# Broad-Band TEM Diode Limiting*

R. V. GARVER†, MEMBER, IRE, AND J. A. ROSADO†

*Summary*—**The bandwidths of two types of limiters operating below diode resonance and one type of limiter operating at diode resonance are calculated. A 2.5-Gc base-band limiter was made providing a low power insertion loss of less than 1 db, a limiting threshold of 10 mw, and a high power isolation of greater than 20 db. A 0.9- to 1.3-Gc matched limiter was made having a VSWR of less than 1.2 for all power levels. The burnout power of these two limiters was calculated to be about 10 watts incident CW power or 1500 watt-microsecond incident pulse energy. Using the diode resonance the calculations indicate that it is possible to make a 5-Gc limiter with 15 per cent bandwidth, less than 1-db low power insertion loss, a limiting level of 10 mw, and greater than 20-db isolation at high power. The bandwidths derived for diode limiting are equally applicable to switching.**

## Introduction

With the development of improved diodes for microwave applications several types of diode limiters have been developed. The first type [1]–[8] is the HF application of the LF technique of two diodes shunting a transmission line. The radio frequency above a certain power level is clipped by the diodes. The second type [9]–[11] is a pumped subharmonic oscillator. When the system beaks into subharmonic oscillation with increasing pump power the amplitude swing corresponding to the pump remains constant. The third type [12]–[14] makes use of parametric frequency converters. With two frequency conversions the output frequency can be the same as the input frequency and the device can provide gain at low power levels. The fourth type [15] of limiter makes use of diode switching techniques in that a diode switch that provides high isolation with diode current conduction functions passively as a limiter when its switching terminals are short circuited.

Limiters can be used as protective devices in protecting mixer diodes from burnout and in reducing receiver saturation from pulse leak-through, or they can be used as levelers for reducing undesirable amplitude fluctuations; *e.g.*, for parametric amplifiers in which the amplifier is radically affected by variations in pump power, for frequency swept oscillators in which amplitude variations obscure information, and with phase detection radar systems in which reflectivity and range variations cause undesirable amplitude modulation. For some of the above uses only broad-band limiters will suffice and for the other uses the availability of broad-band limiters will eliminate the need for custom-designed devices for each narrow-frequency range.

The diode limiters reported here are based on diode switching techniques. Of the many possible ways of making broad-band TEM switches that are limiters, three appear to be the most practical. The first type uses diodes in shunt [1]–[8]. The second type is a matched limiter using a 3-db coupler or circulator [15]. The third type, which is most practical above 2 Gc, utilizes the diode self-resonance.

The definitions used in diode switching will be applied to limiters. The attenuation $\alpha$ of a switch or limiter is the ratio expressed in decibels of incident power to the power transmitted past the device. The insertion loss $\delta$ is the minimum attenuation which occurs at the lowest incident power. The isolation $\eta$ is the maximum attenuation which occurs at the highest incident power. In an ideal limiter the attenuation should change with increasing power in such a way that the output power remains constant for a wide range of input power.

## I. Two Diodes in Shunt

### A. Diodes in Same Plane

Two diodes shunting a transmission line (Fig. 1) will clip the voltage appearing across the matched terminating load according to the voltage at which the diodes begin to conduct. The voltage-current characteristic of the two diodes is shown in the lower portion of Fig. 1. Germanium diodes begin conduction at about 0.3 v; silicon at 0.7 v; and gallium arsenide at about 1.0 v. The power delivered to a 50-ohm matched load would correspondingly be 2 mw, 10 mw, and 20 mw. Biasing the diodes raises or lowers the limiting level. Below the limiting power level the diodes have a low susceptance which will cause very little reflection and thus have a low insertion loss.

The attenuation $\alpha$ of a diode of admittance $Y(=G+jB)$ shunting a transmission line of characteristic admittance $Y_0$ is (see [16])

$$\alpha = 10 \log \left[\left(\frac{G}{2Y_0}+1\right)^2+\left(\frac{B}{2Y_0}\right)^2\right]. \qquad (1)$$

If the diode is purely inductive at high power levels (owing to diode whisker inductance $L_w$) maximum limiting will decrease with increasing frequency as shown by the inductance labeled curves in Fig. 2. If the diode is purely capacitive at low power levels (owing to the zero-bias capacitance of the diode and cartridge capacitance $C_0+C_c$ and assuming zero whisker inductance $L_w$) the insertion loss will increase with increasing frequency (Fig. 2). The limitation on maximum isolation imposed by the diode spreading resistance $R_s$ is shown on the attenuation axis. For two identical diodes placed in parallel in the same plane, as is done at low fre-

* Received March 21, 1962.
† Ordnance Corps, Diamond Ordnance Fuze Laboratories, Washington, D. C.

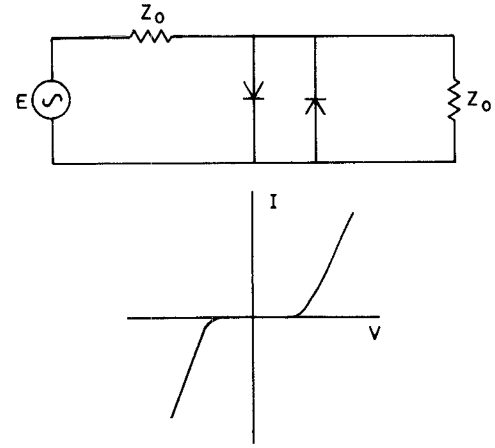


Fig. 1—The schematic diagram of two diodes shunting a transmission line and their combined voltage-current characteristic.

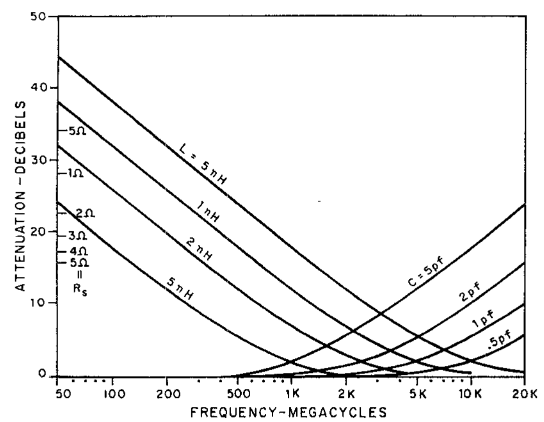


Fig. 2—Attenuation bandwidth of a diode shunting a 50-ohm transmission line as a function of diode inductance, and capacitance, and the limit on maximum attenuation imposed by spreading resistance.

quencies, the figure describes their characteristics if twice the capacitance and half the inductance and spreading resistance are used. Two varactors with $L_w = 2$ nh, $C_0 = 1$ pf, and $R_s = 4\ \Omega$ will give greater than 20-db high power protection from zero to 400 Mc. This curve also describes the maximum possible bandwidth that can be obtained using these diodes. When an inductor is put in parallel with the diodes the center frequency of the low power insertion loss increases in frequency and retains the same bandwidth in megacycles. By placing capacitors in series with the diodes the center frequency of the isolation may be similarly increased in frequency, while bandwidth remains unchanged. It is necessary at higher frequencies to take into account the diode self-resonance, which will increase the effective inductance at high powers and the effective capacitance at low powers.

For pill-type varactors with $L_w = 0.7$ nh, $R_s = 4\ \Omega$, $C_0 = C_c = 1$ pf, the limiter should give a high power isolation of 22 db, this maximum decreasing to 20 db at about 1.1 Gc. The low power insertion loss should increase to about 1.5 db at 1.1 Gc. Much better characteristics can be obtained from these two diodes if they are not located in the same plane.

### *B. Diodes Not in Same Plane*

Assuming the two diodes are pure susceptances, Fig. 3 (from [16]) gives the attenuation as a function of spacing in $l/\lambda_g$ ($l$ is physical distance between diode centers and $\lambda_g$ is the wavelength in the transmission line) and as a function of normalized susceptance for shunt diodes or normalized reactance for series diodes. At low power levels the capacitive susceptance of the diodes is proportional to frequency as is $l/\lambda_g$ for a fixed spacing. Knowing the normalized susceptance at some high frequency, and by selecting $l/\lambda_g$, the parameters follow a straight line to the origin as frequency is reduced to zero. By proper choice of spacing the insertion loss at lower frequencies will be equal to or less than its value at the high frequency. Taking the normalized susceptance of each diode to be 1.2 at the highest frequency and taking the diodes to be in the same plane ($l/\lambda_g = 0$), the maximum insertion loss will be 4 db. If they are spaced $l/\lambda_g = 0.2$ apart at the highest frequency, the maximum insertion loss will be 0.5 db. The insertion loss will follow the straight line to the origin as shown in Fig. 3. In 50-ohm TEM transmission line the normalized susceptance of 1.2 corresponds to a reactance of 41.7 ohms. Fig. 4 shows the frequency at which various capacitances give this reactance. Knowing the diode capacitance, the upper frequency as a function of maximum insertion loss can be determined directly from Fig. 4. For example, if two diodes each have a zero-bias capacitance of 1.0 pf and can be mounted so that the mounting contributes no capacitance, a limiter could be made that has a 4-Gc base-band insertion loss of 1.0 db by spacing them according to Fig. 3. An attempt was made to make a diode mount with zero mounting capacitance by using rectangular coax. It was found that the dielectric constant of the diode ceramic causes as much mounting capacitance in air rectangular coax as does the metal diode end protruding up in strip-line, which in both cases is about 1 pf. Thus the 1-pf diode becomes 2 pf and the base-band limiter can give less than 1.0-db insertion loss only up to about 2.5 Gc.

At high power levels the diodes are inductive. An inductance shunting a transmission line will give a hyperbola with changing frequency in the second quadrant of Fig. 3. Two diodes having a normalized susceptance of −4.5 each will give 20-db isolation if properly spaced and only 13 db if placed in the same plane. The high isolation curves are fairly insensitive to changes in $l/\lambda_g$, and the isolation improves with decreasing frequency. In a calculation similar to that used to obtain the insertion loss curves of Fig. 4, the isolation curves are derived. If in the future it is possible to mount diodes in strip-line without additional mounting capacitance and if the diodes have low whisker inductance and spreading resistance, then base-band limiters can be made to operate over the useful range of TEM transmission lines.

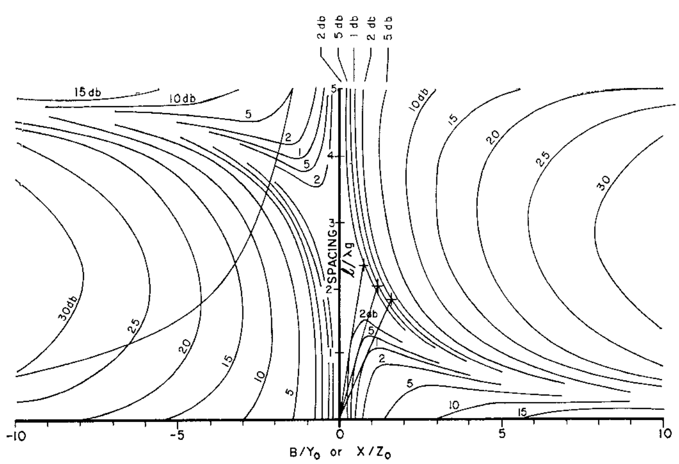


Fig. 3—Attenuation from two diodes as frequency is varied. At low power levels two shunt diodes follow a first quadrant straight line to the origin. At high power levels two shunt diodes follow a second quadrant hyperbola.

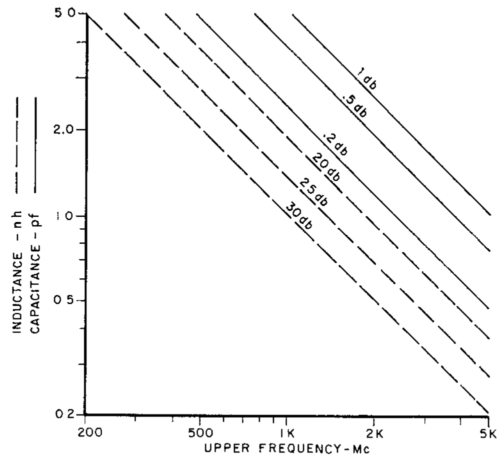


Fig. 4—Insertion loss and isolation of a baseband limiter using two diodes at optimum spacing as a function of upper frequency and diode capacitance and inductance respectively. $Z_0 = 50\ \Omega$.

### C. Precise Calculations to Determine Mount Capacitance

The diodes available today, however, have large $L_w$ and $R_s$; thus the diode resonance and losses must be taken into account. At zero bias, diode resonance will occur between 2 and 10 Gc and has to be taken into account at about one fifth of the resonant frequency for greatest accuracy. Calculations taking the resonance and losses into account can most easily be made as sketched in Fig. 5. The upper equivalent circuit is the same as that used for Fig. 3 but now it includes the losses of the diodes. The equivalent circuit of the diode at low power consists of a whisker inductance $L_w$, depletion layer capacitance $C_d$, spreading resistance $R_s$, and mount capacitance $C_m$ (which includes cartridge capacitance). At high power levels $C_d$ is shunted by conduction current. To facilitate the calculation for attenuation, the $\pi$ equivalent circuit of the length of line $l$ is used [17]. Voltage-current considerations then give the equation for attenuation shown at the bottom of Fig. 5. $Z_1$

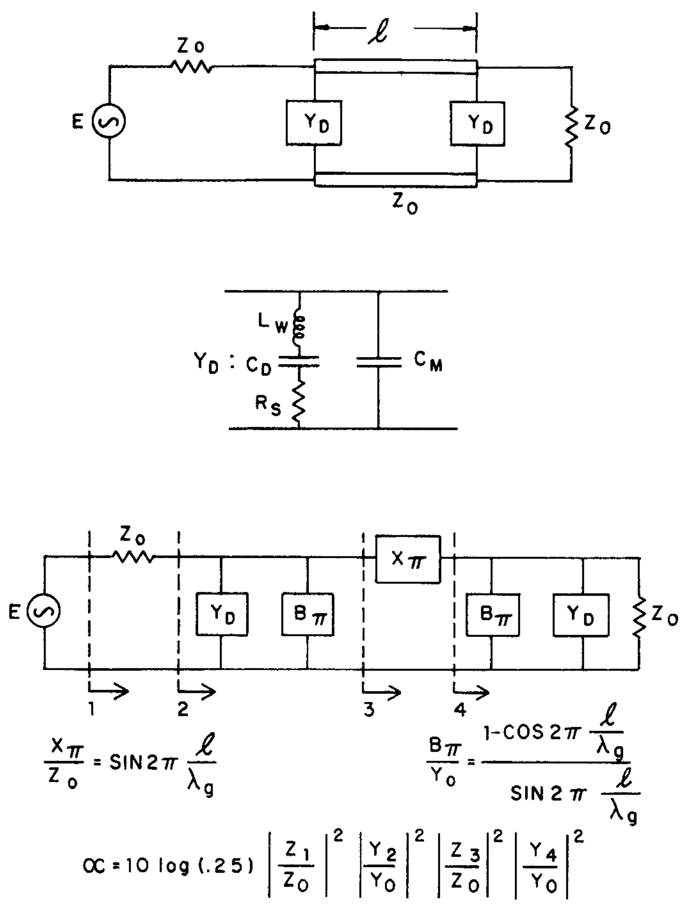


Fig. 5—Equivalent circuits and equations for calculation of attenuation of two diodes shunting a transmission line in which diode losses are taken into account. $|e/\lambda_g| \leq \frac{1}{4}$.

is the impedance of everything to the right of the 1 plane, etc.

Fig. 6 shows the calculated curves for insertion loss from 0 to 3 Gc for various values of $C_m$. The crosses are data from two pill-type varactors (MA4254, $C_0 = 1$ pf) in rectangular coax [18]. Several points should be noted. The inclusion of $R_s$ in the calculation raises the insertion loss in the neighborhood of 2.5 Gc having the effect of filling in the valley in Fig. 3. In fact for $C_m = 0$ pf in Fig. 6 the valley has almost completely disappeared. Note also that if $C_m$ could be sufficiently suppressed this diode would give less than 1-db insertion loss to all frequencies up to 3 Gc. Since the data points approximately follow the theoretical curves as if $C_m$ were about 0.9 pf, $C_m$ is taken to be 1 pf for this type of diode mount.

For $C_m = 1$ pf the insertion loss was calculated for various spacings between the diodes to determine how high in frequency a 1-db limiter could be made with these diodes. These results are shown in Fig. 7. Based on these curves a limiter was made in rectangular coax with a diode spacing of 0.9 inch which gave the insertion loss shown by the crosses in Fig. 7. Another limiter using the same diodes was made in teflon dielectric triplate strip-line. The spacing was reduced by the dielectric to 0.54 inch. Each diode mount consisted of an

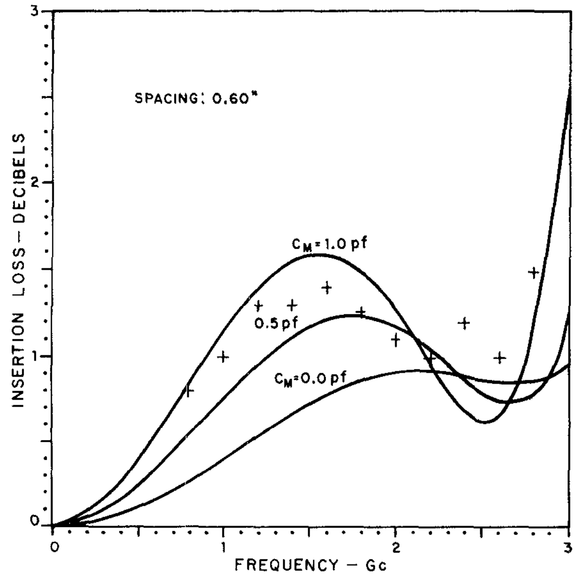


Fig. 6—Calculated insertion loss as a function of frequency and mount capacitance for two diodes spaced 0.6 inch apart assuming $L_w=0.6$ nh, $C_d=1$ pf, $R_s=4$ Ω, and $Z_0=50$ Ω. The data points are from a limiter made in rectangular coax.

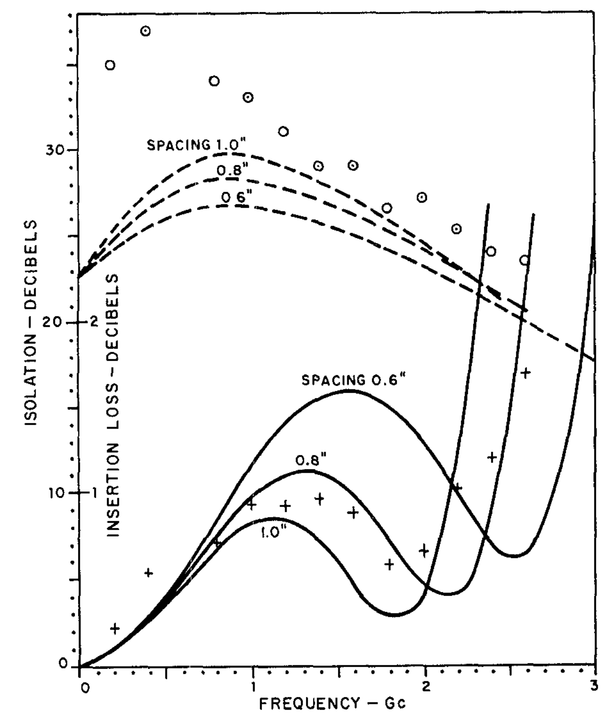


Fig. 7—Calculated insertion loss and isolation as a function of frequency and spacing assuming $L_w=0.6$ nh, $C_d=1$ pf, $C_m=1$ pf, $R_s=4$ Ω, and $Z_0=50$ Ω. The data points are for a limiter made in rectangular coax with 0.9 inch spacing between diodes.

8-32 screw hole tapped to the center strip. An 8-32 screw then secured the pill varactor. The insertion loss was not as well behaved as for rectangular coax but it was less than 1 db for all frequencies up to 2.5 Gc.

The high power isolation of the limiter was also calculated and is shown as dashed lines in Fig. 7 for the various spacings. The high power isolation data points were taken by having the diodes mounted in the same polarity and biasing them into conduction. All points are higher than the theoretical curves. Biasing causes minority carrier injection which gives a forward resistance lower than $R_s$ which is based on majority carrier current only.

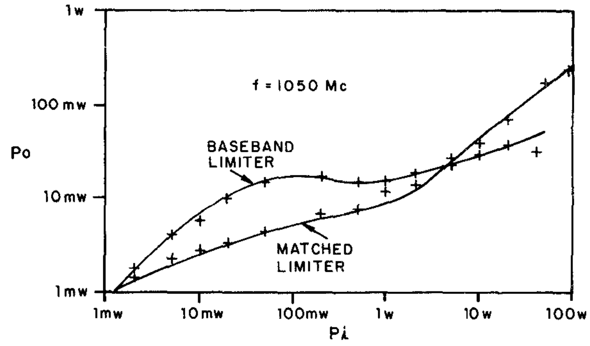


Fig. 8—Output power as a function of input power for the base-band limiter (Fig. 7), and matched limiter (Fig. 12). The pulse length was 0.1 μsec.

Fig. 8 shows the limiting characteristics of the two pill-type varactors. The measurements were made using a 0.1 μsec pulse at 1.05 Gc. No pulse leak-through was observed. Between 50-mw and 5-w input the output was 18 mw ± 1 db. The output is always limited to about 10 mw using silicon diodes shunting 50-ohm transmission line. Other limiting power levels may be obtained by biasing the diodes, using germanium or gallium arsenide diodes, or by changing $Z_0$.

## *D. Power Rating*

Using [2] the power ratings of the diodes used in the limiter of Fig. 8 were calculated. Because of approximations (to make the problem solvable) the results are expected only to indicate the order of magnitude of the burnout power. The mesa is considered to be a silicon cylinder of height $h$ and radius $a$, sitting on a semi-infinite silicon region. The thermal resistance of the mesa cylinder is given by the equation $R_m=h/\pi a^2K$, in which $K$(0.837 w/cm-°C) is the thermal conductivity of the silicon. The thermal spreading resistance is given by the equation $R_{sT}=1/4aK$. The heat is considered to be generated in the junction at the top of the mesa and dissipated only into the semiconductor. The experimental results interpreted with these assumptions in [2] indicate that the silicon junction is burned out at 650°C. A silicon diode at zero bias has a capacitance of about $(f_c/130+0.53)\times 10^5$ pf/cm² where $f_c$ is the cutoff frequency of the diode in Gc. This allows mesa radius to be calculated when $f_c$ and $C_0$ are known.

A diode in shunt can control average incident power $\overline{P_i}$, determined by the equation [16] $\overline{P_i}=\overline{P_d}(Z_0/4R_s+1)$. $\overline{P_d}$ is the power dissipated by the diode. With two diodes it is assumed that the total isolation is at worst caused by conductive mismatch. Fig. 7 shows the base-band limiter giving 30-db isolation at 1 Gc which corresponds to $R_s\approx 1$ Ω (see Fig. 2). Then $\overline{P_i}=13.5\overline{P_d}$. This is confirmed by VSWR measurements which indicate $\overline{P_i}=13.3\overline{P_d}$.

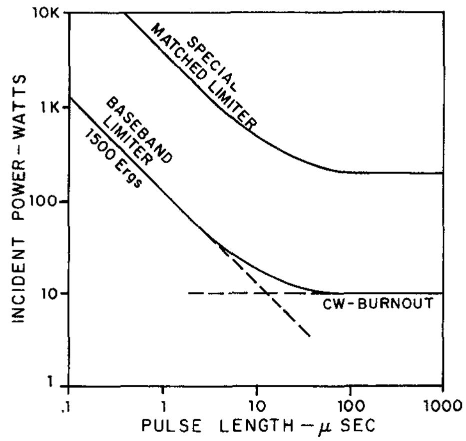


Fig. 9—Estimated burnout power of limiter in Fig. 7 and the special matched limiter (See Section II) as a function of pulse length.

To calculate the limiter burnout power it is assumed that the first of the two diodes gets most of the power. To calculate the diode burnout power, $\overline{P_{BO}}$, it is assumed that $h=0.0038$ cm, $C_0=1$ pf, and $f_c=100$ Gc. Then $R_m=635$, $R_s=198$, and the total thermal resistance is 833°C/w. The semiconductor end of the diode is taken to be $T_e$, the equipment temperature (25°C). Thus, $\overline{P_{BO}}=(650-T_e)/833\approx 0.75$ w and $\overline{P_i}=(0.75 \text{ w})\ (13.3) =10$ w.

In [2] the approximate thermal rise time is calculated. For the above diode 10 per cent of final temperature is reached in 1.3 μsec and 90 per cent is reached in 35 μsec. For pulses of duration less than 1 μsec it is assumed that no heat is dissipated, hence the burnout is determined by heat generated and the curve of Fig. 9 can be drawn.

## II. Matched Limiter Using 3-db Coupler

The second type of TEM diode limiter reported here makes use of a broad-band 3-db 90° coupler, the same as used at $X$ band. In strip-line the equivalent circuit of the configuration is shown in Fig. 10. The incident power is split by the 3-db coupler so that half of the power goes into each diode arm. At low power levels the diodes reflect the incident power, and the phases of the reflected waves add in such a way as to come out the output arm as shown in Fig. 10. At high power levels the diodes are conductive and the power isconveyed into the matched loads with very little reflection and thus little power out the output arm. The attenuation as a function of the diode impedance $Z=R+jX$ is given by the relation

$$\alpha = 10 \log \left[ \frac{\left(\frac{R}{Z_0}+2\right)^2 + \left(\frac{X}{Z_0}\right)^2}{\left(\frac{R}{Z_0}\right)^2 + \left(\frac{X}{Z_0}\right)^2} \right]$$

$$= 10 \log \left[ \left(\frac{2G}{Y_0}+1\right)^2 + \left(\frac{2B}{Y_0}\right)^2 \right].$$

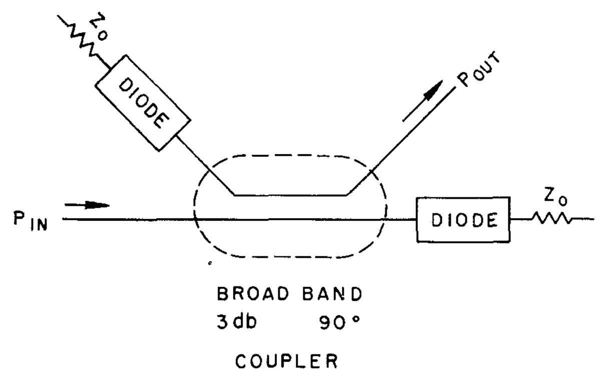


Fig. 10—Schematic diagram for matched limiter using two diodes in a 3-db coupler.

The attenuation as a function of frequency and diode parameters is shown in Fig. 11 for $Z_0=50\ \Omega$. Again these curves describe the bandwidth when the diodes are made to operate at higher frequencies by tuning with parallel inductors and series capacitors. This limiter has two advantages. First, it is matched for all power levels. Second, the high power passed to the matched load can be recombined again in another 3-db coupler and transmitted as with TR tubes.

It was originally thought that this limiter could be made with inexpensive gold-bonded diodes; however, their response time was too slow for pulse limiting and the output was not flat enough for CW limiting. Thus the limiter was made using the same pill varactors as above but now in a strip-line series mount. An inductor consisting of 10 turns of 0.004-inch wire on a 0.016-inch mandrel was used to tune the frequency of minimum insertion loss to 1.1 Gc. A 1- to 2-Gc 3-db coupler was used. For the calculations in Fig. 11 a perfect 3-db coupler was assumed. The 3-db coupler used could contribute as much as 0.5-db insertion loss but allows isolation greater than 30 db. The frequency dependence of the insertion loss of the limiter is shown in Fig. 12. The limiting is plotted in Fig. 8. From 0.9 to 1.3 Gc the VSWR of the limiter was less than 1.2 for all incident power levels.

The average incident power which a series diode can withstand is determined by the equation [16] $\overline{P_i}=\overline{P_d}(Z_0/R_s+1)$. Since the series diode cannot be mounted with the semiconductor end touching metal as in Section I it was necessary to measure the thermal resistance of the strip-line series diode mount. This was easily done by forward-biasing the diode in the mount, with a thermocouple between the "hot" end of the diode and the metal ground plane. The temperature drop divided by the dc power into the diode is the thermal resistance of the mount. The thermal resistance of the Sanders Associates mount MX24 is 71°C/w. This adds to the diode thermal resistance of 833 to give 904°C/w making

$$P_{BO} = \frac{650 - T_e}{904} \approx 0.7 \text{ w}.$$

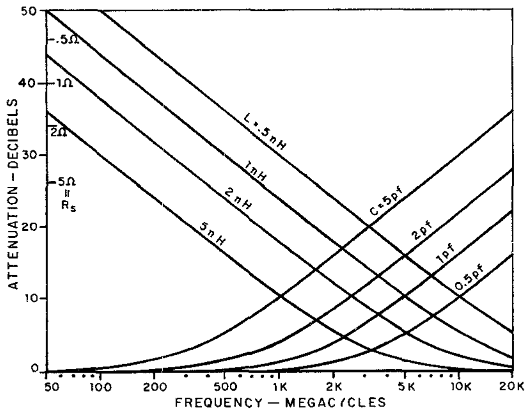


Fig. 11—Bandwidth of attenuation for diodes mounted in a 3-db coupler as functions of diode inductance and capacitance and the maximum attenuation as a function of spreading resistance. $Z_0=50\ \Omega$.

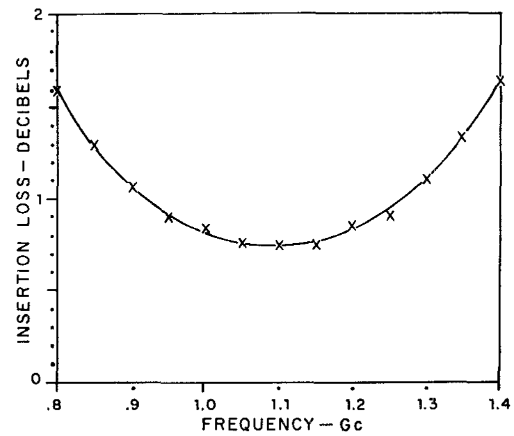


Fig. 12—Low power insertion loss of a matched limiter, using a tuned pill-type varactor in each arm of a 3-db coupler.

Since $R_s=4\ \Omega$

$$\text{then } \overline{P_i} = (0.7 \text{ w})(13.5) = 9.5 \text{ w}.$$

However, only half of the power incident to the limiter goes to each diode; thus this limiter can withstand about 20 w average power. Since the thermal time constants are about the same as for the base-band limiter, this limiter can therefore handle about twice the power shown for the baseband limiter in Fig. 9.

More power could be handled by increasing the mesa size. For example if the mesa radius were doubled,

$$\left.\begin{aligned} C_0 &= 4 \text{ pf} \\ R_s &= 1\ \Omega \\ R_M &= 159°\text{C/w} \\ R_{sT} &= 99°\text{C/w} \end{aligned}\right. \qquad \text{for } f_c = 100 \text{ Gc}.$$

Including mount thermal resistance, the total thermal resistance is 329.

$$\overline{P_{BO}} = \frac{650 - T_e}{329} \approx 1.9 \text{ w}$$

$$\overline{P_i} = (1.9 \text{ w})(51) = 97 \text{ w}.$$

Thus the limiter could withstand 200 w average incident power. The thermal time constants are increased by the larger mesa [2] to 2 $\mu$sec and 47 $\mu$sec respectively for the 10 per cent and 90 per cent points. The anticipated power-handling capability of this limiter is shown in Fig. 9 as the curve for the special matched limiter. The zero-bias capacitance of 4 pf allows the insertion loss to be less than 1 db over a 200-Mc bandwidth according to Fig. 11.

## III. Limiters Using the Diode Resonance

The two broad-band limiting techniques described thus far can be applied at frequencies below that of the diode resonance. With present varactors at zero bias these techniques are useful up to about 2 Gc. To make limiters above 2 Gc, it is necessary that 1) the diodes must be operated at a reverse bias to increase the frequency of the diode resonance (this changes the limiting power level and puts bends in the output curve), 2) diodes with resonance at a higher frequency may be built, or 3) limiters must be made which make use of the diode resonance.

From [16] it can be seen that a series diode operating in switching mode 2 will have the proper switching polarity for limiting. The equivalent circuits of this mode are shown in Fig. 13. At low power, $C_d$ series resonates with $L_w$ causing a low insertion loss. At high power, $C_d$ is shunted by conduction current allowing $L_w$ to parallel resonate with $C_c$. This makes the diode a high impedance in series with the center conductor; thus it reflects incident power.

Considering $R_s$ to be negligible, $C_c$ must equal $C_d$ for the lowest insertion loss and highest isolation to occur at the same frequency [16]. As frequency is varied above or below parallel resonance, the reactance of the resonant structure will be greater than $100X_{\text{res}}$, over a 1 per cent frequency bandwidth, and greater than $10X_{\text{res}}$ over a 10 per cent bandwidth, etc, where $X_{\text{res}}$ is the reactance of $L_w$ or $C_d$ at resonance.

The center frequency of the resonance is the geometric mean of the extremes rather than the arithmetic mean as implied by the use of per cent bandwidth. The reactance of the series resonant structure is less than $X_{\text{res}}/100$ over a 1 per cent bandwidth and less than $X_{\text{res}}/10$ over a 10 per cent bandwidth, etc. The presence of $C_c(=C_d)$ causes a parallel resonance at $\sqrt{2}$ times the frequency of the series resonance which causes the insertion loss to increase a little more rapidly with increasing frequency than it would without $C_c$. But at lower frequencies the insertion loss is slightly decreased by the presence of $C_c$. The over-all effect is to shift the skirts of the resonance down in frequency about as much away from linear as the geometric mean shifts it up for parallel resonance. Using the attenuation equa-

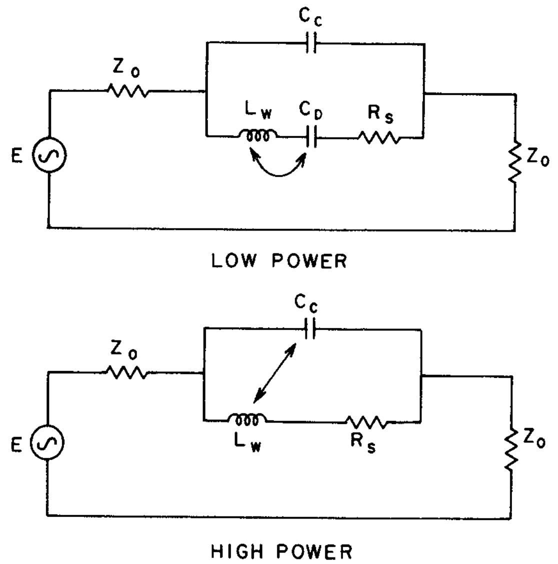


Fig. 13—Equivalent circuit of a resonant diode in series with a transmission line. At low power levels the series resonance between $L_w$ and $C_d$ causes the diode to be a low impedance thus having a low attenuation. At high power levels $C_d$ is shunted by conduction current allowing $L_w$ to parallel resonate with $C_c$. The resulting high diode impedance causes a high attenuation.

tion for a diode in series [16],

$$\alpha = 10 \log \left[ \left(1 + \frac{R}{2Z_0}\right)^2 + \left(\frac{X}{2Z_0}\right)^2 \right],$$

the isolation and insertion loss are calculated as functions of $X_{res}/Z_0$ and bandwidth and plotted in Fig. 14.

$X_{res}$ is determined by $L_w$ and the frequency. Typically, $L_w$ has discrete values of about 5 nh for large standard cartridge diodes, 2 nh for short standard cartridge diodes, and 0.7 nh for pill-type diodes.

The spreading resistance will place limitations on the insertion loss and isolation of a limiter using this resonant structure. At series resonance, the insertion loss will be limited directly by $R_s/Z_0$, since $X/Z_0$ is zero. At parallel resonance the diode impedance will be $R_s + X^2{}_{res}/R_s$ which will limit the maximum isolation. Using the diode cutoff frequency $f_c = (2\pi R_s C_d)^{-1}$ the maximum isolation and minimum insertion loss are plotted as functions of $X_{res}/Z_0$ and $f/f_c$ in Fig. 15.

For the purpose of showing the possibilities of this limiting technique, the problem is considered of making a limiter at 5 Gc with commercially available diodes, having the best possible bandwidth, 1-db insertion loss, and 20-db isolation. The best diodes today have cutoff frequency $f_c = 200$ Gc at maximum reverse bias. At zero bias, $f_c$ is approximately 80 Gc making $f/f_c$ about 0.06. Referring to Fig. 15 it is possible to go from 0.8 db to 29 db for $f/f_c = 0.06$ when $X_{res}/Z_0 = 3.5$. From Fig. 14 it is seen that the reactive component will limit the bandwidth to about 15 per cent with 0.2-db insertion loss and 20-db isolation. Actual calculation of the insertion loss will show that the 0.8 db and the 0.2 db do not add up to 1.0 db but are only about 0.8 db, adding approxi-

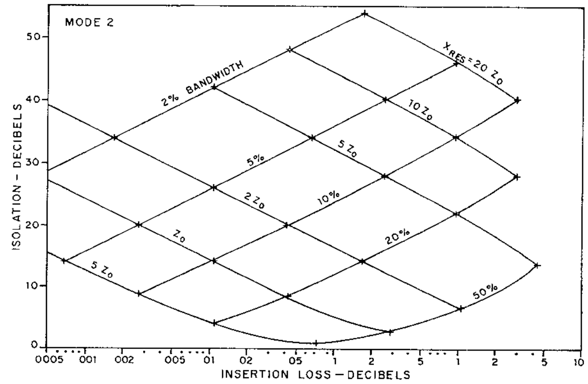


Fig. 14—Minimum isolation and maximum insertion loss of a limiter using a resonant diode in series with the center conductor as a function of bandwidth and $X_{res}$ (reactance of $L_w$ at resonance). These curves also describe the characteristics of a limiter made using resonant diodes in shunt in a 3-db coupler if $Z_0$ is replaced by $Z_0/4$.

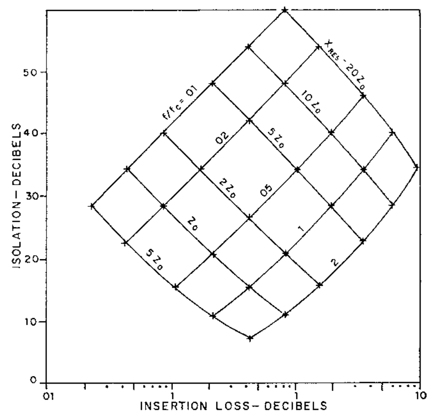


Fig. 15—Maximum isolation and minimum insertion loss of a resonant diode in series with the center conductor as a function of $X_{res}$, center frequency, $f$, and diode cutoff frequency, $f_c = (2\pi R_s C_d)^{-1}$. These curves also describe the characteristics of a limiter made using resonant diodes in shunt in a 3-db coupler if $Z_0$ is replaced by $Z_0/4$.

mately as their squares. The calculated characteristics of the limiter are as follows:

| Bandwidth | | Insertion Loss—Isolation |
|---|---|---|
| 5 Gc | | 0.8 db—29 db |
| 4.88—5.13 Gc | 5 per cent | 0.8 db—29 db |
| 4.75—5.25 Gc | 10 per cent | 0.8 db—24 db |
| 4.63—5.38 Gc | 15 per cent | 0.8 db—20 db |
| 4.5 —5.5 Gc | 20 per cent | 1.1 db—18 db |

A diode with $L_w = 2$ nh used for this limiter will require a zero-bias capacitance of 0.5 pf for the necessary 5-Gc resonance. To accomplish this a capacitive sleeve is built around the diode to make $C_c = 0.5$ pf [16]. Since $X_{res}$ is 63 ohms, $Z_0$ has to be 220 ohms to fulfill the $X_{res}/Z_0 = 3.5$ requirement. In a 50-ohm TEM transmission-line system, transformers of sufficient bandwidth have to be on each side of the diode. For narrow bandwidth the transformers are one-quarter wavelength long and of the mean impedance. In this example the diode is mounted in the center of a half-wavelength (3-cm) 105-ohm line section. It is also necessary to use quarter-wavelength stubs to act as dc ground returns on each side of the diode.

The limiting is expected to be flattest at the center frequency. Away from resonance the reactance variation of $C_d$ with power should cause slight variations in the power output curve. To make a matched limiter, two identical resonant diodes are mounted in shunt in each arm of a 3-db coupler.

The attenuation of a shunting diode in a 3-db coupler is a function of the reflection from the diodes.

$$\alpha = 10 \log \left[ \left( 2 \frac{R}{Z_0} + 1 \right)^2 + \left( 2 \frac{X}{Z_0} \right)^2 \right].$$

Figs. 14 and 15 apply for this configuration if $Z_0/4$ is substituted for $Z_0$. The curves then range from $X_{\text{res}} = 5\ Z_0$ to $X_{\text{res}} = 0.125\ Z_0$. The above example requires $Z_0 = 55$ ohms. The use of $Z_0 = 50$ ohms would not greatly alter the expected results.

Additional flexibility in impedance level is allowed by using multiple identical diodes. Two or more glass-encapsulated varactors could be mounted in the same sleeve, adding spreading resistance, "whisker" inductance, and depletion layer capacitance in parallel, and fabricating one capacitive sleeve. Two diodes, if they have a very low insertion loss and a low isolation, can be placed one-half wavelength apart to have very good additive isolation. The curves of Fig. 3 are cyclic in every $0.5l/\lambda_g$. Their isolations and insertion losses will approximately add. This relationship is useful with a pair of diodes used in any resonant mode as a switch or limiter.

## IV. Summary

Limiters are characterized by the following qualities: limited output power level, insertion loss, dynamic range, bandwidth, response times, power ratings, reflections, generation of harmonics, phase distortion, noise, temperature effects, and reliability.

An ideal limiter has constant output power for a wide range of input power above the limiting threshold level. The spreading resistance of diodes in diode limiters causes the output power to increase gradually with increasing input power instead of remaining constant. Low values of $R_s$ give the greatest degree of flatness. An inductive diode or interaction between multiple diodes cause variations in the power output. The limiting threshold level is determined by the voltage at which the diodes begin conduction, the characteristic impedance of the transmission line in which they are used, and the external dc circuitry of the limiter. The threshold power level can be increased 10 db by using a dc open instead of a short. The range of limiting is no greater than the maximum isolation which, along with the insertion loss, has been described.

The bandwidth (which is the theme of this paper) is portrayed in Figs. 2 and 4 for shunt diodes, Fig. 11 for series diodes in a 3-db coupler or circulator, and Fig. 14 for a resonant diode in series.

When varactors are used in limiters, response times are very small. The low $Q$ of broad-band limiters allows the incident voltage to appear across the diode junction with negligible delay time for buildup. In varactors most of the forward conduction is due to majority carrier current which has no long time delay associated with it. Thus the diode can go into conduction imediately when the power is increased allowing no spike leakage past the limiter. For high forward conduction there will be some minority carrier current which will have little effect on the forward resistance, but when the forward voltage is removed it will take time for the injected minority carriers to recombine. Thus the junction will have higher than normal conductivity which will return to normal in time as the carriers recombine. In a limiter this will cause a dead time in the form of an attenuation decreasing exponentially with time. At some power level no more minority carriers can be generated; thus there will be some saturation power level above which the dead time will not be increased. It should be possible to characterize dead time by plotting linear attenuation in decibels vs logarithmic time. With low voltage high cutoff frequency varactors, recovery times less than 10 nsec are common.

The power ratings of the diodes are estimated in Fig. 9 but should be established experimentally by testing diodes with increasing power up to burnout.

Diode limiters made in 3-db couplers are matched at all power levels with the reflections determined by the quality of the coupler. The other limiters discussed here reflect high powers.

Since the varactor diodes are operated with a dc short to give the lowest power limiting, it is quite possible that significant harmonics are generated. They will not be efficient harmonic generators, however, since most of the incident power is reflected and not converted by the diodes. If the harmonics are troublesome they can be removed by a low pass filter.

Phase distortion has not been considered.

The poor coupling between incident power and power absorbed in the diode is an indication that Johnson noise in the spreading resistance will not be closely coupled to the transmission line; thus the diode should contribute very little noise to the system aside from that addition to the signal-to-noise ratio from its insertion loss.

As the diode is heated at low power levels the zero-bias capacitance will increase slightly resulting in a slight increase in insertion loss. Heating the diode in the isolation state will reduce its power-handling capability as detailed in the power burnout equation.

If the diodes are used well below their burnout power they should provide the high reliability typically called "solid state reliability."

## Conclusion

The purpose of this paper is to show how broad-band TEM diode limiters can be made using TEM diode switching techniques.

Using today's diodes, it is possible to make a 2.5-Gc base-band limiter. If tomorrow's diodes can have "whisker inductance" of 0.2 nh, zero-bias depletion layer capacitance of 0.25 pf, cutoff frequency of 400 Gc, and mount capacitance of 0.25 pf, then two of these diodes can give a 10-Gc base-band limiter which would be described by Fig. 7 if the abscissa were multiplied by 4 and the spacing were divided by 4. The power rating of the limiter need not be decreased if the mesa height is reduced.

Very good broad-band matched limiters can be made using diodes in a 3-db coupler. Calculations indicate a limiter is feasible which will provide 200-Mc bandwidth and handle 200-w CW, or 1 $\mu$sec pulses up to 4-kw peak power. At higher frequencies it was shown that broad-band limiters can be made by controlling the diode resonance.

The details of this analysis are useful in designing diode switches as well as limiters. In that respect it can be considered to be an extension of [16]. In fact, Fig. 4 indicates that if the two diodes used for the base-band limiter are reverse-biased and spaced properly, a very good 4-Gc base-band switch can be made

$$(C_m + C_0 = 1.4 \text{ pf}).$$

It is noted that every diode switch will function either as a limiter or an expander when its biasing terminals are short circuited. Limiters find wide application while expanders do not. Some functions which expanders can perform include increasing the amplitude modulation index, decreasing pulse rise and fall time, pulse-forming, and (when combined with a reflective limiter and a $T$ junction) performing as a broad-band duplexer with very low loss.

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