parison of these results with the theoretical curves of Fig. 6 is not possible, but it can be seen that the general nature of the curves agrees well with the theoretical predictions. There appears to be a slight drift of the center quency of the limiter with increasing power level, which is attributed to a shift in the dc capacitance due to second-order effects, since the amount of this shift was dependent on the stiffness of the dc bias supply.

In general, it seems fair to conclude that excellent agreement is obtained between these measurements and the theoretical formulas developed earlier in this paper.

Conclusions

The passive parametric limiter offers several features, including, a) a sharp limiting threshold; b) flat power output for a substantial dynamic range above the threshold; c) little or no phase distortion; d) conveniently low-threshold power levels; and e) simple construction, with no auxiliary pumps, power supplies, or equipment (other than a simple dc bias supply) being required. Such limiters can be built at essentially any frequency from the audio to the microwave range, and may find employment either in protective applications or in signal-processing applications, particularly those where the phase-distortionless feature is essential. The present work has presented design equations and data for parametric limiters employing varactor diodes, together with experimental results at VHF and microwave frequencies which indicate the performance characteristics of such limiters. The excellent agreement between theory and experiment gives substantial confidence in the accuracy of the design equations.

References

[1] A. E. Siegman, "Phase-distortionless limiting by a parametric method," Proc. IRE, vol. 47, pp. 447–448; March, 1959.
[2] I. T. Ho, "Passive Phase-Distortionless Parametric Limiter," Stanford Electronics Labs., Stanford, Calif., Tech. Rept. 157-2; April, 1961; also Ph.D. dissertation, Dept. of Elec. Engrg., Stanford Univ., Stanford, Calif.; April, 1961.
[3] A. A. Wolf and J. E. Pippin, "A passive parametric limiter," 1960 Internatl. Solid State Circuits Conf. Digest of Tech. Papers, Philadelphia, Pa., pp. 90–91; February, 1960.
[4] R. W. Degrasse, "Low-loss gyromagnetic coupling through single crystal garnets," *J. Appl. Phys.*, vol. 30, pp. 155S–156S; April, 1959.
[5] G. S. Uebele, "Characteristics of ferrite microwave limiters," IRE Trans. on Microwave Theory and Techniques, vol. MTT-7, pp. 18–23; January, 1959.
[6] A. D. Sutherland and D. E. Countiss, "Parametric phase-distortionless *L*-band limiter," Proc. IRE, vol. 48, p. 938; May, 1960.
[7] F. A. Olson, *et al.*, "Parametric devices tested for phase-distortionless limiting," Proc. IRE, vol. 47, pp. 587–588; April, 1959.
[8] W. Shockley, "Theory of p-n junctions in semiconductors and p-n junction transistors," *Bell Sys. Tech J.*, vol. 28, pp. 436–489; July, 1949.
[9] A. Uhlir, Jr., "Varactors," Microwave Associates, Inc., Burlington, Mass., Brochure 59V; February, 1959.
[10] H. Heffner and G. Wade, "Gain, bandwidth, and noise characteristics of the variable-parameter amplifier." *J. Appl. Phys.*, vol. 29, pp. 1321–1331; September, 1958.
[11] J. M. Manley and H. E. Rowe, "Some general properties of nonlinear elements—Part I: General energy relation," Proc. IRE, vol. 44, pp. 904–913; July, 1956.
[12] J. Von Neumann, "Nonlinear capacitance or inductance switching, amplifier, and memory organs," U. S. Patent No. 2,815, 488; December 3, 1957.
[13] E. Goto, "The parametron, a digital computing element which utilized parametric oscillation," Proc. IRE, vol. 47, pp. 1304–1316; August, 1959.
[14] "X-band ferrite limiter," Hughes Aircraft Co., Component Div., Culver City, Calif. Interim Engrg. Repts. I–IV; September, 1959–June, 1960.
[15] E. T. Jaynes, "Ghost modes in imperfect waveguides," Proc. IRE, vol. 46, pp. 416–418; February, 1958.
[16] F. A. Olson and G. Wade, "A cavity-type parametric circuit as a phase-distortionless limiter," IRE Trans. on Microwave Theory and Techniques, vol. MTT-9, pp. 153–157; March, 1961.

Solid-State X-Band Power Limiter*

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***Summary*—An *X*-band solid-state power limiter has been designed and built to protect receiver crystals from high-power microwave pulses in the kilowatt region. This passive and reliable crystal protection has been achieved by utilizing the nonlinear properties of both ferrites and semiconductor diodes. An understanding of the ferrite nonlinear mechanism, which gives rise to the characteristically large leakage spike, has been achieved and quantitatively described. This formulation resulted in an essentially optimized high-power ferrite limiter, whose mode of operation is qualitatively understood. Use of this ferrite limiter for crystal protection requires a fast response, lower threshold secondary-limiting unit, which was developed by using semiconductor diodes for power limiting in a reactive mode of operation. The ferrite and diode limiters were combined in a single device with an over-all insertion loss of 2.0 db and a 200-Mc operating "bandwidth."**

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INTRODUCTION

THERE has been a long standing need in radar systems for a passive solid-state power limiter capable of providing protection for sensitive receiver crystals from high-power microwave pulses. Although present-day gaseous TR tubes perform adequately in many systems applications, their poor recovery time, somewhat questionable reliability, and relatively poor performance under high average power conditions limits their effectiveness in many of the newer radar systems. To determine whether a practical solid-state limiter is possible, extensive investigations have been conducted using both ferrites and microwave diodes as limiters.

A practical limiter must pass low-level signals through the limiter essentially unattenuated and attenuate high-power signals such that no more than approximately a half erg of spike energy and a half watt maximum of flat leakage reach the crystals. In addition, quick recovery time, high-power handling capability, broad-band operation, and reasonable size are important. In this paper, the study of nonlinear properties of ferrites and diodes has resulted in a power limiter which combines the attractive features of each nonlinear element. Though this ferrite-diode limiter does not yet meet all of the requirements of a practical device, the feasibility of constructing such a limiter has been established.

FERRITE LIMITER

Subsidiary resonance absorption in ferrites, first observed by Damon[1] and later explained by Suhl,[2] was first utilized for power limiting by Scovil.[3] An extensive study of this limiter was made by Uebele[4] who pointed out two problems which prevented a practical limiter design. First, a high amplitude spike resulted on the leading edge of the "limited" pulse and, second, the threshold power for limiting was too high for crystal protection. In all other respects the subsidiary resonance limiter is satisfactory and it would be a useful device if the two basic problems were solved. DeGrasse[5] has shown that ferrite instabilities at the main ferromagnetic resonance can also be used for limiting. The ferrite resonator is utilized as a band-pass filter; at low power the Q is high and a low insertion loss results. At high power, spin-wave excitation drastically lowers the resonator Q resulting in high loss. At S-band frequencies, the threshold power is low because subsidiary resonance and main resonance coincide, and these limiters have been effective as low-power limiters. Since Uebele's limiter has the greater potential for a high-power limiter, it was studied in detail and the result is a greatly improved limiter design.

The Leading Edge Spike Problem

The equations of ferrite behavior at high RF power levels, although complicated, present a clear-cut circuit analog. The uniform precession resonant circuit is coupled through a nonlinear reactance to the spin-wave resonant circuits. The reactance presents a negative resistance to the spin waves and amplification results. When the negative resistance exceeds the positive resistance, the spin waves of low (thermal) excitation rise in amplitude exponentially with time, creating a substantial loss to the uniform precession. The equation for this exponential rise is given by Suhl as

$$a_k = a_t e^{j\omega_k t} e^{(|\rho_k||a_0|-\eta)t}, \tag{1}$$

where

a_k = amplitude of kth spin wave,
a_t = measure of thermal level of the spin waves,
a_0 = uniform precession amplitude,
ρ_k = coupling terms from uniform precession to kth spin wave,
η = spin wave damping term.

This equation shows that if the uniform precession amplitude becomes large enough so that $|\rho_k|\ |a_0| > \eta$ (the threshold condition), then the spin-wave amplitude will increase exponentially with time. This increase would continue indefinitely were it not for the reaction of the spin wave back on the uniform precession. Therefore, the variation of the uniform precession can be described as follows: suppose a microwave pulse, of sufficient amplitude to be above the threshold value, is applied to the sample of ferrite. The uniform precession will tend to build up to the value given by the usual low-power theory within a time of order $1/\gamma\Delta H$ or about 10^{-8} sec and, initially, the precession angle is given by the conventional theory. Eventually the spin waves increase in amplitude, reducing a_0 and causing a decline of the exponential rate of increase of the spin waves. Finally, no further increase in the spin-wave level can take place and the steady state is attained with a_0 reduced to a value just below threshold. Fig. 1 gives a qualitative picture of the increase of the spin-wave amplitude and the decrease in output power at the signal frequency.

An exact transient analysis of the loss to the signal circuit would be very difficult to compute. However, by analysis of the beginning of the exponential rise of the spin waves, valuable information can be obtained about the time required for the initiation of loss to the signal

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[1] R. W. Damon, "Relaxation effects in the ferromagnetic resonance," *Rev. Mod. Phys.*, vol. 25, pp. 239–245; January, 1953.

[2] H. Suhl, "The nonlinear behavior of ferrites at high microwave signal levels," PROC. IRE, vol. 44, pp. 1270–1284; October, 1956.

[3] H. E. D. Scovil, private communication.

[4] G. S. Uebele, "Characteristics of ferrite microwave limiters," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-7, pp. 18–23; January, 1959.

[5] R. W. DeGrasse, "Low-loss gyromagnetic coupling through single crystal garnets," *J. Appl. Phys.*, suppl. to vol. 30, pp. 1555–1656; April, 1959.

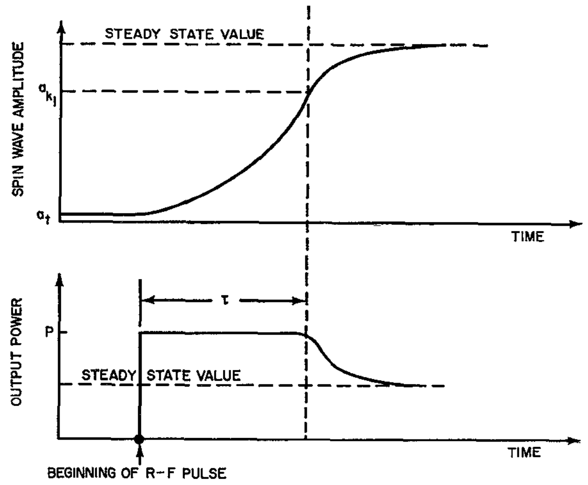


Fig. 1—Build-up of spin-wave amplitude and its effect on ferrite limiter output pulse.

circuit and the factors which enter into the amplitude and duration of the "spike."

Assume that the loss to the signal circuit will become evident when the spin waves reach an amplitude given by a_{k_1} as shown in Fig. 1. The time taken to reach this amplitude is τ if (1) is valid during this time. The logarithm of (1) gives

$$\ln\left|\frac{a_{k_1}}{a_t}\right| = (|\rho_k|\,|a_0| - \eta)\tau = L, \tag{2}$$

where L is a constant equal to the logarithm of $|a_{k_1}/a_t|$. Since $|a_0|^2$ is proportional to the input power P,

$$|\rho_k|\,|a_0| = K\sqrt{P}, \tag{3}$$

where K is a proportionality constant which depends on ρ_k, a materials parameter, and on the conversion from power to field, a ferrite geometry parameter. Eq. (2) can therefore be rewritten as follows:

$$\frac{1}{\tau} = \left(\frac{K}{L}\right)\sqrt{P} - \left(\frac{\eta}{L}\right). \tag{4}$$

The above equation assumes, of course, that the RF input pulse rises to its peak value much more rapidly than τ. The spike will be of the same amplitude as the input pulse and its duration will be given by (4).

An experimental verification of (4) is more difficult than might be anticipated because of the frequency modulation occurring at the leading edge of the usual magnetron pulse. In order to avoid this difficulty, an X-band CW magnetron was used as a power source in conjunction with a ferrite modulator. In this way, it became possible to eliminate frequency modulation and to obtain long pulse lengths (>10 μsec) which are necessary for the evaluation of the constants K/L and η/L. Data were obtained for a Ferramic R-4 ferrite slab, and the plot of $\sqrt{P_i}$ versus $1/\tau$ (see Fig. 2) showed the expected straight-line dependence.

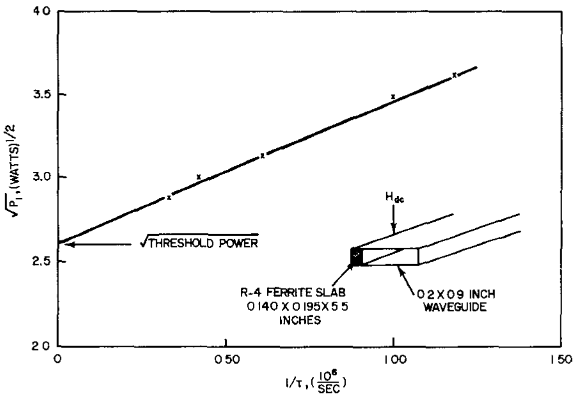


Fig. 2—This graph indicates the dependence of the spin-wave rise-time on the incident power for the slab structure shown.

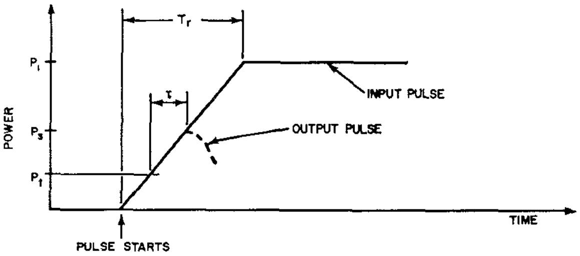


Fig. 3—Formation of the spike for an RF pulse of finite rise-time.

If the RF input pulse rises more slowly than τ, as shown in Fig. 3, then the above analysis must be modified. Here the spike amplitude P_s is no longer equal to the input-pulse amplitude but is smaller and is given by a cubic equation. This equation simplifies into the following:

$$P_s = \left(\frac{L}{K}\right)^2\left[\frac{1}{\tau} + \frac{\eta}{L}\right]^2, \tag{5}$$

$$P_i = \left(\frac{T_r}{\tau}\right)\left[P_s - \left(\frac{L}{K}\right)^2 \times \left(\frac{\eta}{L}\right)^2\right], \tag{6}$$

where the threshold power is given by

$$P_t = \left(\frac{\eta}{K}\right)^2. \tag{7}$$

One can easily show that $dP_s/dP_i = 0$ for large P_i. Keeping in mind that $P_s = P_i$ until $\tau = T_r$, a graph of P_s vs P_i is produced much like the CW power response of a long thin slab, *i.e.*, the spike has a "threshold" value of power and ultimately becomes limited. The spike amplitude P_s was actually measured for the Ferramic R-4 slab of Fig. 2 as P_i was increased to the maximum available from the CW magnetron source. Good qualitative agreement was obtained with (5) and (6) (see Fig. 4), and a modification of the equations using the true pulse shape, instead of the ideal linear rise, gave quantitative agreement.

The four parameters, T_r, L, K and η can be varied to eliminate or suppress the leakage spike. Taken in order, we note first that the pulse rise time, T_r, strongly affects the spike amplitude. Thus, as T_r is decreased, the spike amplitude should increase; in particular, the spike

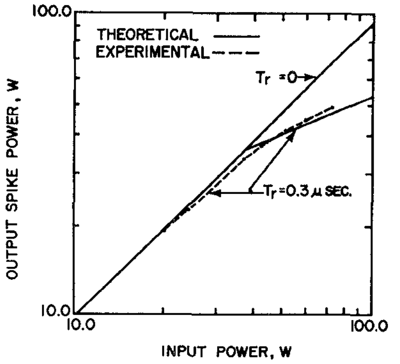
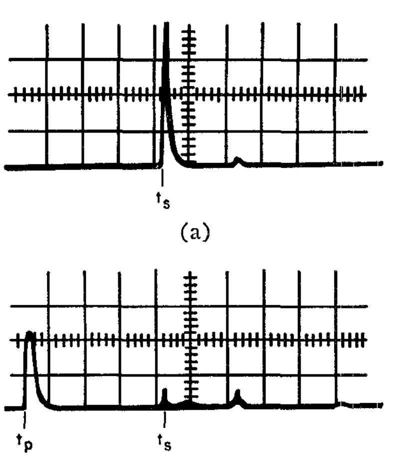


Fig. 4—Variation of spike amplitude with incident power for a spike of finite rise-time.

threshold power should rise as T_r is decreased. It is easy to alter the input pulse rise time, keeping L, K, and η constant, by placing various microwave filters with different Q_{ext} before the ferrite. For the Ferramic R-4 slab structure of Fig. 2, pulses with $T_r = 0.50$ μsec and $T_r = 0.03$ μsec yielded spike threshold power values of $P_i = 20$ w and $P_i = 200$ w, respectively. Simple calculations show good qualitative agreement with (5) and (6). However, the reduction of T_r in a practical device is probably not possible, since most radar systems employ RF pulses with sharp rise-times.

The effect of decreasing the quantity L, the logarithmic ratio of a_{k_1} to a_t, is to reduce P_s. Physically, this is equivalent to raising the ferrite spin temperature a_t or pre-exciting the spin waves. It is actually possible to pre-excite the spin waves by using a "primer-pulse" which, in addition to the signal, also excites subsidiary resonance. The following experiment[6] was performed to show the effectiveness of this pre-pulsing scheme. The pulsed outputs of two X-band magnetrons were synchronized so that a 2.5-μsec pre-pulse would envelope a 0.5-μsec signal pulse. Fig. 5 illustrates the observed output wave shapes with (a) only the signal pulse applied and (b) both the signal pulse and the pre-pulse applied. A decrease in signal spike amplitude can be observed in case (b). Ideally, both pulses should be at the same frequency for optimum effect since then the subharmonic spin-wave frequencies are identical, *i.e.*, the same spin wave is being amplified by both pulses. For different pre-pulse frequencies, more pre-pulse power is expected, and this behavior is borne out by the data plotted in Fig. 6. This pre-pulsing technique demonstrates the decrease expected of P_s with decrease in L, but obviously is not a practical method for spike suppression.

Thus, it appears that only η, the spin-wave damping parameter and K, a measure of geometric RF "field peaking" can conveniently be altered. The parameter η is selected according to which ferrite is used. What effect this selection will produce on the spike amplitude depends on the relative magnitudes of η/L and $1/\tau$ as can be seen in (5). If $\eta/L \gg 1/\tau$, then the spike amplitude is sensitive to changes in η, and at the same time the spike amplitude and the threshold power will be relatively large. On the other hand, if $\eta/L \ll 1/\tau$, the spike amplitude is insensitive to lowering η/L, and the selection of a better ferrite (lower η/L) will not appreciably affect the spike. From the equations, it can be noticed that increasing K always has a decisive effect on spike amplitude and threshold power level. Therefore, the greatest improvement in the spike response of the slab-type limiter can be achieved by selecting experimentally a material with the lowest threshold for instability in a given configuration and by then concentrating on field peaking techniques.



(a)

(b)

Fig. 5—Effect of spike of applying pre-pulse; t_p = start of prepulse, t_s = start of signal pulse, time scale = 0.25 μsec/cm. (a) Without prepulse. (b) With prepulse.

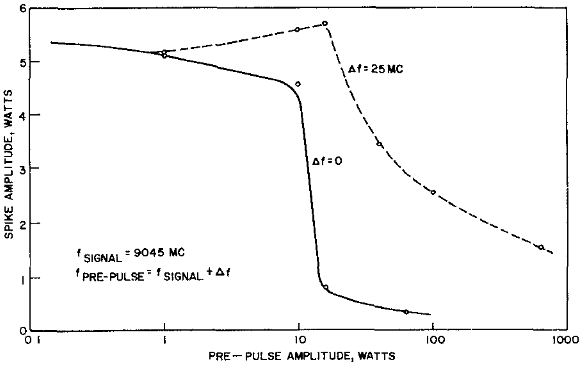


Fig. 6—Dependence of signal pulse spike on frequency and amplitude of prepulse. The signal pulse amplitude is 12 w.

Selection of Ferrite Material and Geometric Configuration

The first step in ferrite limiter design is to select the proper material. A suitable material must exhibit low loss at low-power levels, a low instability threshold, and high loss at high-power levels. Accordingly, threshold power and loss data were taken on ferrite slabs in waveguides and on spheres in a microwave cavity. The result is that for a subsidiary resonance limiter, General Ceramic's Ferramic R-4 was superior to all com-

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[6] This work was jointly done by Dr. P. C. Fletcher and T. S. Hartwick.

mercial and experimental ferrites examined. Since a figure of merit would be difficult to define, we state simply that this material exhibits a low-power loss comparable to Ferramic R-1 (the material used by Uebele), but a threshold power an order of magnitude lower than R-1. It should be noted that R-4 ferrite possesses the lowest threshold power measured. On the basis of Suhl's approximation for the spin-wave damping, $\eta = \gamma\Delta H/2$, the lowest line-width material should exhibit the lowest threshold power and this is not consistent with the broad ($\Delta H > 250$ gauss) line-width of R-4 ferrite. This behavior must be due to an effect produced by the unusual polycrystalline structure of the ferrite.

Typical of the thin-slab data is that shown in Fig. 7 of an 0.080-in R-4-slab 4 inches long placed against the side wall of standard X-band waveguide. The insertion loss was only 0.05 db and no deterioration of performance was noticed at large average power levels. For these reasons, such a configuration could be utilized as a "prelimiter," useful for dissipating the brunt of the average power in a large incident signal.

A low impedance microwave circuit must be selected which will produce a large K value and a reasonably low insertion loss. These requirements tend to be mutually exclusive since the ferrite always contributes a linear loss term. A low impedance circuit with excellent characteristics consists of the tapered ferrite slab, shown in Fig. 8, in reduced height X-band waveguide. Uebele showed that the threshold power for this structure was lowered by an order of magnitude over the threshold power for a thin slab at only a slight compromise in insertion loss. Therefore, this structure was studied further and was utilized in all of the main limiter designs.

Some insight into the operation of Uebele's limiter configuration can be gained by noting two important facts. First, experiments indicate that as the thickness of a thin slab is increased, the threshold power decreases. Second, at an input power just above threshold, Uebele's limiter produced a plateau power output less than the threshold power level, an effect that is hard to reconcile with the absorption mechanism involved. Both of these facts suggest that a transition occurs from a TE_{10} mode to a different mode in the ferrite-loaded waveguide section.

A perturbed TE_{10} mode propagates when H_{dc} is set to subsidiary resonance in a thin slab against the side wall of rectangular waveguide. It has been shown that for thick slabs,[7] appropriately biased with an H_{dc}, a surface mode (or ferrite-dielectric mode) can propagate. The RF field intensity becomes quite large at the ferrite-air interface for this low impedance mode. Thus, one might expect the tapered slab to support some combination of both modes and to actually furnish a gradual transition from TE_{10} mode at the slab ends to the surface mode at the slab center. Precisely this behavior was experimentally observed by probing the electric field in the slab-loaded waveguide section, and plotting the transverse variation of intensity as the length of the slab was traversed. Additional support for the moding effect is furnished by probing the RF field at high power. One expects the central portion of the slab, where the RF magnetic field is most intense, to become unstable first as the power is raised. For a power level just above threshold, the pulse shape is rectangular at the input end of the slab and remains so as the probe is traversed along the length of the slab, until, at the slab center the pulse becomes distorted to a spike and plateau.

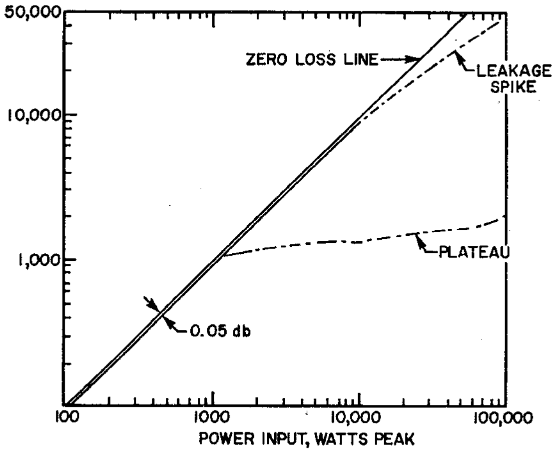

Fig. 7—Power response of a Ferramic R-4 thin-slab limiter.

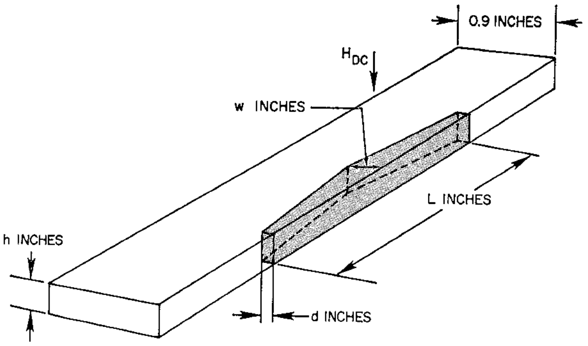

Fig. 8—Limiter geometry showing tapered ferrite-slab structure.

An analytical solution to the tapered-slab problem is difficult even in the absence of high-power effects and is probably insoluble if loss is included. Although the above reasoning is qualitative, the arguments are strongly supported by experiment and provide information to guide the design of the ferrite-waveguide structure.

Ferrite Limiter Construction

Several limiters were constructed using Ferramic R-4 ferrite and the limiter shape of Fig. 8. The limiter dimensions were systematically varied to obtain the lowest value of threshold power for an arbitrary insertion-

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[7] K. J. Button and B. Lax, "Theory of ferrites in rectangular waveguides," IRE Trans. on Antennas and Propagation, vol. AP-4, pp. 531–537; July, 1956.

loss value, usually taken as 1 db. Limiters resulted with an insertion loss of 1 db and a threshold power which varied from about 20 w ($h=0.400$ in) down to 5 w ($h=0.050$ in). Shown in Fig. 9 is the response curve of a typical limiter with dimensions $L=4.0$ in, $h=0.050$ in, $w=0.160$ in, and $d=0.080$ in. The insertion loss is less than 0.9 db over a 300-Mc band and the threshold power is less than 10 w over the same bandwidth.

Although variation of any one dimension affects all parameters specifying the limiter response, certain dependences are noticed. Thus, the main effect of decreasing the waveguide height is to lower the threshold power. The w and d dimensions primarily affect the insertion loss, since the match between modes is altered; small variations about $w=0.150$ in and $d=0.090$ in were generally found to provide a suitable insertion-loss value. The transverse dc magnetic field adjustment is not extremely critical, but large variations affect both the insertion loss and the threshold power.

It is clear that the limiter with the response of Fig. 9 cannot protect crystals over any power range. Even if the threshold power were lowered below the 1-w level, the leakage spike would prevent crystal protection. Since the microwave circuit design has been considerably improved, and is hopefully nearly optimized, it is natural to inquire whether a better ferrite material might aid in suppressing the spike. Eqs. (5) and (6), with the experimental constants $T_r=0.03$ μsec and $\eta/L=3.1$ $(\mu\text{sec})^{-1}$, taken from Fig. 2, yield the result that $\eta/L \ll 1/\tau$ and so it follows that the spike cannot be reduced by using a more nonlinear material. Hence, the slab-type subsidiary resonance limiter is useful for crystal protection only if it is followed by a secondary limiter which will eliminate the leakage spike.

Since the output power of the slab-type limiter is relatively low, one is encouraged to re-inspect single crystal garnet limiters. However, YIG single crystal limiters operating at both subsidiary resonance and main resonance can be ruled out as secondary limiters. At subsidiary resonance the high Q cavity, necessary to obtain a low threshold, creates a high insertion loss and severely reduces the bandwidth. The main resonance limiter, useful at S band where the coincidence of main and subsidiary resonance instabilities creates an extremely low threshold, requires a microwave cavity at X band and is narrow band. In addition, a large spike is observed on the output pulse thus making this type of device unsuitable as a secondary limiter.

DIODE LIMITERS

It has been realized for some time that the nonlinear properties of diodes can be used for passive power limiting at low power levels where diode power dissipation is low. An investigation was therefore undertaken to determine if a diode could handle a ferrite limiter spike in the 100 erg region and a flat leakage of 20 w and thus serve as a secondary limiter.

One type of diode limiter makes use of the diode nonlinearity to convert signal power, above a threshold level, into subharmonic oscillations[8,9] as in the ferrite case. This type of device is narrow band and generally is restricted to limiting power levels below 1 w. A second scheme, better suited to our purposes, utilizes the diode nonlinearity to change the admittance presented to the RF wave, as was shown by Garver and Tseng.[10] The essential advantage of this technique is that the admittance change can be used to switch the high power into a load resulting in little power dissipation in the diode itself. To predict the admittance of a diode as a function of the applied RF power, that is, the limiting characteristic, the equivalent circuit of the diode and mount is needed. Although substantial efforts have been made to determine the equivalent circuit,[11] an accurate model of the diode plus mount has not yet been given.

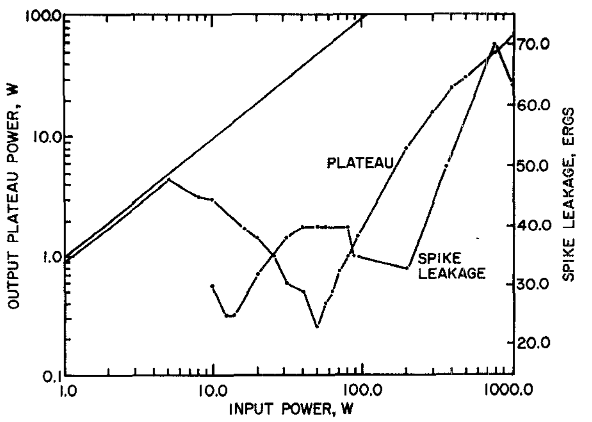

Fig. 9—Power response of Ferramic R-4 tapered-slab limiter

Diode Limiter Configurations

In order to study in more detail the possible limiter circuits and diode limiter spike response, a 1N263 diode limiter was constructed. The diode was mounted across the waveguide at the E-field maximum with the diode axis parallel to the narrow waveguide wall. The diode mount introduced little capacitance and provided a way to apply a dc bias voltage to the diode. A matched load terminated the diode mount and the admittance of the diode plus matched load was measured as a function of RF power. The results, shown on the Smith chart (Fig. 10), clearly demonstrate the power-switching technique. Thus, for input powers below 0.2 mw an extremely large VSWR is presented to the RF wave resulting in nearly total reflection; above this power level

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[8] A. A. Wolf and J. E. Pippin, "A passive parametric limiter," *Digest of Tech. Papers*, Solid-State Circuits Conf., Philadelphia, Pa., pp. 90–91; February, 1960.

[9] A. E. Siegman, "Phase-distortionless limiting by a parametric method," PROC. IRE, vol. 47, pp. 447–448; March, 1959.

[10] R. V. Garver and D. Y. Tseng, "X-band diode limiting," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES (Correspondence), vol. MTT-9, p. 202; March, 1961.

[11] R. V. Garver and J. A. Rosado, "Microwave diode cartridge impedance," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-8, pp. 104–107; January, 1960.

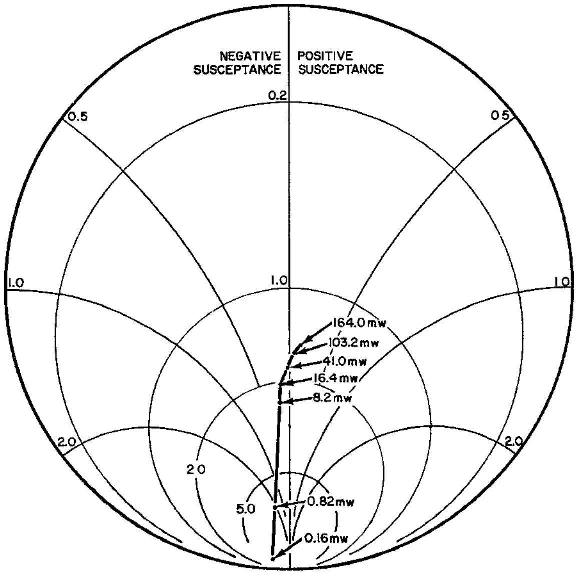

Fig. 10—Variation of admittance with RF power for 1N263 diode and mount.

both the conductance and susceptance change rapidly with power, little power is reflected, and the remainder is split between the diode and matched load. Since the diode conductance is much smaller than the load conductance, and the diode susceptance is also small, most of the incident power is dissipated in the load. This behavior is ideal for power limiting because of the low power dissipation in the diode.

In an actual reflection-type limiter model, means must be provided for separation of input and output signals. This may be accomplished by connecting two diode mounts, terminated in matched loads, to a 3-db broad-band hybrid.[10] The input signal splits at the junction with a 90° relative phase shift, half of the power going to each diode section. At low signal level most of the power incident on each diode is reflected, and the phases are such that the reflected signals add in the hybrid output arm and cancel in the hybrid input arm. At high signal levels the power is largely transmitted to the load and the signal at the hybrid output port is correspondingly reduced. This hybrid limiter was constructed and effectively limited 2.5-μsec signal pulses for up to 5 w amplitude with a threshold power of 1.0 mw and an insertion loss of 0.6 db. Two improvements were made in this design which left the insertion loss unchanged, but increased the power-handling capacity to 15 w. First, a dual-diode mount was built by placing two 1N263 diodes side-by-side in the transverse plane; second, 10-kilohm resistors were placed in the diode dc-return to limit the current flow. Both improvements decrease the current flow across the diode junction and, in this way, less power is dissipated per diode.

This hybrid diode limiter is useful only if it can limit the sharply rising leakage spike, produced by the ferrite limiter. By using the ferrite power limiter as a spike signal source, the spike-limiting response was obtained. The results prove the diode limiter can limit a leakage spike of 0.020-μsec width for as much as 100-w amplitude (20 erg). However, at more than a 10-erg input a gradual diode deterioration was noticed for long exposure times (several minutes). This effect restricts the application of the limiter as a secondary limiter, and points out the need for a limiter diode with a larger power-handling capacity.

Varactor Diode Limiter

Mesa-type varactor diodes possess a large contact area and are able to dissipate a larger amount of power, without diode deterioration, than the 1N263 point-contact type of construction. Accordingly, several varactor diodes, including MA460A, MA460E, MA460F, MA450A, MA450F, and some experimental diodes, were evaluated; only the double-ended MA450F diode, back biased to -5 volts, offered the desired characteristics. At input powers below 500 mw the reflection loss was a low 0.2 db, and the reflection loss increased to 10 db for an input power above 500 mw. This behavior, ideal for constructing a secondary limiter, is probably due to the rapid decrease in resistance associated with the avalanche breakdown. By contrast, the other diodes exhibited very little change in admittance as the power was varied, and were generally lossier.

It is significant that the diode-rectifying structures in the MA460 and MA450 series are electrically equivalent, and that their microwave response differs only because they are enclosed in different types of packages. The MA460 series of diodes which uses a single-ended cartridge package proved to be ineffective for limiting. In contrast, the MA450 diodes with a double-ended package (similar to the 1N263 package) gave consistently better results. The performance difference between the MA450F and the MA450A is due to the change in diode cutoff frequency which probably tunes the diode series resonance.

An experimental limiter was constructed using four MA450F diodes biased to -5 volts, in conjunction with a hybrid section, as discussed earlier. The dc external loading resistors used here, however, were 470 kilohms. It was also found possible to position a tuning screw a distance $\lambda g/2$ behind the diodes and greatly enhance the amount of isolation. This resulted in the power response curve shown in Fig. 11. The low-power insertion loss of this device is 0.4 db, with a threshold power level of 1 w; these values are constant over a 200-Mc band centered about 9300 Mc. For this limiter, the leading edge spike has a width less than 10 nsec. If the avalanche breakdown mechanism is principally responsible for the diode admittance change, then the spike data is not inconsistent with the relatively long time constants appropriate to the avalanche. Since the spike rapidly becomes narrower as power increases, and is also even-

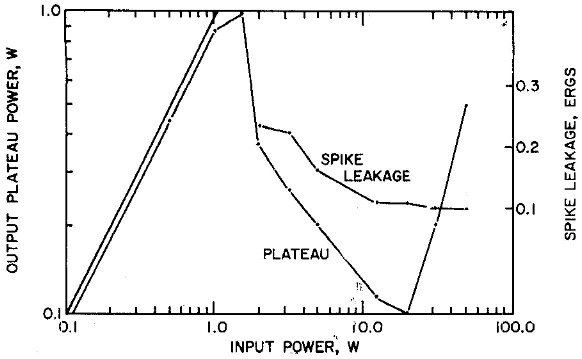

Fig. 11—Power response of MA450F hybrid limiter.

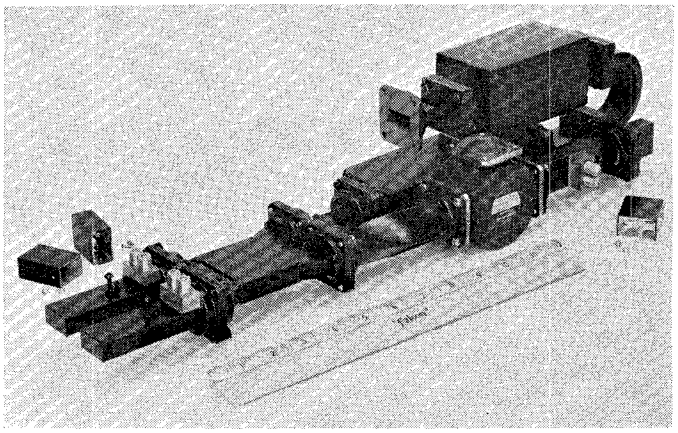

Fig. 12—Photograph of four-section solid-state limiter.

tually limited, its presence does not abort the limiter's effectiveness as a secondary limiter. Again using the ferrite limiter as a spike source, over 15 db of spike limiting was measured for input-leakage spike amplitudes of 1500 w (300 ergs) with no noticeable deterioration of the MA450F diodes.

A three-port circulator for the separation of the input signal from the output signal provides a simple method of obtaining a threshold power just one half that of the hybrid limiter. Here the input signal, incident on the diode section, is either reflected into the output port or transmitted to the load, depending on the signal level. Such a device was built and it exhibited less than 0.4-db insertion loss, and over 10 db of isolation for input powers above 500 mw.

These two limiter configurations provide the secondary limiter characteristics needed to build a passive solid-state limiter useful for crystal protection.

SOLID-STATE LIMITER

Since the diode and ferrite limiters have been constructed with the desired individual characteristics, the problem has been reduced to combining the individual units into a workable solid-state limiter. In doing this, emphasis has been placed on demonstrating the feasiability of eventually constructing a model which will be useful in radar sets. Therefore, the procedure has not been to redesign compatible limiter stages, but rather to adjust the characteristics of the existing limiters so as to obtain limiting over a wide power range with a reasonably low insertion loss, taken here as 2 db.

Accordingly, a four-stage limiter was constructed consisting of a ferrite prelimiter, a ferrite main limiter and two stages of diode limiting. The completed limiter model is shown in Fig. 12. The ferrite prelimiter consists of a tapered slab of ferrite in 0.200-in high X-band waveguide similar to that shown in Fig. 8. For an insertion loss of only 0.3 db, a threshold power of 100 w was obtained.

The main ferrite limiter, which was previously discussed, exhibits the limiting response shown in Fig. 9. At the design frequency of 9300 Mc, the insertion loss is 0.8 db. By combining the prelimiter and main limiter, a plateau power output less than 20 w was obtained. However, upon combining the two limiters, a double leading edge spike resulted and some care had to be exercised in adjusting the magnetic field to minimize this effect.

Following the two ferrite limiters, both a hybrid MA450F diode limiter and a circulator MA450F diode limiter were used in order to reduce the spike and plateau level to a value sufficiently low to provide complete crystal protection. The characteristics of these diode limiters have been described in detail in a previous section. Briefly, these two diode limiters in tandem have an insertion loss less than 1 db over a bandwidth greater than 200 Mc and a threshold power of 500 mw.

In combining the various stages of this limiter, a variety of adjustments are required in order to match the individual characteristics to obtain an optimum device. These adjustments were most easily accomplished by duplexing a low-power signal modulated at 1000 cycles together with the high pulsed signal. Thus, the effect of any adjustments on both low- and high-power performance could readily be determined.

The performance characteristics of this complete power limiter are as follows:

Center frequency—9300 Mc,
Insertion loss—2 db,
Maximum input power (for complete crystal protection)—5000 w,
Bandwidth (over which crystal protection is provided)—200 Mc,
Spike leakage—less than 0.6 erg.

The complete power response is plotted in Fig. 13. In addition to the above measurements, the effectiveness of the limiter in protecting a balanced mixer using 1N23E crystals was tested. No deterioration of the 8.8-db-mixer noise figure could be measured after the limiter-mixer combination was subjected to a 5-kw peak input signal. This test clearly demonstrates the effectiveness of the limiter for crystal protection.

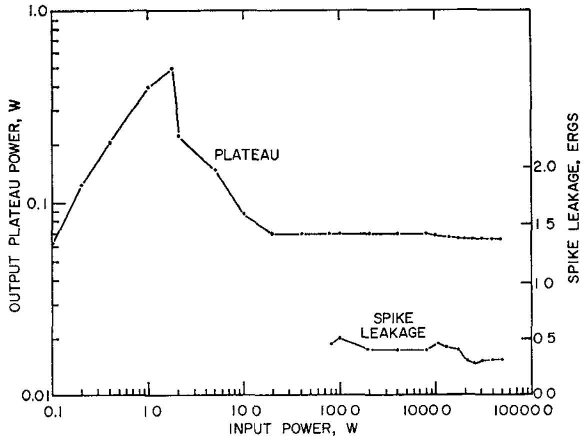

Fig. 13—Power response of four-section solid-state limiter.

Conclusion

The X-band limiter described in this paper indicates the feasibility of building an all solid-state device capable of complete crystal protection. Obviously, a substantial engineering effort is still required before this type of limiter can become practical for system use. Thus, a reduction in size, weight, cost and insertion loss is vital in order for this device to compete successfully with presently available TR tubes. Nevertheless, such improvements will undoubtedly come, particularly in those systems where the present TR tubes are not satisfactory.

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On Spurious Outputs from High-Power Pulsed Microwave Tubes and Their Control*

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Summary—**Excessive spurious outputs from high-power pulsed magnetrons, klystrons and traveling-wave tubes can cause intolerable radiation interference and deleterious effects in a high-power microwave system. The harmonic output from a klystron may vary appreciably with changes in operating conditions. Harmonic outputs from tubes cannot be eliminated but their radiation can be significantly reduced by using filters. If the parasitic or spurious oscillations are very strong, adverse effects such as amplitude and phase instability of the fundamental frequency output may occur. Some of the spurious outputs may be reduced or eliminated by redesign of the tube or its modulator.**

Introduction

THE emission of spurious outputs from microwave tubes has been known for a long time. Spurious outputs are defined as those frequency components other than the fundamental frequency with its normal sideband modulation components.

With steady increase in transmitter power level, receiver sensitivity, and density of radiating equipment, the problem of spurious outputs has taken on greater significance in terms of radiation interference [1], [2]. As the power levels of high-power tubes have increased so have the spurious output power levels. In a microwave system, the presence of spurious power may have deleterious effects such as arcing in chokes, arcing at flanges and undesired leakage through ionized duplexers. In addition, if a large amount of spurious power is generated, the tube efficiency may be decreased; it may cause other harmful effects to the tube and an objectionable amount of amplitude and phase instability may be added to the fundamental frequency output.

The presence of spurious frequencies usually can be detected at the tube output, provided that the spurious frequency is above cutoff of the output transmission line. If the spurious frequency is below cutoff and the spurious signal amplitude is sufficiently large, its presence may be inferred from any phase and amplitude instability of the fundamental frequency output. This instability may adversely affect the system of which the tube is a part without interfering with other nearby systems.

Although the frequencies can be measured with relative ease, the power levels are much more difficult to ascertain [3]–[6]. Spurious outputs other than harmonics are often quite erratic. Since a methodical redesign study of a high-power tube is generally costly,

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