this experiment are shown in Fig. 2.10. Abrupt changes in the slope of $J_n$ versus $V$ should be observed at $V = \chi_e/q$ and
Figure 2.10: The current characteristic (a) and its derivative (b) for a degenerate MIS tunnel diode.
neglect $\chi_m/q$ due to the potential barrier changes from a trapezoid to a triangle (see Fig. 2.4(b) and (c)). The mean barrier height $\bar{\phi_0}$ at $V=0$ can also be obtained from the low-voltage approximated relation

$$\ln(R/d) = C d (m^* \bar{\phi_0})^{1/4}$$

where $R$ is the resistance of the tunnel diode at $V=0$, $C$ is a constant, and $m^*$ is the electron effective mass in the oxide.

The Si-SiO$_2$ barrier height ($\chi_e$) has been found to increase from 0.42 eV at 10 Å to 0.65 eV for 25.5 Å of SiO$_2$ on degenerate p-type Si, and from 0.64 eV at 14 Å to 1.27 eV for 29.3 Å of SiO$_2$ on degenerate n-type Si. The Al-SiO$_2$ barrier height ($\chi_m$) is determined to be about 0.61 eV.

In the case of non-degenerate semiconductor, Card et al. [23,41] have deduced the Si-SiO$_2$ barrier height by fitting the experimental tunnel current at forward bias with the I-V characteristics predicted by MIS theory (see Fig. 2.2(a)). Refering to (2.8), $\phi_s$ can be measured by the extrapolation to the $V$-axis of the (linear) $C^{-1}$-$V$ plot, and $n$ can be obtained from the slope of the [$\ln(J_n)$]-vs-$V$ curve. Therefore, the unknown $\theta_e$ is determined, and $\chi_e$ is found from the relation $\theta_e \sim \exp(-\chi_e^{1/2}/d)$.

Kumar and Dahlke [42] have proposed another curve fitting technique. For the MIS system ($\chi_e = \chi_m = \chi$; large surface state density) shown in Fig. 2.11, the tunnel current can be expressed as
Figure 2.11: Energy band diagrams of a MIS diode at (a) reverse and (b) forward bias, from [42]. Note that the field reversal in (a) is caused by a large negative charge density at the interface.
\[ J_n = A^*T^2 \theta_e \left[ \mathcal{F}_1(-\frac{qV_s}{kT}) - \mathcal{F}_1(-\frac{qV_m}{kT}) \right] \] (2.35)

where

\[ \ln(\theta_e) = -\chi^{1/2}d \left\{ \begin{array}{ll}
(1 - \frac{qV_m}{2\chi})^{1/2} & \text{if } qV_m \leq \chi \\
\frac{\chi}{\sqrt{2qV_m}} & \text{if } qV_m > \chi
\end{array} \right. \]

The measured forward and reverse currents, and the surface Fermi potentials \( V_s \) and \( V_m \) obtained from the measured surface potential \( \psi_s(V) \), are used to calculate the barrier function \( U(V_m) \equiv \ln(A^*T^2 \theta_e) \). The values of \( U \) as functions of \( V_m^{-1} \) and \( V_m \) are shown in Fig. 2.12. The slopes \( s_1 \equiv \chi^{3/2}d/\sqrt{2} \) of \( U(V_m^{-1}) \) at large \( V_m \) and \( s_2 \equiv d/(4\chi^{1/2}) \) of \( U(V_m) \) at small \( V_m \), and the difference of their intercepts with the ordinate \( s_3 \equiv \chi^{1/2}d \) are also indicated. By finding the values of \( s_1, s_2 \) and \( s_3 \), and the intercepts of the slope lines, the parameters \( \chi, d \) and \( A^*T^2 \) can be obtained simultaneously.

A Richardson plot provides an alternate means of determining the barrier height. Ashok et al. [43] have analyzed the I-V characteristics of Au-nGaAs MIS diodes, and the forward current is found to obey the relation

\[ J_n = J_s[\exp\left(\frac{qV}{nkT}\right) - 1] \] (2.36)

\( J_s \) is the extrapolated saturation current and is related to the metal-semiconductor barrier height \( \phi_b \) by

\[ J_s = A^*T^2 \theta_e \exp\left(-\frac{\phi_b}{nkT}\right) \] (2.37)

The inclusion of the ideality factor in (2.37) has been found necessary to explain
Figure 2.12: Curve fitting technique proposed by Kumar and Dahlke [42].
the experimental data. The Richardson plot \( \ln(J_s/T^2) \)-vs-\( 1/nT \) results in a straight line, and the values of \( \phi_b \) and \( A^*\theta_e \) are obtained from the slope and the y-intercept. Once \( \theta_e \) is known, the tunneling barrier height can be found.

Realizing that the previous studies of barrier height have been confined to the tunneling of electrons, Ng and Card [24] proposed a procedure to determine the tunneling barrier of holes for Au-SiO\(_2\)-nSi tunnel junctions with oxide thickness in the range of 20-30 Å. The energy band diagram of the short-circuit junction under optical illumination is shown in Fig. 2.13. For an oxide thickness > 20 Å, a suppression of the short-circuit photocurrent is observed and the hole concentration at the semiconductor surface increases with illumination intensity. In the steady state, the photocurrent \( J_{ph} \) is equal to the summation of the short-circuit hole tunnel current \( J_{sc} \) and the back diffusion current \( J_d \). One can formulate an explicit expression for the hole tunneling probability \( \chi_h \) as

\[
\exp(-\chi_h^{1/2}d) = qD_p\frac{4\pi m^*qk^2T^2}{\hbar^3N_0}L_p\exp\left[\left(\frac{q\psi_0}{kT}\right)\left(\frac{J_{ph}}{J_{sc}} - 1\right)\right]^{-1}
\]

(2.38)

The magnitudes of \( J_{ph} \) and \( J_{sc} \) can be obtained from the J-V characteristics of the diode under dark and illuminated conditions as shown in Fig. 2.14.
Figure 2.13: Energy band diagram of a short-circuit MIS tunnel diode under optical illumination.
Figure 2.14: The J-V characteristics of a MIS tunnel diode under dark and illuminated conditions.
Chapter 3

The Formulation of an Improved Tunnel Junction Model

The analytic model formulated by Tarr et al. [22] is commonly used to calculate the steady-state current in a metal-insulator-semiconductor junction. In this chapter two improvements to this model are proposed. Firstly, the two-band model for $SiO_2$ based on the Franz dispersion relation is replaced by a one-band model. Secondly, the restriction of the tunneling probability factor to a constant value is removed.

3.1 The One-Band Model

The motivation for switching to a one-band model is the success achieved by O’Neill [30] in using it to explain the asymmetry in electron and hole tunnel currents in thin $SiO_2$ layers. In the two-band model, the band structure of $SiO_2$
barrier is assumed to have the shape shown in Fig. 3.1(a), where the complex conduction and valence bands extend into the tunneling energy range. Therefore the evanescent states in the oxide are derived from both the conduction and valence bands.

Recent experimental data on the band structure of SiO₂ in the α-quartz form suggests that the one-band model is more appropriate, and that only the complex conduction band can extend into the energy band gap at reasonably small values of $k_f$ (Fig. 3.1(b)). This constitutes the central postulate of the one-band model, namely: that the evanescent states in the oxide are derived only from the conduction band edge [30].

The switching from the two-band model to the one-band model affects the way which tunnel barriers are represented in the energy band diagram. The corresponding barrier height for holes, $\chi_h$, depends on the model used to represent the oxide (see Fig. 3.2). For the two-band model we have

$$\chi_h = E_{gi} - E_g - \chi_e$$

where $E_{gi}$ is the oxide band gap, $E_g$ is the semiconductor band gap and $\chi_e$ is the silicon electron barrier height. Therefore, the electron and hole tunneling probabilities depend on two adjustable parameters, i.e. $E_{gi}$ and $\chi_e$.

The situation is different in the one-band model represented by the symbolic band diagram of Fig. 3.2(b) which we propose. The evanescent states in
Figure 3.1: Band structures of the SiO$_2$ barrier assumed in (a) the two-band model and (b) the one-band model.
Figure 3.2: Energy band diagrams for (a) two-band and (b) one-band representations of the oxide.
the oxide are derived from the oxide conduction band only, and the electron and hole barriers are related by:

\[ \chi_h = \chi_e + E_g \]  

(3.2)

The electron and hole tunneling probabilities depend on only one adjustable parameter, i.e. the silicon electron barrier height \( \chi_e \). Note that in Fig. 3.2(b), the lower portion of oxide energy band does not represent the insulator valence band. It is, instead, a reflected version of the top part of the energy band configured in such a way as to give the correct barrier shape for holes. Due to the symmetry, the expressions for hole tunneling probability \( \theta_h \) are similar to the set given by (2.17)-(2.19), with \( E_z, E_{ci} \) and \( \bar{E}_{ci} \) replaced by \( E'_z, E'_{ci} \) and \( \bar{E}'_{ci} \).

As mentioned in Section 2.1, the values for \( \chi_e \) reported in the literature vary widely, and the barrier lowering effect is not fully understood at present. In the tunnel junction model of the past [3,22,44], it has been assumed that the unlowered bulk oxide values for electron and hole barrier height (\( \chi_e \sim 3.2 \) eV, \( \chi_h \sim 4.2 \) eV) and the independence of \( \chi_e \) and \( \chi_h \) with tunnel oxide thickness are valid. As pointed out by [45,46], the predictions of tunneling current based on these assumptions disagree with the experimental results. Therefore, to describe the conduction mechanism in a tunnel junction we adopt the one-band model with only one adjustable parameter, namely \( \chi_e \), whose value may be much smaller than the bulk oxide value of 3.2 eV. The barrier height \( \chi_e \) depends
on the oxide thickness and surface condition, and can only be determined by parameter fitting theoretical results to experimental data.

### 3.2 The New Power Series for Tunneling Current

In Section 2.1, the inaccuracy of using (2.5)-(2.9) to approximate the tunneling currents in MIS junction was discussed. We have developed new power series expressions for $J_{cm}$ and $J_{vm}$ which have several advantages over the previous analytic expressions ((2.7), for example). Firstly, the new expression accounts for the energy dependence of the tunneling probability $\theta$, and can be evaluated given the relation between $\theta$ and $E_s$. This means the expressions can be adapted into both one-band and two-band models and used with any shape of tunneling barrier. Secondly, this single expression can accurately predict current in different operating regimes, and is not restricted to particular relative positions of $E_{Fm}$, $E_{eo}$ and $E_{Fn}$. Thirdly, the series are fast converging and calculate, within the constraints of the WKB approximation, to any desired accuracy depending on the number of terms employed.

Using Harrison's independent electron approach [25], the electron tunneling current can be written as (see Fig. 3.3)
The electron tunneling probability $\theta_c(E_x)$ can be expanded using the Taylor series with respect to the conduction band edge

$$
\theta_c(E_x) = \theta_c(E_{co}) + \theta'_c(E_{co})(E_x - E_{co}) + \frac{\theta''_c(E_{co})}{2!}(E_x - E_{co})^2
$$

$$+ \cdots + \frac{\theta^n_c(E_{co})}{n!}(E_x - E_{co})^n \quad (3.4)
$$

where

$$
\theta^n_c = \left. \frac{d^n \theta_c}{dE^n_c} \right|_{E_x=E_{co}}
$$

Substituting (3.4) into (3.3), the electron tunnel current can be written as

$$
J_{cm} = \frac{4\pi q m^*_c}{\hbar^3} \sum_{n=0,1,2}^\infty (I_{nn} - I_{mn}) \quad (3.5)
$$

where

$$
I_{nj} = \int_{E_{co}}^{E_{ci}} \left[ \int_0^\infty \frac{dE_t}{1 + \exp\left(\frac{E_t+E_{x}-E_{xn}}{kT}\right)} \right] \frac{\theta'_c(E_{co})}{j!} (E_x - E_{co})^j dE_x
$$

and $I_{mj}$ is the same with $E_{Fm}$ in place of $E_{Fn}$.

Note that $I_{nj}$ can be further simplified by changing the order of integration.
Figure 3.3: Complete energy band diagram of a MIS diode with tunneling currents indicated.
of the double integral.

\[ I_{nj} = \int_0^\infty \left[ \int_{E_{co}}^{E_{so}} \frac{\theta^i_c(E_{co})}{j!} \cdot \frac{(E_x - E_{co})^j}{1 + \exp\left(\frac{E_x + E_{so} - E_{F_n}}{kT}\right)} \right] dE_t \]

Substituting \( \eta = \frac{E_x - E_{so}}{kT} \), we get

\[ I_{nj} = \int_0^\infty \left[ \int_{E_{co}}^{E_{so}} \frac{\eta^i d\eta}{(kT)^{(j+1)} \theta^i_c(E_{co})} \right] dE_t \]

Since the normalized barrier height \( \frac{E_{so} - E_{so}}{kT} \) is usually very large (~40 for a 1 eV barrier), we could write

\[ I_{nj} = (kT)^{(j+1)} \theta^i_c(E_{co}) \int_0^\infty \left[ \frac{E_{F_n} - E_{co} - E_t}{kT} \right] dE_t \]

\[ = (kT)^{(j+2)} \theta^i_c(E_{co}) \mathcal{F}_{j+1}\left(\frac{E_{F_n} - E_{co}}{kT}\right) \]  

(3.6)

Substituting \( I_{nj} \) and \( I_{mj} \) back into (3.5), we obtain

\[ J_{cm} = \frac{4\pi q m^* k^2 T^2}{\hbar^3} \sum_{n=0,1,2} (kT)^n \theta^n_e(E_{co}) \left[ \mathcal{F}_{n+1}\left(\frac{E_{F_n} - E_{co}}{kT}\right) - \mathcal{F}_{n+1}\left(\frac{E_{F_m} - E_{co}}{kT}\right) \right] \]  

(3.7)

where \( \theta^i_c(E_{co}) \) is defined in (3.4).

Similarly, the hole tunneling current is given by (see also Fig. 3.2)

\[ J_{um} = -\frac{4\pi q m^* k^2 T^2}{\hbar^3} \sum_{n=0,1,2} (kT)^n \theta^n_v(E_{vo}) \left[ \mathcal{F}_{n+1}\left(\frac{E_{vo} - E_{F_P}}{kT}\right) - \mathcal{F}_{n+1}\left(\frac{E_{vo} - E_{F_m}}{kT}\right) \right] \]  

(3.8)

where

\[ \theta^n_v(E_{vo}) = \frac{d^n \theta_v}{d(-E_v')^n} \bigg|_{E_v' = E_{vo}} \]
Notice that the first terms of the power series expressions (3.7-3.8) are exactly equivalent to the $J_{em}$ and $J_{vm}$ expressions obtained in [22]. The importance of the additional terms that have been derived can be readily seen in the MIS system in Fig. 2.2(b) where $E_{Fm} < E_{co}$. Suppose electrons are tunneling from the metal to the semiconductor, and assume that $E_{co} - E_{Fm} \gg 0$, the magnitude of this component of the electron current can be written as (from (3.3))

$$J_{m\to c} = \frac{4\pi q m^* kT}{\hbar^3} \int_{E_{co}}^{E_{ei}} f_0 \left( \frac{E_{Fm} - E_z}{kT} \right) \theta_c(E_z) dE_z$$

(3.9)

The current magnitude can intuitively be understood as the integration of the product of two functions within an interval. The situation can be represented graphically in Fig. 3.4. The tunneling probability $\theta$ is plotted against $E_z$, and the curve is exponentially increasing with $E_z$ (see (2.17)-(2.19)). If we approximate this curve (# 3) using a Taylor series expansion about $E_z = E_{co}$ as in (3.4), then $\theta_1$ is the first term, $\theta_2$ and $\theta_3$ are the summations of the first two and three terms respectively. This implies $\theta_1$ is constant, $\theta_2$ is linear and $\theta_3$ is quadratic, etc. In the case where $E_{Fm1} < E_{co}$ (see curve # 1), the integration of the product of $f_1$ and $\theta_1$ in the interval $[E_{co}, E_{ei}]$ gives a good estimate of $J_{m\to c}$, because of the exponential drop of $f_1$ near $E_z = E_{co}$. This is why the assignment of a constant tunneling probability is satisfactory in this case. As the Fermi level rises above $E_{co}$ to a position $E_{Fm2}$, see curve # 2, the integral of the product of $f_2$ and $\theta$ in the interval $[E_{co}, E_{Fm2}]$ contributes much to $J_{m\to c}$.
Figure 3.4: Graphical representation of the factors appearing in the integrand
of (3.9).
Since \( f_2 \) is flat in this interval, the fact that \( \theta \) is approximated by \( \theta_1 \) can cause large error. We can use \( \theta_2, \theta_3 \), or even higher order curves depending on the accuracy desired. Of course, the more the curve \( f \) shifts to the right, the more terms of the Taylor series will be required, thus more terms of the \( J_{em} \) series (3.7) should be kept.

To investigate how fast the series expression for \( J_{m-\epsilon} \) (3.7) with \( E_{co} - E_{F_n} \gg 0 \) converges, and how accurate the series expressions with one, two, and three terms are compared with the result from numerical integration (3.3) and the expression derived by Simmons (2.9), we present Figs. 3.5 to 3.8. Fig. 3.5 shows the tunnel current from exact numerical integration for six different combinations of oxide thickness, barrier height and shape. We can see that the currents calculated assuming rectangular and trapezoidal barrier only differ by a few \%. Therefore we conclude that the shape of the barrier, whether rectangular or trapezoidal, does not affect the tunneling current much. Figs. 3.6-3.8 show the relative error of the power series expressions and Simmons' expression with respect to the exact numerical solution in the case of a rectangular barrier. Generally, the convergence rate of the series is satisfactory, only three terms are needed to achieve the accuracy obtained by Simmons' expression within the normal operation range where \( E_{F_m} - E_{co} < 25kT \). Note that Simmons' expression is not as good as the series expression in the way that it always underestimates the current at low bias and overestimates the current at high bias, while the
Figure 3.5: Comparison of exact numerical integration of (3.9) for the case of rectangular barriers (solid lines) and trapezoidal barriers (dashed lines).
Figure 3.6: Percentage errors, w.r.t. numerical integration results, of $J_{m\rightarrow e}$ as computed using Simmons' expression and various numbers of terms in the series expression. ($d = 16\lambda, \chi_e = 1.1eV$.)
Figure 3.7: As for Fig. 3.6 but with $d = 16\text{Å}, \chi_e = 2.2eV$. 
Figure 3.8: As for Fig. 3.6 but with $d = 12\,\text{Å}$, $\chi_e = 1.1\,\text{eV}$. 
series expression can predict quite accurately the current at low bias, and by adding more terms the current at high bias can be predicted at any desired accuracy.

The insufficiency of Tarr's expression (3.5) also deserves our attention. For example, in Fig. 3.6 when $E_{Fm} - E_{co} = 25kT$, the actual current by numerical calculation is about 5 times larger than the prediction from the first term of the series, but only 1.5 times larger than the prediction from the sum of the first three terms. The improvement by adding more terms is pronounced.

By comparing Figs. 3.6-3.8, we notice that the series expression in general predicts the actual current more accurately with increasing barrier height and decreasing oxide thickness. This can intuitively be explained by referring to Fig. 3.4. Increasing the barrier height or decreasing the oxide thickness can reduce the slope of the curve #3 near $E_m - E_{co}$, therefore the truncated series expression can approximate the exponential curve #3 better and give rise to a more accurate result.

### 3.3 The Revised Model Formulation

The revised model for the MIS tunnel junction utilizes a one-band representation of the oxide and allows for the energy dependence of the tunneling probability factors. These two improvements to the original model are included in
the new model by replacing the expressions for $J_{cm}$ and $J_{um}$ ((12a) and (13) in [22]) by (3.7) and (3.8).

The expressions for the electron and hole tunneling probabilities at the band edges ($\theta_{cm}$ and $\theta_{um}$ as derived in Appendix II of [22]) are replaced in the new formulation by the energy-dependent probabilities given in (2.17-2.19). Selection of a rectangular, trapezoidal or triangular barrier is now possible. The tunnel probabilities in (2.17-2.19) are incorporated in the tunnel current expressions via the series expansions of (3.4).

All the other equations in [22] are preserved, except for the replacement of $F_{3/2}$ in the expression for semiconductor charge density ((25) in [22]) by $F_{1/2}$ [49]. This Fermi-Dirac integral, and those required in the evaluation of the tunnel currents, are computed by short-series approximations [60].
Chapter 4

Modeling the TETRAN Device

Historically device designers have sought relentlessly to increase the speed of transistor operation. Initially the bipolar transistor was the superior high speed device while the MOS transistor was more useful for low-speed/low-power applications. With continued scaling however, the MOS transistor has the potential to outperform bipolar transistors for very high speed and very high density circuits. This is because in MOS devices, capacitances are more amenable to scaling and the shrinking of vertical dimensions in the MOSFET (junction depth and oxide thickness) is more easily accomplished than in the BJT (emitter depth and base width). Unfortunately, the endless thrust to ever-decreasing device geometries leads to problems in both junction and MOS transistors. A very serious problem encountered in either device is the punchthrough effect. In the bipolar transistor, punchthrough occurs when the collector merges with the emitter, and in the MOS transistor when the drain and source depletion
regions begin to merge.

Recently, Taylor and Simmons [61,62] have proposed the bipolar inversion-channel field-effect transistor (BICFET) which operates on the principles of inversion in heterojunctions. This is a new bipolar transistor concept which has combined the virtues of the bipolar and MOS concepts. The inversion channel replaces the base in the conventional bipolar (or heterojunction bipolar) device, and all of the problems associated with the neutral base region, which include scattering in a very heavily doped layer and the storage of minority carriers, are eliminated in the BICFET. One of the claims for this device is a value for $f_T$ of around 10000 GHz [62].

Following the announcement of the BICFET, a different structure called the TETRAN (see Fig. 1.1) which operates on the same principle, was reported [3,4]. Note that in [4] the device is referred to as a BICFET. This could be confusing because in the BICFET described in [61,62] the region of the emitter is a thick, wide band gap semiconductor rather than a thin tunnel oxide, and the current transport through this region is deemed to be by diffusion or thermionic emission rather than by tunneling. The BICFET of [4] is, in our terminology, a TETRAN.

The operating principle of the TETRAN is quite simple. Application of a reverse bias voltage to the metal emitter leads to depletion of the underlying
semiconductor, see Fig. 4.1. Holes can be injected into the depletion region from the \( p^+ \) contact, the source, so inverting the surface region of the semiconductor underneath the emitter. This increase in hole concentration leads to a redistribution of the voltage drops across the emitter insulator and the depletion region, see Fig. 4.1(b). The increased field in the insulator paves the way for an increase in the electron tunnel current. There will be some tunneling of holes from the semiconductor to the metal but, if this current is less than the enhancement in electron current, the structure will exhibit current gain. Modest gains of around 120 have been measured in operational devices [4]. The d.c. characteristics of this device will be analysed in Section 4.2. Moravvej-Farshi and Green [4] also suggest that, in the TETRAN, the transconductance is lower and the input capacitance is higher than in a BICFET, and they hint that the intrinsic cut-off frequency for the TETRAN is about 600 GHz. This is still a sensational figure, one that has prompted the work in Section 4.3. This, to the present author’s knowledge, is the first detailed analysis of the high frequency capabilities of the TETRAN.

### 4.1 Model Formulation

The basis for the model of the TETRAN is the improved analytical model for the MIS tunnel junction discussed in Chapter 3. This model has been modified to
Figure 4.1: Effect of source current on the potential distribution and charge flow in the MIS junction. (a) $I_s = 0$. (b) $I_s > 0$. 
accommodate a third electrode (the source) and to improve the characterization of the tunneling process. The latter improvements are twofold: firstly, the two-band model for $SiO_2$ based on the Franz dispersion relation has been replaced by a one-band model; secondly the restriction of the tunneling probability factor to a constant value [22] has been removed. The detailed energy band diagram of the device is illustrated in Fig. 4.2. The expressions for electron tunnel current $J_{em}$ and hole tunnel current $J_{vm}$ are given in (3.7) and (3.8), and the electron and hole barriers ($\chi_e$ and $\chi_h$) are related by (3.2). In the model, $\chi_e$ is taken as an adjustable fitting parameter. Any solution for the tunnel current $J_{em}$ and $J_{vm}$ must be consistent with the stipulations of Kirchoff's laws. Regarding voltages, summation of the potential drops shown in Fig. 4.2 yields:

$$\phi_o + \psi_s + \frac{\chi_e}{q} + \psi_I - \frac{\chi_m}{q} - V_{CE} = 0 \quad (4.1)$$

where $V_{CE}$ is the collector-emitter voltage and $\psi_I$ is the potential drop across the oxide given by

$$\psi_I = \frac{d}{\epsilon_I} Q_s \quad (4.2)$$

where $\epsilon_I$ is the oxide permittivity and $Q_s$, the total charge in the semiconductor, is derived in Appendix B of [49] as:

$$Q_s = \text{sgn}(\psi_s) \sqrt{2kT\epsilon_s |N_e| \int_{1/2} (\frac{E_F - E_o}{kT}) - n_o}$$

$$+ N_o \int_{1/2} (\frac{E_o - E_F}{kT}) - p_o + N_D \frac{q\psi_s}{kT} |1/2 \quad (4.3)$$

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Figure 4.2: Energy band diagram for the TETRAN.
where $n_o$ and $p_o$ are the equilibrium concentrations of electrons and holes respectively.

Considering now the need for continuity of charged particle flows across the interface we have, for the case of holes, neglecting recombination/generation in the depletion region [3], impact ionization effects [17], surface state tunneling [36] and trap-assisted tunneling [63]:

$$J_s = J_{vm} + J_d \quad (4.4)$$

where $J_d$ is the current due to diffusion of holes from the source into the quasi-neutral region of the collector (see Fig. 4.2), and is given by

$$J_d = \frac{qD_p n_o^2}{L_p N_D} [\exp(\frac{q\phi}{kT}) - 1] \quad (4.5)$$

Equations (4.1) and (4.4) are two non-linear equations which need to be solved simultaneously. By substituting equation (3.8) for $J_{vm}$ and (4.5) for $J_d$ into (4.4), and equations (4.2) and (4.3) into (4.1), the equations (4.1) and (4.4) can be written in terms of two independent variables $\psi_s$ and $\phi$. Thus, for given values of $J_s$, $V_{CE}$, $\chi_h$ and $\chi_e$ (see (3.2)) solutions for $\psi_s$ and $\phi$ can be obtained. A standard iterative technique based on a generalized secant method was used for this purpose.

Once $\psi_s$ and $\phi$ are known, the electron tunnel current $J_{em}$ can be com-
puted, as well as the terminal currents:

\[ J_E = J_{em} + J_{um} \]  
\[ J_C = J_{em} - J_d \]  

(4.6)  
(4.7)

A computer program has been written to evaluate the steady-state characteristics of TETRAN devices using numerical methods to solve the equations of the above model. This program is listed in Appendix A.

4.2 DC Characteristics

To test the capabilities of the TETRAN model, d.c. current-voltage characteristics were generated for a device structure resembling that used in the experimental work reported in [4]. The physical parameters of the device were: oxide thickness = 16Å, collector doping density = \(7 \times 10^{14}\) cm\(^{-3}\) and collector thickness = 10\(\mu\)m. The device exhibits a current gain of about 120 at a collector current density of around \(10^3A/cm^2\). To simulate the experimental I-V curve in [4], the tunnel model in [22] was first used, i.e. the two-band model with constant tunneling probability. The J-V characteristics are shown in Fig. 4.3, and it is clear there is a strong disagreement with the experimental results shown in Fig. 2 of [4]. The d.c. gain predicted is only about 5 and the predicted collector current is an order of magnitude below that of the experimental current.
Figure 4.3: Prediction of TETRAN characteristics: \( d = 16\text{Å} \), other model parameters as in Table I of [22]. \( \theta_c \) independent of energy as in [22].
On replacing the two-band model in [22] by the one-band model with a lowered barrier height of 0.8 eV, but still using the constant tunnel probability approximation in the calculation, the J-V characteristics shown in Fig. 4.4 are predicted. The parameters used in the simulation are listed in Table 4.1. The collector current level has been raised an order of magnitude higher than the level in Fig. 4.3. However, the gain declines very rapidly as the source current increases. This is due to the fact that the $J_{em}$ and $J_{vm}$ expressions based on constant tunnel probability deviate increasingly from the actual current as the junction is more reverse biased. This shortcoming of the model can be corrected by using the series expressions for $J_{em}$ and $J_{vm}$, in which case the characteristics in Fig. 4.5 are obtained, assuming $\chi_e = \chi_m = 1.1$ eV. The collector current level is about the same as that of the actual device of [4], and the predicted current

\begin{table}
\begin{tabular}{|c|c|}
\hline
\(T\) & temperature \hline
\(E_g\) & silicon bandgap \hline
\(m_{ei}\) & electron effective mass in oxide \hline
\(m_e\) & electron transverse mass in Si \hline
\(m_{hi}\) & hole transverse mass in Si \hline
\(\epsilon_s\) & permittivity of silicon \hline
\(\epsilon_f\) & permittivity of SiO$_2$ \hline
\(\mu_p\) & hole mobility \hline
\(\tau_p\) & hole lifetime \hline
\(n_i\) & intrinsic carrier concentration \hline
\(N_c\) & conduction band density of states \hline
\(N_v\) & valence band density of states \hline
\hline
300K & \hline
1.12eV & \hline
0.5m_e & \hline
0.2m_e & \hline
0.66m_e & \hline
11.9\epsilon_o & \hline
3.9\epsilon_o & \hline
480cm^2V^{-1}s^{-1} & \hline
8\mu s & \hline
1.45 \times 10^{10}cm^{-3} & \hline
2.8 \times 10^{19}cm^{-3} & \hline
1.04 \times 10^{10}cm^{-3} & \hline
\end{tabular}
\end{table}

Table 4.1: Model parameter values for simulation of the TETRAN.
Figure 4.4: Prediction of TETRAN characteristics: same parameters as for Fig. 4.3, except for the use of a one-band representation of the oxide. $\theta_e$ given by one term of (3.4) with $\chi_e = 0.8eV$. 

$J_c \times 10^3 \text{A cm}^{-2}$
Figure 4.5: Prediction of TETRAN characteristics: same parameters as for Fig. 4.4, except for the use of three terms in (3.4) for $\theta_c$, and $\chi_c = 1.1eV$. 

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gain $\Delta J_c/\Delta J_s$ is 100 at a collector current density of around $2 \times 10^3 \text{Acm}^{-3}$. This is in close agreement with the value of 120 measured by Moravvej-Farshi and Green [4] at similar collector current densities. As would be expected, the predicted current gain is sensitive to the value chosen for $\chi_e$. For example, reducing $\chi_e$ to 1.0eV leads to about a 30% increase in gain.

A unique feature of the TETRAN I-V characteristic is the reversal in polarity of the collector current at low collector-emitter voltages. The effect is clearly visible in Fig. 4.5. The value of $V_{CE}$ at which the current reversal occurs is called the cut-in voltage. The predicted value of 0.4V is close to the measured value of 0.6V [4]. No doubt even better agreement could be achieved by adjusting the value of the hole lifetime used to compute $L_p$ in the expression for $J_d$ (4.5), or by including in the model effects due to other currents, e.g., recombination/generation in the depletion region. In the present model the collector current at voltages below the cut-in voltage is due to the dominance of $J_d$ over $J_{em}$.

Turning attention to the low gain region at low source current, it is suggested that the neglect of surface-state tunneling is responsible for the experimental data not being modeled very satisfactorily. By incorporating into the TETRAN model the surface-state tunneling model of Section 2.2, the J-V curve shown in Fig. 4.6 is predicted. It is assumed that all the surface-
Figure 4.6: Prediction of TETRAN characteristics: same parameters as for Fig. 4.5, except for the inclusion of the surface-state tunneling effect, and $\chi_s = 0.85\,eV$. The dashed lines represent experimental curves from [4].
states are acceptor-like. The electron and hole capture cross sections are $\sigma_n$, $\sigma_p \sim 5 \times 10^{-18} \text{cm}^2$, the surface state density is $N_s \sim 5 \times 10^{12} \text{cm}^{-2}$ and the tunneling time constant is $\tau_m \sim 1.2 \times 10^{-8} \text{s}$. All these parameters lie within the range of values obtained from previous experimental work [36,37]. When surface-state tunneling is included, there is an additional leakage charge flow from the hole inversion region in the semiconductor surface to the metal, thus reducing the gain of the TETRAN device. Therefore a lower barrier height of 0.85 eV is needed to fit the experimental curve in Fig. 2 of [4]. The predicted J-V curve agrees closely with the experimental result. The cut-in voltage is found to be 0.5 V, and the maximum gain is about 115.

Another interesting correlation of predicted and experimental data is the effect of the passage of high collector currents. In [4] irreversible reduction of current gain was observed in devices for which $J_C$ was taken above $10^4 \text{Acm}^{-2}$. This effect was attributed to the generation of surface states at the oxide-semiconductor interface. The results of the proposed model suggest an alternative mechanism. The calculations indicate that at these current levels the voltage drop across the oxide is about 0.55V. This corresponds to an oxide field strength of $3.5 \times 10^6 \text{Vcm}^{-1}$, which is about the breakdown field strength of thin oxides [48]. The saturation voltage $V_{CE}$ is observed to be about 1V in the simulation, and is much less than the value of about 5V measured in the experimental results. This discrepancy is most probably due to the emitter contact.
The results of this work on the TETRAN device have been published [49]. The value for electron tunnel barrier height ($\chi_e$) that fits the TETRAN d.c. characteristics best is found to be around 1eV, which is close to the values 0.7 and 0.9eV which have been used by others [30,50] in the analysis of devices with thin tunnel oxides.

4.3 Small-Signal Analysis

The small-signal behaviour of the TETRAN device can be described using the common-emitter hybrid-$\pi$ model which is shown in Fig. 4.7. The parameters are defined as follows:

- **Transconductance** $g_m = \left| \frac{dJ_C}{dV_{SE}} \right|_{V_{CE}}$
- **Common-emitter input capacitance** $C_{se} = \left| \frac{dQ_s}{dV_{SE}} \right|_{V_{CE}}$
- **Collector-source capacitance** $C_{sc} = \left| \frac{dQ_s}{dV_{CE}} \right|_{V_{SE}}$
- **Input resistance** $r_{se} = \left| \frac{dV_{SE}}{dJ_S} \right|_{J_C}$
- **Output resistance** $r_o = \left| \frac{dV_{CE}}{dJ_C} \right|_{J_S}$

The unity current gain cut-off frequency can be expressed as:

$$f_T = \frac{g_m}{2\pi C_{se}\left[1 + 2C_{sc}/C_{se}\right]^{1/2}} \tag{4.8}$$
Figure 4.7: Common-emitter small-signal hybrid-\( \pi \) equivalent circuit for the TETRAN device.
Table 4.2: Simulation results for the small-signal hybrid-\( \pi \) parameters of the TETRAN.

<table>
<thead>
<tr>
<th>( V_{CE} )</th>
<th>( \chi_e )</th>
<th>( J_s )</th>
<th>( g_m ) ((Sm^{-2}))</th>
<th>( C_{se} ) ((Fm^{-2}))</th>
<th>( C_{sc} ) ((Fm^{-2}))</th>
<th>( r_{se} ) ((\Omega m^2))</th>
<th>( r_o ) ((\Omega m^2))</th>
<th>( f_T ) ((GHz))</th>
</tr>
</thead>
<tbody>
<tr>
<td>5V</td>
<td>1.0eV</td>
<td>32Acm(^{-2})</td>
<td>( 2.7 \times 10^8 )</td>
<td>( 1.8 \times 10^{-2} )</td>
<td>( 2.0 \times 10^{-7} )</td>
<td>( 2.0 \times 10^{-7} )</td>
<td>( 1.9 \times 10^{-4} )</td>
<td>2.3</td>
</tr>
<tr>
<td>5V</td>
<td>1.0eV</td>
<td>8Acm(^{-2})</td>
<td>( 0.9 \times 10^8 )</td>
<td>( 2.1 \times 10^{-2} )</td>
<td>( 5.0 \times 10^{-7} )</td>
<td>( 6.5 \times 10^{-7} )</td>
<td>( 1.8 \times 10^{-4} )</td>
<td>0.7</td>
</tr>
<tr>
<td>2V</td>
<td>1.0eV</td>
<td>32Acm(^{-2})</td>
<td>( 1.8 \times 10^8 )</td>
<td>( 1.6 \times 10^{-2} )</td>
<td>( 2.0 \times 10^{-7} )</td>
<td>( 3.0 \times 10^{-7} )</td>
<td>( 1.9 \times 10^{-4} )</td>
<td>1.8</td>
</tr>
<tr>
<td>5V</td>
<td>1.1eV</td>
<td>32Acm(^{-2})</td>
<td>( 1.5 \times 10^8 )</td>
<td>( 1.7 \times 10^{-2} )</td>
<td>( 1.8 \times 10^{-7} )</td>
<td>( 2.0 \times 10^{-7} )</td>
<td>( 3.3 \times 10^{-4} )</td>
<td>1.4</td>
</tr>
</tbody>
</table>

The hybrid-\( \pi \) parameters listed above were computed from the d.c. model (neglecting surface-state tunneling) described in Section 4.2 by examining the changes in terminal currents, terminal voltages and stored charge in the semiconductor in response to small changes in either \( J_C \), \( V_{SE} \) or \( V_{CE} \). The perturbations had a magnitude of 1% of the operative steady-state values.

The results for a range of operating conditions and two different values of electron barrier height, \( \chi_e \), are listed in Table 4.2. With respect to the conditions used to obtain the first row of the Table 4.2, note that a reduction in either \( J_s \) or \( V_{SE} \), or an increase in \( \chi_e \), leads to a reduction in \( f_T \), principally
via a decrease in transconductance. However, the changes are not great and $f_T$ remains in the neighbourhood of 1-2 GHz. This is a far cry from the value of $f_T \approx 600$ GHz previously suggested [4] as being appropriate for a TETRAN of the type under discussion.

Comparing the structure of the BICFET and TETRAN, it is clear that the thin insulator employed in the TETRAN will cause it to have a larger input capacitance than the BICFET. Also, the effective electron barrier height $\chi_e$ is likely to be much larger for a semiconductor-semiconductor interface. Thus $g_m$ for the BICFET should exceed that for the TETRAN. The input capacitance $C_{se}$ usually exceeds $C_{sc}$, certainly in the case of the TETRAN (see Table 4.2), and, therefore (4.8) can be reduced to:

$$f_T = \frac{g_m}{2\pi C_{se}} \quad (4.9)$$

It is quite clear from this equation that the aforementioned differences in $g_m$ and $C_{se}$ will cause $f_T$ in the case of the TETRAN to be inferior to that of the BICFET. The results presented in this section indicate the large extent of this difference in high frequency capability.

Confirmation of the estimate of $f_T$ resulting from the numerical analysis can be obtained by carrying-out an approximate analytical evaluation of $f_T$ via (4.9). This is demonstrated below.

To obtain an estimate for $g_m$ we neglect the contribution of the diffusion
current to $J_C$, see (4.7) and seek an expression for $J_{em}$ which can be readily differentiated. Such an equation appears as equation (A.32) in [3]. Taking the dominant term in this equation and ignoring the voltage dependence of pre-exponential factors we have, in the notation of the present work:

$$J_{em} = K \exp[-\gamma_e (\chi_m - q\psi_I/2)^{1/2}]$$

(4.10)

$J_{em} \approx J_C$ as noted above, and $\psi_I$, the potential drop across the insulator, is the main contributor to $V_{SE}$, thus:

$$g_m = \frac{dJ_e}{dV_{SE}} \approx \frac{dJ_{em}}{d\psi_I} = \left[\frac{q\gamma_e}{4(\chi_m - q\psi_I/2)^{1/2}}\right]J_{em}$$

(4.11)

Taking $\chi_m = 1.1eV$, $d = 16\AA$, $V_{CE}=5$ V and $J_S = 3.2 \times 10^5 \text{Acm}^{-2}$ (i.e. as per row 4 of Table 4.2) we find from the model that $J_C = 3.8 \times 10^7 \text{Am}^{-2}$ and $\psi_I=0.68$ V. Substituting these values into (4.11) gives:

$$g_m = 3.3J_C = 1.1 \times 10^8 \Omega^{-1}m^{-2}$$

As regards the input capacitance, this can be taken as being approximately equal to the capacitance of the ultra-thin emitter oxide, i.e.:

$$C_{se} \approx \frac{\epsilon_f}{d}$$

(4.12)

For $d = 16\AA$ and $\epsilon_f = 3.9\epsilon_0$ we have $C_{se} = 2.2 \times 10^{-2} \text{Fm}^{-2}$. Using these approximate values of $g_m$ and $C_{se}$ in (4.9) yields $f_T=0.8$ GHz. This figure is in good agreement with the results of the numerical analysis given in Table 4.2.
4.4 Summary

Both the d.c. and a.c. performance of the TETRAN device have been carefully studied by numerical simulations. Computer analysis indicates that the TETRAN is a modest gain (~ 100), low current ($J_C$ up to $10^4 A/cm^2$) and reasonably high frequency (~ $2GHz$) device. The performance of the device depends largely on the tunnel oxide parameters. Theoretically, reducing the thickness or barrier height of the tunnel oxide improves the gain and frequency response of the device. However, practically, a stable oxide with a thickness of 16Å in an MIS junction is just about the thinnest that one can obtain. Further reduction of oxide thickness may affect the reliability of the device. Therefore, the performance of the TETRAN is not expected to improve to a large extent by optimizing the oxide parameters, without introducing serious reliability problems.

A possible way of improving the performance of the TETRAN is by changing the emitter material from metal to heavily n-doped polysilicon. Since the Fermi-level in heavily n-doped polysilicon can occur above the conduction band edge, this will increase the emitter current ($J_{em}$), and therefore increase the gain, transconductance and cut-off frequency. Also the SIS junction is likely to have less surface-state traps than the MIS junction. This will increase further the gain of the device, because the leakage current from base to emitter due
to surface-state tunneling will be minimized. With these potential merits the polysilicon TETRAN should be further investigated.
Chapter 5

Modeling the MISET Device

MIS emitter transistors are potentially very useful devices. They are simple to fabricate and have potential for high current gain [5]. They can be viewed as having operating principles similar to heterojunction bipolar transistors. The structure of the MIS emitter transistor is shown in Fig. 1.2. The p-type base region is defined by implanting boron into an n-type epitaxial substrate which forms the collector. The heterojunction emitter is formed by growing a thin thermal oxide (~20Å) on the wafer surface and capping this oxide with a layer of low work function metal such as Mg, or Al. The low work function metal electrostatically induces a thin layer of electrons along the silicon surface underneath it. This causes the tunneling of conduction-band electrons between the metal and silicon to be greatly favored over that of valence-band holes. The first MIS emitter transistor structure was proposed and fabricated by Kisaki [65]. A current gain of 120 and a unity gain cut-off frequency of about 1GHz
were measured. Gains have increased dramatically in recent years. Green and Godfrey [5] have fabricated some operational devices with maximum d.c. gains of 600, 10000, and 25000 with base implant doses $5 \times 10^{12}$, $10^{12}$, and $5 \times 10^{11} \text{cm}^{-2}$ respectively. Theoretical work has not been reported to explain the extremely high gain achieved by these transistors. Also no theoretical (or experimental, for that matter) investigation of the high frequency performance of these transistors has been carried out.

The work described in this chapter seeks to provide a theoretical analysis of the d.c. and high frequency properties of the MISET using the improved tunnel junction model developed in Chapter 3. The theoretical I-V characteristics of the transistors are then compared with the d.c. experimental results in [5], and a general agreement is sought by adjusting the tunnel barrier height. The next step is to consider the effect of small variations of voltages and currents, thus generating a small-signal hybrid-$\pi$ model of the MISET. From the parameters of the a.c. model, it is possible to estimate the unity gain cut-off frequency using a similar approach to that taken in [49]. In order to optimize the transistor for high $f_{\text{max}}$, the trade-off between the base conductance and $f_T$ is also studied.
5.1 Model Formulation

The operation of the MISET is similar to that of the conventional npn BJT in the sense that the supply of holes from the base lead into the p-neutral base region biases the emitter-base potential and controls the flow of electrons from the emitter to the collector. The only difference is that the transport of carriers across the emitter-base junction is through tunneling, rather than diffusion. Since the hole tunneling barrier is always higher by an amount equal to one silicon energy band gap than the electron tunneling barrier (as proposed by the one-band model), the back-injected hole current is greatly reduced and the emitter efficiency is improved. This is analogous to the effect achieved in AlGaAs/GaAs heterojunction devices, where the injected holes are discouraged from diffusing to the emitter by the energy barrier resulting from the larger band gap of the emitter [56].

The energy band diagram of the MISET is shown in Fig. 5.1. All the quantities indicated in the diagram are positive. The two independent variables of the system are $\phi$ and $\psi$. Once $\phi$ and $\psi$ are known, all the potentials and current components in the system can be calculated.

The total charge stored in the semiconductor space-charge region is given by [49]
Figure 5.1: Energy band diagram of a MISET.
\[ Q_0 = -\text{sgn}(\psi_s)(2kT\epsilon_s)^{1/2}[N_C \mathcal{F}_{1/2}(\frac{-q\alpha_2}{kT}) - n_B \exp(\frac{q\phi}{kT}) + N_{\nu} \mathcal{F}_{1/2}(\frac{-q\beta_2}{kT}) - P_B + \frac{qP_B\psi_s}{kT}]^{1/2} \]  

(5.1)

where

\[ q\alpha_2 = E_{gb} - q\phi - q\phi_o - q\psi_s \]

\[ q\beta_2 = q\phi_o + q\psi_s \]

Assuming the absence of fixed charge or surface-state charge at the semiconductor/oxide interface, the potential drop \( \psi_I \) across the oxide is related to the stored charge \( Q_0 \) by

\[ \psi_I = \frac{d}{\epsilon_I} Q_0 \]  

(5.2)

Notice that \( \psi_I \) is a function of the two variables \( \phi \) and \( \psi_s \). As a matter of fact, all the current components in the system are dependent on the variables \( \phi \), \( \psi_s \), or \( \psi_I \). Let's examine each current component in the system. The electron and hole currents are given by the series expressions

\[ J_{em} = A_e^* T^2 \sum_{j=1}^{n}(kT)^j \theta_e^j [\mathcal{F}_{j+1}(\frac{-q\alpha_1}{kT}) - \mathcal{F}_{j+1}(\frac{-q\alpha_2}{kT})] \]  

(5.3)

and

\[ J_{hm} = A_h^* T^2 \sum_{j=1}^{n}(kT)^j \theta_e^j [\mathcal{F}_{j+1}(\frac{-q\beta_2}{kT}) - \mathcal{F}_{j+1}(\frac{-q\beta_1}{kT})] \]  

(5.4)
respectively, where

\[ q\alpha_1 = \chi_m + q\psi_I - \chi_e \]
\[ q\beta_1 = E_{eb} - q\alpha_1 \]

The derivatives of the tunneling probabilities \( \theta'_e \) and \( \theta'_o \) are properly defined in Section 3.2. The effective Richardson constants \( A_e^* \) and \( A_h^* \) are defined as in [22]. Generally, series expressions with three terms will be sufficient to approximate the tunneling currents. However, an exact integration may be needed to accurately compute the electron current under high bias because of the low electron barrier [47,49].

The electron diffusion currents at the depletion edges of the neutral base near the emitter and collector are represented by \( J_{ne} \) and \( J_{nc} \) respectively. They can be given in terms of \( \phi \) and \( V_{CB} \) as

\[
J_{ne} = \frac{qD_{eb}n_{ob}}{L_{eb}} \left\{ \coth \left( \frac{W'_e}{L_{eb}} \right) \left[ \exp \left( \frac{q\phi}{kT} \right) - 1 \right] \right.
- \left. \csch \left( \frac{W'_e}{L_{eb}} \right) \left[ \exp \left( -\frac{qV_{CB}}{kT} \right) - 1 \right] \right\} (5.5)
\]

\[
J_{nc} = \frac{qD_{eb}n_{ob}}{L_{eb}} \left\{ \csch \left( \frac{W'_h}{L_{eb}} \right) \left[ \exp \left( \frac{q\phi}{kT} \right) - 1 \right] \right.
- \left. \coth \left( \frac{W'_h}{L_{eb}} \right) \left[ \exp \left( -\frac{qV_{CB}}{kT} \right) - 1 \right] \right\} (5.6)
\]

where

\[ V_{CB} = V_{CE} + \psi_o + \phi_o - E_{eb}/q + \psi_I + \chi_m/q - \chi_e/q \]
\( n_{eb} \) is the equilibrium electron density, \( D_{eb} \) is the electron diffusion constant and 
\( L_{eb} \) is the electron diffusion length in the neutral base. \( W'_{b} \) is the effective base width, which is voltage dependent due to the basewidth modulation effect.

The base recombination current is obtained as

\[
J_{rb} = J_{ne} - J_{nc} = \frac{q D_{eb} n_{ob}}{L_{eb}} \left[ \coth \left( \frac{W'_{b}}{L_{eb}} \right) - \text{csch} \left( \frac{W'_{b}}{L_{eb}} \right) \right] \cdot \left\{ \exp \left( \frac{q \phi}{kT} \right) - 1 \right\} + \left\{ \exp \left( -\frac{q V_{CB}}{kT} \right) - 1 \right\} \tag{5.7}
\]

The diffusion constant \( D_{eb} \) is related to \( \mu_{eb} \) by the Einstein relation

\[
D_{eb} = \frac{kT}{q \mu_{eb}} \tag{5.8}
\]

The mobility of electrons \( \mu_{eb} \) is dependent of the base doping density \( P_B \) and is given by the empirical relation (valid in the range of \( P_B \sim 10^{18} - 10^{20} \text{cm}^{-3} \) ) [57]

\[
\mu_{eb} = 88 + \frac{1.252 \times 10^3}{1 + 6.984 \times 10^{-18} \times P_B (\text{cm}^{-3})} \text{cm}^2\text{V}^{-1}\text{s}^{-1} \tag{5.9}
\]

The electron diffusion length \( L_{eb} \) is given by

\[
L_{eb} = \sqrt{D_{eb} \tau_{eb}} \tag{5.10}
\]

The electron lifetime \( \tau_{eb} \) also depends upon the base doping density \( P_B \), and is normally related experimentally by [55]

\[
\tau_{eb} = \frac{\tau_{eb}}{1 + (P_B/5 \times 10^{16} \text{cm}^{-3})} \tag{5.11}
\]
In the base-collector depletion region, the generation current is expressed as

\[ J_g = \frac{q_n l_{bc}}{\tau_g} \exp\left(\frac{-q V_{CB}}{2kT}\right) - 1 \]  

where \( l_{bc} \) is the depletion width at the base-collector junction and \( \tau_g \) is the generation lifetime of the carriers.

The hole current due to diffusion from the collector to the base is formulated as

\[ J_d = -q \sqrt{\frac{D_{he}}{\tau_{he}}} p_{oc} \exp\left(-\frac{q V_{CB}}{kT}\right) - 1 \]

where \( D_{he} \) is the hole diffusion constant, \( \tau_{he} \) is the hole lifetime and \( p_{oc} \) is the equilibrium hole concentration at the collector.

To obtain \( \phi \) and \( \psi \), when the voltage applied across the collector and emitter \( (V_{CE}) \) and the base current \( (J_B) \) are given, two non-linear equations need to be solved:

\[ J_{ne} - J_{cm} = 0 \]  

\[ J_B + J_g + J_d - J_{vm} - J_{rb} = 0 \]

These equations represent the conditions for the continuity of electron and hole flows in the transistor, and can be solved using a standard iterative technique based on a generalized secant method. Once \( \phi \) and \( \psi \) are known, all the current components can be computed and the terminal currents can be obtained as:

\[ J_E = J_{cm} + J_{vm} \]
\[ J_C = J_{nc} + J_t + J_d \]  

(5.17)

A computer program has been written to evaluate the steady-state characteristics of MISET devices using numerical methods to solve the equations of the above model. The program is listed in Appendix B.

5.2 DC Characteristics

To test the validity of the model, a set of J-V characteristic curves are generated and compared with experimental results from similar structures reported in [5]. Devices with three different base doping densities were constructed in [5], and each exhibited a different collector characteristic. The experimental J-V curves for three different values of the base implant dose are shown in Fig. 2 of [5]. For the highest base dose \((5 \times 10^{12} \text{cm}^{-2})\), the device has a moderate current gain \((\beta)\) of about 600. For a base dose of \(10^{13} \text{cm}^{-2}\), \(\beta\) increases from about 3000 at low voltages to about 10000 near the punchthrough voltage which is in excess of 25V. For a base dose of \(5 \times 10^{11} \text{cm}^{-2}\), \(\beta\) increases from above 10000 at low voltages to nearly 25000 near the punchthrough voltage of 4V. To simulate these experimental curves, the physical parameters of the device listed in Table 5.1 are used. The d.c. characteristics are especially sensitive to five parameters, namely the electron tunnel barrier height \((\chi_e\text{ is assumed equal to } \chi_m)\), the oxide thickness \((d)\), the base doping density \((P_B)\), the base width \((W_b)\)
| $T$ | temperature | $300K$ |
| $E_{ph}$ | silicon bandgap at base region | $1.1eV$ |
| $m_{el}^*$ | electron effective mass in $SiO_2$ | $0.5m_e$ |
| $m_e^*$ | electron transverse mass in $Si$ | $0.2m_e$ |
| $m_h^*$ | hole transverse mass in $Si$ | $0.66m_e$ |
| $\varepsilon_s$ | permittivity of $Si$ | $11.9\varepsilon_o$ |
| $\varepsilon_f$ | permittivity of $SiO_2$ | $3.9\varepsilon_o$ |
| $\tau_g$ | carrier generation lifetime at base-collector depletion region | $0.2\mu s$ |
| $\tau_{hc}$ | hole recombination lifetime in collector | $0.2\mu s$ |
| $D_{hc}$ | hole diffusion constant in collector | $12cm^2s^{-1}$ |
| $n_i$ | intrinsic carrier concentration | $1.45 \times 10^{10}cm^{-3}$ |
| $N_C$ | conduction band density of states | $2.8 \times 10^{19}cm^{-3}$ |
| $N_V$ | valence band density of states | $1.04 \times 10^{19}cm^{-3}$ |

Table 5.1: Model parameter values for the simulation of the MISET.

and the base intrinsic lifetime ($r_{ph}$ defined in (5.11)). With $\chi_e = 0.8eV$, $d=18\AA$, $W_b=0.2\mu m$, $r_{ph} = 4 \times 10^{-8}s$ and three different values of $P_B$, the collector characteristics shown in Figs. 5.2-5.4 are obtained. The values for the base doping density used in the simulation are deliberately set to $2.5 \times 10^{17}$, $5 \times 10^{16}$ and $2.5 \times 10^{16}cm^{-3}$, which are equivalent to the base implant doses of $5 \times 10^{12}$, $10^{12}$ and $5 \times 10^{11}cm^{-3}$ in the experimental structures with a base width equal to $0.2\mu m$. The simulated curves in Figs. 5.2-5.4 are found to agree closely with the experimental curves in Fig. 2(a)-(c) of [5]. In the case of $P_B = 2.5 \times 10^{17}cm^{-3}$ (Fig. 5.2), the current gain $\beta$ increase from 500 at low $V_{CE}$ ($\sim 1V$) to about 750 at high $V_{CE}$ ($\sim 30V$) when $J_C$ is low ($\sim 8 \times 10^{-3}Acm^{-2}$). For $P_B = 5 \times 10^{16}cm^{-3}$ (Fig. 5.3), $\beta$ increases from 4300 at $V_{CE} \sim 1V$ when $J_C \sim 1.6 \times 10^{-3}Acm^{-2}$, to
Figure 5.2: Common-emitter characteristic of the MISET with $P_B = 2.5 \times 10^{17} cm^{-3}$.
Figure 5.3: Common-emitter characteristic of the MISET with $P_B = 5 \times 10^{16} \text{cm}^{-3}$. 

$\Delta J_B = 1.6 \times 10^{-3} \text{Acm}^{-2}$
Figure 5.4: Common-emitter characteristic of the MISET with $P_B = 2.5 \times 10^{16} \text{cm}^{-3}$. The dashed lines represent experimental curves from [5].
about 7000 at $V_{CE} \sim 20V$ which is near to the base punchthrough voltage. In the case of $P_B = 2.5 \times 10^{16} cm^{-3}$ (Fig. 5.4), $\beta$ increases from 10600 at $V_{CE} \sim 1V$ to 25000 at $V_{CE} \sim 3V$ near base punchthrough, when $J_C \sim 3.2 \times 10^{-4} Acm^{-2}$.

5.3 Small-Signal Analysis

Using the same approach as described in Section 4.3 for the TETRAN, the d.c. model can be used to examine the small-signal parameters which affect directly the high frequency performance of the MISET device. The transconductance is defined as

$$g_m = \left. \frac{dJ_C}{dV_{BE}} \right|_{V_{CE}}$$

(5.18)

For high performance transistor design, the emitter series resistance $R_e$ needs to be minimized because it can degrade the transconductance. In the area of high speed polysilicon emitter transistors, a lot of effort has been expended in attempting to design an emitter contact with minimal series resistance [66]. In the MISET, the emitter series resistance is mainly due to the thin tunnel oxide, and can be expressed as

$$R_e = \left. \frac{d\psi_I}{dJ_E} \right|_{V_{CE}}$$

(5.19)

The base-emitter capacitance $C_{be}$ is also a critical parameter determining
the speed of transistor operation. It is defined as

\[ C_{be} = \frac{dQ_B}{dV_{BE}} \bigg|_{V_{BE}} \]  

(5.20)

\( Q_B \) is the total charge stored in the base region, including the stored charge \( Q_s \) in the space-charge region and the storage charge of minority carriers \( Q_{\text{min}} \) in the neutral base region, i.e.

\[ Q_B = Q_s + Q_{\text{min}} \]  

(5.21)

By differentiating (5.21) w.r.t. \( V_{BE} \), we obtain

\[ \frac{dQ_B}{dV_{BE}} = \frac{dQ_s}{dV_{BE}} + C_{\text{diff}} \cdot \frac{d\phi}{dV_{BE}} \]  

(5.22)

where \( C_{\text{diff}} \) is the base diffusion capacitance and is defined as \( dQ_{\text{min}}/d\phi \). It can be shown that although the magnitude of \( C_{\text{diff}} \) may become comparable to \( dQ_s/dV_{BE} \), the ratio \( d\phi/dV_{BE} \) is usually small, so the second term of the right side of (5.22) can be neglected (see the numerical example later). Therefore, the base-emitter capacitance can be approximated as

\[ C_{be} \approx \frac{dQ_s}{dV_{BE}} \bigg|_{V_{OE}} \]  

(5.23)

The unity current gain cutoff frequency can be expressed as

\[ f_T = \frac{g_m}{2\pi C_{be} \left[ 1 + 2C_{be}/C_{bc} \right]^{1/2}} \]  

(5.24)

In the MISET device, the base-collector capacitance \( C_{bc} \) is mainly a depletion layer capacitance, and is usually small compared to \( C_{be} \). Thus, \( f_T \) can be
approximated as

\[ f_T \approx \frac{g_m}{2\pi C_{be}} \]  (5.25)

All the above small-signal parameters can be computed numerically from the d.c. model by examining the changes in \( J_C, \psi_T \) or \( Q \), in response to small changes in \( J_B \) or \( V_{BE} \).

In order to give an estimate of the high frequency cut-off of the MISET, we consider a particular structure with \( \chi_e = 0.8eV, \ d = 18\AA, \ W_b = 0.2\mu m, \ P_B = 2.5 \times 10^{18}cm^{-3} \) and \( \tau_{eb} = 4 \times 10^{-8}s \). At \( V_{CE} = 2V \), a base current \( (J_B) \) of \( 32Acm^{-2} \) gives rise to a collector current \( (J_C) \) of \( 6.6 \times 10^3Acm^{-2} \). By changing \( J_B \) slightly, i.e. perturbing the steady-state system, we obtain \( d<j>/dV_{BE} \sim 0.08, \ dQ_s/dV_{BE} \sim 1.3\mu Fcm^{-2} \) and \( g_m \sim 2 \times 10^4Scm^{-2} \). The base storage capacity \( C_{diff} \) is estimated by the formula

\[ C_{diff} = \frac{q^2}{kT} \left( \frac{W_b'}{n_{eb}} \right) \exp\left( \frac{q\phi}{kT} \right) \]  (5.26)

where \( n_{eb} \) is the equilibrium electron concentration in the neutral base. This equation gives a value of \( 1.2\mu Fcm^{-2} \) for \( C_{diff} \). Even though the magnitude of \( C_{diff} \) is comparable to \( dQ_s/dV_{BE} \), the small ratio \( d\phi/dV_{BE} \) justifies the use of (5.23) for finding \( C_{bc} \). Also \( C_{bc} \) is found to be about \( 0.018\mu Fcm^{-2} \), which is small compared to \( C_{be} \), thus justifying the approximation of \( f_T \) in (5.25):

\[ f_T \approx \frac{2 \times 10^4Scm^{-2}}{2\pi \times 1.3 \times 10^{-6}Fcm^{-2}} = 2.4GHz \]
Note that the above $f_T$ value is obtained under conditions of high base-emitter bias, where the emitter metal Fermi-level rises well above the semiconductor conduction band and the semiconductor surface is degenerated by a large population of holes. Under these conditions, it is possible to estimate $f_T$ by analytical expressions, such as (4.11) and (4.12). Substituting $\chi_m = 0.8eV$, $d = 18\AA$, $\psi_I = 0.67V$ (found from the simulation), and $J_C = 6.6 \times 10^3Acm^{-2}$ into (4.11) gives

$$g_m = 4.8V^{-1} \times J_C \approx 3.2 \times 10^4\Omega^{-1}cm^{-2}$$

For $d = 18\AA$ and $\epsilon_I = 3.9\epsilon_o$, (4.12) gives $C_{se} = 1.9\mu Fcm^{-2}$. Using (5.25), the unity current gain cutoff frequency $f_T$ is estimated to be about $2.7$GHz, which closely agrees with the value obtained from numerical analysis.

To investigate the dependence of the small-signal parameters such as $\beta$, $g_m$, $R_s$ and $f_T$ on the collector current $J_C$, we plot all these parameters against $J_C$ in Figs. 5.5-5.7. All the parameters have been calculated numerically, since the previous analytical expressions for $g_m$ and $C_{se}$ are only valid in the regime of high $J_C$, and are not satisfactory for low $J_C$. In Fig. 5.5, the gain decreases quite rapidly as the collector current increases, from $\beta \sim 20000$ near $J_C = 1Acm^{-2}$ to $\beta \sim 120$ near $J_C = 0.7 \times 10^4Acm^{-2}$. The plot resolves the large discrepancy between two different reports on the current gain of MISETs: $\beta \sim 20000$ in [5] and $\beta \sim 120$ in [65]. In the former case the measurements of $\beta$ were made at very low $J_C$ (of the order of $1Acm^{-2}$), while in [65] $J_C$ was several orders of
Figure 5.5: Dependence of current gain on collector current of the MISET.
Figure 5.6: Dependence of transconductance and emitter resistance on collector current of the MISET.
Figure 5.7: Dependence of cut-off frequency on collector current of the MISET.
magnitude higher.

In Fig. 5.6, we observe that $g_m$ rises while $R_e$ falls as $J_C$ increases. It is interesting to note that $R_e$ can be higher than $10^4 \mu \Omega cm^2$ at low collector current, which is in general agreement with the extremely high values of emitter series resistance measured in [67]. Experimentally, Moravrej-Farshi [67] has also observed that increasing the collector current level by an order of magnitude or more results in a 2 to 3 fold drop in the measured series resistance, which is essentially the predicted tendency in Fig. 5.6. The predicted $R_e$ at $J_C = 7 \times 10^3 A/cm^2$ is about $35 \mu \Omega cm^2$, that is an order of magnitude higher than typical values for the polysilicon emitter transistor (see [66]). $R_e$ degrades $g_m$, which explains why the $f_T$ for MISETs ($\sim 2 GHz$) is relatively low compared to PETs ($\sim 15 GHz$). Of course by reducing the oxide thickness we can decrease $R_e$, and this is precisely the reason for reducing the interfacial oxide thickness as much as possible in high speed PET design [66].

Fig. 5.7 shows that $f_T$ of the device increases monotonically with $J_C$. The $f_T$ increases from about 10MHz when $\beta \sim 20000$ at low $J_C$, to about 2.4GHz when $\beta \sim 120$ at high $J_C$. The prediction is in good agreement with the experimental result in [65], where an operational device exhibited a gain of 120 and a $f_T$ of about 1GHz at high $J_C$.

In conventional high-speed BJTs, there is a trade-off between the base
conductance and $f_T$, and the base doping density must be carefully chosen to optimize the maximum oscillation frequency $f_{\text{max}}$. This is the frequency at which the forward power gain of a transistor becomes unity, and can be expressed as [68]

$$f_{\text{max}} = \frac{1}{2} \sqrt{\frac{f_T}{2\pi R_b C_{bc}}}$$

(5.27)

where $R_b$ is the base sheet resistance and $C_{bc}$ is the base-collector capacitance of the device. Increasing the base doping density will increase the transit time of minority carriers in the base and therefore decrease $f_T$. Also, $R_b$ will be reduced due to the lower base resistivity. On the other hand, decreasing the base doping density will increase both the $f_T$ and $R_b$. Thus there is an optimal base doping density at which $f_{\text{max}}$ is highest.

To determine whether a similar trade-off exists in MISETs, the effect of the base doping density $P_B$ on $\beta$, $f_T$ and $f_{\text{max}}$ for a particular device at a certain base current was considered. Neglecting the parasitic resistance and capacitance, $R_b$ and $C_{bc}$ are essentially the base-spreading resistance and the base-collector depletion capacitance respectively. For a rectangular base layer with two contacts at two opposite sides, which is the structure assumed in this work, the base-spreading resistance can be calculated as [69]

$$R_b = \frac{1}{12} \cdot \frac{\rho_b h}{W_b l}$$

(5.28)

where $W_b$ is the base width, $h$ is the distance between the two contacts and
$l$ is the length of the contact. $\rho_b$ is the resistivity of the base region, and the dependence of its value on base doping density is shown in [18, p.32].

In Figs. 5.8-5.10, three small-signal parameters $\beta$, $f_T$ and $f_{\text{max}}$ are plotted against base doping density $P_B$ for two values of intrinsic base lifetime $\tau_{ob}$ (see (5.11)). The device parameters are: $\chi_e = 0.8eV$, $d = 18\lambda$, $W_b = 0.2\mu m$, $A_E = h \times l = 5 \times 5\mu m^2$ (emitter area), and the transistor is operating with $J_B = 16Acm^{-2}$ and $V_{CE} = 2V$.

Referring to the case of the larger base lifetime ($\tau_{ob} = 5 \times 10^{-7}s$ as used in the SEDAN III program [55]), $\beta$ and $f_T$ stay more or less constant up to $P_B \sim 7.5 \times 10^{17}cm^{-3}$. $f_{\text{max}}$ rises rapidly up to this point because the base resistance decreases as $P_B$ increases. Beyond $P_B \sim 7.5 \times 10^{17}cm^{-3}$, $f_{\text{max}}$ tends to increase slowly because $f_T$ starts to fall. Even at the high base doping density of $5 \times 10^{18}cm^{-3}$, $f_{\text{max}}$ has still not reached its maximum value because $R_b$ is dropping more rapidly than $f_T$. This suggests that the base regions of MISETs, like GaAs heterojunction transistors, can be doped as heavily as wished in order to improve high frequency performance.

In the case of the short base lifetime $\tau_{ob} = 5 \times 10^{-8}s$, which might be a typical value in actual devices due to excessive base recombination, $f_T$ drops rapidly beyond the doping density of $P_B = 2 \times 10^{17}cm^{-3}$. When it reaches $P_B = 7.5 \times 10^{17}cm^{-3}$, the increase in base conductance cannot compensate for
Figure 5.8: Dependence of current gain on the base doping density of the MISET.
Figure 5.9: Dependence of cut-off frequency on the base doping density of the MISET.
Figure 5.10: Dependence of maximum oscillation frequency on the base doping density of the MISET.
the drop in $f_T$, which is more rapid than that in the case of long base lifetime. Therefore $f_{\text{max}}$ begins to decrease after this point. The result suggests that for a transistor with short base lifetime, there is only a limited advantage to be gained regarding improving the high frequency performance by increasing the base doping density. This is because base recombination generally reduces the gain, transconductance and cut-off frequency of the MISET device. With a short base lifetime, this effect becomes important. The rise of base sheet conductance cannot compensate for the drop of $f_T$ as the base doping density increases, leading to degradation of $f_{\text{max}}$.

5.4 Summary

The MISET device displays the interesting behaviour of having high current gain and low $f_T$ at small bias $V_{BE}$, but low current gain and high $f_T$ at large bias. The reason for this is that the emitter resistance $R_e$ decreases as the bias increases; this is an inherent property of MIS tunnel junction emitters. The cut-off frequency of the MISET is found to be about $2GHz$ at high collector current, close to the $f_T$ predicted for the TETRAN. This is by no means coincidental, since at high bias the relative positions of the Fermi-levels and band edges near the oxide interfaces for both the TETRAN and MISET are essentially identical.

The $f_T$ of the MISET is not likely to be improved by reducing the oxide
thickness or the barrier height, since an oxide thickness of 18 Å is about the reliability limit for a MIS junction. However, the $f_{\text{max}}$ of the MISET device can be optimized by adjusting the base doping density. The simulations indicate that, for reasonable values of base minority carrier lifetime, $f_{\text{max}}$ continues to increase as the base layer doping density increases to the practical limit.
Chapter 6

Modeling the PNP Polysilicon Emitter Transistor

High performance bipolar integrated circuits have significantly benefited from the advent of polysilicon emitter contact technology. Because the polysilicon contact reduces the back-injected emitter current, the emitter efficiency of this kind of device is improved. The resultant increase in current gain can then be traded-off with higher base doping (lower base sheet resistance). The outcome is an increase in switching speed.

Polysilicon emitter transistors (PETs) that have been studied extensively in recent years are mainly of the npn type. The best devices have exhibited a high cut-off frequency of 16GHz [58]. Recently, Maritan and Tarr [70] have fabricated pnp PETs with different surface treatments, and demonstrated that they can exhibit reasonable gain and acceptably low values of emitter series
resistance. These results are extremely important, because they imply that the cut-off frequency of pnp devices might be very high. Although pnp devices will exhibit larger base transit times than npn devices due to the lower hole mobility, they have smaller base sheet resistance compared with npn devices. This should be advantageous in terms of $f_{\text{max}}$. The resistivity of n-type (phosphorus doped) silicon is generally two to three times lower than that of p-type (boron doped) silicon, see [18, p.31]. As long as $f_T$ for pnp PETs is not 2-3 times less than that for equivalent npn devices, the $f_{\text{max}}$ values for these two transistor types will be comparable to each other.

In the area of GaAs heterojunction bipolar transistors (HBTs), theoretical analysis and computer simulations [71,72] indicate that the values of $f_{\text{max}}$ for pnp and npn structures are very close. This feature may open up the possibility of complementary npn/pnp design approaches for circuits such as amplifiers and A/D converters, and provide a solution to the long-standing problem of developing very high speed complementary circuits in III/V materials for low-power applications. If pnp PETs can be optimized to exhibit values of $f_{\text{max}}$ close enough to npn PETs, then silicon bipolar technology might enjoy the advantages of complementary circuit design which have already been foreseen in GaAs HBTs.

In this section, we are concerned mainly with pnp devices with a deliberately-
grown interfacial oxide to which the improved tunneling model developed in Chapter 3 can be applied. Two types of device, one with and the other without post polysilicon-deposition anneals, are simulated by the model and compared to the experimental results in [70]. The device structure of the pnp PET reported in [70] is illustrated in Fig. 6.1. The intrinsic base is formed by ion implantation of phosphorus, and has a depth of 0.3μm after annealing. A thin layer of oxide (10 – 20Å) is chemically grown in the emitter window, then a boron-doped amorphous Si film is deposited and later recrystallized at low temperature. An emitter polysilicon film of about 0.1μm thick is obtained. This device, without any post polysilicon-deposition annealing, gives an effective
emitter Gummel number $G_e$ of $1 - 2 \times 10^{14} \text{scm}^{-4}$, combined with an emitter resistance $R_e$ of about $26 \mu \Omega \text{cm}^2$. If the device is annealed at $900^\circ C$ for 30 minutes after the polysilicon film deposition, a monocrystalline emitter approximately $0.15 \mu m$ deep forms underneath the oxide due to dopant diffusion from the polysilicon into the base region. The annealed device gives the same $G_e$ as the unannealed device, but much lower $R_e$ ($\sim 1 - 2 \mu \Omega \text{cm}^2$). Both types of device can exhibit current gain of up to about 300.

The reasons why pnp PETs can exhibit good gain and low emitter resistance are not obvious. The one-band model [30], which implies the tunneling probability of holes in silicon MIS structures is inevitably smaller than that for electrons, suggests that the ratio of hole emitter current to electron back-injected base current should be small, leading to inferior current gain. Also because of the larger hole tunnel barrier height $\chi_h$ in pnp devices compared to electron barrier height $\chi_e$ in npn devices, the emitter resistance of pnp devices is expected to be large. All these reasons suggest that pnp device performance should be very poor.

In this chapter it is shown that the one-band model, which has successfully predicted the characteristics of TETRANs and MISETs, is capable also of predicting the characteristics of pnp oxidized PETs (PETs with chemically grown emitter oxides). The simulated d.c characteristics of both unannealed
and annealed devices agree well with the experimental d.c. data in [70]. The high frequency performance of these devices is assessed by computing some important a.c. parameters.

6.1 Model Formulation

In the unannealed device, there is no mono-emitter region underneath the tunnel oxide. The energy band diagram is simply as shown in Fig. 6.2. It can be seen that the charge flows and potential drops across the unannealed PET are similar to those existing in the MISET, except that here we have a pnp rather than a npn structure. The similarity is obvious if the band diagram of the MISET (Fig. 5.1) is viewed upside down. Therefore, the method of solution used in the MISET is directly applicable to solving the the case of the unannealed PET. However, two modifications need to be made in order to simulate the device correctly, as the emitter material is polysilicon rather than metal.

Firstly, the effect of minority carrier conduction in poly-Si must be considered. The minority carrier diffusion in polysilicon film can be described by exactly the same equation used for diffusion in mono-silicon. Taking into consideration the boundary conditions for minority carrier concentrations at the metal/poly-Si and oxide/poly-Si interfaces, the electron diffusion current at the
poly-Si/oxide junction can be written as

\[ J_{n1} = \frac{q D_{pol}}{L_{pol}} \coth\left(\frac{W_{pol}}{L_{pol}}\right) |N_C \exp\left(-\frac{\zeta_1 + q \phi_1 - E_F}{kT}\right) - n_{pol}| \] (6.1)

where \( n_{pol} \) is the equilibrium concentration of minority carriers, \( W_{pol} \) is the polysilicon layer thickness, \( D_{pol} \) and \( L_{pol} \) are the effective diffusion constant and diffusion length in the p-doped polysilicon, respectively. In addition to the two non-linear equations (see (5.14) and (5.15)) describing the continuity of electron and hole flows at the mono-Si/oxide junction, another equation \( J_{n1} = J_{tn} \) which represents the electron (minority carrier) continuity at the poly-Si/oxide junction (see Fig. 6.2), must be solved simultaneously.

Secondly, the forbidden energy range for tunneling is no longer coincident with the mono-Si band gap as in the case of the MISET, but is determined by the range of overlap of the poly-Si and mono-Si band gaps. In Fig. 6.2, \( E_{co} \), the energy above which electron tunneling occurs, is in line with the poly-Si conduction band edge, while \( E_{vo} \), below which hole tunneling occurs, is coincident with the mono-Si valence band edge. If the potential drop across the oxide \( \psi_I \) is reversed, then the mono-Si conduction band edge and the poly-Si valence band edge will become \( E_{co} \) and \( E_{vo} \) respectively. The parameters \( \alpha_1 \), \( \alpha_2 \), \( \beta_1 \) and \( \beta_2 \), which determine the tunneling currents (see (5.3) and (5.4)), are defined as the differences between the quasi-Fermi levels with respect to \( E_{co} \) or \( E_{vo} \). The effective electron (or hole) tunneling barrier height is the energy
Figure 6.2: Energy band diagram of the pnp oxidized PET without post-deposition annealing.
difference between $E_{co}$ (or $E_{vo}$) and the middle point of the oxide conduction (or valence) band edge.

In the annealed device a layer of p-doped monocrystalline silicon about 0.15$\mu$m thick exists underneath the thin oxide. The energy band diagram shown in Fig. 6.3 is seen to be quite different from that of the unannealed device. As a result, a different solution procedure is required.

The independent variables chosen to solve the equations describing the system are $\psi_I$, $\phi_1$, $\phi_2$ and $\phi$. These are defined in Fig. 6.3. All the current components and potential drops indicated in the band diagram can be expressed in terms of these four variables. The minority carrier diffusion current in the polysilicon $J_{n1}$ and the tunneling currents $J_{in}$ and $J_{tp}$ are treated in exactly the same way as for the unannealed devices. $J_{rm}$ is the recombination current in the mono-emitter region, and $J_{n2}$ and $J_{nm}$ are the electron diffusion currents across the oxide interface and depletion edge of the mono-emitter respectively. Their magnitudes are given by (5.5)-(5.7), after replacing $-V_{CB}$ and $W'_b$ by $\phi_2$ and the mono-emitter depth, respectively. $J_{rb}$ is the recombination current, while $J_{pc}$ and $J_{pc}$ are the recombination current and hole diffusion currents at the depletion edges of the neutral base. These currents can be related to $\phi$ and $V_{CB}$ by (5.5)-(5.7), except all the minority carrier parameters are required to change to those of holes. The generation current $J_g$ at the base/collector junction is
Figure 6.3: Energy band diagram of the pnp oxidized PET with post-deposition annealing.
given by (5.12), and the recombination current $J_r$ at the base/mono-emitter junction is also given by (5.12), but with $-V_{CB}$ and $l_b$ replaced by $\phi$ and the base/emitter depletion width, respectively. $J_d$ is the electron diffusion current from the collector to the base, and is given by (5.13), after changing the minority carrier parameters to those of electrons.

Finally, there are four non-linear equations to be solved. They are:

$$J_{pe} = J_{tp} - J_{rm} - J_r$$  \hspace{1cm} (6.2)

$$J_{nm} = J_B + J_g + J_d - J_r - J_{r_b}$$  \hspace{1cm} (6.3)

$$J_{n2} = J_{ln}$$  \hspace{1cm} (6.4)

$$J_{n1} = J_{ln}$$  \hspace{1cm} (6.5)

The first equation is concerned with the continuity of electron current at the base depletion edge near the emitter. The other three take care of the continuity of hole flows at the mono-emitter depletion edge and the two silicon/oxide interfaces, respectively.

After $\psi_I$, $\phi_1$, $\phi_2$ and $\phi$ are determined, the terminal currents can be computed by

$$J_E = J_{ln} + J_{tp}$$  \hspace{1cm} (6.6)

$$J_C = J_{pc} + J_g + J_d$$  \hspace{1cm} (6.7)
Computer programs that are used to obtain the steady-state characteristics of unannealed and annealed pnp PETs are listed in Appendix C and D respectively.

6.2 DC Characteristics

The first simulation carried out was for the case of an unannealed device. In Fig. 6.4 the collector and base currents ($J_C$ & $J_B$) are plotted against the base-emitter bias ($V_{BE}$). Some of the physical parameters for this simulation are: electron barrier height ($\chi_e$) = 0.5eV, oxide thickness ($d$) = 10Å, $\xi_1$ = 0.1eV, base width ($W_b$) = 0.3μm, base doping density ($P_B$) = $10^{17}cm^{-3}$ and intrinsic base lifetime ($\tau_{ob}$) = $5 \times 10^{-8}s$. The value of $\xi_1$ used here corresponds to a polysilicon doping density of $10^{20}cm^{-3}$, i.e. about the solid solubility limit for boron in silicon. The value of $\chi_e$ used gave the best fit of predicted and experimental data. Similar values have been used by others in modeling SIS tunnel structures [59]. Since no experimental data for the minority carrier mobility and lifetime in p-doped polysilicon is available, we use here the data for minority carrier transport in n-doped polysilicon. In heavily n-doped polysilicon ($N_{pol} \sim 2 \times 10^{19}cm^{-3}$), the minority carrier mobility and lifetime are estimated to be about $8cm^2V^{-1}s^{-1}$ and $2 \times 10^{-10}s$ respectively [12]. Therefore, in the simulation performed here $D_{pol}$ and $L_{pol}$ are taken to be $0.2cm^2s^{-1}$ and 0.06μm,
Figure 6.4: Gummel plots of the unannealed pnp PET from computer simulation (dashed line) and published experimental data (solid line) [70].
respectively.

The d.c. characteristics are not affected significantly by changing the values of $D_{pol}$ or $L_{pol}$ by one or two orders of magnitude. This implies that the quantum mechanical reflection of electrons at the oxide barrier is the dominant mechanism for preventing back-injection, consequently minority carrier diffusion in the polysilicon is relatively unimportant. It is somewhat similar to the case of the npn polysilicon transistor, where the d.c. characteristics are greatly changed by altering the interfacial oxide thickness, but not nearly so much by changing the grain size of the polysilicon or by using hydrogen treatments for the passivation to the poly-emitter [73].

Fig. 6.4 shows that the simulated curves closely agree with the experimental curves reported in [70], except for the feature that $J_B$ is underestimated in the high forward bias ($V_{BE} > 0.8V$) regime. This discrepancy could probably be removed by including a surface-state tunneling current component in the model. "Kinks" in $J_B$-$V_{BE}$ curves are often observed in practical npn polysilicon emitter transistors, for example in [73,74]. Simulations show that these "kinks", or departures from linearity, can be smoothed-out by increasing the surface-state density [74].

The emitter Gummel number $G_e$ and the emitter resistance $R_e$ can be easily obtained from the simulation. For a conventional pnp transistor in which
the base current is dominated by the back injection of electrons into the emitter, $J_B$ can be related to $G_e$ via

$$J_B = \frac{q n_i^2}{G_e} \exp\left(\frac{q V_{EB}}{kT}\right)$$

(6.8)

This means $G_e$ can be easily computed by determining the base saturation current from the y-intercept of the Gummel plot. Although in polysilicon emitter transistors $J_B$ is given by the tunneling current expression (3.8), rather than by such a simple expression as (6.8), the Gummel number concept is still very useful for evaluating the transistor's d.c. performance. The $J_B$ curve in the Gummel plot deviates from linearity at high emitter-base bias, therefore $G_e$ can only be computed by extrapolating the linear portion of the curve in the low bias region. $G_e$ is found to be about $6.1 \times 10^{14} \text{Scm}^{-4}$, which is in the same order as the experimental result of $1 - 2 \times 10^{14} \text{Scm}^{-4}$. An expression for the emitter resistance $R_e$ has already been given in (5.19). For a high collector current level ($\sim 10^5 \text{Acm}^{-2}$), $R_e$ is computed to be about $15 \mu\Omega \text{cm}^2$, which is close to the experimental value of $26 \mu\Omega \text{cm}^2$.

The annealing step can affect the integrity of the oxide and reduce the tunnel oxide thickness. Also during annealing the dopants from the heavily-doped polysilicon diffuse across the tunnel oxide to the mono-silicon region, pushing the emitter-base junction away from the oxide interface. Therefore, a mono-emitter is formed underneath the tunnel oxide. In the annealed device
the mono-emitter region is 0.15\(\mu m\) deep, while the base width is reduced from the original 0.3\(\mu m\) to 0.15\(\mu m\). The mono-emitter and base doping densities are taken to be \(10^{19} \text{cm}^{-3}\) and \(2.3 \times 10^{17} \text{cm}^{-3}\) respectively. These values are based on SUPREM simulation results of the processing sequence used in [70]. With an electron barrier height (\(\chi_e\)) of 0.5eV, an interfacial oxide thickness \(d\) of 7\(\AA\) is required to give a low \(R_e\) of 2.7\(\mu\Omega cm^2\). The simulations indicate that this reduction of oxide thickness (from the value of 10\(\AA\) used for the unannealed device) is essential to bring \(R_e\) down to the value of 2\(\mu\Omega cm^2\) found in experimental devices.

The result suggests that the interfacial oxide becomes thinner after the high temperature annealing, thus reducing the emitter resistance by almost an order of magnitude. From the simulated Gummel plot, \(G_e\) is computed to be \(1.2 \times 10^{14} \text{scm}^{-4}\), which is in the experimentally observed range of \(1 - 2 \times 10^{14} \text{scm}^{-4}\). The simulated common emitter characteristics for the annealed device are shown in Fig. 6.5, plotted along with the experimental data from [70]. The current gain is 300 at 0.5 \(\times 10^{-3}\) Acm\(^{-2}\), which is in good agreement with the experimental value of 250 at the same current level.
Figure 6.5: Common emitter characteristics of the annealed pnp PET from computer simulation (dashed line) and published experimental data (solid line) [70].
6.3 Small-Signal Analysis

In this section the unity current gain cut-off frequency $f_T$ is used as a figure of merit to compare the a.c. performance of annealed and unannealed devices. By perturbing the steady-state parameters by 1% in the d.c. model, small signal parameters such as $g_m$ and $C_{be}$ can be determined and $f_T$ can be computed by (5.25). In both types of device (annealed and unannealed), the same formula (5.18) is used to obtain $g_m$, but different expressions for $C_{be}$ are needed. For the unannealed device, $C_{be}$ is computed in exactly the same way as for the MISET. The contribution of minority carrier storage capacitance (second term on the right side of (5.22)) is always small compared to charge storage in the depletion region underneath the oxide, therefore $C_{be}$ can be approximated by (5.23).

In the annealed device, $C_{be}$ is the equivalent of two series capacitances $C_{ox}$ and $C_{diff}$, i.e.

$$\frac{1}{C_{be}} = \frac{1}{C_{ox}} + \frac{1}{C_{diff}}$$

(6.9)

where $C_{ox}(= \varepsilon l/d)$ is the oxide capacitance and $C_{diff}$ is the base minority carrier storage capacitance. An expression for $C_{diff}$ is given in (5.26). This can be used in this case with $n_{oe}$ replaced by the base equilibrium hole concentration $p_{oe}$.

The unity current gain cut-off frequencies of both types of device are plotted in Fig. 6.6. For the unannealed device, $f_T$ increases quite rapidly from
Figure 6.6: Dependence of the unity current gain cut-off frequency on the base current for both the annealed and unannealed pnp PETs.
0.12GHz at $J_B = 10^{-3} Acm^{-2}$ to 2.7GHz at $J_B = 35 Acm^{-2}$, while for the annealed device $f_T$ remains constant at about 16GHz regardless of the current level. The dependence of $f_T$ on $J_B$ for the unannealed device is similar to that of the MISET, where $f_T$ can be significantly increased by an increase in $J_B$. This is because in unannealed devices, where there is no mono-emitter region, the device performance is mainly determined by the tunneling mechanism in the emitter tunnel oxide. Since the resistance of the tunnel oxide decreases rapidly as the device becomes more forward-biased, the transconductance $g_m$ also rises rapidly. On the other hand the value of $C_{be}$, which approaches the oxide capacitance $C_{ox}$, only rises slowly. This explains the increase of $f_T$ with $J_B$ shown in Fig. 6.6 for the unannealed device.

In the annealed device, $C_{be}$ mainly depends on $C_{diff}$ due to the large $C_{ox}$. Since both $g_m$ and $C_{diff}$ increase as the bias current increases, a more or less constant $f_T$ results. In the case of conventional BJTs, neglecting the emitter and collector transit times, $f_T$ can be written as

$$f_T = \frac{1}{2\pi} \cdot \frac{g_m}{C_{diff}} = \frac{1}{2\pi} \cdot \frac{1}{r_B} = \frac{1}{2\pi} \cdot \frac{W_b^2}{2D_b}$$

where $r_B$ is the base transit time, $W_b$ is the base width and $D_b$ is the diffusion constant of minority carrier in base. Note that $f_T$ is constant and only depends upon the base width and base doping density. Considering the annealed PETs, $C_{be}$ is essentially dominated by $C_{diff}$ rather than $C_{ox}$, and $g_m$ is controlled by
the bias at the mono-emitter/base junction, therefore a constant $f_T$ is to be expected, as in the case of conventional BJTs.

Generally speaking, the high frequency performance of the unannealed device is much worse than that of the annealed device. At high forward bias, $f_T$ of the annealed device is still several times higher than that of the unannealed device, due to the lower emitter resistance (thinner tunnel oxide) and smaller base/emitter capacitance ($C_{diff} < C_{ox}$).

6.4 Summary

The d.c. performance of both the annealed and unannealed pnp devices is good, in as much as reasonable current gains are possible. This is an unexpected result in view of the low hole tunnel probability predicted by the one-band model. For a tunnel oxide thickness of around 10Å, pnp devices exhibit current gains close to 300, which are much less than that of npn devices (~ 2000) with the same thickness of oxide [7]. Simulations show that as the oxide thickness increases the current gain of the pnp PETs decreases. The current gain can even drop below unity if the oxide thickness is larger than 20Å. This trend is contrary to what is observed in npn devices. It is further evidence of the fact that the low tunnel probability of holes makes tunneling the current limiting mechanism in pnp PETs.
From the high frequency performance study of the annealed and unannealed devices, it is predicted that $f_T$ for the annealed devices will be higher than that for unannealed devices. This result is mainly due to the lower emitter resistance and emitter-base capacitance of the annealed devices. The computed $f_T$ (based on a calculation which includes only the base transit time) for pnp annealed devices, is about 16GHz, which is extremely good. Even if the base transit time only comprises half of the emitter-to-collector delay time in a practical device, then a $f_T$ value of 8GHz should be considered feasible. The highest value reported so far for an experimental pnp PET device is 1.6GHz [75]. Currently, a state of the art npn polysilicon emitter transistor can have a $f_T$ of 16GHz. If a pnp device with $f_T$ approaching 8GHz can be made, than the complementary circuit design approach, which implies low power consumption and high speed, will be feasible in Si technology.
Chapter 7

Conclusion

The new formulation of the MIS tunnel junction model proposed in this thesis is used to simulate several silicon bipolar structures with tunnel oxide emitters. The new formulation represents the tunnel oxide band structure via a one-band model, and also allows for the energy dependence of the electron and hole tunneling probabilities. For the TETRAN device, which has d.c. characteristics that cannot be adequately predicted by an earlier analytic model, the new model generates d.c. curves which agree closely with the experimental data. The unity current gain cut-off frequency is estimated to be about 2GHz, which is far more reasonable than the estimate of 600GHz recently suggested by others. In the modeling of MISETs, the new formulation also correctly predicts the experimental d.c. characteristics of devices with different base doping densities. The simulated current gains, emitter series resistances and unity current gain cut-off frequencies are in good agreement with the experimental data at
selected collector current levels. For pnp polysilicon emitter transistors, simulated results based on the new model fit the experimental d.c. curves very well. Contrary to intuitive expectations, the simulations suggest that these devices can exhibit moderate current gains and cut-off frequency values. The predicted results are in good agreement with experimental data reported in the literature.

The major contribution of the work is a demonstration that the one-band model, which has never been employed previously to characterize transistors with tunneling emitters, leads to an accurate description of carrier transport through thin silicon oxides. The energy dependence of the tunneling probabilities of electrons and holes, which were not considered in previous analytic models, is shown to be crucial to accurate device modeling, especially at high current levels. New power series expressions for the tunneling currents have been formulated in order to accommodate the energy dependent tunneling probabilities in a computationally economic manner. The resulting new model should be easy to incorporate into a more general device simulation program, such as the SEDAN program.

There is now much interest in GaAs devices and it appears that the model developed here may be of use in analyzing some advanced structures utilizing this semiconductor material. The enhancement of gate barrier height in MESFETs [76,77] may be due to the presence of a thin insulating layer. If so, the
gate MIS structure could be investigated by modifying the program to use parameters relevant to GaAs. Another interesting device is the GaAs BICFET [78]. This device is similar to the TETRAN, but has a wide bandgap semiconductor in place of a thin insulator. By considering diffusion or thermionic emission currents, rather than tunnel currents, it may be possible to adapt the present model to analyse BICFETs.
Bibliography


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**Listing of MIS9.S at 22:12:01 on OCT 10, 1988 for CCid=KCHU on 0**

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This program is written to generate the steady-state characteristics of the Tetran.

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**MODIFIED IN LINES 199, 206-7, 43, 62-4, 201, 204, 271**

**BANDGAP OF SiO2 UNSPECIFIED! (LINE 225)**

**ONE-BAND MODEL (LINES 88, 91-7, 175, 178-81, 225, 330-9, AREA1)**

**USE FIRST THREE TERMS OF POWER SERIES FOR HOLE CURRENT ONLY**

**USE EXACT INTEGRAL (IMSL PROQM) FOR ELECTRON CURRENT**

**MODIFY LINES 188, 195-6, 200-7**

**MODIFY LINES 501, 506**

**IMPLICIT REAL*8 (A-Z)**

**INTEGER CNTRL1.DATFL,FDFL.FREE,I,IADNR,IFAIL,ITMAX1,ITMAX2,KR.**

**LABEL,L.R,MINO,NTIMES,NTIME,NT,IER**

**EXTERNAL COMPF, FUNCM**

**DIMENSION NT(100),RHO(100),N1(100),P1(100),SIGMAN(100),**

**SIGMAP(100),CN(100),CP(100),CN(100),IADNR(100),TAUM(100)**

**DIMENSION ACCEST(2),X(2),UPPER(2),LOWER(2)**

**DIMENSION FREE(1),LABEL(15)**

**DIMENSION FDFLA(81),FDFLC(81),FDFLE(81),FDFLF(81),FDFL(81)**

**COMMON /AREA1/ CBAR,CHISE,CHISV,CJCM,**

**COMMON /AREA2/ NT.RHO.N1,P1.CN.CP.TAUM.IADNR,NTDM**

**COMMON /AREA3/ ETAC,ETAMC,ETAMV,ETAV.JVM.JPN.NSURF.PHI.PSII.PSIS,**

**COMMON /AREA4/ CNTRL1.ITMAX1,L.P,NTIME**

**COMMON /AREA5/ CFO1,FDFLA,FDFLC,FDFLE,FDFLF,FDFL(81)**

**COMMON /AREA6/ BRC**

**DATA EPSIO/8.850-12/,HBAR/1.0540-34/,KBOLTZ/1.380540-23/,**

**DATA PSFILE/8.850-12/,HBAR/1.0540-34/,KBOLTZ/1.380540-23/,**

**READ IN DATA DESCRIBING DEVICE**

**READ(KR,5) (LABEL(I),I=1,15)**

**5 FORMAT(15A4)**
READ(KR,FREE) ITMAX1,ITMAX2,ERR1,ERR2,DELY
READ(KR,FREE) CNTRL1
READ(DATFL,FREE) ND,CHISC,PHIM,MISTAR,D,KI,NCONV,AE,AH,OFIX
READ(DATFL,FREE) JOD,JORG,JUPC
DO 100 I=1,100
READ(DATFL,FREE) RHO(I),NT(I),SIGMAN(I),SIGMAP(I),TAUM(I),IADNR(I)
IF(NTRAP .LE. 0.00) GOTO 200
CONTINUE
NTRAP=I-1
READ(FOFL,FREE) (FDFLF(I),FOFLA(I),FDFLC(I),FDFLE(I),FDFLG(I),I=1,81)
ECHO  PRINT
WRITE(10,10) (LABEL(I),I=1,15)
10 FORMAT(15A4)///)
WRITE(15,15) ND,CHISC,PHIM,MISTAR,D,KI,NCONV
15 FORMAT(1X, 'ND=', D10.3,2X, 'CHIS=', F5.3,2X, 'PHIM=', F5.3,2X,
# 'MISTAR=', F5.3,2X, 'D=', D10.3,2X, 'KI=', F8.2,2X,
# 'NCONV=', F7.5)///)
WRITE(20,20) JOD,JORG,JUPC,AE,AH,OFIX
IF(NTRAP .EQ. 0) GOTO 400
WRITE(25,25)
25 FORMAT(2X, 'I=', 3X, 'ETA:', 5X, 'NT:', 7X, 'SIGMAN:', 5X, 'SIGMAP:',
# 5X, 'TAUM:', 5X, 'IADNR:'///)
DO 300 I=1,NTRAP
WRITE(30,30) I,RHO(I),NT(I),SIGMAN(I),SIGMAP(I),TAUM(I),IADNR(I)
30 FORMAT(1X,13.2X,F5.3.D10.3,3(10.3,2X).12)
WRITE(35,35)///)
NORMALIZE POTENTIALS TO KBOLTZ'T/Q AND COMPUTE CONSTANTS
VTHERM=KBOLTZ'T/Q
CTUNL=2.00D0*DSQRT(2.00D0*ME)/HBAR*DSQRT(Q)
CF0=PI*PI/6.00D0
EPSIS=KS*EPSIO
EPSII=KI*EPSIO
NC=NI*DEXPL(EGAP/(2.00D0*VTHERM))*DSQRT(NCONV)
NV=NI*DEXPL(EGAP/(2.00D0*VTHERM))/DSQRT(NCONV)
EGAPI=EGAP/VTHERM
EGAP=EGAP/VTHERM
CHISC=CHISC/VTHERM
CHISV=EGAP+CHISC
PHIM=PHIM/VTHERM
CTNLCM=-CTUNL*D*DSQRT(MISTAR*VTHERM)
CTNLVM=-CTUNL*D*DSQRT(MISTAR+VTHERM)
Listing of MIS9.S at 22:12:01 on OCT 10, 1988 for CC1d=KCHU on G

98 C
99 NNO=N0
100 NDO=N1
101 PHI0=LOG(NC/N0)
102 PSIS0=LOG(ND/N0)
103 CJVM=AM+T
104 CJCM=AE+T
105 CBAR=PHI0+CMISC-PHIM
106 CPSIS=0/EPSII/VTHERM
107 CPSIS=DSORT(2.0D0*KBOLTZ*T*EPSIS)
108 IF(NTRAP .EQ. 0) GOTO 800
109 DO 500 I=1,NTRAP
110 RHO(I)=RHO(I)/VTHERM
111 CN(I)=SIGMAN(I)*VELTH
112 CP(I)=SIGMAP(I)*VELTH
113 NI(I)=NC*DEXPL(RHO(I)-EGAP)
114 500 PI(I)=NV*DEXPL(-RHO(I))
115 C
116 C READ IN VOLTAGE V AND A STARTING ESTIMATE FOR PHI
117 C
118 C 800 READ(KR,FREE,END=1400) V,PHI
119 V=V/VTHERM
120 PHI=PHI/VTHERM
121 C
122 C GIVEN PHI, COMPUTE A STARTING ESTIMATE FOR PSIS
123 CALL FPSIS
124 C CALL SSM TO FIND A SOLUTION FOR THE COUPLED POTENTIAL
125 CALL SSM
126 C AND HOLE CURRENT CONTINUITY EQUATIONS
127 C
128 NTIMES=0
129 X(1)=U
130 X(2)=PSIS
131 DO 700 I=1,3
132 Y(I,1)=U
133 Y(I,2)=PSIS
134 700 Y(I,1)=Y(I,1)+DELY
135 NEWY=.TRUE.
136 CALL SSM(X,F,2.0,ERR1,ITMAX2,COMPF,NEWY,NEWA,NEWB,IFAIL,&800)
137 C CALL SSM AGAIN TO OBTAIN AN ESTIMATE OF THE ERROR IN THE SOLUTION
138 CALL SSM(X,F,2.0,ERR2,ITMAX2,COMPF,NEWY,NEWA,NEWB,IFAIL,&800)
139 C PREPARE FOR OUTPUT OF RESULTS
140 C
141 U=X(1)
142 PSIS=X(2)
143 PHI=DABS(U-U1)/VTERM
144 PSISER=DABS(PSIS-PSIS1)/VTERM
145 PHI=V-U
146 U0=U*VTERM
147 C CALL FOS
Listing of MIS9.S at 22:12:01 on OCT 10, 1988 for CC1d=KCHU on G

158 JVT=0.000
157 JCT=0.000
156 JMT=0.000
159 QSS=0.000
160 IF(NTRAP .EQ. 0) GOTO 1300
161 CVALEN=PHIO-EQAP*PSIS
162 DO 1200 I=1,NTRAP
163 ZETA=CVALEN*RHO(I)*V
164 FM=1.0DO/<1.0D0*DEXPL(ZETA))
165 FTNUM=(NSURF*N(I))*CN(I)*PSURF*P(I)*CP(I)*FM/TAUM(I)
166 FTONM=(NSURF*N(I))*CN(I)*PSURF*P(I)*CP(I)*1.000/TAUM(I)
167 FT=FTNUM/FTONM
168 IF(IAODR(I) .EQ. 1) GOTO 1000
169 QSS=QSS-FT*Q*NT(I)
170 GOTO 1100
171 1000 QSS=QSS+(1.0D0-FT)*Q*NT(I)
172 1100 JVT=JVT+Q*NT(I)*CP(I)*(PSURF*FT-P(I))*(1.000-FT))
173 JCT=JCT+Q*NT(I)*CN(I)*(NSURF*1.000-FT)-N1(I)*FT)
174 1200 JMT=JMT*Q*NT(I)*FT-FM)/TAUM(I)
175 1300 PSIS=CPSIS*(QSS+QSS+QSS+QSS)
176 C ETAMC=-(PHIO*PSIS*V)
177 C
178 C
179 C
180.3 BRC = CHSC + PSII/2.000
180.4 THCM = DEXPL(CNLCM*DSQRT(BRC))
180.5 LOWER(1) = 0.00
181 UPPER(1) = 1.202
181.1 LOWER(2) = 0.00
181.2 UPPER(2) = BRC
181.3 DOUI = OMIFUNCM,LOWER.UPPER.2,20000.0.DO.1.0-3.IER)
181.4 JC=M=JCJM * DOUI
181.5 C
181.6 C
181.7 C
182 CALL FJVM
183 C
184 CALL FJPN
185 C
186 JTOT=JCJM*JVM*JMT
187 JCOLL= JCJM + JPN + JUPC
188 VSE = (ETAV-ETAV) * VTHERM
189 C
190 UO=U*VTHERM
191 VO=V*VTHERM
192 PHIO=PHI*VTHERM
193 PSISO=PSIS*VTHERM
194 PSII=PSIS*VTHERM
195 WRITE(LP,40) VO, JTOT, JCOLL, JUPC, IFAIL, IER
196 40 FORMAT('U=',F8.4,'X', 'PHI=',F8.4,'X', 'PSISO=',F8.4,'X', 'ETAC=',F8.4,'X', 'SOURCES=',F8.4,'X', 'TARGET=',F8.4,'X', 'IFAIL=',I4,'X', 'IER=',I4)
197 WRITE(LP,45) PHIO, PHIER, PSISO, PSISER, PSISE
198 45 FORMAT('PHIO=',F8.4,'X','PHIE=',F8.4,'X','PSISO=',F8.4,'X','PSISER=',F8.4,'X','PSISE=',F8.4,'X')
199 46 FORMAT('ETAC=',F8.4,'X','ETAV=',F8.4,'X','ETAMC=',F8.4,'X','ETAMV=',F8.4,'X')
200 50 FORMAT('PSURF=',F8.4,'X','NSURF=',F8.4,'X','ETAC=',F8.4,'X','ETAV=',F8.4,'X')
201 55 FORMAT('ETAMC=',F8.4,'X','ETAMV=',F8.4,'X')
Listing of MIS9.S at 22:12:01 on OCT 10, 1988 for CCId=KCHU on G

203 WRITE(LP,55) JCM, JVM, JCT, JVT, JMT, JPN
204 55 FORMAT(IX,'JCM=',D11.4,'JVM=',D11.4,'JCT=',D11.4,'JVT=',D11.4,
205 "JMT=',D11.4,'JPN=',D11.4)
206 WRITE(LP,60) THCM, THVM, QS, VSE
207 60 FORMAT(IX,'THCM=',D11.4,'THVM=',D11.4,'QS=',D11.4)
208 C GOTO 800
209 C 1400 STOP
210 C END
211 C BLOCK DATA
212 C IMPLICIT REAL*8 (A-Z)
213 INTEGER CNTRL1, ITMAX1, LP, NTIMES
214 C COMMON /AREA1/ CBAR, CHISC, CHISV, CJCM,
215 # CJVM, CPSII, COS, CTNLCM, CTNLVM, EGAP,
216 # JOD, JORG, JUPC, NC, ND, NNO, NV, Q, PHI0, PNO,
217 # PSISIO, VTHERM, OFIX
218 COMMON /AREA4/ CNTRL1, ITMAX1, LP, NTIMES
219 C DATA EGAP/1.07862456D0/, 0/1.80210-19/, LP/8/
220 C END
221 C SUBROUTINE COMPF(X, F)
222 C THIS ROUTINE COMPUTES THE RESIDUE OF THE POTENTIAL AND
223 C HOLE CURRENT CONTINUITY EQUATIONS
224 C IMPLICIT REAL*8 (A-Z)
225 INTEGER CNTRL1, I, IADNR, ITMAX1, LP, NTIMES, NTRAP
226 C DIMENSION X(2), F(2), TAUM(100), RHO(100),
227 # NT(100), CN(100), CP(100), NI(100), PI(100), IADNR(100)
228 C COMMON /AREA1/ CBAR, CHISC, CHISV, CJCM,
229 # CJVM, CPSII, COS, CTNLCM, CTNLVM, EGAP,
230 # JOD, JORG, JUPC, NC, ND, NNO, NV, Q, PHI0, PNO,
231 # PSISIO, VTHERM, OFIX
232 COMMON /AREA2/ NT, RHO, NT, NI, CH, CP, TAUM, IADNR, NTRAP
233 COMMON /AREA3/ ETAC, ETAMC, ETAV, JVM, JPN, NSURF, PHI, PSII, PSIS,
234 # PSURF, OS, THVM, THCM, U, V
235 COMMON /AREA4/ CNTRL1, ITMAX1, LP, NTIMES
236 C NTIMES=NTIMES+1
237 U=X(1)
238 PHI=V-U
239 PSIS=X(2)
240 C CALL FOS
241 QSS=0.0D0
242 JVT=0.0D0
243 IF(NTRAP.EQ.0) GOTO 300
244 CVEALEM=PHI0-EGAP-PSIS
245 DO 200 I=1,NTRAP
Listing of MISO.S at 22:12:01 on OCT 10, 1988 for CCId-KCHU on G

260 ZETA=CVALEN*RHO(I)*V
261 FM=1.000/(1.000*DEXPL(ZETA))
262 FTNUM=NSURF*CN(I)*P1(I)*CP(I)*FM/TAUM(I)
263 FTDNM=(NSURF*NT(I))*CN(I)*(PSURF+P1(I))*CP(I)+1.000/TAUM(I)
264 FT=FTNUM/FTDNM
265 IF(IADNRU).EQ.1) GOTO 100
266 QSS=QSS+FT*QNT(I)
267 GOTO 200
268 QSS=QSS+(1.000-FT)*QNT(I)
269 JVT=JVT+QNT(I)*CP(I)*(PSURF+P1(I))*(1.000-FT))
270 C
271 300 PSII=CPSII*(Q+QSS+QFIX)
272 F(I)=CBAR*V+PSIS*PSII
273 C
274 CALL FJVM
275 CALL FJPN
276 F(2)=JVM-JPN-JVT
277 C
278 IF(CNTRL1.EQ.0) GOTO 400
279 UO=U'VTHERM
280 PSISO=PSIS*VTHERM
281 WRITE(LP.5) NTIMES,UO,PSISO,F(1),F(2)
282 5 FORMAT(1X,I3.2X,D23.18.2X,D14.7.2X,2(D14.7.2X>)
283 C
284 400 RETURN
285 END
285.05 C
285.1 C DOUBLE PRECISION FUNCTION FUNCM(N,X)
285.15 C IMPLICIT REAL*8 (A-Z)
285.2 INTEGER N
285.3 DIMENSION X(N)
285.35 COMMON /AREA1/ CBAR, CHISC, CHISV, CJCM,
285.4 # CJVM, CPSII, CS, CTNLCM, CTNLVM, EGAP,
285.45 # JOD, JORG, JUPC, NC, ND, NNO, NV, O, PHIO, PNO,
285.5 # PSISO, YTHEM, QFIX
285.55 COMMON /AREA3/ ETAC, ETAMC, ETANY, ETAV, JVM, JPN, NSURF, PHI,
285.6 # PSII, PSIS, PSURF, QS, THVM, THCM, U, V
285.65 COMMON /AREA6/ BRC
285.7 ETTO = X(1) * X(2) - ETAC
285.75 FSO = 0.00
285.8 IF (ETTO .GE. 150.00) GO TO 100
285.81 FSO = 1.00/(1.00*DEXPL(ETTO))
285.82 100 CONTINUE
285.83 ETIM = X(1) * X(2) - ETAMC
285.84 FM = 0.00
285.85 IF (ETIM .GE. 150.00) GO TO 200
285.86 FM = 1.00/(1.00*DEXPL(ETIM))
285.87 200 CONTINUE
285.88 FUNCM = (FSO - FM) * DEXPL(CTNLCM'QSORT(BRC-X(2)))
285.89 RETURN
285.9 END
285.91 C
285.92 C SUBROUTINE FQS
286 C THIS ROUTINE COMPUTES THE CHARGE QS STORED ON THE SEMICONDUCTOR
290  IMPLICIT REAL*8 (A-Z)
291  INTEGER CNTRL1, ITMAX1, LP, NTIMES
292  C
293  COMMON /AREA1/ CBAR, CHISC, CHISV, CJCM,
294  #   CJVM, CPSII, CS, CTNLVM, CTNLVM, EGAP,
295  #   JOD, JORG, JUPC, NC, ND, NNO, NV, Q, PHIO, PNO,
296  #   PSISIO, VTERM, QFIX
297  COMMON /AREA3/ ETAJ, ETAMC, ETAMV, ETAJ, JVM, JVPHII, NPSURF, PHI, PSII, PSIS,
298  #   PSURF, QS, THVM, THCM, U, V
299  COMMON /AREA4/ CNTRL1, ITMAX1, LP, NTIMES
300  C
301  PXN=PSNO*DEXPL(PHI)
302  NXN=NNO*PXN-PNO
303  ETAC=-(PSIS*PHIO)
304  NSURF=NC*FD102(ETAC)
305  C
306  NNSURF=NC*FD302(ETAC)
307  ETAJ=-(EGAP-PHIO-PHI)
308  PSII=UY*FD102(ETAJ)
309  C
310  ARGMT=NSURF-NXN+PSURF-PXN+NO*PSIS
311  C
312  WRITE(5,5) ARGMT
313  5 FORMAT(1X, 'WARNING: SQUARE OF SURFACE FIELD IS NEGATIVE',5X,1D14)
314  ARGMT=0.0DO
315  C
316  IF(ARGMT .GE. 0.0DO) GOTO 100
317  RETURN
318  END
319  SUBROUTINE FJVM
320  THIS ROUTINE COMPUTES THE CURRENT FLOW JVM BETWEEN THE
321  VALENCE BAND AND THE METAL
322  C
323  IMPLICIT REAL*8 (A-Z)
324  C
325  COMMON /AREA1/ CBAR, CHISC, CHISV, CJCM,
326  #   CJVM, CPSII, CS, CTNLVM, CTNLVM, EGAP,
327  #   JOD, JORG, JUPC, NC, ND, NNO, NV, Q, PHIO, PNO,
328  #   PSISIO, VTERM, QFIX
329  COMMON /AREA3/ ETAJ, ETAMC, ETAMV, ETAJ, JVM, JVPHII, NPSURF, PHI, PSII, PSIS,
330  #   PSURF, QS, THVM, THCM, U, V
331  ETAMV=-(EGAP-PHIO-V-PSIS)
332  C
333  BRV = CHISV - PSII/2.0DO
334  BRVSR = DSORT(BRV)
335  THVM=DEXPL(CTNLVM*BRVSR)
336  THVMD1 = -CTNLVM * THVM / (2.0DO*BRVSR)
337  THVMD2 = -CTNLVM * (CTNLVM/BRV) + 1.0DO/(BRVSR**3) * THVM/4.DO
338  JVM0 = CJVM * THVM * (FD11(ETAMV)-FD11(ETAJ))
339  JVM1 = CJVM * THVMD1 * (FD12(ETAJ)-FD12(ETAMV))
340  JVM2 = CJVM * THVMD2 * (FD13(ETAJ)-FD13(ETAMV))
SUBROUTINE FJPN

C

C  THIS ROUTINE COMPUTES THE MINORITY CARRIER HOLE CURRENT FLOWING

C  INTO THE SEMICONDUCTOR

IMPLICIT REAL'S (A-Z)

C

COMMON /AREA1/ CBAR,CHISC,CHISV,CJCM,

#  CJVM,CPSII,COS,CTNLCM,CTNLVM,EGAP,
#  JOD,JORG,JUPC,NC,NO,NNO,NV,Q,PHIO,PNO,
#  PSISI0,VTHERM,QFIX

COMMON /AREA3/ ETAC,ETAMC,ETAMV,ETAV,JVM,JPN,NSURF,PHI,PSII,PSIS,

#  PSURF,QS,THVM,THCM,U,V

PSIS=PSISIO-PHI

IF(PSIS .GT. 0.0D0) PSISR=PSIS

IF(PSIS .LT. 0.0D0) PSISR=0.0D0

JRG=JORG*DSQR(PSISR/PSISIO)*(DEXPL(PHI/2.0D0)-1.0D0)

JD=JOD*(DEXPL(PHI)-1.0D0)

JPN=JRG+JD-JUPC

RETURN

END

SUBROUTINE FPSIS

C

C  THIS ROUTINE COMPUTES PSIS GIVEN PHI. ASSUMING NO SURFACE STATES

IMPLICIT REAL'S (A-Z)

INTEGER CNTRL1,ITMAX1,LP,NTIMES

C

COMMON /AREA1/ CBAR,CHISC,CHISV,CJCM,

#  CJVM,CPSII,COS,CTNLCM,CTNLVM,EGAP,
#  JOD,JORG,JUPC,NC,NO,NNO,NV,Q,PHIO,PNO,
#  PSISI0,VTHERM,QFIX

COMMON /AREA3/ ETAC,ETAMC,ETAMV,ETAV,JVM,JPN,NSURF,PHI,PSII,PSIS,

#  PSURF,QS,THVM,THCM,U,V

COMMON /AREA4/ CNTRL1,ITMAX1,LP,NTIMES

CBARV=-CBAR-V

IF(CBARV .GT. 0.0D0 ) GOTO 100

IF(CBARV .LT. 0.0D0) GOTO 200

PSIS=0.0D0

RETURN

100  PSISLO=0.0D0
Listing of WIS9.S at 22:12:01 on OCT 10, 1988 for CCid=KCHU on G

399  PSISHI=CBARV
400  GOTO  300
401 200  PSISLO=CBARV
402  PSISHI=0.000
403  C
404  300  DO  500  I=1,ITMAX1
405  PSIS=(PSISLO+PSISHI)/2.000
406  CALL FOS
407  PSII=CPSII'OS
408  F1=PSIS+PSII-CBARV
409  IF(F1 .EQ. 0.000) RETURN
410  IF(F1 .GT. 0.000) GOTO  400
411  PSISLO=PSIS
412  GOTO  500
413 400  PSISHI=PSIS
414  500  CONTINUE
415  C
416  RETURN
417  END
418  DOUBLE PRECISION FUNCTION FD1(ETA)
419  C
420  C  FERMI-OIRAC INTEGRAL OF ORDER ONE
421  C
422  IMPLICIT REAL'S (A-Z)
423  INTEGER INDEX
424  C
425  DIMENSION FOFLA(81),FDFLC(81),FDFLF(81),FDLFG(81)
426  C
427  COMMON /AREA5/ CF01,FOFLA,FDFLC,FDFLF,FDLFG
428  C
429  IF(ETA .LT. -4.000) GOTO  100
430  IF(ETA .GT. 4.000) GOTO  200
431  X=(ETA*4.000/0.100+0.500
432  INDEX=X+1
433  ETA0=DFLOAT(INDEX-41)0.100
434  OXETA=ETA-ETA0
435  DXPETA=DEXPL(ETA0)
438  FD1=FDFLF(INDEX)+OXETA*(DLOG(1.000*DXPETA)+
439  #  0.100)/2.000/(1.000+1.000/1.000)
438  RETURN
439  C
440  100  FD1=DEXPL(ETA)
441  RETURN
442  C
443  200  FD1=DEXPL(-ETA)+ETA*ETA/2.000+CF01
444  RETURN
445  END
446  DOUBLE PRECISION FUNCTION FD102(ETA)
447  C
448  C  FERMI-DIRAC INTEGRAL OF ORDER ONE-HALF
449  C
450  IMPLICIT REAL'S (A-Z)
451  INTEGER INDEX
452  C
453  DIMENSION FOFLA(81),FDFLC(81),FDFLF(81),FDLFG(81)
454  C
455  COMMON /AREA5/ CF01,FOFLA,FDFLC,FDFLF,FDLFG
IF(ETA .LT. -4.000) GOTO 100
IF(ETA .GT. 4.000) GOTO 200
X=ETA+4.000/0.100+0.500
INDEX=X+1
ETA0=DFLOAT(INDEX-41)*0.100
DELETA=ETA-ETA0
FD102=DFFLOAT(INDEX)+DELETA*(DFDFLC(INDEX)+DELETA/2.000*DFDFL( INDEX))
RETURN
100 FD102=DEXPL(ETA)
RETURN
200 FD102=-DEXPL(-ETA)*ETA/2.000*CFD1
RETURN
END

DOUBLE PRECISION FUNCTION F0302(ETA)
C
C FERMI-DIRAC INTEGRAL OF ORDER THREE-HALVES
C
IMPLICIT REAL'S (A-Z)
INTEGER INDEX
DIMENSION FDFA(81),FDFLC(81),FDFLF(81),FDFLG(81)
COMMON /AREA5/ CFD1,FDFA,FDFLC,FDFLF,FDFLG
IF(ETA .LT. -4.000) GOTO 100
IF(ETA .GT. 4.000) GOTO 200
X=ETA+4.000/0.100+0.500
INDEX=X+1
ETA0=DFLOAT(INDEX-41)*0.100
DELETA=ETA-ETA0
FD302=DFDFG(INDEX)+DELETA*(DFDFL(INDEX)+DELETA/2.000*DFDFLC(INDEX))
RETURN
100 FD302=DEXPL(ETA)
RETURN
200 FD302=-DEXPL(-ETA)*ETA/2.000*CFD1
RETURN
END

DOUBLE PRECISION FUNCTION FD11(X)
IMPLICIT REAL*8 (A-Z)
DATA PI/3.141592654DO/
Y = X
IF(Y .LE. 0.DO) GOTO 50
Y = -Y
Z = 1.00*DEXPL(Y) -.250052*DEXPL(2.*Y) +.111747*DEXPL(3.*Y)
#+ -.084557*DEXPL(4.*Y) +.040754*DEXPL(5.*Y)
#-.020532*DEXPL(6.*Y) +.005108*DEXPL(7.*Y)
IF(X .LE. 0.DO) GOTO 100
Z = -Z + (X**2)/2.DO + (PI**2)/8.DO
CONTINUE
CONTINUE
FD11 = Z
RETURN
END
C
C
DOUBLE PRECISION FUNCTION FDI2(X)
IMPLICIT REAL*8 (A-Z)
DATA PI/3.14592654D0/
Y = X
IF(Y .LE. 0.00) GOTO 50
Y = -Y
Z = 1.00*DEXPL(Y) - 1.25048*DEXPL(2.*Y) + 0.037842*DEXPL(3.*Y)
Y = -Y
Z = Z + (X**3)/6.00 + (PI**2)*X/6.D0
CONTINUE
FDI2 = Z
RETURN
END

DOUBLE PRECISION FUNCTION FDI3(X)
IMPLICIT REAL*8 (A-Z)
DATA PI/3.14592654D0/
Y = X
IF(Y .LE. 0.00) GOTO 50
Y = -Y
Z = 1.00*DEXPL(Y) - 0.02592*DEXPL(2.*Y) + 0.013661*DEXPL(3.*Y)
Y = -Y
Z = Z + (X**4)/24.D0 + (((PI**2)*X)/12.D0 + 7.*(PI**4))/360.D0
CONTINUE
FDI3 = Z
RETURN
END

DOUBLE PRECISION FUNCTION DEXPL(X)
IMPLICIT REAL*8 (A-Z)
IF(X .GE. 150.000) GOTO 200
IF(X .LT. -150.000) GOTO 100
DEXPL=DEXP(X)
RETURN
DEXPL=0.00
RETURN
DEXPL=DEXP(150.000)
RETURN
END
Listing of MISETS.S at 22:05:34 on OCT 10, 1988 for CCId=KCHU on 0

1. THIS PROGRAM IS WRITTEN TO GENERATE THE STEADY-STATE CHARACTERISTICS OF THE MISETS.
3. NOTE THE DABS IN 129
4. ADD JDO TO AREA1, ALSO ADD JRNB AND CHANGE JN TO JNE IN AREA1
5. ADD ARE4; INCLUDE GAIN FEEDBACK BY BASEWIDTH MODULATION
6. NOTE LEB IN AREA4
7. MODIFY 31, 40, 60, 104, 108, 133
8. MODIFY 82 (GET A CONTINUOUS CURVE)
9. USE EXACT INTEGRAL FOR ELECTRON TUNNEL CURRENT (17, 170, 174-180, 186)
10. MODIFY 82 (GET A CONTINUOUS CURVE)
11. USE EXACT INTEGRAL FOR ELECTRON TUNNEL CURRENT (17, 170, 174-180, 186)
12. IMPLICIT REAL*8 (A-Z)
13. INTEGER FREE, I, J, IFAIL, ITMAX1, ITMAX2
14. LOGICAL NEWY, NEWA, NEWB
15. EXTERNAL COMF, FUNCT
16. DIMENSION FREE(I), LABEL(15), X(2), F(2)
17. COMMON /AREA4/, WPBASE, DELWB, DELWBO, LEB
18. COMMON /AREA5/, XBRC, XNALP1, XNALP2
19. COMMON /AREA1/, XCHIM, XCHIE, XEBG, XPHIO, JDO, JD, JB, QFIX
20. KT, XCHIH, JNO, JRO, JGB, XVBI, XVE, JNE, JRNB, JR, JG
21. COMMON /AREA2/, WP, PB, NC, NV, COS, CPSII, GS
22. COMMON /AREA3/, JIN, JIP, GAMA, ASTOC, ASTOY, XPSII
23. COMMON /AREA4/, WPBASE, DELWB, DELWBO, LEB
24. COMMON / AREA5/, XBRC, XNALP1, XNALP2
25. COMMON /SESSOM/, A(20,22), B(20,22), D(20,22)
26. NC = 2.8025
27. NV = 1.04025
28. DATA FREE/"", NEWY/.TRUE./, NEWA/.FALSE./, NEWB/.FALSE./,
29. 1.45018/. TRUE./, KBOTJ/1.38066D-23/. G/1.802180-19/,
30. EPSIO/8.85418D-12/, PI/3.141592654D0/, ME/0.910950-30/, H/0.62617D-34/
31. NC = 2.8025
32. NV = 1.04025
33. READ(5,5) (LABEL(I), I=1,15)
34. READ(5,FREE) ITMAX1, ITMAX2, ERR1, ERR2, DELY
35. WRITE(6,10) (LABEL(I), I=1,15)
36. FORMAT(15A4)
37. FORMAT(15A4)
38. READ(3,FREE) T, EGB, CHIM, CHIE
39. FORMAT(2X,'T=', F6.2,2X, 'EGB=', F4.2X, 'CHIM=', F4.2X,
40. 'CHIE=', F4.2X)
Listing of MISET5.S at 22:05:34 on OCT 10, 1988 for CCId=KCHU on Q

READ3,FREE) MCDME, MVDME, MIDME
39.3 20 FORMAT(2X,'MCDME=',F4.2,2X, 'MVDME=',F4.2,2X, 'MIDME=',F4.2,2X)
40 READ3,FREE) NCOLL, PB, WBASE, DI, LEB
40.3 25 FORMAT(2X,'NCOLL=',D9.3,2X, 'PB=',D9.4,2X, 'WBASE=',D9.3,2X,
40.6 30  'DI=',D9.3,2X, 'LEB=',D9.3,2X)

WRITE(8.10) (LABEL(I),I=1,15)
47 10 FORMAT('1',20X,15A4///)
48 WRITE(8,15) T, EGB, CHIM, CHIE
50 15 FORMAT(2X,'T=',F8.2,2X, 'EGB=',F4.2,2X, 'CHIM=',F4.2,2X,
51 '# 'CHIE=',F4.2,2X)
52 WRITE(8,20) MCDME, MVDME, MIDME
53 20 FORMAT(2X,'MCDME=',F4.2,2X, 'MVDME=',F4.2,2X, 'MIDME=',F4.2,2X)
54 WRITE(8,25) NCOLL, PB, WBASE, DI, LEB
57 WRITE(8,30) DB, OP, TAUCP, TAUR, TAUU
59 60 'TAUR=',D9.3,2X, 'TAUG=',D9.3,2X)
60 WRITE(8,35) NEPSII, NEPSIS, QFI
61 35 FORMAT(2X,'NEPSII=',F5.2,2X, 'NEPSIS=',F5.2,2X, 'QFI=',F5.2,2X,
61 70 'QFI=',D10.4,///)
62 WRITE(8,40) JB
63 40 FORMAT(2X,'JB=',D10.4///)
64 WRITE(8,50) VTHERM
66 50 FORMAT(2X, 'VTHERM=',D9.4,///)
68 WRITE(8,55) NB, NI, NI / PB
70 KV = KBOLTZ * T
71 VTHERM = KV / Q
72 XEGB = EGB / VTHERM
73 DELWBO = DSORT(2*NEPSI*EPSIO*(NCOLL/(PB*(NCOLL+PB))))*VTHERM/Q)
74 XCHIM = CHIM / VTHERM
75 XCHIE = CHIE / VTHERM
76 XCHII = XCHIE * XEGB
77 MC = MCDME * ME
78 WRITE(8,50) VTHERM
79 50 FORMAT(2X, 'VTHERM=',D9.4,///)
80 MV = MVDME * ME
81 60 FORMAT(2X,'MV=',D9.4,2X, 'XPHIO=',D9.4,2X, 'XSII=',D9.4,2X)
82 MI = MIDME * ME
83 65 FORMAT(2X,'MI=',D9.4,2X, 'ASTOC=',D9.4,2X, 'ASTOV=',D9.4,2X)
84 70 FORMAT(2X,'JDO=',D10.5,2X, 'JOE=',D10.5,2X, 'JRO=',D10.5,2X,
84 75 'JGO=',D10.5,2X, 'DELWBO=',D10.5,2X)
85 WRITE(8,70) NVB
86 70 FORMAT(2X,'XVBI=',D9.4,///)
87 WRITE(8,75) NB = NI * NI / PB
88 75 FORMAT(2X, 'NB=',D9.4,///)
89
Listing of WISE5.S at 22:05:34 on OCT 10, 1986 for CC1d=KCHU on G

80 XPHIO = DLOG(NV/PB)
81 CPSII = DI / (NEPSII*EPSIO) / VTERM
82 CSQ = DSQRT(2*K* EPSIO * NEPSIS)
82.1 GO TO 90
82.2 READ(5, FREE, END=340) VCE
82.5 PHIO = PHI
82.7 PSISO = PSIS
83 GAMMA = 4 * PI * DI / DSQRT(2*MI) / H
84 ASTOC = (4*PI/H) * (Q/H) * (MC/H) * (KT)**2)
85 ASTQV = (4*PI/H) * (Q/H) * (NV/H) * (KT)**2)
86 JDO = Q * DSQRT(DP/TAUCP) * NI * NI / NCOLL
87 JNO = Q * DB * NI * NI / PB
88 JRO = Q * NI * DSQRT2*NEPSIS*EPSIO*VTERM/(Q*PB)) / TAU
90 JGO = Q * NI * DSQRT2*NEPSIS*EPSIO*(NCOLL*PB)/(NCOLL*PB))
91 # VTERM/Q) / TAU
92 DELWBO = DSQRT2*NEPSIS*EPSIO*(NCOLL/(PB*(NCOLL*PB))*/VTERM)
93 XVBI = XEGB * XPHIO * DLOG(NC/NCOLL)
94 #
95 #
96 #
97 WRITE(8,50) VTERM
98 50 FORMAT(2X, 'VTERM=', D9.4/
99 WRITE(8,80) NB, XPHIO, CPSII
100 60 FORMAT(2X, 'NB= , D9.4, 2X, 'XPHIO=', D9.4, 2X, 'CPSII=', D9.4/
101 WRITE(8,65) GAMMA, ASTOC, ASTQV
103 DELWBO = DSQRT2*NEPSIS*EPSIO*(NCOLL/(PB*(NCOLL*PB))*/VTERM)
104 70 FORMAT(2X, 'JDO= , D10.5, 2X, 'JNO= , D10.5, 2X, 'JRO= , D10.5, 2X,
105 VBE = (XEGB - XPHIO - XPSII * XCHIM - XPSII * XCHIE) * VTERM
106 JE = JTN * JTP
107 JC = JN * JG + JG
108 JC = JNE - JRNB * JG + JD
109 JG = JNE - JRNB * JG + JD
110 86 CBC = DSQRT(Q*NEPSIS*EPSIO*NCOLL*PB / (2*(NCOLL*PB)*
111 92 # (VTERM*KVBI + VCE - VBE))
112 # 'JGO= , D10.5, 2X, 'DELWBO= , D10.5/
113 WRITE(8,75) XVBI
114 300 FORMAT(2X, 'VCE= , F5.2, 2X, 'VBE= , D11.5, 2X.
115 86 # 'JG= , D11.5, 2X, 'JC= , D11.5, 2X,
117 75 FORMAT(2X, 'XPHIO= , D9.4/)
118 310 FORMAT(2X, 'PSII= , D11.5, 2X, 'PHII= , D11.5, 2X, 'PHIER= , D11.5, 2X,
119 # 'PSIS= , D11.5, 2X, 'PSISER= , D11.5, 2X, 'DELWBO= , D11.5/
120 83 JRNB, CBC
121 86 330 FORMAT(2X, 'JRB= , D11.5, 2X, 'CBC= , D11.5/)
122 GOTO 80
123 JC = JN + JG + JD
124 JE = JTN + JTP
125 JC = JNE - JRNB * JG + JD
126 # 'PHIO= , F8.3, 2X, 'PSISO= , F8.3, 2X.
127 92 XVBI = XEGB * XPHIO * DLOG(NC/NCOLL)
128 111 #
129 C
130 C
131 C
130 80 READ(5, FREE, END=340) VCE
131 90 PHIO = PHI
132 PSISO = PSIS
133 GO TO 90
134 80 READ(5, FREE, END=340) VCE
135 PHIO = PHI
136 PSISO = PSIS
Listing of MISET5.S at 22:05:34 on OCT 10, 1988 for CC1d=KCHU on G

118.5 INTEGER FREE
117  90 XVCE = VCE / VThERM
118  Y(1,1) = PHI0 / VThERM
119  Y(2,1) = PSISO / VThERM
120  C
121  DO 220 J=2,3
122 COMMON /AREA4/ WBASE, DELWB, DELWBO, LEB
123  DATA FREE/*'*/
124  DO 210 J=1,2
125  Y(J,1) = Y(J,1) * YU. 1)
126  WRITE(6,80) X(1), X(2)
127  FORMAT(2X,'X(1)=',11.5,2X,'X(2)=',11.5/
128 220  Y(J,1) = Y(J,1) * DELY
129  WRITE(8,70)
130  FORMAT(2X.'STEP JST1')
131  NEWY = .TRUE.
132  CALL SSM(X,FZ.0,ERR1,ITMAX1,CONF,NEWY,NEWA,NEWB,IFAIL,240)
133  C
134 240 XIPHI = X(1)
135  XIPSIS = X(2)
136  JD = -JDO * (DEXPL(XEGB-XPHIO-X(2)-XCHIM-XPSII-XCHIE)-1)
137  EXPX = EXPXCB = DEXPL(XVB)-1
138  JRB = JNO * (ITANHY - ISINHY) * (EXPX + EXPXCB)
139  JNE = JNO * (ITANHY * EXPX - ISINHY * EXPXCB)
140  C
141 280 PHI = X(1) * VThERM
142  PSIS = X(2) * VThERM
143  PSII = XPSII * VThERM
144  PHI = DABS(X(1)-XPHI) * VThERM
145  PSIS = DABS(X(2)-XPSII) * VThERM
146  VBE = (XECB - XPHIO - X(2) - XCHIM - XPSII - XCHIE) * VThERM
147  JE = JTN + JTP
JC = JN + JG + JD
JC = JNE - JNJB + JG + JD

CBC = DSORT(Q*NEPSIS*EPSIO*NCOLL*PB / (2*(NCOLL+PB))

C
C WRITE(6,300) VCE, VBE, JE, JC, PHIO, 'SISO, IFAIL
C
C 300 FORMAT(2X,'VCE=',F5.2,2X,'VBE=',D11.5,2X,
C # 'JE=',D11.5,2X,'JC=',D11.5,2X,
C # 'PHIO=',F6.3,2X,'PSISO=',F6.3,2X,'IFAIL=',I5/)
C
C WRITE(6,310) PSII, PHI, PHIER, PSIS, PSISER, DELWB
C 310 FORMAT(2X,'PSII=',D11.5,2X,'PHI=',D11.5,2X,'PHIER=',D11.5,2X,
C # 'PSIS=',D11.5,2X,'PSISER=',D11.5,2X.'DELWB=',D11.5/)
C
C WRITE(6,320) OS, JTN, JTP, JNE, JR, JD
C 320 FORMAT(2X,'OS=',D11.5,2X,'JTN=',D11.5,2X,'JTP=',D11.5,2X,
C # 'JNE=',D11.5,2X,'JR=',D11.5,2X,'JD=',D11.5/)
C
C WRITE(6,330) JRN8, CBC
C 330 FORMAT(2X,'JRN8=',D11.5,2X,'CBC=',D11.5////)
C
C GOTO 80
C
C STOP
C
C END

C
C SUBROUTINE COMF(X,F)
C IMPLICIT REAL*8 (A-Z)
C INTEGER FREE
C
C DIMENSION X(2), F(2), FREE(1)
C COMMON /AREA1/ XCHIM, XCHIE, XEGB, XPHIO, JDO, JD, JB, OFIX,
C # KT, XCHIM, JNO, JRO, JGO, XVBI, XVCB, JNE, JNJB, JR, JG
C COMMON /AREA2/ NB, PB, NC, NV, COS, CPSII, QS
C COMMON /AREA3/ JTN, JTP, GAMMA, ASTOC, ASTOV, XPSII
C COMMON /AREA4/ WBASE, DELWB, DELWBO, LEB
C DATA FREE/*"*/
C
C WRITE(6,80) X(1), X(2)
C CO FORMAT(2X,'X(1)='.D11.5,2X,'X(2)=',D11.5)
C CALL FOS(X)
C WRITE(6,70)
C CO FORMAT(2X,'STEP FOS')
C XPSII = CPSII * (QS+OFIX)
C CALL FJTN(X)
C WRITE(6,80)
C CO FORMAT(2X,'STEP FJTN')
C CALL FJTP(X)
C WRITE(6,90)
C CO FORMAT(2X,'STEP FJTP')
C XVCB = XVCE-XEGB*XPHIO+X(2)+XCHIM*XPSII-XCHIE
C SRXVCB = DSORT(XVBI+XVCB-XEGB*XPHIO+X(2)+XCHIM*XPSII-XCHIE)
C SRXVCB = DSORT(XVBI+XVCB)
C DELWB = DELWBO * SRXVCB
C WBEFF = WBASE - DELWB
C ITANHY = 1/DTANH(WBEFF)
C ISINHV = 1/OSINH(WBEFF)
C EXPX1 = OEXPL(X(1)) - 1
C EXFXCB = DERXPL(XVCB) - 1
JRNB = JNO * (ITANHY - ISINHY) * (EXPX1 + EXPXCB)
JNE = JNO * (ITANHY - ISINHY) * (EXPX1 - EXPXCB)
JM = JNO * DEXP(X(1)) / (DBASE - DELWB)
JR = JNO * DSORT(DABS(X(2))) * (DEXP(X(1)/2) - 1)
JG = JGO * SRXVCB
JD = -JDO * (DEXP(XGB-XPHIO-X(2)-XCHIM-XPSII+XCHIE-XVCE) - 1)
JQ = JGO * SRXVCB
JD = -JDO * EXPXCB
F(1) = (JN - JR - JTN) * 1.0*4
F(2) = JB + JD - JTP - JTP
F(1) = (JNE - JTN) * 1.0*4
F(2) = JB + JD - JTP - JTN
WRITE(8,100) X(1), X(2), F(1), F(2)
FORMAT(2X,'X(1)='D11.5,2X,'X(2)='D11.5)
WRITE(8,110) JTN, JTP, JN, JR, JG
FORMAT(2X,'JTN='D11.5,2X, 'JTP='D11.5,2X, 'JN='D11.5,2X, 'JR='D11.5
WRITE(6,120) OS, XPSII
FORMAT(2X,'OS=',D11.5,2X, 'XPSII=',D11.5)
RETURN
END
SUBROUTINE FQS(X)
IMPLICIT REAL*8 (A-Z)
DIMENSION X(2)
COMMON /AREA 1/ XCHIM, XCHIE, XEGB, XPHIO, JDO, JD, JB, QFIX,
# KT, XCHIM, JNO, JRO, JD, XVBI, XVCE, JNE, JRB, JR, JB
COMMON /AREA2/ NB, PB, NC, NV, COS, CPSII, QS
C
C
C
C
C
C
C
END
SUBROUTINE FJTN(X)
IMPLICIT REAL*8 (A-Z)
DIMENSION X(2), LOWER(2), UPPER(2)
COMMON /AREA1/ XCHIM, XCHIE, XEGB, XPHIO, JDO, JD, JB, QFIX,
# KT, XCHIM, JNO, JRO, JD, XVBI, XVCE, JNE, JRB, JR, JB
COMMON /AREA2/ NB, PB, NC, NV, COS, CPSII, QS
C
C
C
C
C
C
C
RETURN
END
SUBROUTINE FJTN(X)
IMPLICIT REAL*8 (A-Z)
DIMENSION X(2), LOWER(2), UPPER(2)
COMMON /AREA1/ XCHIM, XCHIE, XEGB, XPHIO, JDO, JD, JB, QFIX,
# KT, XCHIM, JNO, JRO, JD, XVBI, XVCE, JNE, JRB, JR, JB
COMMON /AREA2/ NB, PB, NC, NV, COS, CPSII, QS
C
C
C
C
C
C
C
RETURN
END
SUBROUTINE FJTN(X)
IMPLICIT REAL*8 (A-Z)
DIMENSION X(2), LOWER(2), UPPER(2)
COMMON /AREA1/ XCHIM, XCHIE, XEGB, XPHIO, JDO, JD, JB, QFIX,
# KT, XCHIM, JNO, JRO, JD, XVBI, XVCE, JNE, JRB, JR, JB
COMMON /AREA2/ NB, PB, NC, NV, COS, CPSII, QS
C
C
C
C
C
C
C
RETURN
END
SUBROUTINE FJTN(X)
IMPLICIT REAL*8 (A-Z)
DIMENSION X(2), LOWER(2), UPPER(2)
COMMON /AREA1/ XCHIM, XCHIE, XEGB, XPHIO, JDO, JD, JB, QFIX,
# KT, XCHIM, JNO, JRO, JD, XVBI, XVCE, JNE, JRB, JR, JB
COMMON /AREA2/ NB, PB, NC, NV, COS, CPSII, QS
C
C
C
C
C
C
C
RETURN
END
SUBROUTINE FJTN(X)
IMPLICIT REAL*8 (A-Z)
DIMENSION X(2), LOWER(2), UPPER(2)
COMMON /AREA1/ XCHIM, XCHIE, XEGB, XPHIO, JDO, JD, JB, QFIX,
DOUBLE PRECISION FUNCTION FUNCM(N, XE)

IMPLICIT REAL*8 (A-Z)

INTEGER N

DIMENSION XE(N)

COMMON /AREA5/ XBRC, XNALP1, XNALP2

XNETT1 = XE(1) * XE(2) - XNALP1

FN1 = 0.0

IF (XNETT1 .GE. 150.00) GO TO 100

FN1 = 1.00 / (1.00 * DEXPL(XNETT1))

CONTINUE

XNETT2 = XE(1) * XE(2) - XNALP2

FN2 = 0.0

IF (XNETT2 .GE. 150.00) GO TO 200

FN2 = 1.00 / (1.00 * DEXPL(XNETT2))

CONTINUE

FUNCM = (FN1 - FN2) * DEXPL(-GAMMA * (XBRC - XE(2)) * KT)

RETURN

END

SUBROUTINE FJTP(X)

IMPLICIT REAL*8 (A-Z)

DIMENSION XI(2)

COMMON /AREA1/ XCHIM, XCHIE, XEGB, XPHIO, JD, JB, QFIX, KT, XCHIH, JNO, JRO, JGO, XVBI, XVCE, JNE, JRN, JG

COMMON /AREA3/ JTN, JTP, GAMMA, ASTOC, ASTOV, XPSII

XNBET1 = XCHIM * XPSII - XCHIE - XEGB

XNBET2 = -X(2) - XPHIO

BRV = (XCHIH + XPSII/2) * KT

THETVO = DEXPL(-GAMMA * DSQRT(BRV))

THETV1 = GAMMA * THETVO / (2 * DSQRT(BRV))

THETV2 = GAMMA * (GAMMA / BRV + 1 / DSQRT(BRV))^3 * THETVO/4

JTP = ASTOV * (THETVO + (FD1(XNBET2) - FD1(XNBET1)))

# + (KT * THETV1) * (FD2(XNBET2) - FD2(XNBET1))

# + ((KT * 2) * THETV2) * (FD3(XNBET2) - FD3(XNBET1))

RETURN

END
DOUBLE PRECISION FUNCTION DEXPL(X)
IMPLICIT REAL*8 (A-Z)
IF(X .LT. -150.000) GOTO 100
DEXPL=DEXPL(X)
RETURN
100 DEXPL=0.000
RETURN
END

DOUBLE PRECISION FUNCTION FD1(X)
IMPLICIT REAL*8 (A-Z)
DATA PI/3.14159285400/
Y = X
IF(Y .LE. 0.00) GOTO 50
Y = -Y
50 Z = 1.00*DEXPL(Y) - .250052*DEXPL(2.Y) + .111747*DEXPL(3.Y)
# - .064557*DEXPL(4.Y) + .040754*DEXPL(5.Y)
# - .020532*DEXPL(6.Y) + .005108*DEXPL(7.Y)
IF(X .LE. 0.00) GOTO 100
Z = -Z * (X**2)/2.00 + (PI**2)/8.00
100 CONTINUE
FD1 = Z
RETURN
END

DOUBLE PRECISION FUNCTION FD2(X)
IMPLICIT REAL*8 (A-Z)
DATA PI/3.14159265400/
Y = X
IF(Y .LE. 0.00) GOTO 50
Y = -Y
50 Z = 1.00*DEXPL(Y) - .125046*DEXPL(2.Y) + .037642*DEXPL(3.Y)
# - .018183*DEXPL(4.Y) + .012484*DEXPL(5.Y)
# - .007486*DEXPL(6.Y) + .003446*DEXPL(7.Y)
IF(X .LE. 0.00) GOTO 100
Z = Z * (X**3)/8.00 + (PI**2)*X/8.00
100 CONTINUE
FD2 = Z
RETURN
END

DOUBLE PRECISION FUNCTION FD3(X)
IMPLICIT REAL*8 (A-Z)
DATA PI/3.14159265400/
Y = X
IF(Y .LE. 0.00) GOTO 50
Y = -Y
50 Z = 1.00*DEXPL(Y) - .062592*DEXPL(2.Y) + .013681*DEXPL(3.Y)
# - .008796*DEXPL(4.Y) + .012057*DEXPL(5.Y)
# - .010589*DEXPL(6.Y) + .053448*DEXPL(7.Y)
IF(X .LE. 0.00) GOTO 100
Z = Z * (X**4)/24.00 + ((PI*X)**2)/12.00 + 7.* (PI**4)/380.00
100 CONTINUE
RETURN
END
CONTINUE
FD3 = Z
RETURN
END

DOUBLE PRECISION FUNCTION FD102(X)
IMPLICIT REAL*8 (A-Z)
DATA PI/3.14159265400/
IF (X .LE. 0.DO) GOTO 100
IF (X .LE. 2.DO) GOTO 200
IF (X .LE. 4.00) GOTO 300
GOTO 400
100 FD102 = DEXPL(X) - 0.353568'DEXPL(2'X) * 0.192439*DEXPL(3'X)
          - 0.122973*DEXPL(4*X) * 0.077134*DEXPL(5*X)
          + 0.0382828*DEXPL(6*X) * 0.008348*DEXPL(7*X)
GOTO 500
200 FD102 = 0.765147 * X'(0.604911 * X - 0.189885 +
          # X*(0.020307 * XM-0.004380 * X*(-0.000388 *
          # X*0.000133)))
GOTO 500
300 FD102 = 0.777114 * X'(0.581307 + X'(0.206132 +
          # X'(0.017680 + X'(0.004940 + X'(0.000784 -
          # X*0.000036))))
GOTO 500
400 XN2 = 1/(X**2)
FD102 = (X**1.5) * (0.752253 + XN2*(0.828195 +
          # XN2*10.880830 + XN2*(25.7829 + XN2*(-553.838 +
          # XN2*(3531.43 -XN2*3254.05))))
RETURN
END
Listing of PISETP.S at 21:56:53 on OCT 10, 1988 for CC14=KCHU on G

This program is written to generate the steady-state characteristics of the unannealed PNP IETS.

Define all variables

Implicit real's (a-z)
Integer free, i, j, ifail, itmax1, itmax2
Logical newy, newa, newb
External comf

Dimension free(1), label(15), x(3), f(3)

Common /area1/ xchim, xchie, xegb, xphi, jyo, jdo, jb, ofix
  # k, xchim, jno, jro, jgo, xybi, xvce, jne, jrnb, jr, jg
  # xegb, xeta
Common /area2/ nb, pb, nc, nv, cos, CPSII, QS
Common /area3/ jpol, jtn, jtp, gamma, astoc, astov, xpsii
Common /area4/ wbase, delwb, delwbo, leb
Common /area5/ dpol, lpol
Common /se$jom/ a(20,22), b(20), y(22,21)

Data free/*'*/., newy., true.,./., newa., false.,./., newb., false.,./.
  # ni/1.45d16, kbolzt/1.380660e-23, 0/1.602180e-19,
  # epsio/8.85418d-12,
  # pi/3.14159265400, me/0.910950e-30, h/6.626170e-34

Reverse conduction and valence
To model PNP transistor

NV = 2.8D25
NC = 1.04D25

Read iteration specifications from input file 5

Read(5,5) (label(i), i=1,15)
  format(15A4)
Read(5,free) itmax1, itmax2, err1, err2, dely

Read device parameters from input file 3

Read(3,free) T, egb, chim, chie
Read(3,free) mcdme, mvdme, wdom
Read(3,free) ncoll, pb, wbase, di, leb
Read(3,free) db, dp, taucp, taur, tauq
Read(3,free) nepsi, nepsis, ofix
Read(3,free) dpol, lpol, zeta, pegb
Read(3,free) jb
Listing of PISETP.S at 21:58:53 on OCT 10, 1988 for CC1d=KCHU on Q

59 C ECHO DEVICE PARAMETERS TO OUTPUT FILE 6
60 C
61 C WRITE(8,10) (LABEL(I),I=1,15)
62 10 FORMAT('I',20X,5A4//)
63 C WRITE(6,15) T, EGB, PEGB, CHIM, CHIE
64 15 FORMAT(2X,'T='F6.2,2X,'EGB='F4.2,2X,'PEGB='F4.2,2X,
65 4 C 'CHIM='F4.2,2X,'CHIE='F4.2/) 
66 C WRITE(6,20) MCDME, MVDME, MIDME
67 20 FORMAT(2X,'MCDME='D9.3,2X,'MVDME='D9.3,2X,'MIDME='D9.3/) 
68 C WRITE(8,25) NCOLL, PB, WBASE, DI, LEB
69 25 FORMAT(2X,'NCOLL='D9.3,2X,'PB='D9.3,2X,'WBASE='D9.3,2X,
70 C 'DI='D9.3,2X,'LEB='D9.3/) 
71 C WRITE(6,30) DB, DP, TAUCP, Taur, TauG
72 30 FORMAT(2X,'DB='D9.3,2X,'DP='D9.3,2X,'TAUCP='D9.3,2X,
73 C 'TAUR='D9.3,2X,'TAUG='D9.3/) 
74 C WRITE(6,35) NEPSII, NEPSIS, QFIX
75 35 FORMAT(2X,'NEPSII='F5.2,2X,'NEPSIS='F5.2,2X,'QFIX='D10.4/) 
76 C WRITE(6,37) DPOL, LPOL, ZETA
77 37 FORMAT(2X,'DPOL='D9.3,2X,'LPOL='D9.3,2X,'ZETA='F4.2/) 
78 C WRITE(6,40) JB
79 40 FORMAT(2X,'JB='D10.4///)
80 C
81 C NORMALIZE DEVICE PARAMETERS TO UNITS OF KT
82 C ALL VARIABLES STARTING WITH AN X ARE IN KT'S
83 C
84 C KT = KBOLTZ * 1
85 C VTERM = KT / Q
86 C XZETA = ZETA / VTERM
87 C XEGB = EGB / VTERM
88 C XPGB = PEGB / VTERM
89 C XCHIM = CHIM / VTERM
90 C XCHIE = CHIE / VTERM
91 C XCHIH = XCHIE - XEGB
92 C MC = MCDME - ME
93 C MV = MCDME - ME
94 C MI = MIDME - ME
95 C
96 C COMPUTE ALL CONSTANTS
97 C
98 C NB = NI * NI / PB
99 C XPHIO = DLOG(NV/PB)
100 C CPSII = DI / (NEPSII*EPSII) / VTERM
101 C CQS = DSORT(2*KT, EPSII : NEPSII)
102 C GAMMA = 4 * PI * DI * DSORT(2*NI) / H
103 C ASTOC = (4*PI/H) * (Q/H) * (MC/H) * (KT**2)
104 C ASTOV = (4*PI/H) * (Q/H) * (NV/H) * (KT**2)
105 C JDO = 0 * DSORT(DP/TAUCP) * NI * NI / NCOLL
106 C JNO = 0 * DB * NI * NI / PB
107 C JRO = 0 * DI * DSORT(2*EPSII*EPSII*VTERM/(Q*PB)) / TAUR
108 C JGO = 0 * DI * DSORT(2*EPSII*EPSII*VTERM/(Q*PB)) / TAUG
DELWBO = DSQRT(2*NEPSIS*EPSIO*(NCOLL/(PB*(NCOLL+PB)))*VTHERM/Q)
XVBI = XEBB - XPHIO - DLOG(NE/NV)

WRITE ALL CONSTANTS TO OUTPUT FILE 8

WRITE(6,50) VTHERM
FORMAT(2X,'VTHERM=',D9.4/)  
WRITE(6,60) NB, XPHIO, CPSII
FORMAT(2X,'NB=',09.4,2X,'XPHIO=',D9.4,2X,'CPSII=',D9.4/)  
WRITE(6,65) GAMMA, ASTOC, ASTOV
FORMAT(2X,'GAMMA=',09.4,2X,'ASTOC=',D9.4,2X,'ASTOV=',D9.4/)  
WRITE(6,70) J00, JNO, JRO, JGO, DELWBO
FORMAT(2X,'J00=',10.5,2X,'JNO=',10.5,2X,'JRO=',10.5,2X,'JGO=',10.5,2X,'DELWBO=',10.5/)  
WRITE(6,75) XVBI
FORMAT(2X,'XVBI=',09.4///)

READ INITIAL VALUES FOR THE VARIABLES TO BE USED IN THE ITERATION FROM INPUT FILE 5
READ(5,FREE,END=340) VCE, PHIO, PSISO, PHI10
XVCE = VCE / VTHERM

SET UP INITIAL CONDITIONS FOR NUMERICAL SUBROUTINE PACKAGE SSM
Y(1,1) = PHIO / VTHERM
Y(2,1) = PSISO / VTHERM
Y(3,1) = PHI10 / VTHERM
DO 220 I=2,4
  DO 210 J=1,3
    Y(J,I) = YIJ,1)
  220 Y(I,1) = Y(I,1) + DELY
RUN SUBROUTINE SSM
NEWY = .TRUE.
CALL SSM(X,F,3,0,ERR1,ITMAX1,COMF,NEWY,NEWA,NEWB,IFAIL,&240)

XIPHI = X(1)
XIPSIS = X(2)
XIPHI1 = X(3)
NEWY = .FALSE.
CALL SSM(X,F,3,0,ERR2,ITMAX2,COMF,NEWY,NEWA,NEWB,IFAIL,&280)

CONVERT RETURNED VALUES TO VOLTS
Listing of PISETP.S at 21:59:53 on OCT 10, 1988 for CC1d=KCHU on G

175 C
176 260 PHI = X(1) * VTHERM
177 PSIS = X(2) * VTHERM
178 PHI1 = X(3) * VTHERM
179 PSII = XPSII * VTHERM
180 C
181 C CALCULATE DIFFERENCE BETWEEN FIRST AND SECOND CALL TO SSM
182 C
183 C PHIER = DABS(X(1)-XPHI) * VTHERM
184 PSISER = DABS(X(2)-XPSIS) * VTHERM
185 PHIER1 = DABS(X(3)-XPHI1) * VTHERM
186 C
187 C COMPUTE FINAL VOLTAGES AND CURRENT DENSITIES TO BE OUTPUT
188 C
189 C VBE = (XEGB - XPHIO - XCHIM - XPSII) / (VTHERM)
190 194 * VTHERM
191 JE = JTN + JTP
192 JC = JN + JG + JD
193 JC = JNE - JRNB + JG + JD
194 CBO = DSQRT(Q*NEPSIS*EPSIO*NCOLL*PB / (2*(NCOLL*PB) * VTHERM) * VBI - VCE - VBE))
195 IFAIL
196 C
197 C WRITE VOLTAGES, CURRENT DENSITIES AND VARIABLES
198 C DETERMINED FOR THE FINAL SOLUTION
199 C TO OUTPUT FILE 6
200 C
201 C WRITE(6,300) VCE. VBE. JE. JC. PHI. PSISO. PHIIO. IFAIL
202 300 FORMAT(2X,'VCE=',F5.2,2X,'VBE=',D14.8,2X,
203 'JE=',D11.5,2X,'JC=',D14.8,10X,'PHIO=',F8.3,2X,
204 'PSISO=',F8.3,2X,'PHIIO=',F8.3,2X,'FAIL=',I5/)
205 C
206 WRITE(6,310) PHII, PHI. PSIS. PSISER
207 310 FORMAT(2X,'PHII=',D11.5,2X,'PHI=',D14.8,10X,'PHI=',D11.5,2X,PHIIO,PHIIO,'FAIL=',I5/)
208 C
209 WRITE(6,320) JS. JTN. JTP. JNE. JG. JD. JPO
210 320 FORMAT(2X,'JTN=',D14.8,2X,'JTP=',D11.5,2X,'JNE=',D11.5,2X,JD,JD,'JPO=',D11.5/)
211 C
212 WRITE(6,330) JRNB, CBC
213 330 FORMAT(2X,'JRNB=',D11.5,2X,'CBC=',D11.5,/) 214 STOP 80
215 ENDDC
216 C
217 C MAIN SUBROUTINE USED TO CALCULATE
218 C CURRENT DENSITIES USED IN CONTINUITY EQUATIONS
219 C
220 SUBROUTINE COMF(X,F)
221 IMPLICIT REAL*8 (A-Z)
INTEGER FREE
DIMENSION X(3), F(3), FREE(I)
COMMON /AREA1/ XCHIM, XCHIE, XEGB, XPHIO, JDO, JD, JB, QFIX,
#   KI, XCHIH, JNO, JRO, JB, XBI, XV2E, JNE, JRB, JR, JG,
#   XPEGB, XZETA
COMMON /AREA2/ NB, PB, NC, NV, COS, CPSII, QS
COMMON /AREA3/ JPOL, JTN, JTP, GAMMA, ASTOC, ASTOV, XPSII
COMMON /AREA4/ WBASE, DELWB, DELWBO, LEB
COMMON /AREA5/ DQPOL, DPOL
DATA FREE/'"/
C
WRITE(6,60) X(1), X(2), X(3)
FORMAT(2X,'X(1)=',011.5X, 'X(2)=',011.5X, 'X(3)=',011.5X)
C
COMPUTE OS:CHARGE STORED AT OXIDE-SEMICONDUCTOR INTERFACE
C   AND PSI2:VOLTAGE ACROSS THE OXIDE
CALL FOS(X)
WRITE(6,70)
FORMAT(2X,'STEP FOS')
XPSII = CPSII * (QOS+QFIX)
C
COMPUTE JPOL:MINORITY CARRIER CURRENT DENSITY IN POLYSILICON
CALL FJPOL(X)
WRITE(6,75)
FORMAT(2X,'STEP FJPOL')
C
COMPUTE JTN:MAJORITY CARRIER TUNNELING CURRENT DENSITY
CALL FJTN(X)
WRITE(6,80)
FORMAT(2X,'STEP FJTN')
C
COMPUTE JTP:MINORITY CARRIER TUNNELING CURRENT DENSITY
CALL FJTP(X)
WRITE(6,90)
FORMAT(2X,'STEP FJTP')
C
COMPUTE VCB:COLLECTOR-BASE VOLTAGE
XVCB = XVCE-XEGB*XPHIO*X(2)+XCHIM*XPSII-XCHIE*XZETA
C
COMPUTE DELWB:BASE WIDTH-MODULATION
SRXVCB = DSQRT(XVBI*XCVE-XEGB*XPXIO*X(2)*XCHIM*XPSII-XCHIE)
DELWB = DELWBO * SRXVCB

COMPUTE JRN: RECOMBINATION CURRENT DENSITY IN THE BASE

WBEFF = WBASE - DELWB
ITANHY = 1/ DTANH(WBEFF/LEB)
ISINHY = 1/ DSINH(WBEFF/LEB)
EXPX1 = DEXPX(X(1)) - 1
EXPXCB = DEXPX(-XVCB) - 1
JRN = JNO * (ITANHY - ISINHY) * (EXPX1 + EXPXCB)

COMPUTE JNE: MAJORITV CARRIER EMITTER CURRENT DENSITY

JNE = JNO * (ITANHY * EXPX1 - ISINHY * EXPXCB)

COMPUTE JG: GENERATION CURRENT DENSITY IN THE CB JUNCTION

JG = JQO * SRXVCB

COMPUTE JD: DIFFUSION CURRENT DENSITY IN THE CB JUNCTION

JD = -JDO * (DEXPX(XGB-XPHIO)-XCHIM-XPSII+XCHIE-VCVE)-1)

COMPUTE ERROR FUNCTIONS
USING CURRENT DENSITY CONTINUITY EQUATIONS FOR NUMERICAL PACKAGE SSM

F(1) = (JNE - JTN) * 1.D-4
F(2) = JB + JG + JD - JTP - JRN
F(3) = JPOL - JTP

WRITE(6,100) X(1), X(2), X(3), F(1), F(2), F(3)
C00 FORMAT(2X,'X(1)',D11.5,2X, 'X(2)',D11.5,2X, 'X(3)',D11.5,2X,
'H(1)',D11.5,2X, 'H(2)',D11.5,2X, 'H(3)',D11.5,2X)
C #
C WRITE(6,110) JTN, JTP, JG, JPOL
C10 FORMAT(2X,'JTN=',D11.5,2X, 'JTP=',D11.5,2X,
'JG=',D11.5,2X, 'JPOL=',D11.5)/
C #
C WRITE(6,120) QS, XPSII
C20 FORMAT(2X,'QS=',D11.5,2X, 'XPSII=',D11.5/)
SUBROUTINE TO CALCULATE CHARGE STORED AT OXIDE-SEMI INTERFACE

SUBROUTINE FOSIX)
IMPLICIT REAL'S (A-Z)
DIMENSION X(3)
COMMON /AREA 1/ XCHIM, XCHIE, XEGB, XPHIO, JDO, JD, JB, OFIX,
# KT, XCHIH, JNO, JRO, JGO, XVBI, XVCE, JNE, JRN, JR, JG,
# XEPGB, XZETA
COMMON /AREA2/ NB, PB, NC, NV, COS, CPSII, OS

NSURF = NC * FD102(X(2)*XPHIO*X(1)-XEGB)
PSURF = NV * FD102(-XPHI0-X(2))
ARGMNT = NSURF - NB*DEXPL(X(1)) - PSURF - PB * PB*X(2)
IF (ARGMNT .GE. 0.000) GOTO 100
WRITE(6.5) ARGMNT
FORMAT(1X, 'WARNING: SOUR OF SURFACE FIELD IS NEGATIVE'., 5X, D11.4)
ARGMNT = 0.000
100 OS = -COS * DSORT(ARGMNT)
IF (X(2) .LT. 0.000) OS = -OS
RETURN
END

SUBROUTINE TO CALCULATE MINORITY CARRIER CURRENT DENSITY IN THE POLYSILICON

SUBROUTINE FJPOL(X)
IMPLICIT REAL'S (A-Z)
DIMENSION X(3)
COMMON /AREA 1/ XCHIM, XCHIE, XEGB, XPHIO, JDO, JD, JB, OFIX,
# KT, XCHIH, JNO, JRO, JGO, XVBI, XVCE, JNE, JRN, JR, JG,
# XEPGB, XZETA
COMMON /AREA2/ NB, PB, NC, NV, COS, CPSII, OS
COMMON /AREA3/ JPOL, JTN, JTP, GAMMA, ASTOC, ASTOV, XPSII
COMMON /AREA5/ DPOL, LPOL

XBETA1 = XEPGB - X(3) - XZETA
PJ = NV * FD102(-XBETA1)
JPOL = 1.60218D-19 * DPOL * PJ / LPOL
RETURN
END

SUBROUTINE TO CALCULATE MAJORITY CARRIER TUNNELING CURRENT DENSITY USING A THREE TERM SERIES APPROXIMATION

SUBROUTINE FJTN(X)
IMPLICIT REAL'S (A-Z)
DIMENSION X(3)
COMMON /AREA 1/ XCHIM, XCHIE, XEGB, XPHIO, JDO, JD, JB, OFIX,
Listing of PISETP.S at 21:59:53 on OCT 10, 1988 for CC1d=KCHU on G

407 # KT, XCHIH, JNO, JRO, JGO, XVB1, XVE, JNE, JRN8, JN, JG,
408 # XPEGB, XZETA
409 COMMON /AREA3/ JPOL, JTN, JTP, GAMMA, ASTOC, ASTOV, XPSII
410 C
411 IF (XPSII .LT. 0) GOTO 10
412 XNALP1 = -XPSII - XZETA
413 XNALP2 = X(2) • XPHIO • X(1) • XEGB
414 BRC = (XCHIE • XPSII/2) • KT
415 GOTO 20
416 JTN = ASTOC • (THETCO • (FD1(XNALP1)-FD1(XNALP2)))
417 RETURN
418 END
419 C
420 COMMON /AREA3/ JPOL, JTN, JTP, GAMMA, ASTOC, ASTOV, XPSII
421 IF (XPSII .LT. 0) GOTO 10
422 XNBET1 = XZETA • X(3) - XPEGB
423 XNBET2 = -XPEGB - XPSII • XEGB - X(2) - XPHIO
424 BRV = (XCHIH • XPSII/2) • KT
425 GOTO 20
426 RETURN
427 END
428 C
429 COMMON /AREA3/ JPOL, JTN, JTP, GAMMA, ASTOC, ASTOV, XPSII
430 COMMON /AREA3/ JPOL, JTN, JTP, GAMMA, ASTOC, ASTOV, XPSII
431 SUBROUTINE TO CALCULATE
432 MINORITY CARRIER TUNNELING CURRENT DENSITY
433 USING A THREE TERM SERIES APPROXIMATION
434 C
435 SUBROUTINE FJTP(X)
436 IMPLICIT REAL'S (A-Z)
437 DIMENSION X(3)
438 COMMON /AREA1/ XCHIM, XCHIE, XEGB, XPHIO, JDO, JD, JB, QFIX,
439 # KT, XCHIH, JNO, JRO, JGO, XVB1, XVE, JNE, JRN8, JN, JG,
440 # XPEGB, XZETA
441 COMMON /AREA3/ JPOL, JTN, JTP, GAMMA, ASTOC, ASTOV, XPSII
442 IF (XPSII .LT. 0) GOTO 10
443 XNBET1 = XZETA • X(3) - XPEGB
444 XNBET2 = -XPEGB - XPSII • XEGB - X(2) - XPHIO
445 BRV = (XCHIH • XPSII/2) • KT
446 GOTO 20
447 RETURN
448 END
449 C
450 COMMON /AREA3/ JPOL, JTN, JTP, GAMMA, ASTOC, ASTOV, XPSII
451 IF (XPSII .LT. 0) GOTO 10
452 XNBET1 = XZETA • X(3) - XPEGB
453 XNBET2 = -XPEGB - XPSII • XEGB - X(2) - XPHIO
454 BRV = (XCHIH • XPSII/2) • KT
455 GOTO 20
456 RETURN
457 END
458 C
459 COMMON /AREA3/ JPOL, JTN, JTP, GAMMA, ASTOC, ASTOV, XPSII
460 COMMON /AREA3/ JPOL, JTN, JTP, GAMMA, ASTOC, ASTOV, XPSII
461 C
462 C
463 C
EXPONENTIAL FUNCTION

DOUBLE PRECISION FUNCTION DEXP(X)
IMPLICIT REAL*8 (A-Z)
IF(X .LT. -.150.000) GOTO 100
DEXP=EXP(X)
RETURN
END

FERMI-DIRAC FUNCTION OF ORDER 1 USING A SERIES APPROXIMATION

DOUBLE PRECISION FUNCTION FD1(X)
IMPLICIT REAL*8 (A-Z)
DATA PI/3.141592654D0/
Y = X
IF(Y .LE. 0.00) GOTO 50
Y = -.Y
Z = 1.00*DEXP(Y) -.250052*DEXP(2.*Y) -.111747*DEXP(3.*Y)
END

FERMI-DIRAC FUNCTION OF ORDER 2 USING A SERIES APPROXIMATION

DOUBLE PRECISION FUNCTION FD2(X)
IMPLICIT REAL*8 (A-Z)
DATA PI/3.141592654D0/
Y = X
IF(Y .LE. 0.00) GOTO 50
Y = -.Y
Z = 1.00*DEXP(Y) -.125046*DEXP(2.*Y) -.037642*DEXP(3.*Y)
END

FERMI-DIRAC FUNCTION OF ORDER 3 USING A SERIES APPROXIMATION
DOUBLE PRECISION FUNCTION F03(X)
IMPLICIT REAL'S (A-Z)
DATA PI/3.141592654D0/
V =  X
IF(V .LE. 0.00) GOTO 50
Y = -Y
Z = 1.00*DEXP1(Y) - 0.02592*DEXP1(2.*Y) * 0.13861*DEXP1(3.*Y)
# -0.009796*DEXP1(4.*Y) + 0.012978*DEXP1(5.*Y)
# -0.010859*DEXP1(6.*Y) + 0.003448*DEXP1(7.*Y)
IF(X .LE. 0.00) GOTO 100
Z = -Z + (X**4)/24.00 + (PI**2*X)**2)/12.00 + 7.*(PI**4)/
# - 380.00
CONTINUE
FD3 = Z
RETURN
END

DOUBLE PRECISION FUNCTION FD102(X)
IMPLICIT REAL'S (A-Z)
DATA PI/3.141592854D0/
IF (X .LE. 0.00) GOTO 100
IF (X .LE. 2.00) GOTO 200
IF (X .LE. 4.00) GOTO 300
GOTO 400
100
FD102 = DEXP1(X) - 0.353568*DEXP1(2*X) + 0.192439*DEXP1(3*X)
# -0.122973*DEXP1(4*X) + 0.077139*DEXP1(5*X)
# -0.036231*DEXP1(6*X) + 0.008348*DEXP1(7*X)
GOTO 500
200
FD102 = 0.785147 * X*(0.804911 + X*(0.189885 +
# X*(0.020307 + X*(-0.004380 + X*(-0.000366 +
# X*(-0.000036))))))
GOTO 500
300
FD102 = 0.777114 * X*(0.581307 + X*(0.208132 +
# X*(0.017880 + X*(-0.006549 + X*(0.000784 -
# X*(-0.000036))))))
GOTO 500
400
XN2 = 1/(X**2)
FD102 = (X**1.5) * (0.752253 * XN2*(0.928195 +
# XN2*(0.680630 + XN2*(25.7829 + XN2*(-553.836 +
# XN2*(3531.43 -XN2*3264.65))))))
RETURN
END
Listing of PIMSP.S at 21:52:10 on OCT 10, 1988 for CC1d=KCHU on G

C
C
C THIS PROGRAM IS WRITTEN TO GENERATE THE STEADY-STATE
CHARACTERISTICS OF THE ANNEALED PNP PETS.
C
C
C DEFINE ALL VARIABLES
C
C IMPLICIT REAL*8 (A-Z)
INTEGER FREE, I, J, IFAIL, ITMAX1, ITMAX2
LOGICAL NEWY, NEWA, NEWB
EXTERNAL COMF
DIMENSION FREE(1), LABEL(15), X(4), F(4)
COMMON /AREA1/ XCHIM, XCHIE, XEGB, XPHIO, JDO, JD, JB, OFIX,
   # KT, XCHIM, JNO, JRO, JDO, XVBI, XVCE, JME, JNMB, JR, JG,
   # XPEGB, XZETA, XZETAZ, XVIE, JPO, JRM, JPE, JP2
COMMON /AREA2/ NE, NB, PB, NC, NY, COS, CPSII
COMMON /AREA3/ JPOL, JTN, JIP, GAMMA, ASTOC, ASTOV
COMMON /AREA4/ WBASE, DELWB, DELWBO, LEB, LHE, LE, DE
COMMON /AREA5/ DPOL, LPOL
COMMON /SESSION/ A(20,22), B(20), Y(22,21)
C
C DATA FREE/* */, NEWY/, TRUE/, NEWA/, FALSE/, NEWB/, FALSE/,.
   # NI/1.45D16/, KBOLTZ/1.38068D-23/, Q/1.80218D-19/,.
   # EPSIO/8.85418D-12/,.
   # PI/3.141592654D0/, ME/0.910950-30/, H/8.62817D-34/.
C
C REVERSE CONDUCTION AND VALENCE FOR PNP DEVICE
C
C NV = 2.8025
NC = 1.04025
C
C READ ITERATION SPECIFICATIONS FROM INPUT FILE 5
C
C READ(5,5) (LABEL(I),I=1,15)
S FORMAT(15A4)
READ(5,FREE) ITMAX1, ITMAX2, ERR1, ERR2, DELY
C
C READ DEVICE PARAMETERS FROM INPUT FILE 3
C
C READ(3,FREE) T, EGB, CHIM, CHIE
READ(3,FREE) MCDME, MVDME, MIDME
READ(3,FREE) ME, MCOLL, PB, WBASE, DI, LEB, LHE
READ(3,FREE) LE, DE, DB, DP, TAUCP, TAN, TAO
READ(3,FREE) NEPSII, NFPSI, OFIX
READ(3,FREE) DPOL, LPOL, ZETA, PEB
READ(3,FREE) JB
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C ECHO DEVICE PARAMETERS TO OUTPUT FILE

C

WRITE(6,10) (LABEL(I),I=1,15)

 FORMAT('1. 20X, 15A4///)

WRITE(6,15) T, EGB, PEGB, CHIM, CHIE

FORMAT(2X,'T= ',F6.2,2X, 'EGB=',F4.2,2X, 'PEGB=',F4.2,2X,

# 'CHIM=',F4.2,2X, 'CHIE=',F4.2///)

WRITE(6,20) MCDME, MVOME, MIDME

FORMAT(2X,'MCOME=',F4.2,2X, 'MVOME=',F4.2,2X, 'MIDME=',F4.2///)

WRITE(6,25) NE, NCOLL, PB, DI, LEB, LHE, WBASE

FORMAT(2X,' NE=',09.3,2X, 'NCOLL=',D9.3,2X, 'PB=',D9.3,


WRITE(6,30) LE, DE, DB, DP, TAUCP, TAUR, TAUG


# 'TAUCP=',D0.3,2X, 'TAUR=',D0.3,2X, 'TAUG=',D0.3///)

WRITE(6,35) NEPSII, NEPSIS, QFIX

FORMAT(2X,'NEPSII=',F5.2,2X, 'NEPSIS=',F5.2,2X, 'QFIX=',F10.4///)

WRITE(6,40) JB

FORMAT(2X,'JB=',D10.4///)

C

NORMALIZE DEVICE PARAMETERS TO UNITS OF KT

ALL VARIABLES STARTING WITH AN X ARE IN KT'S

C

KT = KBOLTZ * T

VTHERM = KT / Q

XZETA = ZETA / VTHERM

XEGB = EGB / VTHERM

XPGB = PGB / VTHERM

XCHIM = CHIM / VTHERM

XCHIE = CHIE / VTHERM

XCHIH = XCHIE - XEGB

MC = MCDME * ME

MV = MVOME * ME

MI = MIDME * ME

NB = NI * NI / PB

XPHIO = DLOG(NV/PB)

XZETA2 = DLOG(NC/NE)

CPSII = DI / (NEPSII*EPSIO) / VTHERM

COS = DSQRT(2*KT*EPSIO*NEPSIS)

GAMMA = 4*PI*DI/DSQRT(2*MI) / H

ASTOG = (4*PI/H) * (Q/H) * (MC/H) * (KT**2)

ASTOW = (4*PI/H) * (Q/H) * (MV/H) * (KT**2)

JDO = 0 * DSQRT(DP/TAUCP) * NI * NI / NCOLL

JNO = Q * DB * NI * NI / PB

JNO = Q * DB * NI * NI / PB / LEB

JPO = 0 * DE * NI * NI / LHE / NE
JRO = 0 * NI * DSRT(2*NEPSI'EPSIO*((NE*PB)/(NE2*PB))
# "VTERM/Q) / TAUR
JGO = 0 * NI * DSRT(2*NEPSI'EPSIO*((NCOLL+PB)/(NCOLL*PB))
# "VTERM/Q) / TAUR
DELWBO = DSRT(2*NEPSI'EPSIO*(NCOLL/(PI*(NCOLL+PB))*/VTERM/Q)
XVBI = XEGN - XPHIO - DLOG(NC/NCOLL)
XVBIE = XEGN - XPHIO - XZETA2

WRITE ALL CONSTANTS TO OUTPUT FILE 8

WRITE(6,50) VTHERM, XZETA2
50 FORMAT(2X,'VTHERM*'.09.4,2X, 'XZETA2*'.011.5/)
WRITE(6,80) NB, XPHIO, CPSII
80 FORMAT(2X,'NB*'.09.4,2X, 'XPHIO*'.09.4,2X, 'CPSII*'.09.4/)
WRITE(6,85) GAMMA, ASTOC, ASTOV
85 FORMAT(2X,'GAMMA*'.D9.4,2X, 'ASTOC*'.09.4,2X, 'ASTOV*'.09.4/) WRITE(6,70) JDO, JNO, JPO, JRO, JGO, DELWBO
70 FORMAT(2X,'JDO*'.09.4,2X, 'JNO*'.09.4,2X, 'JPO*'.09.4,2X, 'JRO*'.09.4,2X, 'JGO*'.09.4,2X, 'DELWBO*'.09.4/)
WRITE(6,75) XVBI, XVBIE
75 FORMAT(2X,'XVBI*'.09.4,2X, 'XVBIE*'.09.4/) READ INITIAL VALUES FOR THE VARIABLES TO BE USED IN FIRST ITERATION FROM INPUT FILE 5
READ(5,FREE,END=340) VCE, PHIO, PHI10, PHI20, PSIIO
XVCE = VCE / VTHERM

SET UP INITIAL CONDITIONS FOR NUMERICAL SUBROUTINE PACKAGE SSM
Y(1,1) = PHIO / VTHERM
Y(2,1) = PHI10 / VTHERM
Y(3,1) = PHI20 / VTHERM
Y(4,1) = PSIIO / VTHERM

DO 220 I=2,5
DO 210 J=1,4
210 Y(J,1) = Y(J,1) + DELY
220 Y(I,1) = Y(I,1) + DELY

RUN SUBROUTINE SSM
NEWY = .TRUE.
CALL SSM(X,F,4,0,ERR1,ITMAX1,CONF,NEWY,NEWA,NEWB,IFAIL,&240)
XIPHI = X(1)
XIPHI1 = X(2)
XIPHI2 = X(3)
XIPSII = X(4)
NEWY = .FALSE.
CALL SSM(X,F,4,0,ERR2,ITMAX2,CONF,NEWY,NEWA,NEWB,IFAIL,&260)
CONVERT RETURNED VALUES TO VOLTS
PHI = X(1) * VTERM
PHI1 = X(2) * VTERM
PHI2 = X(3) * VTERM
PSII = X(4) * VTERM

COMPUTE BAND BENDING AT EB JUNCTION
PSIS = (XEGB - XPHIO - X(1) - XZETA2) * VTERM
CALCULATE DIFFERENCE BETWEEN FIRST AND SECOND DALL TO SSM

PHIER = DABS(X(1)-XPHI1) * VTERM
PHIER = DABS(X(2)-XPHI2) * VTERM
PHIER = DABS(X(3)-XPHI2) * VTERM
PSIIER = DABS(X(4)-XIPSII) * VTERM

COMPUTE FINAL VOLTAGES AND CURRENT DENSITIES TO BE OUTPUT
VBE = (X(1) + XZETA2 - X(4) - XZETA) * VTERM
JE = JTN + JTP
JC = JNE - JRNB + JG + JD
C = DSORTQ*EPSIS*EPSIO*NColl*PB / (2*(NColl*PB) *

WRITE VOLTAGES, CURRENT DENSITIES AND VARIABLES
DETERMINED FOR THE FINAL SOLUTION
TO OUTPUT FILE 8

WRITE(8,300) VCE, VBE, JE, JC
FORMAT(2X,'VCE=',F5.2,2X, 'VBE=',D14.8,2X, 'JE=',D11.5,4X,

WRITE(8,305) PHIO, PHIO, PHIO, PSIIO, IFAIL
FORMAT(2X,'PHIO=',F6.3,2X, 'PHIO=',F6.3,2X,

WRITE(8,310) PSIS, PHI, PHI, PSI
FORMAT(2X,'PSIS=',D11.5,2X, 'PHI=',D14.8,7X, 'PHI=',D11.5,2X,

WRITE(8,315) PHIER, PHIER, PHIER, PSIIER, DELWB
FORMAT(2X,'PHIER=',D11.5,2X, 'PHIER=',D11.5,2X,

WRITE(8,320) JTN, JTP, JTP, JNE, JR, JD, JPO
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233  # 'JNE=',011.5,3X, 'JR=',011.5,38X, 'JG=',011.5,2X,  
234  # 'JD=',011.5,2X, 'JPOL=',011.5/ 
235  WRITE(6,330) JANB, JRM, JPB, CBC 
236  330  FORMAT(2X,'JANB=',011.5,2X, 'JRM=',011.5,2X,  
237  # 'JP2=',011.5,2X, 'CBC=',011.5///) 
238  GOTO 80 
239  STOP 
240  END 

C MAIN SUBROUTINE USED TO CALCULATE 
C CURRENT DENSITIES USED IN CONTINUITY EQUATIONS 
C C SUBROUTINE COMF(X,F) 
C IMPLICIT REAL*8 (A-Z) 
C INTEGER FREE 
C DIMENSION X(4), F(4), FREE(1) 
C COMMON /AREA1/ XCHIM, XCHIE, XEGB, XPHIO, JDO, JD, JB, OFIX,  
C # K,T, XCHIM, JNO, JRO, JGQ, XVBI, XVE, JNE, JNM, JHR, JG,  
C # XBGB, XETA, XZETA2, XVBIE, JPO, JRM, JPB, JPB2  
C COMMON /AREA2/ NE, NB, PB, NC, NV, COS, CPSII 
C COMMON /AREA3/ JPN, JTP, GAMMA, ASTOC, ASTOV 
C COMMON /AREA4/ WBASE, DELWB, DELWBO, LEB, LHE, LE, DE 
C COMMON /AREAS/ OPOL, LPOL 
C DATA FREE/**/ 

C COMPUTE JPM: MINORITY CARRIER CURRENT DENSITY IN POLYSILICON 
C CALL FJPOL(X) 
C COMPUTE JTN: MAJORITY CARRIER TUNNELING CURRENT DENSITY 
C CALL FJTN(X) 
C COMPUTE JTP: MINORITY CARRIER TUNNELING CURRENT DENSITY 
C CALL FJTP(X) 
C COMPUTE VCB: COLLECTOR-BASE VOLTAGE 
C XVCB = XVCE-X(1)-XZETA2*X(4)*X2ETA 
C COMPUTE DELWB: BASE WIDTH MODULATION 
C SRXVCB = DSORT(XVBI+XVCB) 
C DELWB = DELWBO * SRXVCB 
C
WBEFF = WBASE - DELWB
ITANHY = 1/ DTANH(WBEFF/LEB)
ISINHY = 1/ DSINH(WBEFF/LEB)
ITANHU = 1/ DTANH(LE/LHE)
ISINHU = 1/ DSINH(LE/LHE)

EXP1 = DEXPL(X(1)) - 1
EXP1R = DEXPL(X(1)/2) - 1
EXP3 = DEXPL(X(3)) - 1
EXPXB = DEXPL(-XVCB) - 1

C

COMPUTE JRNB: RECOMBINATION CURRENT DENSITY IN THE BASE

JRNB = JNO * (ITANHY - ISINHY) * (EXP1 * EXPXB)

C

COMPUTE JNE: MAJORITY CARRIER EMITTER CURRENT DENSITY

JNE = JNO * (ITANHY * EXP1 - ISINHY * EXPXB)

C

COMPUTE JPE: MINORITY CARRIER DIFFUSION CURRENT DENSITY AT THE MONOCRYSTALLINE EMITTER DEPLETION EDGE

JPE = JPO * (ITANHU * EXP1 - ISINHU * EXP3)

C

COMPUTE JP2: MINORITY CARRIER EMITTER CURRENT DENSITY

JP2 = JPO * (ISINHU * EXP1 - ITANHU * EXP3)

C

COMPUTE JRM: RECOMBINATION CURRENT DENSITY IN THE EMITTER

JRM = JPO * ((ITANHU - ISINHU) * (EXP1 + EXP3))

C

JN = JNO * DEXPL(X(1)) / (WBASE-DELWB)

C

COMPUTE JR: RECOMBINATION CURRENT DENSITY IN THE EB JUNCTION

JR = JRO * EXP1R * DSRT(XVBIE - X(1))

C

COMPUTE JG: GENERATION CURRENT DENSITY IN THE CB JUNCTION

JG = -JGO * EXPXB * SRTXVCB

C

COMPUTE JO: DIFFUSION CURRENT DENSITY IN THE CB JUNCTION

C
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349  C  JD = -JDO * (DEXP(XEGB-XPHIO-X(2)-XCHIM-XPSII*XCHIE-XVCE)-1)
350  JD = -JDO * EXPXCXB
351  C  COMPUTE ERROR FUNCTIONS
352  C  USING CURRENT DENSITY CONTINUITY EQUATIONS
353  C  FOR NUMERICAL PACKAGE SSM
354  C  F(1) = (JNE - JTN - JRM - JR) * 1.0 - 4
355  F(2) = JPE - JB - JD - JG - JRNB - JR
356  F(3) = JPOL - JTP
357  F(4) = JP2 - JTP
358  RETURN
359  END

388  C  SUBROUTINE FOS(X)
389  IMPLICIT REAL*8 (A-Z)
390  DIMENSION X(3)
391  COMMON /AREA1/ XCHIM, XCHIE, XEGB, XPHIO, JDO, JD, JB, OFIX.
392  COMMON /AREA2/ KT, XCHIH, JNO, JRO, JGO, XVBI, XVCE, JNE, JRNB, JR, JG,
393  COMMON /AREA3/ XPEGB, XZETA
394  COMMON /AREA4/ NB, PB, NC, NV, COS, CPSII, QS
395  COMMON /AREA5/ NSURF, NA, NC, FD102(X(2)*XPHI0*X(D-XEGB)
396  COMMON /AREA6/ [FORM](A,DI1.4)
397  COMMON /AREA7/ ARGMT = NSURF - NB*DEXP(X(1)) + PSURF - PB + PB*X(2)
398  IF (ARGMT .GE. 0.000) GOTO 100
399  WRITE(6,5) ARGMT
400  FORMAT(6,5) 'WARNING: SOURCE OF SURFACE FIELD IS NEGATIVE',
401  5X, DI1.4)
402  ARGMT = 0.000
403  QO = -COS * DSORT(ARGMT)
404  IF (X(2) .LT. 0.000) QO = -QO
405  RETURN
406  END

493  C  SUBROUTINE TO CALCULATE MINORITY CARRIER CURRENT DENSITY IN THE POLYSILICON
494  C  SUBROUTINE FJPOL(X)
495  IMPLICIT REAL*8 (A-Z)
496  DIMENSION X(4)
497  COMMON /AREA1/ XCHIM, XCHIE, XEGB, XPHIO, JDO, JD, JB, QFIX.
498  COMMON /AREA2/ KT, XCHIH, JNO, JRO, JGO, XVBI, XVCE, JNE, JRNB, JR, JG,
499  COMMON /AREA3/ XPEGB, XZETA, XZETA2, XVBE, JPO, JRM, JPE, JP2
500  COMMON /AREA4/ NE, NB, PB, NC, NV, COS, CPSII
501  COMMON /AREA5/ JPN, JTN, JTP, GAMMA, ASTQC, ASTOV
502  COMMON /AREA6/ OPOL, LPOL
503  COMMON /AREA7/ XBETA1 = XPEGB - X(2) - XZETA
504  PJ = NV * FD102(XBETA1)
SUBROUTINE TO CALCULATE
MAJORITY CARRIER TUNNELING CURRENT DENSITY
USING A THREE TERM SERIES APPROXIMATION

SUBROUTINE FJTN(X)
IMPLICIT REAL'S (A-Z)
DIMENSION X(4)
COMMON /AREA1/ XCHIM, XCHIE, XEGB, XPHIO, JDO, JD, JB, OFIX,
# KT, XCHIM, JNO, JRO, JGO, XVBI, XVCIE, JNE, JRN, JR, JG,
# XPEGB, XZETA, XZETA2, XVBE, JPO, JRM, JPE, JP2
COMMON /AREA3/ JPOL, JTN, JTP, GAMMA, ASTOC, ASTOV

IF (X(4) .LT. 0) GOTO 10
XNALP1 = -XPEGB - X(4) - XZETA • XEGB
XNALP2 = -XZETA2
BRC = (XCHIE  • X(4)/2)  • KT
GOTO 20

10 XNALP1 = -XZETA
XNALP2 • -XEGB • X(4) • XPEGB • XZETA2
BRC = (XCHIE  • X(4)/2)  • KT

THETCO = DEXPL(-GAMMA ' DSORT(BRC))
THETC1 = GAMMA ' THETCO/(2 ' DSORT(BRC))
THETC2 = GAMMA * (GAMMA/BRC + 1/(DSORT(BRC))**3) * THETCO/4
JTN = ASTOC ' (THETCO * (FD1(XNALP1)-FD1(XNALP2))
# + (KT*THETC1) ' (FD2(XNALP1)-FD2(XNALP2))
# + (KT**2*THETC2) * (FD3(XNALP1)-FD3(XNALP2)))
RETURN
END

SUBROUTINE TO CALCULATE
MINORITY CARRIER TUNNELING CURRENT DENSITY
USING A THREE TERM SERIES APPROXIMATION

SUBROUTINE FJTP(X)
IMPLICIT REAL'S (A-Z)
DIMENSION X(3)
COMMON /AREA1/ XCHIM, XCHIE, XEGB, XPHIO, JDO, JD, JB, OFIX,
# KT, XCHIM, JNO, JRO, JGO, XVBI, XVCIE, JNE, JRN, JR, JG,
# XPEGB, XZETA, XZETA2, XVBE, JPO, JRM, JPE, JP2
COMMON /AREA3/ JPOL, JTN, JTP, GAMMA, ASTOC, ASTOV

IF (X(4) .LT. 0) GOTO 10
XNBET1 = -XPEGB • XZETA + X(2)
XNBET2 = -XPEGB • X(4) + XZETA2 + X(3)
BRC = (XCHIM • X(4)/2)  • KT
GOTO 20

10 XNBET1 = -XEGB + X(4) + XZETA + X(2)
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465      XNBET2 = -XEG8 * ZZETAZ * X(3)
466      BRV = (XCH8H - X(4)/2) * KT
467      C
468      C
469      20  THETV0 = DEXPL(-GAMMA*DSORT(BRV))
470      THETV0 = GAMMA * THETV0/(2*DSORT(BRV))) 3) * THETV0/4
472      JTP = ASTOV * (THETV0 * (FD1(XNBET2)-FD1(XNBET1))
473      # + (KT*THETV1 * (FD1(XNBET2)-FD2(XNBET1))
474      # + ((KT**2)*THETV2 * (FD3(XNBET2)-FD3(XNBET1))
475      RETURN
476      END
477      C
478      C
479      C
480      C
481      C
482      C
483      C
484      C
485      C
486      C
487      C
488      C
489      C
490      C
491      C
492      C
493      C
494      C
495      C
496      C
497      C
498      C
499      C
500      C
501      C
502      C
503      C
504      C
505      C
506      C
507      C
508      C
509      C
510      C
511      C
512      C
513      C
514      C
515      C
516      C
517      C
518      C
519      C
520      C
521      C
522      C

EXPONENTIAL FUNCTION

DOUBEL PRECISION FUNCTION DEXPL(X)

IMPLICIT REAL*8 (A-Z)

DATA PI/3.14159265400/

IF(Y .LE. 0.0) GOTO 50

Z = 1.00*DEXPL(Y) .250052*DEXPL(2.*Y) +.111747*DEXPL(3.*Y)

# .018557*DEXPL(4.*Y) .045756*DEXPL(5.*Y)

# .020532*DEXPL(6.*Y) +.005108*DEXPL(7.*Y)

IF(Y .LE. 0.0) GOTO 100

Z = -Z * (X**2)/2.00 + (PI**2)/6.00

CONTINUE

FD1 = Z

RETURN

END

FERMI-DIRAC FUNCTION OF ORDER 1

USING A SERIES APPROXIMATION

DOUBEL PRECISION FUNCTION FD1(X)

IMPLICIT REAL*8 (A-Z)

DATA PI/3.14159265400/

Y = X

IF(Y .LE. 0.00) GOTO 100

Z = -Z * (X**2)/2.00 + (PI**2)/6.00

CONTINUE

FD1 = Z

RETURN

END

FERMI-DIRAC FUNCTION OF ORDER 2

USING A SERIES APPROXIMATION

DOUBEL PRECISION FUNCTION FD2(X)

IMPLICIT REAL*8 (A-Z)

DATA PI/3.14159265400/

Y = X

IF(Y .LE. 0.00) GOTO 50

Y = -Y
DOUBLE PRECISION FUNCTION FD02(X)
IMPLICIT REAL*8 (A-Z)
DATA PI/3.1415926540D0/
Y = X
IF(Y .LE. 0.0D0) GOTO 50
Y = -Y
50 Z = 1.0D0*DEXPL(Y) - 0.25048D0*DEXPL(2.*Y) + 0.037842D0*DEXPL(3.*Y)
51 # -0.016181*DEXPL(4.*Y) + 0.012484D0*DEXPL(5.*Y)
52 # -0.007486D0*DEXPL(6.*Y) + 0.002133D0*DEXPL(7.*Y)
53 IF(X .LE. 0.0D0) GOTO 100
54 Z = -Z * (X**4)/24.0D0 + ((PI*X)**2)/12.0D0 + 7.0D0*(PI**4)/360.0D0
55 CONTINUE
FD2 = Z
RETURN
END
Listing of PIMSP.S at 21:52:10 on OCT 10, 1988 for CCId=KCHU on G

581     #    XN2*(0.680839 + XN2*(25.7629 + XN2*(-553.836 +
582     #    XN2*(3531.43 - XN2*3254.85)))))
583     500  RETURN
584     END
PUBLICATIONS


