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Preamble

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THE STARTING CHARACTERISTICS OF TRAPATT OSCILLATORS

BY

KRISHNA KUMAR

A thesis submitted to the Faculty of Mathematical and Physical Sciences at the University of Surrey for the degree of Doctor of Philosophy

October 1980
THIS WORK IS DEDICATED TO

CHITRA, my wife for her patience, understanding and encouragement.

AND

MY LATE FATHER, for his inspiration and motivation.
SUMMARY

The aim of the work described in this thesis was to examine the starting characteristics of Trapatt (Trapped Plasma Avalanche Triggered Transit) oscillators capable of producing high (peak) power at high efficiencies, normally in L and S bands. General approach was aimed to be mainly experimental using deep diffused silicon devices suitable for operation in S band.

To perform the experiments an oscillator circuit, in 7 mm coaxial line was constructed, suitable for using a device in S-4 package. The overall arrangement of the experimental setup was similar to one described by various other researchers for observing the dynamic current and voltage waveforms.

The C-V profiling experiment produced a characteristic which suggested that the depletion layer capacitance of the device does not really saturate for increasing voltage even upto its breakdown voltage. A detailed analysis of the C-V plot and related features was carried out and it was concluded that the actual doping distribution appears to be favouring a graded junction.

Experimentally it was established that the device has negative resistance at VHF and also at d.c. It has also been shown that the device has small signal negative resistance at the operating frequency of the oscillator.

Detailed investigations into the starting characteristics indicated that the oscillations start from the beginning at the final fundamental frequency of the oscillator. For the device and the circuit reported in this thesis no evidence could be found to suggest that Impatt
type oscillations trigger the Trapatt oscillations. As a result of our experiments it has been possible to establish that the rate of growth and the time taken for the oscillations to start depend upon the d.c. drive. There are strong indications that in future Trapatt devices could successfully be employed in various systems as a microwave generator.
ACKNOWLEDGEMENTS

The author is deeply indebted and thankful to Dr. K. W. H. Foulds for his help, guidance and encouragement not only as a research supervisor but also as an understanding person who could appreciate the problems of a mature candidate.

It was a great pleasure to be associated with a really friendly department and to every one of them I am thankful. I am also thankful to Dr. J. L. Sebastian Franco of Madrid and Dr. F. J. Lidgey of Oxford Polytechnic with whom I had many useful discussions.

There are many to whom I am greatly thankful, outside the work area, including Mrs. Britten, Mr. and Mrs. Saxena, Mr. M. K. Verma, Mr. S. K. Sinha and many more friends.

Finally I am thankful to Dr. B. H. Newton, Mr. J. G. Summers and Dr. R. Davies of Philips Research Laboratories, Redhill, for providing the devices and lending me the sampling scope without which the work could not have been completed.

In the end I am thankful to Science Research Council and the University of Surrey for providing the financial assistance for this work.
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CHAPTER 1

INTRODUCTION

Immediately after the second world war the limitations of the conventional valve type devices and their capabilities were realised. The search for suitable solid state materials to replace them resulted in the famous discoveries of Gunn and Avalanche devices around 1950, both being two terminal devices. Benson (1) (Fig 1.1) has reported about various microwave semiconductor devices in the form of a tree, which gives a clear view about them and the materials used to produce them, however this diagram does not give their historical development. Perhaps Shockley (2), in early 1950 realised the potential applications of junction devices when operated beyond their breakdown voltages under reverse biased conditions. However, this work did not gain much attention till about 1958, when W.T. Read (3) published his famous paper on high frequency negative resistance diode. This paper still forms the basis and the starting point of further advanced work on avalanche devices. Most avalanche devices are known as IMPATT (Impact-Avalanche Transit Time). As a result of numerous experiments in the last 20 years oscillators upto 100 GHz, capable of producing upto 10 watts of c.w. power are commercially available. None of the conventionally available microwave devices were able to produce high peak power at relatively low frequency (sub-transit time) and high efficiency. It was the need of such a device which resulted in the discovery by Prager, Chang and Weisbrod in 1967 (4), now known as TRAPATT (Trapped-Plasma Avalanche Triggered Transit).
FIG 1.1 : DEVELOPMENT OF MICROWAVE SEMICONDUCTOR DEVICES
1.1 SALIENT FEATURES OF THE TRAPATT OSCILLATOR

For about one year following the publication by Brager et al., the mechanism of oscillations was not at all understood and hence this type of behaviour was called "The Anomalous Mode". It was called so because of the following important characteristics not normally observed in transit time devices:

(a) Very high d.c. to r.f. conversion efficiencies.
(b) Oscillations at frequencies well below the transit time frequency of the device.
(c) A sudden decrease in the voltage across the device and a consequent increase in the bias current at the onset of the oscillations.
(d) High threshold power for oscillations.

An insight into its operation was obtained in 1968 by Johnston, Scharfetter and Bartelmink through a computer simulation of these devices. As a direct consequence of the results produced by this programme, this device was given the present name.

Simplified semi-analytical treatment of the device physics and its application was given by Clorfeine et al. and De Loach and Scharfetter in 1969 and in 1970. In both the approaches the most important condition required to be satisfied was that the current density through the device must be greater than $eNV_s$ (where $e$ is the electronic charge, $N$ the background doping density in the depletion layer and $V_s$ the saturated drift velocity of the carriers), so as to create a fast moving avalanche shock front. Apart from many simplifying assumptions made by both the authors, they have treated the device as an abrupt junction type. No doubt the devices used in earlier experiments could have been approximated
this way but the same situation does not exist today since most practical devices are supposed to be graded.

Among the useful practical oscillator circuits, the one suggested by Evans (8) appears to be the most adaptable and flexible. The general configuration of the coaxial oscillator circuit is shown in Fig 1.2, which will be discussed in detail in other chapters. As a result of various experiments conducted by Evans some salient circuit details useful in designing and understanding the device – circuit interaction, were known. Evans pointed out that the 50 ohm line is not capable of supplying large current required to initiate the high efficiency oscillations and suggested the need of an extra capacitor in the immediate vicinity of the diode. Trew et al. (9) has given a very good explanation of this effect and has designated it as "charging capacitor". On the basis of detailed experimental results, they concluded that for the best performance of the oscillator circuit there exists an optimum value of this capacitor and for either too large or too small a value of C the performance is greatly degraded. The optimum value of this capacitance is a function of the device area and as a rule larger area devices require larger value of C. The effect of the C values are shown in Fig 1.3, as reported by Trew et al. (9). The design triangle reported by Clorfeine (10) in August 1971 forms a good starting point and does establish certain boundaries and limits for successful operation. Similar work, on the basis of the bias current and bias voltage, has been reported by Mackintosh (11), who has developed a "Mode Chart" formed by a combination of straight lines. The main difference between the two guide lines is that Clorfeine has taken into account the interaction between the device and the circuit whereas the mode chart reported by Mackintosh is mostly circuit oriented.

The most outstanding feature of the Trapatt oscillations is
FIG 1.2: COAXIAL OSCILLATOR CIRCUIT

Fig 1.3: TRAPATT OSCILLATOR WAVEFORMS ILLUSTRATING THE EFFECTS OF LARGE AND TOO SMALL CHARGING CAPACITANCE VALUES.
the conversion efficiency (12, 13, 14) with which these devices can be operated. The device is able to do so because for the most part of the r.f. cycle the current flowing in the device is almost zero. Various ideas have been put forward to explain the general nature of the dynamic current and voltage waveforms, however the exact details of these waveforms can only be understood by dividing the entire period into a number of segments (6). However an essential feature of the oscillations is the overdrive voltage required to create excessive avalanching and thereby allowing heavy conduction current through the device. In Fig 1.4 we have reported the dynamic waveforms measured by Kerzer and Weissglass (15). It can be seen that initially the voltage and the current increase, the voltage reaches to its peak value at about 5 amps, the current still keeps on increasing whereas the voltage begins to fall rapidly. After this period the voltage begins to recover and the current begins to fall. For more than half the total period of the r.f. cycle the current is almost zero and the voltage less than the breakdown voltage.

1.2 REVIEW OF THE STARTING MECHANISMS

In the past various aspects of Trapatt oscillations have been studied by various research workers and most of the details regarding designing and producing a good Trapatt oscillator are well understood. However, one aspect of this type of oscillations is still far from clear i.e. "The Starting Mechanism". Probably Evans (8) in 1969 was the first to suggest that Impatt generated signals are essential to start the Trapatt oscillations. His argument was based on the fact that such devices did not exhibit small signal negative resistance in the required frequency range. However, for similar devices Bower (16) and Snapp (17) have reported that the devices capable of operating in Trapatt mode do possess negative resistance at d.c. and upto the transit time frequency of the device.
Fig 1.4: DYNAMIC VOLTAGE AND CURRENT WAVEFORMS
There are others like Zappert and Lee (18), Bogan and Frey (19) and Blakey (20) etc, who believe that the oscillations from the start are of "Relaxation" type. East et al. (21) have reported some experimental results and conclude that either VHF or Impatt signals may be responsible for triggering the Trapatt oscillations, and the mode of triggering is largely determined by the bias circuit employed. There are various other alternative theories including the parametric amplification as the possible triggering mechanism. Most of these theories will be dealt with in some detail in the sixth chapter of this thesis.

1.3 AIMS OF THE PRESENT THESIS

The main aim of this thesis is to investigate the starting mechanism of the Trapatt oscillations. The theme of the work will be mainly experimental, however necessary theoretical approach will also be undertaken to explain some of the results particularly relating the device and its d.c. characteristics. The devices used for the work were kindly supplied by Philips Research Lab, Redhill. All the the devices were Si PNN+ type, with deep diffused P+. There are six chapters, including the present one. In the second chapter we have tried to predict the doping profile on the basis of the known details and steps involved in the fabrication of the device. For experimental devices the C-V plot is reported and critically analysed. It is concluded that the doping distribution is likely to be graded rather than abrupt. A theory and computer program is developed to calculate the C-V plot for an assumed doping profile derived from the previously predicted one. A comparison of the measured and computed C-V plots suggests that the predicted doping profile appears to be very close to the actual one.

The third chapter is devoted to the d.c. analysis of the device
on the basis of the Poisson and Continuity equations in one dimension. The
detailed program uses different ionization coefficients for holes and
electrons at the room temperature to calculate the voltage across the device
for a known value of the d.c. current through it. The presence or absence
of a knee or turning over point in current – voltage characteristics will
indicate whether or not the device has some differential negative
resistance at d.c., on the basis of the created space charge within the
depletion layer. It is concluded that the device does not manifest any
negative resistance at d.c.

In the fourth chapter, we have reported three results which
show that the device does have d.c. differential negative resistance, negative
resistance at VHF and also small signal negative resistance at the fundamental
frequency of the oscillator. Low frequency negative resistance is brought about
by the rectification of some VHF signals set up within the system. The
growing nature of the VHF oscillations confirm the presence of the negative
resistance. Small signal negative resistance is confirmed as a result of
the reflection type experiment. The experimental evidence of this chapter
will form the basis of the explanation to be given in the sixth chapter.

In the fifth chapter, we have described the details of the
oscillator circuit, including the charging line, tuning slugs, voltage and
current probes. This chapter also deals with some details of the triggering
arrangements of the sampling scope.

The sixth chapter contains a detailed investigation of the
starting mechanism and a comprehensive review of the literature available
on this subject. Detailed results have been reported for one bias setting,
however the results of other similar conditions have been presented in the
form of a table and graphs. It is concluded that for the device and the
circuit described in this thesis the oscillations start from small signal at the fundamental frequency of the Trapatt oscillator.
CHAPTER 2

THE DEVICE AND THE INTERPRETATION OF THE C-V PLOT

High efficiency Trapatt devices have been constructed using either Si or Ge. Si is commercially preferred because of its ability to sustain high input power density to produce high efficiency oscillations. The low breakdown field and small saturated drift velocity in Ge combine to give threshold power densities a factor four smaller than Si \(^{(7)}\) \(^{(22)}\). Blakey has given a very good survey of various suitable device materials and their useful properties. Though the techniques of device fabrication are essentially similar to Impatts, Trapatt devices have some striking differences. On the basis of the drift region these devices may be put in two categories.

(a) Single Drift Region Devices SDR
(b) Double Drift Region Devices DDR

In the literature mainly SDR devices also known as punch through devices have been described. In these devices the depletion layer punches through the substrate before the breakdown occurs. This aspect of the device is shown in Fig. 2.1 for P\(^+\)NN\(^+\) structure. An alternative complementary structure can have P type region in the middle. In these devices a narrow region of relatively high resistivity (4-6 ohm cm), low doping density is sandwiched in between regions of very high doping densities (approx \(10^{19}\) cm\(^{-3}\)) designated as P\(^+\) or N\(^+\).
Fig. 2.1: FIELD PROFILE OF A SINGLE DRIFT $P^+NN^+$ DEVICE
The end regions are highly doped so as to achieve the desired punch through structure in which beyond a certain value of the applied voltage the depletion layer capacitance begins to saturate. This voltage is known as "Punch Through" voltage. The P+NN+ device as shown in Fig 2.1 is said to be punched through if the depletion layer sweeps across the N layer and into the N+ region before the field at the metallurgical junction reaches the breakdown value. For this structure the punch through factor \( F \) is defined as the ratio of \( W_b/W \) i.e., the ratio of the depletion layer width of the one sided abrupt junction diode to the width of the N region in the punch through diode with the same doping concentration. This ratio can also be defined as the square root of the ratio of the breakdown voltage to the punch through voltage \( \sqrt{V_B/V_P} \).

In these devices under reverse bias condition the field reaches to max at P+ N interface where avalanching takes place and electron hole pairs are created. In the structure shown the holes are driven into the P+ region whereas the electrons drift through the N region towards the N+ contact. The conduction current is mainly due to the electrons. In essence carriers drift through a single drift region.

DDR devices have two drift regions as compared to a single drift region in SDR devices. In DDR devices both electrons and holes contribute towards the conduction current. Though DDR type devices were known even in 1959 \( (24) \), its applications to Trapatts was first proposed by Scharfetter \( (25) \) et al in 1970. Perhaps Fisher \( (26) \) was the first to develop a theory for a device similar to DDR. A very nice description of DDR as applied to Impatts is given by Carroll \( (27) \) under the heading of "The Current Fraction" in his book. Various researchers have reported the superiority of DDR over SDR in terms of output power and efficiency,
however, a very good survey and comparison has been given by Seidel (28) et al. Though Seidel mainly talks about those devices which are made by the "Ion Implantation" process in which the low density middle regions have almost constant impurity concentration but these facts are also applicable to devices fabricated by solid state diffusion.

The literature available on DDR Trapatt devices is not enough to enable us to say that the low frequency performance of these devices will be as good as at high frequencies (50-100 GHz). Perhaps Kawamoto (29,30) et al were the first to successfully demonstrate the advantages of these devices which they called "Four Layer Diode Structure". They clearly indicated the existence of two distinct possibilities based on the methods employed for their fabrication. Looking at the structure closely it seems logical and meaningful to arrive at the conclusion that the structure suggested by Liu (31) et al in March 1970 was really of the similar type as employed by Kawamoto etc. However later on in 1973 Liu (32) published a report in which he describes a double ended (P+PNN+) Trapatt device and argues that this structure may be looked upon as the series connection of P+NN+ and NPP+ diodes.

In Fig 2.2 a double drift structure is sketched and it is shown that the avalanche zone is located at the metallurgical junction and lies in both P and N regions. Holes drift towards the P+ contact and electrons towards the N+. A very significant point about this structure is a common avalanche region for two drift regions and thus it is expected that this structure should be more efficient than its counterpart SDR. Scharfetter (25) indicates that there are two possible reasons for this increased efficiency ie increase in d.c. voltage and the reduction in minority carrier storage.
Fig 2.2: FIELD PROFILE OF A DOUBLE DRIFT P⁺PNN⁺ DEVICE

Electric Field in Volt Cm⁻¹

Distance in μm
2.1 THE DEVICE FABRICATION

There is enough literature available giving details of various steps involved in fabrication of these devices, however we will try to give some details under this heading which will form the basis of further work later in the chapter. Perhaps a special mention should be made of Liu's \(^{31}\) paper in which he outlines the need of deep diffusion so as to avoid "burn out" problem encountered in earlier devices made by shallow diffusion process. The starting material is either P or N substrate of very high doping concentration. Depending on the starting material either P or N type layer of predetermined resistivity is epitaxially deposited on the substrate. The thickness of this layer is determined by the diffusion process to be employed to obtain either N or P end regions. From the surface of the epitaxial layer P or N type of impurities are deep diffused at elevated temperature in two steps. To explain these steps and some other details in chronological order a flow chart is given in Fig 2.3. Some numbers are given on the chart to indicate roughly the temperature and time used for various steps. Though the flow chart given in Fig 2.3 outlines a most general scheme, it does not reveal some of the most important features of the device. Various steps of diffusion, their temperatures and times are very carefully chosen and optimised to obtain the required and desired electrical and thermal properties. Two step diffusion ensures a significant out diffusion of Sb in to the N region and this is done to achieve the required value of the breakdown voltage. Two step diffusion produces the Gaussian \(^{33}\) distribution of Boron at P\(^+\)N junction and this together with the out diffusion is likely to produce a P\(^+\)PNN\(^+\) type of structure. Various relevant device details are well documented \(^{14,34}\).

2.2 THEORETICAL PREDICTION OF THE DEVICE DETAILS

For the purpose of theoretical prediction and calculation
Preparation of Sb doped substrate having an average resistivity of 0.005 to 0.01 ohm cm

Epilayer deposited at approximately 1150°C

Epitaxial deposition of P doped N layer having an average resistivity of 4 to 6 ohm cm. The length of the epilayer app. 15.5 μm.

N = Substrate doping density
N₀ = Epilayer doping density.

OX = Length of the epilayer 15.5 μm
AX = Outdiffusion during epilayer dep.

Various steps of cleaning, washing, drying, photo masking, oxidation and etching etc.

Two step diffusion of Boron

(a). Predeposition at 1100°C for ½ hour. It is done to introduce a fixed amount of impurity within a very narrow region less than a μm, close to the surface.

Doping profile after Predeposition

(b). Drive in at app 1110°C for 24 hours and then slow cooling.

Final doping profile:

| N₁ = 2.7 \times 10^{19} \text{ cm}^{-3} |
| N₂ = 5.0 \times 10^{18} \text{ cm}^{-3} |
| N₀ = 1.5 \times 10^{15} \text{ cm}^{-3} |

OX = 15.5 μm
OX = 12.3 μm

X_j is the metallurgical junction.

CX is the total out diffusion of Sb in to the N layer.

Fig 2.3: THE FABRICATION PROCESS OF P⁺N⁺ DEEP DIFFUSED TRAPATT DEVICE
of the device doping distribution, we will have to make some assumptions
and simplifications. We will first of all decide the general nature of
the distribution finally obtained in N-N+ region and then decide the
distribution in P N region.

2.2.1 THE DISTRIBUTION IN N-N+ REGION

The starting doping distribution in the N+ region is $1.0 \times 10^{19}$
cm$^{-3}$ and this region extends for a large distance and hence it can be said
that the impurity concentration far away from the N-N+ interface is indep­
endent of time and is equal to the substrate doping. For all practical
purposes, the net out diffusion of Sb in the N region can be divided in
to three main headings.

(a) Out diffusion during the epitaxial process.
(b) Out diffusion during the pre-deposition of Boron
at the surface.
(c) Out diffusion during the drive-in of Boron
at the surface.

The temperatures involved in all three steps are almost similar
however the drive in time is many times more than the time taken in steps (a)
and (b) above. Typically for our deep diffused devices :-

Time taken for epitaxial growth = 20 min
Time taken for pre-deposition = 32 min
Time taken for drive-in = 1440 min ie 24 hours

Hence for the purpose of the calculation of the doping distri­
bution we will neglect the out diffusions due to (a) and (b) above. Thus
the values to be used for calculations below are the substrate doping
distribution = $1.0 \times 10^{19}$ cm$^{-3}$, drive in temperature and time being 1110°C
and 24 hours respectively.
The impurity concentration at \( N-N^+ \) interface after out diffusion will be \( \text{Sub/2} \) if \( V \sqrt{t/Dt} > 1 \). \( V \) is the film growth rate, \( t \) is the time taken and \( D \) is the diffusion constant for the substrate dopant. In general this condition is always met in fabrication of most practical devices. Using the Fig.13 on page 31 of Sze (35) for Si.

Diffusion coefficient for Sb at 1110°C = \( 2.6 \times 10^{-14} \text{cm}^2 \text{Sec}^{-1} \)

Time of diffusion = 24 hours = 8.64 \ times 10^4 \text{Sec}

Diffusion length, \( 2\sqrt{Dt} \) = 9.54 \times 10^{-5} \text{cm}

\( B_0 \), The Diffusion Parameter = \( 1/2\sqrt{Dt} \)

= 1.05 \times 10^4 \text{cm}^{-1}

\( N_2 \) at \( X_0 = 15.5 \times 10^{-4} \text{cm}, = 5.0 \times 10^{18} \text{cm}^3 \). The out diffusion of \( N_2 \) in to the \( N \) towards the surface away from \( X_0 \) will be given by equation (2.1).

\[
N_d(x) = N_2 \{1 - \text{erf} \left( B_0 (X_0 - x) \right)\} \pm N_o
\]

\( N_o \) is the background doping density in the \( N \) region. In our case since the nature of the impurities in \( N \) and \( N^+ \) is the same, we will retain the positive sign.

\[
N_d(x) = N_o + N_2 \{1 - \text{erf} \left( X_0 - x)B_0 \right) \}
\]

(2.1)

Substituting the values of \( N_o, N_2, B_0 \) and \( X_0 \) we will obtain the doping distribution in \( N-N^+ \) region. This is given by the equation (2.2).

\[
N_d(x) = 1.5 \times 10^{15} + 5.0 \times 10^{18} \{1 - \text{erf} \left( 1.05(15.5-x) \right) \}
\]

(2.2)

\( x \) is in microns.

2.2.2 THE DOPING DISTRIBUTION IN \( P^+ N \) REGION

There are two main steps involved during the diffusion of Boron...
in to the epilayer:-

(1) Pre-deposition at 1075° C for 32 min.

Diffusion coefficient of Boron at 1075° C = \(6.0 \times 10^{-13} \text{ cm}^2 \text{ Sec}^{-1}\)

The diffusion length, \(2\sqrt{Dt}\) = \(0.675 \times 10^{-4} \text{ cm}\)

(2) The Drive in at 1110° C for 24 hours

Diffusion coefficient of Boron at 1110° C = \(3.6 \times 10^{-13} \text{ cm}^2 \text{ Sec}^{-1}\)

The diffusion length, \(2\sqrt{Dt}\) = \(3.53 \times 10^{-4} \text{ cm}\)

As indicated in the Fig 2.3, during pre-deposition a fixed amount of the impurity is contained within a very narrow region very close to the surface. The concentration at the surface approaches the solid solubility of \(3.0 \times 10^{20} \text{ cm}^{-3}\) for Boron in Si. Since the ratio \(\frac{2\sqrt{Dt}}{(2\sqrt{Dt})_{\text{pre-deposition}}} \gg 1\), the extent of the penetration of pre-deposition profile can be neglected, since it can be regarded as negligibly small in comparison to that of the final profile resulting after the drive in diffusion. "Drive in" is the process of reducing the surface concentration and moving the impurities deep into the semiconductor without at the same time increasing the total number of impurities within the material. The final distribution is Gaussian type given by the equation (2.3).

\[
N_a(x) = N_1 e^{-x^2/4Dt} \quad (2.3)
\]

Where \(N_1\) is a quantity determined by the pre-deposition step.

\(4Dt\) is the square of the diffusion length in the final drive in step.

\[
N_a(x) = N_1 e^{-8.0 \times 10^6 x^2} \quad (2.4)
\]

Where \(1/4Dt = 8.0 \times 10^6\)

Determination of \(N_1\):- During the drive in process the time of diffusion
is optimised to have junction at \( x = 12.3 \times 10^{-4} \) cm from the surface, ie at \( x = 12.3 \times 10^{-4} \) cm \( \text{Na}(x) = 1.5 \times 10^{15} \text{ cm}^{-3} \), we make use of this value of the acceptor doping to find the value of \( N_1 \) from equation (2.4):

\[
N_1 = \frac{1.5 \times 10^{15}}{e^{-8.0 \times 10^6 (12.3 \times 10^{-4})^2}}
\]

\[
= 2.7 \times 10^{20} \text{ cm}^{-3}
\]

Now the final expression for the doping distribution of the acceptors in the \( P^+ \) region can be written as in equation (2.5).

\[
\text{Na}(x) = 2.7 \times 10^{20} e^{-8.0 \times 10^{-2} x^2} \quad (2.5)
\]

\( x \) is in microns.

Now we are in a position to calculate the predicted doping distribution in the device, making use of the equations (2.2) and (2.5). In Fig 2.4 we have plotted the doping distribution for a deep diffused \( S \) Band Trapatt device. Under section 2.5 we will arrive at an approximate distribution derived from this one to compute a C-V plot to compare with the experimental plot. The comparison of the two plots will give us an idea about the accuracy of our predicted doping profile.

### 2.3 THE MEASUREMENT OF THE C-V PLOT

Most of the device details can be obtained by the measurement of its depletion capacitance for a known applied reverse bias voltage. The accuracy of the measurement critically depends on the value of the device area. Most commercial capacitance meters and profiling instruments measure the capacitance of the device under test by applying a constant \( \text{rf.} \) voltage and monitoring the imaginary component of the resulting \( \text{rf.} \) current. (ie the component \( 90^\circ \) out of phase with the drive voltage). The imaginary component of the \( \text{rf.} \) current is then directly proportional to the depletion
Fig 2.4: PREDICTED DOPING DISTRIBUTION OF THE DEVICE

\[ Na(x) = 2.7 \times 10^{20} e^{-8.0 \times 10^{-2} x^{2}} \]

\[ Nd(x) = 1.5 \times 10^{15} \{1 - \text{erf}(1.05(X_{o} - x))\} \leq 10^{18} \]
layer capacitance.

For the purpose of our measurement a Boonton Capacitance Meter Model 72B was used, which provides 15v rms at 1 MHz. The measurement was done at Philips Research Lab, Redhill. To set the electrical zero of the meter an open circuit $S_4$ package was used which automatically calibrates the meter for other parasitics etc. The subsequent measurements gave the true value of the depletion layer capacitance. The X-Y recorder was calibrated and then the voltage was scanned from its zero value to the breakdown value. The C-V plot obtained is shown in Fig 2.5 trace a.

From the measured C-V plot it is possible to obtain the doping distribution, punch through voltage, the breakdown voltage and the width of the depletion layer for one sided abrupt and linearly graded junctions.

2.4 THE INTERPRETATION OF THE MEASURED C-V PLOT

The main purpose of this section is to use the measured C-V plot and to calculate the required parameters, however, before we can do this, it is necessary to ensure the exact nature of the junction. The junction can fall in either of the following three categories:

(a) Abrupt junction.
(b) Linearly graded junction.
(c) Graded junction.

For the cases (a) and (b) above exact details are readily available and standard expression and approaches given in any relevant text books may be used, however for the case in (c) we will have to develop the expressions later in the chapter.

For one sided abrupt junction, in which the doping density on one side of the junction is relatively very much higher than the other side in such a way that a step distribution across the junction can be
(a). Measured C-V plot.

(b). Computed C-V plot.

Fig 2.5: MEASURED AND COMPUTED C-V PLOTS
The capacitance per unit area is given by the equation (2.6):

\[ C = \left( \frac{e \varepsilon_0 N}{2 (V_{bi} + V)} \right)^{1/2} \text{PF/cm}^2 \]  

(2.6)

where

\[ e = \text{electronic charge, } 1.6 \times 10^{-19} \text{ Coulomb} \]
\[ \varepsilon_0 = \text{dielectric constant of Si, } 1.06 \times 10^{-12} \text{ Farad/cm} \]
\[ N = \text{Background doping density in cm}^{-3} \text{ in the lightly doped region.} \]
\[ V_{bi} = \text{Built in potential in Volt} \]
\[ V = \text{Applied reverse bias voltage in Volts} \]
\[ A = \text{Area of the device in cm}^2 \]

The equation (2.6) can be rewritten in a more convenient form as in (2.7):

\[ \frac{d(1/C^2)}{dv} = \frac{2}{e \varepsilon_0 N} \]  

(2.7)

The equation (2.7) suggests that a plot of \(1/C^2\) versus \(V\) will be a straight line if \(N\) is constant. The slope of the line gives the doping concentration and the intercept at \(1/C^2 = 0\) gives the built in potential \(V_{bi}\).

The built in potential can be defined as the potential set up in between the depletion layer of a P-N junction when P and N type materials are brought together. This potential is established as a consequence of the alignment of the fermi levels of the two materials. For one sided abrupt junction this potential depends upon the background doping density of the lightly doped side. For one sided abrupt junction the slope of \(\ln C\) versus \(\ln V\) should be \(-1/2\).

**Linearly Graded Junction** :- For a linearly graded junction the doping across the depletion layer varies linearly with a constant slope of the doping density with the distance, the depletion layer capacitance is given by the equation (2.8):

\[ C = \left( \frac{ea (\varepsilon_0)^2}{12 (V_{bi} + V)} \right)^{1/3} \text{PF/cm}^2 \]  

(2.8)
Where $a$ is the doping gradient in cm$^{-4}$ at the junction, the equation (2.8) can be rewritten as in (2.9):

$$d(1/C)^{1/3}/dv = \frac{12}{ea(\varepsilon\varepsilon_0)^2}$$

(2.9)

The equation (2.9) suggests that a plot of $1/C^3$ versus $V$ should be a straight line and the slope of the line should give the doping gradient $a$ at the junction. The slope of $\ln C$ versus $\ln V$ should be $-1/3$ for a linearly graded junction.

The width of the depletion layer in either case is inversely proportional to the capacitance, as in the case of a parallel plate condenser:

$$W = \varepsilon\varepsilon_0 A / C$$

(2.10)

Having developed the basic ideas regarding two types of junction, now we will try to establish as to the exact nature of the junction given by the C-V plot. To begin our interpretation, first of all we look at the Fig 2.6 which is a plot of $\ln C$ versus $\ln V$. On this graph we have three lines, marked (a), (b) and (c). The lines (a) and (b) are showing a slope of $-1/2$ and $-1/3$ corresponding to one sided abrupt and linearly graded junction. Curve (c) is a plot of $\ln C$ versus $\ln (V_{bi} + V)$. It is possible to divide the measured curve (c) into three parts, representing straight lines. The first section of the line extends from 0 to 1.10 volts and is marked "A", the second portion extends from "A" to "B" i.e 1.10 to 27.6 volts and the third section extends beyond "B". Comparing the slopes of these lines with (a) and (b), the following results may be concluded:

1. The slope of the first section is approximately $-1/3$ suggesting that the junction behaves almost like a linearly graded junction between 0 and 1.10 volts.

2. The slope of the second section, the line between 1.10 to 27.6 volts is neither $-\frac{1}{3}$ or $-\frac{1}{2}$, this suggests that the junction
Curve c from the measured C-V plot.

Curve d from the Computed C-V plot.

Fig 2.6: A PLOT OF lnV Vs lnC
is neither linearly graded nor one sided abrupt type.

(3) The slope of the third section, the line beyond
27.7v is approximately $-\frac{1}{2}$, this suggests that the junction behaves
almost like one sided abrupt type beyond this point.

The results outlined above indicate that the doping distribution is progressively changing from linear to abrupt and from abrupt to something totally different, for increasing reverse bias voltage. In other words the distribution in the depletion region is not uniform. The reason for this will be explained later in the chapter.

To investigate more, in Fig 2.7 we have plotted three curves (a), (b), and (c) for $W$, $1/C^2$ and $1/C^3$ versus the applied reverse bias voltage. Neither of the two curves (b) and (c) can be approximated to a straight line, this also, suggests that the junction is neither abrupt nor linearly graded. However the curve (a) can be divided into a set of five straight lines, progressively for each section the rate of change of the depletion layer width goes on reducing for increasing applied reverse bias voltage. This situation is plotted in Fig 2.8, an asymptotic behaviour of the curve suggests that the doping density increases rapidly with respect to the distance from the junction. In other words this curve also brings out the same results as was obtained earlier.

Thus it is concluded that the junction can neither be treated as abrupt nor linearly graded for the entire width of the depletion region. A true analysis should therefore treat the junction as graded in such a way that the doping density keeps on changing away from the junction. In section 2.5 we will develop this general approach, however in the following section 2.4.1, the doping density will be calculated using the expression (2.6) and measured values of C.
Fig 2.7: GRAPH OF W, 1/C² AND 1/C³ DERIVED FROM THE MEASURED C-V PLOT.
Fig 2.8: RATE OF CHANGE OF W WITH APPLIED VOLTAGE.
2.4.1 THE CALCULATION OF THE DOPING DISTRIBUTION FROM THE MEASURED C-V PLOT TREATING THE JUNCTION AS ONE SIDED ABRUPT TYPE

The expression (2.6) can be rewritten for Si in terms of the applied voltage and the measured value of C, the area of the device is taken as $3.0 \times 10^{-4}$ cm$^2$, and is given by the equation (2.11).

$$N = 1.31 \times 10^{14} \frac{C}{(V_{bi} + V)} \text{ cm}^{-3} \quad (2.11)$$

Where

- $C$ is the measured Capacitance in PF
- $(V_{bi} + V)$ is the total applied voltage in Volts

$V_{bi}$ is approximately 0.8 volt

The doping density and the corresponding position is calculated using equations (2.11) and (2.10). A plot of the calculated doping profile is shown in Fig 2.9, the plot indicates the presence of very impractical and peculiar situation between 2.3 and 2.8 microns from the junction. This could mean either of the two things, firstly, may be the profile does vary as calculated or secondly the assumption that the junction behaves like one sided abrupt type is incorrect. The latter of the two appears to be the real situation on the basis of the explanations given in the previous sections.

2.5 THEORETICAL CALCULATIONS OF THE C-V PLOT

In this section we will develop some theoretical expressions for an unknown doping distribution in such a way that we can compute the C-V plot for an assumed and predicted distribution. The principles of general approach are reasonably straight forward and normally applied to an ordinary diode having a junction. Following two conditions, under some
Fig 2.9: Doping distribution for one sided abrupt junction from the measured C-V plot.
suitable boundary conditions, form the basis of the work in this section:

1. The total charge on either side of the junction must be equal, i.e., the condition of the charge neutrality must be maintained.

2. The Poisson's equation \((\frac{d^2V}{dx^2} = -\frac{dE}{dx} = -\frac{\rho}{\varepsilon \varepsilon_0})\) must be satisfied on either side of the junction.

![Fig 2.10: SIGN CONVENTION FOR THE FORMULATIONS.]

Referring to Fig. 2.10 for sign notations to be employed for this section, it is assumed that the distances from left to right are increasing with the junction at \(x = 0\). Thus distances in the P region will be negative and positive in the N region.

Let

\[\begin{align*}
NA(x) &= Na(x) - N_o, \text{ be the effective doping distribution of acceptor impurities in the P region, cm}^{-3} \\
ND(x) &= Nd(x) + N_o, \text{ be the effective doping distribution of donor impurities in the N region, cm}^{-3} \\
A &= \text{be the area of the device in cm}^2 \\
\varepsilon \varepsilon_0 &= \text{be the dielectric constant of the material for silicon this value is } 1.06 \times 10^{-12} \text{ Farad/cm} \\
e &= \text{be the electronic charge, } 1.6 \times 10^{-19} \text{ Coulomb} \\
\rho(x) &= \text{be the variation of the charge density, Coulomb cm}^{-3}
\end{align*}\]
be the total charge

The subscripts \( p \) and \( n \) denote the side of the junction being considered. We will restrict the treatment to one dimension only. On the basis of the conditions listed above, we will write the following expression.

\[
\begin{align*}
\rho_p(x) &= -e NA(x) \quad (2.12) \\
Q_p &= -eA \int NA(x) \, dx \quad (2.13) \\
\frac{dE}{dx} \bigg|_p &= \frac{\rho_p(x)}{\varepsilon_0} = -\frac{e}{\varepsilon_0} NA(x) \quad (2.14) \\
\rho_n(x) &= e ND(x) \quad (2.15) \\
Q_n &= eA \int ND(x) \, dx \quad (2.16) \\
\frac{dE}{dx} \bigg|_n &= \frac{\rho_n(x)}{\varepsilon_0} = \frac{e}{\varepsilon_0} ND(x) \quad (2.17) \\
V_p &= -\int E_p(x) \, dx \quad (2.18) \\
V_n &= -\int E_n(x) \, dx \quad (2.19)
\end{align*}
\]

From the condition of the charge neutrality

\[
Q_p = Q_n \quad (2.20)
\]

The equation (2.20) is related to the area under the doping distribution and two areas on either side of the junction must always be equal. By solving equation (2.20) for a known distribution and the length of the depletion layer in one region, the required length to be depleted in the other region can be found. To find the field and hence the total voltage across the junction the equation (2.14),(2.17),(2.18) and (2.19) are to be solved under the following boundary conditions:

1. at \( x=-x_p \) \( \quad V=0 \quad (2.21) \)
2. at \( x=0 \) \( \quad E_p(x) = E_n(x) \quad (2.22) \)

To explain the general procedure, we will work out the
expression for a simplified case. To this we will try to approximate the predicted profile given in Fig 2.4 by a set of exponential expressions of the form \( Ae^{-Bx} \). The values of \( A \) and \( B \) will be found by fitting these exponentials to the actual predicted curves for the P and N regions.

### 2.5.1 APPROXIMATION TO THE PREDICTED DOPING PROFILE

In this section first of all we will try to derive an exponential expression for the acceptor impurities in the P⁺N region, which actually follows a Gaussian distribution. Next a similar expression will be derived for the donor distribution in N⁺N region, which actually obeys a complementary error function law.

(a) **Approximation to the Gaussian distribution:** The actual distribution is given by \( 2.7 \times 10^{20} e^{-8.0 \times 10^6 x^2} \). At \( x = 12.3 \times 10^{-4} \) cm from the surface where the metallurgical junction is formed, the doping density is equal to the background doping of \( 1.5 \times 10^{15} \) cm\(^{-3}\). Following the sign notation outlined earlier in the section, we will take, \( x = x_j = 12.3 \mu m \) as the reference starting point and will call \( x = 0 \).

\[
Na(x) = Ae^{m_1 x}
\]

at \( x = 0 \), \( Na(0) = 1.5 \times 10^{15} = N_0 \)

therefore \( A = N_0 = 1.5 \times 10^{15} \) cm\(^{-3}\).

From equation (2.23) various values of \( m_1 \) are found for known values of \( Na(x) \) at different values of \( x \) calculated from the Gaussian distribution. A simple arithmetic mean of all such values of \( m_1 \) is taken for the approximate distribution given by (2.23).

\[
m_1 = \frac{\ln \left( \frac{Na(x)}{N_0} \right)}{x} \quad (2.24)
\]

\[
m_1 = \frac{\sum m_1}{N} \quad (2.25)
\]
Following the steps given above in the vicinity of the junction we find that the average value of \( m_j = 1.9 \times 10^4 \text{ cm}^{-1} \) gives a reasonable fit to the predicted curve Fig 2.4. Therefore an approximate expression for the distribution of the acceptor impurities may be written as in (2.26), however for the chosen sign notation in the left hand side of \( x_j \), \( x \) is negative. Therefore equation (2.27) will give the actual distribution in our case.

\[
Na(x) = N_o e^{1.9166 \times 10^4 x} \quad (2.26)
\]

\[
Na(x) = N_o e^{-1.9166 \times 10^4 x} \quad (2.27)
\]

(b) Approximation to the Complementary Error Function Distribution :- The most general form of the exponential expression which can fit the donor distribution due to out diffusion can be written as in expression (2.28)

\[
Nd(x) = N_o (e^{m_2(x-d)} - 1) \quad (2.28)
\]

where

\( d \) is the distance where \( Nd(x) = N_o \), from the Fig 2.4 \( d \) is approx 0.4 \( 10^{-4} \) cm. The value of \( m_2 \) is calculated from the equation (2.29).

\[
m_2 = \frac{\ln((Nd(x)+N_o)/N_o)}{N_o (x-d)} \quad (2.29)
\]

From the actual distribution for various values of \( x \), the values of \( Nd(x) \) are found and substituted in equation (2.29), and thus various values of \( m_2 \) are found. A mean of these is taken as the suitable value. The value of \( m_2 \) which fits the distribution properly is found to be 3.4 \( 10^4 \text{ cm}^{-1} \). Therefore the distribution of \( Nd(x) \) will be given by equation (2.30).

\[
Nd(x) = N_o (e^{3.4 \times 10^4 (x-0.4)} 10^{-4} - 1) \quad (2.30)
\]
2.5.2 THE CALCULATION OF THE C-V PLOT

In this section some useful expressions will be developed to calculate the width of the depletion layer, the field distribution and the voltage across the device. On the basis of these expressions a C-V plot will be computed.

The model taken for calculations is shown in Fig 2.11, the edges of the depletion layer extend from \( x = -x_p \) to \( x = -x_n \). On the diagram the expressions for the acceptor and donor distributions have been given, however, for the purpose of formulation the expressions for effective distributions in the P and N regions given by (2.31) and (2.32) will be used.

\[
NA(x) = Na(x) - N_o = N_o \left( e^{-m_1 x} - 1 \right) \quad (2.31)
\]

\[
ND(x) = N_o + Nd(x) - Na(x) = N_o \left( e^{m_2(x-d)} - 1 \right) + N_o \left( 1 - e^{-m_1 x} \right) \quad (2.32)
\]

The equation (2.33) can be written by combining the condition of the charge neutrality and the expressions (2.13) and (2.16), \( \forall x > d \)

\[
-x_p \int_{-x_p}^{x_n} N_o(e^{-m_1 x} - 1) \, dx = \int_{0}^{x_n} N_o \, dx + \int_{x_n}^{x_n} N_o \left( e^{m_2(x-d)} - 1 \right) \, dx \quad (2.33)
\]

Finally the equation (2.33) can be expressed as the equation (2.34).

\[
e^{\frac{m_1 x_p}{p}} - m_1 x_p = K \quad (2.34)
\]

Where

\[
K = e^{-x_n} + m_1 x_n + m_1 \left( (e^{m_2(x_n-d)} - 1)/m_2 - (x_n-d) \right) \quad (2.35)
\]
Fig 2.11: ASSUMED MODEL FOR THE CALCULATION OF THE C-V PLOT.

\[ N_a(x) = N_0 e^{-m_1 x} \]

\[ N_d(x) = N_0 (e^{m_2 (x-d)} - 1) \]
For a suitably chosen value of $x_n$, the value of $K$ is found from the expression (2.35). The corresponding value of $x_p$ is found by solving the equation (2.34) for the previously calculated value of $K$. Thus the width of the depletion layer $W$ is determined.

The field distribution in the P and N regions are given by the expressions (2.36) and (2.37) which are derived from the equations (2.14) and (2.17). Detailed steps in obtaining the equations (2.36) and (2.37) are given in the appendix "A".

\[
E_p(x) = -\frac{N_0 e}{\varepsilon_0} \left( e^{-\frac{m_1 x}{m_p}} - x - \frac{e}{m_p} - x \right) \quad (2.36)
\]

\[
E_n(x) = -\frac{N_0 e}{\varepsilon_0} \left( e^{-\frac{m_1 x}{m_n}} - x - \frac{1}{m_n} - \{x + \frac{1}{m_1}(e^{-m_1 x} - 1) + \frac{1}{m_2}(e^{m_2(x-d)} - 1) - (x-d)\} \right) \quad (2.37)
\]

To determine the total voltage across the junction the equations (2.18) and (2.19) are used and the integration is performed within the limits $x_n$ to $-x_p$.

\[
V_{x_n} \quad \text{w.r.to} \quad -x_p = \int_{-x_p}^{x_n} E(x) \, dx
\]

\[
= -\int_{-x_p}^{x_n} E_p(x) \, dx + \int_{-x_p}^{x_n} E_n(x) \, dx
\]

\[
= -(v_p + v_n)
\]

The equation (2.38) gives the voltage across the junction $V_T$. This equation has been derived from the equations (2.36) and (2.37).

\[
V_T = \frac{N_0 e}{\varepsilon_0} \left[ \frac{m_1 x_n}{m_1} + \frac{1}{-m_1 x_n - e^{-m_1 x_n} - e^{-m_1 x_p}} \right] - \frac{W^2}{2} - \frac{1}{2} \left( \frac{m_2(x_n-d)}{(e - 1) - m_2(x_n-d) - \frac{m_2^2(x_n-d)^2}{2}} \right) \quad (2.38)
\]
To obtain the C-V plot the equations (2.34) to (2.38) and (2.10) were solved. On the basis of these equations a computer program given in the appendix "B" was developed. The computed C-V plot is shown in the Fig 2.5 (curve marked (b)).

A very good agreement between the computed and the experimental C-V plots exists at the two extreme ends of the applied reverse bias voltages. For voltages up to 25V the computed values of C are lower (W values higher) than the experimental values. This means that the depletion regions for these voltages are more heavily doped than predicted, however the attempts were made to account for this deviation and it was found that the boundary conditions at either end of the applied voltages were not satisfied i.e. the C values at zero bias and W at the breakdown voltage. To obtain a better agreement various C-V plots were computed for different values of \( N_0 \), \( m_1 \), \( m_2 \) and \( d \), the best result was obtained for \( N_0 = 1.5 \times 10^{15} \text{ cm}^{-3} \), \( m_1 = 1.9 \), \( m_2 = 3.4 \) and \( d = 0.3 \mu\text{m} \). It is thought that the region of somewhat strange behaviour is due to the stray effects which may be due to the bonding and heatsinking. It is therefore concluded that the two C-V plots relate to the similar device and therefore the predicted doping distribution should represent the actual distribution in the device.

The curve (d) in Fig 2.6 derived from the computed C-V plot indicates that the junction is neither one-sided abrupt type nor linearly graded. The slope of the curve is \(-0.4275\).

Similar to Fig 2.7, three curves have been plotted in Fig 2.12. These curves (\( W, 1/C^2 \) and \( 1/C^3 \) verses the applied voltage) have been derived from the computed C-V plot. These curves also indicate that the junction is neither abrupt nor linearly graded. The general pattern of these curves is in close agreement with the curves in Fig 2.7.
Fig 2.12: GRAPH OF W, 1/C^2 AND 1/C^3 DERIVED FROM THE COMPUTED C-V PLOT.
Two sets of curves have been plotted in Fig 2.13. Fig 2.13(a) is the predicted doping distribution for which the C-V plot was computed and the curve (b) has been derived from this C-V plot (Fig 2.5(b)), assuming the junction to be one sided abrupt type. It can be seen from the two curves that there is no resemblance between them and this is because of the incorrect assumption. It is therefore essential to know the nature of the junction prior to analysing the C-V plot, otherwise the results derived could be incorrect. This really proves that if a graded junction is treated as abrupt then the calculated doping profile can be totally different and this does explain that the results reported in Fig 2.9 can not be relied upon.

From the analysis given in this chapter it is concluded that the doping profile in our device is graded.
Fig 2.13 : (a) PREDICTED DOPING DISTRIBUTION.

(b) DOPING DISTRIBUTION FOR ONE SIDED ABRUPT JUNCTION
FROM THE COMPUTED C-V PLOT.
CHAPTER 3

D.C. ANALYSIS OF THE DEVICES

The main purpose of this analysis is to show whether the Trapatt device discussed in the second chapter does have d.c. differential negative resistance. In the literature it has been said that Trapatt devices have neither d.c. nor small signal negative resistance at the fundamental frequency. The main question which the previous researchers were trying to answer was how the Trapatt oscillations begin in the absence of any small signal negative resistance. Ultimately they reached the conclusion that the final Trapatt oscillations are triggered by initial Impatt oscillations, which do not require the d.c. negative resistance as such. To support their claims various experimental evidences were produced. It is our intention, therefore, first to establish in this and the next chapter whether our device has differential negative resistance at d.c.

3.1 THE JUNCTION BREAKDOWN

Perhaps the most important phenomena, which forms the basis of almost all the avalanche type microwave semiconductor devices is the junction breakdown, when operated under reverse bias condition. Strutt\(^{(36)}\) has given a detailed description of various breakdown mechanisms which can be present in a junction device. In the discussion here we will be mainly concerned with the breakdown due to "The Impact Ionization or Avalanche Multiplication". The ultimate condition of the avalanche breakdown can be best understood by analysing various scattering mechanisms which take place, under the influence of increasing electric field, between the charge carriers
and the acoustical and optical phonons. Phonons are the quantum mechanical particles defining the wave motion in the lattice. The lattice behaviour begins to change interestingly when the charge carrier temperature overtakes the lattice temperature.

3.2 DRIFT VELOCITY VERSUS THE APPLIED FIELD

The dependence of carrier drift velocity on the applied electric field is given in Fig 3.1. In the beginning the drift velocity increases linearly with the applied field because this velocity is smaller than the thermal velocity. The combined effect of the acoustical phonon and ionised impurity scattering on the mobility is to keep it significantly constant. During this process the carrier remains in thermal equilibrium with the lattice, however, for still higher fields the carriers become "warm" and the drift velocity is found to be proportional to the square root of the applied field \( \sqrt{E} \). For Si this situation begins at about \( 2 \times 10^3 \) Volt Cm\(^{-1}\) and continues upto \( 20 \times 10^3 \) Volt Cm\(^{-1}\). During these limits the carrier temperature is approximately twice the lattice temperature, the carrier velocity is about three times the sound velocity in Si and the mobility drops by 30% of its low field value. The main scattering mechanism during this process is due to the optical phonons. Two situations discussed so far have been marked as region 1 and 2 in Fig 3.1.

For applied fields greater than \( 20 \times 10^3 \) Volt Cm\(^{-1}\) the velocity of electron is found to saturate as shown in Fig 3.1 (region 3). The saturation in the drift velocity for high fields results from the interaction of the carriers with the optical phonons. These are not excited in thermal equilibrium due to the large energy required to do so. The probability of emitting an optical phonon increases when an electron acquires an energy greater than the optical phonon energy \( E_p \). This value for Si is 0.063 ev. Gibbons (38) has shown that the high field mobility is inversely proportional
Fig. 3.1: DRIFT VELOCITY VERSUS APPLIED ELECTRIC FIELD

FOR SILICON
to the external field and the drift velocity given by approximately the square root of the ratio of the optical phonon energy to the effective mass \( \sqrt{E_p/m} \). The concept of saturated drift velocity is a useful material property for avalanche devices.

The energy acquired by the electrons from the external field goes on increasing beyond \( E_p \) and for fields in excess of \( 1.0 \times 10^5 \) Volt Cm\(^{-1}\) an interesting process begins to take place in such a way that the carriers begin to multiply by a process known as "Avalanche Multiplication or Impact Ionization". The ionization can occur at smaller fields but the multiplication will be small. In this process, as the electrons gain more energy the velocity saturation mechanism due to the optical phonon scattering is bypassed and a sudden increase in current is observed. Eventually the electrons gain so much of energy that the collision with the atoms take place, Si to Si bond is broken, and electrons are knocked off the crystal atom and thus an electron–hole pair is generated. In this process electrons are "pulled off" the valence band in the presence of very high fields. The generation rate of electron hole pairs is given by the equation (3.1).

\[
G = \alpha_e n v_e + \beta_h p v_h
\]  

(3.1)

Where \( \alpha_e \) and \( \beta_h \) are the electron and hole ionization rates defined as the average number of electron hole pair generated by an electron or hole in travelling a unit distance in the lattice. Both these factors are strongly dependent on the applied field, \( n \) and \( p \) are the number of initial electrons and holes taking part in the ionization process, \( v_e \) and \( v_h \) are the electron and hole velocities.

3.3 THE AVALANCHE BREAKDOWN

The sequence of events leading to the avalanche breakdown can
be best understood by referring to Fig 3.2(b). The atomic structure of Si shown in Fig 3.2(a) gives an indication of electron-hole movement, when an electric field is applied. There are five distinct steps marked on the band diagram, Fig 3.2(b), namely 1, 2, 3, 2' and 3'. The process starts at the step 1, when an electron is pulled off the valence band leaving behind a hole in its place and thus creating an electron-hole pair. In the presence of high electric field these carriers acquire enough kinetic energy and are able to break Si to Si atomic bonds leading to creation of new electron-hole pairs. The process of gaining enough kinetic energy is depicted by the step 2 and 2'. In the step 3 the energy acquired by the electron in the conduction band is given to the valence electron (on average this energy is 1.5 times the band gap) and thus the valence electron is pulled up in to the conduction band. A similar process is carried out by the energetic hole shown by the step 3'. This new pair also takes part in the ionization process and thus the multiplication of carriers takes place. Since the electrons and holes drift in the opposite directions the current increases very rapidly during the avalanche breakdown. On average during the collision the momentum of the accelerated electron is redistributed among the two electrons and hole as shown in Fig 3.2(c) and by the equation (3.2). The kinetic energy is redistributed as the potential energy of the new pair and three kinetic energies as given by equation (3.3).

The law of conservation of momentum gives

$$m_e v_0 = m_e v_1 + m_e v_2 + m_h v_3$$

(3.2)

The law of conservation of energy gives

$$\frac{1}{2}m_e v_0^2 = \frac{1}{2}(m_e v_1^2 + m_e v_2^2 + m_h v_3^2) + E_g$$

(3.3)

These equations assume that there is no net loss of energy
Fig. 3.2(a) : THE ATOMIC STRUCTURE OF SILICON

Fig. 3.2(b) : SCHEMATIC ILLUSTRATION OF THE BREAKDOWN PROCESS

Fig. 3.2(c) : THE GENERATION OF ELECTRON-HOLE PAIR
in this process. So far only the physical picture of the breakdown process was developed, however, in the next section the field dependence of the ionization rates and the breakdown condition given by the ionization integral will be discussed.

3.3.1 THE FIELD DEPENDENCE OF THE IONIZATION RATES

The exact relationship between the ionization rates and the applied field is fairly complex, various researchers have tried to develop simpler model to explain their approach and results. The analytical treatment of the avalanche devices to a large extent depends on the type of the approximations assumed for the ionization rates. We will deal with an exponential model suggested by Sze and Gibbons (39), though in the literature simpler models such as zero ionization rates up to certain critical field and thereafter a constant value has been assumed by De Loach (7) et al. to develop a physical understanding of the Trapatt device. Three parameter theory of Baraff (40) first proposed in 1962 and later developed by Crowell and Sze (41) in 1966, gives a better approximation of the ionization rates. The parameters involved in the theory are functions of the lattice temperature and the physical properties of the material. The exponential model of the ionization rates for electrons and holes in Si can be expressed as in equation (3.4) and (3.5).

\[ \alpha_e = 3.8 \times 10^6 e^{-1.75 \times 10^6 E^{-1}} \]  \hspace{1cm} (3.4)

\[ \beta_h = 2.25 \times 10^7 e^{-3.26 \times 10^6 E^{-1}} \]  \hspace{1cm} (3.5)

The parameters in the above expressions are measured at the room temperature, however, Crowell's (41) approach can be used to determine
these parameters at any other temperatures. These expressions for $\alpha_e$ and $\beta_h$ will be used later in the chapter to develop a computer program for the d.c analysis of the Trapatt devices.

### 3.3.2 THE AVALANCHE BREAKDOWN CONDITION

In the absence of significant ionization, the reverse saturation current remains constant, however, as the ionization process proceeds this current begins to rise rapidly. The ratio of the reverse current during impact ionization to the reverse saturation current is defined as "The multiplication factor $M". The value of $M$ approaches infinite when the junction breakdown occurs. Miller (42) has suggested an empirical formula to calculate the value of this multiplication factor given by equation (3.6).

\[
M = \left( 1 - \left( \frac{V_{rev}}{V_B} \right)^n \right)^{-1}
\]  

(3.6)

Where

- $V_{rev}$ is the applied reverse bias voltage
- $V_B$ is the breakdown voltage

The values of $n$ for Si are suggested to be between 1.5 to 2 for high resistivity P material and 3.5 to 4 for high resistivity N material. A similar condition can be derived in terms of the ionization rates and the charge density as a result of the impact ionization, however, if equal ionization rates for electrons and holes are assumed the expression obtained for $M$ can be easily understood. Strictly the condition of equal ionization rates is true in the case of GaAs and GaP but not so in the case of Ge and Si. Sze (39) has given an expression for the multiplication factor for unequal ionization rates.

In the absence of recombination, a p-n junction can be represented by Fig 3.3. Assuming equal ionization rates i.e $\alpha = \alpha_e = \beta_h$, 

Fig. 3.3: $W$ is the width of the Depletion layer, $x_p$ and $x_n$ are the boundaries of the depletion layer in P and N regions, $n_o$ is the number of electrons per unit time entering in the depletion layer at $x_p$, $n$ is the number of electrons per unit time leaving the depletion layer at $x_n$, $E$ is the electric field.
the number of electrons generated per unit time within \( x \) and \( x+dx \) can be expressed as in the equation (3.7)

\[
dn_1 = (n_0 + n_1)dx + n_2 dx = n dx
\]

(3.7)

Where

- \( n_0 \) is the number of electrons entering the depletion layer at \( x = x_p \) per unit time.
- \( n_1 \) and \( n \) are the electrons generated between \( x_p \) and \( x \) and \( x+dx \) and \( x_n \).
- \( n = n_0 + n_1 + n_2 \) are the number of electrons leaving the depletion at \( x=x_n \).
- \( n_2 dx \) Number of pairs generated by holes coming from right to left.

The multiplication factor in this case is defined as the ratio of \( n \) to \( n_0 \) and is given by the equation (3.8).

\[
M = \frac{n}{n_0}
\]

(3.8)

Total number of electrons generated within the width \( W \) can be expressed as in equation (3.9).

\[
\int_{x_p}^{x_n} dn_1 = n \int_{x_p}^{x_n} dx = n - n_0
\]

(3.9)

Therefore

\[
M = \frac{n}{n_0} \left( 1 - \int_{0}^{W} dx \right)
\]

(3.10)

For the multiplication factor to be infinite the integral in the denominator of the equation (3.10) must be equal to one. Therefore
the equation (3.11) defines the condition of the avalanche breakdown in a p-n junction.

\[
\frac{W}{f} \alpha \int_{0}^{x} dx = 1 \tag{3.11}
\]

3.4 FORMULATIONS OF BASIC EQUATIONS

The analysis given here is on the basis of the model developed by Bower. The sign conventions are outlined in Fig 3.4. To obtain the d.c. characteristics of the device the Poisson and Continuity equations are simultaneously solved, under appropriate boundary conditions. Equations (3.4) and (3.5) are used to account for electron and hole ionization rates.

The dependence of the drift velocity on the electric field has been assumed to be given by equations (3.12) and (3.13) as suggested by Prior.

\[
v_e = 4.5 \times 10^5 E^{0.1525} \tag{3.12}
\]

\[
v_h = 4.68 \times 10^4 E^{0.4445} \tag{3.13}
\]

The equations (3.12) and (3.13) are applicable for fields \( E > 3.5 \times 10^3 \) Volt Cm\(^{-1}\), for fields less than this value usual linear relationship between the drift velocities and the applied field is assumed.

In the analysis given in this section the main aim is to derive a couple of equations for electron current and the electric field at any point in the depletion layer. The Poisson equation at a distance \( x \) can be expressed as in equation (3.14), and the Continuity equations for electrons and holes as in equations (3.15) and (3.16).

\[
\frac{dE}{dx} = \frac{e}{\varepsilon_0} \left( ND - NA + p - n \right) \tag{3.14}
\]
Fig. 3.4: SIGN CONVENTIONS FOR THE ASSUMED MODEL
\[ \frac{\delta n}{\delta t} = \frac{1}{e} \frac{\delta J}{\delta x} + G \]  \hspace{1cm} (3.15)

\[ \frac{\delta p}{\delta t} = -\frac{1}{e} \frac{\delta J}{\delta x} + G \]  \hspace{1cm} (3.16)

Where

\( N_D \) and \( N_A \) are the donor and acceptor doping densities.
\( p \) and \( n \) are the hole and electron densities.

\( G = a_e |v_e|n + \beta_h |v_h|p \) is the generation rate.

\( J_n = -ev_en = -e|v_e|n \) is the electron current. \hspace{1cm} (3.17)

\( J_p = ev_hp = -e|v_h|p \) is the hole current. \hspace{1cm} (3.18)

\( J_T = -(J_n + J_p) \) is the total current.

\[ |v_e| = +v_e \] because electrons drift in the +ve x direction.

\[ |v_h| = -v_h \] because holes drift in the -ve x direction.

\[ |J_T| = -(J_n + J_p) \]

For d.c conditions \( \frac{\delta n}{\delta t} = \frac{\delta p}{\delta t} = 0 \), therefore equations (3.15) and (3.16) can be rewritten as:

\[-\frac{1}{e} \frac{\delta J}{\delta x} = a_e |v_e|n + \beta_h |v_h|p\]

\[\frac{1}{e} \frac{\delta J}{\delta x} = a_e |v_e|n + \beta_h |v_h|p\]

Substituting the values of \( n \) and \( p \) from equations (3.17) and (3.18) in equation (3.14) the following equation (3.19) is obtained:

\[ \frac{dE}{dx} = \frac{e}{\varepsilon_0} \left( N_D - N_A - \frac{J_p}{e|v_h|} + \frac{J_n}{e|v_e|} \right) \]  \hspace{1cm} (3.19)
Equation (3.20) is obtained by adding and subtracting $J_n/e|V_h|$ to the right hand side of the above equation within the bracket.

$$\frac{dE}{dx} = \frac{e}{\varepsilon_0} \{ ND - NA + \frac{J_n}{e} \left( \frac{1}{|V_e|} + \frac{1}{|V_h|} \right) + \frac{|J_T|}{e|V_h|} \}$$

(3.20)

For d.c conditions the continuity equation after adding and subtracting $\beta_n J_n$ results in equation (3.21)

$$\frac{\delta J_n}{\delta x} = J_n (e - \beta_n) - \beta_n |J_T|$$

(3.21)

The equations (3.20) and (3.21) are integrated between an initial point and some other point in the depletion region under the following boundary conditions. The equation (3.22) gives the total voltage.

The voltage = $\int_{x_i}^{x_f} Edx$

(3.22)

3.4.1 THE BOUNDARY CONDITIONS

The integration is started at some initial point $x_i$ where the field is assumed to be $15.07 \times 10^3$ Volt Cm$^{-1}$ as suggested by Misawa. At this small field it is assumed that practically no avalanche multiplication takes place. The integration is terminated at some point $x_f$ where once again the field is small. Similar conditions are imposed on current at $x_i$ and $x_f$. These boundary conditions are listed below:

at $x = x_i$
- $E(x_i) = -E_c = 15.07 \times 10^3$ Volt Cm$^{-1}$
- $J_p(x_i) = J_T$, Total current
- $J_n(x_i) = J_{ns}$, Electron saturation current

at $x = x_f$
- $E(x_f) \ll E_c$
- $J_p(x_f) = J_T$
- $J_n(x_f) = 1.0 \times 10^{-4} J_T$
3.4.2 THE DEVELOPMENT OF THE COMPUTER PROGRAM

A Fortran program was developed to solve the equations (3.20) and (3.21) for various values of total terminal current $J_T$. $N_0$, $N_1$, $N_2$, $A_0$, $B_0$, $e$, $e_0$, $e_1$, $\mu$, $x_0$ (in the program), $E_c$, $D_y$ (increment in $x$), $D_x$ (increment in the initial point), $x_1$ (the initial point) were provided as data in a file.

For different values of $J_T$, the program calculates the voltage using equation (3.22) and the field distribution between $x_1$ and $x_f$. The equation defining the doping distribution can have any general form. Runge-Kutta method of approximation has been used to calculate the values of field and an iterative process to select a suitable value of $x_1$. The listing of the program is given in the appendix "C".

3.4.3 INTERPRETATION OF THE COMPUTED RESULTS

Computed I-V characteristics is plotted in Fig 3.5 and does not indicate the presence of a "knee" or the "turning over" point, it is therefore concluded that this analysis does not suggest that these devices have negative resistance at d.c. Fig 3.5 indicates that upto $2 \times 10^3 \text{ Amp Cm}^{-2}$ the voltage does not change much but beyond this value almost exponential rise in the voltage is observed.

To demonstrate the effect of increasing $J_T$ on the field and the width of the depletion layer, Fig 3.6 has been plotted for very low and high current densities (100 Amp Cm$^{-2}$ and $20 \times 10^3 \text{ Amp Cm}^{-2}$). The field profile for 100 Amp Cm$^{-2}$ closely resembles with the field profile for a PIN type device. The "sagging" observed in Fig 3.6 at $20 \times 10^3 \text{ Amp Cm}^{-2}$ gives an impression that probably at still higher currents the device may show some negative resistance. As a general rule it can be said that if the field depression in the middle is less than the increase at the edges then the device will exhibit a positive resistance, however, if the lowering
Fig 3.5: COMPUTED I-V CHARACTERISTICS OF THE DEVICE
Fig 3.6: COMPUTED FIELD PROFILE OF THE DEVICE

(a) $J_T = 100 \text{ Amp Cm}^{-2}$

(b) $J_T = 20 \text{ K.Amp Cm}^{-2}$
of the field in the centre is more than the increase at the edges, the
device will exhibit negative resistance. In terms of the area under the
field distribution curve, it can be stated that if the net area decreases
for increasing current the device will exhibit negative resistance. This
was checked by increasing the current density upto $80 \times 10^3$ Amp Cm$^{-2}$ and still
no negative resistance was observed. Because of the doping distribution at
either ends of the junction, the width of the depletion layer keeps on
increasing with the current and thus the dip in the middle does not increase
appreciably. This result is not really surprising since the measured C-V
plot reported in the second chapter (Fig 2.5) did not show a real tendency
of saturation for increasing voltage.

3.5 MODIFICATIONS

It has been suggested by Lee et al.\cite{44} that a considerable
improvement in small signal negative conductance can be achieved if the
values of ionization rates corresponding to 200°C are used. The values
reported by them for Si were incorporated in the program which meant changing
the expressions for $A_p, B_T, P_A$ and $P_B$ ( $A_p = 1.8 \times 10^6 \text{ Exp}(-1.64 \times 10^6 \text{ } \text{E}^{-1}$ and
$B_T = 1.0 \times 10^7 \text{ Exp}(3.2 \times 10^6 \text{ } \text{E}^{-1}$ ). The results obtained were similar to one
already reported except that the voltages were slightly higher than in the
previous case.

Thus it can be concluded that the device does not have negative
resistance at d.c as a result of the space charge effect.
In general an oscillator circuit consisting of an active device, resonant circuit and the load may be represented by Fig 4.1. In the Fig 4.1 \(-R_d + jX_d\) represents the device impedance and \(R_L + jX_L\) the load impedance when the resonant circuit is properly matched to load and the device. If it is assumed that the current \(I\) flows in the circuit, then the condition (4.1) must be satisfied for a steady state operation of a free running oscillator.

\[
I \left( -R_d + jX_d \right) + I \left( R_L + jX_L \right) = 0 \tag{4.1}
\]
\[
-R_d + R_L = 0 \tag{4.2}
\]
\[
X_d + X_L = 0 \tag{4.3}
\]

The equation (4.2) in general determines the amplitude whereas the equation (4.3) determines the frequency of the oscillator.

Fig. 4.1 : TWO TERMINAL NETWORK REPRESENTING THE OSCILLATOR CIRCUIT
In equation (4.1) the resistance of the device is shown negative to indicate that the power is coming out of the device and it is working as a generator. It is a fundamental condition and the device must have negative resistance at the operating frequency.

In the previous chapter attempts were made to show if the device in use had any negative resistance down to d.c. The results of the theoretical analysis were inconclusive and it was decided, therefore, to check up experimentally the presence of negative resistance. The experiments conducted can be classified into three main parts:

1. The Quasistatic Analysis.
2. The Negative Resistance at VHF.

4.1 THE QUASISTATIC ANALYSIS

In this part of the experiment the circuit is first of all adjusted for high efficiency oscillations and then a measurement is made of the average voltage and current prior to the onset of the oscillations. The analysis is dynamic to the extent that it is capable of producing dynamic waveforms and static because use is made of the average voltage and current values. A similar analysis has been reported by P.J. deWard and Snapp. Apparently Snapp's diode "S" appears to be quite similar to our devices.

4.1.1 THE EXPERIMENTAL ARRANGEMENT

The arrangement of the test set up is given in the Fig 4.2. In the main it consists of a coaxial circuit containing a device in package located at one end of the line and various number of slugs some distance away from the diode. In the immediate vicinity of the diode a thin
Fig. 4.2: EXPERIMENTAL ARRANGEMENT FOR t.d.t TRAPATT OSCILLATOR CIRCUIT
film resistor (approx. 0.3 ohm d.c.), to monitor the current through the device and an aerial probe capacitively coupled to the cavity to monitor the \( \frac{dv}{dt} \), are located. The diode is located in a 30 ohm collar, this along with other 30 ohm slugs in the neighbourhood acts like a charging line. The characteristic impedance of the line is chosen to be 50 ohm to properly match with other standard systems to be employed.

The biasing circuit consists of a power supply (Advance, type PP 1 ) and a pulser ( hp 214 ). The power supply provides a constant d.c. voltage \( V_0 \) which acts like a pedestal for the pulse from hp 214. This arrangement was chosen to bias the diode beyond the breakdown voltage in the absence of high power pulser ( Velonex 350 ). The resistor and capacitor combination serves to limit the current from the power supply and to protect the pulser from d.c. voltage. The bias pulses (average voltage and current through the device) were measured using A.C. voltage probe P 6006 and the current probe P 6019 both of Tektronix make. These pulses were displayed on a storage dual trace scope Tektronix-7613. The scope was externally triggered from the pulser hp 214. The pulse is coupled to the coaxial circuit via Microlab, HW-30N bias T which provides d.c. path for the bias pulse and a very high impedance for R.F. output.

The power output at the frequency selected by the tuning slugs, mainly the first one, is measured by a peak power meter ( Narda-66A3A ) via a suitable coaxial detector ( Model 561 ) and -10 dB arm of the Narda (1.7-4.2) GHz coupler. The capacitor in the bias T couples the r.f. to the peak power meter and the Decdb switch. At either end of the switch a spectrum analyser ( hp-8551 ) and a set of units to derive a trigger pulse for the Tektronix sampling scope are connected. The r.f. trigger pulse is connected at the external trigger input on the 7T11 pannel of the sampling scope via the Countdown unit ( hp 1104 ), Tunnel Diode Mount ( hp 1106 ) and a high
pass filter (hp 1109) and a 10 dB attenuator.

The dynamic current and voltage waveforms were displayed on the scope via attenuators and a suitable sampling head (S-4) housed in the vertical amplifier section (7SL1) of the sampling scope. The attenuators are required to avoid any backwards and forward reflections between the sampling heads and the coaxial circuit.

4.1.2 THE PROCEDURE

After applying the bias to the circuit the slugs were adjusted till a clean waveform was displayed on the scope and a maximum power output was obtained. The duration and the repetition rate of the pulse was chosen to be approximately 350 ns and 2 kHz, respectively.

The sequence of events leading to the conclusion that there exists a negative resistance at d.c. will be explained with the aid of a few oscillographs. It can be shown analytically that the average voltage and current measured by the probes (P 6006 and P 6019) is the voltage across the device and the current through the device if the coaxial circuit can be assumed to be lossless.

In Fig 4.4 and 4.5 four oscillographs have been displayed for different bias conditions determined by the d.c pedestal voltage $V_0$. In all these measurements the pulser voltage was unchanged. The variable control was kept at fully clockwise on the 50v range.

A close observation of these pulses reveals the mechanism by which the applied pulse amplitude decreases from its initial value for increasing d.c. voltages. This can be explained by looking at the reverse saturation characteristics of the device plotted in Fig 4.3. This was obtained at zero pulse voltage and the circuit was prevented from any possible oscillations. The d.c. pedestal voltage $V_0$ was gradually increased and the current taken from the power supply via 9.5 kilo ohm was recorded.
Fig 4.3: REVERSE I-V CHARACTERISTICS OF THE DIODE.
For voltages greater than about 86v the current through the device begins to increase rapidly and the pulse voltage is clipped. It is this clipping of the pulse voltage which results in the reduction of the pulse amplitude. In the following section the oscillographs in Fig 4.4 and 4.5 will be described and analysed.

Details of Fig. 4.4 (trace a):- The top and bottom traces are the pulse voltage and current. The actual voltage is obtained by multiplying the reading by 10 and the current scale is 2 mA per mv. These traces correspond to the case when \( V_0 = 0 \)v and the pulse voltage was approximately 50v. There are some ripples at the beginning and the end of the pulse which can be looked upon either as the transient of the input pulse or the reflections setup in the circuit. In either case these ripples are of some importance as suggested by East et al. \(^{(21)}\). These traces indicate almost constant voltage and zero current during the entire period of the pulse.

Details of Fig. 4.4 (trace b):- In this case \( V_0 = 30 \)v and the pulse voltage was 50v and thus the total voltage across the device was 80v. Though still there was no current flowing through the device the condition in this case was significantly different than in the above case. This can be seen by the changes brought about at the beginning and the end of the voltage pulse. What in fact is happening inside the device is that the depletion layer capacitance is continuously decreasing for increasing total bias voltage. This change in capacitance changes the device impedance and hence the phase of the reflection coefficient at the diode. It can be seen that the current begins to go round the bend of the reverse saturation characteristics (Fig. 4.3) which marks the beginning of the rapid increase in current through the device.
Fig 4.4 : THE BIAS PULSES
Details of Fig. 4.5 (trace a):— In this case $V_o = 57$ v and the pulse voltage is clipped to 38 v. Many other traces for $V_o$ greater than 30 v and less than 57 v were also taken which indicated a gradual increase in current through the device. In this case the total voltage across the diode is approximately 92 v and it is almost the limiting value beyond which the oscillations setup and the peak power meter begins to show some r.f. output. Thus in this case there are no oscillations in the circuit. The device is being biased beyond its breakdown voltage.

The current pulse on this trace can be divided into three main segments. From the beginning of the pulse upto 75 ns the current increases from zero to approximately 580 m.A, in two steps. In the first step between zero to 20 ns the current reaches to approximately 480 m.A. and in the second step between 20 to 75 ns it first decreases to 400 m.A. and then increases to 580 m.A. Beyond 75 ns the current through the device decreases monotonically to 400 m.A.

The voltage pulse can also be divided into three periods as above. Initially the pulse voltage increases to about 42 v from zero in first 20 ns. In the next portion the voltage pulse has many ripples which fluctuate between ±1 v and finally the voltage increases from 38 v to 42 v.

It is obvious from the above discussion that beyond 75 ns there exists a negative resistance because during this period the average voltage increases with the corresponding decrease in the current. However the initial increase in the current from 480 to 580 m.A at almost constant voltage just before the appearance of the negative resistance is due to the progressive reduction in the diode impedance. To illustrate this point more clearly an I-V characteristics derived from this trace has been plotted in Fig. 4.6. The existence of a turning over point at about 580 m.A is clearly demonstrated by this characteristics. In this figure points between
(a) $V_o = 57\, V$

(b) $V_o = 65\, V$

Fig 4.5: THE BIAS PULSES
Fig 4.6: THE QUASISTATIC I-V CHARACTERISTICS OF THE DIODE.
20 ns and 70 ns have not been taken into account to avoid any confusion due to reflections. Haitz \(^{(16)}\) has given a detailed explanation of this phenomenon and argues that the final portion of the bias pulse characterised by the increase in voltage and the corresponding reduction in the current is due to the junction temperature of the device and the d.c. negative resistance observed as indicated by Fig. 4.6 may be due to this effect.

Details of Fig. 4.5 (trace b): This trace depicts the nature of the bias pulses when the circuit is oscillating. In this case \(V_o\) was 65v and the maximum voltage across the device was 100v which dropped back to approximately 56v under oscillatory condition. The drop back in the voltage is accompanied by the corresponding rapid increase in current. In actual fact it is this increase in current which results in the voltage drop back. The oscillations seem to be beginning at approximately 720 mA.

Thus it is concluded from this study that the device does appear to have negative resistance at d.c. for bias currents greater than 580 mA and is determined by pulse to pulse temperature rise of the junction.

4.2 THE PRESENCE OF NEGATIVE RESISTANCE AT VHF

Though East et al. \(^{(21)}\) had indicated that one possible mode of triggering the Trapatt oscillations could be the presence of VHF oscillations in the circuit prior to the onset of the high efficiency oscillations, their findings were neither conclusive nor convincing. Here in this section first of all we will very clearly demonstrate the presence of VHF oscillations, their growth and origin, then later on we will make use of these results to explain the starting mechanism of high efficiency Trapatt oscillations.

As far as we have been able to examine and verify, the details
of the experiment reported in this section are unique and have not been reported so far and it is expected that some new work will emerge out of these results.

The experimental setup for this part was essentially the same as reported in Fig. 4.2 except that the coaxial circuit containing various slugs e.t.c were replaced by a line simply containing the device in S-4 package, the current and voltage probe. The rest of the circuit arrangement used was unaltered.

As in the previous case the pulse voltage was maintained at its maximum on 50V range and the d.c. pedestal voltage \( V_o \) was varied to generate oscillographs reported here. To establish the origin of the VHF oscillations Fig. 4.7 is given which is the electrical equivalent of the circuit employed in this section. In the circuit shown in Fig. 4.7 VHF oscillations were seen to be developing at \( V_o = 59V \). These oscillations were coherent, well defined, stable and capable of properly synchronising the trigger pulse to the sampling scope. These oscillations were able to grow as in any circuit giving rise to no doubt as to whether there is any negative resistance or not. Three traces have been shown in Fig. 4.8 to establish the growing nature of the VHF oscillations and their possible physical explanation. The trace a has been taken when the value of \( V_o \) was approximately 62V and the growing oscillations were observed at about 40 ns from the beginning of the pulse. This trace suggests that the negative resistance offered by the device to the circuit is more than the circuit resistance and hence the oscillations are maintained. However this negative resistance reduces as the value of \( V_o \) is reduced as shown in the traces b and c and thus the oscillations are seen to decay.

To establish the origin of these VHF oscillations and their relationship with the circuit various oscillographs have been given in
Fig 4.7: THE COAXIAL OSCILLATOR CIRCUIT
Fig. 4.8: VHF WAVEFORMS

Scale

Y axis 1 cm = 125 m.A

X axis 1 cm = 25 ns

Trace a, \( V_o = 62 \text{v} \)

Trace b, \( V_o = 59 \text{v} \)

Trace c, \( V_o = 57 \text{v} \)
Details of Fig. 4.9 (trace a):— The conditions in this case were identical to one reported above. The cable between the points A and B was 1.1 meter long resulting in VHF oscillations at 48.7 MHz.

Details of Fig. 4.9 (trace b):— In this case an additional 1 meter cable was added between the pulser hp 214 and the 50 ohm (refer to Fig. 4.7). The general nature and the frequency can be seen to be the same as reported above except that the signal is somewhat delayed.

In Fig. 4.10 three traces have been given to establish the origin of these oscillations. The lengths of the cable and the corresponding frequencies are marked on the traces.

It is concluded, on the basis of the traces given in Fig. 4.9 and 4.10, that the frequency of VHF oscillations is determined by the transit time of the pulses between the device and the point marked B in Fig. 4.7. The graph given in Fig. 4.11 suggests that the frequency of these oscillations is inversely proportional to the cable length discussed above. The growing nature of these oscillations clearly demonstrates and confirms the presence of negative resistance at these frequencies.

Finally in the following sections, we will show that our experimental devices in the circuit considered also possess small signal negative resistance at 2 GHz, the fundamental frequency of the Trapatt oscillator.

4.3 AMPLIFICATION OF THE INJECTED SIGNAL

It was established in the previous two sections of this chapter that the device does have negative resistance at d.c. and also at VHF. However it does not mean that this negative resistance will still be available at microwave frequencies, though Bower (16) has demonstrated that
(a) THE LENGTH OF THE CABLE 1.1 meter

\[
\text{VHF} = 48.7 \text{ MHz}
\]

(b) SIGNAL DELAYED BY 1 Meter CABLE

\[
\text{VHF} = 48.7 \text{ MHz}
\]

Fig 4.9 : VHF OSCILLATIONS AT 48.7 MHz
(a) $L = 2.10 \text{ m}$
$f = 29 \text{ MHz}$.

(b) $L = 1.98 \text{ m}$
$f = 32 \text{ MHz}$.

(c) $L = 0.44 \text{ m}$
$f = 66.7 \text{ MHz}$.

Fig. 4.10 : VARIOUS TRACES OF VHF OSCILLATIONS
Fig. 4.11: The graph showing the changes in frequency with length.
a device which possesses negative resistance at d.c. will maintain this nature at least up to the avalanche resonance frequency. (In our case this frequency will be approximately 14 GHz for a transit angle of π radians; \( f = \frac{V_a}{2W} \), \( V_a \) is the saturated drift velocity of the carrier and \( W \) the width of the depletion layer). It is the intention of this section to show that the device does have some negative resistance round about 2 GHz. In the following section of this chapter the method will be outlined and the results will be reported.

### 4.3.1 THE EXPERIMENTAL ARRANGEMENT

In the main the test setup for this part of the experiment was similar to one reported to show the presence of VHF oscillations in the section 4.2, with the additions as indicated in Fig. 4.12. The main additions are, Pulser 2 which is used to provide an amplitude modulated pulse to modulate the r.f. signal from the oscillator. This pulser is externally triggered by the triggered output pulse of the Pulser 1 (hp 214). The modulated r.f. signal to be injected, is fed to the coaxial circuit via the coupler (SL-4590), the circulator (SC-3590) and the coupling capacitor of the bias T. Tektronix scope 7704 is used to ensure that the modulated r.f. and the pulse from the Pulser 1 appear at appropriate times. The sampling scope is triggered by the modulated r.f. signal in the usual way via the high pass filter (hp 1109A), the tunnel diode mount (hp 1106A) and the trigger countdown (hp 1104A). The delay line 7M11 and suitable length of the coaxial cables are necessary to use for the purpose of delaying the signal with respect to the triggering signal.

### 4.3.2 THE TEST PROCEDURE

To begin with it is ensured that the bias pulse arrives at the diode plane during the flat portion of the modulated r.f. signal. In
Fig. 4.12: EXPERIMENTAL SETUP TO VERIFY THE PRESENCE OF NEGATIVE RESISTANCE
other words the bias pulse and the modulated r.f. are aligned as shown in Fig. 4.13 (a) avoiding the ripples seen on the lower trace. The system is ready for use once this adjustment is done.

4.3.3 THE PRINCIPLE

The principle of operation is based on the reflection mechanism at the diode plane. In this section we will deal with this aspect and establish the condition favourable for the existence of the negative resistance. In general the reflection coefficient at the device plane can be expressed as equation (4.4)

\[
\rho_d(t) = \frac{Y_o - Y_d(t)}{Y_o + Y_d(t)} = \frac{V_-(t)}{V_+(t)} = |\rho| e^{j\phi}
\]

where

\( Y_o \) = admittance of the cavity in the immediate vicinity of the diode.
\( Y_d(t) \) = admittance of the device.
\( V_- \) = reflected voltage.
\( V_+ \) = incident voltage.
\( \phi \) = phase angle between the reflected and the incident voltage.

It is evident from equation (4.4) that the total voltage at the diode plane will be given by the equation (4.5):

\[
V_d(t) = V_-(t) + V_+(t)
\]

\[
= V_+(t) \{ 1 + \rho_d(t) \}
\]
Fig. 4.13(a): THE TRACE SHOWING THE BIAS PULSE AND THE MODULATED R.F.

Fig. 4.13(b): THE GRAPH SHOWING THE VARIATION OF THE REFLECTION COEFF.
From equation (4.4) it can be seen that $|P_d(t)| \leq 1$ for the positive values of $\text{Re}(Y_d(t))$ and thus, under this condition the total voltage at the diode will always be less than $2V_+(t)$. However if $|P_d(t)|$ were to be greater than unity then the voltage will be greater than $2V_+(t)$ and this can happen only when $\text{Re}(Y_d(t))$ is negative. In Fig. 4.13(b) a graph is plotted to show the variation of $|P_d(t)|$ for different values of $\text{Re}(Z_d(t))$. The curve clearly demonstrates that $|P_d(t)| > 1$ for $\text{Re}(Z_d(t)) < 0$.

It is concluded from the above discussion that if it can be established that the signal being reflected back from the diode is greater than the incident signal then it must mean that there is a net gain and this can take place only when the injected signal is being terminated into a negative impedance. To investigate into this the calibration and characterization of the circulator is necessary.

4.3.4 THE CHARACTERIZATION AND CALIBRATION OF THE CIRCULATOR

By characterization we mean the performance of the circulator for different frequencies. Ideally the circulator is meant to be used within 2 to 4 GHz band. The input VSWR between the adjacent ports were measured by terminating the other ports into a matched load. These were measured to be 1.3, 3.6 and 20.0 at 2.0, 1.6 and 1.3 GHz respectively. The input VSWR values very strongly suggest that the performance of the circulator is not going to be dependable at the lower end of its frequency limits and it deteriorates very rapidly below 1.6 GHz.

By calibration we mean the determination of the coupling coefficients between various ports of the circulator. These coefficients are required to estimate exactly the strengths of the signal which couples into the coaxial circuit and reflects back from the diode for a known signal. In Fig. 4.14 four such coupling coefficients (only the magnitudes) $K_1$, $K_2$, $K_3$ and $K_4$ along with their conditions of measurement have been defined.
Fig. 4.14: Determination of the Coupling Coefficients

\[
|K_1| = \left| \frac{V_1(\oplus)}{V_3(\oplus)} \right|
\]

\[
|K_2| = \left| \frac{V_2(\oplus)}{V_3(\oplus)} \right|
\]

\[
|K_3| = \left| \frac{V_3(\oplus)}{V_1(\oplus)} \right|
\]
It is not really possible to estimate the phase angles associated with these coupling coefficients, however, it is realised that they play a significant role in the system. It is possible to define these coefficients in terms a 2X2 scattering matrix. Under ideal conditions \( K_1 = K_2' = 1 \) and \( K_2 = K_3 = 0 \).

### 4.3.5 THE EXPERIMENTAL PROCEDURE

In the main experiment the r.f modulated signal is fed via a 10 dB attenuator at the port 3 of the circulator. \( K_1 V_1(\alpha) \) couples in the arm 1 and \( K_2 V_2(\alpha) \) into the arm 2 of the coupler. If arm 1 was not terminated in a matched load then additional \( K_2' K_1 V_1(\alpha) \) would couple into the arm 2 as the signal reflected back from the mismatch in the arm 1. The input signal at the port 3 is measured on the sampling scope. Port 1 is terminated into a movable short and the total signal coming out of the port 2 appearing at \(-10\) dB arm of the coupler is displayed on the sampling scope. It was found that an accurate estimate of the magnitudes of the signal direct from the screen was not possible, therefore, it was decided to trace them by a pen recorder which was suitably calibrated for different positions of the movable short. In Fig. 4.15(a) and (b) traces marked 1 to 21 represent the total signal at port 2 of the circulator for different positions of the short. Trace 4 corresponds to the maximum signal (0.62v) and traces 9, 17 and 21 correspond to the minimum signal (0.52v) for an input signal of 0.63v at 2 GHz. From these traces (1 to 21) a graph was plotted between peak to peak amplitude and its corresponding position. The graph is shown in Fig. 4.16 giving the variation of this amplitude. The maximum and minimum values are neither properly spaced nor well defined. This could be due to some inaccuracies in estimating the values and due to various discontinuities and losses in the system. The graph clearly demonstrate that the
Fig. 4.15(a): Traces showing variation of signal at port 2 for different positions of the short circuit.
Fig. 4.15(b): Traces showing variation of signal at port 2 for different positions of the short circuit.
Fig. 4.16: TOTAL SIGNAL AT THE PORT 2 OF THE CIRCULATOR VERSUS THE POSITION OF THE SHORT CIRCUIT.
the maximum value under the short circuit measurement is approximately 0.62v.

The short circuit plunger was next replaced by the coaxial circuit along with its bias circuit. The pulse voltage was kept at the maximum on 50v range and the pedestal voltage $V_o$ was varied. At about 36v some conduction current begins to flow through the device as the breakdown voltage is approached. Beyond the total voltage of approximately 86v the device ceases to behave like a capacitor and the resulting waveforms are no more sinusoidal. However at about approximately 48v (pedestal voltage $V_o$) the device changes settle down and the sinusoidal nature is restored. In Fig. 4.17 two traces have been given one at $V_o = 36v$ and the other at $V_o = 58v$. The total current at these voltages were zero and approximately 400 mA. Peak to peak voltage at 58v was 0.72v. This suggests that the net gain at 2 GHz was atleast 2.02 dB, because the signal reflected back from the short circuit measurement was estimated to be 0.57v.

The experiment was also repeated by reducing the input signal strength by a factor of approximately 2.2. The gain on the similar basis was estimated to be 1.5 dB. This suggests that the gain of the injected signal is a function of the input signal level.

Thus it is concluded from the experimental results given above that the device has small signal negative resistance at 2.0 GHz.
Fig. 4.17: WAVEFORMS SHOWING THE GAIN OF THE INJECTED SIGNAL.
Trapatt oscillator circuits have been designed and realised in various configurations of standard microwave transmission line systems including coaxial line, microstrip and stripline. However for general experimental purposes the coaxial line circuits offer the maximum flexibility, ease in fabrication and realisation. For our experiments we have used 50 ohms coaxial air line (The diameter of the outer and the inner conductors being 7 and 3 m.m respectively) in which the packaged diode is mounted at one end. This line is suitable upto 20 GHz. In this chapter various details of the oscillator circuit and the explanation of the individual components have been described in detail, though in the fourth chapter a mention was made of the circuit employed.

5.1 THE DIODE AND ITS PACKAGE

In a coaxial line a packaged device can be mounted in either of the two possible arrangements, namely the "end" or "shunt" mounting. In our circuit the diode was mounted at one end of the line in such way that the P+ end of the device was embeded into the central conductor of the line. For accurate characterisation of the circuit it is important to know the details of the package and the device. In our case a coaxial ceramic S-4 package (Ceramic International Corporation, type A 251) is used. The mechanical details and the associated parts of the package are shown in
Fig. 5.1(a). The lumped equivalent circuit of the package up to 12 GHz is shown in Fig. 5.1(b). Corbey (47) and Owen (48) have described in detail the significance of various elements in Fig. 5.1(b). In the diagram the distributed inductance and capacitance is represented by its lumped equivalents. The inductance L is mainly due to the central post also known as pedestal, $C_1$ and $C_2$ are the capacitances between the top hat and the bottom and top of the post through an air gap and the ceramic ring. Getsinger (49,50) and Owen (51) have reported more complicated equivalent circuits, the use of a particular circuit depends upon the application and the required accuracy. For the work reported in this chapter the π network shown in Fig. 5.1(b) has been used. The device is packaged in "flip chip" configuration. The $P^+$ end is metalised using either Titanium-Rhodium Gold or Titanium Gold. For the purpose of contacting 50 μm gold is plated on the metalisation prior to thermal compression bonding of the device on the post. The $N^+$ end of the device is also metalised in the similar manner prior to thermal compression bonding of low inductance 25 μm gold wire in a "criss-cross" fashion. The top hat of the package is welded on the projection. In general "flip chip" mounting is preferred over the "faceup" mounting to obtain better heat distribution and to reduce the current distribution variation. Haitz (46) has shown that the temperature difference between the diode centre and the heatsink is approximately three times higher for the faceup mounted compared with the flip chip mounted diode.

5.2 THE AIR COOLED DETACHABLE HEATSINK

The detachable heatsink is shown in Fig. 5.2(a). It holds the device to provide good electrical and mechanical contact and also serves as an additional heatsink. The mounting surface also provides a reference.
Fig. 5.1(a) : THE S-4 PACKAGE

"Top Hat" Electrically welded around the projection.

Projection.

Diode chip.

Ceramic post.

Gold plated contact and Heatsink.

Fig. 5.1(b) : LUMPED EQUIVALENT CIRCUIT OF THE S-4 PACKAGE
Fig. 5.2(a) : COMPONENTS OF THE OSCILLATOR CIRCUIT

Fig. 5.2(b) : THE OSCILLATOR CIRCUIT
plane for the system. As shown in the Figure the top hat of the package is clamped in the sprung collet which can be tightened by the knurled head at the extreme end of the detachable heatsink. It is not possible to observe the current waveform with this arrangement, however, it can be done by replacing this heatsink with a suitable current probe to be described later in the chapter.

5.3 THE COLLAR AND THE TRIGGERING LINE

It was mentioned in the first chapter that a 50 ohm line is incapable of supplying large charging current required to initiate Trapatt oscillations. The combination of the collar and the triggering line provides the extra capacitance required to supply the large current. The collar and the next three slugs are all of 30 ohm characteristic impedance and form the triggering line. Though the circuit is capable of operating at other impedances, the best optimum results are obtained at 30 ohm as suggested by Summers et al. (14) (Fig. 2). It is also known that this impedance is a function of the junction diameter of the device and for devices of junction area 2 to 3 $10^{-4}$ cm$^{-2}$ this value is 30 ohm. All the slugs used were air filled and 9 mm in length, however, the collar was PTFE ($\varepsilon_r = 2.2$) filled and approximately 6 mm in length. PTFE is chosen to support the central conductor of the line and to achieve the required impedance. Overall general mechanical configuration of these and the tuning slugs is such that a good electrical contact is made with the walls of the coaxial line. A free movement of these slugs should be possible and they should not get disturbed on their own from their positions previously adjusted. To obtain a suitable solution the arrangement shown in Fig. 5.2 (a) and (b) was used.

5.4 THE TUNING SLUGS AND THE SLUG CARRIERS

A slug can be represented by a $\pi$ network, the discontinuity
capacitances at either ends of the slug constitute additional shunt elements. In the oscillator circuit four tuning slugs have been used. All the slugs were 9 mm long (air filled) and their characteristic impedances were 10 and 15 ohms. The shape of these slugs is shown in Fig. 5.2(a). The position of the first tuning slug (10 ohm) starting from the device end approximately determines the oscillator frequency, in general this separation is slightly less than half the wavelength at the operating frequency. The general characteristics of the tuning slugs and the coaxial circuit, when adjusted for high efficiency operation, is found to be that of a low pass filter. Ideally this low pass filter should act like a short circuit at the harmonics of the fundamental Trapatt frequency. Thus any high frequency components are contained within the cavity between the diode and the first tuning slug. These slugs also perform the function of an impedance transformer to match the r.f. load to the rest of the circuit. To achieve proper movement, the slug, finger and C shaped carrier assembly was used, in a split line as shown in Fig. 5.2(a) and (b). The voltage and current waveforms can be observed by replacing the detachable heatsink with the current probe and a different collar to accommodate a voltage derivative probe. These two probes are shown in Fig. 5.3 and will be described in the following section.

5.5 THE VOLTAGE PROBE

The electric field probe which was used to measure the derivative of the voltage across the diode is shown in Fig. 5.3(b). The probe is coupled capacitively to the electric field between the end of the central conductor and the shorting plane. The probe consists of an OSSM semi-rigid cable (outer diameter being 2.15 mm) with an extended central conductor at one end and an OSM connector at the other end. For accurate location
Fig. 5.3(a) : THE CURRENT PROBE

Fig. 5.3(b) : THE VOLTAGE PROBE
radially just above the diode; the stem of the OSM connector was threaded with the corresponding threads in the collar. When screwed into the collar the extended end of the inner conductor does not penetrate too far into the cavity and thus it is assumed that the existing field pattern in the vicinity of the diode is not disturbed.

5.6 THE CURRENT PROBE

A current probe consists of a thin film disc resistance in series with the device. The voltage developed across the resistor is a direct measure of the current through the device. To reduce the losses and hence the degradation of the oscillator circuit it is essential that the d.c value of this monitoring resistance should be as low as possible. In our case this value was 0.3 ohm. General assembly of the probe is shown in Fig. 5.3(a). A very good electrical contact and satisfactory mechanical fitting of this resistance with the device and the circuit is very important. In order to avoid unwanted reflections it is essential to use some suitable attenuators between the current monitoring resistance and the sampling scope.

5.7 THE TRIGGER COUNTDOWN, TUNNEL DIODE MOUNT AND HIGH PASS FILTER

It was mentioned earlier in the fourth chapter that these units are used in the system for external triggering of the Tektronix sampling scope. The facility of the internal triggering available within the scope is not suitable for signals above 1 GHz, and hence the use of these units in observing the Trapatt waveforms becomes essential. A typical setup incorporating these units in conjunction with the sampling scope is shown in Fig. 5.4. This system is capable of displaying signals up to 18 GHz. The only front panel adjustment available is the "stability" control on the
Fig. 5.4: ARRANGEMENT FOR EXTERNAL TRIGGERING OF THE SAMPLING SCOPE
Trigger Countdown unit (hp 1104A). This control is able to adjust the frequency of the tunnel diode oscillator around its centre frequency of 100 MHz in such a way that a stable operation occurs at a submultiple frequency of the input signal. Details of these units are available in the Hewlett Packard manual (52).
It is now known that Trapatt devices are one of the most suitable semiconductor candidates for high power high efficiency applications in L and S bands. The efficiency from a single device up to 81% (53) have been reported, however to achieve this value the researchers have exploited the knowledge they gained from the impedance measurements. Kappelar has successfully employed "active harmonic loading" at the third harmonics to enhance the efficiency and to improve the stability. Power from many such devices has been successfully combined to show their suitability for various system applications in communication and navigational aids. Although the technology and the understanding of the device and its interaction with the circuit has greatly improved since 1967, the starting mechanism of the Trapatt oscillations has not yet been resolved and the vast amount of literature available on this aspect has created more confusion. Various authors from time to time have proposed various mechanisms and to substantiate their claims they have produced either experimental or computed results. Probably they are all justified in their remarks to some varying degree. It is our belief that for accurate physical understanding of this type of oscillations it is absolutely crucial and important to know the necessary and sufficient boundary conditions to start the oscillations. It is this conviction which forms the basis of this chapter.

Perhaps the most comprehensive survey of all such mechanisms in relation to t.d.t oscillators and their possible loci has been given by
Blakey (54), which he calls as "mode chart for starting in t.d.t oscillators". The mode chart is shown in Fig. 6.1. We will follow our independent approach and finally trace the path on this chart. On the basis of the results reported in the fourth chapter a possible route is likely to be via low frequency oscillations at VHF, however the results of the detailed investigations reported in this chapter do not seem to confirm it.

6.1 VARIOUS KNOWN STARTING MECHANISMS

In the first chapter a reference was made of different ways in which the Trapatt oscillations can possibly start. Here in this section we will expand on these ideas and establish the validity and truth of their claims. Among various modes which have appeared in the literature, perhaps, most authors believe that the presence of Impatt type oscillations (55-71) is essential for the ultimate appearance of high power high efficiency type of oscillations. After the initial proclamation of Prager et al. in April 1967 regarding this mode, probably Clorfeine and Hughes (63) were the first to indicate and confirm that for avalanche devices capable of operating in "anomalous mode", which possess neither d.c. nor small signal negative resistance, the presence of Impatt oscillations is necessary. The most common theory appears to be that the transit time oscillations are setup at small currents as a result of avalanche and transit time delays, these oscillations grow as the average current is increased and ultimately large signal Impatt oscillations trigger the Trapatt oscillations. In all such cases the circuit configuration was so chosen that it was possible to tune simultaneously for the Impatt and the Trapatt frequency. In this approach no justification has been given about possible relationship between the two frequencies. It would seem that these two are not necessarily related at all. However there are some including Snapp (17) who believe that the
Fig. 6.1: Mode Chart for Starting in t.d.t Oscillators
Trapatt frequency is subharmonically related to the Impatt frequency. This appears to be more logical but it raises a very serious question. The question which has to be answered is whether the Impatt frequency is generated first and then the Trapatt or some higher harmonics of final Trapatt frequency is present in the system as a result of suitable circuit conditions.

It has been established that the Trapatt devices are very good harmonic generators and an improvement in final conversion efficiency can be obtained by suitably extracting power at these harmonics. It has been reported by Carroll (72) that up to 25th harmonics are present in a Trapatt oscillator. Snapp (73) has obtained 3.8% efficiency at the 12th harmonic (20.1 GHz) and reported that the initiating current for optimum generation of the third harmonic is considerably less than that associated with the efficient fundamental generation. This means that higher harmonics are generated at lower currents and thus it is possible that instead of just the Impatt frequency any suitable harmonics may be present before the high efficiency Trapatt oscillations. Thus it is concluded that higher frequencies present in the system prior to the onset of the Trapatt oscillations are harmonically related.

While discussing this mode, we feel that it will substantiate our claim later in the chapter, if at this stage we report some results by Cripps et al. (74). To confirm the role of Impatt oscillations in initiating the Trapatt mode they operated two devices in push-pull configuration and found that a good diode is able to force a poor Trapatt diode into oscillation when placed in this configuration. Such poor diodes showed no sign of high efficiency oscillations in a single diode conventional circuits. The authors argue that it would appear that whilst many diodes are capable of maintaining a Trapatt type oscillations, fewer possess the characteristics
required to initiate these oscillations by themselves. By placing an X band impatt as the test device in a push-pull circuit and operating it with a good Trapatt diode they were able to confirm their claim. It should be added that the impatt device when operated in a single diode conventional circuit did not oscillate in the Trapatt mode. In Fig. 6.2 (a) and 6.2 (b) the results obtained by Cripps et al. have been reported. Fig. 6.2 (a) displays the bias voltage pulse showing the flat drop in the voltage indicating the presence of Trapatt oscillations. In the beginning of the pulse some high frequency Impatt type oscillations could be observed but the point to be noted is that they are not growing in amplitude. However between these and the stable Trapatt oscillations there is unusually long period of "jitter". The total period before Trapatt oscillations are observed is 400 ns which represents about half the period of the applied pulse. This violates the common theory that a few cycles of Impatt oscillations appear prior to the onset of the Trapatt oscillations. The authors rightly argue the results that it does not suggest that Impatt oscillations play no role on the starting of Trapatt oscillations, but it would certainly seem to suggest that the starting is not simply one strong Impatt oscillations leading immediately to Trapatt oscillations. Fig. 6.2 (b) displays the resulting Trapatt voltage waveform as seen on the sampling scope.

We conclude from the above results that though the authors did observe some Impatt oscillations in the beginning, it almost certainly was not responsible for triggering the Trapatt oscillations, thus it can also be argued that the presence of Impatt type of oscillations is not a precondition to observe the Trapatt type of oscillations. In fact our results do confirm this point of view.

Parametric mode has also been suggested as a possible triggering mode. In these situations a high frequency source of energy is
Fig. 6.2(a): The bias pulse reported by Cripps

Fig. 6.2(b): The dynamic voltage waveform reported by Cripps
required for pumping the nonlinear reactance of the device. It has been suggested by Clorfeine (75) and Snapp et al. (17,76) that such a pump frequency is "self generated" by the device. In a degenerate case this pump frequency \( f_p \) equals exactly twice the signal frequency \( f_s \). If \( f_p \) is near \( 2f_s \), a modulation of \( f_s \) results with the unavoidable generation of new (idling) frequencies in the mixing process. For most efficient energy transfer from \( f_p \) to \( f_s \), power flow at all idler frequencies except \( f_p - f_s = f_i \) must be prevented. Evans and Haddad (77) have given a detailed analytical treatment of this process and have concluded that in proper circuit, parametric frequency conversion may result in a negative conductance at the input and outputs of the converter, therefore, high gain frequency conversion and parametric amplifications are possible.

Though Trapatt oscillations appear to have a superficial resemblance with the parametric operation in so far as both seem to need a high frequency pump and that it predicts the major features to be expected i.e. a fall off of power with frequency and three frequency systems having gains predictable from the analysis of the Manley and Rowe equations. The parametric label does not appear to be appropriate because an order of magnitude more power is sometimes attainable at the Trapatt (low) frequency than is obtainable at the pump frequency. Thus the power frequency scaling laws observed in simple parametric systems seem to be completely violated. It would seem logical to think that those workers who observed the presence of high frequency oscillations prior to the onset of the low frequency ones, were mistakenly driven into their conclusions because of the nonlinear capacitance characteristics of the Trapatt diodes and their previous knowledge of parametric behaviour of such devices. It would seem as if both the Impatt and the Parametric mode are one and the same as far as the starting characteristics of Trapatt is concerned.
There is a growing tendency, among recent researchers, to believe that the Trapatt oscillations start in a conventional relaxation mode. Theoretical analysis and experimental results seem to support this view very strongly. The most distinguishing feature of this type of oscillations is that the frequency of oscillation is determined by the RC time constant in the bias circuit and its value increases with the bias current. One major difference between the relaxation oscillator circuit and the conventional experimental coaxial circuit (t.d.t) as pointed out by Tantraporn (78) and Blakey (20) is the manner in which the energy is stored in two cases. In the first case the energy storage is mainly electrostatic (often mainly in the depletion capacitance of the device), whereas in t.d.t circuit the energy storage is mainly electromagnetic, in the elements of the transmission line. This remarkable difference would lead to a very simple and easily realisable oscillator circuit. Qualitatively this type of oscillators fall into the category of astable multivibrators, in which the conducting and nonconducting i.e. on-off conditions keep on shifting between two active devices and the repetition rate or the frequency is determined by the RC time constants in the two stages. There is very close resemblance between this and the actual Trapatt cycle, however in the later case it is only one device which keeps on changing its conditions and of course the total period is the sum of on and off periods.

Perhaps the most important work on this mode was first reported by Hoefflinger (79) in 1966. By suitably designing the device he was able to observe the oscillations starting from 100 MHz upto 10 GHz. He invoked the idea of "the space charge feedback" as a possible explanation for low frequency oscillations. It is reported that saw tooth wave shaped repetition frequency was determined by the circuit and the driving current, while the shape depended mainly on the circuit. It is suggested that the
cut off frequency of oscillation can be obtained in terms of the depletion capacitance, the value of the negative resistance and the bulk resistance. Hoefflinger's idea seems to be very useful in explaining some of the phenomena observed in the present day Trapatt devices, which may well have some small region of almost intrinsic layer very close to the metallurgical junction.

Ward and Udelson (80) first in 1967 reported the relaxation behaviour of an avalanche diode and later in 1968 Carlson (81) presented some experimental results to confirm Ward's computed calculations. Various salient features of this mode were clearly demonstrated by these two workers. However Ward's simple circuit as shown in Fig. 6.3 (a) was criticised for low efficiency of about 14%, a figure very low in comparison to the values achieved using t.d.t circuits. A major step forward in this direction was the theoretical and experimental work of Yu and Tantraporn (78) and Blakey's (20) computed results. The reported results by these two workers almost completely put a shadow on the previous findings.

The oscillator circuit employed for simulation by Ward and Udelson was such that the Trapatt charging rate was determined by R and the combined external and depletion capacitance of the device. Looking at the circuit, the main drawback appears to be the position of R, which has to serve many functions simultaneously. It acts as r.f. load, a tuning element and dissipates power, therefore a proper choice of its value becomes very difficult as it has to compromise between conflicting requirements. The circuit shown in Fig. 6.3 (b) and (c) is a possible improvement and was suggested by Blakey (20). Though there is some flexibility and improvement in the circuit performance, by isolating the resistance from d.c. and reducing its effect on tuning as in Fig. 6.3 (b), the most significant improvement is obtained when employing the circuit shown in Fig. 6.3 (c). The inductance
Fig. 6.3(a) : WARD AND UDIELSON’S RELAXATION OSCILLATOR CIRCUIT

Fig. 6.3(b) : BLAKEY’S RELAXATION OSCILLATOR CIRCUIT

Fig. 6.3(c) : BLAKEY’S IMPROVED RELAXATION OSCILLATOR CIRCUIT
L is approximately resonant with the series combination of $C_1$ and $C_0$ and thus the desired frequency component may be optimised. The inductance $L$ also acts as a r.f. choke to other higher harmonics. This circuit is almost identical to the one used by Yu and Tantraporn (78) in their pure mode configuration.

Using the circuit configuration of Fig. 6.3 (c) Blakey (20) has indicated that an efficiency upto 30% should be possible for the same devices which produced upto 40% efficiency when used in a t.d.t circuit. The major feature of this circuit is that the Trapatt oscillations can start without delay as compared to t.d.t circuit in which it can take upto half of the pulse duration before stable Trapatt oscillations are observed. Yu and Tantraporn and Blakey both reach the same conclusion. Yu and Tantraporn have demonstrated that the Trapatt action can be initiated within one capacitive charging time, instead of being considerably delayed as is necessary in the conventional circuits. These results clearly demonstrate that the presence and growth of Impatt type of oscillations is not a precondition for ultimate Trapatt oscillations and in fact Yu et al. do reach to this conclusion and comment that it is not necessary to initiate the Trapatt by the Impatt mode. They further say that it is desirable to have resonance mode so that, with synchronisation, Trapatt oscillations can be obtained at lower current levels.

There are some other research workers who have obtained similar mode of oscillations including Bogan and Frey (19), Zappert and Lee (18), Dalman, Zappert and Lee (82) and Shackle (83). Shackle has reported an I-V characteristics displaying the presence of negative resistance and apparently it resembles very closely with the similar curve reported in the fourth chapter. It is also interesting to note that the circuits used by Yu et al. and Blakey reported earlier bear a close resemblance with the circuit reported by Shackle. Bogan and Frey (19)
through their numerical calculations have established that high frequency oscillations often observed prior to the onset of the Trapatt oscillations are shown to be of relaxation type rather than impatt. They have also indicated some other modes of triggering the high efficiency oscillations to be dealt later in the chapter.

The numerical calculations of Matsumura and Abe (84) and the experimental results of Torizuka and Yanai (85) have produced slightly different results than reported so far. Matsumura's results indicate that the anomalous mode is not necessarily triggered by large signal impatt type oscillations. It is suggested that these oscillations start more easily when a period of relaxation oscillation is approximately equal to the time lag of the circuit response. The calculations confirm Torizuka's results that under steady state conditions the voltage waveform of the anomalous mode can be constructed by the superposition of the saw-tooth and the positive spike waves.

Clorfeine and Hughes (86) report yet another possible mode of triggering. They report that d.c. current voltage curves obtained for the silicon devices capable of oscillating in high power anomalous mode do not show a negative resistance characteristic as such, however a d.c. negative resistance is seen to exist as a result of transit time oscillations. Clorfeine points out that the presence of this induced d.c. negative resistance is probably the triggering mechanism for anomalous mode of oscillations.

Perhaps it is an appropriate place to report that O' Callaghan et al. (87) reported in Jan 1970 that the devices capable of oscillating in the anomalous mode do possess d.c. negative resistance. This experimental result is quite opposite to the one reported above. They conclude that although d.c. negative resistance is not necessary condition for anomalous mode operation, it is a good guide to the diode structure required for high
efficiency oscillations, and in general, it can be stated that the presence of d.c. negative resistance is desirable for the operation of the anomalous mode but undesirable for normal transit time oscillations.

So far we have reported and criticised the existing postulates and theories of possible starting mechanisms of the Trapatt oscillations and we can safely conclude that it is not necessary that large signal impatt oscillations must exist in the circuit to trigger the Trapatt oscillations.

Finally we consider that mode of triggering mechanism which was initially considered \(^{(88)}\) and still believed \(^{(55)}\) to be an undesirable feature of the avalanche oscillator circuits namely " the bias circuit oscillation ". Brackett \(^{(88)}\) has given a very good detailed analytical, theoretical, physical and experimental explanation of their nature and origin. It is not required to comment on various results put forward by Brackett as his explanations are very clear, however, it is essential to mention that he concludes that the presence of low frequency instabilities extending from d.c. to several tens and perhaps hundreds of megahertz in Si or Ge devices is an unavoidable fact of life. He has also mentioned that for the type of the devices which were in use at that time these low frequency oscillations should be eliminated otherwise the ultimate device failure through " burn out " is destined. However, nowadays the avalanche devices and their technology has considerably changed and it seems that these low frequency oscillations play a vital role in starting the Trapatt oscillations.

In 1975 East et al. \(^{(21)}\) reported for the first time that VHF oscillations, which are established in the circuit due to a combination of d.c. space charge negative resistance and a.c. rectification negative resistance, is responsible for triggering the Trapatt oscillations. He appears to have confused between two different phenomena i.e. the bias
circuit oscillation and the bias circuit ringing. In the previous case the active device plays a vital role as has been shown by Brackett however in the later case the device plays a secondary role of encouraging the ringing which is established in the circuit when a fast rise pulse is applied to the circuit (19). This ringing causes an over voltage in the bias voltage and therefore yields a faster Trapatt turn on.

Thus in our view the bias circuit ringing had been the sole cause of the VHF oscillations in the results experimentally obtained by East et al. because the modern Trapatt devices do not show any d.c. negative resistance solely due to the space charge effects as claimed by Bower in 1968, however the second explanation of a.c. rectification at VHF appears to be valid. We personally feel that the true triggering mechanism need not be the presence of low frequency oscillations in the system. The validity of the statement will become clear later in the chapter when the results of our experiment will be discussed.

In the following sections of this chapter the details of the experimental approach and the results to establish the starting mechanism will be described.

6.2 DETAILS OF THE EXPERIMENT

The details of the test setup and relevant individual elements of the oscillator circuit have been described in the fourth and fifth chapters. The experimental arrangement chosen for this part of the experiment was exactly the same as in Fig. 4.2. The main aim and object of this investigation was to determine which one or the combination of a few mechanisms mentioned above was responsible for initiating the Trapatt oscillations in our t.d.t circuit. It is absolutely essential to locate the beginning of the pulse to establish the mode of triggering. Using the
arrangement outlined in Fig. 4.2, the signal displayed on the screen of the sampling scope does not contain the front of the pulse because of the delays between the signal and the trigger pulse. To resolve this problem the signal was suitably delayed by introducing a delay line in its path. It was found that the standard G.R.A. cable was very lossy and dispersive and hence the delay line 7M11 available as an optional extra with the Tektronix sampling scope was used. Though this delay line, capable of introducing a total of 150 ns delay, was also lossy and dispersive it was found to be adequate for our experiment. After having located the front of the pulse, we need to know exactly the location of a particular point on the pulse. To be able to do this various coarse and fine delay controls were required to be calibrated. The calibration process is described in the following section.

6.2.1 THE CALIBRATION OF THE DELAY CONTROLS

The calibration described in this section concerns the "delay" controls on each vertical amplifiers (7S11) of the sampling scope and fine "Time Position Control" on the time base (7T11). At the top of the Fig. 6.4(a) we have given a skeleton sketch of the system used for calibration purposes. A sweep generator precisely set at 2 GHz and checked by the frequency counter was used as a reference signal. The scope was externally triggered by this sinusoidal signal.

First of all the "delay" controls were calibrated in the steps of 5 ns. The displayed signal was moved till the point B aligned with the point A, shown in the bottom traces of Fig. 6.4(a), this position of the "delay" control was marked, the control was further rotated again till once again the traces aligned. This control in both the amplifiers offers a maximum delay of 12.9 ns. The delays offered by the coarse and fine "Time Position Controls" were determined by the "Time Position
Fig. 6.4(a) : THE SKETCH AND WAVEFORMS SHOWING THE CALIBRATION PROCESS
Range " selected. The ranges we were normally concerned were either 0.5 μs or 50 ns, and corresponding to these the coarse control produced the equal amount of delay, where as the fine control was capable of producing a maximum of 10% of these. The delay was zero when these controls were fully clockwise. The calibration process of these controls on 7T11 module was exactly the same as in the case of the "delay" control on 7S11 module. The exact position on the pulse also depends on "Stability/Trig. Level" control on 7T11 module.

6.2.2 Generation of the Ringing Signal

It has been pointed out by Bogan and Frey (19) and East et al. (21) that low frequency VHF oscillations can trigger the Trapatt oscillations. In the fourth chapter various oscillographs were given to find out the origin and the circuit dependence of these oscillations. In this section two more oscillographs are given in Fig. 6.4(b) to confirm the previous results.

The top trace in the Fig. 6.4(b) corresponds to the case when the bias circuit at the end of the current and voltage monitoring point was terminated into a 50 ohm (the circuit including the bias T was disconnected). In these traces the damped sinusoidal signals were absent because of the matched termination and the absence of reflections. In the bottom trace the 50 ohm termination was removed and the end of the cable was left open. In these traces VHF oscillations like a damped sinusoid can be seen to be present. Thus it is concluded that the VHF oscillations are setup in the system because of the mismatch at the end of the cable and the available inductance and capacitance in the bias T and the rest of the circuit do not make any significant contribution.
The bias pulses when the circuit was terminated in 50 ohm.

Fig. 6.4(b) : Waveforms showing generation of the ringing signal.

The bias pulses when the circuit was terminated in an open circuit.
There is enough evidence that at relatively lower drive, the VHF oscillations grow in amplitude to the extent where the Trapatt oscillations could be triggered, however no conclusive evidence could be found to support this point of view when looking at the starting mechanism in detail. Therefore it is our belief that the presence of these low frequency oscillations is not a precondition to observe the Trapatt oscillations and in a suitable circuit these oscillations start at the fundamental frequency decided by the position of the first slug and the rest of the tuning elements.

6.3 THE PROCEDURE

Here in this section our main aim is to establish the starting mechanism and the time it takes to start i.e., the turn on time. Attempts will also be made to investigate into the possibilities of reducing the turn on time thus improving the performance of the oscillator circuit. On the basis of the detailed results obtained during this investigation by virtue of our careful selection, it will be shown that a Trapatt device is capable of operating and adjusting over a wide range of operating points. In the past various workers have reported that the design of both, the device and the circuit is very critical and very careful balancing into various aspects of the oscillator performance, is required for proper operation of the Trapatt oscillator. Trew (9) has argued that the microwave circuit plays the most vital role, here in this thesis we report that the device is equally capable of performing a similar function.

In all the measurements, the pulse voltage, duration and the repetition rate were maintained at, fine control fully clockwise on 50v range, 350 ns and 2 KHz, respectively. The d.c. pedestal was slowly raised till some oscillations were seen on the sampling scope and the bias voltage
dropped back to a lower value with the corresponding increase in the bias current, characterising the onset of the Trapatt oscillations. At this setting the r.f. power was also seen to be appearing in the peak power meter. Various tuning slugs were slowly and carefully adjusted for the most clean waveform and the maximum output. Having achieved the optimum slug positions for the best possible oscillator performance at the chosen bias conditions, these positions of the slugs were kept fixed. The d.c. bias pulses, r.f. power output and various traces of the dynamic current and voltage derivative were taken. These measurements were repeated for various bias conditions for proper understanding of the oscillator performance.

In the following sections detailed results will be reported for one bias setting which will explain the general philosophy of our measurements and rest of the results for other settings will be presented in the form of tables and graphs.

6.4 THE RESULTS OF THE EXPERIMENT

In Fig. 6.5 two oscillographs have been given for the bias voltage and the bias current and the bias current and the rectified r.f. output. Various details of these oscillographs are given in Fig. 6.5. The details of the oscillator performance are given below:

Maximum d.c. voltage across the device, \( V_{\text{l}} = 106 \text{v} \)

Bias voltage, \( V_{\text{b}} = 62 \text{v} \)

Dropback voltage, \( V_{\text{l}} - V_{\text{b}} = 44 \text{v} \)

Bias current, \( I_{\text{b}} = 2.55 \text{ Amp.} \)

Pulse width = 350 ns

Pulse repetition rate = 2 KHz.

Frequency = 2.16 GHz.

Output power = 32 Watts
Fig. 6.5: THE BIAS PULSES AND RECTIFIED R.F OUTPUT
Efficiency = 20.6%
Apparent turn on time = 25 ns

The top trace in Fig. 6.6 is the dynamic current through the device and the bottom trace is the dV/dt at the device plane. Both these traces indicate the presence of upto 4th harmonics and were inverted at the sampling scope as displayed by the arrows on these traces.

6.4.1 THE RESULTS RELATING TO THE STARTING CHARACTERISTICS

The results will be analysed for the oscillator settings described in 6.4 above. First of all we begin from the bias pulses given in Fig. 6.5 (a). The Trapatt mode is typically characterised by the drop in voltage across the device with the corresponding increase in current. This point can be seen to exist at about 25 ns from the front of the pulse, where the rate of change of current has significantly increased and more so some sort of fuzz can also be seen, suggesting the beginning of the oscillations including the front end jitter. The current corresponding to this time is approximately 1.2 amp. The bias current appears to be 2.55 amp, when full amplitude high efficiency oscillations are properly setup.

To observe the front of the pulse signals were delayed by 157 ns. The "delay" and "time position" controls were used to move the pulse by a known amount to observe the details of the waveforms at a known position in time with respect to the start of the pulse. Though the signals were very much attenuated at the frequencies of interest (10, 21 and 33 times at 2, 4 and 6 GHz,) the residual amplitudes were enough to detect their relative presence against the background noise and fluctuations. In all the measurements the pen recorder sensitivity was kept at 100 mV/cm, where as the scope sensitivities were chosen to suit the measurements. Because of the fixed amplitude (vertical) available at the output terminals
Fig. 6.6: DYNAMIC CURRENT AND VOLTAGE WAVEFORMS
of the scope and the chosen sensitivity of the pen recorder, the trace scale turns out to be 1 cm on the scope equal to 2 cm on the graph paper.

The Fig. 6.7 to 6.16 can be grouped into three main sections. Fig. 6.7 to 6.13 are the traces of current and dV/dt without any filter, Fig. 6.14 and 6.15 are the traces of current and dV/dt for the waveforms in which the fundamental component has been filtered out and the idea is to see the presence and growth of any high frequency signal. The traces for current and dV/dt given in Fig. 6.16 are the signals without any additional delay other than caused by essential connecting cables between the scope and the location of the current and voltage probes. These traces correspond to that portion of the pulse where high power-high efficiency Trapatt oscillations are present.

To start the analysis of the results reported in Fig. 6.7 to 6.16, refer to Fig. 6.7 which shows the start of the pulse and displays first 40 ns of the pulse containing the waveforms. The trace for the current suggests that some sort oscillations are beginning to setup at 20 ns which do not grow much till 21.25 ns. In other words the current can be seen to be growing fast beyond 21.25 ns, however these initial observations will be carefully investigated in detail. Exactly similar behaviour can be seen on dV/dt trace at 21.75 ns.

Thus it appears that the oscillations start at about 20 ns and begin to grow after a couple of nanoseconds, therefore the pulse upto and around 20 ns will be investigated in great details.

Traces in Fig. 6.8 to 6.10 are the current and dV/dt for the first 20 ns of the pulse in which both the scales have been expanded and enlarged 5 times except the current between 10 and 20 ns period which has been enlarged only two times so as to accomodate the entire pulse during this time interval. Each centimeter on the time scale on these traces
represent 500 ps. There are no signs of any coherent oscillations in traces on Fig. 6.8 and 6.9, however, perhaps there may be some oscillations developing at 18 ns as indicated by the arrow on dV/dt waveform which appears to contain some oscillations at about two times the fundamental frequency. Similar information can not be obtained from the current waveforms probably because of the system noise and the chosen amplitude scale.

Fig. 6.11 corresponds to the pulses between 15 and 25 ns. The first thing which is indicated by the current trace is that the oscillations start at 18.9 ns and begin to grow at 21 ns. Initially the waveforms appear to be rich in harmonics upto about 24 ns and afterwards these components appear to be obscured by the fundamental, this point can be looked upon as the time when a possible mode conversion takes place. Prior to this mode conversion successive change in the phase of the waveform can also be seen.

As regards to the triggering of the Trapatt oscillations, it would seem that the oscillations from the beginning itself are at the final Trapatt frequency but the efficiency of these oscillations appears to be significantly lower than the final value as manifested by their relative r.f. amplitudes. This is not really very surprising as it has already been established that our experimental devices have adequate negative resistance to produce a signal gain of atleast 2 dB at 2 GHz when the input signal level was very small.

Fig. 6.12 and 6.13 are the traces of current and dV/dt between 20.5 to 30.5 ns and indicate the general nature of the growing waveforms and their relationship with each other. These traces also indicate the time when the possible mode conversion takes place and can be seen to be approximately 24 ns. It can be seen on the current trace of Fig. 6.12 that the amplitude of oscillations increases rapidly after 24.35 ns.
Fig. 6.14 and 6.15 are the traces of current and $dV/dt$ during 10 to 30.5 ns, in these traces the fundamental component of the Trapatt oscillations has been filtered out to see the nature of high frequency oscillations. These traces also suggest that the oscillations start at about 18.9 ns.

The results of the experiment described so far clearly suggest that for the bias conditions under discussion the oscillations start at 18.9 ns at the fundamental Trapatt frequency. Various components present in the direct signal shown in Fig. 6.16 can be seen to be present in the signals delayed by 157 ns (refer to Fig. 6.11 between 22 to 24 ns), though their relative amplitudes are greatly reduced. This suggests that if high frequency signals were growing prior to the onset of the Trapatt oscillations we should have been able to see them. In the following section we will make use of the results presented in Fig. 6.12 and 6.15 to calculate the growth rates of various signals to establish their relative growths in the beginning of the oscillations.

6.4.2 DETERMINATION OF THE GROWTH RATES

The growth rates are calculated by estimating the peak to peak variations of various signals. A growing wave at an angular frequency $\omega$ can be represented by an expression of the form given by equation (6.1):

$$V(t) = V e^{\sigma t} e^{j(\omega t)}$$

(6.1)

where

$V$ = Constant amplitude

$\sigma$ = The growth rate

$\omega$ = The angular frequency of the wave

The slope obtained from the logarithmic plot of equation (6.1) will give the growth rate. Fig. 6.12 and 6.15 have been used (between
Fig 6.7: DYNAMIC WAVEFORMS FROM THE BEGINNING OF THE PULSE.
Fig 6.8: WAVEFORMS FROM THE BEGINNING TO 10 ns.
Fig 6.9: WAVEFORMS FROM 5 to 15 ns
Fig 6.10: WAVEFORMS BETWEEN 10 AND 20 ns
Fig 6.11: WAVEFORMS BETWEEN 15 AND 25 ns
Fig 6.12: CURRENT WAVEFORM BETWEEN 20.5 to 30.5 ns

Time in ns
Fig 6.13: VOLTAGE WAVEFORM BETWEEN 20.5 to 30.5 ns
Fig 6.14: WAVEFORMS BETWEEN 10 AND 25 ns
Fig 6.15: WAVEFORMS BETWEEN 20.5 AND 30.5 ns
Fig 6.16: Detailed current and voltage waveforms
the time intervals 22.6 to 25.35 ns and 21.7 to 24.05 ns) to estimate the growth rates at the fundamental and other higher frequencies. A plot between ln (peak to peak amplitude) versus time for the above case is given in Fig. 6.17. The growth rates for the fundamental and higher frequency components have been estimated to be $7.45 \times 10^8 \text{ sec}^{-1}$ and $5.27 \times 10^8 \text{ sec}^{-1}$. These two growth rates clearly demonstrate that for the bias conditions listed in this section the fundamental component grows at least 1.4 times faster than the higher frequency components at the beginning of the oscillations and thus it is concluded that these oscillations start and grow at the final fundamental frequency of the Trapatt oscillator.

6.5 RESULTS OF THE EXPERIMENT FOR OTHER BIAS SETTINGS

In this section the results of the experiment similar to one described above for three more bias settings are given in the table 6.1. Various notations and symbols used in the table have already been described in the text and given in Fig. 6.5(a). The results relevant to the starting mechanism are $t_{on}$ (the time when the oscillations start) and the growth rates.

Table 6.1

<table>
<thead>
<tr>
<th>$V_o$</th>
<th>$V_1$</th>
<th>$V_b$</th>
<th>$I_b$</th>
<th>$f$</th>
<th>$P_{out}$</th>
<th>Eff.</th>
<th>$t_{on}$</th>
<th>Growth Rates</th>
</tr>
</thead>
<tbody>
<tr>
<td>Volts</td>
<td>Volts</td>
<td>Volts</td>
<td>Amps</td>
<td>GHz</td>
<td>Watts</td>
<td>%</td>
<td>ns</td>
<td>$10^8 \sigma_1$</td>
</tr>
<tr>
<td>80</td>
<td>104</td>
<td>60</td>
<td>2.30</td>
<td>2.09</td>
<td>28.5</td>
<td>20.8</td>
<td>22.75</td>
<td>5.7</td>
</tr>
<tr>
<td>70</td>
<td>102</td>
<td>58</td>
<td>2.2</td>
<td>2.10</td>
<td>26.0</td>
<td>18.8</td>
<td>25.0</td>
<td>5.1</td>
</tr>
<tr>
<td>65</td>
<td>100</td>
<td>56</td>
<td>1.9</td>
<td>2.13</td>
<td>23.0</td>
<td>21.6</td>
<td>44.9</td>
<td>3.6</td>
</tr>
</tbody>
</table>

$\sigma_1$ and $\sigma_2$ are the growth rates for the fundamental and harmonics
Fig. 6.17: GRAPH FOR THE GROWTH RATES WHEN $V_0$ WAS 90 VOLTS
On the basis of the results reported here it can be concluded that the oscillator turn on time is a decreasing function of the d.c. drive and the oscillations start from the beginning at the final fundamental frequency of the oscillator.

6.6 CRITICAL CURRENT DENSITY VERSUS START OF OSCILLATIONS

While investigating into the starting characteristics of the Trapatt oscillations some very useful results have been obtained as regards to the current required to initiate the oscillations. It has been suggested that for satisfactory operation of the oscillator circuit the initiating current also known as "critical current" should be in excess of $eNv_s$ and should be at least $1.5(eNv_s)$ \(^{(10)}\). For our devices this current should be at least 1.1 amps. It has been found for four different bias settings that these currents estimated from the dynamic waveforms are 0.6, 0.7, 0.8 and 1.0 amps. Thus we find that the oscillations can start even when the current is only 0.54 times the minimum value suggested earlier. Efforts were made to understand this feature of the oscillator circuit and the only suggestion which appears to be available in the literature is also based on the experimental evidence of Goronkin and Wanuga \(^{(89)}\). They report that low efficiency oscillations start at smaller currents than the high efficiency oscillations. There are strong indications to suggest that our oscillator circuit is also behaving in a similar manner since the r.f amplitudes in the beginning are significantly smaller than the final amplitudes.

A possible explanation to this phenomenon has been given by Yu and Tantraporn \(^{(78)}\), who have reported the presence of similar oscillations at $0.3(eNv_s)$. They associate it with the internal dynamics of the device and suggest that with the specially tailored doping profiles (which they do not disclose) "internal quenching" can be designed, during this process at such low currents the drives are small and the field does not
collapse to almost zero and for most part the carriers drift at their saturated value ($v_s$).

Thus it is concluded from the above discussions that it is not necessary that the current through the device should be in excess of $eNv_s$ to initiate the Trapatt oscillations.

6.7 SUMMARY

In the end we conclude this chapter by giving below a summary of our results and conclusions:

1. The oscillations start at the fundamental frequency of the oscillator.

2. The time to start the oscillations is a decreasing function of the d.c. drive.

3. The initial current to start the oscillations can be significantly smaller than the "critical current".

4. The oscillations begin to grow very rapidly within a couple of nanoseconds from the start of the oscillations.

5. Small and large signal oscillations appear to be separated by some sort of mode transformation characterised by the gradual change in their waveshapes.

6. Initial oscillations are richer in harmonics than the final oscillations.

7. Though the presence of VHF oscillations may be helpful in triggering the Trapatt oscillations, no conclusive evidence could be found to suggest that their presence is a precondition for the ultimate appearance of the Trapatt oscillations.
The main purpose of this chapter is to report some useful and interesting results which have been obtained while investigating into the starting characteristics of the Trapatt oscillations. The results presented here will clearly demonstrate that in due course of time it should be possible to design and realise a Trapatt oscillator which will be capable of operating over a wide range of operating points and still be able to produce stable steady state oscillations due to the availability of more flexible devices.

7.1 TRAPATT A MORE FLEXIBLE DEVICE

It has already been reported earlier that the slug positions were initially adjusted for the best possible operation and thereafter their positions were not altered while the bias conditions were varied over a wide range as given in table 6.1. The consequence of this arrangement can be best understood by measuring the impedance of the coaxial circuit at the device plane, which obviously will be the same at a fixed frequency, since the slug positions and their impedances are kept fixed.

Analytically the circuit impedance at the diode plane can be calculated in terms of the characteristic impedances of various sections of the line and their separations. Gupta (90) has given a detailed treatment of the general approach and the procedure. Thus if the separations of the
various sections of the line and their characteristic impedances are not changed, then the circuit impedance will also remain the same at a particular frequency.

It has been reported that under different bias conditions the frequency does change but only slightly. This change will result in different values of the circuit impedance, when either measured or calculated at the diode plane, even when the mechanical configurations of the circuit are not altered. It has been shown that the device in the circuit considered is capable of properly oscillating over a wide range of the operating points and it can do so only if $|Z_{diode}| = |Z_{circuit}|$. It has already been shown in the fourth chapter that this condition must always be satisfied in an oscillator circuit (refer to the equation 4.1). Thus the device is able to adjust to the circuit changes brought about by the following changes in the bias conditions:

1. The operating bias voltage changes by 11%.
2. The operating bias current changes by 34%.
3. The input d.c. power changes by 48.6%.

While maintaining stable steady state oscillations for the wide range of variations mentioned above, the only changes observed in the oscillator performance were as below:

1. The frequency changes by 3.3%.
2. The oscillator efficiency changes by 15%.
3. The output power changes by 39%.

As a consequence of the changes mentioned above (in the operating conditions), the general behaviour of the oscillator, as regards to its starting characteristics was concerned, remained the same. The only change observed was in $t_{on}$, which decreased by a factor of 2.37 for 11% increase in the bias voltage.
The conclusion we arrive at from the above results is that the device is capable of performing satisfactorily over a range of operating points and the idea that the device merely behaves like a fast switch shunted by a variable non-linear capacitance (91) does not appear to be the most suitable definition of such a flexible device. As an extension of our conclusion regarding the flexibility of the device, we can probably state that in future it should be possible to realise an oscillator circuit in which the device replacement should be possible, still maintaining the oscillator performance within the specified limits.

In the end we conclude that a suitably designed and fabricated device can be very flexible in its performance.

7.2 SUGGESTIONS FOR FURTHER WORK

Experimentally we have shown the presence of VHF oscillations and established their origin. A possible suggestion as to how adequate negative resistance could be available at these frequencies has also been given. It has also been shown that the device does have small signal negative resistance at 2 GHz and a signal gain of at least 2 dB has been obtained. Theoretical work needs to be done, to establish the operating mechanism by which these devices produce the negative resistance at VHF and the microwaves, so that we can understand the wideband negative resistance characteristics of these devices. The results of such an exercise will provide more confidence to our experimental results reported in this thesis regarding the starting characteristics of the Trapatt oscillators.

Finally on the basis of the results presented in this chapter regarding the performance of the device, more work needs to be done to provide
enough confidence to circuit designers and system manufacturers so that the Trapatt oscillators could find a suitable place among other microwave generators.
REFERENCES


52. Hewlett-Packard "Trigger Countdown, models 1104A, 1106A and 1108A Operating Note", Published by Hewlett-Packard Company, Colorado Springs Division, Colorado, U.S.A.


55. Haddad, G.I et al "Basic Principles and Properties of Avalanche Transit Time Devices", IEEE Trans. on Microwave Theory and Techniques, pp 752-
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CALCULATION OF THE FIELD DISTRIBUTION

Referring to Fig. 2.11 and making use of the Poisson's equations in one dimension, the field distribution in the P and N regions were calculated as given below.

For the P region

\[
\left. \frac{dE}{dx} \right|_P = -\frac{e}{\varepsilon \varepsilon_0} NA(x)
\]

\[
x \int_{-x_P}^x \frac{dE}{\varepsilon \varepsilon_0} = - \int_{-x_P}^x N_0 (e^{-m_1x} - 1) \, dx
\]

\[
E_P(x) - E(-x_P) = -\frac{N_0 e}{\varepsilon \varepsilon_0} \left[ -1/m_1 (e^{-m_1x} - e^{-m_1x_P}) - (x + x_P) \right]
\]  

at \( x = -x_P \) \( E(-x_P) = 0 \)

\[
E_P(x) = -\frac{N_0 e}{\varepsilon \varepsilon_0} \left\{ \frac{m_1x_P}{m_1} - \frac{e^{-m_1x}}{m_1} - x \right\}
\]

at \( x = 0 \) \( E_P(0) = E_m \) (The maximum field at the junction)

\[
E_m = -\frac{N_0 e}{\varepsilon \varepsilon_0} \left\{ \frac{m_1x_P}{m_1} - \frac{e^{-m_1x}}{m_1} - x \right\}
\]

For the N region

In this case \( x \) is greater than \( d \) and the Poisson's equation can be written as equation (6).

\[
\left. \frac{dE}{dx} \right|_N = \frac{e}{\varepsilon \varepsilon_0} ND(x)
\]
\begin{equation}
\int \frac{dx}{dE} = \frac{e^{x}}{\epsilon \epsilon_{0}} \int_{0}^{x} \text{NE}(x) \, dx \tag{7}
\end{equation}

at $x = 0$, the field in the two regions will be continuous and therefore,

$$E_{F}(0) = E_{N}(0) = E_{m}$$

$$E_{N}(x) - E(0) = \frac{\epsilon \epsilon_{0}}{e^{x}} \int_{0}^{x} \left(1 - e^{-m_{1}x}\right) \, dx + \int \left(e^{m_{2}(x-d)} - 1\right) \, dx$$

$$E_{N}(x) = E_{m} + \frac{\epsilon \epsilon_{0}}{e^{x}} \left[ \frac{e^{-m_{1}x}}{m_{1}} + \frac{e^{-m_{2}d}}{m_{2}} \right]$$

$$E_{N}(x) = -\epsilon \epsilon_{0} \left[ \frac{-m_{1}x}{m_{1}} - \frac{1}{m_{1}} \left( x + \frac{1}{m_{1}} (e^{-m_{1}x} - 1) + \frac{1}{m_{2}} \left( e^{m_{2}(x-d)} - 1 \right) - (x-d) \right) \right] \tag{8}$$
APPENDIX " B "

0001 REM CALCULATION OF C-V FOR AN EXPONENTIALLY GRADED JUNCTION
0002 DIM F[20,1]
0003 DIM G[20,1]
0004 DIM I[35,2]
0010 PRINT "INPUT M1, M2, N0, D"
0019 INPUT M1, M2, N0, D
0021 LET M1=M1*1000000
0022 LET M2=M2*1000000
0023 LET N0=N0*1E+21
0024 LET D=D-, 000001
0028 LET C1=(!1.8*2*10!-17)/(36*4*ATN(1))
0029 LET C2=(N0+1.8*36*4*ATN(1)*10!-10)/11.8
0030 DEF FNA(U)=EXP(U)-U-K
0031 DEF FND(U)=EXP(U)-1
0041 PRINT
0042 PRINT "D2", "N", "C", "V", "E0*1E-5 "
0044 DIM P[1,71]
0045 LET P2="#=#, #=# #=#, #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=# #=
0305 GOSUB 1000
0310 NEXT L
0320 MAT PRINT F
0330 MAT PRINT G
1000 REM ELECTRIC FIELD DISTRIBUTION
1001 PRINT
1002 PRINT "X", "E KV/CM"
1003 LET T=0
1004 LET O=-EXP(U1)+U1
1005 LET D1=M-D2
1010 FOR X=02 TO -D1 STEP -.1
1015 LET T=T+.1
1020 LET U=M1*X*: 0.000001
1025 LET Z=M2*X*: 0.000001
1040 IF U>U1 THEN GOTO 1070
1050 LET R3=EXP(Z-Z3)-1-(Z-Z3)/M2
1060 GOTO 1080
1070 LET R3=0
1080 LET E=C2*(((O+EXP(-U))/M1)+R3)
1085 GOTO 1100
1090 LET E=C2*(O+EXP(-U)+U)/M1
1100 LET I(T, 2)=E*: 0.00001
1110 LET I(T, 1)=X
1120 NEXT X
1130 MAT PRINT I
1150 RETURN
APPENDIX "C"

DATA

1.0000D+16  2.7000D+21  5.0000D+19  8.0000D+07  1.050D+05  1.600D-18
1.060D-11  5.0000D+03  1.600D+04  1.550D-02  1.507D+05  1.904D-13
8.000000000D-05  10.524000000D-02  5.000000000D-05

DOUBLE PRECISION XP,XN,T(70),U,Z,PA,AR2
DOUBLE PRECISION A,AB,AB2,AB3,AC,AC1,AC2,B1,BA,BB
DOUBLE PRECISION BT,F,JN1,JN,F,PN,PN1,VP1,VSUM
DOUBLE PRECISION K(I),S(I),R(70),X(70),JH(200),JR(200)
DOUBLE PRECISION E(200),VP(200),VN(200),NL(200),Hi(200),hi(200)
DOUBLE PRECISION S15AEP,ND(200)
CALL SRCH$(1,'DATA4',5,1,TYEP,ICODE)
READ(5,7) CNO,CN1,CN2,A0,BO,Q
READ(5,7) EP,CMP,CMN,Y0,EC,JNS
7 FORMAT(6(B10,4))
   READ(5,9) DY,X1,DX
9 FORMAT(3(B21,10))
   WRITE(1,12) CNO,CN1,CN2,A0,BO,Q
12 FORMAT(6(2X,B10,4))
   WRITE(1,13) DY,X1,DX
13 FORMAT(3(3X,B21,10))
14 READ(5,19) JT
19 FORMAT(D6,2)
   WRITE(1,21) JT
21 FORMAT(1H1,10X,D6,2)
   IF(JT,EQ,0.) GO TO 120
XN=0.
   XP=0.
   DO 117 I=1,70
117 IF(I,EQ,1) GO TO 27
   JC=JT*1,E-?C
   IF((R(I-1),GT,0.),OR.(T(I-1),LT,0.)) GO TO 201
   XP=X(I-1)
   IF(XN,NE,0.) GO TO 202
   X(I)=X(I-1)-DX
   GO TO 30
201 XN=X(I-1)
   IF(XP,NE,0.) GO TO 202
   X(I)-X(I-1)+DX
   GO TO 30
202 X(I)=(XP+XN)/2.
   GO TO 30
27 X(I) = X(I)  
30 WRITE(1,31) X(I)  
31 FORMAT(5F6.4) X(I) = D17.10  
DO 116 J=1,200  
IF(J.GT.1) GO TO 38  
Y=X(I)  
39 FORMAT(4H10) E(J)= -EC  
JN(J)=-JNS  
GO TO 82  
38 Y = Y + DY  
DO 76 L=1,4  
IF(L.EQ.1) GO TO 72  
IF(L.EQ.4) GO TO 66  
A = E(J-1) + K(L-1)/2.  
JN1 = JN(J-1) + S(L-1)/2.  
A1 = AD*(Y-DY/2.)*(Y-DY/2.)  
B1 = BO*(YD-YHDY/2.)  
46 F1 = CN0+CN2*(1.-S15AEF(B1, IFAIL))- CN1*(DEXP(-A1))  
IF(A+3.5E3) 48, 48, 63  
48 AB = DABS(A)  
AB1 = 0.4445*DLOG(AB)  
AB2 = 0.1525*DLOG(AB)  
VF1 = 4.68E4*DEXP(A1)  
VN1 = 1.45E6*DEXP(A2)  
53 AB3 = (VP1*VN1)/(VP1*VN1)  
PA = -1.75E6/(3*AB)  
PB = -3.26E6/(3*AB)  
AF = 3.8E6*DEXP(PA)*DEXP(PA)*DEXP(PA)  
BT = 2.55E7*DEXP(PB)*DEXP(PB)*DEXP(PB)  
K(L) = BY*D*F1*JN1+AB3/Q+JT/(Q*VP1) /EP  
S(L) = (AF-BT)*JN1-BT*JT)*DY  
GO TO 76  
63 VP1 = CMP*AB  
VN1 = CHN*AB  
GO TO 53  
66 A = E(J-1) + K(L-1)  
JN1 = JN(J-1) + S(L-1)  
A1 = A0*XY*K'  
B1 = BO*(YD-Y)  
GO TO 46  
72 K(L) = BY*D*(ND(J-1)-NA(J-1)) + JN(J-1+C(N(J-1)-1)+VP(J-1))/  
$ (U(J-1)*VN(J-1)+JT/VP(J-1)) /EP  
S(L) = BY*(JH(J-1)) (AL(J-1)-BE(J-1)) / (E(J-1)) /JT;  
76 CONTINUE  
E(J) = E(J-1) + K(L-1)*2. + K(2)*2. + K(3)*2. + K(6)*2. ...  
JH(J) = JN(J-1) + (3.1)(2.6)*S(J)/(3.3)(3.6) ...  
104 FORMAT(9H10) H(E) = D17.10; 5X HJN(J) = D11.10;  
IF(E(J), GT, 0.) GO TO 123  
83 AR1 = AD*XY*K'  
AR2 = BO*(YD-Y)  
NA(J) = CN1*(DEXP(-AR1))  
ND(J) = CN0+CN2*(1.-S15AEF(AR2, IFAIL))  
BN = 1.75E6/(3.*E(J))  
BB = 3.26E6/(3.*E(J))  
AL(J) = 3.8E6*DEXP(BA)*DEXP(BA)*DEXP(BA)  
BE(J) = 2.55E7*DEXP(BB)*DEXP(BB)*DEXP(BB)  
IF(E(J)+3.5E3) 95, 95, 101  
95 AE = E(J)  
P = 0.4445*DLOG(AE)  
N = 0.1525*DLOG(AE)
VP(J) = 4.63E4*DEXP(P)
VN(J) = 1.45E4*DEXP(N)
GO TO 106
101 VP(J) = -CMPE(J)
   VN(J) = -CMNE(J)
106 IF(J.EQ.1) GO TO 115
   VSUM = VSUN + DTVI:J(J)/E(J-1)/2.
108 FORMAT(2X,4Hn =,D17,10)
   IF(J.EQ.200) GO TO 123
   IF(E(J).LT.0.) GO TO 116
123 U = JN(J)
    Z = DABS(U+JT)
    WRITE(1,127) U, V SUM
127 FORMAT(2X,4Hn =,D17,10)
    IF((U,LT,0.), AND, (Z,LT,JC)) GO TO 124
    IF(I,EQ,70) GO TO 124
    GO TO 121
124 WRITE(1,125) (E(I), M=1,J)
125 FORMAT(4(2X,D17,10))
    GO TO 126
115 V SUM = 0.
116 CONTINUE
121 R(I) = U
   T(I) = R(I) + JT
   WRITE(1,122) R(I), T(I)
122 FORMAT(2X,7Hn,) =,D17,10,5X,7H T(I) =,D17,10)
117 CONTINUE
126 GO TO 14
120 CALL SRC# & (4,'DATA4',5,1,ITYPE,ICOL)
   CALL EXIT
END