LEVEL II

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CONTROLLED AVALANCHE TRANSIT-TIME TRIODE AMPLIFIERS

S-W. Lee

Electron Physics Laboratory
Department of Electrical and Computer Engineering
The University of Michigan
Ann Arbor, Michigan 48109

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<td>The purpose of this work is to study the theoretical effects of avalanche multiplication and collector transit time on microwave controlled avalanche transit-time triode (CATT) devices. The objectives of this report are to obtain a better device model, develop a complete one-dimensional large-signal simulation computer program, calculate the large-signal performance of Class C CATT amplifiers and make a comparison between Class C CATT and Class CBJT amplifiers. The following studies were carried out in order to achieve these objectives.</td>
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20. Abstract (Cont.)

A dc computer program was developed which calculates the dc avalanche multiplication factor vs. base-collector dc bias characteristics. The results provide an estimation of the suitability of various semiconductor materials, optimum collector geometrical structures and doping densities.

Analytical models of dc and small-signal characteristics for Read-type collector structures are given which incorporate both the avalanche multiplication and collector transit-time mechanisms. Contrary to previous findings, the small-signal characteristics indicate that a large avalanche multiplication factor decreases the RF power gain of small-signal Class A CATT amplifiers. The results are given and discussed.

A large-signal computer simulation was developed which consists of three computer programs: the emitter-base computer program (EBCP), the large-signal simulation program (LSSP), and the collector circuit computer program (CCCP). The effects of high impurity doping level in the emitter, high injection level in the base, time-varying width of the neutral base region, carrier-induced drift field in the base, nonzero minority carrier concentration at the edge of the base-collector depletion region in the base, and the feedback hole current are incorporated in EBCP. Computer program LSSP models the semiconductor region through a set of difference equations of the semiconductor equations. A current-conserving boundary condition is given. The simulation includes the velocity-electric field, diffusion-electric field, and avalanche ionization rate-electric field characteristics in the collector region. The computer program CCCP incorporates the displacement current in the collector semiconductor region and the effects of the external load impedance.

Large-signal results of Class C CATT amplifiers are obtained and are presented. Effects of base-collector dc bias, load, collector structure, and operating frequency are discussed. The simulation calculates amplifier output power, gain, and efficiency. It also gives the emitter-base current and voltage waveforms; avalanche multiplication factor; waveforms of voltage across the base-collector depletion region and collector terminal current; and spatial distributions of electrons, holes, and electric field at any time instant. Large-signal output power, gain, efficiency, dynamic range, and inherent bandwidth of Class C CATT and BJT amplifiers are compared and suggestions for further studies are given.
FOREWORD

This report describes the investigation of simulation studies at the Electron Physics Laboratory, Department of Electrical and Computer Engineering, The University of Michigan, Ann Arbor, Michigan. The work was sponsored by the Air Force Systems Command, Air Force Avionics Laboratory, Wright-Patterson Air Force Base, Ohio under Contract No. F33615-77-C-1132.

The work reported herein was performed during the period July 1977 to July 1980 by Dr. Shiu-Wuu Lee. The report was released by the author in September 1980.

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dc When used as a subscript of a current or current density indicates a dc quantity.

E Electric field intensity (V/cm).

$E_B$ Breakdown electric field intensity (V/cm).

$E_E$ Electric field intensity in the high-field region of a head-type collector structure (V/cm).

$E_{sus}$ Minimum electric field intensity required to sustain charge carriers at their saturation drift velocity (V/cm).

$E_{sus_n}, E_{sus_p}$ Minimum electric field intensities required to sustain electrons and holes at their saturation drift velocities, respectively (V/cm).

e Electronic charge ($1.6021 \times 10^{-19}$ C).

f Frequency (Hz).

$f_{max}$ Maximum frequency of oscillation above which operating power gain is less than or equal to 1 (Hz).

G Avalanche generation rate ($cm^{-3} \cdot s^{-1}$).

$G_{in}, G_{out}$ Input and output conductances as defined in Fig. 3.6.

$G_L, G_S$ Load and source conductances as defined in Fig. 3.7.

$G_{L_{opt}}, G_{S_{opt}}$ Optimum load and source conductances for maximum operating gain.

$G_P$ Operating power gain (dB).

$\bar{\beta}_{ij}$ Small-signal conductances as defined by Eqs. 3.54 and 3.55.

$h, w$ Widths of base fingers and emitter fingers, respectively (cm).

I When used as a subscript indicates a difference equation field point.

$I_B$ Base current (A).

$I_{C_1}, I_{C_2}, I_{CDE}, I_{CTE}$ Collector current, collector displacement current, emitter-base diffusion current, emitter-base displacement current, respectively (A).
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Small-signal voltages as defined in Fig. 3.6.

Small-signal voltages as defined in Fig. 3.6.

Junction breakdown voltage, junction built-in potential, collector punch-through voltage, and minimum voltage across the collector depletion region needed to maintain carriers at their saturation drift velocity at all times, respectively (V).

Base-collector de bias, emitter-base input signal, and small-signal emitter-base input signal, respectively (V).

Voltages across the emitter-base terminal, emitter-base depletion region, base-collector junction and the load, the load, and the base-collector depletion region, respectively (V).

Drift velocities of electrons and holes, respectively (cm·s⁻¹).

Saturation drift velocities of electrons and holes, respectively (cm·s⁻¹).

Carrier saturation drift velocity (cm·s⁻¹).

Boundaries of depletion regions as defined in Fig. 3.1.

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Width of neutral base region and separation between emitter-base and base-collector metallurgical junctions, respectively (cm).

Width of high doping region of a LO-HI-LC collector structure (cm).

Width of neutral emitter region and depth of the emitter-base metallurgical junction, respectively (cm).

Width of the whole collector region (cm).

Collector depletion width when \( V_T = V_{SUS} \).

Space coordinates in the collector region and the base region, respectively (cm).
\(X_B, X_B\)

Width of base region beneath the emitter finger and the depth of emitter-base metallurgical junction from the device surface, respectively (cm).

\(Y_{in}, Y_{out}\)

Input and output admittances, respectively.

\(Y_L, Y_S\)

Load and source admittances, respectively.

\(Y_{ij}\)

Device small-signal \(y\)-parameters defined by Eqs. 3.48 and 3.49.

\(\dot{Y}_{ij}\)

Device small-signal \(y\)-parameters defined by Eqs. 3.54 and 3.55.

\(Z_L\)

Load impedance.

When used as a subscript of a voltage or an electric field quantity indicates a dc quantity.

When used as a subscript for a current, a voltage or an electric field quantity represents a small-signal RF quantity.

\(a\)

Ionization rates (cm\(^{-1}\)).

\(a_o\)

Dc ionization rates (cm\(^{-1}\)).

\(a_o\)

Dc common-base current gain.

\(a_{inj}^o, a_{to}^o\)

Dc emitter injection efficiency and dc base transport factor, respectively.

\(g_o\)

Dc common-emitter current gain.

\(\delta\)

Amplifier input tunability.

\(\varepsilon\)

Dielectric permittivity (F·cm\(^{-1}\)).

\(\eta\)

Amplifier efficiency (percent).

\(\eta_c\)

Collector dc to ac conversion efficiency (percent).

\(\theta_D, \theta_T\)

Transit angles of the drift region and the whole collector, respectively.

\(\theta_{inj}, \theta_{sc}\)

Phase angles associated with charge injection and delay of charge injection due to space charge, respectively.

\(\nu_n, \nu_p\)

Electron and hole mobility, respectively.
\( \rho \)  
Resistivity (\( \Omega \cdot \text{cm} \)).

\( \sigma_B \)  
Conductivity in the base region (\( \text{mho} \cdot \text{cm}^{-1} \)).

\( \tau_e \)  
Delay time in responses due to heavy impurity doping level in the emitter region.

\( \tau_T \)  
Transit time in the carrier generation region.

\( \tau_{n^*p} \)  
Lifetimes of electron minority carriers and hole minority carriers in p- and n-type semiconductors, respectively (s).

\( \tau_R \)  
Dielectric relaxation time.

\( \psi \)  
Potential variation in p-n junction depletion region.
CHAPTER I. INTRODUCTION

1.1 Historical Background

1.1.1 Bipolar Junction Transistors (BJTs). Of all semiconductor devices the BJT, an acronym for bipolar junction transistor, is the most important. Its invention brought about an unprecedented growth of research and development in solid-state physics and engineering. Transistors are now key elements, for example, in high-speed computers, in space vehicles and satellites, and in all modern communication and power systems.

The development of point-contact transistors by Bardeen and Brattain\(^1\) was announced in 1948. Then in 1949, Shockley\(^2\) proposed a junction bipolar transistor and laid out the basic theory of this fundamental structure. In the microwave power bipolar transistor area, pioneering work was done by Earley\(^3\) and Pritchard,\(^4\) where they considered high-frequency effects, and Ebers and Moll,\(^5\) Fletcher,\(^6\) and Emeis et al.,\(^7\) who studied high-power effects on transistor operation. Some papers have been devoted to the analysis of the mutual dependence of these effects.\(^8\)-\(^10\) On the other hand, some more recent studies have been published concerning the theory and characterization of microwave bipolar transistors.\(^11\)-\(^15\)

At the inception of the BJT, transistors were able to operate at frequencies up to a few hundred kHz only and the choice of the semiconductor material was restricted to germanium, since the material had been produced with sufficient purity and in single-crystal form. Theory
predicted that they should be able to operate at much higher frequencies, even into the microwave frequency range, by a reduction of their overall dimensions. In particular, it was recognized that the transit time of charge carriers through the device and the rate of change of electrical charge stored within the device would limit the frequency response. In order to improve the performance at higher frequencies, the base width of the BJT must be reduced to reduce the transit time and the active device area must decrease in order to reduce the capacitance or stored charge. These refinements demanded a much tighter control on all three dimensions of the device. Transistor technology has enjoyed many breakthroughs, particularly in the alloy-junction and grown-junction techniques and in zone-refining, diffusion, epitaxial, planar, beam-lead, and ion implantation technologies. These breakthroughs have helped to increase the power and frequency capabilities of transistors, as well as their reliability, by many orders of magnitude.

With the present technology and without electron-beam or x-ray exposure (i.e., 1 µm linewidth), an aspect ratio of 20:1 is theoretically attainable with interdigitated, overlay or mesh structures. The practical limit, however, seems to be approximately 10:1. As to the base layer width, the lowest value achieved under a compromise between minimum base transit time and minimum base spreading resistance is 0.1 µm.

The physical properties of the semiconductor theoretically determine the ultimate electrical performance of the transistor. For example, Johnson showed that the maximum frequency of operation will be proportional to $E_B v_s$, where $E_B$ is the junction breakdown electric field and $v_s$ is the scattering limited carrier drift velocity. As the
size of the device is reduced to achieve high-frequency performance, the voltage must be maintained at a value sufficient to give the required power output. In the limit, a further reduction in device dimension parallel to the electric field direction would be impossible because the electric field would exceed \( E_B \). In practice, the frequency limit derived by Johnson has not been reached. For technological reasons, Si is preferred to Ge and GaAs for microwave bipolar transistors. The technological superiority of Si is due mainly to the ability of silicon dioxide to act as a diffusion mask and the ability to etch very fine patterns in this oxide. The oxides of Ge and GaAs are not as stable as silicon dioxide and for these semiconductor materials chemical vapor deposited silicon dioxide and silicon nitride, when used as a diffusion mask or as an insulating material, produced results inferior to thermally grown silicon dioxide on silicon. Much progress in GaAs technology has been achieved in the last few years. Another reason why Si material is preferred is its good thermal conductivity which is a factor of two better than GaAs. Good thermal conductivity is an important concern, especially in high-power applications.

Bipolar junction transistors have the following advantages which assure their place in the microwave power semiconductor device family:

1. Due to their three-terminal configuration, their application, particularly as amplifiers or switching devices, is much easier and the corresponding circuits much simpler than for two-terminal devices.

2. Due to their operation with both majority- and minority-carrier types, very high local current density can be reached, much higher than in the majority-carrier devices.
3. The operating power efficiency is high, particularly for Class C amplifiers.

4. The operational bandwidth is large, particularly for Class A amplifiers.

5. The power gain in amplifier operation is relatively high.

6. Signal distortion is lower than in two-terminal devices. Noise level is lower than in avalanche diodes.

7. With present Si technology, good output power can be obtained in the X-band frequency range.26

1.1.2 Controlled-Avalanche Transit-Time Triode (CATT) Devices. Diodes, and in particular the IMPATT, which is an acronym for impact ionization avalanche transit time, have relatively simple configurations and operate close to the well-known Johnson8 material parameter limitation. Transistors, however, perform well below the material limit, in spite of considerable effort to optimize their configurations and the great advancement made in Si technology. On the other hand, three-terminal devices have many advantages over diodes as mentioned previously. A new three-terminal device was proposed by Yu et al.27 in 1974. This new device was named CATT, an acronym for controlled-avalanche transit-time triode. It is self-evident from its name that this device utilizes both avalanche multiplication and transit time. In designing a BJT device, avalanche multiplication has always been associated with junction breakdown and was always avoided in amplifier applications. Incorporation of avalanche multiplication into the BJT device is important for several design applications.28-30 For a large number of circuits, transistor junction breakdown is used to provide a reference voltage.
An ability to model such operation is desirable. For other applications, circuit performance under surge conditions must be determined. Possible malfunctions due to second breakdown can only be determined if an avalanche model is first established. Another reason for investigating avalanche multiplication is that when a BJT is biased in the avalanche multiplication region, a negative differential resistance between collector and emitter may exist. This part of the characteristic, commonly named the avalanche region, may be used for fast switching applications. In power amplifier applications, it has been established that BJTs having long collector regions, i.e., large junction breakdown voltages, can be desirable, but it was not until the discovery of the CATT device by Yu et al. that both avalanche multiplication and collector transit time were actively used to advantage in power amplifiers. In 1974, Winstanley and Carroll proposed the IMPISTOR, a transistor with an IMPATT-like collector region, for which Yu et al. have suggested the name CATT. Carroll discussed three possible modes of operation for the avalanche transit-time transistor: (1) the multiplication mode, (2) the negative impedance mode, and (3) the pulse mode. The modes move progressively through the phenomena of avalanche multiplication, IMPATT negative conductance combined with multiplication, and voltage collapse and high current pulses associated with TRAPATT operation. The multiplication mode will be studied in this report. Quang, in 1975, presented a lumped-distributed small-signal equivalent circuit for an IMPISTOR in the negative impedance mode. In 1976, Lefebvre et al. utilized a computer program developed for high-efficiency IMPATT diodes to investigate the influence of a thermionic-type injected current on the dynamic operating conditions and performance of GaAs.
IMPATT diodes. The work showed that interesting results can be obtained in X-band and the practical realization of such a device would be possible by using a CATT operating in the negative impedance mode.

1.2 Basic Properties of CATT Devices

1.2.1 Collector Structures of CATT Devices. The various structures are described in terms of the avalanche region width $w_{av}$, the avalanche region doping density $N_{av}$, the drift region doping density $N_{drift}$, and the drift region width $w_D$. The LO-HI-LO doping profiles also include the total concentration of carriers in the charge clump per unit area $Q_c$. The doping and electric field profiles for various common CATT collector structures are shown in Fig. 1.1 along with the parameters that are used to describe their characteristics.

1.2.2 Principles of Operation of CATT Devices. The CATT device operates in a manner similar to that of a bipolar junction transistor. In an n-type CATT device, whose structure is shown in Fig. 1.2, electrons and holes are injected across the forward-biased emitter-base junction. The majority of the injected electrons, minority carriers in the base region, diffuse across the neutral base region and then the electric field, set up by the base-collector reverse bias, draws them into the collector region. A small percentage of the emitter-injected electrons are lost in the base region through carrier recombination. Electrons that are drawn into the collector region first undergo avalanche multiplication in the high-field portion of the collector region and then drift across the depleted low-field portion. While making a transit across the collector depletion region, a current is induced at the collector terminal. Unlike the bipolar junction transistors whose
FIG. 1.1  DOPING AND ELECTRIC FIELD PROFILES OF VARIOUS CATT COLLECTOR STRUCTURES.
FIG. 1.2 SCHEMATIC STRUCTURAL DIAGRAM OF A CATT DEVICE.
common-base current gain is always less than unity, carrier multiplication in the CATT device results in a current gain of the order of two to ten.

With the proper doping profile, a long collector structure allows the sustenance of a very large RF voltage across the collector depletion region. The CATT amplifiers develop additional power gain through avalanche multiplication and by the use of transit time in the collector. Consequently, the power gain can be much higher than for a bipolar junction transistor having an equivalent emitter-base structure, or for the same gain at a higher frequency.

Structurally, the n-type CATT device is similar to an npn bipolar junction transistor except for two major differences, as shown in Fig. 2. The collector has a high-field avalanche multiplication region in which the emitter injected electrons are multiplied and a long drift region. It should be noted that the long drift region is used to provide proper timing for the avalanche multiplication of the emitter injected electrons, besides providing the collector with a large RF voltage capability. If the load impedance is a high-Q tank circuit whose resonant frequency equals the emitter-base signal frequency, the induced current waveform of a common-base Class C CATT device is such that its fundamental component is automatically 90 rad out of phase with the near-sinusoidal collector RF voltage regardless of the collector transit angle $\theta_q$. The induced current waveform is always centered around the phase angle $3\pi/2$ rad of the RF voltage waveform. On the one hand, if the collector region is narrow, the induced current waveform will be narrow and the collector efficiency, which is defined as the ratio of signal power to dc input, is high. On the
other hand, the emitter injected electrons will enter the avalanche multiplication region at a time when the voltage across the collector region is low and not many electron-hole pairs will be generated. Limited by a narrow induced current waveform and small current multiplication, a CATT device with a narrow collector region will not be able to produce very much RF power. If the collector transit angle is greater than \( \pi \) rad, the emitter injected electrons will enter the avalanche multiplication region at a time when the voltage across the collector region is high and carrier multiplication will be large. A collector transit angle greater than \( \pi \) rad implies that there is conduction current flowing during the positive half-cycle of the collector RF voltage which means energy dissipation instead of power generation. The collector efficiency is poor for large collector transit angle situations. An optimum operating condition seems to be when the collector transit angle is approximately \( \pi \) rad.

It should be noted that the collector RF voltage is initiated by the entering of emitter injected electrons into the collector. This is because the bias voltage, as seen by the collector, drops when the collector current flows in the external load, whereas in the IMPATT diode, the current waveform is initiated by the device voltage which is the superposition of a large RF voltage over a dc bias which is only slightly below the device junction breakdown voltage. If the space-charge effect is ignored, theoretically, a pulse of charge is always injected into the low-field drift region at the time when the phase angle of the IMPATT RF voltage is close to \( \pi \) rad, regardless of the drift region transit angle. The collector voltage and induced current
waveforms of a CATT device and those of an IMPATT are very similar when
the drift region transit angle is approximately \( \pi \) rad and they are dif-
ferent when the drift region transit angle differs significantly from
\( \pi \) rad.

The upper limit on the collector voltage is approximately the
base-collector junction breakdown voltage \( V_B \). Actually, the collector
voltage can exceed \( V_B \) slightly for a short duration of time. The lower
limit on the collector voltage is the voltage needed to sustain electrons
at approximately the scattering limited velocity during its entire
transit across the depleted collector region. For high-power appli-
cations, the optimum base-collector dc bias is such that the collector
can have a large avalanche multiplication factor and a large RF voltage
simultaneously. If the dc bias is increased above the optimum value,
although the current gain would be increased, the amplitude of the
collector RF voltage will be decreased due to the upper limit set by
\( V_B \). If the dc bias is decreased below the optimum value, current gain
would definitely be decreased and possibly the amplitude of the collec-
tor RF voltage would also decrease due to the lower limit set by the
voltage required to maintain carriers at the scattering limited veloc-
ity. At optimum dc bias, the avalanche multiplication factor of a
typical CATT device ranges from two to ten rather than a million as
in the IMPATT diode. Therefore, when the emitter is not injecting
carriers into the collector, the collector current equals approximately
ten times the thermally generated reverse saturation current. A
significant conduction current exists in the collector only when the
emitter-base junction is forward biased. This is why it is called a
controlled-avalanche transit-time triode.
The CATT is a complicated device for several reasons. The seemingly simple avalanche multiplication process as employed, for example, in avalanche photodiodes becomes much more complex in CATT devices due to the large RF voltage swing. The CATT is complex for another reason. The avalanche multiplication generated holes will feed back into the base region and constitute a negative recombination base current component. This phenomenon results in a more uniform emitter current injection and better use of the emitter finger area than for bipolar junction transistors. If the feedback hole current is large enough the polarity of the base current is reversed. Pinch-in phenomena rather than pinch-out phenomena in the base region would occur if the carrier multiplication is large.

When the collector transit angle equals \( \pi \) rad, the signal carriers are injected at or near the time when the RF voltage equals \( V_{T_0} \), where \( V_{T_0} \) is the average value of the base-collector terminal voltage, if the space-charge effect is negligible; the avalanche multiplication process is, therefore, almost independent of the RF voltage amplitude. This is an essential condition for a linear amplifier. It should also be pointed out that nonlinearity in the CATT is due mainly to the exponential turn on of the emitter-base junction as in the BJT. To a certain extent this nonlinearity is alleviated in the CATT due to the fact that the space charge will cause the effective avalanche multiplication to decrease as the input signal level is increased. This leads to a wider dynamic range.

\textbf{1.2.3 Comparison of Collector Efficiencies in BJT, CATT and IMPATT Devices.} For the induced collector current and voltage waveforms
shown in Fig. 1.3, denoted by $I_T$ and $V_T$, respectively, the output power is given by

$$P_{out} = \frac{1}{2\pi} \int_0^{2\pi} I_T(\omega t)V_{RF} \sin \omega t \, dt$$  \hspace{1cm} (1.1)

or

$$P_{out} = \frac{I_{T_{\text{max}}}}{2\pi} \frac{V_{RF}}{2\pi} [\sin(\theta_{\text{inj}} + \theta_T) - \sin \theta_{\text{inj}}]$$  \hspace{1cm} (1.2)

where $V_{RF}$ is the amplitude of the collector RF voltage, $I_{T_{\text{max}}}$ is the maximum induced collector current level and $\theta_{\text{inj}}$ is the phase difference between the injected current pulse and the collector RF voltage. The dc collector current level can be found by averaging the induced current level as follows:

$$I_{T_{\text{dc}}} = \frac{1}{2\pi} \int_0^{2\pi} I_T(\omega t) \, dt$$

$$= \frac{\theta_T}{I_{T_{\text{max}}}} 2\pi$$  \hspace{1cm} (1.3)

By using Eqs. 1.2 and 1.3, the collector efficiency can be expressed as

$$\eta_c = \frac{P_{out}}{P_{dc}}$$

$$= \frac{V_{RF}}{V_T} \frac{\sin (\theta_{\text{inj}} + \theta_T) - \sin \theta_{\text{inj}}}{\theta_T}$$  \hspace{1cm} (1.4)
FIG. 1.3 VOLTAGE AND CURRENT WAVEFORMS IN THE CATT DEVICE.

(a) BASE-COLLECTOR TERMINAL VOLTAGE AND (b) INDUCED EXTERNAL CURRENT.
where $P_{dc}$ is the dc input power. The waveforms of $V_T$ and $I_T$ for BJT, CATT and IMPATT devices are shown in Fig. 1.4. A negative $P_{out}$ implies that RF power is being generated. Both the BJT and CATT devices are operating in Class C configuration and the load is a high-Q tank circuit whose resonant frequency is approximately equal to the signal frequency. The induced current waveform for the BJT is a short current pulse, because its collector region is narrow. A narrow collector region also implies that only a small RF voltage swing is allowed. From the induced current waveforms, it is expected that the CATT efficiency will be lower than that of the BJT due to the large transit angle requirement. The induced current waveforms of the CATT and IMPATT corresponding to various drift region transit angles are shown in Fig. 1.4. The maximum induced current level is assumed to be constant. For the CATT, $\theta_{inj}$ is always given by

$$\theta_{inj} = \pi - \frac{\theta_T}{2}$$

and Eqs. 1.2 and 1.4 become

$$P_{out} = -\frac{I_{max} V_{RF}}{\pi} \sin \frac{\theta_T}{2}$$

and

$$\eta_c = -\frac{V_{RF} \sin(\theta_T/2)}{V_T \theta_T}$$

For the IMPATT, $\theta_{inj}$ is given by
FIG. 1.4 COLLECTOR RF VOLTAGE AND INDUCED CURRENT WAVEFORMS
OF BJT, CATT AND IMPATT DEVICES.
\[ \theta_{\text{inj}} = \frac{\pi}{2} - \theta_{\text{sc}}, \quad (1.8) \]

where \( \theta_{\text{sc}} \) is the injection phase delay due to the space-charge effect.

If the space-charge effect is ignored, \( \theta_{\text{sc}} = 0 \) rad and Eqs. 1.2 and 1.4 become

\[ P_{\text{out}} \approx -\frac{I_{\text{Tmax}} V_{\text{RF}}}{2\pi} \left(1 - \cos \theta_{\text{T}}\right) \quad (1.9) \]

and

\[ \eta_{\text{c}} = -\frac{V_{\text{RF}}}{V_{\text{T}}} \frac{1 - \cos \theta_{\text{T}}}{\theta_{\text{T}}} \quad (1.10) \]

Shown in Figs. 1.5 and 1.6 are plots of \( P_{\text{out}} \) and \( \eta_{\text{c}} \) as functions of \( \theta_{\text{T}} \) at different values of \( \theta_{\text{inj}} \). The quantities of \( P_{\text{out}} \) and \( \eta_{\text{c}} \) are normalized in terms of \( I_{\text{Tmax}} V_{\text{RF}}/\pi \) and \( V_{\text{RF}}/V_{\text{T}} \), respectively. Results indicate that the CATT \( P_{\text{out}} \) and \( \eta_{\text{c}} \) are higher than the IMPATT due to the fact that the induced current waveform of the CATT is always centered at \( \theta = 270 \) degrees. It is also noticed that the CATT 3-dB \( P_{\text{out}} \) and \( \eta_{\text{c}} \) bandwidths are wider than the IMPATT. As \( \theta_{\text{inj}} \) decreases due to the space-charge effect, the IMPATT normalized \( P_{\text{out}} \) and \( \eta_{\text{c}} \) and their 3-dB bandwidths also decrease.

1.3 State of the Art of BJT and CATT Amplifiers

Electrical characteristics and performance of microwave power transistors depend first on the operational mode, i.e., amplifier, oscillator, and switching, and second on the operational class, generally Class C or A, in the amplifier and oscillator cases. On the other
FIG. 1.6 NORMALIZED COLLECTOR EFFICIENCY CURVES OF CATT AND IMPATT DEVICES AT DIFFERENT $\theta_{inj}$. 

$\theta_{inj} = 90^\circ$

CATT

$\theta_{inj} = 75^\circ$

$\theta_{inj} = 60^\circ$

$\theta_{inj} = 45^\circ$

$\% (\sqrt{RF/V_t})$

$\%_{\text{PERCENTAGE}}$
hand, power performances are influenced through the packaging condition. The transistor dynamic characteristics can be seriously degraded through the package inductances and capacitances. Making external input and output matching at microwave frequencies is extremely difficult. The first attempt to eliminate this drawback was to reduce these inductances and capacitances as much as possible. Two more sophisticated solutions were developed recently. The first consists of input and output matching within the transistor packaging, the so-called chip carriers; the second does away with the package and introduces transistors directly in a microwave integrated circuit called MIC. The first type is commercially available; the second is currently under commercial development.

The bipolar power transistors have been developed primarily for Class C applications, because in this class of operation both power and efficiency attain their highest values. However, in recent years the field of application of the BJT has been extended to Class A, with a view to reduce the nonlinearity and noise figures. The output power of the Class A BJT is also characterized by a large bandwidth and gain flatness.

Tables 1.1 and 1.2 give the electrical characteristics of BJTs for Class C amplifier and oscillator operation and the performance of Class C BJT microwave power amplifiers. Similarly, Tables 1.3 and 1.4 give the electrical characteristics of BJTs in Class A amplifier and oscillator modes and the performances of Class A BJT microwave power amplifiers. The results in Tables 1.1 through 1.4 represent the state of the art in 1977. More recently, a Si power BJT for use
Table 1.1

Electrical Characteristics of Microwave Power Bipolar Transistors (Commercially Available) for Class C Amplifier and Oscillator Operations (Teszner and Teszner\textsuperscript{h2})

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Frequency Range (GHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Collector-base breakdown voltage, $B_{VCBO}$ (V)</td>
<td>35-70</td>
</tr>
<tr>
<td>Collector-emitter breakdown voltage with shunted BE junction, $B_{VCER}$ (V)</td>
<td>35-60</td>
</tr>
<tr>
<td>Collector-emitter breakdown voltage (with opened BE junction), $B_{VCEO}$ (V)</td>
<td>15-40</td>
</tr>
<tr>
<td>Emitter-base breakdown voltage, $B_{VCEO}$ (V)</td>
<td>2-4</td>
</tr>
<tr>
<td>Collector-emitter continuous voltage, $V_{CE}$ (V)</td>
<td>12-28\textsuperscript{b}</td>
</tr>
<tr>
<td>Emitter metal finger average current density, $I_{Emfa}$ (A cm\textsuperscript{-2})</td>
<td>$-5 \times 10^4$-$5 \times 10^5$</td>
</tr>
</tbody>
</table>

a. The shunting must be of sufficiently low resistance to limit the minority carrier injection as much as possible.

b. For CW operation; in pulsed-wave operation, $V_{EC}$ is generally increased, up to 40 V.
Table 1.2

Electrical Performances of Microwave Power Bipolar Transistors, with or Without Internal Matching, for Class C Amplifier Operation (Teszner and Teszner)\(^4\)\(^2\)

<table>
<thead>
<tr>
<th>Frequency Range (GHz)</th>
<th>Amplifier transistor chip</th>
<th>Output power in CW operation, (P_{outmax}^{b}) (W)</th>
<th>Amplifier power added efficiency, (\eta^{c}) (percent)</th>
<th>Power gain PG (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>2</td>
<td>4.2</td>
<td>4.5</td>
<td>6</td>
</tr>
<tr>
<td>8</td>
<td></td>
<td>1.2</td>
<td>1.5</td>
<td>3</td>
</tr>
<tr>
<td>10</td>
<td></td>
<td>1</td>
<td>1</td>
<td>1</td>
</tr>
</tbody>
</table>

- **a.** The data given up to 4 GHz concern commercially available transistors, of both types, with or without internal matching.
- **b.** In pulsed-wave operation, an output power level of 150 W (duty cycle - 1 percent) at 1 GHz with \(PG = 10\) dB is obtained with commercially available transistors (\(V_{CE}\) being increased to 40 V).
- **c.** The \(\eta\) values indicated here correspond to the \(P_{outmax}\); however, some higher values have been obtained for lower \(P_{out}\); in particular, at 1 GHz, \(\eta\) goes up to 65 percent and at 2 GHz, up to 60 percent.
- **d.** Internally matched devices.
- **e.** Devices without internal matching.
- **f.** Experimental.
- **g.** Experimental.
- **h.** Experimental.
Table 1.3

Electrical Characteristics of Microwave Power Bipolar Transistors
(Commercially Available) for Class A Amplifier and Oscillator Operations (Teszner and Teszner\textsuperscript{h2})

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Frequency Range (GHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Collector-base breakdown voltage, BV_{CBO} (V)</td>
<td>$\geq 40$</td>
</tr>
<tr>
<td>Collector-emitter breakdown voltage, BV_{CEO} (V)</td>
<td>$\geq 15^a$</td>
</tr>
<tr>
<td>Emitter-base breakdown voltage, BV_{EBO} (V)</td>
<td>$\geq 3$</td>
</tr>
<tr>
<td>Collector-emitter continuous voltage, V_{CE} (V)</td>
<td>$\geq 15$</td>
</tr>
<tr>
<td>Collector-emitter continuous current, I_{CE} (mA)</td>
<td>$\geq 50$</td>
</tr>
</tbody>
</table>

\textsuperscript{a} For BV_{CE} the values approximate BV_{CBO}.
Table 1.4

Electrical Performances of Microwave Power Bipolar Transistors, Without Internal Matching, for Class A Amplifier Operation \(^b\) (Teszner and Teszner \(^4\))

<table>
<thead>
<tr>
<th>Frequency Range (GHz)</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>6</th>
</tr>
</thead>
<tbody>
<tr>
<td>Amplifier transistor chip output power in CW operation, (P_{\text{outmax}}) (W)</td>
<td>6</td>
<td>3</td>
<td>1.2</td>
<td>0.8</td>
<td>0.8</td>
</tr>
<tr>
<td>Amplifier power efficiency, (\eta) (percent)(^c)</td>
<td>-30</td>
<td>-25</td>
<td>-20</td>
<td>-15</td>
<td>-17</td>
</tr>
<tr>
<td>Power gain PG (dB)(^c)</td>
<td>10</td>
<td>10</td>
<td>8</td>
<td>6</td>
<td>4</td>
</tr>
</tbody>
</table>

a. A linear amplifier operating in Class B or AB with active broadband bias circuit has been developed in the laboratory. Compared with Class A amplifiers it would provide for identical I/C (intermodulation ratio) = -20 dB, an efficiency 2 times higher (15.5 vs. 8.5 percent) with PG = 10 dB, \(P_{\text{out}} = 0.4\) W at 4 GHz. This efficiency ratio seems to be increased for lower I/C; thus it becomes -5 for I/C = -30 dB, but with \(\eta\) decreasing to 12 and 2.2 percent, respectively, and proportionately \(P_{\text{out}}\). However, all the efficiencies quoted remain low compared with the efficiencies of commercially available devices given above.

b. The data given up to 4 GHz apply to commercially available transistors. The data given in the last column are for an experimental model, under laboratory development. In both cases the output power is obtained at 1 dB gain compression.

c. The \(\eta\) and PG values correspond to \(P_{\text{outmax}}\).
at X-band has been developed using e-beam and ion implantation. The BJT has a bar size of 0.5 x 1 mm² and consists of four 27.5 x 75 μm² active cells. The combined output power of four cells operating in Class C configuration is nearly 2 W at 8 GHz and almost 1.5 W at 10 GHz.

Experimental data on Class C CATT amplifiers is very limited. The only published results are those by Yu et al. They demonstrated the feasibility of a CATT amplifier both theoretically and experimentally from 0.5 to 4 GHz. An S-band CATT device with an emitter periphery to active base area ratio of 2.3 x 10³ in⁻¹ achieved a gain of 13 dB at a pulsed power output of 12 W and 28 percent power added efficiency when operated at 2 GHz. The gain at this operating point was 11 dB higher than its equivalent BJT transistor. Experimental results indicate that the impedance levels and Q values of the CATT are favorable for impedance matching, power combining and instantaneous bandwidth operation with useful gain.

1.4 Outline of the Present Study

The objective of this study is to investigate the theoretical capability of Class C CATT amplifiers and to compare them with BJT amplifiers. Analytical equations, circuit models and computer simulations are used to determine dc, small-signal, and large-signal behavior.

In Chapter II a computer program is developed which calculates the dc avalanche multiplication factor as a function of collector structure, base-collector dc bias and material parameters. Information concerning optimum collector structure for large-signal operation and suitability of various materials can be obtained.
In Chapter III analytical dc and small-signal models for CATT devices are given. Computer solutions of the analytical models are given and the results are discussed.

In Chapter IV a large-signal computer simulation is developed. The simulation calculates the RF output power and efficiency and many other parameters. Detailed descriptions of the numerical algorithms and the computer programs are given.

In Chapter V large-signal results are obtained for a series of X-band CATT and BJT devices. The computer results for both devices are given, discussed and compared.

In Chapter VI a summary of this study is given and suggestions for further work are described.
CHAPTER II. MATERIAL PARAMETERS AND THE DC AVALANCHE MULTIPLICATION FACTOR

2.1 Introduction

As mentioned previously, an ideal CATT device should have a dc bias point at which both a significant avalanche multiplication factor and a large RF voltage across the base-collector terminal can result. A simple computer program called AVALAN was developed which provides information on the dc avalanche multiplication factor \( M_{Ao} \) as a function of the dc base-collector terminal voltage \( V_{T0} \), the punch-through voltage \( V_{PT} \), the breakdown voltage \( V_B \), the sustaining voltage \( V_{sus} \) and the electric field \( E \) distribution in the collector depletion region. In this chapter the material parameters, the avalanche multiplication factor and the electric field distribution for devices of different materials and various structures are given and comparisons are made. The effects of device temperature and space charge are also described here.

2.2 Material Parameters

The material parameters required to calculate \( M_{Ao} \), \( V_{PT} \), \( V_B \), \( V_{sus} \) and \( E \) are \( \alpha_n \), \( \alpha_p \), \( \epsilon \), \( E_{sus} \), and \( E_{sus} \), where \( \alpha_n \) and \( \alpha_p \) are the electron and hole ionization rates, respectively, \( \epsilon \) is the dielectric permittivity, and \( E_{sus} \) and \( E_{sus} \) are the electric fields required to sustain electrons and holes at their saturation velocities. These values for Si and GaAs are listed in Tables 2.1 and 2.2. Shown in Fig. 2.1 is the dc avalanche multiplication factors of two n-type Si devices, where electrons initiate the avalanche process. Figure 2.2 shows the dc avalanche multiplication factors for two p-type Si devices, where holes initiate the
Table 2.1
Material Parameters for Si \( \varepsilon = 1.04477 \times 10^{-12} \) F/cm,

\[
E_{\text{sub}} = 2 \times 10^6 \text{ V/cm}, \quad E_{\text{sur}} = 6 \times 10^4 \text{ V/cm}, \quad a(E) = A \exp \left[ - \left( \frac{b}{E} \right) \right] \text{ cm}^{-1}
\]

<table>
<thead>
<tr>
<th>Holes</th>
<th></th>
<th>Electrons</th>
<th>Electric Field (kV/cm)</th>
<th>Reference</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>( A_p ) (cm(^{-1}))</td>
<td>( b_p ) (V/cm)</td>
<td>( A_n ) (cm(^{-1}))</td>
<td>( b_n ) (V/cm)</td>
</tr>
<tr>
<td>( T = 27^\circ C )</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>2.25 \times 10^7</td>
<td>3.26 \times 10^6</td>
<td>3.8 \times 10^6</td>
<td>1.75 \times 10^6</td>
<td>200-500</td>
</tr>
<tr>
<td>2 \times 10^6</td>
<td>1.97 \times 10^6</td>
<td></td>
<td></td>
<td>200-530</td>
</tr>
<tr>
<td>5.6 \times 10^5</td>
<td>1.32 \times 10^5</td>
<td></td>
<td></td>
<td>530-770</td>
</tr>
<tr>
<td>2.6 \times 10^6</td>
<td>1.43 \times 10^6</td>
<td>6.2 \times 10^5</td>
<td>1.08 \times 10^6</td>
<td>240-530</td>
</tr>
<tr>
<td>5 \times 10^5</td>
<td>9.9 \times 10^5</td>
<td></td>
<td></td>
<td>530-770</td>
</tr>
<tr>
<td>( T = 200^\circ C )</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>1.0 \times 10^7</td>
<td>3.2 \times 10^6</td>
<td>1.8 \times 10^6</td>
<td>1.64 \times 10^6</td>
<td>200-500</td>
</tr>
<tr>
<td>2 \times 10^6</td>
<td>2.166 \times 10^6</td>
<td></td>
<td></td>
<td>200-530</td>
</tr>
<tr>
<td>5.6 \times 10^5</td>
<td>1.516 \times 10^6</td>
<td></td>
<td></td>
<td>530-770</td>
</tr>
<tr>
<td>2.6 \times 10^6</td>
<td>1.661 \times 10^6</td>
<td>6.2 \times 10^5</td>
<td>1.311 \times 10^6</td>
<td>240-530</td>
</tr>
<tr>
<td>5 \times 10^5</td>
<td>1.221 \times 10^6</td>
<td></td>
<td></td>
<td>530-770</td>
</tr>
</tbody>
</table>

Ionization Rates for Electrons and Holes

\[
a = \frac{1}{\lambda} \exp \left[ \left( (11.5r^2 - 1.17r + 3.9 \times 10^{-4})x^2 + (46r^2 - 11.9r + 1.75 \times 10^{-2})x - 757r^2 + 75.5r - 1.92 \right) \right],
\]

where \( r = \frac{\varepsilon_r}{\varepsilon_i} \), \( x = \frac{\varepsilon_i}{\varepsilon_i e^E} \), \( \varepsilon_r = 0.063 \) eV,

\[
\frac{\varepsilon_r}{\varepsilon_i} = \tanh \left( \frac{\varepsilon_r}{2kT} \right) = \frac{\lambda}{\lambda_o},
\]

\( \varepsilon_i = 1.5 \times (1.16 - 7.02 \times 10^{-4} x r^2 / (T + 1108)) \),

\( \lambda_o = 76 \) \( \AA \) for electrons and \( \lambda_o = 47 \) \( \AA \) for holes.

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Table 2.2

Material Parameters for GaAs (ε = 1.10675 x 10^{-12} F/cm, E_{sun} = 3.3 x 10^3 V/cm,

\[ E_{sun} = 3 x 10^3 V/cm, \quad a(E) = A \exp \left(-\frac{b}{E}\right) \text{ cm}^{-1}. \]

<table>
<thead>
<tr>
<th>n</th>
<th>( A_p ) (cm(^{-1}))</th>
<th>( b_p ) (V/cm)</th>
<th>( A_n ) (cm(^{-1}))</th>
<th>( b_n ) (V/cm)</th>
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<td>2 \times 10^5 x[1+(T-300)x]</td>
<td>5.5 \times 10^5 x[1+(T-300)x]</td>
<td>2 \times 10^5 x[1+(T-300)x]</td>
<td>5.5 \times 10^5 x[1+(T-300)x]</td>
<td>Hall and Leck(^{46})</td>
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<tr>
<td></td>
<td>7 \times 10^{-4}</td>
<td>1.2 \times 10^{-3}</td>
<td>7 \times 10^{-4}</td>
<td>1.2 \times 10^{-3}</td>
<td></td>
</tr>
</tbody>
</table>

\( T = 27^\circ C \)

1  | 3.6 \times 10^8 | 2.9 \times 10^6 | 1.2 \times 10^7 | 2.3 \times 10^6 | Stillman et al.\(^{47}\) |

\( T = 200^\circ C \)

2  | 3.15 \times 10^5 | 8 \times 10^5 | 3.15 \times 10^5 | 8 \times 10^5 | Constant et al.\(^{48}\) |
FIG. 2.1 DC AVALANCHE MULTIPLICATION FACTORS CORRESPONDING TO VARIOUS IONIZATION RATES FOR ELECTRONS AND HOLES AT 27°C FOR TWO n-TYPE SI COLLECTOR STRUCTURES, WHERE ELECTRONS INITIATE THE AVALANCHE PROCESS. (UNIFORMLY DOPED COLLECTOR STRUCTURES)
FIG. 2.2 DC AVALANCHE MULTIPLICATION FACTORS CORRESPONDING TO VARIOUS IONIZATION RATES FOR ELECTRONS AND HOLES AT 27°C FOR TWO p-TYPE Si COLLECTOR STRUCTURES, WHERE HOLES INITIATE THE AVALANCHE PROCESS. (UNIFoRMLY DOPED COLLECTOR STRUCTURES)
avalanche process. The ionization rates of electrons and holes used are those reported by Grant,⁴³ Lee et al.,⁴⁴ and Crowell and Sze.⁴⁵ Grant's ionization rates give the lowest breakdown values. Figure 2.3 shows the dc avalanche multiplication factors of two n-type GaAs devices where electrons initiate the avalanche process. The ionization rates used are those obtained by Hall and Leck,⁴⁶ Stillman et al.,⁴⁷ and Constant et al.⁴⁸ The breakdown values obtained by using Hall and Leck's ionization rates are lower. The impact avalanche ionization rates are very important in determining the characteristics of CATT devices, but there is some dispute about their precise values and electric field dependence.

2.3 Dc Multiplication of Charge Carriers in n-Type and p-Type Si and n-Type GaAs CATT Devices

2.3.1 Derivation of the Analytical Expression for the Dc Avalanche Multiplication Factor of CATT Devices. An n-type CATT device is shown in Fig. 2.4. The electron current \( I_{nB} \) represents the electron carriers injected into the collector depletion region, which originated from the forward-biased emitter-base junction. While the electron particle current is flowing in the positive \( x \)-direction, the electronic current is positive in the negative \( x \)-direction. The dc time-independent continuity equations for electrons and holes in the collector depletion region are

\[
\frac{dJ_n}{dx} - (\alpha_n J_n + \alpha_p J_p) = 0 \tag{2.1}
\]

and

\[
\frac{dJ_p}{dx} + (\alpha_n J_n + \alpha_p J_p) = 0 \tag{2.2}
\]
Fig. 2.3 DC avalanche multiplication factors corresponding to various ionization rates for electrons and holes at 200°C for two n-type GaAs collector structures, where electrons initiate the avalanche process.

(Uniformly doped collector structures)
1. \(- I_{nE}\), Electron Particle Current Crossing the Forward-Biased Emitter-Base Junction

2. \(I_{pE}\), Hole Particle Current Crossing the Forward-Biased Emitter-Base Junction

3. \(- I_{nB}\), Component of \(- I_{nE}\) which has Reached the Base Edge of the Collector-Base Depletion Region

4. Component of \(- I_{nE}\) Lost in the Base Region Due to Carrier Recombination

5. \(- I_T\), Total Collector Particle Current

6. \(I_{pB}\), Feedback Hole Current Due to Avalanche Multiplication

7. \(- I_{cns}\), Electron Reverse Saturation Current

8. \(- I_{cps}\), Hole Reverse Saturation Current

**FIG. 2.4 SCHEMATIC OF THE ELECTRON AND HOLE PARTICLE CURRENT DISTRIBUTIONS IN A NORMALLY BIASED n-TYPE CATT DEVICE.**

(ELECTRON PARTICLE CURRENTS HAVE OPPOSITE SENSES OF DIRECTION TO THAT OF ELECTRONIC CURRENTS)
respectively. When Eqs. 2.1 and 2.2 are added, the result is

\[
\frac{d}{dx} (J_n + J_p) = 0 \tag{2.3}
\]

which indicates that the total particle current \( J_T \) is constant, independent of the space coordinate \( x \) where \( J_T = J_n + J_p \). The following differential equation is obtained from Eqs. 2.1 through 2.3:

\[
\frac{dJ_n}{dx} + (\alpha_p - \alpha_n) J_n = \alpha_p J_T
\]

whose solution is

\[
J_n(x) = \frac{J_T \int_0^x \alpha_p(x') \left[ \exp \int_0^{x'} [\alpha_p(x'') - \alpha_n(x'')] \, dx'' \right] \, dx' + c}{\exp \int_0^x [\alpha_p(x') - \alpha_n(x')] \, dx'} \tag{2.4}
\]

where the constant \( c \) is determined by imposing the proper boundary conditions. The boundary conditions are

\[
J_n(0) = J_{c_{ns}} + J_{nB}
\]

and

\[
J_p(\infty) = J_{c_{ps}}
\]

where \( J_{c_{ns}} \) and \( J_{c_{ps}} \) are the collector reverse saturation currents. By using Eqs. 2.3 and 2.4 and the boundary conditions, the following expression can be obtained:
\[ J_T = \frac{J_{c_{ps}} \exp \int_0^{v_T} (a_p - a_n) \, dx + J_{nB} + J_{c_{ns}}}{\exp \int_0^{v_T} (a_p - a_n) \, dx - \int_0^{v_T} a_p \left( \exp \int_0^{x} [a_p(x') - a_n(x')] \, dx' \right) \, dx} \]

which can be further reduced to

\[ J_T = \frac{J_{c_{ps}} \exp \int_0^{v_T} (a_p - a_n) \, dx + J_{nB} + J_{c_{ns}}}{1 - \int_0^{v_T} a_n \left( \exp \int_0^{x} [a_p(x') - a_n(x')] \, dx' \right) \, dx} \tag{2.5} \]

by using the relation

\[ \frac{d}{dx} \exp \int_0^{x} f(x') \, dx' = \left[ \exp \int_0^{x} f(x') \, dx' \right] f(x) \]

When the emitter-base junction is forward biased, the relation

\[ J_{nB} \gg J_{c_{ps}} \exp \int_0^{v_T} (a_p - a_n) \, dx \]

is true for any base-collector reverse bias voltage up to a value significantly above the breakdown value. Therefore, Eq. 2.5 can be reduced to

\[ J_T = M_A \left( J_{nB} + J_{c_{ns}} \right) \tag{2.6} \]

where the dc avalanche multiplication factor is given by
The expression in Eq. 2.7 is used to calculate the carrier multiplication and the breakdown voltage when the avalanche process is initiated by electrons as in n-type CATT devices.

For p-type CATT devices, where the avalanche process is initiated by holes, expressions for $J_T$ and $M_{A_O}$ can be similarly derived and they are

$$J_T = M_{A_O} (J_{pB} + J_{c_p})$$

and

$$M_{A_O} = \frac{1}{1 - \int_0^{W_T} \alpha_p \left[ \exp \int_0^x [\alpha_n(x') - \alpha_p(x')] dx' \right] dx}$$

2.3.2 Dc Avalanche Multiplication Factor vs. $V_T$ Characteristics

in p-Type and p-Type Si and n-Type GaAs CATT Devices. A distinction can be made on the dc avalanche multiplication factor vs. the base-collector terminal voltage characteristic which depends on the amount of feedback in the avalanche process. This distinction is made clear by investigating three cases of n-type Si CATT devices. The first one is for the case where the ionization coefficients $\alpha_n$ and $\alpha_p$ are assigned realistic values. The second one is where the hole ionization
coefficient $a_p$ is assigned to be equal to the electron ionization coefficient $a_n$, and the third one is where the hole ionization coefficient $a_p$ is set to zero. In p-type Si CATT devices, the second and third cases are when $a_n$ is assigned to be equal to $a_p$ and when $a_n$ is set to zero, respectively. For n-type devices, where in reality $a_n > a_p$, the second case represents a situation where the positive feedback in the avalanche process is artificially increased, and the dc avalanche multiplication rises more rapidly with increasing bias voltage. In this case, to achieve significant carrier multiplication, the bias voltage must be set very close to the breakdown value which, on the other hand, would severely limit the allowed RF voltage amplitude $V_{RF}$. For p-type Si devices, the second case represents a situation where the positive feedback in the avalanche process is artificially reduced since $a_p < a_n$ in reality and $M_{A_n}$ rises slower with increasing bias voltage. The third case represents a situation, for both n-type and p-type devices, where there is no positive feedback in the avalanche process, and multiplication occurs during a single transit of the high field region. The multiplication factor $M_{A_0}$ rises slowest in the third case in which, theoretically, the breakdown voltage is infinite. The various cases described previously are depicted in Figs. 2.5 and 2.6.

When the realistic ionization coefficients formulated by Crowell and Sze are used for Si and those by Hall and Leck are used for GaAs, $M_{A_0}$ vs. $V_T$ characteristics of several CATT devices whose collector regions are uniformly doped at various impurity levels are displayed in Figs. 2.7 through 2.9. The width of the collector regions
FIG. 2.5 CALCULATED $N_A^o$ VS. $V_T$ IN AN n-TYPE Si CATT DEVICE. ($w_T = 5 \times 10^{-4}$ cm, $N_c = 5 \times 10^{15}$ cm$^{-3}$ AND $T = 27^\circ$C)
FIG. 2.6 CALCULATED $N_A/G$ VS. $V_T$ IN A P-TYPE SI CATHODE DEVICE. ($N_A = 5 \times 10^{15}$ cm$^{-3}$ AND $T = 27^\circ$C)
FIG. 2.7 $M_A$ vs. $V_T$ characteristics of n-type Si CATT devices whose collector regions are uniformly doped. ($V_T = 4 \times 10^{-4}$ cm and $T = 27^\circ$C)
Fig. 2.8 $V_{C}$ vs. $V_{T}$ characteristics of p-type Si CNT devices whose collector regions are uniformly doped. ($V_{c} = 4 \times 10^{-4}$ cm and $T = 27^\circ$C).
Fig. 2.9 $M_A$ vs. $V_T$ Characteristics of n-Type GaAs CATT Devices
Whose Collector Regions Are Uniformly Doped. ($V_T =
4 \times 10^{-4} \text{ cm and } T = 27^\circ \text{C}$)
in all cases is \( b \times 10^{-4} \) cm and the CATT devices in Figs. 2.7 through 2.7 are at 27°C. The breakdown voltage of each device can be approximately defined to be the \( V_T \) value at which \( \beta \) reaches 100. The optimum load for a Class C CATT amplifier is a high-Q tank circuit whose resonant frequency is tuned at the signal frequency. Injection of a sharp pulse of charge which traverses across the depleted collector region would result in an induced current waveform and an RF voltage \( V_T \) waveform as shown in Fig. 2.10. It is observed that at \( \theta = 270 \) degrees, \( V_T \) is minimum and the sharp pulse of charge is located spatially near the midpoint of the collector region. An estimation of the allowed minimum \( V_T \) of a uniformly doped collector structure is \( V_{\text{sus}} \) by definition and its value* is

\[
V_{\text{sus}} = \frac{1}{2} w v_{\text{sus}} E(0),
\]

where

\[
w v_{\text{sus}} = E(0)/(e\beta c)
\]

and

\[
E(0) = \frac{1}{2} w_{\text{su}} \frac{e\beta}{c} + E_{\text{su}}.
\]

The electric field profile at \( \theta = 270 \) degrees and when \( V_T \) equals the minimum allowed value \( V_{\text{sus}} \) is shown in Fig. 2.10. The value of \( V_{\text{sus}} \)

* For HI-LO collector structures, the appropriate expressions for calculating \( V_{\text{sus}} \) are given in Section 2.3.3.
FIG. 2.10 INJECTED CURRENT, INDUCED CURRENT, TERMINAL VOLTAGE AND POSITION OF INJECTED CHARGE PULSE VS. PHASE, AND ELECTRIC FIELD PROFILE AT PHASE = 3π/2 AND WHEN $V_T = V_{sus}$ IN A UNIFORMLY DOPED COLLECTOR REGION.
for each device is also indicated in Figs. 2.7 through 2.9. Voltages $V_B$ and $V_{sus}$ roughly represent the upper and lower limits on $V_T$. It is clear from the $M_{AO}$ vs. $V_T$ characteristics that n-type CATT devices are more suitable for making high gain Class C amplifiers. For p-type Si devices, although their breakdown values are slightly higher than those of similarly structured n-type Si devices, their dc bias must be set much closer to the breakdown value in order to achieve significant carrier multiplication which severely reduces the allowed amplitude of $V_T$. Another disadvantage of p-type devices is their higher $V_{sus}$ which again would limit the RF voltage amplitude. For n-type GaAs devices, although their $V_{sus}$ are slightly lower than those of similarly structured n-type Si devices, their $M_{AO}$ vs. $V_T$ characteristics are such that the dc bias must be set closer to the breakdown values.

Because of its favorable $M_{AO}$ vs. $V_T$ characteristic and the advanced Si technology, the investigation of Class C CATT amplifiers was concentrated mainly on n-type Si devices. It should be noted that thus far space-charge effects were ignored in determining the $M_{AO}$ vs. $V_T$ characteristic, $V_B$, $V_{sus}$ and electric field profile. These effects will be examined briefly in Section 2.3.3. It should also be mentioned that the $V_B$ determined from the $M_{AO}$ vs. $V_T$ characteristic is higher than the actual $V_B$ because the effect of the base-collector junction curvature on $V_B$ is ignored.

2.3.3 Effects of Collector Structure on $M_{AO}$ vs. $V_T$ Characteristic of n-Type Si CATT Devices. In this section, the effects of changing the doping density $N_{av}$ in the high doping region of HI-LO
collector structures, the effect of changing the doping density $N_{\text{drift}}$ in the drift region, and the effects of changing the length $w_D$ of the drift region on the dc $M_{A_0}$ vs. $V_T$ characteristic, $V_B$ and $V_{\text{sub}}$ are studied. In actual large-signal operation, $V_T$ can be significantly higher than $V_B$ and slightly lower than $V_{\text{sub}}$, but the voltages $V_B$ and $V_{\text{sub}}$ can serve as a guide for the upper and lower limits of $V_T$. The large-signal power gain, to the first-order approximation, is proportional to the product of $M_{A_0}$ and $V_{\text{RF}}$, where $V_{\text{RF}}$ is the amplitude of the RF base-collector terminal voltage, and its value is dependent on both the $M_{A_0}$ vs. $V_T$ characteristic and the base-collector dc bias. The dc bias is chosen to maximize the product of $M_{A_0}$ and $V_{\text{RF}}$ but it must not be lower than $V_{\text{PT}}$ in order to avoid large collector resistance due to the undepleted high-resistivity collector region.

$V_{\text{sub}}$, the minimum $V_T$ allowed, for HI-LO collector structures is given by

$$ V_{\text{sub}} = \frac{1}{2} \left[ E(0) + E(w_{av}) \right] w_{av} + \frac{1}{2} E(w_{av}) w_{\text{sub}} , $$

(2.11)

where

$$ E(w_{av}) = \frac{eN_{\text{drift}}}{\varepsilon} \left( \frac{w_T}{2} - w_{av} \right) + E_{\text{sub}} , $$

$$ E(0) = \frac{eN_{av}}{\varepsilon} w_{av} + E(w_{av}) $$

and

$$ w_{\text{sub}} = E(w_{av}) \frac{\varepsilon}{(eN_{\text{drift}})} . $$
The doping profile and the electric field profile at a phase angle $= \frac{3\pi}{2}$ and $V_T = V_{sus}$ are shown in Fig. 2.11.

The space-charge effect and the effect of the base-collector junction curvature are ignored. All devices are operated at $270^\circ$C.

2.3.3a Effects of Different Doping Densities $N_{av}$. It can be seen from Fig. 2.12 and the data in Table 2.3 that $V_{RF}$ varies with $V_{bias}$, the base-collector dc bias. At optimum $V_{bias}$, devices with higher $N_{av}$ have lower $V_{RF}$, but $M_{AO}$ is higher for devices with higher $N_{av}$. The device with $N_{av} = 2 \times 10^{16}$ cm$^{-3}$ appears to be most suitable for making a high RF power gain amplifier since it has the highest $M_{AO,RF}$ product. Moreover, the optimum dc bias is lower for devices with higher $N_{av}$ which means lower dc power dissipation in the collector region at equal dc collector current densities. When $N_{av}$ equals $3 \times 10^{16}$ cm$^{-3}$ or higher, $V_{PT}$ is even higher than $V_B$.

When the space-charge effect is incorporated into the analysis, as is done in Section 2.3.4, $M_{AO}$ decreases significantly with increasing dc collector current density while $V_B$ and $V_{RF}$ increase. Therefore, the optimum $V_{bias}$ and $N_{av}$ will be different from the values extracted from Fig. 2.12.

2.3.3b Effects of Different Doping Densities $N_{drift}$. From Fig. 2.13 and Table 2.4, it is clear that devices with higher $N_{drift}$ have higher $V_{PT}$ and $V_{sus}$ and lower $V_B$. Therefore, higher $N_{drift}$ implies lower $V_{RF}$. From the data in Table 2.4, the optimum $V_{bias}$ for a device with $N_{drift} = 5 \times 10^{14}$ cm$^{-3}$ appears to be only slightly below $V_B$. At $V_{bias} = 65$ V, $V_{RF}$ is only 2 V but $M_{AO}$ is approximately 60 and the $M_{AO,RF}$ product is higher than that corresponding to lower $V_{bias}$.
Fig. 2.11 Doping profile of a Hi-Lo collector region and the electric field profile at a phase angle $= 3\pi/2$ and when $V_T = V_{sus}$. 
Fig. 2.12 Dependence of $\mu_A$ vs. $v_T$ characteristic on $N_{av}$ (n-type, Si, $v_{av} = 1 \times 10^{-4}$ cm, $N_{drift} = 2 \times 10^{15}$ cm$^{-3}$, $v_D = 3 \times 10^{-4}$ cm and $T = 23\degree C$).
Table 2.3

$V_{RF}$ and $M_{A_0}$ at Different Dc Biases (Extracted from Fig. 2.12)

Devices: n-Type, Si, $w_{av} = 1 \times 10^{-4}$ cm, $N_{\text{drift}} = 2 \times 10^{15}$ cm$^{-3}$,

$w_D = 3 \times 10^{-4}$ cm, $T = 27^\circ$C and Different $N_{av}$

$N_{av} = 7.5 \times 10^{15}$ cm$^{-3}$

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<tr>
<th>$V_{bias}$ (V)</th>
<th>$V_{RF}$ (V)</th>
<th>$M_{A_0}$</th>
<th>$M_{A_0} \times V_{RF}$</th>
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<td>38.5</td>
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<td>1.15</td>
<td>46</td>
</tr>
<tr>
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<td>36</td>
<td>1.23</td>
<td>44.3</td>
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<td>6.2</td>
<td>37.2</td>
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$N_{av} = 1 \times 10^{16}$ cm$^{-3}$

<table>
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<tr>
<th>$V_{bias}$ (V)</th>
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<th>$M_{A_0}$</th>
<th>$M_{A_0} \times V_{RF}$</th>
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<td>85</td>
<td>5</td>
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(Cont.)
Table 2.3 (Cont.)

\( N_{av} = 2 \times 10^{16} \text{ cm}^{-3} \)

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<td>10</td>
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</tr>
<tr>
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<td>5</td>
<td>12</td>
<td>60</td>
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<tr>
<td>52.5</td>
<td>2.5</td>
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FIG. 2.13 DEPENDENCE OF $N_A^0$ VS. $V_T$ CHARACTERISTIC ON $N_{\text{drift}}$ (n-TYPE, Si, $N_{av} = 2 \times 10^{16}$ cm$^{-3}$, $w_{av} = 1 \times 10^{-6}$ cm, $v_D = 3 \times 10^{-4}$ cm AND $T = 23^\circ$C)
Table 2.4

Data \( V_{RF} \) and \( M_{A_0} \) at Different Dc Biases (Extracted from Fig. 2.13)

Devices: n-Type, Si, \( N_{av} = 2 \times 10^{16} \text{ cm}^{-3} \), \( w_{av} = 1 \times 10^{-4} \text{ cm} \),

\( w_D = 3 \times 10^{-4} \text{ cm} \), \( T = 27^\circ C \) and Different \( N_{\text{drift}} \)

\( N_{\text{drift}} = 5 \times 10^{14} \text{ cm}^{-3} \)

<table>
<thead>
<tr>
<th>( V_{bias} ) (V)</th>
<th>( V_{RF} ) (V)</th>
<th>( M_{A_0} )</th>
<th>( M_{A_0} \times V_{RF} )</th>
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<td>1.58</td>
<td>11.2</td>
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<tr>
<td>65</td>
<td>2</td>
<td>60</td>
<td>120</td>
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\( N_{\text{drift}} = 2 \times 10^{15} \text{ cm}^{-3} \)

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<th>( V_{RF} ) (V)</th>
<th>( M_{A_0} )</th>
<th>( M_{A_0} \times V_{RF} )</th>
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</tbody>
</table>
In real situations, the space-charge effect drastically reduces $M_A$ at high $V_{bias}$ values and the optimum $V_{bias}$ is significantly below $V_B$. From the data in Table 2.4, it appears that the lower $N_{drift}$ the better. This is true in real situations only if the space-charge density does not exceed the collector impurity level significantly. Otherwise, the spatial direction of increasing electric field reverses.

2.3.3c Effects of Different Drift Region Widths. The effects of increasing the drift region width can be seen from the $M_{AO}$ vs. $V_T$ characteristic in Fig. 2.14 and the data in Table 2.5. Longer collectors have higher breakdown values, but $V_{PT}$ and $V_{sus}$ are also higher. A device with $w_D = 7.5 \times 10^{-4}$ cm has the largest $V_{RF}$. Although the device with $w_D = 1 \times 10^{-3}$ cm has the smallest $V_{RF}$ because of its much larger $M_{AO}$, it has the highest $M_{AO} - V_{RF}$ product. The aforementioned space-charge effect will change this result. The drift region width which corresponds to the highest $M_{AO} - V_{RF}$ product in real situations is shorter than $1 \times 10^{-3}$ cm. In actual large-signal operation, the carrier multiplication factor is not only dependent on $V_{bias}$ but is also dependent on $V_{RF}$ and the injection angle $\theta_{inj}$, which is dependent on $\omega$ and $w_D$. The amplifier efficiency, maximum RF power output and RF power gain cannot be understood in terms of $V_{bias}$, $M_A$ and $V_{RF}$ alone. The large-signal operation of Class C CATT amplifiers is discussed in detail in Chapters IV and V.

2.3.4 Temperature and Space-Charge Effects in p-Type Si CATT Devices. The effect of device temperature can be seen from Fig. 2.15. The ionization rates of electrons and holes decrease with increasing temperature and therefore the breakdown voltage increases with increasing temperature.
FIG. 2.1b DEPENDENCE OF $M_{A0}$ VS. $V_T$ CHARACTERISTIC ON $w_D$ (n-TYPE, Si, UNIFORMLY DOPED COLLECTOR REGION, $N_c = 2 \times 10^{15} \text{ cm}^{-3} \text{ AND } T = 23^\circ \text{C}$)
Table 2.5

$V_{RF}$ and $M_A$ at Different Dc Biases (Extracted from Fig. 2.14)

Devices: n-Type, Si, Uniformly Doped Collector, $N_c = 2 \times 10^{15}$ cm$^{-3}$,

$T = 27^\circ C$ and Different $w_D$

$w_D = 2 \times 10^{-4}$ cm

<table>
<thead>
<tr>
<th>$V_{bias}$ (V)</th>
<th>$V_{RF}$ (V)</th>
<th>$M_{A_o}$</th>
<th>$M_{A_o} \times V_{RF}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>8</td>
<td>3.8</td>
<td>1.0</td>
<td>3.8</td>
</tr>
<tr>
<td>25</td>
<td>20.8</td>
<td>1.0</td>
<td>20.8</td>
</tr>
<tr>
<td>30</td>
<td>25.8</td>
<td>1.0</td>
<td>25.8</td>
</tr>
<tr>
<td>35</td>
<td>25</td>
<td>1.01</td>
<td>25.3</td>
</tr>
<tr>
<td>40</td>
<td>20</td>
<td>1.1</td>
<td>22</td>
</tr>
<tr>
<td>50</td>
<td>10</td>
<td>2</td>
<td>20</td>
</tr>
<tr>
<td>55</td>
<td>5</td>
<td>4</td>
<td>20</td>
</tr>
</tbody>
</table>

$w_D = 3 \times 10^{-4}$ cm

<table>
<thead>
<tr>
<th>$V_{bias}$ (V)</th>
<th>$V_{RF}$ (V)</th>
<th>$M_{A_o}$</th>
<th>$M_{A_o} \times V_{RF}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>16.5</td>
<td>9.5</td>
<td>1.0</td>
<td>9.5</td>
</tr>
<tr>
<td>40</td>
<td>33</td>
<td>1.0</td>
<td>33</td>
</tr>
<tr>
<td>45</td>
<td>38</td>
<td>1.0</td>
<td>38</td>
</tr>
<tr>
<td>50</td>
<td>34</td>
<td>1.05</td>
<td>35.7</td>
</tr>
<tr>
<td>60</td>
<td>24</td>
<td>1.25</td>
<td>30</td>
</tr>
<tr>
<td>70</td>
<td>14</td>
<td>2.1</td>
<td>29.4</td>
</tr>
<tr>
<td>80</td>
<td>4</td>
<td>7.8</td>
<td>31.2</td>
</tr>
</tbody>
</table>

(Cont.)
Table 2.5 (Cont.)

$v_d = 5 \times 10^{-4}$ cm

| 42 | 27 | 1.0 | 27  |
| 60 | 45 | 1.0 | 45  |
| 65 | 50 | 1.0 | 50  |
| 70 | 52 | 1.0 | 52  |
| 80 | 42 | 1.12| 47  |
| 95 | 27 | 1.5 | 40.5|
| 110| 12 | 3.2 | 38.4|
| 120| 2  | 15  | 30  |

$v_d = 7.5 \times 10^{-4}$ cm

| 90 | 60 | 1.12| 67.2 |
| 95 | 61 | 1.17| 71.4 |
| 100| 56 | 1.23| 68.9 |
| 120| 36 | 1.8 | 64.8 |
| 140| 16 | 4.25| 68  |
| 150| 6  | 10  | 60  |

$v_d = 1 \times 10^{-3}$ cm

| 150| 20 | 5   | 100 |
| 155| 15 | 6.1 | 91.5|
| 160| 10 | 10  | 100 |
| 165| 5  | 16  | 80  |
Fig. 2.15 Dependence of $M_0$ vs. $V_n$ characteristic on temperature. (N-type, SI, uniformly doped collector and $V_n = 5 \times 10^{-4}$ cm.)
The effect of space charge\(^7,51\) on \(M_{A_0}\) can be seen from Fig. 2.16. The upper limit on dc collector current density is dependent on \(N_{av}\) and \(N_{drift}\). The electric field profile in a typical HI-LO collector is depicted in Fig. 2.17. As \(J_{\text{dc}}\) increases while \(V_T\) is held constant, \(E(w_{av})\) decreases. If the space-charge density is higher than \(N_{drift}\), \(E(w_T)\) is higher than \(E(w_{av})\). If \(N_{av}\) is very high, \(E(w_{av})\) at \(J_{\text{dc}} = 0\) is low and it does not take a space-charge density much higher than \(N_{drift}\) to force \(E(w_{av})\) below \(E_{\text{sus}}\). If \(N_{drift}\) is very low, the electric field in the region \(0 \leq x \leq w_{av}\) decreases rapidly with increasing \(J_{\text{dc}}\) while \(V_T\) is held constant, therefore, \(M_{A_0}\) decreases rapidly with increasing \(J_{\text{dc}}\). Moreover, the sign of \(dE/dx\) is reversed at lower \(J_{\text{dc}}\) which means that \(E(w_T)\) is higher than \(E(w_{av})\). The condition that no significant carrier multiplication occurs near \(x = w_T\) is violated at lower \(J_{\text{dc}}\) for devices with lower \(N_{drift}\). A situation could arise where the collector region near \(x = 0\) becomes undepleted before \(E(w_T)\) is sufficiently high to cause significant carrier generation. This collector current induced neutral region will effectively increase the effective neutral base region width and the base transport factor will decrease. This case is shown in Fig. 2.17 and corresponds to \(J_{\text{dc}} = J_x\). The aforementioned space-charge effects have been demonstrated by applying AVALAN on various collector structures at different dc collector current densities.

2.4 Summary

This chapter contained the ionization rates of electrons and holes in Si and GaAs and comparisons of the dc avalanche multiplication factor vs. \(V_T\) characteristics of n-type and p-type Si and n-type GaAs CATT devices. n-type Si devices are found to have the most suitable
Fig. 2.16 Effects of space charge. (n-type, $S_1$, $W_V = 1 \times 10^{-4}$ cm, $W_{dc} = 2 \times 10^{-5}$ cm, $W_D = 3 \times 10^{-7}$ cm and $T = 23^\circ$C)
Fig. 2.17 Dependence of the electric field profile on space-charge density, $N_{av}$ and $N_{drift}$

$0 < J_1 < J_2 < J_3$
$M_A$ vs. $V_T$ characteristics for making high-gain CATT amplifiers. The effects of collector structural parameters, i.e., $N_{av}$, $N_{drift}$ and $w_D$ on $M_A$ vs. $V_T$ characteristics, were investigated and the effects of temperature and space charge were discussed.
CHAPTER III. DEVICE PHYSICS, DC AND SMALL-SIGNAL ANALYTICAL MODELS

3.1 Introduction

This chapter presents a discussion of the physics of CATT devices, a dc analytic model and a small-signal one. For the dc device model, the analytical expression for the avalanche multiplication factor, which was derived in Chapter II, is employed in describing the current multiplication phenomenon. The effects of high-level injection in the base region are ignored, but are included in the large-signal simulation model presented in Chapter IV. The dc space-charge effects in the collector region are included. The space-charge effects are different under RF operating conditions and they are considered in detail in Chapter IV. The electric field in the collector depletion region is assumed to be sufficient to sustain the carrier saturation velocity everywhere. The small-signal analytical model is derived from a first-order Taylor series expansion of dc quantities and equations about their quiescent values and the linearization of the resulting equations. This procedure, in general, produces coupled nonlinear ordinary differential equations in phasor space for the small-signal variables of interest: particle concentrations, particle currents, terminal voltages, etc. The small-signal analytical model does not only include the current multiplication phenomenon, but also the transit-time effect associated with the base-collector depletion region. The common-base $y$-parameters are derived and the unilateral power gain, maximum frequency of oscillation, operating power gain, transducer power gain, optimum source and load terminations,
Linvill stability factor, and bandwidth are calculated. A brief discussion of the possibility of operating CATT devices as IMPATT diodes with variable equivalent thermally generated current is given.

3.2 DC Analytical Model

3.2.1 Carrier Concentration in BJT and CATT Devices. The standard solution for carrier concentrations and currents in the field-free base region of a uniformly doped BJT or CATT structure under low-level injection was first presented by Shockley in the classic paper that introduced the junction transistor in 1949. Shockley's transistor theory still serves as the foundation for all present-day theories of bipolar junction transistor operation including high-level injection, field-aided base transport, the Webster effect and Kirk effects. A one-dimensional uniformly doped npn structure is shown in Fig. 3.1.

The minority carrier boundary conditions at the base edge of the emitter-base depletion region and at the base edge of the base-collector depletion region first introduced by Shockley are

\[ n(w) = n_0 \exp \left( \frac{eV_{EB}}{kT} \right) \]  

and

\[ n(w) = n_0 \exp \left( - \frac{eV_{EB}}{kT} \right) \]  

respectively, under normal bias conditions, where \( V_{EB} \) is the emitter-base terminal voltage and \( n_0 \) is the thermal equilibrium concentration. Application of these expressions for the minority carrier boundary conditions to BJT and CATT devices with the emitter-base junction forward biased such that \( n(w) \gg n_0 \), the collector-base junction reverse biased such that \( n(w) \ll n_0 \), and with a base width greater than...
FIG. 3.1 ONE-DIMENSIONAL npn TRANSISTOR STRUCTURE.
several Debye lengths\textsuperscript{52} yields a satisfactory description of carrier transport in the base region under low-level injection. The modification required in Eq. 3.1 under high-level injection is discussed in Chapter IV. If the boundary condition required for the analysis of the collector region is the minority carrier current density, Eqs. 3.1 and 3.2 are adequate. If the collector boundary condition required is the minority carrier density, Eq. 3.2 is not appropriate. The requirement that the minority carrier concentration at the base edge of the base-collector depletion region be less than the thermal equilibrium concentration, independent of the minority carrier current density, is disconcerting at best. An expression for the minority carrier concentration at the base edge of the base-collector depletion region, which would result in a smooth transition from diffusion transport to drift transport that the electrons must undergo in traversing the field-free base region to the high-field base-collector region, is given in Chapter IV. For the dc and small-signal analytical models, the collector boundary condition required is the minority carrier current density. The case where the base width becomes comparable to a Debye length has been investigated by Mc Cleer,\textsuperscript{52} but it is of no concern in the modeling of CATT devices.

If a constant base-doping concentration and a low-level emitter injection are assumed, transport of minority carriers in the field-free base region can be described by diffusion alone and the dc electron current density is

\begin{equation}
J_{n_{dc}} = eD_n \frac{dn_{dc}}{dy},
\end{equation}
where \( n_{dc} \) is the dc electron concentration and \( D^*_{n} \) is the low-field electron diffusion constant. Another relationship which governs the base region electron distribution is the time-independent electron continuity equation

\[
0 = \frac{n_{dc} - n_{o}}{\tau_n} - \frac{1}{e} \frac{d}{dy} n_{dc},
\]

(3.4)

where \( \tau_n \) is the electron lifetime. By substituting Eq. 3.3 into Eq. 3.4, a second-order, constant coefficient, homogeneous ordinary differential equation is obtained whose solution is of the form:

\[
n_{dc}(y) - n_{o} = A e^{-(y-w_2)/L_n} + B e^{(y-w_2)/L_n},
\]

where \( L_n = \sqrt{D^*_{n} \tau_n} \) is the electron diffusion length in the base region. The constants \( A \) and \( B \) can be determined from the boundary conditions given by Eqs. 3.1 and 3.2. Once \( n_{dc}(y) \) is found, the electron carrier current is determined from Eq. 3.3. At \( y = w_2 \) and \( w_3 \), the dc electron currents are

\[
\hat{i}_{n_{dc}} = \int_{w_2}^{w_3} \left[ n_{dc}(y) - n_{o} \right] \coth \left( \frac{w_B}{L_n} \right) dy - \left[ n_{dc}(w_3) - n_{o} \right] \sinh \left( \frac{w_B}{L_n} \right),
\]

(3.5)

and

-68-
\[ I_{n_{dc}}^{B} = I_{n_{dc}}^{3} = \frac{e D_{A_{n}}}{L_{n}} \left[ n_{dc}(w_{2}) - n_{o} \right] \text{csch} \left( \frac{\varphi_{B}}{L_{n}} \right) \]

\[ - \left[ n_{dc}(w_{3}) - n_{o} \right] \text{coth} \left( \frac{\varphi_{B}}{L_{n}} \right), \quad (3.6) \]

where \( A_{E} \) is the emitter area, \( w_{B} = w_{3} - w_{2} \),

\[ w_{2} = \left( \frac{2e}{e_{n}^{n_{A}}} \right) \left( \frac{N_{D}}{N_{A} + N_{D}} \right)^{1/2} \frac{1}{V_{EB} - V_{EO}}, \quad (3.7) \]

\[ w_{3} = w_{B} - \left( \frac{2e}{e_{n}^{n_{A}}} \right) \left( \frac{N_{D}}{N_{A} + N_{D}} \right)^{1/2} \frac{1}{V_{BL} - V_{EO}}, \quad (3.8) \]

\( w_{B} \) is the metallurgical base width, \( N_{D} \) and \( N_{A} \) are the emitter and base doping levels, \( V_{EB} \) is the dc emitter-base terminal voltage and \( V_{BL} \) is the built-in potential at the proper junction. Under normal bias conditions the following is obtained:

\[ I_{n_{dc}}^{E} = \frac{e D_{n_{E}}}{L_{n}} \left[ n_{dc}(w_{2}) - n_{o} \right] \text{coth} \left( \frac{\varphi_{B}}{L_{n}} \right) \]

\[ (3.9) \]

and

\[ I_{n_{dc}}^{B} = \frac{e D_{A_{n}}}{L_{n}} \left[ n_{dc}(w_{2}) - n_{o} \right] \text{csch} \left( \frac{\varphi_{B}}{L_{n}} \right) \]

\[ (3.10) \]

The hole current crossing each junction must be calculated in order to find the total current. Unlike the electron current which links both emitter and collector junctions, the hole currents link only one junction because the base hole current can easily enter or leave
the base region owing to the ohmic contact to the external terminal.

The dc hole current at \( y = -w \) can be found from Eq. 3.5 by changing all the electron parameters to the corresponding hole parameters in the appropriate regions and by taking the limit of the resulting expression in the limit as \( w_B \to \infty \). The result is

\[
I_{p_{dc}} = I_{p_{dc}}(-w) = \frac{eD_Ae}{L_p} [p_{dc}(-w) - p_0], \quad (3.11)
\]

where \( D_p \) = the low-field hole diffusion constant, \( L_p \) = the hole diffusion length, \( p_0 \) = the hole thermal equilibrium concentration, and

\[
w = \left( \frac{2e}{e_{H_D}} \right) \left( \frac{N_A}{N_D + N_A} \right)^{1/2} \frac{1}{\sqrt{V_{BI} - V_{EB_0}}} . \quad (3.12)
\]

The total dc emitter particle current is given by

\[
I_{E_{dc}} = \frac{eD_Ae}{L_n} \left[ n_{dc}(w) - n_0 \right] \coth \left( \frac{w}{L_n} \right) + \frac{eD_Ae}{L_p} [p_{dc}(-w) - p_0] . \quad (3.13)
\]

If \( w_B \ll L_n \), Eq. 3.13 can be further reduced to

\[
I_{E_{dc}} = \frac{eD_Ae}{w_B} \left[ n_{dc}(w) - n_0 \right] + \frac{eD_Ae}{L_p} [p_{dc}(-w) - p_0] . \quad (3.14)
\]

A very important circuit parameter is the short-circuit dc current gain from emitter to collector and it can be written as the product of three parameters as follows:
The first factor is the injection efficiency $\alpha_{\text{inj}}$ which gives the fraction of the total emitter current which is made up of the electron carriers. Under normal bias conditions, the dc injection efficiency can be calculated by the following expression:

$$
\alpha_{\text{inj}} = \frac{1}{1 + \left( \frac{D_p}{D_n} \right) \left( \frac{L_B}{L_p} \right) \left( \frac{W_A}{W_B} \right) \tanh \left( \frac{L_B}{L_n} \right)} .
$$

The second factor is the base transport factor which is given by the following expression:

$$
\alpha_{t_o} = \text{sech} \left( \frac{W_B}{L_n} \right) .
$$

The most accurate two-term expression for $\alpha_{t_o}$ is

$$
\alpha_{t_o} = 1 - \frac{L_B^2}{2h^3} \left( \frac{W_B}{L_n} \right) .
$$

The third factor is the collector avalanche multiplication factor. For BJTs it is essentially equal to unity. For CATT devices, it is given by Eq. 2.7. For convenience, Eq. 2.7 is stated again as follows:

$$
N_A = \frac{1}{1 - \int_{0}^{L_B} \alpha_0 \left( \exp \int_{0}^{X} \left[ \alpha_p(x') - \alpha_n(x') \right] dx' \right) dx} .
$$
A good microwave BJT or CATT must, first of all, be a good dc BJT or CATT. From the expressions for \( a_{\text{inj}} \) and \( a_{\text{t}} \), the design requirements on the emitter and base regions can be deduced.

### 3.2.2 Base Spreading Resistance

The cross section of a single-emitter strip in an interdigitated device structure is shown in Fig. 3.2. The base spreading resistance can be calculated by using the following expression:

\[
R_B = \left( \frac{w}{12x_B \sigma_B^k} + \frac{d}{2x_B \sigma_B^k} \right) \frac{1}{\text{number of emitter-base finger pairs}},
\]

(3.19)

where the device structural parameters are defined in Fig. 3.2 and the conductivity of the uniformly doped base \( \sigma_B \) is

\[
\sigma_B = e u_p n_A.
\]

The variation in \( R_B \) due to base region conductivity modulation, which is caused by high concentrations of electron and hole carriers at medium or high-level injection, and base region width modulation, which is caused by the Early effect,\(^3\) has been ignored in the device modeling.

### 3.2.3 DC Computer Model

The dc computer model is shown in Fig. 3.3a. The particle current \( I_{\text{dc}} \) flowing in the diode is given by Eq. 3.13. The collector particle current \( I_{\text{dc}} \) is calculated by supplying input data \( I_{nB} \), \( V_0 \), and device structural parameters to the computer program DCCP, where \( I_{nB} \) is given by Eq. 3.10. When avalanche multiplication occurs in the collector depletion region,
FIG 3.2 CROSS SECTION OF SINGLE-EMITTER STRIPE IN AN INTERDIGITATED TRANSISTOR STRUCTURE.
FIG. 3.3 CALCULATION OF COMMON-BASE COLLECTOR CHARACTERISTICS. (a)

DC DEVICE MODEL AND (b) BLOCK DIAGRAM OF DC PROGRAM DCCP.
electron-hole pairs are created. Electrons, emitter injected and avalanche created, will drift toward the collector contact. Holes, mainly avalanche created, will be transported into the base region and therefore constitute a negative recombination current component.\textsuperscript{28, 54} They reduce the base current which must be supplied externally. If the avalanche multiplication factor is large enough, the polarity of the base current actually reverses because the avalanche multiplication process in the collector depletion region is providing more holes to the base than are necessary to support recombination in the base and emitter.

The common-base configuration is chosen because the consequences of this avalanche multiplication are straightforward in such an operation. The collector particle current simply equals the product of $I_{nB}^{dc}$ and $M_o$. Current $I_{E}^{dc}$ is the parameter of the common-base collector characteristics and it is dependent only on $V_{EB}^o$, the dc emitter-base terminal voltage. Therefore the common-base characteristics reflect primarily a direct avalanche contribution to the collector current. The consequences of avalanche multiplication are more complicated and subtle when the transistor is operating in the common-emitter configuration since base current $I_{B}^{dc}$ is the parameter of the collector characteristics. The avalanche multiplication factor is dependent on $V_{EB}^o$, $I_B^{dc}$ and $I_{E}^{dc}$ are dependent on $I_{nB}^{dc}$ and $M_o$, and $I_{nB}^{dc}$ is the difference between the $V_{EB}^o$-dependent recombination current and $I_{pB}^{dc}$. Therefore, points on an $I_{E}^{dc}$ vs. $V_{EB}^o$ (constant $I_{B}^{dc}$) characteristic curve may correspond to different values of $I_{nB}^{dc}$ and such a characteristic curve does not reflect directly the avalanche contribution to the collector current.
Figure 3.3b shows the block diagram of the computer program DCCP which produces the common-base dc collector characteristics. The results of computer program DCCP are given and discussed in Section 3.4.

3.3 Small-Signal Analytical Model

To consider the general ac case, the general one-dimensional time-dependent electron continuity equation must be solved:

$$\frac{\partial n}{\partial t} = \frac{1}{e} \frac{\partial J}{\partial y} - \frac{n - n_o}{t_n}.$$  \hspace{1cm} (3.20)

The following is obtained using Eqs. 3.3 and 3.20

$$\frac{\partial^2 n}{\partial y^2} - \frac{1}{L_n^2} (n - n_o) = \frac{1}{D_n} \frac{\partial n}{\partial t}.$$  \hspace{1cm} (3.21)

and the solution of Eq. 3.21 is of the following form:

$$n(y,t) = n_{dc} + n_1 e^{j\omega t}.$$  \hspace{1cm} (3.22)

when the sources are sinusoidal with frequency $\omega$. The first term on the right-hand side corresponds to the dc solution. The second term is the ac variation and is in the form of a product of two terms. This is a frequently used method of solving partial differential equations. It is assumed in Eq. 3.22 that the minority carrier concentration varies sinusoidally with the same frequency as that of the driving sources. When Eq. 3.22 is substituted into Eq. 3.21, the following two equations are obtained:

$$\frac{d^2 n_{dc}}{dy^2} - \frac{n_{dc} - n_c}{t_n^2} = 0.$$  \hspace{1cm} (3.23)
and

\[
\frac{d^2 n_1}{dy^2} + \left( \frac{1 + \omega t}{L_n^2 n_1} \right) n_1 = 0 . \tag{3.24}
\]

The boundary condition in Eq. 3.1 is quite general and is not restricted to dc applied voltages. Thus if the emitter-base junction voltage is composed of a dc voltage \( V_{EB_0} \) and an ac voltage \( V_{EB_1} e^{j\omega t} \), the minority carrier concentration on the base side of the emitter-base junction is

\[
n(y, t) = n_0 e^{V_{EB_0}/kT} + n_0 e^{V_{EB_1}/kT} e^{j\omega t} , \quad (3.25)
\]

where the first term on the right-hand side is the dc value and the second term is the ac value. Equation 3.25 is valid provided that

\[
|V_{EB_1}| << \frac{kT}{e} , \quad (3.26)
\]

which represents the small-signal assumption.

Note that the form of the differential equation for the ac electron density in Eq. 3.24 is the same as that for the dc electron density. This suggests that the ac solution can be obtained from the dc solution directly by the use of some simple transformations. The ac electron particle currents at \( y = w_2 \) and \( w_3 \) are found to be

\[
I_{nE_1} e^{j\omega t} = \frac{e^{2D_n A_n}}{kT} \frac{\sqrt{1 + \omega t}}{L_n n_1} x \frac{V_{EB_1} e^{j\omega t} n_{dc}(w_2)}{L_n} \coth \left( \frac{\sqrt{1 + \omega t} n_{EB_1}}{L_n} \right) \left( \frac{\sqrt{1 + \omega t} n_{EB_1}}{L_n} \right) . \tag{3.27}
\]

and
under normal bias conditions. The depletion region boundaries $w_2$ and $w_1$ are determined by Eqs. 3.7 and 3.8 with $V_{EB}$ and $V_T$ set at their respective dc values. The hole current at $y = - w_1$ is given by

$$I_{pE} e^{j \omega t} = \frac{e^{2D} A_E}{kt} \frac{\sqrt{1 + j \omega T}}{L_p} e^{j \omega t} V_{EB} e^{j \omega t} p_{dc}(-w_1). \quad (3.29)$$

Physically, $I_{nB} e^{j \omega t}$ is the time-dependent electron current injected into the base-collector depletion region. The expressions which approximate the physics in the collector region under small-signal conditions are derived next. Finally, the small-signal $y$-parameters of the common-base configuration are derived.

Most of the carrier generation due to avalanche multiplication occurs in a narrow high-field portion of the collector region, which is named the generation region, whose width is $w_{av}$. The remainder of the collector region is the drift region whose width is $w_D$. The basic device equations in the generation region are

$$\frac{\partial n}{\partial t} = \frac{1}{e} \frac{\partial J}{\partial x} + a_s(n + p) \quad (3.33)$$
\[
\frac{3p}{\partial t} = \frac{1}{e} \frac{J}{\partial x} + av (n + p), \tag{3.34}
\]

where \( a_n = a_p = a \) and \( v_{ns} = v_{ps} = v_s \) are assumed. The convention for the direction of particle flow and electric field for n-type CATT devices is shown in Fig. 3.4. The addition of Eqs. 3.33 and 3.34, substitution of Eqs. 3.31 and 3.32 and integration from \( x = 0 \) to \( x = v_{av} \) yields

\[
\frac{v_{av}}{v_s} \frac{dJ}{dt} = \left( E - \frac{1}{n} \right) v_{av} + 2J \int_{0}^{v_{av}} av \, dx,
\]

where \( J_g \) is the total conduction current in the generation region.

With boundary conditions

\[
J_n(0,t) = J_{nB}(t) + J_{cns}, \tag{3.35}
\]

and

\[
J_p(v_{av},t) = J_{cps}, \tag{3.36}
\]

the preceding equation becomes

\[
\frac{dJ_g(t)}{dt} = \frac{2J}{\tau_g} \left( \int_{0}^{v_{av}} av \, dx - 1 \right) + \frac{2\left(J_{nB}(t) + J_{cns} + J_{cps}\right)}{\tau_g}, \tag{3.37}
\]

where \( \tau_g = \frac{\Delta v_{av}}{v_s} \). With the small-signal assumptions

\[
\alpha = \alpha_0 + \alpha_1 e^{j\omega t},
\]

\[
J_g = J_{dc} + J_{g1} e^{j\omega t},
\]

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FIG. 3.4  (a) GENERAL DEVICE COLLECTOR STRUCTURE AND
(b) ELECTRIC FIELD AND CURRENT CONVENTIONS.
\[ E = E_0 + E_1 e^{j\omega t} \]

and

\[ J_{nB} = J_{nB_{dc}} + J_{nB_{1}} e^{j\omega t} , \]

the following can be obtained from Eq. 3.37

\[ J_{nB_{dc}} = \frac{J_{nB} + J_{c_{ns}} + J_{c_{ps}}}{1 - \int_0^{\omega v a} a_o \, dx} \]

(3.38)

the dc equation, and

\[ J_{nB_{1}} = \frac{J_{nB_{dc}} \left[ \int_0^{\omega v a} a_o \, E_0 + J_{nB_{1}} \right]}{1 - \int_0^{\omega v a} a_o \, dx + J \frac{\omega v E}{2}} \]

(3.39)

the ac equation, where the second-and higher-order terms are neglected.

If the collector has a Read-type structure and negligible space-charge effect, both \( E_0 \) and \( E_1 \) are constants. Equation 3.34 becomes

\[ J_{nB} \left( \frac{b}{E^2} \right) \left[ 1 - \frac{1}{M_{A_0}} \right] E_0 + J_{nB_{1}} \]

(3.40)

where \( M_{A_0} = \frac{1}{(1 - a_o w_{av})} \). By using the analytical expression for the ionization rate

\[ a_o = e^{-b/E_0} \]

\[ a_o = A e^{-b/E_o} , \]
Eq. 3.40 can be written as

\[
J_e = \left[ \frac{\ln\left(\frac{v_{av}}{s}\right) - \frac{1}{M_{o}}} b \right]^2 \left( \frac{1}{M_{o}^2} + \frac{\omega t}{2} \right) \frac{E_g + J_nB_1}{J_{nB_1}}.
\]

Equation 3.41 gives the current density at the interface between the generation region and the drift region.

If a saturated drift velocity \( v_s \) and zero carrier generation in the drift region are assumed, the ac conduction current density \( J_D(x) e^{j\omega t} \) in the drift region propagates as an unattenuated wave at this drift velocity

\[
J_D(x) = J_e e^{j\omega (x - v_{av})/v_s},
\]

where the exponential term represents the phase delay. The induced ac collector terminal current density is

\[
J_T = \frac{1}{v_D} \int_{w_{av}}^{w_T} J_D(x) \, dx = \frac{1}{v_D} \frac{\theta_D}{2} e^{-j(\theta_D/2)},
\]

where \( \theta_D = \frac{\theta_D}{v_{av}} \) and \( J_T(t) = J_{T_{dc}} + J_{T_1} e^{j\omega t} \).
The voltage across the base-collector terminal is \( V_T = V_{T_1} + V_{T_1} e^{j\omega t} \), where \( V_{T_1} \) is the ac voltage component. The ac electric field is constant throughout the collector region if the space-charge effects of \( J_D \) and \( J_T \) are ignored (see Fig. 3.5). The ac electric field in the generation region is

\[
E_{T_1} = \frac{V_{T_1}}{w_{T_1}}.
\]  

By combining Eqs. 3.41, 3.43 and 3.44, the following expression for \( I_{T_1} \) can be derived:

\[
I_{T_1} = \left( J_{n_{dc}} + J_{c_{ns}} + J_{c_{ps}} \right) \left( \frac{M_{A_0} - 1}{M_{A_0}} \right) \left[ \ln(w_{A_0}) - \ln\left( 1 - \frac{1}{M_{A_0}} \right) \right]^{2} 
\]

\[
I_{T_1} = \left( \frac{1}{M_{A_0}} + J \frac{w_f}{2} \right) \frac{V_{T_1}}{w_{T_1}}
\]

\[
+ \left( \frac{1}{M_{A_0}} + J \frac{w_f}{2} \right) A_e \sin \frac{\theta_D}{2} e^{-j(\theta_D/2)}.
\]  

The preceding expression is a more complete description of the small-signal device physics than the formulation of Winstanley and Carroll\(^3\) and it is very similar to the expression derived by Quang\(^3\) except that \( M_{A_0} \) is assumed to be infinity in the latter case. The total ac collector terminal current is given by

\[
I_c = I_T + jwC_1 V_{T_1},
\]  

where \( C_1 = \varepsilon A_1 / w_{T_1} \). The small-signal device model is given in Fig. 3.6a and the small-signal circuit model is given in Fig. 3.6b. The circuit
FIG. 3.5 ELECTRIC FIELD PROFILE UNDER RF CONDITIONS.
FIG. 3.6 SMALL-SIGNAL MODELS. (a) DEVICE MODEL

AND (b) CIRCUIT MODEL.
element $C_{TE}$ is the emitter-base transition region capacitance and its value is

$$C_{TE} = \frac{e A_c E}{w + w} \left( \frac{1}{1} + \frac{2}{V_{EB}} \right), \quad (3.47)$$

where $w$ and $w$ are calculated by using Eqs. 3.7 and 3.12. If $I_1$ and $I_2$, the total ac emitter current and the total ac collector current, are written as

$$\tilde{I}_1 = y_1 I_a + y_1 I_b \quad (3.48)$$

and

$$\tilde{I}_2 = y_2 I_a + y_2 I_b \quad (3.49)$$

where $I_1$ and $I_2$ are ac voltages across the terminals, then using Eqs. 3.27 through 3.29 and Eqs. 3.45 and 3.46 yields the following expressions:

$$y_{11} = \frac{e}{kT} \left( \frac{I_{nE} \sqrt{1 + j \omega_{n} \tanh \left( w_B / L_n \right)}}{\tanh \left( \frac{w_B \sqrt{1 + j \omega_n}}{L_n} \right)} \right) + \frac{I_{nE} \sqrt{1 + j \omega_p}}{\frac{I_{nE} \sqrt{1 + j \omega_p}}{1 + j \omega_p}} + j \omega C_{TE}, \quad (3.50)$$

$$y_{12} = 0 \quad (3.51)$$

$$y_{21} = -\frac{e}{kT} \left( \frac{I_{nE} \sqrt{1 + j \omega_{n} \tanh \left( w_B / L_n \right)}}{\sinh \left( \frac{w_B \sqrt{1 + j \omega_n}}{L_n} \right)} \right) \frac{1}{M_{A_0}} \frac{1}{1 + j \frac{\omega_{k}}{2}} + \frac{\sin \frac{\theta_D}{2}}{\frac{\theta_D}{2}} e^{-j(\theta_D/2)} \quad (3.52)$$
and

\[ y_{22} = \frac{(I_{nB} + I_{eNS} + I_{ePS})(M_{A0} - 1) \left[ \ln(w_{avA}) - \ln\left(1 - \frac{1}{M_{A0}}\right) \right]^2}{\left(\frac{1}{M_{A0}} + J \frac{\omega_E}{2}\right)(\omega_{b1})} \]

\[ \cdot \frac{\sin\frac{\theta_D}{2}}{\theta_D} \cdot e^{-J(\theta_D/2)} + J\omega_C. \quad (3.53) \]

If \( \tilde{y} \)-parameters are defined as

\[ \tilde{I}_1 = \tilde{y}_{11} \tilde{V}_1 + \tilde{y}_{12} \tilde{V}_2 \quad (3.54) \]

and

\[ \tilde{I}_2 = \tilde{y}_{21} \tilde{V}_1 + \tilde{y}_{22} \tilde{V}_2 \quad (3.55) \]

it can be shown that

\[ \tilde{y}_{11} = \frac{y_{11}(1 + R_B y_{22})}{1 + R_B(y_{11} + y_{21} + y_{22})}, \quad (3.56) \]

\[ \tilde{y}_{12} = -\frac{R_B y_{11} y_{22}}{1 + R_B(y_{11} + y_{21} + y_{22})}, \quad (3.57) \]

\[ \tilde{y}_{21} = \frac{y_{21} - R_B y_{11} y_{22}}{1 + R_B(y_{11} + y_{21} + y_{22})} \quad (3.58) \]

and

\[ \tilde{y}_{22} = \frac{(1 + R_B y_{11}) y_{22}}{1 + R_B(y_{11} + y_{21} + y_{22})}. \quad (3.59) \]
The input and output admittances, expressed in terms of $\tilde{y}_{ij}$, are

$$Y_{\text{in}} = G_{\text{in}} + jB_{\text{in}} = \tilde{y}_{11} - \frac{\tilde{y}_{12}\tilde{y}_{21}}{\tilde{y}_{22} + Y_L}$$  \hspace{1cm} (3.60)$$

and

$$Y_{\text{out}} = G_{\text{out}} + jB_{\text{out}} = \tilde{y}_{22} - \frac{\tilde{y}_{12}\tilde{y}_{21}}{\tilde{y}_{11} + Y_S}$$  \hspace{1cm} (3.61)$$

where $Y_L$ and $Y_S$ are the load and the source admittances, respectively.

By using the derived $\tilde{y}$-parameters, many small-signal device performance capabilities can be calculated.

The passivity condition\cite{55} is given as follows:

$$\bar{g}_{11} \geq 0$$  \hspace{1cm} (3.62)$$

and

$$\bar{g}_{22} \geq 0$$  \hspace{1cm} (3.63)$$

$$\bar{g}_{11}\bar{g}_{22} - \bar{g}_{12}\bar{g}_{21} - \frac{|\tilde{y}_{21} - \tilde{y}_{12}|^2}{4} \geq 0$$  \hspace{1cm} (3.64)$$

and

$$\bar{g}_{11}\bar{g}_{22} + \bar{g}_{12}\bar{g}_{21} - \frac{|\tilde{y}_{21} + \tilde{y}_{12}|^2}{4} \geq 0$$  \hspace{1cm} (3.65)$$

where $\bar{g}_{1j}$ and $\bar{g}_{1j}$ are the real and imaginary parts of $\tilde{y}_{ij}$. If any of these conditions are violated, the device is active instead of passive.

The passivity condition in Eq. 3.64 can be written as

$$1 - U \geq 0$$

where

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Thus, if $g_{11} \geq 0$ and $g_{22} \geq 0$ and $0 \leq U \leq 1$, the device is passive. If $U > 1$, the device is active. The quantity $U$ is a measure of the degree of activity. The maximum frequency of oscillation $\omega_{\text{max}}$ is defined as

$$\tilde{y}(\omega_{\text{max}}) = 1.$$  \hfill (3.67)

The Linvill stability factor $S$ is

$$S = \frac{|\tilde{y}_{12} - \tilde{y}_{21}|}{2g_{11} g_{22} - g_{12} g_{21} + b_{12} b_{21}}.$$  \hfill (3.68)

When $1 < S < \infty$ or when $S < 0$, the device is potentially unstable, that is, oscillation can occur for certain selected passive terminations. When $0 < S < 1$, the device is inherently stable, that is, no passive terminations can cause oscillation without suitable external feedback. When $S = 1$, the device is critically stable (see Fig. 3.7 for the definitions of various powers). The operating power gain is

$$G_p = \frac{P_{\text{out}}}{P_{\text{in}}} = \frac{|\tilde{y}_{21}|^2 G_L}{|\tilde{y}_{22} + Y_L|^2 \text{Re}\left\{\tilde{y}_{11} - \frac{\tilde{y}_{12} \tilde{y}_{21}}{\tilde{y}_{22} + Y_L}\right\}}.$$  \hfill (3.69)

where Re is the real part of any complex number and $G_L = \text{Re}(Y_L)$. If the amplifier is inherently stable, the maximum operating gain is achieved when the load is

$$G_{L_{\text{opt}}} = \frac{1}{2g_{11} \sqrt{[2g_{11} g_{22} - \text{Re}(\tilde{y}_{12} \tilde{y}_{21})]^2 - |\tilde{y}_{12} \tilde{y}_{21}|^2}}$$  \hfill (3.70)
FIG 3.7 LINEAR ACTIVE CATT CIRCUIT. ($P_{AVS}$ = THE POWER AVAILABLE FROM THE SOURCE
AND $P_{AVO}$ = THE POWER AVAILABLE TO THE LOAD)
and

\[ B_{L_{opt}} = \frac{\text{Im}(y_{12})}{2g_{11}} - \frac{b_{22}}{22} , \]  

(3.71)

where Im is the imaginary part of any complex number. Substitution of Eqs. 3.70 and 3.71 into Eq. 3.60 yields the input admittance \( Y_{\text{in}} \) with \( Y_L = Y_{L_{opt}} \). In order to deliver maximum power to the device the input admittance must be the conjugate of the source admittance; therefore, when \( Y_S \) is given by

\[ G_{S_{opt}} = \text{Re}(Y_S) = \frac{1}{2g_{22}} \sqrt{\left[ \frac{2g_{11} g_{22} - \text{Re}(y_{12})}{} \right]^2 - \left| y_{12} y_{21} \right|^2} \]  

(3.72)

and

\[ B_{S_{opt}} = \frac{\text{Im}(y_{12})}{2g_{22}} - \frac{b_{22}}{22} , \]  

(3.73)

maximum power as well as maximum operating power gain are achieved.

The optimum terminations for a CATT amplifier are usually obtained by adjusting the values of the components in the matching networks. Ease of tunability, or alignability, depends on fractional change in input admittance compared to the fractional change in load admittance and on the fractional change in output admittance compared to the fractional change in source admittance. When the fractional changes in input and output admittances are small, it is not necessary to alternate many times between input and output tuning to achieve the desired condition of operation. In terms of the \( y \)-parameters, the input
Tunability is given by

\[ \delta = \left| \frac{dY_{\text{in}}/Y_{\text{in}}}{dY_{\text{L}}/Y_{\text{L}}} \right| = \frac{|\tilde{\gamma}_{12} \tilde{\gamma}_{21}| |Y_L|}{|\tilde{\gamma}_{22} + Y_L| \left| \tilde{\gamma}_{22} \left( \tilde{\gamma}_{11} + Y_L \right) - \tilde{\gamma}_{12} \tilde{\gamma}_{21} \right|}. \tag{3.74} \]

Tuning is easier as \( \delta \) is made smaller. In order to achieve good tunability, the load admittance may be chosen such that \( Y_L > \tilde{\gamma}_{22} \) in order to make \( \delta < 0.3 \). Then Eq. 3.74 becomes

\[ \delta = \left| \frac{\tilde{\gamma}_{12} \tilde{\gamma}_{21}}{\tilde{\gamma}_{11} Y_L} \right|. \tag{3.75} \]

The output power varies with frequency partly because the device parameters are frequency dependent but primarily because the conditions for a conjugate match at the input and output ports may not be obtained except at the design-center frequency \( f_0 \). The amplifier bandwidth can be approximated under certain conditions from the inherent bandwidth \( B\text{W}^i \) at either port given by\(^{55}\)

\[ B\text{W}^i_{\text{in}} = \frac{2f_0 G_{\text{S}}}{B_{\text{S}}^{\text{opt}}} \tag{3.76} \]

for the input circuit, and by

\[ B\text{W}^i_{\text{out}} = \frac{2f_0 G_{\text{L}}}{B_{\text{L}}^{\text{opt}}} \tag{3.77} \]

for the output circuit. Inherent bandwidth is the name given to the bandwidth that is obtained in the conjugate matching situation where the optimum terminations are determined only by the device parameters. It often happens that \( B\text{W}^i_{\text{in}} \gg B\text{W}^i_{\text{out}} \) and in such cases the overall
amplifier bandwidth is determined by the output circuit. Bandwidths are inversely proportional to $B_L$ and $B_S$. Therefore, the input or output bandwidth can be decreased below the inherent bandwidth by adding shunt capacitances $C_L$ or $C_S$ across the input and output terminals without affecting the optimum termination conductances. Maximum operating power gain is still achieved. When an inherent bandwidth is too small for a particular application, the bandwidth can be increased by raising the termination conductances. This means that additional shunt resistance can be connected across the input or output terminals. The power gain is decreased when the bandwidth is increased.

When the transistor parameters are such that potential instability occurs, a certain output load causes the input conductance to be zero or negative and the power gain to be infinite. A mismatching technique can be used to ensure the stability of the amplifier circuitry.

3.4 Dc and Small-Signal Results

Typical common-base characteristics of CATT devices are shown in Figs. 3.8 through 3.10. Carrier multiplication is evident from the common-base characteristics. Figures 3.8 through 3.10 represent three CATT devices with different doping densities in the avalanche multiplication region. The effects of $N_{av}$ are observed in the different rates of increase in $J_{T_{dc}}$ as a function of $V_{T_{o}}$, different breakdown voltages and different punch-through voltages. The effects of $w_{av}$, $N_{drift}$ and $w_D$ on the common-base characteristics can also be investigated by applying the computer program DCCP.

It can be inferred from Figs. 3.8 through 3.10 that common-base biased CATT devices can be employed to produce current amplification, and the device in Fig. 3.8 is suitable for such applications. The dc
FIG. 3.8 COMMON-BASE CHARACTERISTICS. (n-TYPE, Si, $v_{av} = 1 \times 10^{-4}$ cm, $n_{av} = 2.5 \times 10^{16}$ cm$^{-3}$, $v_{D} = 3 \times 10^{-4}$ cm, $n_{drift} = 2 \times 10^{15}$ cm$^{-3}$ AND $T = 300^\circ$K)
FIG. 3.9 COMMON-BASE CHARACTERISTICS. (n-TYPE, Si,

\[ v_{av} = 1 \times 10^{-4} \text{ cm}, \quad N_{av} = 1.5 \times 10^{16} \text{ cm}^{-3}, \]
\[ v_D = 3 \times 10^{-6} \text{ cm}, \quad N_{drift} = 2 \times 10^{15} \text{ cm}^{-3} \]
AND \( T = 300^\circ K \))

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FIG. 3.10 COMMON-BASE CHARACTERISTICS. (n-TYPE, Si,

\[ v_{av} = 1 \times 10^{-4} \text{ cm}, \quad n_{av} = 2 \times 10^{16} \text{ cm}^{-3}, \]

\[ v_D = 3 \times 10^{-4} \text{ cm}, \quad n_{\text{drift}} = 2 \times 10^{15} \text{ cm}^{-3} \]

AND \( T = 300^\circ K \))

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base-collector bias should be set close to the breakdown voltage as long as the thermal limitation is not exceeded and the load resistor is small (see Fig. 3.8). For applications in voltage amplifications, the device in Fig. 3.9 is more suitable. In fact, a device with a lightly doped, uniform collector structure has even greater capability for voltage amplification. The dc base-collector bias should be set close to the breakdown voltage in Class C operation and set close to 

\((1/2)(V_B + V_{pt})\) in Class A operation. The upper and lower limits of \(V_T\) are \(V_B\) and \(V_{pt}\), respectively. It is assumed that a punched-through or nearly punched-through collector is a required operating condition. The load resistor is large. For applications in power amplification, the device in Fig. 3.10 is more suitable where both a large current multiplication and large \(V_{rp}\) can result with proper dc bias and load resistor applied.

A computer program SSCP was written which is capable of determining the small-signal common-base y-parameters of CATT amplifiers. Typical \(U\) vs. frequency and maximum operating power gain vs. frequency plots are shown in Figs. 3.11 through 3.14. The quantity \(U\) is plotted over the frequency range in which \(U \geq 0\). Note that the device can still be active even if \(U < 0\) as long as either \(g_{11} < 0\) or \(g_{22} < 0\). Maximum operating power gain is displayed over the entire frequency range in which the device is capable of power amplification.

From Figs. 3.11 through 3.14 it is observed that the maximum operating power gain at \(M_A > 1\) is much lower than the maximum operating power gain at \(M_A = 1\), but the gain at \(M_A > 1\) is constant over a very wide range of frequencies. If \(V_{eb} \leq 0.6\) V, avalanche multiplication will increase \(f_{max}\), the maximum frequency at which an operating power
FIG. 3.11 MAXIMUM RF POWER GAIN AND U VS. FREQUENCY. \( \omega_T = 4 \times 10^{-4} \) cm AND \( V_{EB_o} = 0.7 \) V
FIG. 3.12 MAXIMUM OPERATING POWER GAIN AND U VS. FREQUENCY. ($w_T = 4 \times 10^{-4}$ cm AND $V_{EB_0} = 0.8$ V)
FIG. 3.13 MAXIMUM RF POWER GAIN AND U VS. FREQUENCY. ($w_T = 4 \times 10^{-4}$ cm AND $V_{EB_o} = 0.85$ V)
FIG. 3.14 MAXIMUM OPERATING POWER GAIN AND U VS. FREQUENCY. \( (w_T = 4 \times 10^{-4} \text{ cm and } V_{EB_0} = 0.9 \text{ V}) \)
gain greater than unity occurs. From Figs. 3.11 and 3.12 it is observed that at frequencies higher than $f_x$, avalanche multiplication also increases the operating power gain. If $V_{EB} > 0.85 \text{ V}$, $f_{\text{max}}$ decreases with increasing $M_{A_o}$ and the operating power gain at $M_{A_o} > 1$ is smaller than the operating power gain at $M_{A_o} = 1$ throughout the entire active frequency range. The effects of $V_{EB}$ on $f_{\text{max}}$ and the operating power gain are illustrated in Fig. 3.15. Highest $f_{\text{max}}$ at $M_{A_o} = 5$ is achieved when $V_{EB} = 0.7 \text{ V}$. Highest operating power gain at $M_{A_o} = 5$ is achieved when $V_{EB} = 0.8 \text{ V}$. When $V_{EB} \leq 0.7 \text{ V}$, $f_{\text{max}}$ is severely limited by a large value of the emitter space-charge resistance. When $V_{EB} > 0.85 \text{ V}$, $f_{\text{max}}$ and operating power gain are reduced due to large values of $J_{nB}$. From Eq. 3.45, there are two components in $J_{T_1}$. One component represents the carriers contained in $J_{nB}$ undergoing constant avalanche multiplication which is determined by $V_{T_0}$. The second component, which is

$$
A_{E} \left[ J_{nB_{dc}} + J_{c ns} + J_{c ps} \right] \left[ M_{A_o} - 1 \right] \left[ \ln(w_{a v}) - \ln\left[ 1 - \frac{1}{M_{A_o}} \right] \right]^2
$$

$$
\frac{1}{M_{A_o}} + j \frac{\omega T}{2} (b_{T_0})
$$

represents the carriers corresponding to $J_{nB_{dc}} + J_{c ns} + J_{c ps}$ undergoing nonconstant avalanche multiplication which depends on the instantaneous voltage $V_T = V_{T_0} + V_{T_1} e^{j\omega t}$. The second current component is responsible for the decreases in $f_{\text{max}}$ and the operating power gain as

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FIG. 3.15 MAXIMUM OPERATING POWER GAIN VS. FREQUENCY.

\( V_{EB_0} = 0.4 \)  
\( V_{EB_0} = 0.5 \)  
\( V_{EB_0} = 0.6 \)  
\( V_{EB_0} = 0.7 \)  
\( V_{EB_0} = 0.8 \)

\( L = 1 \times 10^{-4} \) cm AND \( M_{A_0} = 5 \)
increases. If the second current component is artificially forced to vanish, the corresponding maximum operating power gain and $U$ are displayed in Figs. 3.16 and 3.17. Note that the power gain and $f_{\text{max}}$ at $M_{A_0} = x$ are always less or equal to operating power gain and $f_{\text{max}}$ at $M_{A_0} = y$, if $x < y$. Figure 3.18 shows that for a given value of $M_{A_0}$, higher $V_{EB}$, up to 0.8 V, implies higher $f_{\text{max}}$ and higher operating power gain. For $V_{EB} > 0.8$ V, maximum operating power gain coincides with that corresponding to $V_{EB} = 0.8$ V. Comparison of previous results and the results obtained by artificially setting the second current component to zero indicates that, for the most part, avalanche multiplication of carriers in CATT amplifiers, when operating in Class A, works toward reducing the operating power gain. The situation is that the value of $V_{EB}$ must not be too small, otherwise, the emitter space-charge resistance will severely reduce $f_{\text{max}}$ and the operating power gain, and if the value of $V_{EB}$ is large, the second current component will dominate which has been shown to degrade the operating power gain. From the expression for $I_T$, the phase difference between $V_T$ and the second current component equals $(\theta_{nB}/2) + \tan^{-1}(\omega T M_{A_0}/2)$. If $-\pi/2 < \text{phase difference} < \pi/2$, energy is being dissipated. The situation in Class A operation at low frequency is depicted in Fig. 3.19. It is assumed that the load is a high-Q tank circuit whose resonant frequency equals the emitter-base input signal frequency. Notice that the amplitude of the first current component, when $V_T$ is large, is less than the value given by Eq. 3.45. The reason for this is that when $I_{nB}$ is maximum, $V_T$ is minimum and the carrier multiplication is at its minimum. When $I_{nB}$ is minimum, $V_T$ is maximum and the carrier multiplication is at its maximum. This
FIG. 3.16  MAXIMUM OPERATING POWER GAIN AND U VS. FREQUENCY WHEN AVALANCHE MULTIPLICATION
OF CARRIERS CORRESPONDING TO $J_{nB_{dc}}$ IS IGNORED. ($w_T = 4 \times 10^{-6} \text{ cm and } V_{EB_o} = 0.7 \text{ V}$)
FIG. 3.17  MAXIMUM OPERATING POWER GAIN AND $U$ VS. FREQUENCY WHEN AVALANCHE MULTIPLICATION
OF CARRIERS CORRESPONDING TO $J_{nB_{dc}}$ IS IGNORED. ($v_T = 4 \times 10^{-4}$ cm AND $V_{EB_o} = 0.85$ V)
FIG. 3.18 MAXIMUM OPERATING POWER GAIN AND $U$ VS. FREQUENCY WHEN AVALANCHE MULTIPLICATION
OF CARRIERS CORRESPONDING TO $J_{nB_{dc}}$ IS IGNORED. ($w_n = 4 \times 10^{-4}$ cm AND $M_{A_o} = 5$)
FIG. 3.19 SCHEMATIC PLOTS OF $J_{nB}$, $V_T$, FIRST COMPONENT OF $J_{T_1}$, AND SECOND COMPONENT OF $J_{T_1}$. -108-
situation can be seen by following the load line in Fig. 3.10. From the phase relationship between $V_T$ and the second current component, it is clear that the second current component corresponds to energy dissipation, therefore, it reduces the operating power gain. The usefulness of carrier multiplication and long transit region can be realized only in Class C operation, assuming a proper load is applied across the base-collector terminal and the width of the charge pulse is much less than $\pi$ rad. The Class C operation of CATT amplifiers was briefly described in Chapter I and the large-signal simulation results are discussed in detail in Chapter IV.

3.5 Operation of a CATT as an IMPATT with Variable Equivalent Thermally Generated Current

There has been considerable interest in the operation of an IMPATT diode with an externally controllable equivalent thermally generated current. Sanderson and Jordan\textsuperscript{56} discussed the operation of an IMPATT acted upon by an electron beam. The electron-beam-produced current is subject to multiplication in the same way as the flow of actually thermally generated carriers. The sum of the electron-beam-produced current and the current due to thermally generated carriers is, by definition, the equivalent thermally generated current. Sanderson and Jordan derived small-signal IMPATT expressions which theoretically predict that an appreciable change in the IMPATT resonant frequency can be induced by varying the electron-beam intensity. Results suggest that frequency modulation of the IMPATT oscillator can be implemented by this method. Forrest and Seeds\textsuperscript{57} and others have studied the controlling of microwave IMPATT oscillators with an optical signal.
The use of an electron beam or an optical signal on an IMPATT is equivalent to the situation where the base collector of a CATT is operated as an IMPATT where the emitter and base simply act as an externally controlled thermal current source. The expressions derived by Sanderson and Jordan can simply be obtained from Eqs. 3.37 by allowing $M_A = \text{cond} J_B(t) = J_B^{dc}$.

### 3.6 Conclusions

In this chapter analytical expressions required for dc and ac small-signal CATT device models have been derived. A simple computer program, DCCP, has been written which can generate the common-base characteristics of various CATT devices at minimal cost. Another computer program, SSCP, has been developed which can find the device small-signal $y$-parameters, unilateral gain, maximum operating power gain, optimum load, inherent input and output bandwidth, Linwill stability factor, amplifier tunability, and the maximum frequency of oscillation. The output results of SSCP indicate that under certain restricted operating conditions $f_{\text{max}}$ increases with increasing $M_{A_o}$ and the operating power gain at $M_{A_o} > 1$ is higher than the operating power gain at $M_{A_o} = 1$ for frequencies higher than a critical value $f_x$. However, for frequencies lower than $f_x$, the operating power gain at $M_{A_o} = 1$ is always much higher than the operating power gain at $M_{A_o} > 1$. The cause for the gain reduction was determined to be the component of $J_T$ due to $V_T$-dependent multiplication of those carriers corresponding to $J_B^{dc}$. This component of $J_T$ represents energy dissipation. Computational results of SSCP and DCCP indicate that Class A operation is not very suitable for CATT amplifiers.
CHAPTER IV. LARGE-SIGNAL COMPUTER MODEL

4.1 Introduction

In this chapter a large-signal computer simulation is developed which is hybrid in the sense that the emitter and base regions are modeled by lumped, nonlinear and time-varying circuit elements whose values are determined by a set of analytical expressions while the collector region is modeled by employing the difference-equations version of the semiconductor differential equations. The analytical expressions for the circuit elements of the lumped-circuit model for the emitter and base regions are derived. The numerical methods used to solve the difference equations are modified versions of techniques which have evolved in the Electron Physics Laboratory in the past decade. Detailed descriptions of the numerical algorithms, computer programs, and computer output are also given.

4.2 Development of Device Simulation

4.2.1 General Description. Figure 2.1 shows the depletion regions of a normally biased CATT device and its circuit model is shown in Fig. 4.1. The emitter-base region of the CATT device is represented by three nonlinear circuit elements similar to those in BJT's: a diode in which the particle current is injected across the forward-biased emitter-base junction whose magnitude depends on the emitter-base junction voltage $V_{EBJ}$, a $V_{EBJ}$-dependent depletion capacitance $C_{TJ}$, and a $V_{EBJ}$-dependent diffusion capacitance $C_{DE}$. Resistor $R_B$ is the equivalent base region spreading resistance. The collector region is represented by a capacitor $C_C$ and a dependent current source $I_T$. The
FIG. 4.1 CIRCUIT MODEL FOR A CATT AMPLIFIER IN THE COMMON-BASE CONFIGURATION.
capacitor \( C_c \) is the collector depletion capacitance and the dependent current source \( I_T \) represents the induced current whose value depends on the amount of charge injected into the depleted collector region from the neutral base region and the magnitude of carrier multiplication. It is assumed that the drift region is always completely or nearly completely depleted. Otherwise a \( V_T \)-dependent resistor must be added in series with the collector circuit representation depicted in Fig. 4.1.

The remainder of this section is devoted to detailed derivation and physical interpretation of each element of the device model.

### 4.2.2 Emitter-Base Circuit Model and Computer Program EBCP

In the bulk of the base region, the space-charge neutrality condition always holds. Thus in the base region of n-type CATT devices,

\[
n(y) + N_A(y) - p(y) = 0 \quad (4.1)
\]

An important observation about current in n-type CATT devices is that there is a negligible flow of holes (majority carriers) in most of the base region between the emitter-base and base-collector metallurgical junctions in the reverse direction. This is true under all bias conditions because there can be only a vanishingly small flow of holes from either n-region. The following expressions can be written for hole current in the longitudinal dimension \( y \) and the electric field:

\[
J_p = 0 = e u_p E(y) - e D \frac{dp}{dy}
\]

and

\[
E(y) = \frac{kT}{e} \frac{1}{p} \frac{dp}{dy} \quad (4.2)
\]
where \( p \), the concentration of holes in the bulk base region, is related to \( n \) and \( N_A \) through Eq. 4.1. In a uniformly doped base region, the electric field is negligible at low injection levels, that is, when the concentration of minority carriers is much lower than the base region impurity doping concentration. Therefore, the emitter-base junction voltage \( V_{EBJ} \) equals the emitter-base terminal voltage \( V_{EB} \). At high injection levels, the electron concentration is significantly higher than its thermal equilibrium concentration and the electric field in the bulk base region can no longer be ignored. A portion of the emitter-base terminal voltage appears across the neutral base region such that \( V_{EBJ} \) is less than \( V_{EB} \):

\[
V_{EBJ} = V_{EB} - \int_{w_2}^{w} E(y) \, dy.
\]

By substituting the expression in Eq. 4.2 into the preceding relation, \( V_{EBJ} \) can be expressed as follows:

\[
V_{EBJ} = V_{EB} - \frac{kT}{e} \ln \left( \frac{p(w_2)}{N_A} \right).
\]  \hspace{1cm} (4.3)

By using Eq. 4.3; the charge neutrality condition at \( y = w_2 \); and the junction law, \( n(w) = n_i(w) \exp \left( \frac{eV_{EB} / kT}{N_A} \right) \), the following expression for \( V_{EBJ} \) is derived:

\[
V_{EBJ} = V_{EB} - \frac{kT}{e} \ln \left[ \frac{1}{2} \sqrt{1 + \frac{l_n^2 \exp \left( \frac{eV_{EB} / kT}{N_A} \right)}{N_A^2}} \right],
\]  \hspace{1cm} (4.4)

where \( n_i \) is the intrinsic carrier concentration. In order to be

\hspace{1cm} -114-
assured that Eq. 4.4 is valid at any injection level, it is necessary to prove first that the junction law is valid at any injection level.

Figure 4.2 shows the space-charge distribution and the voltage distribution $\psi(y)$ in the depletion region of a p-n step junction where a voltage $V$ is applied to the n-region. In the space-charge region of a p-n junction there is a large field strength $E$ and a large carrier density gradient. In the current density equation for holes,

$$J_p = e
\mu_p E - \epsilon \frac{\partial p}{\partial y},$$

the current density $J_p$ is the difference between two large opposing currents so that $|J_p| << e\mu_p |E|$ and $|J_p| << \epsilon \frac{\partial p}{\partial y}$ for most of the space-charge region. Hence a good approximation is

$$- e\mu_p \frac{\partial \psi}{\partial y} - \epsilon \frac{\partial p}{\partial y} = 0$$

or

$$p(y) = p(\nu_1) \exp \left(- \frac{\psi(y)}{kT}\right) \quad (4.5)$$

for $-\nu_1 \leq y \leq \nu_2$. Similarly, for electrons the following is obtained

$$n(y) = n(-\nu_1) \exp \left(- \frac{e[V_{BI} - \psi(y)]}{kT}\right) \quad (4.6)$$

for $-\nu_1 \leq y \leq \nu_2$ where $V_{BI}$ is the junction built-in potential. Here it is assumed that the quasi-Fermi levels for electrons and holes are constant in the depletion region. By using Eqs. 4.1, 4.5 and 4.6, the following expressions are obtained:
FIG. 4.2 (a) SPACE-CHARGE DISTRIBUTION IN A p-n STEP JUNCTION
AND (b) VOLTAGE DISTRIBUTION IN A p-n JUNCTION; A
VOLTAGE - V IS APPLIED TO THE n-REGION. V_{BI} IS THE
JUNCTION BUILT-IN POTENTIAL.
\[ p(-w) = \frac{N_A B + N_B B^2}{1 - B^2}, \]  
(4.7)

\[ n(-w) = \frac{N_A B + N_B}{1 - B^2}, \]  
(4.8)

\[ p(w) = \frac{N_A + N_B B}{1 - B^2} \]  
(4.9)

and

\[ n(w) = \frac{N_A B^2 + N_B B}{1 - B^2}, \]  
(4.10)

where

\[ B = \exp \left[ - e(V_B - V_{EBJ})/kT \right]. \]

Based on the facts that in n-type CATIT devices \( N_D \gg N_A \) and \( B \) is generally much less than one, Eqs. 4.8 and 4.10 can be reduced to

\[ n(-w) = N_D \]

and

\[ n(w) = N_D B \]

\[ = n_o(w) \exp \left( eV_{EBJ}/kT \right) \]  
(4.11)

and they are always valid regardless of the injection level. On the other hand, Eqs. 4.7 and 4.9 are reduced to
\[
p(-w) \equiv N_A B
\]
\[
= p_o(-w) \exp(eV_{EBJ}/kT)
\]
and
\[
p(w) = N_A
\]

if \(N_A/N_D >> B\), which is the low injection level condition. When the injection level is no longer restricted to low injection levels, hole concentrations at \(y = -w\) and \(w\) are given by
\[
p(-w) = \{p_o(-w) + n_o(w) \exp[-e(V_{BI} - V_{EBJ})/kT]\}
\]
\[
\exp(eV_{EBJ}/kT)
\]
and
\[
p(w) = N_A + n(w) .
\]

It has been shown that the junction law for electrons is always valid in an n-type CATT device in which \(N_D >> N_A\) and therefore Eq. 4.4 is valid at any injection level.

It has been shown that, except at low injection levels, there is a finite voltage drop in the bulk base region and there exists a finite electric field. This electric field aids the flow of minority carriers in the bulk base region. An effective diffusion constant can be defined which includes the effect of the drift field. The expression for the electron current density is
\[
J_n = e n E_n + e D_n \frac{dn}{dy} .
\]
Substituting the relation in Eq. 4.2 into the preceding expression yields

\[ J_n = eD \left( 1 + \frac{n}{N_A + n} \right) \frac{dn}{dy} , \]

and when it is evaluated at \( y = w \), the result is

\[ J_{nE} = \frac{eD_n}{w_B} \left( 1 + \frac{l}{N_A} \right) \frac{n_0(w_2)[\exp(eV_{EBJ}/kT) - 1]}{1 + \frac{n_0(w_2)}{n_0(w_2)} \exp(eV_{EBJ}/kT)} , \]

where

\[ w_B = w_3 - w_2 \cdot \]

At low injection levels, the expression for \( J_{nE} \) is

\[ J_{nE} = \frac{eD_n}{w_B} n_0(w_2)[\exp(eV_{EBJ}/kT) - 1] , \]

and at very high injection levels the expression for \( J_{nE} \) becomes

\[ J_{nE} = \frac{2eD_n}{w_B} n_0(w_2) \exp(eV_{EBJ}/kT) . \]

The following can then be written:

\[ I_{nE} = \frac{eA_n D_{nEFF}}{w_B} [n(w_2) - n_0(w_2)] , \quad (4.12) \]

where

\[ D_{nEFF} = D \left( 1 + \frac{1}{N_A} \right) \left( 1 + \frac{n(A)}{n(w_2)} \right) . \]
Depletion region widths $w_1$, $w_2$, and $w_3$ can be calculated by using the following analytical expressions:

\[ w_1 = \left( \frac{2e}{eN_D} \right) \left( \frac{N_A}{N_A + N_D} \right)^{1/2} \frac{1}{\sqrt{V_{BI} - V_{EBJ}}} \]  \hspace{1cm} (4.13)

\[ w_2 = \left( \frac{2e}{eN_A} \right) \left( \frac{N_D}{N_A + N_D} \right)^{1/2} \frac{1}{\sqrt{V_{BI} - V_{EBJ}}} \]  \hspace{1cm} (4.14)

and

\[ w_3 = \frac{\varepsilon E(w_B)}{eN_A} \]  \hspace{1cm} (4.15)

where $V_{EBJ}$ is calculated by using Eq. 4.4 and $E(w_B)$ is provided by the collector simulation computer program CCP which is discussed in Section 1.2.3. The Early effect of base width modulation is included in this device model through Eqs. 4.14 and 4.15.

An expression for the emitter particle current, which is valid at any injection level, can be obtained from Eq. 3.14 by replacing $D_n$, $n(w_2)$, $p(-w_1)$, $V_{EB}$, and $w_B$ with the appropriate expressions that have been derived in this section. The emitter particle current is given by

\[ I_E = eA \left[ \frac{D_{n,\text{EFF}}}{w_B} [n(w_2) - n_0(w_2)] + \frac{D}{w_E} [p(-w_1) - p_0(-w_1)] \right] \]  \hspace{1cm} (4.16)

where
\[ w_E = w_{E_0} - w \text{ if } w_{E_0} - w < L_p, \]

and

\[ w_E = L_p \text{ if } w_{E_0} - w > L_p. \]

The charge \( Q_{EB} \) on each side of the emitter-base junction is

\[ Q_{EB} = eN_D w = eN_w. \]

Using the preceding relations yields

\[
C_{TE} = \frac{dQ_{EB}}{d(V_{BI} - V_{EBJ})} = \frac{A_p}{2} \left\{ 2e \frac{n_D n_A}{e \beta + e} \right\}^{1/2} (V_{BI} - V_{EBJ})^{-1/2}. \tag{4.17}
\]

The displacement current through \( C_{TE} \) is \( I_{CTE} \) and is given by

\[
I_{CTE} = \frac{dQ_{EB}}{dV_{EBJ}} \frac{dV_{EBJ}}{dt} = C_{TE} \frac{dV_{EBJ}}{dt}.
\]

The emitter-base diffusion capacitance \( C_{DE} \) is associated with the storage of minority carriers in the bulk base region. If straight-line minority carrier distribution and \( n(w) = 0 \) are assumed, the total minority carrier charge stored in the neutral base region is

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\[ Q_B = \frac{1}{2} e A \omega n(w) \exp\left(\frac{e V_{EBJ}}{kT}\right), \]

and the current associated with \( C_{DE} \) is

\[ I_{CDE} = \frac{dq_B}{dt} = \frac{dQ_B}{dV_{EBJ}} \frac{dV_{EBJ}}{dt} = C_{DE} \frac{dV_{EBJ}}{dt}, \quad (4.19) \]

where

\[ C_{DE} = \frac{dQ_B}{dV_{EBJ}} = \frac{e}{kT} I_n E \frac{w_B^2}{2D_n E F}. \quad (4.20) \]

In reality \( n(w) \) cannot be zero \(^6\) because nonzero electron current density exists at \( y = w \). Because of nonzero \( n(w) \), an additional charge will be stored in the base. Kirk\(^5\) showed that the fluctuation of this stored charge in response to the input signal will introduce an additional time delay, approximately equal to \( v_B/v_{ns} \), where \( v_{ns} \) is the electron saturated velocity. There is another time delay which is associated with emitter heavy doping. Until recently\(^{64-66} \) it had been assumed that the emitter should be as heavily doped as possible in order to ensure a high emitter injection efficiency. However, high doping will lead to the formation of band-edge tails and impurity level
broadening into an impurity band. DeMan introduced a doping-dependent intrinsic carrier concentration $n_i$ to account for the effect of heavy doping on the classical expressions for the electrical parameters. The validity of this approach was later verified theoretically. Under heavy doping conditions the hole current density injected into the emitter from the base is given by

$$J_p = eD_p \left( \frac{P}{n_i^2} \frac{dn_i}{dy} - \frac{P}{N_D} \frac{dN_D}{dy} - \frac{P}{N_p} \frac{dN_p}{dy} \right).$$

The holes experience an additional drift force proportional to the derivative of the change of the valence band edge. Therefore an additional component of hole current flowing from the base into the emitter of an n-type CATT device is introduced. Additional holes stored in the emitter cause an emitter delay time given by

$$\tau_e = \frac{1}{E_o} \int_0^{w E_o} \frac{N_p^2}{N_D} \left( \frac{y N_D}{D n_i^2} \right) dy,$$

where $B_0 = (D N_D w E_o / D N_p w E_o)$. Therefore, microwave CATT devices are designed to have narrow emitter junction depths not only to reduce the emitter capacitance and the emitter sidewall capacitance, but also to reduce $\tau_e$. Henderson and Scarbrough considered the special case of flat arsenic and boron concentrations in the emitter region. In this case $N_D$ and $n_i$ are both constant and therefore $\tau_e$ is given by

$$\tau_e = \frac{w E_o^2}{2D_B g^2}.$$
By including the two additional time delays, an effective diffusion capacitance can be defined as follows:

\[
C_{DE\text{eff}} = \frac{e}{kT} nE \left( \frac{v_B^2}{2.5^{3D\text{EFF}}} + \frac{w_B}{v_{ns}} + \frac{w_B^2}{p_{o}} \right). \tag{4.21}
\]

The base spreading resistance is calculated by using Eq. 3.19 and for convenience it is stated again below:

\[
R_B = \left( \frac{w}{12x_B \sigma_B^f} + \frac{d}{2x_B \sigma_B^f} \right) \frac{1}{\text{number of emitter-base finger pairs}}
\]

where the device structure is defined in Fig. 3.2. Base region conductivity modulation due to electron and hole carriers and the base region width modulation due to the Early effect are ignored in Eq. 3.19 although these effects can easily be included in the analysis.

The computer program for the emitter-base region EBCF solves the circuit model in Fig. 4.3. It is assumed that \( V_{\text{sig}} \), \( R_B \), and \( I_c \) the total collector current, are known quantities where the current \( I_c \) is provided by the collector computer program CCP. Kirchhoff's voltage law demands

\[
V_{\text{sig}}(t) - V_{EB}(t) - R_B \left( I_c(V_{EB}) + \left[ C_{TE}(V_{EB}) + C_{DE}(V_{EB}) \right] \frac{dV_{EB}}{dt} \right) = 0.
\]

This equation can be written as

\[
\frac{dV_{EB}(t)}{dt} = \frac{1}{C_{TE}(V_{EB}) + C_{DE}(V_{EB})} \left( \frac{1}{R_B} \left[ V_{\text{sig}}(t) - V_{EB}(t) \right] + I_c(t) - I_c(V_{EB}) \right). \tag{4.22}
\]
FIG. 4.3 EQUIVALENT CIRCUIT FOR THE
EMITTER-BASE REGION.
which is the differential equation to be solved numerically by the computer program EBCP. This differential equation is a first-order ordinary differential equation of the form

\[ \frac{df}{dt} = g(t,f) \]

and can be solved by the fourth-order Runge-Kutta method:

\[ f(t + \Delta t) = f(t) + \Delta t \left( \frac{k_1}{6} + 2k_2 + 2k_3 + k_4 \right), \quad (4.23) \]

where

\[ k_1 = g(t, f(t)), \]
\[ k_2 = g(t + \Delta t/2, f(t) + k_1 \Delta t/2), \]
\[ k_3 = g(t + \Delta t/2, f(t) + k_2 \Delta t/2) \text{ and} \]
\[ k_4 = g(t + \Delta t, f(t) + k_3 \Delta t). \]

The assumption that \( I_c(t) \) is a known quantity, at this point, appears artificial because \( I_c \) is known to be dependent on the electron particle current injected into the collector region which is dependent on \( V_{EB} \) and \( V_{EB} \) is the unknown quantity being numerically evaluated. This apparent dilemma is resolved in Section 4.2.4.

### 4.2.3 Collector Region Large-Signal Model and the Collector Simulation Subroutine CCP.

#### 4.2.3.1 Equations to be Solved and the Simulation Technique.

The collector simulation subroutine CCP consists of two parts: a large-signal simulation subprogram LSSP and a collector circuit computer subprogram CCCP. The iteration scheme is illustrated in Fig. 4.4. When CCP is called into operation, \( V_{CB}, J_n B, V_{bias}, Z_L \) and \( C_e \) are known quantities where

\[ V_{CB} = V_{bias} + (I_{ET} - I_c)R_B \quad (4.24) \]
FIG. 4.1 THE COLLECTOR ITERATION SCHEME. (a) $I_T$ IS CALCULATED BY USING LSSP WITH $V_T$ KNOWN AND

(b) $I_{CC}$ AND $V_T$ ARE CALCULATED, WITH $V_{CB'}$, $I_T$, $C_c$ AND $Z_L$ KNOWN, BY USING CCCP.
and

\[ J_{nB} = \text{base transport factor} \times \frac{J_{nE}}{L_n} = \text{sech} \frac{V_B}{L_n} \times J_{nE}. \]  \hspace{1cm} (4.25)

The collector depletion capacitance\(^{12}\) is given by

\[ C = \left( C_i + \frac{R_b}{R_b + R_{bb}}, \frac{C_b}{(\text{number of emitter-base finger pairs})} \right). \]  \hspace{1cm} (4.26)

where

\[ R_b = \frac{d_{RL}}{2x_B \sigma_B}, \]

\[ R_{bb} = \frac{w_{B}}{12x_B \sigma_B}, \]

\[ C_i = \frac{e}{w_T} w_k \]

and

\[ C_b = \frac{2e}{w_T} d_k. \]

The definitions of the device geometrical parameters are included in Fig. 3.2. The computer subprogram LSSP is used to calculate \( J_n \), with \( V_T \) and \( J_{nB} \) as known quantities. The computer subprogram CCCP is used to calculate \( V_T, J_C, J_{cc} \) with \( V_{CB}, J_{Tc}, C_c \) and \( Z_L \) as given quantities. Figure 4.5 shows the block diagram of collector simulation subroutine CCP. The assumption that \( V_{CB} \) and \( J_{nB} \) are known quantities appears
FIG. 4.5 BLOCK DIAGRAM OF LARGE-SIGNAL SUBROUTINE CCP.
artificial because $V_{CB}$ and $J_nB$ are $J_c$ dependent and $J_c$ is the quantity being numerically calculated. This dilemma is resolved in Section 4.2.4.

4.2.3b Subprogram LSSP. The subroutine program LSSP is based on a device simulation program developed at the Electron Physics Laboratory written from the point of view of computational efficiency by Bauhahn and Haddad. The subprogram LSSP obtains a spatially one-dimensional solution of the continuity equations, Poisson's equation, and the current expressions obtained from a first-order solution of the Boltzmann transport equation. Figure 4.6 presents the basic flow of the simulation. If the electric field, carrier concentrations, $V_T$ and $J_T$ are known at a particular time $t$, Fig. 4.6 illustrates how the electric field, carrier concentrations, $V_T$ and $J_T$ at $t + \Delta t$ are obtained. Note that currents are defined to be positive in the direction of the electric field. The simulation is voltage driven, therefore $V_T(t + \Delta t)$ must be given. As shown in Fig. 4.6, the first step is to solve the continuity equations and obtain $p(t + \Delta t)$ and $n(t + \Delta t)$ where $G$ is the generation rate ($C \cdot s^{-1} \cdot cm^{-3}$) and $x$ is the one-dimensional spatial coordinate ($cm$). Carrier concentrations at $t + \Delta t$ can be obtained because the derivatives of $p$ and $n$ with respect to time in the finite-difference approximations include $p(t + \Delta t)$ and $n(t + \Delta t)$ terms. The second step is using Poisson's equation to obtain $E(t + \Delta t)$ to within a constant of integration. This constant is determined by requiring that the integral of the electric field across the collector region equals the base-collector voltage $V_T$ at $t + \Delta t$. At this point, the carrier concentrations and the electric field at $t + \Delta t$ are known. In Step 3 the carrier velocities at $t + \Delta t$ are determined since the electric field at $t + \Delta t$ is known where $v_{ps,ns}$ are the saturated hole and electron
1. \[ e^{\frac{\partial p}{\partial t}} = G[E(x,t)] - \frac{\partial J_p(x,t)}{\partial x} \] (A)
\[ e^{\frac{\partial n}{\partial t}} = G[E(x,t)] + \frac{\partial J_n(x,t)}{\partial x} \] (B)

yields \( p(t + \Delta t) \), \( n(t + \Delta t) \).

2. \[ e^{\frac{\partial E(t + \Delta t)}{\partial x}} = e[p(x,t + \Delta t) - n(x,t + \Delta t) - N_A^n(x)] \]

subject to: \( V_T(t + \Delta t) = \int_0^{V_T} E(t + \Delta t) \, dx \)
yields \( E(t + \Delta t) \).

3. \[ v_{p,n}(t + \Delta t, x) = v_{p,s,n} \left\{ 1 - \exp \left[-E(x,t + \Delta t)\frac{v_{p,n}}{v_{p,s,n}} \right] \right\} \]
yields \( v_{p,n}(t + \Delta t, x) \).

4. \[ J_p(x,t + \Delta t) = e^p(x,t + \Delta t)v_p(x,t + \Delta t) - e^D_p \frac{\partial p(t + \Delta t)}{\partial x} \]
\[ J_n(x,t + \Delta t) = e^n(x,t + \Delta t)v_n(x,t + \Delta t) + e^D_n \frac{\partial n(t + \Delta t)}{\partial x} \]
\[ J_T(t + \Delta t) = \frac{1}{V_T} \int_0^{V_T} [J_p(t + \Delta t) + J_n(t + \Delta t)] \, dx \]
yields \( J_{p,n}(x,t + \Delta t) \).

5. \[ G[E(x,t + \Delta t)] = A_p \exp \left[-\frac{b_p}{E(x,t + \Delta t)} \right] |J_p(x,t + \Delta t)| \]
\[ + A_n \exp \left[-\frac{b_n}{E(x,t + \Delta t)} \right] |J_n(x,t + \Delta t)| \]

FIG. 4.6 EQUATIONS SOLVED AND SEQUENCE OF STEPS.

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velocities (cm-s\(^{-1}\)) and \(u_{p,n}\) are the hole and electron low-field mobilities (cm-V\(^{-1}\)-s\(^{-1}\)), respectively. Step 4 shows the expressions used to calculate the current densities at \(t + \Delta t\) where \(J_{p,n}\) are the hole and electron current densities and \(J_T\) is the collector terminal particle current density which is found by integrating the particle current densities throughout the collector region. It should be pointed out that \(J_T\) is not the total collector current density because it does not include the displacement current, which must be supplied externally. The total current density at any point \(x\) is a constant and can be written as follows:

\[
J_c = J_p(x) + J_n(x) + \varepsilon \frac{\partial E}{\partial t}.
\] (4.27)

Integrating from \(x = 0\) to \(x = w_T\) yields

\[
J_c = \frac{1}{w_T} \int_0^{w_T} [J_p(x) + J_n(x)] \, dx + \frac{\varepsilon}{w_T} \int_0^{w_T} E(x) \, dx
\]  
= \frac{1}{w_T} \int_0^{w_T} J_T(x) \, dx + \frac{\varepsilon}{w_T} \frac{3E}{\partial t}.
\] (4.28)

Step 5 is used to determine the generation rate \(G\) whose expression is given in Step 5 of Fig. 4.6. The values of \(a_n\) and \(a_p\) are determined experimentally and can be calculated by using Tables 2.1 and 2.2. After these five steps of calculation, everything about the collector region at \(t + \Delta t\) is known. A new driving voltage value at \(t + 2\Delta t\) is specified and the cycle is repeated.

### 4.2.3c The Difference Equations in Computer Subprogram LSSP

The difference equations used to solve the differential equations of
Fig. 4.5 are presented here. Figure 4.7 shows the space-time grid used for the differential equations where the index $K$ refers to the time step, the index $J$ refers to the carrier space step, and the index $I$ refers to the field space step.

The continuity equations are approximated by

$$
\frac{(p)^{K+1}_J - (p)^{K}_J}{\Delta t/e} = (g)^{K}_J - \frac{(j^p)^{K}_I - (j^p)^{K-1}_I}{\Delta x}
$$

(4.29)

and

$$
\frac{(n)^{K+1}_J - (n)^{K}_J}{\Delta t/e} = (g)^{K}_J + \frac{(j^p)^{K}_I - (j^p)^{K-1}_I}{\Delta x}
$$

(4.30)

where $\Delta t$ is the time step; $\Delta x$ is the space step; and the generation rate, assuming that only pure avalanche generation occurs in the collector region, is given by

$$
(g)^{K}_J = \frac{1}{2} (g)^{K}_I + \frac{1}{2} (g)^{K-1}_I
$$

(4.31)

where

$$
(g)^{K}_I = (a^p)^{K}_I | (j^p)^{K}_I| + (a^p)^{K}_I | (j^p)^{K}_I|
$$

(4.32)

The time step $\Delta t$ is limited in this case by the dielectric relaxation time $\tau_R$ which is given by

$$
\Delta t \leq \tau_R \frac{e(\nu_n + \nu_p)}{e(\nu_n + \nu_p)}
$$

(4.33)

where $\nu_n$ and $\nu_p$ are the electron and hole mobilities. Poisson's equation becomes

-133-
Fig. 4.7  Space-time mesh used to write the difference equations.

(a) Space mesh. (b) Space-time mesh.
where $E$ is the electric field and $N_D$ is the donor doping density in the collector region. The transport equations for electrons and holes become

$$
\frac{(E)^K_I - (E)^K_{I-1}}{\Delta x} = \frac{(p)^K_J - (n)^K_J + (N_D)_J}{\varepsilon/e},
$$

where $E$ is the electric field and $N_D$ is the donor doping density in the collector region. The transport equations for electrons and holes become

$$
(J_e)^K_I = e(p)^K_J (v_p)^K_I - e(D_e)^K_J \frac{(p)^{K+1}_J - (p)^K_J}{\Delta x}
$$

and

$$
(J_e)^K_I = e(n)^K_J (v_n)^K_I + e(D_n)^K_J \frac{(n)^{K+1}_J - (n)^K_J}{\Delta x}
$$

for $(E)^K_I > 0$ and

$$
(J_p)^K_I = e(p)^K_J (v_p)^K_I - e(D_p)^K_J \frac{(p)^{K+1}_J - (p)^K_J}{\Delta x}
$$

and

$$
(J_n)^K_I = e(n)^K_J (v_n)^K_I + e(D_n)^K_J \frac{(n)^{K+1}_J - (n)^K_J}{\Delta x}
$$

for $(E)^K_I \leq 0$, where $v_n$ and $v_p$ are given by the expressions

$$
(v_p)^K_I = v_{ps} \left[ 1 - \exp \left( - \frac{(E)^K_{1_p}}{v_{ps}} \right) \right]
$$

and

$$
(v_n)^K_I = v_{ns} \left[ 1 - \exp \left( - \frac{(E)^K_{1_n}}{v_{ns}} \right) \right],
$$

respectively, and $D_e$ and $D_p$ are the field-dependent electron and hole.

-135-
diffusion coefficients. The drift terms in the current density equations use "upwind" carriers and are explicit. The reason for the name upwind can be seen in Eqs. 4.35 through 4.38, i.e., in order to calculate the particle current at \((I,K)\), the particle concentration in the direction opposing the particle flow, in the upwind direction, is used. The diffusion terms are implicit. This completes the semiconductor difference equations.

The voltage constraint stated in Step 2 of Fig. 4.6,

\[
V_T(t + \Delta t) = \int_{0}^{V_T} E(t + \Delta t) \, dx ,
\]

(4.41)

can be expressed as follows:

\[
(V_T)^K = \Delta x \sum_{I=1}^{NSTEP} (E)^K_I ,
\]

(4.42)

where \(NSTEP \equiv \frac{A}{w_T/\Delta x}\) and the field index \(I = 1,2,\ldots,NSTEP\).

The electron current density at the base-collector metallurgical junction is assumed to be equal to the electron current density at the edge of the base-collector depletion region since the width of the base-collector depletion region into the base region is very small. Most of the carrier generation occurs in the depleted collector region. The electric field at the base-collector metallurgical junction is always above the electric field necessary to sustain electron and hole carrier saturation velocities. If the electron and hole carrier densities at the base-collector metallurgical junction are defined to be
\[ n_B(t) = \frac{J_{nB}(t)}{v_{ns}} \]

and

\[ p_B(t) = \frac{J_{pB}(t)}{v_{ps}} , \]

respectively, where \( v_{ns} \) and \( v_{ps} \) are the electron and hole carrier saturation velocities and the avalanche multiplication produced hole current \( J_{pB} \) is calculated by the subprogram LSSP, the actual current densities at this boundary will be significantly lower than \( J_{nB} \) and \( J_{pB} \). The reason for this is that there are large magnitude electron and hole diffusion currents flowing in the opposite direction to the drift current components. Figure 4.8 contains the schematic distribution profiles of electron and hole densities in the depleted collector region. A set of "current conserving" boundary conditions will now be derived. When the electric field is pointing in the direction opposing that of the space coordinate \( x \), as in \( n \)-type CATT devices, the following is obtained:

\[
J_p = e p v_p + e D_p \frac{\partial p}{\partial x} 
\]

and

\[
J_n = e n v_n - e D_n \frac{\partial n}{\partial x} 
\]

which can be written as

\[
(J_p)^K = e (p^K (v)^K + e D^K (p) (p)^K - (p)^K) 
\]
COLLECTOR REGION

(a)

(b)

(c)

FIG. 4.8 (a) COLLECTOR REGION, (b) $|E(x)|$ PROFILE AND (c) ELECTRON AND HOLE CONCENTRATION PROFILE.
\[
\begin{align*}
(J_n)_K &= e(n)_K (v_n)_K - e(D)_K (n)_2 - (n)_1 \\
&= e(n)_K (v_n)_K - e(D)_K \left( \frac{(n)_2 - (n)_1}{\Delta x} \right).
\end{align*}
\]

The boundary conditions are obtained from these equations by rearranging the terms and they are

\[
n_B(t) = \frac{J_{nB}(t) + eD(n)_2/\Delta x}{ev_{ns} + eD/\Delta x}, \tag{4.43}
\]

and

\[
p_B(t) = \frac{J_{pB}(t) - eD(p)_2/\Delta x}{ev_{ps} - eD/\Delta x}, \tag{4.44}
\]

where \( n_B(t) \triangleq (n)_1 \), \( p_B(t) \triangleq (p)_1 \), \( J_{nB}(t) \triangleq (J_n)_1 \), \( J_{pB}(t) \triangleq (J_p)_1 \) and the indexes 1 and 2 represent the space point at the metallurgical junction and the space point next to the junction, respectively. The carrier velocities are assumed to be at the saturation values and the diffusion constants have the low field values.

2.3d Subprogram CCCP. The circuit problem to be solved by CCCP is shown in Fig. 4.4. The differential equations to be solved are

\[
\frac{dV_n}{dt} = V_n \tag{4.45}
\]

and
\[ \frac{dV_T'}{dt} = \frac{C_L}{C_c + C_L} \frac{d^2V_{CB}}{dt^2} + \frac{1}{C_c + C_L} \frac{1}{R_L} \frac{dV_{CB}}{dt} - \frac{1}{C_c + C_L} \frac{dV_T}{dt} \]

\[ + \frac{1}{C_c + C_L} \frac{1}{L_L} V_{CB} - \frac{1}{C_c + C_L} \frac{1}{R_L} V_T' - \frac{1}{C_c + C_L} \frac{1}{L_L} V_T', \quad (1.46) \]

and they are of the form

\[ \frac{df_1}{dt} = g(t,f_1,f_2) \]

and

\[ \frac{df_2}{dt} = g_2(t,f_1,f_2). \]

Their solutions can be obtained by using the fourth-order Runge-Kutta method.

4.2.4 Iteration Scheme Which Couples the Emitter-Base Region and the Collector Region Simulations. Previously, the dc...is of EBCP and CCP (LSSP and CCCP) were discussed. The iteration scheme which couples EBCP and CCP is described in Fig. 4.9 which is the block diagram of the complete large-signal computer simulation program.

4.3 Sample Results

Large-signal simulations using EBCP and CCP, determine \( I_P \), \( I_{CE} \), \( I_{CD} \), \( V_{EB} \), \( V_{EB} \), \( J \), \( I_{PB} \), \( I_{CE} \), \( I_c \), \( I_T \), and \( V_T \). From these results the CATT amplifier output RF power, power gain and efficiency can be determined. Figure 4.10a shows the waveforms of \( J \), \( J_T \), and \( V_T \) for the case in which the load is a high-Q tank circuit whose resonant frequency is tuned to the signal frequency. Figure 4.10b
FIG. 4.9 BLOCK DIAGRAM OF THE COMPLETE LARGE-SIGNAL
SIMULATION PROGRAM.
FIG. 4.10 (a) $J_{NB}$, $J_T$ AND $V_T$ WAVEFORMS AND (b) DISTRIBUTION
PLOTS OF ELECTRON DENSITY (MINUS SIGNS), HOLE DENSITY
(PLUS SIGNS), AND ELECTRIC FIELD PROFILE (SOLID LINES).
(xxx REPRESENTS THE REGION WHERE THE ELECTRIC FIELD
HAS REVERSED ITS DIRECTION)
shows the distribution plots of electron and hole densities and the electric field profile at various phase angles. At $\theta = 18$ degrees an insignificant number of electrons is injected into the collector region, at $\theta = 108$ degrees a large number of electrons is injected and they undergo avalanche multiplication, at $\theta = 144$ degrees electron carriers drift across the depleted collector region, and at $\theta = 288$ degrees most of the electron carriers have been collected at the collector contact.

### Improvements Over Previous Large-Signal Simulation of Class C CATT Amplifiers

The computer large-signal simulation program described in this chapter has eliminated the following inadequacies contained in the large-signal simulation by Yu et al.: 33

1. Emitter-base high-injection level effects were not incorporated.

2. Effects of high emitter doping level, minority carrier induced electric field in the base region, and a nonzero minority carrier density at the base-side edge of the base-collector depletion region were ignored.

3. The Early effect was ignored.

4. Diffusion currents in the depleted collector region were ignored.

5. A current nonconserving boundary condition for minority carriers in the depleted collector region was employed.

6. In determining the feedback hole current in an n-type CATT device, a time-averaged carrier multiplication factor was used which
is valid only if the emitter-base injected charge is very narrow and the carriers generated due to the base-collector reverse saturation current are negligible.

4.5 Conclusions

In this chapter, analytical expressions for the circuit model of the emitter-base region were derived. These expressions are valid at any injection level. The Early effect, high emitter doping density effect, and nonzero $n(v^3)$ effect were taken into consideration. The collector region is modeled by employing a difference-equations version of the semiconductor differential equations. Descriptions were given for the numerical methods used. Field dependences of charge carrier drift velocities, diffusion constants and ionization coefficients, and the space-charge effect were included in modeling the collector region. The iteration scheme which couples the emitter-base circuit model and the collector region is described. The iteration scheme is constructed in such a way that the time step size for EBCP, LSSP and CCCP can be independently different from each other. This feature is very convenient and can save much unnecessary simulation cost.
CHAPTER V. LARGE-SIGNAL STUDIES FOR CLASS C CATT AMPLIFIERS
AND COMPARISON WITH CLASS C BJT AMPLIFIERS

5.1 Introduction

In this chapter large-signal results of Class C CATT and BJT amplifiers are obtained from the computer simulation model developed in the previous chapter. This study is concerned with a series of X-band CATT devices with uniformly doped and HI-LO collector structures and a series of X-band BJTs with uniformly doped collector structures. The main distinction between the CATT devices and the BJTs is that the avalanche multiplication factor of the latter is always limited to 1.1 or less. The effects of impurity doping levels and widths of the base and collector regions and the optimum dc bias and load are investigated at various input power levels and operating frequencies. The Class C CATT and BJT amplifiers are compared on the basis of their efficiency, power gain, and the device inherent bandwidth.

5.2 Large-Signal Simulation Results for Class C CATT Amplifiers

5.2.1 General Discussion. The following study shows the effects of avalanche multiplication and wide collector regions. Large-signal simulations of Class C CATT amplifiers with various impurity doping levels and widths of the base and collector regions are carried out. The usefulness and the limitation of carrier multiplication and long collector transit angles are explored. The optimum values of the device parameters are determined. The amplifier inherent bandwidth, dynamic range, power gain, efficiency, optimum dc bias and optimum load are investigated.
The following large-signal simulations used the label DEV:material, type, letter, N\(c\) (cm\(^{-3}\)), \(w_T\) (\(\mu m\)) for devices with uniformly doped collectors and DEV:material, type, letter, \(N_{av}\) (cm\(^{-3}\)), \(w_{av}\) (\(\mu m\)), \(N_{drift}\) (cm\(^{-3}\)), \(w_D\) (\(\mu m\)) for devices with HI-LO collector regions, where the symbols \(N_c\), \(w_T\), \(N_{av}\), \(w_{av}\), \(N_{drift}\) and \(w_D\) have been defined previously. Therefore, DEV:Si,n,A,2 \(\times\) \(10^{16}\),1,1,1 \(\times\) \(10^{15}\),2,9 is the label used for a Si device which has an emitter-base structure A and a collector structure whose avalanche multiplication region is impurity doped at 2 \(\times\) \(10^{16}\) cm\(^{-3}\) and is 1.1 \(\mu m\) wide, and whose drift region is impurity doped at 1 \(\times\) \(10^{15}\) cm\(^{-3}\) and is 2.9 \(\mu m\) wide. The geometrical dimensions and the impurity doping levels of the emitter-base structures simulated are listed in Table 5.1.

5.2.2 Optimum Load. The optimum load for a Class C CATT amplifier is the same as that for a Class C BJT amplifier. To obtain maximum RF power at the fundamental harmonic, the load should be a high \(Q\), parallel RLC tank circuit which, in combination with the collector cold capacitance \(C_c\), has a resonant frequency equal to that of the emitter-base driving signal. With such a load, a typical set of \(J_T(t)\), \(V_T(t)\) and the emitter-base injected charge carrier waveforms is shown in Fig. 5.1 where the effective dynamic multiplication factor \(M_A\) is defined as:

\[
M_A = \frac{\int_0^T J_T(t) \, dt}{\int_0^T \left[ J_{nB}(t) + J_{cns} + J_{cps} \right] \, dt}
\]

(5.1)

and \(\eta\) is the overall amplifier efficiency. The aforementioned optimum load condition can be derived as follows. The average output power is
<table>
<thead>
<tr>
<th>Emitter-Base Structure</th>
<th>$N_D$ (1/cm$^3$)</th>
<th>$W_{E_0}$ (um)</th>
<th>$N_A$ (1/cm$^3$)</th>
<th>$W_{B_0}$ (um)</th>
<th>$X_{B_1}$ (um)</th>
<th>$X_{B_2}$ (um)</th>
<th>$w$ (um)</th>
<th>$d$ (um)</th>
<th>$t$ (um)</th>
<th>$h$ (um)</th>
<th>Emitter-Base Finger Pairs</th>
<th>$A_E$ (cm$^2$)</th>
</tr>
</thead>
<tbody>
<tr>
<td>A</td>
<td>$1 \times 10^{19}$</td>
<td>0.3</td>
<td>$5 \times 10^{17}$</td>
<td>0.3</td>
<td>0.3</td>
<td>0.6</td>
<td>0.75</td>
<td>1.55</td>
<td>30</td>
<td>4</td>
<td>10</td>
<td>$2.25 \times 10^{-6}$</td>
</tr>
<tr>
<td>B</td>
<td>$5 \times 10^{19}$</td>
<td>0.4</td>
<td>$2.5 \times 10^{17}$</td>
<td>0.25</td>
<td>0.25</td>
<td>0.65</td>
<td>0.6</td>
<td>0.2</td>
<td>20</td>
<td>4</td>
<td>14</td>
<td>$1.68 \times 10^{-6}$</td>
</tr>
</tbody>
</table>
FIG. 5.1 (a) $J_T$ AND $V_T$ WAVEFORMS WHEN $R_L$ IS OPTIMUM

($G_p = 9.74$ dB, $n = 55$ PERCENT, $V_{RF} = 31.3$ V
AND $M_A = 1.036$). (b) $J_T$ AND $V_T$ WAVEFORMS WHEN

$R_L$ IS GREATER THAN THE OPTIMUM VALUE ($G_p = 9.15$ dB, $n = 52$ PERCENT, $V_{RF} = 33$ V, $M_A = 1.053$). (DEV:Si.n.B.6 x 10$^{15}$,2; $f = 12.75$ GHz;

$v_{sig}(t) = 1.1 \sin \omega t - 0.191 V$; $V_{bias} = 40$ V)
given by

\[
P_{\text{out}} = \frac{1}{T} \int_{0}^{T} I_c(t)V_T(t) \, dt , \tag{5.2}
\]

where \( T \) is the RF period. Since \( I_c(t) = I_{1c}(t) + C_c(dV_c/dt) \),

\[
P_{\text{out}} = \frac{1}{T} \int_{0}^{T} I_{1c}(t)V_T(t) \, dt . \tag{5.3}
\]

If \( I_{1c}(t) \) and \( V_T(t) \) are expressed in terms of their harmonic components, Eq. 5.3 becomes

\[
P_{\text{out}} = \frac{1}{T} \left[ \sum_{n=1}^{\infty} I_{1c} V_{Tn} \cos \theta_n + \frac{I_{1c} V_T}{2} \cos \theta + \frac{I_{1c} V_T}{2} \cos 2\theta + \ldots + \frac{I_{1c} V_T}{n} \cos \theta_n + \ldots \right] , \tag{5.4}
\]

where \( \theta_n \) is the phase difference between the \( n \)th harmonic components of \( I_{1c}(t) \) and \( V_T(t) \). Under optimum load conditions, current component \( I_{1c} \) sees a nearly pure resistive load and all higher harmonics of \( I_{1c}(t) \) see nearly a zero impedance. Therefore, \( \theta_n \) is approximately \( \pi \) rad and \( V_{Tn} = 0 \) V for \( n \geq 2 \).

The resistor \( R_L \) of the parallel RLC load should be such that an optimum \( V_T(t) \) is produced. For a given dc bias, if \( R_L \) is below the optimum value, a small amplitude \( V_T \) is produced which means less output power. If \( R_L \) is too large, an overly large \( V_T \) will be impressed across the collector region which will cause a decrease in both the amplifier efficiency and the available power to the load. This may be due to two different reasons or to a combination of the two depending on the dc bias. First, an extremely large-amplitude RF voltage causes the electric field in significant portions of the drift region to be depressed below
that necessary to sustain carriers at a saturated velocity during a significant portion of their transit across the drift region. Two consequences follow which reduce the efficiency and the output power. The \( J_T(t) \) and \( V_T(t) \) waveforms corresponding to this situation are shown in Fig. 5.1. The slowing down of charge carriers caused the large depression in the \( J_m(t) \) waveform which is highly detrimental to the efficiency and power output since the depression occurs near the minimum of \( V_T(t) \). The slowing down of the charge carriers also caused an expansion in the width of the \( J_m(t) \) waveform which means lower efficiency. The reduction in efficiency is more severe at higher collector transit angles due to the functional dependence of efficiency on \( \sin \left( \frac{\theta_T}{2} \right) \) where \( \theta_T = \frac{\omega_T}{\nu_s} \). If \( \theta_T = \pi \) rad, an expansion in the width of the \( J_m(t) \) waveform could mean that large conduction current would be flowing during the positive half-cycle of the nearly sinusoidal \( V_T(t) \) which means energy dissipation. Second, having a large-amplitude RF voltage impressed across the collector causes the electric field in the avalanche multiplication region to become extremely high when \( V_T(t) \) is maximum so that a significant pulse of charge, produced by the avalanche multiplication of the thermally generated collector reverse saturation current, is injected into the drift region. This pulse of charge is injected much earlier than the emitter-base injected pulse of charge. Thus, the effective width of the \( J_T(t) \) waveform is significantly increased and therefore the efficiency is reduced drastically. The \( J_T \) and \( V_T \) waveforms corresponding to the second case are shown in Fig. 5.2.

5.2.3 DC Bias. It was mentioned previously that a high-efficiency, high-gain CATT amplifier requires a dc bias such that both a
FIG. 5.2 (a) $J_T$ AND $V_T$ WAVEFORMS WHEN $R_L$ IS OPTIMUM

($G_p = 10.75 \, \text{dB}$, $\eta = 43 \, \text{PERCENT}$, $V_{RF} = 27.2 \, \text{V}$

AND $M_A = 1.107$). (b) $J_T$ AND $V_T$ WAVEFORMS WHEN

$R_L$ IS GREATER THAN THE OPTIMUM VALUE ($G_p = 7.37 \, \text{dB}$,

$\eta = 20.9 \, \text{PERCENT}$, $V_{RF} = 26 \, \text{V}$ AND $M_A = 1.67$).

(DEV:Si,n,B,6 x $10^{15}$,2; $f = 12.75 \, \text{GHz}$; $V_{\text{sig}}(t) =$

$1.1 \sin \omega t - 0.191 \, V$; $V_{\text{bias}} = 45 \, \text{V}$)
large-amplitude $V_T(t)$ and a significant multiplication factor occur. The effects of the dc bias on the amplifier performance can be illustrated by the results given in Table 5.2. Figure 5.3 shows the optimum $J_T$ and $V_T$ waveforms of a typical 4-μm CATT device at different dc biases. The operating frequency is 12.75 GHz. The optimum dc bias is 65 V. At a lower dc bias, i.e., 60 V, both the allowed amplitude of $V_T(t)$ and the multiplication factor are reduced. The output power, power gain and efficiency are also decreased. At a dc bias higher than the optimum value, i.e., 80 V, although the multiplication factor is higher, the allowed amplitude of $V_T(t)$ decreases drastically which leads to lower output power, lower power gain and lower efficiency. The drastic reduction of the amplitude of $V_T$ is due to the fact that when the dc bias is significantly higher than the optimum value, a large-amplitude $V_T(t)$ would cause an injection of a charge pulse at a phase angle much earlier than the emitter-base injected signal charge. This prematurely injected pulse of charge is the result of avalanche multiplication of the thermally generated collector reverse saturation current under an extremely high electric field. The effective width of $J_T(t)$ widens, therefore, efficiency decreases. The device also loses its controlled avalanche characteristics.

When the emitter-base driving signal is varied in its amplitude, the amount of charge carriers injected into the collector region varies as well. It was found that as the amplitude of the driving signal increases the optimum dc bias for optimum gain and efficiency decreases slightly. This is due to the space-charge effect of the injected charge carriers from the emitter-base junction. A larger number of injected carriers results in lower carrier multiplication, and the
**Table 5.2**

Effects of DC Bias on Amplifier Performance

DEV: Si, n, A, 2.2 x 10^16, 1.1, 2 x 10^15, 2.9; \( V_{\text{sig}} = 1.15 \sin \omega t - 0.2 \) V; \( f = 12.75 \) GHz

<table>
<thead>
<tr>
<th>( V_{\text{bias}} ) (V)</th>
<th>RF Voltage Amplitude (V)</th>
<th>( I_A ) (A)</th>
<th>( P_{\text{out}} ) (W/cm²)</th>
<th>( \eta ) (Percent)</th>
</tr>
</thead>
<tbody>
<tr>
<td>57.5</td>
<td>12</td>
<td>6.25</td>
<td>7.5 x 10³</td>
<td>10</td>
</tr>
<tr>
<td>55</td>
<td>17</td>
<td>5.65</td>
<td>8.5 x 10³</td>
<td>16.5</td>
</tr>
<tr>
<td>52.5</td>
<td>24</td>
<td>5.5</td>
<td>8.5 x 10³</td>
<td>25</td>
</tr>
<tr>
<td>50</td>
<td>28.5</td>
<td>5.5</td>
<td>8.5 x 10³</td>
<td>28</td>
</tr>
<tr>
<td>42.5</td>
<td>22.5</td>
<td>4.6</td>
<td>5.25 x 10³</td>
<td>22.5</td>
</tr>
</tbody>
</table>

DEV: Si, n, B, 6 x 10^15, h; \( V_{\text{sig}} = 1.1 \sin \omega t - 0.191 \) V; \( f = 12.75 \) GHz

<table>
<thead>
<tr>
<th>( V_{\text{bias}} ) (V)</th>
<th>RF Voltage Amplitude (V)</th>
<th>( I_A ) (A)</th>
<th>( P_{\text{out}} ) (W/cm²)</th>
<th>( \eta ) (Percent)</th>
<th>( S_P ) (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>80</td>
<td>12.5</td>
<td>2.3</td>
<td>2.4 x 10⁴</td>
<td>10</td>
<td>11.77</td>
</tr>
<tr>
<td>73</td>
<td>30.1</td>
<td>1.83</td>
<td>2.98 x 10⁴</td>
<td>25</td>
<td>14.13</td>
</tr>
<tr>
<td>65</td>
<td>42.4</td>
<td>1.75</td>
<td>2.96 x 10⁴</td>
<td>38.1</td>
<td>15.24</td>
</tr>
<tr>
<td>60</td>
<td>38.7</td>
<td>1.71</td>
<td>2.63 x 10⁴</td>
<td>34.1</td>
<td>14.58</td>
</tr>
<tr>
<td>55</td>
<td>31.8</td>
<td>1.43</td>
<td>2.09 x 10⁴</td>
<td>32.9</td>
<td>13.22</td>
</tr>
<tr>
<td>35</td>
<td>12.9</td>
<td>1.09</td>
<td>6.197 x 10³</td>
<td>21.4</td>
<td>8.05</td>
</tr>
</tbody>
</table>

(Cont.)
Table 5.2 (Cont.)

\[ V_{\text{sig}} = 1.1 \sin \omega t - 0.191 \text{ V}; f = 12.75 \text{ GHz} \]

<table>
<thead>
<tr>
<th>( V_{\text{bias}} ) (V)</th>
<th>RF Voltage Amplitude (V)</th>
<th>( M_A )</th>
<th>( P_{\text{out}} ) (W/cm²)</th>
<th>( n ) (Percent)</th>
<th>( G_p ) (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>55</td>
<td>17.4</td>
<td>3.1</td>
<td>2.985 \times 10^4</td>
<td>20.7</td>
<td>12.99</td>
</tr>
<tr>
<td>52.5</td>
<td>26.4</td>
<td>2.76</td>
<td>3.47 \times 10^4</td>
<td>26.8</td>
<td>14.55</td>
</tr>
<tr>
<td>50</td>
<td>31</td>
<td>2.69</td>
<td>3.38 \times 10^4</td>
<td>25.1</td>
<td>14.86</td>
</tr>
<tr>
<td>47.5</td>
<td>27.8</td>
<td>2.41</td>
<td>2.7 \times 10^4</td>
<td>22.6</td>
<td>13.73</td>
</tr>
<tr>
<td>45</td>
<td>24.4</td>
<td>2.13</td>
<td>2.088 \times 10^4</td>
<td>20</td>
<td>12.43</td>
</tr>
<tr>
<td>40</td>
<td>15.5</td>
<td>1.53</td>
<td>1.23 \times 10^4</td>
<td>17.1</td>
<td>9.65</td>
</tr>
</tbody>
</table>
FIG. 5.3 $J_T$ AND $V_T$ OF A TYPICAL $4 \times 10^{-4}$ cm COLLECTOR

DEVICE OPERATING AT 12.75 GHz AND MAXIMUM ALLOWED
RF VOLTAGE AMPLITUDE. (a) DC BIAS BELOW THE
OPTIMUM VALUE, (b) DC BIAS ABOVE THE OPTIMUM VALUE,
AND (c) AT OPTIMUM DC BIAS. (DEV:Si,n,B,6 x $10^{15}$,4)
importance of carrier multiplication in determining the amplifier gain and output power of a given CATT device at the optimum operating condition relative to that of the RF voltage amplitude is reduced. The variation in optimum dc bias as a function of the emitter-base driving signal level is illustrated in Table 5.3. The space-charge effect is illustrated in Figs. 5.4 and 5.5. Plots of electric field profile, electron and hole distributions at various phase angles, and $J_T$ and $V_T$ waveforms corresponding to two different emitter-base signal levels are shown. The device is DEV:Si,n,B,6 x 10$^{15}$,4.

The profiles of the electric field at various phase angles explain the reduction of $M_A$ as the driving signal level is increased. It is also observed that $V_{sus}$, the minimum allowed $V_T$, is lower due to space charge and therefore the maximum allowed RF voltage amplitude increases slightly with increasing $V_{sig}$, assuming $V_{bias}$ is kept constant.

The optimum dc bias also varies with the operating frequency. It was found that, for a given device, the optimum dc bias increases with increasing operating frequency. This is because as the operating frequency increases the transit angle increases; therefore, charge carriers are injected into the depleted collector earlier in a given cycle when the value of $V_A(t)$ is higher. Avalanche multiplication becomes a more important factor in determining the gain and the RF output power. Avalanche multiplication is a sensitive function of the electric field. The variation of optimum dc bias with operating frequency is illustrated by the results tabulated in Table 5.4.

5.2.4 Effects of $N_{av}$, $w_{av}$, $N_{drift}$ and $w_D$ in HI-LO Collector Structures. The effects of varying $N_{av}$, $w_{av}$, $N_{drift}$ and $w_D$ are examined in this section. The results given in Table 5.5 give the
Table 5.3

Variation of Optimum Dc Bias with Emitter-Base Signal Level

DEV:Si,n,B,1 x 10^{16},h; f = 12.75 GHz

<table>
<thead>
<tr>
<th>$V_{\text{sig}}$ (V)</th>
<th>Optimum Dc Bias (V)</th>
<th>RF Voltage Amplitude (V)</th>
<th>$I_A$</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.05 sin $\omega t$ - 0.1823</td>
<td>55</td>
<td>26.3</td>
<td>3.98</td>
</tr>
<tr>
<td>1.10 sin $\omega t$ - 0.191</td>
<td>51.5</td>
<td>29</td>
<td>2.71</td>
</tr>
<tr>
<td>1.15 sin $\omega t$ - 0.199</td>
<td>51</td>
<td>31</td>
<td>1.75</td>
</tr>
<tr>
<td>1.20 sin $\omega t$ - 0.2083</td>
<td>48.5</td>
<td>34</td>
<td>1.57</td>
</tr>
<tr>
<td>1.25 sin $\omega t$ - 0.217</td>
<td>46</td>
<td>38.1</td>
<td>1.45</td>
</tr>
</tbody>
</table>

DEV:Si,n,B,6 x 10^{15},h; f = 12.75 GHz

<table>
<thead>
<tr>
<th>$V_{\text{sig}}$ (V)</th>
<th>Optimum Dc Bias (V)</th>
<th>RF Voltage Amplitude (V)</th>
<th>$I_A$</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.05 sin $\omega t$ - 0.1823</td>
<td>70</td>
<td>40</td>
<td>2.55</td>
</tr>
<tr>
<td>1.10 sin $\omega t$ - 0.191</td>
<td>65</td>
<td>42.4</td>
<td>1.75</td>
</tr>
<tr>
<td>1.15 sin $\omega t$ - 0.199</td>
<td>64</td>
<td>49.6</td>
<td>1.57</td>
</tr>
<tr>
<td>1.2 sin $\omega t$ - 0.2084</td>
<td>63.5</td>
<td>50</td>
<td>1.37</td>
</tr>
</tbody>
</table>
FIG. 5.4 (a) ELECTRIC FIELD PROFILE AND ELECTRON AND HOLE DISTRIBUTIONS AT VARIOUS PHASE ANGLES. (b) $J_T$ AND $V_T$ WAVEFORMS. ($\text{DEVICES}: \text{Si}, n, B, 6 \times 10^{15}$, $f = 12.75$ GHz; $V_{\text{bias}} = 65$ V; $V_{\text{sig}} = 1.05 \sin \omega t - 0.1823$ V; and $\times \times \times$ REPRESENTS THE COLLECTOR REGION WHERE THE ELECTRIC FIELD IS NEGATIVE)
Fig. 5.5 (a) Electric field profile and electron and hole distributions at various phase angles. (b) $J_T$ and $V_T$ waveforms. (Dev: Si, $nB_i 6 \times 10^{15}$/cm$^3$; $f = 12.75$ GHz; $V_{bias} = 65$ V; $V_{sig} = 1.2 \sin \omega t - 0.2083$ V; and ** represents the collector region where the electric field is negative).
Table 5.4
Variation of Optimum Dc Bias with Operating Frequency
(DEV: Si,n,B,2.2 x 10^{16},1.1,2 x 10^{15},2.9)

\[ V_{\text{sig}} = 1.1 \sin \omega t - 0.191 \, V \]

<table>
<thead>
<tr>
<th>( f ) (GHz)</th>
<th>Optimum Dc Bias (V)</th>
</tr>
</thead>
<tbody>
<tr>
<td>7</td>
<td>50</td>
</tr>
<tr>
<td>10</td>
<td>52.5</td>
</tr>
<tr>
<td>12.75</td>
<td>52.5</td>
</tr>
<tr>
<td>14</td>
<td>55</td>
</tr>
<tr>
<td>18</td>
<td>57</td>
</tr>
<tr>
<td>21</td>
<td>57.5</td>
</tr>
</tbody>
</table>

\[ V_{\text{sig}} = 1.05 \sin \omega t - 0.1823 \, V \]

\[ V_{\text{sig}} = 1.0 \sin \omega t - 0.1736 \, V \]

\[ V_{\text{sig}} = 1.05 \sin \omega t - 0.1823 \, V \]

\[ V_{\text{sig}} = 1.0 \sin \omega t - 0.1736 \, V \]

<table>
<thead>
<tr>
<th>( f ) (GHz)</th>
<th>Optimum Dc Bias (V)</th>
</tr>
</thead>
<tbody>
<tr>
<td>7</td>
<td>50</td>
</tr>
<tr>
<td>12.75</td>
<td>52.5</td>
</tr>
<tr>
<td>21</td>
<td>57.5</td>
</tr>
<tr>
<td>21</td>
<td>55</td>
</tr>
</tbody>
</table>
Table 5.5

Effects of $N_{av}$ on the Performance of Class C CATT Amplifiers

DEV:Si,n,A,$N_{av}$,1,1,2x10$^{15}$,2,9; $V_{\text{sig}}(t) = 1.15 \sin \omega t - 0.2V$; $f = 12.75$ GHz

<table>
<thead>
<tr>
<th>$N_{av}$ (1/cm$^3$)</th>
<th>Optimum Dc Bias (V)</th>
<th>$M_A$</th>
<th>RF Voltage Amplitude (V)</th>
<th>$n$ (Percent)</th>
<th>$G_p$ (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>$1.6x10^{16}$</td>
<td>67.5</td>
<td>4.1</td>
<td>41</td>
<td>33</td>
<td>2.69</td>
</tr>
<tr>
<td>$1.8x10^{16}$</td>
<td>62.5</td>
<td>4.4</td>
<td>36</td>
<td>32</td>
<td>3.83</td>
</tr>
<tr>
<td>$2x10^{16}$</td>
<td>57.5</td>
<td>5.5</td>
<td>33.5</td>
<td>30.5</td>
<td>4.1</td>
</tr>
<tr>
<td>$2.2x10^{16}$</td>
<td>50</td>
<td>5.5</td>
<td>27.5</td>
<td>28</td>
<td>4.62</td>
</tr>
<tr>
<td>$2.4x10^{16}$</td>
<td>45</td>
<td>5.7</td>
<td>21</td>
<td>24</td>
<td>4.16</td>
</tr>
<tr>
<td>$2.6x10^{16}$</td>
<td>40</td>
<td>6.35</td>
<td>15.7</td>
<td>18.5</td>
<td>3.33</td>
</tr>
</tbody>
</table>

DEV:Si,n,B,$N_{av}$,0.6,2x10$^{15}$,3.4; $V_{\text{sig}}(t) = 1.1 \sin \omega t - 0.191V$; $f = 12.75$ GHz

<table>
<thead>
<tr>
<th>$N_{av}$ (1/cm$^3$)</th>
<th>Optimum Dc Bias (V)</th>
<th>$M_A$</th>
<th>RF Voltage Amplitude (V)</th>
<th>$n$ (Percent)</th>
<th>$G_p$ (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>$1x10^{16}$</td>
<td>75</td>
<td>1.2</td>
<td>43</td>
<td>32</td>
<td>13.2</td>
</tr>
<tr>
<td>$2.5x10^{16}$</td>
<td>67.5</td>
<td>1.5</td>
<td>50</td>
<td>41</td>
<td>15.1</td>
</tr>
<tr>
<td>$3x10^{16}$</td>
<td>66.5</td>
<td>1.82</td>
<td>52.7</td>
<td>39.5</td>
<td>15.6</td>
</tr>
<tr>
<td>$4x10^{16}$</td>
<td>57.5</td>
<td>2.2</td>
<td>40.5</td>
<td>30</td>
<td>15.2</td>
</tr>
</tbody>
</table>

(Cont.)
Table 5.5 (Cont.)

\( V_{\text{sig}}(t) = 1.1 \sin \omega t - 0.191 \, V; \) 
\( f = 12.75 \, \text{GHz} \)

<table>
<thead>
<tr>
<th>( N_{av} (1/\text{cm}^3) )</th>
<th>Optimum DC Bias (V)</th>
<th>M_A</th>
<th>RF Voltage Amplitude (V)</th>
<th>( n ) (Percent)</th>
<th>( G_p ) (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>( 1 \times 10^{16} )</td>
<td>75</td>
<td>1.3</td>
<td>45.5</td>
<td>33</td>
<td>13.9</td>
</tr>
<tr>
<td>( 1.75 \times 10^{16} )</td>
<td>70</td>
<td>1.6</td>
<td>50</td>
<td>38.5</td>
<td>15.2</td>
</tr>
<tr>
<td>( 2 \times 10^{16} )</td>
<td>65</td>
<td>1.7</td>
<td>46.5</td>
<td>39</td>
<td>15.1</td>
</tr>
<tr>
<td>( 2.25 \times 10^{16} )</td>
<td>63.5</td>
<td>1.8</td>
<td>44.2</td>
<td>38</td>
<td>15.4</td>
</tr>
<tr>
<td>( 2.5 \times 10^{16} )</td>
<td>62</td>
<td>2</td>
<td>42</td>
<td>34.5</td>
<td>15.4</td>
</tr>
<tr>
<td>( 3.5 \times 10^{16} )</td>
<td>47.5</td>
<td>2.42</td>
<td>25</td>
<td>27.5</td>
<td>12.7</td>
</tr>
</tbody>
</table>

\( 5 \times 10^{15} \) 80  1.35  46.5  31.5  13.8
\( 1 \times 10^{16} \) 72.5 1.5  49.2  37.5  14.8
\( 1.5 \times 10^{16} \) 63.5 1.95  41  34.5  15.3
\( 1.75 \times 10^{16} \) 60  2.15  35  28.5  15
\( 2 \times 10^{16} \) 53.5  2.4  30.5  26  13.8
optimum performances of CATT devices with different $N_{av}$ and $w_{av}$.

For a fixed $w_{av}$ there is a corresponding optimum $N_{av}$. For devices with $N_{av}$ higher than the optimum value, larger avalanche multiplication is achieved at the device optimum dc bias, but the amplitude of $V_T$ is decreased. The resultant efficiency, output power and power gain are lowered. If $N_{av}$ is lower than the optimum value, although the amplitude of $V_T$ may be slightly higher, the avalanche multiplication is lower.

For devices with wider avalanche multiplication regions, the optimum $N_{av}$ would be lower in order to maintain the device capability for large amplitude $V_T$. As long as the avalanche multiplication region is not too narrow or too wide, i.e., 0.6 to 1.2 μm, there is no significant difference in the performance of devices with optimized $N_{av}$.

The effects of drift region width and its doping level have also been examined. Large-signal simulations have been carried out for devices with emitter-base structure A. The operating frequency is kept at 12.75 GHz. The results are summarized in Table 5.6. The results indicate that the longer devices have higher optimum dc biases, higher multiplication and lower efficiencies. Higher multiplication is due to a longer collector transit angle which implies an earlier injection of carriers in a given cycle when the value of $V_T(t)$ is higher. Lower efficiency is due to its dependence on $\sin(\theta_{n}/2)/(\theta_{T}/2)$. The amplitude of $V_T$, output power and power gain increase with increasing $w_D$ until $w_T$ is approximately 5.5 μm. At $f = 12.75$ GHz, the $w_T$, which corresponds to a π-rad collector transit angle is 4 μm. It was indicated previously that a collector transit angle greater than π rad implies a large current flow during portions of the positive half-cycle of $V_T(t)$ and, therefore, a large amount of energy dissipated. However, due to larger carrier
Table 5.6

Effects of Drift Region Width

DEV:Si,n,A,2.2x10^{16},1.1,2x10^{15},w_D; V_{\text{sig}}(t) = 1.15 \sin \omega t - 0.2; 
f = 12.75 \text{ GHz}

<table>
<thead>
<tr>
<th>$w_D$ ((\mu m))</th>
<th>Optimum Dc Bias (V)</th>
<th>n (Percent)</th>
<th>$M_A$</th>
<th>RF Voltage Amplitude (V)</th>
</tr>
</thead>
<tbody>
<tr>
<td>2.1</td>
<td>45</td>
<td>30</td>
<td>3.7</td>
<td>22.5</td>
</tr>
<tr>
<td>2.4</td>
<td>47.5</td>
<td>29.5</td>
<td>4.3</td>
<td>24.6</td>
</tr>
<tr>
<td>2.7</td>
<td>50</td>
<td>28</td>
<td>5.2</td>
<td>26</td>
</tr>
<tr>
<td>2.9</td>
<td>52.5</td>
<td>27.5</td>
<td>5.9</td>
<td>27</td>
</tr>
<tr>
<td>3.15</td>
<td>52.5</td>
<td>27</td>
<td>6.4</td>
<td>29</td>
</tr>
<tr>
<td>3.4</td>
<td>55</td>
<td>26.5</td>
<td>7.3</td>
<td>30.5</td>
</tr>
<tr>
<td>3.65</td>
<td>57.5</td>
<td>24.5</td>
<td>8.5</td>
<td>32.5</td>
</tr>
<tr>
<td>3.9</td>
<td>60</td>
<td>21.5</td>
<td>9.6</td>
<td>33</td>
</tr>
<tr>
<td>4.4</td>
<td>65</td>
<td>15</td>
<td>11.3</td>
<td>33.2</td>
</tr>
<tr>
<td>4.65</td>
<td>67.5</td>
<td>13.5</td>
<td>11</td>
<td>30</td>
</tr>
<tr>
<td>5.15</td>
<td>70</td>
<td>8.25</td>
<td>10</td>
<td>29</td>
</tr>
</tbody>
</table>
multiplication and RF voltage amplitude, the optimum \( w_T \) for maximum output power and power gain is 5.5 \( \mu \text{m} \). As \( w_T \) is increased beyond 5.5 \( \mu \text{m} \), efficiency becomes very low and the maximum allowed RF voltage amplitude starts to decrease. The decrease in RF voltage amplitude is due to the increase in \( V_{\text{sun}} \) with increasing \( w_T \) while the breakdown electric field intensity stays constant. If efficiency and power gain are both considered, the optimum \( w_T \) at 12.75 GHz is - 4.5 to - 4.75 \( \mu \text{m} \).

The drift region doping density should be as low as possible to minimize \( V_{\text{sun}} \), i.e., maximize the allowed RF voltage. But it should be sufficient to sustain the charge carrier density. Otherwise, the space-charge effect will force the electric field behind the charge pulse in the drift region to be depressed below the charge carrier saturation drift velocity sustaining value. As a consequence, charge carriers at the tail end of the pulse will be slowed down and the charge pulse will spread out which, in turn, will cause a decrease in the fundamental harmonic component of \( I_T \). This is one aspect of the deterioration in amplifier power gain and efficiency due to space charge in a lightly doped drift region. Depression of electric field behind the charge pulse, when severe enough, will bring a reduction in the allowed maximum RF voltage amplitude. It can be seen from large-signal simulation results that at a low input signal level the device with the smallest \( N_{\text{drift}} \) has the largest allowed RF voltage amplitude. As the input signal level increases the allowed maximum RF voltage amplitude of the device with the smallest \( N_{\text{drift}} \) decreases noticeably. This is the other aspect of the deterioration in amplifier power gain.
power output and efficiency due to space charge in the lightly doped drift region.

Table 5.7 summarizes the results of large-signal simulations of devices with various values of $N_{\text{drift}}$. The electric field depression, charge pulse spreading and reduction in $V_{\text{RF}}$ due to space charge are illustrated in Figs. 5.6 through 5.9. It is also observed that space charge in a lightly doped drift region may force up the electric field intensity near the collector contact and a significant number of electron-hole pairs will be created. This will further expand the duration of the induced current, increase power dissipation and reduce the efficiency.

5.2.5 Uniformly Doped Collector Structures. Large-signal simulations of devices with uniformly doped collectors have been carried out. The results are summarized in Table 5.8. These results correspond to the operating condition when both the power gain and the efficiency are optimized. The following observations can be made:

1. Shorter devices have higher efficiencies. The higher efficiency is due to a higher $V_{\text{RF}}/V_{\text{bias}}$ ratio and a smaller collector transit angle. At $f = 12.75 \text{GHz}$, however, efficiency reaches a saturation value when the collector region width is reduced to 2 $\mu$m and it will not increase significantly if $w_T$ is further reduced.

2. Longer devices have higher output power and power gain. The higher output power is due to a larger allowed RF voltage amplitude and a higher avalanche multiplication factor. The higher multiplication factor is due to a higher optimum dc bias and an earlier injection of charge carriers into the collector region relative to the $V_T$ waveform.
Table 5.7
Effects of $N_{\text{drift}}$

DEV: Si, n, B, 2.2 x 10^{16}, 1.1, 1.3 x 10^{15}, 2.9; f = 12.75 GHz; $V_{\text{bias}} = 45$ V

<table>
<thead>
<tr>
<th>$V_{\text{Sig}}$ (V)</th>
<th>$M_A$</th>
<th>RF Voltage Amplitude (V)</th>
<th>$\eta$ (Percent)</th>
<th>$G_{P}$ (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>sin wt - 0.1736</td>
<td>6.45</td>
<td>23.3</td>
<td>23.5</td>
<td>15.7</td>
</tr>
<tr>
<td>1.05 sin wt - 0.1823</td>
<td>3.5</td>
<td>24.3</td>
<td>27.5</td>
<td>14.9</td>
</tr>
<tr>
<td>1.1 sin wt - 0.191</td>
<td>2.35</td>
<td>24.7</td>
<td>26</td>
<td>13.7</td>
</tr>
<tr>
<td>1.15 sin wt - 0.2</td>
<td>1.98</td>
<td>24.3</td>
<td>20.2</td>
<td>12.4</td>
</tr>
<tr>
<td>1.2 sin wt - 0.2084</td>
<td>1.68</td>
<td>22</td>
<td>16</td>
<td>10.5</td>
</tr>
</tbody>
</table>

DEV: Si, n, B, 2.2 x 10^{16}, 1.1, 1.3 x 10^{15}, 2.9; f = 12.75 GHz; $V_{\text{bias}} = 47.5$ V

<table>
<thead>
<tr>
<th>$V_{\text{Sig}}$ (V)</th>
<th>$M_A$</th>
<th>RF Voltage Amplitude (V)</th>
<th>$\eta$ (Percent)</th>
<th>$G_{P}$ (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>sin wt - 0.1736</td>
<td>5.5</td>
<td>25</td>
<td>26</td>
<td>15.5</td>
</tr>
<tr>
<td>1.05 sin wt - 0.1823</td>
<td>3.4</td>
<td>27.5</td>
<td>27</td>
<td>14.9</td>
</tr>
<tr>
<td>1.1 sin wt - 0.191</td>
<td>2.45</td>
<td>28</td>
<td>24.5</td>
<td>13.9</td>
</tr>
<tr>
<td>1.15 sin wt - 0.2</td>
<td>1.96</td>
<td>27</td>
<td>18.38</td>
<td>12.2</td>
</tr>
<tr>
<td>1.2 sin wt - 0.2084</td>
<td>1.78</td>
<td>27.5</td>
<td>13.7</td>
<td>10.5</td>
</tr>
</tbody>
</table>

(Cont.)
### Table 5.7 (Cont.)

DEIV: Si$_n$B; $2.2 \times 10^{16}, 1.1, 2 \times 10^{15}, 2.9$; $f = 12.75$ GHz; $V_{bias} = 52.5$ V

<table>
<thead>
<tr>
<th>$V_{sig}$ (V)</th>
<th>$M_A$</th>
<th>RF Voltage Amplitude (V)</th>
<th>$n$ (Percent)</th>
<th>$G_p$ (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.9 sin $\omega t$ - 0.165</td>
<td>6.7</td>
<td>22.5</td>
<td>23.5</td>
<td>12.7</td>
</tr>
<tr>
<td>sin $\omega t$ - 0.1736</td>
<td>4.4</td>
<td>26.5</td>
<td>28</td>
<td>14.65</td>
</tr>
<tr>
<td>1.05 sin $\omega t$ - 0.1823</td>
<td>3</td>
<td>29.5</td>
<td>28</td>
<td>14.88</td>
</tr>
<tr>
<td>1.1 sin $\omega t$ - 0.191</td>
<td>2.4</td>
<td>30.2</td>
<td>23.4</td>
<td>14</td>
</tr>
<tr>
<td>1.15 sin $\omega t$ - 0.2</td>
<td>1.96</td>
<td>29.7</td>
<td>17</td>
<td>12.3</td>
</tr>
<tr>
<td>1.2 sin $\omega t$ - 0.2084</td>
<td>1.71</td>
<td>27.1</td>
<td>13.3</td>
<td>10.4</td>
</tr>
</tbody>
</table>

DEIV: Si$_n$B; $2.2 \times 10^{16}, 1.1, 7.5 \times 10^{14}, 2.9$; $f = 12.75$ GHz; $V_{bias} = 58.5$ V

<table>
<thead>
<tr>
<th>$V_{sig}$ (V)</th>
<th>$M_A$</th>
<th>RF Voltage Amplitude (V)</th>
<th>$n$ (Percent)</th>
<th>$G_p$ (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>sin $\omega t$ - 0.1736</td>
<td>4.37</td>
<td>33.8</td>
<td>30</td>
<td>14.96</td>
</tr>
<tr>
<td>1.05 sin $\omega t$ - 0.1823</td>
<td>3.02</td>
<td>32</td>
<td>27.5</td>
<td>15.06</td>
</tr>
<tr>
<td>1.1 sin $\omega t$ - 0.191</td>
<td>2.35</td>
<td>31.35</td>
<td>19.7</td>
<td>13.72</td>
</tr>
<tr>
<td>1.15 sin $\omega t$ - 0.2</td>
<td>1.79</td>
<td>24.5</td>
<td>14.6</td>
<td>11.2</td>
</tr>
<tr>
<td>1.2 sin $\omega t$ - 0.2084</td>
<td>1.67</td>
<td>25.3</td>
<td>10.6</td>
<td>9.64</td>
</tr>
</tbody>
</table>
FIG. 5.6 (a) $J_T$ AND $V_T$ WAVEFORMS AND (b) ELECTRIC FIELD PROFILE AND ELECTRON AND HOLE DISTRIBUTIONS AT VARIOUS PHASE ANGLES. (DEV: Si, n, B, 2.2 x $10^{16}$, 1.1, 7.5 x $10^{14}$, 2.9; $f = 12.75$ GHz; $V_{sig} = 1.1 \sin \omega t - 0.191 V$; AND xxx REPRESENTS THE REGION WHERE THE ELECTRIC FIELD IS NEGATIVE)
FIG. 5.7 (a) $J_T$ AND $V_T$ WAVEFORMS AND (b) ELECTRIC FIELD PROFILE AND ELECTRON AND HOLE DISTRIBUTIONS AT VARIOUS PHASE ANGLES. (DEV: $nB, 2.2 \times 10^{16}, 1.1, 7.5 \times 10^{14}, 2.9$; $f = 12.75$ GHz; $V_{sig} = 1.2 \sin \omega t - 0.208377$ V; AND *** REPRESENTS THE REGION WHERE THE ELECTRIC FIELD IS NEGATIVE)
FIG. 5.8 (a) $J_T$ AND $V_T$ WAVEFORMS AND (b) ELECTRIC FIELD PROFILE AND ELECTRON AND HOLE DISTRIBUTIONS AT VARIOUS PHASE ANGLES. (DEV: Si, n, B, $2.2 \times 10^{16}, 1.14 \times 10^{15}, 2.9$; $f = 12.75$ GHz; $V_{\text{sig}} = 1.1 \sin \omega t - 0.191$ V; AND $\cdots\cdots$ REPRESENTS THE REGION WHERE THE ELECTRIC FIELD IS NEGATIVE)
FIG. 5.9 (a) \( J_T \) AND \( V_T \) WAVEFORMS AND (b) ELECTRIC FIELD PROFILE AND ELECTRON AND HOLE DISTRIBUTIONS AT VARIOUS PHASE ANGLES. (DEV: Si, n, B, 2.2 x 10^{16}, 1.1, 1, 4 \times 10^{15}, 2.9; f = 12.75 \text{ GHz}; V_{\text{sig}} = 1.2 \sin \omega t - 0.208377 \text{ V}; \text{ AND } *** \text{ REPRESENTS THE REGION WHERE THE ELECTRIC FIELD IS NEGATIVE})
Table 5.8

Large-Signal Simulation Results of CATT Devices with Uniformly Doped Collector Structures

DEV:Si,n,B,N_s,5; f = 12.75 GHz; \( V_{\text{sig}}(t) = 1.1 \sin \omega t - 0.191 \) V

<table>
<thead>
<tr>
<th>( N_c (1/\text{cm}^3) )</th>
<th>Optimum DC Bias (V)</th>
<th>( M_A )</th>
<th>RF Voltage Amplitude (V)</th>
<th>( \eta ) (Percent)</th>
<th>( G_p ) (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>2\times10^{15}</td>
<td>80</td>
<td>1.12</td>
<td>38</td>
<td>22</td>
<td>11.8</td>
</tr>
<tr>
<td>6\times10^{15}</td>
<td>73.5</td>
<td>3.63</td>
<td>57.7</td>
<td>29.9</td>
<td>18.61</td>
</tr>
<tr>
<td>1\times10^{16}</td>
<td>51.5</td>
<td>2.6</td>
<td>27.5</td>
<td>23.2</td>
<td>14.55</td>
</tr>
<tr>
<td>1.25\times10^{16}</td>
<td>45</td>
<td>2.69</td>
<td>24.9</td>
<td>19</td>
<td>13.3</td>
</tr>
</tbody>
</table>

DEV:Si,n,B,N_s,4; f = 12.75 GHz; \( V_{\text{sig}}(t) = 1.1 \sin \omega t - 0.191 \) V

<table>
<thead>
<tr>
<th>( N_c (1/\text{cm}^3) )</th>
<th>Optimum DC Bias (V)</th>
<th>( M_A )</th>
<th>RF Voltage Amplitude (V)</th>
<th>( \eta ) (Percent)</th>
<th>( G_p ) (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>2\times10^{15}</td>
<td>70</td>
<td>1.05</td>
<td>41.5</td>
<td>38</td>
<td>12.8</td>
</tr>
<tr>
<td>4\times10^{15}</td>
<td>70</td>
<td>1.42</td>
<td>45.5</td>
<td>38</td>
<td>13.9</td>
</tr>
<tr>
<td>6\times10^{15}</td>
<td>63.5</td>
<td>1.72</td>
<td>41.5</td>
<td>38.5</td>
<td>15.8</td>
</tr>
<tr>
<td>8.5\times10^{15}</td>
<td>56.5</td>
<td>2.55</td>
<td>36.5</td>
<td>29.5</td>
<td>16.7</td>
</tr>
<tr>
<td>1\times10^{16}</td>
<td>51.5</td>
<td>2.6</td>
<td>29.5</td>
<td>26.5</td>
<td>14.86</td>
</tr>
</tbody>
</table>

DEV:Si,n,B,N_c,3; f = 12.75 GHz; \( V_{\text{sig}}(t) = 1.1 \sin \omega t - 0.191 \) V

<table>
<thead>
<tr>
<th>( N_c (1/\text{cm}^3) )</th>
<th>Optimum DC Bias (V)</th>
<th>( M_A )</th>
<th>RF Voltage Amplitude (V)</th>
<th>( \eta ) (Percent)</th>
<th>( G_p ) (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>2\times10^{15}</td>
<td>45</td>
<td>1.05</td>
<td>36.5</td>
<td>36.5</td>
<td>11.6</td>
</tr>
<tr>
<td>6\times10^{15}</td>
<td>56.5</td>
<td>1.35</td>
<td>36</td>
<td>44</td>
<td>12.3</td>
</tr>
<tr>
<td>1\times10^{16}</td>
<td>47.5</td>
<td>2.11</td>
<td>33</td>
<td>32</td>
<td>11.5</td>
</tr>
</tbody>
</table>

(Cont.)
Table 5.8 (Cont.)

$\text{DEV:Si}_{n}, B, N_{c}^{2}; f = 12.75 \text{ GHz}; V_{\text{sig}(t)} = 1.1 \sin \omega t - 0.191 \text{ V}$

<table>
<thead>
<tr>
<th>$N_{c}$ (1/cm$^3$)</th>
<th>Optimum Dc Bias (V)</th>
<th>$M_{A}$</th>
<th>RF Voltage Amplitude (V)</th>
<th>$n$ (Percent)</th>
<th>$G_{p}$ (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>$2 \times 10^{15}$</td>
<td>33</td>
<td>1.0</td>
<td>28.5</td>
<td>55</td>
<td>10</td>
</tr>
<tr>
<td>$6 \times 10^{15}$</td>
<td>37.5</td>
<td>1.02</td>
<td>29</td>
<td>52</td>
<td>10.2</td>
</tr>
<tr>
<td>$1 \times 10^{16}$</td>
<td>40</td>
<td>1.06</td>
<td>24.4</td>
<td>47</td>
<td>10.2</td>
</tr>
</tbody>
</table>
3. The optimum $N_c$ is approximately $6 \times 10^{15}$ cm$^{-3}$. With a higher $N_c$, the avalanche multiplication factor may be higher at the optimum dc bias, but the permissible $V_{RF}$ is consistently lower. The difference in the allowed $V_{RF}$ is pronounced when $w_T \geq 3.5 \mu m$ and $f = 12.75$ GHz. To maintain a large permissible $V_{RF}$, due to an increase in $V_{sus}$ with increasing $w_T$, the optimum dc bias of devices with $N_c \geq 1 \times 10^{16}$ cm$^{-3}$ actually decreases with increasing $w_T$ if $w_T > 4 \mu m$. The consequences can be a decreased $M_A$ and a decreased permissible $V_{RF}$. Devices with $N_c < 6 \times 10^{15}$ cm$^{-3}$ at their optimum dc bias suffer from a smaller avalanche multiplication factor and space charge is more likely to reduce the permissible $V_{RF}$. The space-charge effects were discussed in Sections 5.2.3 and 5.2.4.

4. Performance of device DEV:Si,n,B,6 $\times 10^{15},$h compares very favorably with that of devices with various 4 μm HI-LO collector structures at $f = 12.75$ GHz and $V_{sig}(t) = 1.1 \sin wt - 0.191 V$. The results are summarized in Tables 5.5 and 5.6. A CATT device with a uniformly doped collector can be fabricated more easily. The dynamic range and inherent bandwidths of devices with HI-LO and uniformly doped collectors are compared and discussed in a later section.

5. It has been shown that $V_{RF}$ should be as large as possible to maximize the power gain and efficiency. When $V_{RF}$ equals the maximum permissible value, there is a considerable amount of charge carriers generated from avalanche multiplication of the charge carriers of the collector reverse saturation current during the time period when $V_T$ is near its maximum value. This is also true for devices with HI-LO collector structures. Therefore, a more exact definition of $M_A$ is
This expression implies that the induced collector terminal current \( J_T(t) \) is due to avalanche multiplication of both the charge carriers of the emitter-base injected current and the collector reverse saturation current, although the former is usually the dominant one.

5.3 Large-Signal Simulation Results for Class C BJT Amplifiers

The operation of Class C BJT amplifiers can be considered as a special mode of operation of the previously simulated three-terminal devices operated as Class C CATT amplifiers. The difference lies in the combination of collector transit angle, base-collector dc bias and amplitude of \( V_{RF} \) which results in a collector multiplication factor of 1.1 or less for BJT amplifiers.

Since avalanche multiplication does not play an important role in determining the performance of Class C BJT amplifiers, only devices with uniformly doped collectors have been simulated. The results of large-signal simulations of various Class C BJT amplifiers operating at 12.75 GHz are summarized in Table 5.9. The following observations can be made:

1. For devices with any \( w_T \), those with the lowest \( N_c \) consistently have the best performance as long as space-charge density in the collector region is not excessively high. This is due to the restriction on \( M_A \) which makes \( V_{RF} \) the only major factor in determining amplifier performance. Higher \( N_c \) implies higher \( V_{sus} \) and \( V_{PT} \). Higher \( V_{sus} \) and \( V_{PT} \) implies higher electric field in the vicinity of the base-collector
Table 5.9
Large-Signal Simulation Results of Class C BJT Amplifiers

DEV: Si, n, B, N, 5; f = 12.75 GHz; \( V_{\text{sig}}(t) = 1.1 \sin \omega t - 0.191 \ V \)

<table>
<thead>
<tr>
<th>( N_e ) (1/cm(^3))</th>
<th>Optimum Dc Bias (V)</th>
<th>( M_A )</th>
<th>RF Voltage Amplitude (V)</th>
<th>( n ) (Percent)</th>
<th>( G_p ) (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>2x10(^{15})</td>
<td>70</td>
<td>1.06</td>
<td>32.9</td>
<td>20</td>
<td>10.89</td>
</tr>
<tr>
<td>6x10(^{15})</td>
<td>36.5</td>
<td>1.1</td>
<td>13</td>
<td>14.5</td>
<td>7.3</td>
</tr>
<tr>
<td>1x10(^{16})</td>
<td>always operates as a CATT amplifier when ( V_{\text{bias}} \geq V_{\text{PT}} )</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

DEV: Si, n, B, N, 4; f = 12.75 GHz; \( V_{\text{sig}}(t) = 1.1 \sin \omega t - 0.191 \ V \)

<table>
<thead>
<tr>
<th>( N_e ) (1/cm(^3))</th>
<th>Optimum Dc Bias (V)</th>
<th>( M_A )</th>
<th>RF Voltage Amplitude (V)</th>
<th>( n ) (Percent)</th>
<th>( G_p ) (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>2x10(^{15})</td>
<td>70</td>
<td>1.05</td>
<td>41.5</td>
<td>39</td>
<td>12.8</td>
</tr>
<tr>
<td>4x10(^{15})</td>
<td>55</td>
<td>1.1</td>
<td>21</td>
<td>40.5</td>
<td>12.55</td>
</tr>
<tr>
<td>6x10(^{15})</td>
<td>35</td>
<td>1.1</td>
<td>12.8</td>
<td>21.7</td>
<td>8.06</td>
</tr>
<tr>
<td>8.5x10(^{15})</td>
<td>32.5</td>
<td>1.1</td>
<td>10.5</td>
<td>13.5</td>
<td>6.53</td>
</tr>
<tr>
<td>1x10(^{16})</td>
<td>always operates as a CATT amplifier when ( V_{\text{bias}} \geq V_{\text{PT}} )</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

DEV: Si, n, B, N, 3; f = 12.75 GHz; \( V_{\text{sig}}(t) = 1.1 \sin \omega t - 0.191 \ V \)

<table>
<thead>
<tr>
<th>( N_e ) (1/cm(^3))</th>
<th>Optimum Dc Bias (V)</th>
<th>( M_A )</th>
<th>RF Voltage Amplitude (V)</th>
<th>( n ) (Percent)</th>
<th>( G_p ) (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>2x10(^{15})</td>
<td>45</td>
<td>1.05</td>
<td>36.5</td>
<td>56.5</td>
<td>11.6</td>
</tr>
<tr>
<td>6x10(^{15})</td>
<td>50</td>
<td>1.1</td>
<td>33.5</td>
<td>47</td>
<td>12</td>
</tr>
<tr>
<td>1x10(^{16})</td>
<td>27.5</td>
<td>1.1</td>
<td>14.4</td>
<td>20</td>
<td>7.5</td>
</tr>
</tbody>
</table>

(Cont.)
Table 5.9 (Cont.)

DEV:Si,n,B,Ne,2; f = 12.75 GHz; V_{\text{sig}}(t) = 1.1 \sin \omega t - 0.191 V

<table>
<thead>
<tr>
<th>N_c (1/cm^3)</th>
<th>Optimum DC Bias (V)</th>
<th>M_A</th>
<th>RF Voltage Amplitude (V)</th>
<th>n (Percent)</th>
<th>G_p (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>2x10^{15}</td>
<td>33</td>
<td>1.0</td>
<td>28.5</td>
<td>55</td>
<td>10</td>
</tr>
<tr>
<td>6x10^{15}</td>
<td>37.5</td>
<td>1.02</td>
<td>29</td>
<td>52</td>
<td>10.2</td>
</tr>
<tr>
<td>1x10^{16}</td>
<td>40</td>
<td>1.06</td>
<td>24.4</td>
<td>47</td>
<td>10.2</td>
</tr>
</tbody>
</table>
junction. When $M_A$ is restricted to 1.1 or less, therefore, devices with higher $N_C$ have a lower maximum allowed $V_{RF}$. This points out one major difference in the design of CATT and BJT amplifiers.

2. Maximum power gain is achieved by the device DEV:Si,n,B,2 x 10^{15},4 while for CATT devices maximum power gain is achieved by devices with collectors 5 μm or longer. As for BJT amplifiers, longer devices have a large collector transit angle which implies that charge carriers injected from the emitter-base junction reach the multiplication region when $V_T$ is higher. Since $M_A$ is restricted to 1.1 or less, therefore, a longer device has a smaller maximum permissible $V_{RF}$ when $v_T \geq 3$ μm. If the collector is intrinsic or extremely lightly doped, maximum power gain would be achieved by a longer device, but the maximum current density would be limited by the aforementioned space-charge effects and therefore the device output power will be limited. The reductions in maximum permissible $V_{RF}$ and power gain with increasing $v_T$ are very pronounced in devices with $N_C \geq 6 \times 10^{15}$ cm$^{-3}$ and $v_T \geq 3.5$ μm. This points out another difference in the design of CATT and BJT amplifiers.

3. Devices with $v_T \leq 2.5$ μm, when operating at optimum conditions, are BJT amplifiers independent of the values of $N_C$. The performance of the device with $N_C = 1 \times 10^{16}$ cm$^{-3}$ is almost the same as that of the device with $N_C = 2 \times 10^{15}$ cm$^{-3}$ at $V_{sig}(t) = 1.1 \sin \omega t - 0.191$ V. At $v_T \leq 2$ μm, a higher $N_C$ may be desirable due to increased current capability of the device.

4. A shorter device has a higher efficiency.

Comparisons between the Class C CATT and BJT amplifiers are carried out in the next section.
5.4 Comparison Between the Class C CATT Amplifier and the Class C BJT Amplifier

5.4.1 RF Power Gain and Efficiency at \( f = 12.75 \text{ GHz} \) and

\[ V_{\text{sig}}(t) = 1.1 \sin \omega t - 0.191 \text{ V}. \]

Large-Signal simulations of devices with \( w_T \) ranging from 2 \( \mu \text{m} \) to 5 \( \mu \text{m} \) and \( n_c \) from \( 2 \times 10^{15} \text{ cm}^{-3} \) to \( 1 \times 10^{16} \text{ cm}^{-3} \), operating at 12.75 GHz and \( V_{\text{sig}}(t) = 1.1 \sin \omega t - 0.191 \text{ V} \) were carried out and the results are summarized in Tables 5.8 and 5.9.

It is observed that a maximum \( G_p \) of 18.61 dB at 29.9 percent efficiency is achieved by DEV:Si,n,B,6 \( \times 10^{15},5 \) operating as a CATT amplifier as compared to a maximum \( G_p \) of 12.8 dB at 39-percent efficiency achieved by DEV:Si,n,B,2 \( \times 10^{15},4 \) operating as a BJT amplifier. An additional 5.8-dB power gain at the expense of 10-percent efficiency reduction can be developed by employing a suitable CATT amplifier. If DEV:Si,n,B,6 \( \times 10^{15},4 \) operating as a CATT amplifier is used, a 15.8-dB power gain at 38.5-percent efficiency can be achieved. Thus a well-designed CATT amplifier can achieve the same efficiency and a 3-dB additional power gain as compared to the maximum gain BJT amplifier at 12.75 GHz. The higher gain is mainly due to a multiplication factor of 1.72 and, to a lesser extent, a slightly larger allowed \( V_{\text{RF}} \). If DEV:Si,n,B,6 \( \times 10^{15},3 \) \( \mu \text{m} \) is operating as a CATT amplifier, a power gain of 12.3 dB can be achieved which is approximately the maximum power gain achievable by the BJT amplifier at 12.75 GHz. However, due to its shorter collector region, CATT amplifier DEV:Si,n,B,6 \( \times 10^{15},3 \) has a higher efficiency than the maximum gain BJT amplifier.

It was noted previously that devices with \( w_T \leq 2.5 \mu\text{m} \), when operating at optimum conditions and \( f = 12.75 \text{ GHz} \), are BJT amplifiers since their multiplication factors are less than 1.1. The BJT amplifiers
with \( w_T \leq 2.5 \) \( \mu \)m have higher efficiencies but lower power gain than the CATT amplifiers.

5.4.2 Dynamic Range. The dynamic ranges of various CATT amplifiers and BJT amplifiers were investigated and the results are discussed in this section. The \( V_{\text{bias}} \) is kept constant at the optimum \( V_{\text{bias}} \) corresponding to \( V_{\text{sig}}(t) = 1.1 \sin \omega t - 0.191 \) V, although it was shown previously that the optimum \( V_{\text{bias}} \) varies slightly with the input signal level.

From Fig. 5.10 it is observed that \( G_p \) and \( \eta \) of both the CATT and BJT amplifiers are low at low input signal levels. This is due to low emitter injection efficiency at low input signal levels at an operating frequency of 12.75 GHz. At high input signal levels, \( G_p \) and \( \eta \) of both the CATT and BJT amplifiers deteriorate. In the case of CATT amplifiers, the deteriorations in \( G_p \) and \( \eta \) are due to space-charge-caused reductions in \( M_A \) and \( V_{\text{fp}} \) and space-charge-caused spreading in the induced current waveform. High injection level effects in the emitter-base region is another cause. In the case of BJT amplifiers, all the aforementioned are causes for reductions in \( G_p \) and \( \eta \) at higher input signal levels except the reduction in \( M_A \). By definition, \( M_A \) is never greater than 1.1 for BJT amplifiers.

By comparing Figs. 5.10a and b, it is seen that DEV:Si,n,B,1 \( \times 10^{16},4 \) has comparable \( G_p \) at low input signal levels, i.e., \( P_{\text{in}} \leq 1 \times 10^{-3} \) W, as does DEV:Si,n,B,6 \( \times 10^{15},4 \). Its \( G_p \) reduces faster with increasing \( P_{\text{in}} \) in the range \( 1 \times 10^{-3} \) W \( \leq P_{\text{in}} \leq 5 \times 10^{-3} \) W than that of DEV:Si,n,B,6 \( \times 10^{15},4 \) due to the fact that \( M_A \) is a more important factor in determining its \( G_p \) and \( M_A \) is very sensitive to any space-charge-caused reduction in the electric field during emitter-base
FIG. 5.10 POWER GAIN AND EFFICIENCY VS. INPUT POWER
CHARACTERISTICS OF VARIOUS CAT AND BJT AMPLIFIERS.
charge injection. For $P_{\text{in}} > 5 \times 10^{-3} \text{ W}$, the rate of reduction in $G_p$ is slower for $\text{DEV:Si,n,B,l } x 10^{16},4$ because $M_A$ is no longer as important in determining $G_p$ and the space-charge effect in the low-field drift region of $\text{DEV:Si,n,B,l } x 10^{16},4$ is not as severe as in $\text{DEV:Si,n,B,6 } x 10^{15},4$. This explains the even faster rate of reduction in $G_p$ of $\text{DEV:Si,n,B,2 } x 10^{15},4$ as shown in Fig. 5.10d. Of the three devices with uniformly doped collectors, $\text{DEV:Si,n,B,6 } x 10^{15},4$ has the best $G_p$ and $\eta$ characteristics. All three devices with HI-LO collector structures, whose $G_p$ and $\eta$ characteristics are shown in Figs. 5.10j through 1, are very inferior to $\text{DEV:Si,n,B,6 } x 10^{15},4$.

Devices whose $G_p$ and $\eta$ characteristics are shown in Figs. 5.10c, f and g through i are BJT amplifiers. It is observed that the amplifier dynamic range is severely reduced when $N_c$ is too low (due to space-charge effects) and shorter devices with properly designed $N_c$ have lower power gains but higher efficiencies and a wider dynamic range. The wide dynamic range associated with a short device, i.e., $w_T = 2 \mu m$, is due to the fact that $M_A \leq 1.1$ at optimum $V_{\text{bias}}$ even at low input signal levels and therefore an almost constant $M_A$, independent of space charge, exists.

5.4.3 Inherent Bandwidth. Plots of $G_p$, $\eta$, $V_{RF}$, and $M_A$ vs. frequency for CATT and BJT amplifiers are shown in Figs. 5.11 through 5.13. The decrease in $G_p$ for a short BJT amplifier, i.e., $\text{DEV:Si,n,B,6 } x 10^{15},2$, can be attributed to well-known causes such as the emitter-base RC frequency cutoff mechanism, base and collector time delay, and collector capacitor charging time. Since $M_A$ is limited to 1.1 or less for the BJT, $V_{RF}$ decreases slightly as frequency is increased. This is shown in Fig. 5.13. This decrease in $V_{RF}$ also plays
FIG. 5.11 CLASS C CATT AMPLIFIER PERFORMANCE VS. FREQUENCY
AT CONSTANT INPUT POWER LEVEL. (DEV: $S_{i,n,B}=6 \times 10^{15} W/m^2$
AND $V_{bias} = 63.5 V$)
FIG. 5.12 CLASS C BJT AMPLIFIER PERFORMANCE VS. FREQUENCY

AT CONSTANT INPUT POWER LEVEL. (DEV:S1,u,B,2 x 10^{15},h

AND V_{bias} = 70 V)
FIG. 5.13 CLASS C BJT AMPLIFIER PERFORMANCE VS. FREQUENCY

AT CONSTANT INPUT POWER LEVEL. (DEV: $Si, n, B, 6 \times 10^{15}, 2$

AND $V_{bias} = 37.5$ V)

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a role in the decrease of $G_p$ and $\eta$. For a longer BJT amplifier, i.e., DEV:Si,n,B,2 x $10^{15}$, the decrease in $V_{RF}$ is more severe in order to keep $M_A \leq 1.1$. Thus, $G_p$ and $\eta$ decrease more rapidly with frequency as shown in Fig. 5.12. The inherent bandwidth of the 2-\mu m BJT is wider than the 4-\mu m BJT. For a CATT amplifier, $M_A$ is allowed to take on any value as long as it is not too high and a stable operation cannot be achieved. The RF voltage amplitude $V_{RF}$ of CATT device DEV:Si,n,B,6 x $10^{15}$ stays constant up to 1.4 GHz. For higher frequencies, $V_{RF}$ decreases with increasing frequency in order to maintain stable operation. This decrease in $V_{RF}$ results in decreases in $G_p$ and $\eta$. As frequency increases, $M_A$ also increases which helps to offset the decrease in $G_p$. This explains the slower rate of decrease in $G_p$ as compared to that of $\eta$. The 4-\mu m CATT amplifier has a much wider inherent bandwidth than the 4-\mu m BJT amplifier.

5.5 Conclusions

Large-signal simulation results for both the Class C CATT and Class C BJT amplifier were presented in this chapter. The optimum dc bias condition, optimum load and their variations with frequency were investigated. The effects of device structural parameters were studied and optimum device structures for both the CATT and the BJT amplifier operating at 12.75 GHz were obtained. Several differences in the design of optimum CATT and BJT amplifiers were derived from the large-signal simulation results. Comparison between the CATT and the BJT amplifiers in terms of gain, efficiency, dynamic range and intrinsic bandwidth were also given.
CHAPTER VI. SUMMARY, CONCLUSIONS, AND SUGGESTIONS FOR FURTHER STUDY

6.1 Summary and Conclusions

The purpose of this study was to investigate the effects of avalanche multiplication and transit time in the collector region of a BJT. Analytical equations, circuit models, and computer simulations were used to determine dc, small-signal, and large-signal behavior of CATT amplifiers.

In Chapter II a dc computer program was developed which determines the dc avalanche multiplication factor vs. $V_{\text{bias}}$ characteristics for any Si or GaAs collector structure. The results provide an estimation of device large-signal performance capability. The optimum collector parameters, i.e., $v_{av}$, $n_{av}$, $N_c$, obtained from the dc computer program correspond well to the results of large-signal simulation. Results also indicate that an n-type Si CATT amplifier is superior to p-type Si and n-type GaAs CATT amplifiers due to its favorable $M_{AO}$ vs. $V_{\text{bias}}$ characteristics.

In Chapter III analytical models of dc and small-signal characteristics for CATT devices with Read-type collector structures were given which incorporated both the avalanche multiplication and the collector transit-time mechanisms. Contrary to previous findings, the small-signal characteristics of Class A CATT amplifiers indicated that a larger avalanche multiplication factor results in a smaller RF power gain and an increase in $M_{AO}$ does not necessarily imply a significant increase in $f_{\text{max}}$. Results were given and discussed. One conclusion was the inapplicability of CATT devices as Class A amplifiers.
In Chapter IV a large-signal computer simulation was developed which incorporated several improvements over the large-signal simulation previously reported, i.e., high injection level effects in the base region, Early effect, effect of high impurity level in the emitter, nonzero minority carrier concentration at the edge of the base-collector depletion region on the base side, minority carrier induced electric field in the base region, current conserving boundary condition for minority carriers in the collector region, and diffusion current in the collector depletion region.

In Chapter V large-signal results of Class C CATT amplifiers were given. Effects of base-collector dc bias, load, operating frequency and collector structures were examined and discussed. The simulation calculates output power, power gain, and efficiency of the amplifier. It also gives emitter-base current and voltage waveforms; avalanche multiplication factor; \( J_T \) and \( V_T \) waveforms; and spatial distributions of electrons, holes, and electric field in the collector depletion region at any time instant. Various Si n-type CATT and BJT Class C amplifiers operating at 12.75 GHz were compared in terms of output power, power gain, efficiency, and dynamic range. Optimum collector impurity doping level and optimum width of the collector region of both the CATT and BJT amplifiers were determined and discussed. Carrier multiplication and long collector transit time do increase power gain, but at the expense of lower efficiency and smaller dynamic range. Avalanche multiplication does help to increase the device inherent bandwidth.

6.2 Suggestions for Further Study

In the course of this study several additional topics which need further exploration were found. They are as follows:
1. A cost-tolerable, two-dimensional, large-signal computer simulation program to account for the nonuniform injection of carriers at the emitter-base junction and the nonuniform injection of carriers into the collector region due to nonuniform emitter-base junction potential across the emitter finger laterally caused by the flow of conventional base current and feedback hole current in a resistive base region.

2. Fabrication of X-band CATT devices and experimental studies of Class C CATT amplifiers.

3. Construction of new models for use at higher frequencies.


5. Thermal limitation studies.

6. Application of CATT devices as high-voltage, high-current drivers.
LIST OF REFERENCES


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