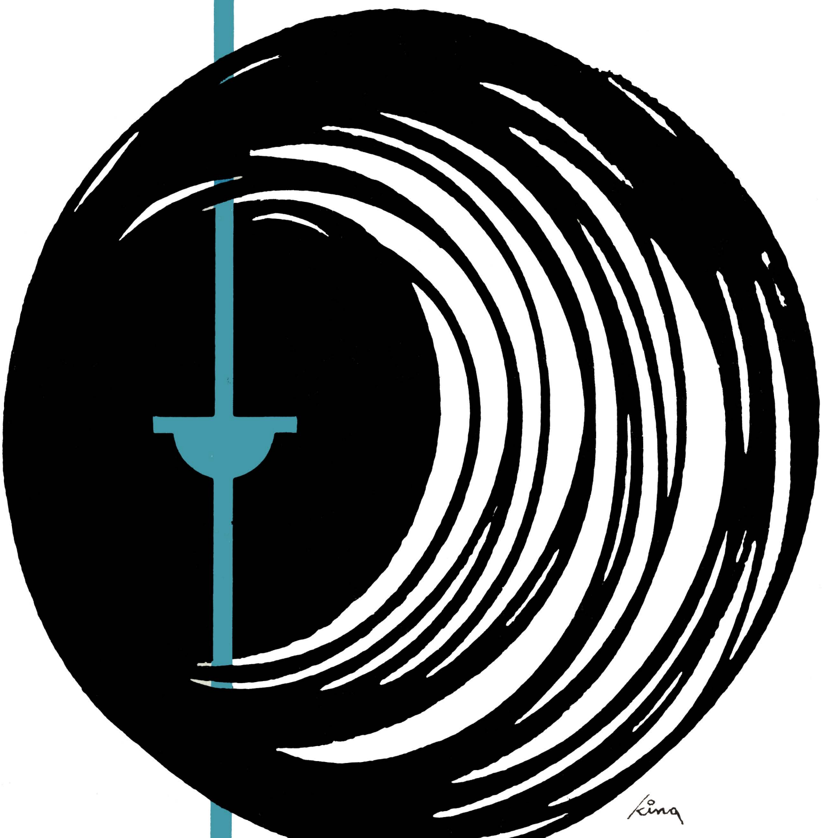
TUNNEL DIODE MANUAL

CIRCUITS APPLICATIONS SPECIFICATIONS



ONE DOLLAR

GENERAL ELECTRIC

TUNNEL DIODE MANUAL

First Edition

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of the Tunnel Diode Manual.

TABLE OF CONTENTS

The circuit diagrams included in this manual are included for illustration of typical tunnel diode applications and are not intended as constructional information. For this reason, wattage ratings of resistors and voltage ratings of capacitors are not necessarily given. Similarly, shielding techniques and alignment methods which may be necessary in some circuit layouts are not included. Although reasonable care has been taken in their preparation to insure their technical correctness, no responsibility is assumed by the General Electric Company for any consequences of their use.

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FOREWORD

The tunnel diode is the newest of continual advancements in semiconductor devices since the introduction of the transistor and silicon controlled rectifier. It is directly responsible for opening up many new areas of application not previously feasible. These additional applications have been made possible by virtue of the tunnel diode's extreme speed (high frequency), stable characteristics that are insensitive to temperature changes, modest power supply requirements, ability to operate in a wide variety of critical environments, low noise level, simplicity, light weight and small size.

Tunnel diodes will not displace transistors or other active components in existing equipment. In some new circuits this may occur; but generally speaking, tunnel diodes should improve the functional worth of other active components by working with them. Tunnel diodes, together with other semiconductor devices, will make possible the practical design of equipment now either impossible or impractical.

Since General Electric first introduced the tunnel diode in 1959, much energy and talent have been directed toward tremendously improving and greatly increasing the versatility of this newest of semiconductor devices.

An example of this effort is the recent introduction of General Electric's "back" diode and "microwave" tunnel diode. This latter device, housed in a small stripline package barely visible to the naked eye, has an inductance of less than .000,000,000,4 henries. This "microwave" tunnel diode package was designed for microwave communications, radar, very high frequency amplifiers, and oscillator applications.

General Electric again takes pleasure in initiating another semiconductor reference, the Tunnel Diode Manual, with the objective of extending the same service to the industry as has the G-E transistor manual since its introduction in 1957.

This manual, then, is one of our contributions to the better understanding of tunnel diodes and their circuitry for use in low power, ultra-high frequency, low cost applications.

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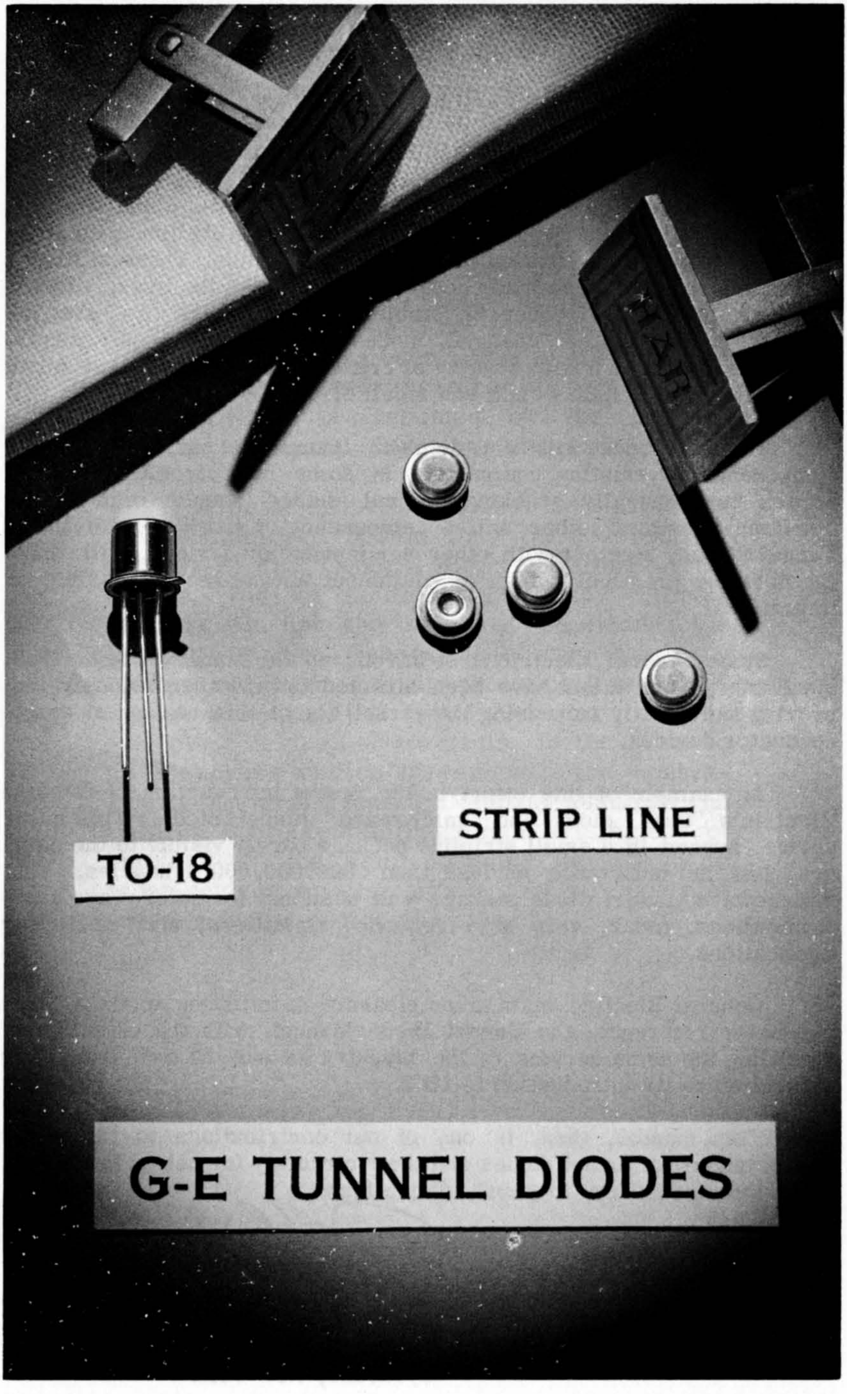

TO-18
STRIP LINE
G-E TUNNEL DIODES

CHAPTER I

THEORY

The tunnel diode is a revolutionary new semiconductor device. Although it can perform many of the functions of conventional devices its principles of operation are entirely different from those of other semiconductor devices and vacuum tubes.

Such conventional amplifying devices as transistors and vacuum tubes depend on emitting a charge carrier into a region where its motion can be influenced by a signal electrode, and on subsequently collecting the charge carrier on an output electrode. The speed of this process is limited by the time it takes a charge carrier, having left the emitter, to traverse the control region, and appear on the collector.

The basic conduction mechanism in tunnel diodes, however, has a theoretical frequency limit of 10^7 megacycles per second which is several orders of magnitude higher than the drift and diffusion mechanism involved in the operation of conventional diodes and transistors.

1.1 The Tunnel Effect

The tunnel diode takes its name from the tunnel effect--a process wherein a particle (obeying the laws of the quantum theory) can disappear from one side of a potential barrier and appear instantaneously on the other side, even though it does not have enough energy to surmount the barrier. It is as though the particle can "tunnel" underneath the barrier.

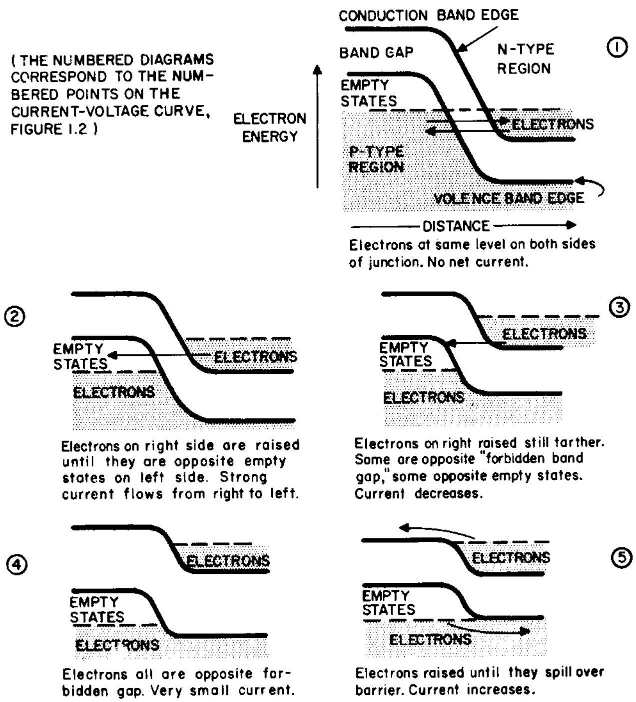
In the case of the tunnel diode, the barrier is the space charge depletion region of a p-n junction. This is the same barrier which prevents the current from flowing in the reverse direction in the case of the ordinary rectifier diode. In the tunnel diode, this barrier is made extremely thin (less than a millionth of an inch) -- so thin, in fact, that penetration by means of the tunnel effect becomes possible This gives rise to an additional current in the diode at very small forward bias which disappears when the bias is increased. It is this additional current that produces the negative resistance in a tunnel diode.

The origin of the tunnel current can be qualitatively understood by considering the changes in the characteristics of a conventional p-n junction diode made with higher and higher concentrations of free carriers in the semiconductor crystal. As the density of the charge carriers is increased, the reverse breakdown voltage decreases. It might be thought that there would be a limiting case when the reverse breakdown voltage was reduced to zero. This is not correct, however; the limit is determined by the solubility of the impurities which determine the carrier concentrations. Experiments have shown that it is possible to dope many semiconductor materials more heavily than is needed to reduce the breakdown voltage to zero. If this heavy doping is used, the diode can still be in the reverse breakdown condition at a slight forward bias. When a larger forward voltage is applied, the diode goes out of the reverse breakdown condition and the current falls to a small value.

1.2 Physical Description of a Tunnel Diode Junction

A semiconductor has a forbidden region where there are no states available for its electrons. This region is called the band gap. The states below this gap (which comprise the valence band) are almost all filled. The states above it (the conduction band) are almost all empty. The number of empty states in the valence band, or electrons in the conduction band, can be controlled by adding either acceptor impurities or donor impurities to the semiconductor crystal. Each acceptor impurity takes one electron out of the valence band, and each donor gives one electron to the conduction band. In this way p-type (empty states in valence band) and n-type (electrons in conduction band) regions can be built into a crystal. The intersection of these two regions is called a p-n junction.

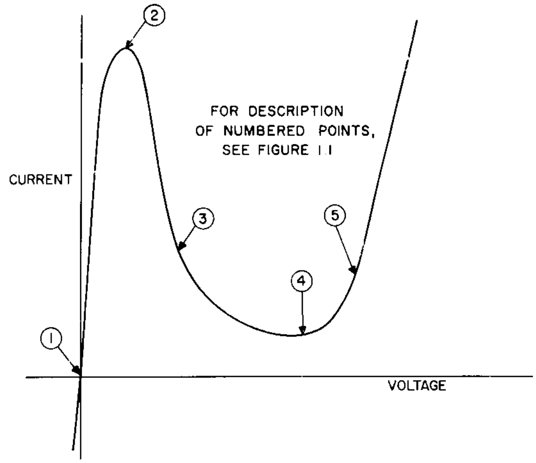
Figure 1.1 and 1.2 represent the conduction and valence bands in



TUNNEL DIODE JUNCTION AT VARIOUS BIAS CONDITIONS

FIGURE 1.1

the vicinity of a junction at different values of applied bias. It can be seen that as the bias is increased the bands which overlap each other at zero bias become uncrossed. Since tunneling is represented by a horizontal transition on this picture, the current decreases as the bands become uncrossed.



TUNNEL DIODE CHARACTERISTIC

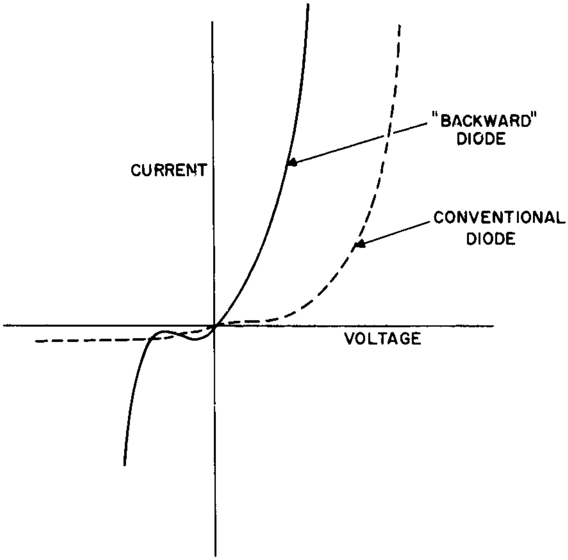
FIGURE 1.2

1.3 Backward Diodes*

In a tunnel diode, the voltages corresponding to points #2 and #4 in Figure 1.2 are essentially determined by the semiconductor material (germanium, silicon, or gallium arsenide) used in the process. The current value corresponding to point #2 is determined, however, by the cross-section area and doping level of the junction. In practice the tunnel junction is made larger in cross-section area than necessary and then etched until the peak current (point #2 in Figure 1.2) has the desired value. Tunnel diodes can be made with peak currents anywhere from a few microamps to several amperes. Tunnel diodes with peak currents of 500 μa and higher are usually used as amplifying or switching elements in circuits that take advantage of the fact that with increasing forward bias, the tunnel current first increases and then decreases.

It will be noted in Figure 1.2 that when voltage is applied to a tunnel diode, in the reverse direction, a large current flows which increases continuously as the voltage is increased. If a negative voltage of the magnitude of point 4 in Figure 1.2 were applied to a tunnel diode, the resulting current would be many hundreds of times higher then when this voltage is applied in the positive direction. A tunnel diode used to take advantage of this large change in current with polarity of applied voltage is called a "backward diode" . The term "backward" meaning that the diode conducts heavily with negative rather than positively applied voltage. Figure 1.3 shows a comparison of germanium "backward" and conventional diodes.

* "Backward" diode is also referred to as "back" diode throughout this edition.



COMPARISON BACKWARD AND CONVENTIONAL DIODE CHARACTERISTIC

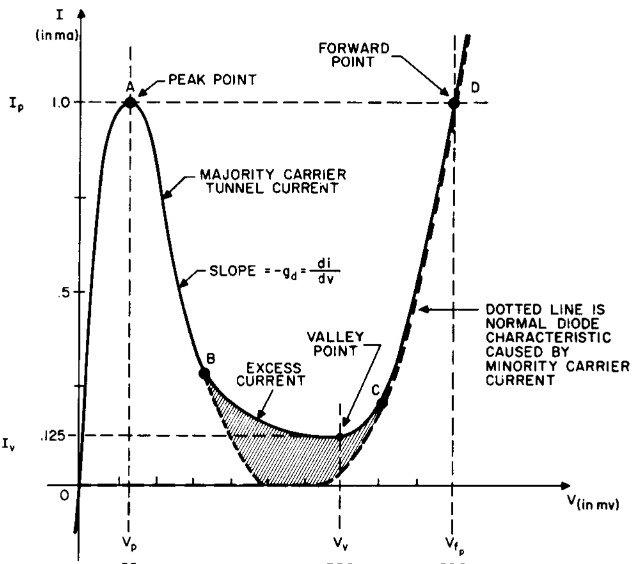
FIGURE 1.3

It can be seen that a backward diode has a lower voltage drop at a given current than a conventional diode. This low forward drop is very advantageous in tunnel diode and some transistor circuits. Backward diodes are designed to have a low (50 μa or less) peak point currents since current flow with forward bias is undesirable in backward diodes for the same reasons that leakage current with reverse applied voltage is undesirable in conventional diodes. In some types of tunnel diode circuits, however, it may be desirable to have a low peak point current and in this case a backward diode can be used.

CHAPTER 2

RATINGS AND CHARACTERISTICS

The voltage-current characteristic of a germanium tunnel diode is shown in Figure 2.1 together with the important DC parameters. The dotted line in this figure shows a normal diode characteristic resulting from minority carrier current. It is seen that the tunnel diode follows this characteristic beyond point C. In the lower voltage region below point C and in the reverse biased state the diode current consists of majority carriers which tunnel through the narrow PN junction.



STATIC CHARACTERISTIC CURVE OF GERMANIUM TUNNEL DIODE

FIGURE 2.1

Electrical Characteristics

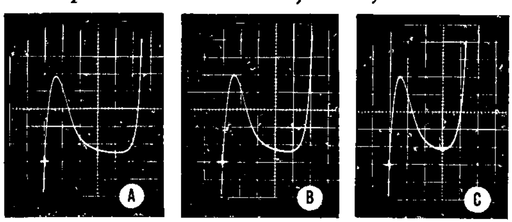
A "relatively" linear negative conductance region exists between point A (the peak point) and point B in Figure 2.1. Between point B and point C, the current is greater than the sum of the theoretical majority and minority currents. The current in this region, identified as the excess current, can not, as yet, be completely explained. Intuitively the excess current or valley current should be low and therefore the highest peak point to valley point current ratio seems desirable. There are some tangible reasons for this also. The greater this ratio, for any given value of peak point current, the greater will be the available output current swing. For example, a tunnel diode with a peak current of one milliampere and a peak to valley current ratio of 8 will have an available current swing of 1.0 - 0.125 = 0.875 ma.

The "peak point current" of a tunnel diode is determined by an etching process and can be held to within $\pm$ 2.5% or better on a production basis. However, the peak point voltage, V_p, valley point voltage, V_v, and forward point voltage, V_{fp}, are determined by the semiconductor material and are largely fixed. For germanium these voltages are respectively 55 mv, 350 mv and 500 mv typical at 25°C.

The magnitude of the negative conductance is equal to the slope di/dv of the voltage current characteristic. For a one milliampere germanium tunnel diode the negative conductance is between 0.006 and 0.010 mhos corresponding to a negative resistance between 100 and 160 ohms.

Temperature Characteristics

Variation of the tunnel diode parameters with temperature is a matter of extreme importance to the circuit designer. Figure 2.2 shows the voltage-current characteristic of a typical germanium tunnel diode at temperatures of -50°C, 25°C, and 100°C.



$T_A = -50°C$ $T_A = +25°C$ $T_A = +100°C$

VOLTAGE-CURRENT CHARACTERISTIC CURVES OF A TYPICAL GERMANIUM TUNNEL DIODE

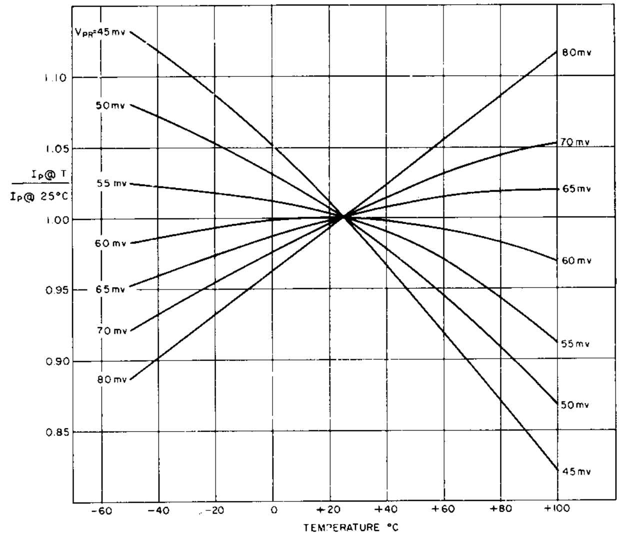
FIGURE 2.2

Note that the peak voltage, valley voltage and forward voltage all decrease with increasing temperatures while the valley current increases with increasing temperature. The peak current may increase or decrease with temperature depending on the doping level and the resistivity of the semiconductor material. This can best be illustrated by Figure 2.3 showing the temperature variations of the peak current for units with different peak voltages. V_p exhibits good correlation with the doping level and hence with the temperature characteristic of the diode.

The parameter V_p is dependent on the dopants and resistivity used. For the diode shown in Figure 2.2, the peak current is a maximum at approximately 25°C and decreases at higher and lower temperatures.

Each application generally has a different temperature problem. For example, in switching circuits the primary concern might be the stability of the peak current since it determines the switching threshold, although the changing forward voltage can affect the amplitude of the output voltage pulse.

In oscillators where direct matching is not required, it may be important only to make sure that at the lowest operating temperatures the device is driven from a voltage source which requires that the re-



VARIATION OF NORMALIZED I_P WITH TEMPERATURE FOR DIFFERENT VALUES OF V_P

FIGURE 2.3

sistance of the source supplying the voltage to the tunnel diode is much less than the negative resistance of the diode. Oscillators have been operated successfully over a temperature range from 4°K to over 573°K, a remarkably wide operating range. In amplifiers where some degree of match between the diode negative conductance and the positive circuit conductances is required, it is obvious that this match must be maintained over the required operating temperature range.

The variation of the important DC parameters between -50°C and +100°C is shown in Figure 2. 4 for 1 ma germanium tunnel diodes of the 1N2939 or 1N2940 types.

CHARACTERISTIC	SYMBOL	IN2939	IN2940
PEAK POINT VOLTAGE	V_P	55 MV	55 MV
TEMPERATURE COEFFICIENT	$\Delta V_P/\Delta T$	60 μV/°C	60 μV/°C
VALLEY POINT VOLTAGE	V_V	350 MV	350 MV
TEMPERATURE COEFFICIENT	$\Delta V_V/\Delta T$	-1.0 MV/°C	-1.0 MV/°C
FORWARD POINT VOLTAGE AT PEAK CURRENT	V_{FP}	500 MV	500 MV
TEMPERATURE COEFFICIENT	$\Delta V_{FP}/\Delta T$	-1.0 MV/°C	-1.0 MV/°C
PEAK TO VALLEY RATIO	I_P/I_V	8 MIN	5 MIN
VALLEY POINT CURRENT TEMPERATURE COEFFICIENT	$\Delta I_V/\Delta T$	+0.75 %/°C	+0.75 %/°C
CONDUCTANCE TO PEAK CURRENT RATIO	g_d/I_P	9.0 MHO/AMP	6.5 MHO/AMP
CONDUCTANCE TEMPERATURE COEFFICIENT	$\Delta g_d/\Delta T$	-0.5 % /°C	-0.5 % /°C
CAPACITANCE TO PEAK CURRENT RATIO	C/I_P	6 Pf/ma	4 pf/ma

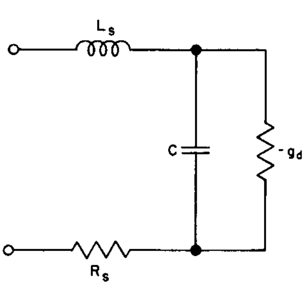
TYPICAL ELECTRICAL CHARACTERISTICS OF GERMANIUM TUNNEL DIODE

FIGURE 2.4

Note that the peak point voltage has a temperature coefficient of -0.06 millivolts/^{o}C and the forward point voltage has a temperature coefficient of -1.0 millivolts/^{o}C as compared with a value of -2.5 millivolts/^{o}C for the forward drop of a conventional diode or transistor. For detailed specifications of individual tunnel diode types, see the specification sheet section of this manual.

Frequency Limitations

The small signal (AC) equivalent circuit for the tunnel diode biased in the negative conductance region is shown in Figure 2.5. The inductance, L_s, in the equivalent circuit is relatively low and is determined primarily by the inductance of the leads. A small amount of series resistance, R_s, is also present which is determined by the bulk resistance of the semiconductor material. The capacity, C, is primarily due to the capacity of the junction although a small portion of the capacity is due to the leads and the package. As the junction capacity is decreased for lower current units, the package capacitance becomes more and more important however. The negative conductance, $-g_d$, in the equivalent circuit is equal to the slope of the voltage-current characteristic at the particular bias point under consideration. The value of the negative conductance can be assumed to be independent of frequency, the chief limitations in the frequency response of the tunnel diode being determined by the parasitic elements in the equivalent circuit (R_s, L_s, C).



TYPICAL VALUES OF 1N2939 PARAMETERS	
SERIES INDUCTANCE, L_s *	6mμh
TOTAL CAPACITANCE, C *	7μμfd
SERIES RESISTANCE, R_s	1.5Ω
NEGATIVE CONDUCTANCE, $-g_d$	.006–.01mho
NEGATIVE RESISTANCE, r_d	100–150Ω
PEAK POINT CURRENT, I_p	1ma
VALLEY POINT CURRENT, I_v	.1ma
PEAK POINT VOLTAGE, V_p	55mv
VALLEY POINT VOLTAGE, V_v	350mv
FORWARD POINT VOLTAGE, V_{f_p}	500mv
* 1/8 INCH LEAD LENGTH	

SMALL SIGNAL EQUIVALENT CIRCUIT AND TYPICAL VALUES OF PARAMETERS

FIGURE 2.5

Two significant frequency figures of merit can be assigned to the tunnel diode:

(a) resistive cut-off frequency, $$f_{ro} = \frac{|g_d|}{2\pi C}\sqrt{\frac{1}{R_s\,|g_d|} - 1} \qquad (2.1)$$

(b) self-resonant frequency $f_{xo} = \frac{1}{2\pi}\sqrt{\frac{1}{L_s C} - \left(\frac{g_d}{C}\right)^2}$ (2.2)

Both these frequencies are derived from the equivalent circuit of Figure 2.5. The resistive cut-off frequency is the frequency at which the real part of the diode impedance measured at its terminals goes to zero. The tunnel diode can not amplify above this frequency. The self-resonant frequency is the frequency at which the imaginary part of the diode impedance goes to zero. It should be pointed out that both frequencies are reduced by external circuit components and therefore the highest possible operating frequency is very circuit dependent. In a transistor package the tunnel diode is limited to frequencies below 1 Kmc, this limit being due primarily to the lead inductance. Microstrip or microwave packaging, owing to its inherently lower inductance, can potentially raise the frequency capabilities by an order of magnitude or more.

Nuclear Radiation Effects

Encouraging results have been obtained from preliminary investigations of the effects of nuclear radiation on the characteristics of tunnel diodes. Under a doseage of 3×10^{14} NVT (90% thermal, 10% fast), no apparent change in the electrical characteristics was observed except for the noise figure which increased by approximately 20% at the point of maximum negative conductance and by 100% near the valley point.

At a doseage of 5×10^{15} NVT, the valley current increased by about 25% while the other DC characteristics had not changed. The noise figure increased by a factor of 3 at the point of maximum negative conductance while the noise figure in the vicinity of the valley point was immeasurably high.

A Battelle Memorial Institute report[16] shows that a 10 ma I_p germanium tunnel diode with an initial valley point current of 1 ma had a valley point current of 3.6 ma at a radiation flux of 2.4×10^{16} NVT (fast neutron/cm^2). Above this flux-density large changes of valley current would occur for small increases in flux. It should be pointed out that although I_v increases while both V_v and V_{fp} decrease with flux density, the characteristic seems to remain unchanged in the reverse direction (important for backward diodes). The same holds true for the positive slope region below the peak. I_p and V_p are totally unaffected while the negative conductance slope does not appear to change radically up to flux levels of 5.5×10^{16} NVT (fast neutrons/cm^2).

Maximum Ratings

The "Absolute Maximum Ratings" of tunnel diodes are primarily dependent on the junction temperature produced by the current flowing through it, the physical size of the junction area and the efficiency with which heat is removed.

The high current density material used in tunnel diodes to provide good g_d/C ratios causes the size of the junction to be very small. A one milliampere unit, for instance, might have a junction diameter of less than 3×10^{-4} inch. A large current flowing through such a small cross-sectional area can rapidly create enough heat to melt the alloys used.

As large currents can flow through the tunnel diode junction even at low forward or reverse voltages, an absolute maximum current specification (forward and reverse) is more meaningful than voltage or dissipation limits per se.

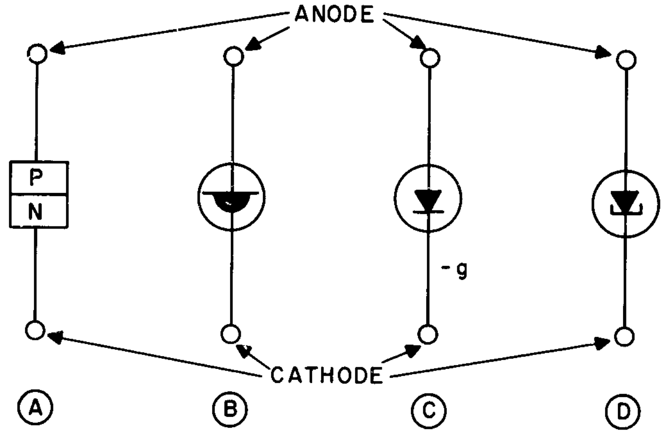
Also given in the "Absolute Maximum Ratings" are the storage and operating junction temperature ranges for the device. Actually the device proper can operate from liquid helium temperatures (4^oK) to the melting point of the alloys. The package might however limit this operating range, as the glass seals might become leaky or even crack at very low temperatures. Even under those conditions, the device might continue to function as it does not necessarily have to be hermetically sealed. Most packages will readily operate over a -50^oC to $+100^oC$ range or better.

Symbols

Several different symbols for tunnel diodes are in current use. The more popular symbols are shown in Figure 2.6 together with the PN junction equivalent. Many of the other symbols which have been used in the literature are not recommended since they are in direct contradiction of AIEE/IRE standards on graphical symbols.

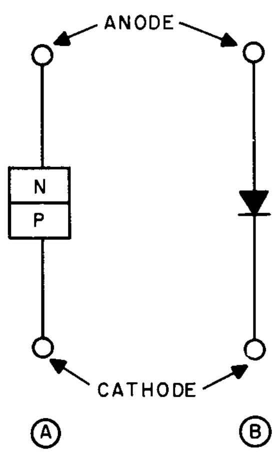
The anode and cathode designations for each of the symbols are indicated in Figure 2.6. If the anode is positively biased with respect to the cathode the tunnel diode will display its negative conductance characteristic and the high voltage injection characteristic. Figure 2.6a shows schematically the PN junction structure of the device. Figure 2.6b shows the symbol which is used exclusively in this manual. The polarity is easily remembered by imagining the arrow on a conventional diode to "tunnel" through the bar after which it appears on the other side of the bar in the blunt, rounded form as shown. This symbol has the advantage that it can be combined with other symbols to indicate complex devices. For example, a PNP transistor with a tunnel emitter can be indicated by adding a semicircle to the conventional symbol for a PNP transistor. The symbol shown in Figure 2.6c indicates a tunnel diode by means of a -g next to the symbol signifying that the device is a negative conductance device. This symbol tends to imply that the device is used for small signal operation and hence is generally used in amplifiers and low power oscillators. Note that the -g can not be used inside the circle without violating IRE standards. The symbol shown in Figure 2.6d is frequently advocated because it contains the letter T in its structure.

The symbol for the back diode (backward diode) is shown in Figure 2.7 together with the PN junction equivalent. Since the back diode can be considered to be a conventional diode with a low reverse breakdown voltage a conventional diode symbol can be used. Note that the anode is connected to the N material and the cathode is connected to the P material since the direction of easy current flow is opposite to that of a conventional diode.



COMMONLY USED SYMBOLS FOR THE TUNNEL DIODE

FIGURE 2.6



PREFERRED SYMBOL FOR THE BACK DIODE

FIGURE 2.7

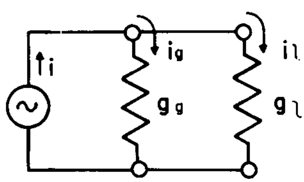
CHAPTER 3

TUNNEL DIODE AMPLIFIERS

Introduction

Since by definition a positive conductance dissipates energy, it follows that a negative conductance generates energy. This is the basis of negative conductance (or resistance) amplifiers.

Consider first the case of an ideal sinusoidal generator, having an internal conductance g_g, delivering a rms current i to a load g_l (see Figure 3.1).



$$P_l \, max = \left(\frac{i}{2}\right)^2 \cdot \frac{1}{g_l} = \frac{i^2}{4 g_g} \quad \left\{ g_g = g_l \right.$$

$$P.G._{AV} = \frac{P_l}{i^2/4 g_g} = 1$$

POWER TRANSFER FROM GENERATOR TO LOAD

FIGURE 3.1

Suppose g_l is varied to maximize the power delivered to it. The latter occurs under a "matched" condition, namely when $g_g = g_l$. Under this condition, the generator current splits evenly between the two branches of the circuit. The maximum power delivered to the load is thus given by:

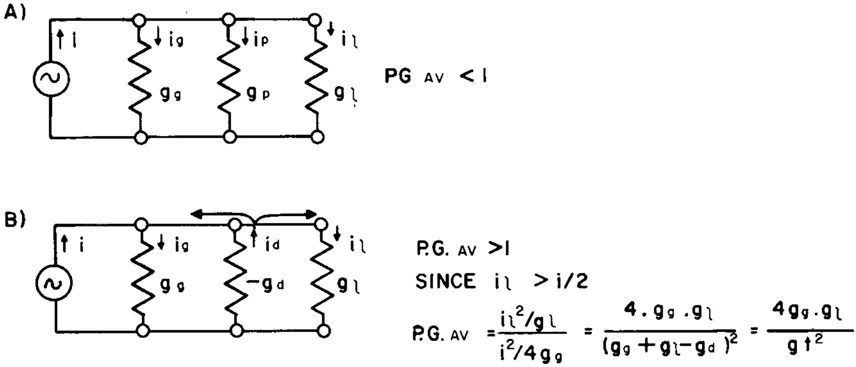
$$P_{l \, max.} = \left(\frac{i}{2}\right)^2 \cdot \frac{1}{g_l} = \frac{i^2}{4 g_g} \tag{3.1}$$

As this is the maximum available power from the generator, one might use this value to find the "transducer gain", defined as the power delivered to an arbitrary load divided by the power available from the generator.

$$P.G._{avail.} = \frac{P_l}{i^2/4 \, g_g} \tag{3.2}$$

This gain can obviously not exceed unity.

Suppose that a positive conductance g is placed in parallel with the generator and load (see Figure 3.2a). Part of the generator current flows into this positive conductance and the transducer gain will be less than unity. If, however, a negative conductance (-g) replaces the positive shunt conductance (as in Figure 3.2b), the current flowing in the added branch is out of phase to that of the previous case and hence current is coming out instead of going into this negative conductance branch. Hence this branch acts as a generator and drives additional current into the circuit. The transducer gain can now exceed unity since the load current i_l can now be greater than $i/2$.



POWER TRANSFER HAVING NEGATIVE CONDUCTANCE IN CIRCUIT

FIGURE 3.2

By proper adjustment of the load, this gain may be infinite as can be seen from the following equation:

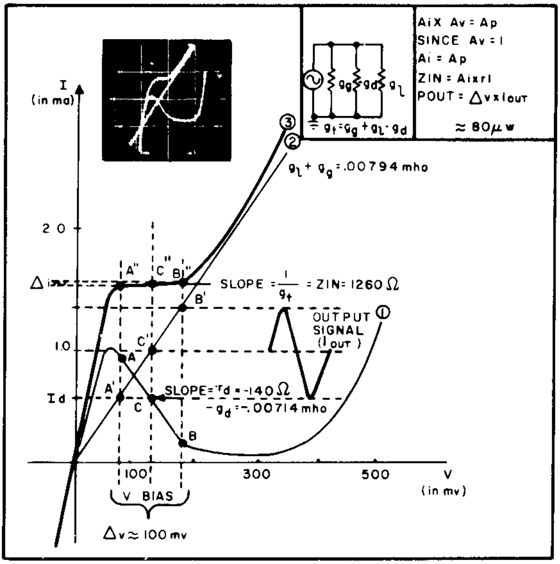
$$P.G._{avail.} = \frac{P_{load}}{P_{av.\ gen.}} = \frac{i_l^2/g_l}{i^2/4g_g} = \frac{4 \cdot g_g \cdot g_l}{(g_g + g_l - g_d)^2} \qquad (3.3)$$

As the sum of the positive conductances $(g_g + g_l)$ is made equal to the negative conductance $-g_d$, the denominator becomes zero and the P.G. infinite. Hence, the degree of "Matching" between $g_g + g_l$ and $-g_d$ will determine the gain. If $g_g + g_l$ is smaller than $-g_d$, the denominator is negative and the circuit unstable. Thus for stable values of gain $g_g + g_l$ must be greater than $-g_d$ and for this gain to be large $g_g + g_l$ must only be slightly larger than $-g_d$. Figure 3.3 shows a graphical representation of the tunnel diode V-I curve, the load line and the resultant parallel impedance.

The tunnel diode capacity is in parallel with the negative conductance and as the frequency increases will become a more and more effective shunt. In a parallel type amplifier, an inductance L is put across the tunnel diode in order to resonate the circuit at a frequency f_o, where:

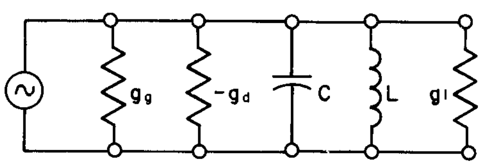
$$f_o = \frac{1}{2\pi\sqrt{LC}} \qquad (3.4)$$

(see Figure 3.4).



GRAPHICAL ANALYSIS OF PARALLEL AMPLIFIER STAGE

FIGURE 3.3



TRANSDUCER GAIN:

$$P.G. = \frac{4\, g_g\, g_l}{g_t^2 + \omega^2 C^2 \left(1 - \frac{\omega_0^2}{\omega^2}\right)^2}$$

AT RESONANCE $\omega_0 = \omega$ AND

$$P.G. = \frac{4 \cdot g_g \cdot g_l}{g_t^2}$$

EQUIVALENT CIRCUIT OF A TUNNEL DIODE "PARALLEL" AMPLIFIER

FIGURE 3.4

The generalized gain equation for this circuit now becomes:

$$P.G._{\text{avail.}} = \frac{4\, g_g\, g_l}{g_t^{\,2} + \omega^2 C^2 \left(1 - \frac{\omega_o^{\,2}}{\omega^2}\right)^2} \qquad (3.5)$$

and since at resonance $\omega = \omega_o$, the available power gain at resonance is still:

$$P.G._{\text{avail.}} = \frac{4\, g_g\, g_l}{g_t^{\,2}} \qquad \text{(as per equation 3.4)}$$

where $g_t = g\ \text{total} = (g_g + g_l - g_d)$ (3.6)

From equation (3.5), it can be seen that gain decreases "off-resonance" and since bandwidth is defined as the difference between the two side band frequencies at which the gain is 3 db down, bandwidth (BW) can be defined as:

$$BW = \frac{\omega_2 - \omega_1}{2\pi} = \frac{g_g + g_1 - g_d}{2\pi C} = \frac{g_t}{2\pi C} \tag{3.7}$$

From equations (3.5) and (3.7), it can be seen that as the gain is increased (by decreasing g_t) the bandwidth decreases, becoming zero at the point where g_t is zero, which coincides with the point of infinite gain (oscillations).

The gain-bandwidth product of the circuit of Figure 3.4 can be expressed fully in terms of the circuit parameters as follows:

$$\sqrt{P.G.} \cdot BW = \sqrt{\frac{g_g\, g_1}{\pi C}} \tag{3.8}$$

if $g_g + g_1$ is approximately matched to $-g_d$, $(g_g + g_1 \approx |-g_d|)$, then:

$$\sqrt{P.G.} \times BW = \sqrt{\frac{g_1\,(g_d - g_1)}{\pi C}} \tag{3.9}$$

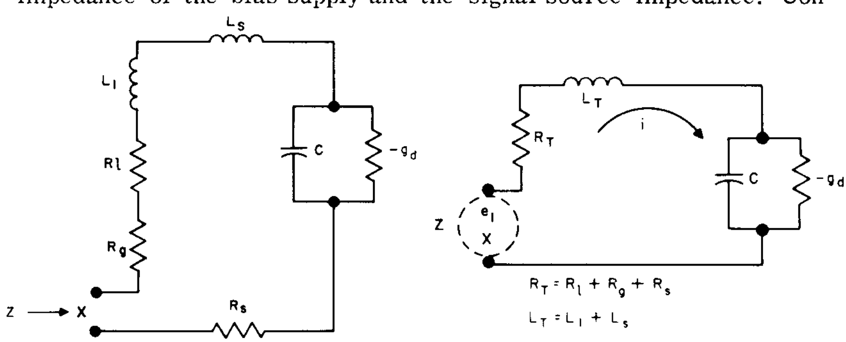
and hence the gain bandwidth product will be largest when $g_1 = g_g = g_d/2$, at which time it is equal to:

$$\sqrt{P.G.} \times BW = \frac{g_1}{\pi C} = \frac{g_g}{\pi C} = \frac{g_d}{2\pi C} \tag{3.10}$$

Other factors influencing the choice of g_g and g_1 for a given g_d are discussed in the section on amplifier noise.

Stability Criteria of Tunnel Diode Amplifiers

Successful linear operation of a tunnel diode amplifier depends on the stability of the complete system, including in particular the internal impedance of the bias supply and the signal source impedance. Con-



(A) SERIES CONFIGURATION OF AMPLIFIER OR OSCILLATOR CIRCUIT

(B) SIMPLIFIED SERIES CONFIGURATION

FIGURE 3.5

sidering the equivalent circuit of Figure 2.5 a basic "series" amplifier circuit can be reduced to that shown in Figure 3.5 where $R_T = R_g + R_1 + R_s$, $L_T = L_s + L_1$, C is the total diode capacity and $-g_d$ the negative conductance of the diode at the operating current and voltage.

To determine the system stability, one can examine the distribution of poles or zeros of the circuit determinant in the complex S-plane.

If the zeros of Z seen at the input fall in the right half of the S-plane, the system is unstable. Conversely, if the zeros fall in the left half side of the S-plane, the circuit is stable. The input impedance is given as:

$$Z_{(S)} = \frac{S^2 L_T C + S\,(R_T C - L_T\,|-g_d|) + (1 - R_T\,|-g_d|)}{SC - |-g_d|} \tag{3.11}$$

and the zeros are:

$$S = -\frac{1}{2}\left(\frac{R_T}{L_T} - \frac{|-g_d|}{C}\right) \pm \sqrt{\frac{1}{4}\left(\frac{R_T}{L_T} - \frac{|-g_d|}{C}\right)^2 - \frac{1 - R_T\,|-g_d|}{L_T C}} \tag{3.12}$$

then S will have a negative real part only if both:

$$\frac{R_T}{L_T} - \frac{|-g_d|}{C} > 0 \tag{3.13}$$

and

$$1 - R_T\,|-g_d| > 0 \tag{3.14}$$

This can be rewritten as:

$$\frac{1}{|-g_d|} > R_T > \frac{L_T\,|-g_d|}{C} \tag{3.15}$$

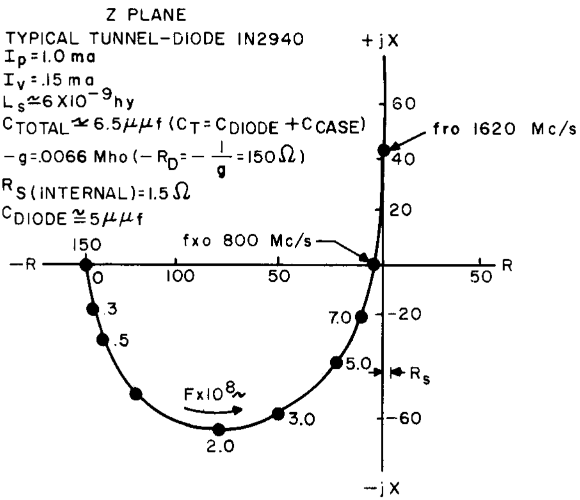
Frequency Limitation

A Nyquist plot of the real and imaginary component of the input impedance divulges two characteristic frequencies. $f_1 = f_{ro}$ is the frequency at which the real part of the circuit impedance goes to zero and $f_2 = f_{xo}$ is the frequency at which the imaginary component of the circuit impedance goes to zero.

$$f_{ro} = \frac{|-g_d|}{2\pi C}\sqrt{\frac{1}{|-g_d|\,R_T} - 1} \tag{3.16}$$

$$f_{xo} = \frac{1}{2\pi}\sqrt{\frac{1}{L_T C} - \frac{g_d^2}{C^2}} \tag{3.17}$$

Figure 3.6 is the Nyquist plot of a typical 1N2940 (device only) where R_T and L_T are replaced by R_S and L_S in the above equation.



CHARACTERISTICS OF A TYPICAL TUNNEL DIODE

FIGURE 3.6

Now several conditions could prevail:

(1) $f_{ro} \le 0$, the device can only switch as $R_T \ge 1/\text{-}g_d$.

(2) $f_{xo} \le 0$, the device can only operate in the relaxation oscillation mode.

(3) If $f_{ro} > f_{xo}$, then at frequencies up to f_{xo}, the circuit sees some real and imaginary values of negative impedance and hence it can oscillate in a relaxation mode.

(4) If $f_{ro} = f_{xo}$, then at this point both the real and imaginary parts of the input impedance are zero, and the circuit once shocked into oscillation by any random noise or transient will yield free sinusoidal oscillations.

(5) If $f_{xo} > f_{ro}$ the circuit will be stable and amplification can take place, since the imaginary term of the input impedance does not go to zero until beyond f_{ro} at which time the circuit is already "cut off".

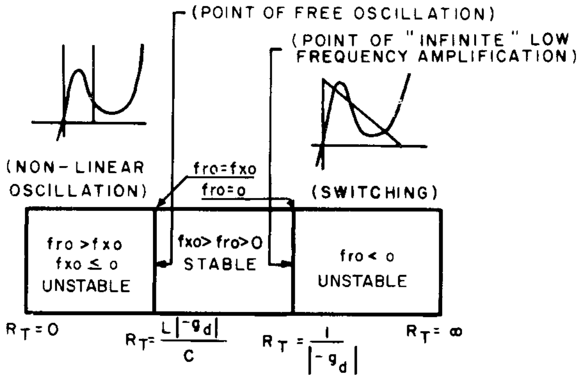
Figure 3.7 is a graphical illustration of the above conditions. In this figure the two roots $R_T = L\, g_d/C$ and $R_T = 1/\text{-}g_d$ are taken from equations 3.11, 3.13, and 3.14.

Summary of Gain and Stability Conditions

The successful design of a tunnel diode amplifier circuit depends on the proper application of the stability criteria. The amplifier circuit must meet the following criteria:

(1) First, the circuit must be biased stably in the negative conductance region from a voltage source ($R_T < 1/\text{-}g_d$ and $f_{ro} > 0$).

(2) Second, the circuit inductance must be smaller than $1 < R_T C/|\text{-}g_d|$ ($f_{xo} > f_{ro}$).



CONDITION FOR STABLE OPERATION SHOWN GRAPHICALLY

FIGURE 3.7

(3) To obtain large values of stable gain the sum of the positive conductances ($g_g + g_l + g_x$) must be always only <u>slightly</u> larger than the negative conductance of the diode. ($g_g + g_l + g_x \geq -g_d$), which also means that the greatest value of gain is obtained very nearly when $R_T = 1/-g_d$.

(4) The greatest gain bandwidth product is obtained when the load conductance is approximately equal to the generator conductance; in which case, they are both approximately equal to one-half of the diode negative conductance. ($g_l = g_g = -g_d/2$). At that point $\sqrt{\text{P.G.}} \times \text{B.W.} = g_d/2\pi C$. This also points out that to obtain greater gain bandwidth product, diodes with the largest possible figure of merit (g_d/C) should be used.

(5) The above condition does not take into consideration amplifier noise figure, temperature and bias variations as well as certain other practical considerations explained a little later.

Noise Considerations

It is generally known that the negative resistance of a tunnel diode exhibits shot noise. A tunnel diode also has a parasitic series resistance exhibiting thermal noise. At very low frequencies (in the audio range) the tunnel diode also exhibits 1/f noise. Despite these noise sources, tunnel diode amplifiers can be made with noise figures lower than 3 db.

Two requirements must be met to obtain a low noise figure tunnel diode circuit. First, the tunnel diode must itself exhibit low noise and secondly, the circuit must be optimized for low noise performance. The tunnel diode's basic performance can be described by a noise figure of merit which is a function only of the diode parameters, its bias point and the operating frequency.

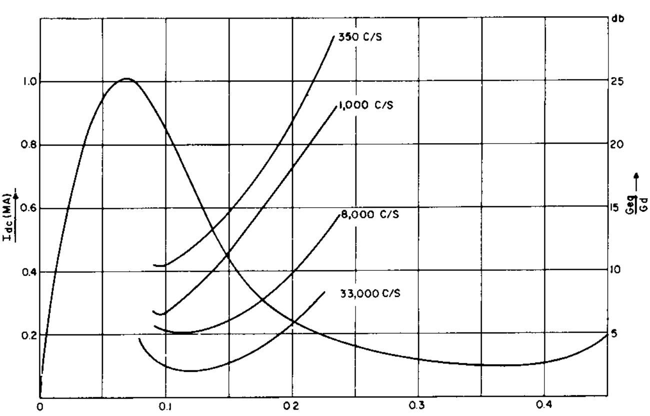
When the diode is used as an amplifier, the circuit noise figure can be made to asymptotically approach the noise figure of merit of the diode. The latter is normally the minimum limit. The amplifier's noise figure is made to approach the figure of merit of the diode by suitable adjustment of circuit parameters.

E. G. Nielsen's[7] paper on the noise performance of tunnel diodes equates the noise figure of merit of the diode as being:

$$F = \frac{\left(1 + \frac{g_{eq}}{g_d}\right)}{\left(1 - \frac{R_S}{|R_D|}\right)\left[1 - \left(\frac{f}{f_{ro}}\right)^2\right]} \qquad (3.18)$$

In this equation the ratio of g_{eq}/g_d assesses the shot noise relative to the diode negative conductance. At frequencies above the audio range, where $I_{eq} = I_{DC}$ (DC bias current through the tunnel diode) g_{eq} is equal to approximately $20\ I_{DC}$. At lower frequencies, I_{eq} is greater than I_{DC} ($g_{eq} = eI_{eq}/2KT$) and the results can be seen in Figure 3.8 where g_{eq}/g_d is plotted at various frequencies thereby exhibiting the 1/f noise contribution of the diode. At 33 Kc/s (which was not quite the break frequency) at a bias of 125 mv, g_{eq}/g_d = 2 db which would yield a total minimum noise figure F = 4.2 db. The ratio of R_S/R_D assesses the parasitic series diode resistance relative to the negative resistance and hence this would assess the thermal noise. At frequencies near the device "cut off" frequency (f_{ro}) where the real part of the negative resistance becomes very small, the ratio of R_S/R_D becomes quite important. Thus the last ratio assesses the frequency dependance of R_D.

With germanium tunnel diodes $R_S/R_D \cong .01$ and $g_{eq}/g_d \cong 1$. Using these figures a device noise figure of close to 3.0 db can be obtained at medium frequencies, but at $f = .707\ f_{ro}$ the minimum would go to 6 db while it would exceed 10 db at $f = .9\ f_{ro}$ .... f_{ro} in this case is the device "cut off" frequency and hence in equation 3.16 R_S replaces R_T.



I/f NOISE IN TUNNEL DIODES

FIGURE 3.8

The noise figure of the amplifier circuit is given by the following equation:

$$N.F. = 1 + \frac{20\, I_{DC}}{g_g} + \frac{T_1\, g_1}{T_g\, g_g} \qquad (3.19)$$

and can be minimized when g_g is made as large and g_1 as small as possible. Referring back to Equations 3.3 and 3.5, the highest gain is obtained when $g_g + g_1 \cong g_d$. Therefore, the optimum value for g_g would make it equal to g_d, with g_1 by necessity becoming very small. Thus to summarize the noise considerations:

1) The minimum attainable circuit noise figure is close to the noise figure of the device as given by equation 3.18, which can be about 3 db.

2) To obtain a diode with such a low noise figure, the ratios of R_S/R_D and $g_{eq}/g_d \cong 20\, I_{DC}/g_d$ must be at a minimum and the operating frequency above the 1/f range and not too close to f_{ro} ($< .707\, f_{ro}$).

3) The circuit must then be optimized for low noise by matching the generator conductance to the negative tunnel diode conductance while the load conductance must be very small. (Example: A generator (r_g) of 100Ω could be driving a (R_D) of -100Ω looking into a load impedance (r_1) of 1000Ω. Assume $C = 5 \times 10^{-12}$. Under the above condition, the gain bandwidth product is not optimized $g_d/2C = 318 \times 10^6$, but could still be adequate $\sqrt{P.G.} \cdot B.W. = \sqrt{g_g g_1/\pi C} = \sqrt{10^{-2} \times 10^{-3} / 3.14 \times 5 \times 10^{-12}} = 201 \times 10^6$.

Practical Design Considerations

Several other factors must be taken into consideration before actually designing an amplifier circuit. These are:

1) The non-linearity of $-g_d$
2) The bias dependance of $-g_d$
3) The temperature dependance of $-g_d$

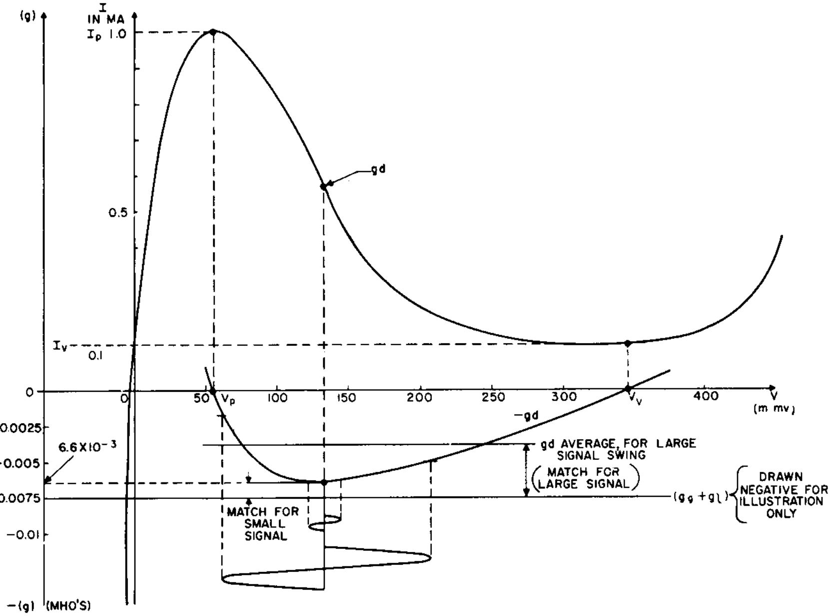
1) Non-Linearity Of The Diode Negative Conductance-

As a result of the non-linearity of $-g_d$ (see g_d vs. V in Figure3.9), the tunnel diode amplifier is endowed with two additional problems.

a) this non-linearity will result in some distortion.
b) another result is a built-in "automatic gain control" effect.

The first problem is evident and fortunately the degree of non-linearity can be kept relatively low. The second problem can best be illustrated by the graph in Figure 3.9.

This graph shows the variation of I and g_d versus voltage. Looking at g_d we see it to be negative between Vp and Vv. Remembering that in an amplifier the resonant gain is obtained through the close matching of the sum of the positive circuit conductances and the negative diode conductance, one might draw these positive conductances as



CONDUCTANCE VERSUS VOLTAGE — CHARACTERISTIC

FIGURE 3.9

a straight line and for illustration purposes only, forget the sign of these positive conductances and draw them as a negative conductance. The distance between line A - B and the negative conductance is the degree of mismatch. Now applying a small signal (say 2mv-rms), one can see that the degree of mismatch is relatively constant over this small voltage range and a certain gain results.

If a larger signal is applied, however, the voltage swing is larger and due to the greater non-linearity of g_d over this range, only an "average" match occurs. Or, one might say the negative conductance is no more equal to $-g_d$ at point OP but is equal to some lesser, average value of $-g_d$. This, of course, increases the distance between the average $-g_d$ line and the $(g_g + g_l)$ line thusly decreasing the gain for larger signal swings.

2) Bias Dependence of the Negative Conductance-

Biasing at the center of the more linear portion of the tunnel diode characteristic (near the inflection point -- point of maximum negative conductance) will allow the greatest voltage swing, hence the greatest dynamic range. For germanium tunnel diodes, this point is around 130 - 140 mv and the more linear portion of the characteristic will be between 80 and 180 mv, thus allowing a maximum signal swing of 100 mv peak to peak.

For high temperature operation, the large signal distortion will increase mainly as a result of the increase in valley current. If the degree of distortion is not acceptable, smaller signal swings will alleviate this problem. The greatest bias problem is that the negative conductance region is not linear. Slight variations in the bias operating point with the resulting change in negative conductance, can cause large changes in gain. It is thus essential to ensure a very stable bias voltage. Some of the methods to obtain a stable, low impedance bias supply are:

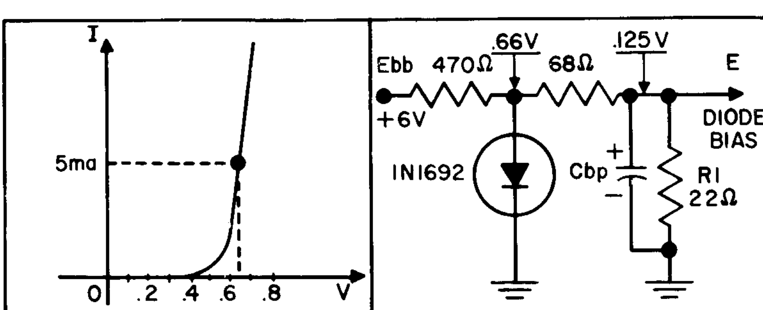
1) The use of mercury cells.
2) The use of forward biased diodes as voltage regulators.
3) The use of breakdown diodes as voltage regulators.

An example of the use of a forward biased diode for bias stabilization is shown in Figure 3.10. Here an inexpensive silicon diode is heavily biased in the forward direction so that it exhibits a low voltage and a low dynamic resistance. A low impedance voltage divider is used to reduce the diode voltage to the value desired for biasing of the tunnel diode. Large variations of supply voltages will cause only minor variations in tunnel diode supply voltage. At first, an apparent shortcoming of this circuit seems to be the fact that this voltage regulator is somewhat temperature sensitive and hence the operating point will shift with temperature. This might be desirable however in view of the next problem.

3) Temperature Dependence of $-g_d$ -

Early work on germanium tunnel diodes indicated that the negative conductance varies roughly linearly with temperature at a rate of $-0.5\%/^{\circ}C$. Such a shift would again cause gain changes in an amplifier circuit. Several methods could be used to compensate for this effect:

a) The use of temperature sensitive resistors in series or parallel with one of the circuit legs to keep the same degree of match.
b) The use of diodes in the bias supply (as per Figure 3.10) to produce small changes in bias resulting in a changing $-g_d$.

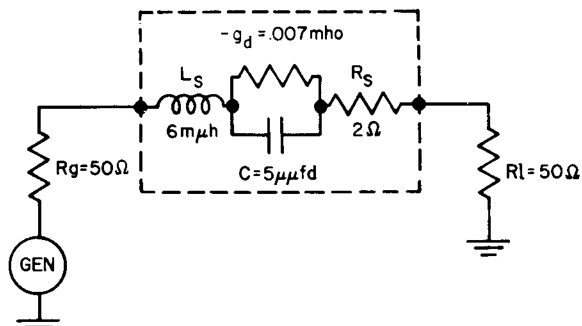


SILICON DIODE VOLTAGE REGULATOR
FIGURE 3.10

The voltage across the 1N1692 decreases at about 2.5 mv/^oC as the temperature is increased. In the tunnel diode, $-g_d$ decreases at 0.5%/^oC as the temperature is increased. Subsequently, if the operating point is chosen slightly beyond the inflection point (say 150 mv or so) as $-g_d$ decreases with increasing temperature, the bias decreases and the operating point moves on the characteristic towards an increased $-g_d$, thereby compensating for the original change. Furthermore, if the 2.5 mv/^oC is too fast a change in bias (overcompensation) then a tunnel diode can be substituted for the 1N1692 which will result in a change of only 1 mv/^oC (see Figure 2.4).

Amplifier Design Procedure

In this circuit (see Figure 3.11), the source is a 50 ohm generator, the load is also 50 ohms while the series resistance (R_S) of the device is 2 ohms. Hence R_T = 50 + 50 + 2 = 102 ohms. Use is made of a 1N2939 having a 5$\mu\mu$fd capacitance and a negative conductance of 7 millimhos (-rd = 143 ohms) at the inflection point.



A.C. SERIES LOOP CIRCUIT

FIGURE 3.11

In order to abide by the previously mentioned stability criteria, the real part of the negative conductance must be made equal to zero at the operating frequency. This also means that the circuit cut-off frequency is made equal to the operating frequency. Hence,

$$R_T - \frac{|-g_d|}{|-g_d|^2 + \omega^2 C^2} = 0, \text{ thus } R_T = \frac{1}{|-g_d|\left(1 + \frac{\omega^2 C^2}{g_d^2}\right)} \qquad (3.20)$$

R_T must be therefore made equal to:

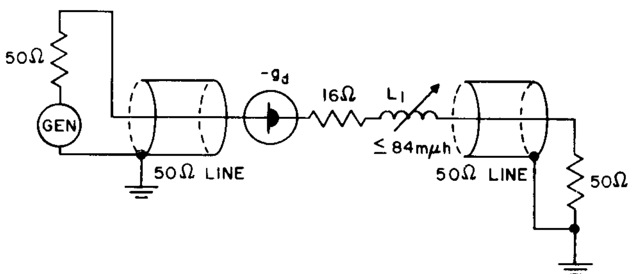
$$R_T = \frac{143}{1.21} \cong 118\Omega$$

Since the present series loop only exhibits a $R_T \cong 102\Omega$, a 16 Ω series resistance must be added to meet the previously outlined gain and stability criteria.

The last part of the AC circuit design procedure is the choice of the tuning inductance L_1. To get the highest value of stable gain L_T total must be only slightly smaller than the oscillation criteria $L_T < R_T C / |-g_d|$ which here must be:

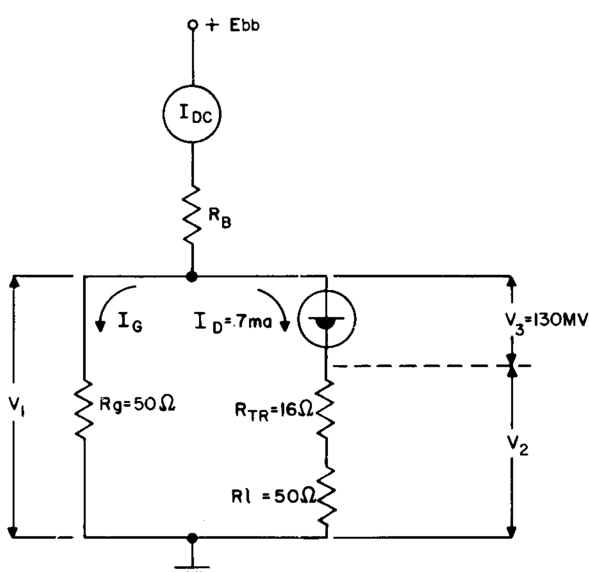
$$L_T < \frac{118 \times 5 \times 10^{-2}}{7 \times 10^{-3}} = 84.3 \text{ m}\mu\text{h}$$

Since 2-12 mμh are inherent in the leads of the device (depending on lead length) and some stray circuit inductance will be found in the circuit, the actual coil (L_1) will have to present a slightly smaller inductance value.



AC CIRCUIT OF 100 Mc/s AMPLIFIER STAGE
FIGURE 3.12

The bias arrangement can be derived in the following manner:



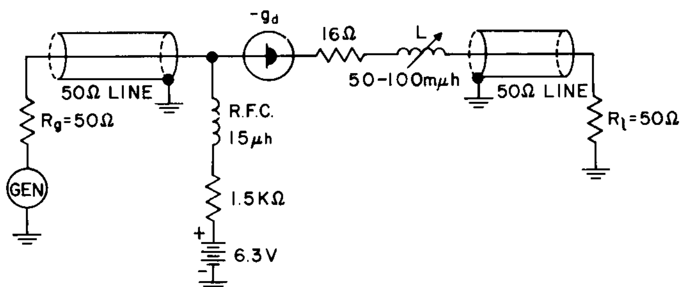
DC BIAS CIRCUIT FOR 100 Mc/s AMPLIFIER STAGE
FIGURE 3.13

Assuming that the inflection point occurs at 130 mv and .7 ma, then V_3 = 130 mv and I_D is .7 ma and V_2 is $(R_{TR} + R_L) I_D = (16 + 50)\ .7 \times 10^{-3}$ = 44 mv; therefore, V_1 = 130 + 44 = 174 mv.

I_G therefore is $174 \times 10^{-3}/50$ = 3.48 ma, and the total DC current $I_{DC} = I_G + I_D$ = 3.48 + .7 = 4.18 ma. If one were to use a 6.3 v battery then $R_B = 6.3 - .174/4.18 \times 10^{-3} = 6.126/4.18 \times 10^{-3}, \cong 1.5K\Omega$. In order to decouple the DC supply from the amplifier by at least a 10 KΩ inductive reactance,

$$L_{RF\ choke} > \frac{X_L}{\omega} = \frac{10^4}{6 \times 10^8} = 15\ \mu h$$

Figure 3.14 shows the complete circuit.



COMPLETE 100 Mc/s "SERIES" AMPLIFIER CIRCUIT

FIGURE 3.14

The measured results were 32 db gain at 100 Mc/s with a 20 Mc/s symmetrical bandwidth. As L_1 is increased toward $L_1 = R_T C / |-g_d|$, the gain increases at the expense of bandwidth magnitude and symmetry. Further refinements in bias and temperature stabilization would complete such an amplifier design.

Conclusions

Tunnel diode "Series" and "Parallel" type amplifiers can be stable, can have low noise, high gain, wide bandwidth providing all stability criterias are satisfied. Some of the static requirements are that the device be properly biased in the negative conductance region from a stiff voltage source, that the sum of the positive conductances just barely exceed the negative conductance (or if the signal swing is greater than a few millivolts -- the average negative conductance). Furthermore, f_{xo} must be made to exceed f_{ro} and be greater than zero. This permits the circuit to be "cut off" at the operating frequency while the imaginary component is still greater than zero, hence the circuit cannot oscillate. Although the circuit is "cut off" (stable) the device in the circuit pumps energy into the load,thus gain can be obtained.

In operation, these conditions must remain satisfied, hence the parameters must remain stable under changes of temperature and supply voltages (or be stabilized for such environmental changes).

Cascading tunnel diode amplifier stages is made difficult by the bilateral properties of this two terminal device. The amplifier is basically a positive feedback type circuit and unilateralization is impossible. Isolation between stages can be obtained by the use of isolators and circulators.

In a 1960 WESCON Paper, Dr. Schaffner[6] described a 450 Mc/s UHF amplifier circuit having 15 db gain, 12 Mc/s bandwidth and 5.5 db noise figure (he quoted further improvements in N. F. down to 4.2 db). The antenna VSWR change was only 1 - 1.12 db, while with an isolator the circuit could withstand a change in 1 - 1.8 db VSWR (with 3 db gain reduction). Several papers in the June and July 1960 issues of the IRE Proceedings dealt with the same general problem. In the July issue, John J. Sie[13] uses two ZJ-56 (1N2939) in conjunction with a SAGE 750 quarter wave stripline coupled Hybrid to obtain stable gain over a very wide frequency range (210 - 628 Mc/s). His results were a gain of 8.2 db ± .6 db with a noise figure of approximately 2 db.

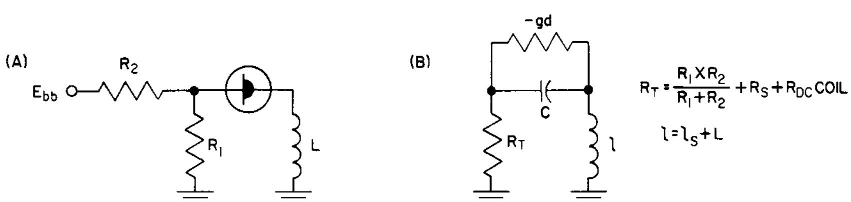
CHAPTER 4

OSCILLATORS

4.1 Tunnel Diode Oscillators

Because of its inherent negative conductance characteristic, the tunnel diode is ideally suited for oscillator operation. The temperature stability of its characteristics is excellent and the frequency stability of the resulting oscillator circuit primarily depends on the associated circuit components.

The simplest tunnel diode oscillator circuit is the "series" type (see Figure 4.1). In this circuit the diode, appropriately biased in the negative conductance region, is put in series with an external inductor.



SIMPLE "SERIES" OSCILLATOR AND EQUIVALENT CIRCUIT

FIGURE 4.1

In this arrangement, the analysis[10] shows that the circuit will oscillate freely in a sinusoidal manner when:

$$\frac{R_T}{|g_d|} = \frac{L}{C} \qquad (4.1)$$

at a frequency:

$$f_o = \frac{1}{2\pi}\sqrt{\frac{1 - R_T \cdot |g_d|}{L \cdot C}} \qquad (4.2)$$

Both the frequency and the stability of this circuit heavily depend on $|g_d|$ and since this conductance is not a constant, but is a time-average value which also varies with voltage and temperature, such a circuit is rather unstable.

Another circuit approach, must less subject to the various causes of instability, can be seen in Figure 4.2. The operating frequency of this "Series-Parallel" circuit is primarily dependent on the L & C of the tank circuit, and can therefore be quite stable. A detailed analysis for this circuit shows that the operating frequency is determined by:

$$f_o = \frac{1}{2\pi} \sqrt{\frac{1}{L(C + C_1)} - \frac{g_d^2}{C_1 (C + C_1)}} \qquad (4.3)$$

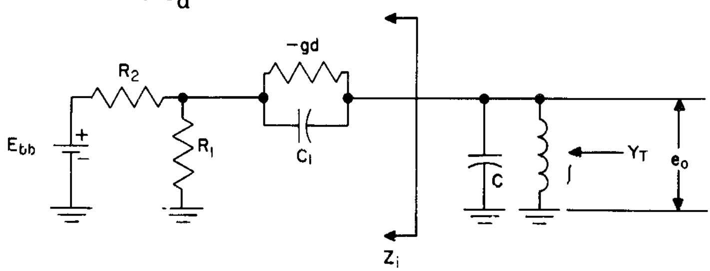
and that for stable sinusoidal oscillations,

$$R_T = \frac{g_d}{\omega^2 C_1^2 + g_d^2} \qquad (4.4)$$

It should be pointed out that several approximations have been made in the analysis. First, the diode series resistance and series inductance have been neglected as they will generally be small compared to the external lumped constants. Secondly, the term g_d, which in reality should be a time-average value where g_d can be expressed in function of time as:

$$g_d(t) = g_o + g_1 \cos \omega t + g_2 \cos 2\omega t + \ldots\ldots$$

is only approximated. Presently, only an "average" value of g_d is given by the manufacturer as it is the only parameter that is currently being specified. While observing the V-I characteristic, the tunnel diode is shunted by a variable resistance. When $R = 1/g_d$, the slope of the negative conductance becomes zero over the "active" range and since R is known, g_d is thus established.



SIMPLIFIED EQUIVALENT CIRCUIT OF "SERIES-PARALLEL" SINEWAVE OSCILLATOR

FIGURE 4.2

Limitations

The use of the tunnel diode as an active device in oscillator circuits involves certain problems. Obviously the first of these is the small amount of power output obtainable from such a circuit. It is apparent from a rough approximation of the V-I characteristics that:

$$P_{o\,max.} = \frac{\#V(I_p - I_v)}{2\sqrt{2}} \approx \frac{I_p}{28} \qquad (4.5)$$

hence it would take 28 amp-I_p to obtain one watt of power output or

#V is a coarse approximation of the "relatively" linear portion of the negative conductance slope - 100mv for Ge & 200mw for GaAs units.

2.8 amps to even get 100 mw. With a 2.8 amp unit, the negative resistance is only around 0.1Ω or less. Driving such a resistance from a stiff voltage source is difficult and inefficient.

Another problem is that such a high power unit has a rather large junction area, hence a relatively large junction capacitance. The magnitude of this capacitance seriously limits the frequency at which this power is available.

Further limitations can be found in coupling this oscillator to a load. The tunnel diode oscillator circuit only delivers a fraction of a volt (rms) to the tank circuit. Generally, as this circuit is coupled into other low impedance semiconductor circuits, a stepdown transformer only serves to lower this output swing.

Finally, the efficiency of such a circuit, driven from present conventional power supplies (relatively high DC voltages) is quite low. An efficient tunnel diode supply would provide a fraction of a volt with a large current capacity -- just the opposite of conventionally used supplies. On the other hand, solar cells and thermoelectric generators could probably be used to advantage in tunnel diode circuits.

Oscillator Design Procedure

Given the frequency and required output power, the choice of diode can be made. For one milliwatt of rms power, a 28 ma peak current unit is required, since $I_p = 28\ P_o$. One consideration in the choice of diode is its capacity and inductance for microwave applications.

Another consideration, is of a purely practical nature. At low frequencies, the size of the inductor becomes electrically and physically large. To minimize physical dimensions, smaller wire is used resulting in larger values of DC resistance. The tunnel diode must remain stably biased in the negative conductance region. Therefore, R_T must be smaller than $1/g_d$, and if the DC resistance of the coil were very large, g_d would have to be very small. This means the use of a very low current tunnel diode since $g_d = \Delta_i/\Delta_v$ and Δ_v is fixed.

For our design example, let us use a 1 Kc/s sinusoidal oscillator having a limited power output of about one microwatt. A rough approximation tells us that a unit with I_p over 28 μa will do the job. As tunnel diodes with $I_p < 1$ ma are not generally available, a backward diode (generally this device can be used as a low current tunnel diode) is utilized. Such a unit is the ZJ69 having a negative conductance of approximately 3×10^{-4} mhos. R_T can now be 1 - 2 KΩ without seriously impairing the DC stability of the circuit.

Once g_d is determined, the choice of R_T can be made. The limits are that $R_T < 1/g_d$, but if R_T is made small, the DC power consumption of the circuit becomes too large. Hence R_T is chosen somewhere between .3 and .7 of $1/g_d$, as this will yield adequate stability with reasonable power consumption.

If we choose $R_T = .3/g_d$, R_T will be about 1000Ω for our example. At this point, C_1 can be calculated, where: (Per Eq. 4.4)

$$C_1 = \sqrt{\frac{g_d\ (1 - R_T \cdot g_d)}{R_T \omega^2}} = \sqrt{\frac{3 \times 10^{-4}\ (1 - .3)}{10^3 \times 4 \times 10^7}} = .071 \text{ mfd}$$

Since:

$$C + \frac{C_1}{1 - R_T \cdot g_d} = \frac{1}{L \cdot \omega^2}$$

if L is chosen somewhat arbitrarily at 100 mh with a $R_{DC} = 100\ \Omega$;

$$C + \frac{C_1}{1 - R_T \cdot g_d} = \frac{1}{.1 \times 4 \times 10^7} = 0.25 \text{ mfd}$$

hence:

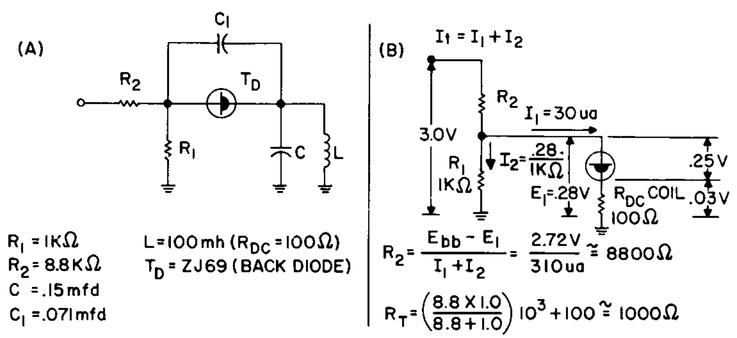
$$C = 0.25 - \frac{.071}{1 - .3} \text{ mfd} = 0.15 \text{ mfd}$$

The DC bias circuit can now be calculated. Assume the circuit supply voltage $E_{bb} = 3.0$ volts. If R_1 (see circuit in Figure 4.3) is 1000Ω, then:

$$R_2 = \frac{E_{bb} - E_{diode} + E_{RDC\ Coil}}{I_{total}} = \frac{3 - (.25 + .03)}{310 \times 10^{-6}} \cong 8800\Omega$$

and I_{total} is: $I_{diode} + E_{diode}/R_1 = 310\ \mu a$

$$R_T = \frac{R_1 \times R_2}{R_1 + R_2} + R_{DC\ Coil} + R_s = \frac{8800 \times 1000}{9800} + 100 + 2 \cong 1000\ \Omega$$



1 Kc/s OSCILLATOR CIRCUIT

FIGURE 4.3

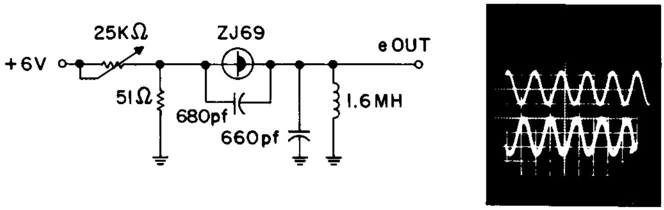
Another example at 100 Kc/s, would simply divide L, C and C_1 by 100 thus:

$$C_1 = 710 \text{ pf}$$

$$C = 1500 \text{ pf}$$

$$L = 1 \text{ mh}$$

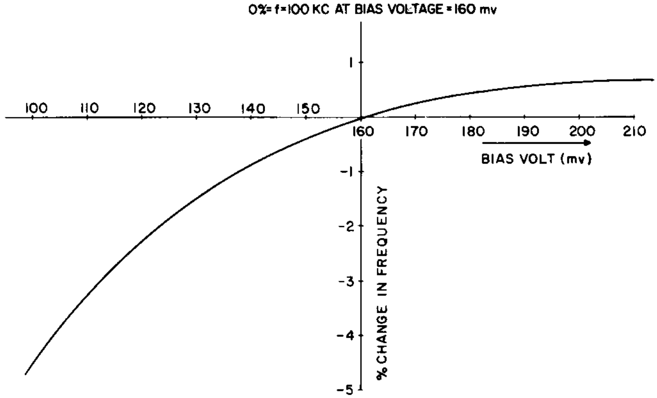
If the same tunnel diode is used, the bias circuit remains the same as in Figure 4.3. At 100 Kc/s, the size of the inductance will easily permit the use of larger peak current tunnel diodes however, and hence the circuit can be readjusted to accommodate a 1 ma I_p conventionally available unit with greater power output. The performance characteristics of a 100 Kc/s oscillator is illustrated in Figures 4.4 through 4.6. Figure 4.4 shows the circuit diagram and the oscilloscope presentation of the output waveform across the tunnel diode (upper) and across the tank circuit (lower).



100 Kc/s SINEWAVE OSCILLATOR CIRCUIT AND OUTPUT WAVEFORMS

FIGURE 4.4

Figure 4.5 shows the frequency variation vs. bias voltage and Figure 4.6 shows the frequency variation vs. temperature characteristic of the diode. In these tests only the tunnel diode was heated in order to establish its contribution to circuit instability.



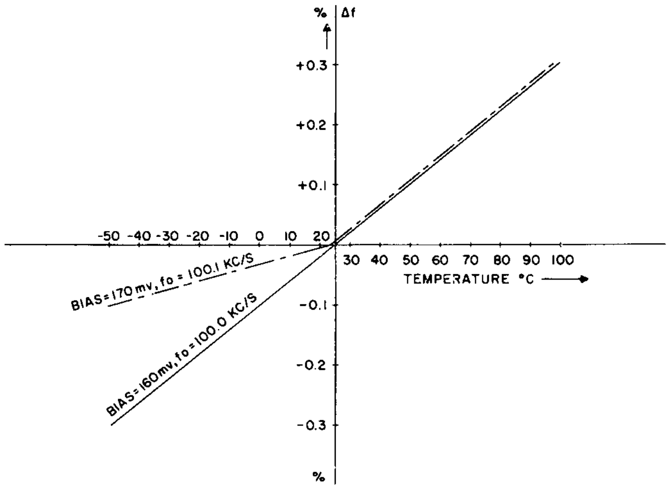
FREQUENCY VARIATION VS. BIAS VOLTAGE AT 25°C

FIGURE 4.5

As can be seen from the data, the frequency stability of this circuit is quite good over a wide range of temperature and bias voltage.

1 Mc/s Oscillator

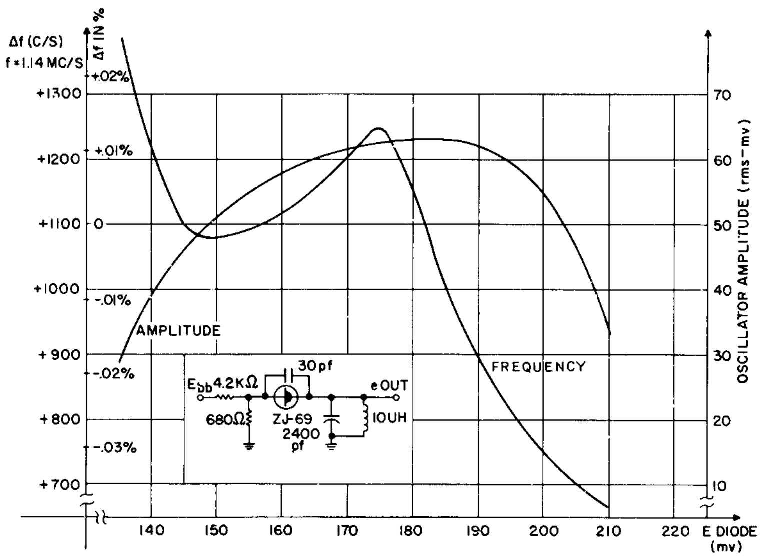
The next circuit, a 1 Mc/s oscillator, was tested and showed considerable improvement, as illustrated in Figures 4.7 through 4.9.



FREQUENCY VARIATION VS. TEMPERATURE
OF 100 Kc/s OSCILLATOR

FIGURE 4.6

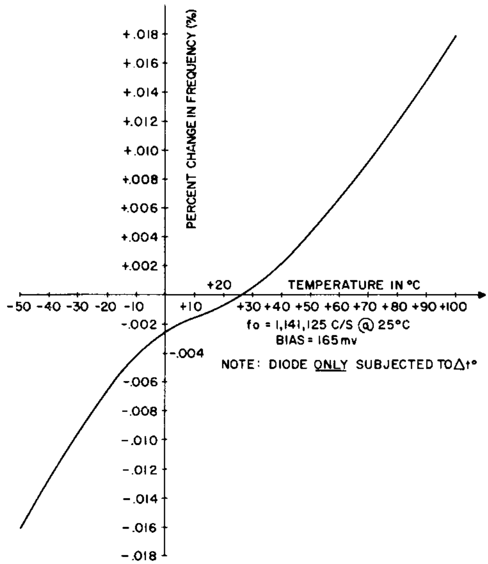
Figure 4.7 illustrates the fact that by changing the bias voltage, the variation of frequency is only in the order of a few hundred cycles out of 1.14 Mc/s, a change in the order of 0.05%. Over the range of 155 to 200 mv, the amplitude of the output voltage varies only from 63 mw-rms down to 55 mv-rms.



FREQUENCY STABILITY VS. BIAS VOLTAGE
OF 1 Mc/s OSCILLATOR

FIGURE 4.7

Figure 4.8 shows the temperature stability of this oscillator circuit over a range of -50°C to +100°C. From this figure it can be seen that the frequency varies only +18 - 16 millipercent over this wide temperature range.



FREQUENCY STABILITY VS. TEMPERATURE OF 1 Mc/s OSCILLATOR

FIGURE 4.8

Frequency Stability

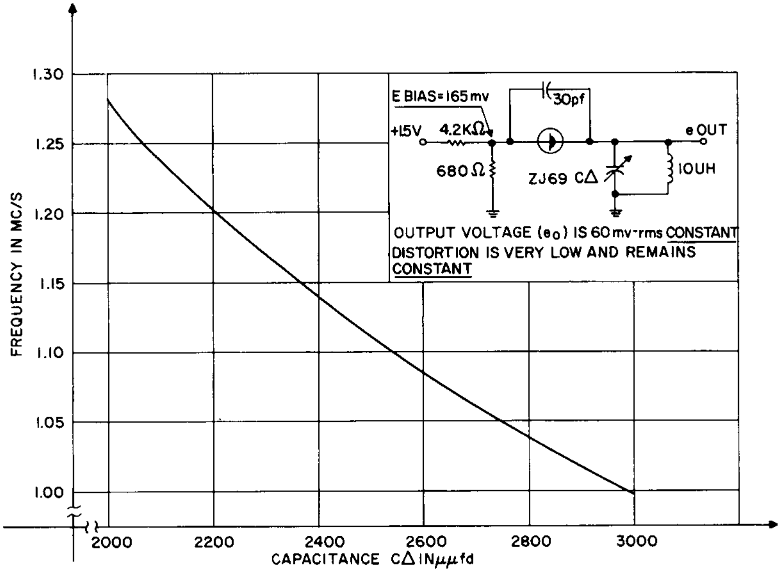
The main variable creating the frequency shift versus bias and temperature is the negative conductance ($-g_d$). As the operating frequency is increased, the variation of frequency (Δf) caused by Δg_d remains the same, but since f changes, $\Delta f/f \times 100$ decreases, thus causing a lower percentage of frequency shift.

Finally, Figure 4.9 shows that varying the capacitor in the parallel tank circuit over an appreciable range accomplishes smooth tuning without change in amplitude or distortion. This latter point is especially important since in the "series" type oscillator circuit a change in l or C only, affects stability and distortion greatly.

Frequency-Power Product

At higher power output requirements (in the milliwatt region) further limitations cause the aforementioned design and the design procedure to become impractical. After having chosen a tunnel diode with a sufficient peak current to furnish the required power output, g_d is essentially established. For example, to get 3 mw of power output (rms), it will take a 100 ma unit to furnish this amount of power. A

100 ma unit has a negative conductance g_d in the order of 1 mho ($-rd = 1\Omega$). If one keeps the product of $R_T \cdot g_d$ at 0.5 then C_1 is established by Equation 4.4.



FREQUENCY TUNING BY MEANS OF C_Δ

FIGURE 4.9

The problem arises in the heretofore somewhat arbitrary choice of L. This inductance becomes much more restrictive as g_d and the operating frequency increase. The inductance is determined by:

$$L = \frac{1}{\omega^2 \left| C + \dfrac{C_1}{1 - R_T \cdot g_d} \right|} \qquad (4.6)$$

The maximum value of L is realized when C becomes zero, at which time:

$$L_{max.} = \frac{1}{\omega^2 \left(\dfrac{C_1}{1 - R_T \cdot g_d} \right)} \qquad (4.7)$$

If one now determines this value by using a higher power example at 1 Mc/s:

$$L_{max.} = \frac{1}{39.6 \times 10^{12} \times .32 \times 10^{-6}} = 78\ m\mu h$$

If a small capacitance is now added and/or if the operating frequency is further increased, L will quickly reach the package limitations.

For example, if f = 10 Mc/s and C is 10 pf, then:

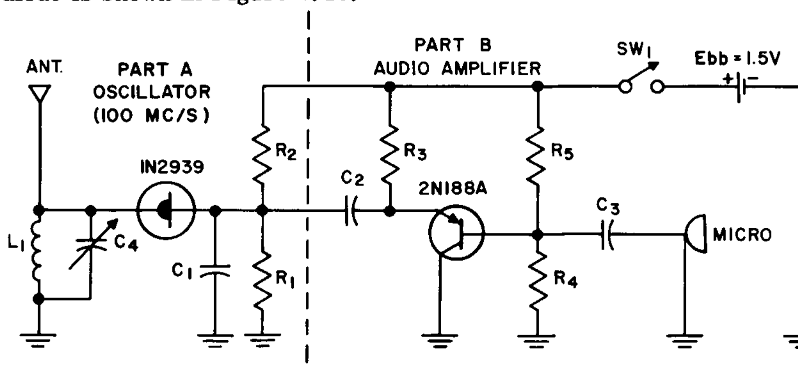
$$C_1 = \sqrt{\frac{1\,(1 - .5)}{.5 \times 39.6 \times 10^{14}}} = .016 \text{ mfd}$$

$$L = \frac{1}{39.6 \times 10^{14} \times 10 \times 10^{-12} + \dfrac{(16 \times 10^{-9})}{.5}} = 7.88 \text{ m}\mu\text{h}$$

For the TO-18 packaged devices, this is the magnitude of the case and lead inductance, and a further reduction of the inductance would be difficult to achieve. Naturally lower inductance tunnel diode configurations as well as some techniques of tuning out some of the inductance (in coaxial or triplate arrangement for microwave applications) might extend these limits somewhat.

FM Transmitter

A simple micropower FM transmitter using the 1N2939 tunnel diode is shown in Figure 4.10.



R_1	22Ω - 1/2 WATT	C_1	.001 CERAMIC DISC	L_1	6T NO. 16 WIRE - 3/8" DIAM. OPEN AIR (L≈.2μh)
R_2	270Ω - 1/2 WATT	C_2	50 μfd ELECTROLYTIC	ANT.	ANTENNA 4 3/4" LENGTH NO. 14 WIRE
R_3	470Ω - 1/2 WATT	C_3	5 μfd ELECTROLYTIC	MICRO	MICROPHONE "SHURE BROTHERS" MODEL NO. 420 OR EQUIV
R_4	10KΩ - 1/2 WATT	C_4	1.5-5.0 μμfd AIR VARIABLE		
R_5	10KΩ - 1/2 WATT				

88-108 Mc/s WIRELESS F.M. MICROPHONE

FIGURE 4.10

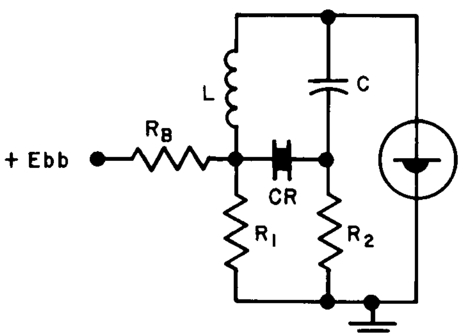
Operation may be best explained by separating the circuit into two portions. Part A is a basic tunnel diode oscillator whose frequency is primarily determined by the resonant circuit in the cathode. Resistors R1 and R2 provide a stable low impedance voltage for the anode of approximately 150 mv. Capacitor C_1 is the RF bypass for the anode.

Part B is a transistor emitter follower stage to amplify the audio signal from the microphone. The amplified audio is fed through capacitor C2 to the anode of the tunnel diode. FM modulation is accomplished by the audio signal instantaneously changing the anode bias. Since the characteristc curve is not perfectly linear in the negative resistance region, the negative conductance changes slightly with bias. As can be seen from the self-resonant frequency equation, f_{xo} is a function of $-g_d$ and therefore the resonance of the circuit is affected. FM deviations of $\pm$ 75 KC are readily obtainable with this type of circuit.

The transmitter shown in the diagram has been successfully used as a wireless portable microphone. Its great advantage is that it allows complete mobility on the part of the speaker, and of course has no wires or cords. When used with an average FM receiver having a sensitivity of 10 μv, an operating range in excess of 100 feet was obtained.

Crystal Controlled Oscillator

The circuit in Figure 4.11 works basically as per the previous description with the exception of the criteria for R_T. R_1 and R_2 are identical and are chosen to be about twice the value required for R_T. As a result, oscillations are not possible "off resonance". At resonance however, the crystal becomes a short circuit and R_1 is in parallel with R_2, essentially halving R_T. This new value of R_T permits the circuit to oscillate freely at a frequency accurately governed by the crystal.



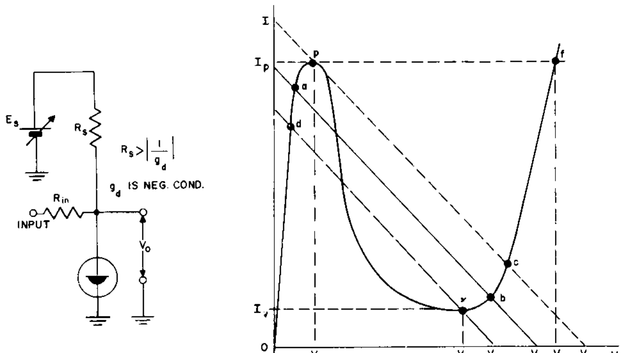
CRYSTAL CONTROLLED OSCILLATOR
FIGURE 4.11

CHAPTER 5

SWITCHES

One of the most promising modes of operation for the tunnel diode is as a switch where it can be used to function as an astable, monostable, and bistable oscillator, or it can be used in performing logic and memory functions. In addition, it can be used with other semiconductor devices to perform a variety of functions.

As discussed in Chapter 3, the stability criteria of the series circuit dictates that the negative resistance portion of the tunnel diode characteristic is an unstable region if the total circuit positive resistance is larger than the negative resistance of the tunnel diode. Consequently, the tunnel diode can only switch through this region to either of its stable states. Figure 5.1a shows a simple tunnel diode bistable switching circuit. This requires that $R_s > |1/g_d|$ where R_s is the load resistance, and $|g_d|$ is the magnitude of the negative conductance. Figure 5.1b shows the conductance load line of R_s superimposed on the tunnel diode characteristic. For a supply voltage V_1,



SIMPLE TUNNEL DIODE BISTABLE CIRCUIT

FIGURE 5.1

the load line intersects the tunnel diode characteristic at points "a" and "b". If the stable point is initially at point "a" and the supply voltage is increased to V_2, the circuit current exceeds the peak current of the tunnel diode, and it switches to point "c". If the supply voltage is reduced to V_3, the circuit current becomes less than the valley current, and the tunnel diode will switch back to point "d". On the other hand, if the supply voltage is held constant at V_1, the tunnel diode can be switched from point "a" to point "b" by a positive current pulse at the input terminals which momentarily increases the tunnel diode current to a value in excess of the peak current. The diode can be switched from point "b" to point "a" by applying a negative current pulse of such magnitude at the input terminals as to momentarily reduce the tunnel diode current to a value less than the valley current.

The low voltage or "off" state of the tunnel diode is the region from zero to the peak voltage, and the current through the diode consists entirely of majority carriers transported across the junction by the tunneling mechanism. The high voltage or "on" state is considered to be the region from the valley voltage to the peak forward voltage, and the current through the diode consists entirely of minority carriers transported across the junction by diffusion.

The speed of switching between the two states is very high and is determined chiefly by the junction capacitance and the amount of charge available from the trigger pulse. If a trigger of minimum amplitude is used in conjunction with a constant current bias source (the load line in Figure 5.1b is horizontal), the rise time of the tunnel diode voltage between the 10 and 90 percent points is approximately given by:

$$t_r = \frac{\left(V_{fp} - V_p\right) C}{(I_p - I_v)} \tag{5.1}$$

where C is the tunnel diode capacitance, V_{fp} and V_p are the forward and peak voltages, and I_p and I_v are the peak and valley currents respectively. Equation 5.1 can also be written:

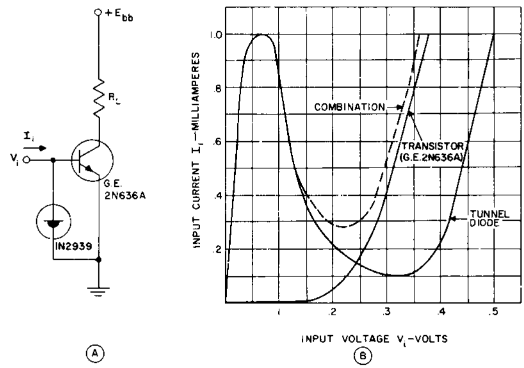
$$t_r = \left(\frac{V_{fp} - V_p}{I_p \frac{\eta - 1}{\eta}}\right) C \tag{5.2}$$

where η is the peak-to-valley current ratio. Since V_{fp}, V_p and $C/I_p - I_v$ are more or less independent of I_p, the rise time is also independent of the peak current, and the rise time can be decreased by reducing $C/I_p - I_v$. This is equivalent to reducing C/g_d. Using the parameters of the 1N2939 listed in Chapter 2, the rise time of this device is calculated to be 3.5 mμsec. which is in close agreement with measured values. Switching speeds for 10 ma germanium tunnel diodes have been measured to be less than 1 mμsec., and the correlation with the calculated values has been excellent.

5.1 Hybrid Circuits

Many simple and practical switching circuits are possible when the tunnel diode is used in conjunction with a transistor or controlled rectifier.

A transistor-tunnel diode hybrid circuit can be formed by paralleling the base-emitter junction of the transistor with a tunnel diode as seen in Figure 5.2a. When the tunnel diode is switched to the high voltage state, the transistor is turned on since the V-I characteristic of the tunnel diode is similar to that of the base-emitter junction of the transistor. A comparison of the tunnel diode and transistor characteristics as well as the parallel combination is given in Figure 5.2b. The net input characteristic can then be analyzed by means of load lines for bistable or astable operation as desired.

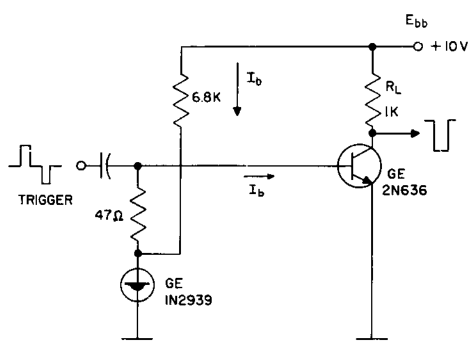


HYBRID TRIGGER CIRCUIT AND CHARACTERISTIC

(Using Germanium Alloy Transistor and Germanium Tunnel Diode)

FIGURE 5.2

A simple bistable circuit or flip-flop is shown in Figure 5.3. In this case, the tunnel diode is biased in the low voltage state by the current I_b which is slightly less than the peak current. Since the transistor is in the "off" condition, the collector is at the supply voltage, E_{bb}. If a positive trigger pulse is supplied at the input such that the tunnel diode current increases above the peak current, the tunnel diode switches to the high voltage state. The tunnel diode will remain in the high voltage state with a major portion of the bias current being diverted into the base of the transistor (how much is diverted can be obtained from the input characteristic shown in Figure 5.2b). The transistor collector voltage will then fall to ground potential if $I_b'h_{FE} > E_{bb}/R_L$.

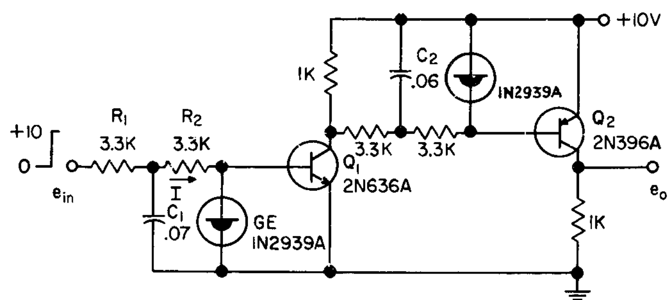


BISTABLE CIRCUIT USING TUNNEL DIODE
AND NPN GERMANIUM ALLOY TRANSISTOR

FIGURE 5.3

Likewise, a negative pulse at the input, which causes the tunnel diode current to fall below the valley current, will switch the tunnel diode back to the low voltage state, thus turning the transistor off. The 47Ω resistor which is in series with the tunnel diode in Figure 5.3 serves to insure that the tunnel diode will be biased above the valley point when in the high voltage state, and it also prevents the loading of the trigger pulses by the tunnel diode.

A simple hybrid time delay circuit which permits any number of consecutive time delays is shown in Figure 5.4. In this case, the bias supply for the tunnel diode has been eliminated, and the input is supplied through a charging circuit. In Figure 5.4, the timing cycle starts with the application of a positive 10 volt step at the input. C_1 charges through R_1, with the current I being proportional to the voltage across C_1. The time constant of the charging circuit is approximately $(R_1R_2/R_1 + R_2)\ C_1$. When the current I reaches the peak current value of the tunnel diode, the tunnel diode switches to its high voltage state and turns Q_1 on. The time delay for a step change at the



TUNNEL DIODE TIME DELAY CIRCUIT WITH TWO CASCADED COMPLEMENTARY STAGES

FIGURE 5.4

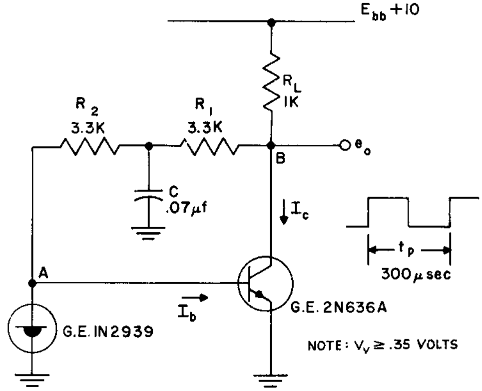
input of e_{in} is given by:

$$t_d \approx R'C_1 \ln \left[1 - \frac{(R_1 + R_2)\, I_p}{e_{in}}\right] \qquad (5.3)$$

where $R' = R_1R_2/R_1 + R_2$. Because the collector voltage of Q_1 falls from +10 volts to a very low value, a similar timing sequence is initiated for the second stage since the second stage is a complementary version of the first stage. At the end of the second timing sequence Q_2 will turn on, and its collector voltage will rise to 10 volts. For the component values shown in Figure 5.4, the time delay of each stage is approximately 120 microseconds. A multiphase oscillator results if an odd number of stages are connected in a closed loop.

Using the time delay scheme described above, a square wave oscillator can be built by returning the input to the collector of the transistor as shown in Figure 5.5. A cycle of operation begins with C charging through R_1 and R_L. The transistor is turned off because the tunnel diode is in the low voltage state. When the current through

R_2 reaches the peak current value of the tunnel diode, the tunnel diode switches to the high voltage stage. The base current I_b then drives the transistor to saturation, shorting the collector and point B to ground. The capacitor then discharges through R_1 and R_2 until the voltage at point A falls below the valley voltage of the tunnel diode, and the tunnel diode then returns to the low voltage state and turns the transistor off. The cycle then repeats itself. The symmetry of the output can be changed if a small current is fed in or taken out of point A. In order that the circuit operate properly, several require-



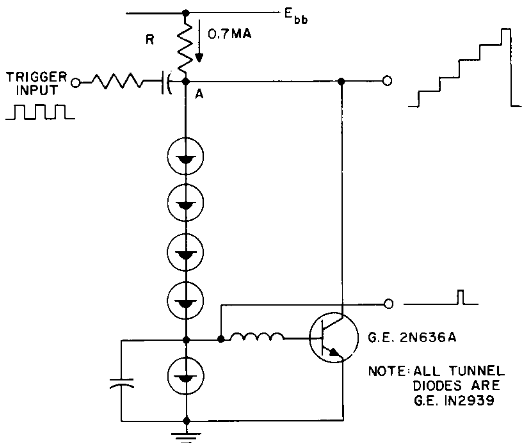
ASTABLE HYBRID OSCILLATOR
FIGURE 5.5

ments must be met:

1. When the transistor is in the off state (the tunnel diode is in the low voltage state), the current through R_2 must exceed the peak current of the tunnel diode before the capacitor finishes charging to its steady-state value.
2. At the valley voltage of the tunnel diode, the product $h_{FE}I_b$ must be larger than I_c or E_{bb}/R_L. This is the same as saying the transistor must remain saturated when the base emitter voltage is equal to the valley voltage of the tunnel diodes.
3. The switching times of the transistor must be less than the time required for one cycle.

With the values of components shown in Figure 5.5, the period is about 300 microseconds.

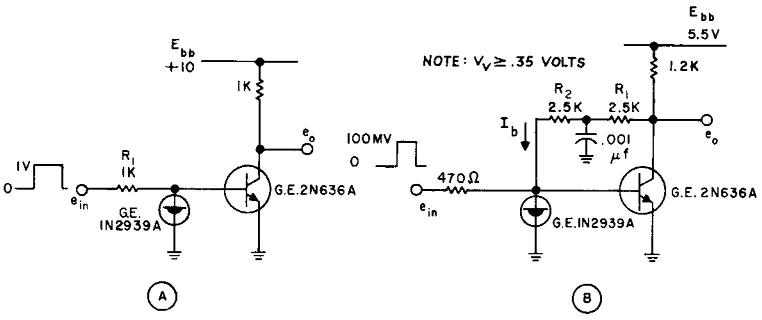
A simple pulse frequency divider of N:1 can be made using the technique shown in Figure 5.6. In this case, five tunnel diodes are connected in series to form a 5:1 frequency divider. They are biased from a current source whose magnitude is less than the peak currents of the tunnel diodes; thus, they are all in the low voltage state. The bottom diode is selected to have a higher peak current than the other tunnel diodes in the circuit. Each time a positive pulse occurs at the input, one of the upper four tunnel diodes switches to the high voltage state. When the fifth pulse occurs, the bottom tunnel diodes switches to the high voltage state and turns on the NPN transistor, which resets the circuit by essentially grounding point A. When point A is grounded, the current through the tunnel diodes is reduced to a value less than the valley current, and the tunnel diodes switch back to the low voltage state. The capacitor across the bottom tunnel diode and the inductor in series with the transistor base delay the signal to the transistor so that complete switching can occur. The waveform appearing across the string of tunnel diodes is a staircase with the risetime of each step being determined by the rise time of the trigger pulse. The operating frequency of the circuit is limited chiefly by the switching speed of the reset transistor.



SERIES CONNECTED TUNNEL DIODES USED FOR 5:1 PULSE FREQUENCY DIVIDER OR STAIRCASE WAVE GENERATOR

FIGURE 5.6

The tunnel diode also makes an excellent level detector. Figure 5.7a shows a simple tunnel diode-transistor level detector which has a very stable switching point, with the stability of the switching point being determined primarily by the stability of the peak current. Since the drift of the peak current with temperature for germanium tunnel diodes is a function of peak voltage, it can be seen from Figure 2.3 that tunnel diodes with peak voltages of 60 mv have the least drift with temperature. If these tunnel diodes are used in Figure 5.7, the drift of the level of switching can be less than $\pm$ 50 mv from -50 to 100°C. If the level of the voltage to be detected is less than a volt, R_1 in Figure 5.7a may be decreased; however, the minimum value it can have is 200 ohms. The circuit in Figure 5.7b allows the detection of voltages down to about 70 mv without the use of a negative supply. I_b is adjusted to be 90 to 95% of the peak current, and the time constant $R_1R_2C/R_1 + R_2$ must be made larger than the width of the pulse for the circuit to operate properly. With the values of components shown, an output is obtained when the input pulse reaches 100 mv, and the circuit will function properly if the pulse widths are between 0.5 and 5 microseconds. A 5.5 volt supply voltage is used since reference diodes

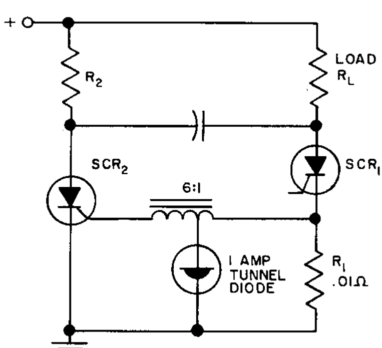


HYBRID LEVEL DETECTOR CIRCUITS

FIGURE 5.7

with this voltage have nearly a zero temperature coefficient. For larger pulse widths, C must be increased. If the circuit is to operate at high temperatures (71°C), a drift in the level of switching will result with the circuit as shown because of the transistor leakage current. For this case, the reference diode should be tied from the collector to ground and the supply voltage raised so that the reference diode is operating well in the breakdown region.

Figure 5.8 shows a tunnel diode-silicon controlled rectifier current limiting circuit for power equipment. In this case, the tunnel diode has a peak current of one ampere. When the load current exceeds the limit value, the voltage across the .01 ohm sensing resistor exceeds the peak voltage of the tunnel diode, and it switches to the high voltage state. The change of voltage across the tunnel diode is stepped up by the auto transformer to a value which is sufficient to fire the controlled rectifier, SCR2. When SCR2 fires, a negative voltage is applied by C_1 across SCR1 which causes SCR1 to turn off in 20 μsec. or less, thus interrupting the load current. The advantage of using the tunnel diode in this application is its ability to be triggered at a very low voltage level with the resultant very low power loss in the current sensing resistor.

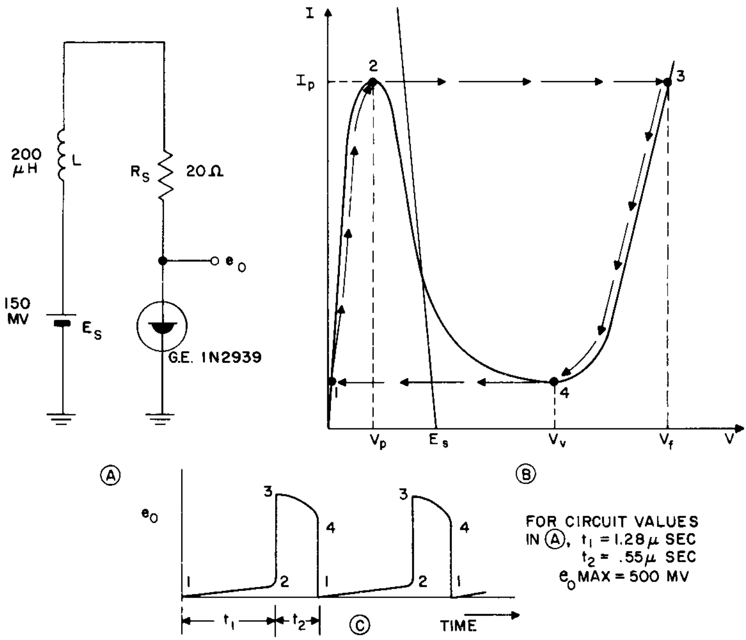


TUNNEL DIODE USED AS CURRENT SENSING ELEMENT IN SILICON CONTROLLED RECTIFIER CIRCUIT BREAKER

FIGURE 5.8

5.2 Astable Oscillators

Figure 5.9a shows a simple tunnel diode relaxation oscillator circuit. For proper operation of the circuit, the DC load line must intersect the tunnel diode characteristic in the negative resistance region as shown in Figure 5.9b. Consequently, the magnitude of E_s must be less than V_v and greater than V_p.



BASIC TUNNEL DIODE RELAXATION OSCILLATOR CIRCUIT

FIGURE 5.9

The circuit operation begins when power is applied. The circuit current increases exponentially, and when it reaches the peak current value of the tunnel diode (point 2 in Figure 5.9b), the tunnel diode switches to the high voltage state (point 3). Since the voltage drop across the tunnel diode is larger than the supply voltage, the circuit current begins to decrease exponentially. However, when the circuit current reaches the valley current of the tunnel diode (point 4), the tunnel diode switches to the low voltage state (point 1). One cycle of operation is completed, and another cycle just like the one just described begins.

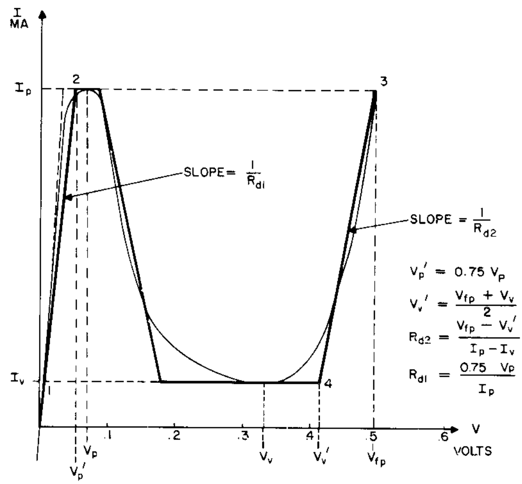
Because the current build up from point 1 to point 2 requires a longer time than when it is decreasing from point 3 to point 4, the output voltage across the tunnel diode will not be symmetrical. This is shown in Figure 5.9c. The trailing edge of the output is quite rounded--this is due to the non-linear V-I characteristic of the tunnel diode in the high voltage region.

In order to avoid a non-linear differential equation in solving for the period of the relaxation oscillator, the V-I characteristic must be linearized since current-voltage relationships of the diode in both the low and high voltage regions are non-linear. This linearization is shown in Figure 5.10. If the linear portion of the diode characteristic in the low voltage region is extended, it will intersect the I_p constant current line at approximately $V_p/2$. Therefore, an "average" value of voltage at this current is $0.75\ V_p$, and the linearized dynamic resistance is then:

$$R_{d1} = \frac{0.75\ V_p}{I_p} \tag{5.4}$$

LINEARIZATION OF THE TUNNEL DIODE CHARACTERISTIC

FIGURE 5.10



In the high voltage state, between V_v and V_{fp}, there is no linear region. Therefore, an "average" value of voltage at the valley current is:

$$V_v' = \frac{V_{fp} + V_v}{2} \tag{5.5}$$

and the linearized dynamic resistance in this region is:

$$R_{d2} = \frac{V_{fp} - V_v'}{I_p - I_v} \tag{5.6}$$

The time required by the circuit current to go from point 1 to point 2 then is given by:

$$t_1 = L/R_{T1}\ \ln\left(\frac{E_s - R_{T1}\,I_v}{E_s - R_{T1}\,I_p}\right) \tag{5.7}$$

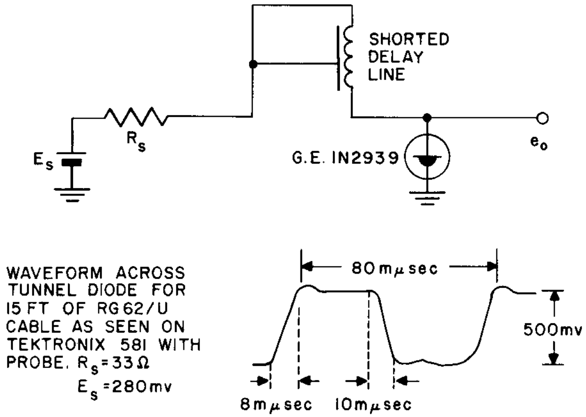
where $R_{T1} = R_s + R_{d1}$, R_s is the external series resistance, and E_s is the supply voltage.

The time required by the circuit current to decrease from point 3 to point 4 is given by:

$$t_2 = L/R_{T2} \ln \left(\frac{V_{fp} + I_p R_s - E_s}{V_v' + I_v R_s - E_s} \right) \tag{5.8}$$

where $R_{T2} = R_s + R_{d2}$. The total period is then $t_1 + t_2$. Correlation between the calculated values using equations (5.7) and (5.8) and actual measurement was within 10% on 0.5, 1 and 2.2 ma germanium tunnel diodes with wide ranges of supply voltage and series resistance.

The period of the simple relaxation oscillator is very sensitive to supply voltage, with the sensitivity being primarily due to t_1. If a shorted delay line is used in place of the inductance, a square wave results, whose frequency is insensitive to supply voltage changes.[17] Such a circuit is shown in Figure 5.11.



RELAXATION OSCILLATOR USING
A SHORTED DELAY LINE

FIGURE 5.11

5.3 Monostable Oscillator

If the supply voltage is decreased or increased such that the DC load line intersects the tunnel diode characteristic in a positive resistance region as shown in Figure 5.12b, a monostable oscillator results. A supply voltage of E_{s1} will bias the tunnel diode in the low voltage state. A positive input current pulse of magnitude I_p-I_a will switch the tunnel diode to the high voltage state (point 3 in Figure 5.12b). The current in the inductance begins to decrease exponentially from I_a until it reaches the valley current at which time the tunnel diode switches back to the low voltage state. The switching time from point 2 to point 3 (rise time) and from point 4 to point 1 (fall time) are less than ten nanoseconds for one milliampere germanium tunnel diodes. The time required for the current to decrease from point 3 to point 4 (the output pulse duration) is given by:

$$t_a = \frac{L}{R_T} \ln \left[\frac{V_{fa} + I_a R_s - E_{s1}}{V_v' + I_v R_s - E_{s1}} \right] \tag{5.9}$$

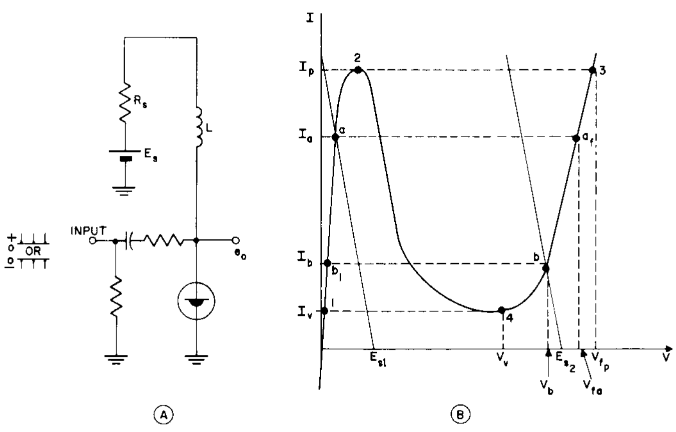
where $R_T = R_{d2} + R_s$ and $V_v' = (V_{fp} + V_v)/2$. R_{d2} is derived as discussed in the above section on astable oscillators. The value of I_a can be determined by measurement, graphical analysis, or it can be derived as:

$$I_a \approx \frac{E_{s1}}{R_s + R_{d1}} \tag{5.10}$$

where R_{d1} is derived as discussed in the above section on astable oscillators. V_{fa} is given by:

$$V_{fa} = V_{fp} - R_{d2}\,(I_p - I_a) \tag{5.11}$$

Correlation between measured and calculated values using equation 5.9 were within 10%. The maximum repetition rate of the trigger pulses is limited by the length of time required for the circuit current to increase from point 1 to point "a" in Figure 5.12b. This time is given by equation 5.7 with the exception that I_p is replaced by I_a.



TUNNEL DIODE MONOSTABLE OSCILLATOR

FIGURE 5.12

If the tunnel diode is biased in the high voltage state (point b in Figure 5.12b), then negative trigger pulses of a magnitude $I_b - I_a$ are required at the input. The output pulse will then be one going from a high voltage to a low voltage, and its width is given by:

$$t_b = \frac{L}{R_T} \ln \left[\frac{R_T I_b - E_{s2}}{R_T I_p - E_{s2}} \right] \tag{5.12}$$

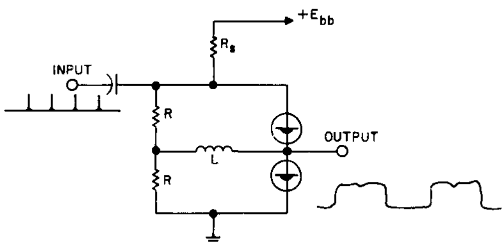
where $R_T = R_s + R_{d1}$. The value of I_b is given by:

$$I_b \approx \frac{E_{s2} - (V_v' - I_v R_{d2})}{R_s + R_{d2}} \tag{5.13}$$

Again, correlation between measured and calculated pulse widths is excellent. The maximum repetition rate is limited by the time required for the circuit current to decrease from point 3 to point b in Figure 5.12b. This time is given by equation 5.8 with the exception that I_v is replaced by I_b and V_v' is replaced by V_b.

5.4 Tunnel Diode Flip-Flop[3,14]

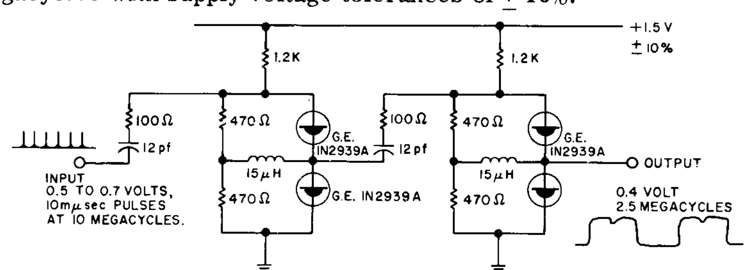
Figure 5.13 shows a tunnel diode flip-flop circuit which requires only trigger pulses of one polarity. The supply voltage is restricted to a magnitude such that only one tunnel diode can be in the high voltage state. The difference between the two tunnel diode currents flows through the inductance. When a positive trigger pulse turns the diode which is in the low voltage state to the high voltage state, the voltage induced in the inductance (because of the decreasing current through it) is of a polarity as to reset the other tunnel diode to the low voltage state. Each pair of trigger pulses completes one switching cycle.



TUNNEL DIODE FLIP-FLOP

FIGURE 5.13

The basic flip-flop circuit can be inter-connected to form a counter as shown in Figure 15.14. With the values of components and input pulses shown, the counter will operate successfully up to 10 megacycles with supply voltage tolerances of $\pm$ 10%.



TUNNEL DIODE TWO STAGE COUNTER

FIGURE 5.14

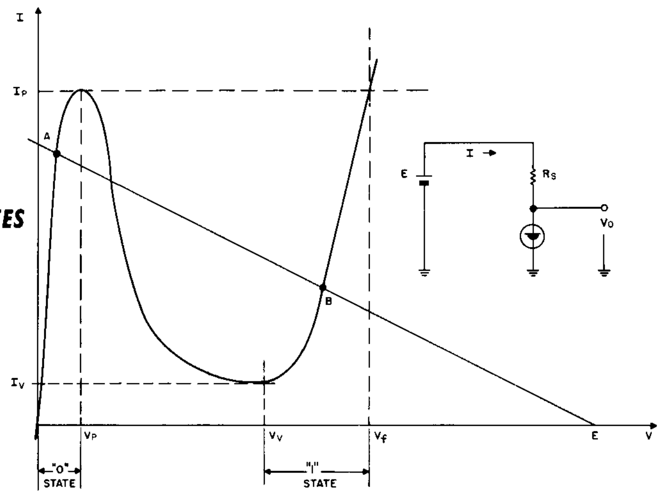
CHAPTER 6

LOGIC CIRCUITS

The fast switching speed of tunnel diodes coupled with their potential low cost and small size make attractive their use in computer logic.

In Chapter 5 it was shown that the tunnel diode can be switched between its low and high voltage states by controlling the current through it. This resultant change in voltage can be used in logic circuits to distinguish between a "1" and a "0". As an example (see Figure 6.1), the low voltage state can be called a "0", and it can have a value ranging from zero to the peak voltage V_p, depending upon the magnitude of the current. The high voltage state can be called a "1", and it can have a value ranging between the forward voltage V_f and the valley voltage, V_v. The time required to switch between the low and high voltage states can be less than a nanosecond.

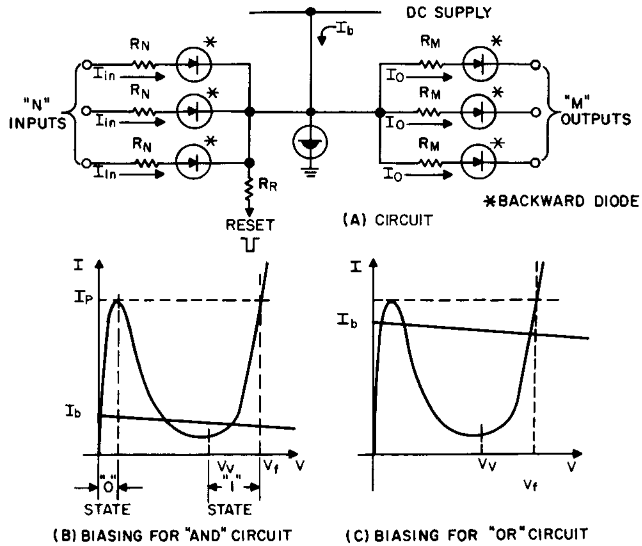
LOGIC VOLTAGE STATES OF THE TUNNEL DIODE
FIGURE 6.1



6.1 Simple Analog-Threshold Logic

A very simple circuit for performing both "and" and "or" logic is shown in Figure 6.2. This has been called analog-threshold logic rather than threshold logic because the inputs are connected together and summed before reaching the threshold device - the tunnel diode[14]. In normal diode threshold logic circuits each input circuit has its own thresholding device-the diode. With analog-threshold logic the allowable tolerance in the amplitude of each signal is a function of the number of inputs: the larger the number of inputs, the smaller the allowable tolerances.

The same circuit configuration (Figure 6.2a) can be used to perform both "and" and "or" functions. In either case, the tunnel diode is normally biased in the low voltage state with the magnitude of the bias current I_b determining whether one or all the inputs must be present in order to drive the tunnel diode to its high voltage state.



ANALOG-THRESHOLD LOGIC

FIGURE 6.2

Once in the high voltage state, the tunnel diode can be returned to the low voltage state by reducing the supply voltage to zero or by a negative reset pulse.

In the analysis which follows, it is assumed that the logic stage is driving and is being driven from similar tunnel diode logic stages. The input coupling resistors R_N may or may not be equal to the output coupling resistors R_M. The number of input circuits or the "fan-in" is N, while the number of output circuits or the "fan-out" is M. Unilateral flow of information is obtained by the use of backward diodes.

"And" Circuit Analysis

The tunnel diode in Figure 6.2 is in the low voltage or "0" state. For an "and" circuit, each of the tunnel diodes driving the "fan-in" resistors must be in the high voltage state for the tunnel diode under consideration to switch to the high voltage state. However, the value of voltage of the input tunnel diodes may vary from V_f to V_v, depending on the number of loads each is driving. In addition, tolerances of supply voltage, resistors, and tunnel diode parameters will affect the values of I_{in}, I_b, and threshold current I_p. Two requirements which must be met for the circuit to function properly are:

$$(N-1)\ I_{in\ max.} + I_{b\ max.} < I_{p\ min.} \tag{6.1}$$

$$(N)\ I_{in\ min.} + I_{b\ min.} > I_{p\ max.} \tag{6.2}$$

When the tunnel diode has switched to the "1" or high voltage state, the bias current must be large enough to supply the load current in addition to the tunnel diode current, and the tunnel diode current must be larger than the valley current, otherwise the tunnel diode will switch back to the low voltage state. Thus, the requirement to be met when the tunnel diode is in the high voltage state is:

$$I_{b\ min.} - M\, I_{o\ max.} > I_{v\ max.} \tag{6.3}$$

If the resistance tolerance is α, the supply tolerance Δ, and the tunnel diode parameter tolerance is β; then, in the limiting case, equations 6.1, 6.2, and 6.3 can be written respectively:

$$(N-1)\left[\frac{V_f\,(1+\beta) - V_p\,(1-\beta)}{R_N\,(1-\alpha)}\right] + I_b\,\frac{(1+\beta)}{(1-\alpha)} = I_p\,(1-\beta) \tag{6.4}$$

$$N\left[\frac{V_v\,(1-\beta) - V_p\,(1+\beta)}{R_N\,(1+\alpha)}\right] + I_b\,\frac{(1-\Delta)}{(1+\alpha)} = I_p\,(1+\beta) \tag{6.5}$$

$$M\left[\frac{V_v\,(1+\beta) - V_p\,(1-\beta)}{R_M\,(1-\alpha)}\right] - I_b\,\frac{(1-\Delta)}{(1+\alpha)} = -\,I_v\,(1+\beta) \tag{6.6}$$

By setting R_N equal to R_M, by solving equations 6.4, 6.5, and 6.6 simultaneously, and by eliminating R, I_b and I_p, the relationship between N and M can be obtained in terms of the tunnel diode voltages, the tolerances, and the peak-to-valley ratio. For example, the above equations are solved for various tolerances using a germanium tunnel diode with $I_p/I_v = 10$, $V_p = .055$ volts, $V_v = .35$ volts, and $V_f = .5$ volts:

(a) Case of zero tolerance ($\alpha = \beta = \Delta = 0$)

$.295N + .295M = 1.1\, I_p R$

N and M are not dependent upon one another and M can be made large by increasing R. The maximum "fan-in", however, is 3. (This is determined by eliminating I_b and I_p when 6.4 and 6.6 are solved simultaneously.)

(b) 2% Tolerance ($\alpha = \beta = \Delta = .02$)

$.193N + .062M = .45$

For M = 1, N can only be 2

(c) 3% Tolerance ($\alpha = \beta = \Delta = .03$)

$0.20N + 0.093M = .45$

For M = 1, N can only be 1

Thus the tolerance problem is very severe for a worse case design of the "and" circuit. In actuality, it is more severe than indicated since ideal backward diodes are assumed and capacitance and overdrive effects have not been considered.

"Or" Circuits

The circuit in Figure 6.2 can perform the "or" function if the bias current is increased sufficiently such that only one input is required to switch the tunnel diode into the high voltage state. If backward diodes are used in the coupling circuits, then the requirements to be met are:

$$I_{in\ min.} + I_{b\ min.} > I_{p\ max.} \tag{6.7}$$

and

$$I_{b\ max.} < I_{p\ min.} \tag{6.8}$$

Equations 6.7 and 6.8 can be written respectively in terms of the tolerances in the limiting case as:

$$\frac{V_v (1 - \beta) - V_p (1 + \beta)}{R_N (1 + \alpha)} + I_b \frac{(1 - \Delta)}{(1 + \alpha)} = I_p (1 + \beta) \tag{6.9}$$

and

$$I_b \frac{(1 + \Delta)}{(1 - \alpha)} = I_p (1 - \beta) \tag{6.10}$$

R can be calculated from 6.9 and 6.10 if the tolerances, peak current and voltages are known. If backward diodes are not used, then the current which flows in the N-1 inputs which are in the low voltage state must be taken into account in equation 6.9.

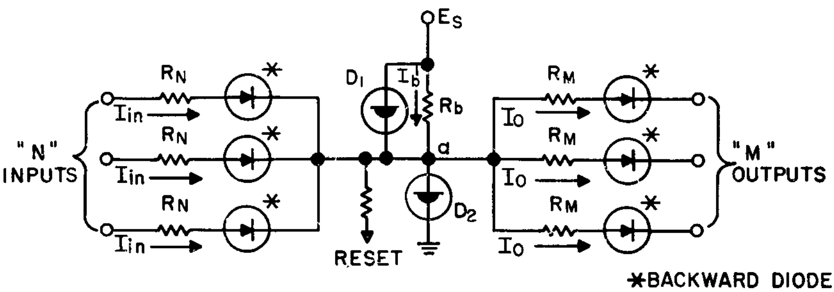
After the tunnel diode has switched to the high voltage state, the "fan-out" equation is the same as given in equation 6.6. If backward diodes are not used in the "fan-in" circuits, then the current drain of the input circuits must be added to equation 6.6.

6.2 Chow's Circuit

Figure 6.3 shows a circuit developed at the G-E Electronics Laboratory by W. F. Chow which reduces the dependence of N on M as compared to the analog-threshold "and" circuit. The difference between this and the previous circuit is the addition of the tunnel diode D_1 across R_b and the reduction of the supply voltage, E_s, to a value which is large enough to permit only one of the tunnel diodes to be in the high voltage state. Thus, $E_s \approx V_v + V_p$. The value of R_b is determined by the number of inputs. Before each logic operation D_2 is set into the low voltage state and D_1 is set in the high voltage state. The bias current in D_2 is then the current through R_b and the valley current of D_1. When the sum of the input and bias currents exceeds the peak current of D_2 and switches D_2 into the high voltage state, D_1 switches to the low voltage state, essentially returning point "a" to the supply voltage. (The dynamic resistance of D_1 in the low voltage stage is much less than R_b.) Consequently, the total output current, MI_o, is limited only by the difference between the peak current of D_1 and the valley current of D_2. It is assumed that backward diodes are used to insure the unilateral propagation of information. The general requirements which must be met to insure that D_2 will only switch to the high voltage state when all the inputs are present are given by equations (6.1) and (6.2). I_b, however, is now equal to $I_v + I_b'$. The requirement to be met after D_2 has switched to the high voltage state is:

$$I_{v2\ max.} + M\, I_{o\ max.} < I_{p1\ min.} \tag{6.11}$$

where $I_{v2\ max.}$ is the maximum valley current of D_2 and $I_{p1\ min.}$ is the minimum peak current of D_1. (The peak current of D_1 and D_2 need not be the same.) In the analysis that follows, it is assumed that the tunnel diode stage under consideration (Figure 6.3) is driving and being driven by similar types of logic circuits.



CHOW'S ANALOG-THRESHOLD LOGIC CIRCUIT

FIGURE 6.3

If the supply voltage tolerance is Δ, the resistor tolerances are α, the tunnel diode parameter tolerances are β, and the bias current tolerance is Ω, then the limit worse case design expressions of 6.1, 6.2 and 6.11 are respectively for the "and" operation:

$$(N-1)\left[\frac{E_s(1+\Delta) - V_p(1-\beta)}{R_N(1-\alpha)}\right] + I_b(1+\Omega) = I_{p2}(1-\beta) \tag{6.12}$$

$$N\left[\frac{E_s(1-\Delta) - 2V_p(1+\beta)}{R_N(1+\alpha)}\right] + I_b(1-\Omega) = I_{p2}(1+\beta) \tag{6.13}$$

$$M\left[\frac{E_s(1+\Delta) - 2V_p(1-\beta)}{R_M(1-\alpha)}\right] + I_{v2}(1+\beta) = I_{p1}(1-\beta) \tag{6.14}$$

where:

$$I_b = I_{v1} + \frac{E_s - V_p}{R_b}$$

and

$$E_s = V_v + V_p$$

If the tolerances are set equal to zero, equations 6.12 and 6.13 can be solved to give the bias current in terms of N and the other parameters. Equations 6.12, 6.13 and 6.14 can be solved to give the relationship between M and N in terms of the tolerances, peak voltage, supply voltage, peak-to-valley ratio, and the ratio of the peak currents of the two tunnel diodes. By making the peak current of D_1 larger than that of D_2, a larger "fan-out" results for a given "fan-in".

As an example, if D_1 and D_2 are respectively 2.2 ma and 1ma germanium tunnel diodes with peak-to-valley ratios of 10, and E_s = .45 volts, V_p = .055 volts and the tolerances are $\alpha = \beta = \Delta = 3\%$ and $\Omega = 6\%$, then the relationship between M and N is:

$$0.64N + 0.29M = 3.5 \tag{6.15}$$

Thus, for M = 1, N = 5; and for M = 3, N = 4.

If the tolerances are set equal to zero, equations 6.12 and 6.13 can be solved for the maximum possible "fan-in":

$$N = \frac{E_s}{V_p} \quad \text{or} \quad N = \frac{V_v}{V_p} + 1 \tag{6.16}$$

For germanium tunnel diodes, the maximum possible "fan-in" (zero tolerance conditions) is 7. As the tolerances are increased, the "fan-in" is reduced.

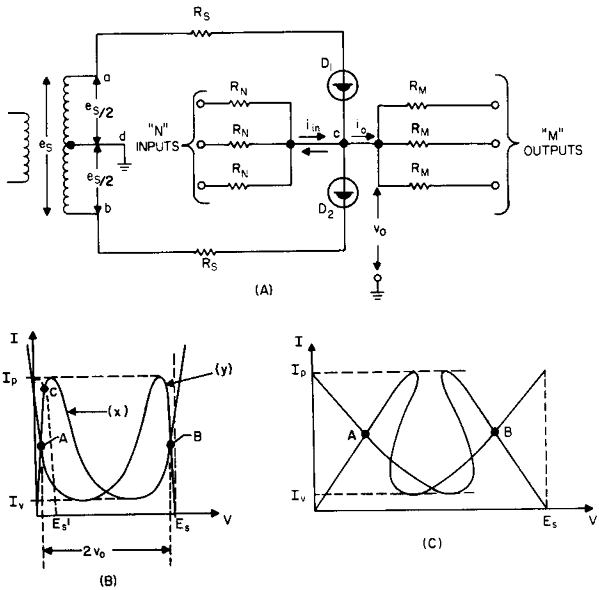
6.3 The Goto or "Twin" Circuit[15]

Binary "1" 's and "0" 's can be represented respectively by positive and negative unit voltages instead of positive unit voltages and small or zero voltages, and the logic operation can be based on an analog sum or majority of the voltages. Thus, the output of a gate used in performing majority logic will be dependent upon whether a majority of the inputs is either positive or negative. Consequently, the number of inputs for the majority gate must be an odd number if ambiguity in the output circuit is to be avoided.

Figure 6.4 shows a circuit which will give both a positive and negative output, and which can be used as a majority gate. The supply voltage, which can be in the form of a square wave or sinusoid, is obtained from the secondary of a center-tapped transformer. The tunnel diodes D_1 and D_2 are in series with the supply, and the input and output voltages are taken with respect to the junction of the tunnel diodes and the grounded center-tap of the transformer.

The tunnel diodes are selected to have closely matched peak currents. During the negative half-cycle (when the point "b" is positive with respect to point "a"), both tunnel diodes are conducting in the reverse direction. Resistors or non-linear devices such as backward diodes must be added in series with the tunnel diode in order to limit the current flow during this half-cycle. Point "c" will be at ground potential since the transformer, tunnel diodes, and resistors form the arms of a balanced bridge. As the supply voltage goes positive (point "a" positive with respect to point "b"), both diodes begin conducting in the forward direction. However, point "c" will remain approximately at ground potential until the current in one or the other tunnel diode exceeds its peak current. Which tunnel diode reaches its peak current first is dependent upon whether the majority of input signals is positive or negative with respect to ground. If the analog sum of the inputs is positive, the signal current will flow into point "c", and its direction is such that it adds to the current in D_2 and subtracts from the current in D_1. Therefore, D_2 will reach its peak current first and will switch to the high voltage state. If the supply voltage is limited such that only one tunnel diode can be in the high voltage state, the other tunnel diode is forced to remain in the low voltage state for the remainder of the half-cycle. The output voltage at "c" then is positive with respect to ground. If the majority of input signals had been negative (current flowing out of point "c"), D_1 would have reached its peak current first and consequently the output voltage at "c" would have been negative for the remainder of the half-cycle. Thus, the Goto circuit produces an output of the same polarity as the majority of the input signals.

Composite V-I characteristics of the Goto circuit are shown in Figure 6.4b for the case where R_S is neglected and in 6.4c where it is taken into account. Thus in Figure 6.4b, tunnel diode (y) is considered as the "load" and its curve is the image of the V-I curve of the other tunnel diode (x) reflected through the current axis and translated along the positive voltage axis by a distance E_S. The stable operating points are the intersections "a" and "b" in the positive resistance regions of the characteristics. As the supply voltage E_S increases from zero, the (y) characteristic or "load" is shifted horizontally along the voltage axis. Thus at a supply voltage E_S', the stable operating point is "c".



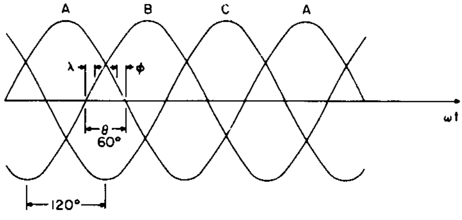
"GOTO" OR "TWIN" CIRCUIT
FIGURE 6.4

The composite characteristic of 6.4c is the same as 6.4b except that each individual characteristic curve is now a composite V-I characteristic of the tunnel diode and resistance R_S.

The Goto gate can also be used to perform "and" and "or" functions. If a positive input is considered a "1" and a negative input a "0", then the circuit in Figure 6.4a will be a two input "and" gate if one of the inputs is returned to a negative bias supply. The reason is that both of the other two inputs must be positive in order to obtain a positive output. Likewise, if one input is returned to a positive supply the circuit becomes a two input "or" gate since only one of the other inputs is required to be positive in order to obtain a positive output.

Backward diodes cannot be used with the Goto circuit to obtain a unilateral flow of information since the inputs and outputs can be of either polarity. One of the most simple ways of achieving this is to use a three phase clock. The over-lapping of the phases must be such that at any instant in time one group of circuits is capable of providing an output, a second group is sensitive to its inputs, and the third group is neither sensitive to the inputs nor is it capable of providing an output. The three states of operation can be called respectively: active, receptive, and passive.[5]

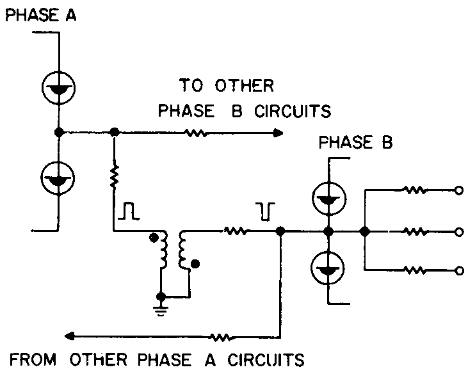
If the three phases of the supply voltage are displaced from each other in time by 120°, then information can be transferred from the circuits supplied by one phase to the circuits supplied by the next phase during the 60° of overlap (see Figure 6.5). Thus the circuits supplied by phase A drive the circuits supplied by phase B; those supplied by phase B drive those circuits supplied by phase C; which in turn drive circuits supplied by phase A. The flow of information is unilateral because during the positive overlap of two phases of the supply voltage, the voltage of the third phase is negative. The tunnel diodes supplied by this phase are inactive during this period since they are biased in the reverse state.



THREE PHASE SUPPLY FOR MAJORITY LOGIC

FIGURE 6.5

Inversion or the logical "not" function can be accomplished by the use of a transformer as shown in Figure 6.6. One of the difficulties encountered in using this method is that the magnitude of the secondary voltage is dependent upon the history of the signals applied to the primary. Thus, if the signal applied to the primary is 101010 (alternate positive and negative pulses), the voltage at the secondary will be the inverse of those pulses. (If the turns ratio is 1:1, the amplitude of the primary and secondary voltages will be the same.) On the other hand if the input is all one's or zero's (all positive or all negative pulses), the secondary voltage will be inverted, but the amplitude will be reduced because the transformer will not pass the DC component of the primary voltage. This variation of output amplitude must be taken into account when using a transformer to perform inversion.



INVERSION OR LOGICAL NOT
FIGURE 6.6

If the Goto or twin gate is to function properly as a logic element, the following conditions must be met:

1. During the receptive period and just prior to the switching of one of the tunnel diodes, the majority input current must be large enough to overcome any expected supply voltage unbalance and tunnel diode peak current mismatch.
2. During the active period when one of the tunnel diodes is in the high voltage state, the "fan-out" current and the maximum amplitude of the supply voltage must be limited so that the current through the tunnel diode in the low voltage state is less than its peak current.
3. The overlap of the phases of the supply voltage must be large enough so that the stage in the active mode can transfer its information to the succeeding stages before the information is lost. The information is lost when the supply voltage falls to a value such that the tunnel diode current is less than the valley current. Overlap is an important consideration when the supply voltage is sinusoidal.

The equivalent circuit in Figure 6.7 is used to establish the circuit requirement for meeting the first of the above conditions. For the polarity of input voltage shown (a positive majority) diode D_2 should switch to the high voltage state. The minimum available input voltage results when there is only a majority of one. For this case v_{in} becomes:

$$v_{in} = \frac{1}{N}\left[e_{np}\frac{(N+1)}{2} - e_{nn}\frac{(N-1)}{2}\right] \qquad (6.17)$$

where N is the "fan-in" or number of inputs, e_{np} the positive input voltage, and e_{nn} the negative input voltage. If $e_{np} = e_{nn} = v_o$ (Figure 6.4b), then v_{in} becomes v_o/N. The source impedance R_{in} is $R_{N/N}$

where R_N is the coupling resistor of each input.

Since D_2 is to switch to the high voltage state, I_2 must reach the peak current of D_2 before I_1 reaches the peak current of D_1. Under limiting conditions I_1 and I_2 become equal to I_{p1} and I_{p2} respectively at the same instant. Therefore, the circuit requirement under these conditions becomes:

$$I_2 - I_1 = I_{p2} - I_{p1} \tag{6.18}$$

If the per unit tolerance in peak current is β, then equation 6.18 becomes:

$$I_2 - I_1 = 2\beta I_p \tag{6.19}$$

where I_p is the nominal value of peak current.

The expression relating $I_2 - I_1$ to the circuit voltages and resistances is given in Figure 6.7. This expression is obtained by solving the loop equations and by assuming the voltage drop across the tunnel diodes to be zero. Notice that the first term of the expression is a function of the supply unbalance, while the second term is a function of the input majority.

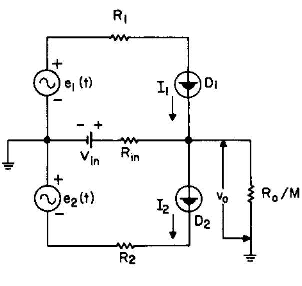
The relationship between the "fan-in" and "fan-out" in terms of the circuit tolerances can be obtained by substituting 6.19 for $I_2 - I_1$ in the equation of Figure 6.7; by letting $e_1 = e\,(1 + \Delta)$, $e_2 = e\,(1 - \Delta)$ and $v_{in} = v_{in}\,(1 - \Omega)$; and by assigning resistance tolerances so as to make the right hand side of the equation as small as possible. If $R_N = R_M$, and $R_N/M + N >> R_1R_2/R_1 + R_2$, then the relationship becomes:

$$2\,I_p\,\beta \approx -\frac{e\,(\alpha + \Delta)(M + N)}{R_N} + \frac{N\,v_{in}\,(1 - \Omega)}{R_N} \tag{6.20}$$

If the Goto circuit is to operate over a wide temperature range, the value of v_{in} at the highest operating temperature should be used in 6.20.

EQUIVALENT CIRCUIT OF THE GOTO STAGE DURING THE RECEPTIVE PERIOD

FIGURE 6.7



$$I_2 - I_1 = \frac{R_1 e_2 - R_2 e_1}{R_2 R_1 + \dfrac{R_{in} R_o (R_1 + R_2)}{R_o + M R_{in}}} + \frac{R_o v_{in}}{R'(R_o + M R_{in}) + R_{in} R_o}$$

$$R' = \frac{R_1 R_2}{R_1 + R_2}$$

$$R_{in} = \frac{R_N}{N}$$

$$v_{in} = \frac{V_o}{N} \text{ FOR A MAJORITY OF 1}$$

SEE FIG. 6.4 (B) FOR V_o

After diode D_2 has switched, the peak value of the supply voltage must be limited so that the current through D_1 does not exceed its peak current. The equivalent circuit of Figure 6. 7 can be used to determine the circuit requirements to meet this condition if v_{in} is set equal to zero. If I_L is the total current which flows through fan-in resistors R_N and the fan-out resistors R_M, and if the current in D_2 is I_2', then

$$I_{p1\ min.} > I_2'{}_{max.} + I_{L\ max.} \tag{6.21}$$

If $R_N = R_M$, $R_1 = R_2 = R$, then under limiting conditions 6.21 can be written:

$$I_{p1}(1-\beta) - I_2'(1+\beta')\left[1 - \frac{R}{R + \frac{R_M(1-\alpha)}{M+N}}\right] = \frac{e_m(1+\Delta)}{R + \frac{R_M(1-\alpha)}{M+N}} \tag{6.22}$$

where β' is the per unit tolerance of I_2' and e_m is the peak value of the supply voltage if a sinusoidal supply is used. From 6.22 the maximum permissible value of supply voltage is given in terms of M, N, the circuit parameters and tolerances. If the peak value of the supply voltage is to be made as large as possible, I_2' should be a minimum.

When a sinusoidal supply voltage is used, sufficient overlap must exist so that the information is passed to succeeding stages before the logic stage under consideration looses its information. The information is lost because at some angle $\emptyset$ the supply voltage has decreased to a value where it is unable to sustain a current through D_2 in excess of the valley current. The angle $\emptyset$ (Figure 6.5) where the logic stage is reset is given by:

$$\text{Sin } \emptyset = \frac{V_v}{e_m\left[2 - \frac{R}{R + \frac{R_M}{M+N}}\right]} + \frac{R\,I_v}{e_m} \tag{6.23}$$

where e_m is the peak value of the supply voltage with respect to ground and V_v and I_v are the valley voltage and currents respectively of the tunnel diode in the high voltage state. Equation 6.23 is written assuming that $R_N = R_M$ and $R_1 = R_2 = R$. For worst case conditions, the tolerances of the circuit parameters are included in 6.23 so as to increase $\emptyset$. The value of e_m can be obtained from 6.22.

Before the logic is reset (as given by $\emptyset$), the succeeding stages must receive the information and switch to the proper state. The angle λ (Figure 6.5) at which they can switch (assuming a positive input) occurs when the supply voltage has increased sufficiently so as to make I_2 equal to I_{p2}. Thus for $R_1 = R_2 = R$, $R_N = R_M$ and $e_1 = e_2 = e_m \text{ Sin } \lambda$:

$$\text{Sin } \lambda = \frac{R\,I_{p2}}{e_m} - \frac{M\,R\,v_{in}}{2\,R_M + (M+N)R} \tag{6.24}$$

For worst case design conditions, the circuit tolerances are included in 6.24 so as to increase λ. Thus, for proper operation of the Goto circuit, $\lambda + \emptyset$ must be less than the overlap angle, θ, as shown in Figure 6.5.

At high frequencies (above a megacycle) or when a square wave is used, the capacitance mismatch of the tunnel diodes becomes important, and it is reflected in the performance of the stage as a mismatch in peak currents.

CHAPTER 7

TUNNEL DIODE TEST CIRCUITS

The measurement of tunnel diode parameters requires considerable care to insure that the test circuit is in a stable state while performing the test. Oscillation or switching is usually avoided by appropriate "loading" or "damping" of the circuit while in some special cases it may even be encouraged in order to measure a specific parameter more accurately.

The parameters to be tested depend on the application. For example, if the device is used as an oscillator or amplifier, the negative conductance $|g_d|$, the capacitance C, the inductance L_s and the series resistance R_s must be known (see the tunnel diode equivalent circuit in Figure 2.5).

As a switching circuit element, it might be more desirable to know I_p, I_v, V_p, V_v, V_{fp} and switching speed (t_r). The following test circuits are designed to measure these parameters. Some of these circuits will yield readings with accuracies of only $\pm$ 10 to 20% and are quite simple in nature. Others are designed for high accuracy and naturally will be more complex.

7.1 V-I Curve Tracer

Observing the tunnel diode V-I characteristic is not always an easy task. Conventional curve tracers usually have enough series resistance in their sweep circuits as to appear as a current source to the tunnel diode under test, that is,

$$R_T > \frac{1}{|g_d|} \tag{7.1}$$

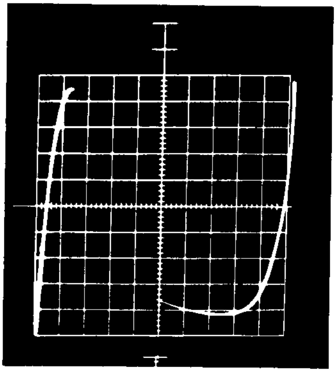
where R_T is the total DC resistance in series with the tunnel diode being tested. As a result, the displayed V-I characteristic lacks the negative conductance portion as shown in Figure 7.1.
This figure shows that as the sweep passes the peak point current, the diode switches very quickly from V_p to some value of forward voltage dependent on the load line, without ever going through the negative conductance region.

This problem becomes even worse when testing larger peak current units as they will exhibit smaller values of negative resistance. The sweep circuit will thus look more like a current source to such units enabling them to "switch" more readily.

Even when the sweep circuit resistance is small compared to $|-rd|$, still another problem might prevent the undistorted display of the V-I curve. Figure 7.2 illustrates this problem which is caused by ex-

V-I CHARACTERISTIC AS DISPLAYED BY CURVE TRACER WITH LARGE INTERNAL SERIES RESISTANCE

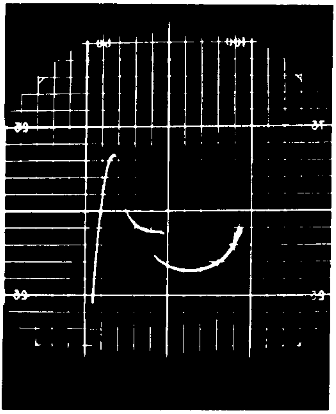
FIGURE 7.1



cessive series inductance in the test circuit. As the sweep goes through the negative conductance region, this inductance will resonate with the diode and circuit capacitances, making the circuit oscillate and causing a display as shown in Figure 7.2.

V-I CHARACTERISTIC OF OSCILLATING TUNNEL DIODE

FIGURE 7.2

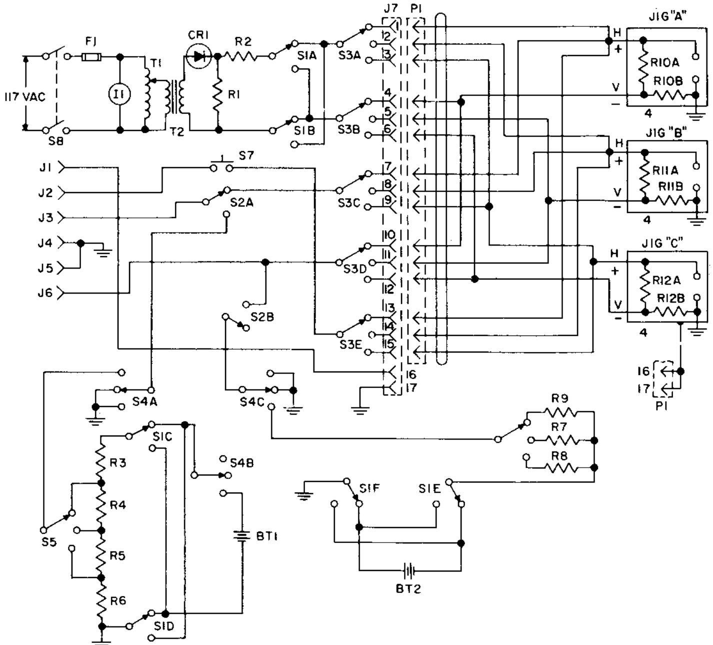


If the total circuit inductance (including the diode inductance L_s) is kept at a minimum, this oscillation will not occur. Actually, if the circuit self-resonant frequency $f_{xo\ (circuit)}$ is made larger than its cut-off frequency, $f_{ro\ (circuit)}$, no oscillations can occur. To meet this criterion, the total circuit inductance must be smaller than:

$$L_{total} < \frac{R_{Total} \times C_{total}}{g_d} \qquad (7.2)$$

Here again larger peak current units will be more troublesome to test, as one can see from equation 7.2 that if $|g_d|$ is increased, the maximum permissible circuit inductance is further reduced.

The main features of a well working V-I curve tracer capable of displaying the full V-I characteristic of tunnel diodes with a wide range of peak currents, are thus:



PART NO.	NAME	DESCRIPTION
BT1	BATTERY	1.34 VDC MALLORY #RM-12R *
BT2	BATTERY	1.34 VDC MALLORY #RM-12R *
CR1	DIODE	IN1695
F1	FUSE	1 AMP
I1	LAMP	NE-5:
J1	BINDING POST	SUPERIOR ELECTRIC TYPE DF30BC *
J2	BINDING POST	SUPERIOR ELECTRIC TYPE DF30RC *
J3	BINDING POST	SUPERIOR ELECTRIC TYPE DF30RC *
J4	BINDING POST	SUPERIOR ELECTRIC TYPE DF30BC *
J5	BINDING POST	SUPERIOR ELECTRIC TYPE DF30BC *
J6	BINDING POST	SUPERIOR ELECTRIC TYPE DF30RC *
J7	SOCKET	AMPHENOL #77-MIP-20, 20 PINS *
P1	PLUG	AMPHENOL #70-20, 20 PINS *
R1	RESISTOR	47Ω, 2W
R2	RESISTOR	20Ω, 10W
R3	RESISTOR	340Ω, 1/2 W, 1/4 %
R4	RESISTOR	500Ω, 1/2 W, 1/4 %
R5	RESISTOR	400Ω, 1/2W, 1 %
R6	RESISTOR	100Ω, 1/2W, 1%
R7	RESISTOR	264.6Ω, .5%
R8	RESISTOR	1353Ω, .5%
R9	RESISTOR	131.4Ω, .5%
S1	SWITCH	6 POLE, 2 POSITION
S2	SWITCH	2 POLE, 2 POSITION
S3	SWITCH	5 POLE, 3 POSITION
S4	SWITCH	3 POLE, 3 POSITION, MOMENTARY CONTACT
S5	SWITCH	1 POLE, 3 POSITION
S6	SWITCH	1 POLE, 3 POSITION
S7	SWITCH	N.O., MOMENTARY CONTACT, PUSH BUTTON
S8	SWITCH	DPST, TOGGLE
T1	TRANSFORMER	POWERSTAT TYPE 10B *
T2	TRANSFORMER	STANCOR #P-6134 *
XF1	FUSE HOLDER	FUSETRON TYPE HJM *
XI1	LAMP HOLDER	DIALCO #95408-931 *
1	ASSEMBLY	
2	CABINET	PREMIER #SFC-502 - 8"X14"X8" *
3	CHASSIS	PREMIER #ACH-426-5"X7"X2" ALUMINUM *
4	CURVE TRACING JIG	FOR DETAILED DRAWING SEE FIG. # 7.5
R10A & B	RESISTORS	10Ω PYROFILM RESISTOR CO # D375 CARBON FILM MICROWAVE RES.
R11A & B	RESISTORS	4Ω PYROFILM RESISTOR CO # D375 CARBON FILM MICROWAVE RES.
R12A & B	RESISTORS	2Ω PYROFILM RESISTOR CO # D375 CARBON FILM MICROWAVE RES.

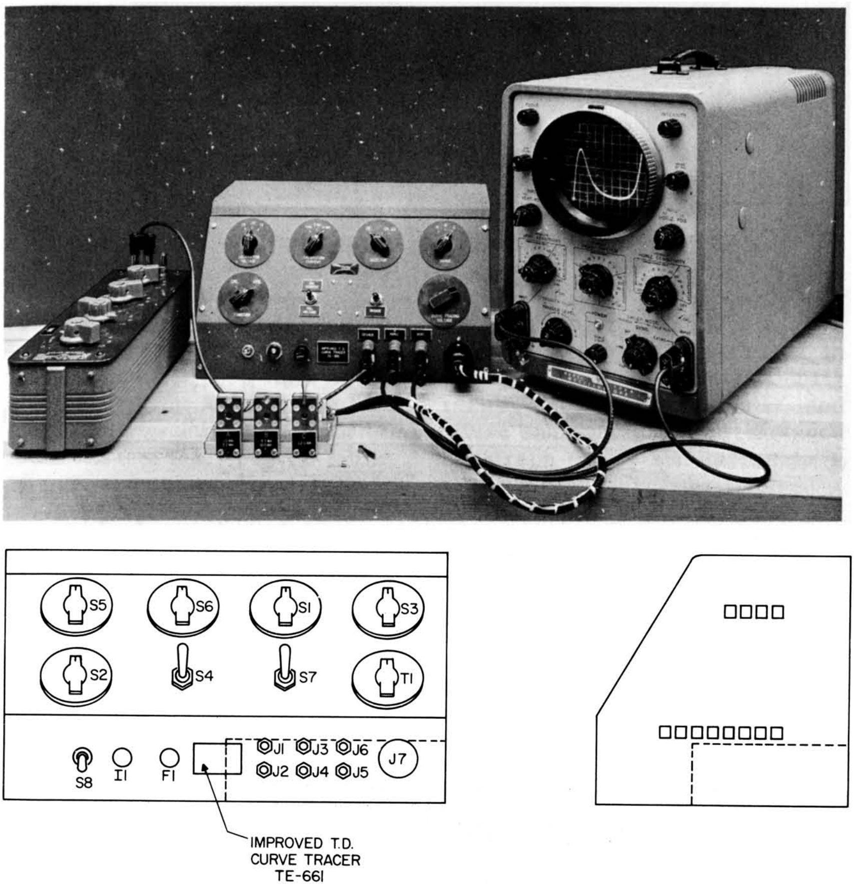
* OR EQUIVALENT

V-I CURVE TRACER CIRCUIT DIAGRAM

FIGURE 7.3

1. A low series resistance sweep circuit and,
2. Low inductance test heads.

The curve tracer circuit shown in Figure 7.3 and pictured in 7.4 covers a range of units from a fraction of one milliampere to 22 ma. With appropriate calibration (see Table of Calibration Procedures), this test set will determine, I_p, I_v, I_p/I_v, V_p, V_v, V_{fp}, V_r as well as to give a rough idea of the positive slopes and the negative

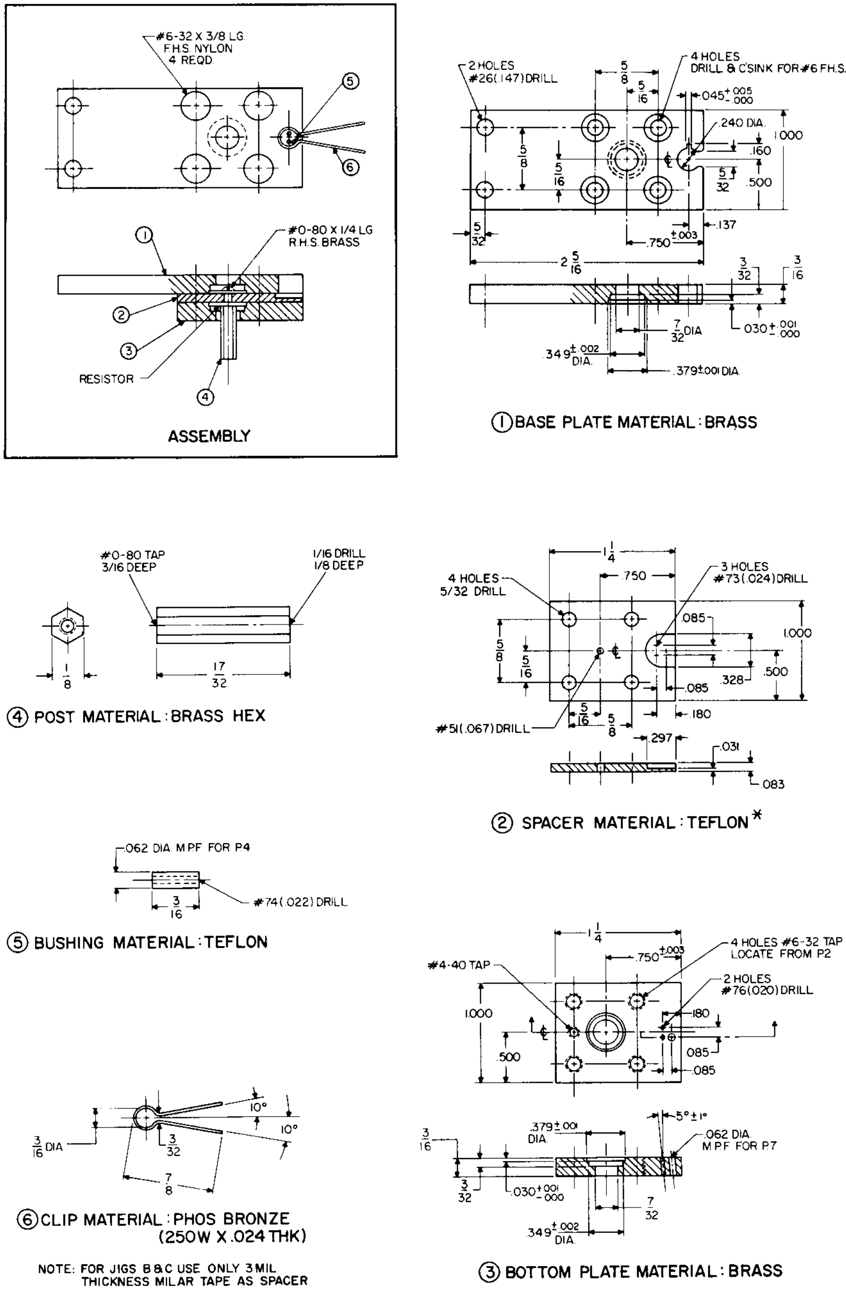


PHOTOGRAPH OF CURVE TRACER TEST SET UP

FIGURE 7.4

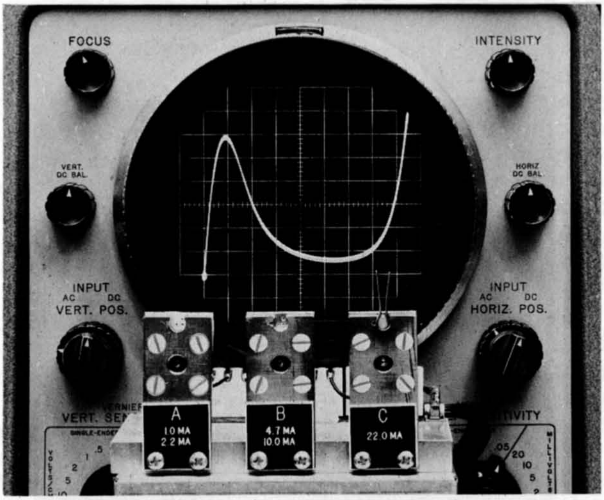
conductance slope. By the use of the shunting decade box, a rough idea can also be obtained of the average value of the negative resistance ($1/g_d = d_v/d_i$).

Figure 7.5 is a mechanical drawing detailing the construction technique of the jigs.



OUTLINE DRAWING OF LOW INDUCTANCE TEST JIGS

FIGURE 7.5



CLOSE-UP PHOTOGRAPH OF JIG AND SCOPE PRESENTATION

FIGURE 7.6

This instrument is made more versatile by the small size, low inductance test heads. Since the bias resistors are incorporated in the test heads, these may be connected with long cables to the main chassis and the decade box. As a result, one can conveniently and stably test tunnel diodes in a temperature chamber at some distance from the test set. Figure 7.6 is a photograph of the V-I curves of a 22 ma germanium tunnel diode.

Operating Instructions for Curve Tracer

The three outputs of the curve tracer are connected to the Vertical and Horizontal inputs of a scope (HP 130B or equivalent) and a Resistance Decade Box. Note: To display the V-I characteristic "right side up", the vertical scope input must be inverted and the ground strap removed from what is then the "hot" lead.

The Selection Switch must be in the germanium (Ge) or gallium arenside (GaAs) position, depending upon the type of unit under test.

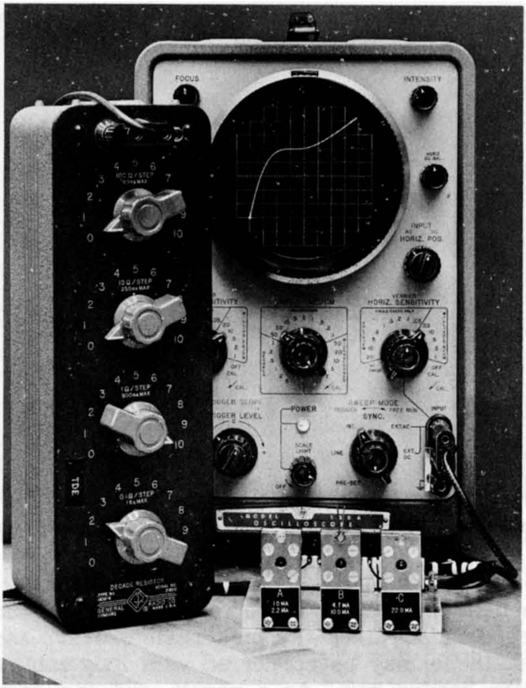
The Range Switch must be in the A, B or C position, depending upon the jig in use.

Jig A is for 1.0 and 2.2 ma units. Jig B is for 4.7 and 10.0 ma units. Jig C is for 22.0 ma units.

The Function Switch must be in operating position.

With the Curve Tracing Voltage Control in the counter-clockwise position, insert the diode. Increase the voltage until the trace is beyond the valley region.

Press the Decade Switch and adjust the decade box until the negative slope is parallel with the horizontal axis. This setting on the decade box is then the terminal negative resistance of the unit being tested. Figure 7.7 shows the resulting curve trace and the setting of the resistance box for a 10 ma unit.



DETERMINATION OF TERMINAL NEGATIVE RESISTANCE
FIGURE 7.7

Scope Calibration

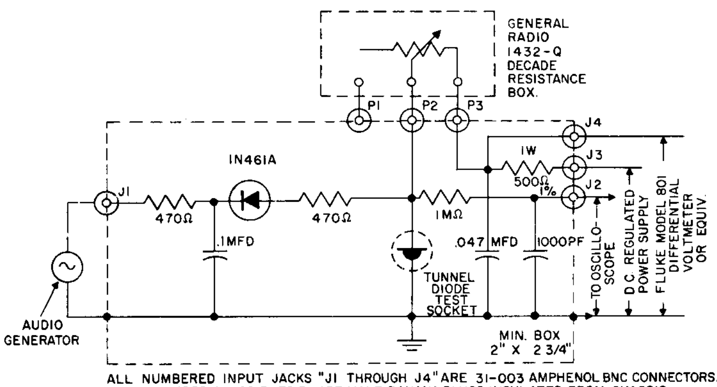
To calibrate the horizontal axis of the scope, the Function Switch must be in the calibration position. In this position, the scope output terminals are shorted and the calibration voltage switch in either the .1v, .5v or 1.0v position. The "Calibrate Voltage-Calibrate Current" switch is then pressed in the "Cal. Voltage" position. The calibrated voltage will then appear across the output terminals.

The same applies to the calibration of the vertical axis, with the exception that, if a 1 ma calibration is desired, the Range Switch must be in the A position and the diode in Jig A removed. If the 5 ma calibration is desired, the Range Switch must be in the B position and the diode in Jig B removed. For 10 ma calibration, the Range Switch must be in the C position and the diode in Jig C removed.

7.2 High Precision Peak Point Current (I_p) Test Set

Some applications require very accurate peak point current control. For instance, a threshold switch or a coincidence gate circuit require a stable switching threshold. To meet this demand, some tunnel diode types are specified with a peak point current rating of plus or minus 2.5%. To test accurately to these limits, the peak current test set must be capable of measuring the test parameters within a fraction of one percent. The following peak point current test set is capable of such accuracy.

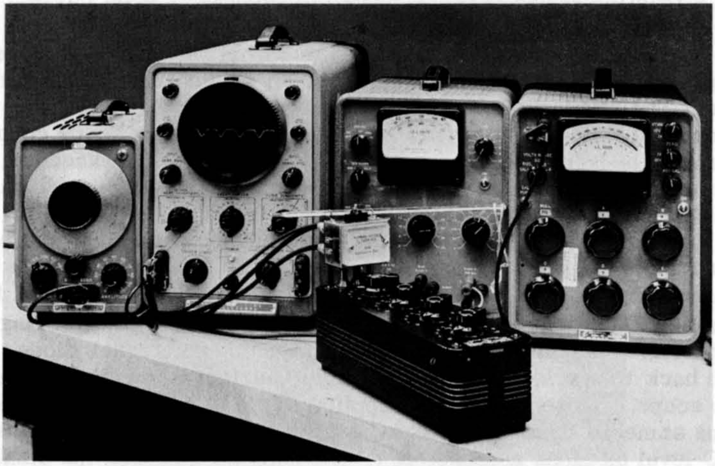
The circuit (see Figure 7.8) is rather simple and consists basically of a tunnel diode driven from an accurately known current source. The DC input is increased to a point just exceeding the peak point current. One can observe this point visually on a scope connected across the tunnel diode. As the peak current is exceeded, the diode switches to its high voltage state which can be observed on the scope as a change in DC level (using a DC scope). At this point, one would still have only a rough approximation of the peak point and hence a small audio sinewave signal is applied. The test set rectifies this signal and applies it to the tunnel diode in a direction subtracting from the applied DC. When the AC signal is large enough to bring the "composite" bias just below the valley point, the tunnel diode will switch back to its low voltage state again, a fact immediately evident on the scope. Subsequently, the applied DC is very gradually reduced and the same is done with the AC signal. The idea is to "fine tune" the DC level as close as possible to the peak point so that the smallest possible AC signal will switch the circuit to its low voltage state. When this is satisfactorily accomplished, the circuit will turn "ON" and "OFF" erratically, but continuously. At this point, a differential voltmeter will very accurately determine this crtical DC supply voltage. The use of a high precision resistance decade box will, by simple use of Ohm's Law, yield the current reading.



PEAK CURRENT TEST SET

FIGURE 7.8

The absolute accuracy of this current reading is about $\pm 0.1\%$, however, the degree of indecision as to the exact point at which the circuit turns "ON" and "OFF" consistently, adds about 0.2% potential judgement error to the reading. It is important, in order to achieve high accuracy, that the circuit be totally shielded, properly grounded and that the power supply be well regulated. Since I_p does change somewhat with temperature, a thermometer can be placed near the test socket for temperature readings. Figure 7.9 is a picture of the complete test set up.



PHOTOGRAPH OF PEAK CURRENT TEST SET UP
FIGURE 7.9

7.3 Tunnel Diode Junction Capacitance Test Set

In previous chapters the tunnel diode equivalent circuit has been analyzed and it can be shown that the apparent capacity looking into the device terminals is:

$$C_{apparent} \cong C_{junction} + C_{strays} - L_s g_d^2 \text{ (when } \omega << g_d/c)$$

In addition, if a bias resistor suitable for biasing the tunnel diode into the negative-conductance region is used, and the inevitable inductance of the leads coupling the measuring node to the biased device is considered, then the capacitance measured by the bridge may be shown to be:[10]

$$C_{measured} = C_j + C_{strays} - L_{leads} (g_{bias} + g_d)^2 - L_s g_d^2$$

Since the inductance of the leads to the bridge can easily be the dominant reactance in the circuit, the method described is seen to be subject to large potential errors even when stringent efforts are made to keep all inductances to the irreducible minimum. Unfortunately this method is necessary in order to measure the variation of junction capacitance with forward voltage in the negative-conductance region.

One further item will be considered before discussing the measurement method to be proposed.

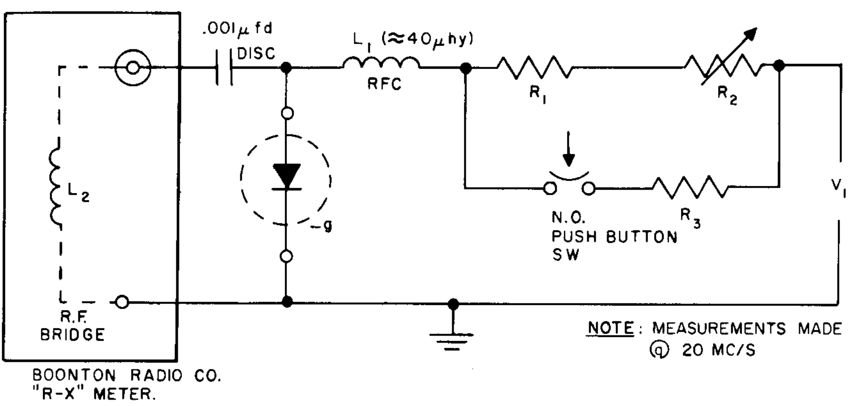
From the stability analysis of the equivalent circuit one finds that for the tunnel diode to be stable in the negative-conductance region,

$$\frac{L_s |-g_d|}{C} < R_s < \frac{1}{|-g_d|}$$

In other words, if the circuit inductance cannot be made less than $R_s C/|-g_d|$ the circuit must oscillate, and any capacitance measured bears only a fortuitous resemblance to the actual capacity. (Even in the valley region the measuring signal will drive the tunnel diode slightly into a region of some negative-conductance for part of the time, and the possibility of oscillations exists.)

Consider an increase of the bias resistance to the limit of $1/|-g_d|$ in order to improve the circuit stability. Then L_s must be less than C/g_d^2 (or conversely $C_j > L_s g_d^2$) or oscillations will persist and the circuit cannot be stabilized. Since the ultimate limits being considered vary as the square of the negative-conductance (a direct function of peak point current) it is apparent that a test circuit which may be stabilized for a low-current tunnel diode will eventually become unstable for some higher current device, due to the limitations of a fixed circuit and/or package inductance.

With the preceding considerations in mind, one sees that the design of a test circuit for measuring the capacitance of tunnel diodes over a large range of peak point currents and g_d/c ratios, can only be accomplished by deliberately violating both stability criteria simultaneously, and forcing stability by using a switching load line to bias the device. It should be also noted that the sum of junction and distributed capacitance can only be measured over a small voltage range around the valley point voltage. Figure 7.10 illustrates the resultant test circuit.



BASIC TUNNEL DIODE CAPACITANCE TEST CIRCUIT

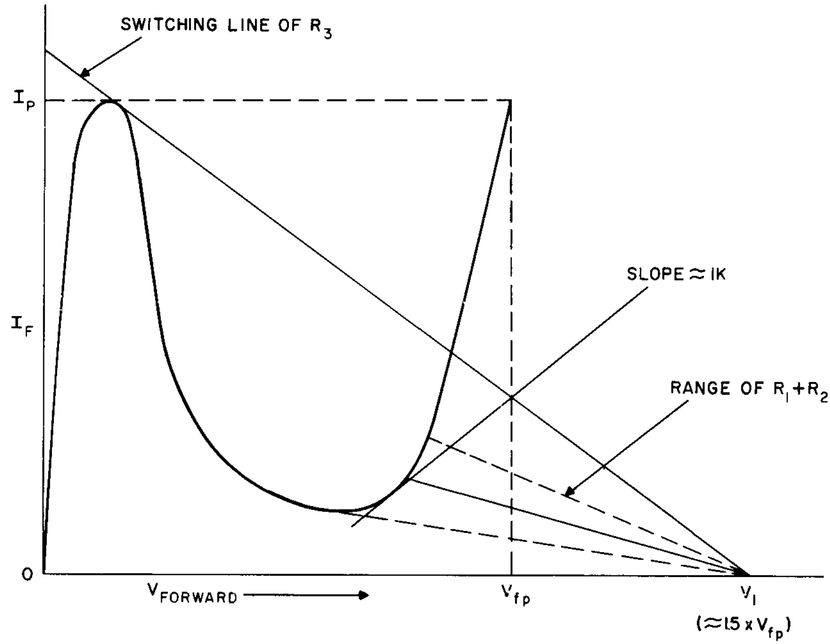
FIGURE 7.10

Figure 7.11 shows the region of operation.

Based on the results of measurements of capacitance variation with voltage made using the former method, one can expect that the junction capacitance will vary approximately as:

$$C \propto \frac{K}{\left(V_{\text{Band Gap}} - (V_{\text{Forward}} + V_{\text{Degeneracy}})\right)^{1/2}}$$

so that the somewhat smaller junction capacitance in the negative



OPERATING REGION AND LOAD LINE
FIGURE 7.11

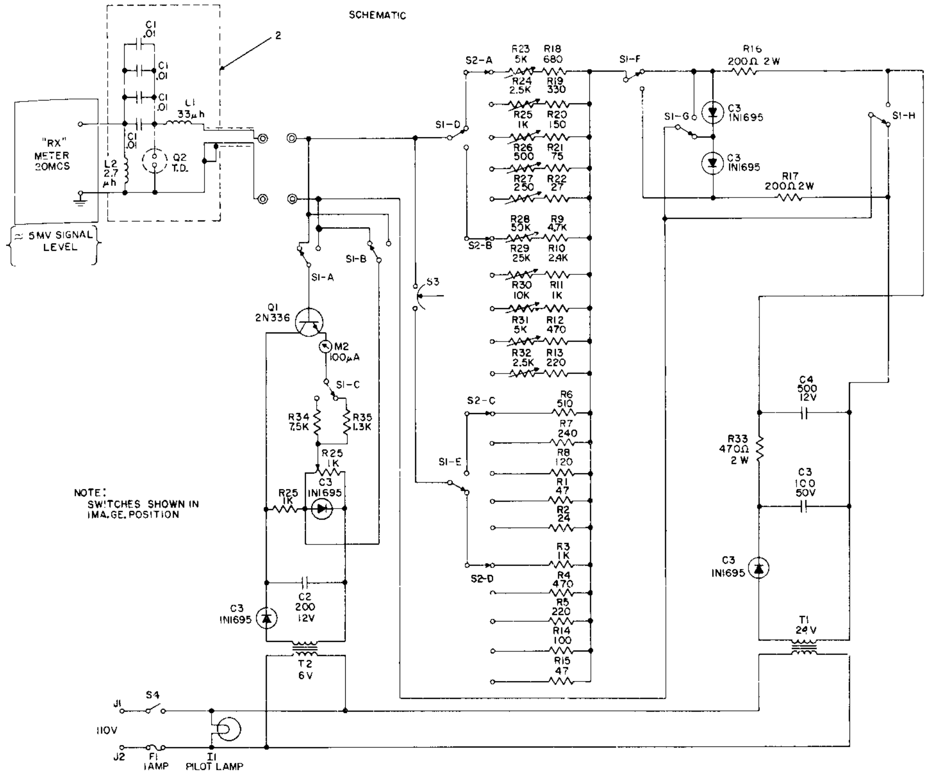
conductance region may be easily calculated. (The case capacitance may be separately measured and subtracted from the measured quantity.)

It cannot be over-emphasized that the measuring signal across the tunnel diode must be kept to less than 10 millivolts to prevent errors due to the non-linearity of the diode characteristics.

The RF choke (L_1) is used to decouple the bias network from the bridge. Since the tunnel diode is switching, R_3 is used to supply a current somewhat greater than the peak point current for the particular range of units being measured, R_1 need only be $> 1/|-g_d|$, but is usually made several times this value to prevent oscillation near the valley region. R_2 is used to adjust the intercept point on the device characteristics so that the bridge measures a parallel positive resistance in the order of 1K ohm or greater. (The bridge lead inductance will still reduce the measured capacity somewhat by Lg_d^2, but if the parallel resistance is 1K ohm or larger it would require 1 μh of lead inductance to cause an error of 1 $\mu\mu$f in the reading. If care is taken in keeping lead lengths to less than one inch, the resultant error will be less than 0.2 $\mu\mu$f and can be neglected.

The supply voltage (V_1) is usually in the order of 0.75v for germanium and 1.5v for gallium arsenide and is supplied by one silicon diode or two in series as required. Figure 7.12 shows a complete test circuit suitable for tunnel diodes in the range of 1 to 25 ma peak currents.

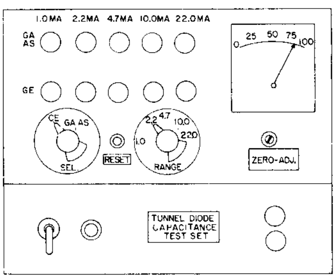
SCHEMATIC



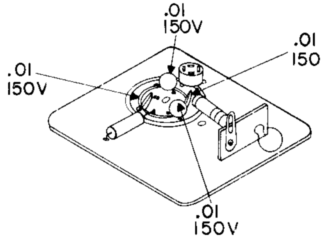
CR-3	RECTIFIER	IN1695
R1	RESISTOR	47Ω
R2	RESISTOR	24Ω
R3	RESISTOR	1 KΩ
R4	RESISTOR	470Ω
R5	RESISTOR	220Ω
R6	RESISTOR	510Ω
R7	RESISTOR	240Ω
R8	RESISTOR	120Ω
R9	RESISTOR	4.7KΩ
R10	RESISTOR	2.4KΩ
R11	RESISTOR	1KΩ
R12	RESISTOR	470Ω
R13	RESISTOR	220Ω
R14	RESISTOR	100 Ω
R15	RESISTOR	47 Ω
R16	RESISTOR	200Ω-2W
R17	RESISTOR	200Ω-2W
R18	RESISTOR	680Ω
R19	RESISTOR	330Ω
R20	RESISTOR	150 Ω

R21	RESISTOR	75Ω
R22	RESISTOR	27Ω
R23	RESISTOR	5KΩ
R24	RESISTOR	2,5KΩ
R25	RESISTOR	1KΩ
R26	RESISTOR	500Ω
R27	RESISTOR	250 Ω
R28	RESISTOR	50K Ω
R29	RESISTOR	25K Ω
R30	RESISTOR	10K Ω
R31	RESISTOR	5 K Ω
R32	RESISTOR	2.5KΩ
R33	RESISTOR	470Ω-2W
R34	RESISTOR	7.5KΩ
R35	RESISTOR	1.3KΩ
C1	CAPACITOR	.01μFD
C2	CAPACITOR	200μFD.-12V.
C3	CAPACITOR	100μ FD 50V
C4	CAPACITOR	500μ FD 12 V
T1	TPANS	24 V. STANCOR P-6469

T2	TRANS	6V. STANCOR P-6469
L1	CHOKE	33μHY
L2	CHOKE	2.7μ HY
S1	SWITCH	GE.-GA.AS. SWITCH 2 POS. 8 CIR.
S2	SWITCH	IpRANGE SWITCH 5 POS. 4 CIR.
S3	SWITCH	N.O. PUSH BUTTON RESET SW.
F1	FUSE	1 AMP
S4	SWITCH	SPST. TOGGLE POSITIVE ACTION BIRNBACH CAT # 6280
J1	TIP JACK	E.F. JOHNSON CAT # 105-520
J2	TIP JACK	E.F. JOHNSON CAT # 105-520
M1	"RX" METER	A+- 20 MCS. TYPE -#250-A
M2	METER	100μ A.
Q1	TRANSISTOR	4JD4C. 2N336
Q2	TEST SOCKET FOR TUNNEL DIODE	
I1	LAMP	NE-51
XF1	FUSE HOLDER	BUSSMAN TYPE HJM-CC



CONTROL PANEL



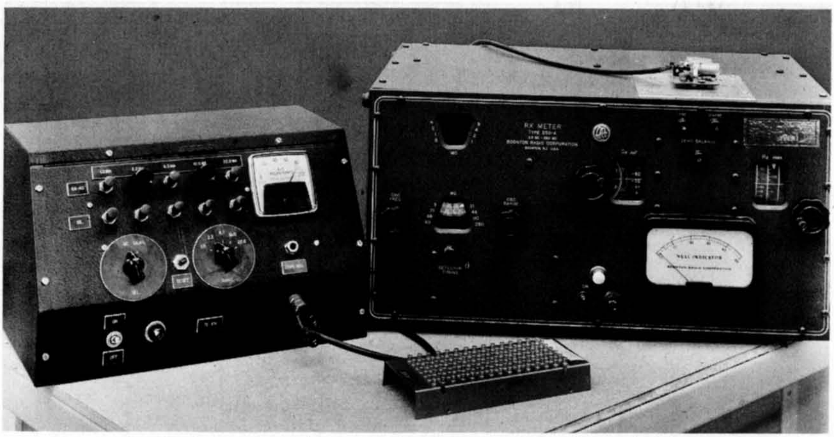
CAPACITANCE BRIDGE

CAPACITY TEST SET CIRCUIT DIAGRAM

FIGURE 7.12

L_2 is used to offset the bridge if capacitances larger than 20 $\mu\mu$f are to be measured (0.27 μh will offset ≈ 20 $\mu\mu$f at the measuring frequency of 20 mc). Note: When the bias is adjusted so that the parallel resistance measured is much greater than 1 - 5Kohms, the measuring signal will tend to switch the device, or the bridge readings will tend to drift, because of the relatively large excursions of intercept voltage with signal, and "averaged" values of R and C will prevent a sharp null.

Figure 7.13 shows a photograph of the test set up.



PHOTOGRAPH OF CAPACITY TEST SET UP

FIGURE 7.13

7.4 R_S Test Set

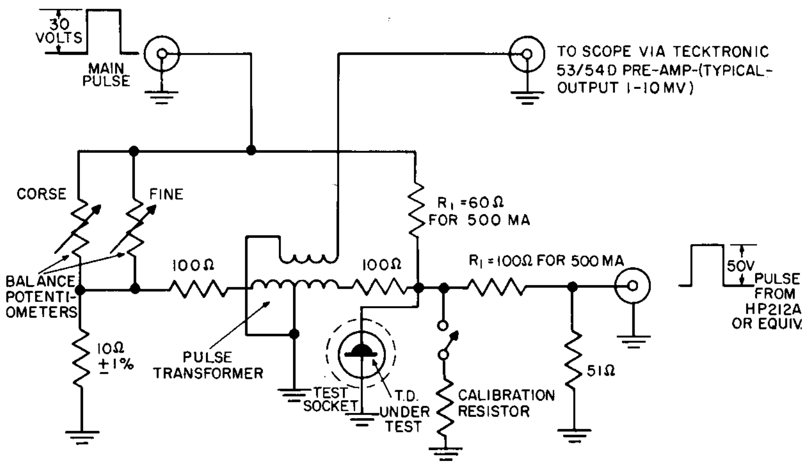
The "Series Resistance" (R_s) is the total internal resistance of the device including contact and lead resistance, as well as the ohmic bulk resistance of the semiconductor material.

In order to obtain a true R_s measurement, it is necessary to measure the incremental slope resistance of the device in its "ohmic" region. This region exists in the high current reverse direction (at least ten times I_p).

The simplest test method would be to apply a large steady state reverse DC current and then measure the incremental slope of R_s by applying a small AC signal. Tunnel diode junction areas are very small however, and the excessive heating which will result at large currents is potentially capable of damaging the units. As a result, a low duty cycle pulse technique was found preferable and this system adopted in the following test set.

A 40 μsec. high current pulse at 60 c/s (0.24% duty cycle) is applied to bias the tunnel diode in the reverse direction. A shorter, lower amplitude pulse is superimposed on top of the main pulse; the latter acting as a pedestal, the former as an "interrogating" or "sampling" pulse.

These pulses are fed into a bridge network containing the tunnel diode as per Figure 7.14. The series R_1 and R_2 in this figure are chosen for a main pulse-to-incremental pulse ratio of 10:1. Since the main pulse feeds both arms of the bridge, the balance potentiometers are used to balance this main pulse out. The output appearing across the pulse transformer is caused by the incremental pulse only. By use of accurate calibration resistors, the pulse output in millivolts can be calibrated directly into ohms for R_s.



R_s TEST BRIDGE

FIGURE 7.14

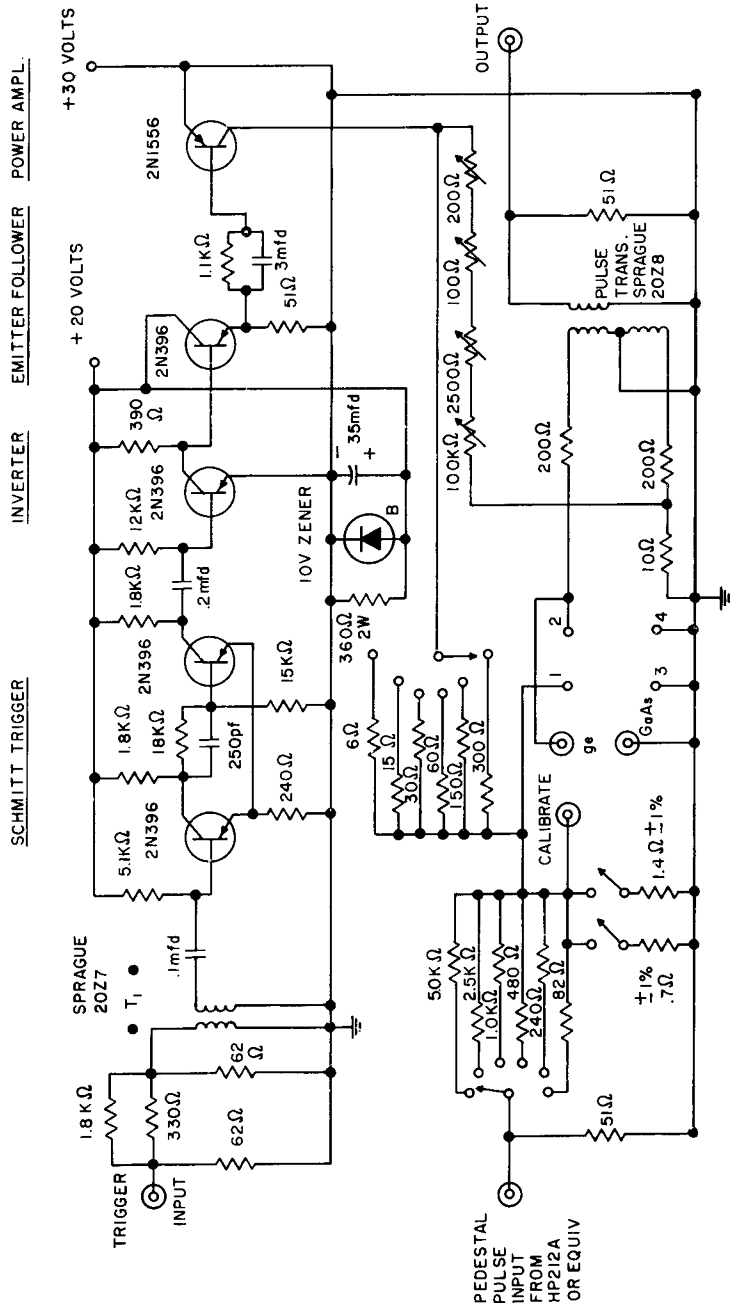
Figure 7.15 is the diagram of this test set.

Test Set Up and Calibration Procedure

The pulse generator used in this "test set up" is a Hewlett-Packard 212A or equivalent. The oscilloscope must be rather sensitive and a Tektronix scope with a 53/54D Pre-Amplifier was used, set up for "External Positive Synchronization". The power supply is set for a 30 volt output. The amplitude of the pulse from the 212A generator should be set to 50 volts when connected to the test set.

1) Place a jumper from the "Calibrate" jack to the Ge (germ.) jack beside the tunnel diode socket. With the tunnel diode removed from its socket, switch one of the calibration resistors into the circuit.

2) Select the desired current test level. Germanium devices can be tested at 100 x I_p to give the most accurate indication of R_s. Because of higher current density and possibly also because of greater power dissipation resulting from a larger voltage at any given current, gallium arsenide devices should only be tested at 20 x I_p.



R_s **TEST SET DIAGRAM**
FIGURE 7.15

3) Switch in one of the calibration resistors and balance the bridge so that the only voltage drop across the bridge at balance is that of the 10 μsec. "interrogation" pulse, using the scope as visual indicator.

This output, divided by the value of calibration resistance used, will now give a calibration in millivolt per ohm.

Operating Procedure

Following calibration, disconnect the jumper and insert a clip lead to make a connection to the diode cap (TO-18 package) in the jack corresponding to the type of device being tested, Ge or GaAs. Next insert the tunnel diode, making sure that the tab for PN devices (such as all General Electric germanium tunnel diodes) point to the left. (Pin 3 of device into socket pin #1). With the tunnel diode in its socket, make the clip lead connection to the cap.

After balancing of the bridge, read the output amplitude of the interrogation pulse.

All that remains is to compute R_s from the calibrated millivolt/ohm reading.

7.5 Measurement of the Device Inductance (L_s)

A precise measurement of "L_s" is a rather difficult measurement to make. The magnitude of this parameter for TO-18 packaged devices is of the order of 1 - 10 nanohenries depending on the method of connection made to the device. For microwave packages (stripline or cartridge types) the device inductance will be an order of magnitude smaller.

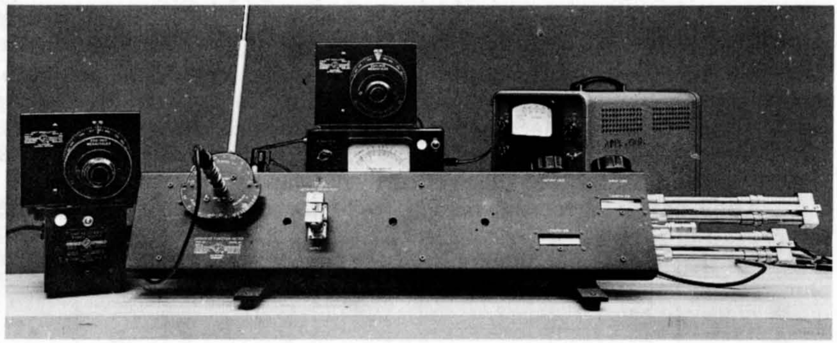
The most necessary item of a L_s test circuit would hence be an extremely low inductance adapter for the unit under test. Also, since the magnitude of this inductance to be tested is so small, a very elaborate test set must be used. The test set used here was the General Radio Transfer-Function and Immittance Bridge, Type 1607-A.

Performing an impedance test at 500 Mc/s, on a "squared off U" piece of wire about 1/4" long, one can bring the plane of measurement slightly above the socket of the transistor mount 1607-P101. By recording the admittance term with the wire protruding 1/8" from the socket, the reference plane is established at this distance from the socket.

Now one can substitute an internally shorted tunnel diode case with the lead length one intends to use plus about 1/4". By subtracting the first reading from the second one, one has established the inductance of the device above the reference plane.

Another method to determine L_s is to measure the tunnel diode impedance at a low frequency when biased in the valley region. This establishes the capacitive reactance since the inductive reactance is negligible at low frequencies. Another impedance measurement at a higher frequency ($\geq$ 500 Mc/s) where the inductive reactance is appreciable will enable the equivalent inductance to be calculated from the measured impedance and previously determined capacitance.

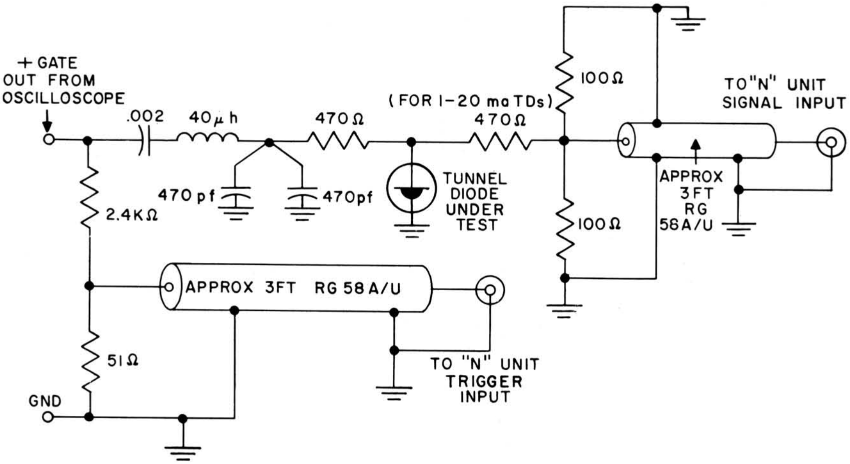
For UHF or microwave application, even for some lower frequency application where L_S must be minimized, a test adapter should be built. General Radio 's "Experimenter" brochure (Vol. 34, Nos. 7 & 8, July-August, 1960) or their Reprint #E109-October, 1960 has an outline drawing of such an adapter for TO-18 packaged devices. These two brochures also contain measurement procedures for most tunnel diode parameters, using the aforementioned Type 1607-A bridge.



PHOTOGRAPH OF L_S TEST SET UP
FIGURE 7.16

7.6 Risetime Measurement

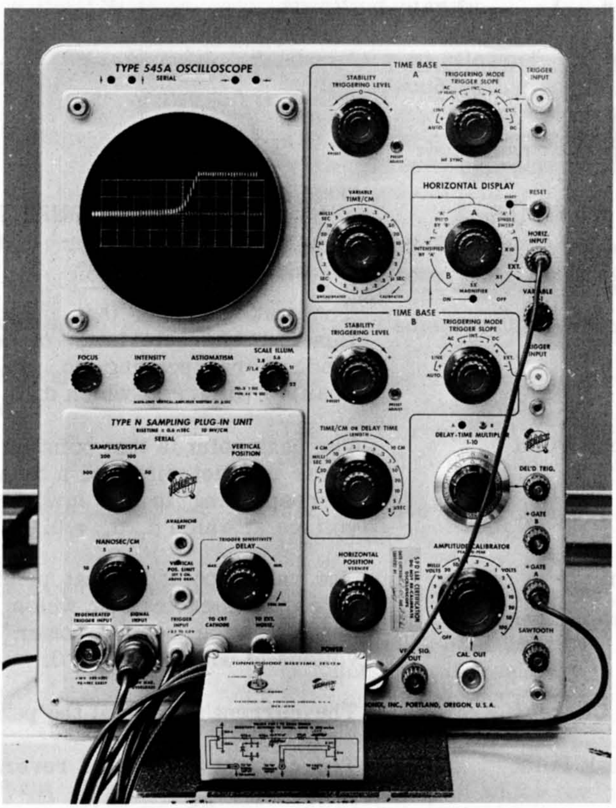
A convenient method of testing switching speed (risetime) is shown in Figure 7.17.



RISETIME TEST SET CIRCUIT DIAGRAM
FIGURE 7.17

A Tektronix plug-in oscilloscope provide both the current ramp source for the tunnel diode and the pre-trigger for the Type "N" sampling scope unit. The use of this sampling scope (or other means of presenting extremely fast speed information) is essential since the risetime of tunnel diodes is much too fast for conventional oscilloscopes.

The Type "N" unit is set up conventionally while the oscilloscope main sweep generator is allowed to "free run" at 1 μsec/cm. The "+ GATE OUT" triggers the "N" unit and also provides a current ramp with a low rate of rise which allows the tunnel diode to switch at essentially its own rate. The rise time of the scope is in the order of 0.6 nanoseconds and should be accounted for in the measurement. Figure 7.18 is a photograph of the test set up displaying a 1.4 nanosecond risetime pulse. This corresponds to an actual circuit risetime of 1 nanosecond since the total risetime is the square root of the squares of the scope and the circuit risetimes.



PHOTOGRAPH OF RISETIME TEST SET UP

FIGURE 7.18

The unit shown in Figure 7.18 is commercially advertised by Tektronix under their part #013-029.

CHAPTER 8

TUNNEL DIODE SPECIFICATIONS

Definition of Terms

Tunnel and Back Diode characteristics are defined with reference to the static characteristic curves of Figure 1 and Figure 2 respectively.

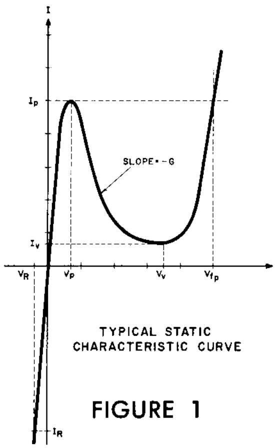


FIGURE 1

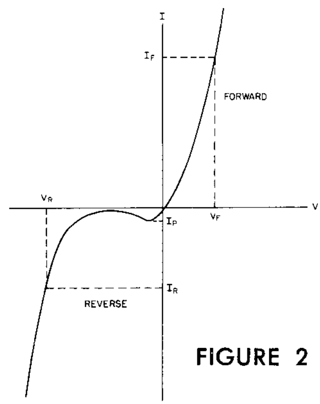


FIGURE 2

Terminology	Symbol	Definition
Peak Point		The peak point is that point on the forward characteristic of a tunnel diode corresponding to the lowest positive voltage at which $d_I/d_V = 0$.
Reverse Peak Point		The peak point is that point on the reverse characteristic of a tunnel diode corresponding to the lowest reverse voltage at which $d_I/d_V = 0$.
Valley Point		The valley point is that point on the forward characteristic which corresponds to the second lower positive voltage at which $d_I/d_V = 0$.
Peak Point Current	I_P	The current flowing at the peak point.
Reverse Peak Point Current	I_P	The current flowing at reverse peak point.
Peak Point Voltage	V_P	The voltage at which the peak point occurs.
Valley Point Current	I_V	The current flowing at the valley point.

Terminology	Symbol	Definition
Valley Point Voltage	V_V	The voltage at the valley point.
Peak-to-Valley Point Current Ratio	I_P/I_V	The ratio of the peak point current to the valley point current.
Forward Peak Point Current Voltage	V_{fp}	The voltage corresponding to that point on the forward characteristic where the current is equal to the maximum specified peak point current.
Forward Voltage	V_F	The voltage corresponding to a point on the forward characteristic at a specified current.
Forward Current	I_F	The current flowing in the first quandrant-conducting region.
Reverse Voltage	V_R	The voltage corresponding to a point on the reverse characteristic curve at a specified current.
Reverse Current	I_R	The current flowing when operated in the third quandrant.
Series Inductance	L_S	The total series equivalent inductance.
Series Resistance	R_S	The total series equivalent resistance.
Capacity	C	The barrier capacity of the intrinsic diode.
Negative Conductance	-G	The negative conductance of the intrinsic diode.
Resistance Cut-off Frequency	f_{ro}	The frequency at which the real part of the diode admittance measured at its terminals goes to zero. The resistance cut-off frequency is given by: $f_{ro} = \frac{g}{2\pi C}\sqrt{\frac{1}{gR_S} - 1}$
Self-Resonant Frequency	f_{xo}	The frequency at which the imaginary part of the diode admittance goes to zero. The self-resonant frequency is given by: $f_{xo} = \frac{1}{2\pi}\sqrt{\frac{1}{L_S C} - \frac{g^2}{C}}$
Pico-farad	pf	Micro-micro farad ($\mu\mu f$) i.e. 10^{-12} farads.

Terminology	Symbol	Definition
Nano-henry	nh	Milli-micro henry ($m\mu h$) i.e. 10^{-9} henries.
Nano-second	nsec.	Milli-micro second ($m\mu sec.$) i.e. 10^{-9} seconds.

The "Tunnel" Diode is a new semiconductor device which makes use of the quantum mechanical tunneling phenomenon thereby attaining unique negative conductance characteristic and very high frequency performance. Its static characteristics are shown in Figure 3.

A "Back" Diode is essentially a "tunnel" diode operated in the reverse direction. For example, you take a "tunnel" diode characteristic curve (Figure 3) and flip it over interchanging quandrants 3 and 4 which results in a "back" diode characteristic curve (Figure 4). The polarity is changed on the back diode since it is operated in the same mode as a conventional diode whose characteristic curve has the conducting region (forward voltage) in the first quandrant and blocking region (reverse voltage) in the third quandrant.

The tunnel diode is designed with emphasis on peak point control and the negative resistance region. The "back" diode is designed with emphasis on forward voltage and its control. In most applications the reverse peak point current of a "back" diode would correspond to a leakage current in a conventional diode. Naturally, the smaller the leakage, the better. The peak point current on back diodes is usually much lower than the peak point current on tunnel diodes. Some "back diodes" are low peak current tunnel diodes and as such do not offer the fine control on conducting voltage that is desired. General Electric Back Diodes are designed for specific diode applications. In order to cover a wide range of applications General Electric offers several types, each of which has minimum and maximum forward voltages for a specific current level of operation.

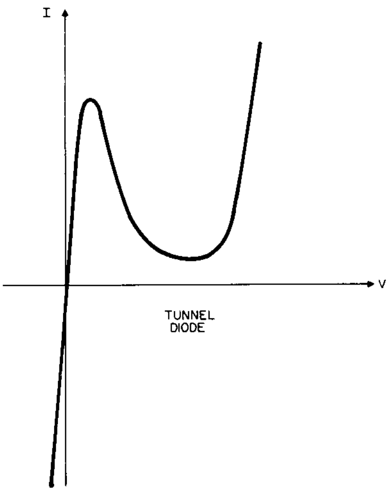


FIGURE 3

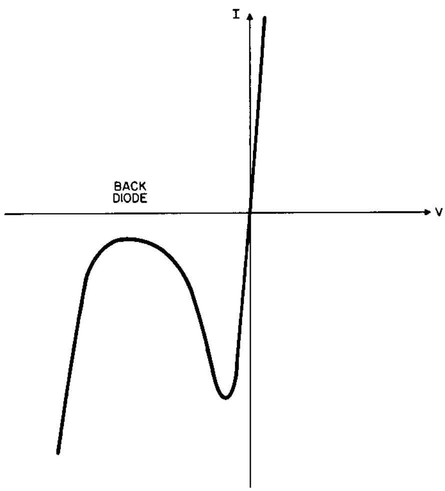


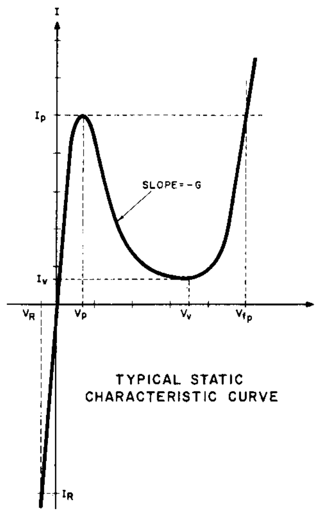
FIGURE 4

TUNNEL DIODE SPECIFICATIONS

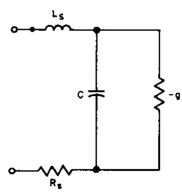
1N2939, 1N2939A

Outline Drawing No. 1

The 1N2939 and 1N2939A are germanium tunnel diodes which make use of the quantum mechanical tunneling phenomenon thereby attaining a unique negative conductance characteristic and very high frequency performance. These devices are designed for low level switching and small signal applications with frequency capabilities up to 2.2 Kmc. They feature a closely controlled peak point current, good temperature stability and extreme resistance to nuclear radiation.



TYPICAL STATIC CHARACTERISTIC CURVE



EQUIVALENT CIRCUIT (BIASED IN NEGATIVE CONDUCTANCE REGION)

SPECIFICATIONS

ABSOLUTE MAXIMUM RATINGS: (25°C)

Current			
Forward (—55 to +100°C)		5	ma
Reverse (—55 to +100°C)		10	ma
Temperature			
Storage	T_{STG}	—55 to +100	°C
Operating Junction	T_J	—55 to +100	°C
Lead Temperature 1/16" ±1/32" From Case for 10 seconds	T_L	260	°C

ELECTRICAL CHARACTERISTICS: (25°C) (1/8" Leads)

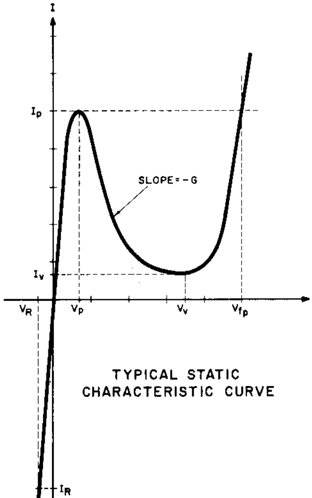
		Min.	Typ.	Max.	
Peak Point Current 1N2939	I_p	0.9	1.0	1.1	ma
Peak Point Current 1N2939A	I_p	0.975	1.0	1.025	ma
Valley Point Current	I_v		0.10	0.14	ma
Peak Point Voltage 1N2939A	V_p	50	60	65	mv
Valley Point Voltage	V_v		350		mv
Reverse Voltage (I_R = 1.0 ma)	V_R			30	mv
Forward Peak Point Current Voltage 1N2939	V_{fp}	450	500	600	mv
Forward Peak Point Current Voltage 1N2939A	V_{fp}	450	500	550	mv
Peak Point Current to Valley Point Current Ratio	I_p/I_v		10		
Negative Conductance	—G		6.6×10^{-3}		mho
Total Capacity 1N2939	C		5.0	15	pf
Total Capacity 1N2939A	C		4.0	10	pf
Series Inductance	L_s*		6		nh
Series Resistance	R_s		1.5	4.0	ohm

*Inductance will vary 1-12 nh (10^{-9} henries) depending on lead length.

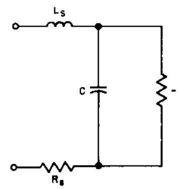
1N2940, 1N2940A

Outline Drawing No. 1

The 1N2940 and 1N2940A are germanium tunnel diodes which make use of the quantum mechanical tunneling phenomenon thereby attaining a unique negative conductance characteristic and very high frequency performance. These devices are designed for low level switching and small signal applications with frequency capabilities up to 2.2 Kmc. They feature a closely controlled peak point current, good temperature stability and extreme resistance to nuclear radiation.



TYPICAL STATIC CHARACTERISTIC CURVE



EQUIVALENT CIRCUIT (BIASED IN NEGATIVE CONDUCTANCE REGION)

SPECIFICATIONS

ABSOLUTE MAXIMUM RATINGS: (25°C)

Current

Forward (—55 to +100°C)		5	ma
Reverse (—55 to +100°C)		10	ma

Temperature

Storage	T_{STG}	—55 to +100	°C
Operating Junction	T_J	—55 to +100	°C
Lead Temperature 1/16″ ±1/32″ From Case for 10 seconds	T_L	260	°C

ELECTRICAL CHARACTERISTICS: (25°C) (1/8″ Leads)

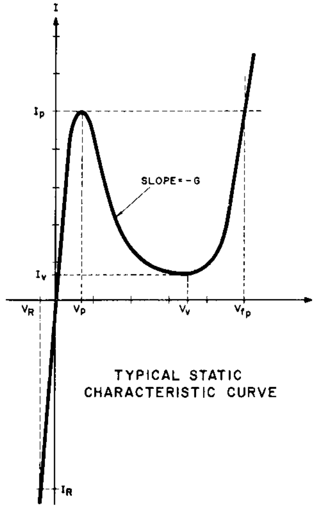
		Min.	Typ.	Max.	
Peak Point Current 1N2940	I_p	0.9	1.0	1.1	ma
Peak Point Current 1N2940A	I_p	0.975	1.0	1.025	ma
Valley Point Current	I_v		0.13	0.22	ma
Peak Point Voltage 1N2940A	V_p	50	60	65	mv
Valley Point Voltage	V_v		350		mv
Reverse Voltage (I_R = 1.0 ma)	V_R			30	mv
Forward Peak Point Current Voltage 1N2940	V_{fp}	450	500	600	mv
Forward Peak Point Current Voltage 1N2940A	V_{fp}	450	500	550	mv
Peak Point Current to Valley Point Current Ratio	I_p/I_v		8		
Negative Conductance	—G		6.6×10^{-3}		mho
Total Capacity 1N2940	C		5.0	10	pf
Total Capacity 1N2940A	C		4.0	7	pf
Series Inductance	L_s*		6		nh
Series Resistance	R_s		1.5	4.0	ohm

*Inductance will vary 1-12 nh (10^{-9} henries) depending on lead length.

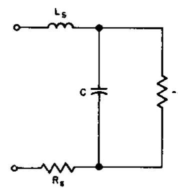
1N2941, 1N2941A

Outline Drawing No. 1

The 1N2941 and 1N2941A are germanium tunnel diodes which make use of the quantum mechanical tunneling phenomenon thereby attaining a unique negative conductance characteristic and very high frequency performance. These devices are designed for low level switching and small signal applications. They feature a closely controlled peak point current, good temperature stability and extreme resistance to nuclear radiation.



TYPICAL STATIC CHARACTERISTIC CURVE



EQUIVALENT CIRCUIT (BIASED IN NEGATIVE CONDUCTANCE REGION)

SPECIFICATIONS

ABSOLUTE MAXIMUM RATINGS: (25°C)

Current

Forward (—55 to +100°C)		25	ma
Reverse (—55 to +100°C)		50	ma

Temperature

Storage	T_{STG}	—55 to +100	°C
Operating Junction	T_J	—55 to +100	°C
Lead Temperature 1/16″ ±1/32″ From Case for 10 seconds	T_L	260	°C

ELECTRICAL CHARACTERISTICS: (25°C) (1/8″ Leads)

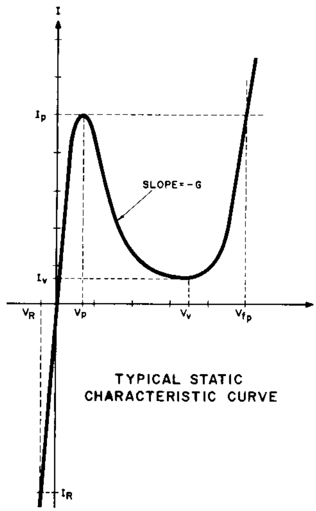
		Min.	Typ.	Max.	
Peak Point Current 1N2941	I_p	4.2	4.7	5.2	ma
Peak Point Current 1N2941A	I_p	4.58	4.7	4.82	ma
Valley Point Current	I_v		0.6	1.04	ma
Peak Point Voltage 1N2941A	V_p	50	60	65	mv
Valley Point Voltage	V_v		350		mv
Reverse Voltage (I_R = 4.7 ma)	V_R			30	mv
Forward Peak Point Current Voltage 1N2941	V_{fp}	450	500	600	mv
Forward Peak Point Current Voltage 1N2941A	V_{fp}	450	500	550	mv
Peak Point Current to Valley Point Current Ratio	I_p/I_v		8		
Negative Conductance	—G		30×10^{-3}		mho
Total Capacity 1N2941	C		15	50	pf
Total Capacity 1N2941A	C		10	30	pf
Series Inductance	L_s*		6		nh
Series Resistance	R_s		0.5	2.0	ohm

*Inductance will vary 1-12 nh (10^{-9} henries) depending on lead length.

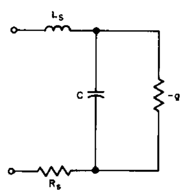
1N2969, 1N2969A

Outline Drawing No. 1

The 1N2969 and 1N2969A are germanium tunnel diodes which make use of the quantum mechanical tunneling phenomenon thereby attaining a unique negative conductance characteristic and very high frequency performance. These devices are designed for low level switching and small signal applications with frequency capabilities up to 2.5 Kmc. They feature a closely controlled peak point current, good temperature stability and extreme resistance to nuclear radiation.



TYPICAL STATIC CHARACTERISTIC CURVE



EQUIVALENT CIRCUIT (BIASED IN NEGATIVE CONDUCTANCE REGION)

SPECIFICATIONS

ABSOLUTE MAXIMUM RATINGS: (25°C)

Current

Forward (−55 to +100°C)		10	ma
Reverse (−55 to +100°C)		20	ma

Temperature

Storage	T_{STG}	−55 to +100	°C
Operating Junction	T_J	−55 to +100	°C
Lead Temperature 1/16″ ±1/32″ From Case for 10 seconds	T_L	260	°C

ELECTRICAL CHARACTERISTICS: (25°C) (1/8″ Leads)

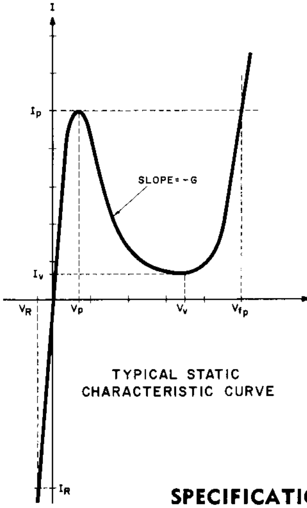
		Min.	Typ.	Max.	
Peak Point Current 1N2969	I_p	2.0	2.2	2.4	ma
Peak Point Current 1N2969A	I_p	2.145	2.2	2.255	ma
Valley Point Current	I_v		.29	.48	ma
Peak Point Voltage 1N2969A	V_p	50	60	65	mv
Valley Point Voltage	V_v		350		mv
Reverse Voltage (I_R = 2.2 ma)	V_R			30	mv
Forward Peak Point Current Voltage 1N2969	V_{fp}	450	500	600	mv
Forward Peak Point Current Voltage 1N2969A	V_{fp}	450	500	550	mv
Peak Point Current to Valley Point Current Ratio	I_p/I_v		8		
Negative Conductance	−G		16×10^{-3}		mho
Total Capacity 1N2969	C		8	25	pf
Total Capacity 1N2969A	C		6	15	pf
Series Inductance	L_s*		6		nh
Series Resistance	R_s		1.0	3.0	ohm

*Inductance will vary 1-12 nh (10^{-9} henries) depending on lead length.

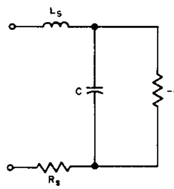
1N3149, 1N3149A

Outline Drawing No. 1

The 1N3149 and 1N3149A are germanium tunnel diodes which make use of the quantum mechanical tunneling phenomenon thereby attaining a unique negative conductance characteristic and very high frequency performance. These devices are designed for low level switching and small signal applications with frequency capabilities up to 1.5 Kmc. They feature closely controlled peak point current, good temperature stability and extreme resistance to nuclear radiation.



TYPICAL STATIC CHARACTERISTIC CURVE



EQUIVALENT CIRCUIT (BIASED IN NEGATIVE CONDUCTANCE REGION)

SPECIFICATIONS

ABSOLUTE MAXIMUM RATINGS: (25°C)

Current

Forward (—55 to +100°C)		50	ma
Reverse (—55 to +100°C)		50	ma

Temperature

Storage	T_{STG}	—55 to +100	°C
Operating Junction	T_J	—55 to +100	°C
Lead Temperature 1/16" ±1/32" From Case for 10 seconds	T_L	260	°C

ELECTRICAL CHARACTERISTICS: (25°C) (1/8" Leads)

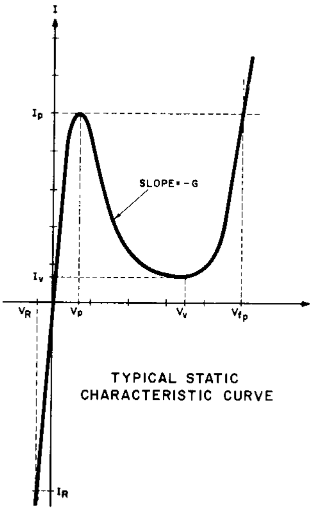
		Min.	Typ.	Max.	
Peak Point Current 1N3149	I_p	9.0	10.0	11.0	ma
Peak Point Current 1N3149A	I_p	9.75	10.0	10.25	ma
Valley Point Current	I_v		1.3	2.2	ma
Peak Point Voltage 1N3149A	V_p	50	60	65	mv
Valley Point Voltage	V_v		350		mv
Reverse Voltage (I_R = 10 ma)	V_r			30	mv
Forward Peak Point Current Voltage 1N3149	V_{fp}	450	500	600	mv
Forward Peak Point Current Voltage 1N3149A	V_{fp}	450	500	550	mv
Peak Point Current to Valley Point Current Ratio	I_p/I_v		8		
Negative Conductance	—G		60×10^{-3}		mho
Total Capacity 1N3149	C		30	90	pf
Total Capacity 1N3149A	C		25	50	pf
Series Inductance	L_s*		6		nh
Series Resistance	R_s		.25	1.5	ohm

*Inductance will vary 1-12 nh (10^{-9} henries) depending on lead length.

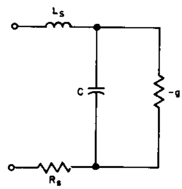
1N3150

Outline Drawing No. 1

The 1N3150 is a germanium tunnel diode which makes use of the quantum mechanical tunneling phenomenon thereby attaining a unique negative conductance characteristic and very high frequency performance. This device is designed for low level switching and small signal applications with frequency capabilities up to 1.3 Kmc. It features closely controlled peak point current, good temperature stability and extreme resistance to nuclear radiation.



TYPICAL STATIC CHARACTERISTIC CURVE



EQUIVALENT CIRCUIT (BIASED IN NEGATIVE CONDUCTANCE REGION)

SPECIFICATIONS

ABSOLUTE MAXIMUM RATINGS: (25°C)

Current

Forward (—55 to +100°C)		100	ma
Reverse (—55 to +100°C)		100	ma

Temperature

Storage	T_{STG}	—55 to +100	°C
Operating Junction	T_J	—55 to +100	°C
Lead Temperature 1/16″ ±1/32″ From Case for 10 seconds	T_L	260	°C

ELECTRICAL CHARACTERISTICS: (25°C) (1/8″ Leads)

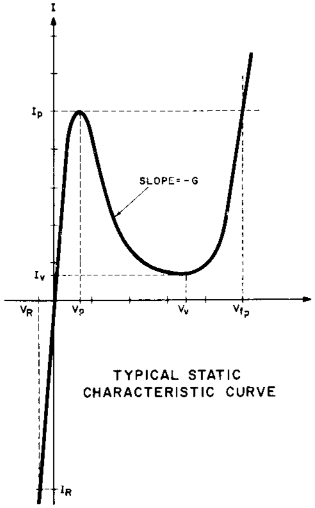
		Min.	Typ.	Max.	
Peak Point Current	I_p	20	22	24	ma
Valley Point Current	I_v		2.9	4.80	ma
Peak Point Voltage	V_p		60		mv
Valley Point Voltage	V_v		350		mv
Reverse Voltage (I_R = 22 ma)	V_r			30	mv
Forward Peak Point Current Voltage	V_{fp}	450	500	600	mv
Peak Point Current to Valley Point Current Ratio	I_p/I_v		8		
Negative Conductance	—G		100×10^{-3}		mho
Total Capacity	C		60	150	pf
Series Inductance	L_s*		6		nh
Series Resistance	R_s		.15	1.0	ohm

*Inductance will vary 1-12 nh (10^{-9} henries) depending on lead length.

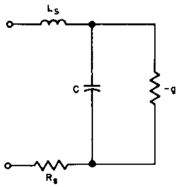
1N3218, 1N3218A

Outline Drawing No. 3

The 1N3218 and 1N3218A are germanium tunnel diodes which make use of the quantum mechanical tunneling phenomenon thereby attaining unique negative conductance characteristics and very high frequency performance. These small stripline type packages are designed for microwave communications, radar, very high frequency amplifiers and oscillator applications. The very low series inductance plus controlled low capacity permits very high frequency performance in the S band.



TYPICAL STATIC CHARACTERISTIC CURVE



EQUIVALENT CIRCUIT (BIASED IN NEGATIVE CONDUCTANCE REGION)

SPECIFICATIONS

ABSOLUTE MAXIMUM RATINGS

Current

Forward (—55 to + 100°C)	5	ma
Reverse (—55 to + 100°C)	10	ma

Temperature

Storage	—55 to + 100	°C

ELECTRICAL CHARACTERISTICS: (25°C)

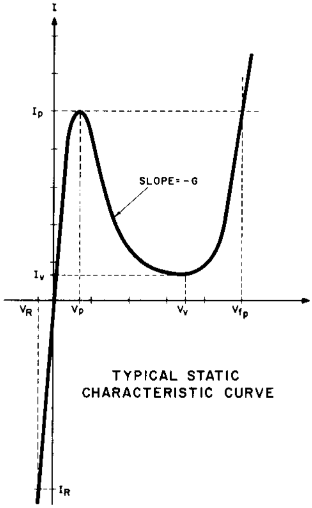
		Min.	Typ.	Max.	
Peak Point Current	I_P	0.9	1.0	1.1	ma
Valley Point Current	I_V		0.13	0.22	ma
Peak Point Voltage	V_P		60		mv
Valley Point Voltage	V_V		350		mv
Forward Voltage ($I_F = I_P = 1$ ma)	V_{fp}	400	500	600	mv
Reverse Voltage ($I_R = 1$ ma)	V_R		20		mv
Negative Conductance	—G	5	9	12	x10^{-3}mho
Series Resistance	R_S		1.5	4.0	ohm
Series Inductance	L_S		0.3	0.5	nh
Total Capacity 1N3218	C	1.5	7	10	pf
Total Capacity 1N3218A	C	1.5	4	5	pf

Color Code: (counter clockwise on flange)
1N3218 Orange—Red—Brown—Gray—Black
1N3218A Orange—Red—Brown—Gray—Brown

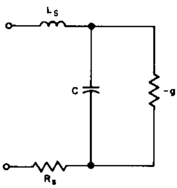
1N3219, 1N3219A

Outline Drawing No. 3

The 1N3219 and 1N3219A are germanium tunnel diodes which make use of the quantum mechanical tunneling phenomenon thereby attaining unique negative conductance characteristics and very high frequency performance. These small stripline type packages are designed for microwave communications, radar, very high frequency amplifiers and oscillator applications. The very low series inductance plus controlled low capacity permits very high frequency performance in the S band.



TYPICAL STATIC CHARACTERISTIC CURVE



EQUIVALENT CIRCUIT (BIASED IN NEGATIVE CONDUCTANCE REGION)

SPECIFICATIONS

ABSOLUTE MAXIMUM RATINGS

Current

Forward (—55 to + 100°C)	5	ma
Reverse (—55 to +100°C)	10	ma

Temperature

Storage	—55 to + 100	°C

ELECTRICAL CHARACTERISTICS: (25°C)

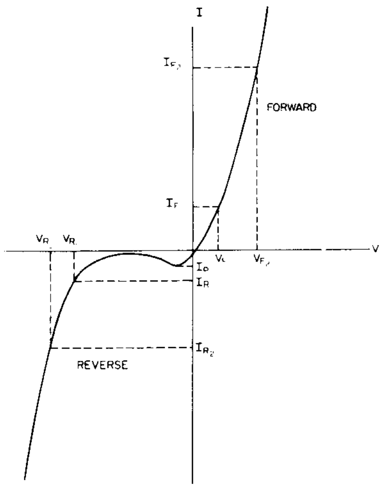
		Min.	Typ.	Max.	
Peak Point Current	I_P	2.0	2.2	2.4	ma
Valley Point Current	I_V		0.28	0.48	ma
Peak Point Voltage	V_P		60		mv
Valley Point Voltage	V_V		350		mv
Forward Voltage ($I_F = I_P = 2.2$ ma)	V_{fp}	450	500	600	mv
Reverse Voltage ($I_R = 2.2$ ma)	V_R		20		mv
Negative Conductance	—G	10	18	25	$\times 10^{-3}$ mho
Series Resistance	R_S		0.7	3.0	ohm
Series Inductance	L_S		0.3	0.5	nh
Total Capacity 1N3219	C	1.5	14	20	pf
Total Capacity 1N3219A	C	1.5	7	10	pf

Color Code: (counter clockwise on flange)
1N3219 Orange—Red—Brown—White—Black
1N3219A Orange—Red—Brown—White—Brown

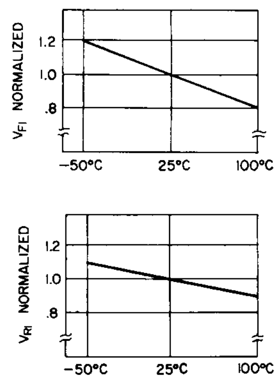
BACK DIODES—GERMANIUM
4JF2A-1, 2, 3, 4, 5, 6, & 7

Outline Drawing No. 2

The General Electric Type 4JF2A is a germanium back diode which makes use of the quantum mechanical tunneling phenomenon thereby attaining very low capacity and low forward voltage drop. This device is designed for high speed computer switching circuits. The very low forward voltage is ideal for use with germanium tunnel diodes. It features stable forward and reverse voltage with temperature.



TYPICAL STATIC CHARACTERISTIC CURVE
BACK DIODE



VOLTAGES VS. AMBIENT TEMPERATURE

SPECIFICATIONS

ABSOLUTE MAXIMUM RATINGS

Current

	TYPE 4JF2A-1	-2	-3	-4	-5	-6	-7	
Forward (−55 to +100°C)	25	15	10	5	5	5	5	ma
Reverse (−55 to +100°C)	10	5	5	5	5	5	5	ma

Temperature

Ambient	−55 to +100	°C
Lead Temperature, 1/16″ ±1/32″ From case for 10 seconds	260	°C

ELECTRICAL CHARACTERISTICS: (25°C) (1/8″ Leads)

		TYPE 4JF2A-1	-2	-3	-4	-5	-6	-7		
Reverse Peak Point Current	I_P	1	.5	.2	.1	.05	.02	.01	ma	max.
Reverse Voltage ($I_R = I_P$ max)	V_{R1}	440	420	400	380	350	330	300	mv	min.
Reverse Voltage ($I_R = 1$ ma)	V_{R2}	440	465	465	465	465	465	465	mv	min.
Forward Voltage at I_{F1}	V_{F1}	80	80	80	80	80	80	80	mv	min.
		100	100	100	100	100	100	100	mv	max.
$I_{F1} =$		10	5	2	1	.5	.2	.1	ma	
Forward Voltage at I_{F2}	V_{F2}	120	130	170	170	170	160	160	mv	typ.
$I_{F2} =$		25	15	10	5	5	2	1	ma	
Total Capacitance	C	15	9	6	6	5	5	5	pf	typ.
		25	15	15	15	15	15	15	pf	max.
Series Inductance	L_s	6	6	6	6	6	6	6	nh	typ.

Outline Drawings

1

DIMENSIONS WITHIN
JEDEC OUTLINE
TO-18

NOTE 1: Minimum tab thickness .005

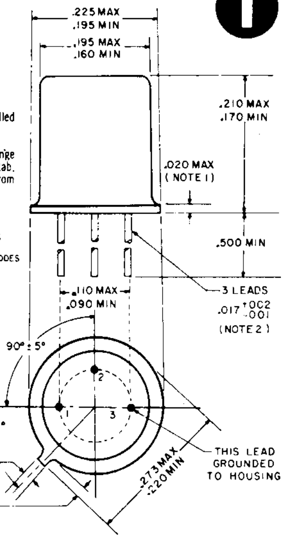
NOTE 2: Lead diameter is not controlled in the area 1 16" from base seat

NOTE 3: Calculated by measuring flange diameter, including tab and excluding tab, and subtracting the larger diameter from the smaller diameter

APPROX WEIGHT: .015 OZ
ALL DIMENSIONS IN INCHES

PIN #1 AND #2 POSITIVE ELECTRODES CONNECTED INTERNALLY TO REDUCE LEAD INDUCTANCE

PIN #3 NEGATIVE ELECTRODE CONNECTED TO THE CASE



2

DIMENSIONS WITHIN
JEDEC OUTLINE
TO-18

NOTE 1: Minimum tab thickness .005.

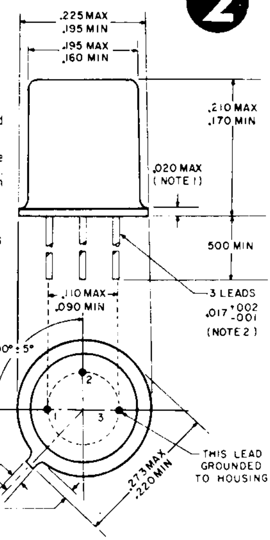
NOTE 2: Lead diameter is not controlled in the area 1 16" from base seat

NOTE 3: Calculated by measuring flange diameter, including tab and excluding tab, and subtracting the larger diameter from the smaller diameter

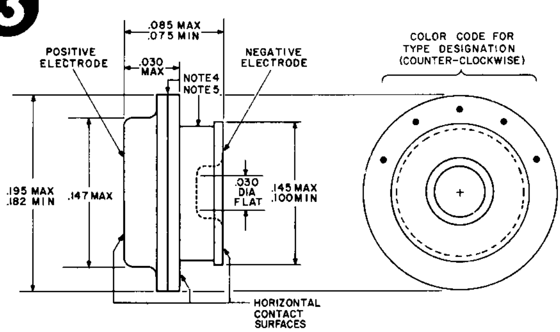
APPROX WEIGHT .015 OZ
ALL DIMENSIONS IN INCHES

PIN #1 AND #2 ARE NEGATIVE ELECTRODES

PIN #3 IS POSITIVE ELECTRODE CONNECTED TO THE CASE



3



NOTES:
1. ALL DIMENSIONS IN INCHES.
2. DIMENSIONS ARE REFERENCES UNLESS TOLERANCED.
3. CONTACT SURFACES TINNED .0003.
4. WELD FLASH ALLOWED (THIS IS NOT A CONTACT SURFACE).
5. INSULATION — DO NOT APPLY CLAMPS.

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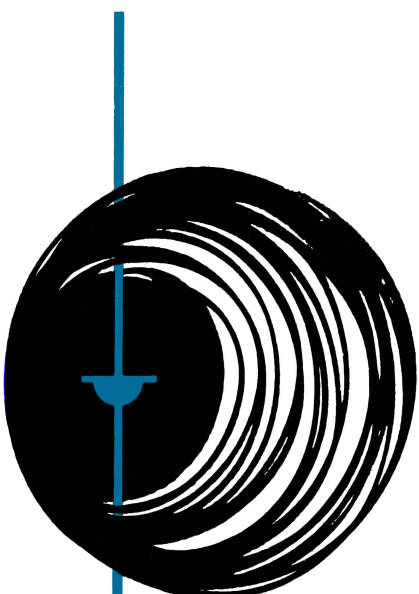

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