



# Schottky Barrier Diode Video Detectors

## Application Note 923

### I. Introduction

This Application Note describes the characteristics of Agilent Technologies Schottky Barrier Diodes intended for use in video detector or video receiver circuits, and discusses some design features of such circuits.

Although a video receiver is typically 35 - 40 dB less sensitive than a heterodyne receiver, nevertheless its simplicity and low cost can outweigh this disadvantage. For example, a video receiver requires no local oscillator power; it generally uses only a single diode; it is capable of being built with a large RF bandwidth; and it is much less critical to design and maintain than a heterodyne receiver. These advantages make it useful as beacon receivers, missile guidance receivers, fuse-activating receivers, countermeasures receivers, and power leveling and signal monitoring detectors.

A typical video receiver circuit is shown in Figure 1. The operation of this circuit is quite simple. At RF, the bypass capacitor ( $C_b$ ) appears as a short circuit, and the input RF choke ( $RFC_1$ ) appears as an open circuit. The input RF filter is optimized to match the signal source impedance to the diode's RF impedance over a specified bandwidth and, ideally, all of the available signal power is delivered to the diode. Due to the nonlinearity of the diode, the video signal is extracted from the modulated RF signal and appears across the video load resistance

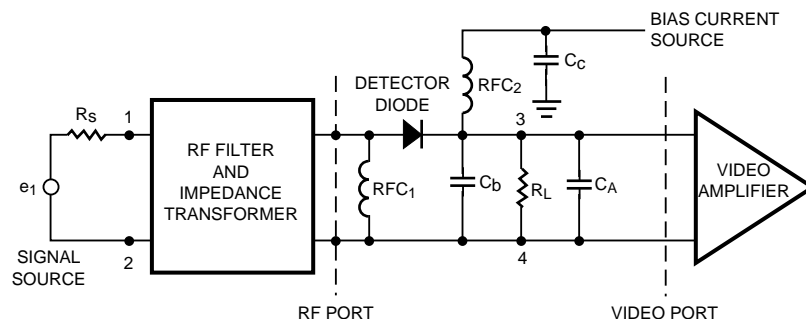


Figure 1. Typical Video Receiver

( $R_L$ ). At video frequencies,  $RFC_1$  acts as a short circuit, while  $C_b$  and the DC bias filter, consisting of  $RFC_2$  and  $C_c$ , both appear as high impedances.  $RFC_1$  also serves as a return path for the DC bias current.

The principal requirements for a video receiver are distortionless recovery of the modulation signal, which is usually a pulse, and maximum RF sensitivity. However, these requirements are competitive and result in the design being, at best, a good compromise between video bandwidth, which determines the fidelity of the detected pulse, and RF sensitivity.

## II. Diode Performance Characteristics

The performance characteristics that are used to describe video detector diodes are Tangential Sensitivity, Video Resistance, and Voltage Sensitivity.

**1. Tangential Sensitivity (TSS)** – This is the lowest signal power level for which the detector will have a specified signal-to-noise ratio at the output of the video amplifier. At Agilent Technologies, the output signal-to-noise ratio is specified to be 8 dB [1]. The units for TSS are dBm or milliwatts. As we shall see later, TSS does not depend entirely on intrinsic diode parameters and many factors affect the measured TSS value for a given diode. For the time being, the most important factors are:

- a) RF Frequency
- b) Video Bandwidth
- c) Diode DC Bias Current
- d) Test Mount or Circuit
- e) Video Amplifier Noise Figure

For an exact TSS specification, the effective video bandwidth should be stated as the lower and the upper 3 dB frequencies of the entire video circuit, including the diode's video resistance. A statement of only the bandwidth of the video amplifier can be misleading because it does not always determine the overall or effective bandwidth of the system. The limitation on overall bandwidth can come both from the circuit between the diode output and the amplifier input or the instrumentation circuit after the amplifier, i.e., oscilloscope or meter.

Because the upper 3 dB frequency is usually several orders of magnitude greater than the lower 3 dB frequency, it is common to state only the upper 3 dB frequency. For example, a usual statement of video bandwidth may be 2 MHz. The implication is that the response of the video circuit of the detector to the modulating signal extends from DC to 2 MHz. In actual practice, the low frequency response very seldom extends down to DC, because this will include the flicker noise contribution of the diode and of the video amplifier, both of which deteriorate TSS. If the diode is expected to be used in a system that will require low frequency response, then the low frequency 3 dB point of the test system must be stated. Failure to do so can lead to gross differences in TSS between the test system and the actual system if the diode's flicker noise corner frequency  $f_N$  is high, i.e.,  $>50$  kHz.

To obtain maximum sensitivity at any given frequency, most [2] detector diodes must be forward DC biased. Bias however introduces shot and flicker (1/f) noise in the diode, and reduces the diode video resistance. These effects exert a competitive influence on TSS – therefore, the bias value must be stated.

Since diodes of different designs can differ widely in their RF impedance characteristics, particularly if package parameters are different or are not sufficiently well controlled, repeatability of performance can only be obtained by a test in a specific mount and at a specific frequency. At Agilent, detector diodes of different types are tested in mounts that have been tuned and “locked” to that type.

**2. Video Resistance ( $R_v$ )** – This parameter is simply the small signal low frequency dynamic resistance of the diode and is dependent on the DC bias current. The value of the bias current used is the same that is used in the TSS test.

$R_v$  consists of the sum of the diode’s series resistance ( $R_s$ ) and the junction resistance ( $R_j$ )

$$R_v = R_s + R_j \quad (1)$$

$R_j$  is obtained by differentiating the diode voltage-current characteristic and is given by:

$$R_j (i) = \frac{nkT}{q(I_d + I_s)} \quad (2)$$

where  $I_d$  is the bias current and  $I_s$  is the saturation current.  $I_s$  is  $\cong 10^{-9}$  amperes for Agilent detector diodes and is negligible when the diode is operated in the optimum bias region.

The nonideality factor  $n$  is different for each diode type. Typical values for Agilent detector diodes are:

Diode Type	$n$
Agilent 5082-2750	1.08
Agilent 5082-2824	1.02

For an  $n = 1.08$  and at room temperature,  $R_j$  can be simplified to:

$$R_j (i) \cong \frac{28}{I_d} \quad (3)$$

where  $I_d$  is in milliamperes and  $R_j$  is in ohms.

**3. Voltage Sensitivity ( $\gamma$ )** – This parameter specifies the slope of the output video voltage versus the input signal power, i.e.,  $V_o = \gamma P_{in}$  of the diode. It is bias, load resistance, signal level, and RF frequency dependent, and all of these conditions must be specified. It is particularly sensitive to signal level which must be kept well within the square law dynamic range of the diode.

### III. Optimum Video Detector Sensitivity

Assuming a diode ideality factor  $n = 1.08$ , and at room temperature, the TSS of a video receiver can be stated as [3]:

$$\text{TSS}_{(\text{dBm})} = -107 + 5 \log B_V + 10 \log I_d + 5 \log \left[ R_A + \frac{28}{I_d} \left( 1 + \frac{f_N}{B_V} \ln \frac{B_V}{f_L} \right) \right] + 10 \log \left[ 1 + \frac{R_S C_{j(i)}^2 f^2}{I_d} \right] \quad (4)$$

where

- $B_V$  = Video bandwidth in Hz
- $I_d$  = Diode dc bias in  $\mu\text{A}$
- $f_N$  = Diode flicker noise corner frequency, Hz
- $f_L$  = Video circuit low frequency 3 dB point, Hz
- $R_S$  = Diode series resistance,  $\Omega$
- $C_{j(i)}$  = Diode junction capacitance, pF, at the bias current  $I_d$

For the Agilent 5082-2750 and 5082-2755 diodes, this is approximately

$$C_{j(i)} = \sqrt{\frac{C_{j(0)}}{1 - 0.1 \log(1300 I_d)}} \quad (5)$$

- $f$  = Operating frequency, GHz
- $R_A$  = Amplifier equivalent series noise resistance,  $k\Omega$

This expression reveals that the only significant diode parameters that affect the detector sensitivity are the three parasitic parameters  $f_N$ ,  $R_S$ , and  $C_j$ . It can also be shown that there is a value of bias current, for a given diode and operating frequency, which results in maximum TSS.

Assuming that the video amplifier contributes negligible noise compared to the diode,  $R_A$  can be neglected in Eq. (5). Differentiating this simplified expression yields an approximate value of the optimum bias as:

$$I_{d(\text{OPT}) (\mu\text{A})} \cong R_{S(\Omega)} [C_{j(i)} (\text{pF})]^2 [f_{(\text{GHz})}]^2 \quad (6)$$

which is valid for  $0 < I_d < 50 \mu\text{A}$ .

The effect of bias on the TSS of several Agilent detector diodes is shown in Figure 2, and the same effect at different signal frequencies for the Agilent 5082-2755 is shown in Figure 3.

It is worthwhile to assess the magnitude of the degradation on sensitivity that is introduced by the various factors in Eq. (5). Assuming the following parameters for the system and the diode:

TEST CONDITIONS:

- VIDEO LOAD IMPEDANCE ( $R_L$ ) = 38  $k\Omega$
- VIDEO BANDWIDTH ( $B_V$ ) = 2 MHz
- EQUIVALENT NOISE RESISTANCE OF VIDEO AMPLIFIER ( $R_A$ ) = 500  $\Omega$

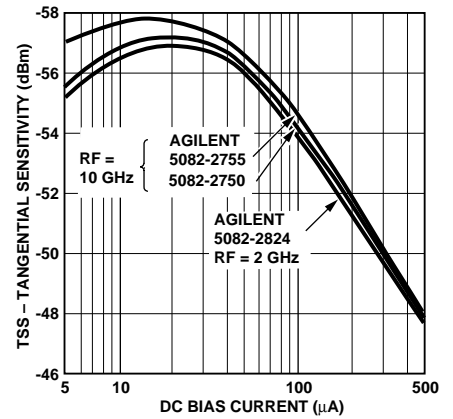


Figure 2. Effect of DC Bias on TSS

TEST CONDITIONS:

- $B_V = 2 \text{ MHz}$
- $R_A = 500 \Omega$
- $R_L = 38 k\Omega$

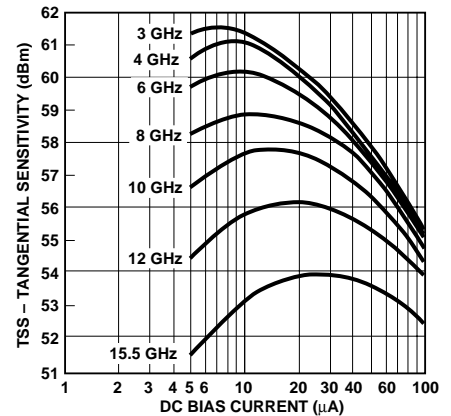


Figure 3. Effect of Bias on TSS at Different RF Frequencies (Typical 5082-2755)

**Diode and System Parameters**

Diode Agilent 5082-2750:  $f_N = 3 \text{ kHz}$ ,  $I_d = 20 \mu\text{A}$ ,  $f = 10 \text{ GHz}$ ,  
 $R_A = 0.5 \text{ k}\Omega$ ,  $B_V = 2 \text{ MHz}$ ,  $R_S = 25 \Omega$

$$10 \log \left( 1 + \frac{R_S C_{j(i)}^2 f^2}{I_d} \right) = 3.0 \text{ dB}$$

$$5 \log \left[ R_A + \frac{28}{I_d} \left( 1 + \frac{f_N}{B_V} \ln \frac{B_V}{f_L} \right) \right] \cong 1.4 \text{ dB}$$

$$10 \log I_d = 13.0 \text{ dB}$$

$$5 \log B_V = 31.5 \text{ dB}$$

$$\text{Total Degradation} = 48.9 \text{ dB}$$

$$\text{Expected TSS} = -107 + 48.8 = -58.1 \text{ dBm}$$

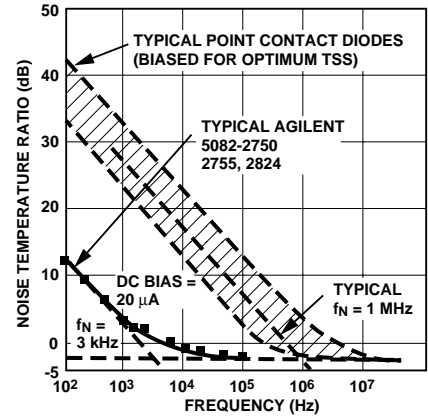
We get the following:

This is typical of the Agilent 5082-2750 diode which is specified at -55 dBm to allow for reasonable variations in diode parameters. The significance of the above is the relative magnitude of the various degradation factors in a particular system. The degradation due to diode flicker and shot noise is particularly low due to the low flicker noise corner frequency for this diode when it is operated at the optimum bias of 20  $\mu\text{A}$ . The typical flicker noise characteristics of the Agilent 5082-2750 and 5082-2824 diodes are shown in Figures 4 and 5. Figure 4 shows the flicker noise characteristic as a function of video frequency at 20  $\mu\text{A}$  bias, and Figure 5 shows the change in the noise corner frequency  $f_N$  as a function of bias. Even at high bias levels,  $f_N$  is sufficiently low, i.e., 30 kHz, by comparison to normal video bandwidths, typically > 300 kHz, that the flicker noise contribution can be considered negligible. This can be seen in Figure 6 in which the factor

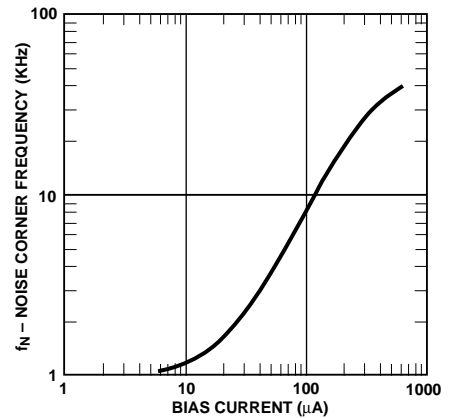
$$5 \log \left[ R_A + \frac{28}{I_d} \left( 1 + \frac{f_N}{B_V} \ln \frac{B_V}{f_L} \right) \right]$$

of Eq. (5) is plotted as a function of  $B_V$  with  $f_L$  fixed at 100 Hz. For most applications of the Agilent Detector Diodes, this factor can be simplified to:

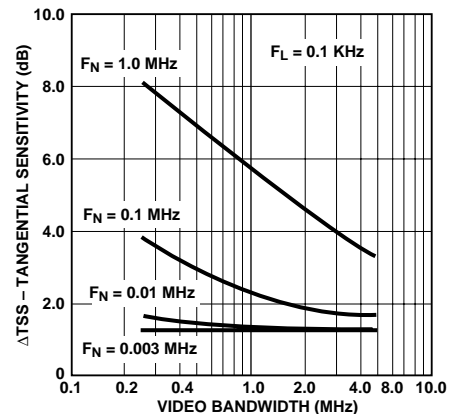
$$5 \log \left[ R_A + \frac{28}{I_d} \right] \quad (7)$$



**Figure 4. Flicker Noise Characteristics of Agilent Detector Diodes**



**Figure 5. Noise Corner Frequency vs. Bias (Typical Agilent 5082-2750, 5082-2755)**



**Figure 6. Amplifier and Diode Noise Contribution as a Function of Video Bandwidth**

#### IV. Bandwidth Requirements

To maintain high sensitivity, the video bandwidth of the video detector should be no greater than necessary to recover the modulation information. The bandwidth required for pulse recovery depends on the nature of the information to be gained from the pulse. For example, in pulse search radar, peak pulse detectability is more significant than pulse shape. In ferret systems, instrumentation, or data transmission systems, the resolution of the fine details of the pulse shape, i.e., risetime and pulsewidth, may be of much greater importance in spite of greater bandwidth requirements.

For maximum pulse detectability, the video bandwidth should be made just sufficient enough to maximize the output signal-to-noise ratio of the detecting system. This condition occurs when the video circuit transfer function is matched to the spectrum of the signal waveform [4,5]; i.e., for a rectangular pulse and a rectangular low pass filter response, the required filter bandwidth is

$$B_V \cong 1/t_w \quad (8)$$

where  $t_w$  = Pulsewidth

For a simple RC filter, which is the usual case at the output of the detector diode, the maximum S/N occurs when the filter's upper 3 dB video frequency is approximately:

$$f_{u(3\text{ dB})} \cong \frac{0.25}{t_w} \quad (9)$$

In other applications, it may be desirable to resolve the shape of the pulse, i.e., risetime, pulsewidth and flat amplitude. To resolve the risetime of the pulse, the upper 3 dB frequency of the video circuit will have to be

$$f_{u(3\text{ dB})} \cong 0.35/t_r \quad (10)$$

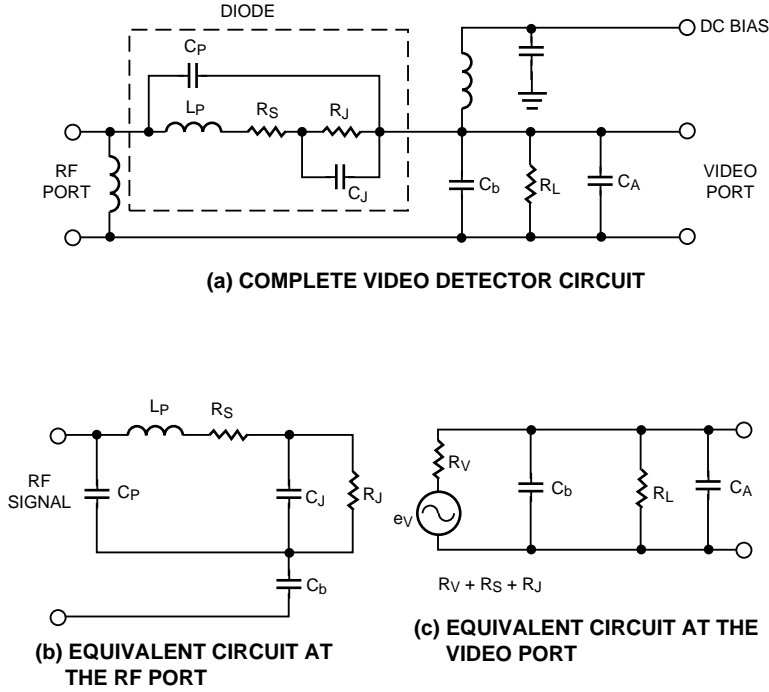
where  $t_r$  = Pulse risetime

To resolve the shape of the pulse to any degree of accuracy, consideration must also be given to how much amplitude droop is permissible. Amplitude droop depends on the low frequency 3 dB point and is given by:

$$f_{L(3\text{ dB})} = \frac{\text{Droop \%}}{600 t_w} \quad (11)$$

## V. Design Considerations for Video Detector

The basic video detector circuit and its equivalent representations at both the RF and video ports are shown in Figure 7.



**Figure 7. Video Detector Equivalent Circuits**

### (A) Video Circuit Design Considerations

In the video circuit,  $R_L$  represents the load or amplifier input resistance, and  $C_A$  represents the amplifier input capacitance as well as all the stray and particularly cable capacitances that may be present in the video circuit. These  $R$  and  $C$  elements will impose a limit on the upper 3 dB cut-off frequency of this circuit, which is given by:

$$f_{u(3\text{ dB})} = \frac{1}{2\pi R_T C_T} \quad (12)$$

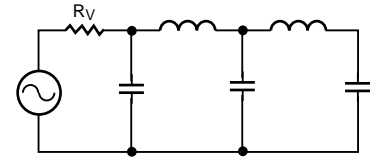
$$\text{where } R_T = \frac{R_V R_L}{R_V + R_L} \quad (13)$$

$$\text{and } C_T = C_b + C_A \quad (14)$$

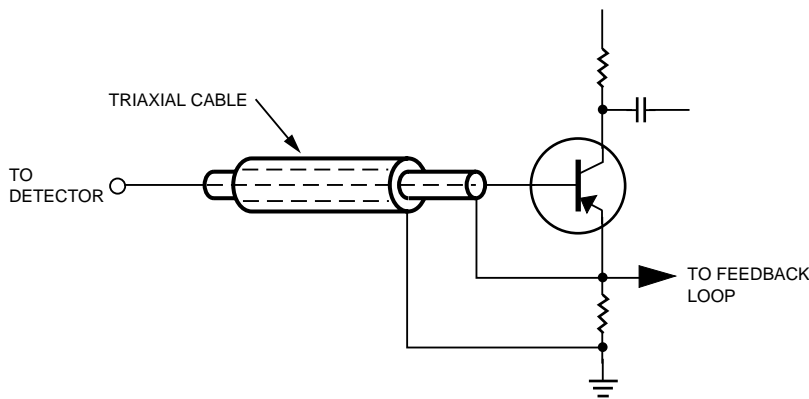
In most video detector designs, this circuit has a greater influence on the effective video bandwidth than the bandwidth of the amplifier. The  $R_T C_T$  time constant can be reduced by reducing all the element values within certain limits. A severe reduction in the value of the RF bypass capacitance  $C_b$  will lead to poor RF/video isolation and a decrease in the signal level delivered to the diode. The reactance of this capacitor at the operating frequency should be kept to less than 10% of the RF impedance of the diode. At low RF frequencies and wide video

bandwidths, this capacitor can be replaced by a low pass filter structure as shown in Figure 8 with a cut-off frequency  $f_c \leq 1/2 f_{rf}$ . Because this filter is required to pass fast pulses, the design should be for flat time delay rather than for equal-ripple.

A reduction in stray and amplifier capacitances is always advisable and in wideband designs is often necessary. One suitable technique which is particularly effective when cable interconnections must be made between the detector circuit and the video amplifier is shown in Figure 9. In this circuit, the original cable capacitance  $C_o$  is reduced by feedback to:  $C_{eff} = C_o (1 - A_v)$  where  $A_v$  is the total gain from the input of the amplifier to the first shield and must be  $< 1$ .



**Figure 8. Low Pass Video Coupling Structure**



**Figure 9. Circuit for Reducing Cable Capacitance**

Alternately, either  $R_L$  or  $R_v$  can be reduced. The amount that  $R_L$  can be reduced is often limited if voltage amplification is desired since the output voltage of the detector is maximized by making  $R_L$  large.  $R_v$  of the diode can be lowered by increasing the bias current. Although this results in reduced sensitivity, as was shown in Figure 2, it may nevertheless be needed to achieve the required video bandwidth. A reduction of video resistance has other beneficial effects. It can, for instance, be adjusted to be the optimum source resistance value for a minimum noise figure of the video amplifier. When this is desired,  $R_v$  should be adjusted so that

$$R_v \cong \sqrt{R_{NS}R_{NP}} \quad (15)$$

where  $R_{NS}$  = Equivalent series noise resistance of amplifier

$R_{NP}$  = Equivalent parallel noise resistance of amplifier

This is usually in the range of 400 - 2000 ohms for most low noise transistor input stages.



### (B) RF Circuit Design Considerations

The RF circuit consists essentially of a filter structure that is designed to match the signal source resistance, usually 50 ohms, to the diode junction resistance,  $R_j$ . The equivalent circuit of the diode at RF is shown in Figure 7b. The presence of  $R_S$  and  $C_j$  introduces a loss of signal which is dependent on frequency and the magnitudes of  $R_S$ ,  $C_j$ , and  $R_j$ . This dependence is given by

$$L_{db} = 10 \log \left[ 1 + \frac{R_S}{R_j} + \omega^2 C_j^2 R_S R_j \right] \quad (16)$$

where  $\omega = 2\pi f$

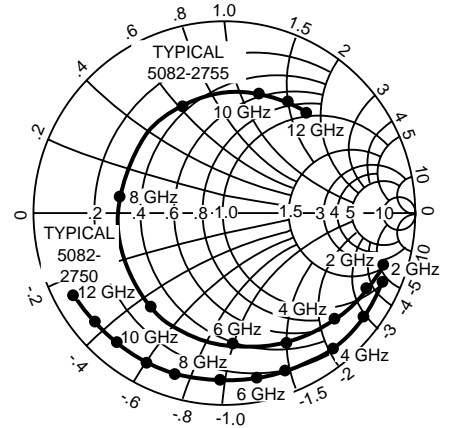
Since  $R_j$  is a function of bias current, i.e.,  $R_j \cong 28/I_d$ , the above can be restated as

$$L_{db} = 10 \log \left[ 1 + \frac{R_S I_d}{28} + \frac{28 \omega^2 C_j^2 R_S}{I_d} \right] \quad (17)$$

from which it is obvious that the RF loss can be minimized at any given frequency by biasing the diode. This effect was included as one of the bias dependent parameters in the TSS expression (Eq. 5), which was then used to determine optimum bias with respect to maximum TSS at a given frequency. Bias also affects the RF impedance of the diode, and the impedance that is obtained at a bias corresponding to maximum TSS is not necessarily optimum for achieving broadband RF matching to the diode.

The typical measured RF impedances for Agilent 5082-2750 and 5082-2755 diodes, which are the same chip in two different packages, are shown as a function of frequency in Figure 10. Figure 10 shows the diode when it is biased at 20  $\mu$ A ( $R_j \cong 1400$ ), which is its optimum bias for maximum sensitivity at 10 GHz. As can be seen, the SWR for either package style is high over most of the frequency range and normally either diode will have to be matched to the source with additional reactive elements. Practical reactances, however well made, will introduce additional losses in the matching structure. Furthermore, the farther these reactances are placed from the diode chip, the more loss will be introduced due to high standing waves and current maxima points due to high standing waves and current maxima points between them. Even well made double-stub tuners can introduce losses as high as 1.5 - 2 dB. The impedance plots of Figure 10 also clearly illustrate the effect that package and chip parasitics can have on the RF impedance of the diode. Assuming that the RF bypass capacitor  $C_b$  is properly chosen and can be neglected, which was the case for the impedance plots, the RF admittance of the remaining circuit can be expressed as

$$Y_{12} = j\omega C_p + \frac{1}{R_S + j\omega L_p + \frac{1}{j\omega C_j + \frac{1}{R_j}}} \quad (18)$$



**Figure 10. Typical RF Impedance of Agilent 5082-2750 and 5082-2755 Diodes at Bias Current of 20  $\mu$ A**

This admittance has two potential resonant frequencies. Neglecting  $R_s$  and  $R_j$ , these resonant frequencies can be expressed approximately as:

$$\omega_S \cong \frac{1}{\sqrt{L_p C_j}} \quad (19)$$

$$\omega_P \cong \sqrt{\frac{1}{L_p C_j} + \frac{1}{L_p C_p}} \quad (20)$$

As can be seen from the plots, a resonant frequency for the high inductance glass package is evident at  $\cong 7.9$  GHz.

Due to the low values of the parasitic elements, the pill packaged device 5082-2750 does not exhibit any resonances up to a frequency of 12 GHz. Such a diode can be brought into match over fairly broad bandwidths by a series length of line and a quarterwave transformer [6].

To simplify broadband matching, the impedance of the diode can be reduced by using more DC bias than would normally be required for maximum TSS. Figure 11 illustrates the magnitude of the effect that bias has on the RF impedance. These plots are made for the same diodes as in Figure 9, but at a bias of  $330 \mu\text{A}$ . The 5082-2750 pill packaged diode impedance is now within a SWR of : 2/1 over most of its useful frequency range. Although there will also be a reduction of sensitivity of about 7 dB, as was shown in Figure 2, nevertheless, the trade-off may be worthwhile in broadband systems.

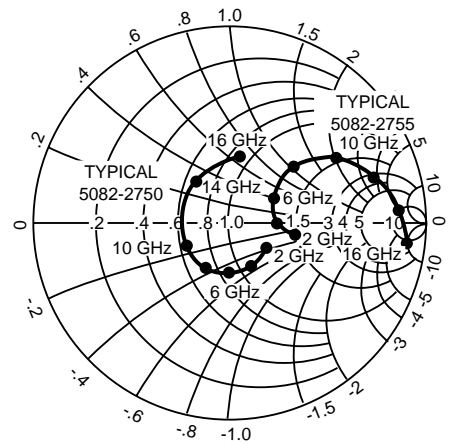
## VI. Dynamic Range

In addition to achieving high sensitivity, video detectors are usually required to have a large dynamic range. Usually this statement refers to the "square-law" range as shown in Figure 12 and is the range over which the output signal voltage is proportional to the input signal power, i.e.:

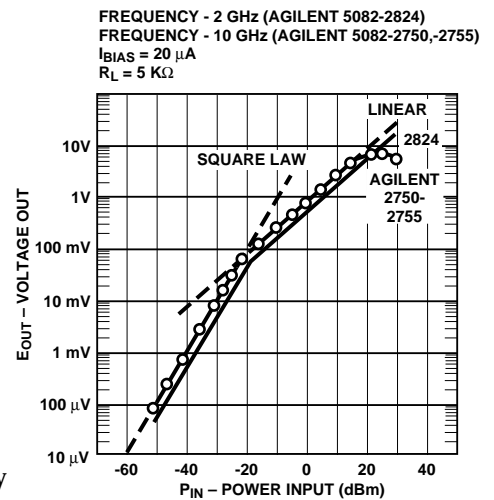
$$V_s = \gamma P_{in} \quad (21)$$

where  $\gamma$  is the diode voltage sensitivity at the specified frequency and bias. Deviation from square-law characteristic occurs when  $\gamma$  departs from a constant value. This will occur when the RF carrier and, consequently, the rectified current become high enough to appreciably affect the quiescent bias current,  $I_d$ . At Agilent, the upper limit of square-law operation is defined to be the power level which is 0.5 dB higher than the power level that would produce the same output voltage if  $\gamma$  had remained constant.

As can be seen from Figure 12, the transfer characteristic of a Schottky Barrier Diode does not simply saturate beyond this point but merely changes from square-law to linear, which then continues to a higher power level before saturation. This range is defined as the linear dynamic range and can be appreciable if the breakdown voltage of the diode is high. For example, the Agilent 5082-2824 detector diode is



**Figure 11. Typical RF Impedance of Agilent 5082-2750 and 5082-2755 Diodes at Bias Current of  $330 \mu\text{A}$**



**Figure 12. Detector Dynamic Characteristics**

rated at a breakdown voltage of 15 V. By contrast, point contact diodes whose breakdown voltage is generally low (2 - 3 V), do not exhibit a broad linear range. For applications that require monitoring of high level RF signals, i.e., radio controlled trigger circuits, high power monitors, or RF proximity fuses, this range is very useful. Applications that are not critical of the change in the transfer characteristics can utilize the very large total (square-law and linear) dynamic range. Finally, in very critical applications where high square-law range is necessary, the video amplifier gain characteristics can be shaped. The range over which such compensation can be achieved will depend very much on the video bandwidth of the amplifier. For bandwidths of 2 MHz or less, the square-law range can be so extended to about +10 dBm with an accuracy of  $\pm 1$  dB.

## VII. Detector Diode Burnout

The destruction of the detector under high signal level conditions merits some discussion. It has been the industry standard to rate detectors according to energy burnout, usually stated as so many ergs. Assuming that energy burnout can be defined sufficiently well, it still remains a very dubious performance requirement to be imposed on the detector. With the exception of possible static discharge which can occur during handling of the diode, total energy stress will very seldom be the kind of stress that a video detector is exposed to.

Video detectors are often designed for broad, RF bandwidths, thus they can be exposed to high peak power radiation from local radar transmitters. A better measure of the video detector burnout is a measure of its capability to handle peak powers in the range of 1 or more microsecond duration, and to expose the diode to this stress for a period of time at a specified repetition rate. Because the diode is generally biased to a junction resistance on the order of 1 - 2 k $\Omega$ , the RF circuit, including the package parasitics, consists of a large impedance ratio matching filter that matches the small-signal impedance of the diode to the source impedance. As the signal level increases, the impedance of the diode changes and produces a mismatch with respect to the matching structure. This mismatch results in a reflection of some of the incident RF power and only a portion of the available RF power is actually dissipated in the diode. The diode, in effect, protects itself. Regardless how much of the incident power is reflected, eventually a power level can be reached where the difference between the incident and the reflected power is sufficient to destroy the diode. This is the peak power dissipation capability of the diode. The amount of peak incident power the diode can handle depends very much on the characteristics of the RF structure between the source and the diode junction. Figure 13 illustrates the typical peak power handling capability of an Agilent 5082-2824 diode using a narrowband tuning structure tuned to the diode at the small-signal (TSS) level. In very wideband systems, where no tuning structures are used and where a tolerable input SWR at small signal levels is obtained by biasing the diode to a lower impedance level, the above situation does not apply. Here an increase in the incident power level brings the diode closer to a matched condition and, consequently, the diode's dissipation ratings apply.

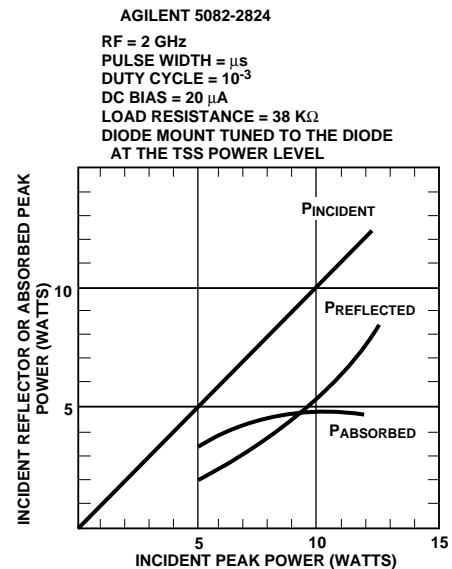
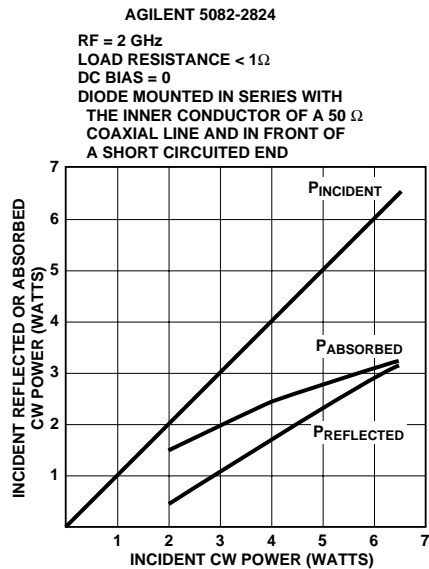


Figure 13. Peak Power Handling Characteristics



In some applications, i.e., power monitoring and modulating, the CW rather than peak power ratings apply. As previously stated, since the impedance of the diode is a function of incident power, the amount of power the diode can handle is greater than its CW dissipation capability. This is shown in Figure 14 for a typical Agilent 5082-2824 detector diode.



**Figure 14. CW Power Handling Characteristics**

## References

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2. Agilent Technologies Application Note 969, "An Optimum Zero Bias Schottky Detector Diode."
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6. Agilent Technologies Application Note 963, "Impedance Matching Techniques for Mixers and Detectors."