1959

Operation of semiconductor junction diodes at very high frequencies

Roy Henry Mattson
Iowa State University

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OPERATION OF SEMICONDUCTOR JUNCTION DIODES AT VERY HIGH FREQUENCIES

by

Roy Henry Mattson

A Dissertation Submitted to the Graduate Faculty in Partial Fulfillment of The Requirements for the Degree of DOCTOR OF PHILOSOPHY

Major Subject: Electrical Engineering

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1959
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The objective of this research was to investigate the a-c electrical characteristics of semiconductor junction diodes at very high and ultra high frequencies for various d-c biasing conditions. An investigation of the electrical characteristics of gas diffused semiconductor junctions under the influence of impinging electromagnetic radiation was performed. As a result of these investigations applications of junction diodes at high frequencies became apparent and were developed.

The procedure followed was to perform an analysis resulting in the small signal high frequency a-c equivalent circuits for alloy, grown and PIN junction diodes. Then the effect of large signals on the equivalent circuits was studied, followed by experimental verification of the theoretical results using commercially available diodes. The reflection characteristics of large area gas diffused PN junction diodes were analyzed. Applications of these semiconductor junction diodes at very high frequencies were invented, and operating systems were tested.

It was predicted that an ideal semiconductor junction diode small signal a-c equivalent circuit was a current sensitive conductance when the diode was forward biased and an open circuit when reverse biased. A voltage dependent depletion layer capacitor placed in shunt with the ideal diode conductance and a shunt leakage conductance as well as a series ohmic body resistance are added to obtain the equivalent circuits of alloy, grown and PIN junction diodes. For alloy junction diodes the depletion layer capacitance varies inversely as the square root of the applied voltage, while the series ohmic body resistance and the shunt conductance are negligible. The grown junction depletion layer capacitance varies inversely as the cube
root of the applied voltage, while the series ohmic body resistance and
the shunt conductance are small. The PIN diode predicted equivalent
circuit is a constant capacitance in parallel with the ideal diode con­
ductance. The predicted equivalent circuit of these diodes when a large
a-c signal is applied to them is the same as the small signal equivalent
circuit for forward d-c bias because of the conductivity modulation effect.
When reverse biased, the alloy and grown junction diodes are voltage sensi­
tive capacitances. The analysis of large area gas diffused PN junctions
predicts that the reflections from the surface of the diode can be con­
trolled.

Measurements of commercially available diodes at very high frequencies
confirmed the predictions. The variation of capacitance with voltage and
the current sensitive conductance were observed. The conductivity modu­
lation effect was also tested.

Various applications of semiconductor junction diodes were predicted,
developed, and tested. Two variable reactance amplifiers which utilize non
linear capacitance to provide a-c gain were designed and constructed, one
operating at 10 megacycles the other at 350 megacycles. The voltage de­
pendent capacitance of PN diodes was used to provide controllable tuning
of transmission lines at high frequencies. PIN junction diodes make excel­
lent switches and a very high frequency antenna switching system was built
and tested. Two electronically controllable shorted stubs were built and
tested. A device for minimizing and controlling reflections from a surface
was designed. The proposed large area PN junction could minimize the
reflection of impinging electromagnetic plane waves at any frequency
between 5 and 500 kilomegacycles. Control was possible through the bias
voltage sensitive depletion layer region. Building and testing the device was not possible because of the lack of facilities.

In this study prediction of the operation of semiconductor junction diodes at very high frequencies was accomplished. Also new uses of semiconductor junction diodes resulted from this investigation.
INTRODUCTION

Many years ago investigators in the field of radio communications observed that metallic points in contact with certain types of crystals had peculiar electrical characteristics. These characteristics proved useful, and many people constructed crystal radio sets using the rectifying contact between the crystal and the metal whisker. This was one of the first applications of semiconductors in the electronics and communications field. The reasons why these devices had desirable characteristics were not clearly understood, and their use decreased as vacuum tubes were perfected.

Much of the progress in employing new materials, such as copper oxide, selenium, and silicon, in the communications field was a product of empiricism even as late as 1930. In 1938 Shockley and Brattain of the Bell Telephone Laboratories initiated research in the field of solid state physics which, together with the works of physicists like Van Vleck, Slater, Sietz, Bozarth, Bragg, Wilson and Mott, expanded the frontiers of knowledge of the solid state (12). The electrical properties of materials were an integral part of this expanding body of knowledge and of greatest interest to the Bell Telephone Laboratories researchers.

During World War II a very necessary part of most radar sets was a small crystal diode used for detection purposes. This crystal rectifier was made from a piece of semiconductor material with a metal whisker in contact with it. The method for making these point contact diodes was a mixture of a little theory and a large amount of empirical data. Although the theory of point contact rectifiers is fairly well understood now, their fabrication still depends on experience to a large extent. Efforts
to perfect point contact diodes for military purposes added greatly to the basic knowledge of solid state physics (30). Until the last year or two such point contact diodes were the only semiconductor devices useful at very high frequencies.

In 1945 Bardeen joined Shockley and Brattain at the Bell Telephone Laboratories, and in 1948 they announced the invention of a tiny amplifying device utilizing semiconductor materials. This was named a point contact transistor, and was a major break-through in the application of semiconductors. The inventors received a Nobel prize for their work. This break-through created tremendous interest in the field of solid state physics, and the resulting increased research activities led to many improvements and discoveries.

Most of the early applications of semiconductors used germanium which, for many devices, becomes inoperative at temperatures slightly higher than room temperature. Research in the use of silicon resulted in better methods of purifying and handling the material and many semiconductor devices were made of this element because of its superior temperature characteristics. Now new semiconductor materials are being studied in many research laboratories.

In 1950 the Bell Telephone Laboratories announced the development of the junction transistor fabricated by a crystal growing technique. Later developments included improved junction diodes, controllable fabricating techniques, power rectifier diodes, and silicon solar cells. Improved semiconductor PN junction diodes were developed using crystal growing, alloying, and diffusion techniques. Each technique resulted in slightly different electrical characteristics and control of electrical charac-
teristics by fabricating methods could be realized to some degree (1). Applications in the power rectifier field followed the invention of silicon power rectifier diodes (22). The invention of the silicon solar cell led to the conversion of light or electromagnetic radiant energy into d-c electrical energy (23).

The semiconductor junction devices previously mentioned, except for point contact diodes, were designed for operation at relatively low frequencies - below 100 megacycles. The high frequency electrical characteristic of PN junction diodes, PIN junction diodes, and gas diffused PN junction diodes had not been investigated previous to this study.
OBJECTIVE AND SCOPE OF THE INVESTIGATION

The objective of this research was to investigate the a-c electrical characteristics of semiconductor junction diodes at very high frequencies and ultra high frequencies for various d-c biasing conditions. Also, an investigation of the electrical characteristics of gas diffused semiconductor junctions under the influence of an impinging electromagnetic radiation was undertaken. As a result of the investigations applications of junction diodes at these higher frequencies became apparent. The applications are discussed in detail.

A discussion of semiconductor materials is presented in the first section of this paper, a treatment of semiconductor junction diodes in the second section, a development of the high frequency characteristics of semiconductor junction diodes in the third section, and finally, particular applications of semiconductor junction diodes are discussed. As equipment to make special semiconductor junctions was unavailable, experimental verification is presented where commercially available devices could be used. Since the commercial devices used were not designed specifically to perform at very high frequencies, results obtained by proper design and fabrication steps should be much better than those presented here.
Most present day semiconductor devices employ either silicon or germanium as the raw material. These materials form a crystal which is basically a face centered cubic structure, but the primitive cell may be regarded as being made up of eight interpenetrating simple cubic lattices (27). This arrangement allows each atom to have four nearest neighbors, and since each atom has four valence or outer electrons these electrons form covalent bonds with an electron from each of the four nearest neighbors. This is represented schematically in the two-dimensional sketch of Figure la. Each circle represents the nucleus of an atom including all filled electron shells, but excluding the four outer or valence electrons which silicon and germanium have since they are from group IV of the periodic table. The four outer electrons are represented by four of the eight dashes arranged around the net four positive electric charges caused by the nucleus and all the filled electron shells. The other dashes represent electrons which are associated with neighboring atoms, and two adjacent electrons represent a covalent bond. Actually Figure la is a simplified physical two-dimensional picture representing a three-dimensional situation, but it is an aid in visualizing operation of semiconductors and semiconductor devices.

Electrical Characteristics of Semiconductors

To obtain a quantitative treatment of the electrical properties of semiconductors it is necessary to investigate the characteristics of motion of the valence electrons. To do this it is necessary to use a quantum mechanical approach which leads to the band theory of solids (5, 32). The results of this approach are the allowable values of momentum and energies
which electrons in the crystal can have. All possible values of electron energies are not allowed due to the periodic nature of the potential inside the crystal caused by the nuclei of the atoms. All wave number are allowed, but the solutions are redundant. The minimum range of wave numbers of interest is found from Brillouin zones. In a particular direction in the crystal it is possible to represent the allowed electron energy values by a sketch Figure lb. Closely spaced allowable electron energy levels exist in the valence band and the conduction band separated by a forbidden band where no allowed energy levels exist. To obtain the distribution of energy levels within a band the momentum space in the direction of interest must be investigated. To determine whether the allowable levels or states are filled with electrons, Fermi-Dirac statistics are used. This is necessary since the electrons act like a degenerate gas because they are so closely spaced, of the order of \(10^{22}\) valence electrons per cubic centimeter. Figure la and lb representing intrinsic semiconductor material can be related to each other in the manner outlined below.

**Intrinsic Semiconductor Material**

A quantitative investigation shows that in a pure or intrinsic semiconductor material there are just as many valence electrons per unit volume as there are available energy states per unit volume in the valence band. If all of the valence electrons in a semiconductor crystal go to the lowest energy states available, they would just exactly fill the valence band. In Figure la this would correspond to a condition where all the covalent bonds are complete. In Figure lb this means all the electrons are in the valence band.
Conduction in a crystal corresponds to a situation where electrons increase their kinetic energy by absorbing some energy from the applied electric field. To do this it must be possible for the electron to move to some slightly higher allowed available energy state, but for the situation just discussed with the valence band filled there are no empty states available except at higher energies in the conduction band. Therefore applying a voltage to a semiconductor sample whose electrons completely fill the valence band will result in no current flow unless the breakdown strength of the sample is exceeded. It is interesting to note that for a metallic conductor the band theory still applies, but either the conduction and valence bands overlap or the valence band is only partially filled with electrons on a per unit volume basis.

In semiconductor materials the distribution of valence electrons is not arbitrary, but is determined by the temperature of the material. The probability, \( P \), that a particular available energy state is occupied by an electron is related to the energy of the available state by the Fermi-Dirac distribution function (32).

\[
P = \frac{1}{1 + e^{(E - E_F)/kT}}
\]

where \( k \) is Boltzmann's constant, \( T \) the absolute temperature, \( E \) the energy of the available state, and \( E_F \) the Fermi level or the energy level which has a probability of one half of being filled. For an intrinsic semiconductor material the Fermi level is midway between the bottom of the conduction band and the top of the valence band. At zero degrees Kelvin the probability of finding an electron in the conduction band is zero as
given by equation 1. At finite temperatures the probability of finding an electron in the conduction band is a function of the gap energy and the temperature, therefore the conductivity and resistivity of a sample of intrinsic semiconductor material are very temperature sensitive. Also at the same temperature, different semiconductor materials have different conductivities because of the difference in their gap energies. For example, germanium has an intrinsic resistivity of 45 ohm - centimeters at 300 degrees Kelvin while the intrinsic resistivity of silicon at 300 degrees Kelvin is 240,000 ohm - centimeters (6).

At a finite temperature for a particular semiconductor material, the number of electrons in the conduction band can be computed using equation 1 and the distribution of states in the conduction band. Each electron in the conduction band corresponds to an electron breaking away from a covalent bond of the type pictured in Figure la. Each of these conduction electrons can absorb energy from the applied field and move under the influence of this field, thereby constituting current flow by the movement of electrons. In Figure la this can be pictured as an electron breaking away from a covalent bond and moving under the influence of an applied field.

However, when an electron moves up into the conduction band of Figure lb, it leaves an empty available state in the valence band. Other electrons can move into this empty state in the valence band. A useful way of visualizing this in terms of Figure la is to think of electrons from adjacent atoms moving into the broken covalent bond caused by the conduction electron. This causes the broken covalent bond to change position in the crystal. Since when the bond was broken a negative charge moved away from the vicinity of the broken bond, there is a net positive charge associated with
the broken covalent bond. It is possible to relate the motion of the broken covalent bond to the motion of a positively charged particle called a hole. Thus in terms of Figure 1b a hole is an empty available state in the valence band, and in terms of Figure 1a a hole is the broken covalent band remaining after a conduction electron leaves the vicinity. Under the influence of an applied electric field the positively charged hole can move constituting current flow by holes. In an intrinsic semiconductor material the current is made up of two components, hole flow and conduction electron flow. The conductivity of a semiconductor material is directly related to the number of conduction electrons and holes as given by equation 2 (27).

\[ \sigma = q \left( n \mu_n + p \mu_p \right) \]  

(2)

where \( \sigma \) is the conductivity, \( q \) the charge on an electron, \( n \) the concentration of conduction electrons, \( p \) the concentration of holes, \( \mu_n \) the mobility of conduction electrons in the particular material, and \( \mu_p \) the mobility of holes.

This discussion of conduction in intrinsic semiconductor material treated the electrons as individuals. Actually this is somewhat misleading since the phenomena discussed are statistical in nature and it is impossible to determine what happens to an individual electron. For example hole and conduction electron pairs are always being generated, and there is a continual recombining of these holes and conduction electrons. On the average, however, there is a concentration of conduction electrons and holes, and this is the important item.
By adding small amounts of impurities to intrinsic semiconductor material, very interesting, controllable, and useful electrical characteristics result. If atoms of an impurity are substituted for atoms of a pure material in the crystal structure, the valence electrons of the impurity atoms determine the electrical characteristics of the sample over a wide temperature range. The intrinsic semiconductor material must be very pure to start with since the amount of impurity added is of the order of one impurity atom for every $10^6$ or $10^8$ pure atoms. The resultant doped or extrinsic semiconductor material can be of two types called N type or P type.

N Type Semiconductor Material

By adding impurity material from the fifth group of the periodic table to intrinsic material, N type semiconductor material is made when the proper procedure is followed. The impurity or donor atoms can be visualized as taking a position Figure 2a, substitutionally in the crystal structure. Note the plus five charge representing the nucleous and all filled electron shells of the group five atom. Associated with this plus five charge are five valence electrons, four in covalent bonds and a fifth loosely coupled to the nucleous. Figure 2b is the same as Figure 1b except that for each impurity atom in the crystal structure an extra electron energy level must be introduced, and these extra impurity levels are found at the top of the forbidden band as pictured. The fact that the extra impurity electrons are loosely bound to the impurity atom means that this electron is easily moved up into the electron band, or away from the effects of the fixed positive charge associated with the impurity atom. Even at relatively low temperatures all of the impurity atoms are ionized, and the conductivity of the
sample of N type material varies only slightly with increasing temperature until temperatures are reached where the concentration of conduction electrons due to the ionized impurities is commensurate with the concentration of thermally generated carriers. Thus over the useful temperature range the conductivity of N type semiconductor material is nearly a constant determined by the concentration of impurity atoms, and conduction is mainly by the majority carrier, conduction electrons, with only a few thermally generated holes adding to the conductivity. In germanium at 300 degrees Kelvin the intrinsic resistivity is 45 ohm - centimeters. At the same temperature the resistivity of an N type sample might be in the order of 0.1 to 1.0 ohm - centimeter. The resistivity of a doped sample is always less than the intrinsic resistivity over the useful operating temperature range.

P Type Semiconductor Material

By adding impurities from group III elements of the periodic chart, P type material can be made. Figure 3a shows a simplified two-dimensional picture of the crystal lattice arrangement. The plus three charge represents the nucleous and all the completed electron shells of an impurity atom. Three of the seven electrons surrounding the nucleous are associated with the impurity atom. This, however, leaves an incomplete covalent bond close to the impurity atom. The electrons of neighboring atoms can easily move into this incomplete bond thereby creating a hole which can move under the influence of an applied electric field. In terms of Figure 3b the incomplete covalent bond may be represented by isolated empty available states just above the top of the valence band into which electrons from
the valence band can move thereby ionizing the impurity or acceptor atom and creating holes in the semiconductor material. The positive holes are free to move, but the negative charges associated with the ionized acceptor atoms are fixed in the lattice structure, causing electrical neutrality in the sample. Over the temperature range of interest the conductivity of the P type material is very nearly constant, and current flow is mainly carried by the majority carriers, holes, with only a relatively few thermally generated hole-conduction electron pairs available for conduction. The realizable resistivities of P type material are the same as for N type material.

Figure 4 shows another way of representing N and P type materials at usable temperatures. For N type material the encircled plus sign indicates the ionized donor atom securely fixed in the crystal lattice structure. There is one of these fixed charges for each of the impurity donor atoms in the material, around $10^{16}$ donor atoms per cubic centimeter. Associated with each of these fixed positive charges is an electron in the conduction band free to take part in electrical conduction. This neutralizes the effect of the fixed charges to give electrical balance. It might be thought that the fixed positive charges would attract the free negative charges, but the effect on the electrons of these charges can be obtained only by using the band theory of solids. These results have been qualitatively discussed previously. Finally in N type material there are some thermally generated hole-conduction electron pairs represented by the plus and minus signs not otherwise accounted for. The concentration of these carriers is less than the concentration of donor atoms in the temperature range of interest.

P type material is also represented in Figure 4, and the fixed charges
associated with each impurity atom are negative. This is because an
electron moves into the available state completing a covalent bond, thereby
creating the negative fixed charge. The carriers in P type material are
mainly holes represented by the plus signs in Figure 4. A free hole is
available for conduction for each of the negative fixed charges. There are
also a few thermally generated hole-conduction electron pairs in the ma­
terial. There are as many positive charges as negative charges in the ma­
terial creating electrical charge balance. Figure 4 represents a useful
simplified way of representing N and P type semiconductor material, but
quantitative information must be obtained from the band theory approach.

High Frequency Effects

In the above discussions a means of visualizing electrical conduction
in intrinsic semiconductor material as well as N and P type semiconductor
material has been developed. Since the interest in this paper is focused
on high frequency effects a discussion of the electrical characteristics at
very high and microwave frequencies of semiconductor material is presented.
Equation 2 relates the conductivity of a material to the carrier mobilities
and carrier concentrations. This equation holds for extrinsic as well as
intrinsic material, but in extrinsic material one or the other of the
carrier concentrations will be very large, thereby controlling the conduc­
tivity. The frequency effects of q, n, and p are nonexistent in the fre­
quency range of interest since q is a constant, and n and p are carrier
concentrations. It is interesting to note that extremely high frequency
electromagnetic energy, frequencies around the optical spectrum for silicon,
can create hole-conduction electron pairs since a photon of such frequency
has enough energy to raise a valence electron to the conduction band if 
the photon is absorbed. This enters into the considerations when deter­
mining the principle of operation of silicon solar cells. Below these 
frequencies the only frequency sensitive components of the electrical 
conductivities are the carrier mobilities $\mu_n$ and $\mu_p$. Over the frequency 
range of interest these are also essentially constants, therefore a sample 
of semiconductor material acts just like a sample of ohmic conducting ma­
terial in the frequency range of interest. It exhibits skin effect at the 
higher frequencies, and this can be computed in the usual way (24). Semi­
conductors differ from conductors in two important ways, one of which is 
important in this study. One effect called the Hall effect is interesting, 
but it has no bearing on the problem in this investigation (28). The im­
portant effect is conductivity modulation.

In a semiconductor material it is possible to introduce extra carriers 
into a sample by various methods. For example if a piece of intrinsic 
germanium is irradiated with light, the conductivity may increase greatly 
due to the light generated hole-conduction electron pairs. These light 
generated carriers increase the $n$ and $p$ in equation 2 and therefore in­
crease the conductivity. If the light source is removed the concentration 
of holes and conduction electrons decreases exponentially to the thermal 
equilibrium value. The time constant associated with the decrease in each 
of the concentrations is called the lifetime of the particular type carrier 
in intrinsic semiconductor material. Thus, the conductivity of intrinsic 
material can be conductivity modulated by shining a light on the sample. 
With P or N type semiconductor there are ways of injecting either majority 
or minority carriers into the sample. These injected carriers can affect
the conductivity of the sample, and when the injecting mechanism is removed there is a finite time during which the effect of the injected carriers is observed. Lifetime of carriers is defined for the particular sample being studied. In intrinsic material light generation of carriers is not the only way to obtain conductivity modulation, but carrier injection can also be utilized. This will be discussed in greater detail later.

In this section discussions concerning the electrical characteristics of semiconductor materials have been presented. The purpose of this presentation is to create a clear picture of these characteristics so that a logical development of the high frequency characteristics of semiconductor junction diodes can be pursued. The next step in this development is the creation of a clear understanding of semiconductor junction diodes.
SEMICONDUCTOR JUNCTION DIODES

Semiconductor junction diodes may be made in a number of ways, and each of the methods results in somewhat different electrical characteristics. PN junction diodes may be fabricated by alloying, growing, or diffusion techniques. PIN junction diodes are fabricated by solid-solid diffusion. Alloying and solid-solid diffusion are very similar with the difference being in the temperature and pressures involved. The alloying and solid-solid diffusion may be envisioned as a migration of impurity atoms into a sample of semiconductor from an impurity solid in contact with the sample. This process can result in a junction. The growing technique creates a PN junction during the growth of a single crystal sample. The junction is created either by introducing impurities into a melt or by changing the rate of growth of the crystal. A PN junction can be created by a gas diffusion process also, where a sample of semiconductor is heated in the presence of a vapor of the proper type of impurity and the impurity migrates into the sample, thereby creating a junction. Each of the methods is used to produce a particularly desirable characteristic. In the investigation of PN junctions, a discussion of the electrical characteristics of the general class of devices can be made using the ideas presented in the previous section. The variation in the electrical characteristics caused by the various fabricating techniques can be pointed out after a general discussion.

PN Junction Diode under Equilibrium Conditions

Figure 4 represents a simplified picture of N and P type semiconductor material. If these two types of material could be brought into perfect
physical contact, there would be a rearrangement of movable charges. This is true because at the moment these materials touch there are more conduction electrons on the N side than the P side, and on the average more electrons move across the PN boundary from the N to the P than from the P to the N. Likewise more holes move from the P to the N than in the other direction. This would cause an unbalance of charges, as pictured in Figure 5b, since the carriers which move across the junction have a good probability of recombining with an opposite type carrier. This situation of charge motion would continue until the unbalance of charges was great enough to cause a built-in potential barrier $V_B$ to be created. Figure 5b shows that the N material becomes positive with respect to the P material, which means the conduction electrons on the N side have difficulty moving over to the P side because they must overcome a retarding potential gradient. For the equilibrium condition the net number of electrons moving from the N to the P type material is zero, and as many holes move from the P to the N as from the N to the P type material giving zero net hole movement across the junction. Therefore the total current flow across the junction for this condition is zero. There is a region at the junction called the depletion region where there are very few carriers, and these carriers experience an electric field which accelerates them. In this region the fixed charges associated with the impurity atoms are uncovered by the previously discussed motion of carriers. These uncovered charges give rise to the barrier potential. The net charge in the semiconductor material is zero maintaining electrical neutrality, but the charges are rearranged in the manner pictured. Figure 5a shows the energy level arrangement across an NP junction under equilibrium conditions. The abscissa
of this curve is distance, in a sense, and the ordinate is electron ener-
gies. The conduction electrons on the N side have trouble climbing the
potential barrier $V_B$, just as the holes on the P side have trouble climbing
down this potential barrier. This is because electrons like to move to the
lower electron energy levels, and holes to higher electron energy levels.
Note that electron energies can be related to potential, but there is a
sign reversal which accounts for the fact that the N side is positive with
respect to the P side. Also the potential $V_B$ is the voltage which gives
the proper electron energy difference in electron volts. Finally, the
Fermi levels of the N and P type materials line up with each other under
equilibrium conditions. The difference in Fermi levels is the external
voltage difference between two points, so a voltmeter connected between the
N and P material under equilibrium conditions will read zero volts rather
than the barrier voltage.

Reverse Biased PN Junction Diode

The simplified picture of a PN junction diode developed in Figure 5b,
and the energy level diagram of Figure 5a give insight as to the charac-
teristics of a semiconductor junction under equilibrium condition. Investi-
gation of figures similar to Figure 5 will give information pertaining to
the operation of PN junction diodes under various biasing conditions.
Figure 6b shows a PN junction diode with an external potential applied to
it. Figure 6a shows the electron energy level diagram with the voltage $V_A$
applied to the junction. This applied voltage tends to make the N material
positive with respect to the P material, which in terms of electron volts
causes the Fermi level on the N side to drop with respect to the Fermi
level on the P side, a distance $V_A$. The electrons on the N side cannot
climb the potential hill $V_A$ plus $V_B$, and the holes on the P side cannot get to the N side. However, a small current does flow through the reverse biased junction due to the thermally generated carriers from both sides of the junction diffusing to the junction and falling through the potential barrier. The current is caused by thermally generated holes from the N side and conduction electrons from the P side. These two components of current in a reverse biased diode compose the saturation current, a very temperature sensitive current of a junction diode. Equation 3 gives the saturation current density as a function of some physical parameters (6).

$$I_s' = I_p + I_n = \left[ \frac{D_p P_0}{L_p} + \frac{D_n N_0}{L_n} \right] q$$

(3)

where $I_p$ and $I_n$ represent the components of current due to thermally generated holes and conduction electrons respectively, $q$ is the electron charge, $D_p$ and $D_n$ are the diffusion constants of holes in the N type material and conduction electrons in the P type material respectively, $L_p$ and $L_n$ are the diffusion lengths for holes in N type and conduction electrons in P type material, and $P_0$ and $N_0$ are the concentrations of thermally generated holes and conduction electrons in the N and P type materials respectively. The saturation current $I_s$ of a PN junction diode is $I_s'$ times the junction area. Thus, when reverse biased, the current through a junction diode is essentially a constant and not related to the applied voltage. To explain this, a consideration of the depletion region is necessary. This leads to the realization that as the reverse bias voltage is varied, the depletion region at the PN junction varies in width. As the width of the depletion layer changes, the number of uncovered fixed charges changes. This allows
for the build up of a voltage across the junction with no steady state
change of current, since the potential caused by the uncovered fixed
charges bucks the externally applied potential. This effect is important
at high frequencies.

Forward Biased PN Junction Diode

Figure 7 shows the electron energy level diagram and a sketch of a
forward biased PN junction diode. Under these conditions the depletion
region of the diode is narrower than for the previously discussed cases.
Figure 7a shows that the applied voltage makes the N side negative. This
causes the Fermi level of the N side to move above the Fermi level on the
P side a distance $V_A$ in electron volts. Now it is apparent that the
electrons on the N side and the holes on the P side have very little trouble
moving over the internal potential barrier, $V_B$ minus $V_A$. The magnitude of
the applied voltage must be less than the built-in barrier voltage to avoid
catastrophic heating caused by heavy current flows. The applied voltage
is in the range of 0.1 to 1.0 volt depending on the semiconductor material
and the current density flowing across the junction. Thus there is a heavy
current flow for a small applied voltage when the diode is forward biased.
The voltage current characteristic of a PN junction diode is sketched in
Figure 8 with the proper directions of applied voltage and current and the
proper symbol for the diode indicated. When the applied voltage makes the
P material positive with respect to the N side, forward bias is assured and
relatively large currents are obtained for a little voltage. When the
applied voltage is negative, the current is essentially a constant, the
saturation current. An equation relating the applied voltage and the
resulting current is given (7):

\[ I = I_S \left[ e^{\left( \frac{qV}{kT} \right)} - 1 \right] \]

where all symbols are as previously defined. At useful temperatures \( \frac{kT}{q} \) is small, about 0.03 electron volt, which indicates that for applied voltages greater than 0.1 volt the current \( I \) is approximately given by equation 5.

\[ I = I_S e^{\left( \frac{qV}{kT} \right)} \]

For applied voltages less than minus 0.1 volt equation 4 is approximately given by equation 6.

\[ I = -I_S \]

Practical Limitations of a PN Diode

Under forward biased conditions the current varies exponentially with the applied voltage. When reverse biased, the current does not change with the applied voltage for an idealized PN junction. Since the semiconductor material is finite in extent and has a finite resistivity, a series ohmic body resistance in a practical diode adds to the voltage drop, especially at high current levels. When a realizable diode is back biased, the current flowing through it does vary with the applied voltage because of the surface leakage paths across the PN junction. Also, at high reverse voltages when the field in the depletion region gets large, the thermally generated carriers passing through the junction are greatly accelerated,
and there is a finite probability that these carriers will collide with and break covalent bonds thereby creating more carriers which are accelerated by the field. The new carriers can get enough energy from the field to break more covalent bonds creating more carriers. Under these conditions an avalanche of carriers is created causing large currents. The reverse applied voltage at which this effect takes place is called the breakdown voltage of the FN junction diode. The breakdown effect is of little interest in this investigation except for secondary considerations.

Variations of Electrical Characteristics Due to the Fabrication Process

Figure 9 shows the relative variation of impurity doping levels as a function of distance for grown and alloyed junction diodes. The ordinate is $N_d - N_a$, the concentration of donor atoms minus the concentration of acceptor atoms. The point where the curves pass through zero is the stoichiometric junction. To the left, where the concentration of donor atoms is greater, the semiconductor is N type material, and where the concentration of acceptor atoms predominates, the material is P type. The relative magnitudes of the impurities is an indication of some of the limitations of particular fabricating techniques. Diffused diodes have impurity concentration curves which usually lie between the limits set by the grown and alloyed type junctions.

The grown junction diode usually has a relatively low impurity concentration in the bulk material, which means the body resistivity of the grown type diode is higher than for the alloy type, and the series body resistance of a grown diode is relatively high. In general this means the current carrying capabilities, which are limited by body heating effects, are
relatively small for a grown type junction. Therefore this type of diode would not be useful as a high current rectifier. On the other hand, the junction is called a graded junction because the variation of impurity level with distance in the vicinity of the junction is relatively small. In this type of diode, fairly large reverse biases can be applied without approaching the breakdown condition because the potential difference is distributed over relatively large distances. Therefore a grown junction diode will have a fairly high breakdown voltage which is desirable in high voltage applications.

An alloy type junction is often called a step junction because the impurity level changes abruptly from an N type to P type. Since the impurity concentrations are high in this type of junction, the series ohmic body resistance is low. Thus high currents can be handled by the alloy junction. On the other hand, a small change in the depletion layer will uncover many fixed impurity charges in the lattice structure. This means the width of the depletion layer for a given voltage is smaller for an alloy diode than for a grown diode. Therefore the electric field intensities across the alloy junction are greater than those across a grown junction. As a result of this, the breakdown voltage of an alloy junction is less than that of a grown junction, and alloy junctions are not very high voltage devices. The diffused junction lies between the extremes outlined above.

**PIN Junction Diode**

For applications where high currents and high voltages are used, neither the alloy nor the grown junction diodes were satisfactory because
of voltage and current limitations respectively. A new type of diode was invented for high power applications (14, 21, 23). This diode is made from a die of very nearly pure intrinsic silicon which has a very high resistivity. A P type and an N type impurity are diffused into the intrinsic material on either side of the die. This makes the PIN or $P^+ \pi N^+$ configuration sketched in Figure 10a. Figure 10b shows the variation of doping level for this diode which is directly proportional to the conductivity as a function of distance through the diode. Thus the N and P regions are high conductivity or low resistivity regions, and the I region is a low conductivity region. With the diode reverse biased very little current will flow through the diode, and the potential difference across the diode will be distributed almost evenly across the nearly intrinsic region which means the diode has a high breakdown voltage. This is desirable, as is a high current rating. When this diode is forward biased the N and P sides offer little resistance to the flow of current. The sandwiched region's resistivity is decreased by the injection of carriers from the N and P sides. This is due to the conductivity modulation effect. Thus when forward biased, the current levels of such a diode can be very high. The dynamic series resistance of the device decreases as the current through it increases because more carriers are injected into the I or $\pi$ region. These devices are used very effectively as low frequency high power rectifiers. For example a Sarkes Tarzian ST 40 x 3 P is rated at 400 volts breakdown and 200 amperes maximum continuous load current with the device being capable of withstanding surge currents as great as 2000 amperes at a temperature of 100 degrees centigrade (26).
Silicon Solar Cell

Silicon solar cells have been devised for a particular application, the conversion of light energy into electrical energy (3, 22). Since this device is also a PN junction diode its characteristics were investigated. Figure 11 shows a sketch of the device, and the impurity level concentration as a function of distance. The device is fabricated from a piece of N type silicon. At elevated temperatures the sample is held in the presence of a boron gas. The gas diffuses into the surface of the sample creating the PN junction shown in Figure 11. The P⁺ and N⁺ contacts are for attaching leads. This junction is fairly close to the surface of the cell so that the impinging photons of light can penetrate into the vicinity of the junction where they create hole-conduction electron pairs which cause the electrical output. For efficient conversion of light to electrical energy the series resistance should be very low so the converted energy is not lost in internal resistance heating. The breakdown voltage for such devices is low since other considerations over-rule any concern for the breakdown voltage. Therefore the device is quite similar to a large area alloy PN junction diode.

This concludes the tutorial discussions of semiconductor junction diodes. The remainder of this material has been independently derived except where references to other work are cited.
Although a diode is inherently a nonlinear device, it is possible to investigate the electrical characteristics of diodes for small signal variations about some operating point using linear circuit techniques. When this is done the diode may be represented by a small signal a-c equivalent circuit. This circuit may be used to compute the a-c currents and voltages in the network including the diode. Thus the analysis of the electrical operating characteristics of semiconductor junction diodes at very high frequencies and small signals is completed when an equivalent circuit is obtained.

Equivalent Circuit of an Ideal Diode

For the purposes of this section, an ideal diode is defined as a diode with a voltage current characteristic as given by equation 4. This equation relates the current through a semiconductor diode to the voltage externally applied to it (27).

\[ I = I_s \left[ e^{\left(\frac{qV}{kT}\right)} - 1 \right] \]  

(4)

where \( I_s \) is the temperature sensitive saturation current given in equation 3, \( kT \) is a voltage equivalent of temperature which is 0.026 volt at room temperature, and \( V \) and \( I \) are the voltage across the diode and the current through it respectively with polarities and current directions defined in Figure 8. The symbol \( V \) is substituted for \( V_A \) for simplification. Equation 5 and 6 represent the V-I characteristic for the forward biased and the reverse biased situations. These equations hold for values of applied voltage greater than 0.1 volt at useful temperatures. Equations 5 and 6
for reverse biases are reproduced below for completeness.

\[ I = I_s e^{\left(\frac{qV}{kT}\right)} \]  \hspace{1cm} (5)

\[ I = -I_s \]  \hspace{1cm} (6)

The ideal diode presents a purely resistive component of impedance to an a-c signal since equation 4 allows for no frequency effects. For a small signal situation the instantaneous voltage applied to the diode will be of the form given in equation 7.

\[ V = V_o + v \cos \omega t \]  \hspace{1cm} (7)

where \( V_o \) is the d-c bias voltage and \( v \) is the peak amplitude of the small a-c signal. For small signal applications \( v \ll V_o \).

Under these conditions equation 5 can be written as equation 8.

\[ I = I_o + \frac{dI}{dV} v \cos \omega t \]  \hspace{1cm} (8)

where \( I \) is the instantaneous current, \( I_o \) is equal to \( I_s e^{\left(\frac{qV_o}{kT}\right)} \), and \( \frac{dI}{dV} \) is a conductance evaluated at \( V \) equal to \( V_o \). Equation 8 is valid for small a-c voltages only since the higher order terms of the Taylor series expansion are neglected. The current \( I \) can be written as shown in equation 9.

\[ I = I_o + i \cos \omega t \]  \hspace{1cm} (9)

where \( i \) is the peak value of the a-c current, and equation 10 gives the relationship between \( i \) and \( v \).

\[ i = \frac{dI}{dV} v = g v \]  \hspace{1cm} (10)
The conductance \( g \) is a small signal a-c conductance relating the a-c current and voltage. The approach outlined above allows the analysis of ideal semiconductor diodes to be performed in two distinct steps. First the d-c bias conditions are analyzed followed by the computation of the a-c operating conditions using the small signal conductance \( g \). The conductance \( g \) for forward biases is a function of the d-c operating point as shown in equation 11.

\[
g = \frac{dI}{dV} \bigg|_{I_0} = \frac{qI_0}{kT} \tag{11}
\]

where all terms have previously been defined. Thus under forward biased conditions, the ideal semiconductor diode presents a conductance, which varies directly with the d-c current flowing through the diode, to a small superimposed a-c signal. At room temperature this conductance is given by equation 12.

\[
g = 38.5 \text{ mhos} \tag{12}
\]

It appears that the conductance \( g \) of an ideal semiconductor diode is in no way related to the geometry of the diode, but the current \( I_0 \) is proportional to the saturation current \( I_s \) which is related to the cross-sectional area of the idealized device. In this manner the conductance is related to the geometry of an ideal semiconductor diode.

When the ideal diode is reverse biased, equation 4 shows that the current is a constant. Since this is true, the small signal a-c conductance is zero. The ideal diode looks like an open circuit to a small applied a-c signal when reverse biased. Therefore the ideal diode looks like a
bias controlled conductance to a small a-c signal when forward biased and an open circuit when reverse biased.

Under bias conditions between the forward and reverse bias, the conditions under which the higher order terms of the Taylor series expansion were neglected may be violated. Thus care must be exercised when analyzing the operation of semiconductor diodes under these bias conditions. For small signals an equivalent impedance can be found which is the slope of the V-I diode characteristic.

The Effect of the PN Junction Depletion Layer

A change in the d-c bias conditions of a non-idealized PN junction diode is accompanied by a change in the depletion layer width, and this change in depletion layer width causes the number of uncovered fixed impurity centers to change. This change is associated with the movement of mobile carriers to the edge of the depletion region, but not across it. The electric field intensity changes across the junction. This change is similar to the change of potential across a capacitance caused by a change in applied voltage. In this manner a PN junction diode depletion layer acts like a capacitance.

When a small a-c signal is applied to a reverse biased PN junction semiconductor diode, the conductance of the ideal junction is zero, but the PN junction presents a capacitive susceptance associated with the depletion layer width variation caused by the small a-c signal. This capacitance may be computed (6). Figure 12 shows the distribution of uncovered fixed charges as a function of distance for a step type PN junction diode. The impurity concentration on the N side is N_d which is less than
the impurity concentration on the P side of the junction. In Figure 12 the integral of the charge density from \(-X_2\) to \(X_1\) should be zero, since the number of uncovered positive charges equals the number of uncovered negative charges. The width of the depletion layer \(X_1\) plus \(X_2\) is a function of the external applied voltage plus the built-in barrier voltage. To obtain the functional relationship, Poisson's equation for the one dimensional case is used (6).

\[
\frac{d^2V}{dx^2} = \frac{qN_a}{\varepsilon} \quad \text{for} \quad x > 0
\]  

(13)

where \(\varepsilon\) is the dielectric constant of the material. Integrating equation 13 gives the electric field intensity as a function of \(x\) for values of \(x\) greater than zero and less than \(X_1\) as given in equation 14.

\[
\frac{dV}{dx} = -E = \frac{qN_a}{\varepsilon} x + C_1
\]

(14)

The boundary conditions are such that the electric field intensity is zero at \(x\) equal to \(X_1\). Therefore equation 15 gives the electric field intensity for \(x\) greater than zero and less than \(X_1\).

\[
-E = \frac{qN_a}{\varepsilon} (x - X_1)
\]

(15)

Doing the same thing for values of \(x\) less than zero results in equation 16 and 17.

\[
-E = -\frac{qN_a}{\varepsilon} x + C_2
\]

(16)

The boundary condition in this case is that \(E\) equal zero at \(x\) equal \(-X_2\).
The potentials to right and the left of the junction can now be obtained by integrating equations 15 and 17 respectively giving equations 18 and 19.

\[
E = -\frac{qN_t}{\varepsilon} (x + x_2)
\]  

(17)

\[
V_R = \frac{qN_a}{\varepsilon} \left[ \frac{x_2^2}{2} - x_1 x \right] + C_3
\]  

(18)

\[
V_L = -\frac{qN_d}{\varepsilon} \left[ \frac{x_2^2}{2} + x_2 x \right] + C_4
\]  

(19)

It is now possible to find the potential of the P material with respect to the N material in terms of \( x_1 \) and \( x_2 \). First define zero potential at the N type terminal. Then using equation 19 it is seen that \( V_L \) the potential at \( x = 0 \) with respect to \( x = -x_2 \) is given by equation 20.

\[
V_L = -\frac{qN_d}{\varepsilon} \frac{x_2^2}{2}
\]  

(20)

Similarly \( V_2 \) the potential at \( x = x_1 \) with respect to \( x = 0 \) can be obtained from equation 18 giving equation 21.

\[
V_2 = -\frac{qN_a}{\varepsilon} \frac{x_1^2}{2}
\]  

(21)

The potential on the P side with respect to the N side is \( V_L \) plus \( V_2 \) given by equation 22 because \( N_a x_1 \) is equal to \( N_d x_2 \).

\[
V_L + V_2 = V + V_8 = -\frac{qN_a}{2 \varepsilon} \left[ 1 + \frac{N_a}{N_d} \right]
\]  

(22)
This potential is the sum of the externally applied voltage plus the built-in barrier voltage. The distance $X_1 + X_2$ is the depletion layer width. As the applied voltage changes, the depletion layer width changes in a very definite ratio determined by the impurity levels on each side of the junction, and there is a voltage dependent change associated with the depletion layer region. A junction per unit area capacitance defined by equation 23 exists.

$$C = \frac{dQ}{d(V_1 + V_2)} = \frac{dQ}{d(X_1 + X_2)} \frac{d(X_1 + X_2)}{d(V_1 + V_2)}$$  \hspace{1cm} (23)$$

The change in charge $dQ$ is the change on either side of the junction, since increasing the number of uncovered charges on one side must be accompanied by an identical change on the other side. Equation 24 gives this change in charge.

$$dQ = -qN_a dX_1 = -qN_d dX_2$$  \hspace{1cm} (24)$$

$$N_a X_1 = N_d X_2$$  \hspace{1cm} (25)$$

Since $X_1$ and $X_2$ are related by equation 25, then using equation 22, equation 26 is obtained which gives the junction capacitance per unit area.

$$C = \frac{dQ}{dX_1} \frac{dX_1}{d(V_1 + V_2)} = \frac{\epsilon}{X_1 (1 + \frac{N_d}{N_a})}$$  \hspace{1cm} (26)$$

Equation 27 may be obtained from equation 22 relating the voltage to the width $X_1$. 
The voltage \( V_1 + V_2 \) is always negative so the resultant width is positive. Substituting equation 27 into equation 26 gives equation 28.

\[
X_1 = \left[ \frac{-\frac{2 \varepsilon (V_1 + V_2)}{q N_a (1 + \frac{N_a}{N_d})}}{2 (V + V_B)(N_a + N_d)} \right]^\frac{1}{2} \tag{27}
\]

Equation 28 gives the junction depletion layer capacitance per unit area as a function of the sum of the applied voltage \( V \) plus the built-in barrier voltage \( V_B \), keeping in mind the polarities of the potentials as shown in Figure 8. Thus a small a-c signal when applied to a PN junction diode sees a capacitive susceptance, and the value of the capacitance is a function of the applied d-c bias voltage since the barrier voltage is a constant. Equation 28 gives the equation relating the per unit area capacitance to the applied voltage for a step type diode.

The grown type of PN junction semiconductor diode has a graded type of impurity distribution in the vicinity of the junction as shown in Figure 9. In this case the relationship between the depletion layer capacitance and the applied d-c voltage plus the built-in barrier voltage differs from that given in equation 28. Figure 13 shows the relationship between the charge density and distance for a graded type junction. Assuming \( -a \) is the slope of the charge density distribution for \( X \) greater than \(-X_1\) and less than \(X_1\) the electric field intensity is given by equation 29 when the boundary condition used is that the electric field equal zero.
at $X$ equal to $X_1$.

$$\frac{dV}{dX} = \frac{-\mathcal{E}}{2\varepsilon} \left( X^2 - X_1^2 \right)$$

(29)

The potential as a function of distance is given in equation 30 with respect to the potential at $X$ equal to $-X_1$.

$$V = \frac{\alpha}{2\varepsilon} \left[ \frac{X^3}{3} - X_1^2 X - \frac{2X_1^3}{3} \right]$$

(30)

The potential at $X_1$ with respect to $-X_1$ is given by equation 31 which shows the potential is negative since the magnitude of $V$ is less than that of the negative built-in barrier voltage.

$$V + V_B = -\frac{2\alpha X_1^3}{3\varepsilon}$$

(31)

The per unit area capacitance is defined by equation 32 for a constant built-in barrier voltage.

$$C = \frac{dQ}{dV} = \frac{dQ}{dX_1} \frac{dX_1}{dV}$$

(32)

The resultant per unit area depletion layer capacitance for a graded PN junction is given by equation 33.

$$C = \varepsilon \frac{2\alpha}{12(V + V_B)} \left[ \frac{a}{12(V + V_B)} \right]^{1/2}$$

(33)

The sum of the applied voltage plus the built-in barrier voltage is always a negative number; therefore $C$ is a real number. Equations 28 and
33 show that the per unit area depletion layer capacitance of a step type junction varies inversely as the square root of the applied voltage, while the depletion layer capacitance of a graded junction varies inversely as the cube root of the applied voltage (6).

Thus it is found that a small signal a-c equivalent circuit of a PN junction diode consists of at least two elements, a bias sensitive conductance \( g \), and a bias sensitive capacitance associated with the depletion layer widening effect. These two elements would appear as parallel elements in an equivalent circuit since the capacitance shunts the conductance. These are not the only elements that need be included in an equivalent circuit because there is a series body resistance and a shunt leakage conductance.

Series Body Resistance and Shunt Leakage Conductance

A PN junction diode small signal a-c equivalent circuit differs from the ideal semiconductor diode equivalent circuit in that a series ohmic body resistance must be included. For a particular diode with constant impurity levels in the N and P materials, equation 3\(^4\) gives the theoretical value of the series ohmic body resistance.

\[
R_s = \frac{\rho_1 L_1}{A_1} + \frac{\rho_2 L_2}{A_2}
\]  

(3\(^4\))

where \( \rho_1 \) and \( \rho_2 \) are the resistivities of the N and P sides, \( L_1 \) and \( L_2 \) are the lengths of the N and P materials, and \( A_1 \) and \( A_2 \) are the cross sectional areas of the N and P sides respectively. The resistivities may be obtained from equation 35.
In some practical cases the series body resistance is very small, and the ohmic contact resistance is appreciable. The ohmic contact resistance is obtained at the point where a non-rectifying contact is made between a semiconductor material and a conducting material. Such contacts are necessary at both ends of a PN junction since the external circuits are composed of conductors. Usually the conductor-semiconductor boundary resistance can be made negligible.

The final element that must be included in a small signal a-c equivalent circuit of a PN junction diode is the leakage conductance across the junction. The value of this conductance is very much related to the fabrication process. Usually a PN diode is etched in an acid solution before final encapsulation. The purpose of this etch is to remove dirt and contamination from the surface of the diode, and the effectiveness of this step will to a large extent determine the leakage conductance of the PN junction diode assuming a good PN junction has been made. Diodes with extremely small leakage conductances can be made.

Small Signal a-c Equivalent Circuit of a PN Junction Diode

The small signal a-c equivalent circuit is shown in Figure 14. The conductance $g$ is the dynamic small signal a-c conductance associated with an ideal semiconductor diode as given by equation 11. This conductance is bias current sensitive. The conductance $g_L$ is the leakage conductance.
across the PN junction. The series resistance \( R_s \) is ohmic resistance as given by equation 34. The capacitance \( C \) is a bias voltage sensitive small signal depletion layer capacity given by equation 28 or equation 33 depending on the type of junction. This circuit represents a PN junction diode small signal a-c equivalent circuit. Each element of the circuit has been developed and related to the fabrication techniques.

PIN Junction Diode Small Signal Equivalent Circuit

Thus far the small signal a-c equivalent circuits of PN junction diodes have been discussed. The PIN diode shown in Figure 10 and discussed previously also has a small signal a-c equivalent circuit. D. A. Kleinman (14) shows that the voltage-current characteristic of a PIN silicon diode is the same as that of the PN junction diode until the current density approaches 200 amperes per square centimeter. Therefore the small signal equivalent circuit for a PIN diode would include a nonlinear conductance \( g \) given in equation 11 just as the ideal PN junction diode does. The series body resistance of a PIN diode is essentially zero because the end terminals of the diode are actually metallic conductors instead of semiconductor material. The leakage conductance of a PIN diode is also essentially zero because leakage paths must extend from the N to the P material around the edge of the intrinsic region, a relatively long path as compared with the leakage path across a PN junction diode. The depletion layer capacitance of a PIN diode can also be small, since the width of the depletion layer is large due to the width of the intrinsic region. However, since the cross section area of a PIN diode is large, tending to cause a large capacitance, the effects may cancel and leave a measurable capacitance. Thus Figure 14
can represent a PIN diode small signal equivalent circuit if $R_g$ is shorted and $G_L$ equals zero. The value of the capacitance $C$ is a function of the width of the intrinsic region, and it will not vary greatly as a function of voltage because in a PIN diode the depletion layer width is a constant equal to the width of the intrinsic region. The resultant equivalent circuit is a small capacitance in parallel with a variable conductance $g$ for the PIN type of diode.

The Effect of Large a-c Signals Applied to Junction Diodes

The operation of semiconductor PN junction diodes when reverse biased at high signal levels has been investigated (31). This treatment shows how the nonlinear capacitance of a PN diode can be used to give low noise gain at microwave frequencies. Large a-c signals are a necessary part of the operation of such amplifiers.

The operation of semiconductor PN and PIN junction diodes at high signal levels and forward bias conditions has not been investigated. However investigations of the switching transients associated with forward biased PN diodes have been made (13). The large signal operation of forward biased junction diodes is closely related to switching considerations because of the minority carrier storage effect. Figure 15a shows a representative switching circuit. With the switch in position 1, a d-c current $I_f$ flows through the diode forward biasing it. At zero time the switch changes to position 2 which tends to back bias the diode. However, due to the minority carrier storage effect, the diode cannot attain a reverse biased condition. Figure 15b shows how the voltage across the diode varies as a function of time. The first phase of the switching transient $T_1$ is
a region where the current flow through the diode is limited by the external resistance in the circuit, and the second phase $T_2$ is where the current decays at a rate determined by the minority carrier lifetime and the dimensions of the diode (13). The point of interest is that there is a finite time $T_1$ where the small signal impedance is zero even though the PN junction diode is presented a voltage that tends to reverse bias it. The time $T_1$ is a function of a number of items, but reasonable values for $T_1$ are between 10 milliseconds and 10 microseconds for forward biased PN diodes.

Consider a forward biased PN junction diode with a large high frequency a-c signal applied to it. Even if the peak value of the a-c signal is many times the d-c bias current, the diode will not necessarily reverse bias for frequencies above 100 megacycles, since the time during which reverse current flows under these conditions is less than 0.01 microsecond. Of course there is a small region between forward and reverse bias where the action of the diode is not so easily explained.

Thus, for a PN junction diode under forward biased conditions, large a-c signals may be applied, and the impedance presented to the signal by the diode does not change as a function of the input power level. In the PIN diode this effect is very pronounced because of the characteristics of conductivity modulation.

**Gas Diffused PN Diodes**

Figure 11a shows the PN junction diode obtained by a gaseous diffusion process. The variation of the net impurity concentration with distance is pictured in Figure 11b. With an external reverse bias voltage applied
to such a junction, it is reasonable to expect three regions to be present in the material as shown in Figure 16. Region I is the P region at the surface of the material with a conductivity determined by the impurity concentration level. The area labeled region II represents a depletion layer region where due to the absence of mobile carriers the conductivity is zero, and there is a static electric field in a direction perpendicular to the surface present in the region. Region III represents the N type region with a conductivity determined by the concentration of donor atoms. An interesting problem to investigate is the impedance presented to an impinging electromagnetic plane wave by the gas diffused junction diode.

Methods of attacking this problem are available (24, Chapter 7). The N type material is an imperfect conductor. The impedance presented to an impinging plane wave by the N material is given by equation 36 since this material is much thicker than the skin depth at the frequency of interest.

\[ \eta_g = (1 + j) R_s \]  

(36)

where \( R_s \) is the surface resistivity or high frequency skin effect resistance per square of the plane conductor of great depth made from the N type material, and its value is given by equation 37.

\[ R_s = \sqrt{\frac{\mu_3 f}{\sigma_3}} \]  

(37)

where \( f \) is the frequency in cycles per second, \( \mu_3 \) is the permeability of the sample, and \( \sigma_3 \) is the conductivity of the sample in mhos per centimeter.

Then the reflection coefficient at the boundary between regions II and III for a plane wave propagating in a direction perpendicular to the surface is
given by equation 38.

\[ \rho_3 = \frac{\eta_3 - \eta_2}{\eta_3 + \eta_2}, \quad \text{where} \quad \eta_2 = \sqrt{\frac{\mu_2}{\epsilon_2}} \tag{38} \]

since region II is assumed to be a perfect dielectric material with permeability \(\mu_2\) and dielectric constant \(\epsilon_2\). The magnitude of the complex reflection coefficient \(\rho_3\) remains a constant throughout region II, but the phase angle changes by an amount which is a function of the distance from the region II-III interface as well as the wavelength of the impinging wave.

Regions II and III present an impedance to region I which can be computed, and is given in equation 39.

\[ \eta = \eta_2 \frac{1 + \rho_3 e^{-2j\beta_2 d_2}}{1 - \rho_3 e^{-2j\beta_2 d_2}} \tag{39} \]

where \(\rho_3\) is given by equation 38 and \(\beta_2\) and \(\eta_2\) are given by equation 40.

\[ \beta_2 = 2 \pi f \sqrt{\mu_2 \epsilon_2}, \quad \eta_2 = \sqrt{\frac{\mu_2}{\epsilon_2}} \tag{40} \]

Thus the impedance terminating region I can be found. Now the problem is to find the impedance presented to an impinging plane wave at the outer surface of the P material. This can also be done since the reflection coefficient \(\rho_2\) at the region I-region II interface is given by equation 41.

\[ \rho_2 = \frac{\eta - \eta_+}{\eta + \eta_+} \tag{41} \]
where \( \eta \) is defined by equation 39, and \( \eta_1 \), the intrinsic impedance of region I, is defined by an equation like equation 36 except the surface resistivity is that of the P type material of region I instead of III. Then equation 42 may be used to determine the input impedance and therefore the reflection coefficient at the surface of the P type material. In this case there is attenuation so this factor must be kept in mind.

\[
\eta_L = \eta_i \frac{1 + \rho e^{-2\gamma d}}{1 - \rho e^{-2\gamma d}} \tag{42}
\]

where \( \gamma \) is the complex propagation constant of the P type material. By using equation 42 in equation 43, the surface reflection coefficient can be found.

\[
\rho_L = \frac{\eta_L - \eta_0}{\eta_L + \eta_0} \tag{43}
\]

where \( \eta_0 \) is the intrinsic or characteristic impedance of air.

It is interesting to note that some control of \( \eta_L \) can be accomplished since it is a function of the depletion layer width \( d_2 \) which in turn is controllable by an externally applied d-c voltage. It is often highly desirable to make the surface reflection coefficient equal to zero in equation 43 because then there would be no component of the impinging electromagnetic wave reflected. Also if the frequency of the impinging wave changed causing a finite \( \rho_L \), it would be desirable to be able to change things in the PN junction diode so as to cause the frequency sensitive reflection coefficient \( \rho_L \) to return to zero. This might be possible in a properly designed gas diffused junction diode since the depletion layer
width is controlled by the externally applied d-c voltage. The design of such a device is presented in the following section treating applications.

Experimental Work

Although it would be highly desirable to fabricate the various types of devices and test and verify each point previously presented, it was not possible to do this. Means of fabricating the devices were not available. Therefore the following approach was taken. Various commercially available devices were purchased and tested. The diodes, in most cases, were encapsulated in such a manner that it was impossible to get to the junction itself. Thus the results of the testing of devices could not be numerically related to the theoretical results already pointed out. However, using the theoretical results, it becomes apparent what type of device would be desirable to perform a particular function at these higher frequencies. Therefore this approach of investigating the diodes is very useful in determining new and unique uses at high frequencies for semiconductor junction diodes.

The frequency range studied was from 100 to 500 megacycles with two exceptions. These exceptions were higher frequencies which were of interest in the study of an electromagnetic wave reflecting from a gas diffused junction diode; and somewhat lower frequencies that were of interest in the case of PIN diodes. One reason for choosing this range of frequencies is that it includes the very high frequency military command broadcast band which extends from 225 to 400 megacycles. Military aircraft use these frequencies to communicate with each other and with installations on the ground. The communication is carried on by a voice frequency, amplitude
modulated, carrier signal with the carrier frequency in the previously mentioned band. Another reason for investigating this frequency range was that little had been done in this area two years ago when this study was undertaken. Since that time, others have been very active doing research especially in the area of variable reactance amplifier applications (31).

The techniques used for measuring impedances of the various diodes are well known. All of the diodes which were to be measured were mounted in series with the center conductor of a 50 ohm coaxial cable with a good short circuit placed a short distance behind the diode. Then the impedance looking into this section of line was measured, using either a Hewlett-Packard model 803A impedance bridge or a slotted line technique. Then the measured impedance was properly rotated on a Smith Chart, and the impedance of the diode was determined. The methods used were straight forward except for the scheme used to introduce bias to the diode. Figure 17 shows in a sketch how d-c bias was applied to the bridge and the diode without allowing either the d-c to be shorted or the radio frequency energy to be radiated in which case the detector might pick it up giving improper readings. The resistor R in the coaxial tee presents a high impedance to the r-f signal as compared with the impedance reflected from the bridge and diode causing the r-f power to go to the diode. Occasionally it was found that if the line lengths were just right the reflected diode impedance at the tee would be very large, resulting in some radiation. When this occurred the insertion of a length of line rectified the problem. The capacitor C was chosen because its a-c impedance was very small. It served to prevent the d-c from shorting out through the loop pick-up of the signal
A useful way to present the measured data is on a $Z-\Theta$ chart where the magnitude and the phase angle of the impedance are plotted as a single point. Such a chart is also a plot of the reflection coefficient which the measured impedance will cause when it terminates a 50 ohm transmission line. The magnitude of the reflection coefficient is obtained by determining the length of the radius vector from the center of the chart to the point in question. The outer edge of the chart represents a reflection coefficient of unity or full reflection. The phase angle of the reflection coefficient is the angle between the positive $x$-axis and the radius vector. Thus from a $Z-\Theta$ chart the impedance of the diode as well as the reflection coefficient of the load can be obtained. This chart is a useful way to present the variation of the impedance of a diode as a function of d-c bias conditions at a single frequency. A second way to present the impedance of a diode is to plot the magnitude of impedance and the phase angle of the impedance as a function of frequency and as a function of bias condition.

Some of the first measurements were made on point contact devices to familiarize personnel with the equipment and techniques as well as to investigate these devices which are as yet unexcelled as microwave rectifiers. Then measurements were made on a graded type junction with a very small cross section. The results of these measurements are plotted in Figure 18. The top curve of Figure 18a shows how the impedance of a reverse biased graded PN junction diode varies with frequency, and the bottom curve of Figure 18b shows how the phase angle of the impedance of the reverse bias diode varies with frequency. It is obvious that the diode looks...
almost like a pure capacitance when reverse biased, since the phase angle is close to \(-90\) degrees and the magnitude of the impedance varies inversely with the frequency. This is expected, since the depletion layer capacitance is the dominant component of the equivalent circuit.

When forward biased to 10 milliamperes the PN junction diode looks like a small impedance, about 10 ohms at all frequencies, while the phase angle of this impedance varies from about 20 to 45 degrees as shown in Figure 18. The positive phase angle of this impedance can to some extent be attributed to the physical length of the diode, but this cannot explain such large angles. It was felt that the diode caused a small discontinuity in the line so that for reverse bias the effect of the discontinuity was not observable, but for forward bias the effect of the discontinuity was to introduce some phase angle. The small a-c resistance when forward biased is \(R_s\) plus the reciprocal of \(g\) from Figure 14. Figure 18 shows that a small area graded junction diode when forward biased has a low impedance, and when reverse biased it looks like a capacitor.

Figure 19 is a plot of how the capacitance of this diode varies as a function of applied reverse bias at a frequency of 100 megacycles. Note that the capacitance of this diode varies nearly inversely as the cube root of the applied voltage as previously predicted. At the lower voltages the phase angle of the diode impedance became rather poor, and it was difficult to obtain the values of capacitance. The predicted results pertaining to a small signal a-c equivalent circuit for a graded type junction were realized to a large extent.

Alloy junction diodes were also investigated, and the results compared favorably with the predicted results. Hughes silicon diodes type 1N459
were measured. Figure 20 is a plot on a Z-Θ chart of the impedance of two of these diodes in parallel as a function of applied bias conditions at 200, 300, and 400 megacycles. It is interesting to note the manner in which the impedance of two of these diodes in parallel varies over a wide range as the bias applied to the diodes changes. A large number of commercial diodes were measured, and this type of diode had a fairly large variation of impedance. In the discussion of applications it will become apparent that a wide variation of diode impedance is useful.

At 200 megacycles with a reverse bias of 175 volts these diodes present an impedance whose magnitude is 190 ohms and phase angle is -90 degrees. As the bias voltage is decreased the magnitude of the impedance decreases while the phase angle stays constant until the voltage becomes 30 volts. At this voltage the phase angle decreases rapidly and the impedance tends to become resistive in nature. Therefore for reverse biases greater than 30 volts the diode acts like a nonlinear capacitance. In this range of voltages \( G_L, R_s, \) and \( g \) in Figure 14 are negligible. Below 30 volts the phase angle of the impedance increases from -90 to -70 with no bias applied, while the magnitude of the impedance changes from 125 to 80 ohms. In the bias region around the zero bias position, the impedance changes considerably with a bias change as is expected. At a 10 milliampere forward bias at 200 megacycles the impedance is 41 ohms at an angle of 10 degrees. At 60 milliampere forward bias the impedance is 17 ohms at an angle of 77 degrees. At this forward bias the conductance \( g \) of the equivalent circuit is a predominant effect along with the discontinuities in the transmission line caused by the diode, diode holder, and encapsulation material.
Figure 20 also shows curves for the impedance of these diodes at frequencies of 300 and 400 megacycles. These curves exhibit characteristics similar to the 200 megacycle curve. If these diodes are to be used as nonlinear capacitances at these frequencies, the diodes must be reverse biased to at least 30 volts.

The PIN junction diodes were rather thoroughly investigated because of a particular application possibility. The diodes tested were Sarkes Tarzian M 500 PIN diodes designed for application as power rectifiers for use at 60 and 400 cycles. The diode is rated at 600 milliamperes and 400 volts. Figure 21a shows the impedance of an M 500 versus frequency for various biasing conditions. The diodes were removed from their cases, and measurements were made in the same manner as for the PN junction diodes. Note that for the no bias and reverse bias conditions, this relatively large cross section area diode had a high value of impedance, greater than 120 ohms, over the entire operating range. Figure 21b shows that the phase angle of the impedance for these operating conditions is somewhat capacitive, indicating that the depletion layer capacity is not zero. By comparing the magnitude of the impedance versus frequency curve for 50 volts reverse bias with the same curve of Figure 18a it is obvious that the PN junction impedance drops much faster with frequency and the small signal circuit is much closer to a pure capacitance. Under forward bias conditions the PIN diode offered very little impedance, less than five ohms when forward biased to 50 milliamperes d-c current with the phase angle of this impedance in the order of 45 degrees.

To test the minority carrier storage effect in these PIN diodes the circuit shown in Figure 22a was arranged. Figure 22b shows the square
wave voltage applied to the diode. During the positive portion of the square wave cycle current I flows through the diode causing the conductivity modulation effect by injecting carriers into the nearly intrinsic region. When the square wave voltage reverses, the diode does not instantaneously reverse bias, but a current flows through the diode in the reverse direction for a storage time labeled $t$ in Figure 22c. This time corresponds to the time $T_1$ in Figure 15b. The peak value of the reverse current is $I_{pr}$, and it is a function of $I$, the resistor $R$, and the applied voltage. With a 10 milliampere soak current $I$ the value of storage time measured was 10 microseconds. The value of $I_{pr}$ was 9 milliamperes. Therefore this PIN diode will not have a chance to reverse bias if a d-c forward bias current of 30 milliamperes is used, and an a-c signal with peak currents of the order of 600 milliamperes is passed through the diode. The minority carrier storage effect is related to charge movement, and the stored charge is the integral of the reverse current with respect to time. Since the measurements showed that a 10 milliampere soak current caused a charge storage of the order of 25 millimicrocoulombs, a 30 milliampere soak current would store at least this much charge. A 600 milliampere current pulse lasting for a half cycle of a 100 megacycle a-c signal corresponds to a charge of about 0.03 millimicrocoulombs not nearly enough to remove the stored carriers. To test this concept further the PIN diode was placed in the circuit shown in Figure 23a. The R's and C's are merely filters to allow a 5 milliampere forward bias to be applied to the PIN diode. Figure 23b shows a plot of output power versus input power up to 1 watt. One watt of a-c power in a 50 ohm system corresponds to a peak current of 200 milliamperes. The results show that the high a-c power is transmitted
through the forward biased PIN diode without causing the diode to reverse bias during any part of the cycle since this would cause the line to flatten out. Perfect transmission is not realized, since the PIN diode has a finite impedance when biased to 5 milliamperes as shown in Figure 21a. This is the reason the power out is not equal to the power in.

It was found that the equivalent shunt capacitance of a PIN diode with no applied bias varied with frequency in the manner shown in Figure 24. This means that the a-c equivalent circuit of an M 500 PIN diode is considerably more complicated than the circuit previously proposed. This can be attributed to the particular diode used.

A question of extreme interest is how good can junction diodes for high frequency operation be made. PN junction diodes have been made to produce a non-linear controllable capacitor. Figure 25a shows the impedance of such a diode as a function of bias at a frequency of 1,000 megacycles (31). Figure 25b shows the impedance of a specially designed PIN diode as a function of bias at a frequency of 1,000 megacycles (31). It is apparent that this diode acts like a nonlinear resistance. These diodes are not commercially available.
APPLICATION OF SEMICONDUCTOR JUNCTION DIODES IN THE VERY HIGH FREQUENCY RANGE

Some applications of semiconductor junction diodes will be treated in this section with experimental verification wherever possible. With the exception of parametric amplification, the applications discussed are new with this study.

Variable Reactance Amplifiers

A very important application of a PN junction diode at very high frequencies is as a nonlinear reactance. As previously mentioned a properly constructed PN junction diode looks like a capacitor in which the magnitude of the capacitance changes as the applied voltage changes even at high frequencies. For a good high Q capacitor the losses of the diode must be small, and therefore the series body resistance $R_s$ must be small. A single figure of merit for such a diode is the cutoff frequency defined by equation 44.

$$f_c = \frac{1}{2 \pi R_s C_m}$$

where $C_m$ is the minimum capacitance of the device, and $R_s$ is the series body resistance.

The usual small signal amplifier utilizes a d-c power source for biasing purposes, and there is a conversion of this d-c power into signal power, thereby giving gain. Such amplifiers are usually noisy because they introduce noise at the output that was not present in the input. The noise figure of such an amplifier is the ratio of the total mean square noise voltage developed across the load to the mean square noise voltage.
developed across the load due to the noise at the input. For amplifiers using vacuum tubes and/or transistors a noise figure of the order of 6 decibels in the very high frequency range would be considered good. Using a parametric or variable reactance amplifier, noise figures of the order of 1.0 decibel are possible (9). A treatment of the theory of operation of variable reactance amplifiers will not be presented here since there are many references available (4, 8, 9, 11, 15, 31). It is interesting to note that the application of variable reactance amplifiers at microwave frequencies comes from relatively recent work which was published in 1958, although the basic ideas are old (29, 31).

The material presented here will be the experimental results obtained on two variable reactance amplifiers. Although the frequency range of interest was between 100 and 500 megacycles, an amplifier was constructed which operated at a signal frequency of 10 megacycles. This was done for two reasons. First, to become familiar with the principles of operation; and secondly, a high frequency unit could not be constructed because of the lack of suitable semiconductor diodes. Recently a good diode was obtained, and a second high frequency amplifier has been built.

Figure 26a shows the circuit of a regenerative type of a variable reactance amplifier which was used in this work. A signal generator feeds a 50 ohm transmission line which in turn feeds a load, in this case a vacuum tube voltmeter. A tuned tank circuit is placed in shunt across the line. A split capacitor feed is employed for impedance matching purposes. This signal tank is tuned to 10 megacycles while the power source, called the pump, is tuned to 37 megacycles, and the idler tank is tuned to the difference frequency 27 megacycles. The pump oscillator develops its
voltage across the PN junction diode since the tanks are low impedances to the 37 megacycle signal. This causes the capacitance of the diode to vary with the pump signal. The conductance presented to the signal tank at the signal frequency is a negative conductance \( (10) \). Therefore the Q of the tank is increased and the negative conductance subtracts from the load conductance thereby giving power gain. Figure 26b shows two experimental curves of power gain versus input level. Note the decrease of gain at the higher signal levels. These results are as expected due to the saturation effect. Curve 1 corresponds to an optimized circuit with a high Q idler tank. Figure 27 shows oscilloscope traces of the response of the circuit as a function of frequency with and without pump power applied to the diode. The response in the bottom picture shows the higher Q and improved response resulting from the application of pump power.

A variable reactance or parametric amplifier operable at higher frequencies has also been built. Figure 28a shows the cylindrical cavity used for the high frequency amplifier. The pump power is capacitively coupled from the pump probe to the varactor diode. The cavity is tuned to the signal by means of the movable plunger. The loop type signal probe generates a transverse electromagnetic wave at the signal frequency. The pump frequency is much higher than the signal frequency, and the cavity is also resonant at a higher frequency, the difference between the pump and signal frequencies, which is the idler frequency. Figure 28b shows the gain of this cavity as a function of signal power at three frequencies. The gain is that gain above a direct connection, therefore it is possible to have a loss through the cavity. Frequencies of 329, 342, and 360 megacycles were arbitrarily chosen. At 342 megacycles the results were
best giving a gain of 28 db with an input signal power of $7 \times 10^{-10}$ watts. The noise figure of this amplifier was estimated as 2 or 3 db. At 360 megacycles the curve drops off at lower power levels contrary to expectations. At 329 megacycles the circuit performed properly, but the system was not optimum. The reason for the variation between curves was that the diode changed position with respect to the pump probe because the tuning plunger was moved. This changed the coupling between the pump and the varactor diode. The curves do exhibit similar slopes and tendencies.

Tuning Using PN Junction Diodes

A second application of PN junction diodes is as electronically controllable tuning elements. This application utilizes the nonlinear capacitance of the PN diode to match loads to transmission lines at very high frequencies. It is possible to match loads to transmission lines in a lossless manner by introducing reactance at proper points along the line so as to cancel out the reactance of the load as well as causing the resistive part of the load to reflect a matched resistance to the desired point. A proposed matching system is shown in Figure 29. A load is pictured at the end of the cable. This load might be an antenna, for example. At a particular carrier frequency this antenna impedance is not usually matched to the 50 ohm coaxial cable. Matching is highly desirable since it means maximum power transfer from the 50 ohm source to the load. Single frequency matching is not difficult to do, but as the carrier frequency is changed, the load impedance changes because of the electrical characteristics of the antenna. Therefore it would be very desirable to have an electronic system which would automatically and continually match the
transmission line and the load at any carrier frequency over a wide range of carrier frequencies. Figure 29 represents such a system, since control of matching is possible by changing the d-c voltages applied to the diodes.

Design procedures have been worked out which are important enough to review briefly (17). A matched system for purposes of this study is defined as one where the resultant standing wave ratio on the transmission line after matching is less than 1.5 to 1. The matching network shown in Figure 29 has three variables which must be specified, the distances $d_1$ and $d_2$ and the number of composite stubs. The optimum number of stubs is four. The distances $d_1$ and $d_2$ depend on the frequencies of interest and the diodes used. The voltages $V_1$, $V_2$, ..., $V_n$ are chosen to give the proper values of depletion layer capacitance to reflect the right admittance through the distance $d_1$ to the line so that matching is accomplished. The details of the procedure followed are outlined in the previously mentioned report (17).

**Switching Very High Frequency Signals using PIN Diodes**

The study of the application of PIN junction diodes to specific problems was quite fruitful with considerable experimental verification. Two problems were attacked: one, an antenna switching problem; and two, an electronically controllable shorted stub problem, which can be directly related to the matching problem previously discussed.

The need for a means of switching radio-frequency power in coaxial cable systems arose in the aircraft antenna field. Although the system resulting from the investigation of this antenna switching problem is discussed, it should be remembered that the results are applicable to a
large number of problems and a much wider frequency range, especially to lower frequencies. To avoid generalities the specific results of the antenna switching project will be used. This work has been reported elsewhere because of its wide range of applications (16, 18, 19, 20).

Omnidirectional radiation patterns on high speed aircraft are difficult to obtain with a single antenna. The radiation pattern of two antennas properly placed on an aircraft is very good; however in the region where signals from both antennas are received simultaneously, interference nulls occur. These nulls interfere with reception and are very undesirable. An antenna switching system which would permit the simultaneous use of both antennas by switching from one antenna to the other would result in a more uniform radiation of signal in all directions.

The principle upon which the switching system operation rests is the sampling principle. A restricted but widely used form of this principle states:

"If a signal that is a time-magnitude function is sampled instantaneously at regular intervals, and at a rate slightly higher than twice the highest significant signal frequency, then the samples contain all of the information of the original signal (2)."

In this particular case, the information consists of voice frequencies up to about 4 kilocycles. The sampling rate then needs to be greater than 8 kilocycles. It was found that sampling or switching rates of the order of 10 kilocycles and greater worked well, but the choice of switching frequency is not critical.

Thus, by means of sampling, it is possible to transmit from one
antenna only part of the time. In particular, assuming a 10 kilocycle sampling rate, it is possible to receive a perfectly understandable signal from a single antenna even though the signal is interrupted for 50 microseconds every 100 microseconds. Interference nulls can therefore be completely eliminated by switching the transmitted signal from one antenna to the other at a fast enough rate so as not to interfere with the voice frequencies. The dual antenna system with the switching action gives an omnidirectional radiation pattern, but in the overlap region of reception stationary nulls no longer exist since only one antenna is transmitting at any given time. Precisely the same situation exists during the period of radio reception. The problem of improving the radiation pattern is thus reduced to a problem of creating a suitable switching system.

Communication sets in aircraft are used for voice communications between aircraft as well as ground to aircraft. The communications set used during this investigation was a transmitter-receiver having a 10 watt peak power output rating. This radio set delivers a voice frequency, amplitude modulated signal with the carrier in the radio frequency spectrum between 225 and 400 megacycles.

Figure 30 shows the proposed system in simple block diagram form. Antennas numbered 1 and 2 are connected to the transceiver by means of a 50 ohm coaxial cable system represented by lines on the figure. Figure 31 shows the heart of the system, the switch devices, in a little more detail.

The operation of this system relies on the nonlinear characteristics of the diodes. A square wave generator operating at the switching frequency results in two square waves 180 degrees out of phase being impressed on resistors $R_1$ and $R_2$. The return path for these square wave voltages
is through the transceiver ground return. Resistors $R_1$ and $R_2$ are the 500 ohm resistors pictured on Figure 31. Thus each square wave is applied across a 500 ohm resistor and one semiconductor diode in series. When the voltage applied to resistor $R_1$ is positive, that applied to $R_2$ is negative. The positive voltage forward biases diode $D_1$ of Figure 30, while diode $D_2$ is reverse biased by the negative voltage applied to it. Then if the diodes are properly chosen, the r-f carrier and side band power emanating from the transceiver sees a low impedance path through $D_1$ and a high impedance path through $D_2$. Therefore the power is channeled to antenna number 1. The 500 ohm resistors keep this r-f power from flowing out into the square wave source since 500 ohms is large compared with the 50 ohm antenna load. The value of these resistors is not critical, but should be greater than or equal to 500 ohms. The filter capacitors allow the r-f power to be delivered to the antennas, but block the switching voltages from the antennas.

When the square wave generator reverses potential, diode $D_1$ reverse biases, and $D_2$ becomes forward biased. The r-f power then flows to antenna number 2. The square wave generator frequency controls the switching rate, which should be greater than 10 kilocycles.

When the transceiver is acting as a receiver, precisely the same switching action takes place.

It should be noted that if two aircraft are communicating with each other, and both have this switching system installed, it is necessary to control the switching rate closely. The reason is that if the switching rate of one differs 100 cycles or more from the other, it is possible to get an audible beat note between the switching frequencies. However, it
should not be difficult to control the switching frequency to one part in 100.

Of course the critical items in this proposed system are the diodes. The commercially available PIN diodes previously mentioned, with impedance characteristics as shown in Figure 21, were used. When forward biased they are low impedances due to the conductivity modulation effect. When reverse biased they are not as high impedances as might be desired, but they are sufficiently good because they are in parallel with and larger than the reflected 50 ohm impedance of the other branch.

These silicon diodes are rated for a maximum d-c current of 600 milliamperes. A 10 watt radio frequency signal in a 50 ohm system corresponds to a current of about 450 milliamperes. The diodes are forward biased to around 20 or 50 milliamperes. This would indicate a peak current of 470 or 480 milliamperes flowing in the diode, which is within the ratings of the device. It is interesting to note that instantaneously the current in the diode may flow in the reverse direction. Currents as large as 400 milliamperes may flow for one half of a carrier cycle in the reverse direction; however the diode will not reverse bias because of the carrier storage effect. This allows the diodes to be operated at low d-c currents when forward biased. These diodes therefore meet the power handling requirements.

Since silicon has good temperature characteristics, these diodes should also operate properly over a large temperature range. Measurements have been made at temperatures ranging from -60 to +100 degrees centigrade. The measurements confirm that the variations of standing wave ratio and forward to reverse power ratio are negligible over this temperature range.
Sustained high temperatures do not affect the diode operation. These diodes are very useful, but they are not the best that might be designed.

With the system connected as shown in Figures 30 and 31, the results in Table I for the standing wave ratio and the power loss were obtained. The standing wave ratio is measured while transmitting into the switching tee with one diode forward biased and the other reverse biased with the coaxial cables terminated in 50 ohm loads representing both antennas.

In the frequency range of interest below 400 megacycles the standing wave ratio is 1.2 or less, and the power loss is less than 1.2 decibels. Much of the mismatch and power loss are attributable to the relatively crude filtering employed. Chronologically, measurements were first made on single diodes, then computations based on these measurements were made. These computations were made to determine how the entire system would operate. Then the system was built and measurements made, and the calculated and measured values were very close. This indicates a basic understanding of the system operation.

Table 1. Standing wave ratio and power loss data

<table>
<thead>
<tr>
<th>Frequency in Megacycles</th>
<th>Voltage Standing Wave Ratio</th>
<th>Power Loss db</th>
</tr>
</thead>
<tbody>
<tr>
<td>200</td>
<td>1.2</td>
<td>1.0</td>
</tr>
<tr>
<td>250</td>
<td>1.1</td>
<td>1.0</td>
</tr>
<tr>
<td>300</td>
<td>1.1</td>
<td>1.0</td>
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<td>1.2</td>
<td>1.2</td>
</tr>
<tr>
<td>450</td>
<td>1.3</td>
<td>1.3</td>
</tr>
<tr>
<td>500</td>
<td>1.4</td>
<td>1.2</td>
</tr>
</tbody>
</table>

The experimental model of the switch was small and rugged giving good results.
The uses of this switch are not limited to this specific application, for it is expected that the switch would operate even more efficiently at lower frequencies. By using different properly designed diodes, very good switches are possible. By arranging a number of these switches in series it should be possible to channel a single source to any one of a number of loads. This constitutes a new improved means of switching very high and ultra high frequency signals, and the band of frequencies being switched can be as wide as is desired, down to frequencies around 100 kilocycles where the carrier storage effect no longer operates and up to as high frequencies as the diode can stand. The switches also can handle high power levels.

An Electronically Controllable Shorted Stub Using PIN Diodes

Another application of PIN junction diodes somewhat related to two of the previous applications came to light as a result of this investigation of the operating characteristics of semiconductor junction diodes. It is possible to match a wide range of loads over a wide range of frequencies by using a triple stub tuner (25).

Such a tuner utilizes three variable shorted stubs properly spaced to do the matching. These stubs can be spaced about an eighth of a wavelength apart at the center frequency of the band of interest. Then almost all impedances can be perfectly matched over a band of frequencies of from 60 to 70 percent of the center frequency. Any particular situation must be studied individually to determine the precise ranges which can be covered.

Thus the matching problem reduces itself to a problem of creating an electronically controllable or variable shorted stub. Then, by using
three of these electronic stubs in a triple stub tuner configuration, it should be possible to match any antenna to a 50 ohm coaxial cable over a frequency range from 225 to 400 megacycles which represents a frequency band of 56 percent of the center frequency.

The electronic stub should have little or no losses associated with it. Also it must be able to withstand relatively large voltages when they are applied to the stub. In other words the voltage standing wave ratio should approach infinity and the current and voltage handling capabilities of the stub should be large. Figure 32 shows a proposed method of controlling the length of the shorted stub in discrete sized steps. If switch \( S_{10} \) is shorted and all other switches are open, the line is \( d_{10} \) long. With \( S_1 \) closed the line is \( d_1 \) long. Therefore the length of the line can be changed by controlling the switches.

A means of switching using PIN diodes has already been proposed. From that concept Figure 33 shows a proposed electronically controllable shorted stub using PIN diodes. When one of the applied voltages is positive and the rest negative, one of the diodes conducts and the radio frequency signal sees a low impedance through the diode and the feed-through capacitor to ground, thereby causing a short on the line. The position of this short can be changed by changing the applied voltages. For instance, with all the diodes reverse biased the line acts like a relatively long shorted stub, while with the first voltage positive the line is short.

The PIN type of diode would be very good for this application since when forward biased it has a very low impedance and good current handling capabilities. However, the M 500 is not a suitable diode because of the relatively low reverse impedance of this diode. The reason this diode
could be used in the previous switching arrangement is that the reverse biased diode had to be a large impedance with respect to a 50 ohm load in that case. In this proposed situation the reverse biased diode must be a very high impedance since it is possible to reflect an open circuit in parallel with the diode impedance. Diodes with the type of impedance as shown in Figure 25b would be suitable, but they are not available for experimentation. Thus experimental verification of the proposed electronically variable shorted stub was carried out using a relatively poor type of diode.

The proposed stub is variable in finite sized steps which necessitates a decision as to how many PIN diodes to use and how to distribute them on a section of line. The answers to these questions are functions of how well the tuning job must be done. A solution to this problem is outlined below.

First determine the highest frequency of interest $f_h$ which in the particular problem outlined is 400 megacycles. Then determine the physical length of a half wave length of the transmission line at this highest frequency as given in equation 45.

$$D_{1} = \frac{\lambda}{2 f_{h}}$$

(45)

where $\lambda$ is the velocity of propagation of the particular transmission line used. Also the physical length of the entire line $D$ can be found by replacing $f_{h}$ by the lowest frequency of interest which is 225 megacycles for the proposed problem in equation 45. In our case with an air dielectric line, $\lambda$ is 30 billion centimeters per second, $D_{1}$ is 37.5 centimeters and
D is 66.7 centimeters. If at the highest frequency it is desirable to break the stub up into n steps in terms of wavelength then the diodes are spaced $D_\perp$ divided by n centimeters apart on the first $D_\perp$ centimeters of the stub. In our case if 10 steps were desired the diodes would be spaced 3.75 centimeters apart which would give 0.05 wavelengths between each diode at 400 megacycles. The spacing of the diodes over the stretch from $D_\perp$ to D can be determined in the following manner. There would be a diode positioned some distance from the shorted end of the stub. This diode would be effective only at the lower end of the frequency band so it should be 0.05 wavelength at 225 megacycles from the end or 6.67 centimeters from the shorted end or 60 centimeters from the input end. The next to the last diode should be 0.1 times 60 centimeters from the last diode or 54 centimeters from the input. Figure 34 shows the spacing between diodes acting as switches to give incremental control of at most 0.05 wavelength at any frequency. In this case 15 diodes are necessary.

These stubs allow finite sized steps in the input impedance of the stub to be obtained. A means of introducing a vernier control on the finite sized steps can be obtained by using a PN junction diode with a bias control in series with the input of the stub. The drawbacks of such a system are the current and voltage handling capabilities of the PN junction diode.

Texas Instrument type 2071 diodes were to fabricate two of these proposed stubs. The standing wave ratios of these stubs were of the order of 3.0 to one. To be perfect the stubs should have an infinite SWR. However the results were good enough to indicate promise for this idea as an electronically tunable element. The measured results are shown in
Figure 35 parts a, b, c, and d corresponding to frequencies of 250, 308, 343, and 400 megacycles respectively. These curves are drawn on admittance coordinates since the stubs are to be used as parallel tuning elements. The various curves are for different frequencies with each of the two stubs shown at each frequency. The points are the input admittances of the stub with one of the diodes forward biased and the rest reverse biased. If the diodes were better switches, the circles on the Smith chart would have larger radii since there would be very little loss in a good switch. The results with the stubs at all frequencies from 250 to 400 megacycles were very similar.

A Proposed Method for Minimizing and Controlling Reflections from a Surface

It is sometimes desirable to control the reflectivity of a surface upon which an electromagnetic wave is impinging. One such case would be in a waveguide where a surface with controllable reflectivity could be utilized as a variable load. A second, perhaps more intriguing case, might be as an external covering of an object which could be made invisible to radar. The latter application is discussed here.

There are methods for minimizing the reflection of impinging electromagnetic energy from a good conducting surface by using a conducting film along with a dielectric layer (24, Section 7.22). This method is good at only one frequency. At the critical frequency the reflections may be negligible, but a change in frequency can cause the reflections to become measurable again. The device proposed in this paper is such that a control of the reflectivity can be utilized to minimize the reflections at any single frequency. Thus a change of frequency can be compensated for, and
the reflections minimized.

The problem is to design a device which will minimize electromagnetic reflections from a surface at any single frequency and to incorporate means of controlling the reflections. In other words, reflections of a plane wave from a conducting surface are to be minimized at a single frequency, and if the frequency of the impinging wave changes, the reflections are again to be minimized by merely varying a voltage.

Figure 36a shows the scheme which minimizes reflections from a perfect conductor (24) Figure 36b shows the transmission line analogy of Figure 36a. In terms of Figure 36b the short circuit is reflected as an open circuit at the impedance $Z_0$. Therefore, the impedance terminating the transmission line is $Z_0$ which causes a reflection coefficient of zero at the load.

In terms of Figure 36a, precisely the same situation exists. The length $l$ is chosen to be a quarter wavelength at the critical frequency. Therefore the impedance reflected to the conducting film from the perfect conductor is infinite. The impedance presented to an impinging plane wave is then the impedance of the conducting film as given by equation 46.

$$Z_i = \eta_2 \coth \gamma_2 d$$  \hspace{1cm} (46)$$

where $\eta_2$ is the intrinsic impedance of the conducting film, $d$ is the width of this film, and $\gamma_2$ is the propagation constant for a plane wave in the conducting film. For $|\gamma_2 d| \ll 1$ equation 46 becomes equation 47, if displacement currents are negligible in the film.

$$Z_i = \frac{\eta_2}{\gamma_2 d} = \frac{l}{\sigma_2 d}$$  \hspace{1cm} (47)$$
where $\sigma_2$ is the conducting film conductivity. Then to be perfectly matched to free space the impedance $Z_4$ must be 377 ohms, which specifies the product $\sigma_2 d$.

This scheme is frequency sensitive since the length $l$ is a quarter wavelength at the critical frequency only. At frequencies other than the critical frequency or odd harmonics of the critical frequency the impedance reflected from the perfect conductor to the conducting film will not be infinite. If the length $l$ in Figure 36a could be controlled this system could be utilized to minimize reflections at any frequency.

It is submitted here that a device somewhat similar to a silicon solar cell could be constructed which would minimize reflections as well as allowing control of the reflections by utilizing the variable width depletion layer as a control mechanism.

Figure 37 shows the cross section of a large area gas-diffused PN junction similar to a silicon solar cell. The region labeled 2 is the P region, region 4 is the N region, and region 3 is the depletion layer or space charge layer. This numbering of regions is similar to that used in Figure 36 for obvious reasons. For a silicon device the dielectric constant is about 12, and the conductivity is a function of the impurity doping concentrations. These concentrations are to be determined. It is assumed that in the depletion layer region the material acts like a good dielectric since there are few carriers in this region. Interest will be focused on frequencies above 5 kilomegacycles and below 500 kilomegacycles. Lower frequencies can be handled by proper design.

Region 2 is a conducting film of P type material whose conductivity $\sigma_2$ and depth $d$ must be chosen as previously mentioned. The product $\sigma_2 d$
should be frequency insensitive. The P type material will meet this con-
dition if the conductivity \( \sigma_2 \) is independent of frequency and if the depth \( d \) is independent of frequency.

The conductivity \( \sigma_2 \) is given by equation 48.

\[
\sigma_2 = q\mu_p \rho
\]

(48)

where \( q \) is the electronic charge and therefore frequency insensitive, \( \mu_p \) is the hole mobility which is frequency insensitive in the range of interest, and \( \rho \) is the concentration of holes which is frequency insensitive also. By doping the P material very strongly the effects of high energy photons, and variations in temperatures can be neglected over wide ranges.

The depth \( d \) is determined by the fabrication process. M. B. Prince (22) shows that boron gas-diffused P layers in silicon of the order of \( 2 \times 10^{-4} \) centimeters are possible. This is a reasonable value for \( d \) since it is small with respect to the wavelength of the frequencies of interest, and therefore would constitute a good conducting film. The depth \( d \) should be frequency insensitive, but the control mechanism proposed in this paper is the variable depletion layer width. Therefore, it is necessary to determine whether the variation of depletion layer width will cause a large change in \( d \).

If the reciprocal of \( \sigma_2 d \) is 377 ohms, and \( d \) is \( 2 \times 10^{-4} \) centimeters, then \( \sigma_2 \) is 13.3 mhos per centimeter. Equation 49 gives the depletion layer widening into the P layer \( W_p \) for reverse biases as a function of the applied voltage \( V_a \), if the acceptor impurity concentration is much larger than donor concentration (6, page 68).
where $\epsilon$ is the dielectric constant and $N_a$ and $N_d$ are the impurity concentrations of the P and N materials respectively. If it is desired that the variation of the depth $d$ be maintained less than 1 percent for applied voltages up to 100 volts, then the ratio of $N_d$ to $N_a^2$ must be less than or equal to $3 \times 10^{-21}$. But $N_d$ is determined by $\sigma_2$ to be $1.65 \times 10^{17}$ atoms per cubic centimeter since $N_a$ equals $p$ in equation 48. Therefore, $N_d$ must be less than $8.16 \times 10^{13}$ atoms per cubic centimeter which is a realizable value, and it meets the requirement that $N_d$ be much less than $N_a$. Thus the P layer design considerations can be realized, and the variation of $d$ with voltage can be maintained negligible.

Two other requirements are to be satisfied by the P layer; that the displacement currents are negligible, and that $|\gamma_2 d| \ll 1$. The displacement currents are negligible if equation 50 holds.

\[
\frac{\sigma_2}{\omega \epsilon} \gg 1 \tag{50}
\]

Let $\sigma_2 = 13.3$ mhos per centimeter, $\epsilon = 12 \times 8.85 \times 10^{-14}$ farads per centimeter, and $\omega = 31.4 \times 10^9$ radians per second, then equation 50 becomes $400 \gg 1$ at 5 KMC and $4 > 1$ at 500 KMC which indicates the analysis is questionable at the higher frequencies. Equation 51 gives the value of the magnitude of $\gamma_2$.

\[
|\gamma_2| = \frac{\sqrt{2}}{\delta} = \sqrt{\pi R \sigma_2 \sigma_2} \tag{51}
\]
Therefore $\gamma_2d$ is of the order of 0.01 at 5 KMC which is much less than unity so this requirement is satisfied. At 500 KMC the requirement appears somewhat tight since $\gamma_2$ is directly proportional to the square root of the frequency. This would limit the high frequency applications somewhat. However, the P layer of the proposed PN junction configuration meets all the necessary requirements of the proposed system over a wide frequency range if $N_a = 1.65 \times 10^{17}$ atoms per cubic centimeter, and if $d$ is $2 \times 10^{-4}$ centimeters.

The width of the depletion layer is given by equation 52 for the case where the impurity doping on the N side is much less than the impurity doping on the P side for a step type PN junction (6).

$$W = \left[ \frac{2e(V_a + V_b)}{qN_d} \right]^{\frac{1}{2}}$$  \hspace{1cm} (52)

where $N_d$ is the impurity donor concentration on the N side, $V_a$ is the applied voltage, and $V_b$ is the barrier voltage. For silicon this becomes equation 53 when the PN junction is reverse biased.

$$W^2 = \frac{1.3 \times 10^7}{N_d} V_a$$  \hspace{1cm} (53)

A relationship between $N_d$ and $V_a$ maximum exists, due to the fact that the breakdown voltage is determined by the N type resistivity. Therefore to determine the maximum width obtainable with a PN junction it is necessary to consider the breakdown voltage of the device.

If the N type material has a resistivity of the order of $5 \times 10^5$ ohm - cm, then the impurity donor concentration is of the order of $10^{10}$ atoms per cubic centimeter which meets the requirements previously mentioned.
Substituting this in equation 53, equation 54 is obtained.

\[ W^2 = 1.3 \times 10^{-3} V_x \]  \hspace{1cm} (54)

A breakdown voltage of the order of 1000 volts is obtained using the almost intrinsic N type material obtained if \( N_d \) is \( 10^{10} \) atoms per cubic centimeter. At 100 volts the width of the depletion layer would be about 0.36 centimeter which represents a quarter wavelength at a frequency of 6.0 kilomegacycles. Thus the assumed donor concentration meets the requirements of the device since the depletion layer width can be made a quarter wavelength at 5 kilomegacycles without exceeding the breakdown voltage.

With a reverse voltage of one volt applied to the junction the barrier width is .036 cm which corresponds to a quarter wavelength at a frequency of 60 kilomegacycles. Thus by changing the applied bias voltage from 1 to 100 volts the depletion layer width changes enough so that the frequency range covered goes from 5 to 60 kilomegacycles. At lower voltages even higher frequencies can be handled. Also since the depletion width need be a quarter wavelength or an odd multiple thereof, the variable aspect need only cover two octaves of frequency and higher frequencies will be handled by the multiplicity of the wavelength. However, it may be better to have the complete control since losses in the dielectric would tend to be troublesome at the higher frequencies.

Thus it would appear on the basis of these calculations that it is possible to create a silicon PN junction whose P layer would correspond to the thin conducting film of Figure 36a, and whose variable depletion layer would correspond to the dielectric of Figure 36a. Now it is
necessary to create a conductor as shown in the figure, and the degree of perfection of this conductor will determine how good the device will be.

The N type material proposed for this junction would not be a good conductor, so a means of lowering the resistivity of the N material must be found. A means of controlling the resistivity of nearly intrinsic material is through conductivity modulation. By flooding the N region outside of the depletion layer with injected carriers, conduction electrons, it should be possible to make the N region close to the space charge region a very good conductor thereby meeting the requirements of the system. In a similar manner the conductivity of the P layer might be controlled, but the complications of such a system may become unduly involved.

To get an idea of how good a conductor the conductivity modulated N region must be, the following computations can be made. The impedance reflected to the conducting film should be infinite ideally, however, if it is large with respect to 377 ohms that is all that is necessary. Assume the reflected impedance is 5000 ohms which is large with respect to the conducting film impedance of 377 ohms. Since the intrinsic impedance of the silicon dielectric is 109 ohms, the per unit impedance of the reflected impedance is 46 + j0. By using a Smith chart it is easily determined that the conductivity modulated N region should present a per unit load impedance of magnitude 0.03 which becomes about 3.3 ohms. Thus the desired surface resistivity of the conductivity modulated N region is found from which its conductivity can be computed to be about 200 mhos per centimeter. At 500 kilomegacycles the conductivity of the material should be 20 mhos per centimeter. These values are within the possible range
using a mechanism of carrier injection.

Figure 38 shows a construction scheme for the proposed device. There are two bias sources in this system, one to control the voltage applied to the junction, and one to control the conductivity of the $N$ region. Voltage $V_1$ controls the width of the depletion layer and voltage $V_2$ controls the conductivity of the $N$ region.

The construction particulars are outlined below. The $N$ type material is to have an impurity concentration of $10^{10}$ atoms per cubic centimeter. The connections to the $N$ material are to be chosen so that conduction electrons are injected from the contacts into the $N$ material. The $P$ layer is to be $2 \times 10^{-4}$ centimeters in depth. The average impurity concentration of the $P$ region is to be $1.65 \times 10^{17}$ atoms per cubic centimeter. The $P$ region must have contacts placed on it to complete the circuit. The fabrication of such a device should not be very difficult if proper equipment is available. Experimental verification of the concepts proposed here is desirable.
ACKNOWLEDGEMENTS

The author is pleased to acknowledge the support of his major professor, Dean George R. Town, and the financial aid of the McDonnell Aircraft Corporation. Also it is recognized that many contributed to this work especially in the construction of systems and the measurement of devices.
REFERENCES CITED


a. Simplified physical picture in two dimensions of intrinsic semiconductor material

Conduction band
Forbidden band
Valence band

Electron energy

b. Band representation of the energy bands of intrinsic semiconductor material

Figure 1. Intrinsic semiconductor material

Conduction band
Forbidden band
Valence band

Electron energies

a. Simplified physical picture of N type semiconductor material

Conduction band
Impurity levels
Forbidden band

Electron energies

b. Band representation of N type semiconductor material

Figure 2. Extrinsic N type semiconductor material

N type material

Figure 3. Extrinsic P type semiconductor material

Conduction band
Forbidden band
Valence band

Electron energies

b. Band representation of P type semiconductor material

Figure 4. Simplified representation of N and P types of semiconductor material with the impurities ionized

P type material
Electron energies
Conduction band
Fermi level
Forbidden band
Valence band

Conduction electrons
Fermi level
Forbidden band
Valence band

Electron energy level diagram for a PN junction diode under equilibrium conditions

Energy level diagram for a PN junction diode under equilibrium conditions

a. Energy level diagram for a PN junction diode under equilibrium conditions
b. Simplified physical picture of a PN junction diode under equilibrium conditions

Figure 5. PN junction diode in equilibrium

Electron energies
Conduction band
Fermi level
Forbidden band
Valence band

Electron energy level diagram for a PN junction diode reverse biased

a. Electron energy level diagram for a PN junction diode reverse biased
b. Simplified physical picture of a reverse biased PN junction diode

Figure 6. Reverse biased PN junction diode

Electron energies
Conduction band
Fermi level
Forbidden band
Valence band

Electron energy level diagram for a PN junction diode forward biased

a. Electron energy level diagram for a PN junction diode forward biased
b. Simplified physical picture of a forward biased PN junction diode

Figure 7. Forward biased PN junction diode

Electron energies
Conduction band
Fermi level
Forbidden band
Valence band

Voltage-current characteristic and voltage and current conventions

a. Voltage-current characteristic and voltage and current conventions
b. Simplified physical picture of a PN junction diode

Figure 8. Voltage-current characteristic and voltage and current conventions

Figure 9. Distribution of impurities in PN junction diodes for different fabricating techniques
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Figure 16. Electromagnetic radiation impinging on a gas diffused diode

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Figure 19. Capacitance of a PN diode as a function of reverse bias at 100 megacycles
Figure 20. Small signal a-c impedance of 2 1 N 459 PN junction diodes in parallel as a function of bias at frequencies of 200, 300, and 400 megacycles.

- The magnitude of the impedance of a PIN diode as a function of frequency for various biasing conditions
- The phase angle of the impedance of a PIN diode

Figure 21. The impedance of a PIN junction diode versus frequency

- Circuit arrangement
- Square wave voltage
- Resultant diode current

Figure 22. Carrier storage effect measurement
a. Measuring circuit

Figure 23. Measurement of the effect of carrier storage

b. Power out versus power in

Figure 24. Shunt capacitance of a PIN junction diode versus frequency

Figure 25. Experimental PN and PIN diodes

a. Circuit diagram of the system used to measure power gain of the 10 megacycle parametric amplifier

Figure 26. The ten megacycle variable reactance amplifier

b. Gain versus input voltage for the 10 megacycle amplifier

Figure 27. At left, the signal tank response with no pump power applied, and at right, the response of the signal tank with pump power applied at the same oscilloscope settings
Signal power in  Pump power in
Signal power out

Vaporator diode
Movable tuning plunger

a. High frequency variable reactance amplifier cavity

b. Gain versus signal power for cavity amplifier

Figure 28. The high frequency variable reactance amplifier

Gain in db

a. High frequency variable reactance amplifier cavity

b. Gain versus signal power for cavity amplifier

Figure 28. The high frequency variable reactance amplifier

Gain in db

Figure 29. A matching network

Figure 30. Block diagram of the proposed switching system

Figure 31. Switching system

Figure 32. Proposed shorted stub

Figure 33. Proposed electronically controllable shorted stub

Figure 34. Diode spacing on the proposed shorted stub
Figure 35. Admittance of electronically controllable shorted stubs

Figure 36. Minimizing reflections from a surface at a single frequency

Figure 37. A large area PN junction diode

Figure 38. Device for controlling the reflections from a surface