inch, and a thickness of 0.125 inch were tested. The gap was varied from 0.030 to 0.070 inch and the length was varied from 0.320 to 0.280 inch without any sign of the breakdown at the peak power of 240 kw. The corners and edges of these chokes were not rounded in any way. These high-power measurements indicate that both types of chokes are able to carry full rated waveguide power.

Remarks

While it is possible to use the serrated choke on waveguides other than rectangular in cross section, it should be noted that since effective gap impedance is actually nonzero, higher order modes are generated at gap.

In considering the use of the new choke for specific applications and configurations, the array of choke pins and its image plane can be interchanged. Also greater utility may be achieved by rotating the plane of the chokes to any other angle about an axis parallel to the longitudinal axis of the waveguide.

Acknowledgment

The initial limited development on this choke, carried out in 1950, was a part of a program sponsored by the Bureau of Ships under Contract No. Nobsr-42419. The greater portion of the data presented here was obtained under a subsequent contract AF30(120)-440 with the Rome Air Development Center.

Microwave Filters Utilizing the Cutoff Effect

P. A. RIZZI†

Summary—**Two band-rejection microwave filters employing the waveguide cutoff effect are discussed. One type utilizes the cutoff property in the series arm of an *E* plane tee to improve the filter's characteristics, while the other utilizes this property in the *E* and *H* arms of a magic tee. Experimental data for both single and multistage filters are presented. Methods of obtaining low standing wave ratios over a broad pass-band are also presented.**

Introduction

IN GENERAL, the design of microwave filters has been treated from an equivalent circuit point of view. That is, low frequency filter theory is used to determine the configuration and from this the microwave analog is constructed. For example, by using resonant cavities and irises, the microwave equivalent of the series, shunt, and ladder type filters have been made.[1–4] In addition, *m* derived filter theory has been applied to microwaves for the design of band-rejection filters.[5] On the other hand, the design of a microwave filter sometimes entails the use of a property peculiar to waveguides. An example of this is the cutoff filter. In normal waveguide theory, the application of the boundary conditions at $x=0$ and $x=a$ (Fig. 1) to Maxwell's field equations reveals the fact that the electromagnetic wave (rectangular TE_{10} mode) will only propagate unattenuated above the frequency f_c, where

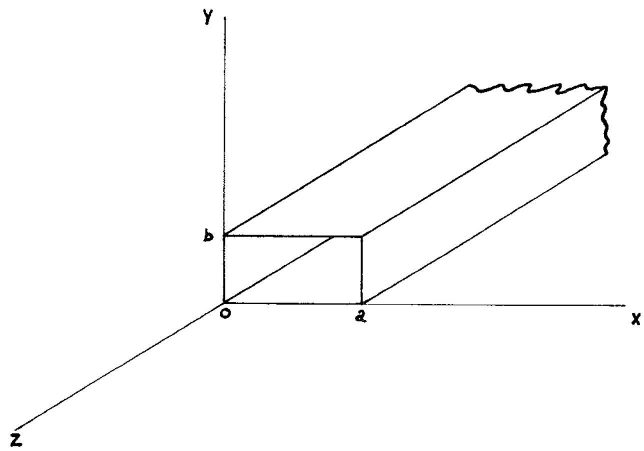

Fig. 1—Rectangular waveguide structure.

$$f_c = \frac{1}{2a\sqrt{\mu\epsilon}}.^{6} \tag{1}$$

For frequencies below f_c, the wave will attenuate at the rate of α nepers per meter, where

$$\alpha = \frac{\pi}{a}\sqrt{1-\left(\frac{f}{f_c}\right)^2}. \tag{2}$$

† Raytheon Mfg. Co., Missile and Radar Div., Bedford, Mass.

[1] R. M. Fano and A. W. Lawson, Jr., "Microwave filters using quarter-wave couplings." Proc. IRE, vol. 35, pp. 1318–1323; November, 1947.

[2] W. W. Mumford, "Maximally-flat filters in waveguide," *Bell Syst. Tel. Jour.*, vol. 27, pp. 684–713; October, 1948.

[3] W. L. Pritchard, "Quarter-wave coupled waveguide filters," *Jour. Appl. Phys.*, vol. 18, pp. 862–872; October, 1947.

[4] J. Reed, "Low-Q microwave filters," Proc. IRE, vol. 38, pp. 793–796; July, 1950.

[5] M. E. Breese and S. B. Cohn, "Diplexing Filters," 1954 IRE Convention Record, Part 8, "Communications and Microwaves," pp. 125–133.

[6] The Rationalized MKS system of units will be used in this discussion.

Thus it is seen that the rectangular waveguide (or, in fact, any closed waveguide) is, by virtue of the cutoff effect, a high-pass filter. Consequently if a high-pass microwave filter is desired which passes frequencies above f_1, one merely sets the width of the waveguide (a) so that f_1 is the cutoff frequency of the waveguide section. That is, let

$$a = \frac{1}{2f_1\sqrt{\mu\epsilon}}\cdot \tag{3}$$

By adjusting the length of the cutoff section, any required attenuation at frequencies below f_1, can be obtained.

Having reviewed the manner in which the cutoff property can be used to design a high-pass microwave filter, we will now show how this effect in conjunction with waveguide tees can be used to construct band rejection and possibly low-pass filters.

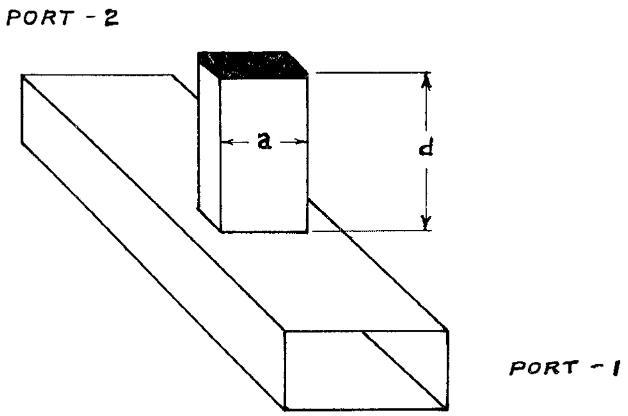

Fig. 2—A single-stage E plane cutoff filter.

The E Plane Tee Cutoff Filter

One method of constructing a band-rejection filter is to make use of the cutoff effect in the series arm of an E plane tee.[7] Fig. 2 shows a single stage E plane cutoff filter. At frequencies below the cutoff frequency of the series arm (1) and above the cutoff frequency of the main guide, the microwave energy will be transmitted from port 1 to 2 with practically no loss. That is, since the series arm is beyond cutoff, the wave will "see" it not as a transmission line but as a small reactive element in series with the main guide. Above the cutoff frequency of the series arm, the section behaves like a normal series tee; namely, a six terminal network. Therefore, by shorting the series arm a distance d from its input terminals, the filter section will reject frequencies at which d is approximately a quarter wavelength long. By choosing the cutoff frequency of the series arm close to the resonant frequency of the shorted series arm, one obtains an attenuation curve which rises very rapidly with frequency. A typical attenuation curve for a single stage filter is shown in Fig. 3.

[7] This can also be done with an H plane tee, but each filter section is physically longer.

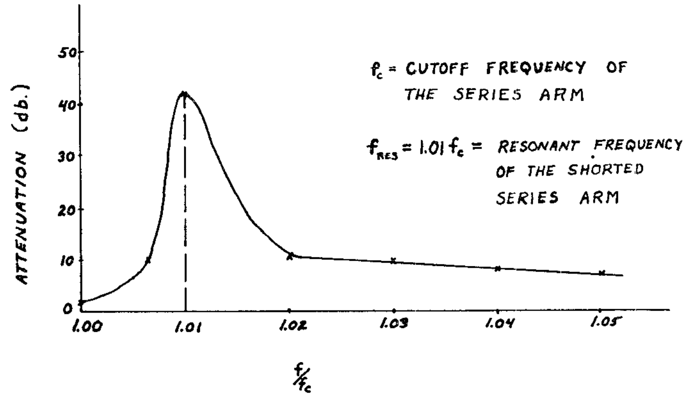

Fig. 3—Attenuation curve for single-stage E plane cutoff filter.

At frequencies below the cutoff frequency of the series arm, the vswr of the filter section is practically constant (Fig. 4). Since the series arm is below cutoff ($f/fc<1$) in the filter's pass band, the vswr is independent of the shorting distance d. This is very important from a practical viewpoint since it allows the problem of good vswr in the pass band and good rejection in the stop band to be treated separately. That is, the filter sections may be spaced so as to give a good impedance match across the pass band. Then the shorts may be adjusted to give good rejection in the stop band without in any way affecting the match in the pass band. Since the tuning of the series-tee cutoff filter, unlike the quarter-wave coupled filter, is independent of the pass band match, the individual stages may be stagger-tuned. Consequently a given rejection over a frequency band can be obtained with a smaller number of filter elements.

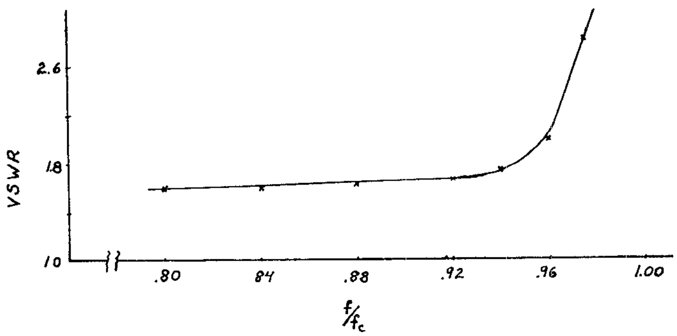

Fig. 4—Vswr for single-stage E plane cutoff filter.

Three types of E plane tee cutoff filters which have been made for application at X band are shown in Fig. 5. The vswr in the pass band and the attenutaion in the stop band for the filter shown in Fig. 5(a) is given in Fig. 6. By spacing the first two filter sections $\lambda/4$ apart, the reflections from the individual sections tend to cancel. Likewise $\lambda/4$ spacing for the next two sections also causes the reflections to cancel. Finally by spacing the two *pairs* $3/4\lambda$ apart, second order cancellation is obtained. That is, at frequencies where the mismatch of each pair becomes bad, the $3/4\lambda$ spacing tends to cancel

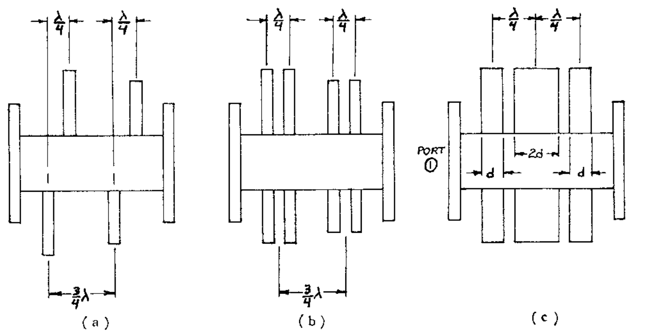

Fig. 5—E plane tee cutoff filters.

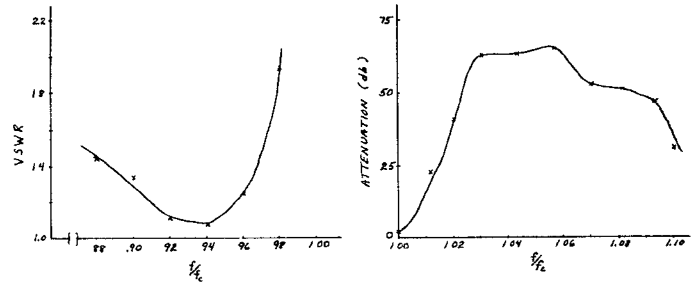

Fig. 7—Vswr and attenuation for filter shown in Fig. 5(b).

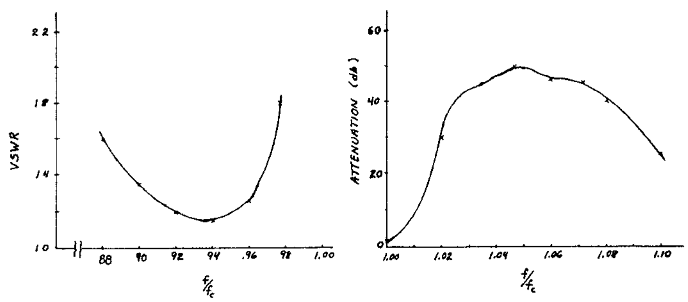

Fig. 6—Vswr and attenuation for filter shown in Fig. 5(a).

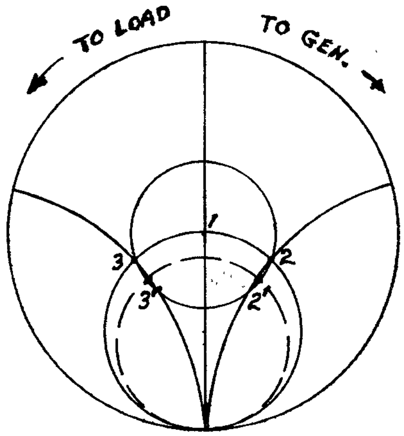

Fig. 8—Matching technique for filter shown in Fig. 5(c).

these reflections and thereby keep the vswr of the complete filter fairly low. The filter shown in Fig. 5(b) also uses the matching technique described above. In this case the number of filter elements is doubled in order to obtain greater attenuation in the stop band. The vswr and attenuation data is given in Fig. 7.

The matching technique used in the filter shown in Fig. 5(c) can be described more conveniently by the use of a Smith Chart. Assuming a generator of internal impedance Z_0 at port 1 and a matched load ($Z=Z_0$) at port 2, the Smith Chart analysis (Fig. 8) proceeds as follows:

1) Starting from the matched load at port 2, the first series stub is encountered. Since this is a small reactive element, impedance is transformed from point 1 along a constant resistance circle to some point 2.

2) Moving approximately $\lambda/4$ toward the generator transforms the impedance to the point 3. Introduction of the next stub which has twice the reactance of the first stub moves the impedance along the constant resistance circle to point 2.

3) Again moving approximately $\lambda/4$ toward the generator on a constant vswr circle moves the impedance to point 3.

4) Insertion of the final stub which has a reactance equal to that of the first stub transforms the impedance to the point 1; that is, a matched condition.

If the operating frequency is now changed, the electrical length between the stubs also is changed. To show that this change in electrical length does not disturb the matched condition to any appreciable degree, let us repeat the above analysis for a decrease in the operating frequency.

1) Starting at point 1, the first stub again transforms the impedance to the point 2.

2) Since the frequency has been decreased, the electrical length between the stubs is now less than $\lambda/4$. Therefore, moving along the first section of line brings the impedance to the point 3′ instead of the point 3. Insertion of the larger stub now transforms the impedance to the point 2′.

3) Since the impedance is at point 2′ instead of point 2, the second length of line (which is also less than $\lambda/4$) moves the impedance to the *point 3.*

4) Insertion of the final stub again transforms the impedance to the center of the Smith Chart.

Hence by virtue of the matching technique used, the change in electrical length between the first and second series stubs is compensated for by the change in length between the second and third stubs; thereby, yielding a good vswr over a considerable frequency band. Fig. 9 shows the vswr and attenuation of this filter [shown in Fig. 5(c)] vs frequency.

As mentioned, individual filter elements were stagger-tuned in all the above filters. This enabled required attenuation to be obtained in shortest possible length. The three filters in Fig. 5 have been designed for X band and are all less than two inches in length.

The Magic-Tee Cutoff Filter

Another method of constructing a band-rejection filter is to make use of the cutoff effect in the E and H arms of a magic tee. Fig. 10 shows a magic tee in which the width of the E and H arms have been reduced to the value "a". At frequencies below the cutoff frequency of

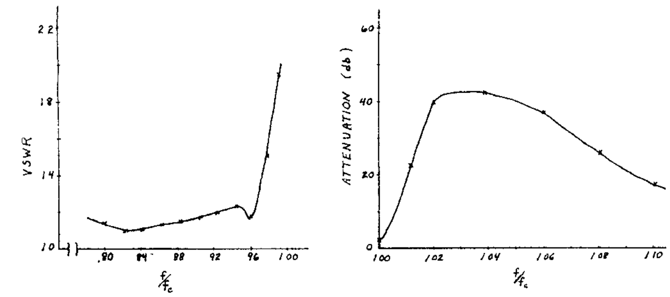

Fig. 9—Vswr and attenuation for filter shown in Fig. 5(c).

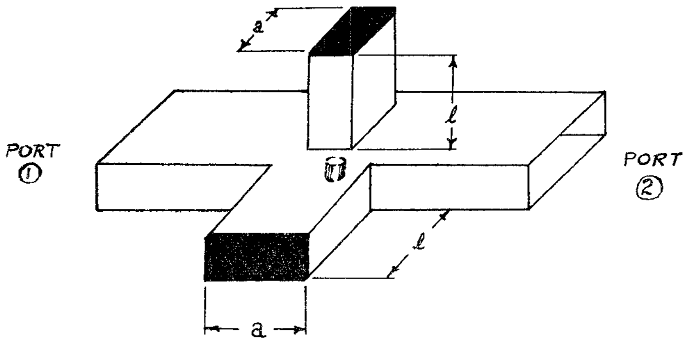

Fig. 10—A magic-tee cutoff filter.

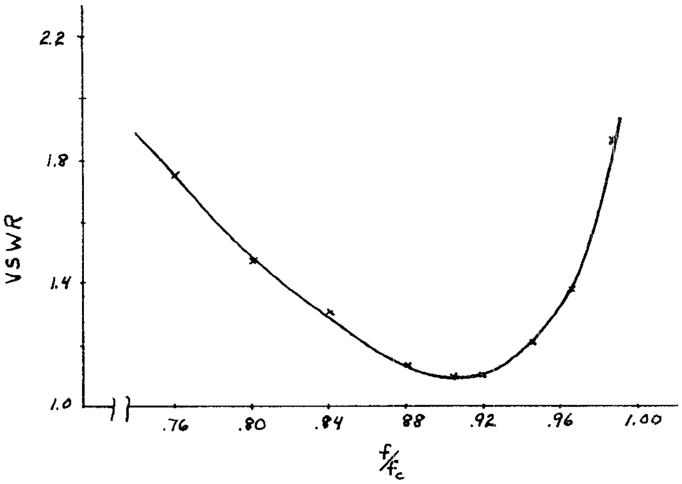

Fig. 11—Vswr for the magic-tee cutoff filter shown in Fig. 10.

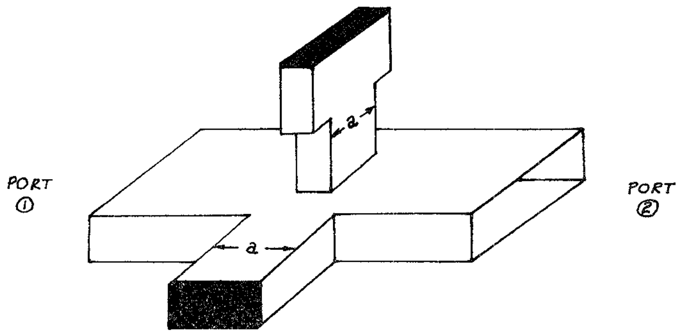

Fig. 12—A broadband magic-tee cutoff filter.

the E and H arms, the microwave energy propagates from port 1 to 2 with practically no loss. In other words, the wave propagating through the main guide will "see" the E and H arms not as transmission lines but as a small reactive effect. The vswr due to this effect can be reduced by placing a small capacitive button in the main guide at the center of the magic tee as shown in Fig. 10. Since the mismatch is corrected at the source of the discontinuity, no long line lengths are involved and consequently a low vswr is maintained over a considerable frequency band as shown in Fig. 11.

Above the cutoff frequency of the E and H arms, the device behaves like an ordinary magic tee. Therefore if shorts are placed in the E and H arms equidistant from their input terminals, the energy entering port 1 will split between the E and H arms and by virtue of the phase relations in a magic tee will all be reflected back to port 1. The only additional requirement on the distances l to the shorts is that they be long enough to attenuate frequencies below the cutoff frequency of the E and H arms before the energy reaches the shorts. Usually 10 db attenuation is adequate to make the shorts "invisible" in the pass band. The required length for any attenuation can be determined from (2).

Since the operation of this filter in the stop band only requires that the electrical length to the shorts in the E and H arms be equal, good rejection should be obtained for all frequencies above the cutoff frequency of the E and H arms. That is, the magic-tee cutoff filter is basically a low pass filter. However, in practice, this wideband rejection is difficult to obtain. The reason is that although the electrical length from the input terminals to the shorts are equal, the actual length of E and H plane stubs are not necessarily equal. For example, the electrical length of the E plane stub is determined by the actual length of the shorted E arm *plus* the distance from the top wall of the main guide to the apparent input terminal of the series arm. This apparent input terminal is located approximately one-quarter of the guide height down from the top wall of the main guide. Similarly, the electrical length of the H arm is a combination of the actual length of the shorted H arm plus the distance from the side wall to the apparent input terminal of the shunt arm. This terminal is located at approximately the center of the main guide. Since the shift in apparent input terminal is different for the E and H arms, different lengths of E and H plane stubs are required to give the same electrical length from the input terminals to the short circuits. Consequently, the frequency sensitivity of the two line lengths are different and adjustment of the line lengths for good rejection at one frequency will not necessarily yield good rejection at all frequencies. To compensate for this, one can add a section of guide of appropriate width and length behind either the E or H arm cutoff section which will tend to cancel the difference in frequency sensitivity between the E and H plane stubs. This idea is similar, in principle, to that presented by H. Sohon.[8] For example, Fig. 12 shows a magic-tee cutoff filter which employs an additional section of guide in the E arm to reduce the difference in frequency sensitivity between the shorted stubs.

[8] H. Sohon, "Wide band phase delay circuits," PROC. IRE, vol. 41, pp. 1050–1052; August, 1953.

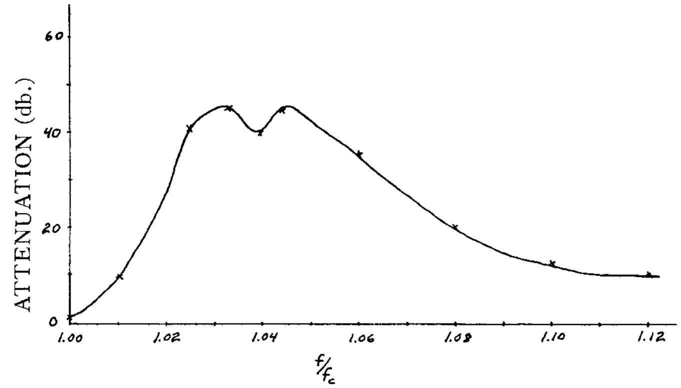

Fig. 13—Attenuation of a magic-tee cutoff filter.

The attenuation vs frequency characteristic of this filter is given in Fig. 13. To obtain greater rejection in the stop band, two or more filters can be cascaded. Another method is to employ a double magic tee; *i.e.*, one with two E arms and two H arms.

Conclusions

Two band-rejection filters employing the waveguide cutoff effect have been discussed. They are the E plane tee cutoff filter and the magic-tee cutoff filter. By appropriate matching techniques, E plane tee cutoff filters have been designed with a vswr of less than 1.5 over 8, 10, and 16 per cent pass bands while maintaining greater than 25 db rejection over 8, 10, and 6 per cent bands, respectively. A magic-tee cutoff filter has also been constructed which has a vswr of less than 1.7 over a 20 per cent frequency band. By proper design of the shorted E and H arms, greater than 10 db rejection has been obtained over a 12 per cent band.

Acknowledgment

The author wishes to acknowledge the assistance of Mr. G. C. Shaw in formulating some of the ideas presented here. He also wishes to express his gratitude to Dr. Ernest Wantuch and Mr. T. S. Saad for their suggestions.

Technique of Pulsing Low Power Reflex Klystrons

J. I. DAVIS†

Summary—**Very little published information is available on pulsing low power reflex klystrons. Since low power reflex klystrons have been generally designed for cw operation as local oscillators, a minimum of effort has been directed toward the development of specific low power pulse reflex klystrons.**

This paper summarizes an effort that has been directed toward pulsing typical low power reflex klystron with a description of the techniques evolved and a summary of the limitations and merits of each technique. Included also is a description of a pulse klystron "priming" technique that minimizes the effects of pulse shortening and leading edge jitter associated with typical pulse operation.

Introduction

IT HAS BEEN apparent for some time to designers of microwave equipment that there has been little or no attempt by the klystron tube manufacturers to design low power klystrons for pulse application. In general, the low power klystrons have been designed for local oscillator use in radar and beacon receivers, and in equipment such as spectrum analyzers. A great deal of effort of late has been devoted toward pulsing high power klystrons for use in generating high speed particles and for use in large anti-jamming radar systems. This has forced the designers of low power beacons and rf signal generators to rely almost wholly on their wit and ingenuity to find suitable techniques for pulsing low power klystrons for optimum pulse fidelity and minimum pulse jitter. In addition to the lack of information from manufacturers on pulsing klystrons there has been very little written in the literature on suitable techniques for pulsing klystrons to yield minimum leading edge jitter and pulse shortening.

Because many of the new guided missile systems have stringent requirements for rf pulsed coded modulation to meet their tactical requirements, a great deal of effort has been expended in advancing the state of the art of pulse circuit design. The circuitry itself has preceded the pulsed rf techniques primarily because of the availability of new components to the circuit designer. These components consist essentially of improved pulse transformers, more reliable thyratrons, and improved hard tube modulator tubes. The basic difficulty of reliable pulsed rf systems has been associated primarily with the pulsing of the klystron itself.

Starting of Pulse Reflex Klystron

It is well known that all electronic oscillators are started by noise or circuit transients associated with the oscillator. This concept applies equally well to the starting of reflex klystron oscillators. However, when one is concerned with pulse operation of reflex klystrons there exists the distinct possibility that the oscillation could have been started by shock excitation of the resonant cavity by the pulsed beam current.

† Litton Industries, Beverly Hills, Calif.