Errata

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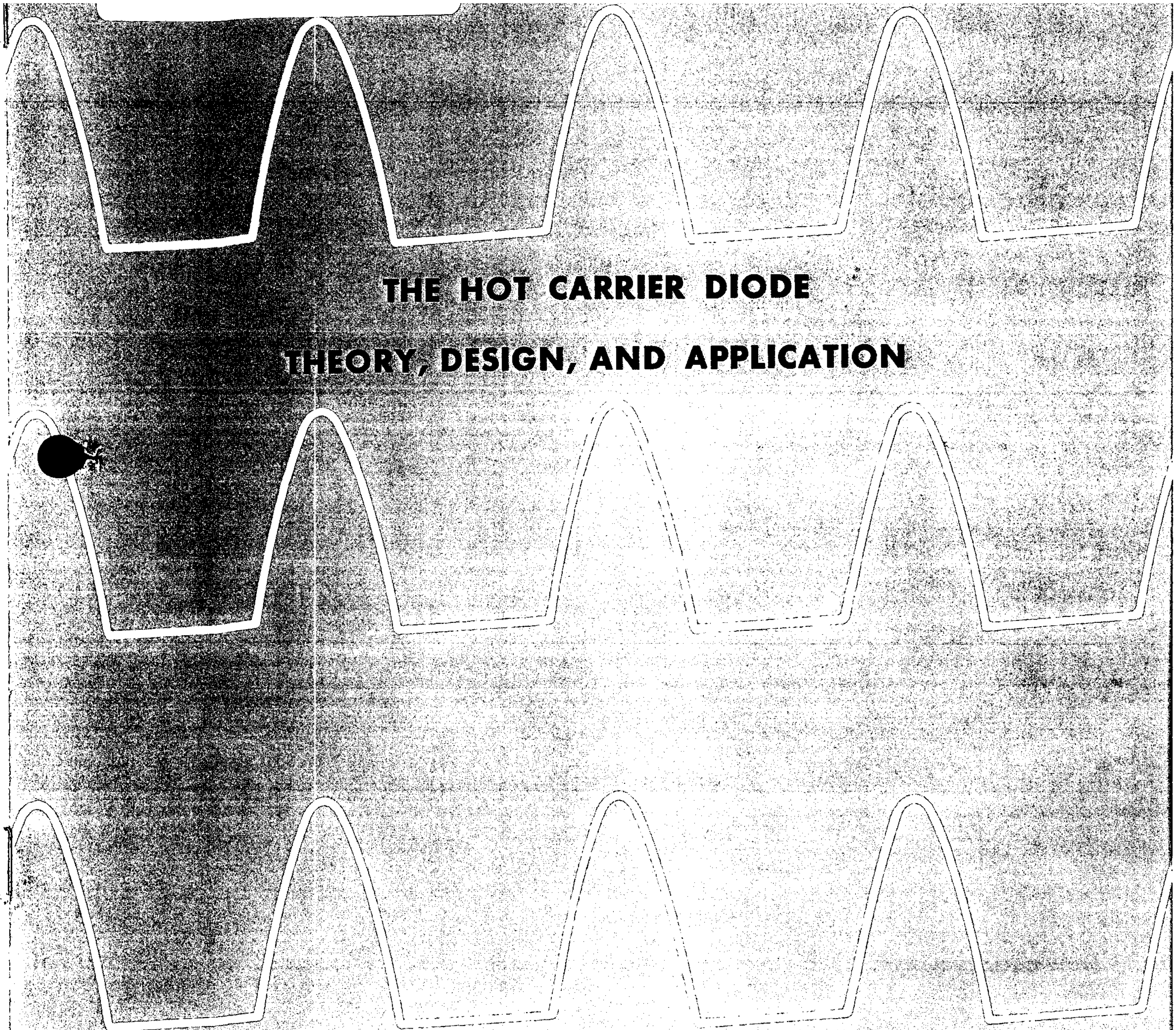

THE HOT CARRIER DIODE

THEORY, DESIGN, AND APPLICATION

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INTRODUCTION

The Hot Carrier diode is a new high frequency and microwave semiconductor device that effectively bridges the gap between the p-n junction diode and the old standby—the point contact diode. It rivals the latter in high frequency performance, and surpasses it in uniformity, reproducibility, and reliability. Unlimited by charge storage phenomena and exhibiting extremely low noise characteristics, the Hot Carrier diode is particularly suitable for fast switching in high frequency computers and as a mixer, detector, and rectifier element extending into the microwave region.

It consists of the practical realization of the theory advanced by Schottky in his work on metal semiconductor devices and is made possible by the vast amount of recent research in the metal-silicon technology; particularly by the availability of very pure semiconductors, by improved techniques of surface cleaning and passivating, and by the epitaxial construction methods. Essentially the Hot Carrier diode is a rectifying metal-semiconductor junction. The metal-semiconductor interface can consist of a variety of metals in conjunction with either n-type or p-type silicon. In general, n-type silicon is preferred because the higher electron mobility permits better high frequency performance. Diodes using evaporated gold, platinum, palladium, silver, and many other metals have been built for a variety of specific applications.

Unlike the p-n junction diode, the Hot Carrier diode is based on majority carrier conduction and in normal operation exhibits virtually no storage of minority carriers. This effect is clearly shown in Figure 1, and results in more efficient rectification at high frequencies. It is similar in concept and in operation to the ideal point contact diode, inasmuch as both employ a Schottky barrier. In practice, however, their characteristics are quite different as can be seen in Figure 2. Whereas the practical point contact diode employs a sharp metal whisker to make contact with the semiconductor element, thereby producing an essentially hemispherical and probably alloyed rectifying junction, the Hot Carrier diode employs a true Schottky barrier consisting of a planar area contact between the metal and the semiconductor element. The planar contact results in a uniform contact potential and uniform current distribution throughout the junction. This results in a lower series resistance, lower noise characteristics, higher power capability, and greater resistance to transient pulse burnout.

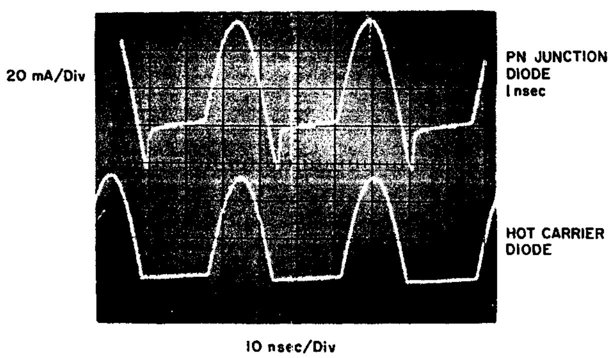

Figure 1. Switching Characteristic of Hot Carrier and P-N Diodes
Top: High Speed P-N Diode with 1 nsec recovery time
Bottom: Hot Carrier Diode
Horizontal: 10 nsec/div.
Vertical: 20 mA/div.
Applied Signal: 30 MHz

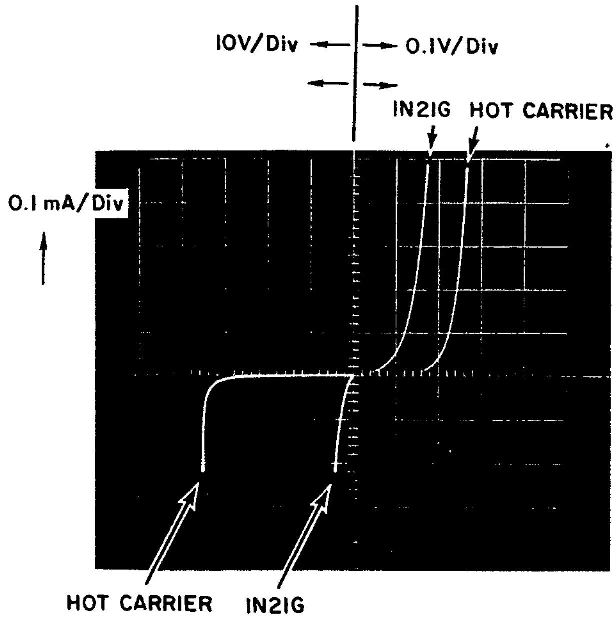

Figure 2. VI Characteristics of Hot Carrier and Point Contact Diodes

THEORY OF OPERATION

The operation of a Hot Carrier diode can be understood most clearly by referring to its appropriate[1] electron energy diagrams shown in Figure 3. These diagrams show the energies of the free electrons in the metal and in the n-type semiconductor under various conditions of bias. At zero bias, Figure 3(a), there must be an electronic equilibrium between the two materials. This is indicated by a constant Fermi level through the metal-semiconductor contact. The bottom of the conduction band in the semiconductor, relative to the Fermi level, can be represented to be at a potential V_ϕ adjacent to the metal and at a potential V_f further in the semiconductor. The potential V_ϕ presents a barrier to the flow of electrons from the metal to the semiconductor and is normally called the barrier potential. It is dependent on the type of metal and semiconductor used, and is also slightly dependent on the reverse bias. The potential V_f is dependent on the doping level in the semiconductor. The difference between V_ϕ and V_f, normally called the built-in or diffusion potential V_b, presents a corresponding barrier to the flow of electrons from the semiconductor to the metal.

The mechanism of electron flow between the two materials is analogous to the *thermionic emission* from a hot cathode into vacuum. The semiconductor in this case acting as the cathode, and the metal as the vacuum. At zero bias there is a constant interchange of electrons between the two materials. The flow of electrons from the semiconductor to the metal results in a current I_s, normally called the diode saturation current. This current is dependent on the area of the junction, the barrier potential and temperature. Since there is an electronic equilibrium the net current across the junction must be zero and there is, therefore, an equal and opposite current $-I_s$ flowing from the metal to the semiconductor.

When the Hot Carrier diode is forward biased (easy current flow) as shown in Figure 3(b), the energy of the electrons in the conduction band of the semiconductor is increased, permitting them to overcome the barrier potential and to be injected into the metal. These electrons are injected with a large kinetic energy, or temperature, compared to the electrons at equilibrium in the metal, hence the name "Hot Carrier."

If the applied external bias is V_e, then the potential barrier on the semiconductor side is reduced by an amount $V_a = V_e - iR_s$ where iR_s represents the voltage drop in the semiconductor substrate resistance, commonly called the diode *series resistance* R_s.

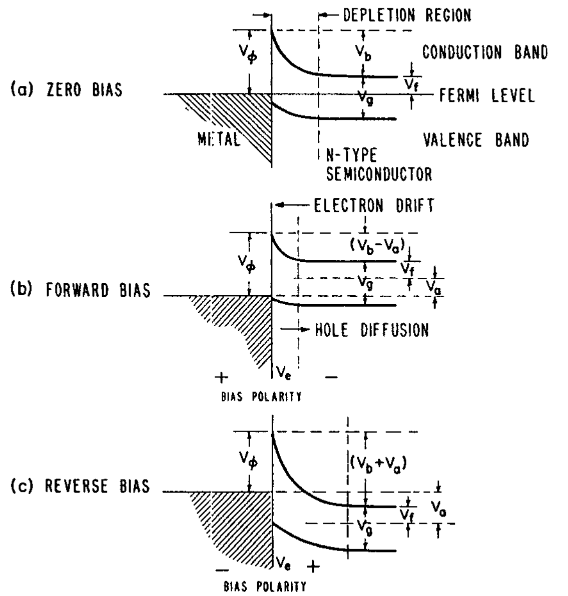

Figure 3. Energy Level Diagram for Hot Carrier Diode

V_ϕ = Barrier Potential. A constant depending on the metal

V_b = Built-in Potential $= V_\phi - V_f$

$$V_f = \frac{kT}{q} \ln \frac{N_c}{N_d}$$

N_c = Effective Density of states in conduction band

N_d = Donor Density

V_g = Gap Potential. A constant depending on semiconductor material

V_a = Applied voltage across the junction

V_e = Applied external bias voltage

This reduction in the barrier results in an increase in the current component from the semiconductor to the metal which is exponentially dependent on the applied bias V_a. The current due to the electron flow from the metal to the semiconductor does not change because the potential V_ϕ on the metal side remains unchanged.

The exponential increase of current continues until finally with large bias voltages the semiconductor conduction band is completely flattened and all the electrons in the conduction band have sufficient energy to flow into the metal. Beyond this point the only mechanism tending to limit the current through the diode is the diode series resistance. This series resistance is essentially ohmic for moderate values of current, resulting in the linear VI characteristics shown in Figure 2 at high bias voltages. The slope of the VI curve beyond this point *should be* equal to the series resistance of the diode. However, it seems to be the nature of semiconductor devices to behave in a more complicated manner. In this case the effective series resistance is modified by the injection of minority carriers (holes) from the metal into the semiconductor substrate. These injected holes have the effect of *reducing* the observed series resistance—a phenomenon which is called *conductivity modulation.* The degree of conductivity modulation depends strongly on the type of metal used to form the junction. Hole injection under forward bias also accounts for the minority carrier storage effects which can be observed when the diode is taken rapidly out of heavy forward conduction. This effect, however, is extremely small compared to the same effect in p-n junction diodes, as seen in Figure 1.

At high current values, corresponding to electron velocities of the order of 10^7 cm/sec which are easily attained in typical Hot Carrier diodes, the series resistance begins to exhibit a non-linear behavior. This is due to electron scattering effects in the lattice structure and is evidenced by a gradual increase in the observed series resistance.

When the Hot Carrier diode is reverse biased, as shown in Figure 3(c), the number of electrons in the conduction band of the semiconductor, having sufficient energy to surmount the barrier, is decreased. A reverse bias of a few tens of millivolts is sufficient to reduce the component of electron current from the semiconductor to the metal to negligible proportions—compared to the component $-I_s$ from the metal to the semiconductor. We therefore expect that the reverse current will be a constant $-I_s$ for all but the smallest reverse voltages. However, the situation is not that simple. In addition to a small *leakage* component of reverse current which behaves in an ohmic fashion, as if a very large resistor were connected in parallel with the diode, a component due to the slight dependence of V_ϕ on applied voltage is also observed. This dependence is extremely small, and is quite negligible for forward bias, but is important for reverse bias. It has been well established experimentally that this phenomenon, called the *image force effect,* is the primary mechanism for reverse current flow in silicon Hot Carrier diodes. The reverse current in the silicon Hot Carrier diode, up to a certain voltage limit, is still many orders of magnitude less than the forward current, despite image force and leakage effects. At the voltage limit referred to above, called the *breakdown voltage,* the current suddenly increases extremely rapidly, often much more rapidly than the onset of heavy conduction in the forward bias direction. *Avalanche multiplication,* as in p-n junction diodes, is usually responsible for this phenomenon.

ELECTRICAL CHARACTERISTICS

Due to the close conformance of the operation of the Hot Carrier diode with theory, its low level VI characteristics can be accurately described by the following equation:

$$i = I_s\left[\left(\exp \frac{qV}{nkT}\right) - 1\right] \tag{1}$$

where I_s is the saturation current, 8×10^{-9} amperes for HPA 2300 series

q is electron charge $= 1.6 \times 10^{-19}$ (coulomb)

T is temperature in degrees Kelvin (°K)

k is Boltzman's constant $= 1.38 \times 10^{-23}$ (joule/°K)

n is diode ideality factor $= 1.05$ for HPA 2300 series

V is the voltage across the diode junction (volts)

Since n is close to unity then at room temperature, ($T = 300°K$), eq(1) can be simplified to:

$$i = I_s\left[\left(\exp \frac{V}{26}\right) - 1\right] \quad (2)$$

where V is in millivolts

The voltage V stated in eq(1) and (2) is the portion of the applied external bias voltage that appears at the diode junction. At dc and low frequencies, this voltage is equal to the previously stated voltage V_a. At microwave frequencies V_a is reduced by the presence of series inductance and junction capacitance. This is clear from the equivalent circuit of the diode shown in Figure 4. The individual values of the circuit parameters are dependent on a specific diode design and in this case are given for the HPA 2300 series diodes.

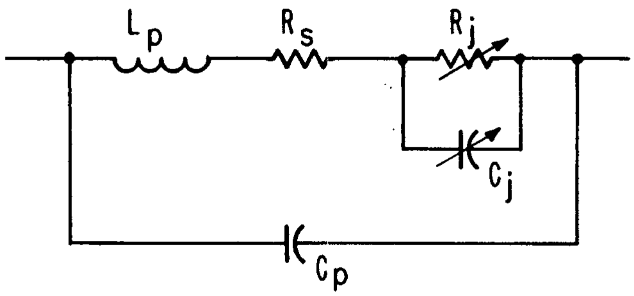

Figure 4. Equivalent Circuit of Hot Carrier Diode

The junction resistance R_j and the junction capacitance C_j are both functions of the current through the diode. The junction resistance can be obtained from eq(2) by differentiating as follows:

$$R_j = \frac{dV}{dI} = \left(\frac{26}{I_s}\right) \exp\left(\frac{-V}{26}\right) \quad (3)$$

For V much greater than 26 or equivalently I much greater than I_s, the following simplification is possible:

$$R_j = \frac{26}{I} \quad (4)$$

where I is in milliamperes

The junction capacitance C_j can be obtained accurately from the depletion layer capacitance expression for a step junction.

$$C_j = \frac{C_{j(0)}}{\left(1 - \frac{V}{V_b}\right)^{1/2}} \quad (5)$$

where $C_{j(0)}$ is the zero bias junction capacitance, typically 0.8 pF
and V_b is the built-in potential ≈ 0.45 electron volts

The series resistance R_s is typically 11 ohms for the HPA 2350. The package capacitance C_p is typically 0.15 pF. The package inductance L_p is typically 3 nanohenries, assuming zero lead length outside the glass envelope.

PHYSICAL AND MECHANICAL CHARACTERISTICS

The physical design and cross-section of the Hot Carrier diode are shown in Figure 5. The diode consists of a heavily doped n^+ silicon substrate approximately 0.020 x 0.020 inches on which an n-type epitaxial layer of a specific resistivity is grown on one surface and an ohmic contact is formed on the opposite surface. Following a thorough cleaning and surface treatment, a matrix of metal "dots" is deposited on the epitaxial surface to form the diode barrier. The type of metal and the geometry of the "dots" is dependent on the desired final diode characteristics—specifically, barrier height, reverse leakage current, series resistance, and barrier capacitance.

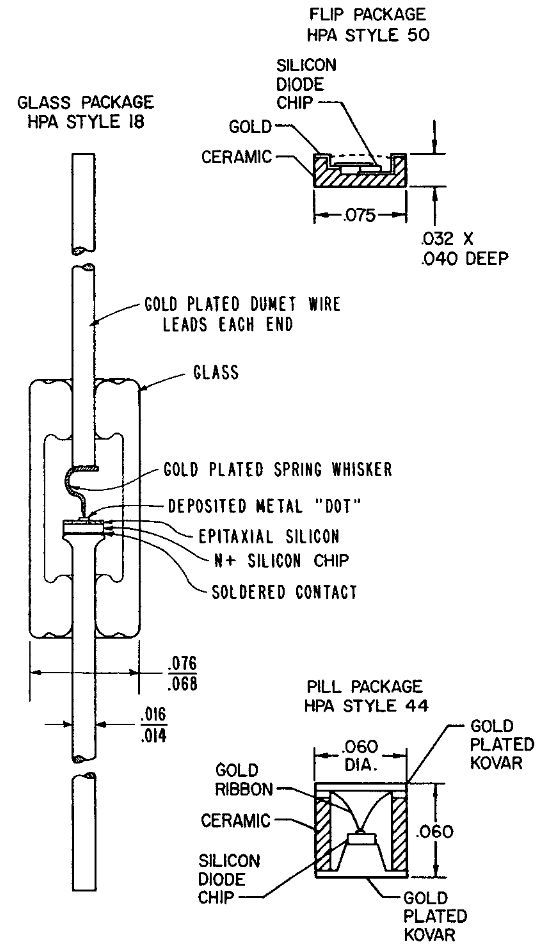

Figure 5. Cross-Sections of Hot Carrier Diodes

This diode chip is then soldered to a pedestal on one of the package leads and an ohmic controlled-pressure contact is made to the top surface of one of the metal "dots" with a gold plated metal whisker. The package is then hermetically sealed in a controlled environment. A matrix of "dots" is deposited to maximize the probability of making contact to a single "dot" on a random try basis. In effect, every diode package contains approximately a gross of potential diodes in addition to the one to which a contact is made.

The overall case size of the diode is kept small, consistent with normal handling and marking requirements. This extremely small size and the use of gold-plated dumet wire leads assure its adaptability to a variety of circuit packaging techniques, particularly printed boards, welded cordwood, and miniature coaxial and strip-line circuits.

The combination of proper materials and precise processing in a controlled environment results in extreme uniformity of diode characteristics and substantial mechanical ruggedness. This has been demonstrated by several thousand device-hours of reliability testing of the Hot Carrier diode in which the failure rate has been found to approach the value of 0.01 failures per million hours at 25°C junction temperature. This is almost three orders of magnitude better than that which is observed[2] for comparable military-type point-contact diodes.

Any reference to a reliability comparison of the Hot Carrier diode with respect to the point-contact type cannot avoid a review of the most common failure mechanism associated with such diodes—that of pulse burnout. The most commonly used term for rating this capability of point-contact diodes is a statement of the energy level (usually in ergs) that is used in the test. Such a simple statement of energy level without reference to the time in which this energy is dissipated and the ultimate effect of this dissipation on the device characteristics is meaningless and does not completely describe the pulse burnout capability of the diode. Degradation is ultimately caused by an excessive temperature rise of the diode junction. When energy is dissipated in a time period that is less than the thermal time constant of the diode (approximately 10 nanoseconds) then the temperature rise of the junction will be a function of the energy level. If, however, the energy is dissipated over a longer period of time, then the device thermal conductivity will diminish the ultimate temperature rise. Under these conditions, the average power dissipation capability of the device applies.

The present test for pulse burnout resistance of the Hot Carrier diode consists of a discharge of three pulses of energy of a specified value through the diode in the forward direction. This energy is obtained from a charged capacitor and the value of the capacitance used is such as to produce a time constant in conjunction with the effective resistance of the diode during the discharge cycle that is no greater than the thermal time constant of the diode. This is the most severe condition of stress and corresponds to the minimum value of energy required to produce burnout.

In addition to this, it is also necessary to define the criterion constituting burnout. In view of the fact that Hot Carrier diodes are intended for both switching and high frequency mixer and detector applications the present criterion for burnout is set as a maximum change of 20% in breakdown voltage. Although this is a suitable criterion for burnout characterization of switching diodes it is not entirely meaningful for mixer and detector applications. In the case of mixer applications, a change in breakdown voltage from a typical value of 30 volts to that of 24 volts does not constitute a burned-out diode since it does not result in an increase in the noise figure of the diode. For mixer application, a noise figure change of 3 dB has commonly been used in the industry as the criterion indicating burnout. Tests to determine the capability of the Hot Carrier diodes with respect to this criterion are currently in progress. There are preliminary indications that the energy levels may be as high as 30 ergs for only a 1 dB change in noise figure.

Results of completed tests, using the 20% change in breakdown voltage as the criterion, and gradually increasing levels of pulse energy are shown in Figure 6. Based on these tests a conservative limit of 5 ergs is currently used as the test level. However, it can be seen that the majority of diodes will withstand levels in excess of 20 ergs.

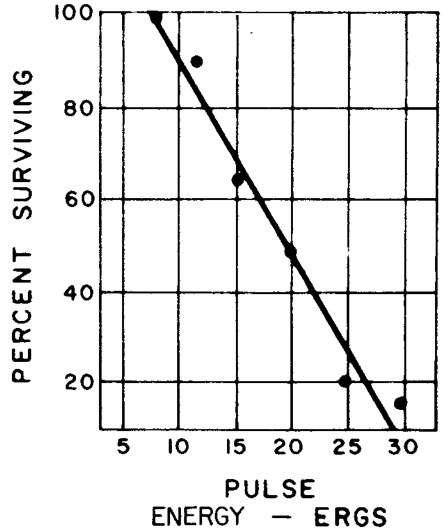

Figure 6. Burnout Resistance of Hot Carrier Diodes

TYPICAL APPLICATIONS OF HOT CARRIER DIODE

Due to its extremely fast turn-on and turn-off characteristics, lack of charge storage, low excess noise, and very uniform forward and reverse characteristics, the Hot Carrier diodes are well suited for a variety of applications extending into the microwave region. In pulse applications they are suitable for fast gating, clamping, sampling, waveform generating, and logarithmic conversion in the fractional nanosecond region. In the microwave area, they are useful as: mixers, detectors, power monitors and rectifiers, limiters, discriminators, harmonic generators, and ultra-fast switches and modulators at nanosecond rates. The excellent uniformity of forward characteristics between diodes permits a relatively easy and economic selection of matched pairs or quads for use in balanced circuit configurations or where accurate tracking between circuits is required. Typical forward matching capabilities are: a forward voltage difference of 20 millivolts maximum between diodes at several current values, and a total capacitance difference of 0.2 pF maximum at zero volts bias. Forward voltage difference of 5 millivolts can be obtained by special selection.

Among these many and varied applications of the Hot Carrier diode, the most interesting, particularly from the point of view of revealing some of the important high frequency attributes of the diode, are those of mixing and detecting. The value of these attributes and their characteristic behavior become most evident in these applications and are consequently covered in greater detail.

MIXER APPLICATIONS

In general the term mixer applies to a circuit that is used for converting an incoming high frequency signal to a signal having a lower frequency. This is accomplished by means of a local oscillator (L.O.) signal and a non-linear resistive element. The L.O. signal and the incoming RF signal are coupled into the non-linear element. Under the influence of the L.O. signal, the non-linear element becomes a time-varying resistance. Under these conditions the diode, insofar as it functions in the linear region, generates a difference frequency, normally called the intermediate frequency (IF), the amplitude of which is proportional to the RF signal amplitude and is independent of the amplitude of the L.O. signal. For a given L.O. frequency, the mixer, insofar as its RF bandwidth permits, will respond equally well to two RF signals—one above and one below the L.O. frequency. One of these is generally referred to as the signal and the other as the image. The non-linear resistive element is usually referred to as the mixer diode. In general the IF can be anywhere from a few hertz to several hundred megahertz. The choice of the IF depends pretty much on the type of information being received and the information bandwidth required for its reception. When the IF is high, for example 30, 60, or 120 MHz, the mixer is generally referred to as a superheterodyne type. When the IF is low, the mixer is usually referred to as the "Zero IF" or "Doppler" type. The latter operation refers to the condition where the transmitted signal also acts as the L.O. and an IF exists only if there is a Doppler frequency shift in the return signal.

The choice of the IF and the characteristics of the diode used in the mixer affect the ultimate sensitivity of the receiving system in which the mixer is used. The sensitivity of any receiving system for a specified signal/noise ratio is:

$$S = -174 + 10 \log B + NF_o + 10 \log (S/N) \qquad (6)$$

where S is the sensitivity of the system in dBm
B is the overall bandwidth of the system in hertz
NF_o is the overall noise figure of the system in dB
S/N is the required signal to noise ratio of the system

The noise-figure of a receiving system consisting of a mixer followed by a high gain IF amplifier, is:

$$\begin{aligned} NF_o &= 10 \log [L_m t_m + (F_{if} - 1)L_m] \\ &= 10 \log L_m(F_{if} + t_m - 1) \qquad (7) \end{aligned}$$

where L_m is the effective mixer conversion loss expressed as a ratio
t_m is the effective mixer noise temperature ratio
F_{if} is the noise factor (a numeric) of the IF amplifier

The terms L_m and t_m, as used in eq(7), are characteristics of the mixer and *are not* the intrinsic conversion loss (L_d) and noise temperature ratio (t_d) of the diode. However, since they are so highly dependent on the latter it is therefore common practice to state these attributes of the diode in *terms* of its performance in a *specific* mixer circuit, and to use the terms L_m and t_m synonymously as characteristics of the diode.

Current practice, and the one used to characterize the Hot Carrier diode, is to measure the noise-figure of the diode in a *Standard* mixer, using a broadband noise source. The Standard mixer is designed to have equal input impedance and conversion loss at the signal and image frequencies. The measurement is made with the IF output of the mixer matched to the IF amplifier input and at a specified L.O. frequency and drive level. The noise figure so obtained is commonly referred to as the double-side-band (DSB) noise figure. This signifies the operation of such a mixer when it is receiving usable signals in both the signal and the image bands. To this measured DSB noise figure a factor of 3 dB is added to convert it to a single-side-band (SSB) value. The latter signifies the noise figure of the same mixer but when it is receiving a useful signal in the signal band only, which is the most common mode of operation. In addition, a correction factor is also added to this value to reflect the effective noise temperature of the broad-band noise source used in the measurement. This correction factor varies with the type of noise source and the degree of match achieved between the noise source and the mixer. Recent calibration tests[3] indicate that the excess noise temperature of some noise tubes, commonly used in coaxial noise sources, is in fact 15.7 dB or 0.5 dB greater than the value of 15.2 dB previously accepted and used in the calibration of automatic noise figure meters. As a result, a correction factor of + 0.5 dB is added to the measured noise figure values of Hot Carrier diodes when stating their noise figure performance as a mixer.

In addition to the noise figure value, the IF source impedance of the mixer and the degree of impedance match (stated in VSWR) that is obtained between the mixer and the L.O. are also stated. These two parameters are measured at the same test conditions as those used for the noise figure measurement, and represent essentially a measure of the IF and RF impedance uniformity of the diode in a specific environment. Typical values of these parameters for presently available Hot Carrier diodes are shown in Table I.

It is obvious that such characterization and specification of the diodes is a useful measure of the mixer performance and uniformity of the diodes, as any in situ test would be. However, these tests leave a great deal unsaid about the intrinsic capabilities of the diode; particularly in other circuit environments.

TABLE I

Typical Mixer Characteristics of Hot Carrier Diodes

HPA Type	Test Frequency GHz	SSB Noise Figure dB	IF Impedance Z_{IF} Ohms	RF Impedance (VSWR)	Series Resistance R_s Ohms	Junction Capacitance $C_{j(0)}$ (pF)
2400	2	6.0	150 - 250	1.5	7 - 11	0.5 - 0.9
2565	3	6.0	100 - 250	1.5	3 - 6	0.3 - 0.7
2511	3	6.0	100 - 250	1.5	3 - 6	0.3 - 0.7
2601	8	6.5	125 - 250	1.5	4 - 7	0.2 - 0.6
2702	9.375	6.5	250 - 350	1.5	4 - 6	0.1 - 0.2

To be able to use the diode and estimate its performance in other mixer configurations, it is necessary to re-examine eq(7). From this expression it is seen that at least one other parameter, either L_m or t_m, must be known. Although either one can be measured, it will be shown that it would be more useful to examine other approaches. It has been shown[4] that t_m is related to the intrinsic noise temperature of the diode t_d by:

$$t_m = \frac{2}{L_m}\left[t_d\left(\frac{L_m}{2} - 1\right) + 1\right] \qquad (8)$$

for the image terminated case
and

$$t_m = \frac{1}{L_m}\left[t_d\,(L_m - 1) + 1\right] \qquad (9)$$

for the image open or shorted case

The diode noise temperature ratio t_d given in this expression is independent of the frequency of operation of the mixer but is dependent on the quiescent bias current through the diode and the IF. The variation of t_d (expressed in dB with respect to unity) with frequency for several values of diode current is shown in Figure 7 for both the Hot Carrier diode and a 1N21G low-noise point-contact diode. From this curve it can be seen that t_d varies inversely with frequency at low frequencies and has a finite and constant minimum value at high frequencies. It can be shown[5] that t_d for the Hot Carrier diode at any frequency can be described accurately by:

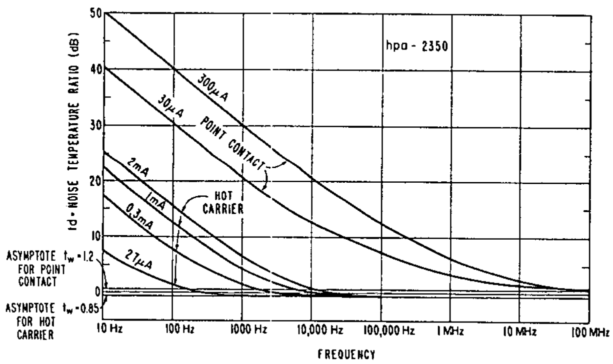

Figure 7. Noise Characteristics of Hot Carrier Diode

$$t_d = t_w + \frac{K_n I_d}{f} \qquad (10)$$

The first term of this equation is the constant term representing the shot noise and thermal noise. The second term represents flicker or 1/f noise. For the average Hot Carrier diode, the constant t_w is typically 0.8 and K_n is typically 1.8 Hz per μA. By integrating over a band, the noise temperature in a given band can be derived and shown to be:

$$t_d = t_w + \frac{K_n I_d}{B} \ln \frac{f_2}{f_1} \qquad (11)$$

where B is the bandwidth $(f_2 - f_1)$
f_2 is the upper frequency
and f_1 is the lower frequency

The noise characteristics of the diode as shown in Figure 7, inasmuch as they are considerably different from those usually observed in point-contact diodes, may require clarification. The reason for the low value of t_w resides primarily in the fact that the diode junction is a planar junction with a uniform contact potential. It has been shown[6] that the average noise current generated by an ideal diode is:

$$\overline{i^2} = 4kTG_jB - 2qI_dB \qquad (12)$$

where k is Boltzman's constant
G_j is the conductance of the barrier $= 1/R_j$
I_d is the diode current
B is the bandwidth

The available noise power from such a junction is given by:

$$P_{na} = \frac{\overline{i^2}}{4G_j} \qquad (13)$$

By substituting the value of G_j given for the Hot Carrier diode eq(3) the resulting available noise power from the Hot Carrier diode is shown to be:

$$P_{na} = kTB\left(1 - \frac{1}{2}\frac{I_d}{(I_d + I_s)}\right) \qquad (14)$$

Since I_d for normal mixer operation is much larger than I_s, I_s can be neglected and it is seen that the available noise power for the Hot Carrier diode approaches one-half the available noise power of a resistor with an equivalent resistance value. The discrepancy between the value of 0.8 given in Figure 7 with that of 0.5 indicated by the above relationship is due to the fact that the diode contains a series resistance, the contribution of which is kTB.

The noise characteristic shown in Figure 7 can be used in conjunction with the proper conversion loss for determining the noise figure of a mixer for any IF frequency and bandwidth. The conversion loss stated in eq(7) is the effective conversion loss realized in a specific mixer. This conversion loss is dependent on several factors and can be considered to consist of the sum of several losses. The first loss is dependent on the degree of match obtained at the RF signal and IF ports. Any departure from an optimum match at either of these ports will result in a loss of RF signal being delivered to the diode, and a loss of available IF signal at the diode being delivered to the IF amplifier. This loss can be expressed as:

$$L_1 = 10 \log \frac{(S_1 + 1)^2}{4S_1} + 10 \log \frac{(S_2 + 1)^2}{4S_2} \qquad (15)$$

where S_1 is the RF VSWR and S_2 is the IF VSWR

The second loss represents a loss of signal power due to the presence of the series resistance (R_s) and junction capacitance C_j in the actual diode. The amount of this loss can be evaluated (see Appendix II) from a consideration of the R_s, C_j, and R_j portion of the diode equivalent circuit shown in Figure 4, and can be expressed as the ratio of the input RF signal power and the power delivered to the junction resistance, or:

$$L_2 = 10 \log \frac{P_{in}}{P_j} = 10 \log \left[1 + \frac{R_s}{R_j} + \omega^2 C_j^2 R_s R_j\right] \qquad (16)$$

The value of R_j used in this expression is the time average value established by the L.O. drive. The minimum value of this loss occurs when R_j is equal to $1/\omega C_j$ and is equal to:

$$L_{2\,(\min)} = 10 \log (1 + 2\omega C_j R_s) \qquad (17)$$

To achieve this minimum at a specific operating frequency, the L.O. drive level must be increased until the required value of R_j is established. This level of drive is the optimum required at that frequency. At 2 GHz it is approximately 1 mW for the HPA 2350. The value of the function in eq(17) increases rapidly for lower L.O. levels and slowly for higher L.O. levels. Consequently the loss due to L_2 is more pronounced at lower L.O. drive levels than it is at higher levels. Lower values of L_2 at low L.O. drive levels can be obtained by introducing a constant dc bias on the diode in the forward direction. Use of dc bias can be beneficial from two points of view. It can be used when a sufficient amount of L.O. drive is not available, or it can be used in conjunction with an intentional decrease in the level of L.O. drive. This is done to decrease the level of noise that is brought in with the L.O. signal. However, care must be exercised in this trade-off not to introduce excessive noise with the dc bias current, as a high level of noise introduced can obviate any real advantage to be gained with the use of dc bias.

The third factor affecting the overall conversion loss is the actual conversion loss of the junction. This loss depends primarily on the forward VI characteristic and partially on the reverse VI characteristic of the diode. If the diode VI characteristic is expressed as:

$$\begin{aligned} i &= K v^{\bar{x}} \quad \text{for } v > 0 \\ &\text{and} \\ i &= g_b v \quad \text{for } v < 0 \end{aligned} \qquad (18)$$

where $\bar{x} = \dfrac{d \log i}{d \log v}$

g_b = the reverse or back conductance of the diode and K = a proportionality factor.

Then it has been shown[7,8] that the minimum available conversion loss that can be realized under various conditions of image termination is a function of x only. For the image terminated case this function is given as:

$$L_{3\,(\min)} = \left(1 + \sqrt{\frac{1 + \gamma_2 - 2\gamma_1^2}{1 + \gamma_2}}\right)\left(\frac{1 + \gamma_2}{\gamma_1^2}\right) \qquad (19)$$

$$\text{where} \quad \gamma_1 = \frac{\left(\dfrac{\bar{x}-1}{2}!\right)^2}{\dfrac{\bar{x}}{2}!\;\dfrac{\bar{x}-2}{2}!} \qquad (20)$$

$$\text{and} \quad \gamma_2 = \frac{\left(\dfrac{\bar{x}-1}{2}!\right)^2}{\dfrac{\bar{x}+1}{2}!\;\dfrac{\bar{x}-3}{2}!}$$

A plot of this function is shown in Figure 8, from which it can be seen that the conversion loss decreases rapidly for increasing values of the diode exponent x. The value of x is the slope of a log-log plot of the diode VI characteristics, and can be determined using the relationship:

$$\bar{x} = \frac{\log \dfrac{i_2}{i_1}}{\log \dfrac{v_2}{v_1}} \qquad \begin{aligned} &\text{where } i_2 > i_1 \\ &\text{and} \quad v_2 > v_1 \end{aligned} \qquad (21)$$

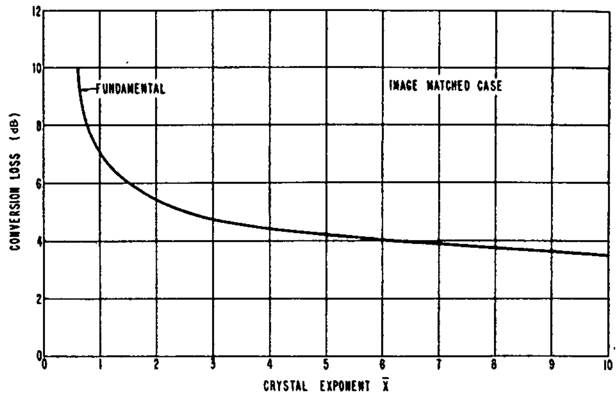

Figure 8. Harmonic Conversion Loss

Such a plot is shown in Figure 9 for both the HPA 2350 Hot Carrier diode and a typical 1N21G point-contact diode. From this plot it can be seen that the exponent $\bar{x}$ is much larger for the Hot Carrier diode (typically 8.8) than for the point-contact diode (typically 4.3), and that it remains high and constant over a larger range of diode current than for the point-contact diode. This indicates that low junction conversion loss can be obtained over a very broad range of L.O. drive. For example, Figure 9 shows that the current range over which $\bar{x}$ is maximum (typically 8.8) for the HPA 2350 is about 0.01 mA to 10 mA, whereas

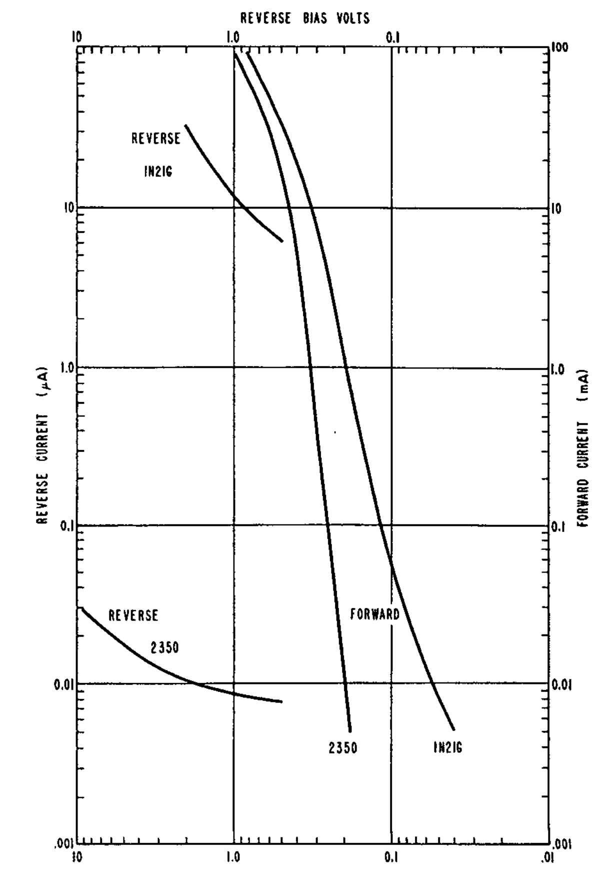

Figure 9. Forward VI Characteristics

the current range over which $\bar{x}$ is maximum (typically 4.3) for the point contact diode is about 0.2 mA to 5 mA. Since diode current is proportional to the L.O. drive level this indicates that the junction conversion loss (L_3) corresponding to this value of $\bar{x}$ remains constant over a very broad range of L.O. drive level.

One implication of this is that a large linear dynamic range can be obtained by operating a mixer at high levels of L.O. drive. The linear dynamic range of a mixer usually means the range of signal power from threshold to some higher value at which a prescribed degree of compression or departure from linearity occurs. The threshold level is determined by the system requirements and mixer noise figure as shown in eq(6). The upper level is determined primarily by the L.O. drive level. For a one dB departure from linearity the upper signal level for the HPA 2350 will be typically 2.5 to 3 dB below the L.O. drive level, and will follow the L.O. level. The dynamic range will therefore increase with the L.O. drive, until it becomes limited by the increasing values of L_2 and L_3 with L.O. drive. For example, at 2 GHz, the noise figure of an HPA 2350 diode at a 20 mW L.O. drive level is typically 0.5 to 0.8 dB higher than it is at the 1 mW level. Consequently an increase of 13 dB in L.O. level results in an increase of 12.5 to 12.2 dB in dynamic range. By contrast the increase in noise figure of a point contact diode is typically 2 dB for the same levels of L.O. power.

Another implication is that low conversion loss can also be obtained at very low levels of L.O. drive. However, operation in this range is restricted by the previously mentioned necessity of maintaining a minimum L_2 loss at the specific operating frequency. It can, however, be obtained through the use of dc bias.

When all of these losses are considered, the following values are obtained for a typical HPA 2350 diode at 2 GHz:

$L_1 = 0.26$ dB for RF VSWR of 1.5/1 and IF VSWR of 1.3/1
$L_2 = 0.9$ dB for $R_s = 11$ ohms and $C_j = 0.8$ pF
$L_3 = 3.9$ dB for $\bar{x} = 8.8$
$L_m = 5.06$ dB (3.2 ratio)

This value of L_m in conjunction with t_w of 0.85 gives a value of $t_m = 0.945$. The resulting mixer noise-figure, for an IF amplifier noise-figure of 1.5 dB, is 6.4 dB. This value corresponds very closely to the measured typical value of 6.5 dB for the HPA 2350 diode. This technique can be used for predicting the expected performance of the diode at other conditions of image termination.

In addition to the characteristics described above, there are two more diode characteristics that have bearing on the practical design of mixers. These are the RF and IF impedances of the diode. Both of these impedances are a function of the L.O. drive.

At typical IF frequencies (30 or 60 MHz) the reactances due to L_p, C_p, and C_j are negligible and consequently the IF impedance is a pure resistance. Typical values of this resistance, at various L.O. drive levels, are given in Figure 10, for HPA 2350 diode. From this data it can be seen that the impedance in the range of optimum L.O. drive of 1 mW is about 180 ohms with a typical range of 150 to 200 ohms. This is about one-half the value normally obtained for point-contact diodes. Since this is in the range of normal impedance levels of low-noise transistor IF amplifiers, a minimum amount of matching is required between the two. This benefit is maximized in the case of balanced mixers where the combined IF impedance of the two diodes is on the order of 90 ohms. In this case practically no matching is required and the two units can be separated, if necessary, using commonly available 93 ohms transmission-line hardware.

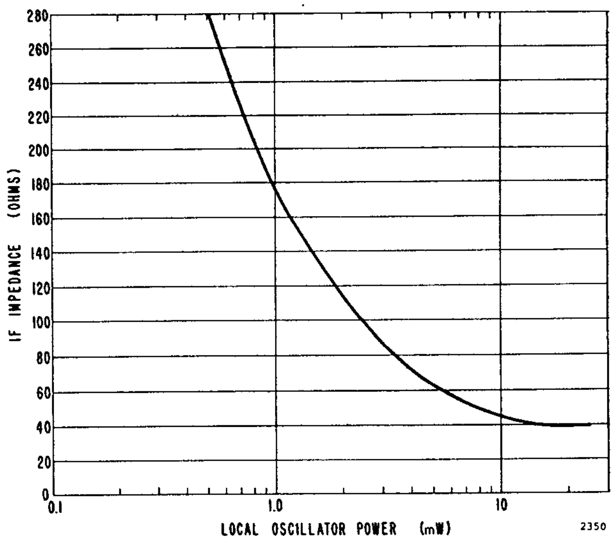

Figure 10. IF Impedance of HPA 2350 Hot Carrier Diode

At the higher RF frequencies, the impedance of the diode as seen by the RF source is influenced by the presence of L_p, C_p, and C_j, and is consequently a complex impedance that depends on the frequency, L.O. drive level, and the output IF load. At a 1 mW signal level and with the diode terminated by a 16 pF RF bypass capacitor, typical RF impedance values for the HPA 2350 in a 7 mm coaxial 50-ohm system are: 41-j42.5 ohms at 2.4 GHz, 52.5-j52.5 ohms at 2 GHz, and 115-j75 ohms at 1 GHz. Generally these values are suitable for a rough first approximation of the required matching structure. In actual practice, the final matching structure is best determined empirically since it is influenced by the actual IF terminating conditions, the geometry of the structure around the diode, the position of the matching structure with respect to the diode and the required signal, image and L.O. terminating conditions.

DETECTOR APPLICATIONS

Detectors are essentially low sensitivity receivers which function on the basis of direct rectification of the RF signal through the use of a non-linear resistive element—a diode. Generally detectors can be classified into two distinct types: the small-signal type, also known as square-law detectors; and the large-signal type, also known as linear or peak detectors.

The small-signal detector operation is dependent on the slope and curvature of the VI characteristic of the diode in the neighborhood of the bias point. The output of the detector is proportional to the power input to the diode, that is, the output voltage (or current) is proportional to the square of the input voltage (or current), hence the term "square law."

The large-signal detector operation is dependent on the slope of the VI characteristic in the linear portion, consequently the diode functions essentially as a switch. In large-signal detection, the diode conducts over a portion of the input cycle and the output current of the diode follows the peaks of the input signal waveform with a linear relationship between the output current and the input voltage.

SMALL-SIGNAL DETECTOR

The most important characteristic of a small-signal detector is its sensitivity. Sensitivity is determined by a number of factors including the RF matching structure, the rectification efficiency of the diode, the output impedance of the diode, the noise properties of the diode, the input impedance of the amplifier, the noise properties of the amplifier, and the bandwidth. The rectification efficiency of the diode is usually stated as either current sensitivity—meaning the ratio of incremental output current to the RF input power, or voltage sensitivity—meaning the ratio of incremental output voltage to the RF input power. The two are interrelated as follows:

$$\beta = \frac{\Delta\, i}{P_{in}} \tag{22}$$

and

$$\gamma = \frac{\Delta\, v}{P_{in}} = \beta R_v \tag{23}$$

where R_v is the dynamic resistance of the diode and is commonly called video impedance or video resistance

By a Taylor expansion of the diode characteristic represented by $i = I_s\,(e_{uv} - 1)$ where $u = q/nkT$, it can be shown (see Appendix I) that the current sensitivity is:

$$\beta = \frac{u}{2}\left[\frac{1 + \left(\frac{Au}{4}\right)^2}{1 + 2\left(\frac{Au}{4}\right)^2}\right] \tag{24}$$

where A is the peak amplitude of the voltage at the diode junction

At low levels of signal the bracketed factor is approximately unity and detection is in the square-law region. A significant decrease in this factor signifies a departure from square-law operation. The maximum signal levels that can be accommodated by the Hot Carrier diodes for departures from square-law by 0.5, 1, and 1.5 dB are shown in Figure 11. Since the ideality factor n of the Hot Carrier diode is 1.05 the low level current sensitivity for these diodes is typically 18.4 $\mu A/\mu W$. For higher values of n, β will be correspondingly smaller. For the point-contact diode the value of n at low levels of signal is typically 1.7-1.9 and the corresponding current sensitivity is 11-10 $\mu A/\mu W$.

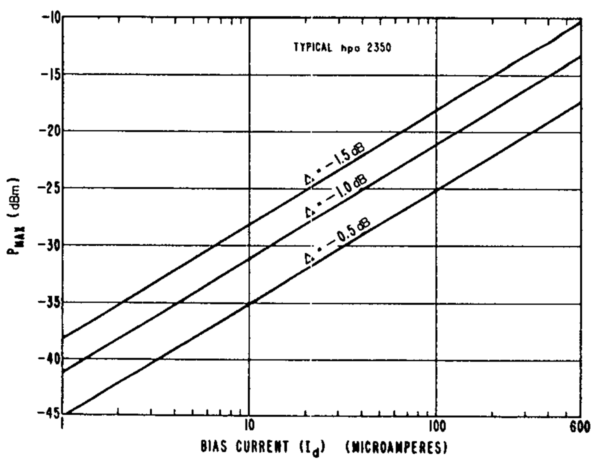

Figure 11. Upper Limit of Square-Law Operation Typical HPA 2350

P_{max} = Maximum signal power input for a departure of Δ (dB) from square law detection

The current sensitivity defined above is based on the junction characteristics of the diode and does not include the effects of the series resistance and the junction capacitance of the diode. These effects can be included by defining a conversion efficiency of the diode as follows:

$$\beta_o = \frac{u}{2}\left(\frac{P_j}{P_{in}}\right) \tag{25}$$

Where the factor P_j/P_{in} represents the ratio of the power delivered to the junction to the input power. It was shown in eq(16) that this factor is dependent on the frequency and the diode bias, and increases rapidly with bias at low bias levels. In normal detector operation, the self bias current is extremely small resulting in low β_o. This situation can be improved by the use of a forward external bias on the diode. The use of external bias has other beneficial effects in that it lowers the RF and the output or video impedance of the diode. The lower RF impedance makes it easier to match the input impedance of the diode to the usually low impedance of the RF source (typically 50 ohms). The lower output impedance permits the use of a lower input impedance in the amplifier and results in an increase in the bandwidth of the output circuit. This is particularly beneficial when low input impedance transistor amplifiers are used in conjunction with the reception of pulse signals. However, as with good food and belly perimeters, there is also a limit as to how much bias can be tolerated. This limit is set by the 1/f noise characteristics of the diode shown in Figure 7 and by the amplifier noise properties. This characteristic indicates that the excess noise generated in the diode will increase with bias. Consequently a trade-off possibility is indicated. The optimum amount of bias needed to assure the best trade-off can be obtained from a consideration of the effect of conversion efficiency and noise generation on the sensitivity of the detector. The sensitivity of a detector is the amount of available signal power (in dBm) that is required to produce a specified signal-to-noise ratio at the output. The various terms currently used for this measure are: The Minimum Detectible Signal (MDS), which corresponds to a signal-to-noise ratio of approximately unity; the Nominal Detectible Signal (NDS), which is *defined* for a signal-to-noise ratio of unity; and the Tangential Sensitivity (TS), which corresponds to a signal-to-noise ratio of approximately 2.5. Although the latter measure is highly subjective and depends upon the operator, it is, however, the one most commonly used by the industry and will be used here.

Under these conditions the ratio of the signal current to the total noise current at the input to the amplifier is equal to 2.5. Using eq(23) and (25), the TS can be stated as (see Appendix IV):

$$TS = \frac{5\, i_n}{u\left(\frac{P_j}{P_{in}}\right)} \tag{26}$$

Where i_n is the total noise current at the input to the amplifier and includes the contributions of: the diode junction given by eq(11), the diode dynamic resistance $R_v = R_j + R_s$, the load resistance R_L and the equivalent noise resistance of the amplifier referred to the input R_a. Using the simplifying assumptions that $R_a/R_L << 1$ and that $R_L > 5\ R_v$ which is usually true, the following expression for i_n can be obtained (see Appendix III):

$$i_n = \sqrt{\frac{4kTB}{R_v}}\sqrt{1 + \frac{R_A}{R_j}} \qquad (27)$$

Where B is the bandwidth of the output circuit, and R_A is an equivalent noise resistance representing the amplifier noise resistance and the flicker noise contribution as shown in Appendix III. Substituting this into eq(26) and rearranging terms, TS can be expressed as follows:

$$TS_{(dBm)} = 10 \log \left[\frac{5\sqrt{4kT}}{u}\ \frac{\sqrt{R_s + R_j}}{R_j}\ \sqrt{\frac{R_j + R_A}{R_j}}\right]$$

$$+ 10 \log \left[1 + \left(\frac{f}{f_c}\right)^2\right] + 5 \log B$$

$$= TS_o + TS_f + 5 \log B \qquad (28)$$

$$\text{Where} \quad f_c = \sqrt{1 + R_s/R_j}\ /\ 2\pi C_j \sqrt{R_s R_j} \qquad (29)$$

The first bracketed term in this expression can be considered as the "zero frequency" tangential sensitivity TS_o. This term is dependent on R_A, which is a constant for a given amplifier and flicker noise coefficient, K_n, and R_j, which is a function of bias current. The variation of TS_o with bias for several values of R_A is shown in Figure 12 and is seen to result in a decrease in sensitivity with increasing bias. The second bracketed term, here defined as TS_f, accounts for the variation in sensitivity as a function of RF frequency and the variation of cut-off frequency with bias. This variation is shown in Figure 13 for several frequencies and is seen to result in an increase in sensitivity with increasing bias. The last term in the expression accounts for the usual variation of sensitivity with bandwidth. The optimum bias current, for any given RF frequency and noise characteristic, R_A, can be obtained by considering the variation of the slopes of TS_o and TS_f with bias. A plot of these slopes is shown in Figure 14. The optimum bias current is found at the intersection of the specified R_A curve with the required frequency curve. The values of TS_o and TS_f corresponding to this bias point are then obtained from Figures 13 and 14 respectively. The final tangential sensitivity for a specified bandwidth is obtained from eq(28). A comparison of the predicted and measured TS of the HPA 2350 diodes at 27 μA bias is shown in Figure 15.

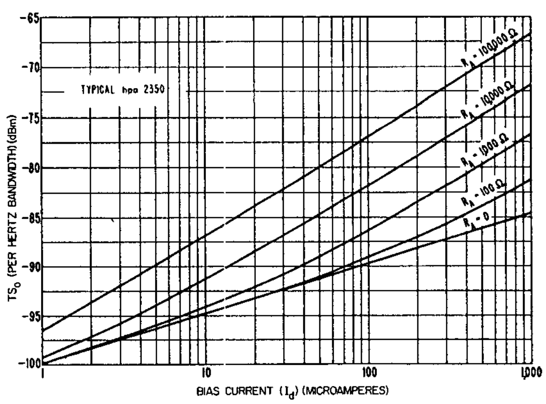

Figure 12. Variation of TS_o with Bias Current Typical HPA 2350

For use in: $TS = TS_o + TS_f + 5 \log B$

R_A = Video amplifier equivalent noise resistance

$$= \frac{(en)^2}{4KTB} + \frac{1.1 \times 10^5}{B} \log \frac{f_2}{f_1}$$

where en = Amplifier noise voltage referred to input
$B = f_2 - f_1$ = Video bandwidth

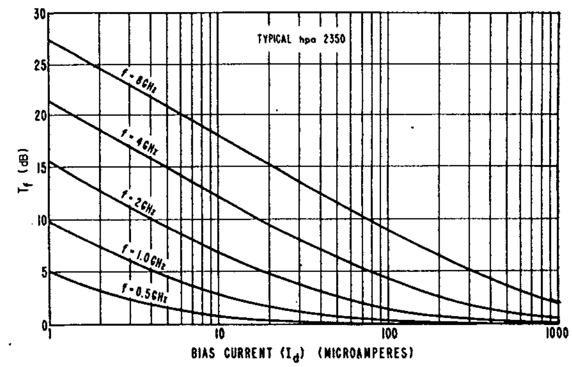

Figure 13. Variation of TS_f with Bias Current Typical HPA 2350
For use in: $TS = TS_o + TS_f + 5 \log B$

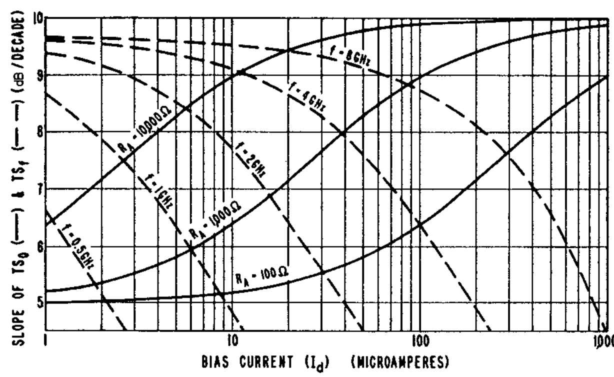

Figure 14. Optimum Bias for Specified Frequency (f) and Amplifier Noise Resistance (R_A) Typical HPA 2350

Note: Intersection of slopes determines optimum bias for a given operating frequency (f) and video amplifier equivalent noise resistance (R_A).

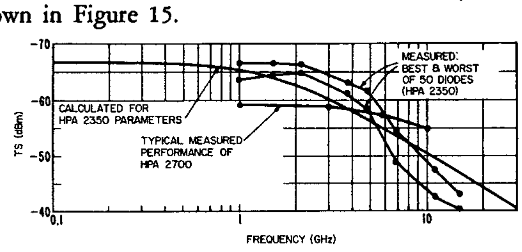

Figure 15. Tangential Sensitivity of Hot Carrier Diode

Test Conditions Bias: 27 μA	HPA 2350	HPA 2700
Amplifier Noise Resistance:	200 ohms	250 ohms
Bandwidth:	100 kHz	500 kHz
Input Resistance:	25 K ohms	2 K ohms
Input Capacitance:	15 pF	20 pF

The output resistance of the diode (R_v) is also established by the choice of bias current inasmuch as it is equal to $R_s + R_j$. For best sensitivity and highest square-law range, the load resistance should be as high as possible relative to the diode output resistance. Therefore, the lower output resistance made possible by the use of bias places less stringent requirements on the input resistance of the amplifier. The output resistance of the diode in conjunction with the capacitance of the output circuit also determines the bandwidth of the output circuit. The lower output resistance of the diode permits the use of higher output capacitance which is needed to maximize the RF power delivered to the diode.

Typical detector performance and parameter values of the HPA 2350 are given in Table II.

TABLE II

Typical[1] Detector Characteristics of HPA 2350 Hot Carrier Diode

Tangential Sensitivity (TS)	— 64 dBm
Video Resistance (R_v)	1000 ohms
Upper Limit of Square-Law	— 27 dBm
Current Sensitivity β	18.4 μA/μW
Noise Corner Frequency	50 Hz
Noise Temperature Ratio @ 1.0 Hz (t_d)	6

NOTE 1: All values are given at 2 GHz with the diode biased at 27 μA and working into an amplifier with a noise resistance of 200 ohms, a bandwidth of 100 kHz, and an input impedance of 25 K ohms shunted by a capacitance of 15 pF.

LARGE-SIGNAL DETECTORS

In large-signal or linear detection, the most important characteristics of the diode are: low series resistance, large linear range of the VI characteristic, high reverse resistance, and high reverse breakdown voltage. The requirements for the RF input circuit for large-signal detectors are the same as for low level detectors and must be designed for a good match or low VSWR. The RF by-pass capacitance at the output of the diode must be high enough to provide a good short at the RF frequency consistent with the RC requirements of the output circuit. The RC time-constant of the output circuit must be sufficiently small to assure that the output voltage follows closely the peaks of the modulated RF signal. The RC time-constant is therefore a function of both the maximum modulation frequency and the index of modulation. Because of the high reverse leakage characteristics of point-contact diodes, the RC time-constant is highly influenced by the reverse resistance of the diode and is dependent on the drive level. The extremely high and constant reverse resistance of the Hot Carrier diodes relieves this restriction and permits the design of the output circuit to be determined on the basis of modulation requirements. The linear-range of the diode must be sufficiently large to assure linear detection at the crest and valley points of the modulated RF waveform as determined by the modulation index. Because of the high levels of drive used in linear detection, the diode is usually substantially back biased. Under these conditions, the ability to operate in the linear region at both the crest and valley points of the modulated waveform is determined by the reverse breakdown voltage. Higher reverse breakdown voltage of the Hot Carrier diode allows operation at higher peak signal voltages thereby assuring linear detection for larger modulation indexes, and/or higher available power levels. For Hot Carrier diodes the minimum RF signal level at which peak detection begins is approximately 0.1 volt RMS, and the upper limit is that at which the peak-to-peak RF voltage equals the reverse breakdown voltage, or $\frac{BV_r}{2\sqrt{2}}$ volts RMS.

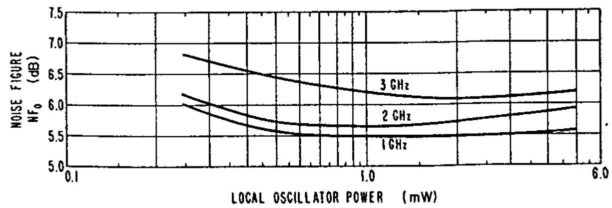

Figure 16. Typical Noise Figure vs. Local Oscillator Power (HPA 2550) $f_{IF} = 30$ MHz. $NF_{IF} = 1.5$ dB

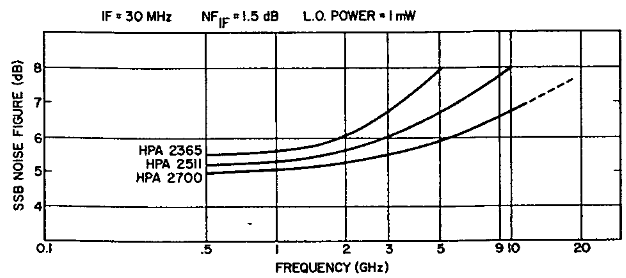

Figure 17. Typical Noise Figure vs. Frequency $P_{LO} = 1$ mW. $f_{IF} = 30$ MHz. $NF_{IF} = 1.5$ dB

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APPENDIX I
Current Sensitivity of Hot Carrier Diode

Any continuous function can be expanded in a Taylor's series of the form:

$$f(x+h) = f(h) + xf'(h) + \frac{x^2}{2}f''(h) + \frac{x^3}{6}f'''(h) + \frac{x^4}{24}f''''(h) + \cdots \quad (1)$$

Ignoring higher order terms, and letting $x = A\cos\omega t$, the incremental result, $f(x+h) - f(h) = \Delta f$, can be written:

$$\Delta f = \left[\frac{A^2}{4}f'' + \frac{A^4}{64}f''''\right] + \left[Af' + \frac{A^3}{8}f'''\right]\cos\omega t + \left[\frac{A^2}{4}f'' + \frac{A^4}{48}f''''\right]\cos 2\omega t + \left[\frac{A^3}{24}f'''\right]\cos 3\omega t + \left[\frac{A^4}{192}f''''\right]\cos 4\omega t \quad (2)$$

Letting $i = f(v)$ and $v = A\cos\omega t$ we can define a power expression as:

$$P = vi^* = [A\cos\omega t]i^* \quad (3)$$

By an application of the trigonometric identity:

$$\cos x \cos nx = \frac{1}{2}[\cos(n+1)x + \cos(n-1)x] \quad (4)$$

It can readily be seen that this has a constant term if and only if n is one. Therefore eq(3) reduces to:

$$P = \frac{A}{2}\left[Af' + \frac{A^3}{8}f'''\right] \quad (5)$$

The only constant term in eq(2) is the first and therefore:

$$\Delta i = \left[\frac{A^2}{4}f'' + \frac{A^4}{64}f''''\right] \quad (6)$$

Current Sensitivity β may then be written in the form:

$$\beta = \frac{\Delta i}{P} = \frac{\frac{A^2}{4}f'' + \frac{A^4}{64}f''''}{\frac{A}{2}\left[Af' + \frac{A^3}{8}f'''\right]} = \frac{1}{2}\frac{f''}{f'}\left[\frac{1 + \frac{A^2}{16}f''''/f''}{1 + \frac{A^2}{8}\frac{f'''}{f'}}\right] \quad (7)$$

Now suppose $i = f(v)$ describes the Hot Carrier diode barrier. Then.

$$i = I_s(e^{uv} - 1) \quad (8)$$

where $u = q/nkT$

Since the eqs(1), (2), (5), (6), and (7) are derivatives of f(x) evaluated at the point $x = h$, let $h = V_o$, so that:

$$\begin{aligned} i' &= f' = I_s u e^{uV_o} \\ i'' &= f'' = I_s u^2 e^{uV_o} \\ i''' &= f''' = I_s u^3 e^{uV_o} \\ i'''' &= f'''' = I_s u^4 e^{uV_o} \end{aligned} \quad (9)$$

Substituting these relationships in eq(7) yields the result:

$$\beta = \frac{u}{2}\left[\frac{1 + \left(\frac{Au}{4}\right)^2}{1 + 2\left(\frac{Au}{4}\right)^2}\right] \quad (10)$$

This expression appears as eq(24) in the text.

Since u is a constant, independent of bias, eq(10) describes square law detection so long as the bracketed factor is approximately 1.0. The power level at which β deviates from square-law can be found from the bracketed term in eq(10):

$$\text{Error (dB)} = \Delta(\text{dB}) = 10\log\left[\frac{1 + 2\left(\frac{Au}{4}\right)^2}{1 + \left(\frac{Au}{4}\right)^2}\right] \quad (11)$$

Since A is the peak voltage on the barrier, the power taken by the junction will be (approximately):

$$P_j \approx \frac{\left(\frac{A}{\sqrt{2}}\right)^2}{R_j} = \frac{A^2}{2}uI_d = \left(\frac{Au}{4}\right)^2\frac{8I_d}{u} \quad (12)$$

Substitution in eq(11) gives the expression for the power level, $P\Delta$, at which an error, Δ(dB), results.

$$P\Delta\ (\text{dBm}) = 30 + 10\log_{10}\frac{8I_d}{u}\left(\frac{10^{\Delta/10} - 1}{2 - 10^{\Delta/10}}\right) \quad (13)$$

Eq(13) is valid only for small (less than 1.5 dB) values of Δ because a limited number of terms were used in eq(1). This expression is plotted in Figure 11 of the text for Δ values of 0.5, 1, and 1.5 dB.

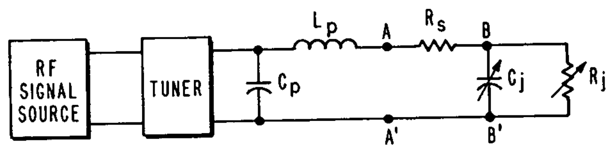
Conversion Efficiency of Hot Carrier Diode

When eq(10) is multiplied by a factor which described the ratio of power entering the junction to power entering the diode, the result is the *Conversion Efficiency*, β_c:

$$\beta_c = \frac{u}{2}\left[\frac{P_j}{P_{in}}\right] \quad (14)$$

APPENDIX II
Cut-off Frequency of Hot Carrier Diode

By referring to the diode RF equivalent circuit shown below



an expression describing the power entering the diode at $A - A'$, to the power entering the junction at $B - B'$, can be derived as follows:

$$\frac{P_{in}}{P_j} = \frac{i^2R_s + \left[i\frac{1/R_j}{\sqrt{1/R_j^2 + \omega^2C_j^2}}\right]^2 R_j}{\left[i\frac{1/R_j}{\sqrt{1/R_j^2 + \omega^2C_j^2}}\right] R_j}$$

$$= 1 + \frac{R_s}{R_j} + \omega^2R_sR_jC_j^2 \qquad (1)$$

This expression appears in eq(16) of the text.

Eq(1) can be further simplified to the form:

$$\frac{P_{in}}{P_j} = \left(1 + \frac{R_s}{R_j}\right)\left[1 + \left(\frac{f}{f_c}\right)^2\right] \qquad (2)$$

in which f is the input signal frequency, and f_c is the cut-off frequency which is described by:

$$f_c = \frac{\sqrt{1 + \frac{R_s}{R_j}}}{2\pi C_j\sqrt{R_sR_j}} \qquad (3)$$

In the above R_s is nearly constant at currents less than five milliamperes, and is approximately 11 ohms for the HPA 2350 diode. The junction capacitance C_j varies according to:

$$C_j = C_{j(0)} \bigg/ \left(1 - \frac{V}{V_b}\right)^{1/2} \qquad (4)$$

$C_{j(0)}$ is the zero-bias junction capacitance, and is approximately 0.8 pF for the HPA 2350 diode. V_b is the built-in potential $\approx$ 0.6 volt for HPA 2350.

The diode expression given in eq(1) of the text can be rearranged to express the junction voltage V in terms of the diode bias current I_d as follows:

$$V = \frac{1}{u}\ln\left(1 + \frac{I_d}{I_s}\right) \approx \frac{1}{u}\ln\frac{I_d}{I_s} \qquad (5)$$

Inserting this into eq(4) above and rearranging gives:

$$C_j = C_{j(0)} \bigg/ \sqrt{1 - \frac{1}{uV_b}\ln\frac{I_d}{I_s}} \qquad (6)$$

Substituting the values:

$$\begin{aligned} u &= q/kT = 38 \text{ volts}^{-1} \\ V_b &= 0.6 \text{ volt} \\ I_s &= 8 \times 10^{-9} \text{ amperes} \\ \ln x &= 2.3 \log_{10} x \end{aligned}$$

The following expression for C_j in terms of the diode bias current I_d is obtained.

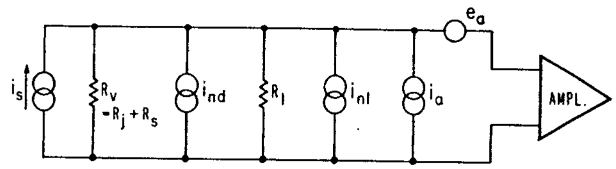
$$C_j = C_{j(0)}/\sqrt{1 - 0.1\log(125\, I_d)} \qquad (7)$$

in which I_d is in microamperes.

Eq(7) in conjunction with eq(3) of the text is used to calculate the variation of cut-off frequency f_c as a function of diode bias current I_d.

APPENDIX III
Noise Analysis

The total noise at the input to the amplifier can be obtained from a consideration of the following equivalent circuit.



Where:

i_s = Rectified Signal Current

i_{nd} = Diode Noise Current $= \sqrt{\frac{4kTB}{R_v}t_d}$

i_{nl} = Load Resistor Noise Current $= \sqrt{\frac{4kTB}{R_l}}$

i_p = Amplifier Input Noise Current
e_a = Amplifier Input Noise Voltage
R_l = Load Resistance (Amplifier Input Resistance)
R_v = Diode Dynamic Resistance at Video Frequencies

From this circuit the total noise current i_n will be:

$$i_n = \sqrt{i_{nd}^2 + i_{nl}^2 + i_a^2 + \left(\frac{1}{R_v} + \frac{1}{R_l}\right)e_a^2} \qquad (1)$$

For the purpose of analysis it is useful to represent the various noise voltages and currents, in a specified band, as equivalent resistances as follows:

$$\begin{aligned} i_{nd}^2 &= \left(\frac{4kTB}{R_v}\right)t_d \\ i_p^2 &= \left(\frac{4kTB}{R_p}\right) \\ i_{nl}^2 &= \left(\frac{4kTB}{R_l}\right) \\ e_a^2 &= (4kTB)\,R_a \end{aligned} \qquad (2)$$

Thus eq(1) may be rewritten as:

$$i_n = \sqrt{4kTB}\sqrt{\frac{t_d}{R_v} + \frac{1}{R_l} + \frac{1}{R_p} + \frac{R_a}{R_v^2} + \frac{2R_a}{R_vR_l} + \frac{R_a}{R_l^2}} \qquad (3)$$

In most applications it is close enough to assume that $1/R_l$ and $1/R_p$ are negligibly small, which leads to:

$$i_n = \frac{\sqrt{4kTB}}{R_v}\sqrt{t_dR_v + R_a} \qquad (4)$$

With the exception of a multiplicative constant, eq(4) is simply the inverse of the Figure of Merit. However, the Figure of Merit is usually specified for R_a = 1.2 K. Such a high value of R_a does not do justice to the low noise properties of Hot Carrier diodes. Consequently a more precise analysis is therefore justified.

Rewriting i_n to represent it as a resistance, R_v, with a noise temperature ratio which includes the amplifier:

$$i_n = \sqrt{\frac{4kTB}{R_v}}$$

$$\sqrt{t_d + \frac{R_v}{R_l} + \frac{R_v}{R_p} + \frac{R_a}{R_v} + \frac{2R_a}{R_l} + \left(\frac{R_a}{R_l} \times \frac{R_v}{R_l}\right)} \qquad (5)$$

If the bias current I_d is high enough to make R_v significantly different from R_j, then the terms in which R_v appears will, except for R_a/R_v, be much less than t_d, which is near or greater than unity. In this term above, therefore, it is significant to retain $R_v = R_j + R_s$. Therefore let

$$\frac{R_v}{R_l} \approx \frac{R_j}{R_l} = \frac{1}{uI_dR_l}$$

$$\frac{R_v}{R_p} \approx \frac{R_j}{R_p} = \frac{1}{uI_dR_p} \qquad (6)$$

$$\frac{R_a}{R_v} = \frac{uI_dR_a}{1 + uI_dR_s} \approx uI_dR_a$$

Making the substitution described in eq(6) gives:

$$i_n = \sqrt{\frac{4kTB}{R_v}} \sqrt{t_b + \frac{2R_a}{R_l} + \frac{1}{uI_d}\left[\frac{1}{R_p} + \frac{1}{R_l}\left(1 + \frac{R_a}{R_l}\right)\right] + \left[\frac{R_a uI_d}{1 + uI_dR_s}\right]} \qquad (7)$$

To examine the effects of flicker noise, the complete expression for t_d as given by eq(11) of the text, i.e.:

$$t_d = t_w + \frac{K_nI_d}{B} \ln \frac{f_2}{f_1} \qquad (8)$$

is substituted in eq(7) resulting in:

$$i_n = \sqrt{\frac{4kTB}{R_v}} \sqrt{\frac{1}{uI_d}\left[\frac{1}{R_p} + \frac{1}{R_l}\left(1 + \frac{R_a}{R_l}\right)\right] + \left[t_w + \frac{2R_a}{R_l}\right] + uI_d\left[\frac{R_a}{1 + R_suI_d} + \frac{K_n}{uB} \ln \frac{f_2}{f_1}\right]} \qquad (9)$$

This expression is complete, and includes such approximations as are appropriate in *any* situation. There are, however, certain specific conditions justifying further approximation.

It is *usually* true that $R_a/R_l \ll 1$, in which case:

$$\left(1 + \frac{R_a}{R_l}\right) \approx 1 \text{ and } \left[t_w + \frac{2R_a}{R_l}\right] \approx 1 \text{ since } t_w < 1 \qquad (10)$$

Using these approximations, eq(9) reduces to:

$$i_n = \sqrt{\frac{4kTB}{R_v}} \sqrt{\frac{1}{uI_d}\left[\frac{1}{R_p} + \frac{1}{R_l}\right] + [1] + uI_d\left[\frac{R_a}{1 + R_suI_d} + \frac{K_n}{uB} \ln \frac{f_2}{f_1}\right]} \qquad (11)$$

To obtain optimum square law range it is usually a standard practice to make $R_l \geq 5\ R_v$. Under this condition the assumption that $\left[\frac{1}{R_p} + \frac{1}{R_l}\right] \approx 0$ is suitable. At bias currents less than 300 μA, an *equivalent noise resistance* R_A can be defined which includes the amplifier noise resistance and the flicker noise contribution as follows:

$$[R_A] = \left[\frac{R_a}{1 + R_suI_d} + \frac{K_n}{uB} \ln \frac{f_2}{f_1}\right] \qquad (12)$$

Eq(11) can now be reduced to:

$$i_n = \sqrt{\frac{4kTB}{R_v}} \sqrt{1 + uI_dR_A} = \sqrt{\frac{4kTB}{R_v}} \sqrt{1 + \frac{R_A}{R_j}} \qquad (13)$$

This expression appears as eq(27) in the text. R_A as a function of bias is also plotted in Figure 12.

APPENDIX IV
Tangential Sensitivity

For an input signal power of P_{in}, the signal current at the input to the amplifier is:

$$i_s = P_{in}\beta_c = P_{in} \frac{u}{2}\left(\frac{P_j}{P_{in}}\right) \qquad (1)$$

where β_c is defined by eq(14) Appendix I. The noise current at the input to the amplifier is i_n. If Tangential Sensitivity is taken to correspond to a signal to noise ratio of 2.5. Then:

$$\frac{i_s}{i_n} = 2.5 \qquad (2)$$

and the Tangential Sensitivity (TS) is:

$$TS = \frac{5\ i_n}{u\left(\frac{P_j}{P_{in}}\right)} \qquad (3)$$

Substituting for (P_j/P_{in}) from eq(2) Appendix II, and for i_n from eq(13) Appendix III,

$$TS = \frac{5}{u}\left(\frac{R_j + R_s}{R_j}\right)\sqrt{\frac{4kTB}{R_s + R_j}}\sqrt{1 + \frac{R_A}{R_j}}\left[1 + \left(\frac{f}{f_c}\right)^2\right]$$

$$TS = \frac{5}{u}\frac{\sqrt{R_j + R_s}}{R_j}\sqrt{4kT}\sqrt{\frac{R_j + R_A}{R_j}}\left[1 + \left(\frac{f}{f_c}\right)^2\right]\sqrt{B} \qquad (4)$$

$$TS_{(dBm)} = 10 \log\left[\frac{5}{u}\sqrt{4kT}\frac{\sqrt{R_j + R_s}}{R_j}\sqrt{\frac{R_j + R_A}{R_j}}\right] + 10 \log\left[1 + \left(\frac{f}{f_c}\right)^2\right] + 5 \log B$$

$$TS = TS_o + TS_f + 5 \log B$$

This equation appears as eq (28) in the text. The variations of TS_o and TS_f with diode bias current are also shown in the text as Figures 12 and 13 respectively.

Because both TS_o and TS_f are functions of bias current, the minimum value of TS can be obtained by differentiating eq(4) with respect to bias current and setting the differential equal to zero. In which case:

$$TS_o' = -TS_f'$$

In practice the minimum value of TS can be obtained from an intersection of the slopes of TS_o and TS_f. These are given in Figure 14 of the text.