with frequency. This effect may be attributed to discontinuity effects at the junctions between the re-entrant and strip line sections. In view of the large plate spacing compared to wavelength ($b/\lambda=0.405$ at 5.0 Gc in the re-entrant section), it is not surprising that discontinuity effects are severe. The poor directivity is also attributable to the discontinuities.

In order to decrease the coupling of the re-entrant section, the outer characteristic impedance, Z_{o1}, was reduced from 84.7 to about 76.3 ohms by reducing the plate spacing from 0.956 inch to 0.832 inch. This produced the desired triple loop appearance of the coupling and main line response curves. The center loop was 0.75 db, the lower loop 0.3 db, and the upper loop 0.7 db. The directivity was almost unaffected, however.

A number of changes were then made in the end sections to cancel discontinuity effects and increase the directivity. The final data for the 1–5-Gc hybrid coupler is shown in Fig. 10. Photographs of the coupler disassembled and assembled are given in Figs. 11 and 12.

Conclusion

The advantages of the re-entrant cross section have been verified experimentally in both single-section and three-section 3-db hybrid-coupler models. For cross-section heights in the range of 0 to 0.16λ, the simple design formulas have proved to yield very good accuracy. Beyond 0.16λ the coupling appears to increase gradually, but nevertheless good performance was achieved even with a cross section height as large as 0.4λ, after a minor adjustment of dimensions.

In the case of a three-section design for 5:1 bandwidth, a center section coupling of -1.28 db is required. This very strong coupling is easily achieved by the re-entrant cross section, and may be maintained accurately in production.

Correspondence

On the General Relation Between α and Q*

The relation between the quality factor Q and the attenuation constant α of a transmission line has been known as follows:

$$\alpha = \frac{\beta}{2Q}$$

where β is the phase constant. Recently from the following relation of propagation constant at resonance

$$\Gamma(\omega_0) + \frac{\partial \Gamma}{\partial \omega}\Delta\omega \simeq i\beta(\omega_0),$$

where

$$\Gamma(\omega_0) = \alpha(\omega_0) + i\beta(\omega_0).$$

Yeh[1] derived a general relation between Q and α, namely,

$$\alpha = \frac{v_p}{v_g}\,\frac{\beta}{2Q},$$

where v_p, and v_g are the phase velocity and group velocity of the wave respectively. This general relation can be derived very simply from the generally accepted definition of α and Q.

General definition of Q applicable to waveguide as well as to ordinary transmission line is as follows:

$$Q = \omega\,\frac{\text{energy stored per unit length}}{\text{power lost per unit length}}.$$

The attenuation factor α in the range of propagation is given by

$$\alpha = \frac{1}{2}\,\frac{\text{power lost per unit length}}{\text{power transmitted}}.$$

If we realize that the power transmitted is equal to the energy stored per unit length multiplied by v_g (group velocity) instead of v_p (phase velocity) then it is readily seen that

$$\alpha = \frac{1}{2}\,\frac{\omega}{v_g Q}.$$

Then from the relation

$$v_p = \frac{\omega}{\beta},$$

we obtain

$$\alpha = \frac{v_p}{v_g}\,\frac{\beta}{2Q}.$$

H. P. Hsu
Elec. Engrg. Dept.
Essex College
Assumption University of Windsor
Windsor, Ontario, Canada.

* Received February 26, 1963.

[1] C. Yeh, "A relation between α and Q," Proc. IRE (*Correspondence*), vol. 50, p. 2145; October, 1962.

A Varactor Frequency-Modulated AFC Reference Cavity*

Summary

A varactor frequency-modulated microwave cavity is described where the application of a periodic square wave voltage to a varactor serves to electronically detune the cavity. There results a discriminator characteristic which makes the device applicable as an AFC reference. Experimental results are given and discussed.

Introduction

When an AFC loop is used to stabilize the frequency of a microwave signal source, a stable frequency reference is needed. A frequency modulated (periodically detuned) microwave cavity has been developed which performs this function by giving an AC error voltage whose magnitude and phase are proportional respectively to the magnitude and direction of the frequency shift of the source from the reference frequency of the cavity. The cavity thus has a discriminator characteristic. The periodic detuning of the cavity is accomplished through the application of a square wave voltage to a nonlinear reactance device (a varactor[1] or *pn*-junction diode) coupled to the cavity.

The varactor frequency-modulated microwave reference cavity is basically a high-Q

* Received February 26, 1963.

[1] A. Uhlir, Jr., "The potential of semiconductor diodes in high frequency communications," Proc. IRE, vol. 46, pp. 1099–1115; June, 1958.

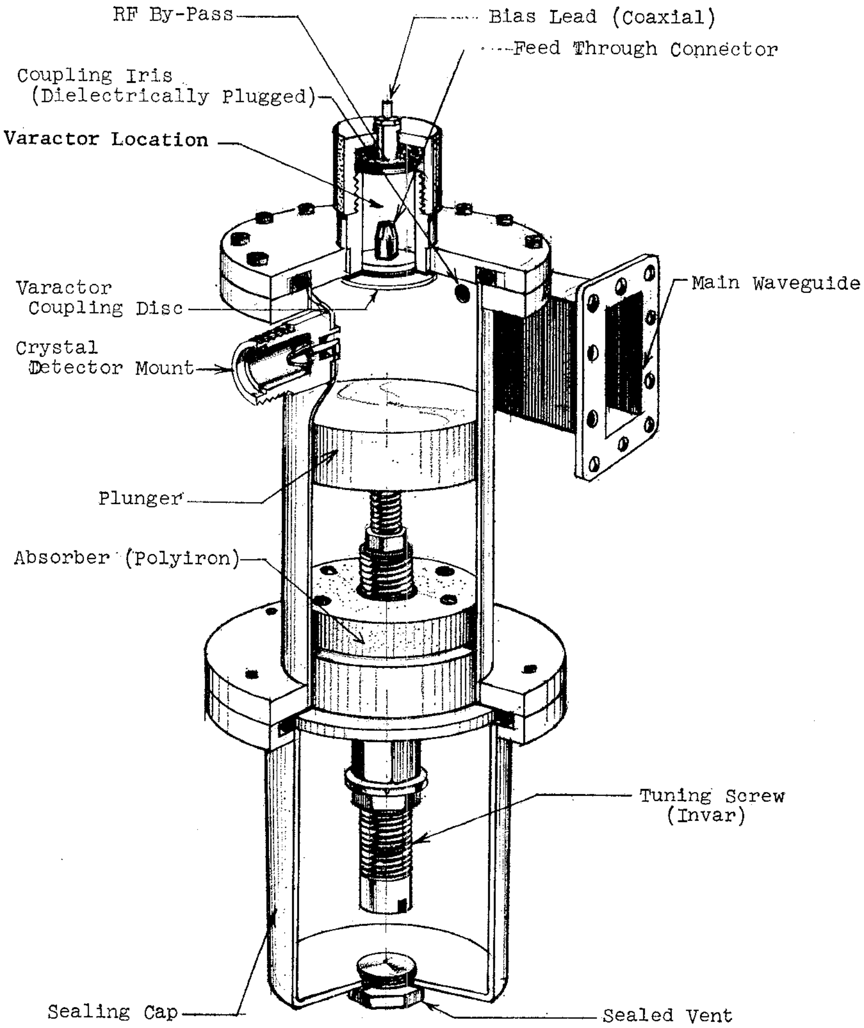

Fig. 1—Varactor frequency modulated AFC reference cavity.

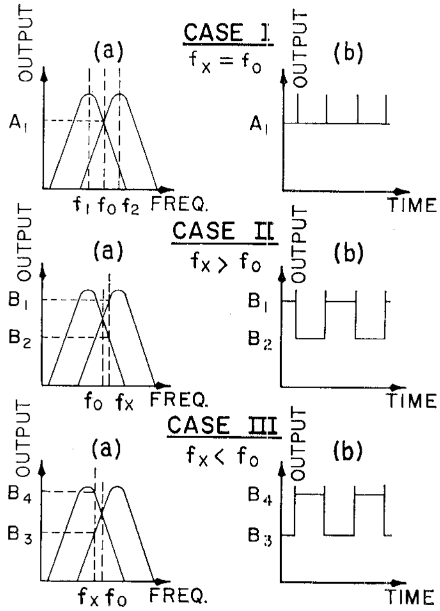

Fig. 2—Diagram of cavity function.

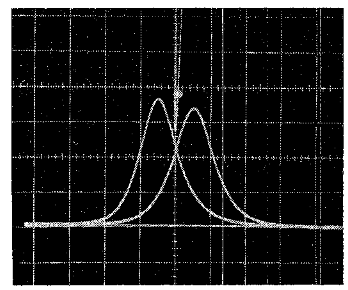

Fig. 3—Oscillogram of cavity response (separation of peaks 6 kMc).

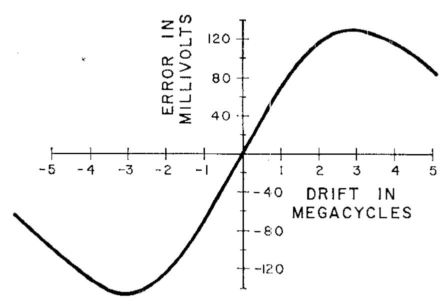

Fig. 4—Discriminator response (error voltage vs frequency drift of klystron).

cylindrical cavity operating in the TM_{011} mode. A *pn*-junction diode (varactor), in an external coaxial mount, is reactively coupled to the cavity by a small post-supported disk which is connected to the varactor through a glass feedthrough in the center of the top end plate of the cavity. This is illustrated in Fig. 1. Frequency modulation of the cavity is accomplished by periodically perturbing the resonant condition of the cavity with a small reactance change in the varactor, when a periodic bias voltage in the form of a square wave is applied to the varactor. A 5.5-v square wave bias will result in two resonant frequencies of the cavity separated by approximately 5 to 7 Mc.

The cavity as now designed operates over the frequency band of 6575 to 7125 Mc. The output is approximately 60 mv per Mc. To extend the frequency range to 5875–8500 Mc slightly different design optimization would be required. For a communications application means of temperature compensation has been designed into the cavity.

Description of Cavity Operation (Method of Frequency Modulation)

The functional operation of the cavity is illustrated in Fig. 2. A square wave voltage of any convenient low frequency is applied to the varactor, and corresponding to each of the two voltage levels there is a resonant frequency of the cavity, designated by f_1 and f_2. If a microwave signal of frequency f_x is introduced into the cavity through the coupling iris and then amplitude detected with a single crystal detector at the output of the cavity (lightly coupled to the cavity), there will result a detected envelope which is a square wave. The amplitude of this square wave will be zero if f_x corresponds to the crossover frequency f_0 of the two cavity resonant response curves. If f_x differs from the crossover frequency f_0, then the detected amplitude will increase with the magnitude of the frequency difference between f_0 and f_x, and there will be a 180° phase reversal in the detected square wave which will give indication of whether f_x is above or below f_0. This square wave voltage is detected and used in conjunction with a synchronous (phase sensitive) detector for use in an AFC application.

Fig. 3 shows an oscillogram of the frequency modulated cavity responses. Here the klystron source and the scope are both swept with a 1-kc sawtooth voltage, while the cavity is periodically detuned by a square wave voltage bias of approximately 80 cps. With a 5-v *pp* square wave, the frequency separation between the two resonant peaks shown is in the order of 6 Mc.

Fig. 4 illustrates the discriminator characteristic of the frequency modulated cavity. This curve is the actual error voltage generated when the frequency of the source deviated from the crossover frequency f_0. At f_0 the error voltage is zero. At a given frequency shift the error is the difference between the two cavity response curves (shown in Fig. 3). When this device is used in an AFC application, it is seen that the rate of correction will be proportional to the amplitude of the error voltage, which in turn is proportional to the deviation of the source frequency from f_0. Further, the sense of correction will correspond to the polarity of the error. Note that the frequency pull-in range is not limited to the frequency separation between the peaks of the two cavity responses.

Design Consideration for Cavity Modulation

The design of a microwave resonant cavity which could be cyclically tuned by the application of a periodic voltage to a solid state device suggested the choice of the TM_{011} mode for several reasons. 1) The electric field is normal to the end plates of the cavity. A slight perturbation of this field at the center of the endplate where the electric field is maximum will most effectively detune the cavity. A disk mounted above the center of the endplate (see Fig. 1) provides such an effective method of cavity perturbation. A post supporting the disk to the endplate provides a dc return for the varactor.

The input waveguide is iris coupled to the cavity so as to set up the desired TM_{011} mode. The size of the iris is designed to give approximately 30-db decoupling from a one-watt source. This input level is sufficiently low to give ample output without subjecting the varactor to a large enough RF field to excite self oscillation in the diode.

The output from the cavity is obtained from a crystal detector mounted in an external coaxial mount. It is very lightly loop coupled to the cavity to give adequate output level without causing the cavity frequency to be susceptible to change in crystal characteristic, either through replacement or because of environmental change.

Results and Conclusions

It has been experimentally verified that the cavity is free from multimoding within the design frequency band of 6575 Mc to 6875 Mc. This stands in confirmation of the mode selection and method of excitation previously discussed.

The cavity can be tuned to any given frequency within the above band to within 10 kc. Fig. 3 demonstrates that the frequency separation between the two resonant frequencies is approximately 6 Mc when the applied square wave bias is 5.5 v *pp*. The slight dissimiliarity in the Q of the curves is attributable to the different loading of the cavity by the varactor for two different bias conditions. However, the symmetry of the two curves is close enough to give the symmetrical error curve shown in Fig. 4.

From the discriminator-like error curve, it is seen that the normal output level close to a crossover is approximately 60 mv per Mc with the detector terminated in a low impedance of 300 ohms. Further, measurements show that there is ample error voltage for "pull-in" in an AFC system for frequency shift of source out to ± 12 Mc.

The symmetry of the error curve permits the use of a frequency modulated carrier as the signal source, provided the major portion of the spectral energy of the source is located within the range of symmetry. A severe limiting test has shown that with a single tone frequency modulation, using a modulation frequency of 1.2 Mc and a frequency deviation of 2.5 Mc, the resulting frequency shift in the cavity crossover frequency is only 200 kc. In a microwave communication application, 600 multiplexed voice channels, each considered to be deviated 200 kc rms, gives rise to this same statistical peak deviation of 2.5 Mc; and a klystron source modulated with a white noise equivalent of such a 600-voice channel signal causes no detectable shift in the cavity crossover. This test shows that the linearity of the error curve is adequate for an operating AFC. For a CW source, the symmetry requisite in the error characteristic is of no consideration.

Tests performed to determine the characteristics of the cavity under conditions of a change in ambient temperature from $-20°C$ to $+65°C$ have shown that a frequency stability of better than ± 0.005 per cent is obtainable.

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A. Farokhrooz
J. B. Linker, Jr.
Communication Products Dept.
General Electric Company
Lynchburg, Va.

Nonlinear Biasing Resistors for Microwave Tunnel-Diode Oscillators*

By using a nonlinear rather than a linear stabilizing resistor in tunnel-diode oscillator and amplifier circuits, the dc power dissipation in the resistor may be reduced by a factor of 3 for typical germanium tunnel diodes, and by a factor of 6 for typical gallium arsenide tunnel diodes. At the same time ac loading by the resistor is reduced. Such nonlinear stabilizing resistors may consist of reverse- or forward-biased heavily doped *pn* junctions.

A typical tunnel-diode microwave oscillator circuit[1] is shown schematically in Fig. 1. In order to bias and stabilize the tunnel diode to an operating point on its negative current-voltage characteristic a biasing resistor R_1 is used. The size of R_1 is a compromise between two conflicting requirements. On the one hand, R_1 should be small enough so that sufficient stabilization of the operating point is obtained. This means that the combined characteristic shown dashed in Fig. 2 of R_1 and the tunnel diode should have a positive slope throughout. On the other hand, R_1 should be as large as possible to minimize the dc power loss in the resistor, measured by the shaded area in Fig. 2.

The purpose of this report is to point out the advantage of using a *nonlinear biasing resistor* with a current-voltage characteristic as indicated by the curve R_2 in Fig. 2 and to show how this can be achieved. Such a nonlinear biasing resistor offers a better compromise between the two conflicting requirements than a linear resistor. Particularly, the power dissipation in the resistor, indicated by the double-crosshatched area in Fig. 2, may be considerably reduced compared to a linear resistor, while still maintaining sufficient stabilization, *i.e.*, a positive slope of the combined characteristic.

The characteristic of R_2 may be obtained by a *pn* junction in three different ways, namely,

a) a forward-biased diode junction. This has several disadvantages: slow response because of storage of injected minority carriers, either voltage offset if the bandgap is appreciable or very large areas and therefore also high capacitance and poor temperature stability.
b) a forward-biased tunneling junction. This also requires low-bandgap semiconductor material for very low voltage operation as with germanium tunnel diodes. It may be used for gallium arsenide tunnel diodes.
c) a reverse-biased tunneling junction. This is the preferred alternative.

The determination of the reduction in dc power loss is straightforward (see Appendix I). Table I shows the ratio of power loss in a linear biasing resistor P_1 to the power loss in a nonlinear biasing resistor P_2 for tunnel-diode oscillators using germanium

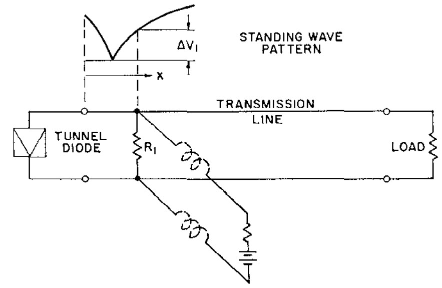

Fig. 1—Schematic picture of biasing arrangement for tunnel-diode oscillator with biasing resistor R_1.

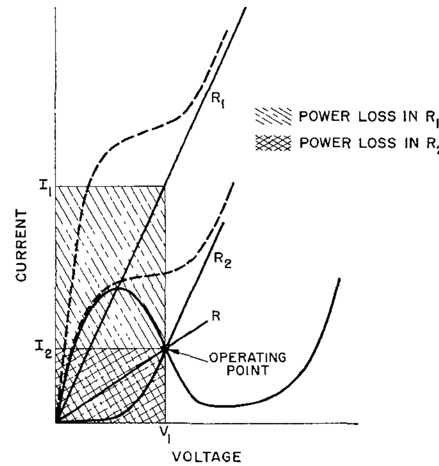

Fig. 2—Current-voltage characteristic of tunnel diode with two alternatives for load resistor. DC power losses for the two alternatives are indicated.

* Received December 26, 1962; revised manuscript received March 6, 1963.

[1] The figures are all lumped.