Second-Harmonic Effects in Tuned Reflectometers*

The theory of operation of the tuned microwave reflectometer has been well documented in the literature[1-4] and its operating characteristics have been described in great detail.[5] According to reflectometer theory, as a short slides in the output waveguide of a tuned reflectometer a pattern of detector response vs short displacement similar to that shown in Fig. 1 will be observed. The over-all slope of the pattern is due to attenuation in the output waveguide, while the regular variations are due to the less than ideal tuning of the reflectometer. In the practical operation of a reflectometer the pattern shown in Fig. 2 can be observed. Such a pattern obviously will result in errors in the determination of the magnitude and positions of maxima and minima. It is the purpose of this note to explain this pattern and the errors it can produce and to suggest a simple remedy.

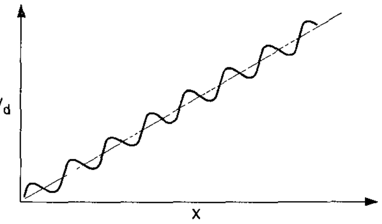

Fig. 1—Detector output vs short position for a non ideal reflectometer. Over-all slope is due to waveguide attenuation.

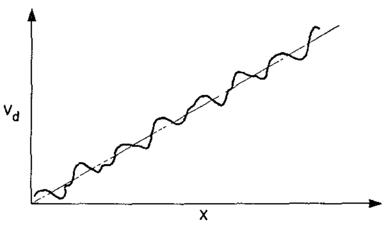

Fig. 2—Detector output vs short position with second-harmonic signal present in main waveguide.

For convenience in detecting low-level signals, a reflectometer may use a crystal diode as a detector in combination with some sort of differenential voltmeter as an indicator. Unlike the slotted-section standing-wave machine, the detector of the reflectometer is tightly coupled to the main waveguide in which the measurements are being made. It is thus possible to inject a signal at the detector and interfere with the signal present in the main waveguide. Tuned crystal mounts are often excellent harmonic generators; a well-designed harmonic generator using a crystal can deliver a second-harmonic signal that is as much as 10 db below the fundamental. The crystal detector of a tuned reflectometer is thus perfectly free to generate a second-harmonic signal and inject it, through the associated directional coupler, back into the main waveguide of the system. Neither the directional coupler nor a detector isolator, because of its construction, can be expected to stop the second-harmonic from entering the main waveguide of the reflectometer. Experiments with an *S*-band reflectometer have shown that in the main waveguide it is possible to have a second-harmonic signal that is only 15 db below the level of the fundamental.

On the basis of normal linear circuit theory a signal with a period greater than that of its fundamental component is not possible. The standing-wave pattern shown in Fig. 2 apparently contains a fundamental component and a component that is a non-integral multiple of the fundamental. This is completely possible in a waveguide system where only a fundamental and its second frequency harmonic are present, for the second frequency harmonic is not the second spatial harmonic. The usual expression relating the waveguide wavelength λ_g to the free-space wavelength λ and the waveguide cutoff wavelength λ_c is

$$\lambda_g = \left[\left(\frac{1}{\lambda}\right)^2 - \left(\frac{1}{\lambda_c}\right)^2\right]^{-1/2}. \qquad (1)$$

Doubling the frequency of a signal halves its free-space wavelength. Therefore the waveguide wavelength of a second harmonic signal is

$$\lambda_{g_2} = \left[\left(\frac{4}{\lambda}\right)^2 - \left(\frac{1}{\lambda_c}\right)^2\right]^{-1/2}.$$

Combining (1) and (2), an expression for the ratio of the second-harmonic waveguide wavelength to the fundamental waveguide wavelength is

$$\frac{\lambda_{g_2}}{\lambda_g} = \frac{1}{2}\left[\frac{1-(\lambda/\lambda_c)^2}{1-(\lambda/2\lambda_c)^2}\right]^{1/2}. \qquad (3)$$

The behavior of the ratio λ_{g2}/λ_g as a function of the ratio of fundamental free-space wavelength λ to the waveguide cutoff wavelength λ_c is plotted in Fig. 3. From this plot it is obvious that the second-harmonic wavelength will never be as much as half the fundamental wavelength because most waveguides are operated at a free-space wavelength of at least half the cutoff wavelength.

For the sake of simplicity, let the output of the detector V_d be a constant times the magnitude of the standing-wave pattern $V_d = KV$, and the fundamental standing-wave pattern be of a $1 + \sin x$ form. To avoid a direct discussion of the crystal characteristic, let the ratio of the fundamental to the second-harmonic signal it generates be k. The amplitude of the second harmonic is thus k times the magnitude of the fundamental at any point. Under these assumptions, the detector output can be written

$$V_d = KV\left[1 + \sin\frac{4\pi x}{\lambda_g}\right] + KkV\left[1 + \sin\frac{4\pi x}{\lambda_g}\right]\left[1 + \sin\frac{4\pi x}{\lambda_{g_2}}\right] \qquad (4)$$

and rewritten

$$V_d = VK\left[(1+k) + (1+k)\sin\frac{4\pi x}{\lambda_g} + k\sin\frac{4\pi x}{\lambda_{g_2}} + k\sin\frac{4\pi x}{\lambda_g}\sin\frac{4\pi x}{\lambda_{g_2}}\right]. \qquad (5)$$

Eq. (5) shows that the output of the detector can be normally expected to contain a dc component, a fundamental component, a second-harmonic component and a cross-product term. The specific case of $k = 0.25$ and $\lambda/\lambda_c = 0.4$ is plotted in Fig. 4.

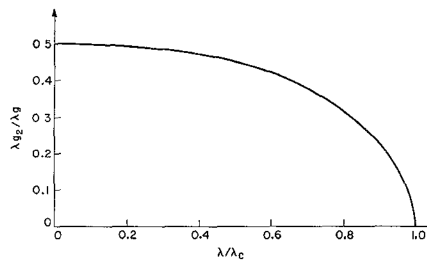

Fig. 3—Ratio of second-harmonic waveguide wavelength to fundamental waveguide wavelength as a function of the ratio of fundamental free-space wavelength to cutoff wavelength.

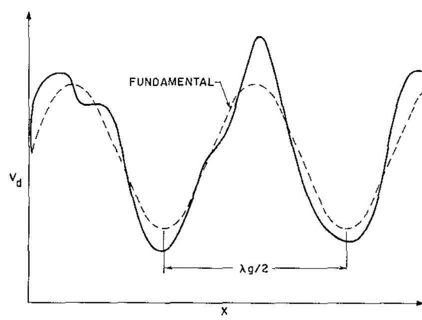

Fig. 4—Illustration of the effect for $k = 0.25$ and $\lambda/\lambda_c = 0.4$.

From Fig. 4 it is evident that the presence of the second-harmonic signal will not only cause an error in the observed magnitude of the maxima and minima, but will also cause an error in locating their positions. These errors can both be evaluated through the use of (5). The amplitude error can be directly evaluated while the phase error must be evaluated through differentiating (5).

In summary, the effect illustrated in Fig. 2 can be directly attributed to the generation of a second-harmonic signal by the crystal

* Received by the PGMTT, October 10, 1961.

[1] A. C. MacPherson and D. M. Kearns, "A new technique for the measurement of microwave standing-wave ratios," Proc. IRE, vol. 44, pp. 1024–1030; August, 1956.

[2] R. W. Beatty and D. M. Kerns, "Recently developed microwave impedance standards and methods of measurement," IRE Trans. on Instrumentation, vol. I-7, pp. 319–321; December, 1958.

[3] G. F. Engen and R. W. Beatty, "Microwave reflectometer techniques," IRE Trans. on Microwave Theory and Techniques, vol. MTT-7, pp. 351–355; July, 1959.

[4] R. W. Beatty, G. F. Engen, and W. J. Anson, "Measurement of reflections and losses of waveguide joints and connectors using microwave reflectometer techniques," IRE Trans. on Instrumentation, vol. I-9, pp. 219–226; September, 1960.

[5] W. J. Anson, "A Guide to the Use of the Modified Reflectometer Technique of VSWR Measurement," Natl. Bureau of Standards, Boulder, Colo., NBS Rept. 6095; April 13, 1960.

detector. Experiments at S band have shown that errors of as much as 10 db in amplitude and 4 to 5 degrees in phase can be produced in an otherwise perfect system when the crystal detector is carefully tuned. It is possible to evaluate such errors directly. However, experimentally, it is far simpler to eliminate the undesirable second-harmonic component by placing a low-pass filter ahead of the crystal detector.

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A High-Capacitance Parametric Diode for Use at Low Frequencies*

For operation of parametric amplifiers at low frequencies and at low-impedance levels a parametric diode of very large capacitance is necessary. In general a capacitance which has an impedance of the same order as the source impedance is required, so that for a 600-ohm system at a frequency of 5 kc the static capacitance C_0 should be

$$C_0 \simeq \frac{1}{\omega_s R_g} \simeq 0.05 \ \mu\text{f}.$$

The capacitance variation with bias should also be large.

A capacitance for this purpose may be assembled from a number of silicon rectifiers of the type used in TV receiver power supplies. The capacitance as a function of voltage for a typical 0.5 ampere silicon rectifier (such as the 1N1763, 1N2094, OA210, etc.) measured at $1/\pi$ Mc and with an applied RF level of 10 mv (rms) is as follows:

Bias (volts)	0	−1	−2	−3	−6	−12
Capacitance ($\mu\mu$f)	74.7	45.8	38 3	30.8	26.3	20.0

An assembly of 220 silicon rectifier wafers of this type, connected in parallel, and measured under the same conditions, gave a C-V curve as in Fig. 1. The cutoff frequency of this diode, measured at −3 volts, was 10 Mc.

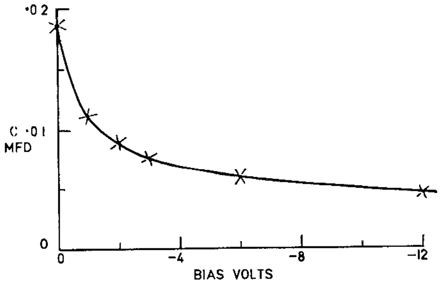

Fig. 1.

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This diode, when used in an appropriate 3-frequency parametric amplifier using high Q inductors ($f_p \simeq 50$ kc $f_s \simeq 5$ kc), shows all of the classical characteristics of parametric amplifiers, both qualitatively and quantitatively.

This note is intended primarily to draw the attention of microwave engineers to the possibility of using such diodes in a low-frequency electrical parametric amplifier for demonstration and teaching purposes.

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A Nonreflective Hybrid Stop-Band Filter*

A novel waveguide stop-band filter employing a four terminal hybrid circuit and two lossy cavity resonators has been developed. With this circuit, adjustable frequency sensitive loss can be accomplished with very little reflection similar to directional filter type operation. The circuit utilizes the diplexing characteristics of two adjacent ports of a standard 3-db short slot forward-wave directional coupler. Two identical cavities, each filled with lossy dielectric material, produce the necessary loss. Fig. 1 shows a schematic diagram of the device.

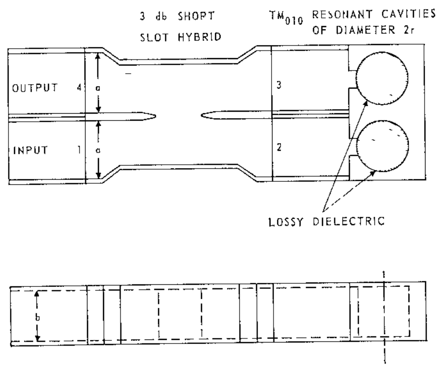

Fig. 1—Stop-band waveguide filter.

For explanatory purposes, first consider a short circuit placed across terminals 2 and 3 of the hybrid in place of the cavities, each short being equidistant from the hybrid. Energy entering port 1 of the hybrid will split equally in power to terminals 2 and 3, and the phase of the signal at terminal 3 will lag terminal 2 by 90 degrees. Upon reflection from the short circuit, energy from port 2 will phase add at port 4 with the reflected energy from port 3. At port 1, destructive interference will occur between the reflected energies from ports 2 and 3. Thus, within the relatively broad passband of the hybrid, all the input energy at port 1 will appear at port 4, with a minimum of loss; and no mismatch will be exhibited at port 1. By replacing the shorts with two identical cavities filled with a lossy dielectric material, the amount of transmitted power from terminals 1 to 4 can be made frequency sensitive again with no reflection appearing at the input terminal. The cavities were designed to resonate in the TM_{010} cylindrical cavity mode. Within a few per cent, this resonant frequency may be stated as follows:

$$f_0 = \frac{4.521}{r\sqrt{\epsilon_r}} \text{ kMc},$$

where r is the cavity radius in inches and ϵ_r is the relative dielectric constant of the lossy dielectric material.

A typical transmission and reflection coefficient plot as a function of frequency is shown in Fig. 2. The dielectric material used

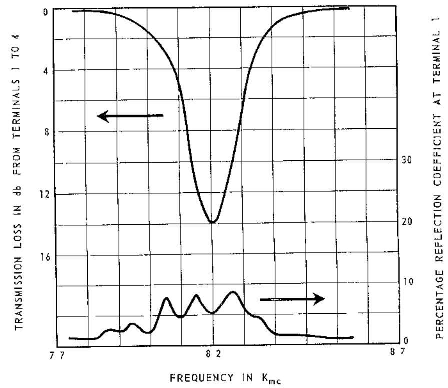

Fig. 2—Transmission and reflection characteristics of a typical stop-band filter.

was a forsterite compound. The loss was 14 db at the filter center frequency of 8.2 kMc. The loaded Q of this filter was about 75 and the reflection coefficient less than 10 per cent over the frequency band of the large X-band hybrid employed. The amount of loss at the center frequency at which the cavities are tuned can easily be controlled by varying the amount of coupling between the waveguide and cavity by adjustment of the cavity iris geometry or by varying the composition of the dielectric material. With this circuit technique, stop-band waveguide filters exhibiting up to 20-db loss at the center frequency can be realized simultaneously with very little mismatch, *i.e.*, with less than a 10 per cent reflection coefficient.

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